ELECTRONICS TECHNICIAN 3 & 2

BUREAU OF NAVAL PERSONNEL

NAVY TRAINING COURSE   NAVPERS 10195

PREFACE

This training course is written for men of the U. S. Navy and Naval Reserve who are interested in qualifying for Electronics Technician Third and Second Class. Combined with the necessary practical experience, this training course will aid you in preparing for the advancement-in-rating examination.

The qualifications for advancement are listed in the Manual of Qualifications for Advancement in Rating, NavPers 18068-B. Because examinations for advancement in rating are based on these qualifications, you should refer to them for guidance.

This training course was prepared by the Training Publications Division, Naval Personnel Program Support Activity, Washington, D. C., for the Bureau of Naval Personnel. Technical assistance was provided by the Electronics Technician School, Naval Training Center, Great Lakes, Illinois; Naval Ship Systems Command; and other activities cognizant of electronic equipments and the duties of Electronics Technicians.

THE UNITED STATES NAVY

GUARDIAN OF OUR COUNTRY

The United States Navy is responsible for maintaining control of the sea and is a ready force on watch at home and overseas, capable of strong action to preserve the peace or of instant offensive action to win in war.

It is upon the maintenance of this control that our country's glorious future depends; the United States Navy exists to make it so.

WE SERVE WITH HONOR

Tradition, valor, and victory are the Navy's heritage from the past. To these may be added dedication, discipline, and vigilance as the watchwords of the present and the future.

At home or on distant stations we serve with pride, confident in the respect of our country, our shipmates, and our families.

Our responsibilities sober us; our adversities strengthen us.

Service to God and Country is our special privilege. We serve with honor.

THE FUTURE OF THE NAVY

The Navy will always employ new weapons, new techniques, and greater power to protect and defend the United States on the sea, under the sea, and in the air.

Now and in the future, control of the sea gives the United States her greatest advantage for the maintenance of peace and for victory in war.

Mobility, surprise, dispersal, and offensive power are the keynotes of the new Navy. The roots of the Navy lie in a strong belief in the future, in continued dedication to our tasks, and in reflection on our heritage from the past.

Never have our opportunities and our responsibilities been greater.

CONTENTS

READING LIST

NAVY TRAINING COURSES

Basic Electricity, NavPers 10086-A
Basic Electronics, NavPers 10087-B
Introduction to Electronics, NavPers 10084
Basic Handtools, NavPers 10085-A
Blueprint Reading and Sketching, NavPers 10077-B
Mathematics Vol. 1, NavPers 10069-C
Mathematics Vol. 2, NavPers 10071-A
Mathematics Vol. 3, NavPers 10073
Standard First Aid, NavPers 10081-B

OTHER PUBLICATIONS

Naval Ships Technical Manual, Chapter 9670

USAFI TEXTS

United States Armed Forces Institute (USAFI) courses for additional reading and study are available through your Educations Services' Officer.* The following courses are recommended:

A788	Introduction to Electronics I
A789	Introduction to Electronics II
C166	Advanced Algebra
C176	Plane Geometry I
C177	Plane Geometry II
C188	Trigonometry

*"Members of the United States Armed Forces Reserve components, when on active duty, are eligible to enroll for USAFI courses, services, and materials if the orders calling them to active duty specify a period of 120 days or more, or if they have been on active duty for a period of 120 days or more, regardless of the time specified on the active duty orders."

CHAPTER 1

ADVANCEMENT

This training course has been prepared for men of the Navy and of the Naval Reserve, who are studying for advancement to the rates of Electronics Technician 3 and Electronics Technician 2.

The Electronics Technician qualifications used as a guide in the preparation of this training course are those contained in Revision A of the Manual of Qualifications for Advancement in Rating, NavPers 18068.

Chapter 2 of this training course contains information on the Navy 3-M system, records, and reports as they pertain to the Electronics Technician. Chapter 3 discusses the operation and use of the radio interference measuring set, spectrum analyzer, absorption wavemeter and echo box. Chapter 4 covers radio receiving sets AN/SRR-11, 12, and 13, AN/WRR-2, and AN/WRR-35.A. This chapter also includes a brief discussion of SSB converter CU-591A/URR.

Chapters 5 and 6 describe radio transmitters with special circuits of the AN/SRT-14, 15, and 16, such as antenna tuning and keying circuits being discussed. A brief discussion of shore-based transmitters is contained in chapter 6. Chapter 7 discusses various multicouplers used by the fleet; the types of switching systems used for the transfer of equipment control, antenna selections, and radar information distribution.

Chapter 8 presents teletype terminal equipment and facsimile equipment used throughout the Navy. The basic principles of communications multiplexing are presented in chapter 9 along with a discussion of telegraph terminal set AN/UCC-1(V). Electronic Aids to Navigation are presented in chapter 10 and 11, using the TACAN system AN/SRN-6, radio direction finder equipment AN/URD-4, and LORAN receiving set AN/UPN-12 as representative equipments. Radar is covered in chapter 12 and radar repeaters are discussed in chapter 13. Special

purpose equipment, such as electronic countermeasures, inertial navigation, and satellite communication equipment is presented in chapter 14. Chapter 15 is devoted entirely to Satellite Navigation which covers the overall Navy Satellite Navigation System and the AN/SRN-9 as the representative equipment.

The remainder of this chapter gives information on the enlisted rating structure, the Electronics Technician rating, requirements and procedures for advancement in rating, and references that will help you in working for advancement and in performing your duties as an Electronics Technician. Therefore, it is strongly recommended that you study this chapter carefully before beginning intensive study of the remainder of this training course.

Throughout this training course, the term "cycle," or "cycles per second," or the abbreviation CPS is used to express electrical frequency in cycles per second. Recently, in effecting world-wide standardization of various units of measurement, the term "Hertz" abbreviated H_z) has been adopted to express cycles per second and will be used in later revisions of this training course.

THE ENLISTED RATING STRUCTURE

The two main types of ratings in the present enlisted rating structure, are general ratings and service ratings.

GENERAL RATINGS identify broad occupational fields of related duties and functions. Some general ratings include service ratings; others do not. Both Regular Navy and Naval Reserve personnel may hold general ratings.

SERVICE RATINGS identify subdivisions or specialties within a general rating. Although service ratings can exist at any petty officer level, they are most common at the PO3 and PO2 levels. Both Regular Navy and Naval Reserve personnel may hold service ratings.

THE ELECTRONICS TECHNICIAN RATING

Electronics Technicians maintain, repair, calibrate, tune, and adjust electronic devices and equipment. Electronics Technician ratings are included in the personnel allowance for practically all Navy ships including repair ships and tenders. The rating was established in 1948, and includes service ratings ETR (radar) and ETN (communications) at petty officer 3 & 2 levels. For advancement to ET1, the candidate must be qualified in both service ratings. Enlisted classification codes for the rating are listed in the Manual of Navy Enlisted Classifications, NavPers 15105 (revised). Classification codes identify men with special training and/or qualifications on equipments such as special radars, AEW systems, automatic landing systems, communications security devices equipment, inertial navigation equipment, ground controlled approach systems, TACAN systems and electronic standards equipment.

ADVANCEMENT IN RATING

Some of the rewards of advancement in rating are easy to see. You get more pay. Your job assignments become more interesting and more challenging. You are regarded with greater respect by officers and enlisted personnel. You enjoy the satisfaction of getting ahead in your chosen Navy career.

But the advantages of advancing in rating are not yours alone. The Navy also profits. Highly trained personnel are essential to the functioning of the Navy. By each advancement in rating, you increase your value to the Navy in two ways. First, you become more valuable as a specialist in your own rating. And second, you become more valuable as a person who can train others and thus make far-reaching contributions to the entire Navy.

HOW TO QUALIFY FOR ADVANCEMENT

What must you do to qualify for advancement in rating? The requirements may change from time to time, but usually you must:

1. Have a certain amount of time in your present grade.
2. Complete the required military and occupational training courses.
3. Demonstrate your ability to perform all the PRACTICAL requirements for advancement by completing the Record of Practical Factors,

NavPers 1414/1. In some cases the Record of Practical Factors may contain the old form number, NavPers 760.

4. Be recommended by your commanding officer, after the petty officers and officers supervising your work have indicated that they consider you capable of performing the duties of the next higher rate.

5. Demonstrate your KNOWLEDGE by passing a written examination on (a) military requirements and (b) occupational qualifications.

Some of these general requirements may be modified in certain ways. Figure 1-1 gives a more detailed view of the requirements for advancement of active duty personnel; figure 1-2 gives this information for inactive duty personnel.

Remember that the requirements for advancement can change. Check with your division officer or training officer to be sure that you know the most recent requirements.

Advancement in rating is not automatic. After you have met all the requirements, you are ELIGIBLE for advancement. You will actually be advanced in rating only if you meet all the requirements (including making a high enough score on the written examination) and if the quotas for your rating permit your advancement.

HOW TO PREPARE FOR ADVANCEMENT

What must you do to prepare for advancement in rating? You must study the qualifications for advancement, work on the practical factors, study the required Navy Training Courses, and study other material that is required for advancement in your rating. To prepare for advancement, you will need to be familiar with (1) the Quals Manual, (2) the Record of Practical Factors, NavPers 1414/1, (3) a NavPers publication called Training Publications for Advancement in Rating, NavPers 10052, and (4) applicable Navy Training Courses. The following sections describe them and give you some practical suggestions on how to use them in preparing for advancement.

The Quals Manual

The Manual of Qualifications for Advancement in Rating, NavPers 18068 B (with changes), gives the minimum requirements for advancement to each rate within each rating. This manual is usually called the "Quals Manual," and the qualifications themselves are often called

ACTIVE DUTY ADVANCEMENT REQUIREMENTS

REQUIREMENTS *	E1 to E2	E2 to E3	#† E3 to E4	# E4 to E5	† E5 to E6	† E6 to E7	† E7 to E8	† E8 to E9
SERVICE	4 mos. service—or completion of recruit training.	6 mos. as E-2.	6 mos. as E-3.	12 mos. as E-4.	24 mos. as E-5.	36 mos. as E-6. 8 years total enlisted service.	36 mos. as E-7. 8 of 11 years total service must be enlisted.	24 mos. as E-8. 10 of 13 years total service must be enlisted.
SCHOOL	Recruit Training.		Class A for PR3, DT3, PT3. AME 3, HM 3			Class B for AGC MUC, MNC.		
PRACTICAL FACTORS	Locally prepared check-offs.	Records of Practical Factors, NavPers 1414/1, must be completed for E-3 and all PO advancements.						
PERFORMANCE TEST			Specified ratings must complete applicable performance tests before taking examinations.					
ENLISTED PERFORMANCE EVALUATION	As used by CO when approving advancement.		Counts toward performance factor credit in advancement multiple.					
EXAMINATIONS **	Locally prepared tests.	See below.	Navy-wide examinations required for all PO advancements.				Navy-wide, selection board.	
NAVY TRAINING COURSE (INCLUDING MILITARY REQUIREMENTS)		Required for E-3 and all PO advancements unless waived because of school completion, but need not be repeated if identical course has already been completed. See NavPers 10052 (current edition).					Correspondence courses and recommended reading. See NavPers 10052 (current edition).	
AUTHORIZATION	Commanding Officer		U.S. Naval Examining Center			Bureau of Naval Personnel		

* All advancements require commanding officer's recommendation .
† 1 year obligated service required for E-5 and E-6; 2 years for E-6, E-7, E-8 and E-9.
# Military leadership exam required for E-4 and E-5.
** For E-2 to E-3, NAVEXAMCEN exams or locally prepared tests may be used.

Figure 1-1.—Active duty advancement requirements.

INACTIVE DUTY ADVANCEMENT REQUIREMENTS

REQUIREMENTS *	E1 to E2	E2 to E3	E3 to E4	E4 to E5	E5 to E6	E6 to E7	E8	E9
TOTAL TIME IN GRADE	4 mos.	6 mos.	15 mos.	18 mos.	24 mos.	36 mos.	36 mos.	24 mos.
TOTAL TRAINING DUTY IN GRADE †	14 days	14 days	14 days	14 days	28 days	42 days	42 days	28 days
PERFORMANCE TESTS				Specified ratings must complete applicable performance tests before taking examination.				
DRILL PARTICIPATION	Satisfactory participation as a member of a drill unit.							
PRACTICAL FACTORS (INCLUDING MILITARY REQUIREMENTS)	Record of Practical Factors, NavPers 1414/1, must be completed for all advancements.							
NAVY TRAINING COURSE (INCLUDING MILITARY REQUIREMENTS)	Completion of applicable course or courses must be entered in service record.							
EXAMINATION	Standard Exam	Standard Exam or Rating Training.	Standard Exam required for all PO Advancements.				Standard Exam, Selection Board. Also pass Mil. Leadership Exam for E-4 and E-5.	
AUTHORIZATION	Commanding Officer		U.S. Naval Examining Center				Bureau of Naval Personnel	

* Recommendation by commanding officer required for all advancements.
† Active duty periods may be substituted for training duty.

Figure 1-2.—Inactive duty advancement requirements.

"quals." The qualifications are of two general types: (1) military requirements, and (2) occupational qualifications.

MILITARY REQUIREMENTS apply to all ratings rather than to any one particular rating. Military requirements for advancement to third class and second class petty officer rates deal with military conduct, naval organization, military justice, security, watch standing, and other subjects which are required of petty officers in all ratings.

OCCUPATIONAL QUALIFICATIONS are requirements that are directly related to the work of each rating.

Both the military requirements and the occupational qualifications are divided into subject matter groups; then, within each subject matter group, they are divided into PRACTICAL FACTORS and KNOWLEDGE FACTORS. Practical factors are things you must be able to DO. Knowledge factors are things you must KNOW in order to perform the duties of your rating.

In most subject matter areas, you will find both practical factor and knowledge factor qualifications. In some subject matter areas, you may find only practical factors or knowledge factors. It is important to remember that there are some knowledge aspects to all practical factors, and some practical aspects to most knowledge factors. Therefore, even if the Quals Manual indicates that there are no knowledge factors for a given subject matter area, you may still expect to find examination questions dealing with the knowledge aspects of the practical factors listed in that subject matter area.

In summary, then, the written examination for advancement in rating may contain questions relating to the practical factors and to the knowledge factors of both the military requirements and the professional qualifications. If you are working for advancement to second class, remember that you may be examined on third class qualifications as well as on second class qualifications.

You are required to pass a Navy-wide military/leadership examination for E-4 or E-5, as appropriate, before participating in the occupational examinations. The military/leadership examinations for both levels are given quarterly. Candidates are required to pass the applicable military/leadership examination only once. Each of these examinations consists of 100 questions based on information contained in the Manual of Qualifications for Advancement in Rating, NavPers 18068-B and Training Publications for Advancement in Rating, NavPers 10052.

The Navy-wide occupational examination for pay grades E-4 and E-5 will contain 150 questions related to occupational areas of your rating.

The Quals Manual is kept current by means of changes. The occupational qualifications for your rating which are covered in this training course were current at the time the course was printed. By the time you are studying this course, however, the quals for your rating may have been changed. Never trust any set of quals until you have checked it against an UP-TO-DATE copy in the Quals Manual.

Record of Practical Factors.

Before you can take the servicewide examination for advancement in rating, there must be an entry in your service record to show that you have qualified in the practical factors of both the military requirements and the occupational qualifications. A special form known as the RECORD OF PRACTICAL FACTORS, NavPers 1414/1 is used to keep a record of your practical factor qualifications. This form is available for each rating. The form lists all practical factors, both military and occupational. As you demonstrate your ability to perform each practical factor, appropriate entries are made in the DATE and INITIALS columns.

Changes are made periodically to the Manual of Qualifications for Advancement in Rating, and revised forms of NavPers 1414/1 are provided when necessary. Extra space is allowed on the Record of Practical Factors for entering additional practical factors as they are published in changes to the Quals Manual. The Record of Practical Factors also provides space for recording demonstrated proficiency in skills which are within the general scope of the rating but which are not identified as minimum qualifications for advancement.

If you are transferred before you qualify in all practical factors, the NavPers 1414/1 form should be forwarded with your service record to your next duty station. You can save yourself a lot of trouble by making sure that this form is actually inserted in your service record before you are transferred. If the form is not in your service record, you may be required to start all over again and requalify in the practical factors which have already been checked off.

NavPers 10052

Training Publications for Advancement in Rating. NavPers 10052 (revised), is a very important publication for anyone preparing for advancement in rating. This bibliography lists required and recommended Navy Training Courses and other reference material to be used by personnel working for advancement in rating. NavPers 10052 is revised and issued once each year by the Bureau of Naval Personnel. Each revised edition is identified by a letter following the NavPers number. When using this publication, be SURE that you have the most recent edition.

If extensive changes in qualifications occur in any rating between the annual revisions of NavPers 10052, a supplementary list of study material may be issued in the form of a BuPers Notice. When you are preparing for advancement, check to see whether changes have been made in the qualifications for your rating. If changes have been made, see if a BuPers Notice has been issued to supplement NavPers 10052 for your rating.

The required and recommended references are listed by rate level in NavPers 10052. If you are working for advancement to third class, study the material that is listed for third class. If you are working for advancement to second class, study the material that is listed for second class; but remember that you are also responsible for the references listed at the third class level.

In using NavPers 10052, you will notice that some Navy Training Courses are marked with an asterisk (*). Any course marked in this way is MANDATORY—that is, it must be completed at the indicated rate level before you can be eligible to take the servicewide examination for advancement in rating. Each mandatory course may be completed by (1) passing the appropriate enlisted correspondence course that is based on the mandatory training course; (2) passing locally prepared tests based on the information given in the training course; or (3) successfully completing an appropriate Class A course.

Do not overlook the section of NavPers 10052 which lists the required and recommended references relating to the military requirements for advancement. Personnel of ALL ratings must complete the mandatory military requirements training course for the appropriate rate level before they can be eligible to advance in rating.

The references in NavPers 10052 which are recommended but not mandatory should also be studied carefully. ALL references listed in NavPers 10052 may be used as source material for the written examinations, at the appropriate rate levels.

Navy Training Courses

There are two general types of Navy Training Courses. RATING COURSES (such as this one) are prepared for most enlisted ratings. A rating training course gives information that is directly related to the occupational qualifications of ONE rating. SUBJECT MATTER COURSES or BASIC COURSES give information that applies to more than one rating.

Navy Training Courses are revised from time to time to keep them up to date technically. The revision of a Navy Training Course is identified by a letter following the NavPers number. You can tell whether any particular copy of a Navy Training Course is the latest edition by checking the NavPers number and the letter following this number in the most recent edition of List of Training Manuals and Correspondence Courses, NavPers 10061. (NavPers 10061 is actually a catalog that lists all current training courses and correspondence courses; you will find this catalog useful in planning your study program.)

Navy Training Courses are designed to help you prepare for advancement in rating. The following suggestions may help you to make the best use of this course and other Navy training publications when you are preparing for advancement in rating.

1. Study the military requirements and the occupational qualifications for your rating before you study the training course, and refer to the quals frequently as you study. Remember, you are studying the training course primarily in order to meet these quals.

2. Set up a regular study plan. It will probably be easier for you to stick to a schedule if you can plan to study at the same time each day. If possible, schedule your studying for a time of day when you will not have too many interruptions or distractions.

3. Before you begin to study any part of the training course intensively, become familiar with the entire book. Read the preface and the table of contents. Check through the index. Look at the appendixes. Thumb through the book without any particular plan, looking at the

illustrations and reading bits here and there as you see things that interest you.

4. Look at the training course in more detail, to see how it is organized. Look at the table of contents again. Then, chapter by chapter, read the introduction, the headings, and the subheadings. This will give you a pretty clear picture of the scope and content of the book. As you look through the book in this way, ask yourself some questions: What do I need to learn about this? What do I already know about this? How is this information related to information given in other chapters? How is this information related to the qualifications for advancement in rating?

5. When you have a general idea of what is in the training course and how it is organized, fill in the details by intensive study. In each study period, try to cover a complete unit—it may be a chapter, a section of a chapter, or a subsection. The amount of material that you can cover at one time will vary. If you know the subject well, or if the material is easy, you can cover quite a lot at one time. Difficult or unfamiliar material will require more study time.

6. In studying any one unit—chapter, section, or subsection—write down the questions that occur to you. Many people find it helpful to make a written outline of the unit as they study, or at least to write down the most important ideas.

7. As you study, relate the information in the training course to the knowledge you already have. When you read about a process, a skill, or a situation, try to see how this information ties in with your own past experience.

8. When you have finished studying a unit, take time out to see what you have learned. Look back over your notes and questions. Maybe some of your questions have been answered, but perhaps you still have some that are not answered. Without looking at the training course, write down the main ideas that you have gotten from studying this unit. Don't just quote the book. If you can't give these ideas in your own words, the chances are that you have not really mastered the information.

9. Use Enlisted Correspondence Courses whenever you can. The correspondence courses are based on Navy Training Courses or on other appropriate texts. As mentioned before, completion of a mandatory Navy Training Course can be accomplished by passing an Enlisted Correspondence Course based on the Navy Training Course. You will probably find it helpful to take

other correspondence courses, as well as those based on mandatory training courses. Taking a correspondence course helps you to master the information given in the training course, and also helps you see how much you have learned.

10. Think of your future as you study Navy Training Courses. You are working for advancement to third class or second class right now, but someday you will be working toward higher rates. Anything extra that you can learn now will help you both now and later.

SOURCES OF INFORMATION

Besides training courses, NavPers 10052 lists official publications on which you may be examined. You should not only study the sections required, but should become as familiar as possible with all publications you use.

One of the most useful things you can learn about a subject is how to find out more about it. No single publication can give you all the information you need to perform the duties of your rating. You should learn where to look for accurate, authoritative, up-to-date information on all subjects related to the military requirements for advancement and the occupational qualifications of your rating.

PUBLICATIONS YOU SHOULD KNOW

Electronic technical publications include various handbooks, bulletins, and manuals published and distributed by the Ship Systems Command, and manufacturers' technical manuals. The Requisitioning Guide and Index of Forms and Publications, NavSandA 2002, furnishes a complete list of NavShips technical publications along with instruction for ordering copies.

One bulletin of great importance to the ET is Electronics Information Bulletin (EIB), NavShips 0967-001-3(---), published bi-weekly for naval electronics activities. A complete file of these bulletins should be maintained.

This bulletin lists field changes and corrections that must be made in instruction books and other publications that are used in the maintenance of electronic equipment. It also lists electronics publications that become available, and gives valuable suggestions, from case histories, for servicing electronic equipment.

The recently revised Electronics Installation and Maintenance Book (EIMB) furnishes data applying to all types of electronics equipment including the theory of operation of basic

electronic circuits. The EIMB is organized and issued as 12 separate handbooks, each with its own NavShips number. A complete listing of all handbooks, stock numbers, and most recent changes can be found in current issues of the EIB.

Another NavShips publication of importance to the ET is the Naval Ships Technical Manual, NavShips 250-000. This manual is the basic doctrine publication of the Naval Ship Systems Command. Chapter 9670 (formerly chap. 67) is of particular importance to the ET and should be a part of the ET library.

The Naval Ship Systems Command Technical News is a monthly publication which contains interesting and useful articles on all aspects of shipboard engineering. This magazine is particularly useful because it presents information which supplements and clarifies information contained in the Naval Ships Technical Manual and because it presents information on new developments.

The manufacturers' technical manuals that are furnished with most electronic equipment are valuable sources of information on operation, maintenance and repair. The manufacturers' technical manuals that deal with Naval Ship Systems Command equipment are usually given NavShips numbers.

Some publications are subject to change or revision from time to time—some at regular intervals, others as the need arises. When using any publication that is subject to change or revision, be sure that you have the latest edition. When using any publication that is kept current by means of changes, be sure you have a copy in which all official changes have been made. Studying canceled or obsolete information will not help you to do your work or to advance in rating; it is likely to be a waste of time and may even be seriously misleading.

TRAINING FILMS

Training films available to naval personnel are a valuable source of supplementary information on many technical subjects. A selected list of training films that may be useful to you is given in Appendix I of this training course. Other films that may be of interest are listed in the United States Navy Film Catalog, NavWeps 10-1-777. This catalog, published in 1966, supersedes three earlier publications: the former catalog with the same title but numbered NP 10000-A; the Supplement, NavWeps 10-1772; and the Navy Classified Film Catalog, NavPers 10001-A.

As you know, procedures sometimes change rapidly. Thus some film become obsolete rapidly. If a film is obsolete only in part, it may still have sections that are useful, but it is important to note procedures that have changed. If there is any doubt, verify current procedures by looking them up in EIMB, EIB, Naval Ships Technical Manual, or other applicable sources.

CHAPTER 2

MAINTENANCE

Maintenance, as defined by the Naval Ships Technical Manual, is the function of retaining material in, or restoring it to, a serviceable condition. Its phases include servicing, repair, modification, modernization, overhaul, and testing.

PREVENTIVE MAINTENANCE

Preventive or scheduled maintenance is defined as those measures taken periodically to achieve maximum efficiency in performance, to ensure continuity of service, to reduce major breakdowns, and to lengthen the useful life of the equipment or system. This form of maintenance consists principally of cleaning, lubrication, and periodic tests aimed at discovering conditions which, if not corrected may lead to malfunctions.

To realize optimum results from the regular functional tests, a careful record of the performance data on each equipment must be kept. The value of these records may be demonstrated in a number of ways. Comparison of data taken on a particular equipment at different times may reveal slow, progressive drifts that may be too small to show up significantly in any one test. While the week-to-week changes may be slight, they should be followed carefully so that necessary replacements or repairs may be effected before the margin of performance limits is reached. Any marked variations should be regarded as abnormal, and should be investigated immediately. Another advantage in keeping systematic records of performance and servicing data is that maintenance personnel develop a more rapid familiarization with the equipment involved. The accumulated experience contained in the records serves as a guide to swift and accurate troubleshooting.

THE 3-M SYSTEM

The Navy Maintenance and Material Management (3-M) System has been implemented in the Navy as an answer to the ever-present problem of maintaining a high degree of material readiness. Although the (3-M) System is designed to improve the degree of readiness, its effectiveness and reliability are dependent upon you, the individual. The accuracy with which you perform your work, along with the neat and complete recording of required data on the prescribed forms, is one of the keys to the degree of readiness of your ship and therefore is a reflection of your success as a petty officer.

The two basic elements of the 3-M System are the Planned Maintenance System (PMS) and the Maintenance Data Collection System (MDCS). The Planned Maintenance System provides a uniform system of planned preventive maintenance. The Maintenance Data Collection System provides a means of collecting necessary maintenance and supply data, in a form suitable for rapid machine processing. A Man-hour Accounting System is used aboard repair ships and tenders in conjunction with the Maintenance Data Collection System.

Preventive maintenance should not be confused with corrective maintenance. Preventive maintenance is a scheduled check on select parts of a piece of equipment. Corrective maintenance is the repair of equipment.

As a third or second class ET, you will be concerned with both the Planned Maintenance System (PMS) and certain portions of the Maintenance Data Collection System (MDCS) as discussed in this chapter.

THE PLANNED MAINTENANCE SYSTEM

Planned maintenance reduces casualties and saves the cost of major repairs. The PMS is

designed to define the minimum planned mainte- nance required, to schedule and control per- formance, to describe the methods and tools to be used, and to aid in the prevention and detection of impending casualties.

In establishing minimum planned mainte- nance requirements for each piece of equip- ment, the Naval Ships Technical Manual, manu- facturers' technical manuals, and applicable drawings are critically reviewed. If the planned maintenance requirements are found to be un- realistic or unclear, they are modified or com- pletely revised before being incorporated into the PMS.

It is possible that the planned maintenance prescribed in the PMS may conflict with that prescribed in other documents such as the Naval Ships Technical Manual. Should this happen, it must be remembered that the PMS supersedes and takes precedence over exist- ing requirements set forth in various technical publications inso/far as preventive maintenance is concerned.

The Planned Maintenance
System Manual

A master Planned Maintenance System Man- ual is tailored to each department of a specific ship. It contains the minimum planned main- tenance requirements for each maintainable component installed in that particular depart- ment. The department Master Manuals are normally kept in the department offices and are primarily used by the department heads in planning and scheduling the maintenance requirements in their departments.

The departmental Master Manual contains a section for each division or maintenance group within the department. Each divisional section contains a table of contents and a Maintenance Index Page (MIP) for each sys- tem, subsystem, or component involved.

Applicable portions of the PMS Manual (referred to as Space Manuals) are kept in the working space for the equipment involved and serve as a ready reference to the planned maintenance required. Each MIP contains a brief description of the maintenance require- ments and the frequency with which mainte- nance is to be effected. The frequency code is: D—daily, W—weekly, M—monthly, Q—quar- terly, S—semiannually, A—annually, C—over- haul cycle, and R—situation requirement. The frequency codes for the daily, weekly, monthly,

quarterly, semiannual, and annual planned main- tenance actions are self-explanatory. Certain planned maintenance actions are performed once each operational cycle between shipyard over- hauls, during a specified quarter of the cycle, and are designated by the code C. The code R is used to identify those planned maintenance actions which are to be performed prior to getting underway, after a specified number of hours of operation, or to meet other require- ments that only arise during a specific situa- tion (prior to overhaul, for example).

Figure 2-1 shows a Maintenance Index Page (MIP) taken from a typical PMS Manual. In- formation to be found on the MIP includes the system or component involved, a short descrip- tion of each maintenance requirement, the main- tenance frequency code plus a consecutive num- ber starting with "1" for each frequency code assigned, the rate or rates recommended to perform the maintenance, the average time re- quired to perform the maintenance, and related maintenance requirements. The related main- tenance listed is additional planned maintenance which can be completed prior to, in conjunction with, or immediately after the scheduled main- tenance.

Because shipboard application of the PMS will vary slightly from one ship to another, the information found on the MIPs regarding rates recommended to perform the maintenance and the average time required for the task re- quires certain clarification. The maintenance taska are actually performed by personnel available and capable, regardless of what rate is listed on the MIP. The average time re- quired, as listed on the MIP, does not take into account the time required to assemble the necessary tools and materials nor the time required to clean the area and put away the tools at the end of the task.

SCHEDULING OF PLANNED MAINTE- NANCE.—For each division or maintenance group, a cycle schedule which provides a visual display of the planned maintenance requirements (based on the operational cycle of the ship between shipyard overhauls) is displayed in the departmental office. The information to be found on the cycle schedule (fig. 2-2), for any particular division or maintenance group in- cludes the MIP number (column 1) from the PMS Manual, a listing of all the equipment within that particular group for which planned maintenance is required, and the specific quarter

System, Subsystem, or Component					Reference Publications				
AN/USM-105A, 105V Oscilloscope					NAVSHIPS 93658(A)				

Bureau Card Control No.					Maintenance Requirement	M.R. No.	Rate Req'd.	Man Hours	Related Maintenance
CT	290ATG2	85	BA38	W	1. Clean air filter.	W-1	ETSN	0.1	None
CT	290ATG2	85	BA39	M	1. Verify speed of blower motor.	M-1	ETSN	0.1	None
CT	290ATG2	85	BA40	Q	1. Clean and inspect. 2. Lubricate the blower motor.	Q-1	ETSN	0.4	Q-2
CT	290ATG2	85	BA41	Q	1. Test operation.	Q-2	ET3	0.4	None
CT	290ATG2	85	BA42	S	1. Inventory accessories. 2. Test operation. 3. Deliver oscilloscope to calibration activity.	S-1	ETSN	0.2	Q-2
					These maintenance cards were prepared for this equipment in which the following field changes have been accomplished: 1 Of these, the following field changes affect the maintenance actions: none New maintenance requirement cards and maintenance index pages will be made available as future field changes are accomplished that affect the prescribed planned maintenance.				

MAINTENANCE INDEX PAGE
OPNAV FORM 4700-3 (4-64)

BUREAU PAGE CONTROL NUMBER   T-3/2-85

98.171

Figure 2-1.—Maintenance Index Page.

EQUIP PAGE	TYPE CLASS MAINTENANCE GROUP COMPONENT	SCHEDULE AS INDICATED				EACH QUARTER
		QUARTER AFTER OVERHAUL				
		1 5 9	2 6 10	3 7 11	4 8 12	
(1)	COMMUNICATIONS					
C-1	AN/URR-35A					M-1
C-2	TED					M-1
C-3	AM/1365-URT					M-1
C-11	AN/WRR-2		S-1		S-1	M-1, Q-1
C-12	MAY-1	S-1		S-1		

CYCLE SCHEDULE   OPNAV FORM 4700-4 (4-64) 0107-766-4000   C-40996

98.172

Figure 2-2.—Cycle schedule.

in which the semiannual, annual, and overhaul cycle planned maintenance actions are to be performed. The cycle schedule also lists the quarterly and the situation requirement planned maintenance actions which must be scheduled, as well as the monthly planned maintenance requirements.

The cycle schedule is used by the department head, in conjunction with his division officers and leading petty officers, to make out the quarterly planned maintenance schedule. The quarterly schedule (fig. 2-3) displayed adjacent to the cycle schedule in a holder known as the maintenance control board, gives a visual display of the ship's deployment schedule and the planned maintenance to be performed during that particular quarter.

The quarterly schedule has thirteen columns, one for each week in the quarter, for the scheduling of maintenance throughout the 3-month period.

At the end of each week, the leading petty officer of the division or maintenance group will update the quarterly schedule by crossing out (with an X) the preventive maintenance performed. If a planned maintenance action is not completed during the week it is scheduled, the leading petty officer will circle the action on the quarterly schedule. The uncompleted maintenance is then rescheduled (as shown in figure 2-3) for another week within the same quarter, normally as soon as the operating schedule of the ship will allow.

At the close of each quarter, the quarterly schedule is removed from its holder and retained on board as a record of the planned
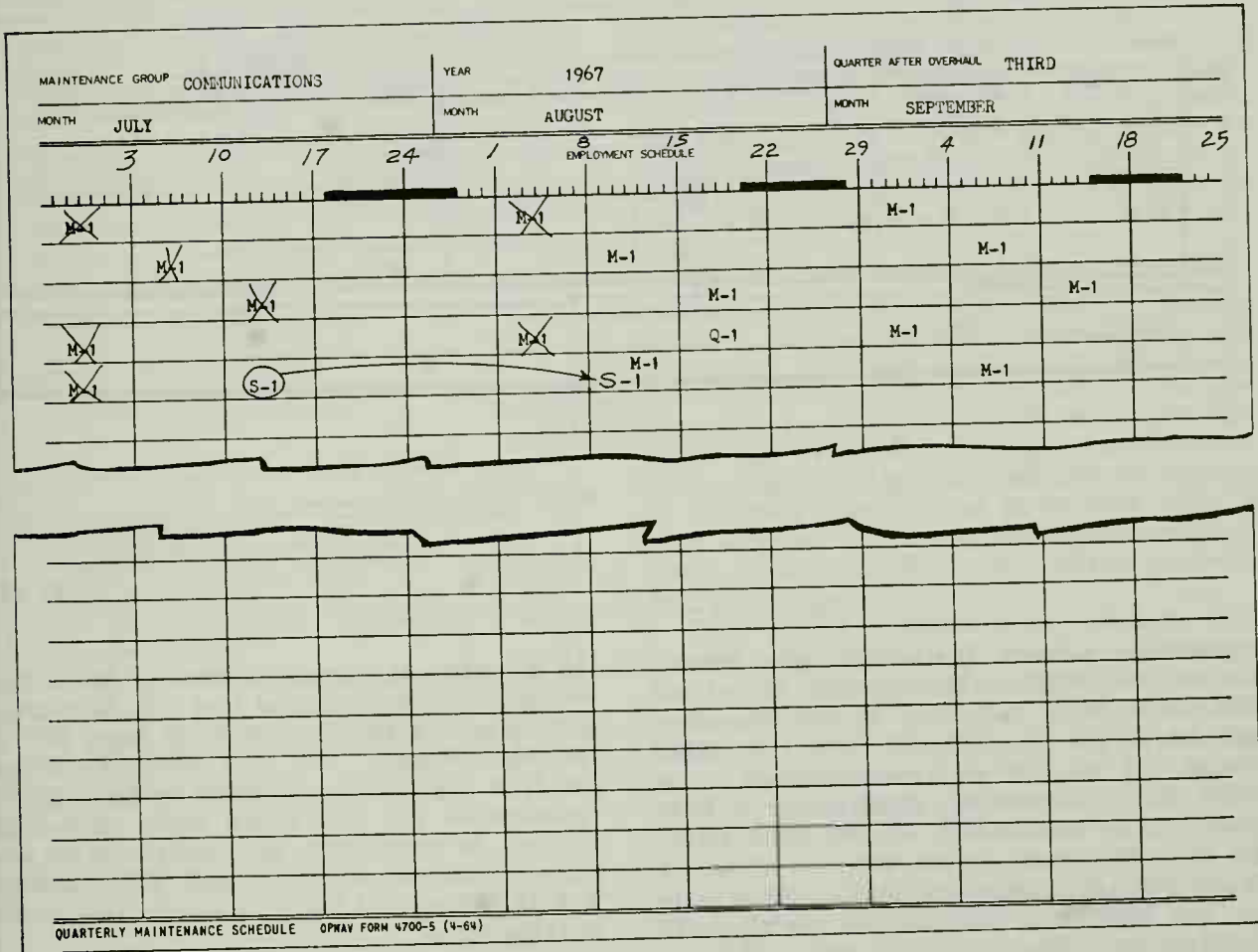
98.173

Figure 2-3.—Quarterly maintenance schedule.

maintenance completed. This record may be discarded at the beginning of the second quarter after the next shipyard overhaul.

The quarterly schedule is also used by your leading petty officer to make out a weekly planned maintenance schedule which is posted in the space where you work. The weekly schedule of planned maintenance should not be considered as the total work for the week. This schedule only covers the scheduled planned maintenance and is in addition to the other routine work, upkeep, and corrective maintenance to be done.

The weekly schedule provides a list of the components in the working area, the appropriate page number of the PMS Manual, and spaces for the leading petty officer to use in the

assignment of planned maintenance tasks to specified personnel. The daily and weekly planned maintenance actions are preprinted on the forms and the other maintenance actions are written in by the leading petty officer as required. When the leading petty officer is assured that a maintenance task has been completed, he crosses out the maintenance requirement number on the weekly schedule. If for some reason a task cannot be completed on the day scheduled, the leading petty officer will circle the maintenance requirement number and reschedule it for another day. The status of scheduled maintenance is readily available by looking at the weekly schedule.

A sample weekly schedule is shown in figure 2-4. In addition to the weekly

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GROUP	RADAR			WORK SCHEDULE FOR WEEK OF		AUGUST 14-20				
COMPONENT (1)	MAINTENANCE RESPONSIBILITY (2)	PAGE (3)	MONDAY	TUESDAY	WEDNESDAY (4)	THURSDAY	FRIDAY	SAT/SUN	OUTSTANDING REPAIRS AND P.M. CHECKS DUE IN NEXT 4 WEEKS (5)
AN SPS-10	CRIDER, RD2	R-6			W3 D1	D1	D1	D1	
AN SPS-T1	WILLIAMS, ET1	R-7		(M3)		→M3			
VK-5	DWYER, ET3	R-8			S1				A-1
COFFEE MESS	BAUMBACH, SA		D1	D1	D1	D1	D1	D1	

SAMPLE WEEKLY WORK SCHEDULE

FORMAT INCLUDES ALL DAILY & WEEKLY CHECKS PREPRINTED

WHERE USED: Posted in Maintenance Group work space in aluminum holder.　WHO USES: Maintenance Group Supervisors and maintenance personnel.

Figure 2-4.—Weekly work schedule.

98.175

maintenance actions, the leading petty officer has assigned Williams the responsibility of requirements M-1, M-2, and M-3 on Tuesday, and Dwyer the responsibility of S-1 on Wednesday. In the far right-hand column, it is noted that maintenance requirement A-1 is listed under outstanding repairs (due within the next 4 weeks). In the schedule shown in figure 2-4, all maintenance scheduled for Monday was completed, but Williams was unable to complete the scheduled requirement M-3 for Tuesday, and the action is rescheduled for Thursday of the same week.

THE MAINTENANCE REQUIREMENT CARD.—The Maintenance Requirement Card (MRC) 5 by 8 inches, has the planned maintenance task defined in sufficient detail to enable assigned personnel to perform the task (see fig. 2-5). A master set of MRCs is maintained in the departmental office and those applicable to the equipment with which you are concerned are maintained in your working space. If a card in the working space becomes lost or mutilated, a new card can be made from the master set and used until a feedback report is sent in and a new card obtained.

The MRC is one of the primary tools of the PMS with which you will be concerned. Suppose that on a Monday morning Dwyer looks at the weekly schedule and finds that he is assigned the maintenance action identified as M-1. The weekly schedule indicated that this particular maintenance action is listed on page T-3 of the PMS Manual. The MRC which describes the task assigned is identified by the number combination T-3 M-1 in the upper right-hand corner. In preparation for performing the assigned task, Dwyer would pull MRC number T-3 M-1 (fig. 2-5) from the set of cards in the working space.

The MRC identifies the component involved; gives a brief description of the maintenance requirement; lists the safety precautions to be followed; and lists the tools, parts, and materials needed to accomplish the task. This information is listed to enable you to be ready to perform completely the prescribed maintenance prior to actually working on the equipment involved. The procedure listed on the MRC is standardized and is the best known method of performing that particular task. Any related maintenance requirement listed on the MRC should be done at the same time or in conjunction with the assigned task for the purpose of time conservation.

The 16-digit number on the lower right-hand side of the MRC is the bureau card control number which can also be found on the MIP. Each MRC has a bureau card control number, which must be referred to in any correspondence concerning the card.

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SYSTEM Communications and Control	COMPONENT AN/USM-105A, 105V Oscilloscope	M.R. NUMBER T-3 M-1	
	MAINT. SIGNI- FICANT NO.		
SUB-SYSTEM Test Equipment	RELATED M.R. None	RATES	M/H
		ETSN	0.1
M.R. DESCRIPTION 1. Verify speed of blower motor.		TOTAL M/H 0.1	
		ELAPSED TIME 0.1	

SAFETY PRECAUTIONS

1. Observe standard safety precautions.

TOOLS, PARTS, MATERIALS, TEST EQUIPMENT

1. Stroboscope, TS-805A/U or equivalent

PROCEDURE

1. Verify Speed of Blower Motor.
   a. De-energize the oscilloscope.
   b. Remove air filter from rear of cabinet.
   c. Energize stroboscope and adjust to measure 1750 rpm.
   d. Energize the oscilloscope.

CAUTION: Do not operate the oscilloscope longer than
         necessary to perform step c with filter
         removed.

   e. Measure speed of blower motor using the strobos-
      cope.

NOTE: If speed of motor has dropped to 1600 rpm, or
     less, the motor should be disassembled, cleaned,
     and lubricated.

   f. Return equipment to normal condition.

(Right column codes: CT 290A TG2 85 BA39 M)

LOCATION   Date 2 August 1965

98.176

Figure 2-5.—Maintenance requirement card.

On some ships, two or more divisions may have identical equipment. When this occurs, each division will have separate (but identical) MRC cards for the equipment.

Feedback Report

The PMS Feedback Report, OPNAV Form 4700-7 (fig. 2-6), is designed to report any discrepancies or suggested improvements in the PMS as installed aboard ship. The report is to be filled out by the man who discovers the discrepancy or suggests an improvement, signed by anyone designated by the commanding officer, and mailed via the type commander to the appropriate field office listed on the reverse side of the originator's copy of the form. Atlantic Fleet ships use the Norfolk address and Pacific Fleet ships use the San Diego address.

When submitting a feedback report, be sure it is filled out completely and legibly. Handwritten copies are acceptable but a ballpoint pen must be used to ensure that all copies are legible. Instructions for filling out the feedback report are listed on the back of the form and are discussed here for the purpose of clarification.

You are required to complete the "TO" line with the appropriate (San Diego or Norfolk) address. In the "FROM" line insert the hull number of your ship (DDG 11, DD789, AD 36, etc.); and in the "VIA" line indicate the type commander. The date will be the day you discover the discrepancy or suggest an improvement. The serial number indicates the number of feedback reports sent in by your ship. If your ship has previously forwarded five feedback reports, the next report will be numbered 6.

The blocks concerning the system, subsystem, component, M.R. number, and the bureau card control number should be filled in with information from the appropriate MIP and/or MRC. If the MRC or MIP is missing and the bureau card control number is not available, the equipment should be identified by its name and its APL/CID or AN number. The APL/CID number identifies the parts list for the equipment and is found in the index to the Coordinated Shipboard Allowance List (COSAL). The "Technical Publication" block in the "Discrepancy" section of the report is intended to include all Naval Ship Systems and Naval Ordnance Systems Command publications. In the event that any of these are in error, identify the publication number, volume, revision, date, change number, page paragraph, and/or figure.

Prior to the distribution of the report (as indicated on the original and each of the four carbon copies), check it for completeness. The discrepancy should always be explained clearly and a suggestion for its correction must always be offered.

THE MAINTENANCE DATA
COLLECTION SYSTEM

The Maintenance Data Collection System is designed to provide a means of recording maintenance actions in substantial detail so that a great variety of information may be collected
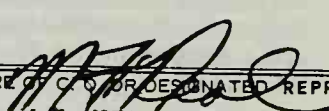
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INSTRUCTIONS ON BACK OF GREEN PAGE

FROM: DD-712

TO: Bushiips/Buweps Maintenance
Management Field Office
Norfolk, Virginia 23511

SERIAL #: 6

DATE August 16, 1967

VIA: ComCruDesLant

SUBJECT: PLANNED MAINTENANCE SYSTEM FEEDBACK REPORT

SYSTEM Communication
and Control

COMPONENT AN/USM-105A, 105V
Oscilloscope

SUB-SYSTEM
Test Equipment

M.R. NUMBER T-3 M-1

BU. CONTROL NO. CT 290A TG285 BA39 M

DISCREPANCY

[X] M. R. Description

[ ] Safety Precautions

[ ] Tools, Etc.

[ ] Missing Maintainance
Requirement Card (MRC)

[ ] Equipment Change

[ ] Missing Maintainance
Index Page (MIP)

[ ] Technical

[ ] Procedure

[ ] Typographical

[ ] Technical
Publications

[ ] Miscellaneous

MRC procedure is to adjust stroboscope to measure
blower motor speed at 1750 RPM.

AN/USM-105A aboard has Field Change #4, requiring
a blower motor speed of 2400 RPM.

Recommend rewording of MRC procedure:

Item #1.c. - Energize stroboscope and adjust
to measure 2400 RPM.

NOTE: If speed of motor has dropped to 2250 RPM
or less, the motor should be disassembled,
cleaned, and lubricated.

SIGNATURE OF C.O. OR DESIGNATED REPRESENTATIVE
M.T. NEAL LT, USN.

THIS COPY FOR: **ADDRESSEE** I

OPNAV Form 4700/7 (10-65)

Figure 2-6.—PMS Feedback Report, OPNAV 4700-7.

40.101

ncerning maintenance actions and the per-
frmance of equipment involved. In addition,
te system provides data concerning the initial
cscovery of a malfunction, how the equipment
ralfunctioned, how many man-hours were ex-
ynded, which equipment was involved, what
ipair parts and materials were used, what de-
lys were incurred, the reasons for delay, and
te technical specialty or work center which
yrformed the maintenance.

Codes must be used in recording maintenance
ations in order to allow converting the informa-
tn to a language that can be read by automatic
cta processing machines. As a third or second
rass Electronic Technician, you will be required
t fill out various maintenance forms, using the
ydes listed in the Equipment Identification
tode (BIC) Manual. The forms are sent to a
tta processing center where the coded infor-
iation is punched into cards which are machine
yocessed to produce the various reports for
te in maintenance and material management.

The reports that are produced by the auto-
iatic data processing machines will be ac-

curate and useful only if the information is
clearly and accurately entered on the main-
tenance forms. Therefore it is very important
that the codes which you enter on the forms be
accurate and clearly written.

Maintenance Data Collection
System Forms

The MDCS forms on which you may be re-
quired to make coded entries include OPNAV
Form 4700-2B (Shipboard Maintenance Action),
OPNAV Form 4700-2C (Work Request), and
OPNAV form 4700-2D (Deferred Action). De-
tailed descriptions of the entries to be made
on these forms are listed in the EIC Manual
and in chapter 3 of the 3-M Manual (OPNAV
43P2). Brief descriptions of these entries are
also listed here to help you become familiar
with the forms, which are shown in figures
2-7 through 2-10.

Block A, Ship Name and Hull No./Activity:
Enter the name and hull number of your ship--
for example, USS Now (DD 746).

17.81B

Figure 2-7.—Shipboard Maintenance Action, OPNAV 4700-2B(8-64).

17.81D

Figure 2-8.—Deferred Action, OPNAV Form 4700-2D (8-64), sheet 1.

Block 1, Administrative Organization: Enter the 4-letter/number code (from section II of the EIC Manual) to which your ship is assigned (if your ship is assigned to DESRON 7, you would enter D070).

Block 2, Unit Identification Code (UIC): Enter the UIC number for the ship or activity for which the maintenance was performed. This entry is obtained from NavCompt Manual, Volume 2, chapter 5.

Block 3, Maintenance Control Number: This number is used to relate all documents submitted on a specific job and is filled in by the maintenance control section of your ship before the documents leave the ship. The numbers will be assigned consecutively beginning with 0001 and ending at 9999. Upon reaching 9999, the series will be started again at 0001.

Block 4, Date: Enter the day, month, and year the maintenance is accomplished; or, for a work request, enter the date the request is submitted.

Block B, Type Availability: This block is left blank on all forms except the Work Request,

OPNAV Form 4700-2C. On the work request, this block will be filled in with a single-letter code, taken from section IX of the EIC Manual, to identify the type of availability in which the repair activity will perform the maintenance.

Block 5, Equipment Identification Code: Enter, from section X of the EIC Manual, the appropriate 7-letter/number code which identifies the system, subsystem, component, and/or lowest designated assembly on which maintenance is requested or performed. Caution must be exercised to ensure that the EIC number correctly identifies the exact item on which maintenance is requested or performed.

Block 6, Work Center: Enter the 3-letter/number code from section III of the EIC Manual which identifies the department, rating, shop, or group which performs the maintenance actions on the equipment identified in the equipment identification code, block 5.

Block 7, Assisting Work Center: Leave blank. When information is required in this block it will be filled out by your supervisor

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17.81C

Figure 2-9.—Work Request, OPNAV 4700-2C (8-64).

in accordance with the EIC Manual and chapter 3 of the 3-M Manual.

Block 8, Repair Activity Unit Identification Code (UIC) obtained from NavCompt Manual, Volume 2, chapter 5.

Block 9, How Mal/MRC: If the action is a result of a malfunction, list the 3-digit code from section IV of the EIC Manual which best describes the trouble. If the action is planned maintenance (MRC), enter a 3-letter/number combination code which identifies the frequency check on the equipment. For example, if the situation is an M-1 PMS action, enter M01. For all other actions three zeroes (000) will be entered.

Block 10, Discovered: Enter the appropriate code from section V of the EIC Manual which identifies when the maintenance requirement was discovered.

Block 11, Action Taken: Enter the code from section VI of the EIC Manual which best describes the maintenance action taken. "Not Repairable This Ship" (NRTS) codes will be used only by repair activities.

Block 12, Units: Enter the number of identical items, identified in block 5, on which the same maintenance action was attempted, performed, or requested. If identical alterations or significant maintenance actions are accomplished on identical equipment bearing different serial numbers, each equipment will be reported on a separate document.

Block 13, Man-hours: On completion of the maintenance action, enter the total man-hours (to the nearest tenth) expended by all shipboard personnel. If the task was completed in 1-1/2 hours, for example, you would enter "0015". Enter a "0" in any space not used.

Block E, Blank: Leave blank, (explained later).

Block 14, Serial Number: When several equipments aboard a ship have identical EIC's, Block 14 is used to identify the specific equipment on which maintenance is performed. This information is required for accountability at the System Command level and for equipment history purposes at the local level. Serial numbers

will be used for electronics and ordnance equipments. For example, the serial number of the radar, the radio, the computer, or the test equipment will be reported; the serial numbers of the major components of these equipment, however, will not be reported even though all work is done in a component such as a radar transmitter assembly.

Blocks 15 thru 19 (block 19 not shown) apply only to OPNAV 4700-2C, Fig. 2-9.

Block 15, T/A: Leave blank.

Block 16, Requesting Work Center: Enter the letter/number code that identifies the department, rating, shop, or group that is requesting assistance from the repair activity. Work center codes may be found in section III of the EIC Manual.

Block 17, Desired Completion Date: Enter the day, month, and year that the outside repair activity must complete maintenance in order for the requesting activity to meet its operational commitments. If this date were 28 Aug. 1967, for example, you would enter 28087.

Block 18, Service: Enter the single-letter code from section VII of the EIC Manual which identifies the type of assistance the requesting ship will furnish the repair activity in completing the requested maintenance.

Block 19, Scheduled Start Date: This entry, which is for repair activities only, is the estimated starting date of the requested maintenance.

Block D, Leave Blank.

Block E and 20, Equipment Downtime and Equipment Operating Time: These entries are required only for equipment identified by green pages in the EIC Manual. Detailed instructions for these blocks may be found in section I of the EIC Manual and chapter 3 of the 3-M Manual.

Block 21, Alteration Identification: If an alteration is to be accomplished, this block will contain the alteration identification from the authorization directive.

The Shipboard Maintenance Action Form (fig. 2-7) (OPNAV Form 4700-2B) is a single-sheet document used to record the completion of planned maintenance actions, corrective

Figure 2-10.—Deferred Action, OPNAV 4700-2D (8-64), sheet 2.

17.81D

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maintenance actions and authorized alterations that have been performed at the shipboard level by shipboard personnel. All planned maintenance actions except daily and weekly planned maintenance actions must be recorded on this form in addition to checking them off on the weekly and quarterly schedules. Routine preservation such as chipping, painting, and cleaning should not be reported.

Any corrective maintenance actions that are deferred because of the ship's operations, lack of repair parts, or the requirement of outside assistance are reported on the Deferred Action Form (fig. 2-8)(OPNAV 4700-2D). The first sheet of this 2-sheet document is used to record and report the reason for deferral and the second sheet is used to report the completion of the maintenance action. If a corrective maintenance action must be deferred, this form is required at the time a malfunction is noted. A sample sheet 1 of the Deferred Action Form which could have resulted is shown in figure 2-8. The action taken code, block 11, and the man-hours expended, block 13, on sheet 1 of the Deferred Action Form apply only to the action taken man-hours involved at the time of the discovery of the malfunction.

Figure 2-9 shows a Work Request (OPNAV Form 4700-2C) to have the motor starter rewound. In completing block F (Description/Remarks) of the work request, note that:

1. The component must be identified by name.

2. The CID number of the component must be listed.

3. The alterations number must be given, if applicable. If not applicable, it must be listed as N/A.

4. A description of what is wrong with the equipment and what you desire the repair activity to do must be given.

The signature blocks G, H, and J are to be signed by three leading petty officers, one from each duty section, who are familiar with the work to be done by the repair activity. Signature block K will be used by the commanding officer or his authorized representative to indicate command approval of the request for repair assistance.

After the repairs have been completed by the repair activity, "sheet 2" of the Deferred Action Form (fig. 2-10) is completed by entering the date of completion in block 4, the action taken code in block 11, the additional man-hours

in block 13, and appropriate remarks in block F. The form is then signed by the man doing the work and by his supervisor, and is then submitted to the data processing center.

The information contained in blocks A, 1 and 2 of OPNAV Forms 4700-2B, 4700-2C, and 4700-2D will be identical on all forms for any particular ship; if desired, this information may be preprinted in by the ship. Each maintenance action is assigned its own unique maintenance control number; if a maintenance task requires more than one form, as in the example using figures 2-8 through 2-10, the numbers will be identical.

Personnel attached to repair activities are required to record certain information on a Work Supplement Card (not shown) for each repair job. Work supplement cards will be explained to you by your division petty officer if you are assigned to a repair activity.

Material Usage and Cost Data

The documentation of material usage and cost data on maintenance transactions requires the joint effort of the supply and maintenance personnel on board your ship. Economy of effort and the elimination of duplicate recordings are highly desirable.

Any time that repair parts or materials are drawn from the supply department for shipboard maintenance, a NAVSANDA Form 1250 is used to request the materials from supply. Maintenance personnel are required to furnish the work center code and the EIC taken from the EIC Manual, the component identification (CID) number, the maintenance control number, the name of the part, quantity required, and the stock number when submitting this form to the supply department. The supply personnel will then complete the form; the cost is documented from this form if the ship does not have data processing equipment aboard. If the ship has data processing equipment, a DD Form 1348 is used instead of the Form 1250.

When the material to support a maintenance action is obtained from outside the normal supply channels or from pre-expended material bins, the reverse side of the appropriate OPNAV form (OPNAV 4700-2B, 4700-2C, or 4700-2D) is used. The reverse sides of OPNAV 4700 series forms are essentially the same. The reverse side of OPNAV Form 4700-2B is shown in figure 2-11. On this side of the form, when appropriate, you will enter the CID number, the

Figure 2-11.—Reverse side of OPNAV 4700-2B.

source code taken from section VIII of the EIC Manual, the cognizance symbol taken from supply publications to identify the supply account and the inventory manager of the item used, the Federal Stock Number, the reference symbol taken from circuit diagrams or the name of the part being replaced, the unit of issue of the material used, quantity used, and the unit cost (except on pre-extended material) of the item.

THE MAN-HOUR ACCOUNTING SYSTEM

The Man-hour Accounting System, sometimes referred to as Exception Time Accounting (ETA), is designed and intended for use by repair activities in conjunction with the Maintenance Data Collection System. It is basically a management tool and accounts for deviation from a normal working day.

As a third and second class petty officer assigned to a repair activity, your only concern with this system will be when you are absent from your assigned working space for more than 20 minutes at a time. If this occurs, your leading petty officer will hand you a Daily Exception Card on which you will be required to fill in the date of change and the number of hours concerned to the nearest tenth of an hour, and then return the card to your supervisor.

In the example card shown (fig. 2-12) Jones, K. P., ET3 (004 code) assigned to the electronics repair shop (67A code) for direct labor (01 code) was absent for one and a half hours. This absence was to take care of personal affairs and is therefore a non-duty absence as shown by the reverse side of figure 2-12.

MAINTENANCE HISTORY REPORT

The Maintenance History Report (MDC-5) will serve as a record of equipment history. Figure 2-13 shows an example of the Maintenance History Report (MDC-5) which will be received by each ship on a monthly basis.

FRONT SIDE

LABOR SUB-CODES

PRODUCTIVE SUPPORT

10.1 Work Center Supervision
10.2 Workload Planning/Control
10.3 Clerical
10.4 Drafting
10.5 Analysis
10.6 Maintenance Technical Training
12.1 Ship Equipment Maintenance
12.2 Repair/Weapon's Dept. Equip Maint
12.3 Repair/Weapon's Equip. Cleaning & Preservation

DELAYS

20.1 Awaiting Work
20.2 Awaiting Parts/Material
20.3 Awaiting Transportation
20.4 Awaiting Assistance
20.5 Inclement Weather

DUTY ABSENCE

21.1 Quarters for Muster/Special Sea Detail

21.2 Department Watch
21.3 Ship/Base Watch
21.4 Condition Watch
21.5 Military Training
21.6 TAD
21.7 Mess Cooks/Compt. Cleaners
21.8 Vehicle/Boat Operations
21.9 Personnel/Zone Inspection

NON-DUTY ABSENCE

22.1 Medical Absence
22.2 Personal Affairs
22.3 Leave
22.4 Special Liberty
22.5 Unauthorized Absence
22.6 Confinement
22.7 Non-Judicial Punishment

REVERSE SIDE

17.81E

Figure 2-12.—Daily Exception Card, OPNAV 4700-2E (1-65).

The report provides a comprehensive maintenance manhour and material history of all actions (different MCNs) received from a reporting activity monthly. It embodies a three line printout with the first line basically identifying the maintenance action and manhours charged, the second line, equipment time and intermediate repair activities, and the third, line material usage. The report is organized at the EIC equipment level and is further summarized at the subsystem and system levels. Ship totals for actions, active maintenance time, and manhours are also included. The report is a basic tool for the analysis of recurring

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COMCRUDESLANT

USS DAVIS

NAVY MAINTENANCE AND MATERIAL MANAGEMENT INFORMATION SYSTEM
MAINTENANCE SUPPORT OFFICE, MECHANICSBURG, PA.

PAGE DM03-0001

MONTHLY MDC-S   MAINTENANCE HISTORY REPORT FOR ACTIONS RECEIVED IN JUL 1967   05 AUG 1967

D-RADAR IFF, ELECTRONIC NAVIGATION SYSTEMS

| MCN | SERIAL/NOUN | MALFUNCTION | WHEN DISCOVERED | ACTION TAKEN | W/C | MAN-HRS | DATE |
| | REF SYMBOL | | EQUIP TIME ACTIVE MAINT TIME ALTERATION ID | CID/APL/AEL/AN | REPAIR ACTY | QTY UI | |

EIC DM00000 RADAR DISPLAY, STANDARD
EIC DM03000 AN/SPA-4A

4751 * * * NO MAINTENANCE EVENT CARD RECEIVED * * *

FUSES   EA   06057

5741 AN/SPA4A#1718   VOLTAGE, INCORRECT   DURING EQUIPMENT OPERATION   OER   10   1.0   19067

58299615   DURING EQUIPMENT OPERATION

9N   5920 280 4431   REPAIR, USE OF SPARE PARTS   OER   8.0   08067
                          NO REPAIR CARDS RECEIVED

                          REPAIR, USE OF SPARE PARTS   OER   MAN-HRS   9.0
                          NO REPAIR CARDS RECEIVED

EIC DM03050 POWER SUPPLY, HIGH VOLTAGE
5441 # 1729   OUTPUT, INCORRECT   MAN-HRS   9.0

EIC EQUIPMENT TOTALS ACTIONS 3
EIC SUB-SYS.   TOTALS ACTIONS 3

DA/MO/YR ACTION WAS TAKEN OR PART WAS ISSUED

TOTAL MAN-HOURS CHARGED TO A MCN

QUANTITY AND UNIT OF ISSUE OF REPAIR PARTS USED

WORK CENTER ACCOMPLISHING THE MAINTENANCE ACTION

ACTION TAKEN BY REPORTING ACTIVITY

REPAIR PART IDENTIFICATION BY FSN (INCLUDING COG), PART NUMBER, OR NOUN NOMENCLATURE

WHEN THE PROBLEM WAS DISCOVERED

COMPONENT IDENTIFICATION/ ALLOWANCE PARTS LIST/ALLOWANCE EQUIPAGE LIST/ ARMY-NAVY NUMBER

DESCRIBES THE TROUBLE ON OR IN THE EQUIPMENT

EQUIPMENT IDENTIFICATION CODE IDENTIFIES TYPE OF EQUIPMENT REPORTED

SERIAL NUMBER IDENTIFIES EACH ITEM OF EQUIPMENT

MAINTENANCE CONTROL NUMBER SEQUENCED FOR EACH ACTION

162.1

quipment and structural maintenance problems n terms of both manhour and material usage, n addition to furnishing a monthly recapitulaion of equipment history.

When fully implemented the PMS and MDCS systems are intended to supercede any other records or reports previously required under other maintenance systems. In the meantime, ships and stations will find they have to operate under several systems simultaneously. It is for this reason that the other maintenance systems, such as POMSEE and CSMP, are explained in this chapter.

A good explanation of which system to use, if you are in the middle of a changeover, can be found in EIB 707 of 27 May 1967, and by reference to type commanders' instructions. It is anticipated that some of the forms previously described will also be changed before the system is fully implemented throughout the Navy, so it is a good idea to always check the latest EIB for up to date information regarding maintenance procedures.

MAINTENANCE STANDARDS BOOKS

Tests and maintenance procedures prescribed in the Maintenance Standards Books provide maintenance and operation personnel with systematic and efficient methods for checking equipments and for performing scheduled preventive maintenance. These books contain a series of reference standard tests (fig. 2-14). These tests, when performed on properly operating equipment, provide quantitative results (reference standards), which collectively represent the design performance of the specific serial numbered equipment or system for which the book is assigned. When these tests are accomplished as scheduled (weekly, monthly, quarterly, etc.), the test results are compared against the reference standards to determine the condition of the equipment. Charts (time schedules) are provided for logging the results of each test so that a performance history of the equipment can be developed (fig. 2-15). In addition to these functional tests, the book also contains the specific procedures and schedule necessary to provide for mechanical and electrical maintenance for the equipment or system (fig. 2-16).

The maintenance checkoff is gradually being superseded by the Planned Maintenance System. If the Planned Maintenance System (PMS) portion of the 3-M system has been implemented,

recording under the POMSEE program is no longer a NavShips requirement. This does not mean that the Maintenance Standards Books can be discarded: Maintenance Standards Books must be retained to make up a complete series of records, even though no new entries are to be made. Directives as to exactly which records are required are promulgated by type commanders, or similar level for shore activities.

PERFORMANCE STANDARDS SHEET

The Performance Standards Sheet (fig. 2-17) establishes the standards for the optimum performance of the equipment which must be met upon installation. The sheet contains a table of standards (table I), for operational performance and a list of standards for equipment measurements (table II). The information on the sheets may be used as the basis for acceptance tests and electronics material readiness inspections.

The standards for operational performance is a tabulated nontechnical description of satisfactory equipment performance which could be tables of expected radar ranges of detection, maximum radio ranges, and voltage ranges. The operational performance values are obtained from design data and by actual field measurements under normal conditions while the equipment is operating satisfactorily in the location of its intended use.

Special information relative to such operational performance as range and bearing accuracy and discernible characteristics of signals may be given on the performance standards sheet in another table. This special information is obtained from actual measurements as well as from design data.

The list of standards for equipment measurements, Table II, provides a record of measurements (standards) that indicates satisfactory overall equipment performance. The test results entered in the Maintenance Standards Book may be compared at any time with the data given in the Performance Standards Sheet to determine if the equipment or system meets the standards as specified.

TECHNICAL MAINTENANCE

Technical maintenance is normally limited to maintenance consisting of replacing unserviceable subassemblies or parts and the alignment, testing, and internal adjustment of equipment.

QUARTERLY STEPS   NAVSHIPS 94715.42   AN/WRR-2, -2A

Q1 thru Q8   AN/FRR-59, -59A

OPERATING CONDITIONS AND CONTROL SETTINGS:

Equipment in full operation and conditioned for A1 reception.
O. L. THRES: OFF

| STEP NO.	ACTION REQUIRED	READ INDICATION ON	REFERENCE STANDARD
Q1	Record over-all sensitivity of Mode A1 at low end of 2-4 mc band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	*PROCEDURE: Connect signal generator to the ANT IN jack (J957). Adjust generator for a 2-mc unmodulated signal and set output at 5 μv. Set BAND selector to 2-4. Tune receiver to 2 mc and adjust the ANT COMP and HF ADJ controls for maximum indication on the RESONANCE meter. Reduce generator output to zero and adjust RF GAIN for a -2 db indication on the LINE A OUTPUT meter. Set generator output to 5 μv and adjust generator frequency for a maximum indication on the RESONANCE meter. Readjust generator output for a +18 db indication on the LINE A OUTPUT meter. Record the generator output in microvolts.		
Q2	Repeat for high end of band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Tune receiver to 4 mc and repeat step Q1.		
Q3	Record over-all sensitivity of Mode A1 at low end of 4-8 mc band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Set BAND selector to 4-8. Tune receiver to 4 mc and repeat step Q1.		
Q4	Repeat for high end of band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Tune receiver to 8 mc and repeat step Q1.		
Q5	Record over-all sensitivity of Mode A1 at low end of 8-16 mc band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Set BAND selector to 8-16. Tune receiver to 8 mc and repeat step Q1.		
Q6	Repeat for high end of band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Tune receiver to 16 mc and repeat step Q1.		
Q7	Record over-all sensitivity of Mode A1 at low end of 16-32 mc band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Set BAND selector to 16-32. Tune receiver to 16 mc and repeat step Q1.		
Q8	Repeat for high end of band.	Signal Generator AN/URM-25	μv $(\overline{1.5 \text{ max.}})$
	PROCEDURE: Tune receiver to 32 mc and repeat step Q1.		

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Figure 2-14.—Sample page from Maintenance Standards Book, reference standards tests.

TIME SCHEDULE
QUARTERLY STEPS

Q1 thru Q8

Q19 and Q20

NAVSHIPS 94715.42

AN/WRR-2, -2A
AN/FRR-59, -59A
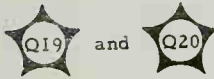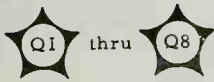
TIME SCHEDULE: Record and initial.

1ST YEAR OF OPERATION

1. Date			2. Date			3. Date			4. Date		
No.	μv	Init.	No.	μv	Init.	No.	μv	Init.	No.	μv	Init.
Q1											
Q2											
Q3											
Q4											
Q5											
Q6											
Q7											
Q8											

1ST YEAR OF OPERATION

1. Date			2. Date			3. Date			4. Date		
No.	kc	Init.	No.	kc	Init.	No.	kc	Init.	No.	kc	Init.
Q19											
Q20											

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Figure 2-15.—Sample page from Maintenance Standards Book, time schedule.

| AN/WRR-2, -2A | NAVSHIPS 94715.42 | UNSCHEDULED |
| AN/FRR-59, -59A | | (U8) thru (U10) |

OPERATING CONDITIONS AND CONTROL SETTINGS:

Equipment de-energized.
AC power plug (P1808) removed from POWER IN jack (J1808).
Drawers extended (in turn) from cabinet.
Decks tilted or separated to gain access.

STEP NO.	ACTION REQUIRED
(U8)	Clean equipment
	PROCEDURE: Clean cabinets and chassis with vacuum cleaner. Check air filter; clean and recharge if necessary (see Quarterly Step Q29).
(U9)	Perform mechanical inspection.
	PROCEDURE: Note action of drawer mechanisms. Operate all controls, observing mechanical action. If control stick or binds DO NOT FORCE MOVEMENT. Find the cause (bent shaft, loose mounting, etc.) and correct it. Observe action of counter mechanisms and all gearing. Note condition of lubricant and replace if necessary. Turn blower fan by hand, observing freedom of rotation. Inspect switches for arcing damage. Burnish contacts by rotating switch, if necessary. Lock all tube shields and/or tube clamps.
(U10)	Perform electrical inspection.
	PROCEDURE: Inspect all electrical components, wiring and cabling. Look and feel for loose connections, kinks, strain, and damaged insulation; correct defects or report them to a qualified technician. Look for evidence of overheating; take corrective action if any is found. Replace bulging or leaking capacitors, after first removing any residue.

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Figure 2-16.—Sample page from Maintenance Standards Book, preventive maintenance steps.

CORRECTIVE MAINTENANCE PROCEDURES

To perform effective corrective maintenance the ET must have a good working knowledge of the basic principles of electricity and electronics. The only way to acquire this knowledge is by diligent study. The ET must also be thoroughly familiar with the theory of operation of the equipments that he must service. A knowledge of the theory of operation can be acquired through a study of the equipment instruction book. This knowledge should be broadened to include other equipments at the earliest opportunity. As a matter of fact, ETs are generally rotated on the various electronic equipments so that their knowledge will be broadened and they will therefore be more valuable to the fleet.

Skill in the use of test equipment (and hand-tools) is also necessary for effective corrective maintenance. Skill in the use of test equipment comes with practice and with careful study of the instruction book that comes with each piece of test equipment. The ET should take advantage of every opportunity to learn more about every type of electronic test equipment used aboard ship. The ability to use test equipment effectively is an absolute must for every well-trained ET.

Chapter 3 of this training course introduces some of the special type test equipments. Additional valuable information (both on test equipment and methods of troubleshooting) is contained in Test Methods and Practices, NavShips 900,000.103.

Tests (opens, grounds, and shorts) and measurements (currents, voltage, resistance, frequency, power, etc.) enable the technician to

6 April 1964   NAVSHIPS 94715.32   AN/WRR-2, -2A
                                                 AN/FRR-59, -59A

PERFORMANCE STANDARD SHEET

for

RADIO RECEIVING SETS AN/WRR-2, -2A, AN/FRR-59, -59A

TABLE I - OPERATIONAL PERFORMANCE

The many variables involved, such as radio propagation conditions at the time of the test, power output of the transmitter being received and the type of antenna installation, preclude definitive predictions of operation ranges. To achieve optimum reliable ranges, frequency selection should be made in accordance with DNC-14, as corrected by half-hourly radio propagation predictions given over WWV and WWVH.

TABLE II - STANDARDS FOR EQUIPMENT MEASUREMENT

MEASUREMENT	STEP	STANDARD
Bandwidth (IF, 1.0 kc)	Q19 and Q20	0.8 kc min. (at 6 db) 2.4 kc min. (at 60 db)
Receiver Sensitivity (Sr)	Q1 thru Q8	1.5 μv max.
Performance Figure (PF)*	Calculate	1.5 max.

* PF = Sr (μv)

Total time required to perform Table II measurements - 1 hour.

All steps refer to tests in the Maintenance Standards Book, NAVSHIPS 94715.42

This Performance Standards Sheet supersedes NAVSHIPS 93550.32

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Figure 2-17.—Performance standards sheet.

diagnose troubles so that repairs may be made. In many instances tests and measurements will indicate conditions that may be corrected before an actual breakdown occurs. Thus, tests and measurements (particularly measurements) are important both in preventive and corrective maintenance.

The purpose of any type of electronic test equipment is to measure accurately certain circuit values or to indicate certain circuit conditions. Each of these measurements or indications is used to determine the operating condition of electronic or electrical equipment. The accuracy with which measurements are

made depends on the type of instrument used, its sensitivity, its rated accuracy, its useful range, and the care that the technician uses in making the measurement.

The exact procedure for making tests and measurements is given in the technical manuals (instruction books) that accompany the various electronic equipments.

It should be emphasized that the mere taking of measurements means little unless they can be properly interpreted. For example, the presence of a voltage across a grid resistor in an audio amplifier would mean little to an ET unless he could interpret this in terms of a possible leaky coupling capacitor. In this connection, the necessity for a knowledge of basic circuit operation must be emphasized.

Before any attempt is made to interpret the results of measurements, an understanding of how the equipment operates should be acquired by a careful study of the applicable technical manual or instruction book.

The actual voltage, resistance, and current measurements that should be obtained are indicated in the circuit diagrams, charts, or in the maintenance standards books. The same is true of waveform measurements. Any deviation from the standard values (beyond the tolerance limits) means that some component is not doing the job that it should. By applying effect-to-cause reasoning, the defective component may be located. This is, of course, a job for a skilled technician.

Maintenance personnel must try to find the source of the trouble causing the equipment failure, particularly when the trouble is a recurrent one. The recurrence of a fault usually indicates that the EFFECT, not the CAUSE, has been remedied.

There are numerous ways to isolate a fault to a component of a system, depending on the type of equipment. The technician must, first of all, know what each component does before he can know that it is not functioning properly.

Assume, for example, that the display on a certain radar repeater is faulty. The trouble could be in the repeater or it could be in one of the circuits that feeds into the repeater. If the other repeaters are working properly, it is probable that the trouble is in the repeater having the faulty display or in its power supply or in the transmission lines or switching system.

The best way (the most economical in time and effort) to isolate a fault is by using a logical troubleshooting method.

Keep in mind throughout this study, that by far the largest section of the average instruction book is the one devoted to corrective maintenance (troubleshooting). This is the section written especially for the ET and when used with the six-step logical troubleshooting procedure (described later) can be for him the most valuable part of the book. The section on the theory of operation, however, is also very important and should be studied with care.

Good troubleshooting is not a talent with which a person is born. It is, however, a skill that can be acquired by anyone with a suitable electronics background. One can become a good troubleshooter if he has:

1. Sufficient electronic knowledge to learn, or to be taught how an equipment works.
2. Suitable skill in reading and interpreting data contained in the equipment's technical manual.
3. Suitable skill in operating test equipment and interpreting test readings.
4. Learned how to troubleshoot in a logical manner.

Logical troubleshooting is a time-proven procedure used by all accomplished technicians. Most of them have applied the procedure so often that they no longer pay attention to its fine points. Through habit and years of experience they may have forgotten its specific details, but the procedures are there and have remained the same.

Probably no two technicians would explain the procedure alike, but all would agree that logical troubleshooting consists of a series of sequential steps based on valid electronic decisions that systematically narrows the trouble down to the faulty part. Some would list the procedure in three or four steps, others would count a dozen, fifteen, or more. Regardless of the number, the principle would be the same. Six steps have been chosen as the easiest method of learning and applying this procedure. The steps in their sequential order are:

1. Symptom recognition
2. Symptom elaboration
3. Listing the probable faulty functions
4. Localizing the faulty function
5. Localizing the faulty circuit
6. Failure analysis

SYMPTOM RECOGNITION

The first step in any troubleshooting problem is recognition of a trouble indication. Recognizing a trouble condition in an equipment is not always easy to do since all conditions of less than peak performance are not always apparent. Lack of targets on a scope, timing error in a loran set, and a decrease in signal-to-noise ratio in a receiver are just a few of the hundreds of examples occurring throughout the Navy. Each of these is a trouble symptom that requires recognition and elimination.

There are many ways in which the existence of a trouble can be detected by a technician. The obvious troubles will undoubtedly be reported by the operator. These usually include complete or almost complete malfunction of the equipment. Troubles that are not easily noticed are those that cause a gradual decrease in equipment performance. For example, a 125-mile radar that is reaching only 105 miles, a 100-watt transmitter that is putting out only 87 watts, or a multimeter that provides readings that are 10% off, are equipment faults that are difficult to recognize because there are no visible or audible indications (built-in) that say they exist.

Since a ship must depend on full-performance equipment, the hidden trouble symptoms must be found, the cause of the trouble located, and the repair made. If the technician makes a point of looking for them every time he touches an equipment most of the decreasing performance symptoms can be easily recognized. Often he can compare the performance between two similar equipments. He can actually make the Performance Standards sheet checks contained in the POMSEE program. By using the POMSEE books, he can verify any change in performance since the last time he tuned, calibrated, or aligned the equipment. Also while troubleshooting, he may look for and probably find symptoms that signify decreasing performance and future breakdown if left unnoticed. Trouble symptoms can be recognized if the technician will only look for them.

SYMPTOM ELABORATION

Breaking out test equipment and equipment prints and proceeding headlong into troubleshooting on just the original identity of a trouble symptom is a very shaky premise. It could also be an unnecessary expenditure of energy. A dead scope, a hum in a receiver, a zero reading on a panel meter, or a missing transmitter pulse, by itself, is not sufficient identification of a trouble symptom. There is a tendency among less efficient technicians to attempt a solution of a troubleshooting problem before they have completely defined it.

The procedures involved are dependent upon the available aids designed in the equipment and the nature of the original symptom. The aids include front panel controls and built-in performance measuring indicators. Additional information can be obtained about any malfunction as the result of a systematic front panel check. If the technician has a fair knowledge of how the equipment works, manipulation of appropriate controls and switches and corresponding checks of equipment meters and scopes will reveal to the technician how the trouble is affecting the entire equipment. From these clues he is able to narrow down the probable areas of the equipment that could contain the trouble.

LISTING THE PROBABLE
(FAULTY FUNCTIONS)

The third step requires that the troubleshooter make an educated guess as to the probable cause of the trouble. From the elements of the trouble symptom, as he has identified it, he determines its most logical locations. Locations are to be confined to the major subdivisions (major or functional units) of an equipment. Educated guesses are made from the knowledge of how the equipment works and a study of the equipment's functional block diagram.

For example, using a malfunctioning radar that has no targets on the PPI but the transmitter and modulator indications read good, the educated guesses could include: (a) remote indicator unit, (b) receiver unit, (c) low-voltage power supply, and (d) duplexer. Making an educated guess that a tube is bad (just because the greater percentage of all equipment troubles are caused by bad tubes) is not acceptable. The purpose here is to use valid reasoning to isolate all probable, technically sensible functional areas which may contain the trouble. It may well be that the specific trouble is a bad tube, but wholesale tube substitution takes a lot of time and quite often introduces additional troubles, particularly in those circuits that operate close to critical tolerances.

Even the accomplished technician may not be able to list all the functional units that

are probable sources of the trouble. However, with the exception of a very obvious trouble, a multi-unit equipment will have many functional units that are probable sources of a trouble.

LOCALIZING THE FAULTY FUNCTION

In this step one of the educated guesses must be selected for testing. It is not necessarily the one that was thought of first nor the one that past experience suggests as being the most attractive. The selection of the functional unit to be tested (or verified) first should be based not only on priority of validity but also on the difficulty in making the necessary tests. Under some circumstances, a troubleshooter might elect to test the second best educated guess rather than the first because the latter might involve testing difficulties that should be initially avoided or require tampering with circuit parts that might later prove to be unnecessary. Like all the others, this step in the troubleshooting procedure places emphasis on common sense thinking rather than the resultant action.

After selecting the order in which the listed units will be checked, the troubleshooter proceeds to verify his first selection. This check normally is made at the output test point of the suspected unit. The test equipment reading is compared with the desired signal contained in the technical manual. No output is relatively easy to recognize. A distorted or nonstandard output, however, should be carefully verified before arriving at a technical conclusion.

If the technician does his mental work properly, manual work in gaining access to test points and using the test equipment can be limited to a bare minimum. This procedure is opposed to trial-and-error methods where the technician searches from point to point with test prods, hoping to locate the faulty test reading that identifies the trouble. Not only does the illogical technician waste valuable time, but his unwillingness to rely on his technical knowledge indicates that he will be very lucky if he finds the trouble.

Upon completing the verification of the probable faulty unit selected, the technician will have arrived at one of several conclusions. The test may verify that this is the unit in which the trouble lies; or that the trouble could be in this unit plus another unit(s) from which it

receives signal or control voltages; or that the trouble is not in this unit at all; or that the output looks suspicious and further verifying tests need to be made.

Whatever the conclusion, the technician has discovered information that can be used to substantiate or eliminate suspected units or provide evidence for adding another. Tests of suspected unit outputs are continued until the single faulty unit is identified. At that point, the technician has narrowed down the trouble to a fraction of the total number of circuits and parts in the equipment. If at this time the proper procedure was carried out the search can be confined to the functional area isolated. NOTE: There are some equipments such as communications receivers, that cannot be easily divided into functional units. If this type of equipment is involved, steps 3 and 4 above, can be eliminated from the overall troubleshooting procedure.

LOCALIZING THE FAULTY CIRCUIT

After the faulty unit has been isolated, the next step is to identify the faulty circuit. The same narrowing-down procedures are used here as before. The unit is mentally subdivided into circuit groups by function, and valid technical reasoning is employed to select those that might probably contain the trouble. Using this procedure the technician can find the faulty circuit without going through the unnecessary time-wasting chore of test-point to test-point checking from one end of the unit to the other.

The technician works from the servicing block diagram of the unit. He then applies the information obtained from the preceding steps regarding the nature of the trouble. In narrowing down the trouble to a single functional group of circuits, the process used is called "bracketing." In this process brackets are placed, mentally or in pencil, around the area in which the trouble lies. Initially, a bracket is placed at the input(s) to the units that are known to be good and at the output(s) known to be bad. As each deduction is made and verified by a test, the input or output bracket is moved to the next point in the block diagram where the test was made. In this manner the closing brackets systematically narrow the fault to a single circuit.

In selecting a point on the servicing block diagram to which one of the brackets is to be

moved, the technician must consider two things: (1) the faulty characteristics of the improper output signal and (2) the types of signal paths contained in the unit. The waveshape of a signal contains characteristics—voltage, time, bandpass, noise content, frequency, etc.—that can be measured or observed. When these characteristics are in accordance with the designed standards, the signal is considered to be good. Bad signal characteristics that are improper can reveal clues that will help to identify a circuit group whose function is to originate or control that portion of the waveshape. For example, the output of a unit is supposed to be a sawtooth waveform with six pulses, equally spaced on its slope. If the pulses are there but the slope is insufficient or improper, the sawtooth generating and shaping circuits would be suspected. If the proper slope is there but there is insufficient number of pulses, the pulse generating or controlling circuit groups may contain the trouble.

Types of signal paths contained in the unit are the other items to be considered before moving a bracket. There are four general types: linear, switching, convergent/divergent, and feedback. In a linear signal path, the signal is processed through circuits that are connected in series. When identification of the faulty circuit group is difficult or impossible, brackets can be moved to successively smaller half-points in the linear string. Signals from two or more circuit channels that meet at a common point or a signal that leaves a common point into two or more channels are examples of convergent/divergent paths. Moving a bracket to the common point (after making the appropriate test) will separate the bad from the good signal paths. In the same manner, a test and bracket at the point where signal paths are connected by a switch will reveal the same information. The remaining type (feedback loops) provides a means of bracketing a group of circuits in the narrowing-down process.

There are no hard-and-fast step-by-step procedures in bracketing. But there are some realistic general rules.

1. Examine the characteristics of the faulty output to determine the circuit group function that either generates or controls the improper characteristic.

2. Study the servicing block diagram to determine the least number of bracket moves that will isolate the faulty circuit. Such moves will be dependent upon the types of signal paths

contained in the unit and the electronic functions of circuit groups that may be responsible for distortions contained in the unit's output.

3. Move only one bracket at a time after verifying the suitability of the signal by making a test.

4. When the test does not reveal sufficient information for a valid bracket move, make another educated guess.

5. The determination of which bracket to move is dependent upon circuit configuration within the unit and the smaller number of circuits that will be enclosed.

The servicing block diagram can serve as the instrument for the completed bracketing process. In some cases it may be necessary to refer to a schematic diagram for bracketing or testing information. In any event there is sufficient diagram information available in the technical manual to support the bracketing procedure and preclude the wastefulness of unreliable circuit to circuit checking. This step is completed when the technician has isolated the trouble to a single circuit and verified that the output of this circuit is the cause of the distortion read at the output of the unit.

FAILURE ANALYSIS

The troubleshooting procedure thus far has narrowed the trouble to a single circuit consisting of a tube (or transistor) and a few electronic components. If there is no output from the circuit, it may be permissible to resort to testing of tube pin numbers. However, such checks can be minimized if there is an output that can be examined for distortions that will reveal the circuit part that is most likely at fault. Quite often the waveform will identify the malfunction to be in the grid, cathode, plate, or screen portion of the circuit. Such a study should be made before any of the parts are checked.

When the faulty part has been identified, it should not be replaced until the technician can SUBSTANTIATE THAT IT IS CAUSING THE ACTUAL TROUBLE. A suspected open resistor, shorted capacitor, detuned coil, or weak tube may not be the reason or the only reason causing the faulty output of the circuit. If the technician replaces the part without an adequate technical reason, he may (when replacing the part) not have cured the trouble and he may yet cause further trouble. ANALYZE THE FAILURE BEFORE MAKING THE REPAIR.

It can now be seen that the six-step troubleshooting procedure is designed to isolate a

trouble in an orderly manner. Success in using the procedure is dependent upon the technician's knowledge of electronics, the equipment under test, and his skill in using the technical manual and test equipment. The process is no more complicated than the ability to subdivide an equipment into progressively smaller functional areas, such as functional units into functional circuit groups, to a circuit, and finally a part or an adjustment within the circuit. It is the only logical way to troubleshoot any equipment, and it is more reliable and faster than any other method.

RECORDS AND REPORTS

Each activity engaged in the operation of electronic equipment is required to maintain cards for recording the results of inspections of equipments, and records of any tests, repairs, and field changes made. These cards provide a comprehensive record of the items concerned. They are kept up to date and available for inspection at all times and are in-tegrated into preventive maintenance programs such as the Current Ships Maintenance Project (CSMP).

All records and reports discussed in the following paragraphs are required except where the 3-M System has been implemented AND where the type commander has indicated that they have been superseded by 3-M data products or procedures.

ELECTRONIC EQUIPMENT MATERIAL HISTORY

The material history of electronic equipment consists of the Electronic Equipment History (Unit Location Card), NAVSHIPS 536-1 (8-66), (fig. 2-18) and the Electronic Equipment History Record Card, NAVSHIPS 536 (Rev 8-66) (fig. 2-19). One set of electronic equipment history cards is to be prepared and maintained for each electronic equipment (or set) on board. The two cards (NAVSHIPS 536 and 536-1) are companion items. The initial or "top" card is the Electronic Equipment History

Figure 2-18.—Electronic equipment history unit location card.

1.2

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Figure 2-19.—Electronic equipment history record card.

1.2

(Unit Location Card), NAVSHIPS 536-1, which is to be used as an inventory list. It provides space for tabulating all units of an equipment, together with serial numbers and installation locations. The Electronic Equipment History Record Card, NAVSHIPS 536, is used to record all failures and other work accomplished on an equipment. The NAVSHIPS 536 card has been arranged for chronological listing of all entries, regardless of the unit in which work was done. However, separate cards may be maintained for each unit, as deemed necessary by the individual activities.

All cards for a complete equipment are to accompany that equipment when it is removed from the ship or activity. In cases where an individual unit of an equipment is removed or substituted, a separate history record card, containing information pertinent to that unit, is to accompany the unit upon removal. If a chronological listing is used, it will be necessary to transcribe applicable data to a separate card.

NOTE: The two history record cards described and illustrated in figures 2-18 and 2-19 supersede the Electronic Equipment History Record Card, NAVSHIPS 536 (Rev. 9-48), and are to be used on all new equipment installations. It is not mandatory that existing history records be revised to incorporate these new cards. Such revision may be accomplished at the option of the individual Electronics Material Officer.

The following instructions are to be followed in filling out the forms:

1. NAVSHIPS 536-1; This form is to be completed as soon after equipment installation as possible. The heading of the card should be typed, but entries on the body of the card may be typed or written in ink or indelible pencil as shown in figure 2-19. Individual column entries should be made in accordance with the following guidelines.

Equipment Model Designation: All letters and numbers should be included to indicate the specific equipment. For instance, AN/SPS-10D should not be entered SPS-10.

Equipment Serial Number: This is the main equipment serial number and can be found on the nameplate bearing the overall equipment model designation. (In the case of the AN/SPS-10D, it is located on the receiver-transmitter unit.) Do not use the unit serial. Where it is definitely established that an equipment does not bear a serial number, an asterisk (*) is to be entered in this space.

Technical Manual (NavShips) Number: Enter the applicable NAVSHIPS number in this space. For commercial manuals, enter the abbreviation "Coml" and any other identifying numbers or data.

T. M. On Board: This space is to be checked when the final technical manual is received. If only the manuscript form of the manual is on board, the block should be left blank and a pencil notation inserted to indicate the status of the edition that has been received (whether manuscript is preliminary or advance). The pencil notation is then erased when the final edition is received.

Card No.: This space is to indicate the card serial. The numbering should follow, in consecutive order, that of the Unit Location Card(s). If there is only one Unit Location Card, the number to be entered on the first History Record Card is "2"; all numbering of additional cards filed for this equipment is to follow in sequence. If individual cards have been prepared, letter suffixes may be used to identify additional cards for each unit, e.g., 2A, b, etc.

Name of Contractor: Enter the name of the contractor, in full, as given on the overall equipment nameplate or the technical manual.

Contract No.: This is to include all letters and numbers given on the equipment nameplate.

Date Installed: This refers to the date the equipment was installed. If the installation was accomplished over an extended period of time, the completion date is to be entered.

Installing Activity: This is the name of the activity or organization which actually installed the equipment.

Unit Type No.: This column is to list the official type (A-N) designations of all units that comprise the complete equipment. If commercial equipment, the manufacturer's designation is to be entered. All letters and numbers are to be included.

Name of Unit: These are the official names of the units that make up the equipment.

Unit Serial No.: This column is to list the serial numbers of the units.

Location: Enter the unit locations in this column, using the name and compartment number. If the unit is portable, indicate the space where it is normally stored.

2. NAVSHIPS 536: Since this form is used to list work performed on the equipment or failures that occur, it is not necessary to prepare the card until the first problem is corrected. As shown in figure 2-18, the heading of the card should be typewritten, but entries on the body of the card may be typed or written in ink or indelible pencil. Individual column entries should be made in accordance with the following guidelines:

Equipment Model Designation, Name of Equipment, and Equipment Serial No.: These entries should be identical to those made in the heading of the companion Unit Location Card.

Card No.: This space is to indicate the card serial. The numbering is to follow, in consecutive order, that of the Unit Location Card(s). If there is only one Unit Location Card, the number to be entered on the first History Record Card is "2"; all numbering of additional cards filed for this equipment is to follow in numerical sequence.

Failure Date: This is the date of a failure, the correction of a failure, or other work involving maintenance or repair.

Symptoms: This column is to be used to describe the symptoms exhibited by the equipment to indicate failure or faulty operation. The entry should be brief, but should describe the symptom in such a manner that it can be readily recognized by others.

Corrective Action Taken: This and the preceding column are most important. In this column enter the corrective measures taken to eliminate the fault described in the symptoms column. The entry can be brief, but should be as clear as possible. The clearer the information in this column, the more valuable it will be to the ship, the Navy Department, and the manufacturer. The cause of the problem is to be listed together with the parts replaced or adjustments made to correct the deficiency. Some activities may also wish to record in this column such information as the name and rate of the person who actually accomplished the work, the man-hours consumed, and the signature of the division officer. Such entries are optional.

Unit Type No.: This is to list the unit or units in which the corrective action was taken.

As shown in figure 2-19, the type number may be abbreviated.

Unit Serial No.: This is to include the serial numbers to the units involved.

Although time consuming, it is still possible to keep a complete and accurate material history by having each man who performs a repair make the entry in the material history as soon as he finishes the job. To ensure that proper entries are being made, the Electronic Material Officer or leading ET should check them once a week for completeness and correctness. Spot check should indicate whether or not entries are being omitted. Large ships often require repairs to be logged in a work book for checking by a supervisor, who then enters the appropriate information in the material history. This procedure is successful where there is sufficient personnel; however, workbooks are not required by regulations. It is suggested that ETs on small ships log the repairs directly into the material history. In many cases it has been found that the ETs fail to log repairs made by personnel other than ship's force, that is, those accomplished by tenders, shipyards, and civilian technicians. These, too, must be recorded in the material history.

CURRENT SHIP'S MAINTENANCE PROJECT

The purpose of the current ship's maintenance project (CSMP) is to provide a current record of maintenance, modifications, and repairs to be scheduled and finally accomplished by ship's personnel or by repair activities afloat or ashore.

The CSMP consists of the following three cards:

NAVSHIPS 529—Repair Record Card (blue)
NAVSHIPS 530—Alteration Record Card (pink)
NAVSHIPS 537 - Record of Field Changes (white)

As a repair is required or an alteration is authorized, the work is scheduled by filling out an applicable CSMP card and placing it so the top line of the card is in view alongside the proper history card, NAVSHIPS 536 (Rev 8-66). These CSMP cards are of distinctive color, which facilitates the indication of outstanding work when the history is examined.

When the item of work has been completed, entries are made in two places. One entry of the completion is made on the CSMP card,

and the other entry is made on the history card, NAVSHIPS 536 (Rev 8-66).

Upon completion of work, the Record of Field Changes, NAVSHIPS 537, remains with its NAVSHIPS 536 (Rev 8-66). However, NAVSHIPS cards 529 and 530 are removed from their position adjacent to the history card and filed in a "completed work" section of the CSMP.

The Repair Record Card, NAVSHIPS 529 (blue), and the Alteration Record Card, NAVSHIPS 530 (pink) are identical as to block descriptions, therefore, the description of the entries apply to both cards. Their distinctive colors aid in distinguishing repair from alteration projects.

Repair Record Cards and Alteration Record Cards are retained for a period of 2 years, following which these cards may be destroyed at the discretion of the commanding officer. When ships are decommissioned or placed out of service during this period, the cards are retained onboard.

If the equipment is transferred, these cards are transferred with it.

Field changes scheduled and later made to any portion of an electronic equipment are recorded on the Record of Field Changes, NAVSHIPS 537, which is filed in the binder adjacent to the history card for the equipment to which the change is applicable. Completion of filed change information is also entered on NAVSHIPS 536 (Rev 8-66).

This record is of paramount importance. Without modification, an equipment may be dangerously out of date and subject to numerous serious difficulties. Without a record of field changes it is difficult to determine what modifications, if any, have been made. The information recorded on these cards is essential for routine maintenance, for troubleshooting, and for ordering maintenance parts for the improved equipment.

Figure 2-20 shows the Record of Field Changes Card, NAVSHIPS 537. The spaces for equipment model designation, serial number, date installed, and card number are filled in by typing or writing with ink or indelible pencil.

The columns headed "No.," "Title of Field Changes," and "Authority for Change" are completed in numerical order for all changes affecting a specific equipment. Field changes that affect certain equipments are made known in the Electronic Installation and Maintenance Book (EIMB). Enter applicable extracts from

AN/SRN-6A	63	3 March 1965				1
Equipment Model Designation	Serial Number	Date Installed		U. S. GOVERNMENT PRINTING OFFICE 16—52957-2		Card No.
NO.	TITLE OF FIELD CHANGE		AUTHORITY FOR CHANGE	CHANGE MADE BY	DATE OF CHANGE	
1	Reduction of radiated noise.		EIMB	BRADY, ET1	5/4/66	
2	Replaces Heater Stabilized Spectrum Filter with a Temperature Compensated Spectrum Filter					
			EIMB	NA		

Figure 2-20.—Record of Field Changes (NAVSHIPS 537).

1.4

EIMB on the NAVSHIPS 537 card to indicate exactly its EIMB source and language. The Electronic Information Bulletin (EIB) lists field changes and is also an authority for making such changes.

Do not use obsolete entries on NAVSHIPS 537. Approximately once a year, the Naval Ships System Command determines which EIBs become obsolete because their items are published elsewhere in permanent form (in EIMB and Instruction Book publications). For example, the initial 200 copies of EIB, were named Repair Information Bulletin (RIB) and were canceled (as summarized in EIB 476) along with EIBs from serial numbers 1 through 380. Until further notice, serial numbers above 600 shall be considered authoritative and directive in nature for announcing field changes that are active.

Many field changes are issued directly from the Naval Ships System Command to a ship. Whatever its source, list the authority for such changes in the column provided for it on NAVSHIPS 537.

EQUIPMENT FAILURE AND
PERFORMANCE RECORDS

The Naval Ships System Command must receive accurate reports from the fleet concerning equipment performance and failures, in order to evaluate its reliability and maintainability. Reports are also necessary to determine whether new equipments are meeting their design capabilities and operation requirements.

Reported failures are tabulated on IBM cards, and regular summaries are made to show at a glance the number and types of failures of any part of any equipment.

The importance of reporting failures and their causes cannot be too highly stressed, particularly if failures occur under actual operating conditions. The reports are to be filled in completely and in conformity with the instructions accompanying the card. Many reports received by the Naval Ships System

Command are valueless because they do not give the essential information required by the form or because the information given is incomplete. When indicating the model or type of equipment, include all significant nomenclature, letters, and digits.

Electronic Equipment Failure/
Replacement Report, DD-787

Failures to electronic equipment are reported according to NAVSHIPS Instruction 10550.73. The Electronics Failure Report form, DD-787 (fig. 2-21) and the Electronic Equipment Operational Time Log (discussed later) are submitted only upon failure of special equipments. The initial list of sepcified equipments is shown in the previously referenced NAVSHIPS instruction 10550.73. Additions and deletions will appear in the EIB, which must be checked in each biweekly issue because listed additions will change your reporting requirements.

The DD-787 failure/replacement form is arranged so that all failure information relating to one equipment can be entered on the form. This should simplify the maintenance paperwork which resulted from earlier (obsolete) requirements of listing only a single part on a single card. The DD-787 failure/replacement form, combined with the Operational Time Log, NAVSHIPS 4855, provides the necessary data for accurately showing: (1) mean-time-between failures, (2) mean-time-to-repair, (3) down-time, (4) availability, (5) failure rates, and (6) replacement (consumption) rates.

The report forms are packaged between manila tag covers and flaps, with 50 sets per package. The covers and flaps have printed codes and instructions for accurately completing the entries on the DD-787 form.

Electronic Equipment Operational
Time Log, NAVSHIPS 4855

The Electronic Equipment Operational Time Log (fig. 2-22) serves a twofold purpose. First,

Figure 2-21.—Electronic Equipment Failure/Replacement Report, DD-787.

15.2

15.3

Figure 2-22.—Electronic Equipment
Operational Time Log, NAVSHIPS 4855.

it is used to show accurate time-base data
for figures-of-merit, which concern all failure-
rate or replacement-rate calculations, or which
concern other calculations that show reliability
and maintainability factors. Second, it shows
period of actual operation versus inoperative
periods; such period values are known tech-
nically as equipment population figures.

Without operation time or population figures
that are reasonably accurate, any one or more
figures of merit are not significant. There-
fore, one realizes how the operation time is
essential in evaluating other reports, especially
the previously described DD-787.

The operational time log form is relatively
simple to complete, especially when an equip-

ment is supplied with time meters. Only five
or six entries need to be made on the first day
of a month (depending upon the number of
meters), and three more on the last day of the
same month; a total of 8 or 9 block entries
(on a single form) are to be completed per
equipment per month.

Equipments without time meters require
somewhat greater effort. The format for pack-
aging and using these forms is explained in
detail on the covers (similar to the DD-787).

Electronics Performance and
Operational Report, NAVSHIPS 3878

The Naval Ships System Command must keep
tab on new (and converted) equipments to eval-
uate their usefulness. This is accomplished with
the Electronic Performance and Operation Re-
port, NAVSHIPS 3878, shown in figure 2-23.

It is not desired that reports be submitted
on all equipments. Detailed instructions for the
preparation and submission, and a listing of
equipment requiring a report are contained in
NAVSHIPS Instruction 9670-20E. When appli-
cable, NAVSHIPS 3878 is submitted monthly
to the Naval Ships Systems Command.

The NAVSHIPS 3878 reports are essential
to keeping the Naval Ships System Command
informed on equipment performance and op-
eration. Because they provide first-hand in-
formation on equipment under actual operating
conditions and report the maximum ranges ob-
tained, they are extremely valuable in evaluat-
ing the electronics maintenance program, en-
forcing manufacturers' guarantees, evaluating
installation adequacy, improving equipment op-
eration and safety, and improving equipment
design.

The NAVSHIPS 3878 report contains a place
for general remarks on the back of the form
(not shown). Indicated here is any pertinent
information not included elsewhere on the form
such as detailed information on any unusual
difficulty encountered in operation; exceptional
maintenance required; and suggestions for im-
provement in design, tests, and new applica-
tions. The forwarding of suggested improve-
ments is not to be construed as authority to
modify the equipment in any way. Nor does the
forwarding of this report eliminate the separate
requirement for forwarding the Electronics
Failure Report (DD 787).

ELECTRONIC PERFORMANCE & OPERATIONAL REPORT
NAVSHIPS 3878 (Rev. 4-60)   *Submit original only to Bureau - No forwarding letter required*

REPORT-BUSHIPS-6679-5

FROM: USS RANGER (CVA-61)
(Ship name, type and hull no.)

☐ LANT
☐ FLEET
☒ PAC

TO: CHIEF, BUREAU OF SHIPS (CODE   )

REPORT CLASSIFICATION: UNCLASSIFIED
DATE: 1 March 1966

REPORTING PERIOD: FROM 1 FEB 66 TO 28 FEB 66

TYPE AND MODEL OF EQUIPMENT: AM-1365/URT Amplifier
SERIAL NUMBER: 383

FIELD CHANGES TO DATE: ACCOMPLISHED None   NOT ACCOMPLISHED None

HOURS DURING PERIOD OF THIS REPORT: OPERATED 180   NOT IN OPERATING CONDITION 564

PERFORMANCE FIGURE (PF) & TECHNICAL EVALUATION: ☐ OUTSTANDING ☐ GOOD ☐ SATISFACTORY ☒ UNSATISFACTORY

OPERATIONAL EVALUATION: ☐ OUTSTANDING ☐ GOOD ☐ SATISFACTORY ☒ UNSATISFACTORY

RADAR

PEAK POWER OUTPUT (PT)		AVER. VSWR IN TRANSMISSION LINE		AVER. ECHO BOX RING TIME		MIN. DISCERNIBLE SIGNAL (PMDS)	
dbm				YDS			dbm
MAX. RANGE TARGETS DETECTED MI	MI	MI		MAX. ALTITUDE AT RANGE DETECTED MI		MI	MI
MAX. ALTITUDE TARGETS DETECTED FT	FT	FT		RANGE AT MAX. ALTITUDE DETECTED FT		FT	FT
TARGET CLASS. TYPE - DETAIL (SEE REVERSE)				TARGET CLASS. TYPE - DETAIL (SEE REVERSE)			
MAXIMUM RELIABLE RADAR RANGE			MI	MINIMUM RELIABLE RADAR RANGE			YDS

SONAR

SOURCE LEVEL (LS) db//uBAR	RECEIVING SENSITIVITY db//VOLT/uBAR	SEA STATE	PROCEDURE USED		
NOISE LEVEL db//VOLT	~5 KNOTS	10 KNOTS	15 KNOTS	20 KNOTS	25 KNOTS	30 KNOTS
MAXIMUM RANGE SONAR TARGETS DETECTED AND TRACKED	RANGING YDS		LISTENING YDS	SOUNDING FATHOMS	
TARGET CLASSIFICATION TYPE AND DETAIL					
BT PATTERN					
OWN SHIP'S SPEED.	KTS		KTS	KTS	

COMMUNICATIONS

PERCENT OF TIME OUT OF CONTACT WHILE WITHIN RANGE (IF ANY)	ANTENNA SYSTEMS	INTERFERENCE (Frequencies, intensity, and sources)
0 %	No problems	No problems

POWER OUTPUT		AVERAGE VSWR	REL RANGE	RECEIVER SENSITIVITY	
Voice 100 WATTS		1.5:1	40 miles	NA	UVOLTS

ELECTRONIC WARFARE

MAXIMUM RANGE AND ALTITUDE TARGETS DETECTED	MI	FT	MI	FT	MI	FT
TARGET CLASSIFICATION TYPE AND DETAIL (SEE REVERSE SIDE)						
MAXIMUM RELIABLE RANGE AND ALTITUDE	MI	FT	MI	FT	MI	FT
TARGET CLASSIFICATION TYPE AND DETAIL (SEE REVERSE SIDE)						
MAX. RANGE SONAR TARGETS DETECTED	BT PATTERN YDS		MAX. RELIABLE SONAR RANGE	YDS	BT PATTERN	

35.83

Figure 2-23.—Electronic Performance and Operation Report, NAVSHIPS 3878.

SHIP ELECTRONICS INSTALLATION RECORD, NAVSHIPS 4110

The Ship Electronics Installation Record, NAVSHIPS 4110, furnishes an up-to-date inventory of all electronic equipment aboard each ship to interested fleet and shore activities and is one of the records which will not be replaced by the 3-M system.

To be effective, the NavShips 4110 must be extremely accurate and must be submitted promptly. To indicate the importance and purpose of the document, the NavShips 4110's are specifically used:

(1) By the Naval Ships System Command for determining the equipments to be installed during overhauls and conversions, and the equipment to be procured in the future.

(2) By Naval Shipyards in overhaul planning.

(3) By the Electronic Supply Office to determine repair parts allowances.

(4) To prepare Coordinated Shipboard Allowance Lists (COSAL).

Instructions for preparing, revising, and submitting NAVSHIPS 4110 are contained in NAVSHIPS 900, 135D.

ELECTRONIC MATERIAL IDENTIFICATION

Numbers currently used by the Navy to identify and catalog electronic repair parts are Federal Stock Numbers (FSN). These are the numbers which should be used on all supply and inventory documents.

The FSN used throughout the Department of Defense utilizes a 4-3-4 digit pattern prefixed by a numeric-alphabetical designator called a dual cognizance symbol (fig. 2-24). The numeric-alpha combination identifies the general category of the material, the funding, the Navy Inventory Control Point having cognizance, or the Navy Retail Office having administrative responsibilities for the items under the managership of a Defense Supply Center.

Following the dual cognizance symbol is a four-digit number which indicates the Federal Supply Class. The first two digits represent the Federal Supply Group. In the example shown in figure 2-24, the "59" specifies that the item is an Electrical/Electronic Equipment Component. The second two digits, "20" together with the number "59" show the Federal Supply Class in which fuses and lightning arrestors have been classified. The last seven digits—the 3-4 part of the pattern—are the specific item identifier. They are called Federal Item Identification Numbers (FIINs) and are serially assigned. The same FIIN is never used to describe more than one line item, and once assigned, that number is never reassigned, even though the item identified by the number may have been dropped from the supply system.

Where FSNs are not available, other numbers must be used to identify the repair parts. Such other numbers may be reference numbers or manufacturers' part numbers. Each company or concern which manufacturers material for Government use is assigned a five-digit "manufacturer's code". The use of a reference number, part number, or manufacturer's code is important in the identification of a part when the FSN is unavailable.

U. S. NAVY ELECTRONIC SUPPLY OFFICE

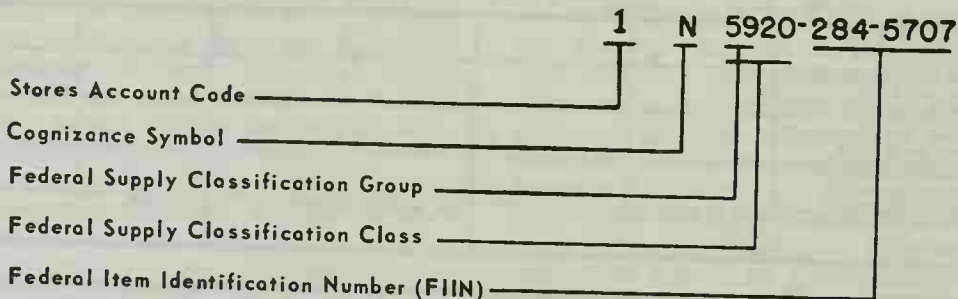The U. S. Navy Electronic Supply Office (ESO) is located at Great Lakes, Illinois,

Figure 2-24.—Breakdown of typical Federal Stock Number.

107.13

where inventory control is exercised over assigned electronic material. ESO has been designated by the Secretary of the Navy to serve as the inventory control point for Navy Ships System Command electronic repair parts (less nuclear items) required in the maintenance, upkeep, and overhaul of equipments in ships and at shore stations, other than those for which control is specifically retained by the System Command.

A part of ESO's primary mission of supplying activities in the electronics distribution system with repair parts is to prepare Coordinated Shipboard Allowance Lists (COSAL) for ships and Electronic Repair Parts Allowance Lists (ERPAL) for shore stations, develop and maintain stock numbered parts lists (referred to as SNITs or Stock Number Identification Tables and APLs or Allowance Parts Lists) for individual equipments, and participate in the initial provisioning of parts for new electronic equipments.

The U. S. Navy Electronics Supply Office stocks no material. It controls and directs the flow of assigned electronic material through the electronics supply segment of the Navy Supply System.

MAINTENANCE OF STOCK LEVELS

The necessity for a consumer activity to maintain sufficient stocks to meet its requirements is obvious. Less obvious, but also important, is the necessity to avoid overstocking. Overstocking not only increases the dollar value of an activity's inventory and occupies storage space unnecessarily, but stocks which may be needed elsewhere are immobilized.

Certain parts may require replacement so often that there may be a tendency to consider them as a part of shop stores. In such a case, an increase in the quality of manufacture, an improvement in design, or obsolesence may leave the shop store with a 6-year instead of a 90-day supply. For this reason alone it is desirable to keep shop stores to a minimum.

REQUISITIONS

In order to have on hand replacements for parts which fail, activities operating electronic equipments must make their requirements known to the supply system. This is done by means of requisitions forwarded to the appropriate supporting supply activity. The supply officer is responsible for requisitioning materials, however, he must be told what materials and quantities are needed.

After a requirement is identified it must be transmitted to the supply department in some acceptable form. Ultimately, documentation must be prepared and processed which accomplishes the recording and accountability requirement.

Ashore and on ships with data processing equipment, which include most aircraft carriers, the DD Form 1348 is used as the request and issue document. An example of a properly prepared DD Form 1348 is shown in figure 2-25. The shaded areas on the form indicate which information is furnished by the customer, with the balance of the information being completed by supply department personnel.

On ships operating under manual supply procedures, which include the majority of all ships, the NAVSANDA Form 1250 is used as the request and issue document. An example of a properly prepared NAVSANDA Form 1250 is shown in figure 2-26. The shaded portion of this document indicates which information is furnished by the customer.

EQUIPAGE CUSTODY

Equipage is the term used to designate nonconsumable material. This type of material has greater value than consumable supplies and more functional importance, such as signal generators, multimeters, and oscilloscopes.

A custody signature on Equipage Stock Cards and Custody Record, NAVSANDA Form 306 (fig. 2-27) is mandatory for equipage under Naval Ships System Command cognizance, as listed in NAVSUP Manual, Volume III; and equipage items which the commanding officer determines should have a custody signature.

A copy of the Equipage Stock Card and Custody Record must be furnished to the department head at the time he signs the original custody card. The duplicate copy becomes the department head's record of equipage requiring custody signature for which he is responsible. As equipage is received or expended, appropriate entries are made on all equipage records.

Figure 2-25.—DD Form 1348.

43.24

Figure 2-26.—NAVSANDA Form 1250.

17.84

When the department head issues the equipage, he should have that person sign a sub-custody receipt; however, this does not relieve the department head of the responsibility for the equipage.

SURVEY REPORTS

A survey is the procedure required by U.S. Navy Regulations when naval property must be condemned as a result of damage,

EQUIPAGE STOCK CARD AND CUSTODY RECORD								

NAV. S. AND A. FORM 306 (REV. 8–54)

CUSTODY SIGNATURE REQUIRED ▶ ☒ YES ☐ NO   CARD NO. Op 11

DEPARTMENT Operations	ALLOWANCE 2	STOCK NO. 2RX6625-785-6500	UNIT Ea	UNIT PRICE 1333.00	ALLOWANCE LIST NO. 78730709 GROUP PAGE LINE

Oscilloscope, AN/USM-105A

DATE	VOUCHER NO.	RECEIVED FROM ISSUED TO	QUANTITY RECEIVED	QUANTITY ISSUED	BALANCE	I acknowledge receipt of quantity of this article as indicated
7/24/67	079825	NSC NORVA	1		1	*R.D.Williams*

U. S. GOVERNMENT PRINTING OFFICE   16—44372–5

7.3

Figure 2-27.—Equipage Stock Card and Custody Record, NAVSANDA Form 306.

obsolescence, or deterioration, or is acknowledged as nonexistent as a result of loss, theft, or total destruction. A survey is made on Survey Request, Report and Expenditure, NAVSANDA Form 154.

Formal surveys (fig. 2-28) are made by a commissioned officer or a survey board appointed by the commanding officer. A formal survey is required at all times when it appears that responsibility for lost or damaged material may be placed on a person or persons in the naval service. Also, formal surveys are required for certain classes of material or articles designated by the bureau or office concerned. In the absence of instructions from higher authority, the commanding officer determines whether a formal or an informal survey is required.

Informal surveys will be made by the head of the department having custody of the material to be surveyed. A survey will be made when a material condition becomes known or when instructions by higher authority direct a survey.

ADDITIONAL REPORTS

Type commanders and other authorities may require reports in addition to those required by the Navy Department. Instructions concerning such reports may be promulgated via letter, message, or other official means. The Electronics Technician, therefore, must examine all official correspondence regarding electronics forwarded to his activity in order to keep abreast of current instructions.

ELECTRONIC SAFETY PRECAUTIONS

There is great diversity in the dangers inherent in the technician's work. These dangers range from those encountered while working

SURVEY REQUEST, REPORT AND EXPENDITURE
NAV. S. AND A. FORM 154 (REV. 8-58)

DATE	NUMBER
1 Aug. 1967	5219852180245

ACTIVITY

(52198) USS Jones Ingram (DD-938)

ORIGINATOR (Signature and title)
Robert C. Sword
R.C. Sword, LTJG., USNR, Operations Off.

REQUEST FOR SURVEY

ITEM	STOCK NO. AND DESCRIPTION	QUANTITY	UNIT PRICE	TOTAL VALUE
1	Oscilloscope OS-8E-U 2F 6625-649-9248	1	$120	$120

REASON FOR SURVEY	ACCOUNT IN WHICH CARRIED (APA, NSA, etc. or nonstores)	OTHER DATA (Source, date of receipt, etc.)
Damaged beyond economical repair	Nonstores	NSC, Norfolk 10/25/67

ACTION BY COMMANDING OFFICER OR DELEGATE

TYPE OF SURVEY	SURVEY TO BE MADE BY	SIGNATURE (C. O. or delegate)	(DATE) 1 Aug. 1967
[X] FORMAL	1. LTJG. R.B. Louer		
	2. LTJG. H.B. Barnes	*R.E. Spallone*	
[ ] INFORMAL	3. LTJG. S.M. Lindquist	R.E. Spallone, CDR., USN	

SURVEY REPORT AND RECOMMENDATION

Condition: Damaged beyond economical repair

Cause: Dropped on deck while being moved during heavy seas; considered unavoidable

Responsibility: None

RECOMMENDATIONS

[X] EXP. FROM RECORDS   [ ] TRANSFER TO B-27D   [ ] TRANSFER TO MAT'L COND. CODE ($_____)   (Est. repair cost)   [ ]

ITEM(S) SURVEYED IN ACCORDANCE WITH NAVY REGULATIONS BY: [Signature(s)]

(1) LTJG. R.B. Louer	(2) LTJG. H.B. Barnes	(3) LTJG. S.M. Lindquist

REVIEW OF SURVEY REPORT

[X] APPROVED [ ] DISAPPROVED	SIGNATURE (C. O. or delegate) *R.E. Spallone* R.E. Spallone, CDR., USN	DATE 5 Aug. 1967	FORWARD TO (Bureau)

ACCOUNTING DATA   BUREAU APPROVAL

APPROPRIATION	EXPEND. ACCOUNT
Nonstores Account	

DISPOSED OF AS INDICATED	SIGNATURE R.C. Cairo, LT., USN, Supply Officer	DATE 7 AUG. 1967

Figure 2-28.—Survey Request, Report and Expenditures, NAVSANDA Form 154.

7.6

aloft to those of nuclear radiation. The technician is also exposed to the dangers of working with high voltage, RF radiation, volatile and combustible materials, and with power tools, etc. The list is almost endless. It must be stressed, however, that these dangers are potential dangers; dangers that can be eliminated, or at least minimized, if the proper safety precautions are observed.

Safety precautions are not stagnant; rather, they are continuously changing. A complete list of safety precautions for today may be inadequate for tomorrow. Advances in technology are often accompanied by new dangers, necessitating new safety precautions. In addition, continuing research in the area of safety periodically brings to light heretofore hidden dangers.

Only by the following actions can you ensure that you are aware of all of the dangers which may be encountered in your work and that you know all of the necessary safety precautions that must be observed:

a. Ensure that all the publications pertaining to safety are aboard. (A complete list of these publications is at the end of this discussion.)

b. Maintain these publications in a current status by making the corrections and inserting the changes as they are received.

c. See that your name is on the routing list for all periodicals that may contain articles on safety; such as the EIB or the Navy Ship System Command Technical News.

SOURCES OF INFORMATION

Unfortunately, all of the precautions necessary to ensure the safety of the technician cannot be found in any one publication. They all exist, however, in the many publications, periodicals, and directives of the various offices and bureaus of the Navy Department. It is the purpose of this discussion to provide a list of authoritative sources of safety information.

OPNAV Notice 5100 of 4 November 1964 states that the Chief of Naval Operations will periodically publish an index of publications, pamphlets, periodicals, and directives, issued by the bureaus, system commands, and offices of the Navy Department, which contain safety precautions applicable to the operating forces. The index is issued as a series of OPNAV Notices under the 5100 subject classification number, and provides an up-to-date list of all published material pertaining to safety precautions. Every Electronics Technician should check this list and ensure that all publications that are applicable to his division are readily available to himself and to others within the division.

The following list of publications was taken from the OPNAV Notice stated above, and includes those of most importance to the technician:

Title	Number
Naval Ships Technical Manual Chap 9670, Sec V	NavShips 250,000
Electronic Installation and Maintenance Book (EIMB) General Handbook	NavShips 0967-000-0110
EIMB Installation Standards	NavShips 0967-000-0130
Electronics Information Bulletin (EIB)	NavShips 900,022A
Instruction	NavShips 6260.5
Instruction	BuMed 6200.5
Various Equipment Technical Manuals	
Standard First Aid Training Course	NavPers 10081-B
Department of the Navy Safety Precautions for Shore Activities	NavSo P-2455
Type Commanders' or Commanding Officers' Directives	

CHAPTER 3

USE OF SPECIAL TEST EQUIPMENT

The purpose of this chapter is to better acquaint the technician with the purpose and practical use of special test equipment. The test equipment discussed is by no means all that will be encountered; however, it is representative of special types of test equipment that you will find aboard most ships and at shore stations.

RADIO INTERFERENCE MEASURING SET AN/PRM-1A

Radio Test Set AN/PRM-1A (fig. 3-1) can be used for radio interference surveys to determine the source of radiated or conducted interference from any source within its frequency range. Such interference may be generated by gasoline engine generators, motors or electronic equipment. Field intensity measurement surveys may be made with Radio Test Set AN/PRM-1A for adjusting directive antennas or for exploring radiation patterns, where the field intensity may vary over a wide range of values. The equipment may also be used as a sensitive radio frequency microvoltmeter. The test set is portable and is suitable for operation aboard naval ships, at shore stations, in the field, in aircraft, and in military vehicles.

TYPES OF RADIO INTERFERENCE

Radio interference, sometimes called "noise", disrupts the utility of radio frequency receivers. Interference includes many forms of disturbances and may be roughly classified as continuous and discontinuous.

Continuous disturbances include random interference such as thermal agitation. Atmospheric interference, or static, originating in lightning discharges in local storms and in more distant tropical storms, resembles random interference in that the impulses are frequent and overlap, with sharp peaks exceeding the average level. If interference impulses follow each other rapidly and are not clearly distinguishable, then the interference is random.

Discontinuous interference is made up of sharp pulses, the frequency determining the character of the interference. If the pulses are relatively infrequent and clearly separated, the interference is termed impulsive. Impulsive interference may be generated by an internal combustion engine ignition system, power line discharges, electrical switching operations, electronic equipment, and by other electrical and electromechanical devices.

Because of the wide frequency range covered by continuous and discontinuous interference, the magnitude of the interference will depend upon the bandwidth of the measuring equipment. The bandwidth characteristics of the AN/PRM-1A are taken into consideration and are utilized in obtaining average, quasi-peak, and peak values of interference. Shipboard electronic interference is described in the Electronic Installation and Maintenance Book, NavShips 0967-000-0150.

FIELD INTENSITY MEASUREMENTS

Field intensity is the value of the electric field at a given point and is measured in terms of volts-per-meter. One volt-per-meter is equivalent to a potential of one volt induced in a vertical rod antenna having an electrical length of one meter.

There are two principal methods of making field intensity measurements. One method is the substitution method, in which a loop antenna is connected to a receiver, and metering circuits give an indication of the receiver output. The receiver is tuned to the desired signal. The loop is rotated to give maximum signal, and the relative value of the received signal is
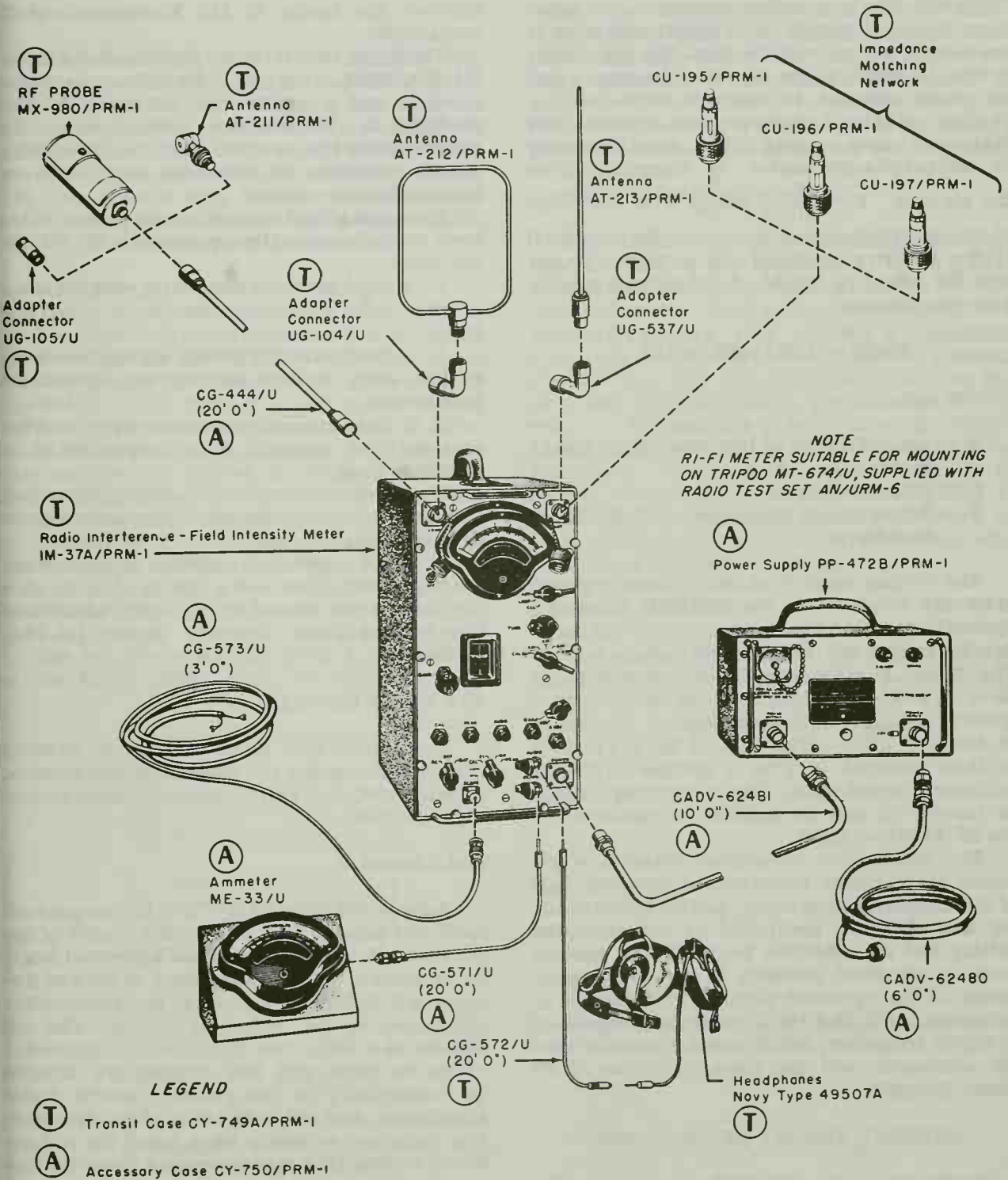
162.2

Figure 3-1.—Radio Test Set AN/PRM-1A, Connection Diagram.

recorded. The loop is then rotated to give minimum received signal, and a signal generator is inserted in series with the loop. The signal generator is tuned to the received frequency and its output adjusted to obtain the same receiver reading of signal input as was obtained with maximum loop signal. The field intensity in microvolts-per-meter is computed from the formula $E = \dfrac{V}{H(\text{eff})}$, where V is the value of the substitute signal in microvolts and H(eff) is the effective height of the antenna. To obtain the effective height of the antenna use the following formula;

$$H(\text{eff}) = 1.351 \ ANF \ x \ 10^8$$

Where

- A = area of 1 turn of loop winding in square inches
- N = number of turns
- F = frequency in kilocycles, and H(eff), is in meters.

The second method of field intensity measurements employed in the AN/PRM-1A uses a specially designed sensitive receiver with built-in attenuators and a calibrated voltage source. The Radio Interference-Field Intensity Meter IM-37A/PRM-1 is tuned to the desired frequency and the calibrated voltage is applied to the r-f input. The gain of the i-f section is then adjusted to give a standard value of calibrator signal input. The calibrating voltage is turned off and the antenna is connected to the RI-FI Meter input.

The attenuators (discussed later) are adjusted for a meter indication at the upper half of the indicating meter dial, and the meter reading thus obtained, multiplied by the attenuator setting and the effective height of the antenna used, is the field intensity in microvolts-per-meter of the received signal. This method is advantageous in that the accuracy is independent of signal frequency, and depends primarily upon the attenuator and the linearity of the RI-FI meter circuits.

GENERAL THEORY OF OPERATION

Radio Test Set AN/PRM-1A is basically a meter intended for intensity measurements of all types of radio frequency energy in the low and medium radio frequency spectrum between the limits of 150 kilocycles and 25 megacycles.

The Radio Interference-Field Intensity Meter IM-37A/PRM-1 (fig. 3-1) includes a sensitive receiver and a vacuum-tube voltmeter (VTVM) section. It also contains internal means for standardizing the receiver gain, thus permitting direct readings in indicated microvolts or microvolts-per-meter.

By special input circuit design, provision has been made for use of the equipment in the following ways:

1. A field intensity measuring equipment operating from a vertical antenna of convenient size.

2. A field intensity measuring equipment operating from a loop antenna with directional properties.

3. A field intensity measuring equipment operating from a small probe responsive to an induction field.

4. A two-terminal voltmeter making both symmetrical and asymmetrical measurements with respect to ground.

A set of eight charts contain simplified operating instructions and a family of calibration curves that are plotted for each individual Radio Interference-Field Intensity Meter IM-37A/PRM-1.

R-F INPUT DEVICES

Input devices for Radio Test Set AN/PRM-1A (fig. 3-1), include a rod antenna, a loop antenna, a loop probe, an r-f probe, and impedance matching networks.

Rod Antenna

Antenna AT-213/PRM-1 is a telescopic vertical rod antenna with an effective length of approximately one-half meter and a physical length of 41 inches. The voltage range of the rod Antenna AT-213/PRM-1 is from two microvolts-per-meter to two volts-per-meter. The rod antenna is a capacitive pickup device, and measurements made with this antenna are affected by conductivity of the ground, nearby metal structures, and the body effect of the operator. For optimum accuracy when using the rod antenna, comparison measurements using the loop antenna and the rod antenna must be made at the frequency in use. The correction factor so obtained is then applied to the rod antenna meter reading.

Loop Antenna

Antenna AT-212/PRM-1 is a shielded loop with rectangular shape and an equivalent diameter of 8 1/4 inches. The loop is a solenoid winding consisting of 2 turns of 13 strand #36 wire, with a voltage range from 10 microvolts-per-meter to 100,000 microvolts-per-meter. The loop antenna may be used as a hand-held probe connected to the end of the 20-foot cable CG-444/U used with the probe antennas.

Loop Probe

The hand-held probe Antenna AT-211/PRM-1 is connected to the LOOP receptacle on the Radio Test Set by means of the 20-foot cable CG-444/U. The loop is a universal winding 1/2" effective diameter, consisting of 13 turns of 15/44 wire. It is enclosed in a phenolic case and can be rotated on its base to utilize the directional characteristics of the loop. The voltage range of the loop probe antenna is from approximately 1,000 microvolts-per-meter to approximately 10 volts-per-meter.

R-F Probe

The r-f Probe MX-980/PRM-1 permits use of the RI-FI meter as a two-terminal volt-meter for measurement of conducted interference on a line. The unit is connected between one side of a line and ground or two lines, balanced or unbalanced to ground. In use, short insulated jumpers are connected to the probe input and the free ends of the jumper are connected to one side of the line and to ground.

Impedance Matching Networks

The three impedance matching networks, CU-195/PRM-1, CU-196/PRM-1, and CU-197/PRM-1 are physically identical except for circuit impedance. They are connected to the ANT receptacle where they present a capacitive reactance equivalent to that of the rod antenna. The 50-ohm unit, CU-197/PRM-1, and the 20-ohm unit, CU-196/PRM-1, are intended for use where it is desirable to terminate a transmission line, an antenna, or other circuits in these resistance values at the input to the RI-FI Meter. The unterminated unit, CU-195/PRM-1, is intended for use with signal generators or other devices whose output is not meant to be terminated in a low resistance.

CONTROL CIRCUITRY

The signal channel, as is evident in the simplified Block Diagram (fig. 3-2), closely

ANTENNAS	DESCRIPTION
AT 2-3/PRM-1	41" ROD
AT-212/PRM-1	LOOP
AT-211/PRM-1	LOOP PROBE
MX-980-PRM-1	LINE PROBE

IMPEDANCE MATCHING NETWORK	
CU-195/PRM-1	GENERAL PURPOSE
CU-196/PRM-1	20-ohms
CU-197/PRM-1	50-ohms

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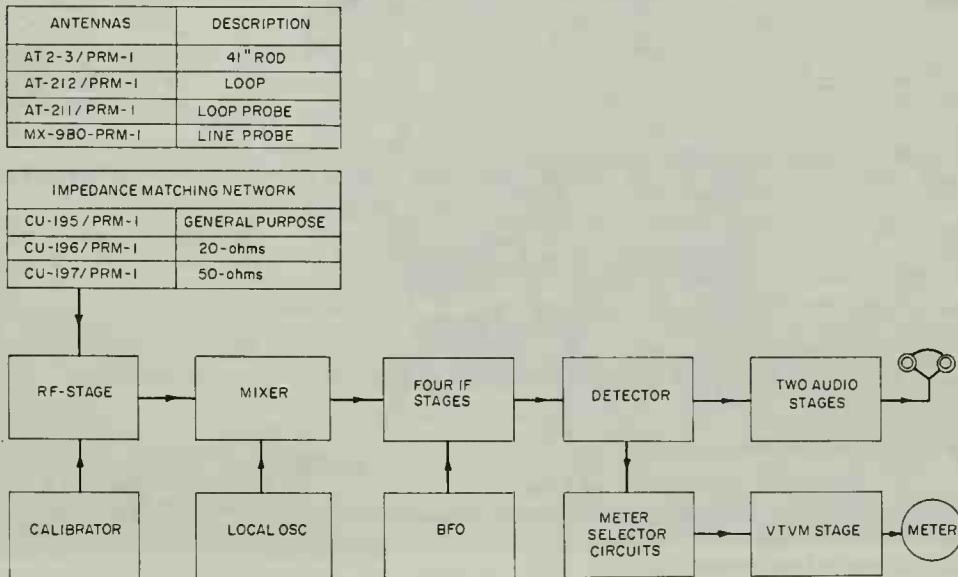
Figure 3-2.—Radio Test Set AN/PRM-1A, Simplified Block Diagram.

resembles a superheterodyne receiver in its r-f, i-f, and a-f portions, but differs from most superheterodyne receivers in its provision for attenuation and measurement of detector output.

R-F and I-F Attenuator

The built-in step attenuator provides the following step ratios: 1, 10, 100, 1,000, and 10,000. Provision is made to obtain these ratios by discrete combination of attenuation in the antenna circuit and the first i-f stage.

R-F Stages

The r-f signal or interference (as picked up by the antenna or probe) is delivered to the LOOP or ANTENNA receptacle (fig. 3-3) at the r-f input. The r-f signal is amplified in the r-f stage (fig. 3-2), and mixed with the local oscillator frequency in the mixer stage, to produce the intermediate frequency. The use of two intermediate frequencies (455 KC and 1600 KC) is used to provide continuous

162.4
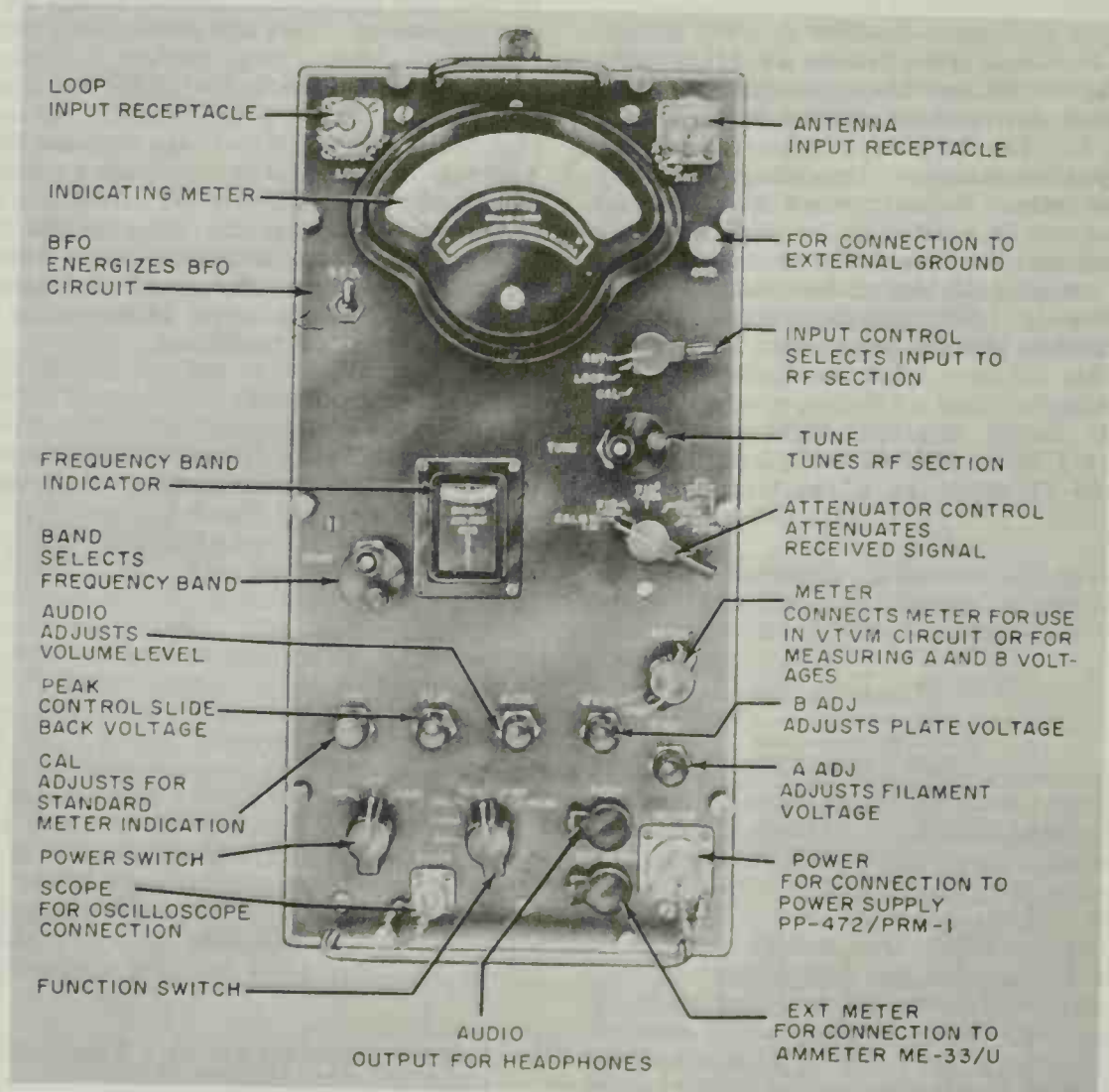
Figure 3-3.—Radio Interference-Field Intensity Meter IM-37A/PRM-1, Panel Controls and Receptacles.

frequency coverage from 150 KC to 25 MC. On all bands except Band 7, the local oscillator frequency is above the input signal frequency. See table 3-1.

I-F Amplifier Assembly

The i-f amplifier assembly consists of four i-f stages and a BFO (beat frequency oscillator). The first two i-f stages are deciding factors in determining the operational characteristics of the VTVM at the receiver output. In this connection, operating potentials of the two stages are arranged so as to obtain logarithmic deflection on a two-decade meter. All i-f stages have separate tuned circuits for each i-f frequency, the appropriate circuits being selected by the Band Selector Switch (fig. 3-3).

Two intermediate frequencies are used in the seven frequency bands as listed in table 3-1. The table also shows how the local oscillator is tuned to produce the i-f (difference) signal.

Table 3-1.—Listing of I-F Frequencies and Local Oscillator Frequency

Band	I-F Frequency	Local Oscillator Frequency
Band 1	455 KC	Above signal frequency
Band 2	1600 KC	Above signal frequency
Band 3	455 KC	Above signal frequency
Band 4	455 KC	Above signal frequency
Band 5	1600 KC	Above signal frequency
Band 6	1600 KC	Above signal frequency
Band 7	1600 KC	Below signal frequency

The third i-f stage (not shown) is similar to the preceding two stages with the exception of the addition of a gain control. The fourth i-f stage (not shown) is a power driver stage for increasing the dynamic range, utilizing a pentode power amplifier tube.

The beat frequency oscillator is used for aural reception of CW signals. The audio signal is produced by heterodyning the i-f signal frequency and the BFO frequency. The BFO switch (fig. 3-3) must be placed in the BFO position, to apply filament voltage to the BFO. When the switch is in this position, the indicating meter is disabled and no readings can be made. The BFO is not to be used for zero beating the i-f signal.

Detector Stage

The detector stage demodulates the i-f signal utilizing a linear diode detector circuit. The diode output branches into two circuits - to the first audio stage and to the AGC and VTVM circuits.

VTVM STAGE

Two pentodes are employed in the VTVM stage, which is a circuit adaptation of a balanced VTVM circuit. The d-c voltage to be measured is applied to the first stage and upsets the balance of the stage, causing the indicating meter to deflect upscale and indicate the magnitude of the unbalance.

The use of this balanced VTVM circuit counteracts the effect of supply variation since supply variations will affect both tubes equally. Balance is indicated by a zero reading of the meter when no signal input is present.

Provision is made in the VTVM circuit for connecting an external meter in series with the panel-mounted meter.

WEIGHTING CIRCUITS

The term "weighting" as applied to the AN/PRM-1A specifically refers to the time constant introduced in the detector and AGC circuit. This "weighting" effectively determines the time versus voltage characteristics of the indication obtained.

The weighting circuits are selected by the function switch and serve to properly weight the voltage supplied to the indicating meter for the three types of measurements performed by the equipment. The weighting circuits also determine the AGC time constant.

Field Intensity Circuit

The AGC and VTVM circuits see a voltage source which charges slowly and discharges slowly so that the meter indication continuously shows an average of the input signal over 600 milliseconds. This is accomplished by the time constant of an RC circuit.

The meter reading for field intensity is proportional to the field intensity of radiated signals, or proportional to the averaged value of r-f interference. The reading will be independent of modulation in the case of modulated CW signals due to a filter preceding the Weighting Circuit.

The Quasi-Peak Circuit

In the quasi-peak function, the time constants of the detector weighting circuits are approximately one millisecond charge and 600 milliseconds discharge. This is arranged by charging a capacitor through the relatively low resistance of a diode and discharging the capacitor slowly, through a resistive load. The meter indication then is near the peak value of an interfering signal, including pulse signals with PRF as low as 300 per second.

The Peak Circuit

In the peak function, the circuit is similar in some respects to the field intensity circuit. The VTVM and AGC circuits are the same, but an additional bias circuit has been added to enable the measurement of peak signal values. The peak value measured is the "peak" as seen by the second detector and not the true peak of the signal at the antenna input. The PEAK function of the function switch acts to apply a d-c voltage in parallel with the signal input. The circuit is similar to an adjustable delayed AGC system, since the d-c voltage is also applied to the detector. The d-c voltage is manually adjusted to the threshold of diode rectification. This measurement technique is similar to that employed in the "slide-back" voltmeter.

POWER SUPPLY

The Power Supply PP-472B/PRM-1, (fig. 3-1) is designed for operation from any a-c power source within the limits of 105-125 and 210-250 volts, and at any frequency between 50 and 1600 cycles per second. It provides a regulated positive 75 volts output for B+ and a 1.1 volts for filament power. When a 230 volt power source is used, the line voltage is reduced to 115 volts by action of an autotransformer. Radio Test Set AN/PRM-1 uses self-contained plate and filament voltage batteries for limited periods of operation in the field. At all times, including periods when the a-c power supply is used, two 4.5 volt and two 1.5 volt bias voltage batteries are required.

RADIO INTERFERENCE CONSIDERATIONS

An open, flat terrain is preferred for making field intensity measurements to avoid the possibility of absorption by dense growths of trees or steep ridges. Also, the face of a nearby cliff may act as a reflector and produce local variations in the intensity of the interference measured.

To isolate the radiated field from the induction field of a source of interference, make radio interference measurements of a radiating source at distances greater than twice the wavelength of the radiated signal. As the amplitude of the induction field energy varies inversely as the square of the distance from the source, the effects of the induction field are negligible at distances greater than two wavelengths.

Interference measurements made near power lines are subject to inaccurate results because of possible reradiation from the power lines. Every attempt should be made to avoid locating the equipment near closed loops present in iron frame buildings and topside structures aboard ship. If possible, avoid locating in the vicinity of underground pipes, trolley cables, or rail lines.

When the equipment is installed aboard ship or in a vehicle, it should be remembered that the metal body or frame tends to distort the pickup pattern of the dipole antenna. In addition, closed loops formed by wiring or structural members will affect the field pattern. In some cases, these errors can be determined with the equipment mounted in a designated location. Correction factors so obtained may then be applied to compensate for pickup pattern distortion.

When a variety of equipment, including radio receivers, is employed at any one site, a check

for interaction between equipments should be conducted.

The ideal site for making interference surveys is an open, flat terrain at a considerable distance (1000 ft or more) from buildings, electric lines, fences, and other wave obstructing devices. Ideal sites are rare in the more populated sections of the country. Therefore, it is good practice to check a proposed location by making signal measurements of the desired signal at several points in the vicinity. If the same value of field intensity is obtained at each of the points, any one may be considered satisfactory. If it is necessary to use an unsatisfactory site, a series of readings should be recorded at a number of different points in the neighborhood of the selected site, and detailed notes on the site conditions should be appended to the recorded data.

MAKING AN INTERFERENCE SURVEY

An interference survey of a suspected area should begin with a series of measurements, at the frequencies under investigation, made with the appropriate pickup. It is customary in general surveys to make measurements at five frequencies in each band of the equipment with sufficient overlap to ensure complete coverage of all bands. If a station or service is found operating on the frequency selected for measurement, move sufficiently away in frequency to avoid sidebands of the station. The survey report should, however, note the frequency and field intensity of signals from the station.

Certain types of interference surveys require scanning an entire band for evidence of peaks or minima. Any such peaks or minima found must be measured and noted in the survey report.

During the survey, the received signals should be aurally monitored at a volume level that will enable identification of the signal or interference.

Three-point fixes on each source of interference enable the geographic location to be determined by triangulation. Investigate each interference source, moving the RI-FI meter as close as is practical to the source to ascertain the distant fix. When the location has been narrowed down to a small area, use the loop probe to further determine the location of the offending equipment or power line.

If the interference is found to be a power line or other current-carrying conductors, use the RI-FI meter as a two-terminal voltmeter to determine the amount of conducted interference on the line. Standard safety precautions should be considered when connecting to power lines of any equipment, which may be energized from a switch not under the control of the operator of the radio interference measuring set. All connections should be made with the power disconnected from the lines of the equipment under test.

Standardizing Gain

The operator must standardize the gain of the Radio Interference-Field Intensity Meter IM-37A/PRM-1 prior to taking a series of measurements on any one frequency. The operator can detect any radical change in sensitivity of the unit because this condition will necessitate a CAL control setting at a point greatly different from that normally used to obtain the standard gain.

Measuring Sine Wave Signals

Before making any type of measurements, the equipment must be standardized as just discussed. After this procedure has been accomplished the measurement of sine-wave voltages should begin by turning the function switch to the FI position for a rms reading of a CW signal, or to the QP or PEAK positions when the carrier plus modulation is to be measured.

Adjust the position of the attenuator control until the meter reads within the scale range, preferably in the upper portion of the meter scale.

The signal being measured should be close to the frequency of gain standardization, as described above. Otherwise, it will be necessary to standardize the gain at the new frequency before a measurement can be accurately made.

To determine the signal strength, read the indicating meter in microvolts, and multiply this reading by the correction factor obtained from the charts. This product is then multiplied by the attenuator control setting.

OTHER APPLICATION OF THE RI-FI METER

The use of the RI-FI meter previously discussed represents a single test application of the equipment. With a knowledge of the function

and circuits of the operating controls considered earlier, the ET should recognize the fact that the RI-FI meter makes itself applicable to many test situations.

Measurement as a null instrument in conjunction with r-f bridges and slotted lines, which operate in the 20 to 400 mc frequency range, is possible. Extremely short leads, preferably shielded, must be used and care must be taken to assure proper impedance matching.

Checking the gain of a radio frequency stage in radio equipment by measuring the voltage input to the stage and the signal output of the stage, is another use. It should be noted that the RI-FI meter shunts approximately 50 ohms across the circuit and may upset the stage under test unless appropriately isolated.

You may also check the voltage of any 50-ohm transmission line up to one volt, provided the transmission line is properly terminated. The voltage rating (d-c or a-c power frequencies) that may be applied to the input without damage to the input circuits should not be exceeded.

Another use is determining the loss in a coaxial line connected to a signal generator by comparing the signal level at the input and output ends of the line. The appropriate impedance matching network must be used between the line and the RI-FI meter. To minimize the effects of standing waves, the higher attenuation positions ($X10^2$, $X10^3$, and $X10^4$) should be used when practicable.

FREQUENCY STANDARD AN/URQ-9A

In recent years, dependence on limited quartz-supply sources stimulated instrumentation for crystal-saving devices. Although prompted by crystal-saving, this development also made possible substantially smaller equipments, with multi-channel features; and furnished a means of obtaining increased stabilities.

During the past few years, emphasis in crystal-synthesis development has shifted from crystal-saving capability to precision-frequency generation. For example, in electronic systems today, a synthesizer is used in generating frequencies from a single crystal, internally available or externally supplied. Phase coherence is a prerequisite of performance.

A synthesizer should contain a reference frequency with a stability of at least 1 part in 10^8 part per day, and a means of translating the reference frequency to the required output frequencies. To achieve ideal performance (that is, to deliver one frequency at a time with no spurious output whatsoever), an ideal synthesizer should not contain elements such as oscillators, mixers, and multipliers. Amplification alone should be permissible; and amplifier linearity should be absolute. In reality, the mathematical functions of addition, subtraction, multiplication, and division must be performed in the translation of frequencies.

The method of frequency division has an advantage over that of frequency multiplication. For example; assume that 2 mc and 16 mc are required to be generated; that, in one case, a 16 mc signal is available and that energy is required at 2 mc; and that, in the other case, a 2 mc signal is available when it is necessary to generate 16 mc. When a 2 mc signal must be generated from a 16 mc signal, and a frequency division of 8 is employed, the frequencies developed are harmonically related to the output with a wide frequency separation. However, the multiplication of a 2 mc signal by 8 results in a 16 mc signal with sideband energies at 14 and 18 mc, which require fine tuning circuits to prevent interaction.

Most frequency synthesizers essentially contain a main variable-frequency oscillator (VFO), which will be disciplined by two secondary oscillators of lower frequency, for example, a 10 kc VFO and a 1 kc VFO. All three oscillators are phase locked to the stable crystal-controlled reference oscillator. Thus the stability of the synthesizer output frequencies equal that of the reference oscillator.

The Navy has in use today, transmitters and receivers with frequency synthesizers that require a highly accurate and stable frequency standard so that adjustments can be satisfactorily performed. One of these frequency standards is the AN/URQ-9 with a stability of 1 part in 10^9 parts per day.

The Frequency Standard AN/URQ-9 (fig. 3-4), is a highly stable, multiple-purpose frequency standard designed for continuous-duty use aboard ship and at shore facilities. It provides three output frequencies (5 mc, 1 mc, and 100 kc) at a minimum of 1 volt.

The frequency standard can be used for laboratory frequency measurements and to drive precision timing devices such as a time

TEST METER

TEST SWITCH
FUNCTION CARD

TEST
SWITCH

OUTER OVEN
HEATER
(INDICATING LAMP)

BAT. POWER SOURCE
(INDICATING LAMP)

AC POWER SOURCE
(INDICATING LAMP)

100 KC OUTPUT
(CONNECTOR)

I MC OUTPUT
(CONNECTOR)

5 MC OUTPUT
(CONNECTOR)
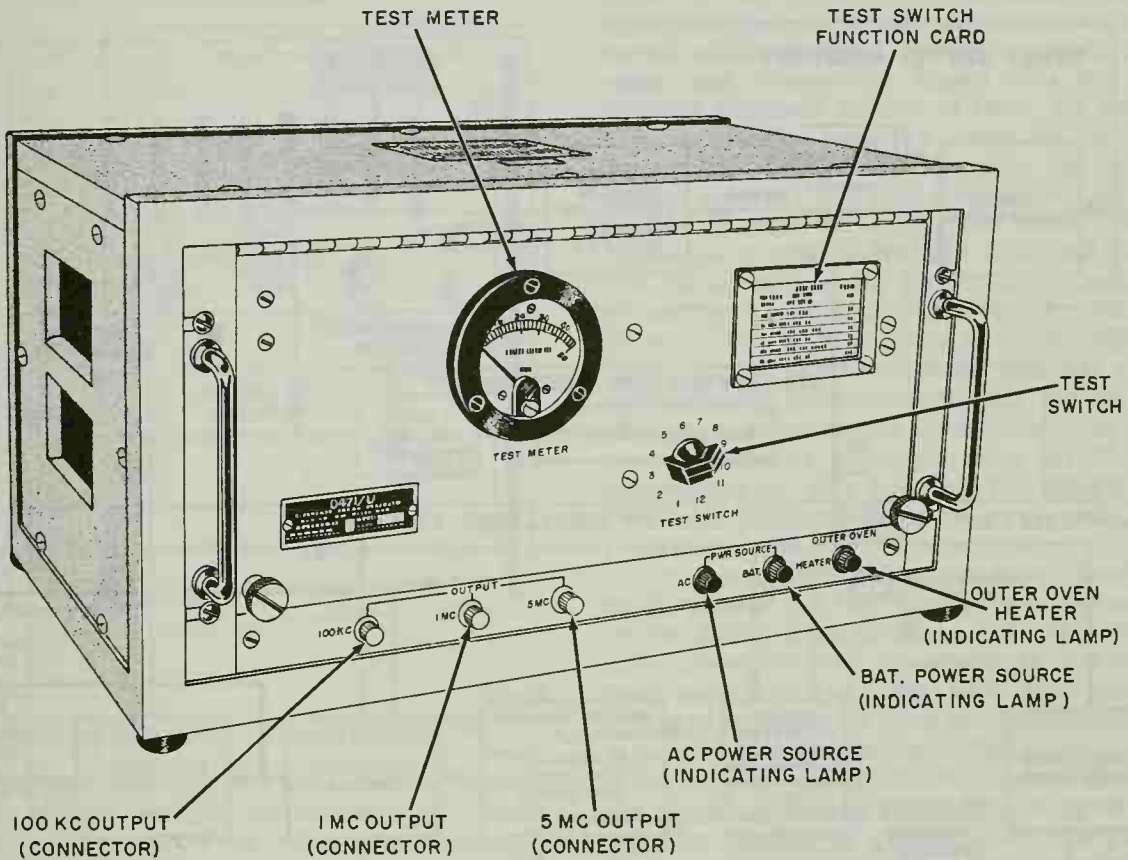
162.5

Figure 3-4.—Frequency Standard, Front Panel Control Locations.

comparator. It can also be used as a standby oscillator unit for other frequency/time-base standards.

The equipment is designed to operate from a nominal 115 volt, 60 cps, single-phase external power source capable of providing 240 watts (approximately 2.0 amps) during periods of maximum battery charging. A battery, which is built into the equipment, is automatically switched into the circuit to maintain operation in the event the external power source fails or is disconnected. When fully charged, the battery is capable of operating the frequency standard for two hours.

The frequency standard consists of three major assemblies (fig. 3-5), a radio-frequency oscillator assembly, a power supply assembly, and a standby battery assembly.

The r-f oscillator assembly contains the crystal, frequency-determining circuit, and amplifiers. The oscillator-amplifier generates a

5.0 mc signal which is fed to the 5.0 mc to 1.0 mc frequency divider. There the signal is amplified and a portion fed to output connectors on the front and back panels of the set. The other portion of the 5.0 mc signal is reduced to a 1.0 mc signal by the frequency divider circuits. The 1.0 mc signal is then amplified and a portion is fed to output connectors on the front and back panels of the set. The other portion of the 1.0 mc signal is fed to the 1.0 mc to 100 kc frequency divider where it is reduced to a 100 kc signal. The 100 kc signal is amplified and fed to output connectors on the front and back panels of the frequency standard.

The power supply assembly and the regulator-converter in the r-f oscillator assembly provide the regulated and unregulated voltages required by the set. If the external power source fails, or the power supply assembly develops a trouble, the set

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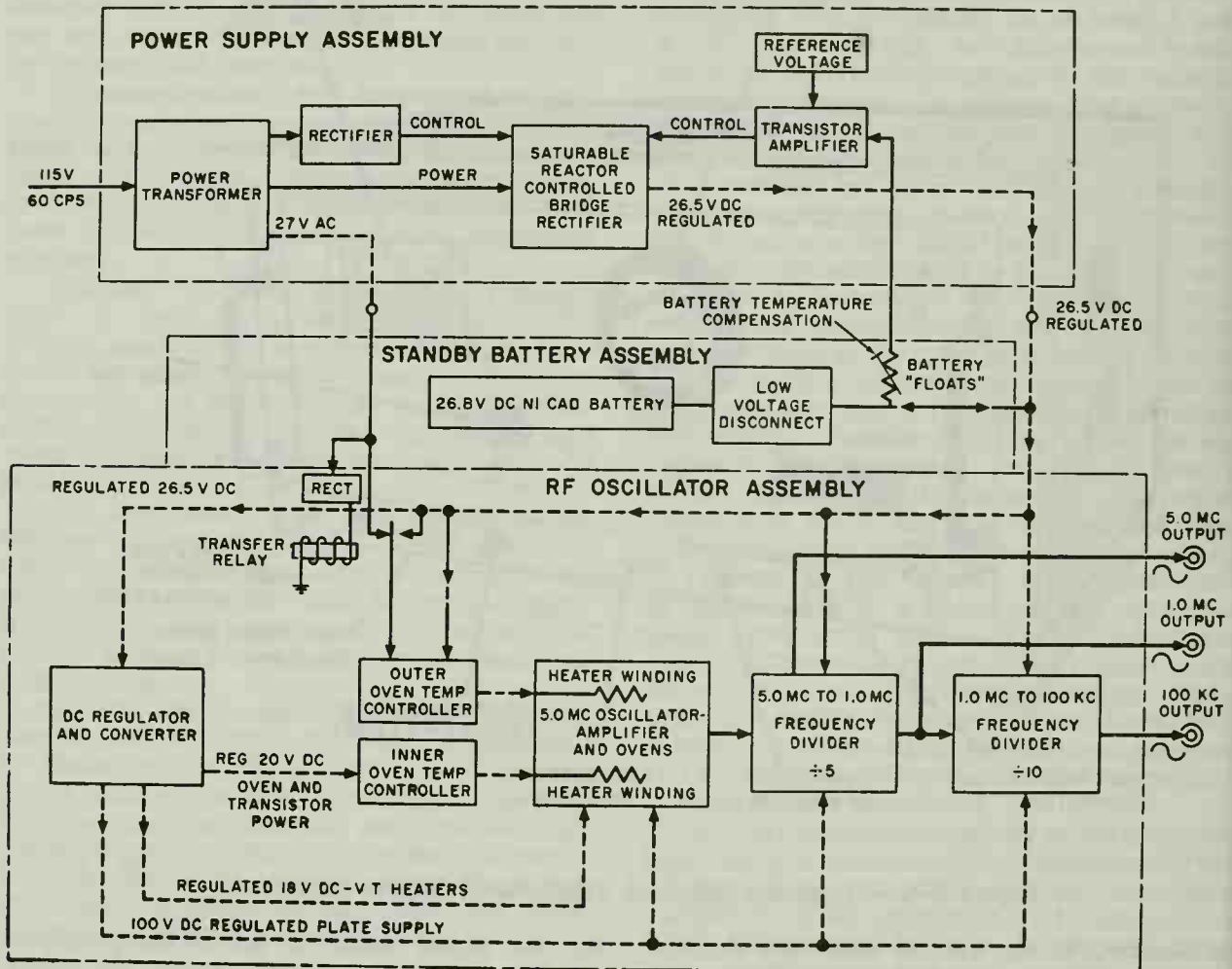
Figure 3-5.—Frequency Standard, Functional Block Diagram.

162.6

automatically switches over to battery operation. The frequency standard will operate from the battery for approximately two hours.

The frequency standard will be received by a calibration laboratory prior to being installed at the using activity. The calibration laboratory technicians will perform initial operation and calibration tests on the set and it will then be transferred to the using activity without being turned off (operating on the standby battery). Once the frequency standard is in operation and is calibrated, it must not be allowed to stop operating as this will cause the ovens to cool and will alter the operating frequency of the crystal. If the set does get turned off, it should be returned to the calibra-

tion laboratory for re-calibration; or, if the set is installed at a shore facility or aboard a ship that is in port, it may be allowed to run for one week and checked by the calibration activity for stability and correct frequency before being placed in service.

ELECTRONIC COUNTER 524D

A COUNTER type frequency meter (fig. 3-6), Electronic Counter 524D (Hewlett-Packard), can measure frequencies from 10 cps to 10.1 megacycles and display the readings in digital form on an eight-place indicating system. In addition to making direct frequency measurements, the counter can measure periods (0 cps to 100
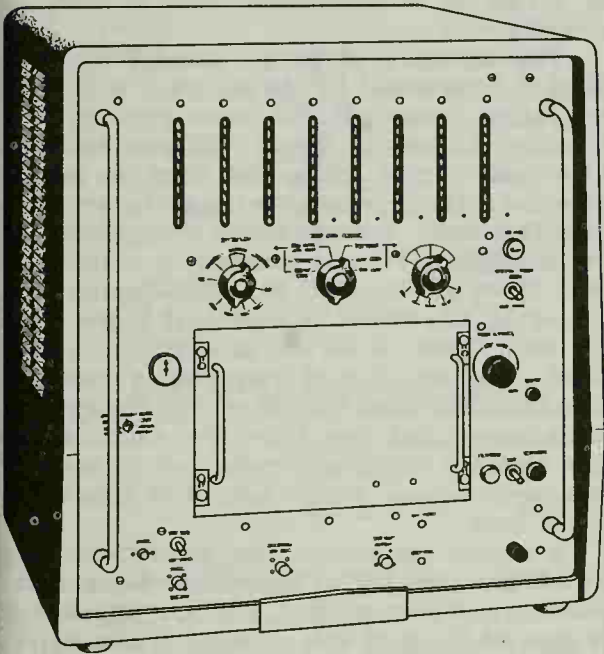
70-35

Figure 3-6.—Electronic Counter, 524D.

kc), frequency ratios, and total events. A self-check feature enables an operator to verify instrument operation for most types of measurements. The internal oscillator is stable within 5 parts of 10^8 per week. Thus these counters make good secondary frequency standards.

To increase the range of measurement, seven accessory plug-in units (not shown) are available. Frequency Converter Units, Models 525A, B, and C, increase the frequency range from 10.1 to 100 mc, 100 to 220 mc, and 100 to 510 mc respectively. Video Amplifier unit 526A increases the basic set sensitivity to 10 mv in the range of from 10 cps to 10.1 mc; Time Interval unit 526B permits measuring time intervals from 1 μsec to 10^7 seconds; Period Multiplier unit 526C extends the period measurement range up to 10,000 periods of unknown frequency; and Phase unit 526D permits measuring phase angle with an accuracy approaching $\pm 0.1^0$. In addition to the plug-ins, the Model 540B Transfer Oscillator extends, as a companion instrument, the frequency range up to 12.4 megacycles (10^9 accuracy).

To measure frequency the basic circuit arrangement of the Electronic counter is shown in figure 3-7. For frequency measurement the signal is fed through a Signal Gate to a series of digital type counters. A precision time interval obtained from the Time Base Section opens and closes the Signal Gate for an extremely accurate period of time, for example, 1 second. The counters count the number of cycles entering through the gate during the 1-second interval and then display the total. The answer is read directly as the number of kilocycles occurring during the 1-second interval. The period of time the Signal Gate remains open is set by the FREQUENCY UNIT switch (not shown). For each position of the FREQUENCY UNIT switch the illuminated decimal point is automatically positioned so that the answer is always read directly in kilocycles. The answer is automatically displayed for a period of time determined by gate time or the setting of the DISPLAY TIME control on the front panel, whichever is greater.

To measure a period or time interval the application of the two signals reverses as shown by the dotted lines in figure 3-7. The period or time interval to be measured is connected to open and close the Signal Gate while one of the standard frequencies from the Time Base Section is passed through the Signal Gate to the counters. When measuring period, one cycle of the incoming signal opens the gate, the next cycle closes it. The number of cycles of the standard frequency from the Time Base that occurred during the period are then indicated on the counters. The standard frequencies obtained from the Time Base have been selected so that the answer to the measured period will always be displayed in direct-reading units of time: seconds, milliseconds, or microseconds.

Provision is also made in the circuit to permit measurement of the average of 10 periods
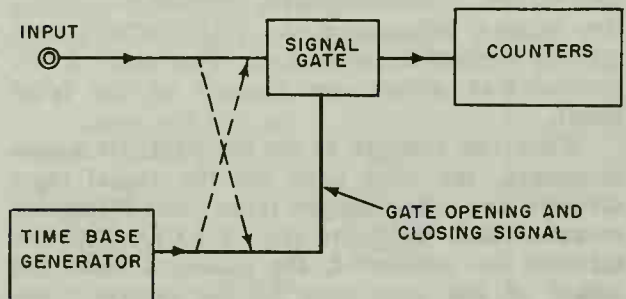
70.36

Figure 3-7.—Basic block diagram of the 524D.

of the unknown frequency. Higher accuracy can thus be obtained than with single period measurements.

The accuracy of frequency measurements is determined by an internal oscillator and by a possible error of ± 1 count that is inherent in the gate and counter type of instrument. At low frequencies, greater accuracy can be obtained by measuring the period of the signal than by measuring the frequency directly.

The block diagram (fig. 3-8, A) shows the circuit arrangement of the basic counter when measuring frequencies in the range of 10 cps to 10.1 mc. To measure frequencies up to 510 mc, one of three frequency converter units is required (fig. 3-8, B). As stated above, the 525C Frequency Converter unit is used between 100 and 510 mc. In these frequency converters the input signal is mixed with a harmonic of 10 mc so that the difference between the signal and the harmonic is not more than 10.1 mc. The difference frequency is counted and displayed. By adding the count displayed by the counter to the known 10 mc harmonic the input signal frequency is determined.

All three frequency converters have tuning systems to indicate the correct mixing frequency. However, if the mixing frequency is within 1 mc of the unknown frequency, there is a possibility of two answers, for you may not know whether to add or subtract the displayed reading from the mixing frequency. In such cases, make additional measurements using the two adjacent mixing frequencies to determine the unknown frequency. When making the final measurement choose a mixing frequency which is at least 100 kc away from the unknown.

When measuring frequency, the counter will count sine waves, rectangular waves, and positive pulses. To measure the frequency of negative pulses, adjustment of a FREQUENCY sensitivity control is necessary. This control is a screwdriver adjustment located on the front panel.

When the counter is set for PERIOD measurements, the time base and the signal input circuits are interchanged from their frequency measurement positions (fig. 3-8,C). With the circuits so connected, the counters count the output of the time base for the period of the unknown input signal. Thus the standard frequencies generated in the time base are used as units of time to measure the unknown period

in terms of microseconds, milliseconds, or seconds.

The accuracy of period measurements is largely determined by the accuracy with which triggering occurs at the same point on consecutive cycles of signal voltages having a slow rate-of-rise. Note that when the signal-to-noise ratio improves, the triggering accuracy also improves. Averaged over ten periods, the single-period error is reduced by a factor of ten. If you use the 526C Period Multiplier unit, the error is reduced an additional factor of ten for each factor of ten you extend the measurement. The accuracy of triggering is considerably improved when the waveforms being measured have a fast rise time. For example, you can obtain a significant reduction in error if you apply square waves instead of sine waves to the input.

In order to follow the slowest-changing waveforms, the period measurement input circuits are direct-coupled and are adjusted to trigger at the zero-volt crossing of a negative-going voltage. Thus any d-c component in the input signal will shift the triggering level so that the maximum slope no longer occurs at the zero-volt level, resulting in a loss of accuracy. If the d-c component is large enough, there may be no triggering at all. An external generator can be used in place of the time base generator for period measurements.

The counter can be used to measure the RATIO of two frequencies. The higher frequency is passed through the signal gate to the counters and is counted for a period of time determined by either one period or ten periods of the lower frequency, which controls the opening and closing of the gate (fig. 3-8,D).

Ratio measurement accuracy is determined by the same factors as period measurement accuracy: consistency of triggering by the lower input frequency and the inherent error of ± 1 count of the higher frequency. The 526C Period Multiplier unit is used to reduce the error by extending the number of periods of the lower frequency over which the measurement is made. For each factor of ten the measurement is extended, the error is decreased by a factor of ten.

Although the time base generator is not used during ratio measurements, you cannot make ratio measurements if the time base generator is not operating. The counter has a holdoff circuit which disables the signal gate if the time base generator fails.

A FREQUENCY MEASUREMENTS BELOW 10 MEGACYCLES

B FREQUENCY MEASUREMENTS ABOVE 10 MEGACYCLES

C PERIOD MEASUREMENTS

D RATIO MEASUREMENTS

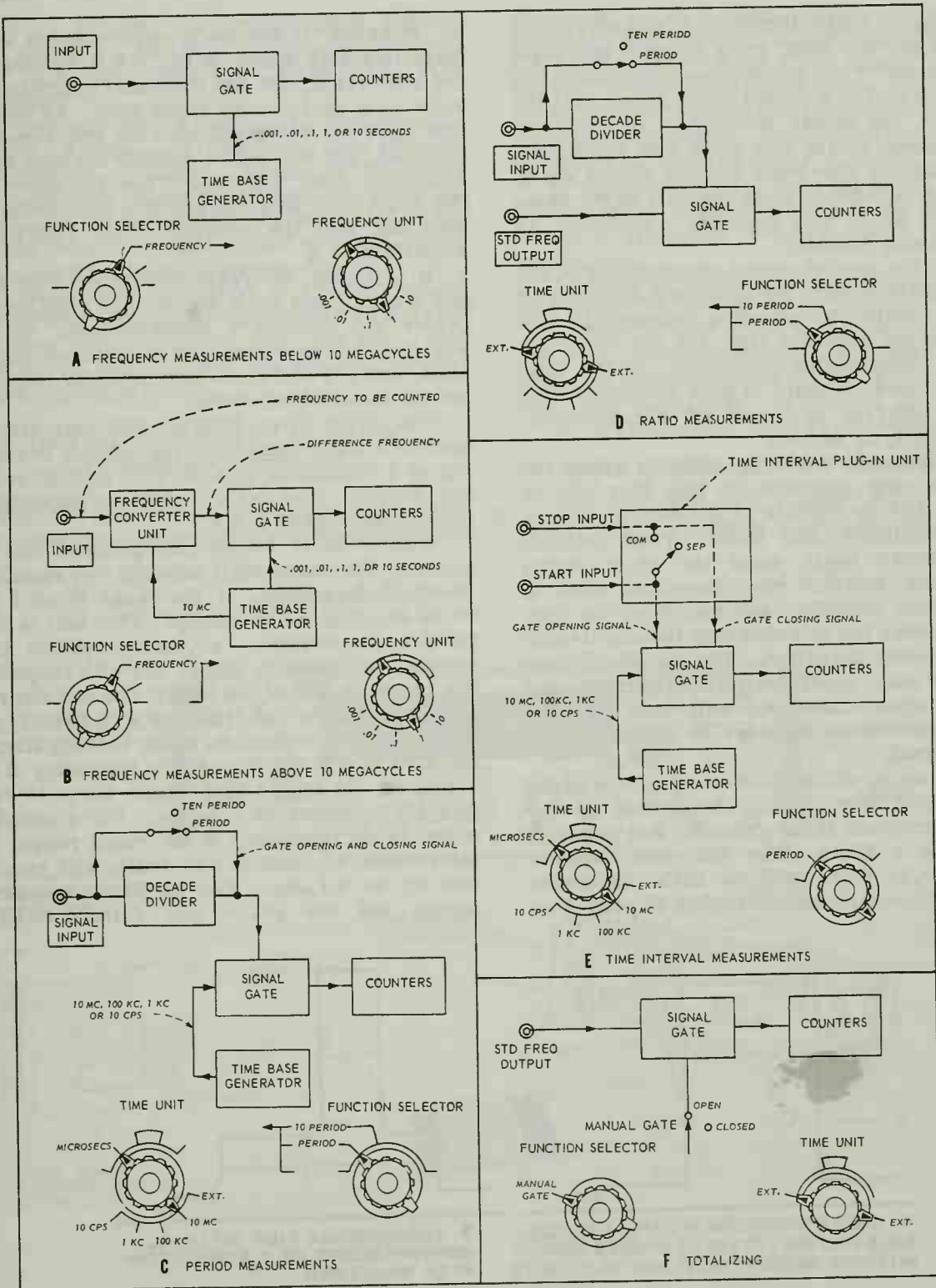**E** TIME INTERVAL MEASUREMENTS

F TOTALIZING

70.37

Figure 3-8.—Test measurement, block diagram.

To make TIME INTERVAL measurements (fig. 3-8,E), the 526B Time Interval unit must be installed. Time interval measurements are similar to period measurements except that the points on the signal waveforms at which the measurement starts and stops are adjustable. The adjustable threshold feature allows you to make measurements from one part of the same waveform or to use separate waveforms as start and stop signals.

As in the case of period measurements, the input signals control the opening and closing of the gate while the standard frequencies are passed to the counters (fig. 3-8,E). Thus the accurate frequencies generated in the time base are used as units of time to measure the unknown interval in terms of microseconds, milliseconds, or seconds.

The threshold-selecting controls adjust the start and stop channels so that they will be actuated only by signals of predetermined polarity, amplitude, and slope. Time interval measurements begin when the start signal crosses the selected start threshold value in the selected direction and end when the stop signal crosses the selected stop threshold value in the selected direction. The threshold controls are only approximately calibrated, and in some applications you will have to take special precautions in order to obtain the desired interval.

If you use an uncomplicated waveform as the start and/or stop signal, the setting of the threshold controls is not critical. For example, if you use a sharp pulse like that shown in figure 3-9,A, there will be little difference whether the measurement begins at point A or

B. However, if you use a more complex waveform like that shown in fig. 3-9,B, to measure the interval X, set the threshold controls near zero as a preliminary adjustment. As you adjust first the start and then the stop threshold controls, you will notice definite changes in the measured time interval. Thus you know that the start and stop thresholds are above the step and that the indicated time interval is actually X.

It is highly desirable to examine both start and stop signals on a d-c coupled oscilloscope before you attempt a measurement. In this way you can determine that no spurious signals exist, and you will know how carefully you must set the threshold controls.

The 526B Time Interval unit may also be used as a high-speed totalizer capable of counting at a maximum rate of 10.1 million events per second. The basic circuit arrangement is indicated in figure 3-8,F.

With a 526D Phase unit plugged into the counter, the phase angle between two signals of identical frequency, in the range from 1 cps to 20 kc, may be measured. This unit is useful for investigating, at various points in a circuit, the phase a signal has with respect to the phase it had at the input. Connect the reference signal to the REFERENCE INPUT, and the signal whose phase is under investigation to the UNKNOWN INPUT. If the frequency of the signal is 400 cps ± 4 cps, phase angle is read directly in tenths of a degree. For a signal of some other frequency in the rated range, the information is read in time units, with resolution up to 0.1 μsec. For all phase measurements, set the phase unit PHASE/PERIOD

A   SIMPLE WAVEFORM SHOWING TWO POSSIBLE VOLTAGE LEVELS AT WHICH A TIME INTERVAL MEASUREMENT MAY BE STARTED

B   TWO POSSIBLE TIME INTERVAL MEASUREMENTS OF A SINGLE COMPLEX WAVEFORM

70.38
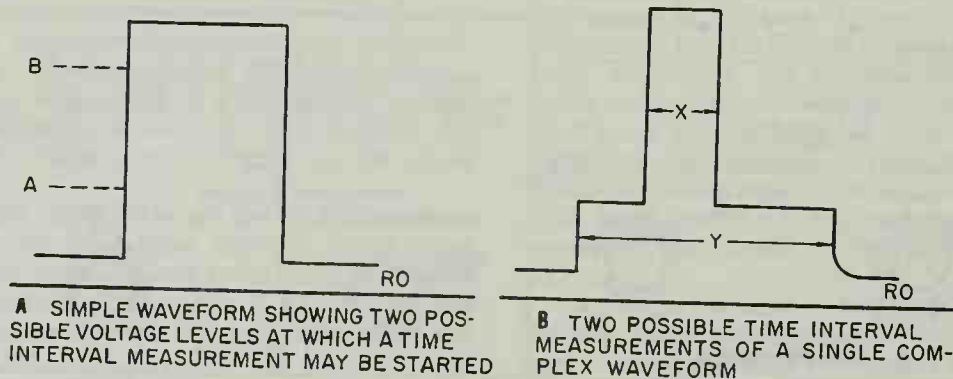
Figure 3-9.—Time interval waveforms.

switch to PHASE, the REFERENCE LEAD/LAG switch to the type of measurement desired, and the counter FUNCTION SELECTOR to PERIOD.

In general, circuit action for a phase measurement is similar to that for a time interval measurement. Trigger circuits in the Phase unit supply the pulses which open and close the signal gate in the counter. Arrangement of the circuits will be similar to that shown in figure 3-8,E, for time interval measurements.

A recommended method of TUNING RADIO RECEIVERS USING A FREQUENCY COUNTER has been included in the EIB, No. 569. This method will soon become the accepted procedure for all such tuning.

DIGITAL READOUT ELECTRONIC COUNTER AN/USM-207

The AN/USM-207 is a portable, solid-state electronic counter for precisely measuring and

displaying on an 9-digit numerical readout the frequency and period of a cyclic electrical signal, the frequency ratio of two signals, the time interval between two points on the same or different signals, and the total number of electrical impulses (totalizing). The counter also provides the following types of output signals:

1. Standard signals from 0.1 cps to 10 mc in decade steps derived from a 1 mc frequency standard, frequency dividers, and a frequency multiplier;

2. Input signals divided in frequency by factors from 10 to 10^8 by a frequency divider;

3. Digital data of the measurement in four-line binary-coded-decimal form with decimal point and control signals for operation of printers, data recorders, or control devices; and

4. A 1 mc output from a frequency standard.

GENERAL DESCRIPTION OF AN/USM-207

The AN/USM-207 (figure 3-10) consists of a major counter assembly, two plug-in

162.7

Figure 3-10.—Digital readout electronic counter AN/USM-207.

assemblies which install in recesses on the front and rear panel, and a group of accessory cables and connectors stored in the detachable front cover.
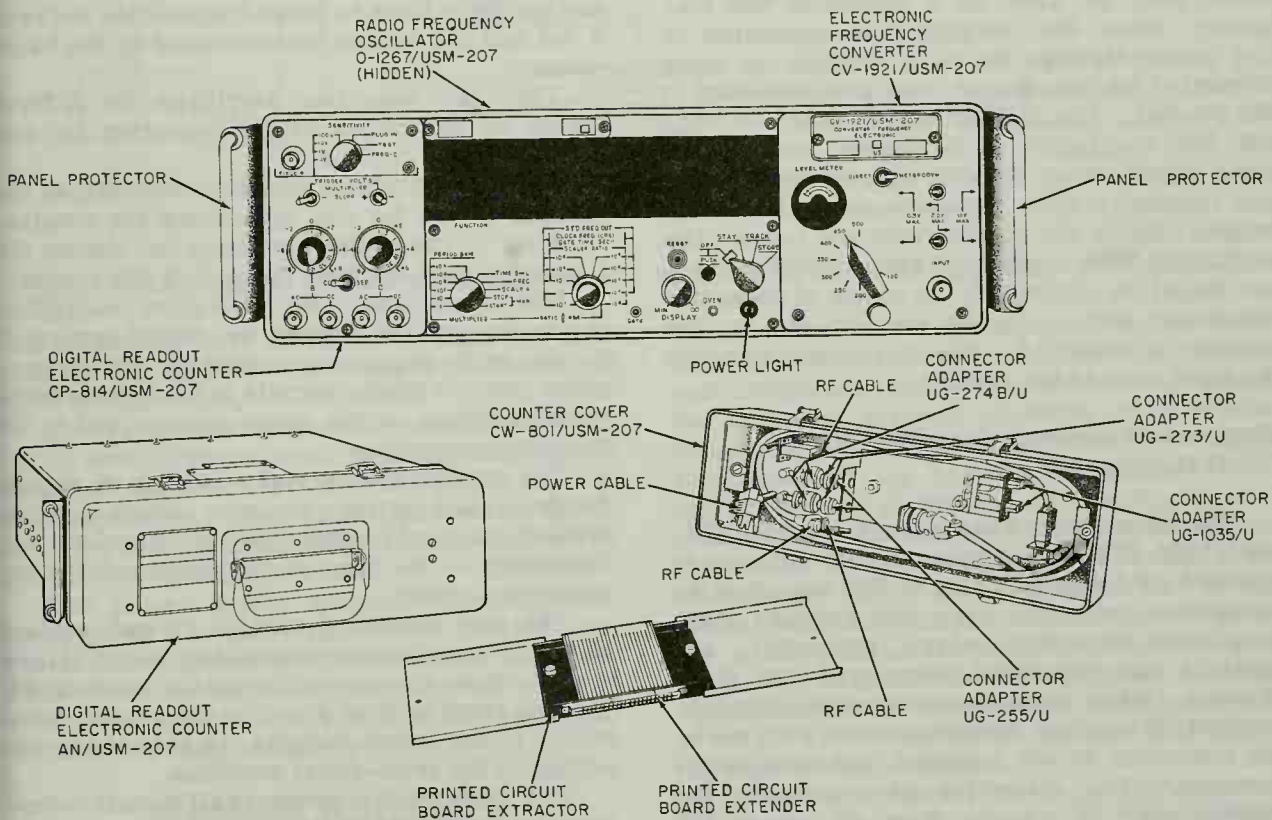
Digital Readout Electronic Counter CP-814/USM-207

The major assembly Digital Readout Electronic Counter CP-814/USM-207 contains the input amplifiers; gate control; display; reset and transfer control; frequency multipliers; time base dividers; decade and readout boards; numerical display tubes; decimal point and units indicators; power supply and regulator; and controls associated with these circuits.

Radio Frequency Oscillator O-1267/USM-207

The Radio Frequency Oscillator O-1267/USM-207 plug-in assembly develops a 1 mc signal and includes its own power supply. The oscillator includes the 1 mc output receptable which may be used as a source of that frequency when the oscillator is connected to a-c power through the basic counter or when connected to the power line independently of the counter. The counter may be operated without the oscillator in totalizing, scaling the input signal, time interval with external clock, and frequency ratio measurements. For other measurements the counter does not require the oscillator when a separate external 100 kc or 1 mc signal is connected. In either of these two situations the oscillator may be left in the counter or removed. The oscillator plugs into the right rear of the counter.

Electronic Frequency Converter CV-1921/USM-207

The Electronic Frequency Converter CV-1921/USM-207 plug-in assembly permits measurement of frequencies up to 500 mc using the heterodyne principle. The unit consists of the broadband amplifier, mixer, multiplier, and controls and indicators associated with these circuits. When measurements other than heterodyne frequency measurements are made, the converter is not required, but need not be removed. The converter also permits the measurement of signals from 35 mc to 100 mc with a greater sensitivity than available with the basic counter. The converter plugs into the right front of the counter.

FUNCTIONAL DESCRIPTION

Figure 3-11 is the overall functional block diagram of the counter. To make a measurement, requires two types of information; a count signal, and a gate control signal. These two signals may be generated within the instrument or they may be supplied from outside sources. The type of measurement the counter will make depends upon the relationship of these two signals. In any function the instrument counts the count signal for a period of time determined by the gate control signal. Routing of these signals within the instrument is accomplished by logic circuits. These logic circuits are controlled by means of the front panel controls.

The radio frequency oscillator (O-1267/USM-207) generates a signal of precise frequency for use throughout the counter or to provide a precise 1 mc signal for use outside the equipment.

The electronic frequency converter accepts radio frequencies between 100 mc and 500 mc and converts them to radio frequencies between 5 mc and 100 mc for measurement by the basic counter.

The "A" amplifier amplifies the A input signal or the output of the converter for use throughout the counter.

The "B" amplifier amplifies and shapes the B input signal for use throughout the counter.

The "C" amplifier amplifies and shapes the C input signal for use throughout the counter.

The 10 mc and 1 mc multiplier multiplies the frequency and shapes the signal generated by the radio frequency oscillator. It also provides precise timing signals to the various functional sections of the basic counter and to the frequency converter.

The scaler consists of a series of decade dividers and gating systems which provide divided standard frequencies and control signals depending on the type of measurement the instrument is making.

The gate control generates the gate control signal. This signal determines the length of time that the count decades will count the count signal.

The count control provides the proper count signal to the count decades, as selected by the setting of the front-panel switches.

The cycle control produces all signals necessary to display the measurement results on the readout and to recycle the counter.

The count decades count the count signal when permitted to do so by the gate control.
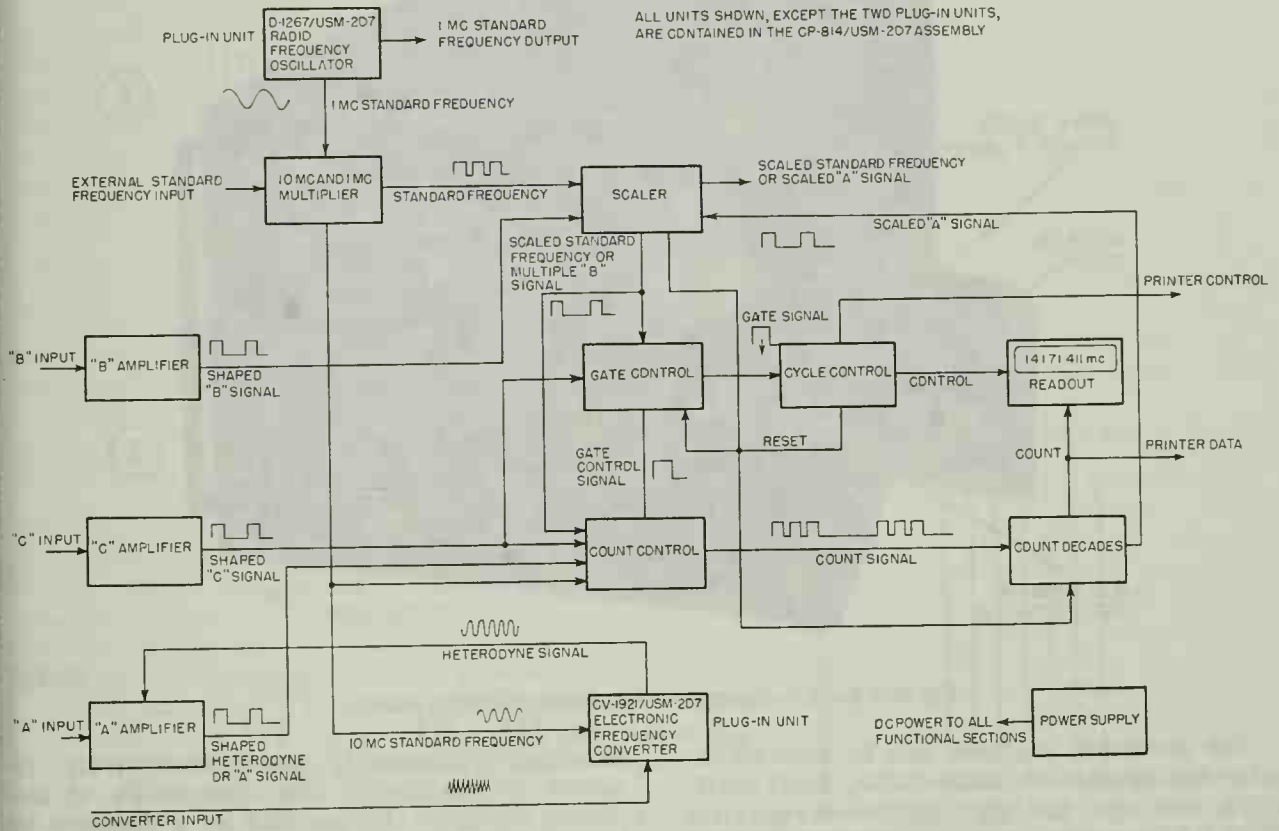
162.8

Figure 3-11.—Digital readout electronic counter AN/USM-207 overall functional block diagram.

The result of their counting becomes the final reading displayed by the readout at the end of each measurement.

The readout receives binary-coded-decimal (BCD) data from the count decades, decodes this data into decimal form and drives the readout indicator tubes. The readout also contains memory circuits which function when the counter is operated in the "Store" mode.

The power supply supplies all d-c power required by the basic instrument and the converters and consists of seven d-c supplies. Five of these supplies (+18 volt, +12 volt, +6 volt, and -12 volt) are regulated and two (+180 volt and +45 volt) are unregulated.

SPECTRUM ANALYZER

A spectrum analyzer is a device that sweeps over a band of frequencies to determine (1) what frequencies are being produced by a specific circuit under test and (2) the amplitude of each frequency component. To accomplish this, the spectrum analyzer presents, on an oscilloscope display, a pattern in which the relative amplitudes of the various frequencies of the spectrum are plotted on the vertical, or Y axis, while the frequencies themselves are plotted on the horizontal, or X axis, of the cathode-ray tube. The overall pattern of this display indicates the proportion of power present at the various frequencies within the spectrum. A representative spectrum analyzer equipment is the AN/UPM-33, which consists of the TS-148/UP Spectrum Analyzer with carrying case and accessories.

The control panel of the TS-148/UP spectrum analyzer is shown in figure 3-12. The circuits of this unit will be discussed in this section since they represent to a great extent spectrum analyzer circuits in general.

WAVE GUIDE
CHOKE FLANGE

ACCESS
DOOR

A.C. POWER
RECEPTACLE

Figure 3-12.—Spectrum analyzer control panel.

20.343

The spectrum analyzer can be used to examine the spectra of magnetrons, local oscillators, test sets, and other equipment operating within its frequency range. The frequency meter is used to measure accurately the frequencies of any of the above equipments, or to set the frequency of radar and beacon local oscillators in radar sets. The spectrum analyzer can also be used as an f-m oscillator to tune T/R boxes and ATR boxes in transmitters.

COMPONENTS OF A SPECTRUM

A common conception of the output of a pulsed oscillator is a single frequency, which is turned on and off for periods of standard duration. This action is similar to the output of a conventional c-w telegraph transmitter. The output of the pulsed radar oscillator does not consist of a fundamental frequency that is turned on and off, but must be considered as a fundamental frequency that is pulse modulated by the waveform of the trigger pulse.

Effects of Modulating Pulses

When modulated, any fundamental frequency will produce a fundamental frequency with sideband frequencies, which collectively is called a spectrum. The distribution of the power on these frequencies is a function of the modulation. Normally, modulation is plotted on an amplitude and time basis as shown in figure 3-13B, C, and D.

Assume F to be the fundamental frequency of the oscillator. Waveform A shows the fundamental frequency plotted as amplitude against time. The number of periods occurring within one second determines the frequency of the oscillation. The amplitude is represented as proportional to the distance between the negative and positive peak of one cycle.

In a spectrum pattern, this same frequency and amplitude would be represented as F in figure 3-13, E. Points along the horizontal axis represent frequency, which increases from left to right, while distance along the vertical axis above the baseline represents amplitude. Thus, two methods of diagramming the results of amplitude-modulating a carrier frequency are shown.

Assume that F_p1 is a modulation frequency applied to the fundamental frequency, F. This is normally represented on an amplitude-versus-time basis. This same type of modulation is shown by two lines on the spectrum pattern.
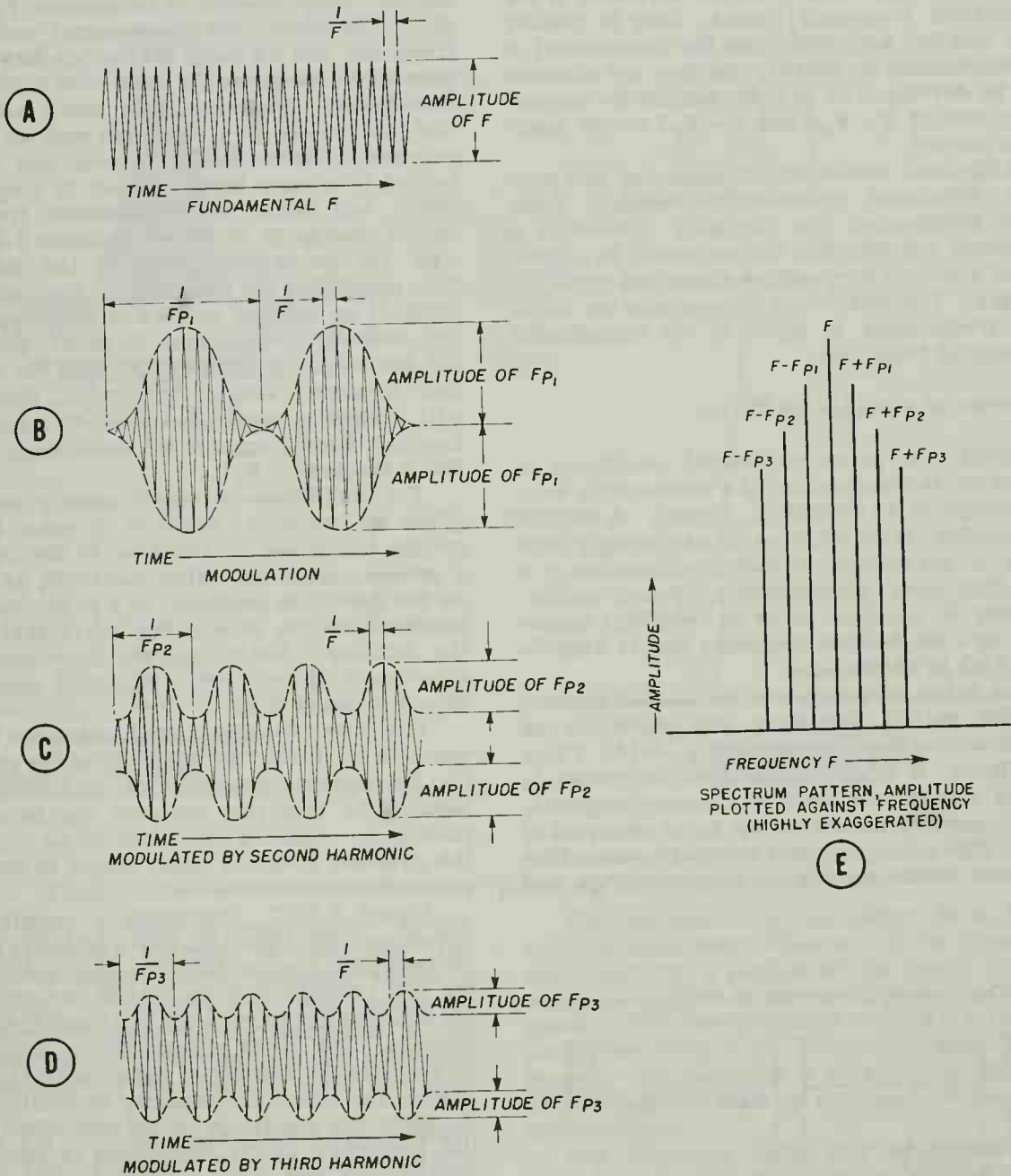
92.136

Figure 3-13.—Comparison of amplitude versus time to amplitude versus frequency.

These lines are marked F+Fp1 and F-Fp1 in figure 3-13,E. The reason for this is that the modulated wave actually represents the results of heterodyning two different frequencies. These frequencies are effectively present in the mod-ulated output, and can be detected by suitable receivers. One of the frequencies is the sum of the two, and the other is the difference. The amplitudes of the new frequencies are each half the amplitude of the modulating frequency.

67

Assume also that a second harmonic of the modulating frequency exists. This is usually of a smaller amplitude than the fundamental of the modulating frequency. Another set of waves will be developed as at C for amplitude-versus-time, and at $F + F_p2$ and $F - F_p2$ on the spectrum pattern.

Additional modulating frequencies will produce additional sideband frequencies. Since these frequencies are normally present in a harmonic relationship, the net result is a number of different frequencies above and below the carrier. The difference between any two adjacent frequencies is equal to the fundamental modulating frequency.

Spectrum of a Pulsed Oscillator

Usually a pulse modulated oscillator is pulsed by the application of a rectangular wave of voltage to an oscillator circuit. A narrow rectangular wave contains an exceedingly wide range of harmonics, including harmonics of a very high order. Consequently, a pulsed oscillator may be assumed to be an oscillator modulated by a modulation frequency that is exceedingly rich in harmonics.

The pulse frequency and the basic frequency of the pulsed oscillator are shown on an amplitude-versus-time graph (fig. 3-14). From this figure, it might appear that the output is simply a pure c-w wave of constant amplitude and frequency, turned on for brief intervals of time. This concept cannot be true because of the presence of the modulation frequency F_p1, and
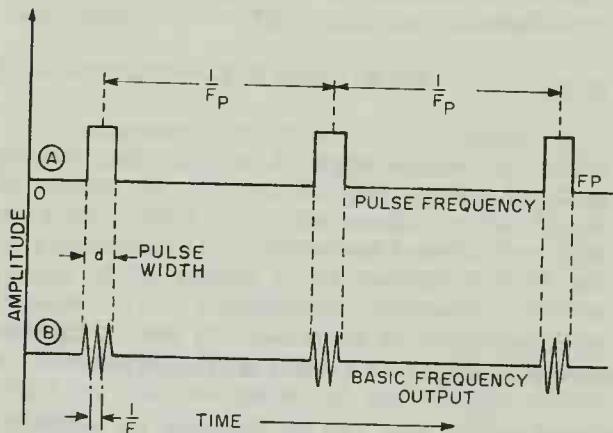
the very large number of harmonics, F_p2, F_p3, etc. Therefore, the fundamental modulation frequency and its many harmonics may be considered to modulate the oscillator to produce a fundamental frequency with many sidebands. The net result is a spectrum such as was developed in figure 3-13,E, except that it is extended to a very large number of frequencies above and below the fundamental frequency. Such a spectrum is shown in figure 3-15 along with the pulse that produced the spectrum. It is shown that the output of the oscillator consists of an infinite number of lines representing different frequencies. However, because of the harmonic relationship between the modulation frequency and its harmonics, these lines will always be separated by a distance on the baseline that is equal to the fundamental modulation frequency, F_p1.

The amplitude-versus-frequency plot provides an envelope, which is of value in estimating the power distribution in the output of a pulsed oscillator. This spectrum, as plotted on the spectrum analyzer, is a power spectrum produced by the square law characteristics of the detector. The true power spectrum, which represents the voltage (amplitude) squared is shown in figure 3-15.

The power spectrum emphasizes the importance of confining the majority of the power at the fundamental frequency of the pulsed oscillator. The spectrum analyzer can be used in turning up a pulsed oscillator so as to provide the greatest range of power output in the bandpass circuits of a receiver.

Figure 3-15,C, represents a sample of the spectrum. The figure shows the effective voltage derived from the oscillator frequency spectrum. We can conclude that the pulsed oscillator output contains both frequency and amplitude components. The effect of amplitude modulation is to increase the number of sidebands in the spectrum. The effect of frequency modulation is to increase the amplitude of the side lobes, since the frequency of the oscillation is effectively shifted back and forth between points on each side of its true fundamental frequency.

BLOCK DIAGRAM

Basically, the spectrum analyzer consists of a superheterodyne receiver with a frequency-modulated, r-f oscillator (fig. 3-16). The analyzer r-f oscillator is modulated with a sawtooth voltage, which causes the oscillator to

Figure 3-14.—Output of a pulsed oscillator.

92.137

OSCILLATOR FREQUENCY

REPETITION FREQUENCY = fp

MINOR LOBE

FIRST ZERO

$\frac{3}{d}$ $\frac{2}{d}$ $\frac{1}{d}$ O $\frac{1}{d}$ $\frac{2}{d}$ $\frac{3}{d}$

(A) THEORETICAL VOLTAGE SPECTRUM OF A PULSED OSCILLATOR WHEN
d = PULSE WIDTH IN TIME
fp = PULSE FREQUENCY
E = PULSE VOLTAGE

d
E
$\frac{1}{fp}$
TIME

(B) WAVEFORM OF PULSE VOLTAGE

(C) EFFECTIVE VOLTAGES DERIVED FROM ABOVE SPECTRUM

FREQUENCY INCREASING

(D) ACTUAL POWER (E²) SPECTRUM ENVELOPE OF SPECTRUM (A) (PLOTTED TO DIFFERENT SCALE) AS SEEN ON TS-14B/U

92.138

Figure 3-15.—Relationship between a voltage spectrum and a power spectrum.

operate over a range of frequencies. As the frequency of the analyzer r-f oscillator increases, it beats with the incoming signal to produce an i-f signal for various frequency components present in the spectrum of the received signal. These signals are presented on a cathode-ray oscilloscope as the spectrum pattern.

The r-f input is applied to a variable attenuator. The attenuator controls the spectrum amplitude, and consists of two carbonized resistance cards inserted in the r-f input end of the waveguide. Varying the contour of these cards varies the degree of attenuation. Maximum attenuation is obtained when the cards are flexed together in the center of the waveguides. The attenuator controls the amplitude of the r-f signal either entering the analyzer, or leaving the analyzer when it is used as a signal generator.

The analyzer r-f oscillator is a reflex velocity-modulated klystron. It is frequency modulated by a portion of the sweep voltage, which is applied to the oscilloscope deflection plates. The sweep voltage creates a maximum frequency swing in the oscillator output of 40 to 50 mc. The analyzer r-f oscillator output is matched to the load by the fixed attenuator in the waveguide.

The incoming signal and the output of the r-f oscillator are mixed at the crystal mixer. The output is applied to the input circuit of the tuned i-f amplifier section to produce the intermediate frequency of 22.5 mc.

An absorption-type frequency meter is mounted in the waveguide at a point between the crystal and the r-f oscillator. The meter consists of a cavity with dimensions that can be varied by a front panel control. Each time the frequency of the r-f oscillator passes through
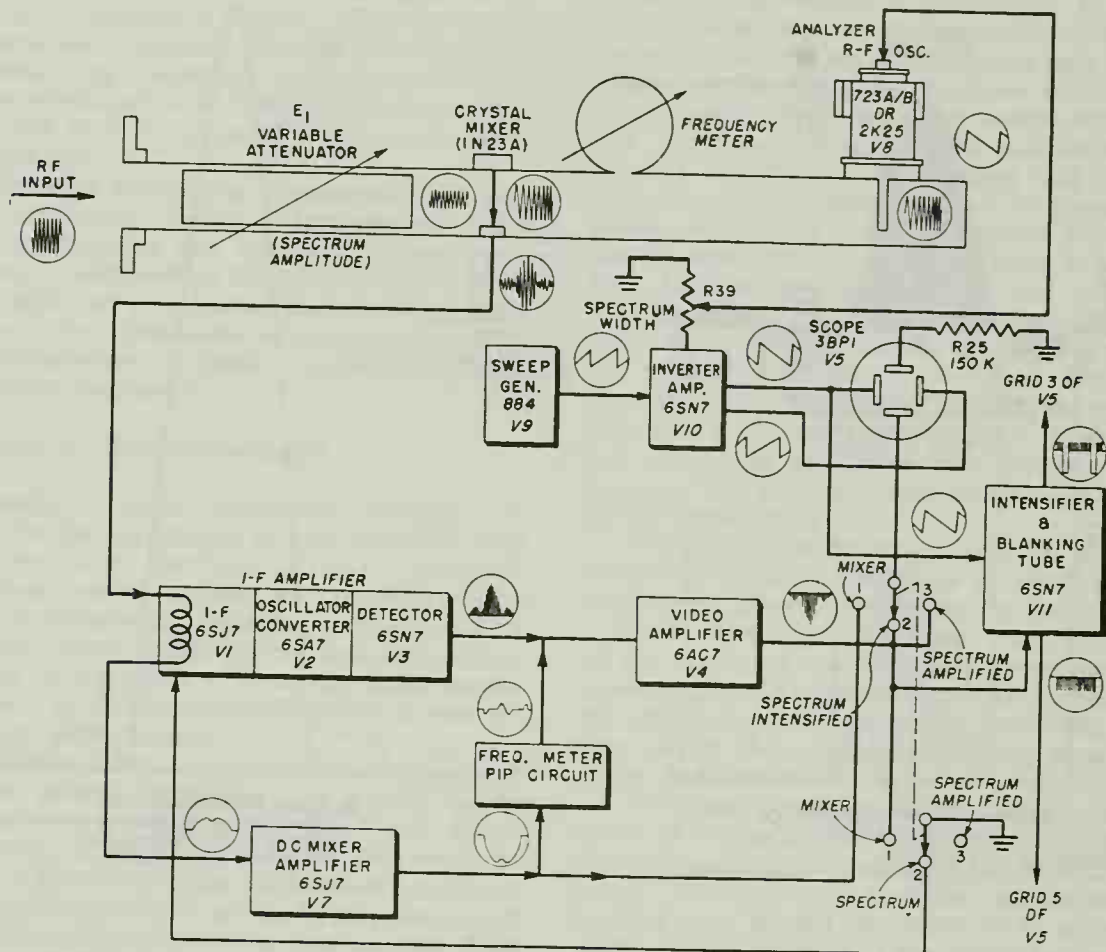
92.139

Figure 3-16.—Spectrum analyzer, block diagram.

the resonant frequency of the frequency meter, the meter absorbs some of the r-f power. This causes a sharp decrease in the amplitude of the output of the r-f oscillator that is applied to the crystal. The change in amplitude is caused to appear on the scope as a sharp pip, which is used as a frequency marker. The pip occurs when the frequency of the r-f input coincides with the frequency of the meter.

The output of the crystal mixer is applied to the input circuit of the 22.5-mc intermediate frequency amplifier. The output of the i-f amplifier section is applied to the oscillator-converter.

The oscillator-converter contains a local oscillator operating at a frequency of 19.5 mc. The intermediate frequency is changed to three megacycles in this stage by beating the 19.5-mc frequency against the 22.5 mc from the crystal mixer. The sharply tuned plate circuit of the oscillator-converter produces a 50-kc circuit bandpass. The output is applied to the detector stage.

The detector section functions as an infinite impedance detector. The rectifier output of the detector is applied to the video amplifier.

The video amplifier stage has two inputs. One is the output of the detector, and the other the frequency meter pip obtained from the d-c amplifier through a differentiating circuit. The output of the video amplifier is coupled to the vertical deflecting plates of the oscilloscope. A portion of the video output is used to trigger the intensifying stage.

70

When the selector switch on the front panel is in the MIXER position (position 1), the d-c mixer amplifies the output of the crystal mixer and applies it through the selector switch to the vertical plates of the cathode-ray tube for test purposes. In the other two positions of the selector switch, the output of the d-c mixer is applied to the video amplifier. The end pips of the waveshape appearing at the video amplifier grid are derived by differentiation in the plate circuit of the d-c mixer amplifier. The center pip corresponds to the point of resonance of the frequency meter.

The sweep generator is a gas tube relaxation oscillator that generates a sawtooth output at frequencies between 10 and 30 cps. The output of the sweep oscillator is used to drive the sweep inverter amplifier.

The sweep inverter amplifier provides push-pull deflection voltages for the horizontal plates of the cathode-ray tube. It also supplies modulating voltage to the analyzer r-f oscillator and trigger voltage for the blanking section.

When the selector switch is placed in either of the SPECTRUM positions (positions 2 or 3 on the front panel), the video amplifier signal triggers the intensifier tube. The output thus produced is coupled to the intensifying grid of the cathode-ray tube.

The input voltage from a section of the sweep inverter amplifier is used to trigger the blanking tube. The output of this tube is applied as blanking voltage to eliminate the return trace that would otherwise appear on the scope screen.

CIRCUIT OPERATION

The attenuator, analyzer r-f oscillator, frequency meter, and crystal mixer circuits of the spectrum analyzer are discussed schematically due to the unique manner in which they have been assembled. The more common oscilloscope circuits such as the sweep generator, and the intensifier, video, sweep inverter, and blanking and d-c mixer amplifiers have been omitted. Also this receiver section including the i-f, oscillator-converter, and detector circuits are not treated further because they have been covered in Basic Electronics NavPers 10087-B.

Attenuator

A schematic diagram of the waveguide assembly circuit and the analyzer oscillator is shown in figure 3-17. Radio frequency signals from an outside source enter the waveguide and pass through the variable attenuator to the
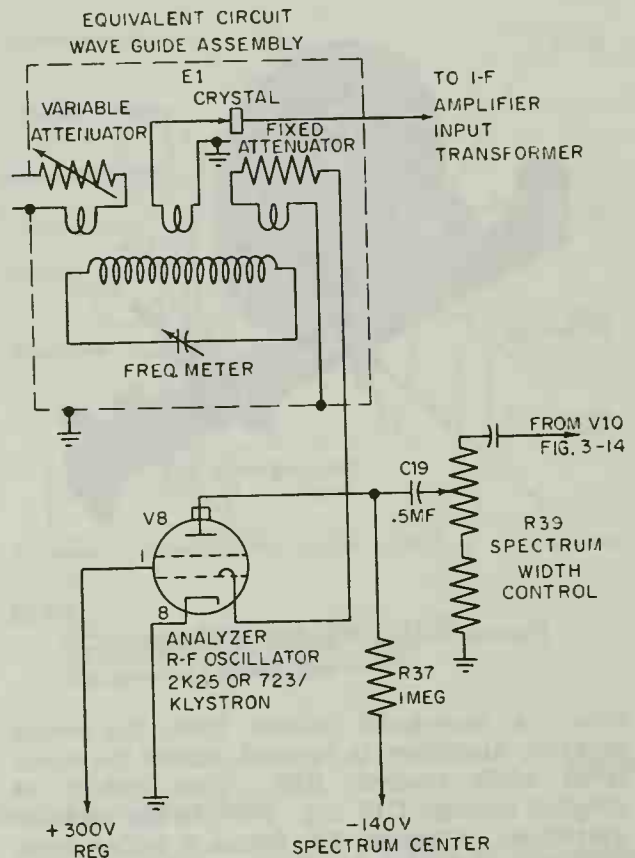
109.85

Figure 3-17.—Wave guide assembly and analyzer oscillator, schematic.

crystal mixer. The attenuator (fig. 3-18) controls the amount of signal voltage that reaches the mixer, and therefore controls the amplitude of the pattern on the screen. The control knob for the variable attenuator is marked, SPECTRUM AMPLITUDE, on the front panel.

The attenuator consists of two carbonized coated resistance strips, which lie flat against the walls of the waveguide. The ends are fastened down but still allow the strips to slide slightly as they are bowed into the center of the guide. As the strips move inward, they intersect more and more of the r-f energy and thus cause the attenuation to increase. The range of the attenuator is from 70 db to 3 db.

Analyzer R-F Oscillator

The analyzer r-f oscillator, V8, uses a Type 2K25 or 723A/B Shepard-Pierce reflex klystron

FREQUENCY METER

CRYSTAL HOLDER

DIAL ASSEMBLY

RIGHT HAND SCREW

X115

STOP ASSEMBLY

ATTENUATOR SHAFT

FIXED ATTENUATOR

WAVEGUIDE

ADJUSTABLE PLATE

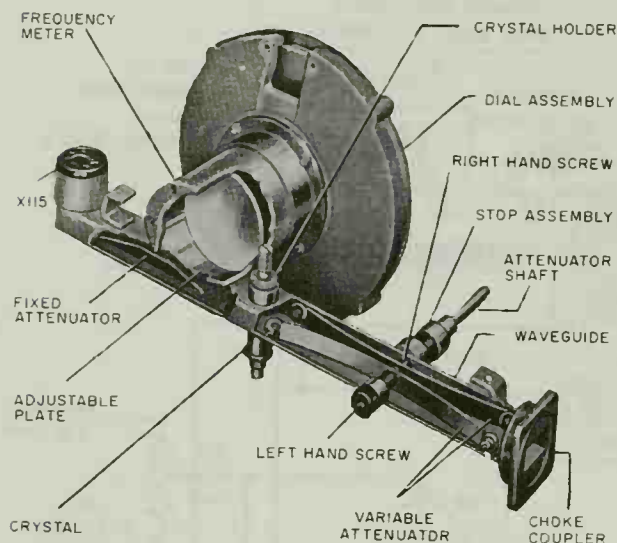LEFT HAND SCREW

CRYSTAL

VARIABLE ATTENUATOR

CHOKE COUPLER

109.86

Figure 3-18.—Waveguide assembly, cut-away view.

tube. A saw-tooth voltage from the sweep inverter amplifier is applied across the spectrum width control, R39. This voltage is coupled through C19 (fig. 3-17) to the repeller electrode. Klystron V8, forms a reflex-type, variable-cavity oscillator. The tube contains an electron gun to emit electrons into the cavity. The cavity is maintained at +300 volts regulated. The repeller electrode is connected to a -140 volt source through R37.

The electrons from the cathode are accelerated as they pass through the positive grids. The a-c potential of the grids (caused by the movement of electrons in and around the grids) changes at the frequency of the oscillator. If a group of electrons arrive at the first grid when it is positive with respect to the second, the electrons are accelerated and pulled away from the electrons that immediately follow. If the second grid is still negative when the group of electrons reach it, the group will be slowed and bunched closer together.

If the second grid is not negative, the electrons pass through it to the negative repeller where they are repelled back into the cavity. If the electrons arrive back at the cavity in phase with the a-c potential of the cavity, they reinforce the bunching effect, giving up energy to the cavity.

The result of this action is a velocity-modulated stream of electrons flowing in the

cavity. As the bunched electrons pass through each grid, it becomes alternately positive and negative. The frequency of this oscillating action is determined by the spacing between the grids, the volume and shape of the cavity, the space between the repeller and the cavity, and the voltage on the repeller. Assuming that the d-c potentials on the repeller and the cavity are constant, the voltage and spacing of the repeller act to control the transit time of the electrons, which fixes the frequency of the oscillation.

The analyzer r-f oscillator may be tuned in either of two ways. One way is by means of a mechanical screw (fig. 3-19). The screw, operated by the oscillator frequency control on the front panel, places a stress on a pair of struts on the side of the tube. As the struts are lengthened they expand the cavity to increase the spacing between the two grids and vary the capacity in the resonant cavity.

The second method used to tune the analyzer r-f oscillator involves varying the voltage on the repeller electrode. The spectrum center knob on the front panel is a fine frequency control, which varies the d-c potential on the repeller. The spectrum center potentiometer is not shown on the schematic. Actually, the klystron will only operate within definite ranges of repeller voltage. The ranges are called modes. The analyzer r-f oscillator normally has a d-c potential of -140 volts applied to the repeller electrode, and makes use of the modes that appear in the -100 to -180-volt range.

It is necessary to frequency modulate the analyzer r-f oscillator V8 in order to display on the scope screen each frequency component of the signal under test. The pulse coupled into the waveguide through the variable attenuator contains many frequencies. The klystron frequency is caused to vary by sweep modulation. The two varying frequencies, when heterodyned at the crystal mixer, will produce a 22.5-mc output at various times or at various points along the frequency sweep of the r-f oscillator. Figure 3-20 shows the spectrum, and how the two frequencies are heterodyned to produce the spectrum components.

The signal at the beginning of the sweep represents the lowest frequencies in the lower sideband (fundamental frequency minus the higher order harmonics) and is followed in turn by the higher frequencies in the lower sideband (fundamental frequency minus the higher order harmonics) until the fundamental frequency is reached. The succession of i-f signals continues
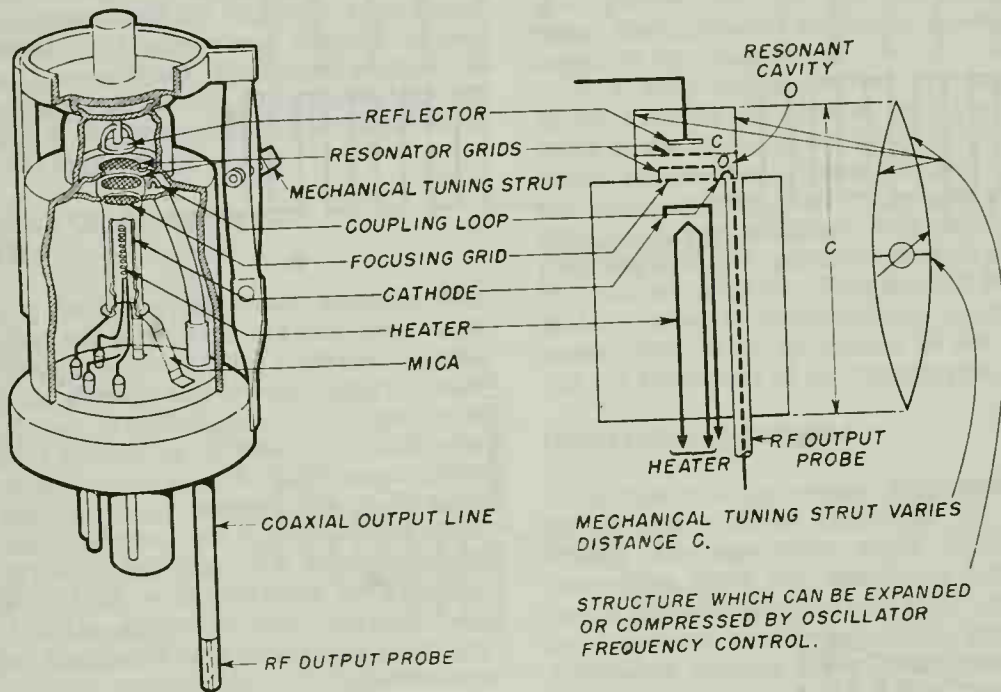
109.87

Figure 3-19.—Analyzer r-f oscillator tube.

through the upper sidebands until the highest frequency is reached. Since the trace on the scope is synchronized in time with the frequency swing of the analyzer r-f oscillator, the sideband frequencies will be spread out from left to right as shown in figure 3-20.

The output of the analyzer r-f oscillator is passed from the coupling loop inside the cavity of the klystron, V8 (fig. 3-17), through a fixed attenuator, to the crystal mixer. The fixed attenuator isolates the analyzer r-f oscillator from the rest of the waveguide so that it is unnecessary to make any adjustments on the mixer when tuning over the frequency range. A portion of the oscillator output is used by the crystal in the heterodyning process. The portion not used in this process moves past the variable attenuator and out into the waveguide. This portion of the r-f is of particular importance when the analyzer is used as a signal generator.

Frequency Meter

The power from the r-f oscillator goes past the frequency meter where a small portion of the energy is absorbed by the cavity of the meter.

The frequency meter is a high Q cylindrical cavity-type meter with a mechanically variable length. The frequency range of the meter is 8470 to 9630 mc.

Each time the frequency of the analyzer r-f oscillator sweeps through the frequency to which the frequency meter is tuned, the meter absorbs some of the r-f power. The absorption results in a sharp reduction of the analyzer r-f oscillator component of the crystal current. This reduction, or pip, is applied to the d-c mixer amplifier and is used to indicate the coincidence of the frequency meter with unknown frequencies on the scope screen.

Crystal Mixer

The crystal mixer is used to rectify the heterodyne frequency produced by mixing the f-m output of the analyzer r-f oscillator with the r-f input signal in the waveguide. When the input signal is 22.5 mc above or below the frequency of the analyzer r-f oscillator, the i-f amplifier responds to the output of the crystal.

The crystal is located between the frequency meter and the variable attenuator (fig. 3-18).

73

PULSES UNDER TEST

8883.0MC 8884.0MC 8885.0MC 8886.0MC 8887.0MC

9023.0MC 9024.0MC 9025.0MC 9026.0MC 9027.0MC

1 ms

MODULATING SAWTOOTH

FUNDAMENTAL

8860.5MC 8861.0MC 8861.5MC 8862.0MC 8862.5MC 8863.0MC 8863.5MC 8864.0MC 8864.5MC

9000.5MC 9001.0MC 9001.5MC 9002.0MC 9002.5MC 9003.0MC 9003.5MC 9004.0MC 9004.5MC

R-F OSCILLATOR
FREQUENCY SWING
44.0MC

LOWER SIDEBANDS

UPPER SIDEBAND

E out →

$\frac{1}{1000}$
SEC

SPECTRUM COMPONENTS
$F - F_{Pn}$ etc.

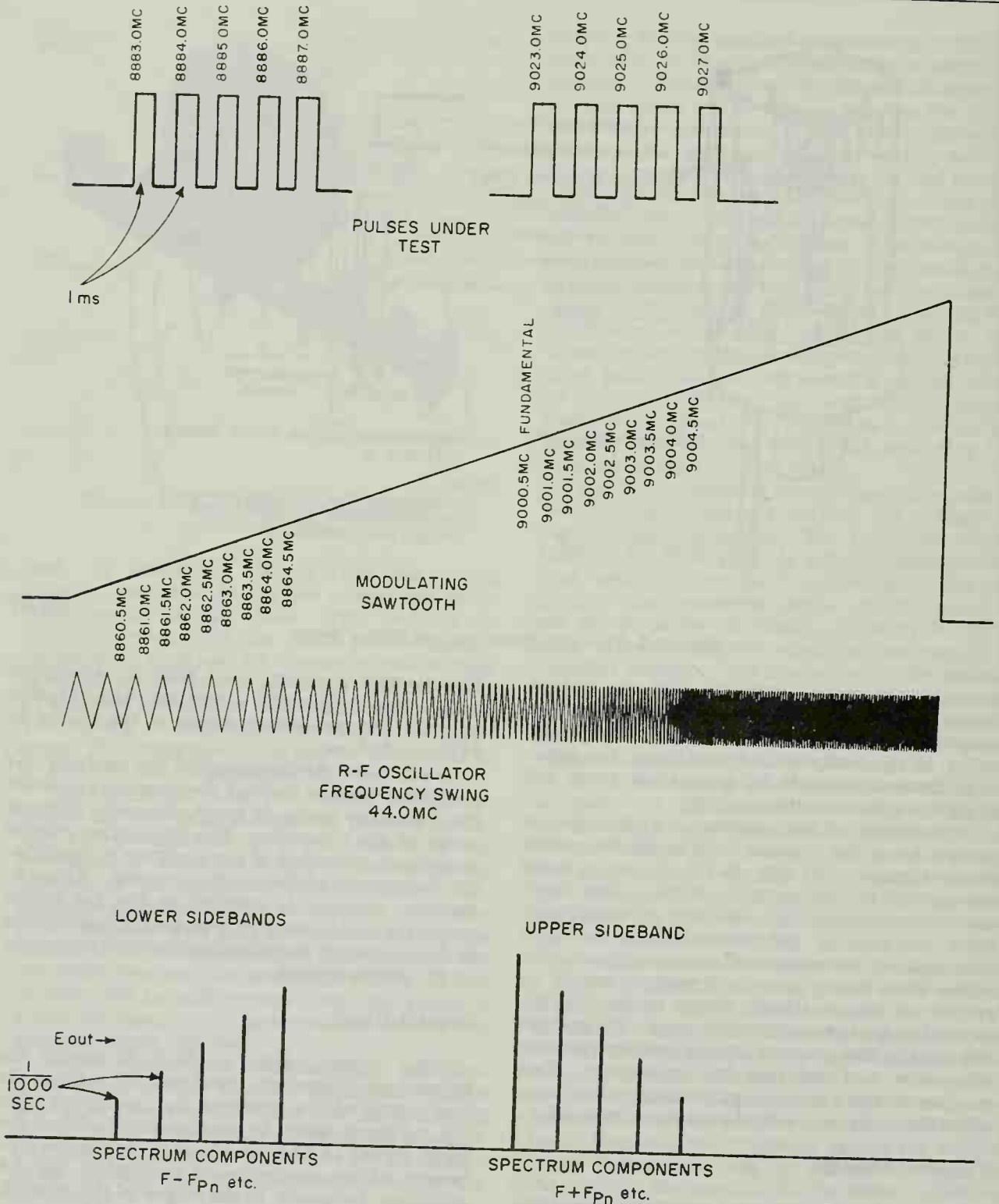SPECTRUM COMPONENTS
$F + F_{Pn}$ etc.

Figure 3-20.—Method of producing spectrum frequencies.

109.88

The current through the crystal represents the rectified components of the signals. The output of the crystal is transformer coupled to the input of the i-f amplifier. The signal contains a rectified component that is proportional to the power output of the analyzer r-f oscillator with respect to the frequency.

PROCEDURE FOR OBSERVATION OF SPECTRA

Assuming that the spectrum analyzer has been installed and turned on, the selector switch should be placed in the MIXER position. In this position of the switch, adjustments may be made on the analyzer r-f oscillator. The spectrum width control should be turned in a clockwise direction until its setting is nearly maximum.

After all necessary adjustments have been made on the analyzer, and the desired width of the scope trace is set, place the selector switch in the SPECTRUM or SPECTRUM AMPLIFIED position. Use the spectrum amplified position if there is no likelihood of interference from any adjacent radar equipment. If interference occurs, spectrum position should be used.

Point the waveguide, which is accessible through the side door shown in figure 3-18, at the radar antenna or to the magnetron output. Coupling the waveguide too close to the source will burn out the attenuator and crystal mixer in the analyzer. Likewise, if the coupling is too close, the probability of magnetron pulling is increased.

An antenna horn and fittings (component parts of the TS-148/UP) may be used if the former method does not supply satisfactory results. This method can be used to pick up the output of the magnetron at distances up to 100 ft if the two antennas are pointing directly at each other.

The operator should never attempt to increase the amplitude of the spectrum trace by increasing the coupling until it is ascertained that the amplitude cannot be increased by adjusting the spectrum amplitude control (attenuator) or by placing the selector switch in its SPECTRUM AMPLIFIED position, or both. To do so would burn up the resistance cards in the attenuator and burn out the crystal, even though the equipment is turned off.

The antenna horn should be pointed directly to the source under test. In the case of an oscillator, it is seldom necessary to remove the oscillator cover. The magnetron and local oscillator spectrums may be observed simultaneously if a proper location for the antenna horn is found that produces sufficient attenuation to the output of the magnetron.

If a very stable method of coupling is desired, remove the antenna horn and connect the cable to the directional coupler or wave selector which is inserted at some convenient place in the transmission line of the radar equipment. The spectrum analyzer will then receive approximately one percent of the power in the directional coupler. This method is very satisfactory when it is necessary to check the antenna while it is in motion to see if it is pulling the frequency of the magnetron.

Observation of Patterns

Adjust the spectrum amplitude control to obtain the desired amplitude on the screen. Rotate the spectrum width control counterclockwise until the spectrum gains sufficient detail to make observation easy. If more pulses are desired in the spectrum, turn the sweep frequency control counterclockwise to a lower sweep frequency.

Frequency Measurement

The frequency meter pip that is visible when the selector switch is in the MIXER position (fig. 3-21) represents the actual frequency of
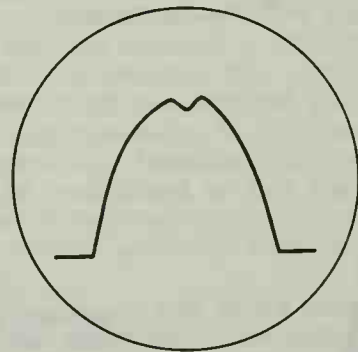
NO SIGNAL PRESENT EXCEPT
THAT OF ANALYZER R-F
OSCILLATOR

92.142

Figure 3-21.—Frequency meter pip. (Selector switch in MIXER position.)

the frequency meter, as read on the frequency meter dial. When the selector switch is in either of its other two positions for frequency measurements, the reading on this dial does not represent the frequency being measured. The unknown frequency must be computed from the dial reading by adding or subtracting the 22.5-mc intermediate frequency of the analyzer.

Whether the intermediate frequency is added or subtracted depends on the image the frequency meter pip is made to coincide. Two images will appear for each component of the spectrum (fig. 3-22). The frequency of the analyzer r-f oscil-
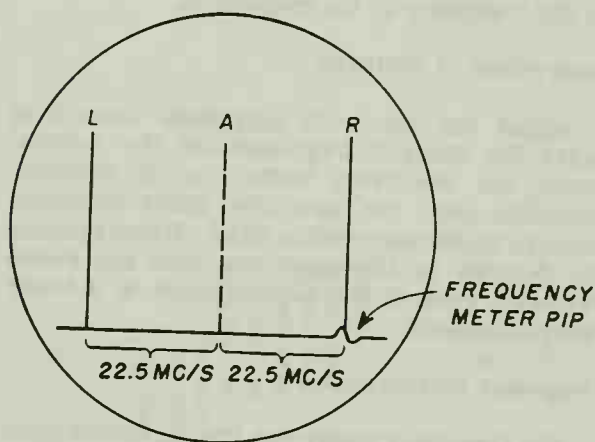

Figure 3-22.—Right-hand and left-hand patterns.

92.143

lator increases as the sweep progresses from left to right across the screen. This action causes an image to be produced below and above the 22.5-mc intermediate frequency. In most cases, both images can be made to appear on the screen although, in actual practice, one is centered on the screen so that the other does not appear.

If the image selected for frequency determination is the one on the right, it is necessary to subtract 22.5 mc from the dial reading. The remainder thus obtained represents the frequency of the unknown signal. If the left image is selected, it is necessary to add 22.5 mc to the dial reading. Since the two images appear above and below the true frequency by 22.5 mc, it is obvious that the two readings may be made directly from the meter, and the average taken to obtain the unknown frequency.

Another very accurate method may be used if the unknown signal is stably coupled to the analyzer. The method consists of tuning the frequency meter until its pip is equidistant from the two images (fig. 3-23). While tuning the meter, both images must be carefully observed. At the instant the resonant frequency of the frequency meter equals the frequency of the unknown signal, it will be possible to see both images dip simultaneously. The reduction in amplitude occurs because the frequency meter absorbs energy from the unknown signal frequency. In this case, the reading on the frequency dial represents the actual frequency of the unknown signal.
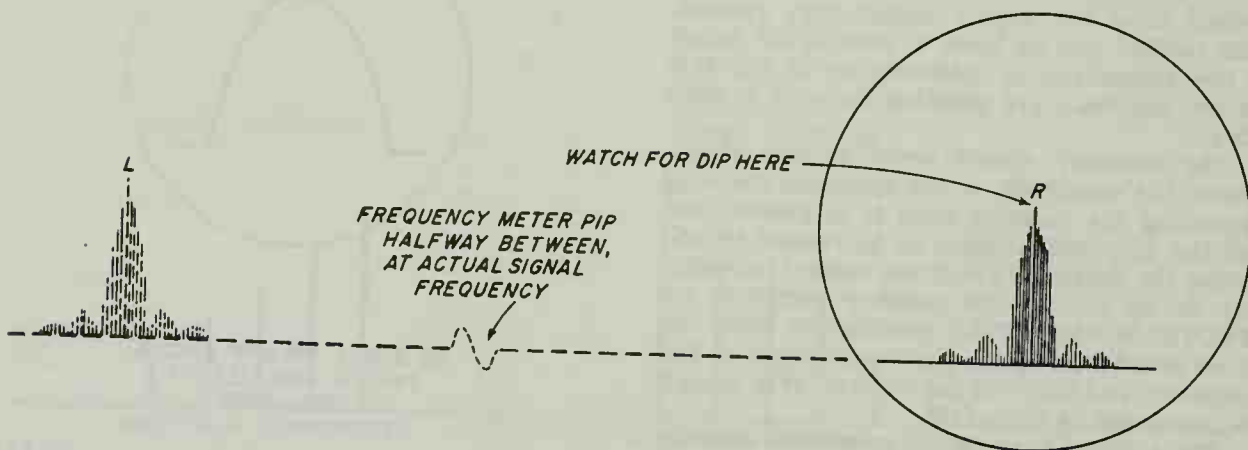

Figure 3-23.—Determination of frequency by tuning frequency meter for dip in spectrum.

92.144

Measurement of Radar Local
Oscillator Frequency

There are two similar methods for measuring the frequency of a radar local oscillator. The choice of methods depends upon whether the intermediate frequency is 30 mc or 60 mc.

Assume that the radar intermediate frequency is 60-mc, and that the local oscillator frequency is above that of the magnetron. It should be noted that two frequencies 60 mc apart cannot be seen on the same scope trace. This is true because the full width of the i-f above and below the analyzer radio frequency is only 45 mc wide (fig. 3-24). In order to see both patterns simultaneously, a unique use is made of the fact that two spectrum images appear for each signal. In practice, the spectrum center control is used to make the right-hand spectrum of the magnetron appear on the left side of the scope. The left-hand pattern of the radar local oscillator should then appear to the right of this pattern. The pattern represented by the dotted lines in the figure does not appear on the scope, but is shown to illustrate the frequency relationship of the two images that do appear.

The accuracy with which the radar local oscillator tracks the magnetron can be determined by measuring the difference between the frequencies of the two images on the screen. This difference should be 15 mc (using the lower magnetron image as the reference). To determine the actual difference, rotate the frequency dial to set the frequency meter pip under the center of the magnetron spectrum, and note the dial reading. Then set the pip under the radar local oscillator pip, and again note the dial reading. If the difference between the dial readings is 15 mc, the radar local oscillator (or AFC discriminator) has been accurately set to follow the excursions of the magnetron frequency.

If the frequency of the radar local oscillator is set 60 mc below the magnetron frequency, the only effect is that the images will be reversed on the screen (fig. 3-24B).

If the intermediate frequency is 30 mc above, the procedure is similar to that first discussed, except that the position of the images is reversed (fig. 3-24,C). This is true since by time relationship, the left-hand spectrum of the radar local oscillator is made to appear on the left side of the screen, while the spectrum of the magnetron now appears on the right. The difference in the frequencies is measured just as it was in the foregoing method, and a frequency difference of 15 mc indicates that the local oscillator is operating at a frequency 30 mc above the magnetron.

If the radar local oscillator operates below the magnetron frequency, the position of the images will again be reversed. That is, the radar local oscillator spectrum will be on the right and the magnetron spectrum on the left.

SPECTRUM ANALYZER USED AS
A SIGNAL GENERATOR

The analyzer may be used as a c-w or f-m signal generator, depending upon whether the spectrum width control is set at either of its extreme positions. If it is set completely counterclockwise, the percent of frequency modulation is at a minimum (about 3 mc), and the output is essentially a c-w signal. If the control is turned all the way clockwise, the percent of frequency modulation is maximum, and the output has a width of from 40 to 50 mc, depending upon the particular tube in the analyzer r-f oscillator. The output power ranges between 1 and 2 milliwatts.

THE ECHO BOX TS-275/UP

Good radar performance is of vital importance because radar is the eye that the Navy depends on to detect enemy ships and planes long before they are detected by other methods. It has failed to accomplish this purpose if attacking enemy craft are detected too late for effective countermeasures to be taken. It is therefore extremely important that radar installations be maintained so that they always operate close to their maximum efficiency, and technicians should spare no effort in making this possible.

The ECHO BOX is one of the most important single test instruments for indicating the overall radar system performance. This results from the fact that the echo-box indication reflects the combined relative effectiveness of the transmitter as a transmitter of energy and the receiver as a receiver of energy.

The echo box, or resonance chamber, consists basically of a resonant cavity, as indicated in figure 3-25,A. The resonant frequency of the cavity is determined by the size of the cavity (the larger the cavity, the lower the frequency); and this, in turn, is determined by
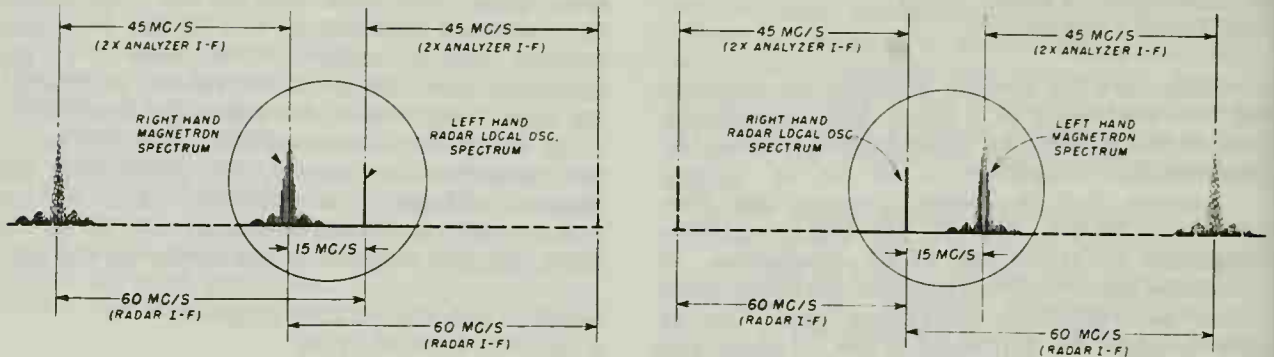
A Measuring the frequency of a local oscillator that is 60 mc above the magnetron.

B Measuring the frequency of a local oscillator that is 60 mc below the magnetron.
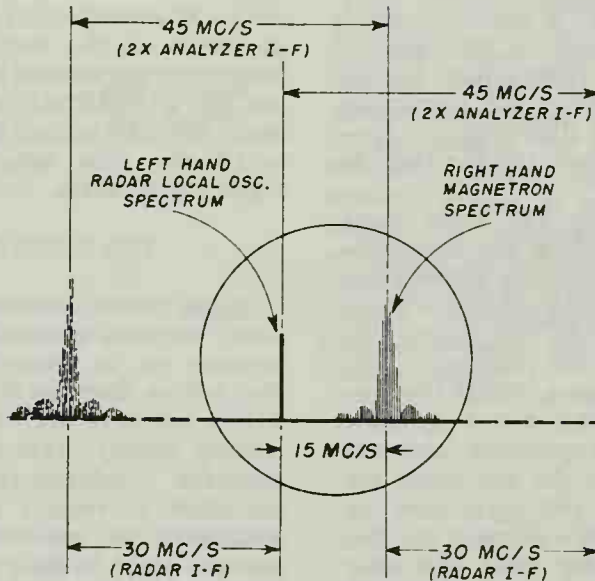
C Measuring the frequency of a local oscillator that is 30 mc above the magnetron.

92.145-.147

Figure 3-24.—Radar local oscillator frequency measurements.

the position of the plunger. The accurately calibrated tuning mechanism controls the position of the plunger and indicates on a dial the resultant resonant frequency, or data that permits the technician to determine the frequency with great accuracy by consulting a set of curves.

Energy is coupled into the cavity from the directional coupler (or pickup dipole) by means of an r-f cable connected to the input loop. Energy is coupled out of the cavity to the rectifier and the microammeter by means of the output loop. The amount of coupling between

TUNING MECHANISM

INPUT LOOP FROM DIRECTIONAL COUPLER

PLUNGER

RESONANT CAVITY

OUTPUT LOOP TO METER

A SIMPLE ECHO BOX

CRYSTAL RECTIFIER

RFC

OUTPUT LOOP

METER

B OUTPUT CIRCUIT

C

DIRECTIONAL COUPLER

ANTENNA

ECHO BOX

TR BOX

RECEIVER

MAGNETRON
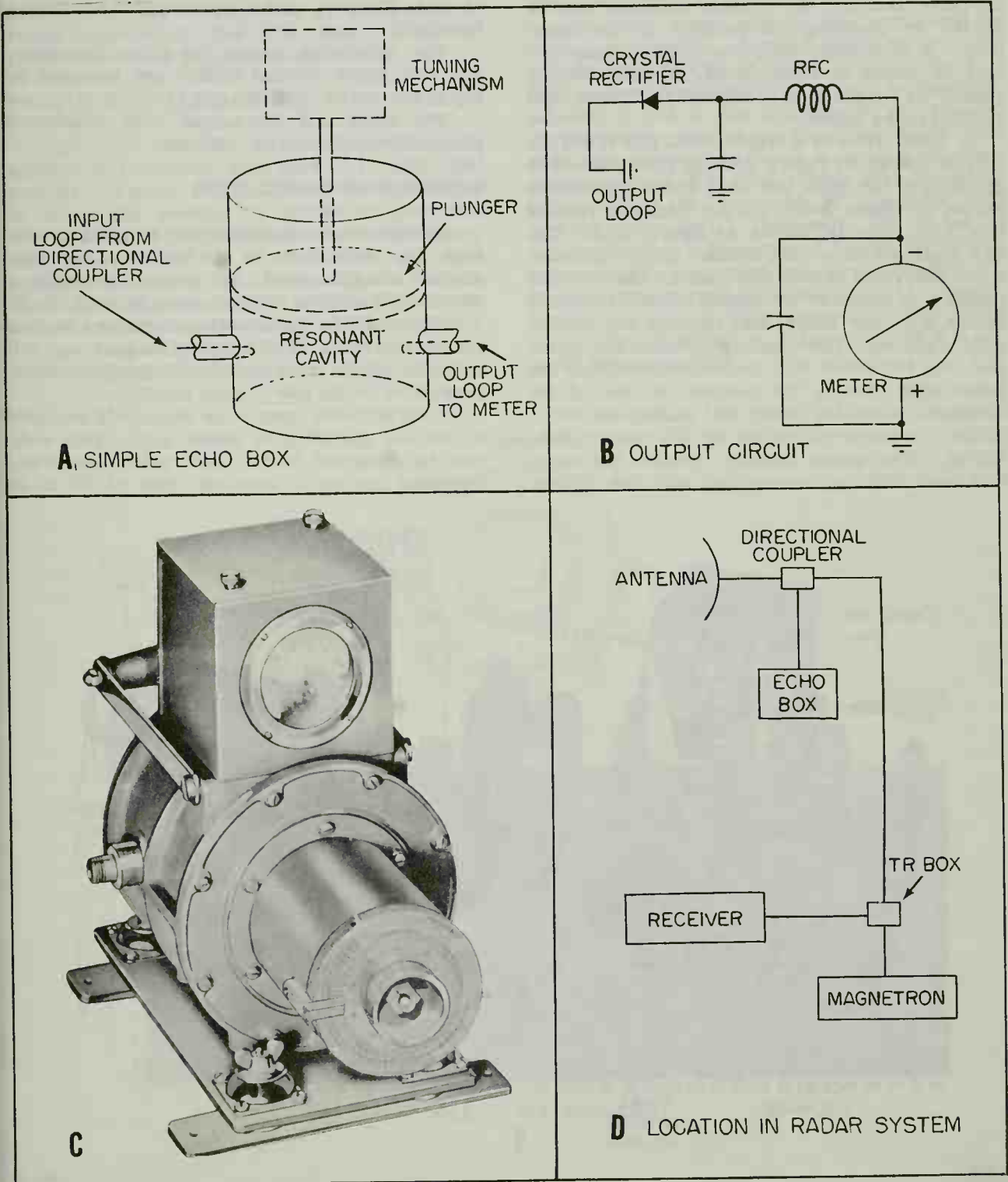
D LOCATION IN RADAR SYSTEM

55.86

Figure 3-25.—Echo box.

79

the echo box and the crystal rectifier can be varied by changing the position of the output loop. A schematic diagram of the output circuit is shown in figure 3-25, B. The energy picked up by the loop is rectified, filtered, and applied to the meter.

A front view of a typical echo box (TS-275/UP) is shown in figure 3-25, C; the method of connecting the echo box in a radar system is shown in figure 3-25, D. An exploded view of this echo box is shown in figure 3-26. The box consists of a cast bronze cavity cylinder with removable bronze end plates. The movable plunger is actuated by means of the adjusting screw and the inner dial through the watertight bellows. The gearing between the inner dial and the outer dial is so arranged that the outer dial travels the amount of one of its divisions while the inner dial makes one revolution (10 major divisions or 100 minor divisions). The gears merely operate the outer dial and have no connection with the driving

of the plunger, and therefore do not cause backlash.

The indicating meter, the filter capacitor, and the spare crystal holder are mounted on top of the cavity cylinder.

The input and the output loop connectors project from the cavity cylinder.

METHODS OF CONNECTING

Either of two methods may be used to connect the echo box to the radar. A pickup dipole may be used, but more generally a directional coupler is used, as in figure 3-25, D.

Much of the information contained in this instruction book is of a general nature and will be very helpful to anyone desiring practical information on the use of echo boxes.

A directional coupler is commonly included in the r-f plumbing of radar sets. This echo box is designed to be used with directional couplers having a coupling loss of 20 to 35
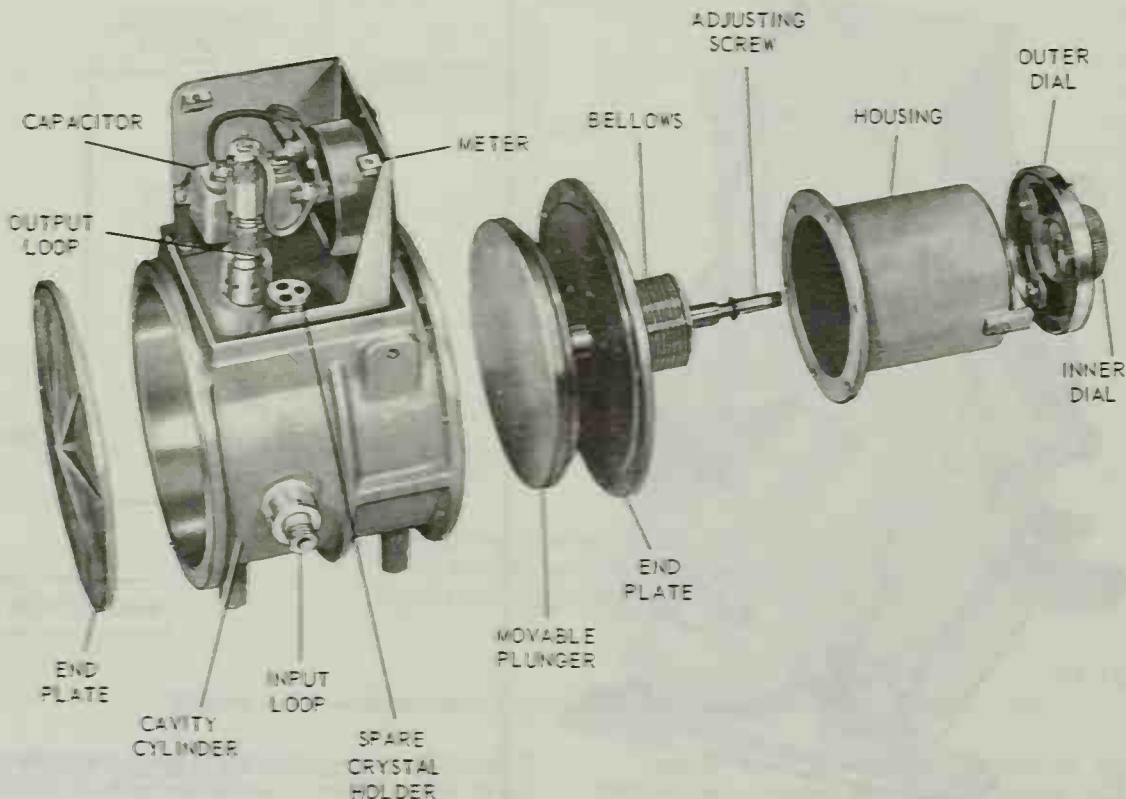

70.46

Figure 3-26.—Exploded view of the TS-275/UP echo box.

decibels on the usual radar in the frequency range covered by this echo box. The exact value depends on the radar.

FREQUENCY CALIBRATION

The echo box dial is read by reading the middle dial number and then the inner dial number. Figure 3-27, A, shows an example in which the reading is three two point six five (32.65). Each major division on the inner dial is one-tenth of a revolution, and each smaller division is one-hundredth of a revolution. Each division of the middle dial represents one complete revolution of the inner dial.

The frequency calibration curve in figure 3-27, B gives the tuning rate of the echo box

with any setting of the tuning dial. The method of using the curve may be best explained by giving an example. Assume that a radar under test is supposed to be tuned to a frequency of 3607 mc. (The direct frequency calibration, marked in red on the instrument, is provided on the outer portion of the outer dial.) The echo box is found to be in resonance when the tuning control is at 27.5 (middle dial divisions) rather than the expected 26 (opposite 3607 mc on outer dial). Obviously, the radar is transmitting on a frequency 1.5 inner dial revolutions higher than intended. From figure 3-27, B, the tuning rate is found to be 15.5 megacycles per (inner) dial revolution. The radar is thus transmitting on a frequency 15.5 x 1.5, or 23.25 mc too high.

A

B

70.47
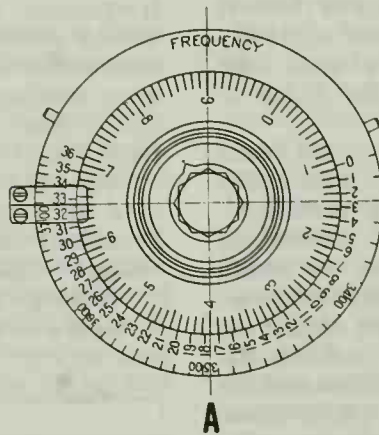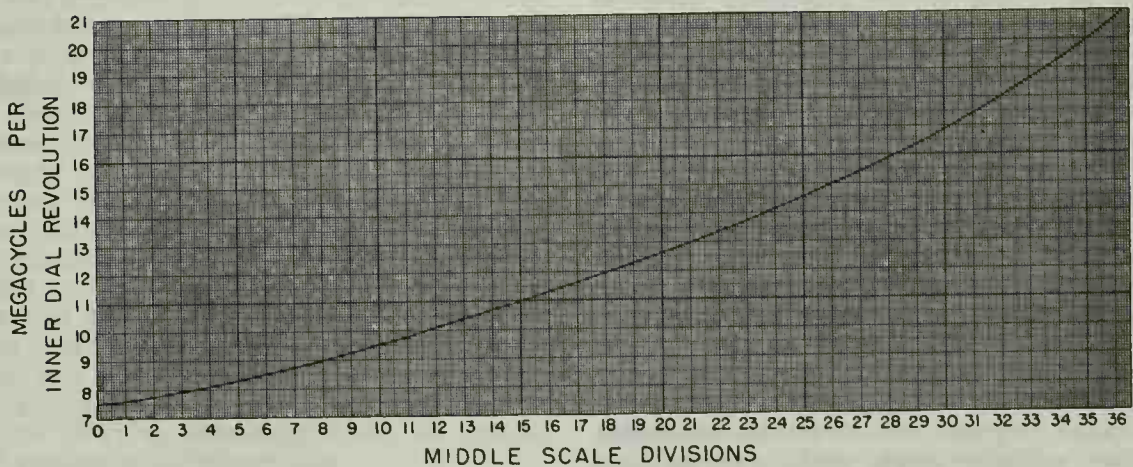
Figure 3-27.—Reading the dial.

RINGTIME

Some of the energy generated by the radar transmitter is picked up by the echo box via the directional coupler. This energy excites oscillations in the echo box that persist for some time after the end of the radar pulse, much in the fashion of an echo that persists in a large room after a loud noise. As this echo dies down, a part of it is fed back into the radar receiving system, again via the directional coupler. This causes a saturated signal to appear on the radar indicator, which is known as RINGING. The longer this ringing extends the better the performance of the radar—that is, the more powerful the transmitter is and/or the more sensitive the receiver is.

The length of time the echo box SHOULD ring under the particular conditions of the test (called the EXPECTED RINGING TIME, or RING-TIME) may be compared with the ringing time observed, to determine whether the radar is performing well or not.

The ringtime to be expected on a good radar depends on the particular type of radar being tested; on the way the echo box is installed—that is, for example, whether a directional coupler or a dipole is used, and on the length and type of cable used; on the individual ringing ability of the particular echo box employed; on the frequency of the radar; and on the temperature of the echo box at the time of the test. Corrections are made for all of these factors according to the procedure given in the technical manual for the echo box being used.

An echo box without correction may be used for the purpose of detecting a CHANGE in the performance of a radar. The ringtime is simply noted and compared from day to day. It should be recognized that these readings do not permit the comparison of a particular radar with a standard of performance, and thus to tell whether more may be expected from a radar than its past performance would indicate.

Ringtime Measurements

Because ringtime measurements constitute the most valuable single feature of the echo box, it is essential that they be carried out properly and with due regard for the necessary precautions. Ringtime measurements are made on the A-scope or on the PPI; both methods are discussed later.

In measuring the ringtime, the technician should make sure that it is the echo-box ringtime and not some fixed-target echo or block of echoes that is being received. This condition can be determined by adjusting the radar gain control and noting if there is a back and forth movement of the ringtime on the scope. The echo box echo will change in range; fixed target echoes, however, will not change in range, only in amplitude.

In order to obtain accurate results, every ringtime measurement should be repeated at least four times, and the readings averaged. Care must be taken to ensure that all readings are accurate. If two or more technicians use the same echo box, they should practice together until their ringtime measurements agree.

Radars have a tendency to drift slightly in frequency. When this occurs, the echo box becomes detuned and accurate ringtime measurement is difficult. Therefore, it is necessary when making ringtime measurements for longer than a very few minutes to retune the echo box from time to time.

The radar antenna should not be pointed at a mast or other nearby obstruction because proximity effect may cause the transmitter to change frequency.

A-scope presentation of ringtime is indicated in figure 3-28, A. The receiver gain should be

A. A-Scope

B. PPI
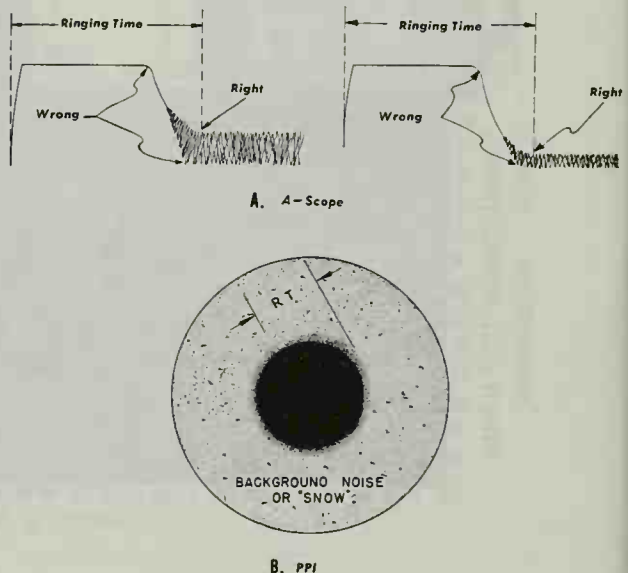
55.121(70)A

Figure 3-28.—A-scope and PPI presentation of ringtime.

set so that the 'grass' or noise is one-quarter to one-third the total saturated signal height on the A-scope. When this is done, a good pattern results, such as either of those shown in the figure. In the event that no 'grass' can be seen, the gain of the radar i-f is inadequate and repairs should be undertaken.

The exact end of the ringtime occurs at the furthest point to the right at which the TOP of the 'grass' is noticeably above the general level of the rest of the 'grass.' Do NOT judge ringtime by the BOTTOM of the 'grass' or by the end of the saturated portion of the ringtime because these items are influenced by the receiver gain setting and other factors.

Setting the gain too high or too low may make it difficult or impossible to read the ringtime with accuracy. (It is essential that 'grass' be present.)

An A-scope indicator measurement of ringtime is usually best performed when the radar antenna is stopped.

PPI presentation of ringtime is indicated in figure 3-28, B. In this instance, the same general principles apply as did in the case of the A-scope presentation.

The following procedure should be followed. With the radar antenna rotating, set the receiver gain at a minimum and adjust the intensity (bias) so that there is a very slight radial trace on the PPI indicator. Increase the receiver gain until the PPI-indicator area seems to be just half covered with flecks of snow.

A PPI ringtime pattern, with proper receiver gain adjustment (and the radar antenna rotating), is shown in figure 3-28. In this case, the echo box is used with a directional coupler.

It should be emphasized that the end of the ringtime signal is NOT at the place where the bright or saturated part of the signal ends, but where the fainter portion of the signal disappears into the background noise. Therefore, when reading the ringtime on a PPI indicator, be sure to observe to the extreme edge of the grass and NOT JUST TO THE END OF THE BRIGHT PORTION OF THE PATTERN. Read to the last point at which the 'snow' is unusually bright. As indicated in figure 3-28, B ringtime (RT) is measured from the center of the pattern to the outer edge.

SPECTRUM ANALYSIS

Every time a radar transmitter generates an r-f pulse, it produces a certain amount of r-f energy in the form of electromagnetic waves. Not all of these waves, however, are of the same frequency; in fact, only a small portion of them have exactly the same frequency as that to which the transmitter is tuned. The rest of the radiation is at slightly higher or slightly lower frequencies, forming the sideband frequencies. This is the natural result of pulse modulation and cannot be avoided.

Actually, the radar energy is distributed more or less symmetrically over a band of frequencies, as illustrated in figure 3-29, A. This frequency distribution of energy is known as the SPECTRUM. An analysis of its characteristics may readily be carried out with the aid of the echo box.

When properly performed and interpreted, a spectrum analysis will disclose maladjustments and troubles that would otherwise be difficult to locate. It is important, therefore, that the technician who uses the echo box be able to carry out a spectrum analysis and understand the results.

When a spectrum analysis is to be made, the tuning control of the test set (fig. 3-25, C) is first turned until a maximum output meter deflection is obtained, then the tuning control is turned slowly from a point well below this maximum to a point well above it.

While this is being done, the output meter readings are noted for various settings of the tuning control. It is good practice to cover the frequency range desired by turning the tuning knob slowly in the same direction to each new position, not by turning it back and forth. This is done to minimize any possible error due to backlash. A reading should be taken about every 0.02 revolution of the tuning knob.

Finally, an accurate graph is constructed with the meter readings plotted against the tuning control dial settings. The resulting graph should resemble one of those shown in figure 3-29.

A radar transmitter in satisfactory condition should give a spectrum curve similar to curve A or curve B. Good curves are those in which the two halves are symmetrical and contain deep, well-defined minimum points on both sides of the main peak.

A curve without deep minima, as in curve C, indicates that the transmitter output is frequency modulated during the pulse. This may be due to the application of a negative pulse to the magnetron that does not have sufficiently steep sides or flat extremities. It may also be due to
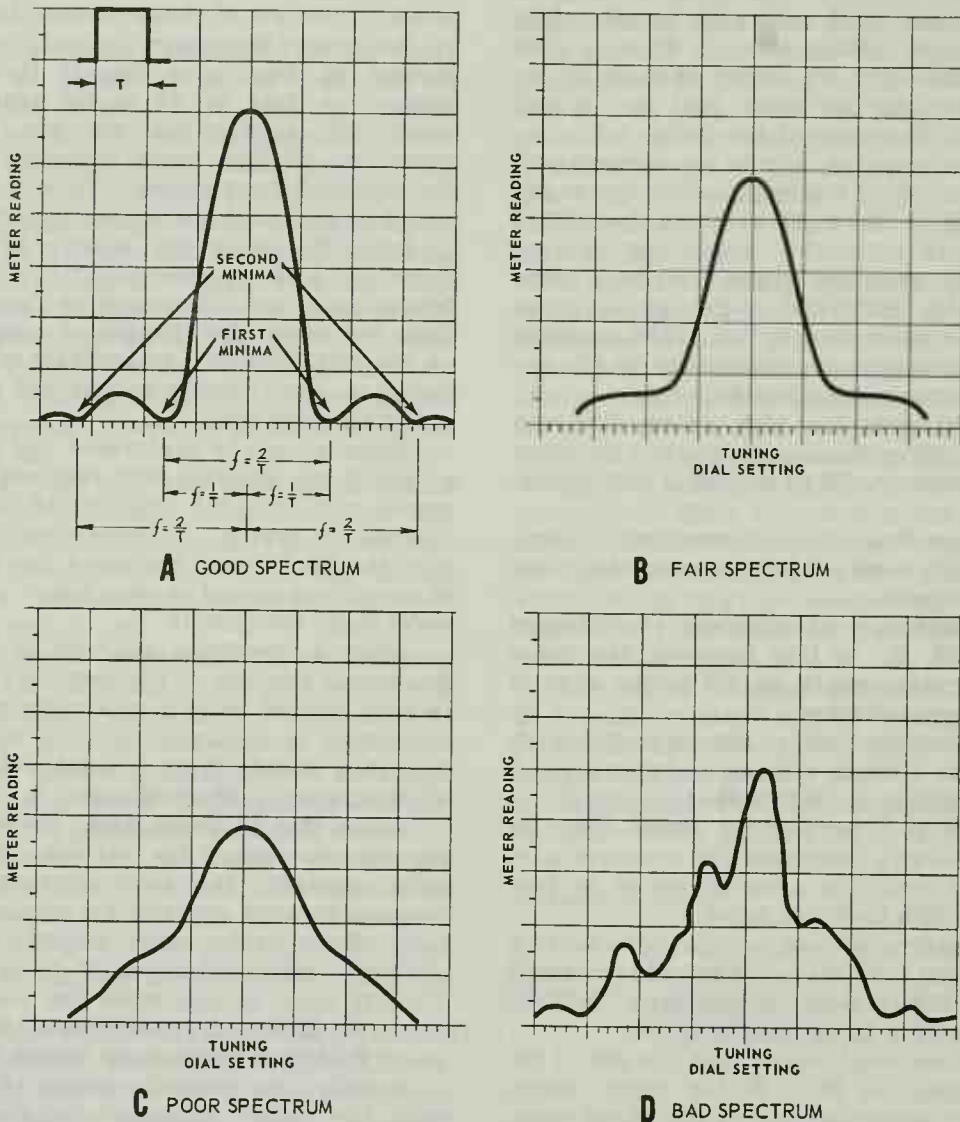
Figure 3-29.—Typical radar spectra.

70.48

a transmitter tube that is unstable or is operated with improper voltage, current, or magnetic field.

When the spectrum is extremely irregular, as in curve D, it is an indication of severe frequency modulation. This will probably cause trouble in the receiver automatic frequency control as well as general loss of signal strength. When the spectrum has two large peaks, quite far apart, it indicates that the transmitter tube is double moding, perhaps because of unwanted

standing waves in the transmission line or a bad transmitter tube. A faulty spectrum can often be improved by adjustment of the transmission line stubs or by replacement of the transmitter tube. Standing waves may be due to a faulty line connection, a bad antenna rotating joint, or obstructions in the line.

In the case of a good or fair spectrum curve with sharply defined minima on both sides of the main peak, the distance between these two minima is proportional to the duration of

the transmitted pulse. Because the duration of the pulse determines the distribution of power in the sideband frequencies, the pulse length may be found from the spectrum graph. The procedure is to determine the distance in megacycles between the minima on either side of the main peak. These minima are separated by a frequency (in megacycles) equal to $\frac{2}{T}$, where T is the pulse length in megacycles and $\frac{1}{T}$ is the number of sideband frequencies contained in either upper or lower sideband (from the carrier frequency to the first minimum on either side of the carrier). Expressed as an equation,

$$\text{pulse length in microseconds} = \frac{2}{\text{distance between minima in megacycles}}$$

Suppose, for example, that the echo box is being used to check the pulse length of a radar. The graph of the spectrum is plotted, and the tuning distance between the minima is from 17 divisions on the middle dial (fig. 3-27, A) to 17.17 divisions (17 divisions on the inner dial). From figure 3-27, B, it is found that for the particular frequency range in question (3480-3490 mc) the echo box tunes at a rate of 11.65 mc per revolution (corresponding to 17 divisions on the middle dial). The frequency span between minima is

$$11.65 \times 0.17 = 1.98 \text{ mc.}$$

Applying the equation given in the last paragraph, the pulse length in microseconds $= \frac{2}{1.98} = 1.01$ microseconds. The value thus calculated can be readily compared to the standard value for a radar of the type under test by reference to the radar manual. Any great change in the test value compared to the standard value indicates an improper pulse length.

The shorter the pulse length, the wider will be the frequency band that the signals occupy. This effect will appear on the graph as a wide span between the first minima of the spectrum curve.

An abnormally narrow spectrum shows that the transmitted pulse is too long. Such a pulse could result in a long ringtime and high power

reading on the echo box output meter, thus falsely indicating superior system performance.

Power Output

At the time of installation, the meter reading on the echo box, TS-275/UP, should have been set between 40 and 80 divisions (by adjusting the orientation of the output loop) while the echo box was tuned to resonance with the radar transmitter.

The output meter reading is closely proportional to the average radar power picked up by the echo box and to the transmitter pulse length, when the echo box is tuned to the maximum output signal of the spectrum. If the pulse length is long, the spectrum curve is consequently high and narrow, and the meter reading is high. Where the pulse length is shorter, the spectrum curve is flatter and the meter reading will be lower.

The power output of a radar is generally good if the transmitter current is normal. Loss in the transmission line may cause loss of power, and in the event that low power is observed at the antenna of the radar by means of a pickup dipole and echo box, the transmission line may be suspected. Because of the high initial cost of radar transmitting tubes, the echo box and its accessories should be checked carefully before discarding such a tube. Transmitter tuning stubs are NEVER adjusted for maximum power output, as indicated at the echo box. These stubs provide a transmitter frequency adjustment, the net effect of which is observed at the receiver indicator rather than at the echo box. The correct procedure for checking the radar frequency is described in this chapter.

OTHER TESTS

GENERAL PROCEDURE.—A variety of radar equipment checks can be carried out with the aid of the echo box. The exact nature of these tests, as well as the detailed methods of procedure, may vary to some extent between different types and models of radar equipment. The typical procedures given in the following paragraphs will be useful in establishing test routines to be followed in radar maintenance. Practice and experience may suggest variations as the technician becomes more familiar with the use of the echo box.

In the testing procedures outlined in the following paragraphs it is assumed that the echo

box is properly installed, and that the expected ringtime and output meter reading are known for the particular radar under test.

All measurements should be recorded in the radar log and/or on forms provided for this purpose. One such form for ringtime, from the Maintenance Standards Books, Part II Preventive Maintenance Checkoff for Radar Set AN/SPS-8, NavShips 91522.41, is included in figure 3-30. The figure includes complete
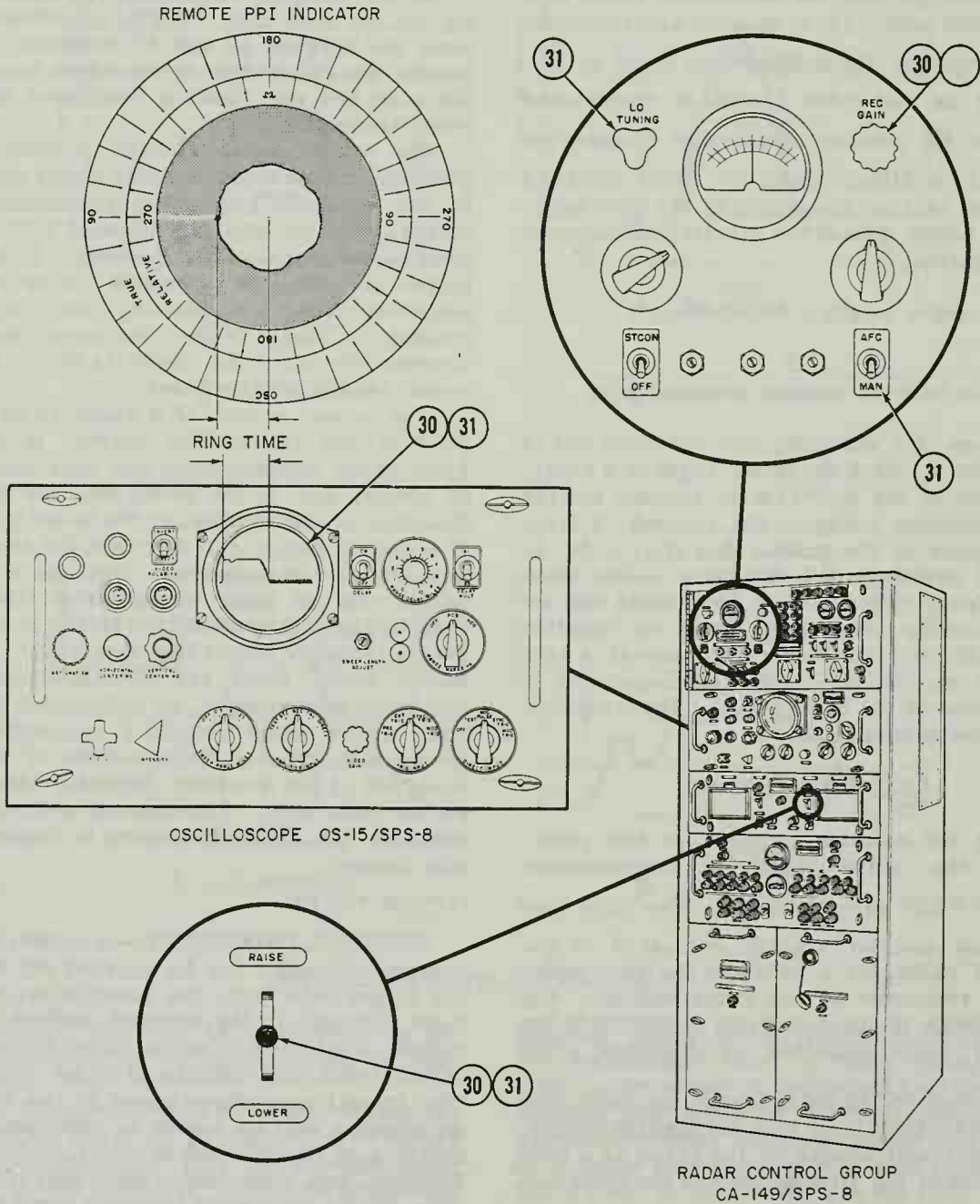

REMOTE PPI INDICATOR

RING TIME

OSCILLOSCOPE OS-15/SPS-8

RADAR CONTROL GROUP
CA-149/SPS-8

36.100(70)A

Figure 3-30.—Steps for recording ringtime.

instructions for making the test. Another form for listing echo box meter readings from the same publication is included in figure 3-31.

As a preliminary step in all tests, the radar equipment should be allowed to warm up fully to the normal operating temperature. The directional (or bidirectional) coupler or the pickup dipole (when used) should be correctly coupled to the echo box. All antijamming provisions (provisions to reduce the effects of the enemy jamming the radar) and the sensitivity time control, if provided, should be turned off.
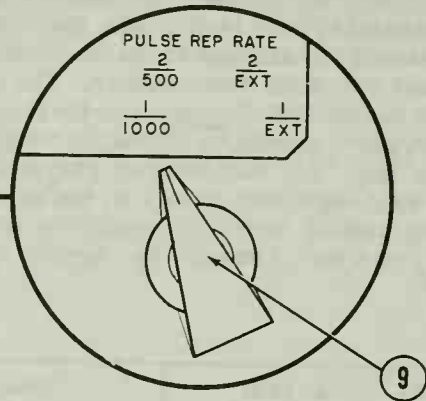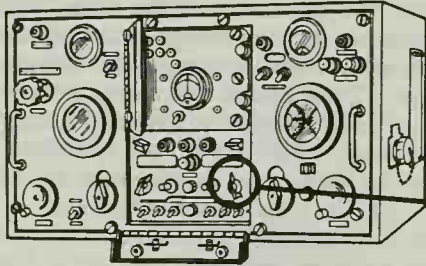
Overall Performance

Adjust the echo-box tuning knob for a maximum reading of the output meter, indicating that the echo box is tuned to resonance with the radar. Then adjust the radar receiver local oscillator frequency for maximum ringtime on the indicator. Measure the ringtime as accurately as possible, preferably by taking an average of at least four readings. Ringtime is most conveniently measured on an A-scope with the antenna stopped, and on a PPI-scope

STEP NO.	ACTION REQUIRED	PRELIMINARY ACTION	READ INDICATION ON	PERFORMANCE STANDARD
(30)	Record radar ring time with AFC.	Set REC GAIN (R7142) for one-half inch of grass on Oscilloscope screen. Tune Echo Box with RAISE-LOWER switch (S7404) for maximum ring time indication on Oscilloscope screen. The ring time is obtained directly in yards on the remote PPI indicator. Rotate Antenna and set the range ring slightly beyond the solid echo portion of pattern as shown on illustration page.	Range Dial Remote PPI Indicator.	_____ YDS (3200 or more)
(31)	Record radar ring time in manual tuning position.	Same as Step (30) except AFC-MANUAL switch (S7104) in MANUAL position, and LO TUNING (R7138) varied for maximum ring time.	Range Dial Remote PPI Indicator.	_____ YDS (3200 or more)

STEP NO	(30)	(31)		(30)	(31)		(30)	(31)		(30)	(31)		(30)	(31)		(30)	(31)	
Month	JAN 19_			FEB 19_			MARCH 19_			APRIL 19_			MAY 19_			JUNE 19_		
Week	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init
1																		
2																		
3																		
4																		
5																		
Month	JULY 19_			AUG 19_			SEPT 19_			OCT 19_			NOV 19_			DEC 19_		
Week	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init	Yds	Yds	Init
1																		
2																		
3																		
4																		
5																		

36.100(70)A

Figure 3-30.—Steps for recording ringtime—Continued.
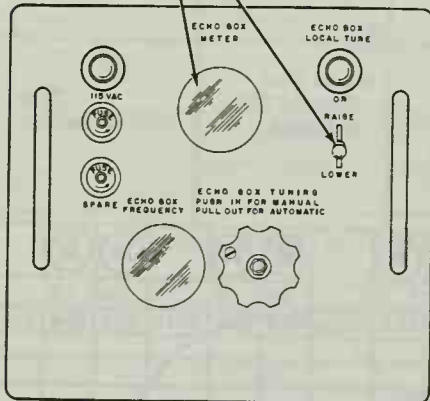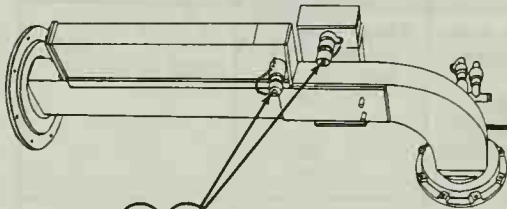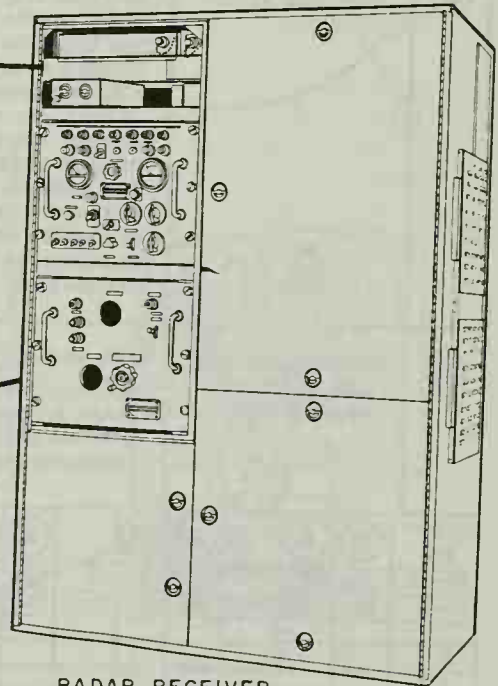
36.100(70)B

Figure 3-31.—Steps for recording transmit/receive ratio.

with the radar antenna rotating. Record the ringtime on the forms provided. Compare this figure with the corresponding value of the performance standard.

If the output meter and ringtime measurements are both satisfactory (compared with the expected values), the radar transmitter and receiver are both functioning well. If the meter

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STEP NO.	ACTION REQUIRED	PRELIMINARY ACTION	READ INDICATION ON	PERFORMANCE STANDARD
⑧	Record Echo Box ratio as an indication of transmission line voltage standing wave ratio (VSWR) with a 1-microsecond pulse. (These ratios are inversely proportional; if actual VSWR is desired see interpolation curves in Section 6 of NAVSHIPS 91522 (A)).	Connect echo box to J3007 on transmitter arm of bidirectional coupler. Vary RAISE-LOWER switch (S3501) for maximum reading on Echo Box Meter. Note reading. Connect echo box cable to J3008 on receiver arm of bidirectional coupler. Vary RAISE-LOWER switch (S3501) for maximum reading on Echo Box Meter. Note reading. Divide the first reading by the second reading and record the ratio.	Echo Box Meter ‾‾‾‾‾‾‾‾ First reading ‾‾‾‾‾‾‾‾‾ Second reading	 ‾‾‾‾‾‾‾Ratio (1.1 or larger)
⑨	Record Echo Box ratio as an indication of transmission line voltage standing wave ratio (VSWR) with a 2-microsecond pulse. (These ratios are inversely proportional; if actual VSWR is desired see interpolation curves in Section 6 of NAVSHIPS 91522 (A)).	Throw PULSE REP RATE switch (S6205) to 2/500 and repeat Step ⑧	Echo Box Meter ‾‾‾‾‾‾‾‾ First reading ‾‾‾‾‾‾‾‾‾ Second reading	 ‾‾‾‾‾‾‾Ratio (1.1 or larger)

STEP NO.	1st QUARTER 19__			2nd QUARTER 19__			3rd QUARTER 19__			4th QUARTER 19__		
	Ratio	Initial	Date	Ratio	Initial	Date	Ratio	Initial	Date	Ratio	Initial	Date
⑧												
⑨												

36.100(70)B

Figure 3-31.—Steps for recording transmit/receive ratio—Continued.

reading is satisfactory but the ringtime is low, the radar receiver is the probable source of the trouble. Service the receiver, consulting the appropriate instruction book for the detailed procedure.

Transmitter Power

The echo-box output meter reading is closely proportional to the average energy radiated from the radar on a particular frequency. The measurement of relative transmitter power is, therefore, a direct and simple procedure. Tune the echo box to resonance and then stop the radar antenna. Record the maximum reading on the output meter. This measurement, compared with the corresponding value on previous tests, gives an index of transmitter power. If the meter reading is satisfactory the radar

power output is good. If the meter reading and ringtime are low, the transmitter power output is low, and a spectrum analysis should be made.

Radar Frequencies

To check the TRANSMITTER FREQUENCY, adjust the echo-box tuning knob for maximum deflection of the output meter, and stop the antenna. Read the tuning-knob scale and determine the transmitter frequency by reading the red calibration on the outer portion of the face of the outer dial, or by referring to the echo-box frequency calibration curve (fig. 3-27, B). If the transmitter frequency is found to be different from the frequency intended, the cause may be the transmitter tube or the transmitter tuning adjustments. Reference should be made to the appropriate instruction book.

To check the LOCAL OSCILLATOR FREQUENCY, the echo box is disconnected from the directional coupler (or pickup dipole). It is then coupled directly to the local oscillator output. (A special connector may have to be improvised to fit the local oscillator output of certain radars. It is desirable to insert 10 to 20 db of attenuation between the local oscillator and the echo box.) Adjust the echo-box tuning control for maximum deflection of the output meter. Read the tuning-control scale and, referring again to the echo-box calibration data, record the local oscillator frequency. The echo-box meter reading is likely to be excessive if attenuation is not inserted, and the output loop (fig. 3-25, A) should be adjusted to protect the meter (the loop may be adjusted with a wrench).

The DIFFERENCE BETWEEN THE TRANSMITTER AND THE LOCAL OSCILLATOR FREQUENCIES is easily checked by the curve in figure 3-27, B. It is necessary to have not only the correct frequency difference but also to have the local oscillator operating on the proper side (above or below) the transmitter frequency. This frequency difference is, in each case, identical with the frequency of the radar i-f amplifier. Using the tuning curve in figure 3-27, B, the technician can readily check the frequencies and determine whether the difference between them is correct.

To TUNE THE LOCAL OSCILLATOR, couple the echo box to the local oscillator. From the echo-box calibration data, find the echo-box setting for the correct local oscillator frequency, and adjust the echo-box tuning control ac-

cordingly. Then adjust the local oscillator, referring to the radar manual for the correct procedure, until the echo-box output meter shows maximum deflection. The oscillator is then approximately tuned to the correct frequency. Final tuning of the oscillator should be such as to produce maximum ringtime and proper radar crystal current.

Erratic Transmitter Operation

Adjust the echo-box tuning control for maximum deflection of the output meter with the echo-box connected to the directional coupler, and stop the antenna. If the transmitter is operating normally, a good ringtime pattern will be displayed on the A-scope (fig. 3-28). If the ringtime is erratic, or extra background noise traces appear in the pattern, then the transmitter may be multiple moding (transmitting on two or more distinct frequencies) or failing to fire on every pulse. This may be due to faulty pulsing; transmission-line troubles, especially arcing; or other causes. Examine the spectrum in order to help localize the trouble. Refer to the appropriate instruction book for help in correcting the trouble.

Transmitter Pulling

Magnetron frequency pulling results from a change in the loading as the antenna is rotated. As a check for magnetron pulling, turn off the automatic frequency control on the radar receiver. Adjust the echo-box tuning control for maximum deflection of the output meter. With the radar antenna rotating, observe the pattern on the PPI indicator. If the transmitter is operating normally, a good ringing pattern will be seen, as in figure 3-32, A. If there are blank spaces on the PPI indicator or if the ringtime is reduced on certain azimuths, as in fig. 3-32, B, then the transmitter frequency is being pulled. In such a case, the output meter reading will also fluctuate as the radar antenna is rotated slowly. Transmitter frequency measurements at several azimuths may also be used to confirm this difficulty. The pulling may be caused by a bad rotating joint or by a reflecting surface near the antenna. Refer to the appropriate instruction book for corrective measures.

Automatic Frequency Control (AFC)

To determine whether the local oscillator is following the transmitter when it is pulled,
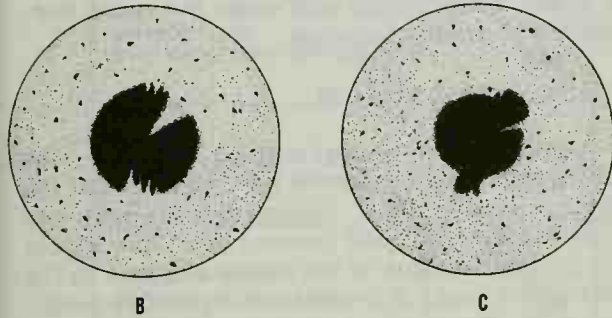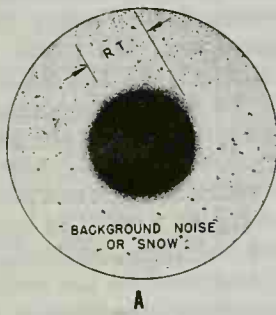
55.121(70)B

Figure 3-32.—Indication of magnetron pulling.

first stop the antenna on an azimuth where the ringtime pattern is broken, and then retune the echo box to resonance. Rotate the radar antenna and again examine the PPI pattern. If the ringtime is now good on the azimuth at which the echo box was retuned (fig. 3-32, C), the AFC is in operation on that azimuth and the local oscillator is following when the transmitter is pulled. As may be seen in part C, the ringtime may have now decreased in those azimuths where it was originally good. If the AFC does not follow, the pulling may be excessive or the AFC may be at fault. If the ringtime is greatly decreased at certain azimuths, and the AFC does not follow, the radar must be considered inoperative at those azimuths and should be so reported.

A simple procedure will show whether the AFC is locked on the proper frequency. Stop the radar antenna and tune the echo box for maximum meter reading. Turn off the AFC switch and tune the local oscillator for maximum ringtime, thus putting the local oscillator in proper tune. Turn the AFC switch on again. If the ringtime decreases, even slightly, the AFC is locking on the wrong frequency, or is failing to lock. The proper instruction book

should be consulted for corrective procedure. The probable cause of the AFC failure is a bad spectrum or the fact that the local oscillator is tuned to a frequency on the wrong side of the transmitter frequency.

T-R Box Recovery

The time required to permit T-R (transmit-receiver switch) recovery is determined by the time it takes the T-R switch to deionize after each transmitter pulse. It is usually defined as the time required for the receiver to return to within 6 db of normal sensitivity after the end of the transmitter pulse. T-R recovery time is the factor that limits the minimum range of a radar because the radar receiver is unable to receive signals until the T-R switch is deionized. In various radar sets, the recovery time may vary from about 3 to 20 μs.

To test for T-R box recovery, adjust the echo-box tuning control for maximum deflection of the output meter, and stop the radar antenna. Adjust the A-scope for a good ringtime pattern, such as curve E in figure 3-33. Slowly and gradually reduce the radar receiver gain setting, or better, detune the local oscillator. A pattern will result, such as curve D in figure 3-33, having the same relative shape as curve E. Further slight reduction in gain setting will produce another pattern, such as curve C, again similar in shape to curve E. Continue until a change occurs in the slope of the curve, as in curve B. This point of change marks the
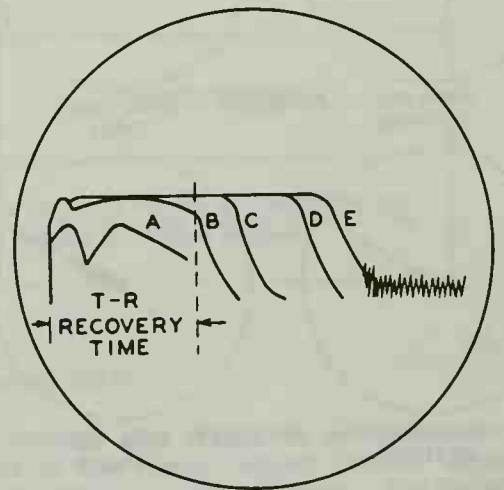
55.123

Figure 3-33.—Checking T-R box recovery.

T-R box recovery time of the radar, as indicated in the figure. For a good surface search radar, the T-R recovery time should correspond to a range of one mile or less.

If the gain control is reduced still further, a greatly distorted pattern will appear, such as curve A in figure 3-33. This curve shows that the T-R box has not recovered. Refer to the appropriate instruction book for corrective procedures. If the above procedure does not produce a series of curves (as indicated) giving a T-R recovery point, and if the ringtime is short, then it is probable that the T-R recovery time is much too high (greater than the ringtime) and a new T-R tube is needed. Check the keep-alive current. The keep-alive current should be negative and between 0.1 and 0.2 ma.

Receiver Recovery

Adjust the echo-box tuning control for maximum deflection of the output meter, and stop the radar antenna. Then detune the echo box, and adjust the radar receiver gain control until the indicator shows a pattern similar to one of those illustrated at the left of figure 3-34. Now retune the echo box to resonance and again observe the indicator pattern. If the receiver

recovery time is normal, the background noise will reappear immediately after the end of the ringtime pattern, and this noise will be approximately as strong as the noise previously observed with the echo box detuned. If the receiver recovery is slow, the noise will be weak and will not reappear for some time after the end of the ringtime pattern (see the right-hand portion of figure 3-34). In extreme cases of receiver nonrecovery, normal background noise may not reappear on the indicator at all. Receiver nonrecovery is usually an i-f tube or video defect which will make the radar susceptible to enemy jamming.

Transmission Line Loss

Using the directional coupler, tune the echo box to resonance. Record the ringtime and the output meter reading. Disconnect the echo-box from the coupler and connect it to the dipole (the proper location of the dipole is given in the instruction book; it is located at a position somewhere in the antenna radiation field). Again tune the echo-box to resonance and again record the ringtime and the output reading. Comparing these measurements, while allowing for the normal difference due to the difference between the coupler attenuation and the antenna space loss (losses are indicated in tables in the echo-box instruction book), an indication can be obtained of the loss in the radar r-f transmission line. If there is unusual loss in the radar transmission line, there will be greater differences than usual in the ringtime and also in the output meter reading. Repair or adjustments may then be undertaken.

Intermittent defects in the transmission line can often be found by rapping on the line while observing the echo-box meter reading.

Rapid Troubleshooting Chart

When the technician has become familiar with the test procedures and measurements, the echo-box may be used for rapid troubleshooting. Radar troubles may be more readily checked with the aid of the cause-and-effect chart of figure 3-35 which is essentially a summary of the information given on the use of the echo-box as a test instrument.

FREQUENCY-POWER METER TS-230

Frequency-Power Meter TS-230 measures the power (and frequency) of unmodulated and
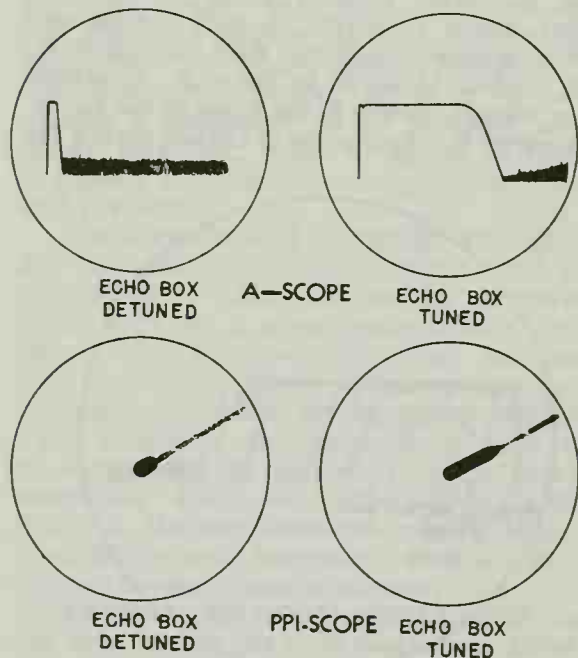

ECHO BOX DETUNED   A—SCOPE   ECHO BOX TUNED

ECHO BOX DETUNED   PPI-SCOPE   ECHO BOX TUNED
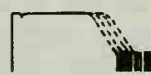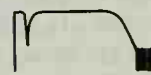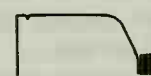
55.119

Figure 3-34.—Receiver nonrecovery.

EFFECT	APPEARANCE ON		PROBABLE CAUSE
	RADAR INDICATOR	ECHO BOX METER	
RINGTIME — NORMAL ECHO BOX METER — NORMAL			RADAR PERFORMANCE SATISFACTORY.
RINGTIME — LOW ECHO BOX METER — NORMAL			RECEIVING TROUBLE: DETUNED MIXER OR LOCAL OSCILLATOR, BAD CRYSTALS, EXCESSIVE I-F NOISE, ADJUSTMENT OF PROBES IN MIXER CAVITY, DETUNED T/R BOX.
RINGTIME — LOW ECHO BOX METER — VERY LOW			LOW POWER OUTPUT-CHECK SPECTRUM.
RINGTIME — LOW ECHO BOX METER — LOW			TROUBLE PROBABLY IN TRANSMITTER AND RECEIVER AND/OR TROUBLE IN TRANSMISSION LINE.
RINGTIME — ERRATIC ECHO BOX METER — STEADY			ECHO BOX DETUNED. BAD PULSING, DOUBLE MOOING TRANSMITTER, OR LOCAL OSCILLATOR POWER SUPPLY TROUBLE. CHECK SPECTRUM.
RINGTIME — ERRATIC ECHO BOX METER — ERRATIC			FAULTY TRANSMISSION LINE OR CONNECTION — CONDITION WORSE WHEN LINE IS VIBRATED.
END OF RINGTIME SLOPES GRADUALLY, POSSIBLY EXCESSIVE RINGING. GRASS APPEARS COARSE. ECHO BOX METER-STEADY AND SATISFACTORY.			OSCILLATING I-F STAGE
PRONOUNCED DIP IN RINGTIME AT END OF PULSE.			FAULTY T/R TUBE
RINGTIME-SLIGHTLY LOW POOR OR BAD SPECTRUM.		POOR SPECTRUM	TRANSMITTING TROUBLE
BLANK SPACES OR ROUGH PATTERN ON PPI RINGTIME INDICATOR. ECHO BOX METER READING VARIES AS ANTENNA IS ROTATED.			FREQUENCY PULLING OF TRANSMITTER DUE TO BAD ROTATING JOINT OR TO REFLECTING OBJECT NEAR RADAR ANTENNA.

55.122

Figure 3-35.—Troubleshooting chart.

pulsed signals in the range from 8500 to 9600 mc ± 4 mc. It measures average power within the limits of 0.1 and 1000 mw (-10 to + 30 dbm). A front view of the meter is shown in figure 3-36.

A general idea of how the power-measuring circuit in the meter works can be obtained from the following consideration. The thermistor (fig. 3-37) is actually the heart of the power-measuring circuit. This circuit operates on
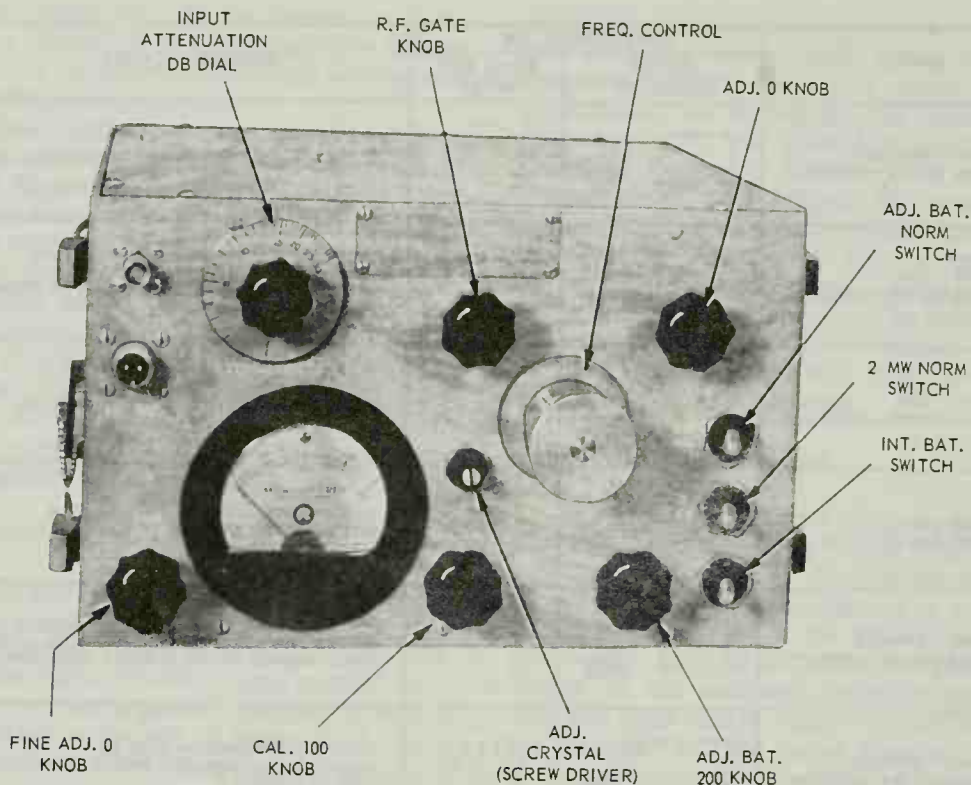
70.49

Figure 3-36.—Front view of Frequency-Power Meter TS-230.

the basic principle of applying the r-f power to be measured to a thermistor and observing the heating effects of that power in changing the thermistor resistance. A thermistor has a high negative temperature coefficient; that is, its resistance decreases rapidly as its temperature increases.
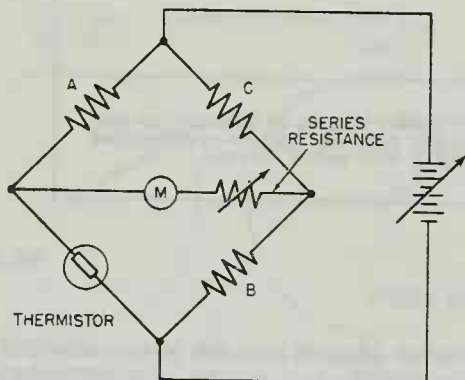
70.50

Figure 3-37.—Thermistor bridge circuit.

The thermistor is used in one arm of a balanced bridge so that any change in its resistance can be detected and measured. The thermistor is placed so that it will absorb r-f energy from the r-f field without applying r-f voltage directly to the bridge. Thus the d-c meter, M, is not subjected to an r-f voltage and the bridge is isolated electrically from the r-f source.

The power required to bring the thermistor to the right resistance for balancing the bridge varies from about 2 to 20 mw, depending on the individual thermistor and the surrounding temperature. This type of bridge is balanced when the meter reads zero. The condition of balance may be expressed mathematically as:

$$\frac{\text{thermistor resistance}}{\text{resistance of A}} = \frac{\text{resistance of B}}{\text{resistance C}}.$$

In the simplified circuit, the bridge is balanced by varying the d-c supply to the bridge, thereby heating the thermistor with enough d-c power to bring it to the right temperature so

that its resistance will balance the bridge (meter reads zero). When the bridge has been balanced, the addition of power (either d-c or r-f) to the thermistor will, through the associated heating effect, unbalance the bridge and cause the meter to move up scale. The meter sensitivity is adjusted so that 1 mw of added power in the thermistor will give a meter reading of 100 (center of dial). Over a limited range (up to 1 mw) the meter reading is directly proportional to the added power.

When measurements are made, the bridge is first balanced by applying the correct amount of d-c power to the thermistor; the resistance in series with the meter is then adjusted so that 1 mw of added power from the d-c supply will cause the meter to read 100. The d-c power is then reduced until the meter deflection is again zero. The r-f power to be measured is applied to the thermistor and the meter deflection again noted. One milliwatt of r-f power is equivalent to one milliwatt of d-c power and will cause the meter to read 100 divisions.

A correct impedance match between the thermistor and the waveguide in the meter is obtained so that substantially all of the r-f power will be absorbed—that is, will be converted to heat energy.

A functional block diagram of the equipment is shown in figure 3-38, A. R-f power is fed to the waveguide by means of an r-f cable adaptor.

The guillotine db input attenuator moves an energy-absorbing element (carbon-coated blade) into the waveguide. The position of the blade is calibrated on the attenuator db dial in terms of loss in decibels.

The thermistor is mounted between top and bottom faces at the center of the waveguide and parallel with the electric lines of force. The bottom end of the thermistor is grounded to the waveguide face, and the top end connects through an r-f by-pass capacitor, which consists of a plastic disc between the outside face of the waveguide and thermistor mount. The d-c power-measuring circuit is connected between this post and ground. Short-circuited coaxial stubs are in the top and bottom faces of the waveguide—one at each end of the thermistor. The bottom stub is tunable; also, the reflector plate at the end of the waveguide section can be tuned by means of a screw. The stub and reflector (in combination with the thermistor adjustment) serve to match the thermistor to

the impedance of the waveguide, thus making the standing-wave ratio satisfactory.

The r-f gate is used for making a check of zero balance in the bridge without disturbing the setting of the attenuator. When the r-f gate plunger in the waveguide is pushed in, nearly all the r-f power will be reflected and will not reach the thermistor.

The thermistor may be heated with power from the battery in the battery case. This is the most convenient method for locations remote from external power, especially when the tests are made during a short period of time. If the internal battery is used for extended periods, its voltage decreases and frequent adjustments are needed during tests. External batteries may be connected to the external battery jack.

In order to measure power, the frequency-power meter is calibrated (as outlined in the instruction book) and the ADJ ZERO control (in the main control circuit from the battery to the bridge network) is set to the position that makes the meter read zero. The frequency-power meter is connected to the radar system; the radar transmitter is turned on and the input attenuator adjusted until a reading of 100 is obtained on microammeter, M. The input attenuator reading then represents the level in db above 1 mw at the input of the waveguide. If the input attenuator dial is turned toward zero and the meter reads progressively less than 100, the meter reading will represent the power absorbed by the thermistor in hundredths of a milliwatt (one division equals 0.01 mw).

To obtain the average power output in the radar waveguide, the input attenuator reading is added to the db loss of the directional coupler (not shown in the figure) between the radar waveguide and the frequency-power meter. The loss in the directional coupler is stamped on the coupler; if a horn pickup antenna is used, the accompanying instruction book will give instructions on figuring losses. The power in watts corresponding to the total dbm (db, above 1 mw) is read directly from a table (not shown). This represents the average power of the radar output.

The meter reading indicates only the average levels of pulsed power throughout the pulse cycle. The peak power can be computed from the average power (Peak-to-Average power conversion method) if the duty cycle ratio is known. The duty-cycle ratio is obtained from the radar technical manual or by assuming a

A FUNCTIONAL BLOCK DIAGRAM

B WAVEMETER, FREQUENCY – POWER METER
TS – 230   EXPLODED VIEW

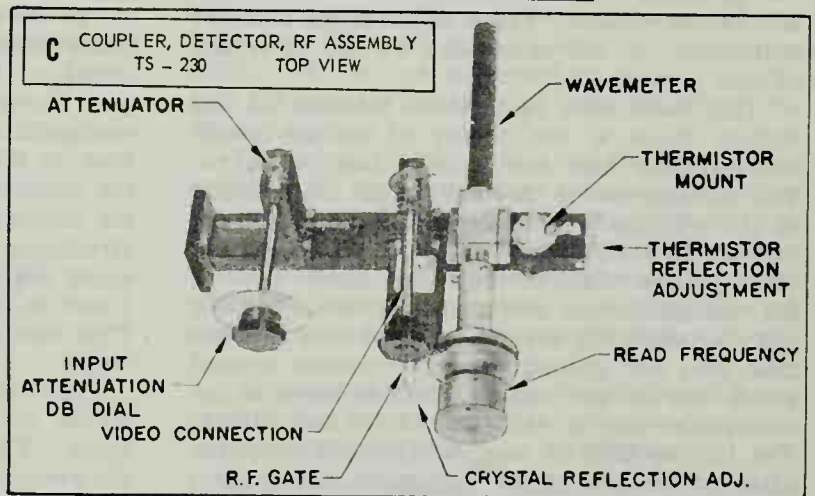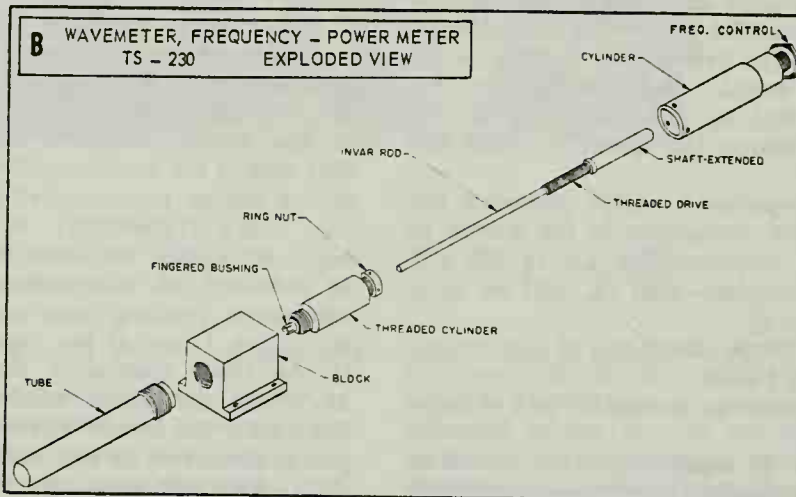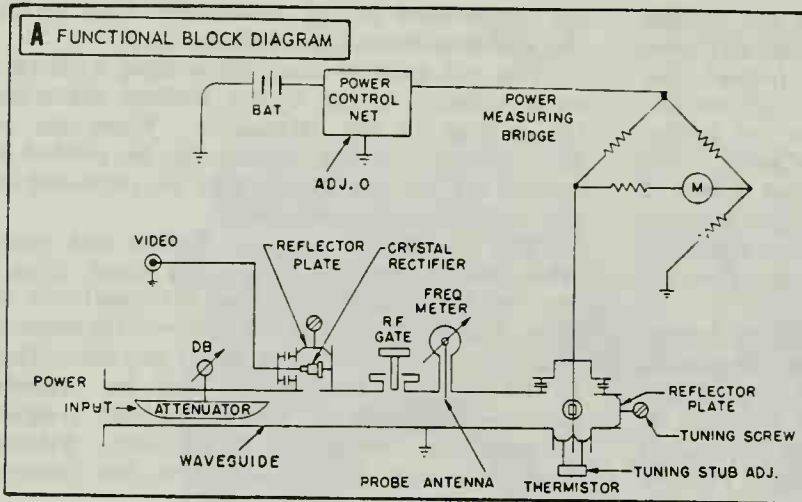C COUPLER, DETECTOR, RF ASSEMBLY
TS – 230   TOP VIEW

70.51

Figure 3-38.—Functional block diagram and r-f plumbing of frequency-power meter.

rectangular pulse shape and using the following formula:

$$\text{Duty-cycle ratio} = \frac{\text{pulse length}}{\text{pulse repetition rate}}$$

Where pulse length is in seconds and the pulse repetition rate is given in number of pulses per second.

In the preceding formula, a perfect square-wave pulse is assumed. For this reason, duty-cycle ratios given in instruction books for particular radars may differ somewhat from these computed values. The instruction book value should be used if it is available.

The peak power is obtained by dividing the average power by the duty-cycle ratio. As an example, let it be assumed that the average power of a radar transmitter, as determined by the frequency-power meter, is 25 w. Assume also that the transmitter radiates pulses 0.5 μs long 600 times per second. The calculations are as follows:

1. The duty-cycle ratio (on a square-wave basis) is the product of the pulse length in microseconds and the repetition rate in pulses per second divided by 10^6. That is,

$$\text{duty-cycle ratio} = .5 \times 10^{-6} \times 600 = 3 \times 10^{-4}$$

2. The peak power is equal to the average power divided by the duty-cycle ratio, or

$$\text{peak power} = \frac{25}{3 \times 10^{-4}} = 83,000 \text{ w}$$

$$\text{peak power} = 83 \text{ kw}$$

TEST SET TS-147/UP

Test Set TS-147/UP (fig. 3-39) is a portable microwave signal generator used for testing and adjusting beacon equipment and radar systems that operate within the band of 8500 to 9600 mc. The set measures the power and frequency of external pulsed and continuous wave signals and supplies output signals that are continuous wave (CW) or frequency modulated (FM). The test set also has a provision to allow for external modulation to produce a pulsed RF signal.

Frequency modulation is produced by applying a sawtooth sweep to a reflex klystron. The amplitude (and slope) of the sawtooth can be continuously varied to give frequency excursions over a range of zero (CW) to over 40 mc.

Average power measurements are made by a temperature compensated thermistor bridge type wattmeter and a calibrated r-f attenuator. The test set can measure the power level of the test signals generated within the instrument and of external signals supplied to the test set. Power is measured in dbm in the ranges of -42 to -85 dbm and from -7 to 45 dbm for test signals supplied by the test set. The range of + 7 to + 30 dbm is for external power supplied to the test set. The accuracy of these measurements is within 1.5 db. These ranges are for the calibration given at the point of the r-f connector and the attenuation of connecting cables and coupling devices must be added to these values.

Frequency measurements are made by an absorption frequency meter that uses a thermistor bridge wattmeter as a tuning indicator. The range of the frequency meter is 8470 to 9630 mc. The accuracy is within 2.5 mc over this range. The frequency difference between any two signals not more than 60 mc apart can be measured to within 1.0 mc. At 9310 mc, the frequency meter accuracy is calibrated to within 1.0 mc.

FUNCTIONAL BLOCK DIAGRAM DESCRIPTION

A block diagram of Test Set TS-147/UP is given in figure 3-40, and the arrangement of r-f plumbing is illustrated in fig. 3-41. The test set receives and measures the power level and frequency of incoming signals and supplies and measures the power level and frequency of outgoing modulated and CW signals.

Normally positive video trigger pulses are applied to the TRIGGER IN connector where they are detected by a crystal diode. The detected r-f pulse is fed to a three stage resistance-coupled amplifier which has a voltage gain of approximately 500. The negative output pulses are capacitively fed to the primary winding of the blocking oscillator transformer and inversion takes place, thus delivering a positive pulse to trigger the blocking oscillator sweep circuit. The output of the blocking oscillator is fed to an integrating network wherein the actual sawtooth waves are formed.

When it is desired to modulate the test set with pulses (square waves, etc.), an external pulse generator producing a negative pulse of
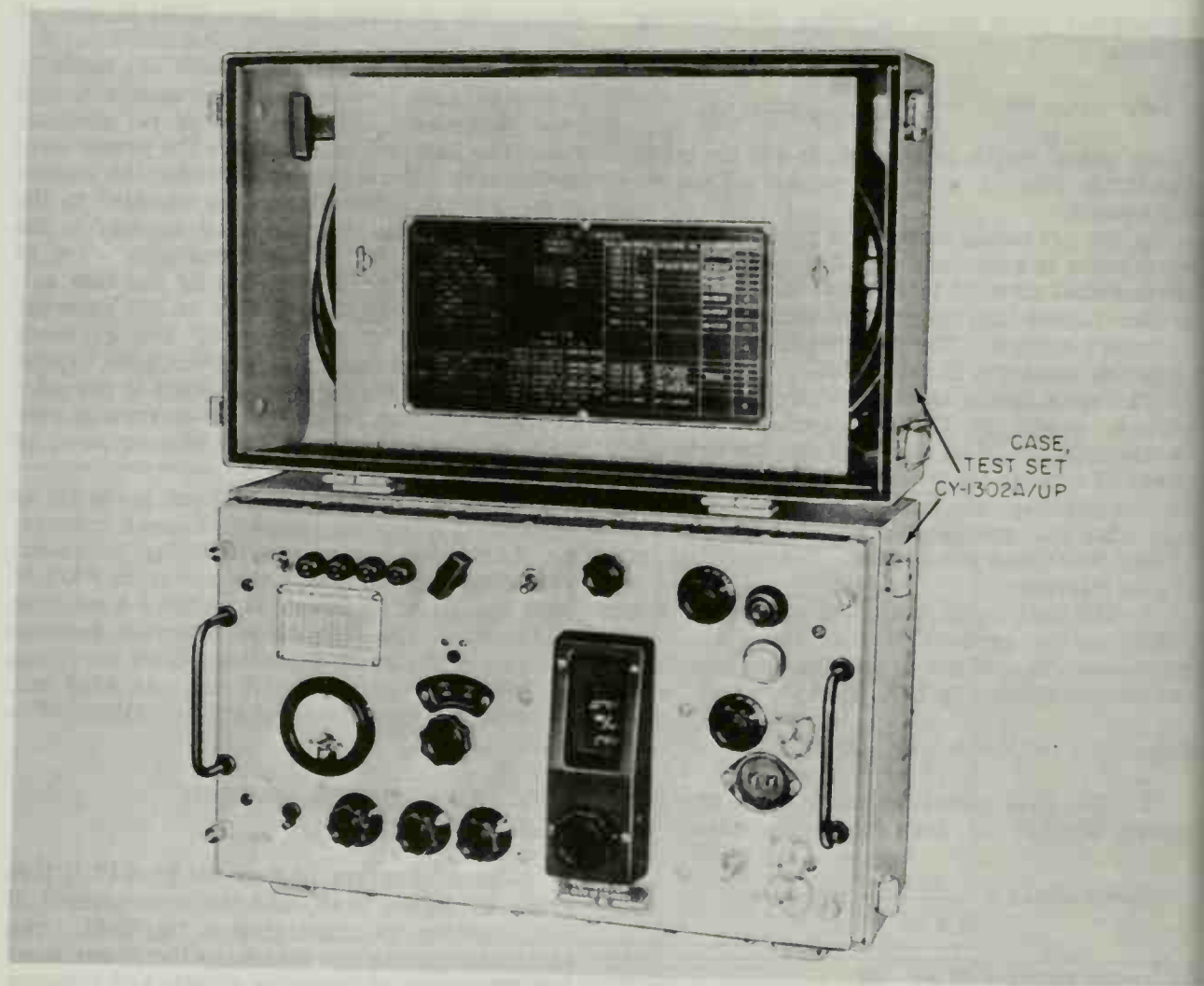
162.9

Figure 3-39.—Test Set TS-147 /UP complete equipment.

approximately 50 volts can be connected to the EXT MOD connector which puts the external modulation onto the reflector of the klystron oscillator.

The oscillator uses a 2K25 reflex klystron tube and is a velocity modulated microwave oscillator, and can be tuned mechanically by varying the volume of the resonator chamber and electrically by varying the reflector voltage.

The Power Set attenuator is a rectangular strip of glass coated with a resistive material that dissipates some of the energy, and is placed longitudinally in the wave guide. Since the electric field strength varies from maximum

at the center to zero at the edge, the power dissipated by the glass strip increases as it is moved toward the center of the wave guide.

The frequency meter assembly consists of a cavity resonator with a dial and a T-section of wave guide that contains the RF cut-off switch. This switch is simply a sliding door that opens or closes the entrance that connects with the oscillator wave guide.

The frequency meter itself is a cylindrical cavity resonator, the volume of which may be varied by moving a plunger in or out and is calibrated to read directly in megacycles/10 (tens of megacycles). That is, each reading
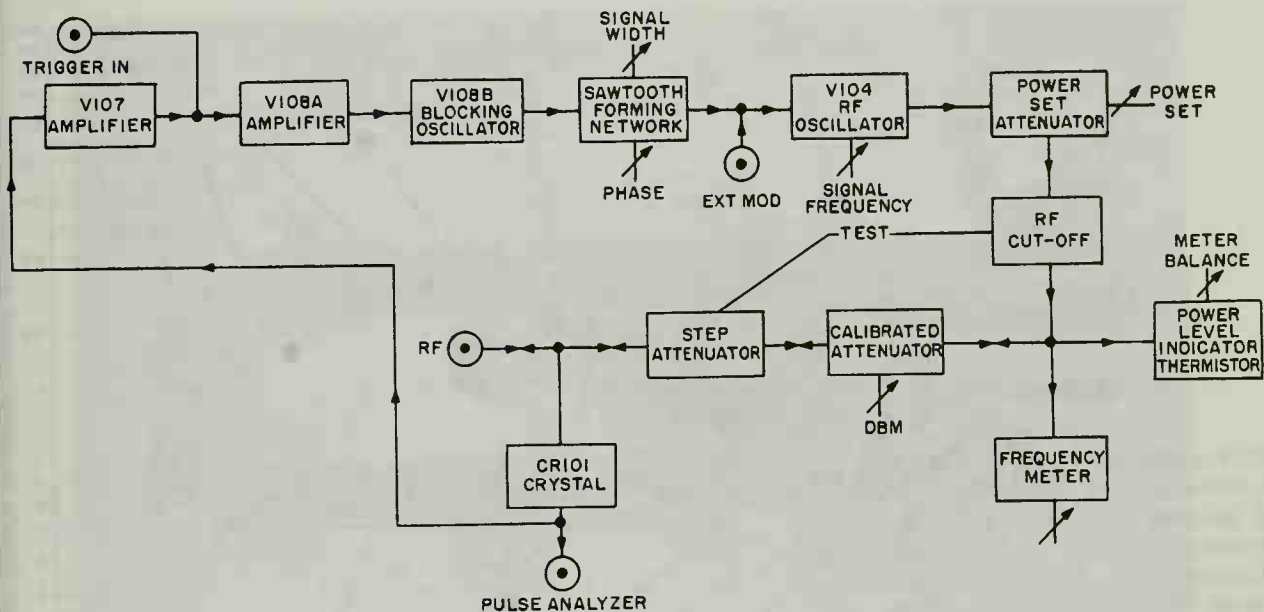
162.10

Figure 3-40.—Block Diagram of Test Set-147 /UP.

must be multiplied by ten to give the frequency in megacycles.

The test set is provided with attenuators so that external sources of r-f signal levels may be measured up to 30 db above 1 milliwatt average; also, so that the internal signal genera- tor may deliver signals at levels down to a value of 85 db below 1 milliwatt. Each attenuator is a thin strip of glass coated with resistive ma- terial and mechanically mounted so that its position in a section of wave guide can be varied.

The step attenuator inserts a fixed 5 db loss or 40 db loss depending upon the position of its front panel control. The variable DBM assem- bly at the counterclockwise setting has a loss of 2 db; and when it is in series with the 5 db minimum of the step attenuator the total loss of both units is 7 db. The dial of the DBM assembly is engraved to show 1 db markings.

The power measuring device of the test set (power level indicator thermistor) is a thermistor bridge wattmeter. It is composed of a Wheatstone bridge circuit having a thermis- tor element in one arm which varies its resis- tance with temperature. A thermistor decreases in resistance as the temperature increases. When current passes through the thermistor, a small amount of power is dissipated in it, which raises the temperature, thus lowering the

resistance. This current may be used by the dc bridge voltage, the r-f field around the ther- mistor, or both. When the dc current in the circuit and the effects of the ambient tempera- ture are properly controlled, the thermistor is used to establish a reference level of power.

The thermistor bridge circuit is designed to read a reference power level of 1 milliwatt and the two attenuators are calibrated. There- fore, by reducing an unknown incoming r-f signal to a power level of one milliwatt, by means of the calibrated attenuators, the power input at the r-f connector can be calculated. Similarly the power output from the klystron oscillator to the attenuator and the thermistor bridge circuit can be adjusted to one milli- watt. The attenuation from that point to the r-f connector is then determined from the settings of the front panel controls:

The frequency of the output signal is meas- ured by the frequefcy meter. When its cavity is tuned to the frequency of the test signal, part of the energy is absorbed, causing a dip in the bridge meter reading and a decrease in the power output of the test set. If the test set is frequency modulated and the frequency meter is tuned within the swept frequency band, power will be absorbed each time the sweep reaches the resonant frequency of the
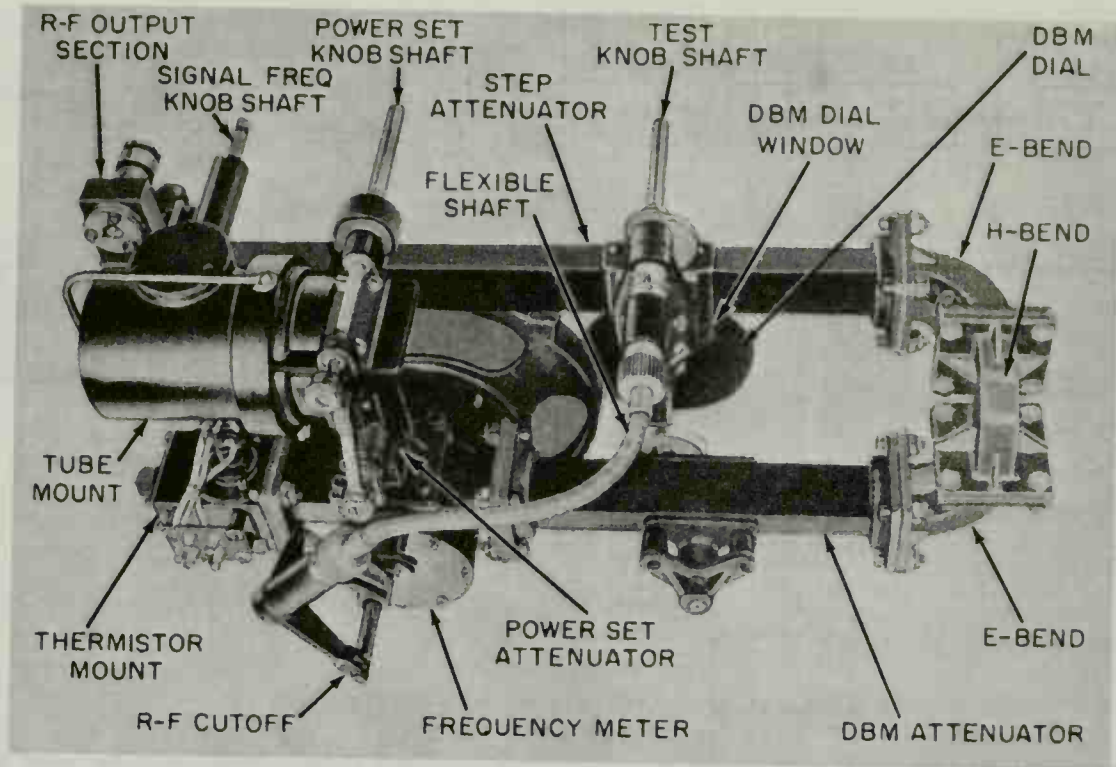
162.11

Figure 3-41.—RF plumbing arrangement.

meter. The resulting decrease in power output will appear as a small dip in the frequency response curve seen in a synchroscope of oscilloscope.

The frequency of input signal is measured in the same way as the frequency of output signal. The energy to be measured is fed into the r-f connector, and the frequency meter absorbs part of the power at resonance causing a dip in the meter reading.

The following example will illustrate the principles of power measurement. Assume that a certain radar transmitter with a built-in directional coupler having a 20-db attenuation is connected to the test set with an r-f cable having a 3.5 db loss.

The test set power-measuring bridge is first properly calibrated. When r-f power is applied, the dbm dial reads 17.5 dbm when the bridge R-F power meter indicates 1 mw. The total db attenuation with reference to 1 mw is 20 db + 3.5 db + 17.5 db = 41 dbm (1-mw reference).

The average power output of the radar transmitter in watts is 41 db above 1 mw. From the dbm-to-watts conversion chart (fig. 3-42), the corresponding average power output of the radar system in watts is approximately 12.5 w.

The peak power is equal to the average power divided by the duty-cycle ratio. The duty-cycle ratio (for a square wave) is equal to the product of the pulse length in microseconds and the pulse repetition rate, PRF, in pulses per second, divided by 10. Expressed as a formula, the peak power is

$$\text{peak power} = \frac{\text{average power}}{\text{duty-cycle ratio}}$$

$$= \frac{\text{average power}}{\text{pulse length x (PRF)}}$$

$$= \frac{\text{average power x } 10^6}{\text{pulse length x (PRF)}}$$

Where the peak power and the average power are in watts.
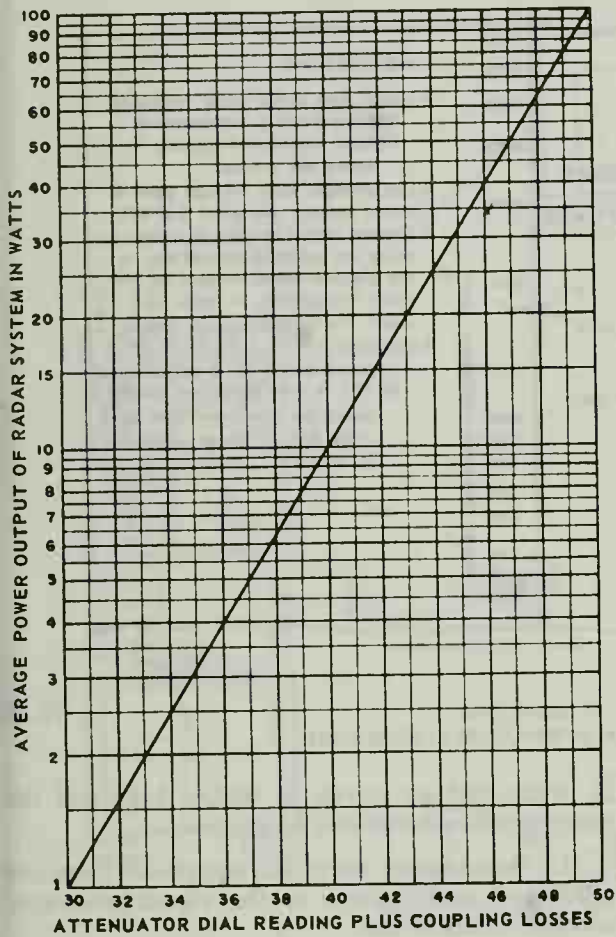
70.55

Figure 3-42.—Dbm-to-watts conversion chart.

If the pulse length is 0.8 μs, the pulse repetition rate is 1706 pulses per second, and the average power is 12.5 w, the peak power will be

$$\text{peak power} = \frac{12.5 \times 10^6}{0.8 \times 1706} = 9150 \text{ w (approx)}$$

A chart for converting average power to peak power is illustrated in figure 3-43. From this chart the db value is indicated for the example being considered as 28.6 db at point C. From this value the peak power is found from the formula,

$$\text{db} = 10 \log \frac{\text{peak power}}{\text{average power}}$$

$$28.65 = 10 \log \frac{\text{peak power}}{\text{average power}}$$

$$\frac{\text{peak power}}{\text{average power}} = 733$$

$$\text{peak power} = 733 \times 12.5 = 9150 \text{ w} \atop \text{(approx)}$$

RADAR PERFORMANCE FIGURE

Although ringtime measurements are valuable in indicating the overall performance of a radar system, they are not as precise as Radar Performance Figure (RPF) measurements. However, neither of these measurements indicates much about how efficient the waveguide or antenna is in performing its function. Ringtime measurements are relatively easy to make; RPF measurements are more difficult to make.

The maximum range of a radar system depends on several factors—for example, (a) transmitter power, (b) receiver sensitivity, (c) the performance of the waveguide and antenna, (d) the effectiveness of the target in reflecting radar energy, and (e) atmospheric conditions.

The first two of the factors listed are especially significant insofar as equipment performance is concerned and are used in determining the RPF of a radar system. The RPF of a radar system is the ratio of the peak (pulse) power, P_p of the radar transmitter to the power of the minimum discernible signal (P_{mds}), expressed in decibels (db) or in decibels with a reference of 1 mw (that is, dbm). Expressed mathematically,

$$\text{RPF (db)} = 10 \log \frac{P_p}{P_{mds}}$$

Because P_{mds} is a very small fraction, its log has a negative sign, and therefore

$$\text{RPF (db)} = 10 \left[\log P_p - (+\log P_{mds}) \right]$$

$$\text{RPF (db)} = 10 \quad \log P_p + 10 \log P_{mds}$$
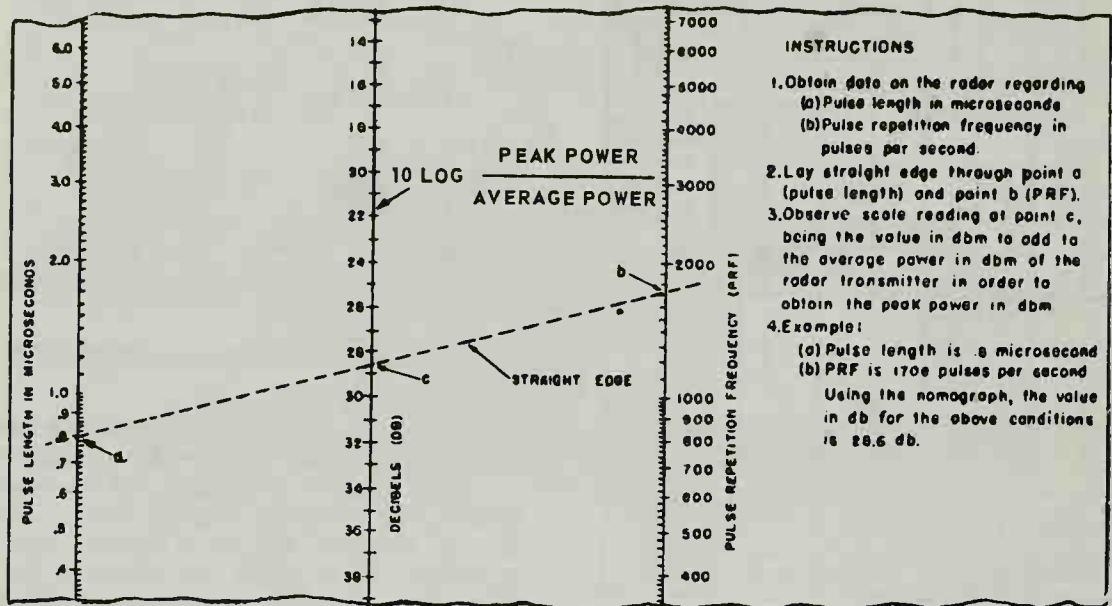
70.56

Figure 3-43.—Average-to-peak power conversion chart.

If the power reference is 1 mw, the equation becomes

$$RPF \text{ (dbm)} = P_p \text{ (dbm)} + P_{mds} \text{ (dbm)}$$

To determine the RPF of a radar system it is necessary first to determine the transmitter average power by means of a power meter (previously discussed) and to convert the average power in dbm to peak power in dbm. The various losses must be taken into consideration in determining the power output. It is also necessary to determine the power of the minimum discernible signal.

The measurement of the minimum discernible signal (MDS) in dbm involves the use of a signal generator and an oscilloscope. The exact method of making the measurement depends on the radar set being tested. For example, the procedure for making MDS measurements on Radar Set AN/SPS-10 is given in the Maintenance Standards Book for the AN/SPS-10.

An external pulse-modulated signal and a method of viewing this signal is required. These requirements are satisfied by the use of Signal Generator TS-419/U and Oscilloscope OS-8B/U or equivalents. The test equipment is connected as shown in figure 3-44 and the correct procedures are as follows:

1. Deenergize the radar equipment (depress S102) and apply power to the signal generator and the oscilloscope.

2. Adjust the signal generator to the assigned transmitter frequency, set the FUNCTION switch to ZERO SET and adjust the ZERO SET control until the meter indication is zero.

3. Set the FUNCTION switch to CW and adjust the POWER SET control until the meter pointer is at the power set marker.

4. Set the FUNCTION SELECTOR switch to the RATE x 10 position, the PULSE WIDTH control to 2 and the PULSE RATE control to 30.

5. Make certain that all connections to the test equipment are correct, and that the oscilloscope is adjusted for maximum clarity of presentation.

6. Depress MAIN POWER ON switch (S101) on the radar set, and set MANUAL-AFC switch to MANUAL. The RADIATION switch (S104) is OFF.

7. While observing the pulse-modulated signal on the oscilloscope, slowly decrease the

102

RADAR RECEIVER—TRANSMITTER
RT—141A/SPS—6

S101 S102

S104

TERMINAL BOARD E1S8

SIGNAL FREQUENCY

ZERO SET

FUNCTION
SWITCH

OUTPUT
ATTENUATOR

RF OUT

PULSE
WIDTH

SIGNAL GENERATOR
TS—419/U

POWER   SYNC   SYNC
ON—OFF   IN   OUT

OSCILLOSCOPE
OS—8/U

TRIGGER PULSE J107

70.60
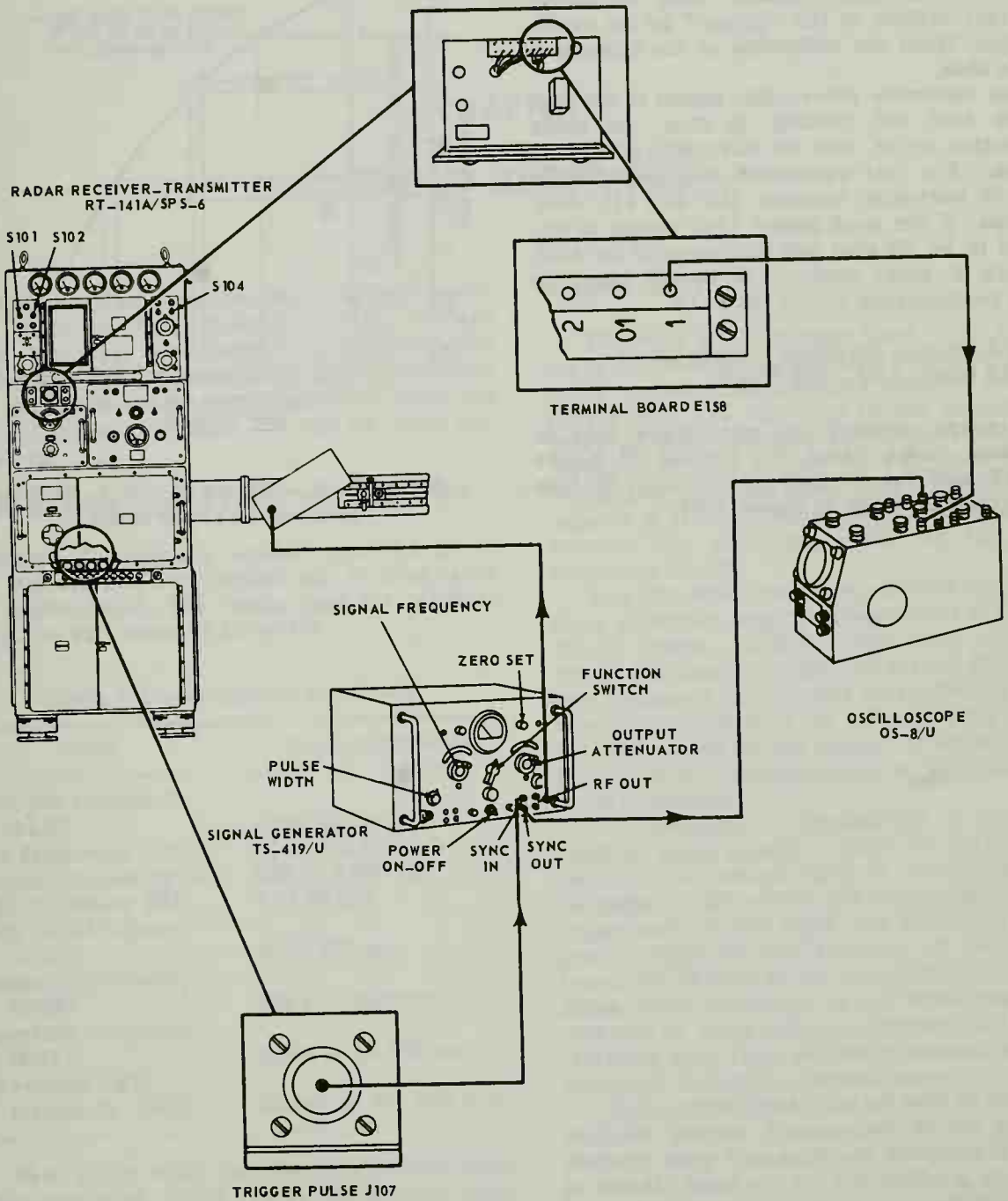
Figure 3-44.—Test setup for making MDS measurements.

output of the signal generator with the OUT-PUT ATTENUATOR control until the signal is barely visible in the "grass" on the oscilloscope. Read the indication of the attenuator dial in dbm.

The minimum discernible signal is the sum of the test set reading in dbm, the cable attenuation in db, and the waveguide connector in dbm. For this equipment, the power of the MDS is normally between 100 and 115 dbm.

Thus, if the peak power (Pp) output is assumed to be 90 dbm and the power of the MDS (that is P mds) assumed to be 105 dbm, the radar performance figure (RPF) is

RPF (dbm) = Pp (dbm) + Pmds (dbm)
RPF (dbm) = 90 + 105 = 195

A graph showing the percentage loss in maximum radar range for various db losses in RPF and for various percentages of loss in ringtime is shown in figure 3-45.
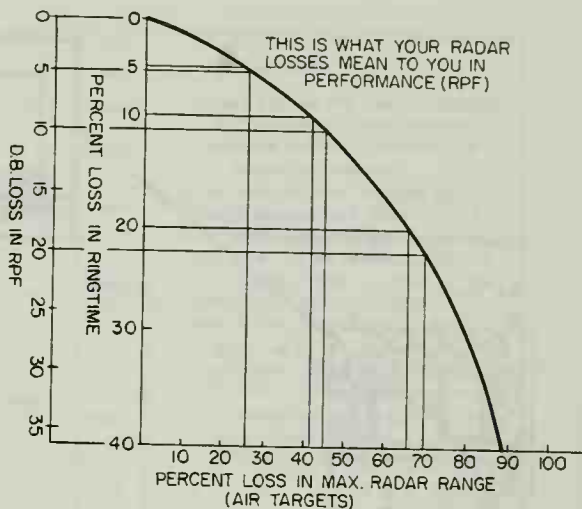
70.61

Figure 3-45.—Effects of loss in RDF and ringtime on radar range.

CHAPTER 4

RADIO RECEIVERS

This chapter describes several representative types of Navy radio receivers. Modern Navy receivers use rugged components, that utilize circuits to make them easy to operate and maintain. They are capable of receiving several types of signals and can be tuned accurately.

RADIO COMMUNICATIONS FREQUENCIES

Radio frequencies extend from 10 kc to 300,000 mc and are divided into various bands for convenience. The bands used for military purposes are shown in table 4-1.

Table 4-1.—Frequency Spectrum

Band	Frequency range
Very low frequency (VLF)	10 to 30 kc
Low frequency (LF)	30 to 300 kc
Medium frequency (MF)	300 to 3,000 kc
High frequency (HF)	3 to 30 mc
Very high frequency (VHF)	30 to 300 mc
Ultrahigh frequency (UHF)	300 to 3,000 mc
Superhigh frequency (SHF)	3,000 to 30,000 mc
Extremely high frequency (EHF)	30,000 to 300,000 mc

Each of the bands have certain transmission characteristics, which are peculiar to that band. Occasionally the characteristics within a band may differ, but generally they are consistent. The terms ''long range, medium range, and short range'', correspond to 1500 miles, 200 to 1500 miles, and under 200 miles, respectively (table 4-2).

WHERE USED

The very low frequency band (from 10 to 30 kc) is not covered by shipboard transmitters, because transmitting antennas required for these frequencies are too long to use aboard ship. However, shore stations, such as NAVCOMMSTA Washington (NSS) employ the VLF band for fleet broadcasts. Stations using such frequencies are capable of transmitting signals through magnetic storms that ordinarily blank out the higher frequency bands.

The direction-finding equipment used by the Navy generally employs frequencies in the 30-to 300-kc range. Reliable long-range communication can usually be obtained in the LF band with low frequency equipments and antennas aboard ship. Frequencies in the LF band (like those in the VLF band) do not depend on skywaves, and provide stable communications with little seasonal variation.

Transmission of frequencies in the 300- to 3000-kc range usually employs the use of groundwaves even though skywave transmission and reception of MF waves are also possible. At the upper end of this band, the ionosphere has a great effect on the skywave. If the correct frequency is used at the proper time, transmissions over relatively large distances can be obtained by using this band. Propagation characteristics vary from season to season, and from nighttime to daytime transmission.

Commercial broadcast as well as some Navy stations employ frequencies in the MF band. Modern Navy transmitters designed to operate in the MF band are the AN/SRT-14, 15, and 16. Special circuits for these transmitters are described in chapter 6 of this training course.

The HF band includes frequencies from 3 to 30 mc. The skywave is used for long-range transmission in this band. Propagation characteristics of frequencies in this range also

Table 4-2.—Frequency Transmission Characteristics

| Band | Range | | Power required | Antenna length requirements |
	Day	Night		
VLF	Long	Long	Very high	Very low
LF	Long	Long	Very high	Long
MF	Medium	Long	High to medium	Long
HF (3 to 10 mc)	Short	Medium to long	Medium	Medium
HF (10 to 30 mc)	Long	Short	Low	Short
VHF	Short	Short	Low	Very Short

change with the time of day and the season of the year.

The HF band is divided into two parts (see table 4-2) to show the characteristics of each part.

The antennas used in the 30- to 300-mc band require less space than lower frequency antennas and thus have an advantage for shipboard use. In addition, the power required for transmission in the VHF range is relatively low.

For frequencies greater than 30 mc, the surface wave component of the ground wave is rapidly attenuated through the dielectric constant of the earth. For this reason, the direct wave (that component of the groundwave traveling directly from the transmitting antenna to the receiving antenna) is the primary component of VHF propagation. The distance covered by a given transmission in this range however, is not always limited to the distance of the horizon, or the line-of-sight distance from the transmitter. Research has shown the early line-of-sight theory to be frequently in error because radio waves of these frequencies may be refracted.

From this discussion, it is easy to see that several factors must be considered when a certain frequency is to be selected for transmission. The Electronics Technician has little to say about the choice of this frequency. A knowledge of the factors discussed, however, should be helpful in determining why that particular frequency was chosen.

RADIO RECEIVING SETS AN/SRR 11, 12, 13, AND 13A

The AN/SRR-11, 12, 13, and 13A receivers represent equipments widely used in the Navy and cover the frequency range from 14 kc to 32 mc. They consist of several major assemblies,

each physically independent of the other. Individual subassemblies, however, are mounted with each assembly, and they plug into an appropriate socket of their major assemblies. When it is necessary to make repairs quickly, a defective assembly or subassembly can be replaced with a spare one.

The AN/SRR-11, 12, 13 and 13A receivers (fig. 4-1) are designed for tabletop mountings. Each is self-contained in a metal case with operating controls, switches, meters (tuning and output), phone jack connectors, and main tuning dial assembly mounted on the front panel. The receiver chassis is mounted on rails in the receiver cabinet and can be withdrawn and positioned 45° or 90° (up or down) for ease in servicing.

The frequency range of each receiver is divided into five bands. Continuous tuning is available throughout each band. The frequency range of the AN/SRR-11 receiver is from 14 to 600 kc (low frequency), that of the AN/SRR-12 receiver is from 0.25 to 8 mc (medium frequency), and the range of the AN/SRR-13 receiver is from 2 to 32 mc (high frequency).

The AN/SRR-11 receiver is provided with circuits and switching to receive A1 (c-w telegraphy) signals, A2 (m-c-w telegraphy) signals, and F1 (telegraphy by frequency-shift keying) signals. In addition, the AN/SRR-12 and 13 receivers are capable of receiving A3 (radiophone) emissions.

The basic receiver represented by the AN/SRR-11, 12 and 13 sets is a double super-heterodyne having two stages of r-f amplification ahead of the first mixer and oscillator.

A first intermediate frequency amplifier (which is essentially the second converter) follows, and its output is fed to the second i-f assembly. Signals of A2 (m-c-w) and A3 (phone) emissions are rectified in a diode
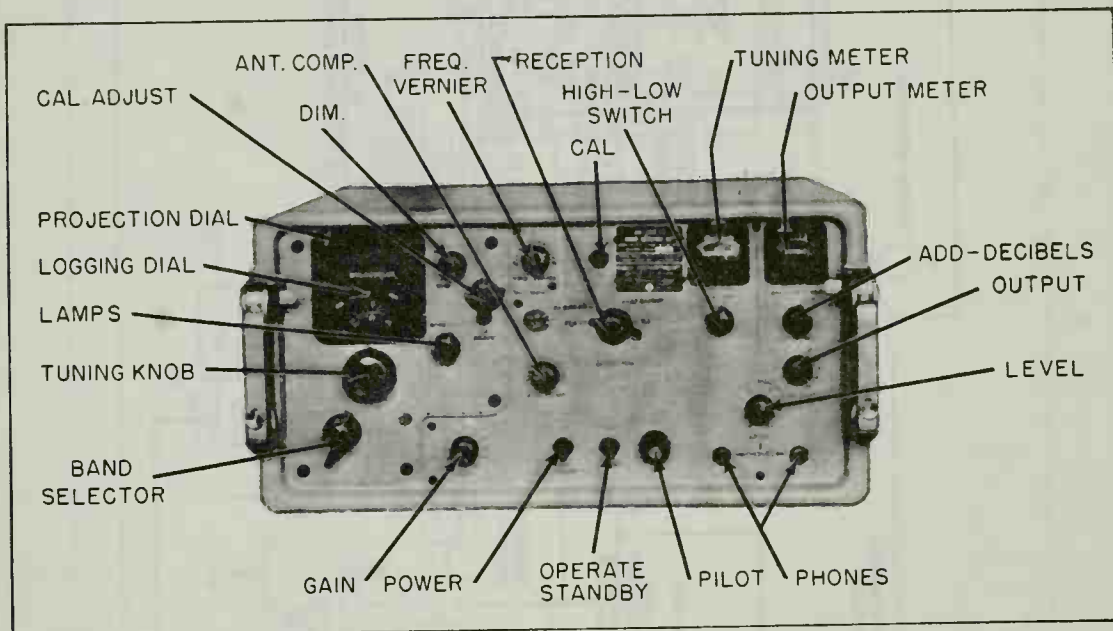
A AN/SRR 11

B AN/SRR 12-13-13A

1.159
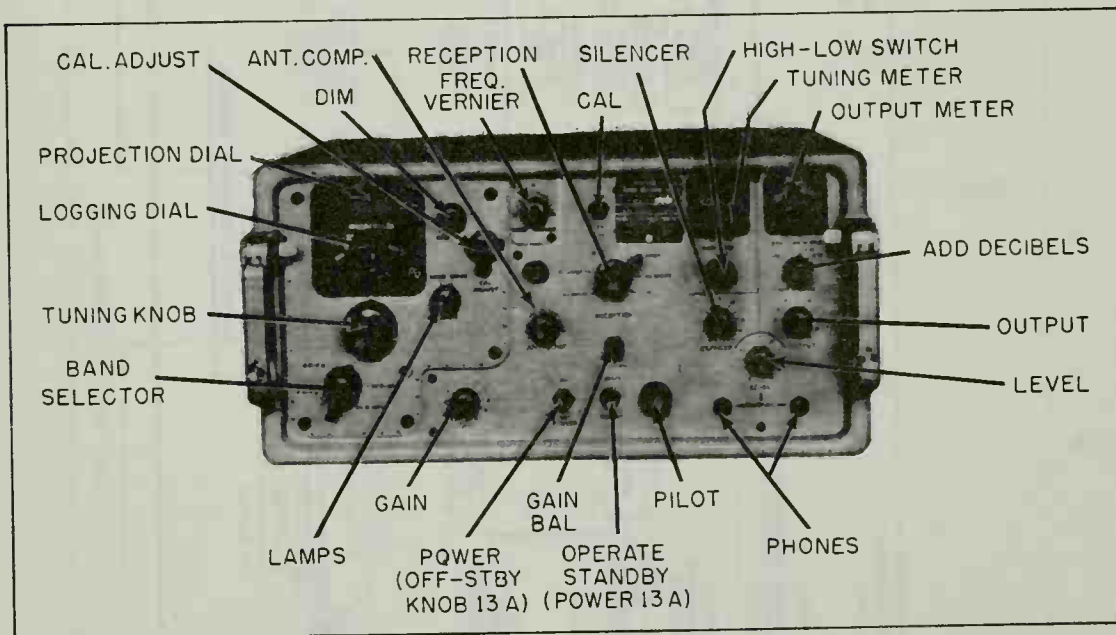
Figure 4-1.—Radio receiving sets AN/SRR-11, 12, 13, and 13-A front panel controls.
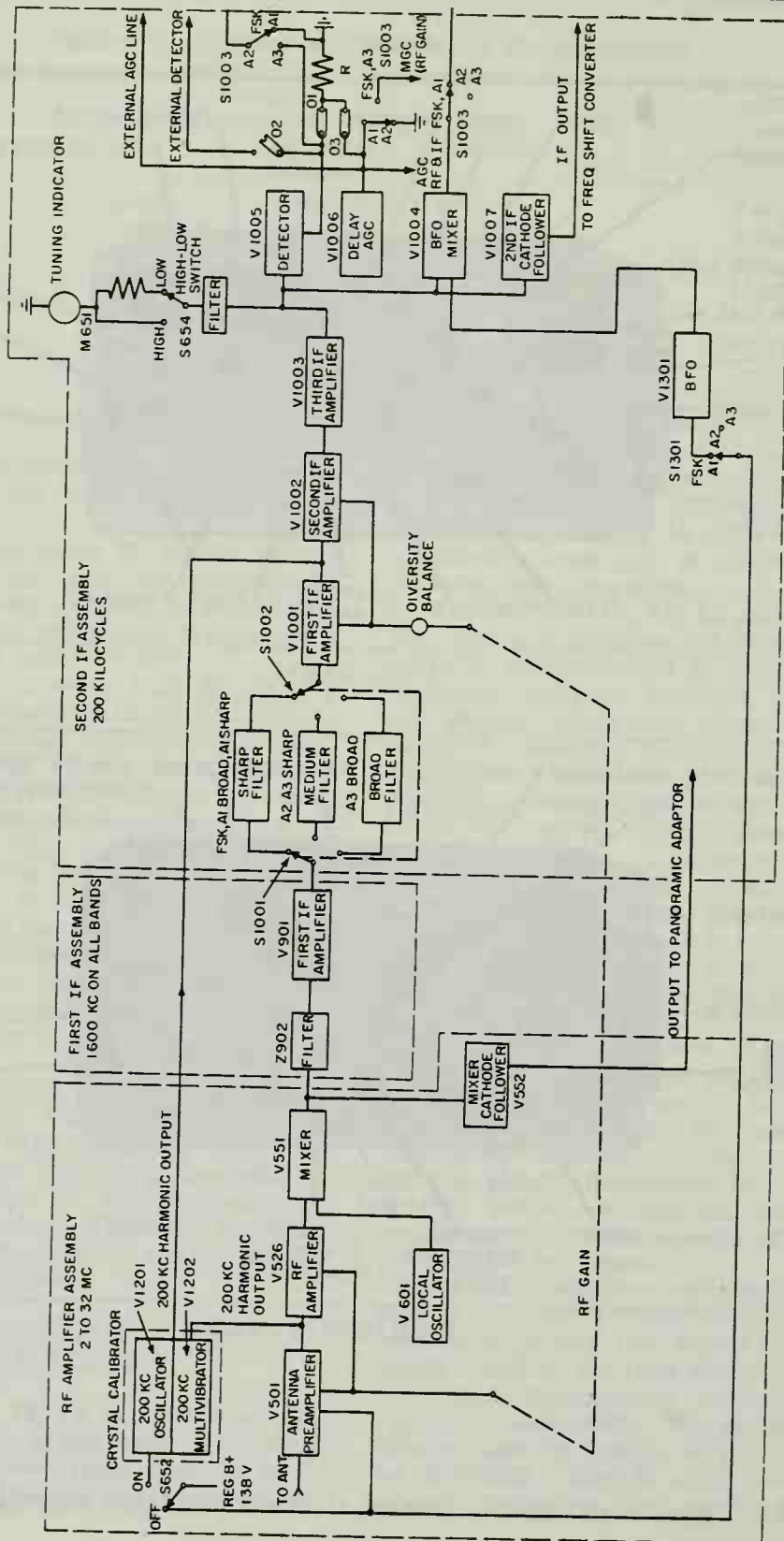
Figure 4-2. —Block diagram of radio receiving set AN/SRR-13A.

50. 31
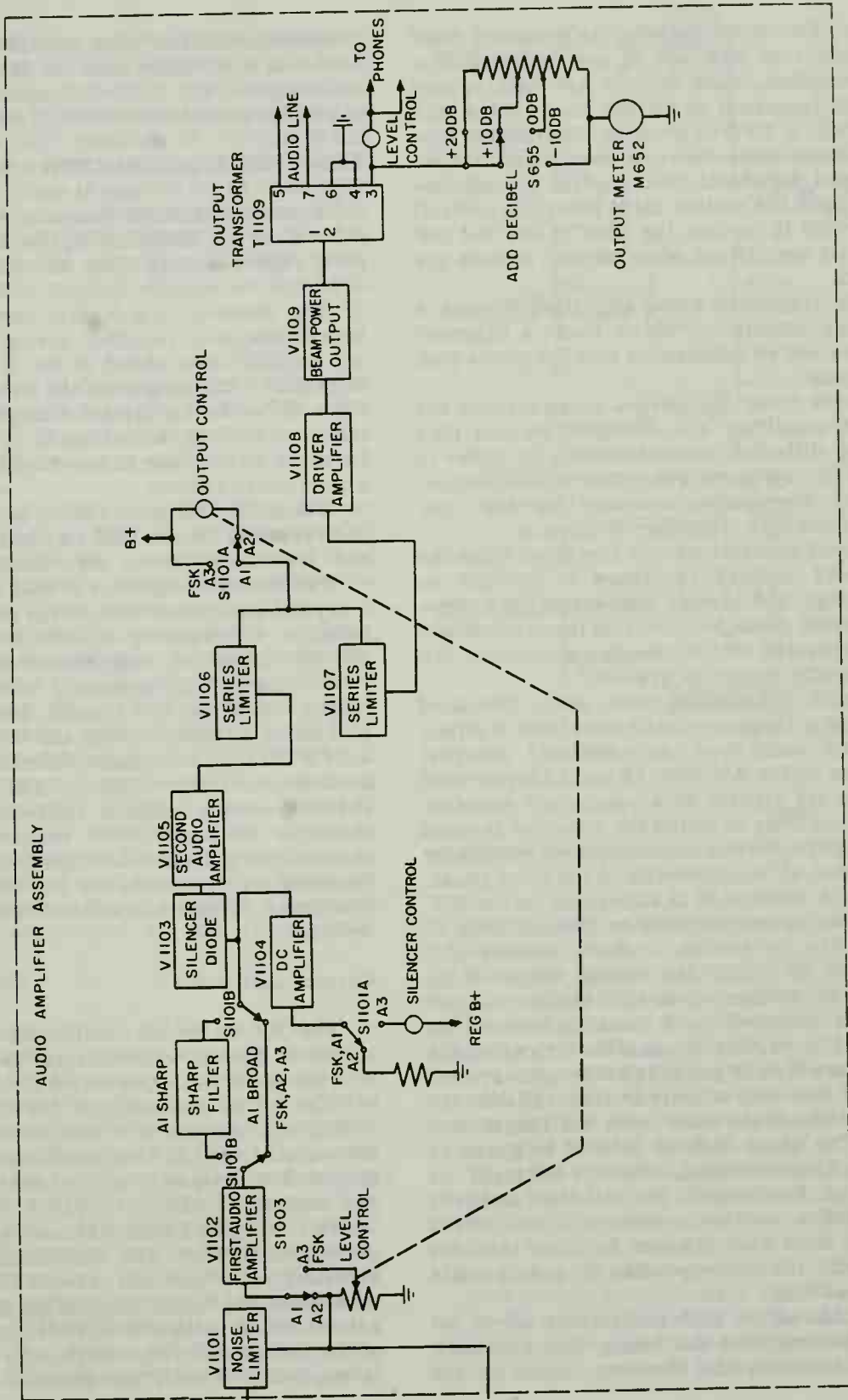
Figure 4-2.—Block diagram of radio receiving set AN/SRR-13A—Continued

50. 31

detector. The diode detector is bypassed when the receiver is switched to c-w or frequency-shift reception. Such signals are heterodyned in a beat frequency oscillator mixer stage with those from a BFO to produce a difference frequency in the audio range. A manual gain control or delayed automatic gain control circuit (depending upon the setting of the reception control) is provided to reduce the gain of the r-f and second i-f amplifiers when strong signals are received.

Audio frequencies are amplified through a four-stage amplifier, which feeds a balanced audio line and an unbalanced line for phone jack connections.

Circuits from the antenna stage through the first i-f amplifier are designed so that they will have different characteristics in order to provide for optimum reception on the several bands of frequencies covered by the low, medium, and high frequency receivers.

External connections are furnished from the AN/SRR-12 and 13 receivers to provide an AGC voltage and circuit connection to a common external diode load, so that these receivers may be operated with a companion receiver in a dual diversity reception system.

External connections are also furnished for feeding a frequency-shift converter of either the i-f or audio type. An additional external connection on the AN/SRR-12 and 13 is provided to supply i-f signals to a panoramic adapter.

The frequency to which the receiver is tuned appears projected on a translucent screen (tuning dial) located at the upper left of the front panel. The dial is calibrated in kilocycles on the AN/SRR-11 and in megacycles on the AN/SRR-12 and 13. The projection system consists of a glass disk on which the tuning range of the receiver is calibrated in five scales. A light source is mounted in a housing behind the tuning dial. A portion of one of the optical scales on the glass disk is projected through a system of lenses and onto a mirror that reflects the image of the scale back upon the translucent screen. The glass disk is rotated by gears as the tuning knob is turned. When the setting of the band switch is changed, the dial light and lens housing are moved up or down by a cam driven from the band switch gears to align the lens system with the corresponding frequency scale on the glass disk.

The CAL adjust knob and locking screw located to the right of the tuning dial provide a means of shifting the frequency scale on the translucent screen when recalibrating the receiver in conjunction with the crystal controlled calibrator. This action is accomplished by a slight horizontal movement of the lens system.

FUNCTIONAL DESCRIPTION AN/SRR-13A

A functional block diagram discussion (fig. 4-2 of the AN/SRR-13A is given as a description of a representative Navy communications receiver.

The receiver is a double-conversion superheterodyne a-m receiver having two stages of r-f amplification ahead of the first mixer and oscillator. The output of the first mixer is fed via a filter to the first i-f amplifier which is tuned to 1600 kc on all bands. A second output from the first mixer is provided as an output to a panoramic adaptor.

The local oscillator (V601) (a Hartley) in the r-f assembly tracks 1600 kc higher than the signal frequency on all five bands (2-32 mc).

The first i-f amplifier (V901) also serves as a crystal controlled Pierce type oscillator operating at a frequency of 1400 kc to produce a 200-kc output to the second i-f assembly.

The second i-f assembly includes 3 filters (sharp, medium, and broad), 3 stages of amplification (V1001, V1002, and V1003 at 200 kc, a BFO (V1301), and its associated mixer (V1004), a diode detector (V1005), AGC delay diode (V1006), and a cathode follower (V1007) to supply a frequency shift converter or other accessory equipment. Link connections are also provided for connecting the AN/SRR-13A with a companion receiver in a dual diversity reception system.

Crystal Calibrator

The 200-kc crystal calibrator (V1201-1), is a cathode coupled multivibrator locked to 200 kc by a 200-kc crystal. One output rich in harmonics of 200 kc is fed through the r-f amplifier (V526) and combined in the mixer (V551) with the signal from the local oscillator (V601). The combination frequency signal goes through S1001 and associated filters to V1001 then to V1002 where it combines with a second output from the crystal calibrator. The combination of the two signals gives a beat note at every 200-kc interval throughout the tuning range of the receiver. (The second output is taken from the common cathodes of the 200-kc multivibrator, and its 4th harmonic is applied to V1002, the second i-f assembly.)

The main tuning dial can thereby be calibrated independently of external signals if the CAL switch (S-652) is turned to the ON position.

The crystal calibrate switch is connected so that in the OFF position, B + voltage is not applied to the crystal calibrator. In the ON position, B + voltage is applied to the calibrator and is removed from the antenna preamplifier (V501) and BFO (V1301) thereby preventing interference with the crystal calibrator from either externally applied signals or the BFO.

Audio Section

Signals at 200 kc from the plate of V1003 are applied to the diode detector, V1005. The audio output of V1005 is applied to the noise limiter V1101 when the reception control is in the A2 or A3 position.

The diode detector voltage developed across R is applied through link 03 to the delay AGC diode, V1006, when the reception control is in the FSK or A3 position. Delayed AGC is effective only for F1 and A3 reception.

Links 01, 02, and 03 are provided for applying a rectified signal voltage and AGC voltage to a second AN/SRR-12 or 13 designated as the companion receiver for dual diversity reception. With the diversity connection, the diode load resistor and the delayed AGC circuit of the local receiver are made common to the companion receiver. In diversity operation the gain of both receivers must be balanced. This action is accomplished by regulating the amount of gain in V1001 and V1002 of each receiver by adjustment of the Diversity Gain Balance control, a recessed screwdriver adjustment on the front panel.

Silencer Circuit

The silencer circuit (V1103 and V1104) is active in the audio assembly when the reception control is in the A3 sharp or A3 broad position to prevent noise from being further amplified by succeeding stages in the absence of signals.

The silencer diode V1103 (fig. 4-3) is connected in series between the first and second audio amplifiers V1102 and V1105. Plate voltage on the diode V1103 depends on the magnitude of plate current through the d-c amplifier V1104. The cathode voltage of the diode V1103 is determined by voltage divider R1115 and R1116. Grid voltage of the d-c amplifier V1104, is determined by the setting of the silencer potentiometer R656 (mounted on the front panel
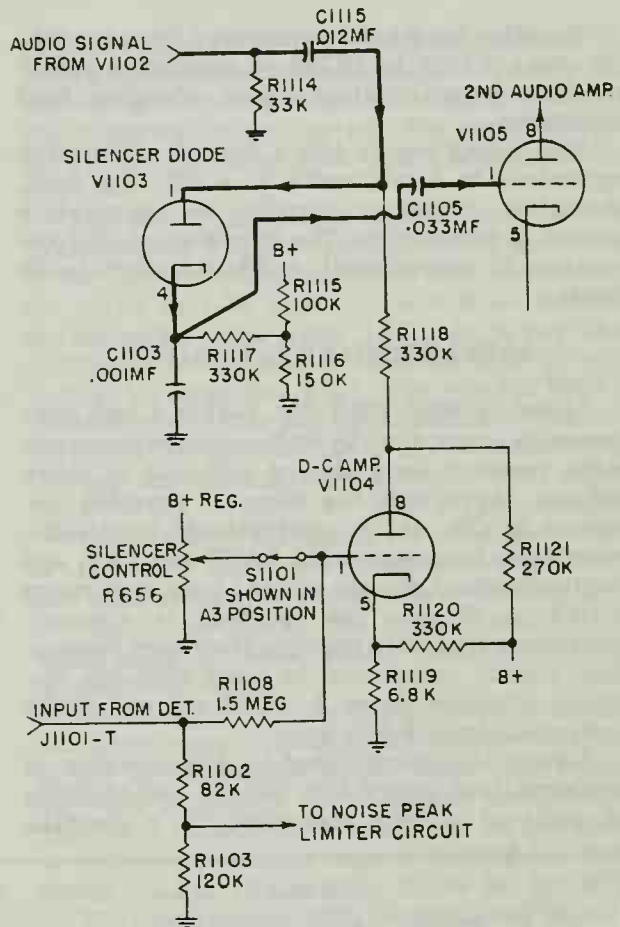
50.38

Figure 4-3.—Silencer circuit, simplified schematic.

of the receiver) and by the average d-c level of the detector output.

When the grid voltage of V1104 is positive, the plate of V1103 becomes negative with respect to its cathode, and the diode cannot conduct. When the grid voltage of V1104 is reduced (as by the presence of a strong carrier), the triode plate current is reduced and the plate of V1103 becomes positive with respect to its cathode, thus causing the diode to conduct.

When the reception control is set to any other position but A3 sharp or A3 broad, the silencer control is inoperative because switch S1101A returns the grid circuit of V1104 to ground instead of to the silencer control.

Two series diode limiters (V1106 and V1107, fig. 4-2), are combined to clip equally on positive and negative portions of the audio cycle when the reception control is set for A1 or A2.

Negative feedback is employed from the output stage V1109 to V1108 to maintain a nearly constant output voltage under changing load conditions.

The output meter has a zero db calibration equivalent to 6 milliwatts in a 600-ohm load. Multiplier resistors provide an appropriate amount of attenuation. The power supply for the receiver is conventional, and is not shown in the figure.

RADIO RECEIVER R-390A/URR

Model R-390A/URR (fig. 4-4) is a high performance, exceptionally stable, general-purpose radio receiver used aboard ship and at shore stations throughout the Navy. It provides reception of CW, MCW, conventional amplitude-modulated, frequency-shift RATT and FAX, and single-sideband signals within a frequency range of 0.5 to 32 mc. The receiver is a superheterodyne type with multiple frequency conversion. Double conversion is used when the receiver operates from 8 to 32 mc and triple conversion from 0.5 to 8 mc.

Tuning is accomplished by the insertion of powdered-iron cores into the r-f and variable i-f coils at a rate controlled by a complex mechanical arrangement of gears, shafts, and cams. The frequency is indicated by a counter-type indicator that is accurate to within 300 cps, an accuracy that permits use of the receiver as an accurate frequency meter.

RECEIVER BLOCK DIAGRAM

Figure 4-5 shows the path of the receiver signals from the antenna input to the audio output and i-f output.

Radiofrequency signals are fed into the receiver by either a balanced two-wire antenna such as a doublet, or an unbalanced antenna such as a whip or long wire. Antenna relay K101 disconnects and grounds the antenna during standby calibration, or break-in operation. When K101 is not energized, the balanced antenna is connected to the input of one of the antenna transformers, T201 through T206, which is selected by the megacycle change switch. The transformers are tuned by the megacycle change and the kilocycle change tuning controls. The output of these tuned circuits is fed into r-f amplifier V201. If an unbalanced antenna is used, the signal is fed directly to the grid circuit of the r-f amplifier.
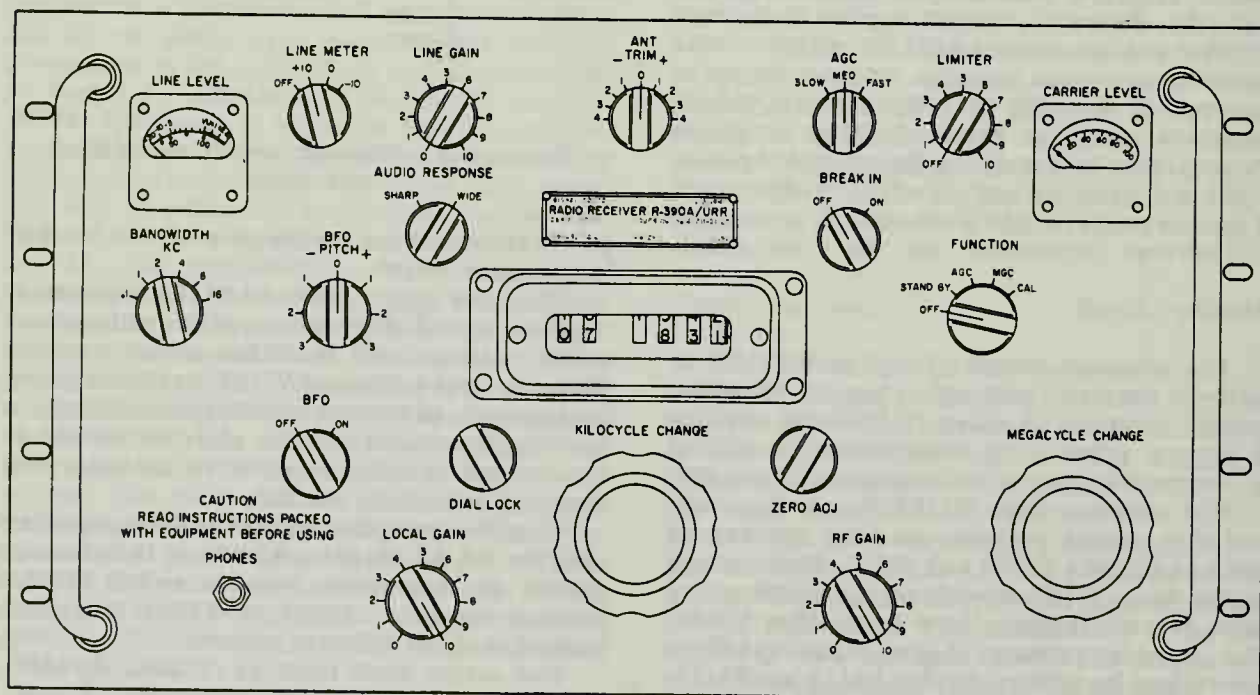

13.68

Figure 4-4.—R-390A/URR radio receiver.

The calibration circuit, consisting of 200-kc crystal calibration oscillator V205A, 100-kc multivibrator V206, and 100-kc cathode follower V205B, injects 100-kc markers into the input circuit of r-f amplifier V201. When the function switch is in the CAL position, B+ is connected to the calibration circuit.

Radiofrequency amplifier V201 amplifies the signals from the antenna before they are fed to first mixer V202. Six tuned circuits are selected by a switch that is connected to the megacycle change control. The tuned circuits are adjusted by the kilocycle change and megacycle change controls. The frequency range of tuned circuits Z201 through Z206 is 0.5 to 32 mc.

Signals from 0.5 to 8 mc are coupled from r-f amplifier V201 and fed into the input circuit of first mixer V202. The 8- to 32-mc signals from V201 are switched around the first mixer and fed directly into second mixer V203. When the receiver is operated between 0.5 and 8 mc, the first mixer combines the r-f signals with a 17-mc signal from first crystal oscillator V207. The output (sum) frequency is the first variable i-f signal, and its frequency varies from 17.5 to 25 mc. Tuned circuit Z213 is a triple-tuned device that is permeability tuned as the megacycle change and the kilocycle change controls are operated.

The 17.5- to 25-mc output of the first mixer stage (when the receiver is tuned from 0.5 to 8 mc) or the 8- to 32-mc output from r-f amplifier V201 (when the receiver is tuned from 8 to 32 mc) is selected by the operation of the megacycle change control and fed into second mixer V203. Also fed into V203 is a signal from crystal oscillator V401. The output frequency of V401 is changed as the megacycle change control is operated so that the difference between signal frequency and oscillator frequency is always between 2 and 3 mc. When the frequency reading on the receiver dial is between 0.5 and 0.999 mc, V401 feeds a 20-mc signal into V203. The output frequency of V203 thus varies downward from 3 to 2 mc, except on the 0.5- and 1-mc band, where the output frequency of V203 varies between 2.5 and 2 mc. This is the second variable i-f signal. Another set of triple-tuned coils, Z216, tunes the output of V203 as the kilocycle change control is operated.

The 3- to 2-mc output of the second mixer and its tuned output circuit Z216 is fed into the input circuit of third mixer V204. This stage beats the 3- to 2-mc signal with a continuously variable signal from variable-frequency oscillator V701. This precision oscillator has accuracy comparable with a frequency standard and is controlled by the kilocycle change control. The output of V204 is a fixed frequency of 455 kc. This is the third i-f.

The 455-kc signal is fed into or around 455-kc crystal filter Y501, as determined by the setting of the bandwidth control. When this control is in the .1-kc or 1-kc position, the 455-kc signal is fed through crystal filter Y501, and in any other position it is bypassed around the filter directly to first i-f amplifier V501. The output of V501 is fed into one of four mechanical filters. The output from the selected filter is fed successively through the second, third, and fourth i-f amplifiers (V502, V503, and V504, respectively).

The output of the fourth i-f amplifier, V504, is fed into half-wave diode detector V506B and audiofrequency voltage is produced. This a-f signal is fed to the limiter stage V507, which clips off noise peaks that might render reception difficult.

Beat-frequency oscillator V505 generates and feeds signals variable in frequency from 452 to 458 kc to detector stage V506B. The resultant beat frequency is adjustable continuously from zero to 3000 cps whether the BFO is above or below the carrier signal. Setting the BFO switch to the ON position turns on the BFO.

The audio output from limiter V507 is fed to the first a-f amplifier V601A. This stage amplifies the audio signal and passes it through or around 800-cps bandpass filter FL601, depending upon the setting of the audio response switch. The audio signal is then fed to a-f cathode follower V601B. This stage feeds the audio signals to the local and line audio channels. The local audio channel consists of local a-f amplifier V602A and local a-f output tube V603. This audio source is used for 600-ohm headsets or loudspeakers. The line audio channel is similar to the local audio channel, consisting of line a-f amplifier V602B and line a-f output tube V604. The output provided matches a balanced 600-ohm line.

Intermediate-frequency signals (455 kc) present at the input to fourth i-f amplifier V504 are also fed to the input of i-f cathode follower V509B. This stage provides a 50-ohm source of 455-kc signals for use with a frequency-shift converter for RATT operation. The i-f output connector for this type of service is located on the rear panel of the receiver. Tube V509B also feeds 455-kc signals into the AGC circuit.
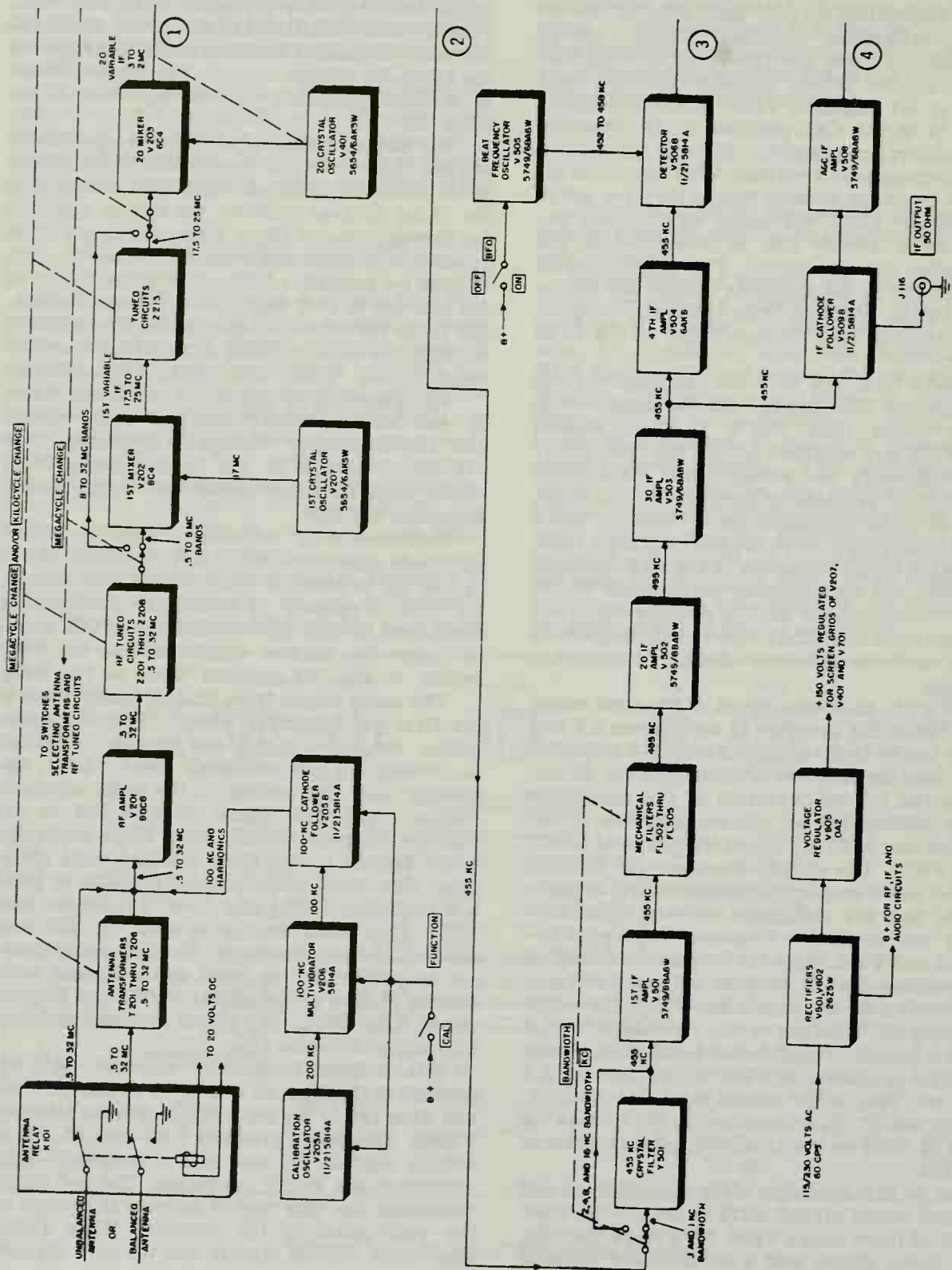
Figure 4-5. — Block diagram of radio receiver, R-390A/URR.

50.39

50.39

NOTE
REFERENCE SYMBOLS DENOTE:
101-200: MAIN FRAME
201-400: RF SUBCHASSIS
401-500: CRYSTAL OSCILLATOR SUBCHASSIS
501-600: IF SUBCHASSIS
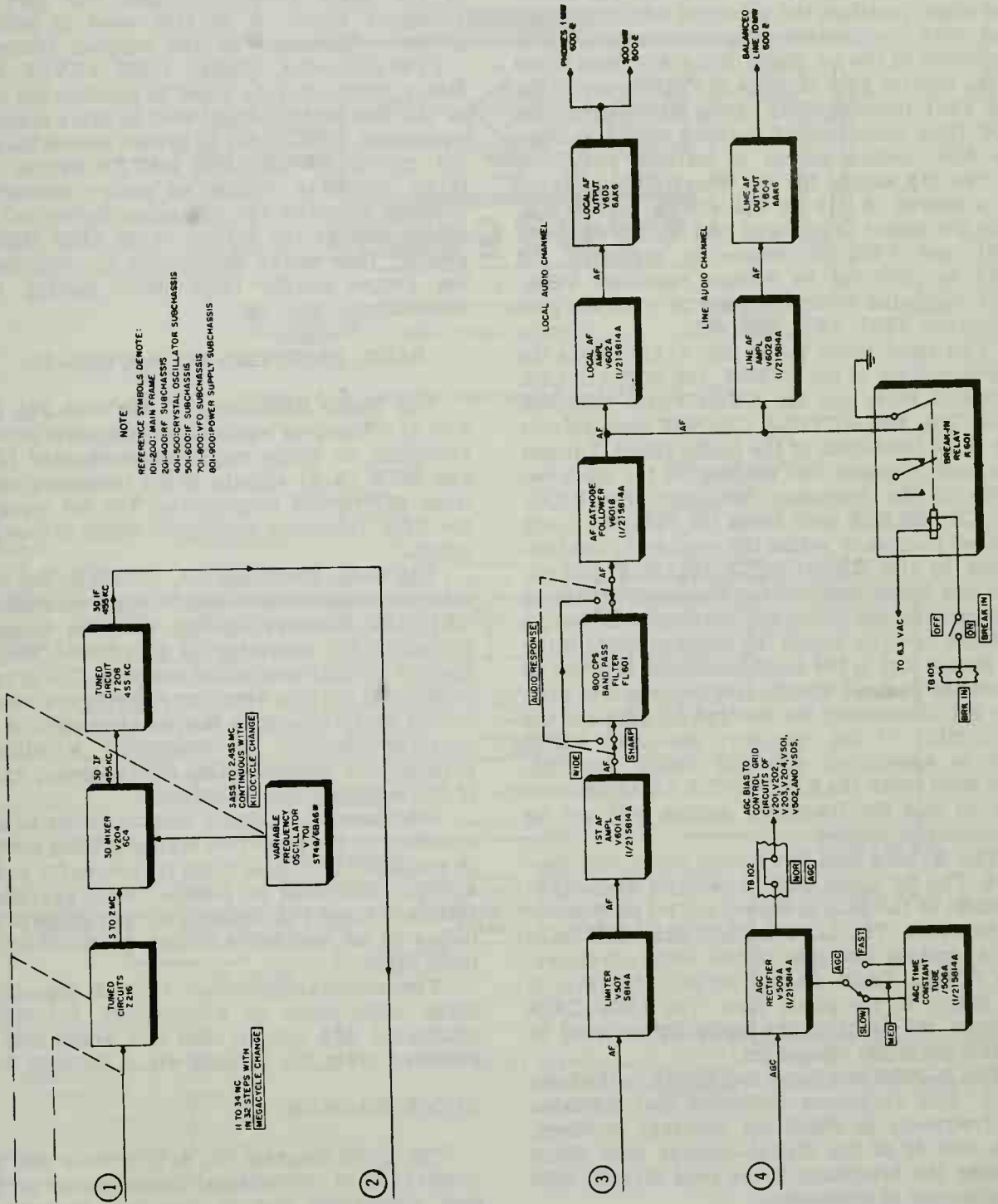701-800: VFO SUBCHASSIS
601-900: POWER SUPPLY SUBCHASSIS

Figure 4-5. – Block diagram of radio receiver, R-390A/URR—Continued

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The first tube of this circuit is AGC i-f amplifier V508. This stage amplifies the 455-kc signal and feeds it to AGC rectifier V509A. The rectifier stage rectifies the i-f signal into direct current, with amplitude in proportion to the average amplitude of the i-f signal. This AGC bias is fed to the control grid circuits of V201 through V204 and V501 through V503. Tube V506A provides AGC time constants of various durations when the AGC switch is set to various positions.

The B+ supply for the receiver is powered by a source of 115 or 230 volts a-c at 60 cps. This a-c power is changed into d-c by rectifiers V801 and V802. A source of regulated 150 volts is provided by voltage regulator V605. This regulated voltage is used as a screen grid supply for V207, V401, and V701.

Two front panel knobs (fig. 4-4) provide the tuning control of the R-390A, the MEGACYCLE CHANGE knob and the KILOCYCLE CHANGE knob. The MEGACYCLE CHANGE knob selects any 1-mc bandwidth of the tuning range. Turning this knob changes the reading of the first two digits of the frequency indicator. The KILO-CYCLE CHANGE knob tunes the receiver to any desired frequency within the megacycle band selected by the MEGACYCLE CHANGE control. The last three digits of the frequency indicator dial provide the kilocycle reading. The tuning controls actually adjust the tuned circuits in the r-f stages and in the local oscillator in order to select the desired station frequency and to provide simultaneously the desired i-f signal to the i-f portion of the receiver. The DIAL LOCK knob is associated with the tuning controls. This knob locks the KILOCYCLE CHANGE control so that the frequency setting will not be accidentally changed.

The R-390A has three front panel gain controls. The RF GAIN control permits manual adjustment of the gain of the r-f and i-f portions of the receiver. The LOCAL GAIN and LINE GAIN knobs control the gain of the audio circuitry. The LOCAL GAIN control adjusts the level of the output to the phone jack. The LINE GAIN controls the level of the audio output used to operate terminal equipment.

The R-390A has three indicators on the front panel. The frequency indicator dial indicates the frequency to which the receiver is tuned. This dial is of the digital-counter type which permits the frequency to be read directly with little chance of misreading.

The CARRIER LEVEL indicator is a meter which measures the level of the r-f signal appearing at the input of the receiver. The operator will find this meter valuable in tuning to the exact frequency which will give the strongest signal. It is also used to indicate proper adjustment of the antenna trimmer.

The indicator labeled LINE LEVEL is a meter which may be used to monitor the level of the line audio output used to drive terminal equipment. This meter is placed across the output circuit by the LINE METER switch. The three available values of meter sensitivity (voltage required for full-scale deflection) are determined by the setting of the LINE METER switch. This meter is valuable in maintaining the proper output level when making tape recordings.

RADIO RECEIVING SET AN/URR-35A

The Radio Receiver Set, AN/URR-35A (fig. 4-6) is a receiver equipment designed to provide reception of voice amplitude-modulated (A-3) and MCW (A-2) signals in the frequency range from 225 to 400 megacycles. The set replaces the RDZ receiving equipment which is now obsolete.

The Radio Receiving Set, AN/URR-35A contains a dual conversion (double-superheterodyne) VHF/UHF receiver chassis which is designed primarily for operation as a pretuned, single-channel crystal-controlled receiver. The crystal holder and various receiver controls are located behind small doors to the left and right of the receiver chassis. By employing a suitable crystal, any channel within the frequency range of the receiver may be selected.

Continuously variable manual tuning is also a feature of the receiver. A single tuning control is employed for tuning to any frequency for either crystal-controlled or manual tuning operation. Either of these two methods of operations is selected by an oscillator switch on the receiver front panel.

The receiver produces a 10-db signal-to-noise ratio when an 8-microvolt r-f signal modulated 30% with a 1000 cps audio note is received from the antenna via a 50-ohm line.

BLOCK DIAGRAM

The block diagram (fig 4-7) reveals that the receiver is of conventional design except for the dual conversion feature. However, because of the high frequencies at which the receiver operates (225 to 400 mc) many special circuits

RECEIVER CABINET

FRONT PANEL SECTION
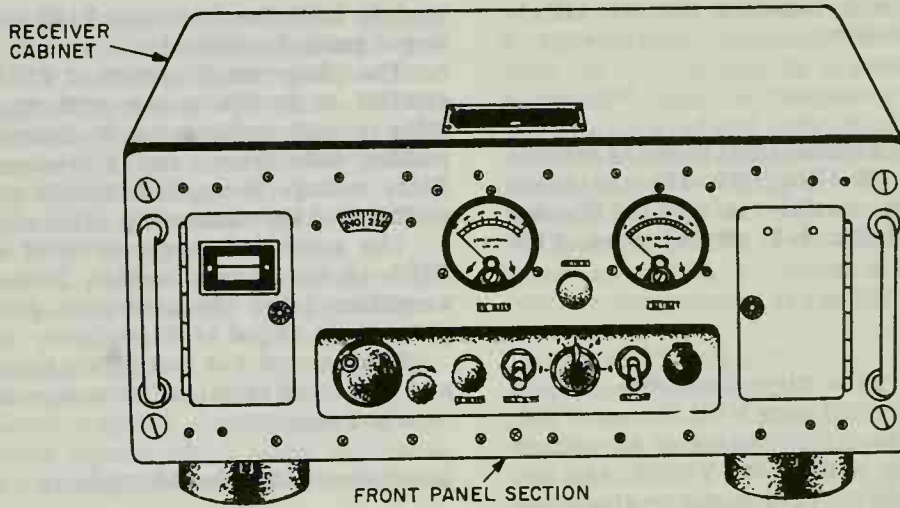
32.58

Figure 4-6.—Radio receiving set AN/URR-35A.

IF/AF SECTION

INPUT METER, M501

18.6 MC I F OUTPUT

1.775 MC I F OUTPUT

PRESELECTOR

ANTENNA J 404

1ST R-F AMPLIFIER V101,V102

2ND R-F AMPLIFIER V103,V104

1ST MIXER V105

1ST I-F AMPLIFIER V201

2ND MIXER V203A

2ND I-F AMPLIFIER V204

3RD I-F AMPLIFIER V205

1ST LOCAL OSCILLATOR V106A

1ST DOUBLER V106-B

2ND DOUBLER V107

TRIPLER V108 V109

SCAN CHANNEL AMPLIFIER V202

2ND OSCILLATOR V203B

AGC

SCAN OUTPUT, J403

OSCILLATOR-MULTIPLIER CHASSIS

AGC

OUTPUT METER, M502

A-F DETECTOR V206A

NOISE LIMITER V206B

SILENCER DIODE V208B

1ST A-F AMPLIFIER V209A

2ND A-F AMPLIFIER V209B

A-F OUTPUT V210

HEADPHONE JACK, J 501

OUT

IN

OUT

IN

NOISE LIMITER SWITCH, S202

SILENCER SWITCH S501

AGC-SILENCER AMPLIFIER V207

AGC-SILENCER DIODE V208A

TO ALL STAGES

POWER SUPPLY V301,T301 V302,V303

LOW-PASS FILTER F-218/URR-35

AUDIO OUTPUT J402

POWER INPUT J 401

AGC

ALIGN

S201 RECEIVE

32.61

Figure 4-7.—Radio receiver, R482/URR-35, block diagram.

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are employed which improve the overall efficiency of the receiver.

Antenna Input

The signal input connections from the antenna to the receiver (R-482A/URR-35) are made through a 52-ohm coaxial cable to the 50-ohm antenna jack, J404 (fig. 4-7) on the bottom of the low-pass filter.

R-F Amplifiers

The grid and plate circuits of the two r-f amplifier stages, comprising V101 through V104, (fig. 4-7), and the grid circuit of the mixer stage, comprising V105A and V105B, are arranged in push-pull, and are tuned by a balanced type five-gang air dielectric capacitor, C101-(not shown).

Concentric type trimmer capacitors are used for aligning the high frequency end of the tuned circuits in the r-f amplifier and mixer stages. Each of the trimmers consists of a metal block mounted on one stator section, and a partially threaded rod which extends from a bracket on the other stator section into the center bore of the block. An insulating tube is contained in the bore of the block which serves as the dielectric. The block and rod constitute, respectively, the stator and rotor plates of the trimmer capacitor. The capacity of the trimmer is varied by turning the rod to adjust the amount of the projection into the block.

Alignment at the low frequency end of the tuning range is accomplished by padder inductances.

Each of the inductors consists of two parallel round rods which are connected by a shorting bar. Positioning the shorting bar, by turning the adjustment screw, varies the inductance of the padder coil.

The push-pull arrangement of V101 and V102 approximately halves the interelectrode capacitance that would be presented by single-ended operation, thus permitting a greater range of frequency coverage with higher circuit efficiency range.

Operating bias for the first r-f amplifier tubes (V101 and V102) is obtained from the AGC circuit. The screen grids are bypassed to ground by stray or distributed capacitances between the socket pins and/or socket wiring and the chassis. Capacitors in the cathode circuit tune out the inductance of the cathode leads, and therefore

tend to hold the V101 and V102 cathode nearer to r-f ground potential.

The plate tuned circuit of V101 and V102 is similar in design to the grid circuit, and contains a tank inductance, a tuning capacitor, a padder inductance, and a trimmer capacitor. Plate voltage is applied through a dropping resistor, and the center-tap of the tank inductance.

The amplified output of V101 is coupled to V103 in the upper section of the second r-f amplifier (also operated push-pull); and V102 couples its output to V104.

The second r-f amplifier, comprising V103 and V104, is identical in design to that of the first r-f stage.

Local Oscillator and Multipliers

The oscillator-multiplier section of the preselector (fig. 4-7) generates a local injection signal which is 18.6 mc higher in frequency than the received signal. The basic oscillator frequency is generated in a cathode-coupled crystal-controlled circuit, comprising V106A and V106B. The oscillator circuit arrangement is that of a Butler oscillator. The local oscillator signal is obtained from an oscillator stage, V106A, followed by three stages of frequency multiplication (V106B through V109).

The V106 oscillator functions as a crystal-controlled circuit when the OSC switch (not shown) is in the CRYSTAL position. The crystal, Y201 (not shown), operates in a series resonant mode to establish the frequency of the feedback voltage from the cathode of V106B to the cathode of V106A. A capacitor in series with the crystal is used to resonate with the inductance of the crystal leads, so that zero phase shift exists between the two cathodes.

The plate output of the oscillator stage, V106A, is coupled to the grid of the first doubler, V106B. The grid input signal drives the V106B grid positive with respect to its cathode during the most positive portion of each input cycle. Thus, the V106B grid draws a current during this period, and C137 (not shown) charges negative toward the grid.

When the grid input signal drops below a value which does not permit the V106B grid to continue to draw a current, C137 partially discharges to establish a grid leak bias between the grid and cathode of V106B.

A test point (not shown) is provided at the grid of the first doubler (V106B) to facilitate measurement of the d-c bias on the tube. The

measured voltage is indicative of the amount of drive from the oscillator. The doubler output is developed across the plate components, and transformer coupled to second doubler, V107A.

The second doubler stage employs V107A and V107B in a push-push circuit arrangement (grids in pushpull, plates in parallel).

The output of the second doubler (V107) is coupled to the tripler stage.

The push-pull circuit arrangement of the tripler stage (V108 and V109) and the center-tapping of the coupling coil, L113 (not shown), suppress the amplification of even order harmonics in this stage. Thus, the push-pull arrangement minimizes the amplification of undesired spurious signals which might otherwise be amplified and radiated by the receiver antenna.

The V108 and V109 tripler plate circuits are tuned to a frequency which is three times their grid input frequency. Thus, the multiplier chain produces a total multiplication of 12, and a resulting output frequency which is 18.6 mc higher than the receiver input frequency.

First I-F Amplifier

The output of the mixer stage (V105A and V105B) is coupled to the grid circuit of the first i-f amplifier stage, V201. Amplifier V201 functions as a conventional i-f amplifier, except that the i-f gain control, R233 (not shown), is connected in the plate and screen circuits as opposed to its more common location in the cathode. The effect of varying R233 is to change the plate and screen voltages, and thus change the gain of the first i-f stage.

The V201 output signal is applied to the primary of the first i-f output transformer. This transformer is shielded, and double-tuned to resonate at the 18.6 mc intermediate frequency by the adjustable iron slugs.

Second Oscillator and Second Mixer Stages

The second oscillator, V203B, is crystal controlled to produce the second heterodyning frequency required for the operation of the receiver. The second mixer stage, V203A, receives the 18.6 mc i-f signal from V201 at its control grid (through a suppressor resistor), while a signal 1.775 mc below the V201 output from the second oscillator, V203B, is applied to the second mixer (V203A) cathode.

The V203B oscillator circuit is connected as a conventional crystal-controlled oscillator, with the crystal and the crystal holder applying a parallel resonant frequency voltage between the control grid and cathode of V203B.

The 18.6 mc signal from V201, and the 16.825 mc input from V203B are both applied between the control grid and cathode of the second mixer, V203A. The mixing of the two input signals takes place in V203A to produce the sum and difference frequencies at the V203A plate.

Second and Third I-F Amplifier Stages

The second and third i-f amplifier stages, V204 and V205, respectively are connected as conventional i-f stages. The components and circuitry of the two stages are almost identical. The essential difference between the two i-f amplifier stages is that the gain of the second i-f amplifier stage, V204, is controlled by the AGC voltage, while the third i-f amplifier stage, V205, is self-biased. The i-f output from V205 is rectified and filtered in the detector stage, V206A.

Scan-Channel Amplifier

A panoramic adapter may be used with the receiver when it is desired to view the received signals. The scan-channel amplifier, V202, is provided to amplify the signal which is to be fed to the adapter. The stage functions as a conventional amplifier which receives the 18.6 mc i-f output of V105. The V202 output, is applied to the 50-ohm scan jack, J403, on the low-pass filter.

Detector and Noise Limiter

The stage to which the detector (V206A) output is coupled is determined by the setting of the silencer and noise limiter switches, S501 and S202 respectively. With these switches in the position shown, the detector output is fed through a noise limiter, V206B, and a silencer diode, V208B. Tube V206A functions as a conventional diode detector stage, and V206B functions as a series diode noise limiter.

AGC-Silencer Amplifier

The AGC-silencer amplifier, V207, is a conventional amplifier circuit used to amplify

the output from the i-f amplifiers to a level which is suitable for use in the AGC and silencer circuits. A -9-volt fixed bias potential from the power supply is applied to the V207 grid.

The V207 plate load consists of a parallel resonant tank which is tuned to the intermediate frequency (1.775 mc). The amplified output of V207 is fed to the AGC-silencer diode, V208A.

AGC-Silencer Diode

The AGC-silencer diode, V208A, functions to produce the negative voltage for the AGC and silencer circuits. The V208A output is also used to assist in aligning the receiver when the align-receive switch, S201, is placed in the ALIGN position.

The conducting and nonconducting condition of V208B is determined by the combined action of the -AGC silencer amplifier, V207, and the AGC-silencer diode, V208A. When both of the switches (S501 and S202) are in the OUT position (opposite to the setting shown) the detector (V206A) output is applied through the switch (OUT) contacts to the first audio amplifier, V209A.

Audio Amplifier Stages

The audio signal is amplified in three stages of audio amplification comprising V209A, V209B, and V210. The audio output, V210, is applied to the headphone jack, J501, and to output meter M502. The output signal is also applied through the low-pass filter to the audio output jack, J402.

SINGLE SIDEBAND CONSIDERATIONS

Because single sideband has become increasingly important in Navy applications, it is necessary that Navy technicians acquire an understanding of the basic principles of operation. If the technician understands the principles of operation, he will experience little difficulty in utilizing the instruction books or technical manuals in corrective maintenance work.

The advantages of SSB cannot be realized without the use of specially designed components and circuitry. First of all, there is the problem of frequency stability, especially when the carrier is totally suppressed. This means that the oscillators in the transmitter and in the receiver must not drift more than a few cycles.

Actually, the permissible frequency variation for SSB systems is $\frac{1}{100}$ of that for an AM system.

In one type of double-sideband generations, filters of extreme selectivity are needed. Linear power amplifiers, which are difficult to design, are also needed.

Another problem, when SSB equipment is used on high-speed aircraft, is that of Doppler shift. This is especially noticeable at the higher radiated frequencies.

Basic information on generating SSB signals, filters, and oscillators has been covered in Basic Electronics, NavPers 10087-B. The EIMB 0967-000-0000, also contains information on SSB circuitry with a complete discussion on theory and troubleshooting on individual typical sections of receivers and transmitters.

A comparison of the frequency and power relationships between a 50-watt, single-sideband transmission and conventional AM transmission show that in SSB, only one sideband (upper or lower) with a peak-envelope power (PEP) of 50 watts is needed to transmit intelligence because the carrier is surpressed or not radiated. A 1-kc tone on a 1-mc carrier would result in a radiated signal of (1 mc ± 1 kc) either 999 kc or 1001 kc. In some systems both sidebands may be utilized independently.

Whereas the same intelligence transmitted by an AM system requires that the radiated signal include the carrier and an upper and a lower sideband frequency for each frequency in the modulating signal. For example, if a 1-mc carrier is modulated by a 1-kc tone, the radiated signal will include the 1-mc carrier, the lower sideband frequency (1 mc - 1 kc = 999 kc), and the upper sideband frequency (1 mc + 1 kc = 1001 kc). If the modulating signal contains many frequencies, there will be, of course, many frequencies in the sidebands. In this system of transmission, none of the transmitted intelligence is contained in the carrier; therefore all of the power put into the carrier is wasted insofar as transmitting intelligence is concerned. Likewise, because duplicate information is contained in each of the two sidebands, the intelligence content of the transmitted signal could be recovered from one sideband only.

In a conventional AM system where both sidebands and the carrier are transmitted, the power in the sidebands is dependent upon the amount of modulation. For 100% modulation the power in the sidebands is equal to one-half that in the carrier. Thus, a conventional AM

transmitter with 100-watts carrier power will have 50 watts in the sidebands (25 watts in the upper sideband and 25 watts in the lower sideband) at 100% modulation, making the total power transmitted 150 watts. It can be seen, then, that two-thirds of the total radiated power in a conventional AM system (assuming 100% modulation) is in the carrier and is therefore not useful in conveying intelligence.

When the r-f signal is demodulated in the conventional AM system, the audio output is a combination of the upper and lower sidebands. In this conventional type of detection (known as coherent detection) the audio output is proportional to the power contained in the two sidebands.

In a single-sideband system, only one sideband is transmitted and therefore the audio output of the SSB receiver is proportional to the power contained in the one sideband.

It therefore becomes apparent that an SSB transmitter and an AM transmitter will perform equally (same signal-to-noise ratio) under ideal propagating conditions, if the total sideband power of the two transmitters is equal. Considering the relationship between sideband power and carrier power in a conventional AM system, it is evident that an SSB transmitter will perform as well as an AM transmitter of twice the power rating under ideal propagating conditions. Thus, a single-sideband transmitter rated at 50 watts will produce the same signal intelligence level at a receiver as a conventional AM transmitter rated at 100 watts of carrier power.

As propagating conditions become less than ideal, the SSB system will show even a greater advantage over an AM system. An AM transmission is subject to deterioration under poor propagation conditions because all three components of the transmitted signal (the upper sideband, the lower sideband, and the carrier) must be received exactly as transmitted to realize perfect reception. Because there is only one component in the transmitted signal for an SSB system, it is not so affected by poor propagating conditions. Studies have shown that the SSB system will give from zero to nine decibel improvement under various conditions of propagation when the total sideband power in SSB is equal to that in conventional amplitude modulation.

It has been proven that under average conditions, the SSB system shows about a 3 db advantage over the AM system. In other words, in normal use, an SSB transmitter rated at 100

watts (PEP) will give equal performance with an AM transmitter rated at 200-watts carrier power.

As far as bandwidth is concerned (assuming one sideband only), the SSB system requires only about one-half the frequency spectrum that the conventional AM system requires.

The advantages of SSB over the conventional AM system may be summarized as follows:

1. The SSB transmitter will perform as well as an AM transmitter of twice the power rating under ideal propagating conditions. Under average conditions there is also an additional 3 db advantage of a SSB system over an AM system having the same sideband power.

2. If only one sideband is used, the SSB system requires only one-half as much r-f spectrum as the AM system.

3. The SSB transmitting system uses smaller units than comparable AM units because less power is required.

4. By virtue of less power in the antenna, lower voltages are required, with attendant reduction of potential breakdown.

5. The SSB system is subject to less noise interference because the bandpass is narrower.

RADIO RECEIVING SET AN/WRR-2

The AN/WRR-2 radio receiving set (fig. 4-8), is designed for shipboard installation and is intended primarily for the reception of single sideband transmissions with full carrier suppression in the frequency range of 2 mc to 32 mc. The set will also receive conventional AM signals of various types including CW, MCW, voice, facsimile, and frequency shift teletype. Simultaneous use of both upper and lower single sideband channels for receiving two different types of intelligence is possible. This capability however, does not extend to concurrent reception of single sideband and AM signals.

The receiver will net with any transmitter capable of operating in the same frequency range, provided the signal is one of the previously mentioned types and the receiver is compatible with existing shipboard auxiliary equipments which process facsimile and frequency shift teletype data.

The receiver consists of Electronic Frequency Converter CV-920/WRR2 (converter drawer) and Intermediate Frequency-Audio Frequency amplifier AM-2477/WRR2 (demodulator drawer) in individual, joined cabinets.

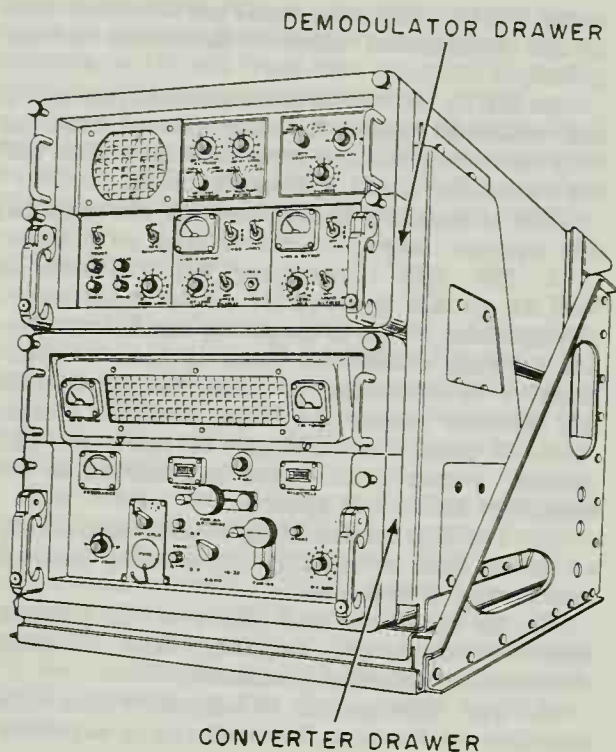The demodulator drawer is above the converter drawer.

DEMODULATOR DRAWER

CONVERTER DRAWER

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Figure 4-8.—Radio receiving set AN/WRR-2.

Signal voltages applied at the receiver antenna (fig. 4-9), are fed to the converter circuits through a coupling device that offers protection from excessively high signal voltages. In the converter drawer the frequency range is divided into four tuning bands. A selection switch and direct reading counters facilitate band selection and tuning. Tuning from 2 mc to 32 mc may be either incremental or continuous. In the incremental method a crystal-controlled internal frequency standard permits accurate settings in increments of a whole kilocycle. Cancellation of oscillator frequency drift is achieved and a frequency stability of one cycle in 10 mc per day is obtained.

Cancellation of the effects of oscillator drift is achieved as follows: The output of the H-F oscillator (V301 fig. 4-9A) is injected into two mixer stages, V151 and V251. The incoming signal is applied to one stage (V151), and a locally generated reference frequency from the crystal oscillator is applied to the other stage (V251). The outputs of the mixers are applied to separate i-f amplifiers, and both i-f amplifier outputs are applied to another mixer stage,

V502. Thus any drift in oscillator frequency affects the outputs of the first two mixers by equal amounts, thereby maintaining the output of the second mixer (V502) constant.

Continuous tuning must be used for frequencies which terminate in a fraction of a kilocycle, and this method reduces slightly the stability of the receiver.

After the process of tuning, triple frequency conversion and i-f amplifications are performed in the converter drawer. The i-f signal is passed to the demodulator drawer where further i-f and AF amplifications are performed. This drawer contains three detector amplifier sections. One processes conventional AM signals and has a BFO for CW, facsimile, and frequency shift teletype signals. The other two sections detect and amplify signals in the upper and lower sideband channels. Each section has its own AGC and silencer circuits.

PRESELECTOR

The preselector consists of two stages of r-f amplification and the first conversion mixer. The AGC voltage developed in the detector amplifier section is applied to V51 and V101. Manual r-f gain control is also applied to these stages. The r-f gain control is a dual section control which also adjusts the gain of i-f amplifiers V1602 and V1603 in the AM unit and V1003 (not shown) in the USB unit.

Mixer V151 receives signals from the high frequency oscillator (V301) and r-f amplifier (V101) and heterodynes these inputs to produce the first i-f signal (1625-1725 kc) to the tunable i-f filter.

TUNABLE I-F FILTERS

The tunable i-f filter is a continuously tunable four-section filter which follows the preselector.

The four sections are tuned from 1625 kc to 1725 kc by a four-section variable capacitor. The filter selects the first conversion frequency from the output of the preselector mixer, V151.

INJECTION I-F AMPLIFIER

The injection i-f amplifier section develops the converter's final intermediate frequency (80 kc) from the outputs of the tunable i-f filter (1625 kc-1725 kc), the interpolation oscillator (V401) (580-kc-680 kc), the harmonic mixer V251 (825 kc), and the synthesizer (140 kc).

The V501 output is applied to mixer V502 which also receives a signal of 1405 kc to 1505 kc from V507. This latter signal is derived from the addition in V506 of a signal in the range of 0 to 680 kc from the interpolation oscillator and the 825 kc signal from i-f amplifier V505.

The 220-kc output from V502 is applied to V503 where it is mixed with the 140-kc signal from the synthesizer via V508. Two AGC circuits in the injection i-f amplifier regulate the signal voltages in their loops. One controls the output of V505, and the other the output of V508.

80-KC FILTER

The 80-kc filter (FL 1301) accepts the 80-kc intermediate frequency developed by mixer V503 (third conversion mixer) and passes it to the AM and USB detector amplifier sections of the demodulator. The filter has a center operating frequency of 80 kc with a band width of 12 kc. The output of FL1301 is applied to the AM detector section via i-f amplifier V160-1 and the upper sideband detector amplifier via V1003.

The lower sideband detector amplifier receives its input from the output of V1003.

UPPER SIDEBAND DETECTOR AMPLIFIER

Two detector amplifiers comprise the single sideband portion of the receiver, making possible either upper or lower sideband reception, or both simultaneously.

An input from the 80-kc filter is applied to a three-stage i-f amplifier V1003, V1004, and V1005 (fig. 4-9B). The output of V1003 is applied to both V1004 and the lower sideband detector amplifier.

The output of V1004 is applied to V1005 which in turn has two outputs. One is to demodulator amplifier V1006 and the other is to the i-f AGC amplifier V1001. The output of V1001 is applied to two diode rectifiers (not shown) and V1002A. The rectified output from one of the rectifiers goes to silencer V1002 B and to Resonance Meter M601. The other rectifier output goes to V1002-A and is fed back to provide gain control for both i-f amplifier V1005 and demodulator amplifier V1006. Another output from the second rectifier is applied to AGC combiner V1103.

The output of V 1006 is mixed in a balanced ring demodulator Z-1001 with an 80-kc carrier signal. The audio frequency representing the difference between the signals (from V1006 and

the 80-kc carrier) is amplified by V1008 and goes to the line A phones jack via AF filter FL1801.

AM DETECTOR AMPLIFIER

As illustrated in figure 4-9 the AM detector amplifier consists of 80-kc i-f amplifiers V1601, V1602, V1603, and V1604, -4 and AGC circuit, a detector, a noise suppression circuit, several audio amplifiers, and BFO. The 80-kc signal is amplified by the four stages of i-f amplifiers and detected by CR-1603, a silicon diode (not shown). After detection the audio signal is amplified in a 4-stage AM amplifier. The output of the last stage (V1609) is supplied as one of two possible outputs of the line A output connector.

An AGC circuit provides gain control for the last three i-f stages. This control is delayed AGC allowing weak signals to be received more readily before AGC takes over, while also limiting strong signals to a level below the saturation point of the detector. The AGC voltage is also applied to the tuning meter M601 which indicates resonance in the front end tuned circuits (preselector). The AGC voltage is also supplied to silencer V1608B.

The silencer V1608 has in its plate circuit a coil of a single-pole double-throw relay. In addition, two voltages are impressed on the grid input of the silencer, one positive from a potentiometer, the other an AGC voltage. With no signal at the receiver input, the AGC voltage is inoperative and only the potentiometer voltage is applied to the silencer. This action causes V1608 to conduct, energizing the relay coil and actuating the relay armature to ground the input to the second AF amplifier stage and block all output from subsequent stages. When a signal is received, the AGC circuit sets up a voltage at the silencer grid that overrides the position voltage to produce a reduction in plate current sufficient to deenergize the relay coil. This action returns the armature to its normally open position, removing the ground from the second AF stage so that the signal can pass to the final audio amplifier.

LOWER SIDEBAND DETECTOR AMPLIFIER

Tubes V1105 and V1106 comprise a two-stage i-f amplifier receiving an 80-kc signal from V1003 of the upper sideband detector amplifier. Tube V1106 has two amplified 80-kc outputs, one
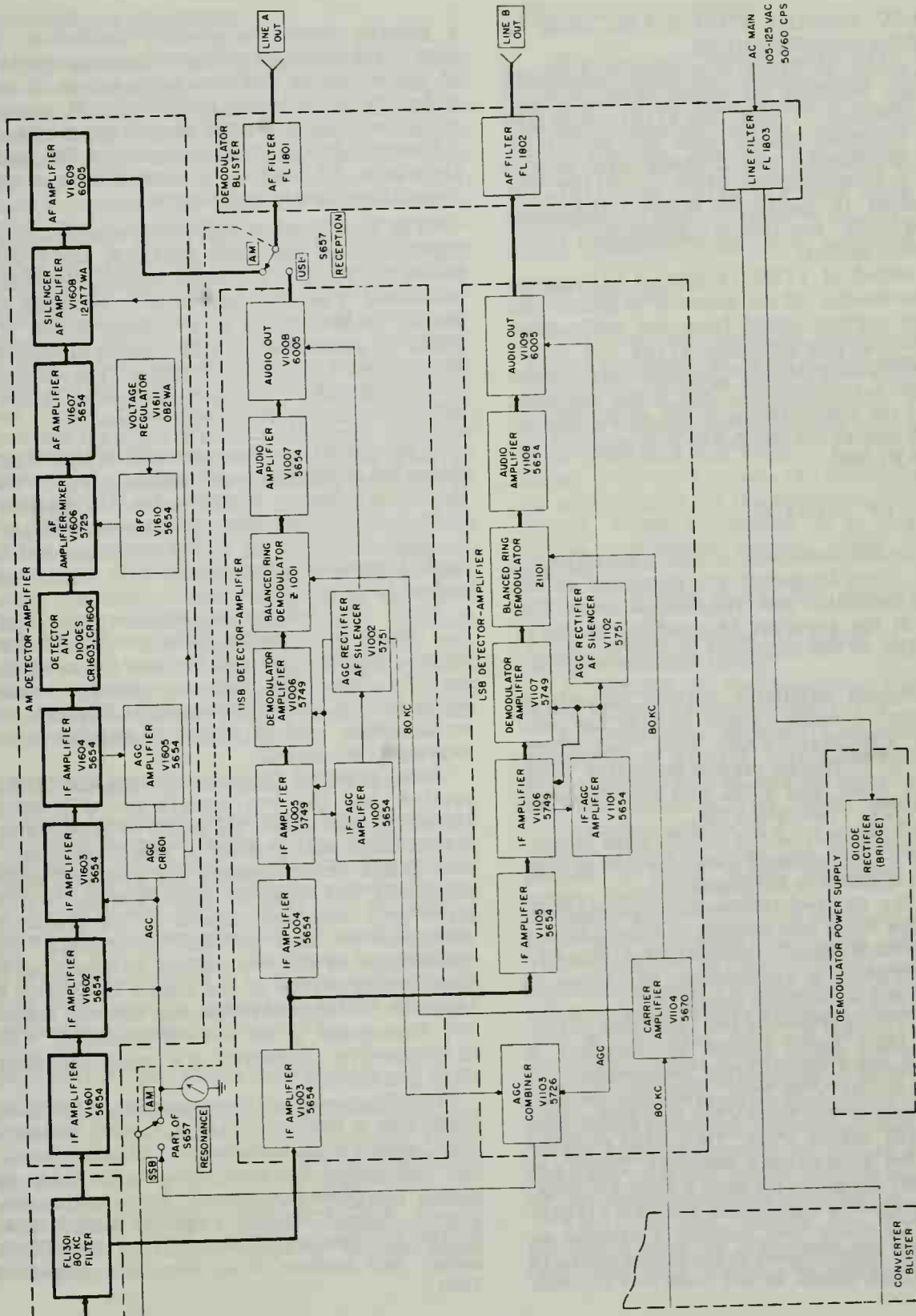
Figure 4-9. — Radio receiving set AN/WRR-2, block diagram.

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Figure 4-9. —Radio receiving set AN/WRR-2, block diagram. —Continued

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to i-f AGC amplifier V1101 and the other to demodulator-amplifier V1107.

The output of V1101 is applied to a silicon AGC diode rectifier the output of which is applied to silencer V1102B. The V1101 output also goes to two other silicon diode AGC rectifiers. One has two outputs, one of which is fed back to provide gain control for V1106 and V1107. The other output is fed to combiner V1103 which furnishes AGC for the r-f amplifiers in the preselector section.

The output of V1107 is applied to balanced ring demodulator Z1101 where it is mixed with the 80-kc carrier signal from the synthesizer section via carrier amplifier V1104.

The difference between the 80-kc carrier and the 80-kc i-f (containing the lower sideband signal) is the audio frequency which is amplified by V1108 and V1109 and fed to line output B via AF filter FL1802.

CRYSTAL OSCILLATOR

The crystal oscillator section generates the standard 1-mc frequency for operation of the spectrum generator and frequency dividers. The stability of the receiver is related directly to the stability of the oscillator (1 part in 10 million).

The crystal oscillator section consists of three major circuits. These include the oven oscillator amplifier (V701, V702, and V703), (fig. 4-9, A) a Pierce crystal oscillator V704, amplifiers V705 and V706, and the frequency dividers V707, V708, and V709.

The temperature of the crystal oven is controlled by a proportional method which is continuous as opposed to thermostatic.

When the desired temperature is reached, bridge arms come into balance and provide only enough feedback signal to maintain that temperature. A deviation in temperature will cause an imbalance and subsequent correction.

Tube V701 operates as the oscillator stage. When its plate tuned circuit is operating, a positive feedback signal from the oven heaters is applied to the V701 grid producing oscillations at about 2 kc. Tube V702 is a buffer amplifier which drives V703. Tube V703 supplies power to the oven heating elements. There are four of these elements forming a bridge circuit. Two terminals are supplied by V703 via a transformer. The other two terminals provide an output signal proportional to the temperature of the oven. This signal is fed back to V701 with

a polarity suited to produce oscillation. The size of the feedback voltage determines the level of heater power and oven temperature. In case of failure of the heater control circuit, contents of the oven are protected against overheating by a thermal switch connected across the heater terminals. Its contacts are normally open: when closed they divert heater power.

The 1-mc crystal and related components are also parts of the oven assembly but are not functional parts of the oven oscillator-amplifier assembly. The complete oven is mounted in a Dewar flask similar to a vacuum bottle. The bottle is protected by a fiberglass blanket and is mounted in a steel container. This assembly is a single unit. Components cannot be removed for individual repair or replacement.

The crystal oscillator-amplifier circuit produces the standard 1-mc frequency from which, through frequency division, the 100-kc output of the crystal oscillator section is derived.

The oscillator is a Pierce type, and the basic frequency is determined by a crystal connected in the plate circuit.

Tube V705 follows the oscillator and amplifies the signal. Buffer V706 provides further signal amplification and isolates the oscillator circuits from the frequency divider section which follows. The 1-mc amplifier V705 includes an amplitude control circuit to provide an unvarying 1-mc signal.

When a signal from an external standard is applied, and the two signals are of identical frequency and phase, the phase or temperature meter M602 will produce a steady reading. If they differ in frequency, the resultant phase difference will cause the meter reading to fluctuate at a rate determined by the frequency difference or beat between the signals. The coincidence of the two signals is measured by timed observations of the meter fluctuation in response to the beat frequency. Internal standard adjustment of the crystal oscillator V704 is obtained by adjustment of a tuning capacitor, C719 (not shown).

The frequency divider consists of V707 and V708 with a buffer V709. Tube V707 obtains a 1-mc signal from the crystal oscillator amplifier and makes the first division to 500 kc by means of a plate tank circuit, tuned to this frequency. Tube V708 has a similar circuit tuned to 100 kc. Divider V708 amplifies the 100-kc signal and applies it to harmonic amplifier V201.

HARMONIC AMPLIFIER

The harmonic amplifier and mixer produce a 100-kc spectrum from the frequency range of 2.9 mc to 32.9 mc when driven by the 100-kc signal from the crystal oscillator section. Basically, the harmonic amplifier and mixer sections consist of harmonic generator V201, tuned amplifier stage V202, and an output mixer V251. The mixer combines the generated spectrum with a signal from the high frequency oscillator V301 for incremental tuning at 100-kc intervals. The tuning control of the HF oscillator and harmonic amplifier and mixer sections are ganged for simultaneous operation.

The 100-kc signal from the crystal oscillator section is applied to harmonic diode CR201 (not shown). The signal clipping action of CR201 provides the control grid of harmonic amplifier V201 with a wide range of harmonics from the 100-kc input signal. The operating frequency is selected by the tank circuit in the plate of V201. This tank circuit is tuned simultaneously with the tuned circuits of the HF oscillator sections. The output of V201 is applied to a second harmonic amplifier, V202, where again the signal amplitude is increased and applied to mixer V251. Mixer V251 combines this selected harmonic signal with a signal from the HF oscillator to produce a difference frequency of 825 kc. The plate load of V251 is located in the injection i-f amplifier section.

HIGH FREQUENCY OSCILLATOR V301

High frequency oscillator V301 covers a frequency range of 3.725 mc to 33.725 mc in four bands. Its outputs are used in the preselector for the first frequency conversion and in the harmonic mixer V251 for the 825-kc signal used by the injection i-f amplifier section. The high frequency oscillator is a modified grid tuned Armstrong circuit with cathode feedback. One output is provided to mixer V251 and another to preselector mixer V151.

INTERPOLATION OSCILLATOR V401

Interpolation oscillator V401 produces frequencies in the 580-kc to 680-kc range. These are applied to the injection i-f amplifier and the synthesizer sections. The synthesizer receives its signal through the TUNING switch S702 located on the crystal oscillator. Tube V401 is a triode connected in a grid tuned Armstrong circuit.

FREQUENCY DIVIDER

The frequency divider (V451, V452, and V453) produces frequency division from 100 kc to 20 kc. The 100-kc input signal is supplied by the frequency dividing circuit in the crystal oscillator sections, and the 20-kc output signal is applied to the synthesizer section for further division. Tubes V452 and V453 form a locked oscillator similar to those in the crystal oscillator frequency divider V707 and V708.

SYNTHESIZER

The very high operating stability of the AN/WRR-2 receiver can be attributed largely to the receiver's crystal-controlled, temperature-compensated master oscillator V704. Frequency dividers, harmonic generators, and harmonic filters process the original locally generated signal to provide signals of frequencies related to submultiples of the master oscillator frequency. These are picked off at stage outputs in the synthesizer section which perform as follows:

1. Contribute to the final conversion to 80 kc.

2. Provide a fixed i-f signal of 140 kc for use in continuous tuning.

3. Furnish an 80-kc carrier signal to the lower sideband amplifier via V1104.

4. Trigger a blocking oscillator for the production of precise 1-kc incremental pulses in conjunction with an interpolation oscillator of variable frequency in its 100-kc range.

Harmonic generator V801 in the synthesizer section receives from the frequency divider section a 20-kc signal processed from the master oscillator V704.

Filter FL-801 picks off the basic 20-kc signal, T802 the seventh harmonic (140 kc), and T803 the fourth harmonic (80 kc), and all three signals are passed to associated circuits in the synthesizer and lower sideband detector amplifier. In the lower sideband detector amplifier, the 80-kc signal is amplified and directed to balanced ring demodulators in both upper and lower sideband detector amplifiers to serve as injected carriers for the demodulator circuits.

Block T802 supplies its 140-kc signal only when the receiver is set for continuous tuning by the continuous tuning switch S702.

Filter FL-801 passes the 20-kc signal through a pair of frequency divider circuits (V802 and V803) with a 1-kc output. This signal

excites pulse shaping circuit V804 to trigger the next stage blocking oscillator, V805.

This oscillator generates a harmonic spectrum in 1-kc incremental steps, making available output frequencies at least as high as the 820th harmonic.

To use this harmonic spectrum with the variable frequency output of interpolation oscillator V401, a bandpass filter with a 100-kc bandwidth in the range of 720 kc to 820 kc (not shown) is connected to the output of blocking oscillator V805.

When a particular harmonic of 1 kc in the above mentioned range combines in mixer stage V806 with a properly related signal from the interpolation oscillator, a difference frequency of 140 kc is produced in the output of the mixer stage. This 140-kc signal is amplified in V807 and passed on to i-f amplifier V508 in the injection i-f section. Mixing the 140 kc with a 220-kc signal generated in the injection i-f results in an 80-kc i-f (final conversion) output from the injection i-f section.

Harmonic generator V801 gets its 20-kc signal from the frequency dividers via V453. Frequency dividers V802 and V803 are locked oscillators which divide the 20-kc signal from the harmonic generator and filter FL-801 by 4 (V802) and then by 5 (V803) to 1 kc which is applied to pulse shaper V804 for use as an accurately controlled trigger signal.

Triggered by pulse shaper V804, blocking oscillator V805 produces sharp pulses ranging in 1-kc steps from a fundamental 1 kc to beyond its 820th harmonic. A portion of this spectrum, 720 kc to 820 kc, provides harmonics to be mixed in the injection i-f amplifier with signals from variable output interpolation oscillator V401 to produce a difference frequency of 140 kc for use in the development of the final receiver i-f of 80 kc.

The pulse shaper V804 is a dual triode. The control grid of V804A receives from the frequency divider a 1-kc signal, the positive half of which is clipped by grid limiting, and the negative half by the tube being driven into cutoff. The 1-kc output of V804A is a square wave which is differentiated into narrow positive and negative going spikes which correspond to the leading and trailing edges of the square wave. These spikes are amplified and inverted by V804B.

The positive spikes provide a 1-kc sync pulse which locks V805 in by occurring at a time slightly earlier than the free-running frequency of V805.

Triode section V805A acts as a triggering tube for blocking oscillator V805B. The output of V805B is applied through bandpass filter FL-803 (not shown) to amplifier mixer V806 to combine with another input signal from the interpolation oscillator or from T802.

Amplifier mixer V806 produces a 140-kc signal for later use in establishing the final conversion signal of 80 kc. One of the inputs to V806 is the 1-kc harmonic in the range of 720 kc to 820 kc from blocking oscillator V805. During incremental tuning the other input is a signal in the range of 580 kc to 680 kc from interpolation oscillator V401. The difference frequency is 140 kc applied to bandpass filter FL-804 (not shown).

During continuous tuning, the 140 kc is picked off directly from filter T802 and passed to FL-804 (not shown). At the continuous setting of the tuning switch S702, plate voltage is removed from blocking oscillator V805B so that the only signal now applied to V806 is the 140-kc signal from T802. The 140-kc signal is fed to i-f amplifier stage V807 the output of which goes to the injection i-f amplifier section.

CONVERTER POWER SUPPLY

The receiver has two power supplies, converter (fig. 4-9, A) and demodulator (fig. 4-9, B). The converter power supply serves all circuits except those of the three detector amplifiers, upper and lower sideband, and AM. Voltage and current-stabilizing circuits are incorporated in the converter power supply to ensure steady inputs for voltage sensitive components in accordance with stability requirements. A full-wave bridge rectifier comprising silicon diodes provides the necessary high voltage output.

DEMODULATOR POWER SUPPLY

Operation of the demodulator power supply is similar to that of the unregulated portion of the converter power supply. The demodulator power supply is a conventional bridge type rectifier connected to a choke input filter. Functional sections of the demodulator get their filament voltages from a 6.3-v a-c winding of the power supply transformer.

RADIO RECEIVER R-1051/URR

The R-1051/URR (fig. 4-10) is a super-heterodyne receiver capable of receiving lower
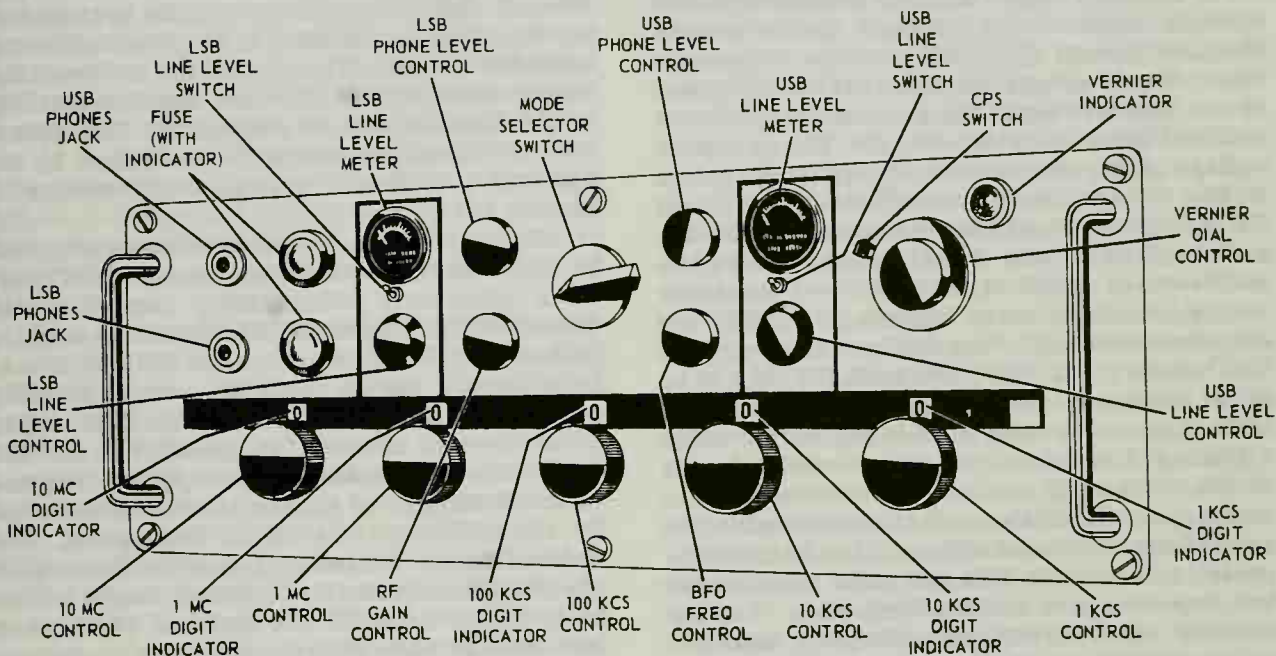
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Figure 4-10.—Radio Receiver R-1051/URR.

sideband (LSB), upper sideband (USB), independent sideband (ISB), frequency shift keyed (FSK), (conventional) amplitude modulated (AM), and continuous wave (CW) transmissions in the 2.0- to 30.0 megacycle frequency range. The ISB mode of operation allows two different types of intelligence to be received simultaneously, one on the LSB channel and one on the USB channel. FSK reception is obtained by using suitable ancillary equipment, such as Teletype Converter-Comparator AN/URA-17 or AN/URA-8. The R-1051/URR may also receive tone-modulated continuous wave (MCW), compatible amplitude modulated (compatible AM), and facsimile (through the use of suitable ancillary equipment) transmissions. The R-1051/URR may be oper-. ated in conjunction with a transmitter, as a transmitter-receiver in systems such as Radio Set AN/WRC-1. In this application either simplex or duplex operation is possible. The R-1051/URR may also be used as a separate, self-contained receiver requiring only a headset, antenna, and a nominal 115-v a-c primary power source for full operation.

Radio Receiver R-1051/URR is a triple-conversion superheterodyne receiver, tunable over the high frequency range from 2 to 30 mc. Tuning of the R-1051/URR is accomplished digitally by five controls (MCS and KCS) and a switch (CPS) located on the front panel (fig. 4-10). A display window directly above each control provides a digital readout of the digits to which the controls are set. The displayed frequency can be changed in 1 kc increments. The front panel switch allows the operating frequency to be changed in 500-cps increments. This tuning provides 56,000 discrete frequencies in which the receiver is locked to a very accurate frequency standard. Each 1-kc increment can be continuously tuned through by selecting the VERNIER position of the CPS switch. When using the vernier, the full accuracy of the frequency standard is sacrificed. Over the frequency range, the input sensitivity for an audio output signal-plus-noise to noise ratio of 10 db is better than 1 microvolt for ISB, LSB, USB, and FSK, 2 microvolts for CW, and 4 microvolts for AM.

BLOCK DIAGRAM

A received signal from the antenna passes through closed relay contacts in the antenna overload circuit (fig. 4-11), to the r-f amplifiers. Should a signal in excess of 15 volts appear at the receiver input, the antenna overload circuit will open the relay contacts. The excessive voltage is thereby prevented from being applied to the r-f amplifiers, which form a part of the RF Amplifier Electronic Assembly. Within the r-f amplifiers, the signal passes through a double-tuned input circuit, two r-f amplifier stages, a single-tuned interstage circuit, and output circuits. All of the resonant tuned circuits are tuned by the MCS, 100 kc (KCS), and 10 kc (KCS) frequency controls on the front panel. The MCS controls operate a code generator, which activates a motor-driven turret containing 28 strips. Each strip contains a tuned transformer and a portion of the capacitance required by each of the four tuned circuits. For each megacycle increment, a different tuned transformer and capacitor are switched into place. The remaining tuned circuit capacitance is mechanically switched into the four tuned circuits by the 100 kc (KCS) and 10 kc (KCS) controls on the front panel. These capacitors are located on circular boards stacked in the center of the

turret. The tuned circuits provide the frequency selectivity required to prevent undesired off-channel signals from distorting the desired signal by cross-modulating or overloading the r-f amplifier stages. The gain of the r-f amplifier stages ensures that the weak-signal sensitivity of the R-1051/URR is maintained. The gain of both r-f amplifier stages is controlled by the application of an AGC voltage from the step AGC circuit.

The output from the r-f amplifiers is applied to the mixers, which form a part of RF Translator Electronic Subassembly located in the Translator/Synthesizer Electronic Assembly (not shown). The mixers consist of three transistor mixer stages, with interstage coupling provided by selective filters. The first mixer receives the injection frequencies from MC Synthesizer Subassembly. The injection frequency is determined by the megacycle band selected by the MCS controls on the front panel. The desired output frequency from the first mixer always falls within two frequency bands, either 19.5 to 20.5 mc or 29.5 to 30.5 mc. The high or low band is also determined by the megacycle band to which the R-1051/URR is tuned. The output from the first mixer is gated through the appropriate 20 or 30-mc filter. This signal is mixed in the second mixer stage with the
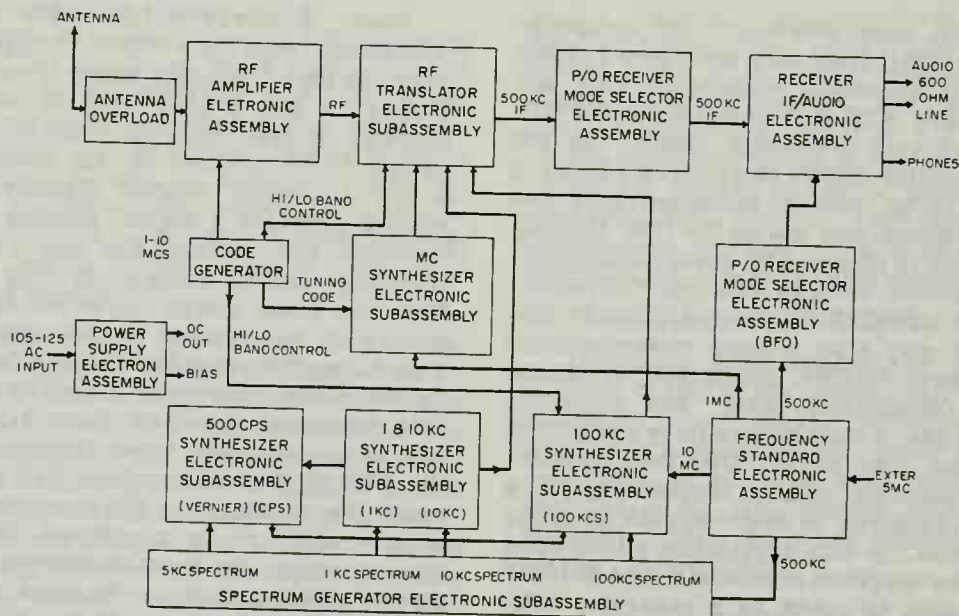

Figure 4-11.—Radio Receiver, R-1051/URR, simplified block diagram.

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njection frequencies supplied from the 100-KC synthesizer Electronic Subassembly. The desired frequency band from the second mixer is 2.8 to 2.9 mc. This signal is coupled through a 2.85-mc filter to the third mixer. The injection frequencies for the third mixer are supplied from the 1- and 10-KC Synthesizer Electronic Subassembly. The output from the third mixer is a 500-kc i-f signal. Signal flow from the antenna input through the output of the third mixer is identical, regardless of the selected mode of operation.

The 500-kc i-f output from the third mixer is applied to the mode gates. Three parallel paths are presented to the signal. The path that passes through the LSB mechanical filter (also used in ISB) is not gated since it has an independent output from the Mode Selector Electronic Assembly. Because the outputs from the USB mechanical filter (also used in FSK and ISB) and the AM mechanical filter (also used in CW) are paralleled for a common output, the input paths to these two filters must be gated so that only one path is open at any given time. Application of the correct gating potentials is determined by the mode of operation selected at the front panel.

The output from the LSB filter is applied to the i-f amplifiers in the IF/Audio Amplifier Electronic Assembly. The common output of the USB and AM filters is also applied to the i-f amplifiers in the TF/Audio Electronic Assembly.

The operating d-c voltage is applied to the proper electronic assembly according to the mode of operation selected at the front panel. In the ISB mode of operation, a d-c operating voltage is applied to both i-f amplifiers. AGC voltage from the step AGC circuit controls the overall gain of the i-f amplifiers by varying the attenuation of the input and the gain of the second i-f amplifier stage. The input to the step AGC circuit is derived from the output from the second i-f amplifier stage.

The output from the i-f amplifiers is applied to the detector circuits, consisting of a product detector and an AM detector. Depending on the mode of operation selected at the front panel, either the balanced product detector or the AM detector is powered by a d-c operating voltage. The product detector demodulates the USB, LSB, FSK, and ISB signals. In these modes of operation, a 500-kc injection, originating at a multiplier-divider in Frequency Standard Electronic Assembly, is applied to the product

detector for carrier re-insertion. This 500-kc injection passes through the 500-kc gate in the Mode Selector Electronic Assembly with little attenuation in these modes of operation. In AM and CW modes, this gate presents a high attenuation, since no carrier re-insertion is required by the AM detector. In the CW mode of operation, the BFO assembly in Mode Selector Electronic Assembly is turned on, and a variable 500-kc output is applied to the input of the AM detector in the IF/AF Assembly. The output frequency from the BFO circuit is controlled by the BFO FREQ control on the front panel.

The audio derived from the detector circuits in the IF/AF Assembly is applied to the USB LINE LEVEL control on the front panel which controls the audio level prior to its application to the audio amplifiers. The LSB LINE LEVEL control sets the audio level from the product detector in the IF/AF Assembly. Each Receiver IF/Audio Amplifier Electronic Assembly has two outputs. One is a 600-ohm remote output, which is applied to a connector at the rear of the case. The second output is to the PHONES jacks on the front panel. The PHONES output passes through a PHONE LEVEL control on the front panel, which adjusts the phone signal amplitude without altering the level of the remote output. Each remote output is monitored at the front panel by a LINE LEVEL meter, which has two scale ranges controlled by the LINE LEVEL switch on the front panel.

The step AGC circuit, which forms a part of the Receiver IF/Audio Amplifier Electronic Assemblies, controls the gain of the r-f amplifiers and i-f amplifiers according to the received r-f signal strength. The output from the i-f amplifiers is applied to the step AGC circuits, where it is converted to a d-c voltage that is applied to both the r-f and i-f amplifiers. One level (step) AGC voltage is used for voice and another level AGC for tone reception. The gain of the r-f and i-f amplifiers may be manually controlled by applying a d-c voltage on the AGC lines with the RF GAIN control. This manual action overrides the normal AGC voltages.

A combination of error cancelling loops and phase-locked loops is used in the frequency synthesizer circuits of the R-1051/URR to ensure that the injection frequencies applied to the mixers are correct. The 5-mc frequency standard produces an accurate, stable reference frequency upon which all frequencies used in the R-1051/URR are based. The circuit is housed in

an oven assembly maintained at a nearly constant temperature of 85° C by the oven control circuit. The accurate output from the 5 mc frequency standard is applied to a switching and compare circuit. An external 5-mc frequency may also be applied to this circuit. The switching and compare circuit routes the internal or external 5-mc signal to the multiplier-divider circuits or to the compare circuit. The compare circuit compares the internal 5-mc frequency with the external 5-mc frequency for an indication of the accuracy of the internal frequency standard. The 5-mc output from the switching and compare circuit is applied to the multiplier-divider circuit, where it is converted to frequencies of 500 kc, 1 mc, and 10 mc. These three outputs are used in the mixing processes required to produce the injection frequencies used in the r-f conversion process. The 500-kc output is also applied to the 500-kc gate circuit for insertion into the product detector for demodulation. The 5-mc frequency standard, oven control, multiplier-divider, and switching and compare circuits comprise the Frequency Standard Electronic Assembly.

The injection frequencies used in the first frequency conversion in the mixers circuit are generated within the MC Synthesizer Electronic Subassembly. This circuit consists of a phase-locked crystal oscillator that is automatically tuned to produce 1 of 17 frequencies between 2.5 mc and 23.5 mc. The output is applied to the high frequency mixer. The output frequency depends on the setting of the front panel MCS controls. The MC Synthesizer Electronic Subassembly employs a phase-locked loop to ensure the accuracy of the mc injection frequencies. The 1-mc output from the multiplier-divider in the Frequency Standard Electronic Assembly is applied to the spectrum generator to produce a spectrum of frequencies spaced at 1-mc intervals between 1 mc and 25 mc. The output from the spectrum generator and the output from the mc oscillator are mixed. Any error in output from the mc oscillator is detected and an error voltage is produced. This error signal is applied to the mc oscillator to lock it to the correct frequency. The accuracy of the oscillator output is the same as that of the 5-mc frequency standard.

The injection frequencies used in the second frequency conversion in the mixers circuit are generated within the 100-KC Synthesizer Electronic Subassembly. This circuit consists of a crystal oscillator, the output from which is 1 of

10 frequencies spaced at 100-kc intervals between 4.553 and 5.453 mc. The output frequency is determined by the setting of the front panel 100-kc (KCS) control. If a low band injection frequency is required, the 17.847-mc output from the 17.847-mc mixer is additively mixed in the lo-band mixer with the output from the 100-kc oscillator (4.553 mc to 5.453 mc, in 100-kc steps) to provide a frequency in the 22.4 to 23.3-mc range. If a high-band injection frequency is required, the 27.847 mc output from the 27.847-mc mixer is additively mixed in the high-band mixer with the output from the 100-kc oscillator (4.553 mc to 5.453 mc, in 100 kc steps) to provide a frequency in the 32.4- to 33.3-mc range. In either case, the resultant frequency is applied to the midfrequency mixer.

The 100-KC Synthesizer Electronic Subassembly employs an error cancelling loop to ensure the accuracy of the 100 kc-injection frequencies. The 500-kc output from the multiplier-divider is applied to the 100-kc spectrum generator to produce a spectrum of frequencies spaced at 100-kc intervals between 15.3 mc and 16.2 mc. The output from the 100-kc oscillator (4.553 mc to 5.453 mc, in 100-kc steps) is applied to the 10.747-mc mixer, where it is mixed with that spectrum point of the 100-kc spectrum which will result in an output of 10.747 mc. The 10.747-mc signal is additively mixed with the 7.1-mc output from the 7.1-mc mixer to produce the 17.847-mc signal, which is used in one of two mixing processes. It is mixed with the 100-kc oscillator output to cancel any oscillator frequency error and produce the low-band injection frequencies, or it is mixed with the 10 mc-output from the multiplier-divider. This mixing produces a 27.847-mc signal, which is mixed with the 100-kc oscillator output to cancel any oscillator frequency error and produce the high-band injection frequencies. The high or low band of injection frequencies is determined by the voltage level on the high/low band control line output from the code generator. If an error were present in the 100-kc oscillator output, it would be cancelled in this mixing scheme. This is accomplished as follows. Assume that the output from the oscillator should be 4.553 mc, but is 200 cycles high (4.5532 mc), and that the desired frequency output is 22.4 mc (in the low band). The subtractive mixing of the oscillator output with whichever 100-kc spectrum point will produce an output as close as possible to 10.747 mc, results in a 10.7468-mc output (15.3 mc - 4.5532 mc = 10.7468-mc).

This signal is then additively mixed with the 7.1-mc signal, producing a 17.8468 mc output. The 17.8468-mc signal is then additively mixed with the oscillator output (17.8468-mc +4.5532 mc = 22.4 mc), resulting in the desired 22.4-mc output. Assume that the output from the oscillator should be 4.953 mc, but is 300 cycles low (4.9527 mc), and that the desired frequency output should be 32.8 mc (in the high-band). Subtractively mixing the 100-kc spectrum point (15.7 mc) with the 4.9527-mc signal results in an output of 10.7473 mc. This signal is then mixed with the 7.1-mc signal, resulting in a frequency of 17.8473 mc. The 17.8473-mc signal is further mixed with the 10-mc signal to obtain a frequency of 27.8473 mc, which is additively mixed with the 4.9527-mc output from the oscillator to obtain the required 32.8-mc output. Therefore, it can be seen that any error existing in the output from the 100-kc oscillator will be cancelled, resulting in the exact 100-kc injection frequency required.

The injection frequencies used in the third frequency conversion in the mixers circuit are generated within the set and 10-KC Synthesizer Electronic Subassembly. This circuit consists of two crystal oscillators, each of which has 10 possible output frequencies. The output from the 1-kc oscillator (1.850 mc to 1.859 mc in 1-kc steps) is determined by the setting of the front panel 1-kc (KCS) control, and the output from the 10-kc oscillator (5.25 mc to 5.16 mc, in 10-kc steps) is determined by the setting of the front panel 10-kc (KCS) control. The outputs from the two oscillators are subtractively mixed to provide one of 100 possible output frequencies spaced at 1-kc intervals between 3.301 mc and 3.400 mc. The output is applied to the low frequency mixer.

Any error existing in the 1- and 10-kc oscillators is cancelled in the following manner. The 100-kc pulses from the 100-kc spectrum generator are applied to the 10-kc spectrum generator producing an output from 3.82 to 3.91 mc in 10-kc increments. The 10-kc spectrum generator also produces 10-kc pulses which are applied to the 1-kc spectrum generator to produce a spectrum of frequencies spaced at 1-kc intervals between 0.122 mc and 0.131 mc. The output from the 10-kc oscillator (5.25 mc to 5.16 mc, in 10-kc steps, is additively mixed with whichever spectrum point of the 10-kc spectrum will result in a frequency of 9.07 mc. The output from the 1-kc oscillator (1.850 mc

to 1.859 mc in 1-kc steps) is additively mixed with whichever spectrum point of the 1-kc-spectrum will result in a frequency of 1.981 mc. The 1.981-mc and the 9.07-mc signals are then subtractively mixed, producing the 7.089-mc signal, which contains the errors of both oscillators. The 1-kc spectrum generator also produces 5-kc pulses, which are applied to the 5-kc spectrum generator to produce an output, consisting of two spectrum points, 110 kc and 115 kc. These spectrum points are used to lock the output frequency of the 500-cps phase-locked oscillator to 110 kc or 115 kc when desired. With the front panel CPS switch in the 000 position, the output from the phase-locked oscillator is 110-kc and is locked to that exact frequency by the 110 kc spectrum point applied to the phase detector. This 110-kc signal is divided by ten and applied to the 7.1-mc mixer, where it is additively mixed with the 7.089-mc output from the 7.089-mc mixer. The resulting 7.1-mc signal is then applied to the error loop of the 100-KC Synthesizer Electronic Subassembly. Therefore, if an error exists in the 1- or 10-kc oscillators, the same error will exist in the 100-kc injection frequencies. This error is then cancelled in the low frequency and midfrequency mixers of the mixers circuit in the following manner. Assume that the output from the 10-kc oscillator should be 5.25 mc, but is actually 5.2502. Also assume that the output from the 1-kc oscillator should be 1.852 mc but is actually 1.8521 mc. Subtractively mixing these two frequencies results in an injection frequency to the low frequency mixer of 3.3981 mc, rather than the desired 3.3980 mc. Therefore, a 100-cycle error exists in the injection signal. The additive mixing of the 5.2502-mc signal and the 10 kc spectrum point (3.82 mc) results in a frequency of 9.0702 mc. The additive mixing of the 1.8521-mc signal and the 1-kc spectrum point (0.129 mc) results in a frequency of 1.9811 mc. Subtractively mixing the 9.0702-mc and the 1.9811-mc signals results in a frequency of 7.0891 mc. The 7.0891-mc signal is mixed with the 11 kc signal from the divide-by-ten-circuit resulting in a frequency of 7.1001 mc, which is mixed with the 10.747-mc signal to produce a frequency of 17.8471 mc. If the output from the 100-kc oscillator is assumed to be 4.553 mc, then the 100-kc injection frequency would be 22.4001 mc. The 100-kc injection is then also 100 cycles high. Therefore, when the 1-and 10-kc injection frequency of 3.3981 mc (which is 100 cycles high) is

subtractively mixed in the low frequency mixer with the output from the mid frequency mixer (which is 100 cycles high), the error will be cancelled. Therefore, since any error that existed in the 1- and 10-kc injection also exists in the 100-kc injection, the error is cancelled during the translation process.

The R-1051/URR can be tuned to 0.5-kc increments using the 000 and the 500 positions of the CPS switch as well as any frequency inbetween, using the VERNIER position of the CPS switch and the VERNIER control, both located on the front panel. When the CPS switch is in the 000 position, the phase-locked oscillator output in the 500-cps oscillator circuit is locked to 110 kc. Therefore, when the 11.0-kc signal (after division by ten) is mixed with the 7.089-mc error frequency, a frequency of 7.1000 mc is obtained. When the CPS switch is in the 500 position, the output from the phase-locked oscillator is locked to 115 kc. Therefore, when the 11.5 kc (after division by ten) is mixed with the 7.089-mc error frequency, a frequency of 7.1005 mc is obtained. Therefore, the 100-kc injection frequency will be 500 cps greater. Thus, the output from the midfrequency mixer may be varied in 500 cps increments. When the CPS switch is in the vernier position, the output from the phase-locked oscillator can be varied between 108 and 122 kc. As a result, the 7.1000-mc error frequency can be varied between 7.0998 and 7.1012 mc. Thus, the output from the midfrequency may be varied continuously between any two 1-kc increments.

The r-f signal from the antenna is converted to the 500-kc i-f as follows: Assume that the front panel controls are set for a frequency of 13,492,500 cps. The 1-mc injection corresponding to the selected mc digits (13) is 16.5 mc in the high-band. The 16.5-mc is additively mixed in the high frequency mixer with 13,492,500 cps producing 29,992,500 cps which is filtered and applied to the midfrequency mixer. Since the mc digits (13) are in the high-band and the CPS switch is in the 500 position, the 100-kc injection frequency corresponding to the 100-kc digit (4) will be 32.8005 mc. The midfrequency mixer subtractively mixes the 29,992,500 cps and the 32.8005 mc thereby producing a frequency of 2,808,000 cps, which is filtered and applied to the low frequency mixer. The 1- and 10-kc injection is that frequency of the 10-kc oscillator corresponding to the 10-kc digit (9) minus that frequency of the 1-kc oscillator corresponding to the 1-kc digit (2). This results in an injection frequency (5.16 mc minus 1.852 mc)

of 3.308 mc. The 3.308 mc is subtractively mixed with the 2,808,000 cps producing the 500-kc i-f. Similarly, any frequency between 2 and 30 mc may be translated into the 500-kc i-f.

The operating voltages for all circuits in the R-1051/URR are produced by Power Supply Electronic Assembly (fig. 4-11). The 105 to 125-v a-c primary power is converted to d-c voltages of 110 volts (r-f amplifier tubes plate and screen supply) -30 volts (r-f amplifier tubes bias), and 28 volts (general use). The 28 v DC is also regulated to 20 v DC for use in all semiconductor circuits of the R-1051/URR.
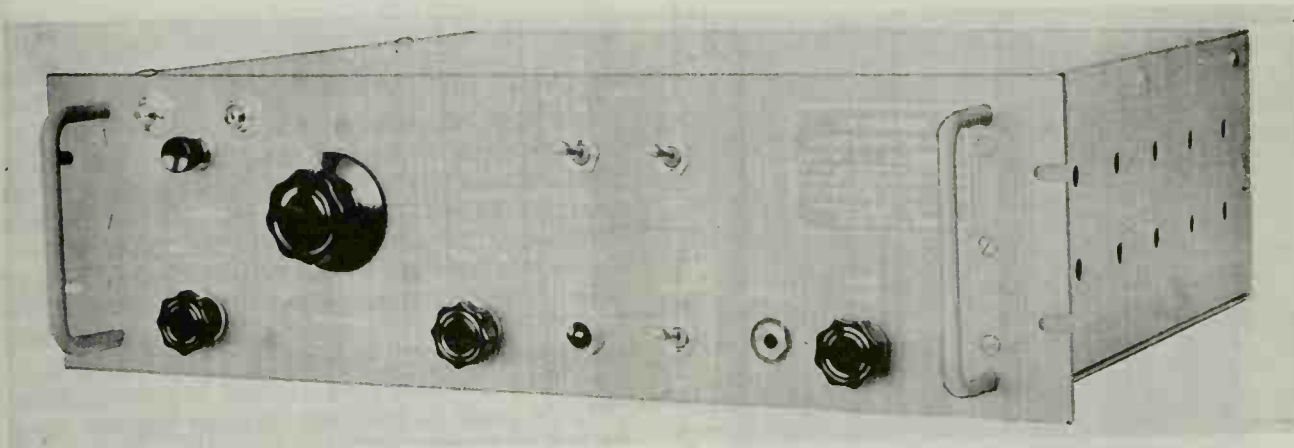
SSB CONVERTER, CV-591A/URR

The CV-591A/URR single sideband converter (fig. 4-12) is used to convert standard communication receivers such as the R-390/URR for SSB use.

The overall selectivity of most receiving systems is greatly sharpened, rejecting unwanted adjacent signals or interference with no detrimental effect to the desired signal. The tuning of single-sideband signals is greatly simplified since the final tuning is done at the converter, not the receiver. A mechanical and electrical bandspread tunes over the i-f bandpass. This effective vernier easily tunes SSB or exalted carrier AM signals within cycles of correct tone. Either sideband is selectable, either with the bandpass tuning feature or by inverting the oscillator separation. Continuous Wave, MCW, and FS signals are easily tunable with the bandspread feature.

For extreme stability, the first oscillator is switched to crystal control for both upper and lower sideband positions.

The local or remote tuned VFO feature of the converter permits operation with any receiver having an i-f nominally centered at 455 kc. When the oscillator is switched to crystal control and the proper crystals inserted however, most any receiver i-f may be accommodated.

All operational controls are located on the front panel. These controls are similar in function and effect to those found on any receiver. The BANDSPREAD control tunes the converter over a limited frequency range. A MANUAL/XTAL switch sets the first oscillator to either variable or fixed crystal operation. The BFO, AVC and AUDIO GAIN controls perform similar functions as on a receiver.

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Figure 4-12.—SSB converter, CV-591A/URR, front view.

Terminals at the rear panel provide simple connections for remote control of the main features of the converter without modifications or the use of additional lines or tones.

By this means it is possible to remotely or locally tune the converter across the receiver i-f passband, select sidebands with a remote indication of which sideband is in use, and still retain all of the remote control features of the remote control system. The converter may be used with remote control system AN/FRA-19(V) or AN/FRA-501, without modification to the system.

BLOCK DIAGRAM

The block diagram (fig. 4-13) presents a simplified outline of the functions of the converter and how they are accomplished.

The AVC Amplifier and Rectifier (V1), provides two kinds of control on the first amplifier. A fast time constant is used to control signals which have an ever-present component; i.e., AM, MCW, FS. A slow time constant will hold the amplifier gain steady for a longer period for SSB and CW signals. The gain of the amplifier is wide open when the AVC is OFF.

The first mixer (V3) (1st product detector), mixes the incoming signal with the first oscillator to place the signal in proper position relative to the bandpass filter. Since the filter extends from 17.3 kc to 20.5 kc and the 17-kc BFO follows, the signal carrier position must be placed at 17 kc to produce the correct tone.

The first oscillator (V7), provides the tunable frequency source for correctly tuning the

signal in the bandpass filter. It may be either variable or crystal controlled. When variable it is tuned with the front panel BANDSPREAD control or with the reactance control for remote operation.

The reactance modulator (V8) is placed across the first oscillator tuned circuit and varies the reactance to change frequency. A 4.5-v DC change in output to the reactance stage is sufficient to vary the oscillator 3 kc. The voltage is obtained from the BFO channel of the Remote Control Receiver System.

As the r-f gain control on the remote control system is turned to reduce the gain of the receiver the required voltage out of the remote control system performing this action reaches a certain level. When the level is such that the receiver is fully silent, its negative level will be sufficient to cut off the relay driver (V9A) and deenergize sidetone relay K3 (not shown). This action in turn trips the sideband two-step relay K2.

Operation of the relay driver or the front panel push-button advances the sideband selector relay K2 one position. This changes the first oscillator frequency from upper to lower sideband selecting frequency or vice versa.

After the first mixer, the filter (Z1) passes a band of frequencies between 17.3 kc and 20.5 kc. The skirts of the selectivity curve provide the sharpness required for attenuation of undesirable adjacent signals.

The signal passed by the bandpass filter to the second detector (V4) is demodulated in either of two ways. If a carrier is normally present
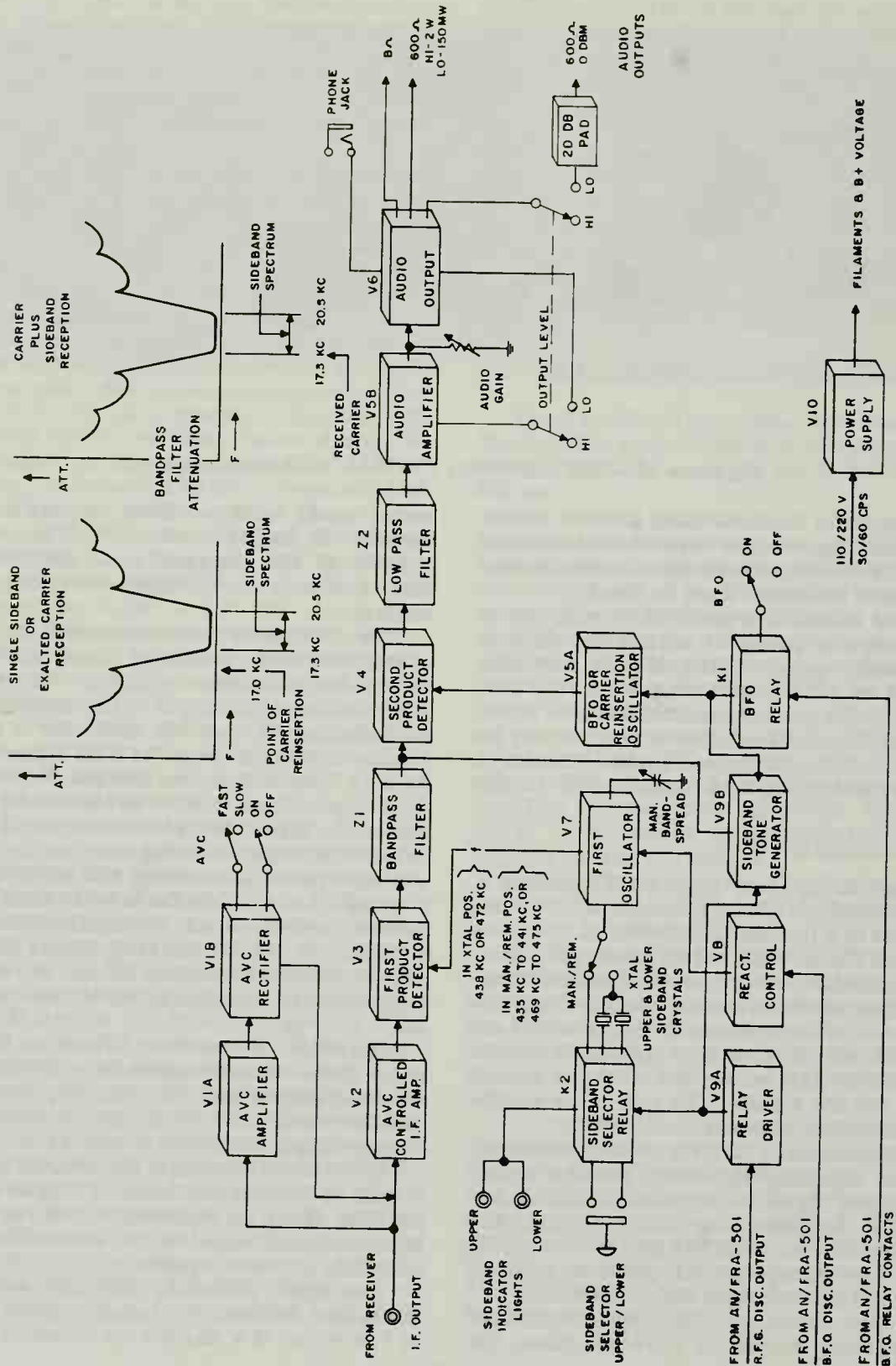
Figure 4-13.—SSB converter CV-591A/URR, block diagram.

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nd allowed to pass through the filter, such as
AM, normal demodulation will take place. If
o normal carrier is present, such as SSB, a
carrier is reinserted.

The BFO or carrier reinsertion oscillator
(V5A) is a stable 17-kc oscillator which provides
he reinserted carrier for SSB and exalted AM,
and produces the beat note for CW and FS signals.

The BFO relay (K1) turns the BFO on or off
by proper actuation of the relay. The front panel
BFO switch or a remote paralleled switch ac-
tuates the relay.

Sideband tone generator (V9B), by proper
manipulation of the r-f gain control of the re-
mote control system, generates a tone, high or
low, which will indicate to the remote operator
which sideband (upper or lower) is in use. The
tone becomes strong enough to be heard only
after the r-f gain has succeeded in fully silenc-
ing the normal receiver action.

Following the second mixer, the low pass
filter (Z2) attenuates any products above 5 kc.
The resultant is the desired audio.

The audio amplifiers (V5B, V6) provide
three levels of output. The output level control
in the HIGH position permits the full 2 watts of
the amplifier to be available for the 600-ohm
line or 8-ohm loudspeaker. In the LOW position,
the gain of the amplifier is reduced so that 150
milliwatts are available for the 600-ohm line or
8-ohm loudspeaker, and 0 dbm or 1 milliwatt
for 600-ohm telephone line. A phone jack on the
front panel permits monitoring of the output
without disabling the speaker.

The power supply (V10) is self-contained and
provides the necessary a-c filament and d-c B+
voltages. A regulated +150 v stabilizes the first
oscillator and BFO.

RADIO RECEIVER SET AN/FRR-60(V)

The AN/FRR-60(V) (fig. 4-14) is a general
purpose Diversity Radio Receiver system de-
signed mainly for use at Navy shore communi-
cation receiver sites. This receiver is the com-
panion equipment for the AN/FRT-39/40 shore
transmitters (discussed in chapter 6). The re-
ceiver covers the radio frequency range of 2
to 32 mc. The equipment is capable of reception
in the following modes of operation:
    (1) SSB (suppressed carrier).
    (2) SSB (with carrier).
    (3) DSB (suppressed carrier).
    (4) DSB (with carrier).
    (5) FSK.
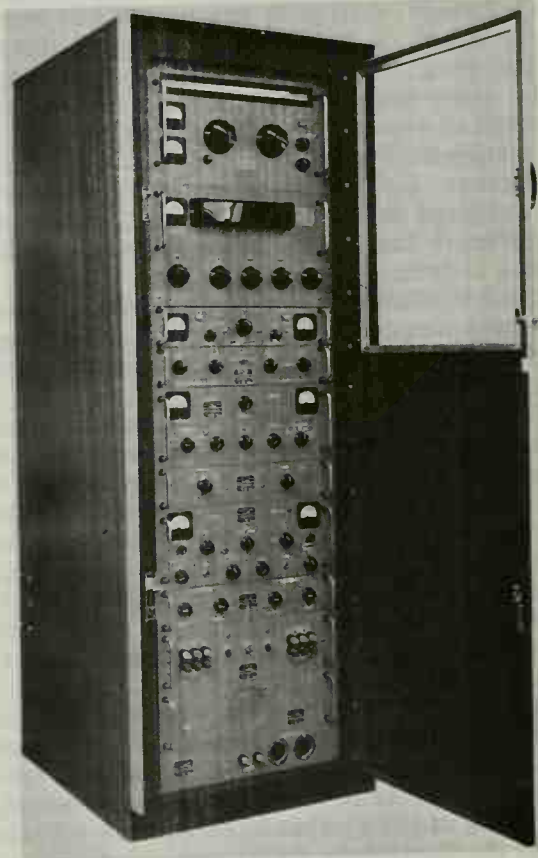
1.348

Figure 4-14.—Diversity Radio Receiver,
AN/FRR-60(V).

    (6) FAX.
    (7) CW.
    (8) MCW.
    (9) Pulse modulation.
    (10) Phase modulation.
The 2- to 32-mc range covered by the AN/
FRR-60(V) is divided into eight continuous
bands. Continuous coverage is provided in either
synthesized (controlling the receiver VFO with
an external frequency standard) or non-synthe-
sized type of operation. In the synthesized type
of operation a frequency standard output is
provided over the entire frequency range of the
receiver in 100-cycle steps. One complete
AN/FRR-60 by itself can only be operated in
nondiversity; however, when two complete AN/
FRR-60's are operated together and their AGC
lines are interconnected, the equipment may be
operated in either space diversity or frequency
diversity. In space-diversity operation the two

receiver antenna inputs are connected to separate antennas which are spaced several wavelengths apart. In frequency-diversity operation the two receiver antenna inputs are connected to the same antenna but the two receivers are operated on separate frequencies which are being transmitted simultaneously with identical intelligence. Both types of diversity operation are designed to overcome the difficulties encountered in long-range communications which are due to fading.

The AN/FRR-60(V) comprises various modular units mounted in a single rack, and is used in fixed-station or mobile communications systems.

The receiver is capable of receiving input signal variations of 70 db without the use of AGC and signal variations of 100 db with the use of AGC. These variations will not affect the usability of receivers output signal. Further, the receiver will continue to provide a useable output signal over a dynamic range of 150 db. Front panel controls are provided so that the AGC voltage can be adjusted to cope with various signal conditions.

Front-panel meters are provided on the receiver to facilitate the monitoring of r-f input signals, audio output signals, automatic frequency control drift, carrier level, i-f output, and locking of the synthesizer frequency with the HFO of the Continuous RF Tuner.

An internally generated low-level r-f alignment signal, which is provided by the synthesizer, facilitates accurate and rapid tuning of the complete system in the absence of any received radio signal. The locally generated signal is also useable as a maintenance "tool" for checking the alignment of the AN/FRR-60(V).

A forced-air cooling system using electric blowers incorporated within the equipment cabinet. Washable air filters are used to filter out external dust.

CONTINUOUS RF TUNER TN-376/UR

The Continuous RF Tuner (fig. 4-15) provides for frequency coverage from 2 to 32 mc in eight bands and displays the tuned frequency on a 14-inch slide-rule type scale. The tuner will accept synthesized control voltage from the synthesizer unit 0-941/UR for extreme frequency accuracy and stability. The RF Tuner converts the input r-f frequency to their lower first i-f of 1.75 mc. The TN-376/UR obtains its operating power from an external power supply unit, PP-3341/FRR-60(V).

The front panel controls and instruments of the RF Tuner (see fig. 4-16) are arranged to permit ease of tuning through any of the eight r-f bands to the desired frequency within the range of 2 to 32 mc, non-synthesized continuous coverage operation or synthesized operation in 100 cycle steps, and alignment and sensitivity checks of the tuner. All operating controls are identified by front panel markings. The following numeral and panel designations list the controls, instruments and the function of each (see fig. 4-16).

(1) MEGACYCLES. The megacycle indicator displays the r-f band selected by operation of the band control knob; refer to designation number (3).

(2) TUNE. The tuning knob moves the slide-rule pointer along the megacycle indicator to the appropriate frequency desired. This control is fitted with a locking device; refer to designation number (7).

(3) BAND. The band control rotates the illuminated MEGACYCLE indicator to the desired r-f band. The r-f bands are arranged as follows:

BAND 1	2-3 mc
BAND 2	3-4 mc
BAND 3	4-6 mc
BAND 4	6-8 mc
BAND 5	8-12 mc
BAND 6	12-16 mc
BAND 7	16-24 mc
BAND 8	24-32 mc

(4) SYNC IND. The synchronizing indicator lamp lights to indicate that the synthesizer is in synchronization with the HFO in the r-f tuner unit.

(5) NOISE SILENCER/OFF/ALIGNMENT SIGNAL. In the NOISE SILENCER position, this three position control activates the noise silencer circuits; disables the alignment signal generator, and connects the antenna (ANT) input jack to the first r-f amplifier circuit. In the OFF position, the control disables the output of the noise silencer, disables the alignment signal generator, and connects the ANT input jack to the first r-f amplifier circuit. In the ALIGNMENT SIGNAL position the control activates the alignment signal generator, disables the output of the noise silencer, and connects the calibration (CAL) input jack to the first r-f amplifier circuit.
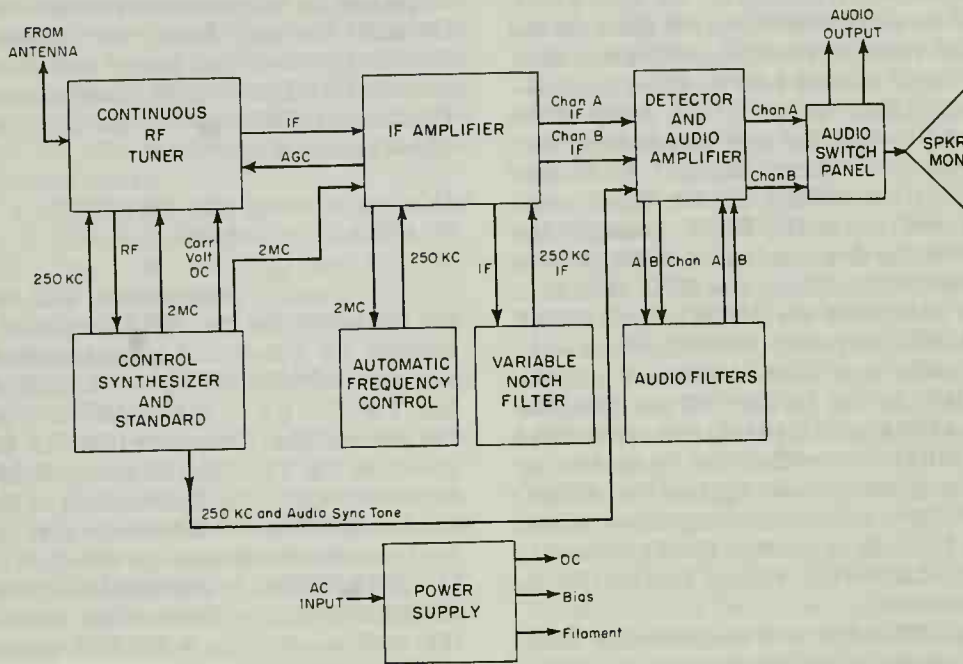
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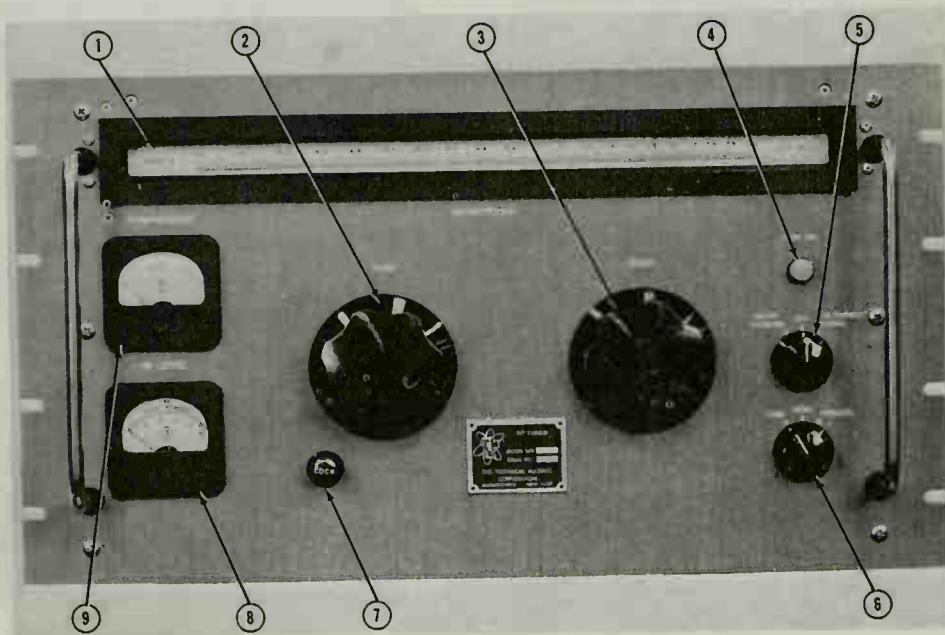Figure 4-15.—AN/FRR-60(V), simplified block diagram.

Figure 4-16.—Continuous RF Tuner, TN-376/UR.

(6) TUNE/SYNC/OPERATE. In the TUNE position, this three position control grounds the d-c correction voltage from the synthesizer unit, causing the HFO circuit to run free, and de-energizes the AUDIO SYNC TONE relay of the audio frequency amplifier unit so as to remove the audio sync tone from channel A. In this position the SYNCHRONIZE meter should read zero (center scale). In the SYNC position the control grounds the d-c correction voltage from the synthesizer unit, causing the HFO circuit to run free, and energizes the AUDIO SYNC TONE relay of the audio frequency amplifier unit so as to inject an audio sync tone in channel A of the audio amplifier. In the OPERATE position, the control causes the synthesizer circuits to control the HFO circuits for synthesized operation in 100-cycle steps, and de-energizes the AUDIO SYNC TONE relay of the audio amplifier unit so as to remove the audio sync tone from channel A.

(7) LOCK. Locks the tuning control (2) in the desired position.

(8) RF LEVEL. The r-f level meter indicates the strength of the antenna input signal or the alignment signal in db above a reference level of $1 \mu V$.

(9) SYNCHRONIZE. The sync meter indicates the amount and polarity of the d-c voltage required to keep the HFO in sync with the synthesizer output. When the system is out of synchronization, the meter will read zero (center scale). This meter functions only when the Continuous RF Tuner is operating in the synthesized mode of operation.

CONTROL SYNTHESIZER AND STANDARD 0-941/UR

The Control Synthesizer and Standard (fig. 4-15), monitors the HFO frequency of the Continuous RF Tuner and provides correction voltage to maintain the free-running oscillators in the TN-376/UR to a stability of 1 part in 10^8 in 24 hours. The frequency that is to be synthesized in the TN-376/UR by the Synthesizer unit is displayed on the front-panel of the 0-941/UR in 1-inch illuminated numerals. A change of synthesized frequency in 100-cycle increments is accomplished by means of manually controlled detented switches located on the front panel of the equipment. The 0-941/UR obtains its operating power from an external power supply unit PP-3341/FRR-60(V).

The controls and indicators of the Control Synthesizer and Standard (fig. 4-17) are arranged to permit the selection of any frequency from 2
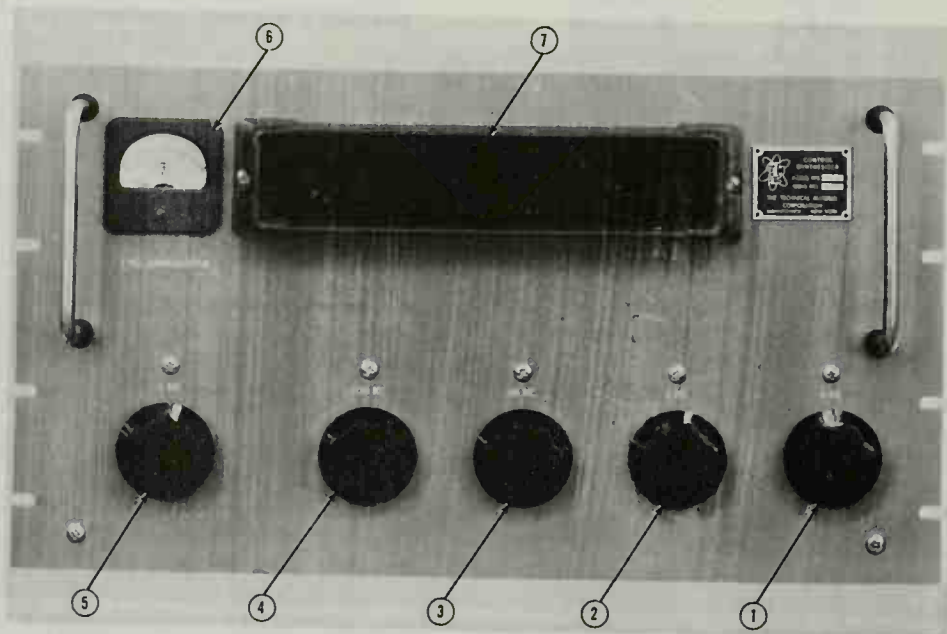

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Figure 4-17.—Control Synthesizer and Standard, 0-941/UR.

to 32 mc in 100-cycle steps. By rotation of the frequency selector switches the used frequency is indicated by the nixies (solid state or vacuum tube, luminous numerical indicating devices) on the front panel. The following numeral and panel designations list the controls, instruments and function of each (see fig. 4-17).

(1) 0.1 KC. The 0.1 kc control tunes the synthesizer in 100-cycle steps.

(2) 1 KC. The 1-kc control tunes the synthesizer in 1,000-cycle steps.

(3) 10 KC. The 10-kc control tunes the synthesizer in 10-kc steps.

(4) 100 KC. The 100-kc control tunes the synthesizer in 100-kc steps.

(5) 1 MC. The 1-mc control tunes the synthesizer in 1-mc steps.

(6) 1-MC COMPARATOR. The 1-mc comparator meter indicates the frequency error of the internal 1-mc frequency standard.

(7) NIXIE FREQUENCY INDICATOR. It indicates (using nixie units) the frequency that the equipment is tuned to.

AUTOMATIC FREQUENCY CONTROL

UNIT C-4099/FRR-60(V)

The Automatic Frequency Control Unit (AFC) (fig. 4-15), accepts a 250-kc input signal from the IF Amplifier Unit, AM-3295/FRR-60(V), and provides automatic frequency control which will compensate for a frequency drift in the received transmission. The AFC unit automatically synchronizes to a received signal within ± 50 cps when the received signal is suppressed as much as 30 db at 1 microvolt above noise threshold and will remain synchronized for approximately ± 1,000 cps of drift at a maximum drift rate of 10 cps/second. The C-4099/FRR-60(V) obtains its operating power from an external power supply unit, PP-3341/FRR-60(V).

The AFC unit (fig. 4-18) is normally operated with the CARRIER selector switch in the OSC position. With the switch in the OSC position, the receiver product detector receives a corrected 250-kc injection frequency from the AFC product detector oscillator. Also, in this position, the AFC responds to correct the carrier drift at the speed rated (10 cps/second maximum): the 250-kc injection frequency will be in phase with the carrier input within 1 cycle (or 360°). In cases where phase is of primary importance, it may be elected to place the CARRIER SELECTOR switch in RCC (reconstructed carrier) position. This effectively strips the 250-kc carrier of its sidebands, amplifies the carrier, and feeds this carrier back into the receiver product detector as the injection frequency. The following numeral and panel designations list the controls, instruments and the function of each (see fig. 4-18).

(1) DRIFT. The DRIFT meter indicates the total drift of the receiver i-f carrier. A center scale reading indicates zero drift. The meter scale is color coded as follows:

Color	Drift (approx)
Green	500 cps
Yellow	500 cps to 1 kc
Red	1 kc or more

(2) ALARM. The alarm lamp will light when the carrier frequency has drifted off center frequency by more than ± 750 cps.

(3) SENSITIVITY. The SENSITIVITY controls the gain of the carrier amplifier stage. The sensitivity may be increased or decreased depending on the ratio of signal to noise.

(4) RESET. The RESET control re-centers the AFC oscillators when the operator is required to tune the receiver to another frequency or resynchronize the receiver due to a drifting signal.
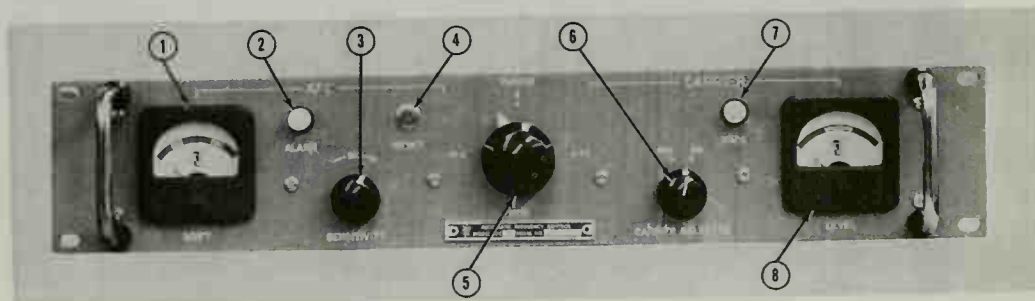
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Figure 4-18.—Automatic Frequency Control, C-4099/FRR-60.

(5) TUNING KCS. The TUNING KCS control tunes the AFC unit's converter injection oscillator to enable the operator to synchronize to the received signal.

(6) CARRIER SELECTOR. The carrier selector control is used to select either the product detector oscillator (OSC) or the reconstructed carrier (RRC) position. The function of both positions was explained in a previous paragraph.

(7) FADE. The FADE lamp will light when the received signal fades below a predetermined level.

(8) LEVEL. The LEVEL meter indicates the strength of the received carrier.

IF AMPLIFIER AM-3295/FRR-60(V)

The IF Amplifier unit (fig. 4-15), accepts a 1.75-mc input signal from the Continuous RF Tuner, TN-376/UR, processes the input signal through manually controlled selectable bandpass filters, and converts the input signal to 250 kc for further demodulation in the Audio Amplifier, AM-3298/FRR-60(V). Rear panel facilities are provided for connecting the Notch Filter, F-711/FRR-60(V).

The i-f amplifier unit (fig. 4-19) is designed to operate with a high degree of versatility. The two i-f channels are identical and the controls for each channel are functionally grouped. By the use of two IF BANDWIDTH KC front switches, either a symmetrical upper or lower sideband may be switched to channel A or B. This type of versatility is helpful, especially during independent sideband operation where voice information may be transmitted on one sideband and multichannel information on the other sideband. When information is transmitted in this manner, sideband orientation of information need not be maintained at the transmitter. The following numerical and panel designations list the controls, instruments and the function of each (see fig. 4-19).

(1) and (3) OUTPUT. The OUTPUT meter indicates the output level of the channel.

(2) MANUAL GAIN. The MANUAL GAIN controls the overall gain of the i-f amplifier.

(4) and (8) IF BANDWIDTH KC. The IF BANDWIDTH KC controls the bandpass of the i-f amplifier and selects either the upper or lower sideband in single-sideband operation, or both sidebands in double-sideband operation.

(5) and (7) AGC DECAY. The AGC DECAY controls the amount of time it takes the AGC to decay after a signal starts to fade.

(6) AFC ON-OFF. The AFC ON-OFF activates or disables the AFC circuits.

VARIABLE NOTCH FILTER F-711/FRR-60(V)

The Variable Notch Filter (fig. 4-15), accepts a 250 kc input signal from the IF Amplifier and, when used, will attenuate an interfering signal within ± 8 kc of the 250-kc input signal. Filtering by the F-711/FRR-60(V) can be engaged or disengaged by means of front panel
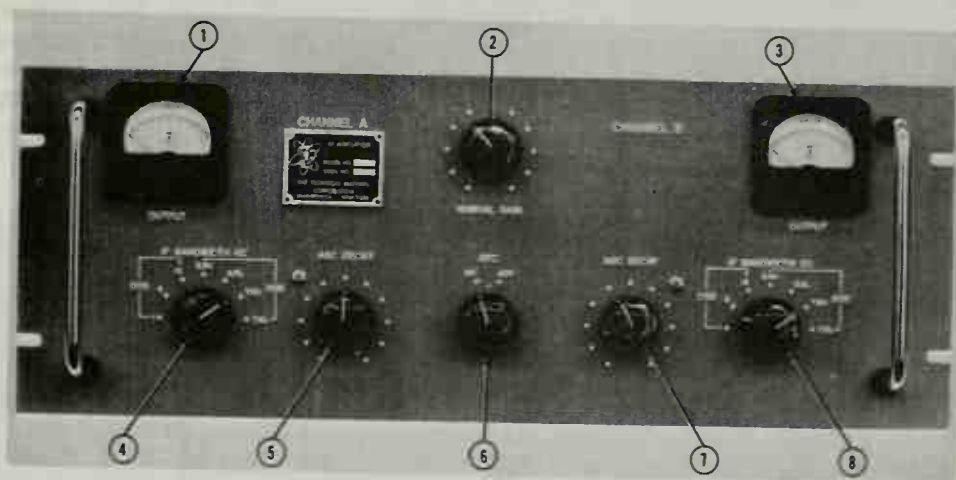

Figure 4-19.—Intermediate Frequency Amplifier, AM/3295/FRR-60(V).

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ntrols. Signals from the Variable Notch Filter
e returned to the IF Amplifier for the purpose
bandpass filtering. The F-711/FRR-60(V) ob-
ins its operating power from an external power
pply unit, PP-3341/FRR-60(V).

The Variable Notch Filter (fig. 4-20) con-
ols are located on the front panel and allow
e operator to select an audio filter to tune an
wanted signal into a notch filter for elimina-
on from the desired signal. If an i-f signal
ists with no interfering signal, the notch may
turned off. The following numeral and panel
esignations list the controls and the function of
ch (see fig. 4-20).

(1) NOTCH ADJUST. The NOTCH ADJUST
ntrol is used to vary the center frequency of
e notch filter either plus or minus from the
nter i-f by as much as ±8 kc.

(2) NOTCH OFF-ON. The NOTCH OFF-ON
ntrols the 250-kc i-f input to the filter. In the
FF position the signal bypasses the filter.

DETECTOR AND AUDIO AMPLIFIER
AM-3298/FRR-60(V)

The Detector and Audio Amplifier unit (fig.
4-15), accepts dual 250-kc input signals from
the IF Amplifier unit, demodulates these signals
and provides dual audio channel outputs. Facili-
ties are incorporated on the rear panel to chan-
nel the output signals to a passive dual audio
filter. The AM-3298/FRR-60(V) obtains its
operating power from an external power supply
unit, PP-3341/FRR-60(V).

The Detector and Audio Amplifier (fig. 4-21)
is divided into two identical channels (A and B).
Each channel has identical controls which are
functionally grouped. The following numeral and
panel designations list the controls and instru-
ments and the function of each (see fig. 4-21).

(1) POWER. The POWER control switch has
two positions. The STANDBY position is used to
disconnect the operating voltages from the unit.

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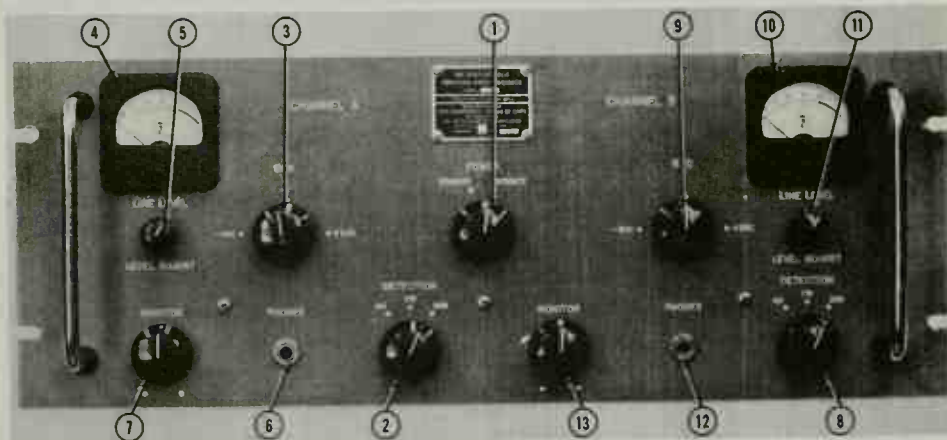
Figure 4-20.—Variable Notch Filter, F-711/FRR-60(V).

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Figure 4-21.—Detector and Audio Amplifier, AM-3298/FRR-60(V).

The OPERATE position is used when the equipment is in operation and connects the operating voltages to the unit.

(2) and (8) DETECTION. The DETECTION control selects the mode of detection and is capable of operating in AM, CW, or SSB, depending on the position of the switch.

(3) and (9) BFO. The BFO control is used in the CW position and will vary the BFO ± 5 kc (maximum) about the center i-f of 250 kc.

(4) and (10) LINE LEVEL. The LINE LEVEL meter indicates the audio level across the output line.

(5) and (11) LEVEL ADJUST. The LEVEL ADJUST control varies the audio output level across the output line.

(6) and (12) PHONES. The PHONES jack is used to connect a standard headset to the unit for the purpose of monitoring the output signal.

(7) and (13) MONITOR. The MONITOR control is used to vary the level of the signal in the monitor phone jack headset.

AUDIO FILTER F-712/FRR-60(V)

The Audio Filter (fig. 4-15) requires no operating voltages other than the signals fed to it by the Detector and Audio Amplifier, AM-3298/FRR-60(V). The F-712/FRR-60(V) accepts dual audio signals from the audio amplifier and is adjustable by front panel controls which vary the upper and lower frequency cutoff point of each audio filter to suit operational requirements. Front panel controls have been engineered to minimize confusion when establishing audio bandpass.

The Audio Filter (fig. 4-22) is used to pass a particular band of frequencies in the audio range for either Channel A or Channel B. An example would be to pass a 2-kc band with the low end of the band starting at 0.5 kc and ending at the high end with 2.5 kc. Set the low cutoff control to 0.5 kc and the high cutoff control to 2.5 kc. As another example, to pass all frequencies above 2.5 kc set the low cutoff control to 2.5 kc and set the high cutoff control to the OUT position. The following numeral and panel designations list the controls and the function of each (see fig. 4-22). (1) and (3) LOW CUTOFF. The control is used to set the low cutoff frequency for the indicated channel. (2) and (4) HIGH CUTOFF. The control is used to set the high cutoff frequency for the indicated channel.
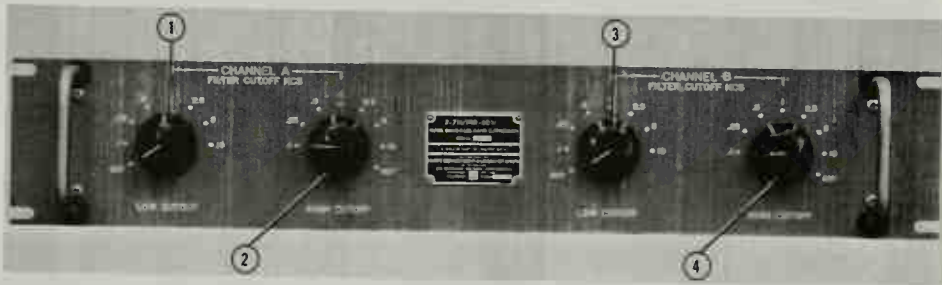

Figure 4-22.—Audio Filter, F-712/FRR-60(V).

Figure 4-23.— Power Supply, PP-3341/FRR-60(V).

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WER SUPPLY PP-3341/FRR-60(V)

The Power Supply (fig. 4-15), provides for gulated B+, regulated bias voltage, and filant voltages for one complete rack of AN R-60(V) units.

The Power Supply (fig. 4-23) contains no nt panel controls. On the front panel are licator lamps and indicator fuses. There are o indicator fuses located on the top panel t behind the front panel.

DIO SWITCH PANEL SB-1865/FR

The Audio Switch Panel (fig. 4-15), filters d interconnects the two audio output channels (Channels A and B) to a terminal strip for further wire transmission to local or remote terminal equipment such as single-channel or multi-channel teletypewriter converter equipment. The Audio Switch provides front panel controls to apply Channel A or Channel B audio signals to the loudspeaker, disconnect audio signals from the loudspeaker, adjust the audio level of signals applied to the loudspeaker, and combine or separate the AGC of both receivers when used in diversity/non-diversity.

SPEAKER PANEL LS-491/FR

The Speaker Panel (fig. 4-15), houses a single 4-inch permanent-magnet (p-m) speaker that is used to monitor the receiver outputs.

CHAPTER 5

RADIO TRANSMITTERS, PART I

This chapter presents a block diagram discussion of the AN/SRT-14, 15, and 16, the AN/WRT-2 Radio Transmitting Set, and the SSB Transceiver AN/URC-32.

RADIO TRANSMITTING SETS AN/SRT-14, 15, and 16

Radio Transmitting Sets AN/SRT-14, 15, and 16 provide radio-frequency energy at any frequency from 0.3 to 26 mc for the transmission of voice, facsimile, and c-w or frequency shift telegraph communication. A manual frequency selection process permits all three sets to cover the entire frequency range in steps of 10 cycles at the 100-watt output level. The AN/SRT-15 and AN/SRT-16 can also operate at the 500-watt carrier level in the frequency range from 2 to 26 mc.

The AN/SRT-14 with its major components consists of a 100-watt transmitter bay and the antenna tuning equipment. The Radio Transmitting Set AN/SRT-15 consists of a transmitter bay and booster and the antenna tuning equipment. The booster provides the necessary facilities for producing the 500-watt operating output level. Therefore, the set may transmit an r-f carrier of either 100 or 500 watts. The Radio Transmitting Set AN/SRT-16 (fig. 5-1) consists of two transmitter bays, each of which is equipped with its antenna tuning units. The dual transmitter is therefore capable of transmitting two carriers simultaneously. One of the outputs is limited to the 100-watt level, while the other may be transmitted at either 100 or 500 watts.

Either of the transmitter sets may be arranged for remote operation with start-stop control, keying, and radiophone. Standard Navy six-wire radio transmitter control circuits are used for all remote control operation.

OVERALL BLOCK DIAGRAM

In the discussion of the AN/SRT-14, 15 and 16 in this chapter, it is sometimes best to use abbreviations for major components of the sets. These major components and abbreviations are:
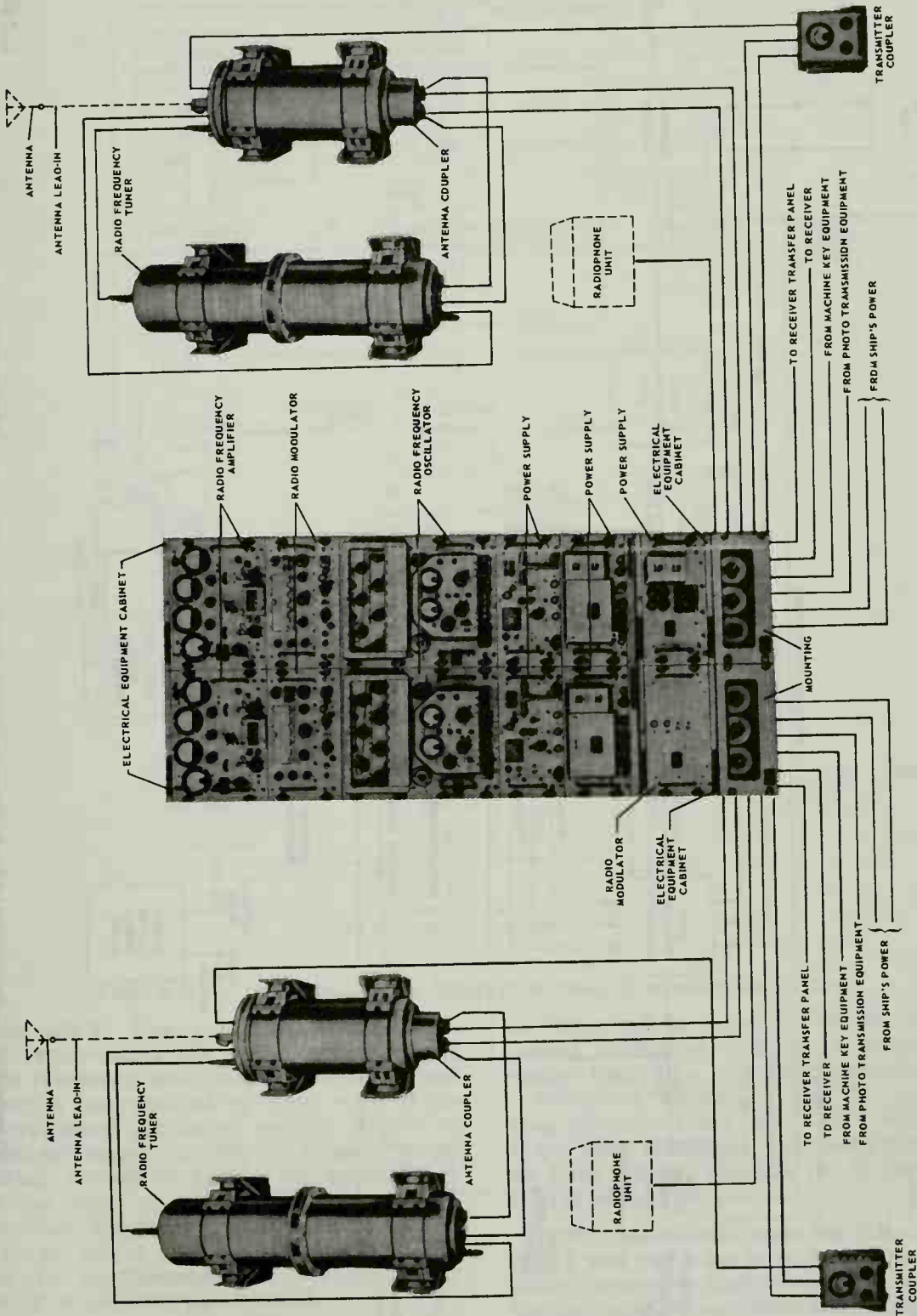
Radio-Frequency Oscillator (RFO)
Low-Voltage Power Supply (LVPS)
Medium-Voltage Power Supply (MVPS)
High-Voltage Power Supply (HVPS)
Low-Level Radio Modulator (LLRM)
High-Level Radio Modulator (HLRM)
Radio-Frequency Amplifier (RFA)
Antenna Tuning Unit (ATU)
Load Adjusting Unit (LAU)

Because of the similarity of the three transmitters, a single functional block diagram is shown in figure 5-2. The diagram is used to show the relationship existing between the major units and the progression of the signal through the set.

Radio-Frequency Oscillator

The radio-frequency oscillator is the master frequency source for the transmitting set. It contains three individual oscillators and several supporting units including frequency converters, step generators, and frequency multipliers, which are necessary to produce the required frequencies for the transmitter. The radio-frequency oscillator unit consists of 14 sub units, which in this discussion, will be referred to by their common name and unit number, as shown in the radio-frequency oscillator block diagram of figure 5-3.
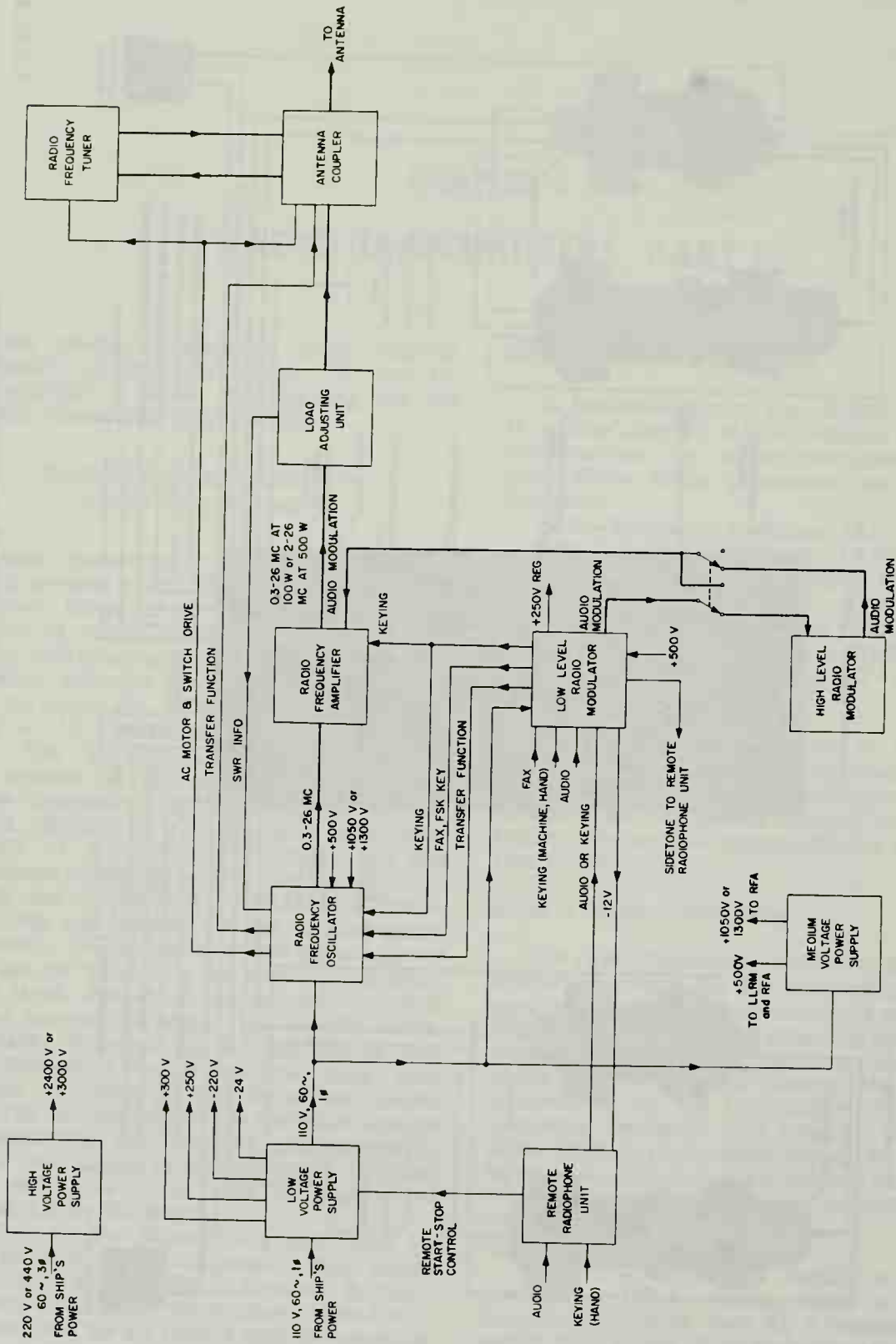
UNIT 1.—The crystal oscillator (unit 1) produces a 100-kc output, which is fed to a frequency multipler (unit 2), a frequency converter (unit 5), the 10-kc step generator (unit 6),

Figure 5-1.—Radio Transmitting Set AN/SRT-16, relationship of units.

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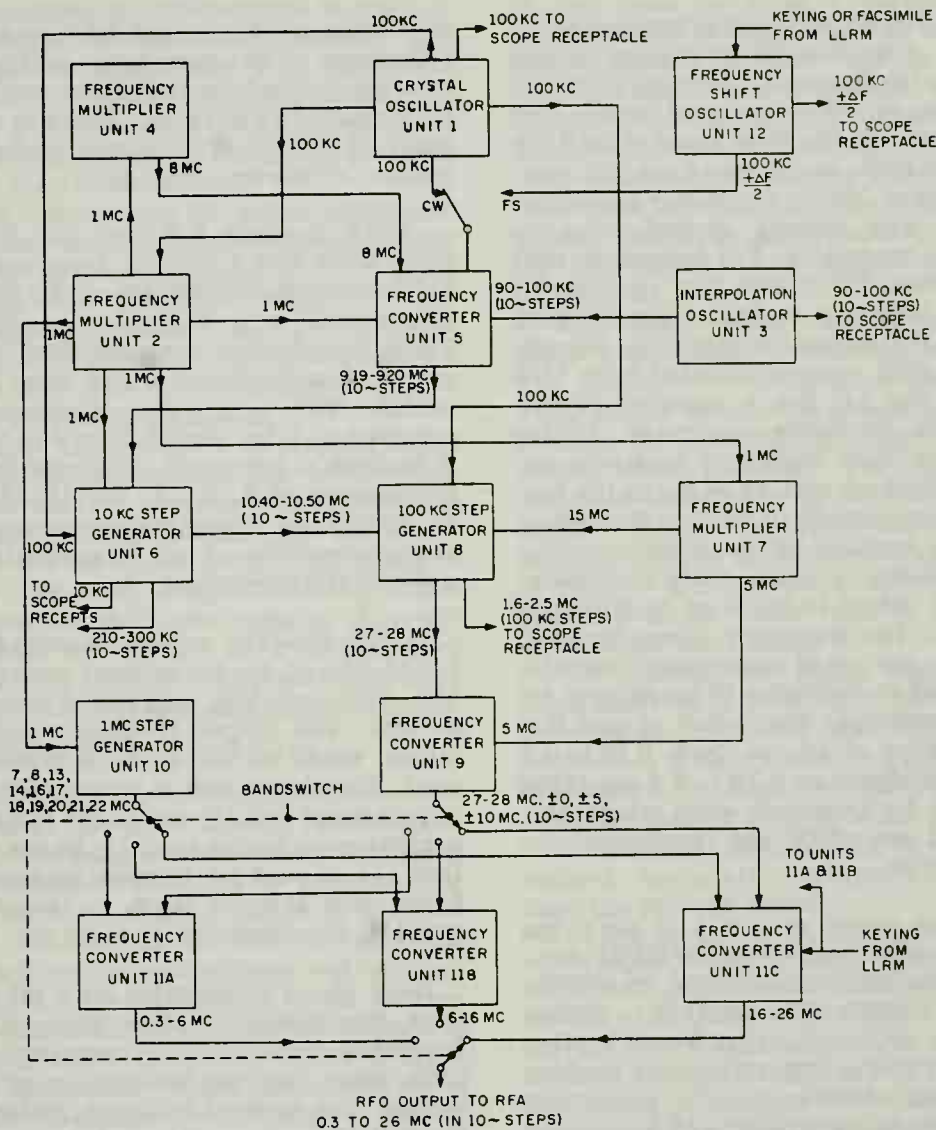Figure 5-2.—Functional block diagram of Radio Transmitting Set AN/SRT-14, 15, and 16.

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Figure 5-3.—Functional block diagram of radio-frequency oscillator.

d the 100-kc step generator (unit 8). This utput synchronizes the step generators, and rough frequency multipliers, provides higher equencies as required by other units of the adio-frequency oscillator. Another portion of e 100-kc output is fed to a zero adjust dicating circuit to enable the operator to neck the output of the interpolation and fre- uency-shift oscillators. Finally, a portion of e 10°-kc output of the crystal oscillator is d to an oscilloscope test receptacle for ervicing or checking as required.

UNIT 2.—In the frequency multiplier (unit , the 100-kc output of the crystal oscillator

is increased to 1 mc and fed to another fre- quency multiplier (unit 4), a frequency con- verter (unit 5), a 10-kc step generator (unit 6), a frequency multiplier (unit 7), and a 1-mc step generator (unit 10). Unit 4 multiplies the 1-mc input frequency from unit 2 and produces an 8-mc output, which is fed to the frequency converter (unit 5).

UNIT 5.—In addition to the 1-mc input from unit 2 and the 8-mc input from unit 4, the fre- quency converter (unit 5) receives outputs from the interpolation oscillator (unit 3) and either the crystal oscillator or the frequency-shift oscil- lator. The interpolation oscillator produces an

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output of 90 to 100 kc, adjustable in steps of 10, 100 and 1000 cps. The major portion of this output is fed to the frequency converter (unit 5). Another portion is fed to an oscilloscope test receptacle located on the front panel of the RFO. The frequency-shift oscillator (unit 12) generates an output of 100 kc, frequency modulated in accordance with signals applied from the low-level radio modulator. For frequency-shift keying operations, the output may be shifted a maximum of ± 1000 cps or $\Delta f/2$ from the knob set frequency. For facsimile operation, the output frequency shift may be adjusted from +500 to +2000 cps (or Δf) for a positive 20-volt signal from the facsimile equipment. During either frequency-shift keying or facsimile operation, the output of unit 12 is fed to the frequency converter (unit 5) to replace the 100-kc signal normally supplied by the crystal oscillator (unit 1). Another portion of the frequency-shift oscillator output is fed to an oscilloscope test receptacle. The frequency converter (unit 5) consists of a series of conventional mixers, tuned filters, and an amplifier to heterodyne its four input frequencies. The output of unit 5 is adjustable in steps of 10 cps from 9.19 to 9.2 mc for c-w and phone, or 9.19 to 9.2 mc ± 1000 cps (maximum) for frequency-shift telegraphy, and 9.19 to 9.2 mc +2000 cps (maximum) for facsimile operation.

UNIT 6.—The output of unit 5 is fed to the 10-kc step generator (unit 6) whose major component is a phase-locked oscillator. The 10-kc step generator receives, in addition, a 100-kc signal from the crystal oscillator and a 1-mc signal from unit 2. The 100-kc input is reduced to a 10-kc signal which is used to phase lock the 210- to 300-kc oscillator with the output frequency of unit 1. Both of the above signals are applied to oscilloscope test receptacles. A portion of the oscillator frequency is then mixed with the input from unit 2 and unit 5 to produce a final output from unit 6 of 10.4 to 10.5 mc (in steps of 10 cps), which is fed to the 100-kc step generator (unit 8).

UNIT 7.—The frequency multiplier (unit 7) receives a 1-mc signal from unit 2 and multiplies it to provide first a 5-mc signal for the frequency converter (unit 9) and, secondly, a 15-mc signal for the 100-kc step generator (unit 8). Unit 8 uses the 100-kc input from the crystal oscillator (unit 1), to produce an output of 1.6 to 2.5 mc, adjustable in 100-kc steps.

A portion of this output may be fed to an oscilloscope. The remaining portion is mixed with the inputs from unit 6 and unit 7 to produce an output of 27 to 28 mc, which is adjustable in steps of 10 cps. This output is fed to the frequency converter (unit 9).

UNIT 9.—Unit 9 mixes the 27- to 28-mc input from unit 8 with the 5-mc input from unit 7 to produce frequencies of 27 to 28 mc ($+0$, $+5$, or $+10$ mc), adjustable in 10-cps steps. These frequencies are amplified by remainder of unit 9 and then fed to one wafer of the bandswitch. The 1-mc step generator (unit 10) receives a 1-mc signal from unit 2 and passes it through a harmonic generator that produces frequencies of 7, 8, 13, 14, 16, 17, 18, 19, 20, 21, and 22 mc. These frequencies are amplified by the remainder of unit 10 and fed to the second wafer of the bandswitch.

UNIT 11.—The outputs of units 9 and 10 are fed to one of the three final converters (units 11A, 11B, and 11C), only one of which is used at a time. The output frequency desired determines which of the three converters will be used. The chosen unit is connected to the outputs of unit 9 and unit 10, and to the radio-frequency oscillator mounting (unit 14) by the bandswitch. Unit 11A is used for frequencies between 0.3 and 6 mc, unit 11B for the 6- to 16-mc range, and unit 11C for outputs of 16 to 26 mc.

UNIT 14.—The mounting (unit 14) serves as a rack into which the other units of the radio-frequency oscillator are connected. Unit 14, in turn, plugs into the transmitter group cabinet wiring. The control indicator, which mounts on the unit 14 front panel contains controls for antenna tuning and an indicator for measuring the standing wave ratio existing on the r-f transmission line.

Radio-Frequency Amplifier

The radio-frequency amplifier accepts from the radio-frequency oscillator a signal in the frequency range from 0.3 to 26 mc. The radio-frequency amplifier consists of three stages: the buffer, the intermediate power amplifier (IPA), and the power amplifier (PA). The signals from the radio-frequency oscillator are received at a level of approximately 0.1 watt, and amplified to either 100 or 500 watts. During 100-watt operation, the radio-frequency

plifier amplifies all frequencies in the 0.3-
26-mc range. In 500-watt operation, only
gnals within the range from 2 to 26 mc are
nplified in the radio-frequency amplifier.
lection of a frequency between 0.3 and 2 mc
uring 500-watt operation automatically switches
e output to the 100-watt level.

The RFA output may be either modulated
audio from the low-level radio modulator
high-level radio modulator, or it may be
eyed on and off by keying voltages from the
w-level radio modulator. During frequency-
uift keying and facsimile operation, the r-f
gnal from the RFA is frequency modulated
the radio-frequency oscillator.

The tuning of the RFA is accomplished
anually by three controls: the bandswitch,
e tune IPA Control, and the tune PA control.
o enable the RFA to tune through the entire
inge from 0.3 to 26 mc, the total range of
equencies has been grouped into six bands.

The function of the buffer stage is to isolate
e RFO from the effects of a varying load
aused by keying signals received by the RFA
om the low-level radio modulator (LLRM).

ow-Level Radio Modulator (LLRM)

The LLRM (fig. 5-4) accepts voice, tele-
raphy (hand or machine keyed), or facsimile
ignals from the external circuits and amplifies
nd shapes them as required to modulate the
arrier. A serivce selector switch is provided
or manual selection of the mode of operation.
he unit contains audio amplifying and modula-
ng circuits for the modulation of the carrier
t the 100-watt level. During 500-watt operation,
he low-level radio modulator feeds a high-
evel radio modulator, which boosts the audio
ignal to the required level. Either a carbon
r dynamic microphone may be used to apply
oice signals to the input.

The term "low-level," as used in the unit
itle, implies that this particular modulator is
sed as the modulating source for the RFA
uring the 100-watt operation. Among the cir-
uits contained in the low-level radio modulator
re an audio amplifier chain, a squelch circuit,
he automatic gain circuit, a keying multi-
ibrator, and circuits that control the on-off
ondition of the transmitter.

AUDIO INPUT.—The audio input signal to
he LLRM is received locally through a handset
r microphone jack. This signal is coupled to
he local-remote switch, S1, in the local position.

If the input is from a carbon microphone,
-12 volts will be supplied. With dynamic micro-
phone input, the -12 volts is disconnected.
When the audio input is applied, it is impressed
across a transformer (through contacts of the
local-remote switch, S1) to a voltage divider.

To compensate for the lower signal ampli-
tude from the dynamic microphone, the voltage
divider output is taken from a higher voltage
tap by means of a dynamic-carbon switch
(not shown).

The audio signal is fed to a two-stage
audio preamplifier consisting of V1 and V2A.
The output of V2A feeds a cathode follower,
V2B, whose function is to match the impedance
between the preamplifier stage and the clipper
stage, which follows. The cathode follower
output is also fed to the automatic gain control
(AGC) circuit. The latter circuit provides a
negative voltage (under strong signal conditions)
to the V1 grid, causing this stage to have a
constant output level over a large range of
input signal amplitudes.

By applying the audio signal input to the
suppressor grid of V1 and the AGC voltage to
the control grid, less AGC voltage is required
to produce a highly sensitive AGC circuit. This
represents an improvement over the conven-
tional system of AGC in which both signals are
applied to the control grid. In the conventional
method, more AGC voltage is required to pro-
duce the desired effect.

AUTOMATIC GAIN CONTROL (AGC).—The
cathode of the AGC rectifier, V6A, is approxi-
mately +20 volts (no signal) and the plate
voltage is +12 volts. Therefore, when a positive
peak of the audio signal exceeds 8 volts (with
respect to the no-signal voltage) the AGC
rectifier V6A, will conduct, and any additional
peak signal voltage will be developed across a
capacitor.

The resulting signal, which approaches a
d-c voltage is fed through the AGC switch in
the ON position to the control grid of V1. The
polarity and magnitude of this voltage increases
the bias on V1, thereby decreasing its gain.
The higher the signal appearing across the out-
put of the cathode follower, the larger will be
the negative bias built up. The gain of V1 is con-
trolled by the AGC rectifier so that a 30-db
range of input will cause less than a 10-db
range of output.

CLIPPER STAGE.—Limiting of the audio
signal amplitude to prevent overmodulation of

Figure 5-4.—Low-level radio modulator, block diagram.

the carrier is accomplished in V3, a twin diode clipper. A potentiometer designated CLIPPER SYMMETRY may be adjusted so that the positive and negative peaks of the audio signal are clipped at the same level. The clipper output is coupled to the audio amplifier, V4A.

AUDIO AMPLIFIER.—V4A receives the output of the clipper, amplifies the signal, and couples this signal to the bandpass filter network, Z1 (not shown). The filter is designed to pass frequencies in the range from 200 to 4500 cycles and assist in cutting out high-frequency harmonics of the audio signal introduced by the clipper action in the previous stage. The circuit also eliminates noise frequencies outside the pass band of the filter.

With the service selector switch, S4, in any position other than phone, the output of V4A will be grounded. The service selection switch, S4, is divided into front and rear sections. The front section is designated F on the switch contacts, and the rear section, R. In phone operation, the keying relay is energized by depressing the press-to-talk button. This action removes the ground on the handset. The signal then passes to the input of V5A.

In addition to the audio signal being coupled to the grid of V5A, there is a negative bias applied from the squelch circuit, V7 and V6B, which is sufficient to cut off V5A if the level of the audio is not high enough. The output of V5A is coupled to the phase inverter, V5B, and to the sidetone amplifier.

SQUELCH AND SQUELCH RECTIFIER CIRCUITS.—A squelch circuit is provided to prevent the transmission of noise in the absence of an audio input signal (when the press-to-talk button is depressed). The squelch circuit prevents amplification by biasing the V5A grid beyond cutoff. The squelch circuit consists of a one-shot multivibrator, V7, a crystal rectifier, CR1, and a limiter, V6B. The audio signal coupled to the grid of V7A must be large enough to overcome the grid-to-cathode bias to effect the operation of the squelch multivibrator. The multivibrator output pulses are coupled to the squelch rectifier, CR1. The output of CR1 is filtered and applied to the V5A grid. This action raises the grid potential of V5A above cutoff, allowing it to amplify the audio signals being received from V4A.

The time constant of the filter circuit is sufficiently long to allow the bias voltage to be essentially constant and at a level needed to keep the tube operating in the linear portion of its characteristic curve.

The diode limiter, V6B, acts to prevent the application of a positive potential to the V5A grid. If the positive pulses from the multivibrator are of high enough amplitude to make the resultant voltage applied to V5A grid positive with respect to ground, the plate of V6B becomes more positive than its cathode. The diode therefore conducts heavily, limiting the voltage applied to the grid of V5B to a maximum of zero volts with respect to ground. Peak positive voltages above the ground level are developed across resistor R29, which is in the plate current path of V6B.

THE SIDETONE AMPLIFIER.—The output of the audio amplifier, V5A, is monitored by the sidetone amplifier, V4B.

The output of the sidetone amplifier, V4B, is coupled by an output transformer to three monitoring points: the handset receptacle, the sidetone jack, and the remote radiophone unit.

A 1000-cps signal from the audio oscillator, V16, is applied to the grid of V4B in hand-key or machine-key operation during the key down (mark) condition for monitoring purposes.

AUDIO OUTPUT CIRCUIT.—The output signals of V5A are fed to the phase inverter, V5B, which produces two signals of equal amplitude, but of opposite phase. These two signals are coupled to a push-pull cathode follower, V8, which acts as a driver for the push-pull power amplifiers, V9 and V10, that constitute the final stage of amplification of the audio amplifier chain. In the AN/SRT-14, the final audio output (50 watts) is fed directly to the radio-frequency amplifier. It should be recalled that the AN/SRT-15 and 16 provide two levels of audio output, depending on the power level at which the transmitter is operating. For 500-watt operation, the output (6 watts) of the LLRM is fed to the HLRM.

PHASE INVERTER.—The circuit components of the phase inverter, V5B, are such that two waveforms, equal in amplitude but opposite in phase, will be developed at the output. The waveforms are taken from the cathode and plate, respectively.

PUSH-PULL CATHODE FOLLOWERS.—V8A and V8B (push-pull cathode followers) act as

drivers for the push-pull power amplifiers, V9 and V10. Since the power amplifier stage is operated class AB2, the cathode followers V8A and V8B are required to supply energy to a low input impedance during the portion of the cycle when grid current flows in V9 and V10.

The audio amplifier chain is not required to amplify audio signals during hand-key and machine-key operation. Therefore, the cathode followers and power amplifiers perform a different function during these two modes of operation. The fact that the power amplifiers use +500 volts for a plate supply is utilized to provide a dumping (loading) circuit to eliminate transient voltage variations imposed on the power source when the transmitter is keyed. The dumping control is fed to the grids of the cathode followers, V8A and V8B, when in either the HAND or MACHINE position.

The +500-volt power supply provides plate voltage for the IPA and the screen voltage (during 500-watt operation) to the PA in the radio-frequency amplifier. In space condition the RFA is keyed off, and IPA plate current and PA screen current are at minimum, while in mark, they are at maximum. If these sudden changes in load current drawn from the 500-volt supply are not modified, transient voltages will appear across the power supply filter chokes.

At the start of a mark condition, the polarity of the transient voltage in the power supply filter will be such as to oppose the sudden rise in current, thus causing a dip to appear in the power supply output voltage. Conversely, at the start of a space condition, the transient voltage will be aiding, opposing the sudden decrease in current, thus increasing the supply output voltage.

The function of the dumping circuit is to provide a load during space condition only, to compensate for the reduction in load current when the IPA is keyed off. Thus, the load to the 500-volt supply will be approximately equal during mark and space due to this elimination of transient variations. The method employed to keep the power supply current from decreasing during a space is to supply a keying voltage from the keying circuit of -65 volts during mark and zero volts during space to the grids of the push-pull cathode followers. This voltage is obtained from the cathode of the dumping cathode follower, V11B. The input to the cathode followers is reproduced at the cathodes, which are directly tied to the grids of the power amplifiers.

POWER AMPLIFIERS.—The no-signal grid bias of the power amplifiers, V9 and V10, applied by way of the V8 cathode followers is approximately -40 volts. With a screen voltage of +30 volts and a plate supply of +500 volts, the value of bias is such that the stage operates as a class AB2 push-pull amplifier. Operation at this level allows high-power outputs with high plate efficiency and low distortion.

The modulation transformer is provided with two secondaries. The 50-watt secondary is used only during 100-watt operation of the transmitter, and the 6-watt secondary is used during 500-watt operation.

The AN/SRT-14 operates at the 100-watt level only and is not controlled by the 500-watt disable line because of the absence of the high-voltage power supply. With the AN/SRT-15 and 16 in 500-watt operation the circuit of the 6-watt secondary is completed through the high voltage supply.

KEYING CIRCUIT.—In hand-key operation the keying circuit accepts keying signals that determine the on-off condition of the carrier. In machine-key operation, keying signals are sent to the keying circuit from a teletypewriter or other keying equipment which also controls the on-off condition of the carrier. In both hand-key and machine-key operation an audio oscillator, V16, produces an audio signal for aural monitoring of the keying signal. In FSK operation, keying signals are received by the keying circuit from machine-key equipment, and signals are produced that control the operation of the frequency-shift oscillater in the RFO. Signals are received from an audio oscillator, which phase-modulate the carrier to minimize the effects of selective fading. In phone operation, audio signals are fed to the audio circuits when the press-to-talk button on the microphone is depressed, which also keys the carrier on and off.

The keying circuit of the LLRM consists of the keying and press-to-talk relay buffer, V11A; keying and control, V12; keying multivibrator, V13; dumping cathode follower, V11B; limiter, V14; keyer, V15; and audio oscillator, V16. The outputs of the keying circuit and their functions are listed in table 5-1.

Power Supply in LLRM(+250 V)

The power supply contained in the LLRM provides a regulated +250 volts, which is used

Table 5-1.—Functions of Keying Circuits
Outputs

Output	Delivered to	Function
F keying signal	Buffer in RFA and output of RFO	To key the carrier
SK keying signal	FS oscillator in RFO	Control frequency of FSO
Keying signals to circuit	LLRM	Load compensation for +500-volt power supply
00-cycle signal	FS oscillator in RFO	Phase-modulate the FSO
000-cycle signal	Sidetone amplifier in LLRM	Monitoring of keying signals

in the criticial keying and audio circuits of the LLRM and RFO. The power supply consists of rectifier V22, regulator control tube V19, and and voltage regulator V20. The a-c power is supplied from the low-voltage power supply through the main power switch. The circuits contained in this power supply are generally the same as those considered in Basic Electronics, NavPers 10087 (revised).

-12 VOLT POWER SUPPLY.—A metallic bridge rectifier, CR2, is used in the -12-volt power supply and is connected so as to yield a negative output. The rectifier receives a 16.5-volt rms signal from an input transformer. The -12-volt output is filtered and supplied to the dynamic-carbon switch for use as operating voltage for the carbon microphone. The output is also fed to the remote radiophone unit where it serves as the operating voltage for a carrier-on idicator, the control relay, and the carbon microphone.
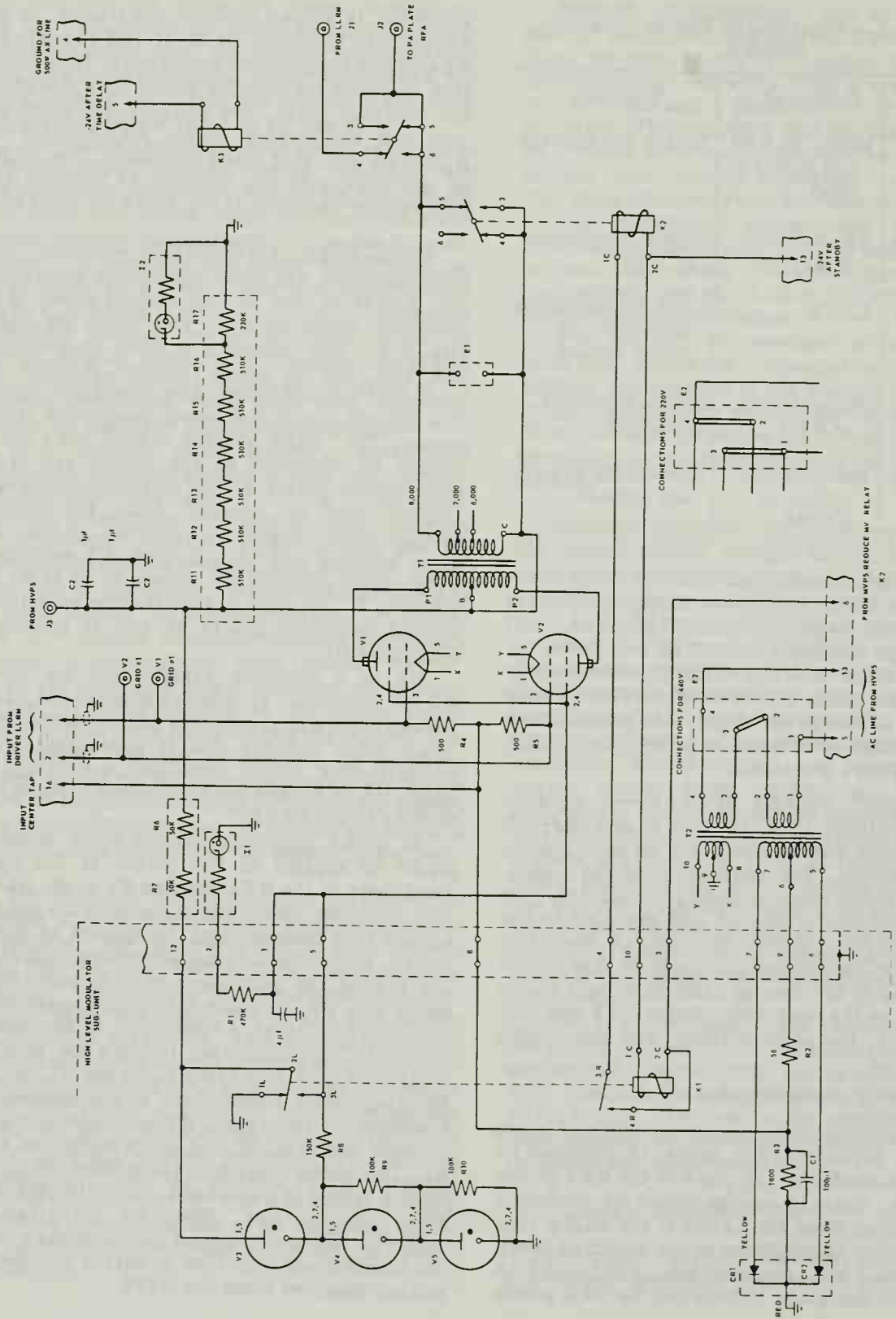
High-Level Radio Modulator

During phone operation at the 500-watt level, an audio signal of 250 watts is required to amplitude-modulate the plate of the RFA power amplifier. This is accomplished by feeding a 6-watt signal from the LLRM to the HLRM (fig. 5-5) where it is amplified to the required level. During hand-key, machine-key, FSK, and FAX operation, the plate voltage for the RFA power

amplifier is routed through the HLRM. In the AN/SRT-16 the output of the HLRM is fed to the RFA in one transmitter group only. This means that only one transmitter group may operate at the 500-watt level. The AN/SRT-15, which has only one transmitter group, may employ the HLRM in 500-watt phone operation. The HLRM is not used with the AN/SRT-14 because this transmitter operates at the 100-watt level only.

PUSH-PULL AMPLIFIERS.—During 500-watt operation, the audio modulating signal from the LLRM is fed to the control grids of the push-pull amplifiers, V1 and V2 in the HLRM (fig. 5-5). The signals are equal in amplitude but 180° out of phase, making possible the push-pull amplification. The amplified signals are applied across the output modulation transformer, T1. Plate voltage (+2400 volts during phone) is delivered to the center tap of the transformer primary. The GROUND FOR 500-w PHONE line is supplied from the medium voltage power supply to relay K1. If the -24v after standby line is activated, K1 energizes. This action supplies the screen grids of V1 and V2 with a +360-volt potential. The potential is applied through contacts 2L and 3L of the now energized K1.

R1 is a dropping resistor for the +350-v screen indicator, I1, which indicates the HLRM screen supply is present. In phone operation the -24-v after standby line is energized by the press-to-talk button such that it controls the operation of K2 and the application of the screen voltage to the HLRM.

In phone operation, +2400 volts from the HVPS is applied to the plates of the power amplifiers in the RFA through T1 in the HLRM. The 250-watt audio signal, which also appears across this transformer, is applied to the PA plates. Protection against excessively high voltage across the secondary of T1 is provided by spark gap E1. In other than phone operation at the 500-watt level, +3000 volts is applied directly to the PA plates through the shorting contacts 4 and 5 of relay K2 in the HLRM. The shorting action takes place in the deenergized condition of the relay. K2 becomes energized through contacts 3R and 4R of energized relay K1. The latter relay energizes when the press-to-talk button is depressed, activating the -24v-after standby line. When K2 energizes and removes the short across the secondary of T1, the transformer is placed in series with the high voltage supplied from the HVPS.

Figure 5-5.—High-level radio modulator, simplified schematic.

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When the transmitter is operated at the 100-watt level, the GROUND FOR 500-WAX line is open and K3 is not energized. Therefore, the LLRM connected to J1 passes its output through contacts 4 and 5 of K3 and to J2, which connects to the power amplifier plates in the RFA. In 500-watt operation, the GROUND FOR 500W AX line is completed, and K3 is energized after the time delay. This action applies the output of the HLRM to the power amplifiers through contacts 6 and 3 of energized relay K3. In the AN/SRT-15, the output of the HLRM to the PA plates occurs in one transmitter group only. In the other group, the plate potential and the modulating voltage is received from the LLRM directly. As the output of the HLRM is a permanent interconnection, the transmitter group to which this output is fed in the AN/SRT-16 will be the group that can be placed in 500-watt operation.

In order to generate a bias voltage to operate the audio amplifiers at the desired class AB2 point on its characteristic curve, a -50 volt bias source is included in the HLRM.

Power Supplies

There are three separate power supplies other than the one previously discussed; low-voltage power supply, medium-voltage power supply, and the high-voltage power supply.

LOW-VOLTAGE POWER SUPPLY.—The LVPS receives 110 volt a-c from the ship's supply and distributes this power to other units. It contains rectifier circuits that provide B+ voltage of +300 volts, +250 volts, -24 volts for control circuits and motor lines, and -220 volts for bias.

MEDIUM-VOLTAGE POWER SUPPLY.—The MVPS receives 110 volts from the low-voltage power source, and supplies +500 volts and +1300 volts, respectively. The +500 volts is applied to the low-level radio modulator and the radio-frequency amplifier. During phone operation, the +1300 volts is reduced to +1050 volts and applied to the radio-frequency amplifier. The medium-voltage power supply contains a relay, which ensures that the 500-volt source is activated before the 300-volt source in the low-voltage power supply comes on. This action is necessary since the 300 volt-source is primarily a screen voltage supply and will cause damage to the tubes it serves if plate voltage has not been applied.

HIGH-VOLTAGE POWER SUPPLY.—The HVPS used only in the AN/SRT-15 and 16 during 500-watt operation is provided with a separate input of 220 volt, or 440 volt, three-phase, 60-cycle a-c from the ship's supply. This circuit has a d-c output of +3000 volts that replaces the 1300-volt output of the medium-voltage power supply during 500-watt operation. During phone operation, the 3000 volts is reduced to +2400 volts.

RADIO TRANSMITTING SET AN/WRT-2

Radio Transmitting Set AN/WRT-2, is capable of delivering an average power output of 500 watts and a peak envelope power (PEP) of 1000 watts in 1—kc steps (or continuously) over a frequency range of 2.0 to 30.0 megacycles. The set provides for emission of machine (MACH) or break-in (C-W) hand-keyed telegraphy, frequency-shift-keyed (FSK) teletype, compatible amplitude-modulated speech, single sideband suppressed carrier (SSB), independent sideband (ISB), and facsimile by using appropriate terminal equipment. The AN/WRT-2 may be installed aboard ships, submarines, and at shore stations. The major components of the set (fig. 5-6) are the transmitter group, (radio-frequency amplifier, radio-frequency oscillator, electrical frequency control, amplifier power supply, and power supply) and the radio frequency tuner.

The radio transmitter group contains all of the equipment required for transmission by machine or break-in c-w, independent sideband (ISB), single sideband (SSB), a-m phone, and/or frequency-shift-keying (FSK), except such accessories as the key (machine or hand), antenna, and remote operating components. Facsimile transmission requires additional terminal equipment.

The transmitter may be operated from a three-phase, 60 cps ship's supply of 115 v, 220 v, or 440 volts. The transmitter group is coupled to an antenna (not shown) through a radio-frequency tuner and associated antenna control circuits.

Provision is made for a total of six audio inputs. Front panel handset jacks are furnished for local phone operation of the equipment with a handset. Remote phone- and/or hand-key operation is also possible.

Provision is made for interconnections to teletype and telegraphy equipment and to a
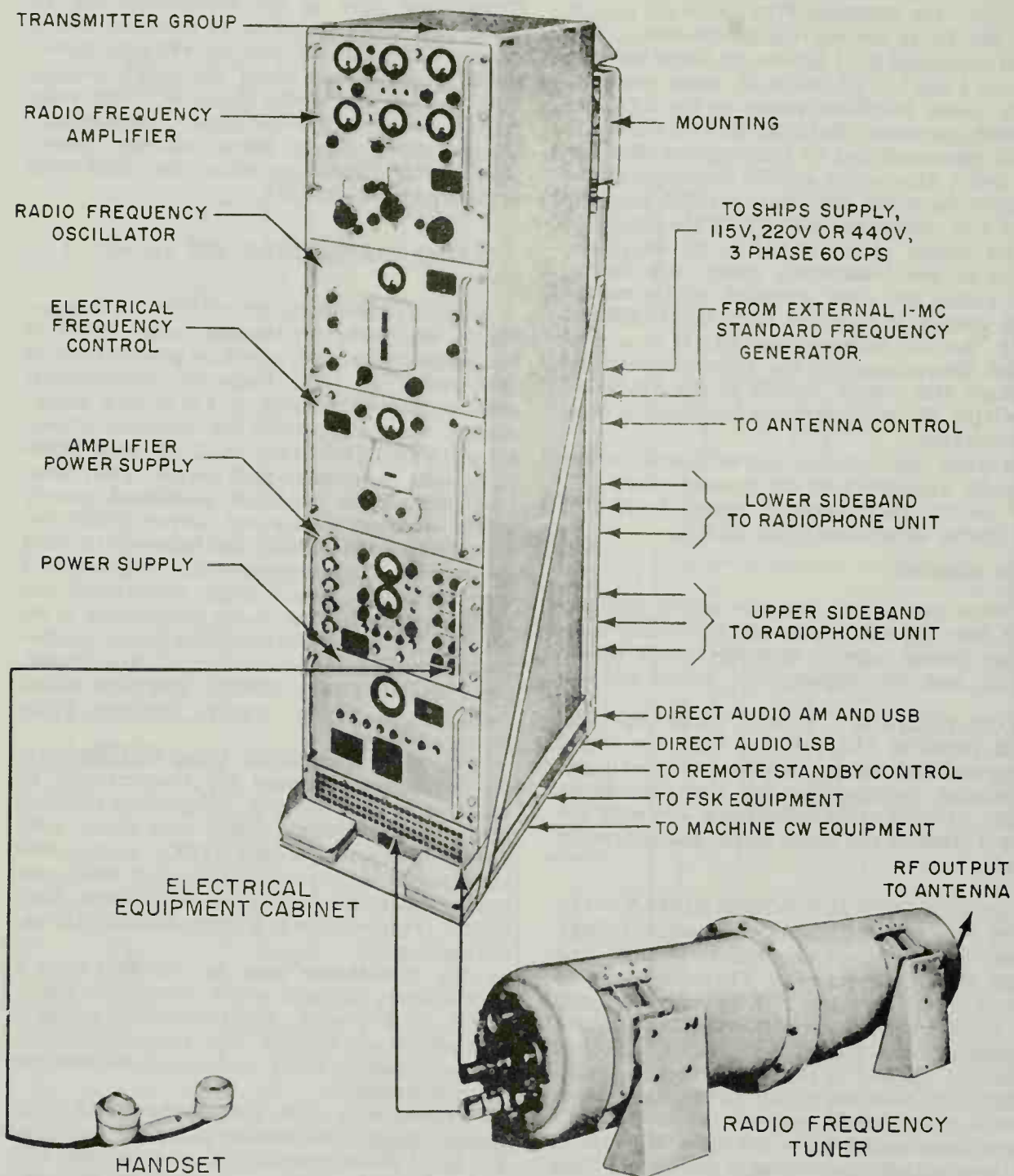
TRANSMITTER GROUP

RADIO FREQUENCY AMPLIFIER

RADIO FREQUENCY OSCILLATOR

ELECTRICAL FREQUENCY CONTROL

AMPLIFIER POWER SUPPLY

POWER SUPPLY

MOUNTING

TO SHIPS SUPPLY, 115V, 220V OR 440V, 3 PHASE 60 CPS

FROM EXTERNAL 1-MC STANDARD FREQUENCY GENERATOR.

TO ANTENNA CONTROL

LOWER SIDEBAND TO RADIOPHONE UNIT

UPPER SIDEBAND TO RADIOPHONE UNIT

DIRECT AUDIO AM AND USB
DIRECT AUDIO LSB
TO REMOTE STANDBY CONTROL
TO FSK EQUIPMENT
TO MACHINE CW EQUIPMENT

RF OUTPUT TO ANTENNA

ELECTRICAL EQUIPMENT CABINET

HANDSET

RADIO FREQUENCY TUNER

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Figure 5-6. Radio Transmitting Set AN/WRT-2, relationship of units.

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...mote transmitter standy control. An in-
...rnal dummy load is provided for transmitter
...neup.

...VERALL BLOCK DIAGRAM

The AN/WRT-2 consists of the following
...nctional sections: r-f generating, power am-
...ifier, modulating, primary power, and low-
...oltage power supply, as shown in figure 5-7.

The r-f generating section produces the r-f
...gnal in conventional master oscillator-
...equency multiplier circuits and applies it to
...e power amplifier. The master oscillator is
...aved to the assigned operating frequency by a
...nique frequency control circuit treated in de-
...il later in this chapter.

The power amplifier section raises the r-f
...gnal input power from the r-f generating sec-
...on to the desired operating power level. The
...utput of the power amplifier is fed to the an-
...enna via the r-f tuner circuits.

The modulating section accepts audio sig-
...als or d-c keying signals (for C-W or FSK
...peration) and suitably transforms these sig-
...als to modulate or control the r-f energy gen-
...rated by the r-f generating section. The out-
...ut of the modulating section is applied to the
...put mixer in the power amplifier section.

The low voltage section contains all the cir-
...uits which supply low d-c voltages to the
...ransmitting set. The high voltage supply is
... separate section and supplies the plate voltage
...or the power amplifier tubes.

The primary power section supplies all a-c
...oltages for the transmitting set. The trans-
...ormers in the primary power section have taps
... provide input voltages of 115 v, 220 v, and
...40 v at 60 cps.

...-F Generating Section

The complete frequency coverage of Trans-
...mitting Set AN/WRT-2 is accomplished in 12
...ands. The master oscillator, V301 (fig. 5-8),
...hich is located in the radio-frequency oscilla-
...or chassis, generates r-f signals from 1.5 mc
... 7.5 mc. These signals are amplified by V302.
...or bands 1 through 6, the fundamental fre-
...quencies (1.5 to 7.5 mc) of the master oscil-
...ator are fed directly to the power amplifier
...section through S302. For bands 7 through 9,
...he fundamental frequencies of bands 4 through
... (3.75 to 6.75 mc) are doubled by the first
...doubler, V303, and fed through S302 to the
...ower amplifier section. For bands 10 through
...2, the fundamental frequencies of bands 4
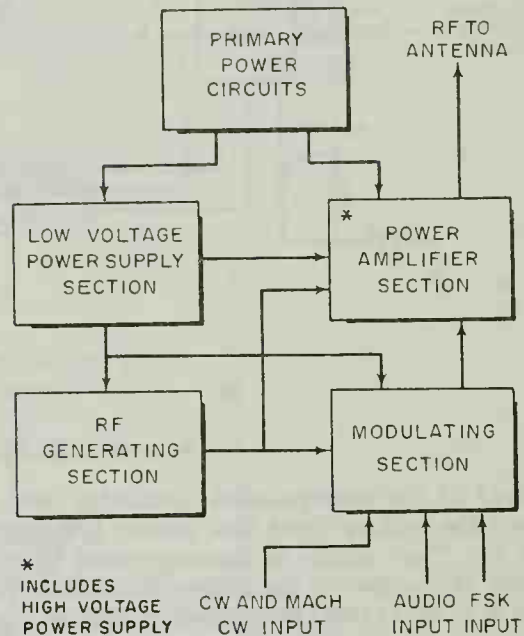
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Figure 5-7. Overall block diagram

through 6 are multiplied four times by the ac-
tion of the first and second doublers, V303 and
V304, before being applied to the power ampli-
fier section.

In addition to supplying the power amplifier
section input, the S302 output is also fed to the
electrical frequency control circuit (fig. 5-9).
This circuit supplies a frequency-correction
voltage to the master oscillator, V301, which
keeps the oscillator operating at the selected
frequency with the stability of 1 part in 10^8 cps.

The control of the master oscillator fre-
quency is accomplished by two independent op-
erations. First, the output of the master oscil-
lator (via amplifier V302 and the frequency
doublers) is compared with that of an interpola-
tion oscillator in a master oscillator phase
detector. Any difference between these two
frequencies causes a correction voltage to be
applied to the master oscillator (MO), and
thereby forces the oscillator to return to the
proper operating frequency.

Second, when operating in 1 kc steps, the
interpolation oscillator is frequency stabilized
in order to provide improved accuracy in cor-
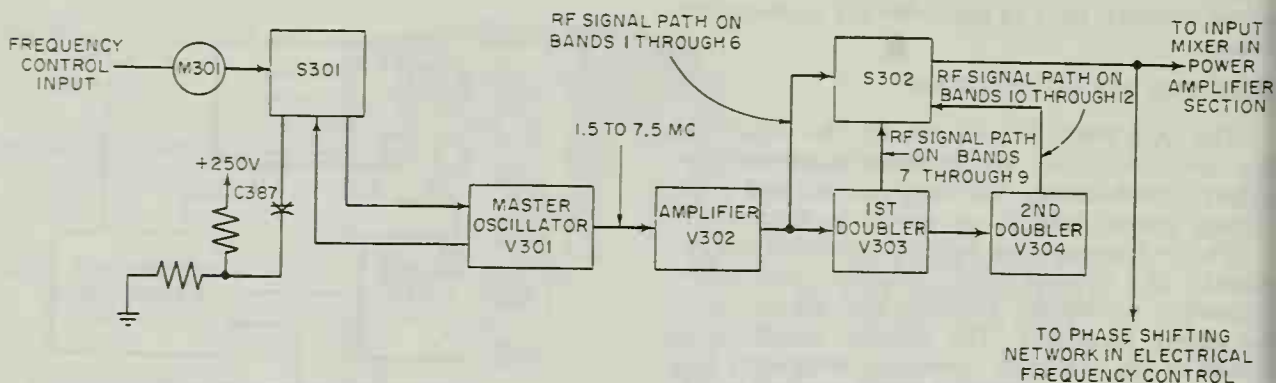recting the master oscillator frequency. The

Figure 5-8—Radio-frequency oscillator.

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control of the interpolation oscillator (IO) involves the locking-in of this circuit to multiples of 1 kc. This action is accomplished by comparing the output of the interpolation oscillator with a 1-kc crystal stabilized reference signal to develop an interpolation oscillator correction voltage. The correction voltage is applied to the interpolation oscillator and holds it at the required frequency. Thus, the controlled stability of the IO results in an added stability of the MO.

The front panel of the radio-frequency oscillator and electrical frequency control chassis showing the indicators and controls is shown in figure 5-10. A more detailed block diagram of the electrical frequency control circuit is shown in figure 5-11.

The r-f input from the radio-frequency oscillator circuit (fig. 5-8) is applied to the frequency control circuits through a phase-shift network consisting of FL603 (fig. 5-11). Sealed unit FL603 contains two independent-phase-shift networks. The networks are so designed that the signal coupled to the top balanced mixer always leads the signal coupled to the bottom mixer by 90°. The outputs of the phase-shift network (FL603) are therefore sine and cosine functions of the r-f signal.

A second input to the mixers is received from the 1-mc oscillator and divider circuits from T608.

Crystal Oscillator (1 mc)

The 1-mc crystal oscillator (Q1301 and Q1302) is used as a frequency standard to provide maximum frequency stability of 1 part

in 10^8. The crystal operates in a precisely controlled constant-temperature oven (not shown). The 1-mc output is amplified by 1-mc amplifiers Q1303 and Q1304 and applied to a 10:1 frequency divider circuit consisting of transistors Q1305 through Q1308, and Q1313 through Q1316. The first stage of the divider circuits divides the 1-mc frequency in half. This 500-kc signal is applied to the 500-kc selector in the modulator section, the 1-mc phase detector, and to the rest of the 10:1 divider circuits. The final output of the frequency divider is a 100-kc signal.

The 100-kc signal is amplified by 100-kc amplifier Q1317 and buffer amplifier V605 and applied to T608. The signal appearing at terminal T3, of T608 is coupled to saturable core reactor L614 where a harmonic-rich peaked waveform is produced. The 100-kc signal and its harmonics are applied to the balanced mixer stages where, due to the simultaneous application of the master oscillator signal, sum and difference frequency components of the two applied signals are produced. The sum and difference frequencies (upper and lower sidebands) contain the r-f oscillator error frequency component.

Sideband Generating Method

The phase-shift method of generating sideband (sum and difference) frequencies is employed. The r-f input signal is shifted 90° in phase by FL603 and applied to the balanced mixers as previously discussed. The mixers are connected in a push-pull circuit arrangement. The combined action of the mixers and
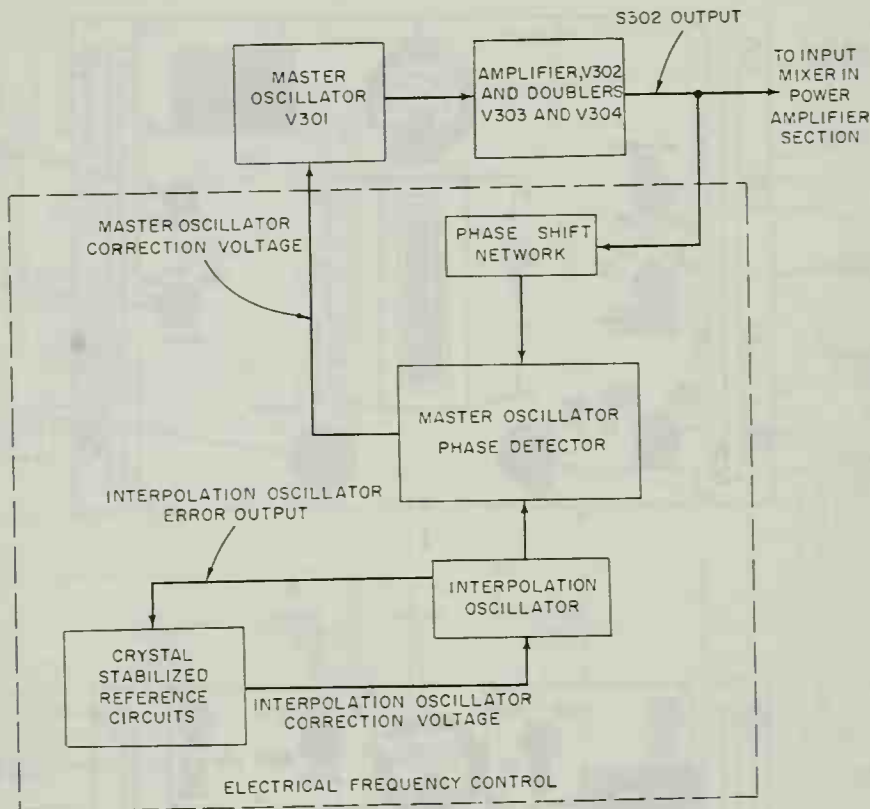
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Figure 5-9.—Electrical frequency control circuit, simplified block diagram.

phase shift networks FL1 and FL2 (at the mixer outputs) assure that the desired sideband frequencies are added and the other sideband frequencies are canceled in the output. This action produces only one of the sidebands at the grid of i-f amplifier V601A, and the need for narrow band filters is eliminated.

For example, a master oscillator (MO) frequency of 2460 kc is heterodyned with 2400- and 2500-kc harmonics of the L614 output to produce 2460 - 2400 or 60-kc, and 2500 -2460 or 40-kc sidebands in each of two mixers. The 40-kc sidebands are shifted 180°, and cancel at the output of the FL1 and FL2. The 60-kc sidebands are brought into phase (0°), and the amplitudes add at the V601A grid.

When the MO frequency is 2440 kc there are 2500 -2440 or 60-kc, and 2440 - 2400 or 40-kc sidebands produced. In this example, the 60-kc and 40-kc phase relations in the two mixers are reversed from their original relation. For the two 60-kc sidebands the reversal is due to the fact that in the first example the 2460-kc MO frequency is above the 2400-kc harmonic and in the second example the 2440-kc MO frequency is below the 2500-kc harmonic. Thus the 40-kc sidebands are shifted to cancel and the 60-kc sidebands are shifted to add.

The acceptable sideband signal is then passed through five cascaded amplifiers, V601A, V601B, V602A, V602B, and V603A. A band-pass filter (not shown) between the first and second i-f amplifiers limits the response of the i-f amplifier strip to the frequency range of 50 to 100 kc. The plate output of the fifth i-f amplifier, V603A, is applied to the master oscillator-phase detector (MOPD). The cathode output of V603A is applied to the 100-kc phase detector for purposes described later.

The selected sideband signal is heterodyned in the master oscillator-phase detector with a signal input from the interpolation oscillator (IO) to produce the master oscillator error voltage. For comparison with the interpolation oscillator output, the range of which is 50 to 100 kc in 5 bands, the sideband frequencies
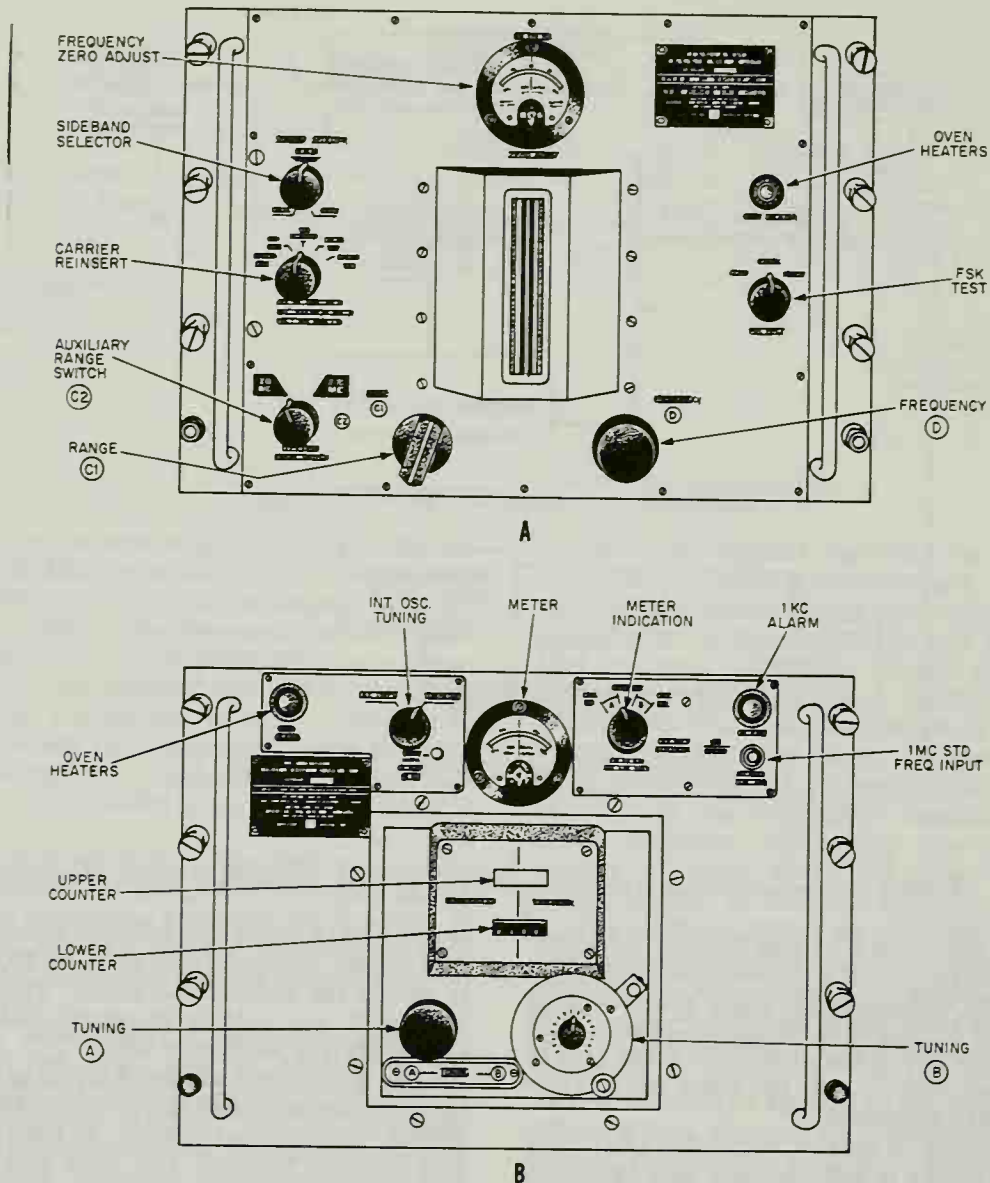
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Figure 5-10.—Pictorial diagrams AN/WRT-2. A. Radio frequency oscillator.
B. Electrical frequency control (front panels).

generated at the balanced mixers must be limited to the range of 50 to 100 kc. Therefore, when the last 10-kc digits of the sideband frequency are in the range of 0 to 50 kc, the output of one of the balanced mixers is reversed by mechanical means and the sideband phase relationships are maintained as in the first example. A relay (not shown) which is energized by the IO tuning A control, located on the electrical frequency control (fig. 5-10, B), controls the output switching when the last 10-kc digits of the sideband frequency are less than 50 kc.

Master Oscillator-Phase Detector

The master oscillator-phase detector (MOPD) provides a d-c correction voltage

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Figure 5-11.—Electrical frequency control, detailed block diagram.

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stabilizing signal for the master oscillator (MO) when the oscillator is not operating at the proper frequency. The polarity depends upon the direction of the error and the amplitude upon the amount of error. The MOPD (fig. 5-12) consists of crystal diodes CR606 through CR609, current limiting resistors R653 through R656, and transformers T604 and T605. The i-f signal from the fifth i-f amplifier V603A is applied (via T604) to the junction of crystal diodes CR606-CR607, and CR608-CR609. The signal from the interpolation oscillator (IO) is applied to the T605 secondary which is connected between points B and D of the circuit. The MOPD output is a d-c correction voltage developed across C652.

When the MO is locked on frequency, the i-f signal is 90° out of phase with the IO signal (i-f signal leading) and no correction voltage is fed back to the MO circuit.

Let us assume that the amplitudes of both the i-f and IO signals (fig. 5-12) are 2.0 volts peak-to-peak. The polarities shown on the bridge (fig. 5-12) are for the first quarter-cycle of the waveforms. The waveforms represent the voltage variations to ground at the four corners of the bridge.

The IO signal is used as a reference frequency. Figure 5-13 illustrates the relative phase of the IO and i-f signals for zero error (i-f leading the IO by 90°), lagging error (i-f leading the IO by 45°) and leading error (i-f leading the IO by 135°). Also shown is the conducting period for each diode (CR606 through CR609) for one complete cycle.

Phase error is generally limited to a few degrees or even a fraction of a degree. In most cases, large errors exist only when the equipment is first turned on or when the circuit is not properly aligned. A large error (45°) is assumed here because it lends itself better for a simple circuit analysis.

When the i-f leads the IO by exactly 90° (the zero error condition), the phase detector will not produce a d-c correction voltage output across C652. Instead, an a-c voltage is produced. The manner by which this voltage is produced is indicated in figure 5-14, A.

At t = 0°. the i-f voltage is positive maximum and the IO voltage is zero. Diodes CR607 and CR609 are conducting as shown by the arrows representing electron flow. Tracing around path FCDGF, the voltage from F to C is equal and opposed to the drop from C to D, so that there is no charge accumulated on C652 (C652 is relatively large).

Likewise, at t = 45° (fig. 5-14, A) the arrows and polarities indicate the applied voltage and conducting diodes. The series aiding voltages in the lower half of the T604 secondary and the right half of the T605 secondary cause a current which charges C652 negative to ground.

At 90° (fig. 5-13) the i-f voltage is zero and the IO voltage is positive maximum. Diodes CR609 and CR608 are conducting; and again there is zero voltage across C652. Note when either the i-f voltage or the IO voltage is zero, the output voltage is zero. At 135° (fig. 5-14, A), the series aiding voltages in the lower half of the T604 and left half of the T605 secondaries cause a current flow which charges C652 positive to ground. A similar analysis at 180°, 225°, 270°, and 315° will further indicate that the C652 output voltage for zero error is alternately negative and positive to ground, and no d-c error voltage is fed to the RFO when the IO and i-f signals are exactly 90° out of phase.

Now consider the phase detector action to produce a d-c correction voltage when the i-f signal leads the IO signal by only 45° (lagging error, fig. 5-14B). At zero degrees no voltage is developed across C652. At 67.5°, the series aiding voltages in the lower half of the T604 secondary and the right half of the T605 secondary cause a current which charges C652 negative to ground. Likewise, at 90°, C652 is still charged negative to ground. However, at 135°, no charge is accumulated on C652.

At 135° (fig. 5-13) diodes CR608 and CR609 are conducting and the output voltage on C652 is positive to ground (135° to 180°, fig. 5-14, B). However, the output voltage is predominantly negative (0° to 135°) for a lagging phase error (i-f leading the IO by only 45°). Thus, the voltage fed to the RFO for this error condition will be a negative d-c voltage with an a-c component (pulsating direct current).

A leading phase error between the IO and i-f signals exists when the i-f leads the IO by more than 90°. This condition will produce an a-c voltage across C652 which is predominantly positive to ground. Figure 5-14, C shows the polarities, conducting diodes, and C652 charge at 112.5° and 292.5 ° of the reference IO signal.

The C652 output (fig. 5-12) is fed through a filter (L621-R692), switch section S302H, and the selected portion of a voltage divider to a voltage-sensitive capacitor (C387). The filter removes the a-c component of the correction voltage.
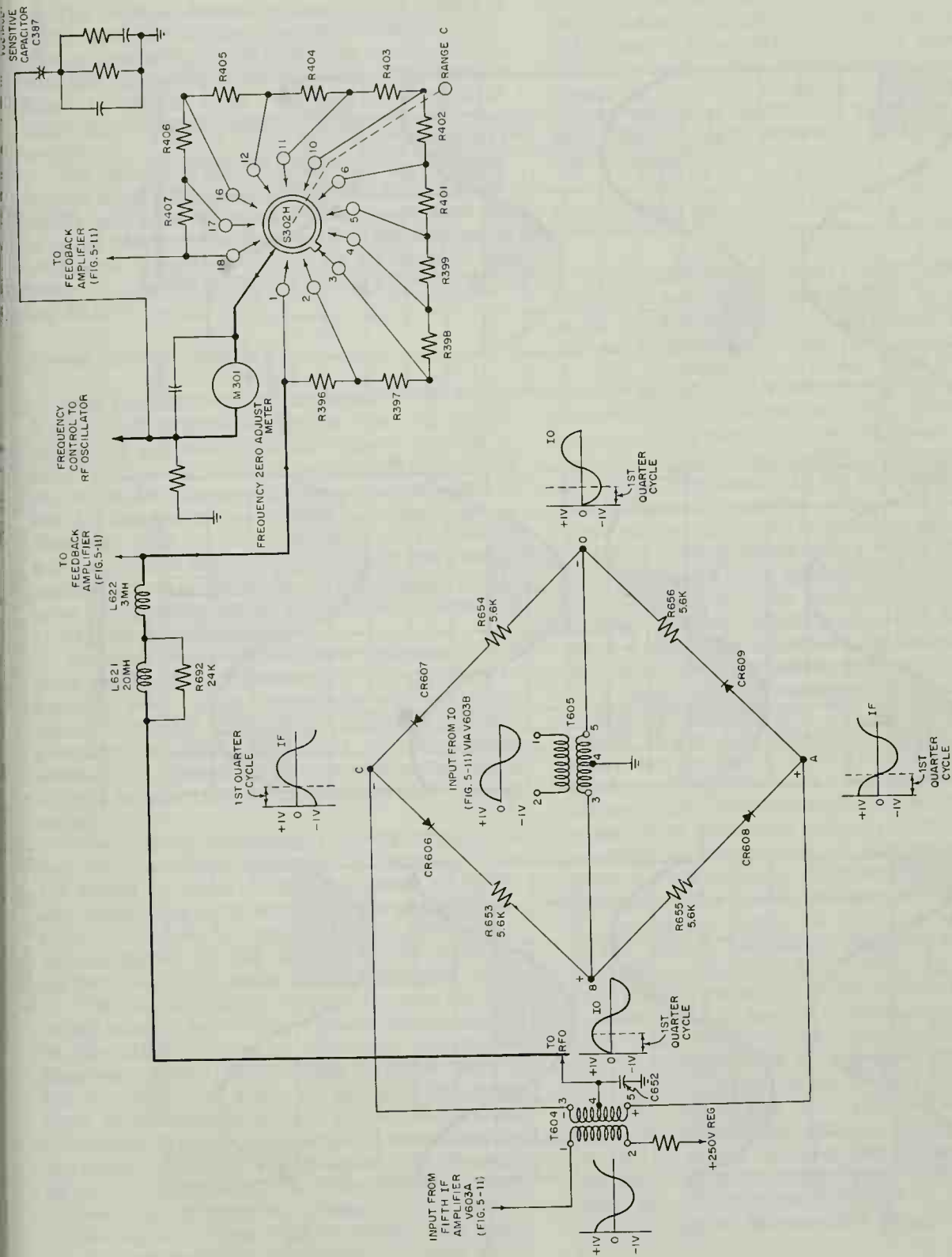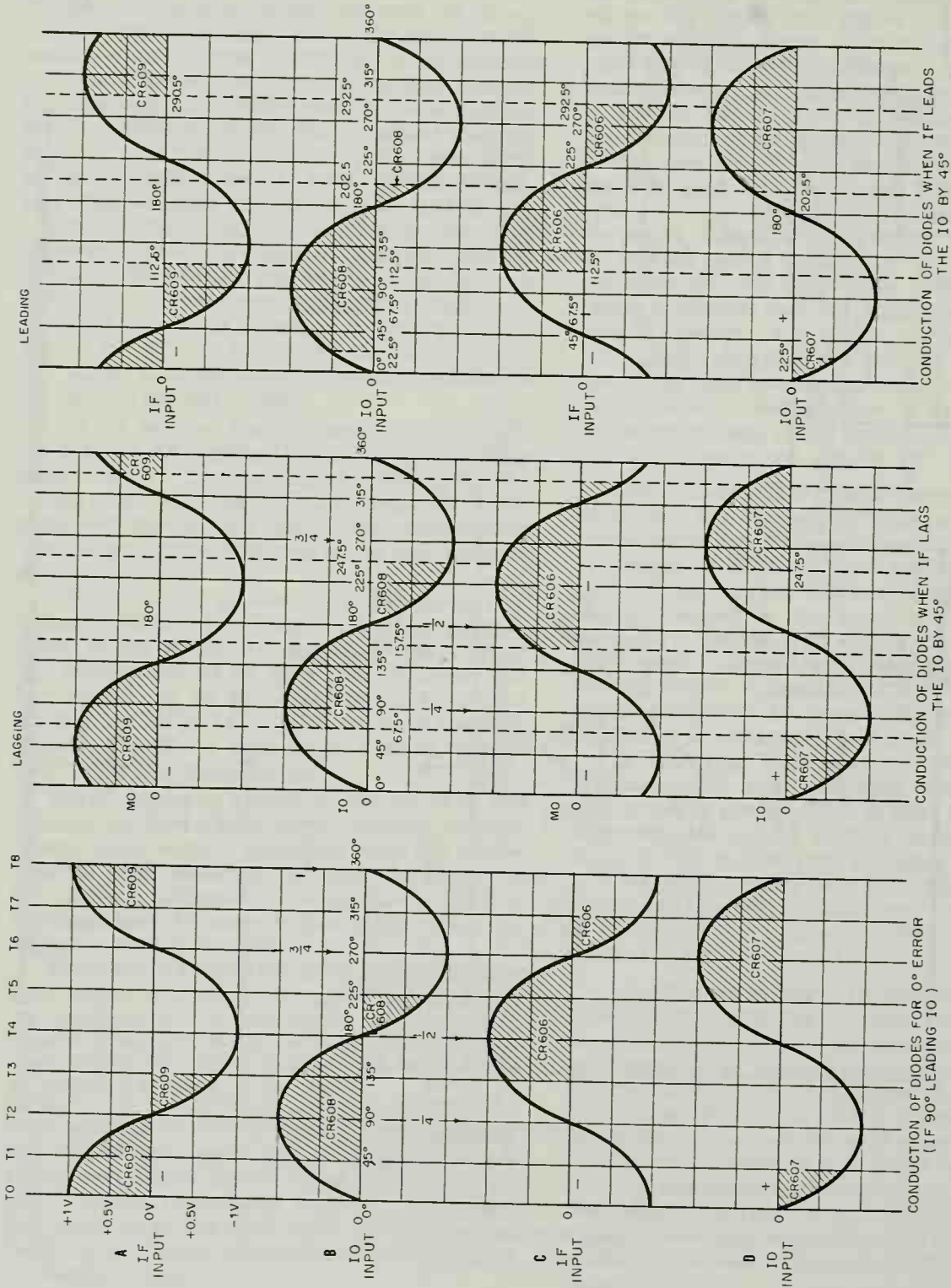
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Figure 5-12.—Master oscillator-phase detector analysis. Simplified diagram with waveforms.

Figure 5-13.—Master oscillator-phase detector analysis. Conduction of diodes.

32.284

The voltage-sensitive capacitor is placed in shunt with the master oscillator tuned circuit. The capacitance of C387 changes in accordance with the magnitude of the d-c correction voltage from the MOPD. The change in the oscillator tank capacitance corrects the oscillator frequency.

The action of S302H and the voltage divider (comprising R396 through R399 and R401 through R407) corrects the amount of error voltage fed back on the selected band so that the amount of oscillator frequency change on all bands is the same for a given d-c error voltage developed at the MOPD.

Phase Detector (100 kc)

The fundamental 100-kc signal at L614 (fig. 5-11) is of greater amplitude than any of the generated harmonics. If the mixers are not in exact balance, it is possible for the 100-kc signal to be fed through the balanced mixers and the i-f stages to the master oscillator-phase detector. This action would cause an erroneous d-c correction signal to be fed from the master oscillator-phase detector to the master oscillator and thereby cause the oscillator to operate at an undesired frequency.

Because the i-f band pass will accept and pass the 100-kc fundamental, it is necessary to keep the mixer stages extremely well balanced. Aging of circuit components and ambient temperature variations inevitably introduce some imbalance; therefore an automatic balancing circuit is provided to null the unwanted 100-kc signal.

The 100-kc balancing network consists of two 100-kc phase detectors. A sample of the i-f signal is taken from the cathode of V603A and applied to both of the 100-kc phase detectors. As shown in figure 5-11, a 100-kc reference signal is also applied to the phase detectors via T608 from buffer amplifier V605.

The 100-kc signal output of one of the balanced mixers is 90° out of phase with the output of the other mixer as discussed previously. Because of this relationship, a phase shift at the detector input shifts the phase of the reference signal applied to one of the 100-kc phase detectors. When no i-f error is introduced, no correction voltage is developed. The presence of an i-f error input from the V603A cathode causes a d-c correction voltage to be developed at the 100-kc phase detector output. The phase of the i-f input from V603A indicates which

mixer is unbalanced. The polarity of the 100-kc phase detector output voltage is proper to provide bias to the mixer which is passing the 100-kc signal. This action causes the 100-kc signal in the two balanced mixers to again cancel and thus nullify any master oscillator error which might otherwise be caused by the 100-kc signal.

Interpolation Oscillator

Interpolation oscillator V606 provides a stabilized reference signal which is compared in the master oscillator-phase detector with the signal delivered through the i-f amplifiers.

The comparison i-f signal from V603A and the IO signal from V606 (via V603B) are fed to the MOPD. As shown earlier, when the two signals are exactly 90° out of phase, no correction voltage is fed back to a voltage-sensitive capacitor (not shown) in the MO circuit and the MO frequency remains unchanged. When any other phase difference is present, due to a shift in the frequency of the MO, a correction signal of proper polarity is applied to the MO to bring this circuit back to the proper frequency. FREQUENCY ZERO ADJUST meter (fig. 5-10, A) in the r-f oscillator indicates the magnitude and polarity of the correction current.

Frequently it is necessary to determine the interpolation frequency. The IO circuit is designed so that only signals in the 50- to 100-kc range are produced, although the frequency applied to the MOPD can be varied over the 2- to 30-mc range. In order to determine the IO frequency it is important to interpret the frequency readings on the front panel counters of the frequency control group correctly. The following information is useful for interpretation of the IO frequency: When the two last place digits of the master oscillator frequency are greater than 50 kc (such as 2355, 2460, 1670, or 1990 kc), the IO frequency is the same as that indicated by the last two digits on the upper counter (fig. 5-10, B). If the MO frequency is 2460 kc, the i-f and also the IO frequencies are 60 kc.

However, when the two last place digits of the MO output are less than 50 kc (such as 2345, 2440, 1630, or 1910 kc) the IO frequency is 100 kc minus the last two digits of the frequency indicated by the lower counter. Thus, when the MO frequency is 2430 kc, the i-f signal and the IO frequencies are 100 - 30 = 70 kc.

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32.284
Figure 5-14.—Master oscillator-phase detector analysis. A. Zero error. B. Lagging error.
C. Leading error.

The d-c correction voltage for the MO contains some harmonics of the comparison i-f signal frequency. These harmonics are removed by a 10-kc low-pass filter inserted between the MOPD and voltage-sensitive capacitor in the master oscillator. However, low frequency spurious signals may still be present in the output of the low-pass filter. In order to remove such spurious signals, feedback amplifiers V607A, V607B, and V608A (fig. 5-11) have been inserted in this circuit. Any spurious signals appearing at the output of the low-pass filter

are fed to grid 7 of first feedback amplifier V607A. The first, second, and third feedback amplifiers amplify the spurious signals. The V608A is applied as a degenerative feedback to the MO input to cancel the spurious signal.

The d-c correction voltage is not affected by the feedback circuit because the amplifiers do not pass the d-c component of the correction voltage. Meter M601 is used to zero the IO tuning. This meter is also used to measure the amount of deviation between the 1-mc crystal oscillator (Q1301, Q1302) and an external 1-mc standard frequency.

In order to keep the frequency control loop as stable and rapid-acting as possible, it is necessary to keep the gain in the a-c loop containing the 10-kc filter and feedback amplifiers approximately constant. However, at high operating frequencies a given d-c control signal to the voltage-sensitive capacitor in the master oscillator produces a greater amount of frequency correction than that which occurs at low frequencies.

In order to overcome this change in sensitivity, the input to V607A is shunted by a capacitive feedback network which is connected between the V607B cathode and the V607A control grid. The feedback network (not shown) is contained in the frequency compensating attenuator. The voltage at the V607A input thus lags the applied voltage. As the operating frequency increases, the voltage developed across the input network decreases. This action, in effect, lowers the input impedance to the feedback amplifiers and slows down the correction response as the operating frequency increases.

A third feedback signal from the feedback amplifier applies a regenerating control signal from the V607B cathode through a filter network (not shown) to the V607A control grid. This signal permits the feedback amplifier circuit to oscillate at approximately 2 cps when the oscillator control circuit is not locked-in at the proper operating frequency. The oscillator sweeps across a limited frequency range until the control circuits lock-in. When the frequency is correct the regenerative feedback drops to zero and oscillations in the feedback amplifiers are no longer sustained.

It was stated earlier that frequency error in the interpolation oscillator frequency is corrected by comparing the IO frequency with that of the 1-mc crystal oscillator reference frequency. The crystal oscillator frequency is first converted to a 1-kc signal. This frequency is compared with the IO frequency to produce a correcting voltage when operating in 1 kc steps.

The 100-kc pulse signal, which is derived from the 1-mc crystal oscillator, Q1301 and 1302 (fig. 5-11), is applied from L614 to buffer amplifier V604A. The 100-kc input is divided 100 times to produce a 1-kc output by the combined actions of two 5:1 dividers (V604B and V610A) and a 4:1 divider V610B.

The 1-kc signal from V610B is delivered to the 1-kc phase detector circuit consisting of T616 and crystal diodes CR632 and CR633. Here it is compared with the signal from the IO and a d-c correction voltage is developed. This correction voltage is applied to sweep oscillator and d-c amplifier circuit V609A.

When the IO circuit is not locked-in at a multiple of 1 kc, a positive-going error voltage signal is applied to the grid of sweep oscillator and d-c amplifier V609A. The plate voltage decrease of V609A is applied to the grid of sweep oscillator V609B. The resultant plate output of V609B is applied through an isolation network as a regenerative input to the V609A grid. This action produces a sweep voltage which is fed from the cathode of V609A through M601 to the interpolation oscillator phase detector reactance modulator which causes the IO to sweep over a narrow band and "hunt" for the proper lock frequency. Meter M601 provides an indication of the amount of the correction voltage.

SSB TRANSCEIVER AN/URC-32

One of the Navy's most versatile modern communication equipments is the AN/URC-32 (fig. 5-15). It is a transceiver operating in the 2- to 30-mc range, with a transmitter peak envelope power (PEP) of 500 watts.

The AN/URC-32 is designed chiefly for single-sideband transmission, and for reception on either the upper or lower sidebands, or on both sidebands simultaneously, with separate audio and i-f channels for each sideband. In addition to single-sideband operation, provisions are included for a-m (carrier reinserted), CW, or FSK operation.

The frequency range of 2 to 30 mc is covered in four bands. The desired operating frequency is selected in 1-kc increments on a direct-reading frequency counter. Frequency accuracy and stability are controlled by a self-contained frequency standard.
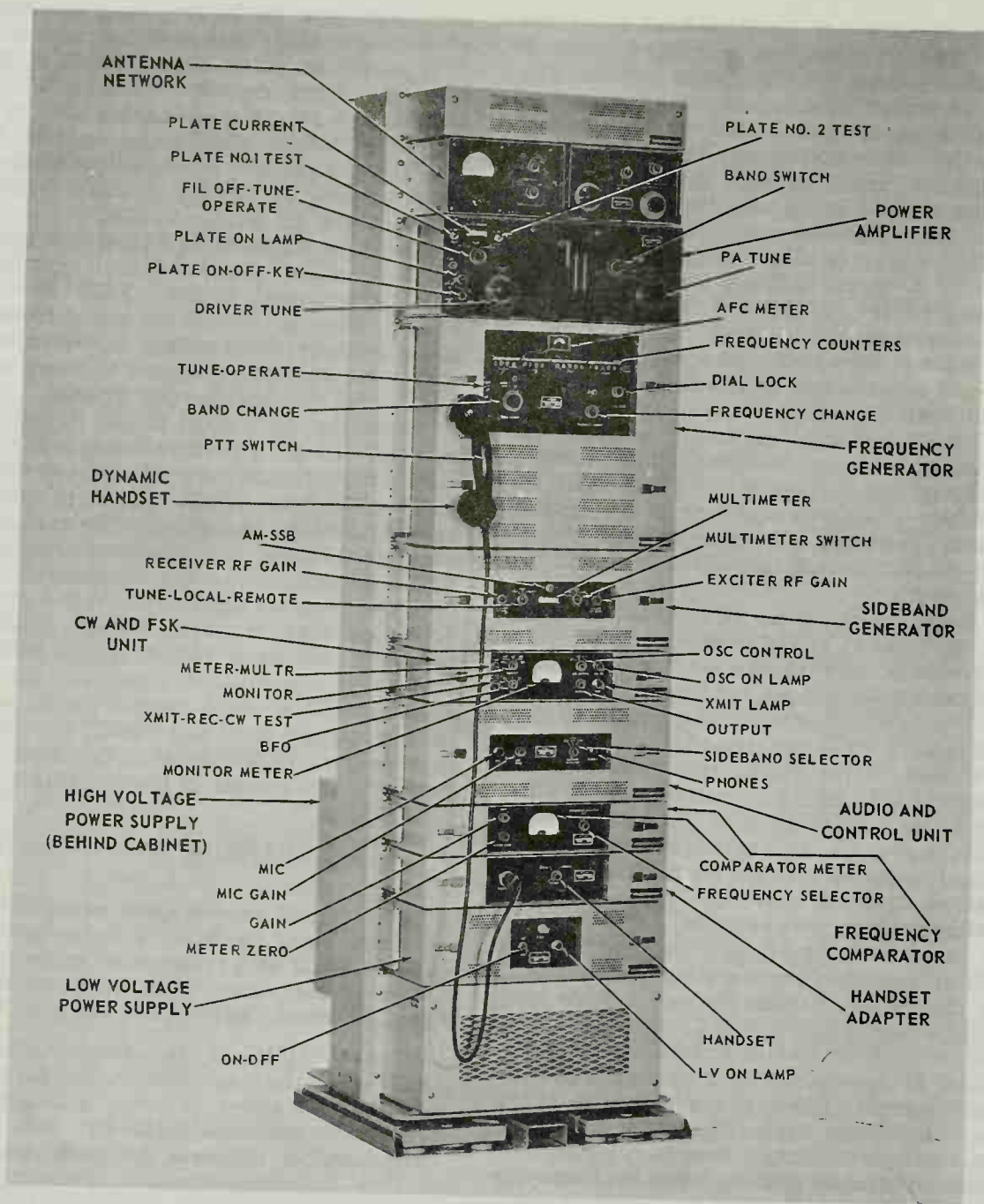
169

Figure 5-15.—Radio Set, AN/URC-32, relationship of units and operating controls.
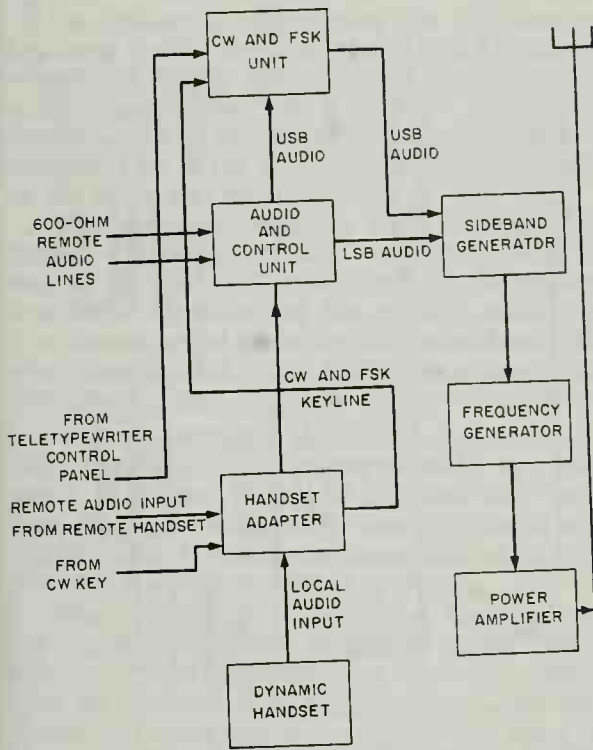
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BLOCK DIAGRAM

The transmitter of the AN/URC-32 (fig. 5-16) produces voice, CW, or FSK modulated signals on a single-sideband r-f carrier, or a compatible amplitude modulated r-f carrier.

Voice input signals from the dynamic handset or input signals from a remote control unit are

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Figure 5-16.—Radio Set AN/URC-32, transmit function block diagram.

applied to the handset adapter, permitting the operator to select either the local or remote audio input. Teletypewriter signals are applied directly to the CW and FSK unit, which provides separate audio tones for the mark and space conditions. These frequencies are later converted to the required frequency shift signals for FSK transmission.

The output from the handset adapter is amplified in the audio and control unit. Two separate audio input paths to the audio and control unit are provided through the 600-ohm remote audio lines.

The audio and control unit amplifies the audio signal and feeds it to the sideband generator. During single-side-band voice operation, the audio and control unit output is fed through a selector switch in the CW and FSK unit. For CW or FSK operation, the CW and FSK unit supplies audio tones to the sideband generator.

The sideband generator converts the audio input to the selected sideband of a 300-kc intermediate frequency. The modulated 300-kc output is fed to the frequency generator. This unit

provides the necessary number of heterodyning processes (while preserving the signal intelligence) to produce the selected carrier frequency in the 2- to 30-mc range. The output signal is amplified in the power amplifier to the required peak envelope power of 500 watts and fed to the antenna.

During receive operation (fig. 5-17) the antenna input signal in the range from 2 to 30 mc is heterodyned in the frequency generator so that the output will be a modulated 300-kc signal. This signal is detected and amplified in the sideband generator, further amplified in the audio and control unit, and fed to the speaker.

During CW reception, the CW and FSK unit supplies a 300.550-kc signal to the sideband generator as a beat frequency for the received signal. The beat frequency can be changed over a range of ±1 kilocycle.

AUDIO AND CONTROL UNIT

The audio and control unit (fig. 5-18) is a dual-channel amplifier, which can provide audio inputs from two 600-ohm balanced lines, a 600-ohm unbalanced line, or a high-impedance microphone. In the normal AN/URC-32 installation, the 600-ohm balanced lines and the dynamic microphone input are not used.
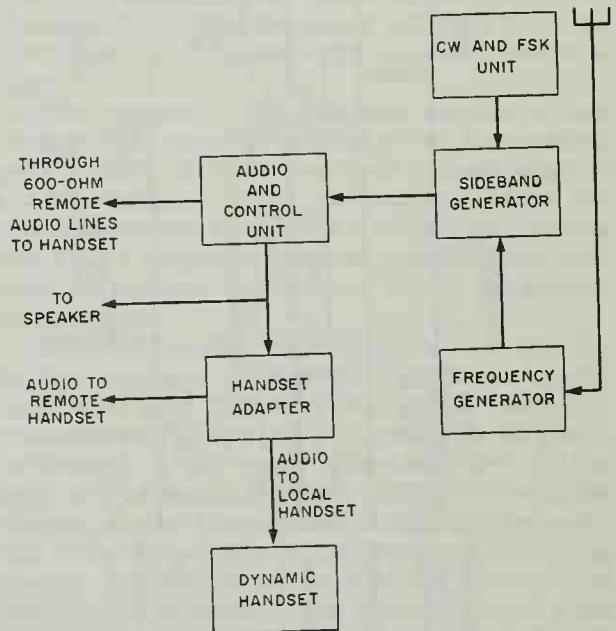
32.137

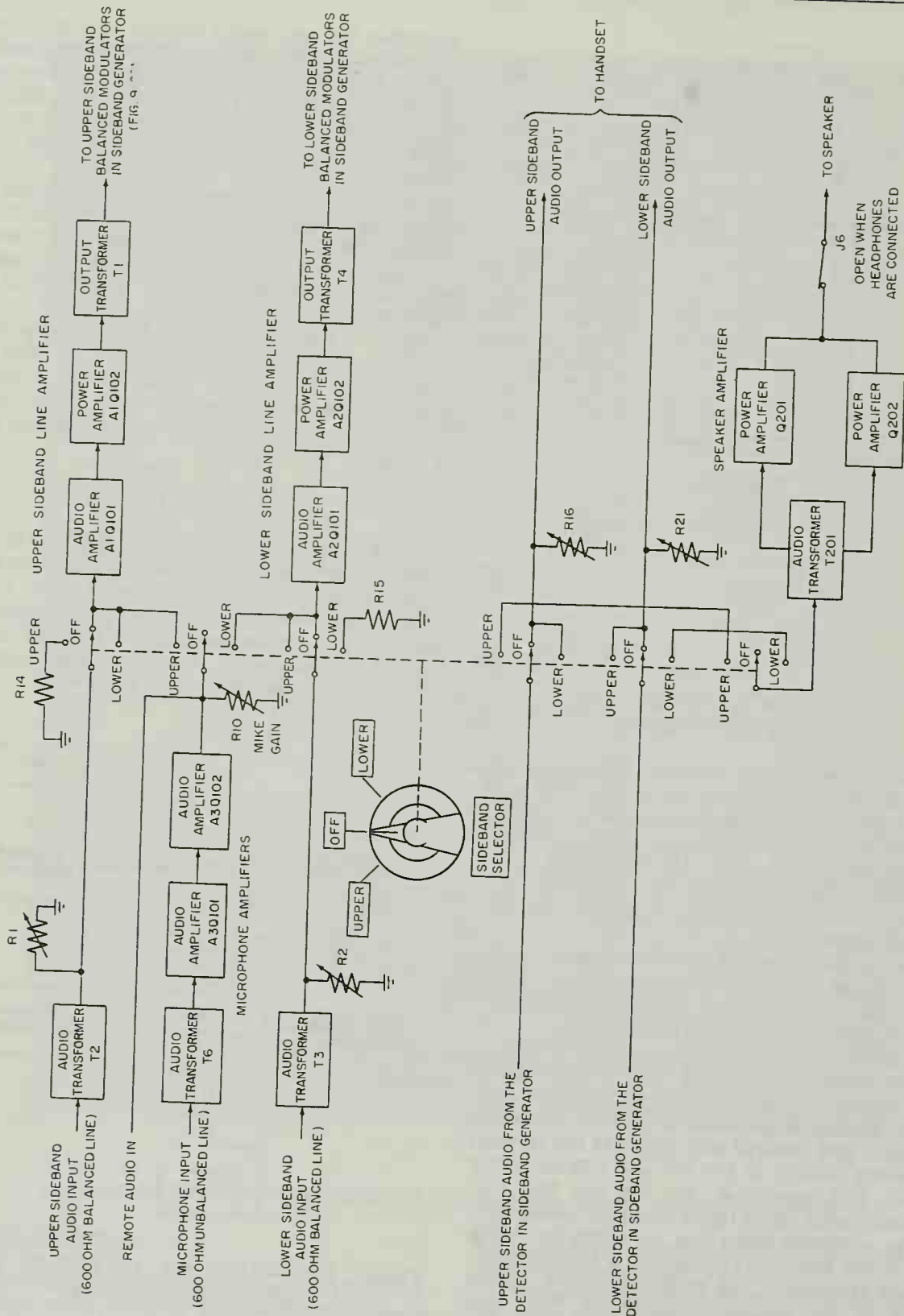Figure 5-17.—Radio Set AN/URC-32, receive function block diagram.

Figure 5-18.—Audio and control unit block diagram.

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On transmit, when using the 600-ohm unbalanced input, the audio signal from the handset adapter is fed to the audio and control unit via audio transformer T-6. This input, after amplification, can be applied to the upper sideband line amplifier or to the lower sideband line amplifier in the audio and control unit.

The sideband selector switch controls the signal transmission and reception. With the switch in the OFF position, the microphone amplifier circuits and the remote audio input are disconnected from the line amplifiers. This also connects the upper and lower sideband 600-ohm audio line inputs to the line amplifiers.

With the sideband selector switch in the UPPER position, the microphone audio or remote audio is fed into the upper sideband line amplifier. This also selects the upper sideband audio output from the sideband generator and applies it to the speaker amplifier circuits.

The reverse of this action happens when the sideband selector is placed in the LOWER position. When earphones (not shown) are plugged into the phone jack on the front panel, the audio output normally fed to the speaker is removed.

The upper and lower sideband line amplifiers are controlled by the upper and lower sideband audio inputs. With the sideband selector switch in the OFF position, the two sideband amplifiers can be used either individually or simultaneously.

Assume the audio input is on the upper sideband. This input is coupled by audio transformer T2 and the sideband selector switch (in the OFF position) to the first upper sideband line amplifier A1Q101 for amplification. The amplified output of A1Q101 is coupled to the power amplifier A1Q102 for further amplification. The output is fed via T1 to the upper sideband balanced modulator in the sideband generator. The gain of the sideband line amplifier is controlled by antenuator R1. Operation of both the upper and lower sideband line amplifiers is the same.

When the sideband selector switch is in the UPPER position, the upper sideband audio is removed from the audio amplifier A1Q101 and is correctly terminated by R14. Either the microphone amplifier circuit or the remote audio is then connected to the audio amplifier A1Q101.

With the sideband selector switch in the LOWER position, the lower sideband audio is removed from A2Q101 and correctly terminated by R15. The microphone amplifier circuits are then connected to the audio amplifier A2Q101.

SIDEBAND GENERATOR

The sideband generator (fig. 5-19) translates audio frequencies to intermediate frequencies during transmit condition, and intermediate frequencies to audio frequencies during receive condition.

Transmit Function

The balanced modulator, carrier generator, and transmitter gain control (TGC) operate during transmit condition. The audio input to the sideband generator is taken from the audio and control unit and applied to the sideband generator via T3 and T4. Audio input transformers T3 and T4 couple the upper sideband and lower sideband audio inputs to the balanced modulators.

The balanced modulators modulate a 300-kc carrier to produce separate and distinct upper and lower sideband signals with the carrier suppressed. The 300-kc carrier is produced in the carrier generator by tripling the 100-kc reference oscillator signal from the frequency generator.

The balanced modulator contains two 300-kc balanced modulators. Because of a frequency inversion in the r-f tuner, the lower sideband balanced modulator is followed with an upper sideband filter, and the upper sideband balanced modulator is followed with a lower sideband filter.

The outputs of the balanced modulator are connected in parallel and fed to the transmitter gain control. The TGC circuit is controlled by a TGC voltage, which is received from the power amplifier unit. This circuit maintains the 300-kc i-f output voltage at a sufficiently low level to prevent overdriving any of the subsequent stages.

The 300-kc SSB signals are fed via line TX to the r-f tuner and power amplifier. During tune and a-m transmit conditions only, the unmodulated 300-kc carrier generator output is reinserted in the upper sideband signal at the input of V1A in the r-f tuner. Reinsertion of the carrier at the transmitter eliminates reinsertion at the receiver and the necessity for having special equipment to receive the transmitted signal. The absence of the lower sideband does not affect the quality of the received signal. However, only one sideband plus carrier is transmitted, and the received signal is considerably stronger than it would be for double sideband a-m operation.
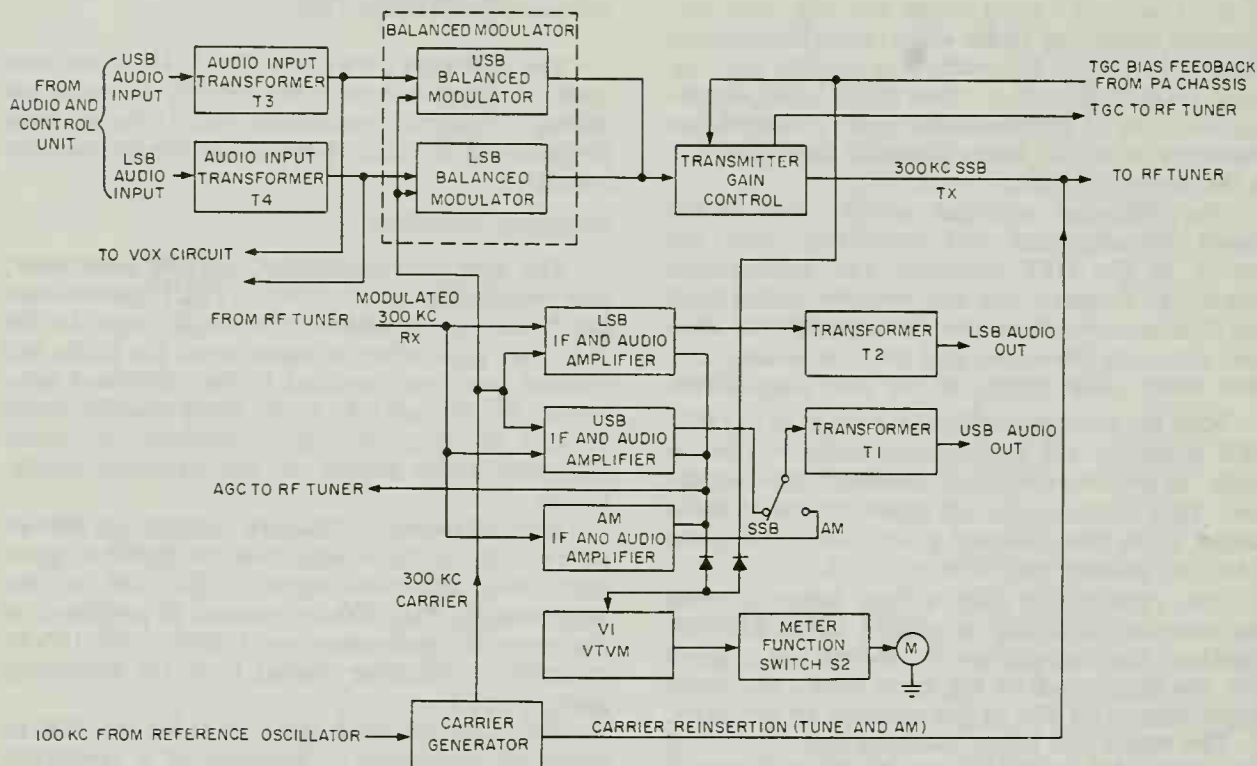
32.146

Figure 5-19.—Sideband generator, block diagram.

Receive Function

The i-f audio amplifier (lower sideband, upper sideband, and a-m) operate only during the receive condition to amplify the modulated 300-kc i-f signal from the r-f tuner (via line RX). These units also demodulate the signal and amplify the detected audio. A 300-kc carrier is reinserted into the lower sideband and upper sideband i-f/a-f amplifier from the carrier generator.

When the front panel SSB a-m switch is in the a-m position, the lower sideband and upper sideband i-f/a-f amplifiers are disabled. The carrier generator (a-m receive only) also is disabled. The audio output from the a-m i-f/a-f amplifier is fed through the SSB a-m position to the upper sideband audio output lines via transformer T1. The a-m i-f/a-f amplifier is disabled when the SSB a-m switch is in the SSB position.

CW AND FSK UNIT

The CW and FSK unit (fig. 5-15) enables the AN/URC-32 transceiver to be operated in the CW and FSK modes of operation. On FSK transmit operation (tone modulation), the CW and FSK unit converts the keying input from a tele-typewriter current loop to audio tones of 1575 cps for space (no loop current) and 2425 cps for mark (loop current). On CW transmit operation the unit provides a keyed audio tone of 1000 cps or 1500 cps as selected by the OSC control switch on the front panel.

During FSK receive operation, the CW and FSK unit provides a bfo (beat-frequency oscillator) signal required for FSK reception. This signal is centered on 300.550 kc, and is variable approximately 1 kc above or below this frequency.

The function of the CW and FSK unit is determined by the position of the oscillator control switch, which is located on the front panel. In
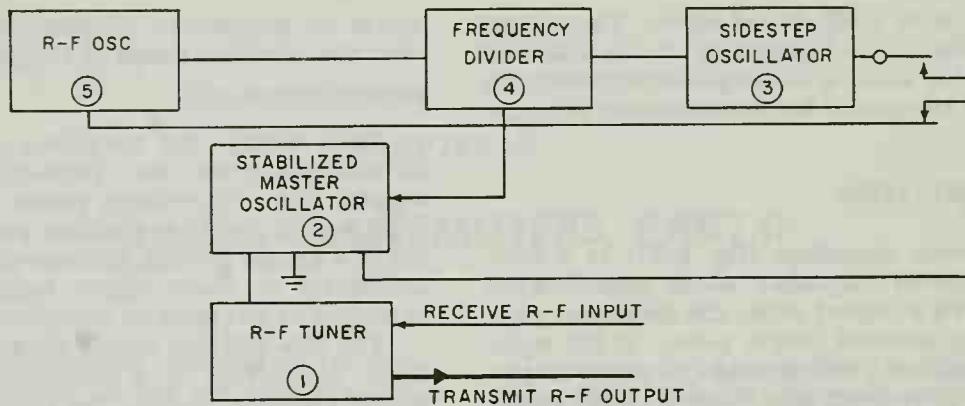
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Figure 5-20.—Frequency generator, simplified block diagram.

the OFF position, a section of the switch disables the CW and the FSK unit by removing the B + voltage (+130 v) from the circuits. Another section connects the USB transmit audio input line from the audio and control unit to the USB transmit audio output line. This line is connected to the USB balanced modulators of the sideband generator, which, in turn, feeds the transmit circuits. Thus, in the OFF position, the CW and FSK unit circuits are deenergized, and the voice input signals from the handset are transmitted.

FREQUENCY GENERATOR

The frequency generator produces the desired radio frequency as determined by the setting of the BAND CHANGE and FREQUENCY CHANGE controls. This unit also provides frequency control and r-f amplification for either transmitted or received signals.

The frequency generator (fig. 5-20) consists of a main chassis and five plug-in units. These units are:

1. R-f tuner.
2. Stabilized master oscillator.
3. Sidestep oscillator.
4. Frequency divider.
5. Reference oscillator.

The r-f tuner of the frequency generator is an i-f to r-f translator during transmit condition, and an r-f to i-f translator during receive condition. During the transmit condition, the r-f tuner accepts the 300-kc single sideband signal from the balanced modulators of the sideband generator via the TGC circuit, and translates it to the desired frequency (in 1-kc steps) ranging from 2.0 to 30.0 mc.

During receive condition, the r-f tuner accepts and amplifies the selected signal from the antenna (as indicated on the band dial) and
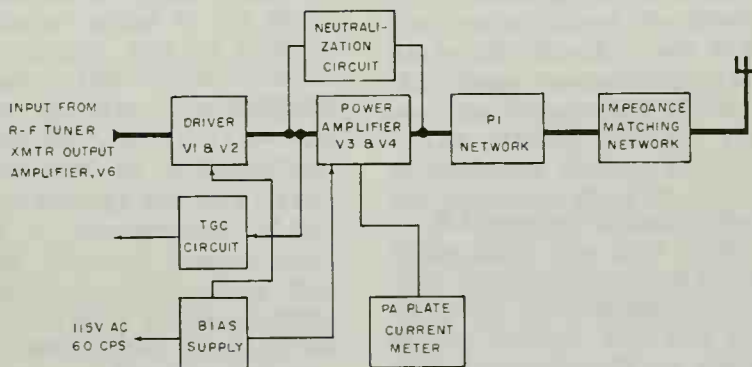

32.161

Figure 5-21.—Power amplifier, block diagram.

translates it to a 300-kc i-f signal. This signal is fed to the i-f/a-f amplifier of the sideband generator (or a-m i-f/a-f amplifier depending on the type of reception) for demodulation and amplification.

POWER AMPLIFIER

The power amplifier (fig. 5-21) is a two-stage r-f power amplifier which amplifies the 0.15-watt PEP signal from the frequency generator to a nominal output power of 500 watts PEP. It contains a driver stage, a power amplifier stage, a transmit gain control (TGC) rectifier, a bias and filament supply, and the necessary control and interlock circuits.

The driver and power amplifier plate circuits are manually tuned through the frequency range of 1.7 to 31.7 mc in four bands. The power amplifier plate circuit uses a tuned pi network to obtain an unbalanced 50-ohm output impedance over the complete range of frequencies.

POWER SUPPLIES

Plate voltage for the power amplifier tubes is obtained from the +2000-volt high-voltage supply. The high-voltage power supply utilizes eight series-connected silicon rectifiers in each leg of a bridge circuit. Silicon rectifiers operate efficiently at much higher temperatures than selenium or germanium rectifiers.

The 400-volt d-c supply consists of 12 rectifiers connected in a conventional full-wave bridge rectifier circuit. The output is fused and filtered by a conventional LR filter.

A low-voltage power supply provides several outputs of relatively low voltage as follows: +250 volts, +130 volts, unfiltered +28 volts partly filtered +28 volts, regulated +28 volts and -90 volts.

CHAPTER 6

RADIO TRANSMITTERS, PART II

This chapter contains information on non-conventional circuits of two transmitters discussed in chapter 5. In addition, two shore-based transmitters and two radio sets are presented in a functional block diagram discussion.

SPECIAL CIRCUITS AN/SRT-14, 15, and 16

Non-Conventional circuits concerning the transmitter and antenna tuning equipment for the AN/SRT-14, 15, and 16 are discussed below. Reference to the block diagrams in chapter 5 of this training course will be helpful.

500-W OPERATE RELAY AND 500-W DISABLE LINE

The 500-watt operate relay, K1 (fig. 6-1) control the GROUND FOR 100-watt line, the GROUND FOR 500-watt line, and the GROUND FOR 500-watt AX line.

The GROUND FOR 500-watt line and GROUND FOR 500-watt AX line allow circuit completion for several control circuits. They are responsible for turning on control circuits in the high voltage power supply, for connecting the output of the low-level radio modulator to the input of the high-level radio modulator, for switching the input to the PA plate from the low-level radio modulator output to the high-level radio modulator output, and for supplying screen grid voltage to the high-level radio modulator during phone operation. The GROUND FOR 500-watt line and GROUND FOR 500-watt AX line are completed when K1 is energized. The completion is made through contacts 8R and 9R, and 1 L and 2 L, respectively, of K1.

The GROUND FOR 100-watt line is completed when K1 is not energized. This line completes the control circuits in the medium voltage power supply, which then supplies the required PA plate potential for 100-watt operation.

To place the transmitter in 500-watt operation, the Push for 500-watt button, S3, is depressed. If the -24 volts after time delay relay, K5 (to be discussed later), has timed out (contacts 2-8 closed), depressing the 500-watt button will apply the -24 volts to the winding of K1. The necessary ground for completion of this circuit is accomplished through a series of protective devices, constituting the 500-watt disable line. The circuit must be complete before the transmitter will switch to the 500-watt level. The protective devices included in the 500-watt disable line are the bandswitch, S2N; the disable 500-watt pushbutton, S4; the Radio Frequency Oscillator (RFO) unit 14 door interlock, S5; the time delay relay, K5; the thermostat, S6; the blower switch, S7; the overload relay, K4 (contacts 2R and 3R); and the plate thermostat, S8. If any one of these controls or protective devices is not closed, the ground will not be completed to K1, and the transmitter can operate at the 100-watt level only. When K1 energizes, the -24 volt after time delay connects to the windings of K1 through its contacts, 1R and 2R, allowing S3 to be released.

The 500-watt disable line contains a switch, S2N, which is ganged to the bandswitch control. The switch allows the circuit to be completed in all bands except 1 and 3. It should be recalled that those frequencies up to and including 2 mc may be transmitted at the 100-watt level only.

The disable 500-watt pushbutton, S4, which is normally closed, may be depressed to open the circuit as desired.

A transparent door covers the control knobs of the RFO. The door contains a switch, S5. This door interlock controls the open and closed condition of the 500-watt disable line, and, if open during normal 500-watt operation, the transmitter output will be reduced to the 100-watt level.
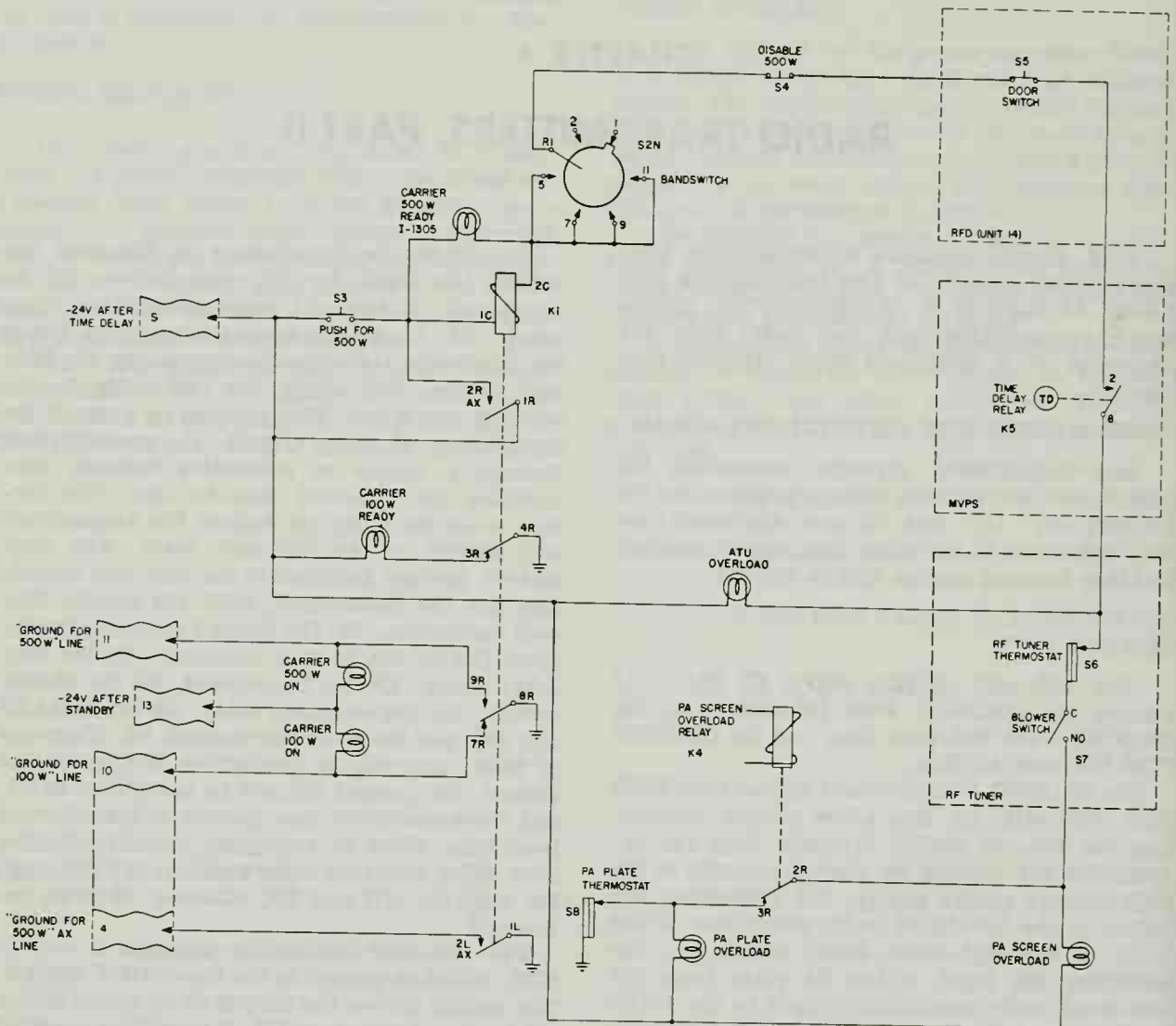
Figure 6-1.—500-watt operate relay and 500-watt disable line simplified schematic.

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The heaters of the tubes in the high-voltage power supply are allowed to warm up before power is applied to the plate transformer. The contacts, 2 and 8, of the time delay relay, K5, are held open until the time delay is over. After this time, the contacts (2 and 8) are closed to complete the circuit.

Thermostat S6 is normally closed. When excessive heat is radiated by tuning coils in the r-f tuner, S6 will open the circuit.

Blower switch S7 is a centrifugal switch operated by the blower motor in the r-f tuner. The circuit is closed through S7 only if the blower is operating.

The contacts, 2R and 3R, of the PA screen overload relay are normally closed. When K4 is energized, due to excessive PA screen grid current, contacts 2R and 3R are opened, and the ground is removed from the 500-watt disable line.

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Four indicator lamps controlled by the action of K1 indicate the power level of the transmitter and also whether or not the transmitter is on the air.

Contacts are provided on the antenna switching relay, K2 (fig. 6-2), for connecting a receiver to the antenna when the transmitter is off the air. When the transmitter is on the air, the output of the RFA is coupled to the antenna system, and the receiver input is grounded.

During hand-key operation (using S5, in fig. 6-2) the circuit for K2 is completed from the -24 volt after time delay, through the winding of K2, into the low-level radio modulator, through contacts 1L and 2L of relay K6, and through the closed contacts of the test key to ground. Energizing the antenna switching relay, K2 actually operates the keying relay, K6, in the LLRM. This action shifts the ground on K2 from 1L and 2L of K6 to the contacts 2L and 3L of K6. C1 is provided to continue the ground on K2 during the time of the changeover. The antenna relay, K2, therefore becomes energized when the test key is closed, corresponding to a mark, moving its contacts, 3L and 2L (fig. 6-2),

to the transmit position. When the hand-key is up, corresponding to a space, there is no ground for relay K2 and the antenna is switched to the receive position.

In machine c-w, frequency-shift keying, and facsimile operation, a permanent ground is placed on the keying line (through the keying equipment), which keeps K2 energized.

TRANSMITTER COUPLER
LOAD ADJUSTING UNIT (LAU)

The function of the load adjusting unit (fig. 6-3) is to improve the impedance match between the 50-ohm output impedance of the radio-frequency amplifier and the impedance presented by the antenna tuning equipment. The improved matching permits the final PA stage of the radio-frequency amplifier to operate at maximum plate efficiency, and improves the transfer of power to the antenna system. The load adjusting unit consists of the standing-wave ratio monitor and the impedance transformer circuits.
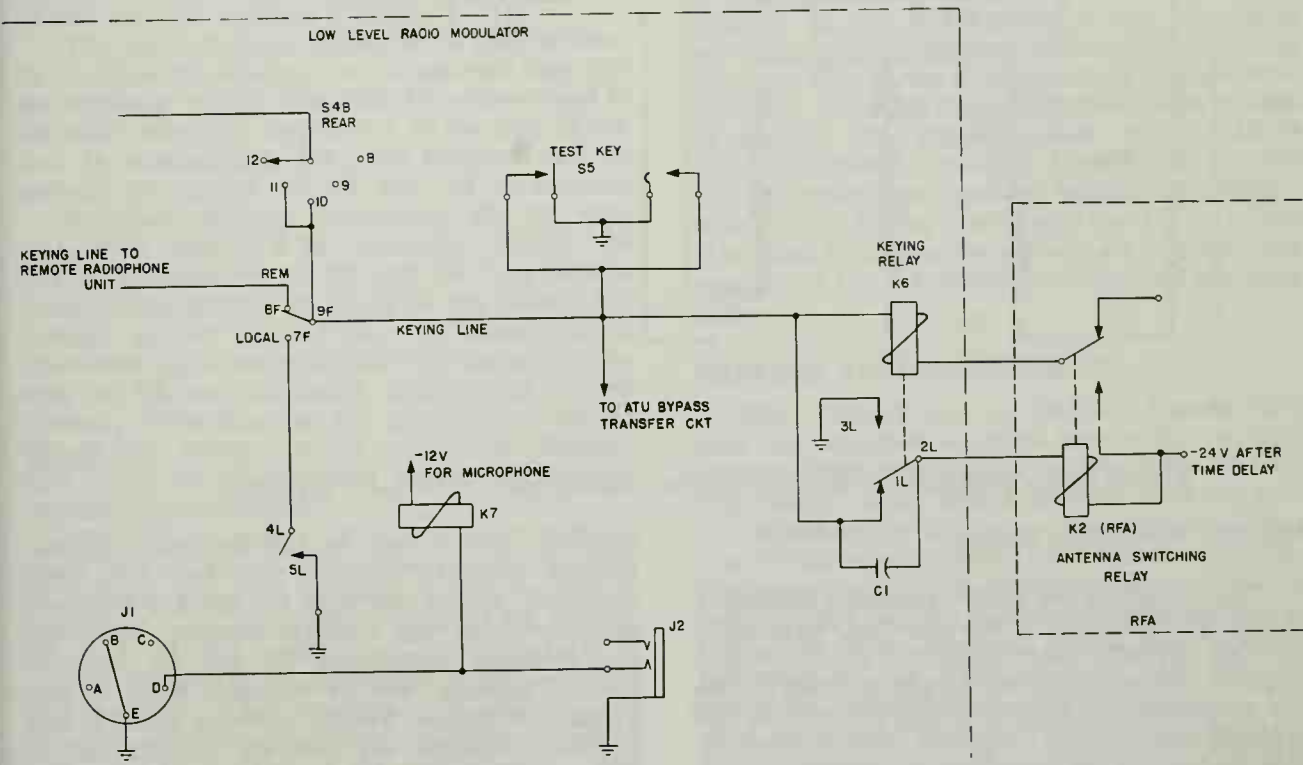

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Figure 6-2 —Antenna switching relay and associated circuits.

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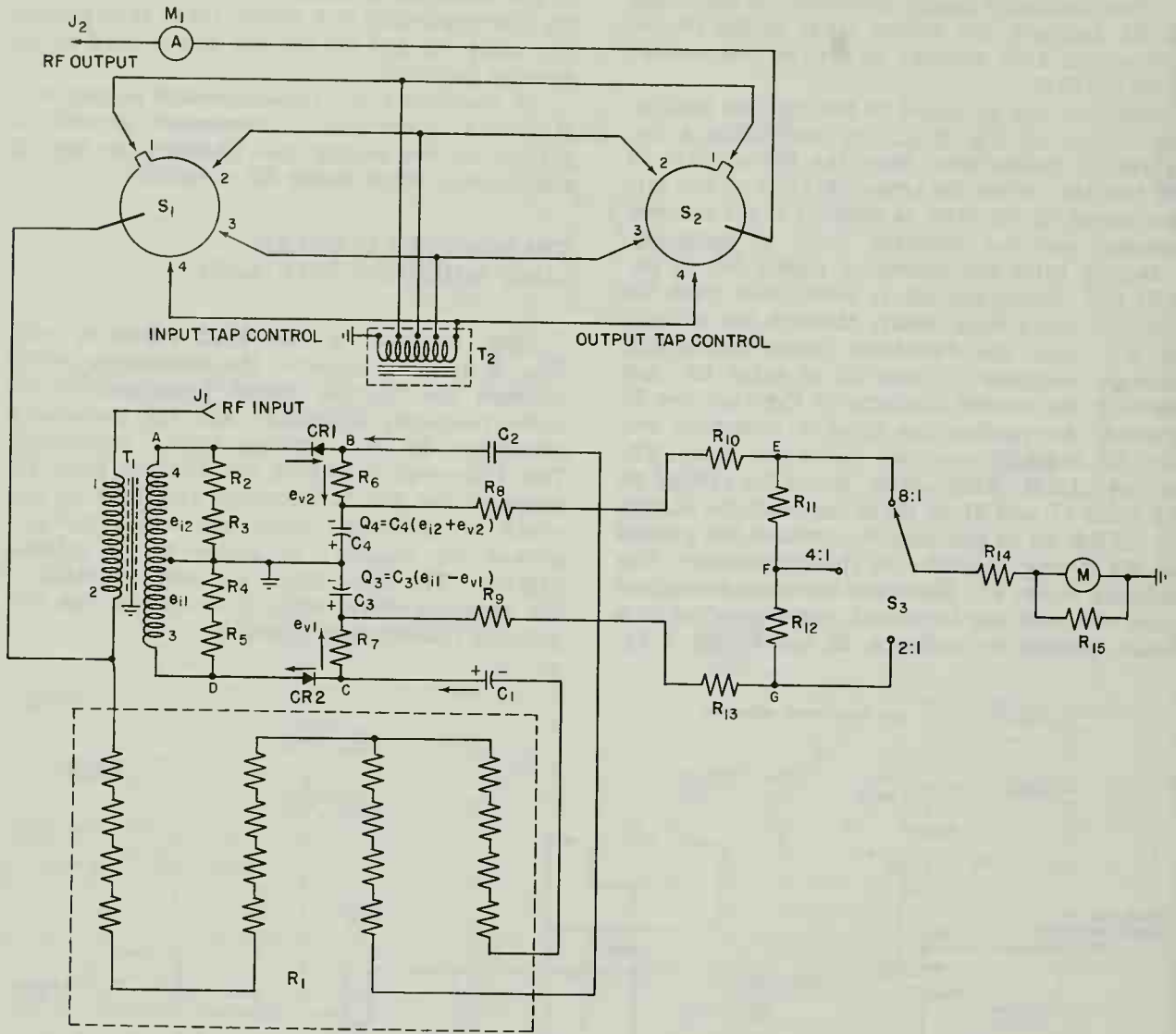
Figure 6-3.—Transmitter coupler (load adjusting unit), schematic diagram.

Standing Wave Ratio Monitor

The output of the radio-frequency amplifier (RFA) is connected to the standing-wave ratio monitor circuit, which functions to detect and monitor the standing-wave ratio present on the r-f transmission line. The components of the standing-wave ratio monitor circuit form a unit designated the transmitter coupler.

The r-f transmission line from the RFA is connected through J1 to the standing-wave ratio

monitor circuit and to the impedance transformer circuit. Voltages e_{v1} and e_{v2}, proportional to the r-f line voltage are obtained directly from the voltage divider, $R1$, and d-c blocking capacitors C1 and C2, respectively. Voltage e_{v1} is between point C and ground. Voltages e_{v2} is between point B and ground. Voltages e_{i1} and e_{i2} proportional to the current in the r-f line, are obtained from the lower and upper halves, respectively, of the secondary winding of the transformer,

T1. The center tap of the secondary is grounded, and ev_1 and ev_2 are in phase with each other and are of equal magnitudes.

Circuit ABCD supplied by the T1 secondary and also by a portion of the line voltage via C1 and C2 includes capacitors C3 and C4, which acquire a charge through the action of crystal diodes CR1 and CR2 and the voltages ei_1, ev_1 and ei_2, ev_2.

On one half-cycle, CR1 and CR2 are conducting, and on the other half cycle they are nonconducting. During the time the diodes conduct, ei_1 and ei_2 attempt to establish a charging current clockwise around circuit ABCD so as to charge C3 and C4 with the polarity shown. During this time ev_1 opposes the charge on C3, and ev_2 aids the charge on C4.

As stated before, ei_1 and ei_2 are proportional to the line current; ev_1 and ev_2 are proportional to the line voltage. Now the quantity of charge in C3 is proportional to $ei_1 - ev_1$; the quantity of charge in C4 is proportional to $ei_2 + ev_2$. Thus, the voltage on C3 is proportional to the difference of two quantities, and C4 is proportional to the sum of two quantities. These quantities are proportional to line current and line voltage.

The ratio of these quantities is proportional to the line impedance; for a matched line with no standing waves, the ratio is proportional to the characteristic impedance of the line. If the line is mismatched, the ratio is upset, and the amount of charge on C3 and C4 is altered.

A voltage divider comprising R8, R9, R10, R11, R12, and R13 is connected across the series combination of C3 and C4. The voltage across this divider is equal to the sum of the voltage across C3 and C4. The meter, M, is connected effectively between the common junction of C3 and C4 and a point on the voltage divider, depending on the position of switch S3. In this bridge circuit the current through the meter is proportional to the unbalanced condition of the bridge.

The values of R10, R11, R12, and R13 are such that if the standing wave on the r-f line is 8:1, the potential at point E with respect to ground is zero. If the SWR CALIBRATE switch, S3, is in the 8:1 position, the SWR BALANCE meter, M, would give an indication at the center null position. If the standing-wave ratio is less than 8:1, with the switch in this position the meter will read to the left of the null position. Conversely, if the standing-wave ratio is greater than 8:1, with the switch in this position the meter would indicate a reading to the right of the null position. The voltage at point F with respect to ground will be zero when the standing-wave ratio is 4:1. Similarly, a 2:1 ratio will give a zero reading at point G. R14 in series with the meter is a current-limiting resistor.

Impedance Transformer

A method for improving the match between the characteristic impedance output of the RFA and that impedance presented to the RFA by the antenna and antenna tuning equipment is provided through the use of the impedance transformer. Matching is accomplished through the use of a tapped autotransformer T2 (fig. 6-3) which has been introduced into the r-f transmission line. The output of the standing-wave ratio monitor may be connected to one of the taps of T2 as shown. The input connections are made through switch S1, the input tap control. The output of the transformer selected by the position of the tap on the output tap switch, S2 is connected through the r-f ammeter M1, to the jack, J2, which is the output to the antenna coupler.

With the input and output tap controls set at position 4, the full winding of the impedance matching autotransformer will be shunted across the line. This is the normal setting of these controls. If, after all tuning procedures have been completed, the standing-wave ratio balance meter indicates a reading higher than 2:1, the controls should be reset for optimum impedance match. The tuning controls of the PA stage are also used in making the adjustment for maximum reading on the r-f ammeter of the load adjusting unit.

ANTENNA TUNING EQUIPMENT

The AN/SRT-14, 15, and 16 transmitting sets are designed to work into a 35-foot whip antenna Navy type C66047 or into a 60- to 130-foot single wire with a 40-foot down lead. If the impedance presented by the antenna is other than 50 ohms, energy will be reflected back along the transmission line, reducing the energy being radiated and causing voltage nodes to exist on the line. The standing-wave ratio measurement indicated by the standing-wave ratio balance meter represents the extent of the mismatch. The ideal matching condition of the antenna load to the RFA output would yield a standing-wave ratio of 1:1. This would occur when the reactive component of the antenna is reduced to zero and the

resistive component is equal to 50 ohms. Acceptable tuning of the antenna load by the antenna tuning equipment is achieved if the standing-wave ratio can be reduced to at least 4:1 for any frequency in the transmitter range.

If the antenna length is an exact multiple of quarterwave lengths corresponding to the frequency in use, the antenna will be resonant and will present zero reactance to the transmission line. It should be understood, however, that since the antenna length is constant, the resonant condition will not remain constant for all frequencies. Further, at even multiples of quarter-wavelengths, the resistive component (for end fed) is high, while at odd quarter-wavelength multiples, the resistive component presented by the antenna (at the same feed point) is low.

The function of the antenna tuning equipment is to make the antenna length, together with the selected tuning component, appear at the feed point as some odd multiple of a quarter-wavelength for all frequencies within the tuning range from 0.3 to 26 mc.

Antenna Coupler

A schematic diagram of the antenna coupler is shown in figure 6-4. The loading switch, S1, permits additional components to be placed into the transmission line to increase its effective length when the length of the main tuning coil is insufficient to achieve the required tuning. This is particularly true at low frequencies when the antenna length should be increased. At frequencies where the effective length of an antenna is an even multiple rather than the desired odd multiple of a quarter-wavelength, a capacitor may be switched in the line. Without this action, poorer match would exist between the antenna and the RFA because the antenna impedance would be further separated from the characteristic impedance of the line.

The input of the r-f tuner is permanently connected to the loading switch of the antenna coupler. If one of the loading components is required in the tuning process, the required component is switched into the transmission line through the two-section loading switch, S1. If the loading components are not required, the transmission line is connected directly through the switch shown in the figure. Loading switch

S1 has six positions: one for direct connection and five (ABCDE) for different combinations of loading components. The components selected may be easily located after proper rotation of the switch. The antenna coupler loading control, S2, supplies operating voltage to an actuator, B1, in the antenna coupler. The actuator then functions to position the loading switch as required to select the proper loading component. Table 6-1 shows the components selected by the different positions of the antenna coupler loading control.

Table 6-1—Components Selected in Each Position of the Loading Switch, S1.

Positions of antenna coupler loading switch, S2	Loading components selected by S1
Direct	None
A	C1 in series
B	L2 in series
C	C1 in series, C3, C2, shunted across the line
D	L1 in series
E	C3, C2, shunted across the line

For example, assume that the antenna coupler loading switch, S2 (fig. 6-4), is placed manually in position A. This action establishes a circuit from one side of the 110-volt a-c line through 11-14 of S2, through contacts NO and C of microswitch S4, to one side of the actuator winding, B1. The other side of the actuator winding is directly connected to the other side of the 110-volt a-c line. The closed contacts NO and C of S4 allow the actuator to rotate until the roller arm of S4 falls into the notch of the positioning cam. This action will open the contacts of S4, breaking the a-c supply to the actuator. The magnetic actuator is ganged to S1 in such a way that the switch is caused to rotate to the A position.

At certain frequencies, the actual antenna impedance is such that no tuning is required to meet the standing-wave ratio readcng of 2:1 required at the r-f output of the radio frequency amplifier. At other times, it is necessary to use one or more of the tuning components described in the discussion of the

CONTROL INDICATOR   ANTENNA COUPLER

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Figure 6-4.—Simplified circuit of antenna coupler.

loading switch in the above paragraphs. To meet these conditions the antenna bypass switch, S3, is introduced into the transmission line.

The antenna bypass switch is controlled by the antenna transfer switch on the control indicator. The switch is a double-pole double-throw toggle switch with three positions. These positions are designated TUNER IN, BYPASS, and REMOTE. The REMOTE position is not shown in the figure. S3 is positioned by a magnetic actuator, B2. One side of the a-c line is permanently connected to one side of the actuator. Relay K1, when energized, completes the line voltage circuit allowing the actuator to energize, which, in turn, rotates the positioning cam. In this manner, the antenna bypass switch is caused to position itself in accordance with

the position chosen by the antenna transfer switch. Relay K1 receives its operating voltage from the -24 volt after time delay line.

With the antenna bypass switch in the BY-PASS position (as shown), relay K1 is not energized, and the r-f output from the load-adjusting unit is directly connected to the antenna through J1 and contacts 4 and 1 of S3. When it is desired to have the tuning elements in the line, the antenna transfer switch is placed in the TUNER IN position. S3 is therefore rotated so that the r-f energy now passes through contacts 4 and 3 of S3, and then out through connector J2 to the r-f tuner output. The output of the r-f tuner is permanently connected to the loading switch, S1. The loading switch selects the desired tuning component. The output of S1

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is connected to contact 2 of bypass switch S3, through contact 1 and to the antenna.

When S3 is switched from TUNER IN to BY-PASS, or vice versa, the transmission line is momentarily interrupted, removing the antenna load. When the antenna load is removed, even for a short time, it becomes necessary to cut off the radio frequency energy at the IPA to prevent surges in the transmitter output caused by this removal. Grounding cam A (fig. 6-5) is fixed to the shaft of the bypass switch, S3, and controls the open or closed condition of microswitch S2. The configuration of the cam is such that it removes the ground from the R9, R10 junction in the IPA control grid circuit during the time that switching takes place. The absence of this ground will cause the IPA to cut off, which, in turn, cuts off the output of the transmitter.

R-F Tuner

The antenna coupler applies its output to the transformer switch, S1 (fig. 6-6) in the r-f tuner. This switch, when connected so that circuits are completed through the 1-2 contacts and the 3-4 contacts, will place the 9:1 impe-

dance transformer in the transmission line. The purpose of the transformer is to step up the signal to the required level. In the other position of the switch (as shown), the impedance transformer is bypassed by the connection made between contacts 2 and 4 to the main tuning coil of the r-f tuner.

Magnetic actuator B1, responsible for positioning switch S1, operates under the influence of the transformer switch, S2, through the contacts, NO and C, of S3 to one side of the winding of actuator B1. The other side of the actuator is connected directly to the other side of the line. Actuator B1 rotates until the roller actuator of S3 falls into the notch of the positioning cam. At this time, the NO and C contacts of S3 open, breaking the line voltage circuit. The transformer switch, S1, is now in the position that eliminates the impedance transformer from the circuit. Placing S2 in position 14 will again energize the circuit through S4. B1 will then cause the positioning cam to rotate until the actuator for S4 falls into the notch of the cam. This corresponds to the position of switch S1, which places the impedance transformer in the circuit.
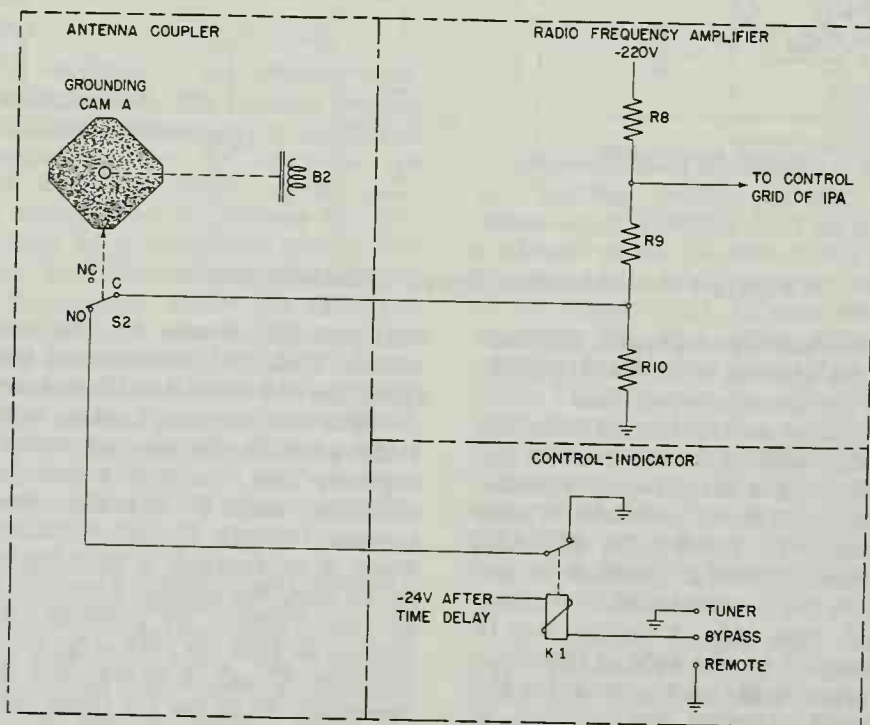
Figure 6-5.—Grounding cam circuit, simplified schematic.

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Figure 6-6.—Impedance transformer (r-f tuner) simplified schematic.

The transmission line is coupled to the main coil (fig. 6-7) of the r-f tuner by a single loop conductor coil, L1, mounted on the shorting ring. The length and configuration of the single loop coil is such that optimum coupling exists above 1 mc if a standard Navy 35-foot whip antenna is used. The resistive component of the antenna impedance is so low at a frequency of 1 mc or lower that the impedance transformer will be necessary to bring the radiated signal to the minimum allowable limit.

The main tuning coil of the r-f tuner is equipped with an adjustable sliding short. The tuning coil is the main tuning component in the r-f tuner.

If the impedance transformer (fig. 6-6) is not required in the antenna tuning process, the r-f tuner input to its loading components is connected to the main tuning coil, L2. This

tuning coil (fig. 6-7) is a section of a helical center conductor transmission line and is adjustable as forestated, by a sliding short. The short is adjusted so that the effective length of the main coil, together with the effective antenna length, is an odd multiple of a quarter-wavelength at the particular frequency in use. This is the point of resonance or zero reactance.

The position of the sliding short may be adjusted either manually by an adjustment control or remotely by a drive motor, B1. The UP, DOWN, and SLOW pushbuttons control the speed and direction of the motor.

ATU Bypass Transfer Circuit

The antenna tuning system may be bypassed in the absence of a transmitter carrier, allowing the antenna to be switched to a receiver. The

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Figure 6-7.—Main tuning coil (r-f tuner), simplified schematic.

receiver may be tuned to a different frequency than that of the transmitter. The presence or absence of a carrier is indicated by the condition of the keying line in the low-level radio modulator.

The antenna bypass transfer circuit is mounted in the LLRM for convenience of location only. The circuitry of the antenna bypass transfer circuit is shown in figure 6-8. It is undesirable to have the antenna switching attempt to follow the speed of the hand-key signals. The antenna bypass transfer circuit therefore functions to provide a signal that will place the antenna tuning system in the circuit at the time of the first hand-key signal, and keep the system in contact with the antenna until after the last hand-key signal has been transmitted.

The relay, K2 (fig. 6-2), when energized completes the signal path through contacts 3 and 4 to the antenna tuning system, which keeps the antenna and antenna tuning equipment in the line. Conversely, when deenergized, the contacts open, and the antenna tuning system is bypassed. Reference to figure 6-2 shows that the keying line during key-up condition (S5 contacts open) is at a potential of -24 volts, while in key-down condition (S5 contacts closed), the keying line is at ground potential.

The control grid of the antenna transfer control tube, V22B and the cathode of the limiter, V22A, receive a -10-volt potential when V22A is nonconducting.

The -10 volts is taken from the junction of R61 and R62.

Since the plate of the limiter section of V22A is connected to the keying line, its potential during key-up condition is -24 volts. With the cathode of V22A at -10 volts and the plate at -24 volts, the tube does not conduct. V22B is also held in the cutoff condition by the -10 volts at its grid. No current therefore passes through the winding of K2, and contacts 3-4 are open. In this condition, the transfer control circuit is open and the antenna tuning system is bypassed.

Capacitor C60 connected from the plate of V22B to the cathode of V22A is charged to approximately 260 volts. A charge in excess of the plate supply voltage is evident since the cathode, to which one side is tied, is connected to a point of -10 volts with respect to ground.

On arrival of the first mark (key-down) signal of a hand-key transmission, the keying line is at zero potential, and the limiter, V22A, conducts, which reduces the bias of V22B, allowing it to conduct. Relay K2 energizes, closing the antenna transfer control circuit, which, in turn, removes the bypass from the antenna tuning equipment.

When the keying signal changes from mark to space (key-up), the V22A limiter plate again returns to -24 volts, which renders the tube non-conducting. As V22A begins to move toward cutoff, capacitor C60 begins to recharge. Since the grid of V22B is going more negative, a reduction in plate current is evidenced. The recharge of C60 opposes the tendency of the grid to assume the negative 10 volts and V22B moves slowly toward cutoff. The relay, K2 will remain energized until the current through V22B is reduced to 0.5 milliampere.

The time constant for the charging of C60 and, likewise, the rate at which the grid of V22B returns to the -10 volt potential is determined by the value of K2 winding resistance; the resistance of the B supply; and the value of C60, R63, R64, and R62. Resistor R64 serves as the time constant adjust control, which varies the charge time of C60 from 0.5 to 6 seconds. The relay, K2, can therefore be made to remain energized for any length of time, within this limit, after the final mark signal.

KEYING CIRCUIT

The keying input circuit, containing service selector switch, S4A, selects the input to
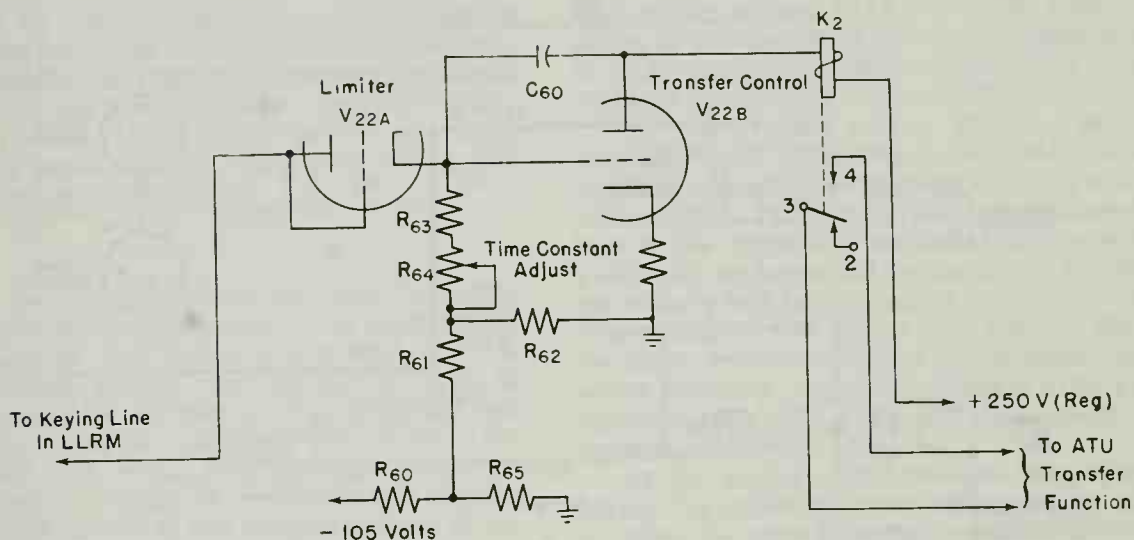
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Figure 6-8.—ATU bypass transfer circuit, simplified schematic.

the keying circuit. In hand-key and phone opera-tion, the keying signals are introduced to the LLRM through the action of the keying relay, K6 (fig. 6-9). With the local-remote switch S1, in the REM position, the keying line is completed to ground in a radiophone unit through contacts 8F and 9F of S1 (if the key is in the MARK or DOWN position). When the key is up, there is no ground. Therefore, the transmitter is keyed on by a mark, and off by a space in c-w operation. The antenna switching relay, K2, is also operated to switch the antenna to the transmitter when the ground is supplied. When in phone operation, the press-to-talk button also supplies ground to the keying line when it is depressed.

If the phone is connected locally, S1 is in the LOCAL position, and the ground from the press-to-talk button also supplies ground to the keying line when it is depressed.

If the phone is connected locally, S1 is in the LOCAL position, and the ground from the press-to-talk button operates K7 from the +12-volt supply. With K7 operated, ground is fed to K6 through contacts 5L and 4L of K7 and contacts 7F and 9F of S1.

The test key, S5, parallels the keying line so that depressing S5 also supplies a ground to operate K6. With the service selector switch, S4, in either FSK or FAX position and the local

remote switch, S1, in the REM position, a per-manent ground is placed on the keying line through contacts 10 and 6 or 11 and 6 of S4B, from the keying equipment.

Two gas tube voltage regulators V17 and V18 supply a regulated -105 volts and +105, respectively, for use in the keying circuit. The -105 volts is connected to contact 4R of relay K6, and the +101 volts is connected to contact 6R. In hand-key operation, when the key is depressed (mark), K6 operates, and +105 volts is connected through contact 6R and 5R of K6, resistor R40, contacts 8 and 6 of switch S4A, contacts 8R and 10R of S6 in the operate position through R41 to ground.

When the key is up (space), K6 is not ener-gized, and -105 volts is now applied through con-tacts 4R and 5R of K6 and through the same cir-cuit as stated above. Resistors R40 and R41, acting as a voltage divider across either V18 or V17, provide either +85 volts or -85 volts for mark and space, respectively, to the grid of the buffer stage, V11A during hand-key operation. In phone operation, the same control of the buffer stage is accomplished through contacts 12 and 6 S4A except that the press-to-talk button replaces the hand key. The result of this action applies +85 volts to the grid of the buffer when the press-to-talk button is

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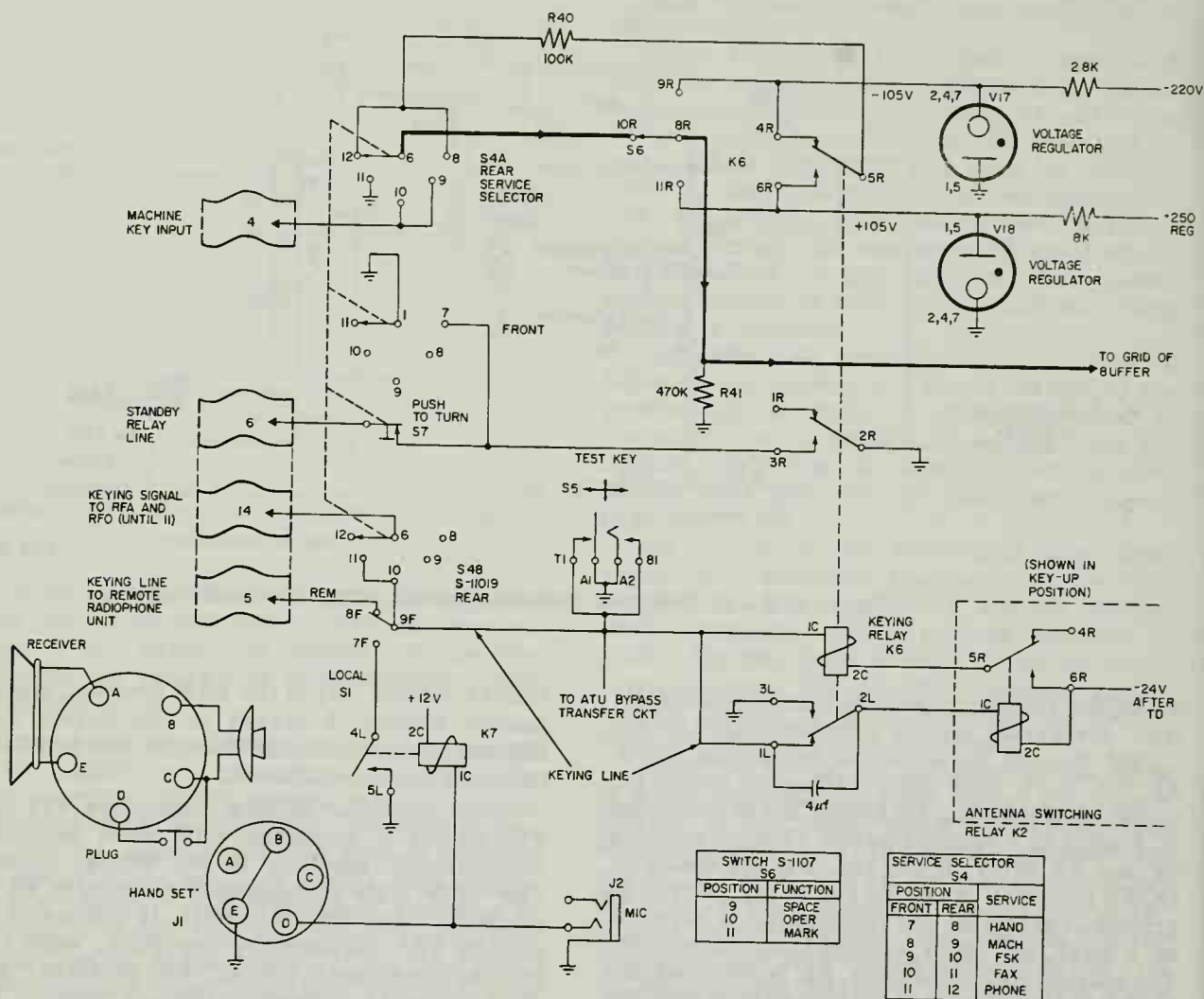
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Figure 6-9.—Keying input circuit, simplified schematic.

depressed and -85 volts to the grid when the button is released

A permanent mark or space signal can be produced at the grid of the buffer by the action of S6. Setting this switch to mark connects +105 volts to the grid of the buffer through contacts 11R and 8R of S6. In the space position. S6 connects -105 volts through its contacts 8R and 9R to the buffer grid.

In addition to providing keying voltages for the keying circuit, the keying relay, K6, also controls the action of the standby relay. K1, located in the low-voltage power supply (not shown).

The operation of K1 provides the -24 volt after time delay, which controls the operation of the MVPS, HVPS, and HLRM. The latter two units are found in the AN/SRT-15 and 16 only. The standby relay, K1, is unable to follow the keying signals, making it necessary to keep the relay energized at all time during hand-key operation. In other modes of transmission, the transmitter cannot go on the air unless K6 is energized, supplying the ground to operate K1. In machine-key, FSK, and facsimile transmission, a ground is supplied over the keying line from the remote radiophone unit that

energizes K6 and provides a standby operate control for K1. In phone operation, K6 is operated only when the press-to-talk button is depressed, either locally or remotely. Therefore, the output circuits of the power supplies controled by K1, which, in turn, is controlled by K6, are completed in phone operation only when the press-to-talk button is depressed.

Buffer

The purpose of cathode follower V11A (fig. 6-10) is to isolate the action of the keying input signals from the keying multivibrator, V13. During weak signal the cathode of V11A is power biased to +40 volts by the voltage divider connected from the regulated +105 volts to the regulated -105 volts. The keying voltages for the different modes of operation are applied at the grid of V11A, and are reproduced at the cathode. When the negative input exceeds approximately -45 volts, V11A will cut off, with the result that the maximum amplitude of a negative keying signal appearing at the cathode is about 40 volts.

Keying Multivibrator

The keying control tube, V12, and keying multivibrator, V13A, receive the output of the cathode follower. In the space condition V13B conducts. The voltage developed across R 48 is sufficient to cut off V13A even though a small positive voltage is developed between the V13A grid and ground. The keying control tube, V12B, ensures that V13B conducts and V13A is cut off during space. The plate of V12B is connected in the grid circuit of V13A. If V12B were not included in the circuit the junction of the resistors R46 and R47 would be about +5 volts during space. The cathode of V12B during neutral keying is tied to the cathode of the buffer, V11A, through contacts 2 and 1 of the neutral polar switch, S8, in the NEUTRAL position. During space this point is at either +4 volts for neutral machine keying or -40 volts for hand keying. As these values are at or below the normal +5 volts to which the plate of V12B is tied, V12B conducts, effectively tying the junction of R46 and R47 to the space voltage appearing at the buffer cathode. Therefore, the grid circuit of V13A during space is at a lower (less positive) potential than the grid of V13B, causing the desired condition of having V13A cut off and V13B conducting. When a spacing signal is applied the same action occurs in polar keying, except that the keying control, V12B, is connected

so that its cathode goes to about +4 volts when the grid of the buffer, V11A, is at -30 volts. This action is accomplished by connecting the cathode of V12B to a tap on potentiometer R 44 through contacts 2 and 3 of S8 in the POLAR position.

The polar space control R44, is a part of a voltage divider consisting of R43, R44, and R45, which is connected between the +105 volt and -105 volt supplies. The tap of the potentiometer is tied to the cathode of the buffer. In space, with a -30-volt keying signal on the grid of the buffer its cathode will be a potential of about -27 volts. The setting of R44 will be such that the tap will be about +4 volts. With the multivibrator in the space condition, the plate voltage of V13A will be approximately +235 volts, and the plate of V13B approximately +100 volts.

As the keying signal changes from space to mark, the potential of the buffer cathode starts to rise. As this potential reaches +5 volts in neutral keying or -27 volts in polar keying, the keying control tube, V12B, no longer conducts and has no effect on the grid of V13A. The potential at the latter grid rises as a result of this action, but does not rise sufficiently high to overcome the common cathode bias developed across R48. As the mark potential rises higher keying control tube, V12A, comes into action. The cathode of V12A is tied to the grid of V13B. The plate of V12A is connected to the tap of potentiometer R45, the cathode resistor of the buffer circuit. One end of the buffer cathode resistor is tied to the -105 volt source.

The potentiometer, R45, is known as the neutral-polar-mark control. In order for V12A to conduct, its plate potential must be higher than that at the grid of V13A. As the mark voltage increases at the cathode of the buffer, the voltage of the tap of R45 will become higher than the grid of V13A, allowing V12A to conduct. When V12A conducts, it effectively makes the potential at the grid of V13A the same as that developed at the tap of R45. The tap is set so that as the mark signal reaches +30 volts at the grid of the buffer, the voltage at the tap, and also at the grid of V13A will be sufficient to overcome the bias potential developed across the common cathode resistor, causing V13A to conduct. The tube remains in conduction throughout the marking impulse. Therefore, in the mark condition, the plate of V13A is at a potential of approximately 70 volts, while the plate of V13B has a potential of about +235 volts.

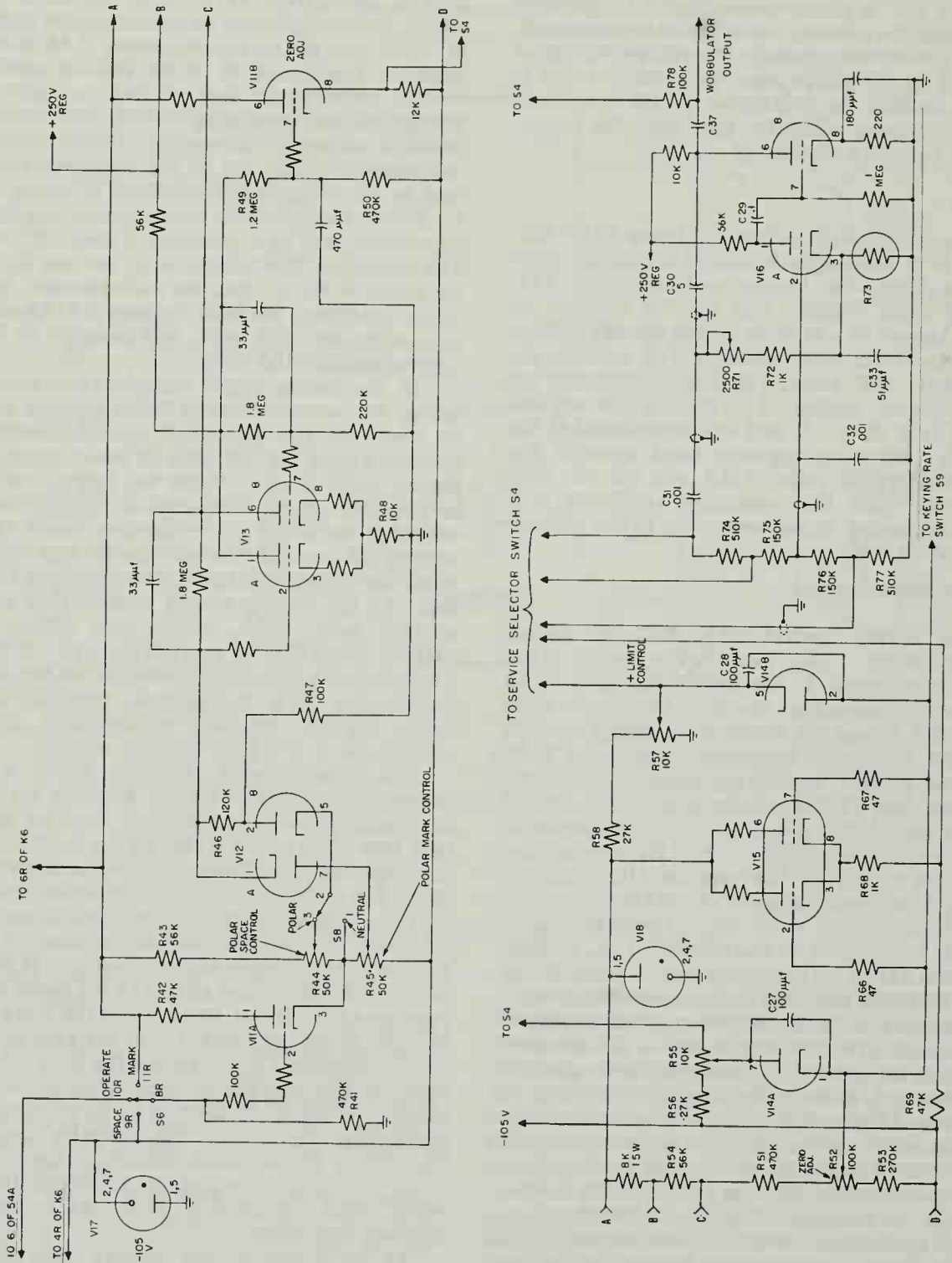As the keying voltage changes from mark to space, the potential of the grid of V13A must be

Figure 6-10.—Low-level radio modulator keying circuit, simplified schematic.

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lowered enough to cause a reduction in the tubes current flow, which will initiate the cumulative flip-flop action. During the mark condition, the junction of R46 and R47 (to which the plate of V12B is tied) is about +11 volts. Therefore, when the keying signal reduces to a point where the cathode of V12B (which is tied to the cathode of the buffer in neutral keying and to the polar space control in polar keying) goes below +11 volts, V12B conducts, lowering the grid potential of V13A. In neutral keying when the keying signal has reduced to about +7 volts, V12B will conduct a sufficient amount to lower the V13A grid below the saturation point, and the flip-flop action takes place. In polar keying the polar space control, R44, is set so that when the negative or space keying signal reaches about -28 volts, the keying control, V12B, conducts enough to lower the grid potential of V13A below the saturation value, and the multivibrator returns to the space condition.

The keying multivibrator, V13, has two outputs. One of the output voltages is developed in the plate circuit of V13A, and the other, which is out of phase with reference to the first, is developed in the plate circuit of V13B. The keying voltage developed in the plate of V13A is fed to the dumping cathode follower, V11B, for use in the push-pull cathode followers of the audio amplifier chain of the LLRM. A keying signal of zero volts for space and -65 volts for a mark is delivered to V11B from a voltage divider in the plate circuit of V13A. The dumping cathode follower provides a low impedance to the input of the push-pull cathode followers. The time constant in the output circuit of the dumping cathode follower is such that the waveshape is not destroyed. The output is taken directly from the cathode of V11B. The output signals are connected to the push-pull cathode followers through the service selector switch, S4 in the HAND and MACHINE positions. In frequency-shift keying and fascimile operation, the transmitter is functioning continuously, and no load compensation is required. In phone operation, load compensation is not required since the +500-volt supply is turned off when the press-to-talk button is up by the action of the keying relay, K6.

The keying signal from V13B, which is +25 volts for mark and -30 volts for space, is delivered to the limiter, V14A and V14B. This range of voltage is obtained from a voltage divider connected from the plate of V13B, through R51, zero adjust control R52, and R53, to the regulated -105 volts. The keying signal is developed across a value of capacitance that is selected by the keying rate switch, S9. The time constant involved in charging and discharging this capacitance (not shown) gives the required rise and decay time to the keying signal wave-shape. In HAND and MACH positions of the service selector switch, a second set of wave-shaping capacitors are connected in parallel with the first set.

Limiter

The plate of the limiter, V14A, is tied to the tap of a potentiometer, R55, known as the limit keyer out control. Potentiometer R55 is adjusted for -10 volts at the tap. Therefore, any negative (space) keying signals in FSK operation are clipped at a maximum of 10 volts by the action of the limiter diode, V14A. In other modes of operation the ground is removed from R55 by the service selector switch, S4, and the plate of V14A is tied to the -105 volts.

In hand-key, machine-key, and phone operation, the cathode of the limiter, V14B, is grounded through service selector switch, S4 for hand, mach, and phone. In these positions of the switch, potentiometer R57 in the cathode of V14B is shorted through the switch contacts. In FSK operation, the short on the cathode of V14B is removed. The cathode is then connected to the tap of the potentiometer R57, which is part of a voltage divider from +105 volts to ground Potentiometer R57 is the (+) limit control, and is adjusted for 10 volts at the tap. In this condition, V14B clips positive (mark) keying signals at +10 volts. Therefore, positive (mark) signals in hand-key, machine-key, and phone operation are clipped at zero, whereas in FSK operation, they are clipped at +10 volts.

Keyer

The keying signals, as supplied by the limiters, are fed to the grids of the keyer V15. This stage utilizes a twin triode operating in parallel. When a mark keying signal (zero volts) is applied to the grid of the keyer, the cathode is driven to about 2 volts positive. Since the stage is operated class A, the mark signal as applied to the buffer of the RFA should not go above zero volts. For this reason the output is taken at the junction of R68 and R69 to reduce the mark voltage from +2 volts to zero.

The output of zero volts for mark and -30 volts for space is the same in hand key, machine key, and phone operation. These signals are used to cut off the buffer in the RFA and the output stage of the RFO in the KEY UP (space) condition for the two keying modes or the press-to-talk button UP in phone operation. Both the buffer in the RFA and the output of the RFO are keyed off to assure that the key-up radiation from the transmitter group is below the minimum allowable level of 400 picowatts

In FSK operation the output for the keyer, which is+10 volts for mark and -10 volts for space, is fed to the radio frequency oscillator to frequency modulate the frequency-shift oscillator in the RFO. Contacts 4R and 6R (of K6, fig. 6-9 control the keying circuits that supply the FSK output to the FSO in the RFO. The signals are fed from the junction of R68 and R69 (fig. 6-10) to the service selector switch S4. Other positions of S4 are grounded since the frequency-shift oscillator is not used in hand-key, machine-key, or phone operation.

Audio Oscillator

A Wien bridge circuit, V16 (fig. 6-10), functions as the audio oscillator and provides two outputs. In FSK operation the output is 200-cycle signal whose purpose is to phase modulate the frequency-shift oscillator in the RFO whenever it is necessary in order to prevent the effects of selective fading. The amplitude of the 200-cycle signal is sufficient to produce a one-radian phase shift in the carrier. This amount of shift can be varied between zero and one radian by a PHASE MOD control (not shown). In hand-key and machine operation, the audio oscillator produces a 1000-cycle signal during the key down (mark) condition that is fed to the sidetone amplifier in the audio chain for monitor purposes.

With the service selector switch, S4, in the HAND or MACH position, R74 is shorted out and R76 is grounded through contacts of S4. Therefore, in hand-key or machine-key operation, the reactive half of the bridge is composed of C31, R75, R76, and C32.

Since R73 is the cathode resistor of V16A, a degenerative feedback voltage is obtained across R73 in the resistive half of the bridge, A regenerative feedback signal obtained in the reactive half of the bridge is applied to the grid of V16A. At a frequency determined by the components of the reactive half of the bridge the regenerative voltage will be in phase with the signal at the grid of V16A. This voltage will be equal to or slightly larger than the degenerative voltage.

The relative amplitude between the two voltages is controlled by the setting of the AUDIO OSC FEEDBACK control R71. The resonant frequency of the circuit, when all the components of the reactive half are used, is 200 cycles. This frequency is used during FSK operation. In hand-key and machine-key operation, with R74 and R77 removed by the action of S4, the resonant frequency becomes 1,000 cycles. In facsimile and phone operation the regenerative feedback in the reactive half of the bridge is grounded through contacts of S4, which disables the circuit.

In the cathode circuit of V16A, R73 (a ballast lamp) is used instead of a resistor to stabilize the amplitude of the oscillation. If the amplitude tends to increase, the current through the lamp also increases. The increase in temperature of the lamp filaments cuases the lamp resistance to increase, thereby increasing the amount of degenerative feedback. The gain of V16A is reduced in such a manner as to maintain the output voltage at a nearly constant amplitude. If the amplitude of the voltage is kept small, the oscillator tube will operate in the linear portion of its characteristic curve, producing a sinusoidal output.

In FSK operation the 200-cycle signal is fed to a FSK PHASE MOD control (not shown). The control contains an on-off switch that controls the application of the 200-cycle signal. The voltage developed across this potentiometer may be adjusted to produce a one-radian phase shift in the output of the frequency-shift oscillator, varying around the fundamental output frequency of 200 cps. The amount of shift introduced into the frequency-shift oscillator is multiplied throughout the remaining units of the RFO. The resulting wobbulating action presented by the 200-cycle shifting causes the transmitter output frequency to shift accordingly.

In hand-key or machine operation, the 1000-cycle output is coupled through resistor R78 and contacts of the service selector switch S4 to the grid of the sidetone amplifier in the audio chain. In hand-key operation a parallel connection, through contacts of S4 is made to keying relay, K6. In space, K6 is not energized, and the 1000-cycle output is grounded. In mark, K6 is energized and the ground is removed, allowing the signal to be applied to the grid of the sidetone amplifier.

SPECIAL CIRCUITS AN/WRT-2

The circuit discussion of the AN/WRT-2 transmitter is limited to a treatment of non-conventional transmitter and antenna circuitry. The circuit operation of the MOPD was treated earlier in Chapter 5 of this training course.

MASTER OSCILLATOR

The circuit of the master oscillator is shown in figure 6-11. The circuit arrangement is that of a Colpitts oscillator.

The oscillator functions the same on all brands. The component changes necessary to produce the required output frequency at the oscillator is accomplished by switch S301. Because of doubling or quadrupling in subsequent stages the transmitter can be tuned over more than on band while S301 is in a given position.

All of the master oscillator frequency determining elements are mounted in a temperature-controlled oven (not shown). These elements consists of voltage-sensitive capacitor C387 (a variable crystal capacitor), L304, and all of the capacitors in the grid circuit of V301. Tuning within the selected band is accomplished by frequency Ⓓ capacitor C313. Variable capacitor C308 provides a tuning control for the high end of the frequency band during alignment. Variable coil L304 is used to set the low frequency end of the band.

A gear train (not shown) controls the oscillator tuning. The RANGE Ⓒ1 (fig. 5-10A in chapter 5) and FREQUENCY Ⓓ controls are the two driven tuning controls for the oscillator. The RANGE Ⓒ1 control selects the frequency band, and the FREQUENCY Ⓓ control tunes the oscillator to the frequency within the selected band.

Frequency Multiplier Circuits

The frequency range of each of the twelve combinations of master oscillator-frequency multiplier circuit arrangements, the position of S302, and the associated band is shown in the table in the middle of figure 6-11. For simplicity, only the master osciallator tanks circuits of bands 4, 7, and 10 are shown in the simplified schematic since the MO operates at the same frequency for each of these bands. Switch S302 selects the proper arrangement of tuned circuits in the multiplier stages to produce the desired transmitter output frequency.

The master oscillator signal is coupled to the multiplier stages (or PA) via T301. Note that on bands 1 through 6, switch section S302C applies a ground to terminal 4 of T301 and the oscillator input is fed from terminal 3 of T301 through any one of the contacts 1 through 6 of S302G (depending on the selected band) to the input mixer stage in the power amplifier and to the master oscillator-phase detector.

On bands 7 through 9 (S302 positions 10 through 12), the ground is removed from terminal 4 of T301 and applied to the center-tap of R308. This action permits the oscillator input voltage to be developed in the 5-6 and 7-8 windings, respectively, of T301, and applied to the grid of first doubler V303. (Doubling takes place in the grid circuit of V303 due to the diodes CR301 and CR302.) Simultaneously, S302E applies a ground to terminal 4 of T302 at the first doubler output. Thus, a frequency output of V303 which is twice the oscillator frequency is fed from terminal 3 of T302 through the selected contact (10, 11, or 12) of S302G to the output circuit.

Finally, on bands 10 through 12 (S302 positions 16, 17, and 18), the ground is removed from terminal 4 of T302 and applied to the arm of R313. The R308 arm remains grounded by the action of S302C. This action permits the first doubler output to be again doubled in the grid circuit of the second doubler V304. The second doubler plate output is fed through S302F to the primary of T203. The output is fed from the T303 secondary through contacts 16 of S302 G, to the output circuit. Similar circuits (not shown) with separate output coupling transformers are used on bands 11 and 12.

Transformer T303, FREQUENCY Ⓓ tuning capacitor C341C and trimmer capacitor C376 tune the V304 doubler plate tank. Capacitor C341B tunes the first doubler plate tank with trimmer c361. Resistors R308 and R313 are used as balance potentiometers for their respective doublers. When these resistors are properly set, the associated crystal diodes conduct equally on peaks of the applied r-f voltage.

MODULATING SECTION

The modulating section accepts audio signals or d-c keying signals and suitably transforms them to modulate or control the r-f energy generated by the r-f generating section. The circuits of the modulating section are contained in two

POSITION OF S302	BAND	OSCILLATOR FREQ.-MULTIPLIER OUTPUT MC
1	1	1.5–2.0
2	2	2.0–2.75
3	3	2.75–3.75
4	4	3.75–5.0
5	5	5.0–625
6	6	6.25–7.5
10	7	7.5–10
11	8	10–12.5
12	9	12.5–15
16	10	15–20
17	11	20–25
18	12	25–30
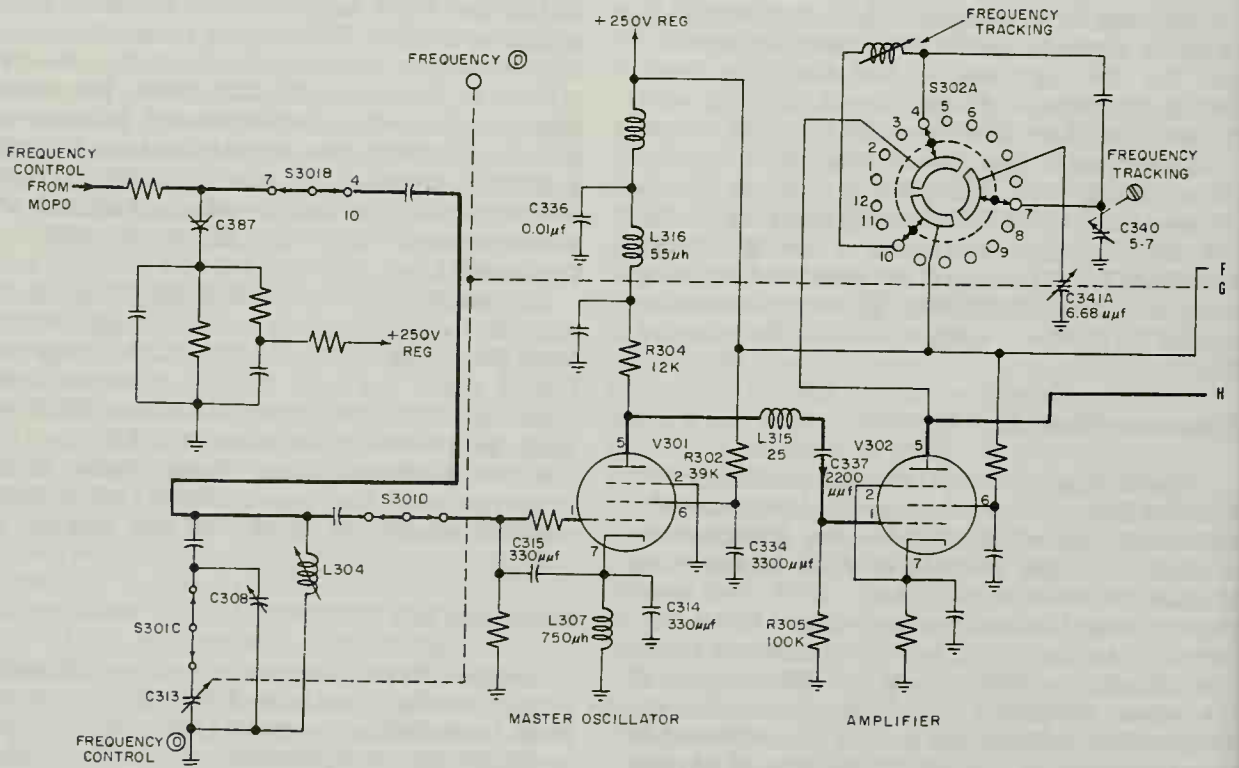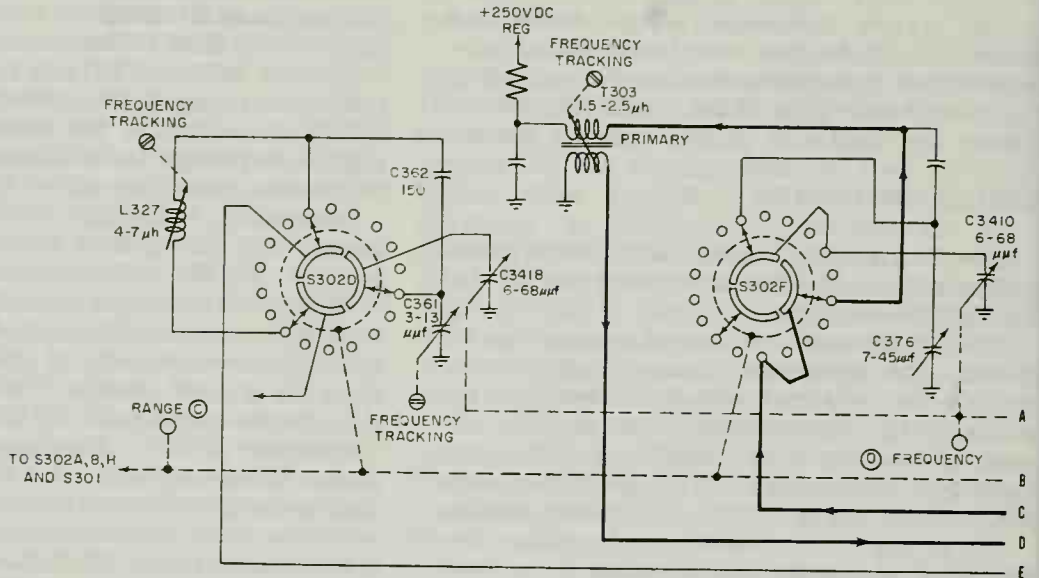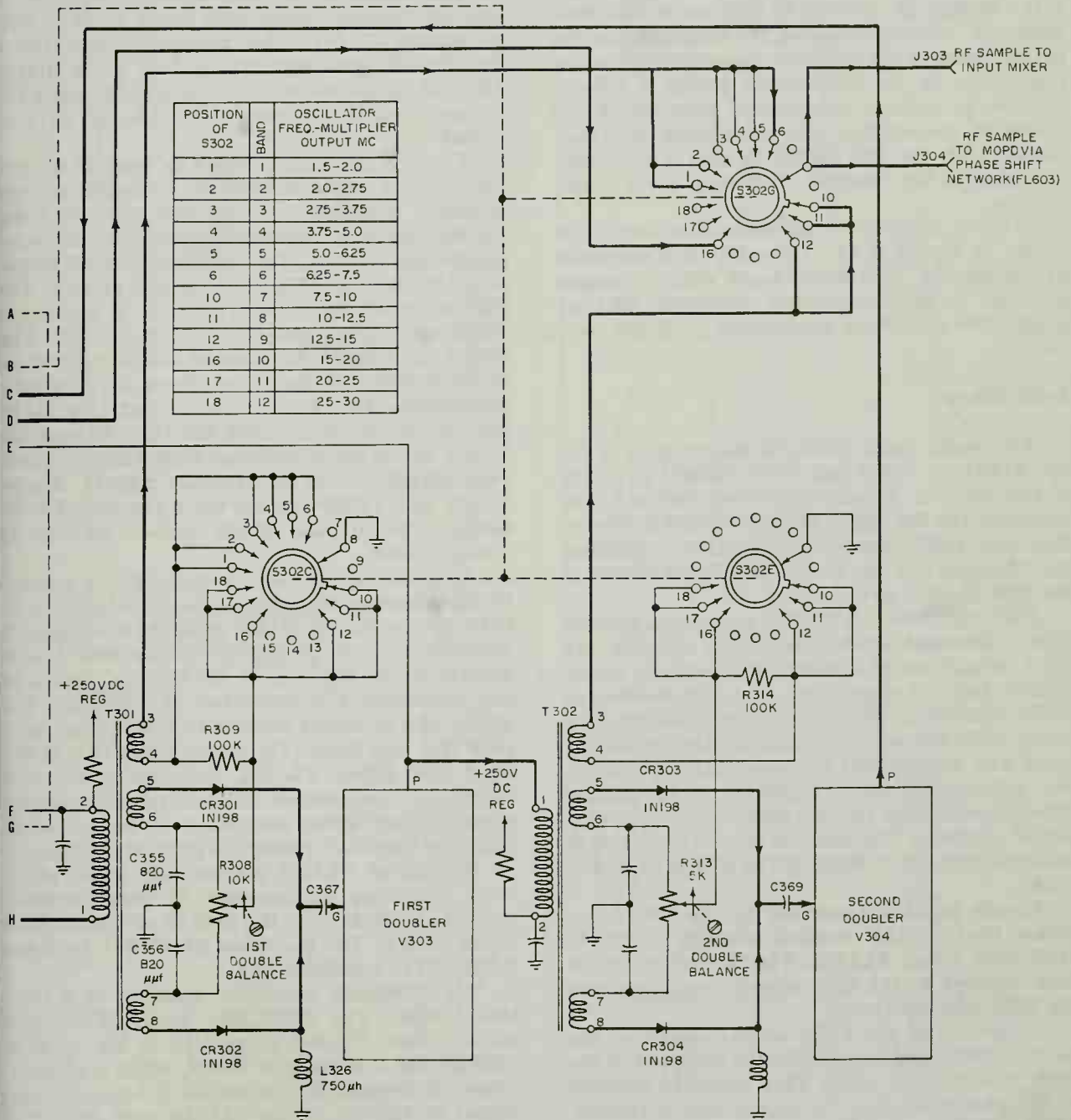
32.285

Figure 6-11.—Master oscillator-frequency multiplier, simplified schematic.

units of the transmitter: the radio frequency oscillator and the amplifier-power supply.

The filter method of generating single sidebands is used in the modulator. With this method, sideband operation must be accomplished at a low frequency level since accurate filter design is difficult to achieve at high frequencies. After generation, the sidebands may be heterodyned to the desired transmitter operating frequency. In the transmitter group, a 2.0-mc carrier is applied to balanced modulators for sideband generation. The sidebands are then mixed with the MO signal to obtain sum frequencies at the desired frequency level of operation.

A block diagram of the modulation section is shown in figure 6-12. Three types of emission are available with audio input signals: single sideband (A3a), independent sideband (A3b), or compatible amplitude modulation (USB and carrier).

Audio Stage

The audio input from the microphone is fed via T1407 or T1408 and filter FL1401 to either of two idential speech amplifier channels. The channels are the upper sideband (USB) and lower sideband (LSB) channels, respectively. Because the channels are the same, only those blocks in the USB channel are shown.

For amplitude modulation operation, only the upper sideband speech amplifier circuits are used and only one sideband plus the carrier (compatible AM) is transmitted. For single sideband (SSB) operation either the upper sideband or lower sideband speech amplifier circuits may be used. For independent sideband (ISB) operation, two independent audio signals may be applied simultaneously to the two separate speech amplifier circuits. The desired type of operation is selected through EMISSION SELECTOR switch S508.

Speech signals generated by the USB telephone handset are coupled through T1407 to band-pass filter FL1401. The filtered signal is then applied to the USB speech amplifier and the USB-AGC control.

A portion of the audio signal output of the second USB speech amplifier is diverted to the USB sidetone amplifier. This signal is fed back to the phone handset for monitoring purposes. TEST TONE switch S511 substitutes the output of the 425-cycle test oscillator, Q1411 and Q1412, for that of the audio when a single-tone

test signal is desired. Meter amplifier Q501 and Mod level meter M502 provide a visual indication of the speech amplifier output levels. Meter M502 is connected in the output circuit by MOD LEVEL switch S512.

The output of the USB speech amplifier is applied through switch S508 to the grids of USB modulator V1201. The modulator contains a twin-triode tube with its control grids driven 180° out of phase by an audio signal, while the common cathodes receive a 2.0-mc carrier signal.

The 2.0-mc signal which is applied to both the carrier reinsert cathode followers and the sideband modulators is derived from the 1-mc crystal-stabilized reference oscillator in the r-f generating section. The oscillator output is applied to the 10:1 frequency divider circuit. The 500-kc output from binary No. 1 is applied to 500-kc amplifier stage V1207A (fig. 6-12). The output of V1207A is coupled to two half-wave diode rectifier circuits (not shown) for harmonic generation. Tuned tanks at the control grids of the 2.0-mc selector V1208 and 1500-kc selector V1209 select the 2.0-mc and 1500-kc harmonics. respectively, from the rectifier outputs. Stages V1208 and V1209 amplify their respective harmonic. The purpose of the 1500-kc selector is treated later.

As stated above, the 2.0-mc input is applied to the common cathodes of the twin-triode USB balanced modulator V1201 while the audio speech amplifier output is applied to the twin triode control grids 180° out of phase. The plates of the modulator are connected in push-pull. The action of a balanced modulator cancels the carrier (2.0 mc) input. The signal appearing at the input USB filter, FL1201, consists of sum and difference frequencies and a carrier frequency signal of negligible magnitude if the push-pull balanced modulator stage is properly balanced.

USB filter FL1201 allows only upper sideband frequency components to pass (without severe attenuation) to the grid of sideband combiner V1203. The bandpass of Fl1201 is from 2,000,300 to 2,008,000 cps.

The sideband combiner consists of a twin triode tube. The 2,000,300- to 2,008,000-cps output from FL1201 is applied to the control grid of the A section of V1203, while a similar range of frequencies below the 2.0-mc carrier signal is applied to the V1203B grid from the LSB filter FL1202 (not shown). The plates of V1203A and V1203B have separate plate loads in the carrier reject filter FL1203. The filter
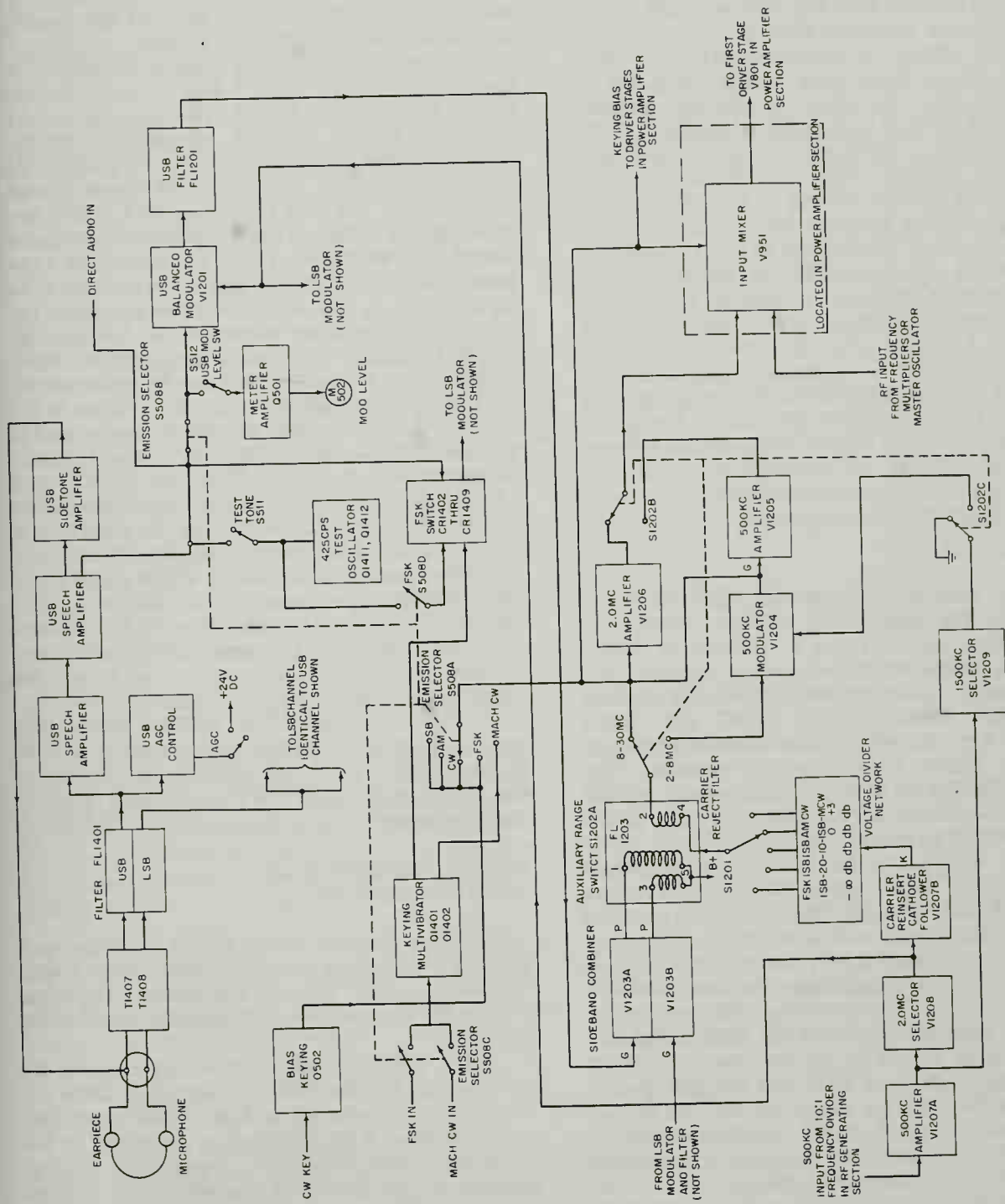
32.286

Figure 6-12.—Modulating section, block diagram.

(filter components not shown) rejects the 2.0 mc carrier at the V12-3 output. The common secondary coil connected betwjen terminals 2 and 4 of FL1203 receive both the USB and LSB signal components as separate signals. These outputs are fed to either the 2.0-mc amplifier, V1206, or to the 500-kc modulator V1204, depending on the transmitter operating frequency, and the position of auxiliary range switch section S1202A.

The 2.0-mc carrier signal from V1208 is also applied to carrier reinsert cathode follower V1207B. The cathode circuit of V1207B contains a voltage divider network from which various voltage levels of the 2.0-mc carrier signal can be selected by CARRIER REINSERT switch S1201. Varying degrees of carrier reinsertion are required to facilitate signal recovery at the receiver. A higher level of carrier reinsertion is possible during c-w operation than during AM operation because the peaks of the speech signals limit the amount of drive that can be applied to the power amplifier stages. Attenuation settings of $-\infty$, -20, -10, 0, and $+3$ db are provided. For AM, MACH c-w, and c-w operation it is necessary to transmit a carrier signal.

Keying Circuits

Three classes of d-c keying signals are accepted by the modulating section: frequency-shift neutral keying from a teletype or multiplex unit, machine or c-w keying signals from an automatic coder, or hand-keyed c-w signals.

In FSK operation, the S508C and S508D contacts are closed. Switch section S508B is open and S508A is in the FSK position. The FSK or MACH c-w input signals are applied through S508C to a keying multivibrator, Q1401 and Q1402. The output of the 425-cps test oscillator Q1411 and Q1412 is applied through S508D to the FSK switch. This circuit comprises CR1402 through CR1409 and is a symmetrical electronic switch which is driven by keying multivibrator Q1401 and Q1402.

When a space signal is on the line, the keying multivibrator causes the FSK switch to route the 425-cps signal to the LSB modulator. When a mark signal is on the line, the keying multivibrator changes state and the FSK switch transfers the 425-cps signal to the USB modulator. Thus, the mark signal is 425 cycles above the carrier frequency and the space signal is 425 cycles below the carrier frequency.

During c-w operation, the c-w keys actuates a bias keying stage, Q502, the output of which is applied through S508A to the grid of the 2.0-mc amplifier, V1206, to the grid of the 500-kc amplifier, V1205, and to the grids of the mixer and driver stages of the radio frequency amplifier (shown later). This action cuts off the transmitter carrier when the key is open.

As stated earlier, the master oscillator in the RFO operates from 1.5 to 7.5 mc. The use of the fundamental MO frequencies for band 1 through 6 (2.0 to 8.0 mc) necessitates the inclusion of the 1500-kc selector, V1209 and the 500-kc modulator, V1204. On the lowest transmitter rane, the MO frequency is 1.5 mc. Because of the action of S302 the master oscillator output on the lowest range bypasses the frequency multiplier stages and is fed directly to an input mixer. A 500-kc signal must also be introduced into the input mixer if the transmitter output frequency is to be 2 megacycles, which is the lowest frequency that can be transmitted.

When the transmitter output is to be in the 2.0- to 8.0-mc range, auxiliary range switch sections S1202A and S1202C apply the sideband combiner output (via FL1203) and the output of 1500-kc selector V1209 to the 500-kc modulator V1204. The 500-kc modulator is identical in operation to the USB modulator described earlier. The push-pull output of V1204 suppresses the 1500-kc signal and a tuned tank in the grid circuit of the 500-kc amplifier V1205 selects only the difference-frequency components generated by the modulator, plus or minus the modulation (intelligence) contained in the original 2.0-mc signal. The 500-kc signal then passes through auxiliary range switch S1202B to the mixer in the PA section and serves as the injection frequency to the PA mixer throughout the frequency coverage of bands 1 through 6.

POWER AMPLIFIER AND
ANTENNA SECTION

A block diagram of the power amplifier and antenna sections is shown in figure 6-13. These sections include driver and power amplifier stages, tuning and coupling circuits, and the r-f monitor.

After mixing in the input mixer stage of the modulating section, the r-f signal contains the modulation component and is in the frequency range from 2.0 to 30.0 mc as selected by the operator. The output of the mixer is amplified in the first and second driver stages, V801 and V802, in the power amplifier section. The smplified r-f is then applied to the power amplifier stages, V804 through V807.
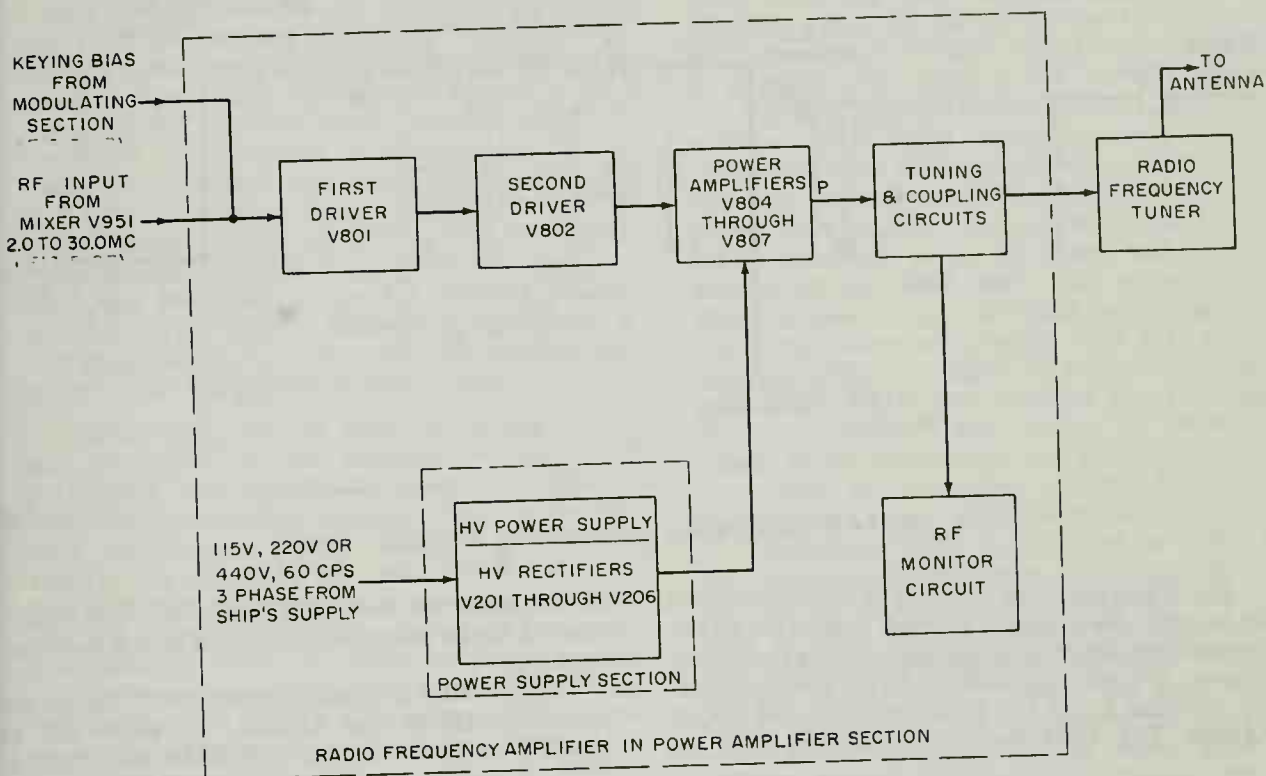
32.288

Figure 6-13.—Power amplifier section, block diagram.

The power amplifier stages consist of four 4CX300 tetrode tubes connected in parallel (not shown). Plate voltage for these tubes is supplied by high voltage rectifiers V201 through V206. The power amplifier output is fed through the tuning and coupling circuits and the radio frequency tuner to the antenna.

A portion of the output from the tuning circuit is fed to the r-f monitor circuit, which consists of a modulation monitor for measuring the modulation percentage, and a reflectometer for measuring the output power level and voltage standing wave ratio on the line which feeds the antenna. A circuit analysis of the r-f monitor is treated later.

Because the circuits of the driver and power amplifier stages are conventional in most respects, the circuit analysis of these stages is not treated in this discussion. The antenna tuning and coupling circuits used in the AN/WRT-2 are not conventional and are therefore treated with the aid of schematic diagrams.

Tuning and Coupling Circuits

The plate output of the power amplifier tubes is fed into the tuning and coupling circuits by C827 (fig. 6-14). The resonant frequency of tuning components L803 through L805 and C836 and C837 is adjusted by a PA TUNING Ⓖ control and PA COUPLING Ⓗ control located on the radio frequency amplifier front panel. These components form a pi network low-pass filter at the operating frequency. The PA tuning control adjusts all five tuning components simultaneously while the Ⓗ control adjust only C837 and L805.

R-F Monitor

As forestated, the r-f monitoring circuit contains a modulation monitor circuit which indicates percentage of modulation, and a reflectometer circuit which measures the transmitter power output and the VSWR on the line of the antenna. Figure 6-15 is simplified schematic of the r-f monitor circuit.

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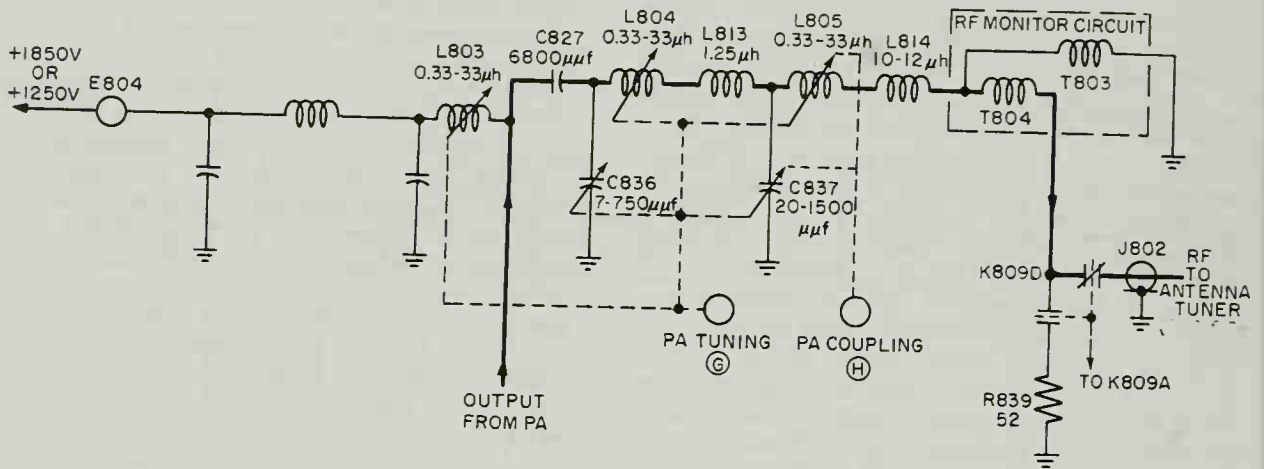
Figure 6-14.—Tuning and coupling circuits.

32.289

The T803 primary winding is connected from the antenna feed line to ground. A portion of the transmitter output is developed across the transformer. The voltage induced in the T803 secondary winding is of the same phase as the output voltage. The T804 secondary winding is of the same phase as the output current. The vector sum of these voltages is applied to the detector circuits.

When READ SWR switch S807 is in its closed position (as shown), 17 volts AC is applied to the cathodes of CR806 and CR812,
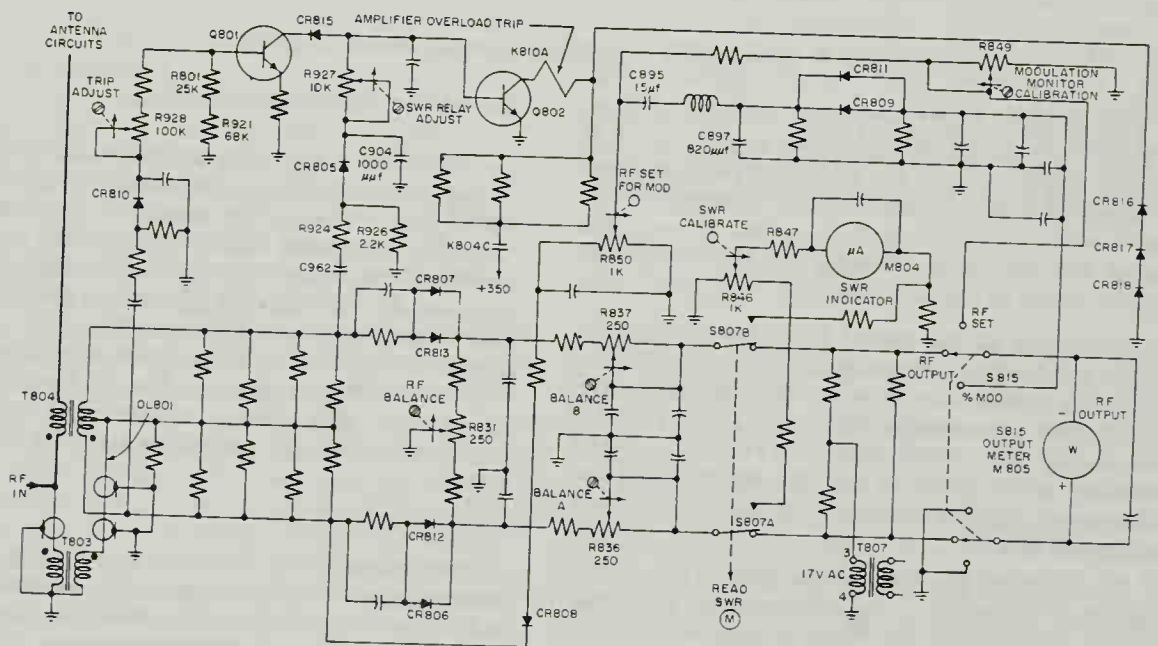
Figure 6-15.—R-f monitoring circuit, simplified schematic diagram.

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and CR807 and CR813. The 17 volts AC places a bias on these diodes so that they operate as square-law detectors, their d-c output voltage being proportional to the square of the applied voltage. The output of one set of diodes is proportional to incident power and the output of the other set of diodes is proportional to reflected power due to a mismatch of impedances. R-F OUTPUT meter M805, connected to the two output lines through OUTPUT METER switch S815, measures the difference between the incident and reflected power or the true power. BALANCE A control R836 and BALANCE B control R837 provide for zero adjustment and upper scale calibration of meter M805.

READ SWR switch S807 is spring loaded and normally in the position shown. R-F BALANCE control R831 is used as a means of balancing conduction irregularities in the two sets of diodes, CR806, CR812, and CR807, CR813. To measure the standing wave ratio, the switch is pushed to its momentary position, and the power monitoring circuit is removed. The removal of the 17 volts AC from the sets of diodes permit the diodes to become linear detectors. SWR INDICATOR meter M804 is connected into the monitor circuit when switch S807 is closed to its upper contacts and the SWR is indicated.

For monitoring modulation, an r-f sample of the transmitter output is obtained from the bottom of T804 and applied to CR808. A negative DC voltage is developed across R-F SET FOR MOD control R850. When OUTPUT METER switch S815 is in R-F SET position a portion of the d-c voltage across R850 is applied to OUTPUT METER M805, and R850 is adjusted so that M805 deflects to its R-F SET marker. When switch S815 is in %MOD position, a pulsating voltage is applied from the arm of R850 through C895. The d-c voltage from R850 is rectified by parallel diodes CR80'} and CR811 and filtered to remove the r-f and permit an a-f signal to be applied to M805.

When a known modulation % is applied to a carrier, OUTPUT METER switch S815 is placed in R-F SET position and MOD MON CAL potentiometer R849 is adjusted to indicate the known modulation %. Thereafter, when switch S815 is placed in % MOD position, and the transmitter emission is of the A3 type, M805 indicates percentage of modulation.

To avoid possible equipment damage, SWR protection stages (Q801 and Q802) are provided to remove the high voltage from the PA stage when the SWR becomes excessive. When no SWR exists on the feeder line to the antenna the base voltage of Q802 (developed across R926 in parallel with Q801) is 180° out of phase with the voltage applied to the base of Q801. This voltage relationship exists because the base signal input for Q801 and Q802 is obtained from opposite ends of the T804 secondary. The 180° phase shift in Q801 causes the signal from the Q801 collector to be in phase with the R926 voltage at the Q802 base. The voltage across R926 increases as the SWR increases. When the ratio becomes greater than 4:1, the collector current of Q802 is sufficient to energize relay K810; and initiates the removal of all d-c voltages. The setting of SWR relay adjust R927 determines the amount of voltage that must be developed across R926 before the overload relay, K810 energizes.

Transistor Q801 is provided in order that the overload circuit may still function properly at low power levels. The collector resistance of Q801 is a function of the base-emitter bias developed across R801 and R921. When operating at low power levels, the base bias of Q801 decreases and therefore its collector resistance increases. At the same time the voltage across R926 has decreased, but the base-emitter bias for Q802 does not decrease substantially due to the increased collector resistance of Q801. Thus, an excessive SWR will still cause relay K810 to energize even when the power output level is reduced. The setting of SWR ALARM TRIP ADJUST potentiometer R928 determines the amount of base-emitter bias applied to Q801.

Antenna Tuner

The radio frequency tuner increases or decreases the effective physical length of the antenna by removing or inserting inductance in series with the antenna. As shown in figure 6-16 drive motor B3301 and 2-speed drive MP-3301 drive a sliding short up and down the main coil L3302 which can be switched in series with the antenna via antenna transfer switch S3301. A pictorial illustration of L3302 is shown in figure 6-17. The arm of R3301 is geared to the mechanical drive mechanism and electrical contact is made to POSITION INDICATOR meter M806 on the front panel of the RFA. The meter indicates the relative position of the sliding short on the main coil.
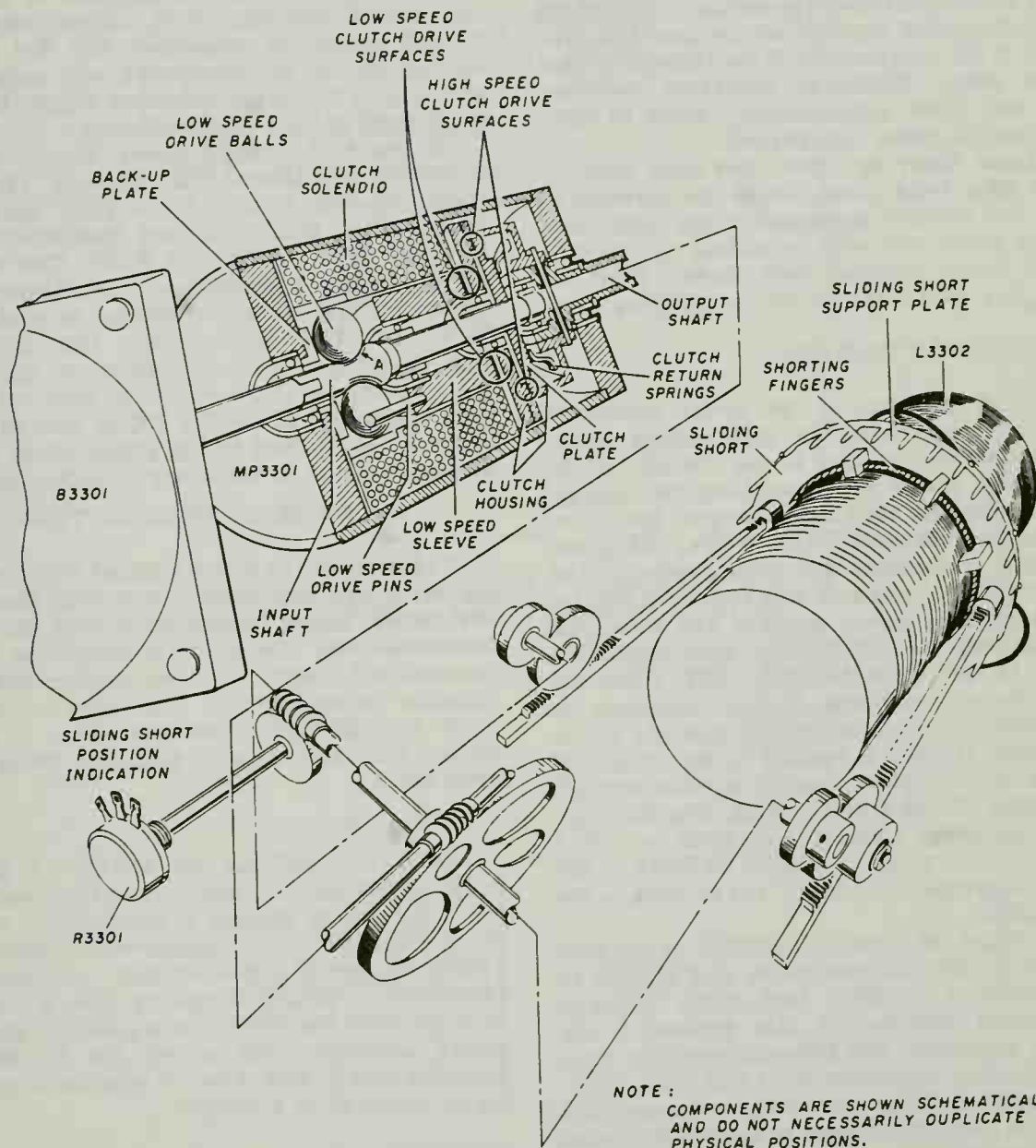
32.291

Figure 6-16.—Radio frequency tuner, schematic diagram.

Fixed capacitance may also be inserted in series or in parallel with the antenna for tuning purposes.

Switch S806 (COUPLER ANTENNA, not shown,) provides 115 volts AC for the actuator L3303 (fig. 6-16) when placed in the DIRECT,

A, or B position. These positions correspond to positions 1, 2, and 3 on S3301. A tuner control switch on the front panel applies 115 volts AC to the actuator when the bypass function is desired. In this position (as shown) relay K3302A is energized. When the tuner

LOW SPEED
CLUTCH DRIVE
SURFACES

HIGH SPEED
CLUTCH DRIVE
SURFACES

LOW SPEED
DRIVE BALLS

BACK-UP
PLATE

CLUTCH
SOLENDIO

OUTPUT
SHAFT

SLIDING SHORT
SUPPORT PLATE

L3302

CLUTCH
RETURN
SPRINGS

SHORTING
FINGERS

B3301

MP3301

CLUTCH
PLATE

SLIDING
SHORT

CLUTCH
HOUSING

LOW SPEED
SLEEVE

LOW SPEED
DRIVE PINS

INPUT
SHAFT

SLIDING SHORT
POSITION
INDICATION

R3301

NOTE:
COMPONENTS ARE SHOWN SCHEMATICALLY
AND DO NOT NECESSARILY DUPLICATE
PHYSICAL POSITIONS.

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Figure 6-17.—Antenna tuner coil, L3302.

components are to be used, the tuner control switch is placed in the TUNER-IN position, and K3302A deenergizes. This again permits the desired tuning components to be selected by S806.

Switches S3307 and S3301, driven by actuator L3303, place C3302 and C3303 in parallel with the antenna when switch S806 is in the B position. When S806 is in the A position, C3301 is placed in series with the antenna, and

when in the DIRECT position, only L3302 is in series with the antenna. The setting of COUPLER Ⓙ switch S806 depends on the operating frequency of the transmitter. If the transfer switch control relay, K3302, is energized, switches S3307 and S3301 automatically return to position 5 and the tuner is bypassed.

Antenna tuner up, down, and slow controls on the RFA front panel enable the operator to select the desired movement of the L3302 arm (sliding short) and also to control the speed of tuning. Limit switches (not shown) limit the upper and lower travel of the shorting arm.

POWER SUPPLY SECTION

A block diagram of the power section is shown in figure 6-18. The high voltage rectifiers supply d-c voltage to the PA tubes. The low voltage power supply supplies the remaining d-c voltages needed for proper operation, such as the +350-volts -350 volts, -24 volts, +12-volts, -12-volts, 250-volts regulated, +24 volts regulated, and a -6 volt regulated supply.

The primary power input of 115 volts, 220 volts, or 440 volts, depending upon the ship's supply, is applied to the T501, T201, T202, and T203 primary windings. Tapped terminals on T501 provide an output of 220 and 115 volts. The output of T501 is applied to the circuits of the low voltage power supply section through FILAMENT POWER switch S502. The 220 volts is applied to the primaries of T502 and T503 while the 115 volts is applied to T617 in the electrical frequency control and to T204 in the PA circuits.

The output of T502 is rectified by the plus and minus 350-volt rectifiers, and applied to the circuits of the RFA. Part of the output of the 350-volt rectifier is also applied to the 250-volt regulator. The 250-volt regulator consists of series regulators V321 and V322, regulator amplifier V323, and voltage reference tube V324. The regulated output is applied to stages in the electrical frequency control circuit and r-f oscillator.

The output of T503 is rectified by the -24-volt rectifiers, and fed to the various circuits throughout the transmitter which require this potential. A second output of T503 is applied to the 12-volt rectifier. The negative output of the 12-volt power supply is fed to the speech amplifier and control circuits. The positive output is fed only to control circuits in the amplifier-power supply and mike circuits.

The output of T617 is rectified by the 24-volt and 6 volt regulators, respectively. The 24-volt output is regulated and fed to the circuit of the r-f generating and modulating sections. The 6-volt regulated supply is used by circuits of the 1-mc oscillator.

In the high voltage power supply, bridge connected rectifiers V201 through V206 get their filament voltage from T204 and their plate voltage from delta-wye connected transformers T201, T202, and T203. The secondaries of the high voltage transformers are tapped so that a lower voltage is applied in TUNE and 100-watt operation than for 500-watt operation. The output from the high voltage rectifiers (1250 v DC in tune and 100-watt operation, or 1850 v DC in 500-watt operation) is applied to the plates of the power amplifier tubes in the power amplifier section.

SHORE BASED TRANSMITTERS

Transmitters at shore stations are basically the same as those used aboard ship. Some are low output power transmitters used for local communication. The greatest difference is the transmitters used for shore-to-ship and long distance communication. The power requirement for this type of communication is much higher than that aboard ship, 10 kw up to 2 megawatts.

AN/FRT-5/6

Radio Transmitting Set AN/FRT-5, figure 6-19, is designed for shore-based communication. The set is capable of providing an output of 15 kilowatts of continuous-wave radio-telegraph, frequency-shift-telegraph, or facsimile transmission on any frequency from 4.0 mc to 26.0 mc. With the addition of a power supply and power amplifier (not shown) the AN/FRT-5 becomes an AN/FRT-6 and is capable of supplying an output of 40 kilowatts.

General Description

The equipment is composed of the following four major units as shown in figure 6-19; RF unit T-226/FRT-5; Power Supply PP-490/FRT-5; Power Transformer TF-122/U and Power Control C-587/FRT-5.

Power Supply PP-490/FRT-5 and RF Unit T-225/FRT-5, the two main enclosures, are installed side-by-side in one cabinet and constitute the main portion of the transmitter.
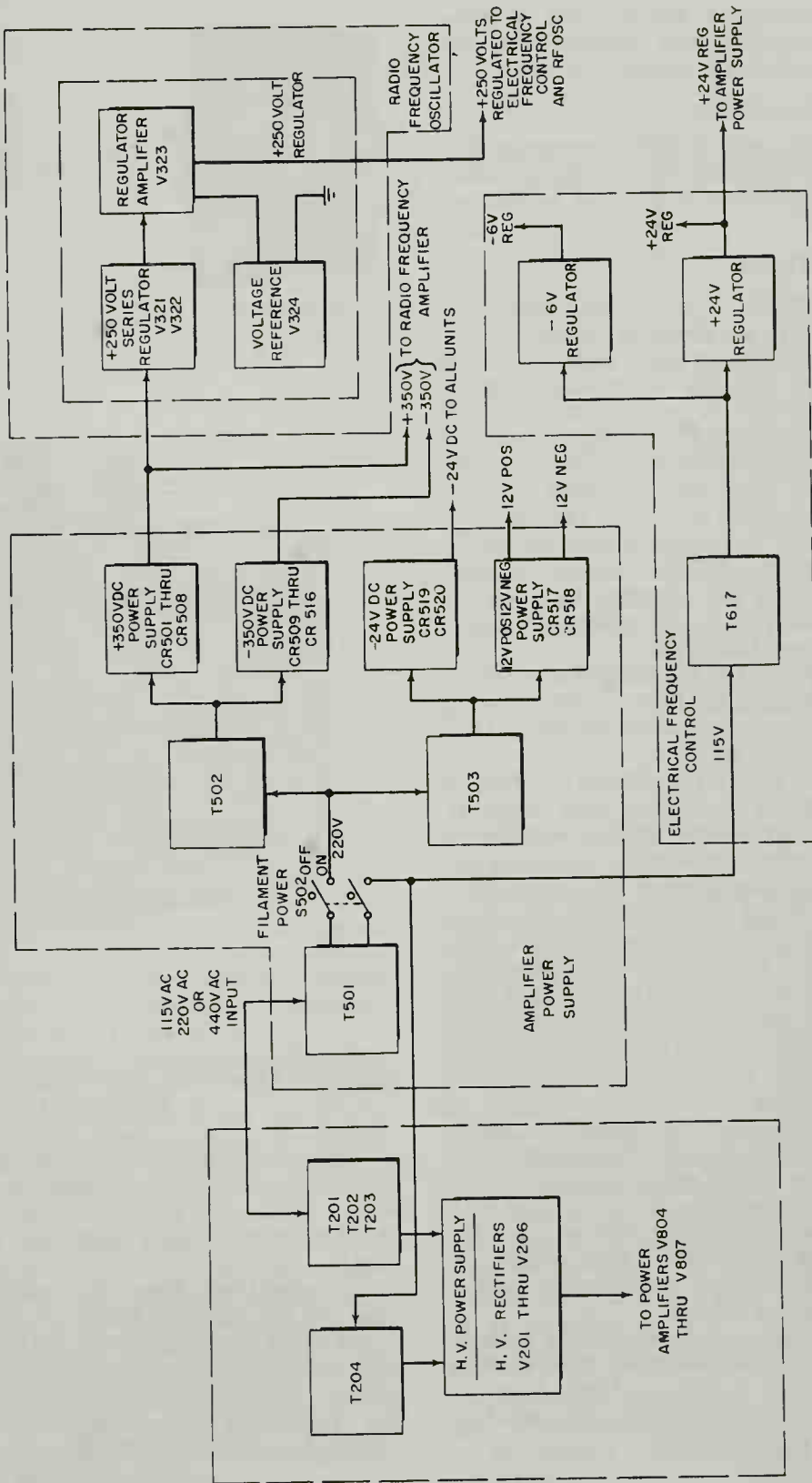
Figure 6-18.—Power supply section, block diagram.

32.293

The other two units, which might be considered as auxiliaries, are located near, but outside of, the main transmitter enclosure.

Functional Block Diagram

The major and minor units are shown in a functional block diagram (fig. 6-20); a brief description of each unit is given in the following paragraphs.

RF Unit T-225/FRT-5

RF Unit T-225/Frt-5 is the right-hand one of the two main bays, as shown in figure 6-19. It contains integral rack and panel type equipment and permanently mounted equipment as indicated by the dotted line in figure 6-20. The rack and panel type equipment is mounted in a standard equipment rack which is pivoted to provide easy access to the rear of all units mounted thereon.

The RF Oscillator O-91 ( )/FRT-5 is a stabilized variable frequency oscillator which derives its stability from a 100-kc crystal standard. Utilizing a permeability tuned oscillator, it provides a frequency stabilized output in the range of 2 mc to 4.5 mc.

RF Oscillator O-92 ( )/FRT-5 is a crystal-controlled 10-channel oscillator, With proper crystals, the oscillator provides an output frequency of 2 mc to 4.3 mc.

The Frequency Shift Keyer KY-45 ( )/FRT-5 is used to provide a frequency shift which is linear with respect to the keying voltage. It causes the transmitter to emit one frequency for a "mark" signal and another for a "space" signal. Because of its linearity with respect to the keying voltage or current, it may also be used for "photo" and teleprinter transmissions.

Power Supply PP-454 ( )/FRT-5 is a dual supply furnishing plate and filament voltages to RF Oscillator O-91 ( )/FRT-5 and Frequency Shift Keyer KY-45 ( )/FRT-5. It also supplies a regulated bias to the keyer.

There are six identical interchangeable, plug-in servo control amplifiers mounted in a common enclosure on the equipment rack in the r-f bay. One of these units is associated with each of the six tuned servo-positioned circuits. in RF Unit T-225 ( )/FRT-5. Each contains a voltage amplifier, a relay control tube, and a pair of motor-control relays.

The servo power supply furnishes voltage to all of the servo amplifiers, and drive motors located in RF Unit T-225 ( )/FRT-5. It is mounted on the equipment rack, immediately below the servo amplifier unit.
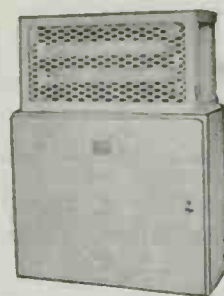
POWER CONTROL
C-587( )/FRT-5

TRANSFORMER
TF-122/U
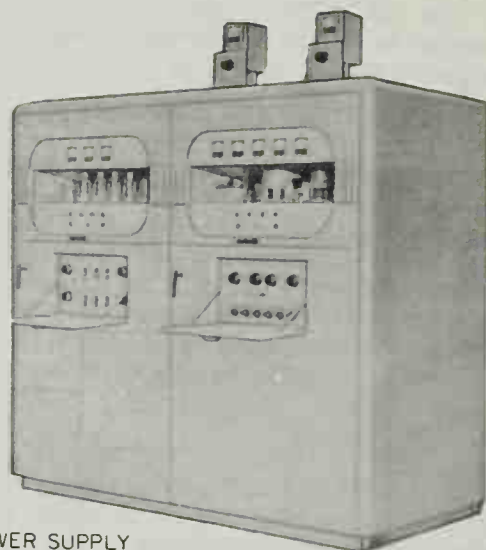
POWER SUPPLY
PP-490( )/FRT-5

RF UNIT
T-225( )/FRT-5

162.15
Figure 6-19.—Radio Transmitting Set AN/FRT-5, major units.

The remaining unit in the equipment rack is the patch panel which contains the two oscillator output jacks, the keyer input and output jacks, the transmitter input jack and several spare jacks.

An electronic keyer is mounted on the right side of the control panel of RF Unit T-225 ( )/FRT-5. It receives the keying impulses and accomplishes off-on keying of the transmitter proper. Provision is also made for keying of the transmitter by negative, positive, polar negative, or polar positive impulses for "mark".
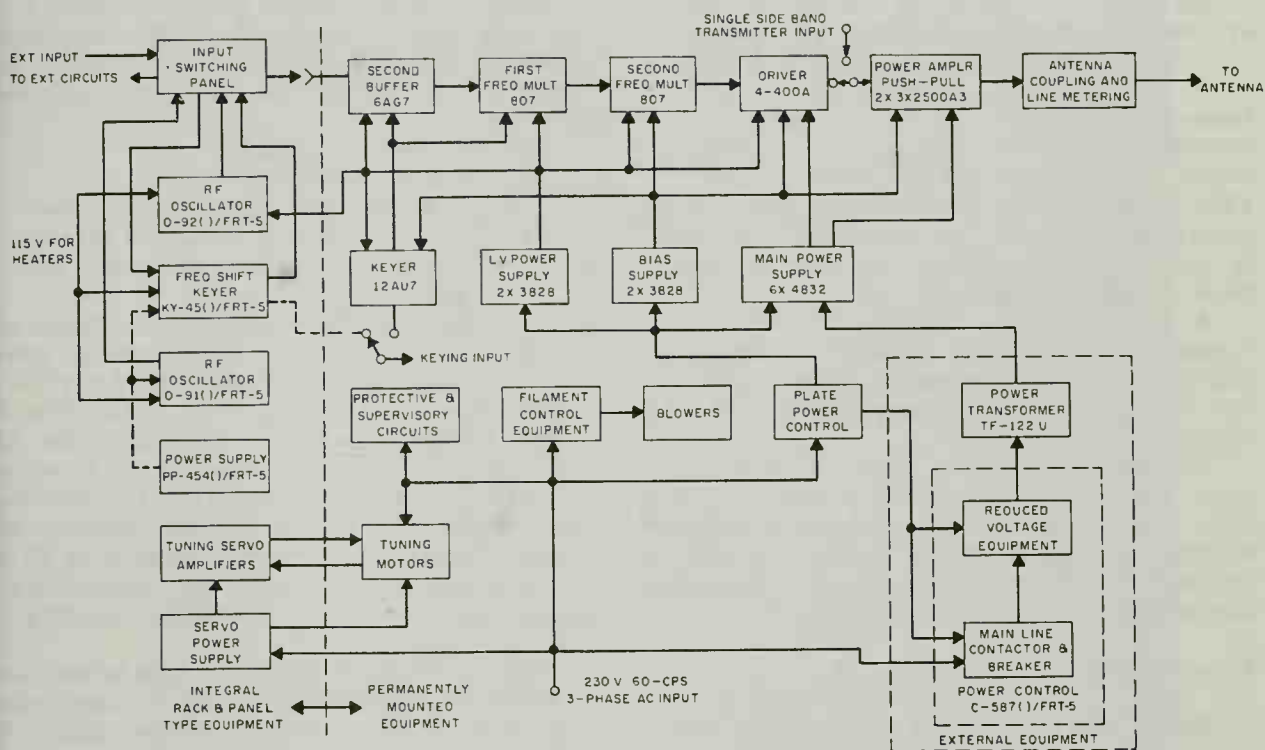
162.16

Figure 6-20.—Functional Block Diagram Radio Transmitting. Set AN/FRT-5.

The control panel carries the basic transmitter tuning and checking controls and is located behind the lower front door. Easy access to these controls is provided by a small drop door located at the top center of the main lower door as shown in figure 6-19.

The buffer amplifier, frequency multipliers, and driver unit are contained within a subassembly which is mounted to the left of the control panel. Inside are located a broad-band buffer-amplifier and two frequency-multiplier stages. The driver tube is mounted on top of the assembly.

The power amplifier is located on the top deck of RF Unit T-225 ( )/FRT-5. It is a push-pull, cross-neutralized, grounded-cathode amplifier with both grid and plate circuits resonant and continuously variable from 4 mc to 26 mc. Cooling air for these tubes is forced into the grid circuit, which is mounted in an enclosure, by a blower in the lower rear of the cabinet, and air flows upward through the tube sockets to cool the radiators.

An alternate provision is available for operating the power amplifier as a linear amplifier driven by a single sideband suppressed-carrier exciter. The grid circuit of the power amplifier is arranged to be modified so that it presents a 200-ohm balanced load to the exciting single side-band transmitter. Approximately 500 watts of peak exciting power are required.

The coupling network occupies the top of RF Unit T-225 (B)/FRT-5. It is a resonant circuit continuously variable across the entire frequency range. Magnetic coupling from the plate tank coils to the inductors of this coupling network is varied by the positioning of the coupler on rails or tracks suspended from the top of the cabinet.

Power Supply PP-490 (B)/FRT-5

The Power Supply PP-490 ( )/FRT-5 consists of the lefthand bay of Radio Transmitting Set AN/FRT-5. It contains the following units and circuits:

A vertical panel inside the lower front door of Power Supply PF-490 ( )/FRT-5 mounts all of the control, overload and timer relays which are associated with both Power Supply PP-490 ( )/FRT-5 and RF Unit T-225 ( )/FRT-5.

A d-c potential of 600 volts, positive in respect to ground, for the operation of the frequency multiplier and buffer stages, the driver screen, and RF Oscillator O-92 ( )/FRT-5 is provided by a common low voltage supply. Rectification for the low voltage supply is accomplished by two type 3B28 xenon-gas filled rectifier tubes.

A d-c potential of 400 volts, negative in respect to ground, for biasing the frequency multipliers, the driver, the power amplifier, and the off-on keyer is provided by the bias supply.

Six type 4B32 tubes mounted on the tube shelf are rectifiers for the power amplifier plate voltage supply. The high voltage transformer for this supply is mounted outside of the unit. The filter components for the supply are located on the floor of the bay.

Power Control C-587 (B)/FRT-5

Power Control C-587 (B)/FRT-5 (fig. 6-19), includes the line input terminals for the 230-volt power supplying all parts of the transmitter, except the high voltage plate supply transformer. It houses the following components; a fused three-phase circuit to the transmitter for filaments; blowers, etc.; a combination overload protection breaker and start-stop contactor; and a "delta-wye" switch for the PA stage plate transformer primary. Also included are a set of series line resistors for tune-up purposes, contained in a ventilated cage on top of the cabinet.

Power Transformer TF-122/U

Power Transformer TF-122/U is a separate unit mounted apart from the cabinets. It is a three-phase, dry-type, air-cooled unit and supplies the high voltage for the power amplifier plate voltage supply.

AN/FRT-39/40

The radio transmitter set AN/FRT-39 (fig. 6-21) is a general purpose radio communication transmitter capable of providing 10,000 watts peak envelope power (PEP) output throughout a frequency range of 2 mc to 28 mc. The principle function of the equipment is to provide long range communications from shore-to-ship or point-to-point by the single-sideband type of operation. The equipment may also be used for the following types of transmission:

(1) CW (keyed carrier)
(2) Frequency-shift carrier
(3) Single-sideband suppressed carrier
(4) Double-sideband suppressed carrier
(5) Independent sideband (separate intelligence)
(6) Single-or double-sideband (with carrier)

With the addition of two cabinets, power amplifier and power supply, the AN/FRT-39 becomes an AN/FRT-40 (Fig. 6-22). The PA stage and antenna tuning controls of the AN/FRT-39, are slightly modified and become the IPA stage of the AN/FRT-40. The final PA section of the AN/FRT-40 consists of a power amplifier, class AB, capable of 40,000 watts (PEP), together with ample power supply, relay and control, protective circuits and meters.

The AN/FRT-39 is constructed in two basic sections, the main frame and the auxiliary frame. The main frame is located to the right of the auxiliary frame and houses the power amplifier (PA), the intermediate power amplifiers (IPA), the main power supply and high voltage section, the PA loading and tuning controls, the relay and indicator control panels and the meter panel. The auxiliary frame is located to the left of the main frame and houses all of the sideband exciter equipment, exciter power supply equipment, and other control equipment for the various modes of operation.

Auxiliary Frame Components

The block diagram of the AN/FRT-39 (fig. 6-23) illustrates the auxiliary frame components to the left of the dotted line in the figure.

The sideband exciter (CBE), will accept two channels of intelligence each having a bandwidth of 7.5 kc and processes them for modulation of an intermediate sideband carrier (250 kc) by means of two balanced modulators. The 250 kc carrier is generated in the regenerative divider network, which is part of the radio-frequency amplifier (CHG). The two 7.5 kc intelligence channels are used to modulate the upper sideband (channel 1) and the lower sideband (channel 2).
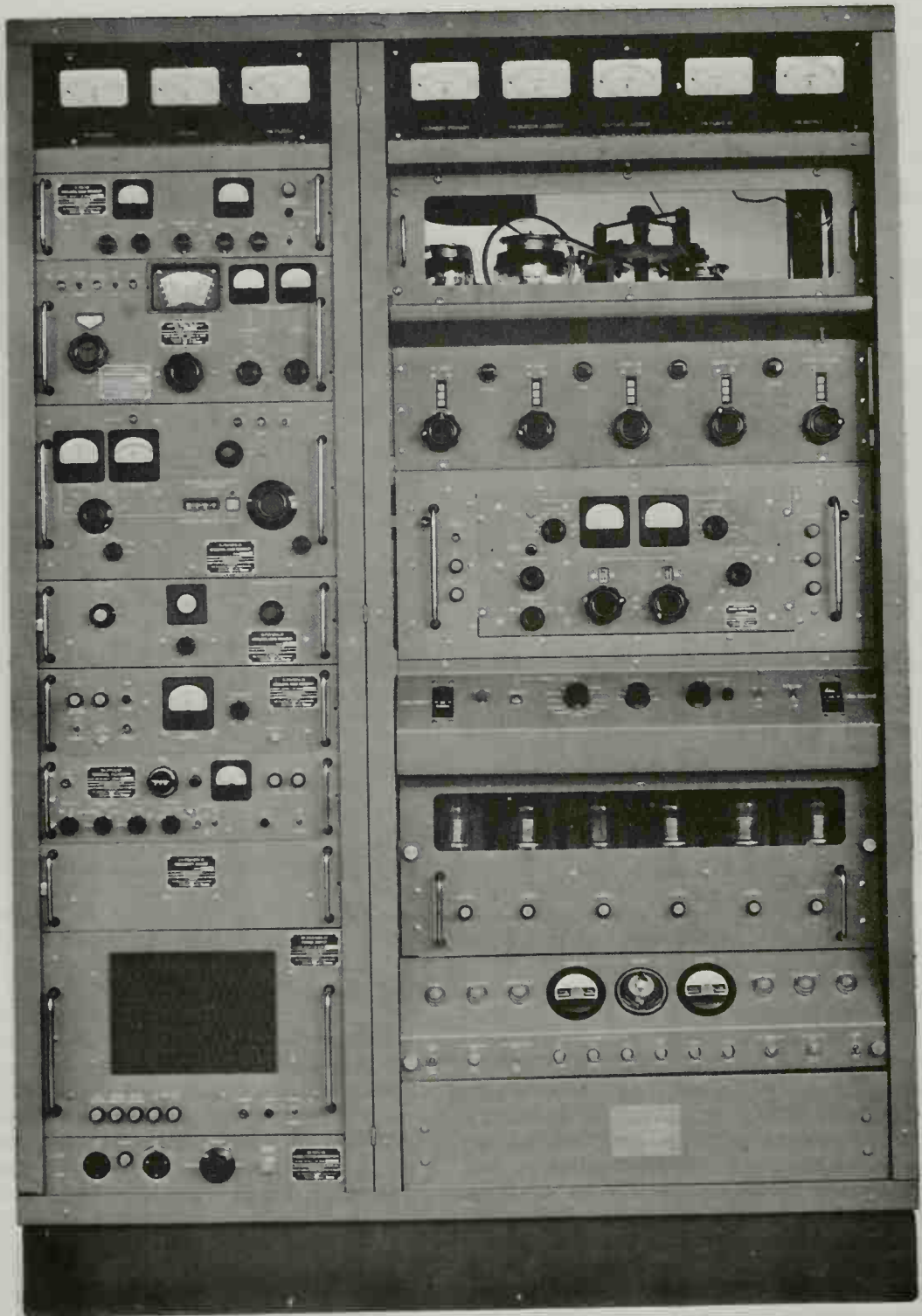
1.330

Figure 6-21.—Radio Transmitting Set (synthesized), AN/FRT-39.

162.17
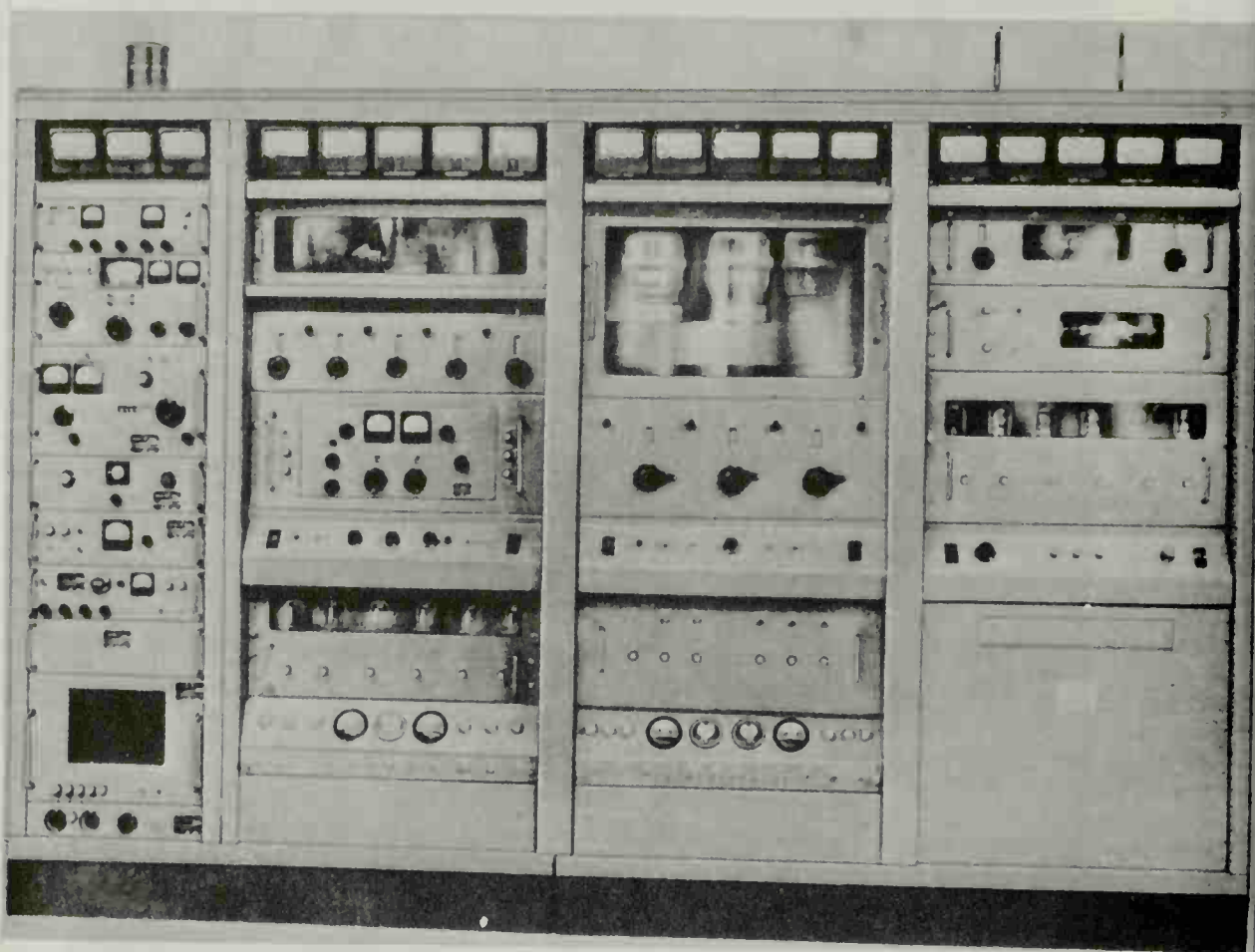
Figure 6-22.—Radio Transmitting Set (synthesized), AN/FRT-40.

The power supply (CPP-1) is mounted behind the auxiliary frame and supplies 200 volts unregulated and 150 volts regulated d-c outputs.

The radio frequency amplifier (CHG), is provided with two signal inputs: the sideband exciter, and the controlled master oscillator. The sideband exciter frequency is 250 kc ± 7.5 kc and the controlled master oscillator frequency is from 2 mc to 4 mc. The input frequencies are modulated by a balanced modulator with the 2 mc to 4 mc input and sideband exciter frequency, and subsequently heterodyned by additional precise high frequencies in order to bring the final r-f output into the 1.75 mc to 33.75 mc range.

The additional precise high frequencies are generated, by a crystal oscillator using different crystals (8 to 15 mc) for various bands, within the r-f amplifier unit and are synthesized by the accurate 1 mc standard.

The controlled master oscillator (CMO) supplies precise frequencies to the r-f amplifier in the 2 mc to 4 mc range. The accuracy of the output frequency is derived from the precise 1 mc standard.

The controlled oscillator (CLL) provides the low frequencies in 100 cps steps for stabilizing the controlled master oscillator. These frequencies are in the range of 510 kc to 519.9 kc and are fed to the master oscillator's phase detector to frequency stabilize the master oscillator's 2 mc to 4 mc output in the 100 cycle and 1000 cycle steps.
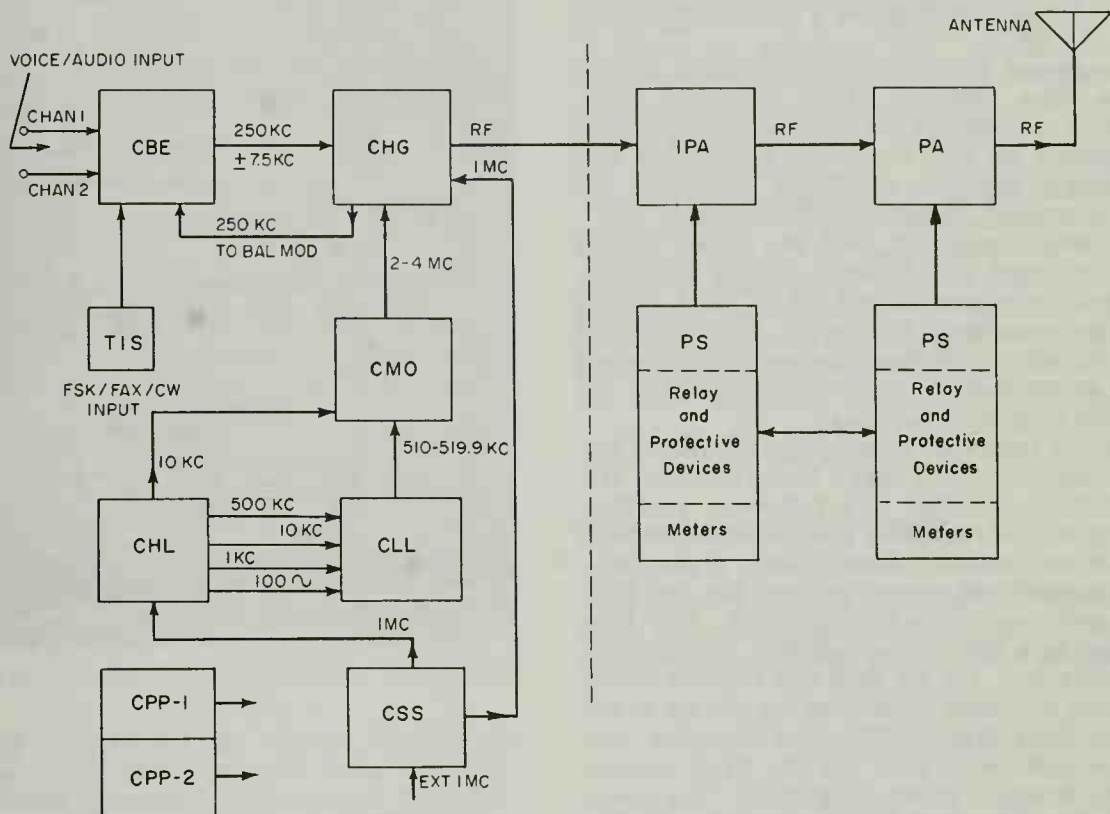
162.18

Figure 6-23.—AN/FRT-39 block diagram (synthesized).

The primary standard (CSS) is a precision oscillator within a temperature controlled oven. The oscillator has a frequency stabilization of one part in 10^8 per day. The frequency output of the primary standard is 1 mc and is used to regulate all of the frequency generating oscillators.

The divider chain (CHL) is used to count down the primary standard 1 mc output by using flip-flop multivibrators. The output frequencies available from the divider chain are: 500 kc, 10 kc, 1 kc, and 100 cps.

The power supply (CPP-2) provides:
(1) +380 volts unregulated
(2) -400 volts unregulated
(3) +160 volts regulated
(4) +75 volts regulated
(5) -6 volts regulated
(6) 6.3 (AC) volts regulated

The tone intelligence unit (TIS) accepts three types of d-c signals (FSK, CW, FAX) and converts them into audio frequency output signals

for sideband transmission via the sideband generator unit.

Main Frame Components

The block diagram of the AN/FRT-39 (fig. 6-23) illustrates the main frame components to the right of the dotted line in the figure.

The r-f amplifier which consists of a first amplifier stage (class A), a second amplifier stage (class A), and an intermediate power amplifier (class AB1), is commonly called the IPA. The input to the first amplifier stage is supplied by the sideband exciter (CBE) as explained earlier. The IPA is manually tune loaded and includes a multimeter with switching capabilities for direct reading of d-c and r-f voltages or currents. A second meter is mounted on the front panel for the purpose of reading the IPA plate current.

The IPA is rated at 1 kw r-f output and can be fed to the PA or independently to an antenna

by bypassing the PA when it is desirable to operate the transmitter at a maximum power output of 1 kw.

The power amplifier (PA), consists of one vacuum tube, rated at 10 kw, operating class AB1 in a grounded grid configuration. Tuning and loading of the PA output network is accomplished with the PA TUNE capacitor, PA LOAD capacitor, tapped inductance coils, OUTPUT balance capacitor, and the output loading inductors which are tuned by ferrite slugs. PA switching is accomplished by a band switch which progressively shorts out sections of the tuning inductance. Either balanced or unbalanced output is available by the use of jumpers on the OUTPUT LOADING inductors.

The r-f amplifier power supply consists of a high voltage rectifier and a bias rectifier. The high voltage rectifier is a full-wave rectifier and supplies unregulated plate voltage to both the first and second r-f amplifiers. It also supplies unregulated screen voltage for the IPA, first, and second r-f amplifiers. The bias rectifier is a half-wave rectifier and supplies -300 volts for the PA grid and PA bias relay. The relay provides protection for the PA in the event of bias failure. The bias rectifier also supplies -150 volts bias for the first, second, and intermediate power amplifiers. The power supply also supplies filament voltage for the IPA's.

The PA high voltage power supply circuits consist of high voltage rectifier tubes, circuit breakers, control equipment, 50 and 60 cycle power transformers, filament transformer, choke coils, high voltage capacitors, blower motors and other miscellaneous components. The power supply provides 7.5 kv PA plate voltage, 1.2-kc PA screen voltage, PA filament voltage, and 3-kc IPA plate voltage.

The relay and indicator control panels are equipped with control circuits for the purpose of disabling the transmitter when serious or abnormal conditions prevail. The relays are also used to apply voltage to the various circuits in proper sequence.

RADIO SETS

Radio Sets generally consist of both transmitter and receiver with associated power supplies and control units. The transmitter and/or receiver may or may not have the capability of being operated seperately. Radio Sets have the advantages of simplicity of operation, ease of maintenance, small size, and are adaptable to
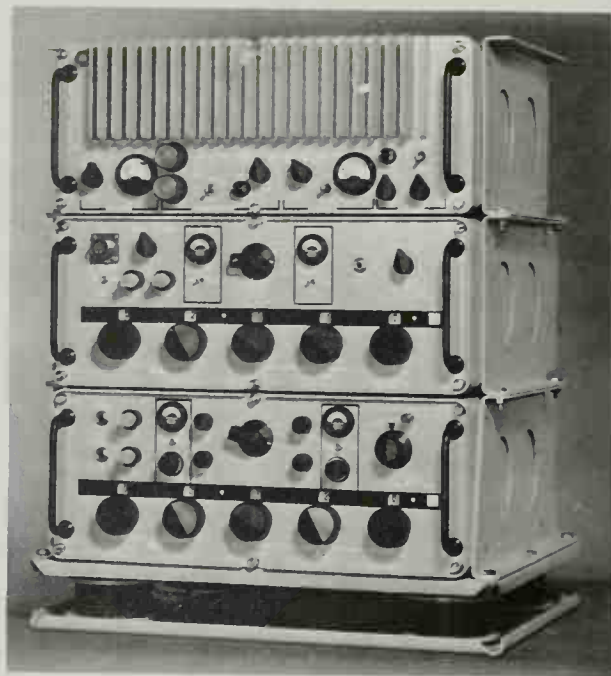
76.61
Figure 6-24.—Radio Set AN/WRC-1. Top unit: AM-3007/URT RF Amplifier; center unit: T-827/URT transmitter (exciter); bottom unit: R-1051/URR receiver.

both ship and shore communication. Although most transmitter-receiver sets are limited to low output power, due to size, the versatility of such equipments offer wide use as radio links in communication networks throughout the Navy.

AN/WRC-1

The AN/WRC-1 (fig. 6-24) is a single sideband radio set capable of transmitting on any one of 56,000 channels, spaced in 0.5 kc increments, in the 2.0 mc to 29.9995 mc frequency range. Vernier (continuous) tuning enables reception on any frequency in the 2.0 mc to 30.0 mc frequency range.

The AN/WRC-1 is capable of transmitting and receiving upper sideband (USB), lower sideband (LSB), continuous wave (CW), compatible amplitude modulated (compatible AM, frequency shift keyed (FSK), and independent sideband (ISB) signals in either a simplex or duplex operation. Tone-modulated continuous wave (MCW), standard amplitude modulated

(AM), and facsimile reception may also be made with the AN/WRC-1.

The AN/WRC-1 consists of Radio Receiver R-1051/URR, Radio Transmitter T-827/URT, RF Amplifier AM-3007/URT, and Interconnection Box J-1265/U. The relationship of these units, the Antenna Coupler CU-927/UR, and associated auxiliary equipment is illustrated in a functional block diagram in figure 6-25.

Radio Receiver R-1051/URR

The Radio Receiver R-1051/URR employs a digital tuning scheme for automatically tuning to any one of 56,000 operating channels, with vernier tuning providing continuous tuning throughout the frequency range. Since the R-1051/URR contains its own power supply, it may be operated as an independent unit, as was previously discussed in Chapter 4 of this training manual.

Radio Transmitter T-827/URT

The Radio Transmitter T-827/URT is a low level transmitter, which produces a nominal 0.1

watt r-f output of USE, 1SB, LSB, CW, FSK, or compatible AM r-f signal, making it capable of driving the AM-3007/URT. Like the R-1051/URR, the T-827/URT employs a digital tuning scheme for automatically tuning to any one of 56,000 channels in 500-cps steps in the 2.0 mc to 29.9995 mc frequency range. All circuits of the T-827/URT (except two r-f amplifier stages) utilize solid-stage devices. These circuits are assembled into plug-in electronic assemblies, some of which are interchangeable between the R-1051/URR and the T-827/URT. The frequency generation circuits, which are referenced to an ultra-stable master frequency standard with a stability better than 1 part in 10^8 per day provide an extremely stable transmitter output.

RF Amplifier AM-3007/URT

The RF amplifier AM-3007/URT is a two-stage power amplifier. With an r-f input of 0.1 to 0.25 watt, it will produce an output of

162.19
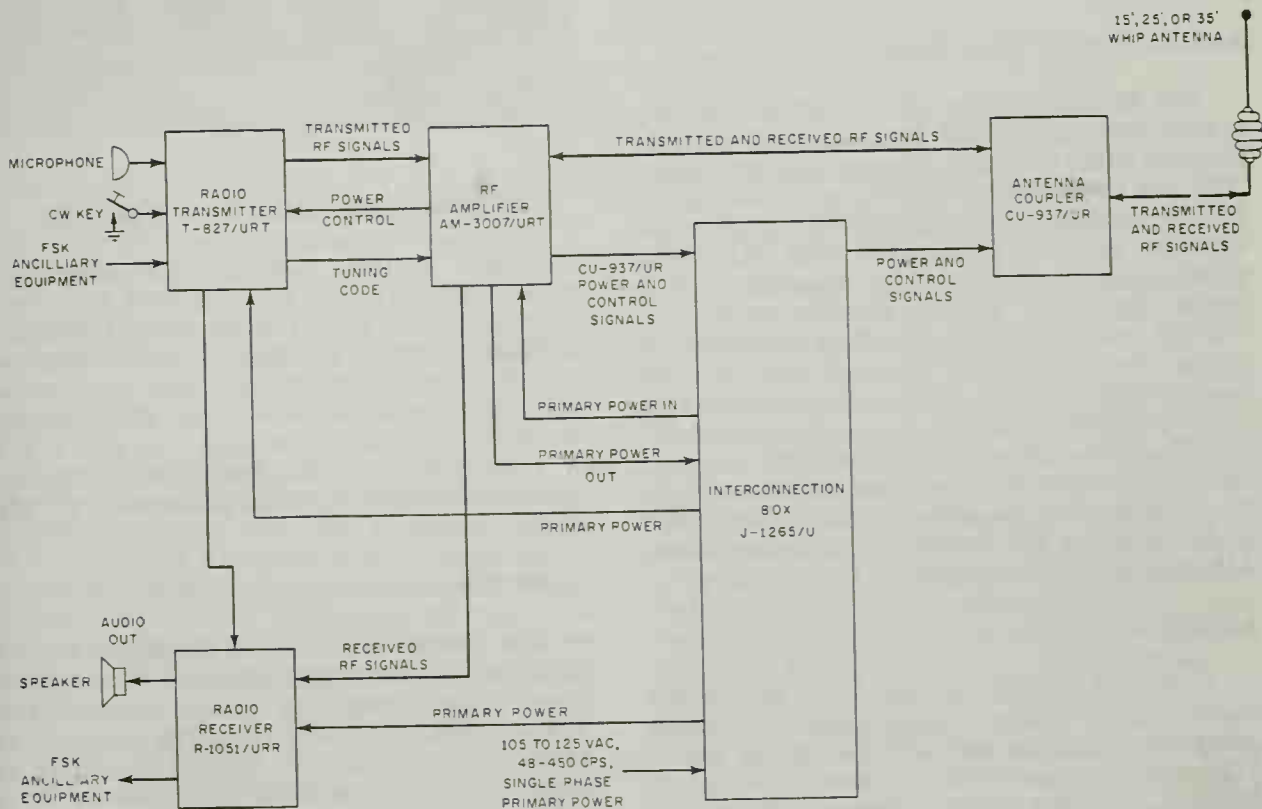
Figure 6-25.—Radio Set AN/WRC-1, functional block diagram.

100 watts PEP for SSB and ISB, 25 watts AM carrier, or 50 watts average CW or FSK into a 50-ohm load. In the AM and SSB modes, the AM-3007/URT is a linear amplifier. In the CW and FSK modes, the AM-3007/URT operates more nearly class C to increase efficiency. The AM-3007/URT is automatically tuned by a five-wire code from the T-827/URT. This code controls a motor that positions a turret containing broadband coils. These broadband coils act as tuned inter-stage and output circuits for the two amplifier stages. An inverse feedback loop is used in the AM-3007/URT to improve linearity and decrease intermodulation distortion.

Interconnection Box-J-1265/U

The Interconnection Box J-1265/U contains no active electronic circuits, but it furnishes the physical interconnection components for the system. Programming for the CU-937/UR for either the 15-, 25-, or 35-foot whip antenna, is accomplished by jumpering various terminals on two of the seven terminal boards in the interconnection box.

Antenna Coupler CU-937/UR

The Antenna Coupler CU-937/UR matches the system antenna (15-foot, 25-foot, or 35-foot whip antenna) to the 50-ohm r-f output of the AM-3007/URT (not used with submarine, multicoupler, or 50-ohm antenna systems). The CU-937/UR operates at power levels up to 100 watts PEP, while maintaining a 50-ohm input impedance at a better than 1.5:1 voltage standing wave ratio (VSWR) over the 2.0 mc to 30.0 mc frequency range. The CU-937/UR is tuned entirely from the AM-3007/URT. A tuning code is generated in the AM-3007/URT to course tune the CU-937/UR. Final fine tuning to bring the VSWR to better than 1.5:1 is then accomplished with controls on the front panel of the AM-3007/URT. The VSWR is indicated by the RF OUTPUT meter on the AM-3007/URT front panel.

AN/SRC-20

Radio Set AN/SRC-20 (fig. 6-26), is designed for shipboard or fixed station operation. This set provides amplitude modulation (AM) on any of 1750 channels spaced 0.1 mc apart in the 225 mc to 399.9 mc range with a minimum power output of 16 watts into a 50-ohm resistive load. Nineteen of the 1750 channels can be present.
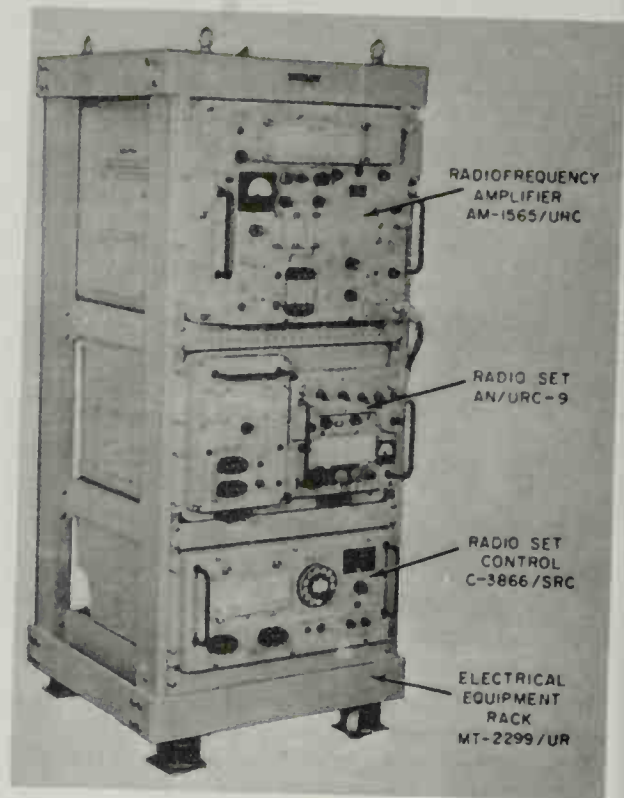
162.20
Figure 6-26.—Radio Set AN/SRC-20.

Complete control, including the selection of preset channels, can be exercised from up to a maximum of four remote control points. In addition, circuits are incorporated which permit the connection of two sets for two-way automatic retransmission. The radio set consists of three units as shown in figure 6-26. A functional block diagram showing interconnections between units is illustrated in figure 6-27.

Radio Frequency Amplifier AM-1565/URC

The Radio Frequency Amplifier AM-1565/URC is an automatically tuned, fixed station linear power amplifier which operates from 225 to 399.9 megacycles. The cathode and plate tuned circuits of the r-f amplifier are high Q coaxial cavities, servo-tuned to the selected frequency. Because the r-f amplifier is linear, interference due to transmitter intermodulation is minimal. An internal power supply
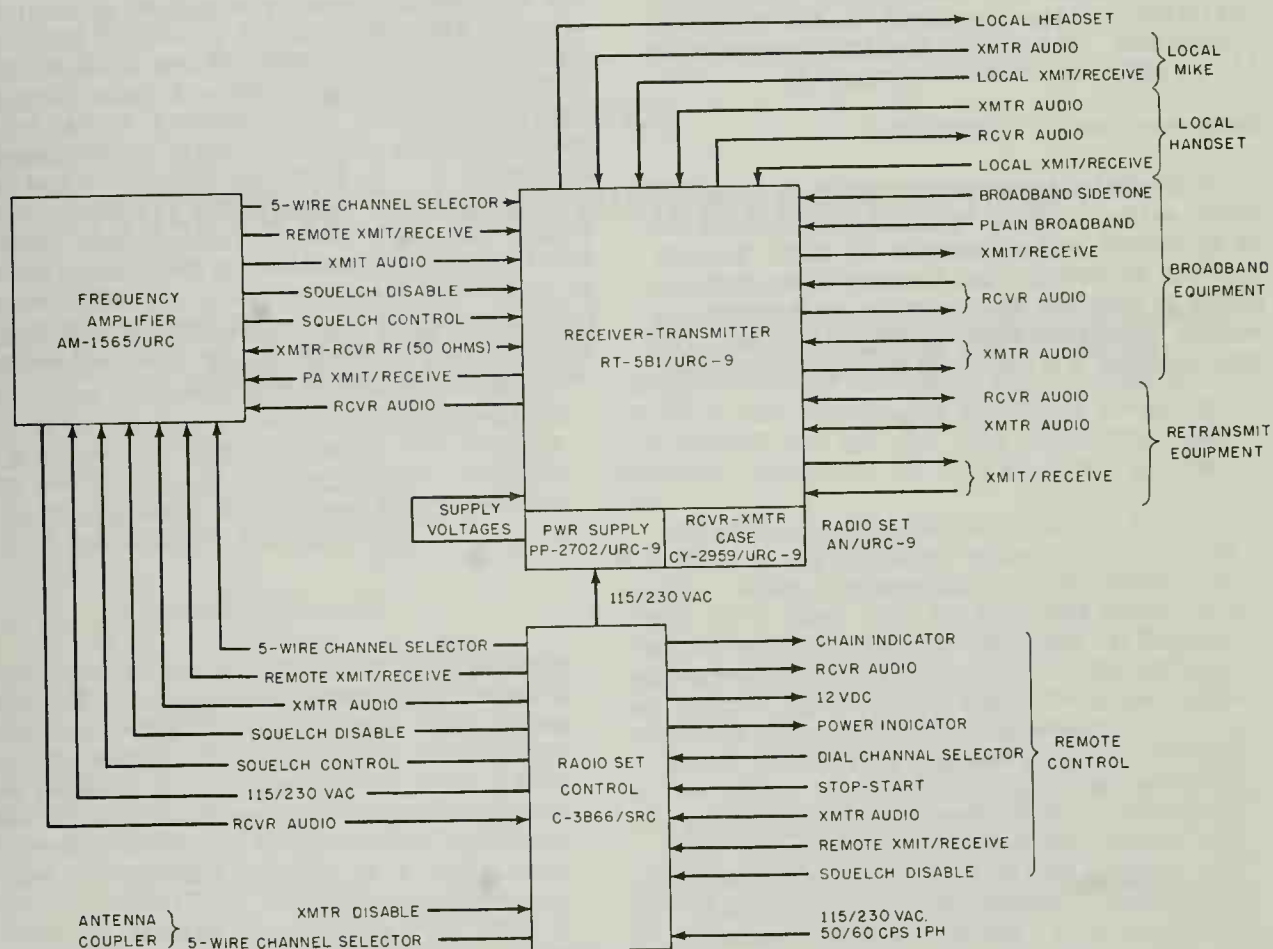
162.21

Figure 6-27.—Radio Set AN/SRC-20, functional block diagram.

operates from a 115/230-volt, 50-60 cps power source.

Channel selection is accomplished by an "Autopositioner". The r-f amplifier is continuously tunable over the frequency range, with provisions for presetting 19 channels for remote or local selection. The twentieth channel allows manual tuning. All channel information is made available to the associated exciter. A remote control unit may be used to control the r-f amplifier.

Radio Set AN/URC-9

Radio Set AN/URC-9 is the major unit of the AN/SRC-20 and consists of Receiver-Transmitter RT-581/URC-9 and Power Supply PP-2702/URC-9. The unit functions as a triple-conversion, superheterodyne receiver during standby operation. When the microphone push-to-talk switch is pushed, a series of T/R (transmit-receive) relays convert the unit to a transmitter which uses many of the same r-f, i-f, and audio circuits used on receive. Three crystal-controlled oscillators provide stable r-f and i-f frequencies on both transmit and receive. Receiver-Transmitter RT-581/URC-9 is automatically tuned by standard "auto-positioners", which are controlled on the front panel or by remote control. Any of 19 preset channels from 225.0 mc to 399.9 mc can be selected from the front panel or the remote control. Any one of the 1750 channels within the frequency range can be automatically selected using the manual frequency control on the front panel.

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Power Supply PP-2702/URC-9 provides all operating voltages required by the receiver-transmitter. The supply nominally operates on 115 or 230 volts, 50 or 60 cycles AC.

Radio Set Control C-3866/SRC

Radio Set Control C-3866/SRC is a device which enables a radio operator to select any one of 10 preset radio channels on Radio Set AN/SRC-20. Remote control from up to four distance points is possible when correct combinations of remote dialing and audio equipment are used. The unit contains a front-panel telephone-type dial, and the relays necessary to operate an internal stepping relay for channel selection. Nineteen squelch level potentiometers are available for setting the squelch level of each radio channel. The C-3866/SRC also contains three relay power supplies and the primary power control equipment for the Radio Set AN/URC-9 and the power amplifier of the Radio Frequency Amplifier AM-1565/URC. A local-remote switch transfers control functions to the parallel group of up to four remote stations, and audio transformers convert the unbalanced inputs from Radio Set AN/SRC-20 to the balanced output required by the remote stations.

CHAPTER 7

MULTICOUPLERS AND DISTRIBUTION SYSTEMS

Although there are many new types of multi-couplers in various power handling, shapes, and sizes, essentially all work on the same theory. Thus if the technician works on multicouplers not described in this chapter, he will find the same basic principles apply. Several UHF and HF multicouplers are discussed with the reason for, and operation of, multicouplers.

MULTICOUPLERS

Because of space limitations and the large number of communications equipments on board ships, it is impractical, if not impossible, to use separate antennas for the various equipments. It is possible to use one antenna over a wide range of frequencies provided the antenna system is properly tuned and coupled to the equipment so that the antenna appears as a nonreactive (pure resistance) load. It is also possible to use a coupling system that allows operation of several transmitters and/or receivers at the same time. It is not, however, normally possible to use an antenna for both transmitting and receiving simultaneously.

One satisfactory approach to the problem of antennas is provided by multicouplers. The multicouplers that are discussed in this chapter are representative of those in current use on naval vessels.

VHF/UHF MULTICOUPLERS

One antenna coupler widely used is the CU-255/UR. When assembled into a group of two to six units, they provide a system for operating two to six UHF transmitters and/or receivers into or from a single antenna. One coupler is required for each transmitter or receiver. Figure 7-1 shows a group of six multicouplers installed in a rack.

The equipment was designed to operate in the frequency range between 230 and 390 mc. Although it can be used with any transmitter or receiver that covers the frequency range, the AN/GRC-27 and AN/GRC-27A transceivers are normally used with the coupler group in figure 7-1.

All of the couplers in the group can be tuned manually to any frequency within the limits given above. When they are used with the equipment listed each tuner may be manually preset to a channel within the band frequencies. It then becomes possible for the operator to dial any preset channel automatically either from the transceiver or on a remote channel selector.

Each CU-255/UR antenna coupler consists of two major components: a coupling cavity or r-f section and an automatic drive mechanism.

The importance of matching impedances has been mentioned previously in this chapter, and the r-f section of the multicoupler is essentially an impedance matching device. It is capable of matching an antenna feedline impedance of 50 ohms to antenna input impedance that mismatch the feedline impedance as much as 2.5 to 1. Thus, the multicoupler provides a means of maximum power transfer between the transmitter and the antenna. The same matching action takes place when the system is used for receiving.

A meter on the front panel of each multicoupler indicates when the unit is correctly tuned. The meter is a part of a reflectometer and indicates the magnitude of the power reflected back from the coupler. When the tuning meter shows zero, there are no standing waves on the transmission line, and the impedances are matched. A discussion of a reflectometer will be given later in this chapter.

The internal arrangement of a multicoupler is shown in the perspective view of figure 7-2. The drive mechanism that permits manual and
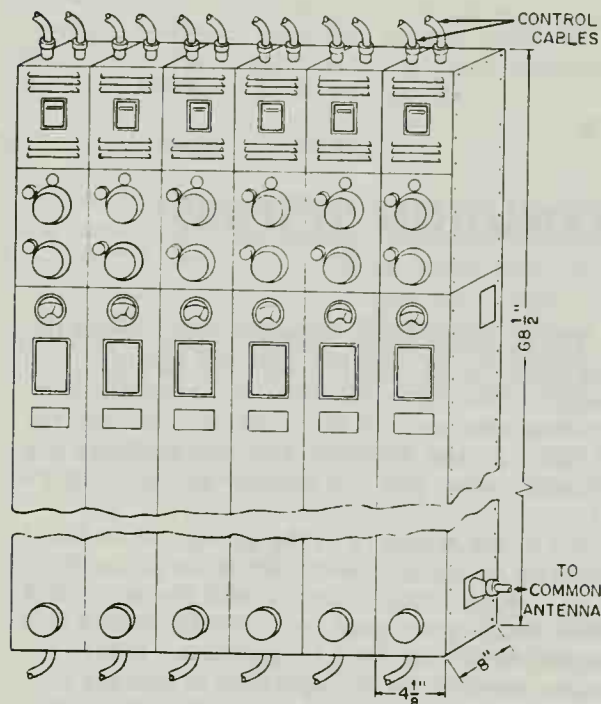
1.266

Figure 7-1.—Group of six type CU-255/UR antenna couplers.

automatic tuning of the coupler fits into the top of the r-f section to form a long thin unit. It can be removed for servicing, when necessary. The normal frequency coverage is from 230 to 390 mc, and the median range is from 227 to 396 mc.

Any device inserted into a transmission line will cause some loss, called the insertion loss. The multicoupler introduces an insertion loss of about 1 db over its tuning range, however the power gained by matching impedances more than compensates for the small insertion loss.

The type CU-332/UR multicoupler covers the same frequency range as the CR-255/UR multicoupler, and both types may be used for either transmitting or receiving. The only difference in the two units is that the CU-332/UR has provision for manual tuning only.

Other multicouplers covering frequencies between 225 and 400 mc are the CU-274/UR, CU-284/UR, CU-355/UR, CU-377/UR, CU-691/U, and CU-692/U.

The CU-274/UR was designed primarily for use with radio transmitter TDZ and radio receiver RDZ, but it may be used with other UHF
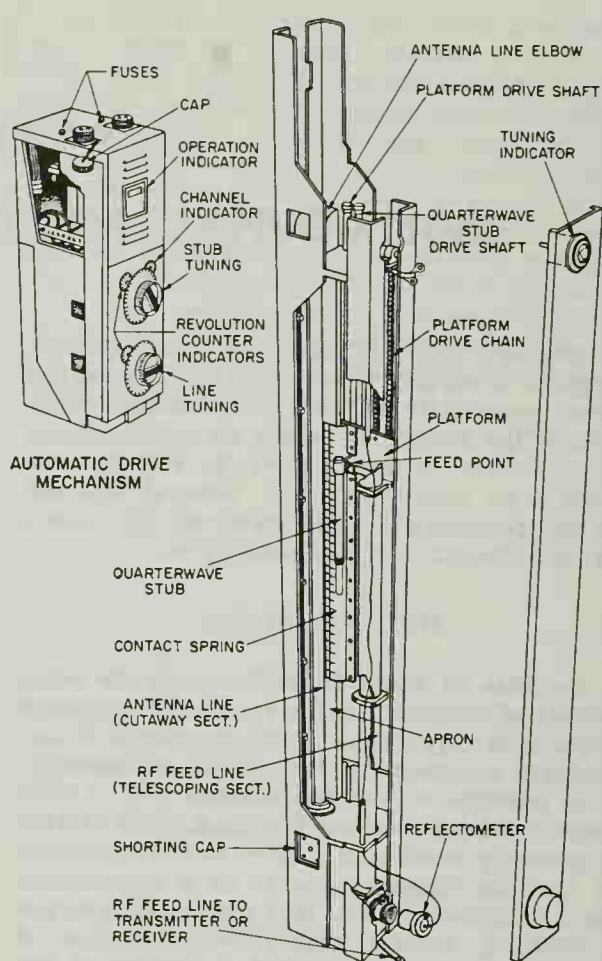
1.267

Figure 7-2.—Perspective view of CU-255/UR multicoupler.

equipments such as the TED, AN/URR-13, AN/GRC-27, AN/GRC-34 and AN/GRC-35. The coupler provides for simultaneous operation of four UHF radio channels into a single antenna in the frequency range of 225 to 400 mc. The four cavities are independently tuned either automatically or manually. For automatic tuning each cavity must first be pretuned to 10 corresponding frequency channels in the frequency range of the attached receiver or transmitter, and thereafter it is automatically retuned by an auto-tune mechanism. Manual tuning of each cavity is possible when the clutch-release knob is released. Any one of the channels may be used interchangeably for transmitting and receiving.

The CU-284/UR multicoupler permits simultaneous operation of two UHF transmitters or receivers into a single antenna. It is primarily designed for installation on submarines and other small vessels. The coupler is manually tuned and consists of two capacitively tuned resonant cavities, each of which has a 50-ohm terminal impedance. The CU-284/UR is similar to the CU-274/UR except that it can be tuned manually only and accommodates two radio channels instead of four.

The CU-355/UR multicoupler has provisions for the simultaneous operation of four UHF transmitters and/or receivers into a single antenna on the frequencies between 225 and 400 mc. Either manual or automatic tuning may be used with each of the four cavities. Each cavity has its own power monitor and standingwave indicator. A 12-position switch on the front panel provides the necessary switching to read the power or standing-wave ratio at any one of the four input connections. The operating frequencies of the units connected to the multicoupler must be separated by one megacycle or more, depending on the characteristics of the auxiliary equipment.

The CU-377/UR permits the simultaneous operation of two UHF transmitters or receivers into a single antenna in the frequency range of 225 to 400 mc. This coupler is similar to the CU-284/UR.

The CU-691/U provides for the simultaneous operation of four transmitters or four receivers into, or from, a single antenna. This coupler has four cavities that are tuned manually. Before the cavity can be tuned, it is necessary to loosen the clutch release.

The CU-691/U multicoupler is completely shielded and does not produce r-f interference. A meter on the coupler indicates when maximum energy is being transferred while transmitting. There is no meter indication when receiving.

The CU-692/U permits two separate transmitters and/or receiver combinations to operate simultaneously into a common antenna. Except for the number of channels and physical dimensions, the CU-692/U is identical to the CU-691/U.

H-F MULTICOUPLERS

A group of high-frequency antenna couplers (AN/SRA--3, 14, 15, and 16) permits the simultaneous operation of one to four transmitters into the same antenna on frequencies between 2 and 26 mc. These couplers are made up into four-channel groups. Each group operates in one of the following bands: 2 to 6 mc, 4 to 12 mc, 6 to 18 mc, or 9 to 26 mc. To obtain complete coverage from 2 to 26 mc, four coupler groups and four broad-band antennas are needed.

AN/SRA-15

The type AN/SRA-15 (6-18 mc) antenna coupler is representative of the h-f group. It has provisions for the simultaneous operation of four transmitters, each rated at 500 watts, into one broad-band antenna. The coupler covers the range between 6 and 18 mc and will operate into any antenna provided the standing-wave ratio is not more than 3 to 1. All four transmitters can be operated at the same time as long as a separation of 10% or better is maintained between carrier frequencies.

Each antenna coupler consists of four drawers and a cabinet to enclose them. Each drawer provides a means of coupling a transmitter to the common antenna. Figure 7-3 shows the drawer and cabinet arrangement of the AN/SRA-15.

A flexible coaxial line extends from the antenna line connector, shown at the lower right side of the cabinet in figure 7-3, to the top coupler unit in the cabinet. The coaxial line ends at this point and is short circuited. Where the line passes through a coupler unit, a section of the outer braid on the cable is removed. The center conductor is also cut, and the ends are connected to two terminals on the coupler unit. A one-turn loop has its ends connected to these terminals so that it is in series with the center conductor at each coupler unit. These loops provide inductive coupling between the antenna and the inductance in the tuner tank circuit.

The tank capacity in each tuner is tapped, and forms an impedance type divider that always presents a 50-ohm input impedance when the unit is correctly tuned. The combination of tunable tank and variable coupling forms a transformer that matches the input impednace of the antenna to the 50-ohm line that leads to the transmitter. The tank circuits have high Q, providing the necessary selectivity for isolation between channels.

Reflectometer

Each coupler unit has a reflectometer, located in the transmission line from the

transmitter, to indicate when the tuner is correctly adjusted by indicating the amount of reflected energy on the line. Thus, any mismatch between the coupler and the 50-ohm transmission line is readily apparent, and it can be corrected by returning the coupler.

Figure 7-4A is the movable section of the reflectometer separated from the fixed section. The schematic diagram (fig. 7-4B) shows how the parts are connected in the moving section. A cable connects the meter to the circuitry of the moving section.

The reflectometer housing is made in the form of a T, and the center conductor of the line passes through the top of the T. The moving part of the bridge slides into the short arm of the T so that the disc is close to the center conductor and capacitively coupled to it. There is also inductive coupling between the loop to which the disc is connected and the center conductor. Because it is necessary to maintain the capacity at the same value when a test is in progress, an adjustable coupling screw passes through the cover of the moving section. The end of this

70.28

Figure 7-3.—Antenna milticoupler AN/SRA-15.

220

1.268

Figure 7-4.—Reflectometer.

When the reflectometer is in use, the voltage applied to the detector is the vector sum of the components because of inductive and capacitive coupling. As the angle formed by the loop and axis of the feedline changes, the amplitude of the inductive component varies from zero to maximum as the capacitive component remains unchanged. The phase of the inductive component is reversed by turning the loop through 180 degrees. Therefore, at some orientation of the loop the currents of the two components will cancel.

The same cancellation can be obtained by leaving the loop parallel to the feedline axis and adjusting the position of the capacitive plate by turning the adjustment screw in or out until cancellation occurs. This will happen when there is no reflected wave on the feedline because (1) the capacitive coupling is very small, (2) the loop inductance is negligible compared to the resistance of R1 (82 ohms), and (3) R1 is noninductive.

After cancellation has been obtained, as indicated by a zero meter reading, the movable section is clamped in position. Thereafter, zero meter reading indicates the impedances are matched and maximum power is being transferred.

There is some hazard in using this method, however, because a mistuned coupler can cause a high standing-wave ratio, which would produce enough reflectometer current to burn out the microammeter and crystal diode.

A more practical method of using the reflectometer is to tune the transmitter and coupler for minimum standing-wave reflection and then use the reflectometer for final fine tuning. Usually zero readings will not be obtained, but a minimum of two to six microamperes can be accepted as an indication of correct tuning.

The multicoupler has two controls. One operates the input coupling capacitor and the tank tuning capacitor which are ganged together, and the other operates the coupling-loop position in relation to the tank inductance. The last-named control is used to adjust the loading, and the first-named control is used to tune the tank to resonance.

All four antenna coupler types have the same general design. However, the AN/SRA-16 (9-26 mc) uses capacitance coupling instead of inductive coupling to the transmission line, and parallel operation instead of series operation as is done in the other three couplers. The input circuitry and the method of control, tuning

screw rests on a shoulder that is a part of the fixed section and maintains constant spacing regardless of how the movable section is turned. Thus, the movable section can be oriented to adjust the pickup loop in relation to the center conductor of the coaxial line.

Normally, the plane of the U-shaped loop (fig. 7-4B) is parallel to the axis of the feedline. One end of the loop is connected to ground through R1, and the other end is connected to a diode crystal detector. Resistor R2 is a current limiting resistor, and C1 is an r-f bypass for the meter. The meter has a 200-microampere movement and is located on the panel of the coupler.

indication, and operation are essentially the same for all units.

Operation

When a multicoupler installation is first put into operation, the transmission line is first terminated with a noninductive, 50-ohm load at the point where it will connect to the antenna couplers. The power-handling capability of the resistor (load) must be such that it can handle the entire output of the transmitter. The transmitter is then operated into the load, and the controls on the transmitter are adjusted for best operation. Tuning for the various channels to be used is set up at this time and, after the transmitter adjustments are completed, the controls need not be touched again unless it becomes necessary to change frequency.

The noninductive load is then disconnected from the transmission line and the line connected to the antenna connector on the multicoupler cabinet. The antenna couplers are adjusted for minimum standing-wave ratio at each operating frequency while the transmitter is being operated. When a minimum standing-wave ratio has been obtained at each frequency, the antenna and coupler are furnishing a 50-ohm load to the transmitter. It is theoretically possible to obtain a standing-wave ratio of 1 to 1, but it is not always possible to eliminate reflections in actual practice. In any event, the tuning of the multicoupler should be done carefully to get the meter reading as close to zero as possible. Otherwise power will be wasted in reflected waves.

AN/SRA-23

Antenna coupler group AN/SRA-23 permits the simultaneous operation of up to eight 500-watt transmitters covering the frequency range of 2 to 27 mc in one broad-band antenna system, with a minimum frequency separation between channels of approximately 10 percent. One antenna coupler group is required for each transmitter in the system. The group consists of five major components as shown in figure 7-5. The three antenna couplers are similar in principle of operation, but their circuit components differ to provide different frequency ranges. A trunking and switching unit (fig. 7-6) is also supplied for combining four coupler groups into a system. This provides a switching system for selection of the desired coupler, dummy antenna load, and the desired broad-band antenna. A four-group sys-

tem can be expanded into an eight-group system by interconnecting two four-group systems back-to-back.

Although the coupler was developed for use with 500-watt transmitters, it is capable of handling transmitters with 1000-watt outputs. For operation with 1000-watt transmitters, the incident power meter M401 (fig. 8-12) must be recalibrated for full-scale deflection with 1000 watts of r-f power instead of 500 watts. Electronics Installation and Maintenance Book, NavShips 0967-000-0110, contains instructions to accomplish this calibration.

FILTER TYPE MULTICOUPLERS

When a number of receivers are to be operated from a common antenna, some isolation between the input connections must be provided to prevent interference between sets and to keep losses at a minimum. The antenna filter assemblies used for this purpose are the AN/SRA-9 and AN/SRA-12. Since the filters are the same except for frequency coverage only one, the AN/SRA-12, will be discussed.

Receiver Filter Assembly

A receiving-antenna distribution system, using a filter assembly, is shown in figure 7-7. This type of distribution system makes possible the multiple operation of a maximum of 28 radio receivers from a single antenna. It is generally preferable however, to limit the total number of receivers to 7.

This filter assembly or "multicoupler" provides 7 r-f channels in the frequency range from 14 kc to 32 mc. Any or all of these channels may be used independently of, or simultaneously with, any of the other channels. Connections to the receivers are made by means of coaxial patch cords and a patch panel.

An external view of the filter assembly is shown in figure 7-7,A. Separation of the frequency range into channels is accomplished by combinations of filter subassemblies, which plug into the main chassis. Each filter subassembly consists of complementary high-pass and low-pass filter sections, the common crossover frequency (F_c) of which marks the division between channels.

The filters not only guard against interference at frequencies falling outside the channel being used, but also prevent receivers connected to alternate rows of jacks from interacting with each other when their tuning and trimming adjustments are made.

CONTROL C-2530/SRA-23

COUPLER CU-680/SRA-23

COUPLER CU-679/SRA-23

COUPLER CU-678/SRA-23

BLOWER
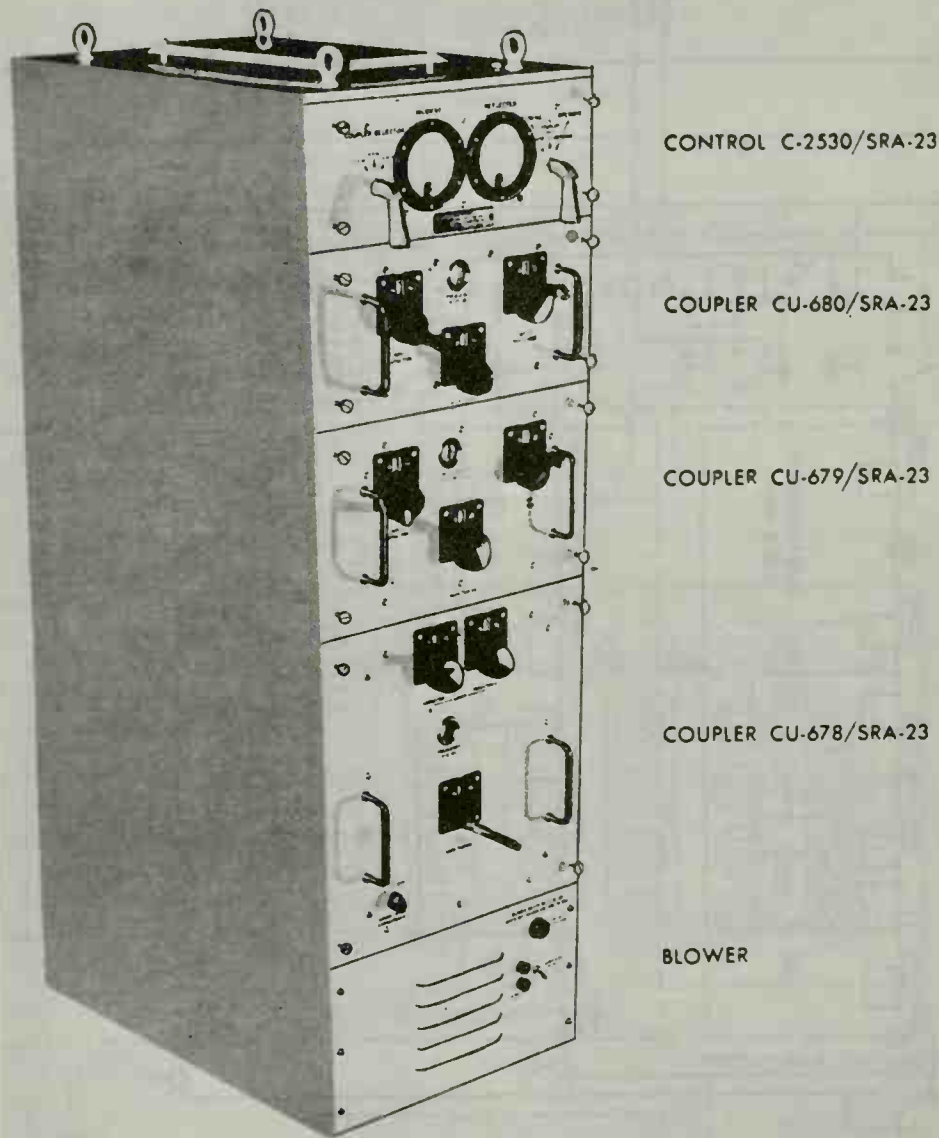
109.40

Figure 7-5.—Antenna coupler group AN/SRA-23.

A set of nine filter subassemblies is available, any six of which may be used at one time. The filter subassemblies are sealed units consisting of inductors and capacitors and are terminated in 4-terminal plugs, which are designed to engage octal receptacles on the main chassis. The subassemblies have numbers stamped on them that indicate their crossover frequencies. These numbers can be viewed through windows in the front panel. The six subassemblies that are used are assembled in the order of decreasing frequencies from left to right, as viewed from the front of the panel.

The filter panel (fig. 7-7,B) contains 1 antenna input jack, 28 output jacks, 21 decoupling resistors, and 6 octal sockets. The antenna input jack and the 28 output jacks are all Navy type-49120 r-f connectors. (In the AN/SRA-12
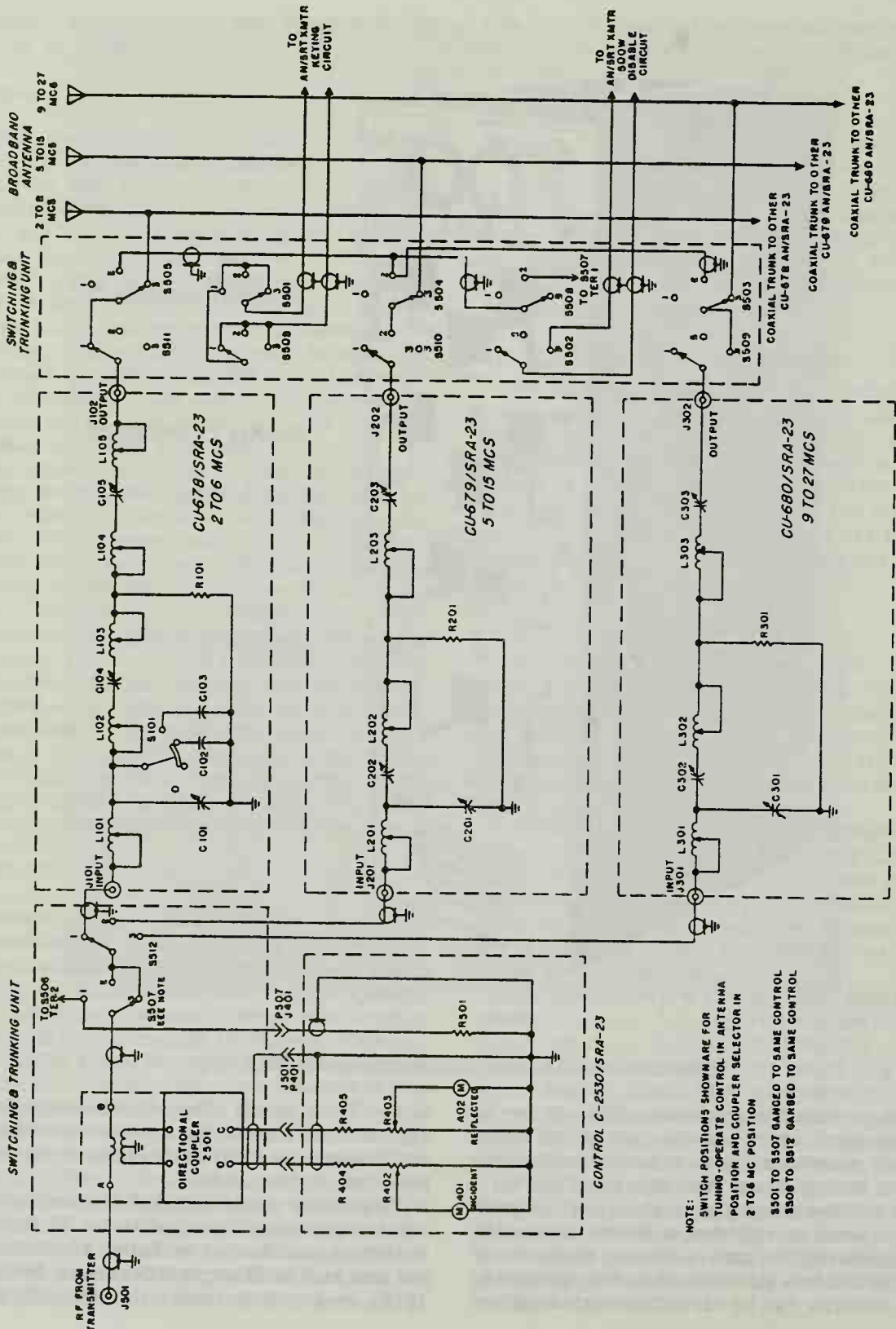
Figure 7-6. — Antenna coupler group AN/SRA-23, simplified schematic.

109.41

224
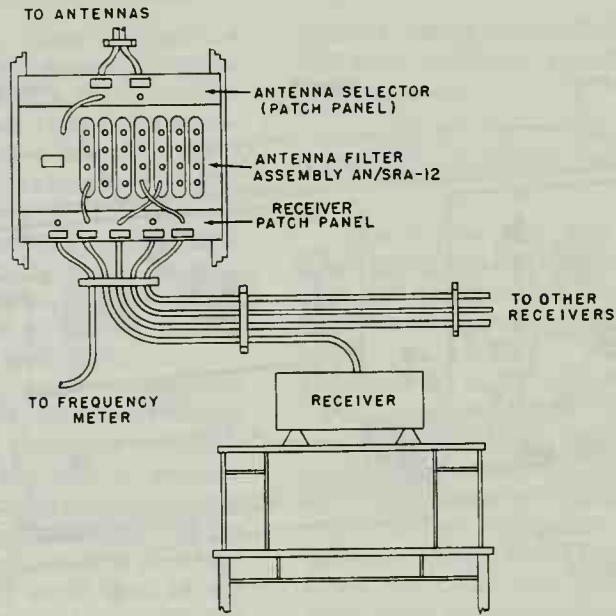
A EXTERNAL VIEW
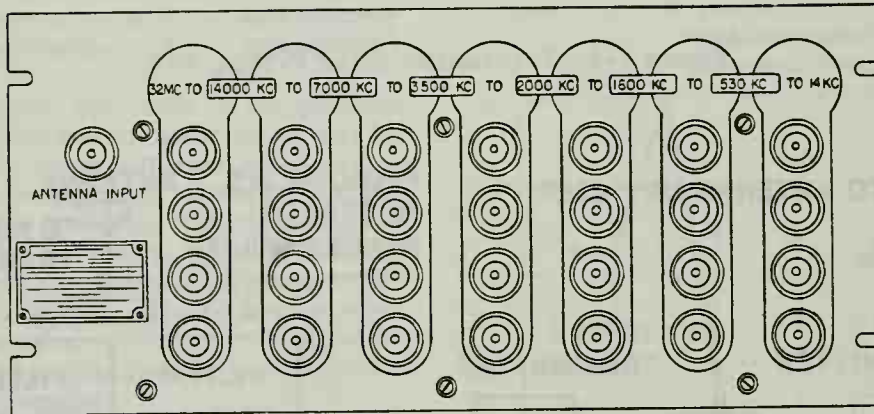
B FILTER ASSEMBLY DETAIL
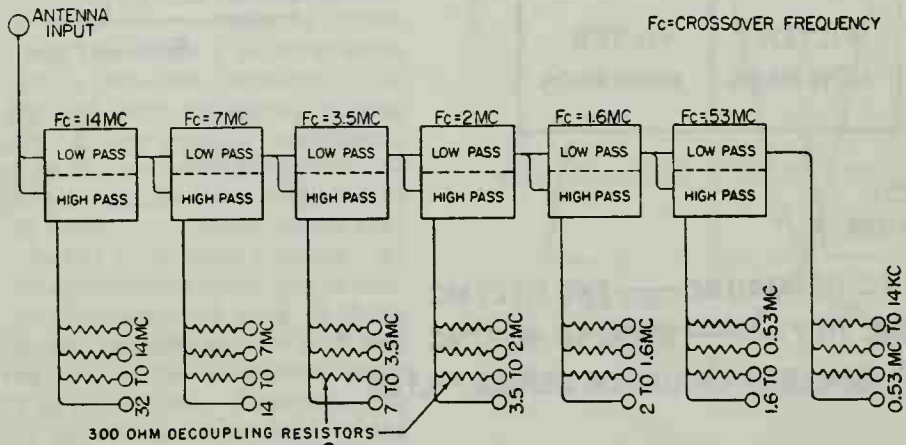
C SCHEMATIC DIAGRAM

1.115

Figure 7-7.—Receiving-antenna distribution system, using antenna filter assembly.

225
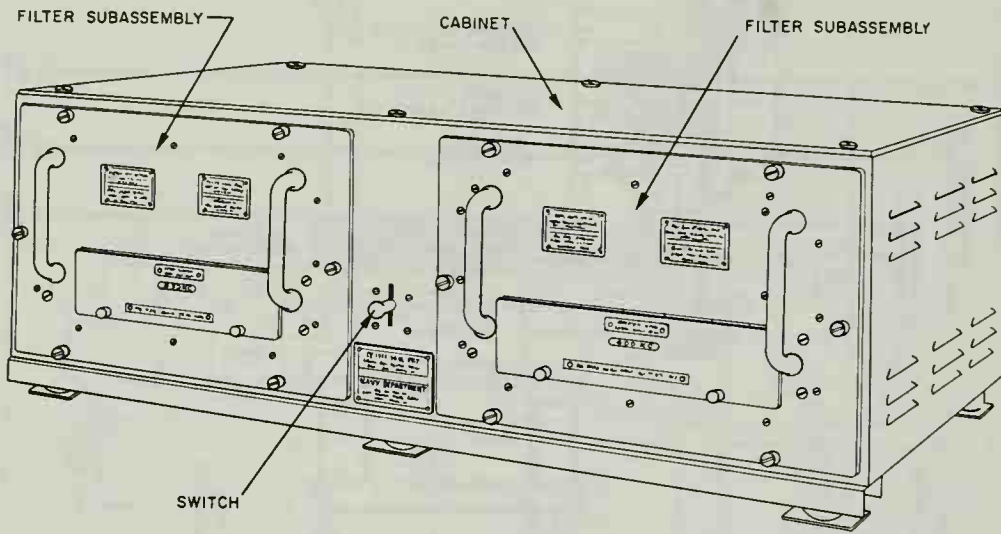
Figure 7-8.—Transmitter filter F-162A/SRT.

109.42
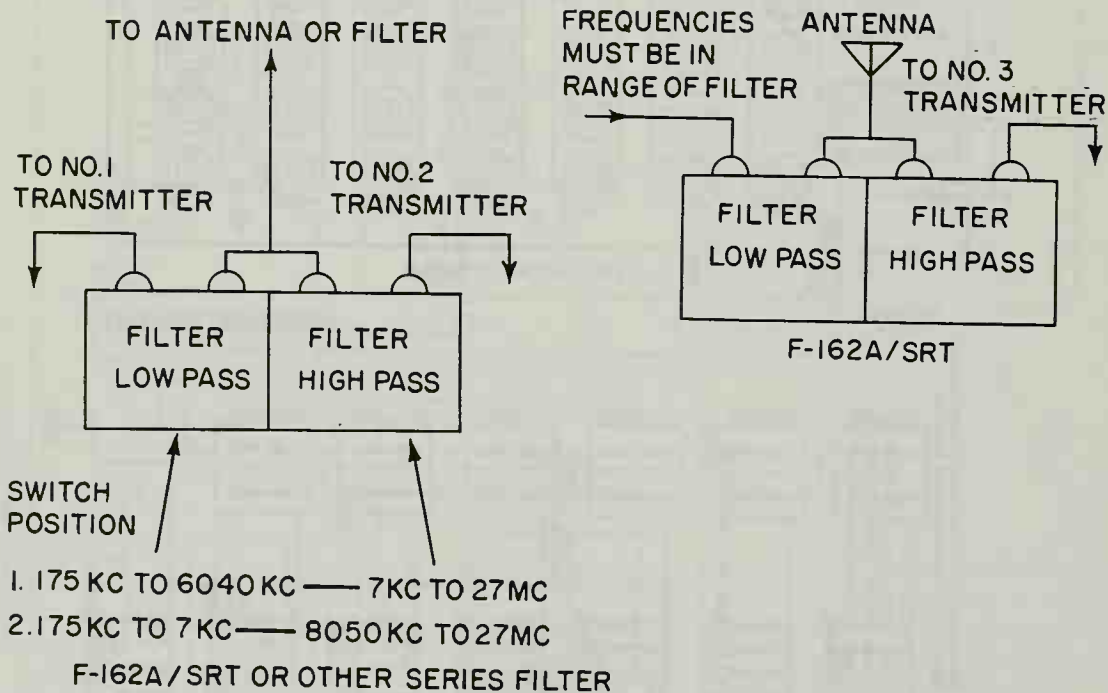
Figure 7-9.—Representative F-162A/SRT installations.

162.22

Receiving Filter Assembly, these connectors have been changed to receptacle connectors UG-1111/U (and plug connectors UG-968/U) which are improved quick-disconnect type r-f connectors.) The filter subassemblies plug into octal sockets in the rear of the main chassis (not shown).

To keep the losses to a minimum, the input and output of the filter assembly should be terminated in 180 ohms, however, only a slight reflection loss (of the order of 1.0 db) results when the input is terminated in 70 ohms.

Because Navy communications receivers generally operate throughout frequency bands that exceed the widths of the channels normally provided by the filter subassemblies, a given receiver must be connected to the particular row of output jacks that provides the signals of the desired frequency. For example, if a receiver is tuned from some frequency in the 7- to 14-mc band, to some frequency in the 14- to 32-mc band, the patch cord would have to be moved from the output of the 14- to 7-mc subassembly unit to the output of the 32- to 14-mc subassembly unit.

When necessary, the bands of frequencies available in a given row of output jacks may be changed either by using different combinations of filter subassemblies or by removing various subassemblies and inserting "shorting plugs," provided with the equipment. These shorting plugs provide continuity between successive subassemblies, as required when testing or when a subassembly is removed for any reason, without the necessity for changing the position of the remaining subassemblies.

The red-painted jacks at the bottom of each row are directly connected to the subassemblies and should be used whenever maximum signal strength is desired. The other 3 jacks in each row are decoupled by 300-ohm resistors and are best suited for use with relatively strong signals (fig. 7-7, C). In the ideal arrangement, only one receiver is connected to each vertical row or jacks, and that receiver is connected to the bottom jack in each row. This means that seven receivers are fed from each antenna. At frequencies somewhat removed from the crossover points, the performance of each of these seven receivers should be comparable with that obtained if a given one of the receivers were connected directly to an antenna. Likewise, the performance of 21 receivers connected to the "decoupled jacks" should be comparable with the performance of 3 receivers decoupled in like

manner, using conventional patch panels (with a 600 ohm isolation resistor between jacks) on a given antenna.

Transmitting Filter Assemblies

The filters that will now be described have been designed for use with shipboard transmitters and antennas that are unbalanced to ground. This series includes Electrical Filter Assemblies F-159/SRT, F-160/SRT, F-161/SRT, F-162/SRT, and F-163/SRT.

A representative filter assembly of this series, the F-162A/SRT, is shown in figure 7-8. Each filter assembly consists of low-pass and high-pass subassemblies. They are used with transmitters operating in the frequency range between 175 kc and 27 mc. Single filter assemblies (fig. 7-9) can be used alone or combined with other subassemblies to operate two or more
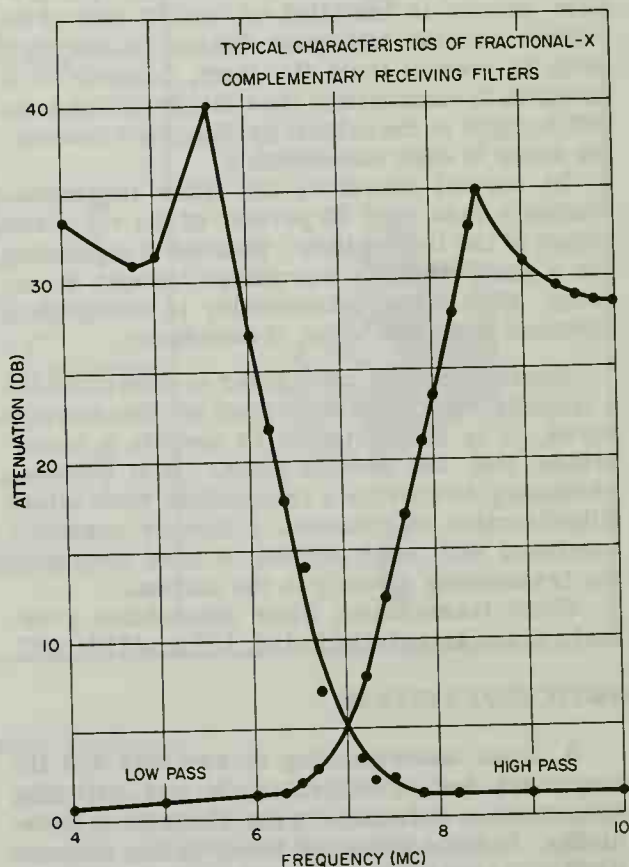
109.43

Figure 7-10.—Filter curves.

227

transmitters (using different frequencies) into a single antenna.

Because a low-pass filter and a high-pass filter are used in each filter assembly, there is a definite crossover frequency for each filter. In the vicinity of the crossover frequency there is a band of frequencies, called the precluded band, which cannot be used because appreciable power would be lost in the filter. When frequencies in the vicinity of the crossover frequency are to be used, it is necessary to change the crossover point. By removing the crossover point around as the precluded band is approached, it is possible to obtain continuous coverage of the 175-kc to 27-mc frequency range. A switch in the center of the front panel on all assemblies except the F-160/SRT, permits shifting of the crossover frequency.

Figure 7-10 shows curves for both low-pass and high-pass filters for the F-162A/SRT.

The filter assemblies are constructed in the form of plug-in drawers. Normally, the low-pass drawer is installed on the left side of the cabinet and the high-pass drawer on the right side, as viewed from the front, however it is possible to interchange the subassemblies from left to right in the cabinet by changing a connecting strap in each subassembly.

In normal operation, the filter assemblies dissipate less then 20 percent of the r-f power output of the transmitter. Because it is possible for a fault, such as a grounded antenna, to develop, each filter subassembly is designed to dissipate over 500 watts, if necessary.

Because the low-pass filter section provides a termination for the high-pass section and vice versa, it is not advisable to operate a transmitter into one section alone. If it becomes necessary to operate a transmitter when either filter section is removed, a jumper assembly furnished with each cabinet is used to connect the transmitter directly to the antenna.

Other transmitting filter assemblies presently in use include the F-156, 157, and 158/URT.

SWITCHING SYSTEMS

A basic understanding of switches and the functional use of switchboards and switching systems is a necessity for the Electronics Technician. In some instances, the switching systems used with electronic equipment are more difficult to understand than the operation of the equipment itself. In general, however, although switching systems are becoming more complex,

the methods of operation are being simplified. Switchboards make use of multisection switches that are not difficult in themselves to understand, but the entire switching function performed by switchboards, or by combinations of switchboards, may be somewhat involved. In this chapter a typical receiver switchboard and a typical transmitter switchboard are treated.

Simplified diagrams, showing how switchboards are connected to receivers, transmitters, radiophone units, and accessory equipment, are included. For completeness, block diagrams of the switching circuits between Radio Central, Radio II, and Radio III in one type of installation, are also included.

Finally, the more complex radar data switching system is described with the aid of block diagrams and simplified switching circuits.

Bear in mind that in many of the switching systems described there are a large number of possible circuit arrangement. In this chapter a great deal of simplification is employed; and in general, only the less complex arrangements are included. Enough basic information is included however, to give the prospective ET- the
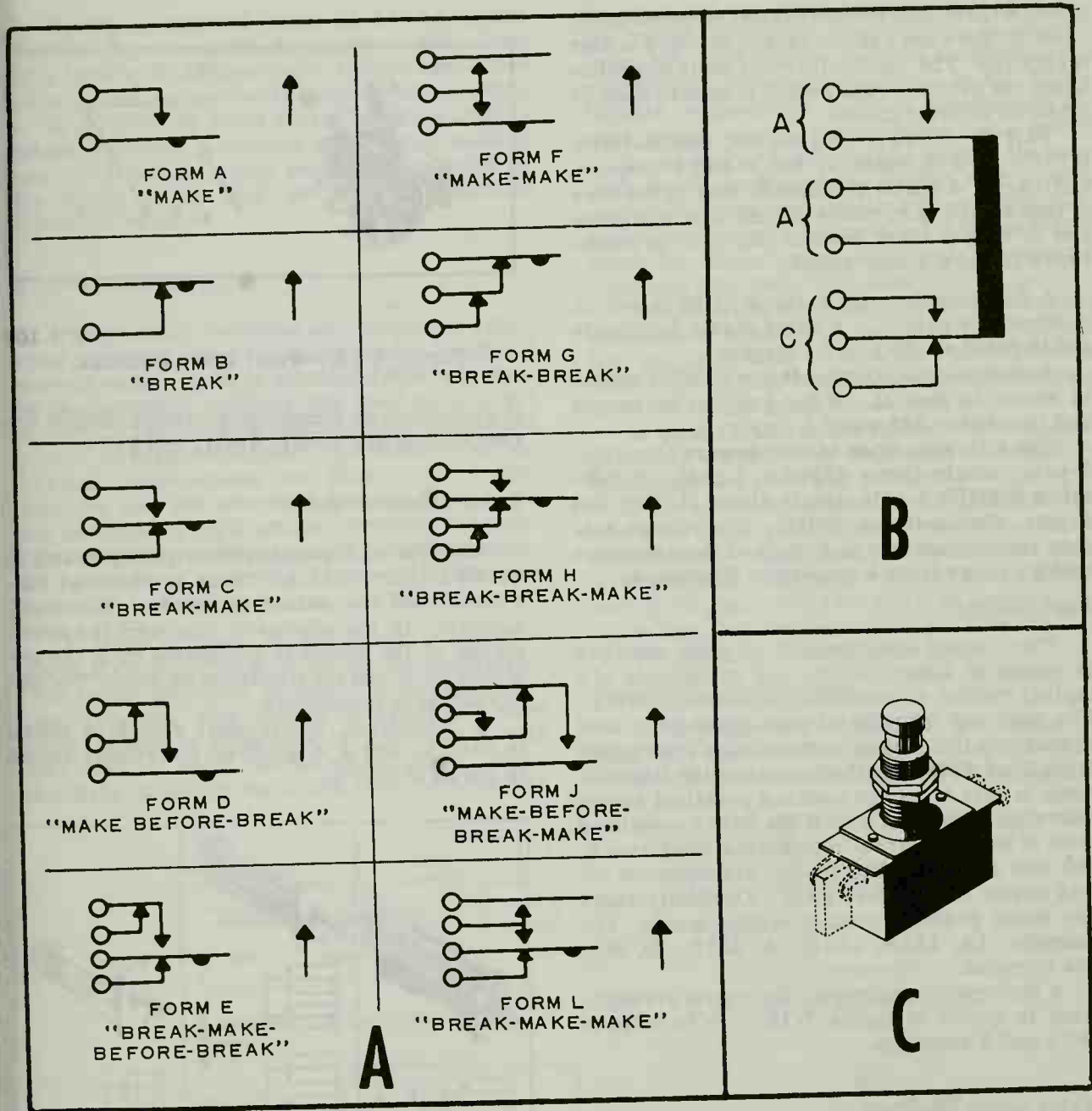

1.98
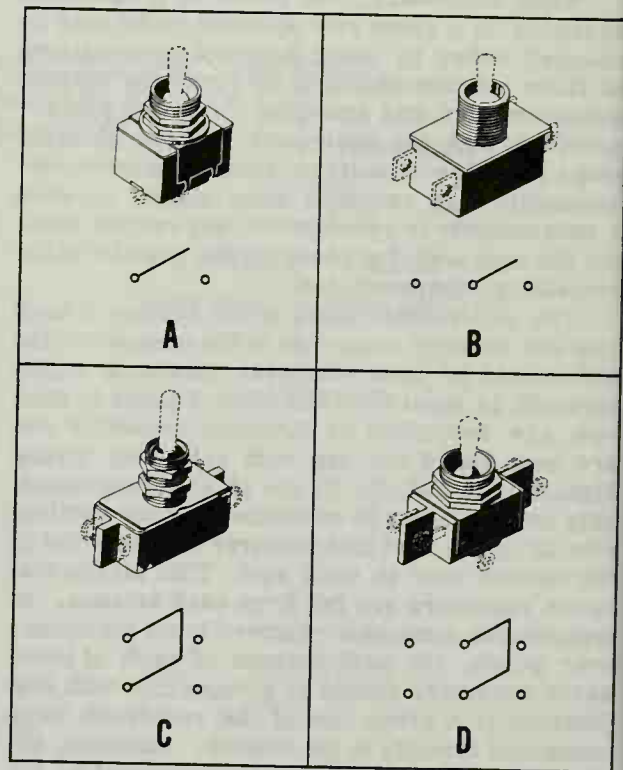
Figure 7-11.—Toggle switches.

FORM A "MAKE"

FORM F "MAKE-MAKE"

FORM B "BREAK"

FORM G "BREAK-BREAK"

FORM C "BREAK-MAKE"

FORM H "BREAK-BREAK-MAKE"

FORM D "MAKE BEFORE-BREAK"

FORM J "MAKE-BEFORE-BREAK-MAKE"

FORM E "BREAK-MAKE-BEFORE-BREAK"

FORM L "BREAK-MAKE-MAKE"

A

B

C

1.99

Figure 7-12.—Push switches.

necessary background for further study on the subject.

TYPES OF SWITCHES

Some of the more common types of switches used with electronic equipment are illustrated and described briefly in this portion of the chapter. There are many variations of each type of switch; however, only a few representative examples are included here.

Toggle Switches

Representative examples of toggle switches are shown in figure 7-11. In part A is shown

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a single-pole, single-throw (SPST) toggle switch, rated at 200 v and 2 amperes, and having 2 solder terminals. The schematic diagram is shown beneath the switch. This switch is used to open or close an electric circuit.

Part B shows a single-pole, double-throw (SPDT) switch, rated at 250 v and 1 ampere, and having 4 screw terminals. One of the uses of this switch is to turn a circuit on at one place and to turn it off at another place. It is sometimes called a 3-way switch.

A double-pole, single-throw (DPST) switch is shown in part C. It has 4 solder terminals and is rated at 250 v and 1 ampere.

A double-pole, double-throw (DPDT) switch is shown in part D. It has 6 solder terminals and is rated at 125 v and 3 amperes.

The following types of switches are also used: 3-pole, single-throw (3PST); 3-pole, double-throw (3PDT); 4-pole, single-throw (4PST); and 4-pole, double-throw (4PDT). The voltage ratings range from 20 v to 600 v, and the amperage ratings range from 1 ampere to 30 amperes.

Push Switches

The contact arrangement of push switches is shown in figure 7-12A, and an example of a typical contact arrangement is shown in part B. The type and quantity of each basic form used to make up the contact assembly are determined from Part A. Part B illustrates how the illustrations in part A may be used in a practical switch assembly. Thus, in part B the switch contains a total of three separate basic forms: two forms A, and one form C. The contact arrangement for this switch is therefore 2A1C. Obviously, there are many possible contact arrangements. For example, 1A, 1A1A, 1A1B, 2A, 2A1B, 1B, etc., are common.

A push switch enploying a 2A contact arrangement is shown in figure 7-12, C. It is rated at 250 v and 3 amperes.

Wafer Lever Switches

A wafer lever switch is shown in figure 7-13. It is a double-pole, triple-throw (DP3T) type of switch rated at 110 v and 0.150 ampere. It locks in position and is nonshorting, that is one circuit is opened before the next circuit is closed. In switches of this type the action may be locking or momentary, and the contacts may be shorting (for example, one circuit remains closed until an instant after the next circuit is closed; then
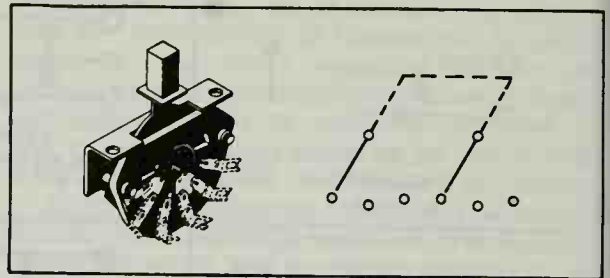
1.100

Figure 7-13.—Wafer lever switches.

it is opened) or nonshorting. Other contact arrangements are DPDT, 4PDT, and 6P3T.

Lever Pileup Switches

One type of 2-position lever pileup switch is shown in figure 7-14,A. There are 9 solder terminals, and the switch is rated at 48 v and 1 ampere. In the schematic diagrams the downstroke of the switch is designated by 2, the upstroke by 1, and OFF position by zero. The No. 2 position is momentary.

A 3-position, 21-terminal switch is shown in part B, and a 2-position, 9-terminal switch is shown in part C.
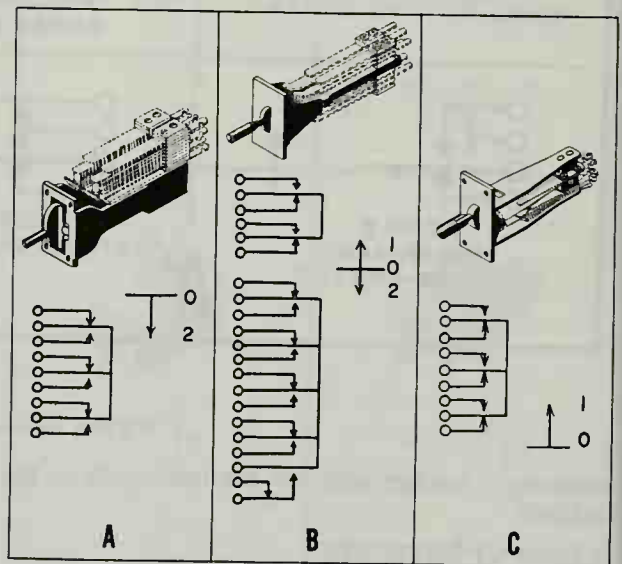
1.101

Figure 7-14.—Lever pileup switches.

230

Nearly a hundred types of lever pileup switches are available for various uses. They may have up to 75 terminals, and the associated switch contacts may be arranged in various ways.

In addition to lever pileup switches, rotary pileup switches (activated by a rotary motion) and the jacktype pileup switches (activated by the thrust of the plug) are used in automatic telephone systems.

Thermostatic Switches

Thermostatic switches are designed to either open or close when the temperature reaches a certain value. A large number of different types of thermostatic switches are used by the Navy to control the temperatures in compartments and rooms, to regulate dampers, to maintain constant crystal temperatures in some radio transmitters; they are also used in many other heating and cooling applications. Switches of a given type may have different contact arrangements, be operated at different temperature range, or have different voltage or current ratings.

A large number of thermostatic switches employ a bimetallic strip as the active element. The basic operating principle is illustrated in figure 7-15A. One side of the bimetallic strip is brass and the other side is iron (other metals may also be used) welded to the brass. When the strip is heated, the brass expands more than the iron and the strip bends downward to open the switch contacts. Thus, power is removed from the load when a certain temperature is exceeded. Although not shown in the figure, various refinements such as adjustments and snap action may be incorporated in the switch.

In mercury thermostatic switches the mercury itself completes the circuit as it expands upward between two metallic contacts. The basic operating principle is illustrated in figure 7-15,B. When the temperature decreases, the mercury contracts and opens the circuit.

The principle of operation of gas thermostatic switches is illustrated in figure 7-15,C. When the gas is heated, it expands the bellows and closes the switch; when it is cooled, it contracts the bellows and opens the switch.

In each of the illustrations, only the basic principle of operation is shown. Certain refinements are generally added.

Figure 7-15,D, shows the thermostatic switch used in the crystal oven of one type of radio transmitter. The contact arrangement is for a single-pole, single-throw operation, and the switch is rated at 115 v and 0.75 ampere. The operating temperature is 170.6° F to 179.6° F, the temperature differential being 9°.

Rotary Pileup Switches

Rotary pileup switches are so constructed that they open and/or close one or more electrical circuits; the contacts are arranged in a

A BIMETALLIC THERMOSTATIC SWITCH

B MERCURY THERMOSTATIC SWITCH

C GAS THERMOSTATIC SWITCH

D THERMOSTATIC SWITCH USED IN RADIO TRANSMITTERS

1.103
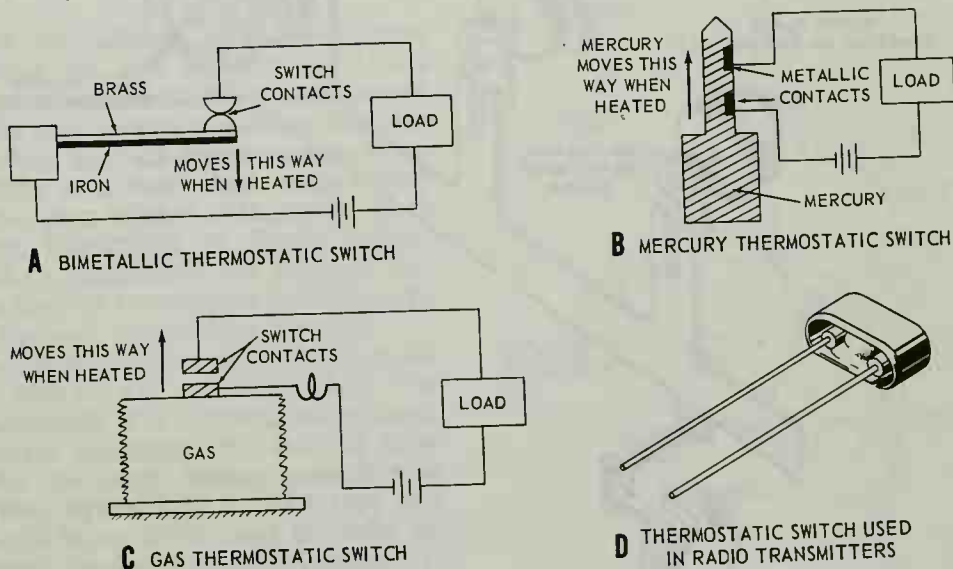
Figure 7-15.—Thermostatic switches.
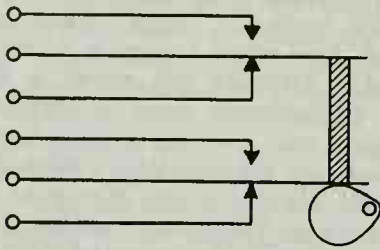
1.104

Figure 7-16.—Rotary pileup switches.

leaf, or pileup, fashion and they are actuated by a rotary motion.

One type of rotary pileup switch is illustrated in figure 7-16. As may be seen in the figure,

there are six terminals. When the armature is moved upward by the rotary motion of the switch knob, two circuits are opened and two other circuits are closed.

This type of switch has numerous applications in low-voltage signal circuits.

Rotary Selector Switches

Rotary selector switches, or nonpileup rotary switches, have many applications in electronic equipments. They may be made up of any number of sections, decks, or wafers depending upon the switching functions required. There are hundreds of possible contact arrangements.
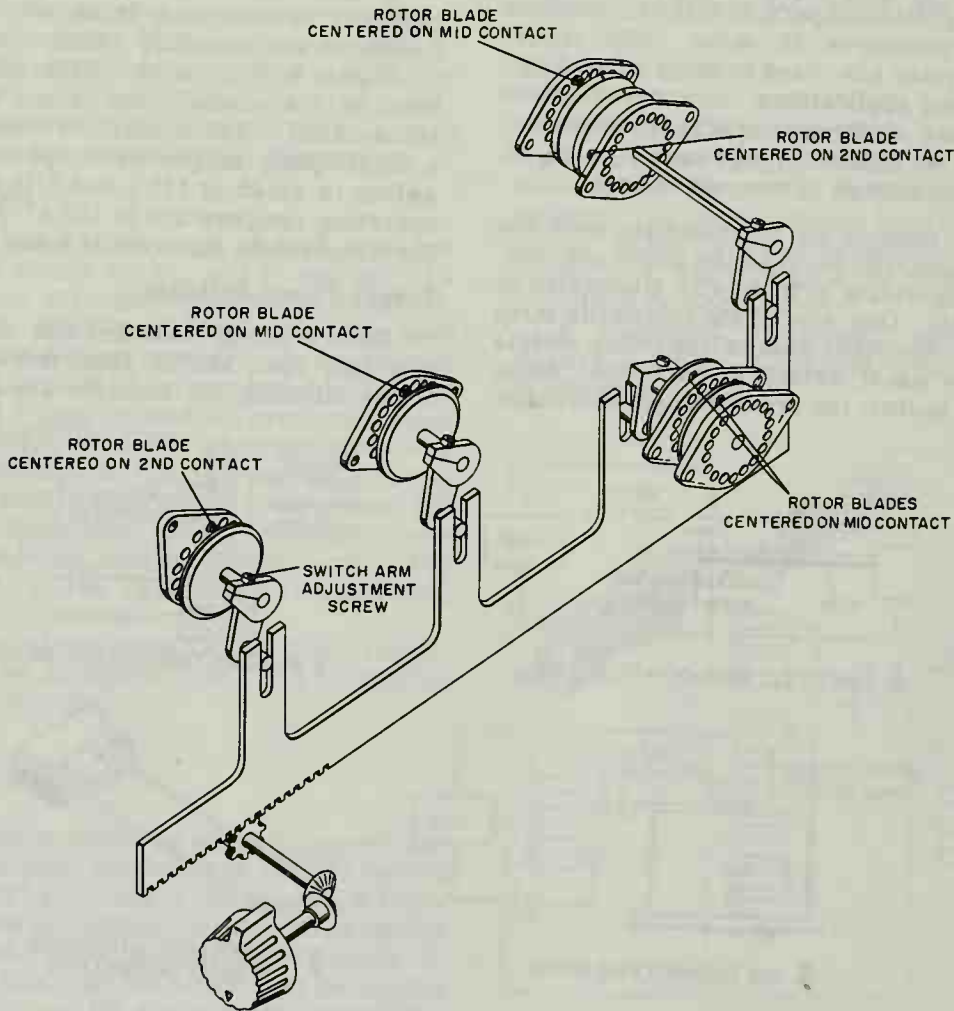
Figure 7-17.—Rotary selector switches operated by gear and sliding bar.

70.62

In many applications the switches are mechanically ganged to operate simultaneously from a single control (fig. 7-17).

One deck of a rotary selector switch is shown in figure 7-18. The code letters are also included at the leads extending from the various terminals. Short clips are indicated by the letter, X. Dummy lugs are indicated by the letter, D. Nonshorting rotary teeth (or blades) are indicated by crossed lines. The word, nonshorting, means that the width of the rotor tooth is less than the distance between adjacent contact clips. This means that as the rotor is turned, one circuit will be opened before the next one is closed. The shorting type is shown without the crossed lines. In this case the rotary tooth is wider than the distance between adjacent clips, and therefore as the rotor is turned, one circuit is closed before the preceding one is opened. Clips that are insulated from their associated lugs are indicated by the letter, S; long clips thus insulated are indicated by the letters, YS. When a clip is thus insulated, the lug may be connected to a clip on the reverse side. A through electrical connection is indicated when the rotor (or a portion of it) is shown in black. This means that the section of the rotor so marked in the figure is connected through to a bottom rotor, not shown.

Telephone Type Jacks

The contact assemblies on telephone type jacks are divided into two categories: (1) the plug contact assembly and (2) the pileup contact assembly. The plug contact assembly includes the contact springs (and all others making electrical connection with them) that make direct contact with the plug when it is inserted. Both categories are illustrated in figure 7-19,A.

Figure 7-19,C, illustrates the common varieties of plug contact assemblies used on jacks and indicates the code designation for each.

The proper contact arrangement for jacks is determined with the aid of figure 7-19,B. The plug contact assembly (J7) is distinguished from the pileup contact assembly (B). and the code designation for the plug contact assembly is determined from figure 7-19,C. The type and quantity of each basic form used to make up the pileup contact assembly are determined. In the case of part B of the figure the proper contact arrangement designation for the jack is "J7-1B."

Jacks are used in plug panels, teletype panels, phone units, and in many other applications.

RADIO RECEIVER BAND-SWITCHING CIRCUITS

Band-switching circuits in radio receivers commonly employ rotary selector switches. In order to trace the various circuits through these switches, it is necessary to pay careful attention to the schematic diagrams. The band-switching circuit of a receiver that is relatively easy to follow is included in figure 7-20.

The 4-position switch, S101B, in the input grid circuits of the oscillator tube, V105, selects 1 of 4 crystals. An additional set of contacts on S101B provides for grounding the three crystals that are not in use at any given time. The operating crystal is connected directly to the grid of V105. In the switch position, 1, shown in the figure, crystal X106 is connected between the grid of V105 and ground; the other three crystals are shorted to ground. In position 2, X107 is connected to the grid, and crystals X106, X108, and X109 are grounded.

The 4-position switch, S101A, mounted on the same shaft as crystal-selector switch S101B,
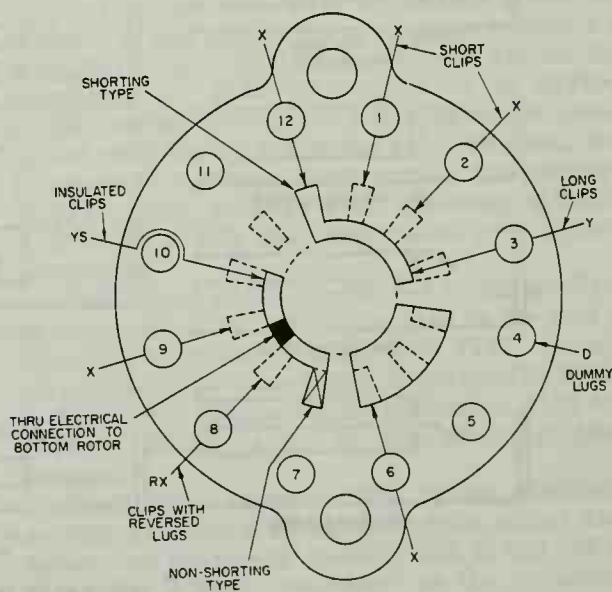
1.105

Figure 7-18.—Deck of rotary selector switch, showing contact arrangement and code letters.

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connects filament potential to the proper channel-indicating lamp through S102. In the position shown in the figure, S101A provides circuit continuity for lamp (1); the other three lamps are open at the switch contacts.

Rheostat R117 enables the operator to control the brilliancy of the channel-indicator lamp and the dial lamps E109, E110, and E111.

The band-switching circuits of the AN/SRR-11 (fig. 7-21) are somewhat more involved, however, careful circuit tracing will enable one to follow the signal path.

Band switching for radio receiving set AN/SRR-11 is accomplished through the use of four 2-section, and two single-section rotary selector switches. The 5-position band selector on

A TELEPHONE-TYPE JACK HAVING PILE-UP AND PLUG CONTACT ASSEMBLIES

B TYPICAL TELEPHONE-TYPE JACK

J1 – TIP AND SLEEVE

J6 – TIP, SLEEVE AND EXTRA SLEEVE CONTACT

J10 – TIP, RING AND SLEEVE

J2 – TIP, RING AND SLEEVE

J7 – TIP, RING AND SLEEVE

J11 – TIP AND SLEEVE

J3 – TIP AND SLEEVE

J4 – TIP AND SLEEVE

J8 – TIP AND SLEEVE

J12 – DOUBLE JACK, TIPS AND SLEEVES

J5 – TIP AND SLEEVE

J9 – TIP, RING AND SLEEVE

J13 – TIP AND SLEEVE, EXTRA SLEEVE CONTACT

C BASIC PLUG CONTACT ASSEMBLIES

Figure 7-19.—Telephone-type jacks.

1.106

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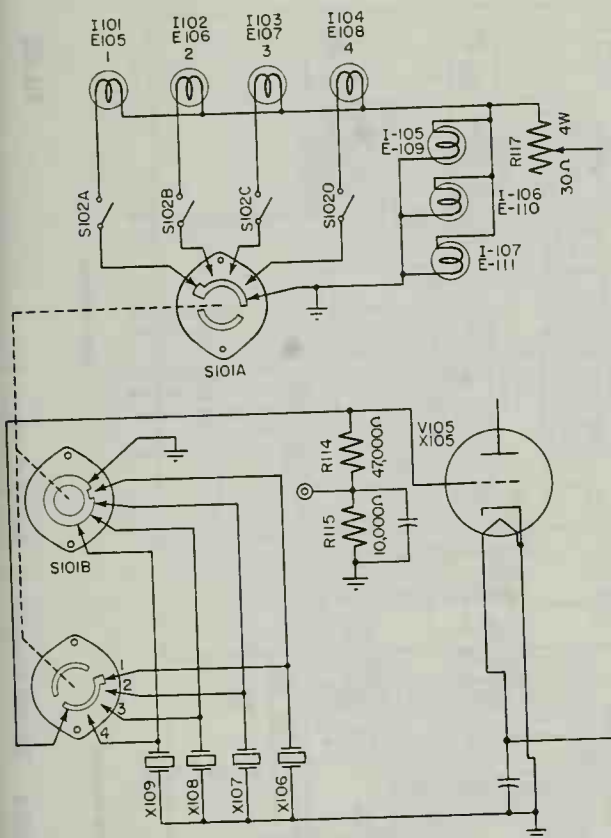
1.107

Figure 7-20.—Band-switching circuit of
a receiver.

the front panel of the receiver is geared to a sliding bar similar to figure 7-17, and operates the selector switches to connect the appropriate components for the frequency band selected.

These selector switches connect the appropriate input and output transformers for the selected band, and connect the ganged sections of tuning capacitor C251 across the selected transformers to maintain tuning within the band. They also serve to short out transformers used for bands lower than the selected band, thereby preventing any absorption of the signal on the band in use.

Antenna matching links 0101 and 0102 (fig. 7-21) are provided to match antenna transformers T101 through T105 to either a high or low impedance input. With the links in the low impedance position the antenna is connected across the primary of the selected transformer

with capacitor C106 connected between the center tap and the low side of the secondary. With the links in the high impedance position (as shown in fig. 7-21), the antenna is connected between the tap and the low side of the selected transformer secondary.

Antenna compensation switch S102 and capacitor C103, operated by the antenna compensation control on the front panel of the receiver, compensate for variations in antenna capacitance.

The wafers of all selector switches (fig. 7-21) are shown in band 1 position. The signal from the antenna is through C102, matching link 0101, C107 to S101A, and out of S101A via contact 3, through R101 to pin D of T101 in the first r-f amplifier section. The output signal voltage developed by the autotransformer action of T101 is fed back into S101A through contact 15 and out to the grid of the first r-f amplifier V101. S101A also connects the antenna section of tuning capacitor C251D across the secondary of T101. S101B is used to short out transformers associated with lower frequency bands than the selected band, and therefore is not used in band position 1.

The output of V101 (terminal H on J103) is fed to S126A in the second r-f amplifier section, out of S126A via contact 9 to pin B on the primary of r-f coupling transformer T126. The output from pin D on T126 is fed into S126A via contact 3, and out to the grid of the second r-f amplifier V126. C251C is connected across the secondary of T126 by contact 15 of S126A. S126B serves to short out transformers associated with bands below the selected band, and to connect capacitor C139 across the primary of T127, T128, or T129 when operating on bands 2, 3, or 4. S126B, therefore, is not used in band position 1.

The output of the second r-f amplifier V126 at terminal H on J128 is coupled to the grid of mixer V151 through transformer T151. Transformer T151 is connected in the circuit by S151A in the same manner as described for T126 and S126A. S151B also is not used for band position 1.

Oscillator assembly Z201 in the oscillator section generates 60 kc above the signal frequency on frequency bands 1 and 4, and 200 kc above the signal frequency on the remaining bands. Oscillator signals for band 1 operation are from pin D on L201, into S201A via contact 9, out of S201A through R201, to the grid of V201. Oscillator feedback is from pin E on

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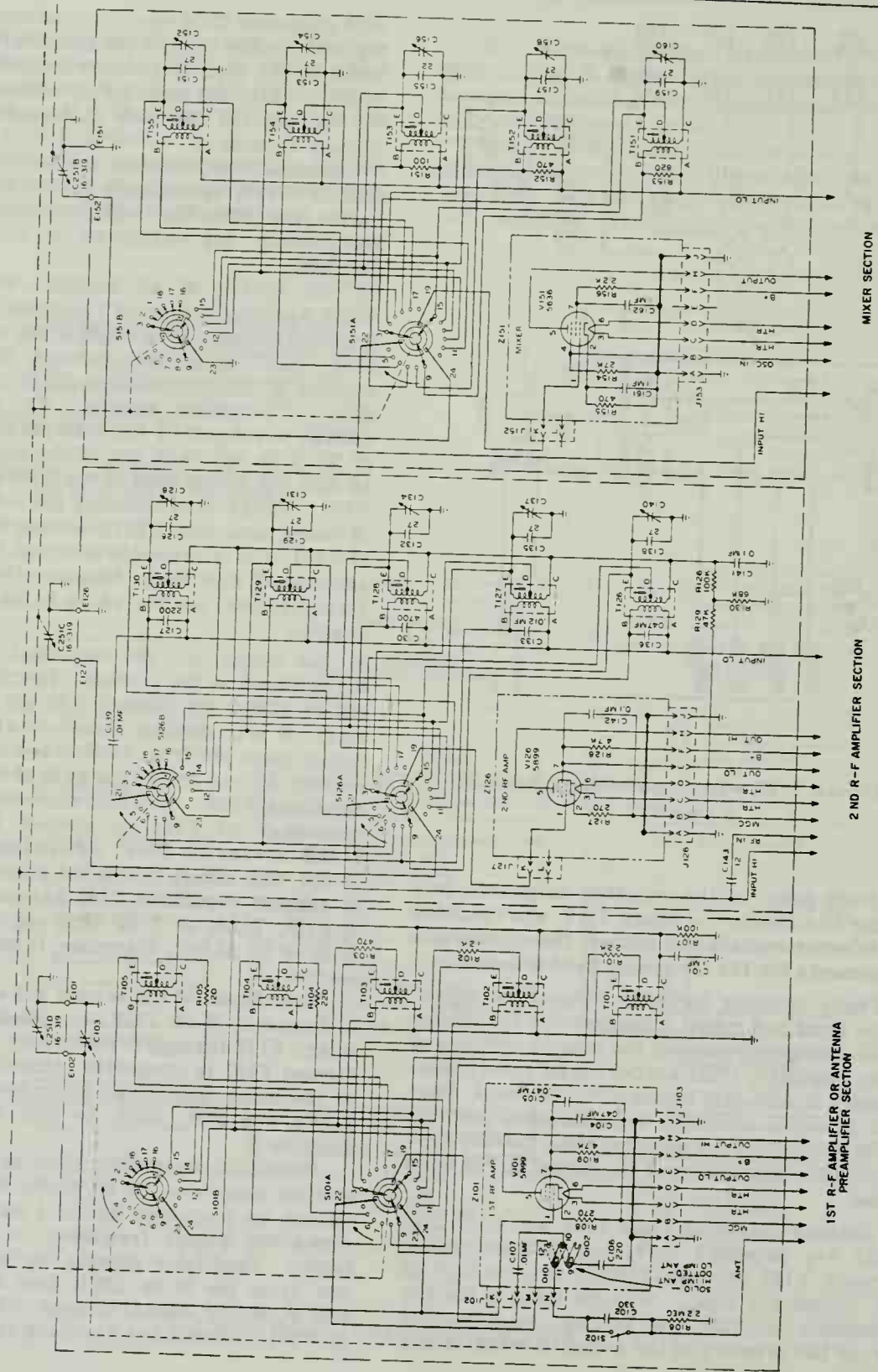
Figure 7-21A.—Band-switching circuits – Radio Receiving Set AN/SRR-11.
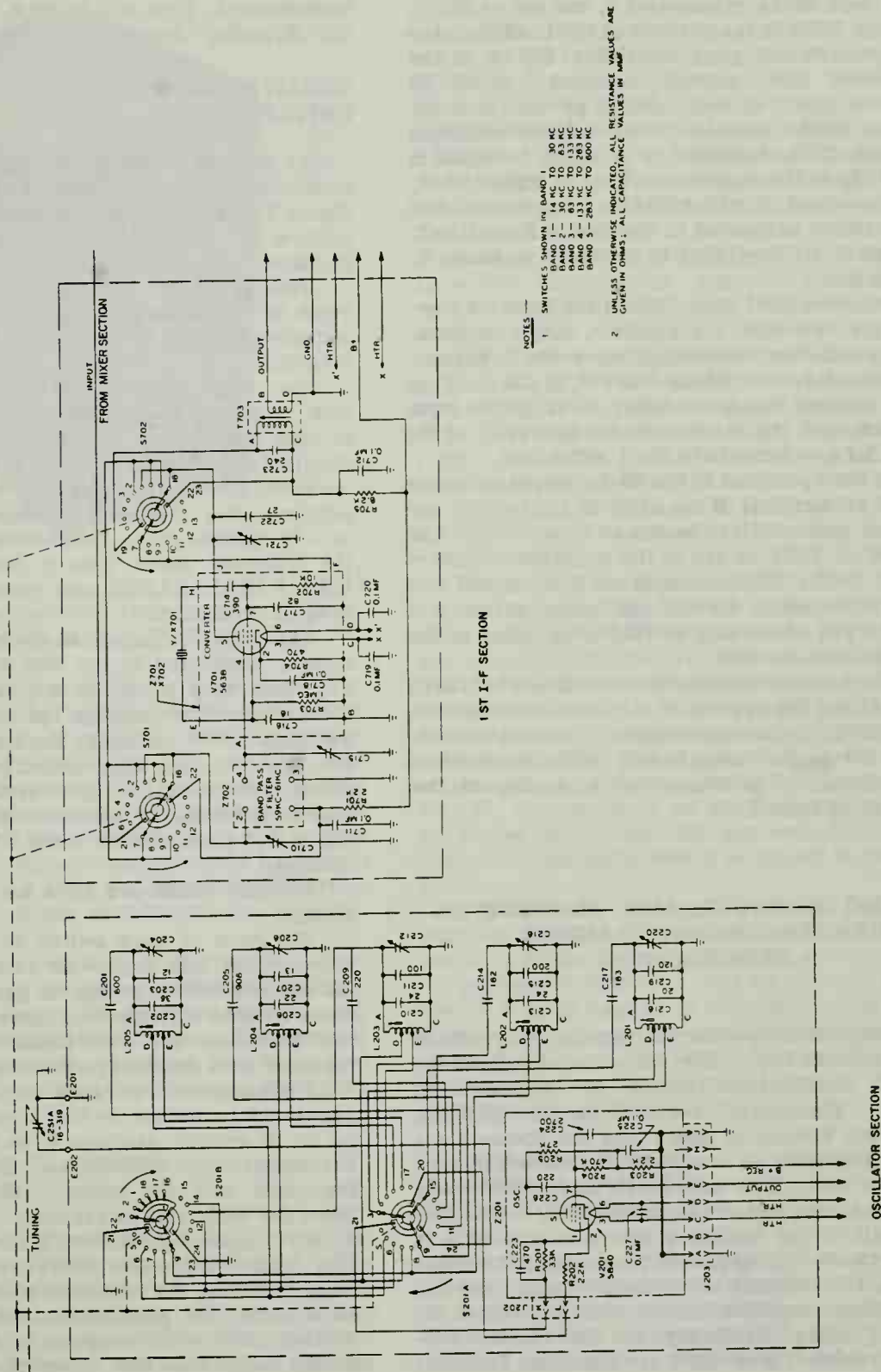
Figure 7-21B.—Band-switching circuits – Radio Receiving Set AN/SRR-11.—Continued.

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L201 into S201A via contact 3, and out of S201A through R202 to the cathode of V201. S201A also connects tuning gang capacitor C251A in the oscillator tuned circuit. Contact 9 on S201B grounds pin D of coil L202 to prevent absorption of band 1 signals. The oscillator output is through C226, terminal E on J203, terminal B on J153, to the suppressor grid of mixer V151.

The output of mixer V151 during bands 1 and 4 operation is applied to the first i-f amplifier. The first i-f amplifier is bypassed on bands 2, 3, and 5.

Switches S701 and S702 in the first i-f section are two-pole, six-position, single-section, rotary selector switches. They serve to bypass the first i-f assembly in bands 2, 3, and 5. They also connect bandpass filter Z702 in the input circuit, and the B+ supply for operation of the first i-f assembly in bands 1 and 4.

In band position 1, the 60-kc output of mixer V151 at terminal H on J153 is fed to S701 and out on contact 16 to bandpass filter Z702. The output of Z702 is fed to the suppressor grid of V701. The 200-kc output of V701 is fed into S702 via contact 16 and out to the primary of T703. The secondary of T703 is the output to the second i-f assembly.

The circuits for band position 2 may be traced by rotating the arrows in all wafer switches one position in the direction shown. To trace the circuits for the remaining bands, rotate the switches the number of positions corresponding to the band to be traced.

RADIO RECEIVER AND TRANSMITTER REMOTE-CONTROL TRANSFER SWITCHBOARDS

Radio remote-control transfer plug panels have become too cumbersome to be used in shipboard radio installations or modern Navy ships. Therefore, control panels utilizing switches instead of plugs and patchboards are being installed in new construction and in conversion jobs. Two unit-constructed panels (one for receivers and one for transmitters) now provide all of the facilities that were available in three types of plug panels (the receiver transfer panel, the transmitter transfer panel, and the radiophone transfer panel), and in addition afford greater flexibility in the remote-control system. These units are Receiver Transfer

Switchboard, Type SB-82/SRR, and Transmitter Transfer Switchboard, Type SB-863/SRT.

RECEIVER TRANSFER SWITCHBOARD

An external view of the Receiver Transfer Switchboard, Type SB-82/SRR, is shown in figure 7-22. A simplified schematic diagram of a portion of the internal switching arrangement is shown in figure 7-23.

The receiver switchboard has 5 vertical rows of 10 double-pole, single-throw (ON-OFF) switches that are continuously rotatable in either direction.

One side of each switch within a vertical row is wired in parallel with the same sides of the other nine switches within that row. Similarly, the other side of each switch is wired in parallel horizontally with the corresponding sides of each of the other four switches in a horizontal row. This method of connecting the switches gives rise to the term CROSS-MAT PARALLELING, and permits a high degree of flexibility.

The audio output from the receiving equipments, connected to the five vertical rows of switches, may be fed to any or all of the remote stations by closing the proper switch or switches. For example, the audio output from the L-F receiver may be fed to remote radiophone station No. 1 by closing switch 16; it may be fed to all of the remote stations by closing all of the switches in the left-hand vertical column.

Shielded leads are used for the audio circuits.

The knob of each switch is marked with a heavy white line to provide visual indication of the communication setup. In general, there are more remote stations than there are receivers, and therefore the switchboards are normally mounted in a vertical position (as in fig. 7-22). This arrangement permits the outputs from 5 receivers to be fed to the 5 vertical rows and up to 10 remote stations to be fed from the 10 horizontal rows of switches. Switchboards are furnished with the knobs in the OFF position when the white line is vertical.

If it becomes necessary to employ a system where there are 10 receivers and 5 remote stations, the switchboard may be mounted in a horizontal position and each switch knob rotated 90° with respect to the shaft. All shafts have two flat sides for setscrews so
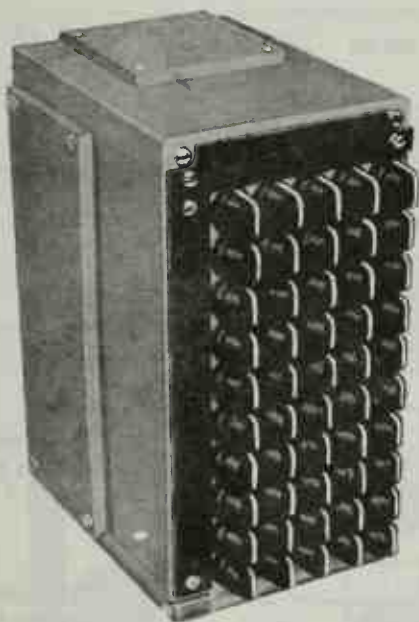
36.69

Figure 7-22.—External view of the Receiver
Switchboard type SB-82/SRR.

that this change can be made if desired. The
purpose of rotating the switch knobs with respect
to the shaft is to standardize the setup so that
the switches will be in the OFF position when
the white line is vertical. To further standardize
all installations, receivers are always connected
to the vertical rows of switches, and remote
stations are always connected to the horizontal
rows of switches after the orientation (vertical
or horizontal) of the switchboard has been de-
termined.

It should be noted that only the receiver audio
output circuit is connected to the switchboard.
This is true also of the receiver transfer plug
panel used in earlier installations. Transmitter
transfer switchboards, however, handle several
other types of circuits in addition to audio cir-
cuits.

TRANSMITTER TRANSFER
SWITCHBOARD

Transmitter Transfer Switchboard type SB-
863/SRT has replaced type SB-83/SRT. The
SB-863/SRT (fig. 7-24) has ten 20-position
rotary selector switches, in two vertical
columns. Each rotary switch corresponds to a
remote control station, and each switch position

(1 through 19) corresponds to a controlled
transmitter. Thus switching control is pro-
vided for up to 10 remote control stations, and
19 transmitters. When more than 10 remote
stations, or 19 transmitters, are to be con-
nected, additional transfer switchboards may be
installed. Position 20 of each rotary switch is
provided for connections to an additional trans-
fer switchboard to control extra transmitters.
The switches consist of 12 wafers which serve
to connect the start-stop indicator, keying,
12-v d-c microphone, carrier control, and
carrier indicator circuits for the various trans-
mitters.

Any of the remote stations may be connected
to control any of the transmitters in the system.
For example, to connect remote station No. 1
(fig. 7-25) to control the TED transmitter,
rotary switch No. 1 (S1) is placed in position 1.
To control the AN/GRC-27 from remote station
No. 1, switch S1 is placed in position 2. Switch
2(S2) is used for remote station No. 2 in the
same manner.

The remote stations assigned to each rotary
switch, and the transmitters assigned to posi-
tions 1 through 19, are engraved on the engrav-
ing plates when the switchboard is installed
(fig. 7-24). If an extra switchboard is installed,
switch position No. 1 on the second switchboard
is assigned to transmitter No. 20, position No.
2 to transmitter No. 21, etc. Thus, if remote
station No. 2 is to have control of transmitter
No. 22, switch No. 2 on the first switchboard
is placed in position 20, and switch No. 2 on
the additional switchboard is placed in position
3.

Figure 7-26 illustrates possible intercon-
nections between the various switchboards lo-
cated in Radio Central, Radio II, and Radio III.

Increased flexibility may be obtained by the
use of a large number of switchboards in Radio
Central and a smaller number of switchboards
in Radio II and Radio III. Most of the space
on two bulkheads may be taken up by the
switchboards in Radio Central.

SWITCH CONNECTIONS
BETWEEN RADIOPHONE
UNIT AND TRANSMITTER

Generally, a radio transmitter connects
through the transmitter transfer switchboard
to a radiophone unit. However, in figure 7-27
the transmitter transfer switchboard has been
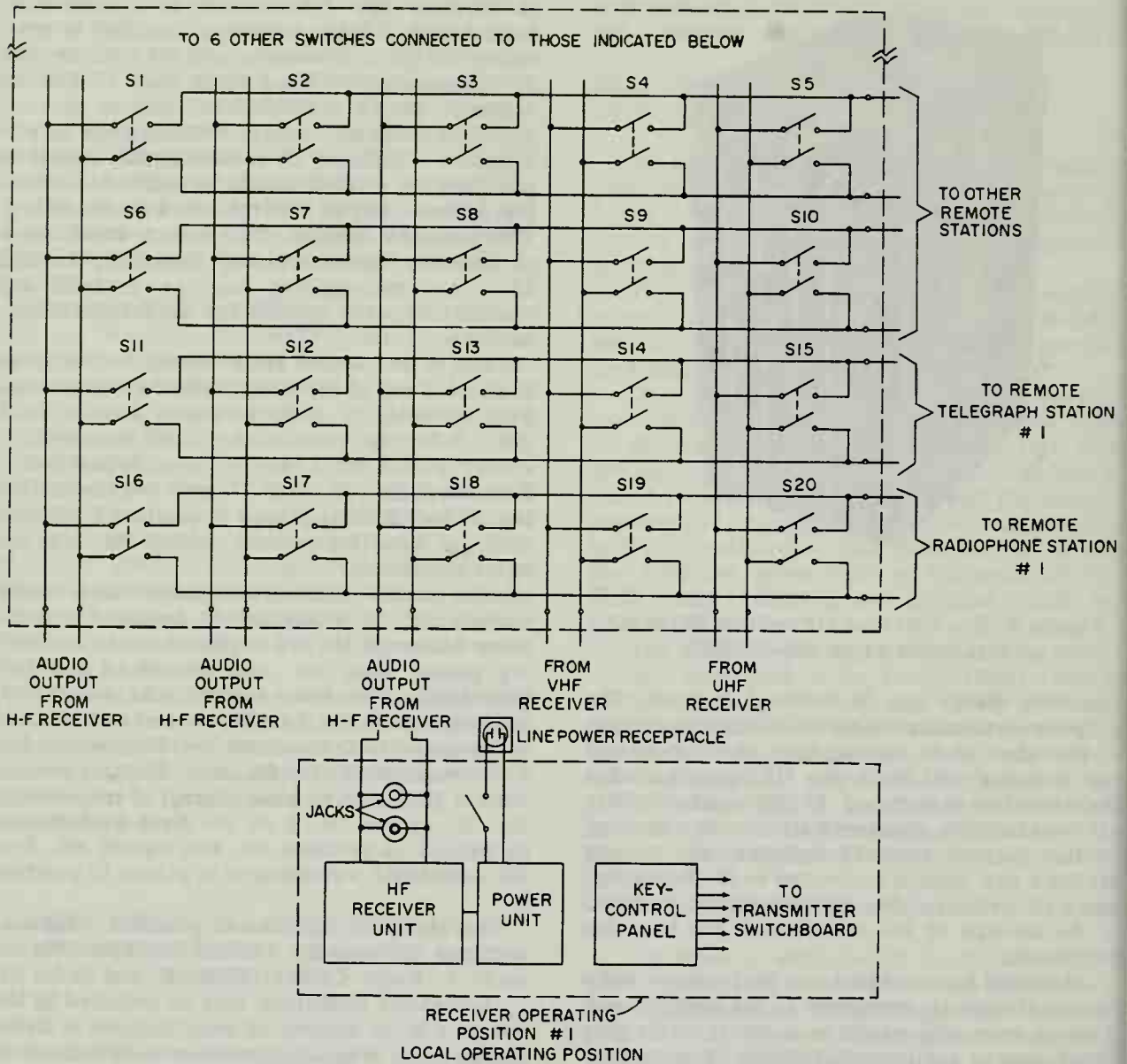omitted and only a simplified diagram of the

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Figure 7-23.—Simplified schematic diagram of the receiver transfer switchboard type SB-82/SRR.

transmitter (Navy Model TED-8), showing the control circuits, is included.

The line power switch, S1, is a 2-pole, single-throw switch in the 60-cycle, single-phase, power-line circuit and must be closed before the equipment can be started with the start-stop switch at the transmitter or RPU. In an emergency, power can be quickly removed from the entire equipment by throwing the line power switch to the EMERGENCY OFF position.

The local start-stop switch, S2, in the transmitter is a momentary contact (nonlocking) switch used to start or stop the equipment after S1 has been closed. Momentary closure of S2 on the START side energizes the equipment start-stop relay, K1, which locks itself in the closed position through contacts K1A,
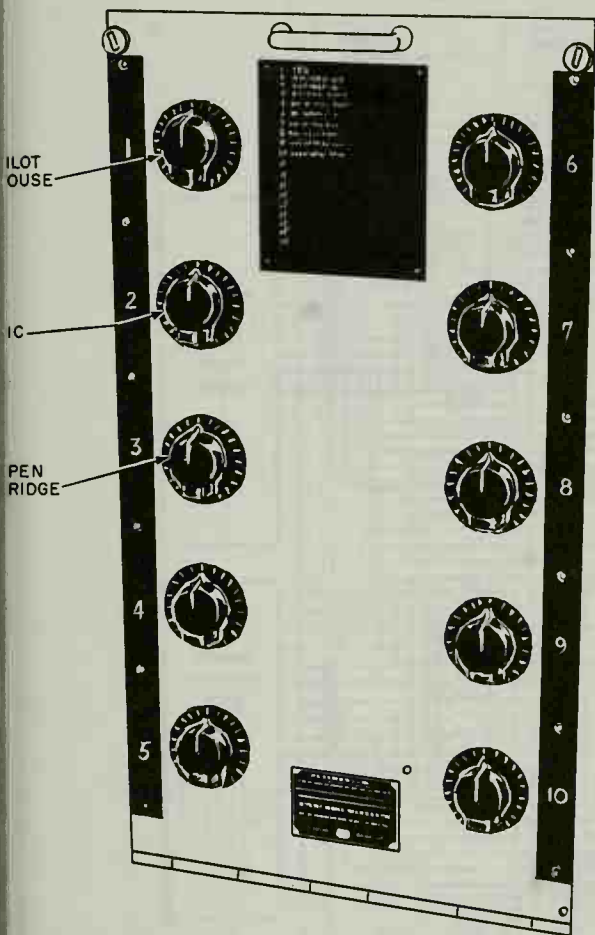
70.64

Figure 7-24.—Transmitter transfer
switchboard-SB-863/SRT.

he thermal cutout relay contacts, and resistor
l. Relay K1 remains energized until its coil
s shorted when S2 is moved to the STOP posi-
ion.

The local remote switch, S3, affords local
or remote control of the operation of the trans-
mitter. When S3 is set to LOCAL, S3(a) con-
nects the parallel combination of the local
handset, the microphone jack, and the local
carrier-control switch in series with the input
windings of the input transformer, T1 (winding
1-2 in series with winding 3-4), thus providing
ocal control of the carrier. Indicating and
control functions are disconnected from all
remote units by operation of S3(a), S3(b), S3(c),
and S3(d).

When S3 is turned to the REMOTE position,
the following sequence of events takes place:

1. Switch S3(a) connects the T2 secondary
and the microphone or key jack in the RPU
as a parallel combination across the transmitter
T1 primary (1-2 winding in series with the
3-4 winding) and disables the local carrier-
control circuits. The circuits through the hand-
set microphone in the RPU are from the -12-v
supply (terminal 7), the T2 primary, terminals
C and D of the handset extension jacks by way
of the push-to-talk switch (not shown in the RPU)
and ground return at terminal 8. The circuits
through the handset receiver in the RPU are
from terminals A and B of the handset extension
jacks, the earphone-level potentiometer, and
terminals 13-14 to the associated receiver via
the receiver transfer switchboard (not shown).

2. Switch S3(b) connects the common lead
to the start-stop circuit to the stop switch in
the RPU.

3. Switch S3(c) connects one side of the
115-v a-c line to the RPU power indicator by
way of terminal 4. The other side of the 115-v
line is always connected directly to the other
side of the power indicator.

4. By means of S3(d), -12 v is applied to
the RPU for microphone current.

The audio monitoring circuit involves the
circuits connected to terminals 5, 6, 10, 13, and
14 on the terminal board. The audio monitoring
circuit enables the operator to listen to the
output of the receiver when the transmitter is
off and to listen to speech "sidetone" when the
transmitter is being used. Circuits are pro-
vided so that this may be done either locally
or at a remote location. The audio signal
from the receiver is brought into terminals 13
and 14.

When the transmitter carrier is off, relay
contacts K2(a) are closed (K2 is deenergized),
and the audio signal from the receiver is fed
via K2(a) to the earphone-level volume control
at the transmitter and to the 7-8 winding of T1.
The receiver may be monitored locally by
plugging the earphones into the earphone jack
at the transmitter, because this jack is con-
nected across a part of the earphone-level
volume control.

When the transmitter carrier is on, relay
K2 is energized and contacts K2(a) are open.
This condition disconnects the receiver audio
signal circuit from the earphone-level control
at the transmitter, the associated earphone jack,
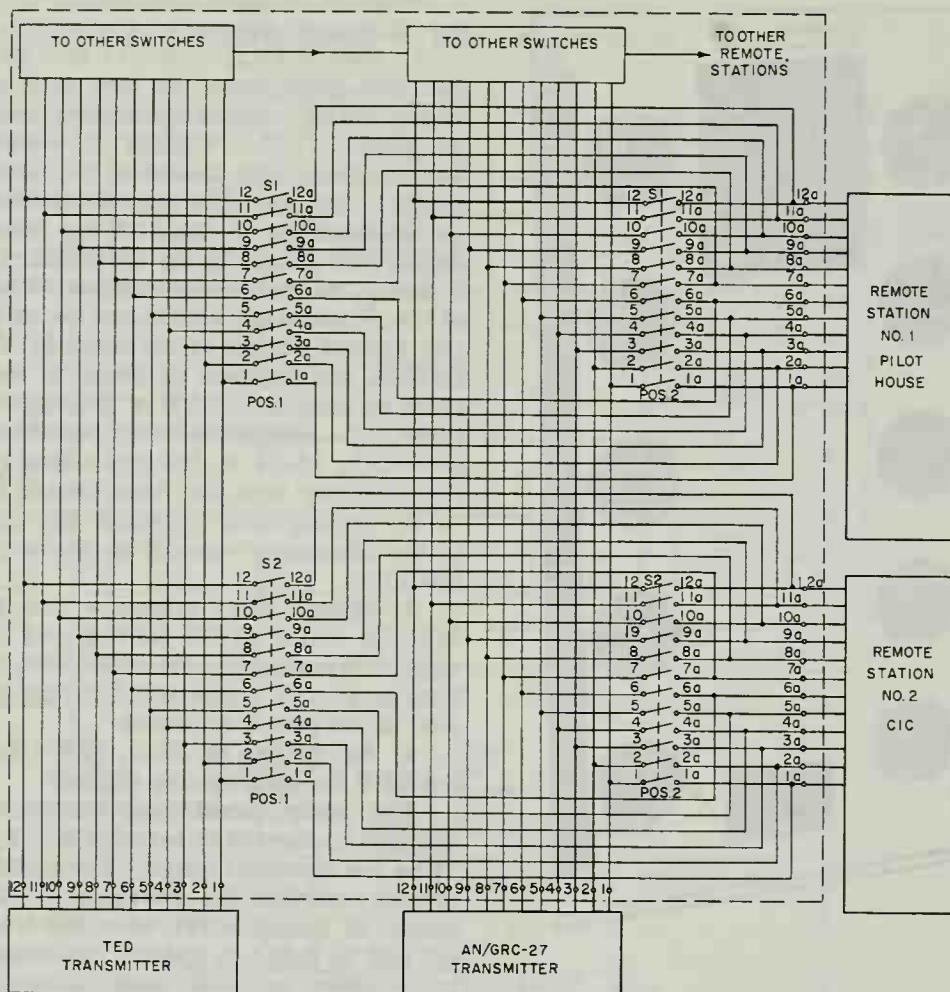the receiver handset plug terminals A and B,

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Figure 7-25.—Simplified schematic diagram of transmitter transfer switchboard SB-863/SRT, showing first two positions of rotary switches 1 and 2.

and the 7-8 winding of T1. The transmitter sidetone signal is picked up by the secondary, 7-8, winding of T1 and impressed across the earphone-level control for local monitoring. The remote monitoring signal is picked up from the control lines on which the signal is normally present during voice operation.

When S5 is in the PHONE position and the local microphone handset press-to-talk switch is operated, keying relay K3 is energized. This action indirectly energizes carrier control relay K2, which disconnects the receiver from T1, as previously described. Thus, the receiver is disconnected when the transmitter carrier

is on. The same action takes place in remote operation, except that relay K4 in the RPU energizes K3 by applying ground to terminal 3 of the 3-4 winding of T1 via terminals 8 and 11 and contacts K4(a).

DISTRIBUTION OF RADAR INFORMATION

The distribution of radar information is somewhat involved, and complex switching equipment is needed. There are various types of switching gear and various combinations of radar equipments and radar repeaters.

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Figure 7-26.— Interconnections between transfer switchboards.

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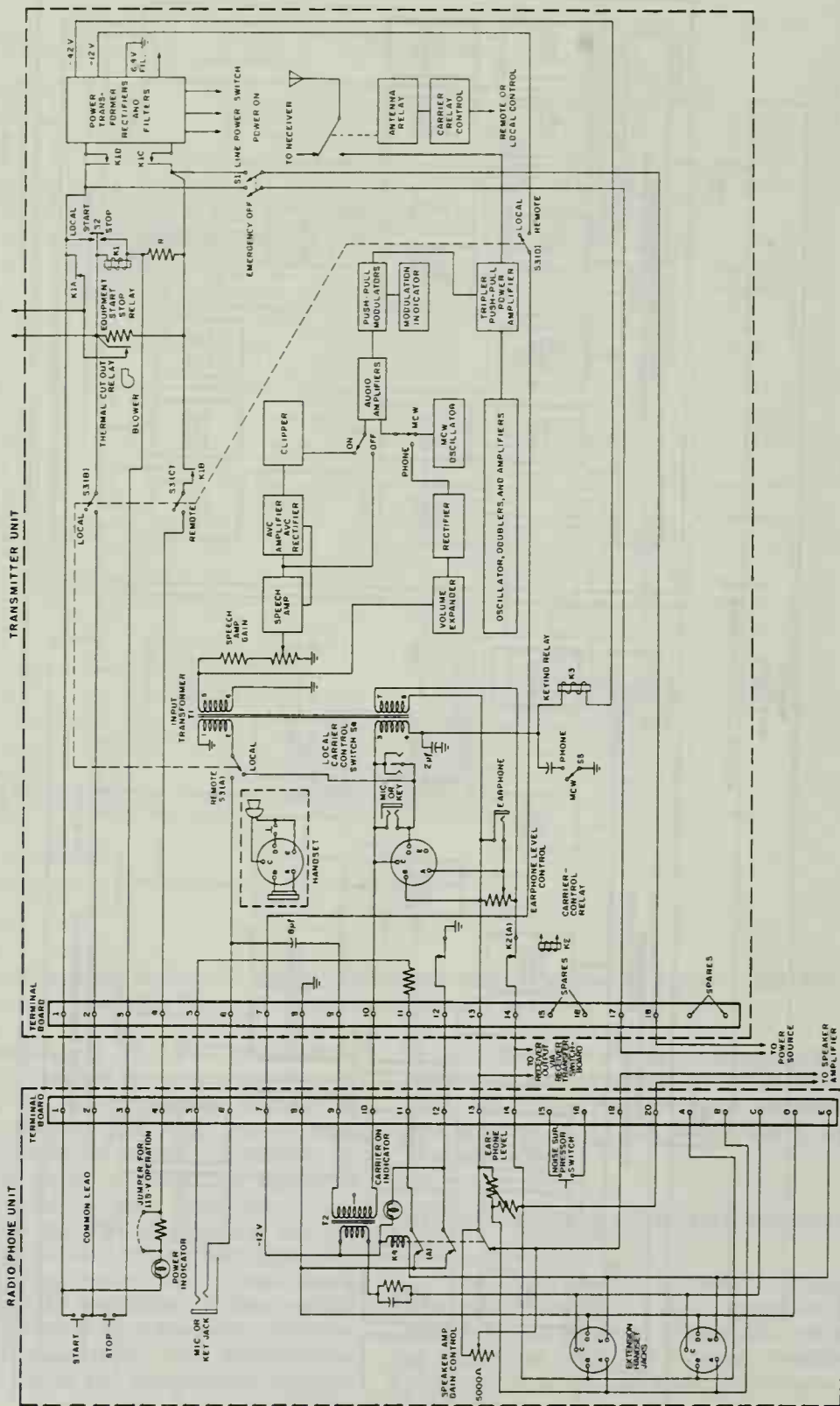
Figure 7-27.— Switch connections between radiophone and transmitter.

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For the purpose of this chapter, a brief discussion of Data Switching Group OA-496/SSA is included. Included with this switching group is Video Amplifier Assembly AM-518/SSA, Rotary Switch SA-243/SSA, and Rotary Switch SA-247/SSA (fig. 7-28). For remote servo operation, Rotary Switch SA-243/U (not shown) is also needed at other repeaters.

As indicated in the block diagram (fig. 7-29), the purpose of the equipment is to distribute radar information to remote indicators throughout the ship. The number of data switching groups actually required in any particular installation depends on the number of signal sources and the number of repeaters to be served. Only a relatively simple system is shown in this figure.

Each of the 10 radar selector switch units in the data switching group provides for the selection (remote or local) of any one of seven

AMPLIFIER ASSEMBLY AM-518/SSA

DATA SWITCHING GROUP OA-496/SSA   ROTARY SWITCH SA-243/SSA   ROTARY SWITCH SA-247/SSA
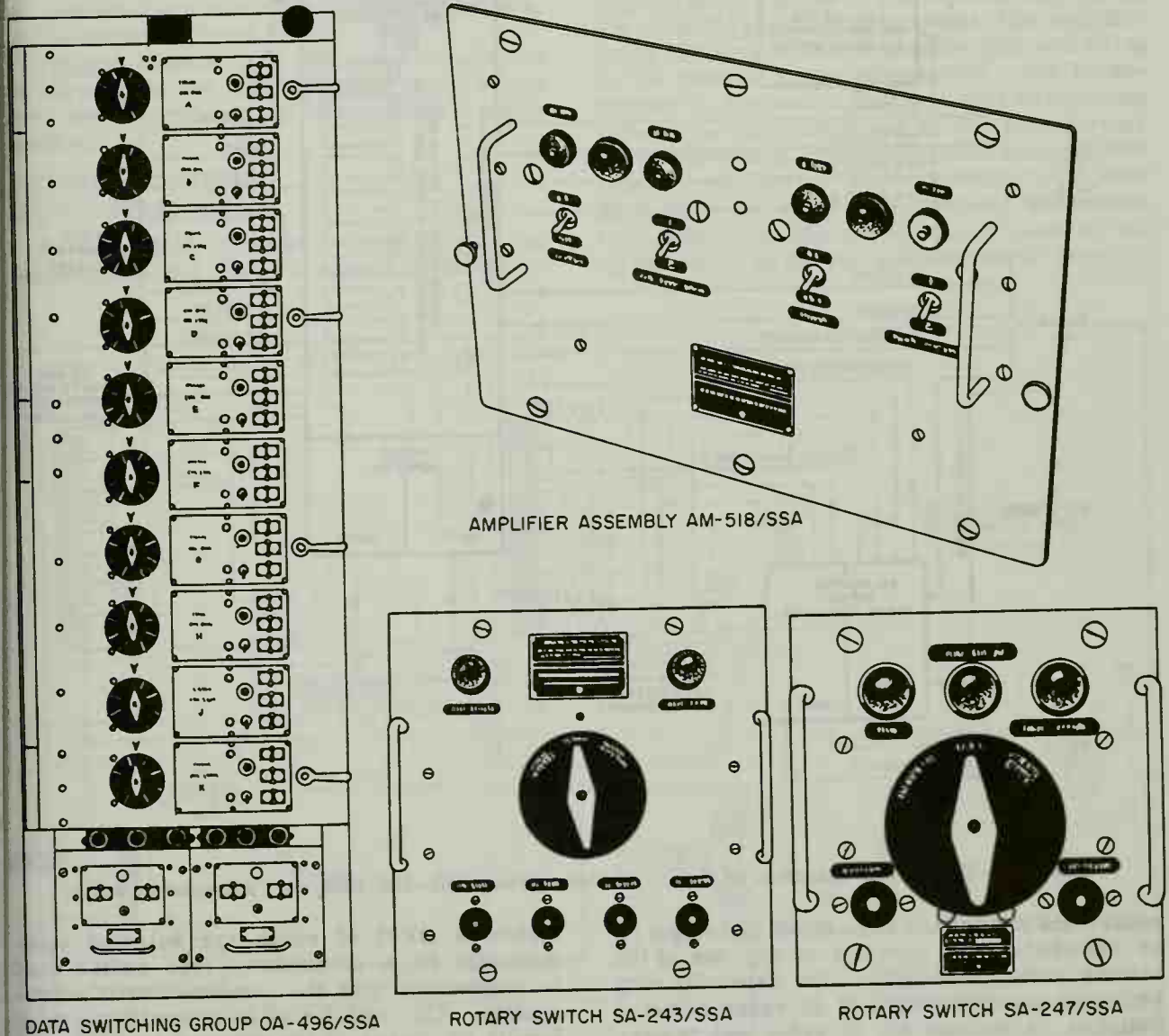
70.68

Figure 7-28.—Data switching group OA-496/SSA, Amplifier Assembly AM-518/SSA, and Rotary switches SA-243/SSA and SA-247/SSA-front panels.
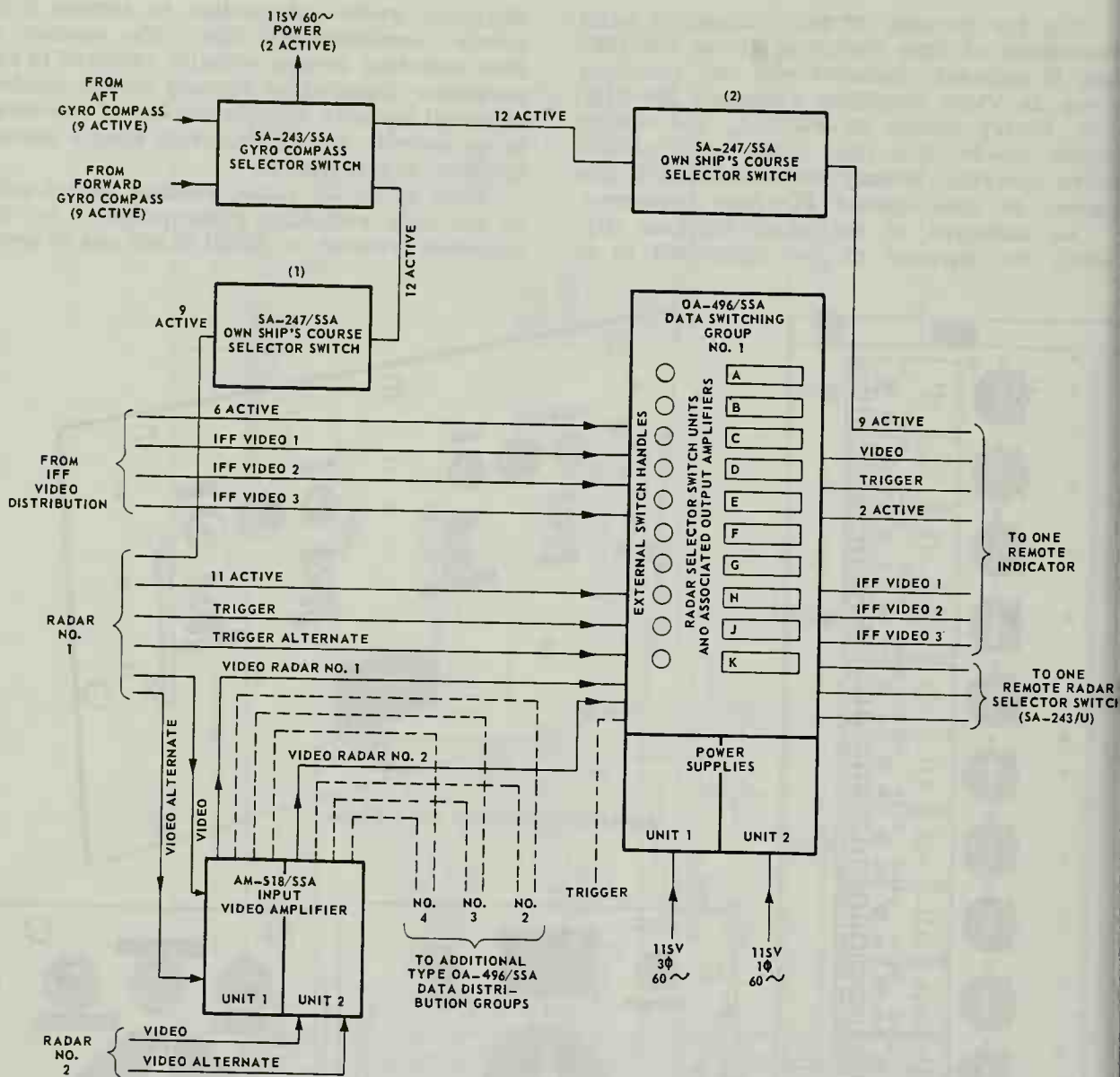
Figure 7-29.—Block diagram of data switching group OA-496/SSA, and associated units.

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radar-data inputs (only 2 are shown in the figure), as selected by the operator at any one of ten remote radar repeaters. The assembly data switching group consists of 10 radar selector switches (A through K), 10 video and trigger cathode follower subassemblies, and 2 power supplies.

Each selector switch has 25 data sections (decks), the 8 positions (7 active positions; the eighth is OFF) of which are selected automatically by a commutator-type switch used in connection with the remote-control servo-system. The switch may be manually operated locally by means of the external switch handle or remotely by means of the remote selector switches. Each switch section has seven input circuit contacts and one output contact from the common arm.

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A 3-section video and trigger switch is installed on the back end of the 25-section switch.

Video signals from the radar receivers are fed through Video Amplifier Assembly AM-518/SSA to the data switching group. This amplifier is capable of providing an essentially flat frequency response for any input signal over a range from 100 cycles through 10 mc. The assembly consists of two amplifiers, each providing for one regular and one alternate input, and four outputs so that the input signal may be delivered to four data switching groups, one is shown in figure 7-29).

Trigger voltages and IFF video signals are fed directly to the data switching group where they are distributed to the desired remote location.

SWITCHING-CONTROL SERVOSYSTEM

A simplified schematic diagram of the switching-control servosystem is shown in figure 7-30. The system consists of a 3-phase motor, which drives the 8-position, rotary type switch through a gear train; and a split phase motor, which controls the operation and direction of rotation of the 3-phase motor by means of switch S502.

Two multitapped, single-phase autotransformers, one in the data switch selector servo unit and one in the remote selector switch unit, provide the required control power to the split-phase motor. These transformers are energized from 1 phase of the 3-phase supply. At the remote selector switch unit, the operator turns the switch to the position that will bring in the desired radar information. The transformer in the switch unit develops an opposing voltage to that developed by the transformer in the servo unit. The voltage contributed by the transformer in the servo unit is zero when S503 is in the H, or OFF position. The potential difference (the potential not balanced out by the action of the two transformers) is applied

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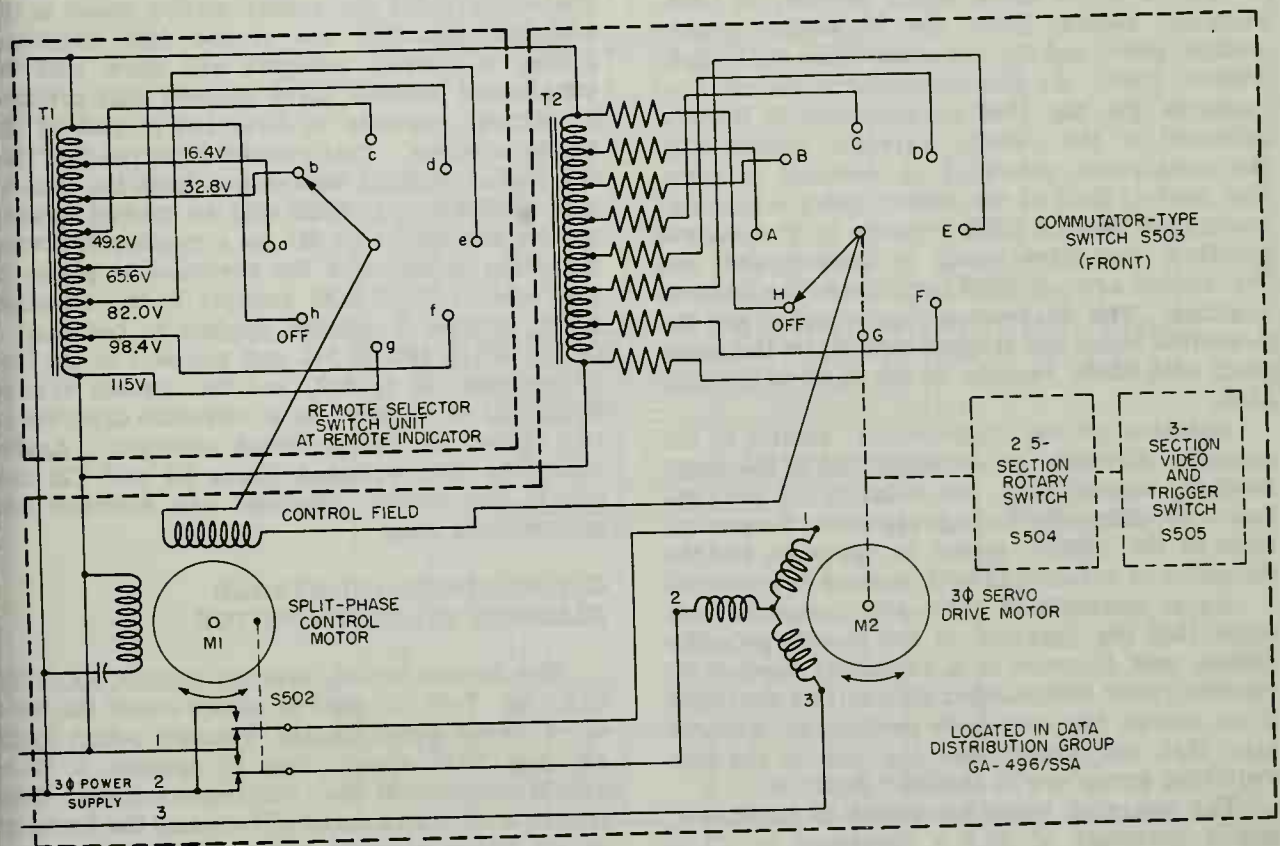
Figure 7-30.—Switching-control servo system.

to the control winding of the split-phase motor and causes it to develop a torque in one direction or the other, depending on the condition of unbalance. A vane on the shaft of the control motor actuates the sensitive switches, S502 (upward or downward against spring action), depending upon the direction in which the torque is developed. The switches apply 3-phase power to the 3 windings of the 3-phase drive motor. The drive motor, M2, drives S503 to zero the voltage on the control field of M1, thereby removing the torque developed by M1 and opening S502 to stop the drive motor.

The drive motor is reversed by reversing the torque on M1 to operate S502 in the opposite direction. This action interchanges two line leads to M2 to reverse its direction of rotation. The reversal of torque on M1 is brought about by the reversal of the control field of M1 with respect to the constant field of M1. The relative phase of the control field depends upon the relative magnitudes of the voltages at T1 and T2 that are applied to the control field.

The 3-phase drive motor drives the commutator switch, S503; the 25-section rotary switch, S504; and the 3-section video and trigger switch, S505. As the commutator switch arm contacts the tap that corresponds to the one selected at the remote selector switch unit, the unbalanced potential is reduced to zero. The control field of the control motor is then de-energized, switch S502 returns to the neutral position, the drive motor is deenergized, and the switch arm of S503 remains in the desired position. The 25-section rotary switch and the 3-section video and trigger switch, on the same shaft with S503, remain in the desired position also.

Rotation of the commutator switch in the opposite direction is accomplished in the same manner, except that the polarity of the unbalanced difference voltage applied to the control field of the control motor is opposite, and the direction of rotation of both motors is opposite.

As an example of automatic operation, assume that the operator at the remote selector switch unit (located at a repeater) desires to receive radar information that will be available if he moves his switch to position b. Assume also that his switch and the one in the data switching group are in the OFF position.

The operator turns his switch to position b, and a potential of 32.8 v developed by T1 is applied to the control field. The opposing voltage in T2 is zero. The control motor turns (for example, clockwise) because of the phase relationship of the control and the constant fields. The motion of the control motor moves S50 downward (for example) and applies phase to terminal 1 and phase 2 to terminal 2 of the 3-phase drive motor. The drive motor rotates switch S503 through position A. The 16.4 v developed by T2 at position A oppose the 32.8 v developed by T1, but a net potential difference of 16.4 v is still applied to the control field of M1. Motor M1 continues to rotate clockwise and M2 moves the contact of S503 to position B. In this position, the voltages developed by T1 and T2 are equal and opposite and no voltage is applied to the control field of M1. Motor M1 stops, S502 springs open, power is removed from M2, and the contact of S503 remains in position B.

The example just given applies to all cases where the remote controller calls for higher position in the alphabet than the initial position of the setting at the data switching group.

For the opposite case, where the remote controller calls for a dial setting lower in the alphabet than the one at the data switching group, a similar analysis will show that the unbalanced control field voltage will produce a current opposite in direction to that of the example given. This results from the fact that the higher voltage now comes from the taps on T2. Accordingly, S502 will be moved upward by the movement of M1 in a counterclockwise direction because of the reversal in phase of the control field with respect to the constant field. Phase 2 will be applied to terminal 1 of the drive motor M1 and phase 1 of the line to terminal 2 of M2, and the contact arm of S503 will be rotated in a direction opposite to that given in the previous example. Again, when the two voltages (from T1 and T2) are equal, the switch positions are matched and both motors stop.

GYROCOMPASS AND STABLE ELEMENT SELECTOR SWITCH

The Gyrocompass Selector Switch, SA-243/SSA (fig. 7-31) is used to select either the forward or aft gyrocompass synchro output (both 1X and 36X speed) and to provide a 78-v synchro electrical zero reference voltage. This type of switch is also used to select the forward or aft stable element output. The switch consists of a 20-pole, 3-position type switch controlled by a relay. Electrical zero reference
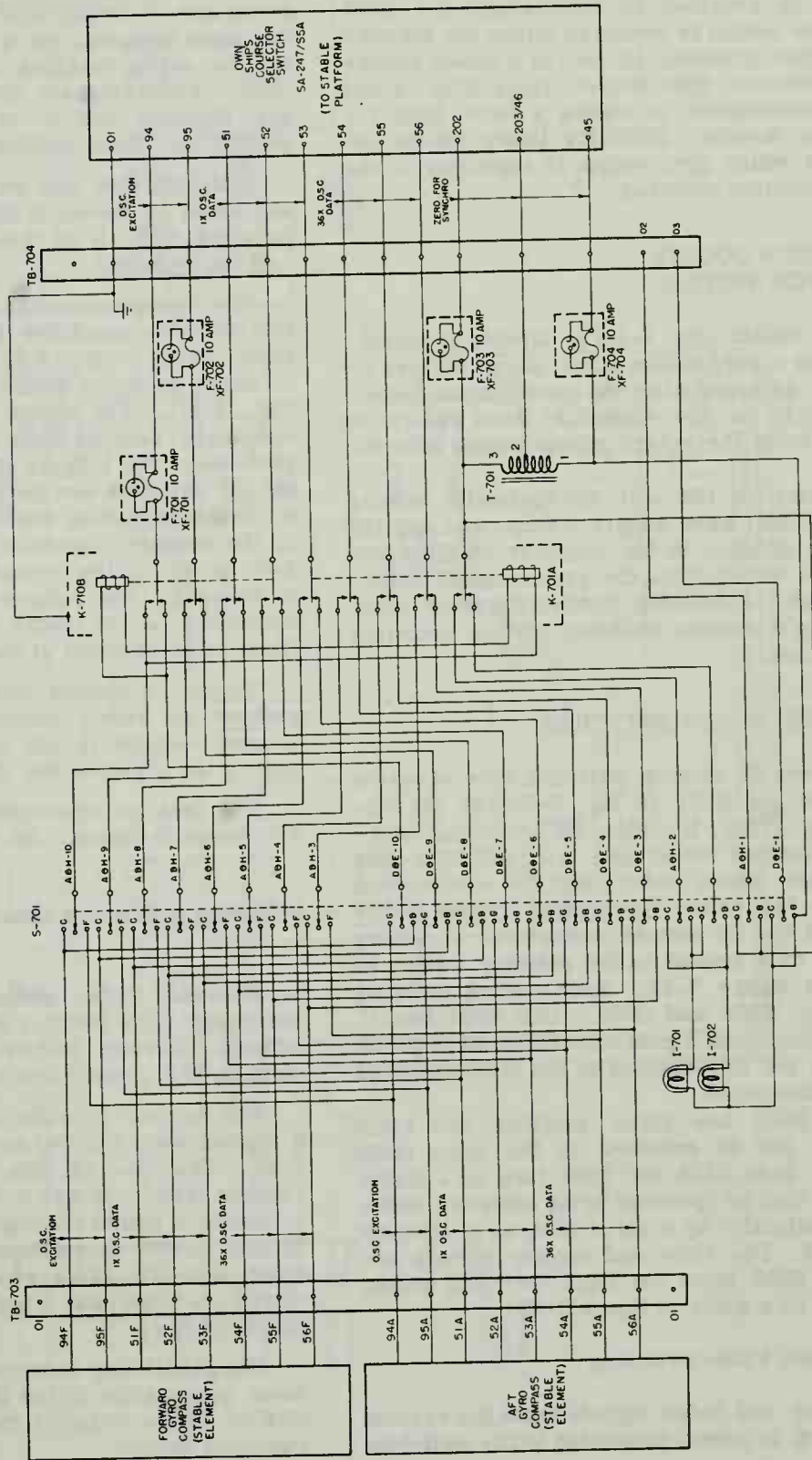
Figure 7-31.— Gyrocompass selector switch.

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voltage is provided by autotransformer T701 when the switch is turned to either the forward or aft gyro position. In case of a power failure of the selected gyro source, relay K701 (A and B) is deenergized, providing a signal from the alternate source. Indicator lights inform the operator which gyro output is appearing on the switch position selected.

OWN SHIP'S COURSE SELECTOR SWITCH

This switch (fig. 7-32) is capable of selecting either a gyro output signal (from the fore or aft gyro, as selected by the gorocompass selector switch) or the electrical zero reference voltage from the rotary gyrocompass selector switch.

A relay in the unit automatically selects the electrical zero supply voltage and applies it to the synchro in the repeater on failure of the gyro output from the gyrocompass selector switch. A flashing light on the front of the own ship's course selector switch indicates gyro failure.

RADAR SELECTOR SWITCHES

The ten 25-section switches (one of which is shown as S504 in fig. 7-33) in the data switching group, transfer the incoming information (except video and trigger) from any one of the seven radars and the information from the IFF video distribution system to any or all of the 10 remote indicators. A wiring diagram of a typical radar selector switch is shown in figure 7-33. Each switch includes two units, S504 and S505. Unit S504 has 25 data sections and 8 positions referred to previously in the introduction to the distribution of radar information.

Unit S505 has three sections and eight positions and is mounted on the same shaft as S504. Both S504 and S505 turn as a single unit and may be operated by an external handle or automatically by a servo unit, as previously described. The video and trigger signals are fed into S505 from the input terminal group, indicated as a block in figure 7-33.

Trigger and Video Switching

Trigger and video signals from the various radars (up to seven) connected to the switching

group are fed to the input terminal group. From the input terminal group, trigger voltages are fed to seven sections of deck section 1 of S505. Video signals from the seven radars are likewise fed to seven sections of deck section 3. Deck section 2 serves as ground.

Although only one switch (made up of S504 and S505) is shown in figure 7-33, it should be recalled that 10 of these switches are wired into the system.

For example, assume that after being amplified in video amplifier AM-518/SSA, the video signal from radar 1 is fed through the input terminal group to switch terminal 1-80 of S505 (fig. 7-33). The same signal is also fed to terminals 1-80 of S505 in the nine remaining switches. The outputs (from terminal 180) of the 10 switches are fed independently through the output amplifier associated with each switch to the repeater associated with each amplifier. Any or all of the repeaters may be switched to radar 1. The radars connected to the other terminals of the radar selector switch may likewise be selected at the repeater.

Trigger switching follows the same general pattern as video switching, except that the trigger voltage is not amplified before it is fed to the input of the data switching group.

The detailed connections of one switch unit are shown in figure 7-34.

Switching of IFF and Other Data

Whereas S505 handles video and trigger switching, S504 handles such additional data as antenna bearing, relative bearing IFF video signals, IFF control functions, etc.

The various circuits feeding into and out of a typical switch (S504) are illustrated in figure 7-35. The type of data passing through the various terminals are written on the terminals to simplify signal tracing. The parallel nature of the circuit connections should be kept in mind, and the same reasoning that was employed in video and trigger switching may be employed here.

The paralleling of switch contacts bearing the same information in the 10 switches is accomplished in the parallel circuits section of the switching group.
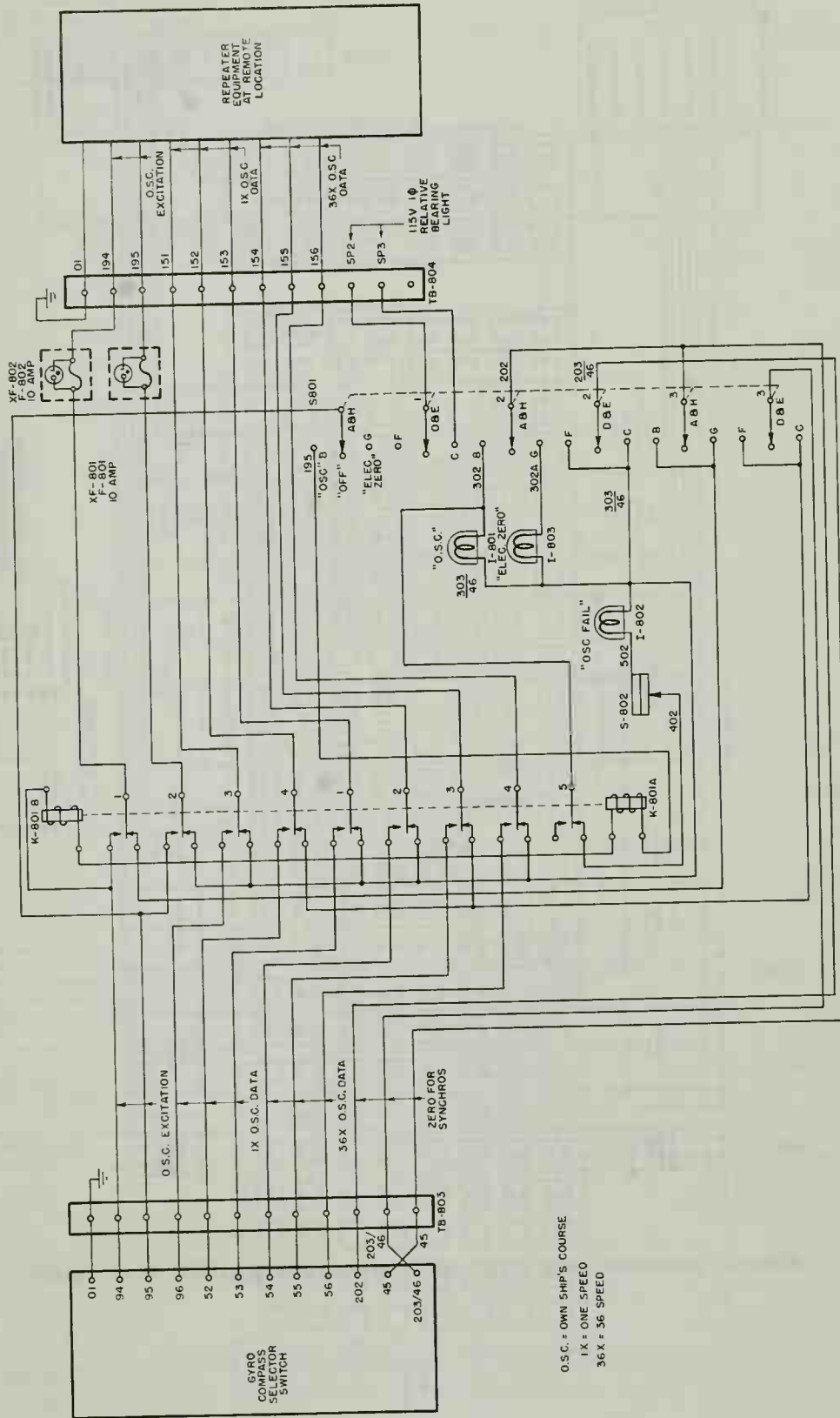
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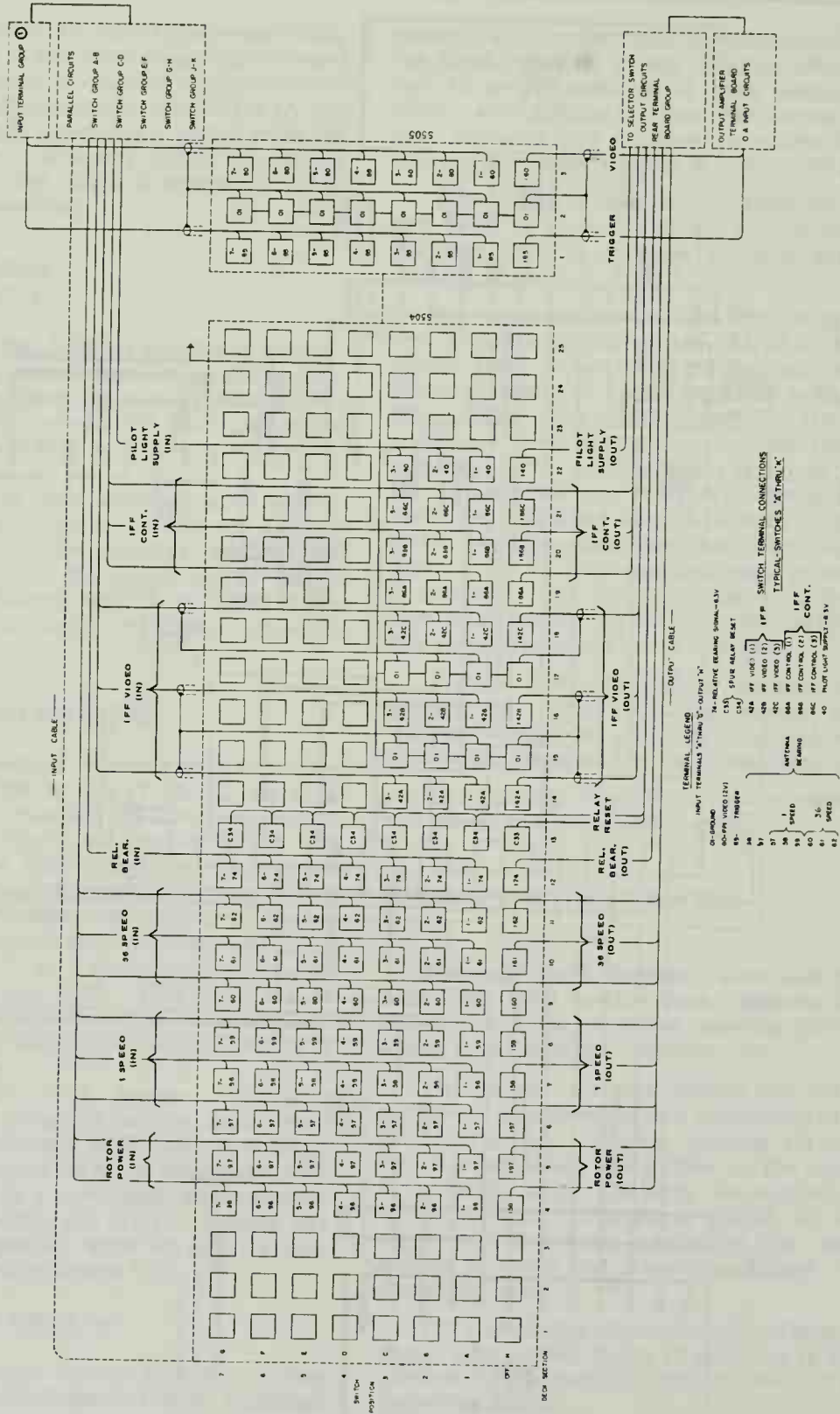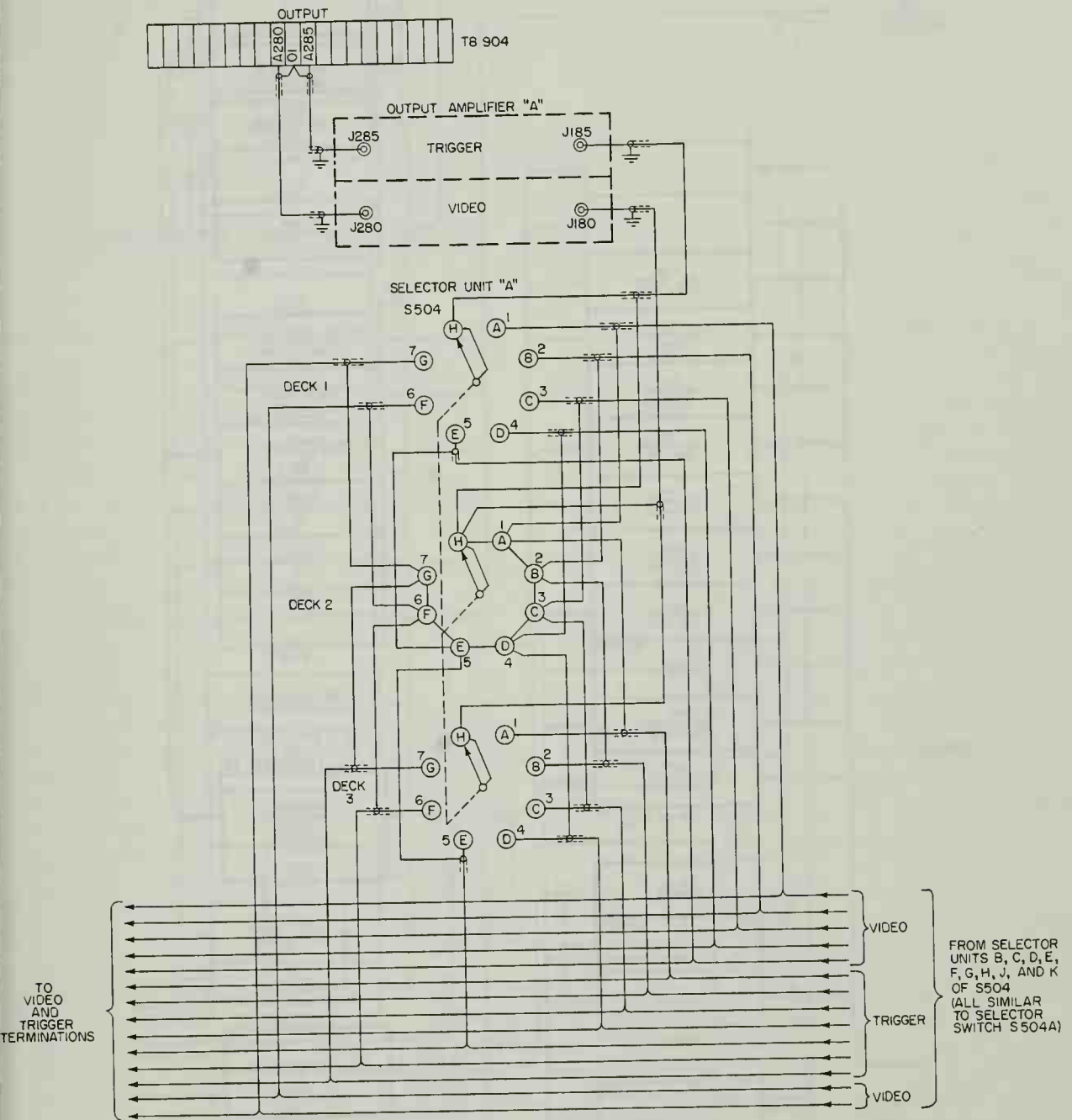Figure 7-32.— Own ship's course selector switch.

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Figure 7-33.— Radar selector switch in the data switching group.

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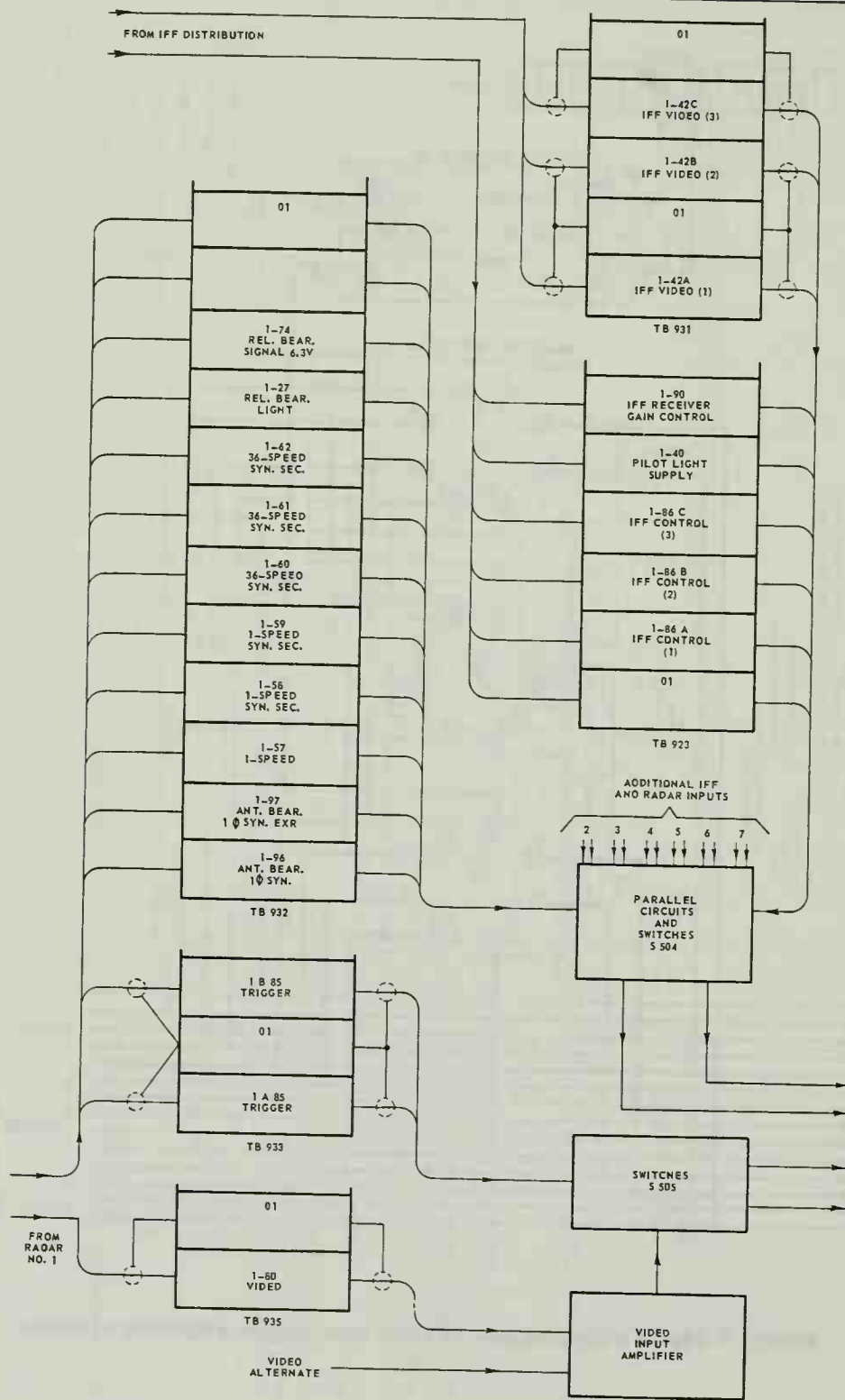Figure 7-34.—Wiring diagram of video and trigger switching circuits.

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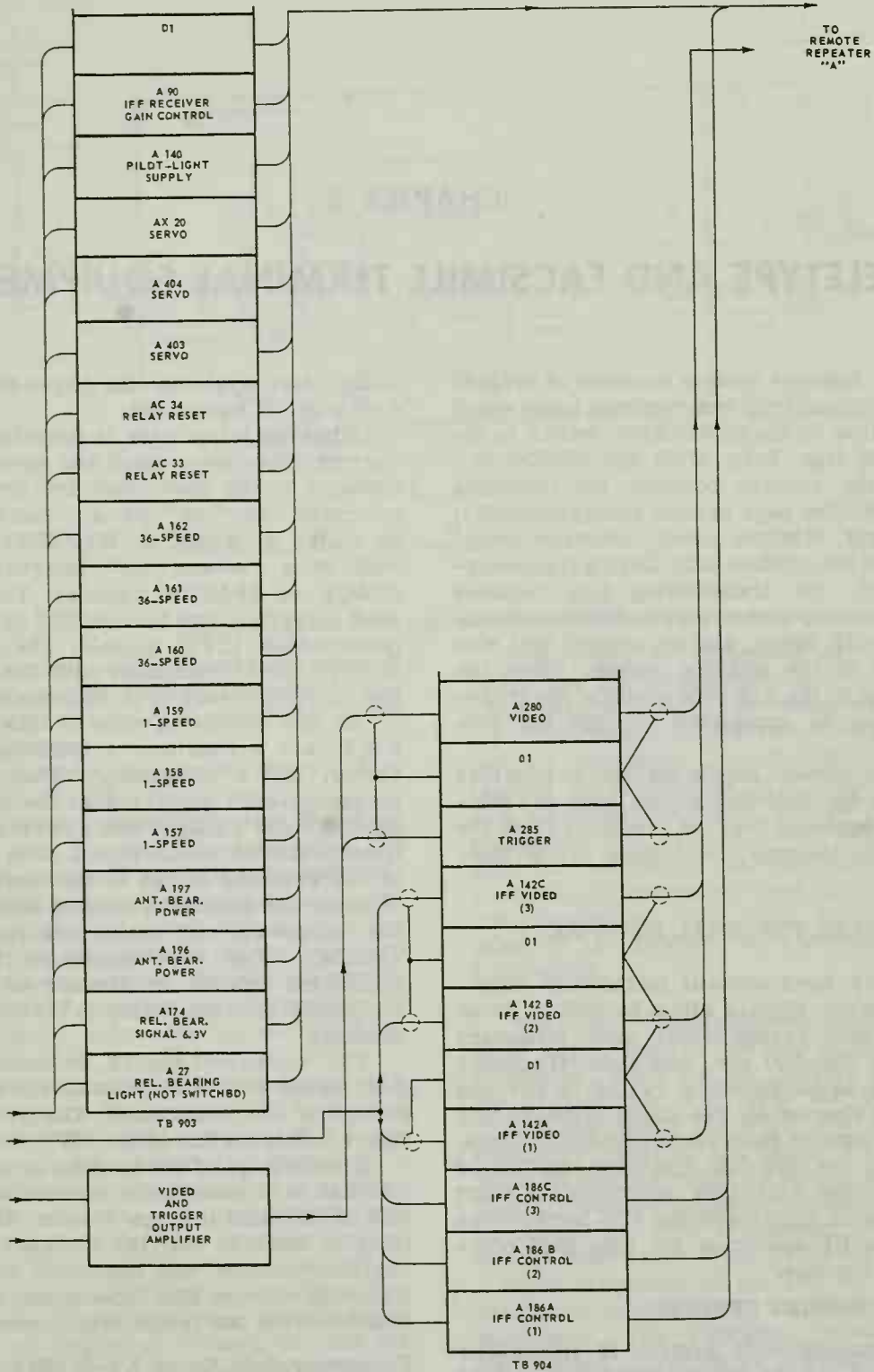Figure 7-35. — Diagram showing route of signals from radar to repeater through the radar switchboard

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Figure 7-35. — Diagram showing route of signals from radar to repeater
through the radar switchboard — Continued.

CHAPTER 8

TELETYPE AND FACSIMILE TERMINAL EQUIPMENT

A radio teletype system consists of series-connected transmitting and receiving loops which are controlled by the SEND-REC. switch in the control unit (fig. 8-1). With the SEND-REC. switch in the receive position, the receiving loop includes the page printer (teletypewriter), power supply, teletype panel, converter-comparator, and the control unit. During frequency-shift keying, the transmitting loop includes the page printer, power supply, teletype panel, frequency-shift keyer, and the control unit with the switch in the position shown. When the switch is set to the S/R tone position, the transmitting loop is completed through the tone converter.

A single power supply pp-765/U provides current for the selected series loop. Six teletypewriter systems may be operated from the power supply through connections to the teletype panel.

TELETYPE TERMINAL EQUIPMENT

The Navy uses several methods of transmitting teletype signals afloat including: Radio Frequency shift keying (FSK), audio frequency shift keying 700-500 cps, and audio frequency shift keying in multiplexing (which is covered in detail in chapter 9). The above methods may be used for long or short range communications. Historically the 700-500 cps tone modulated was used in the VHF/UHF spectrum for short ranges (line of sight) and the FSK method was used in the HF spectrum for long haul communication circuits.

FREQUENCY-SHIFT SYSTEM

The frequency-shift system is illustrated in figure 8-2. At the transmitting end of this system (fig. 8-2,A) are a teleprinter; a frequency-shift keyer unit, which is built into the transmitters; and a transmitter (XMIT). In some older systems, the keyer unit is a separate piece of equipment.

When the teleprinter is operated, the direct-current teleprinter mark and space signals are changed by the keyer unit into frequency-shift intervals. An "on" or a "current" interval is called a MARK or MARKING impulse. An "off or a "no-current" interval is called a SPACE or SPACING impulse. The frequency-shift intervals are transmitted as carrier frequency-shift (CFS) signals. The carrier shift is very small compared with the frequency of the carrier; it may be of the order of 850 cycles.

On the receiving side of this system (fig. 8-2,B) are a receiver, a frequency-shift converter, and a teleprinter. When the carrier frequency-shift signal enters the receiver, it is detected and changed into a corresponding frequency-shifted audio signal. The audio output of the receiver is fed to the converter, which changes the frequency-shifted audio signal into the direct-current mark and space teletype signals. The tone converter in the tone-modulated system is similar to the carrier frequency-shift converter in the frequency-shift system.

The marks and spaces, designated as M and S in figure 8-3, are generated in various code groups of five units each. The group shown in figure 8-3 is for the letter "H".

A knowledge of the specific groupings is incidental to a basic understanding of the operation of the radio teletype system. The important thing to know is that the succession of direct-current "marks" and "spaces" in fixed-timed intervals conveys both intelligence and synchronization from one teleprinter to another.

Frequency-Shift Keyer KY-75/SRT

The KY-75/SRT frequency shift keyer (fig. 8-4) used aboard ship is a calibrated frequency shift exciter in which signaling is accomplished
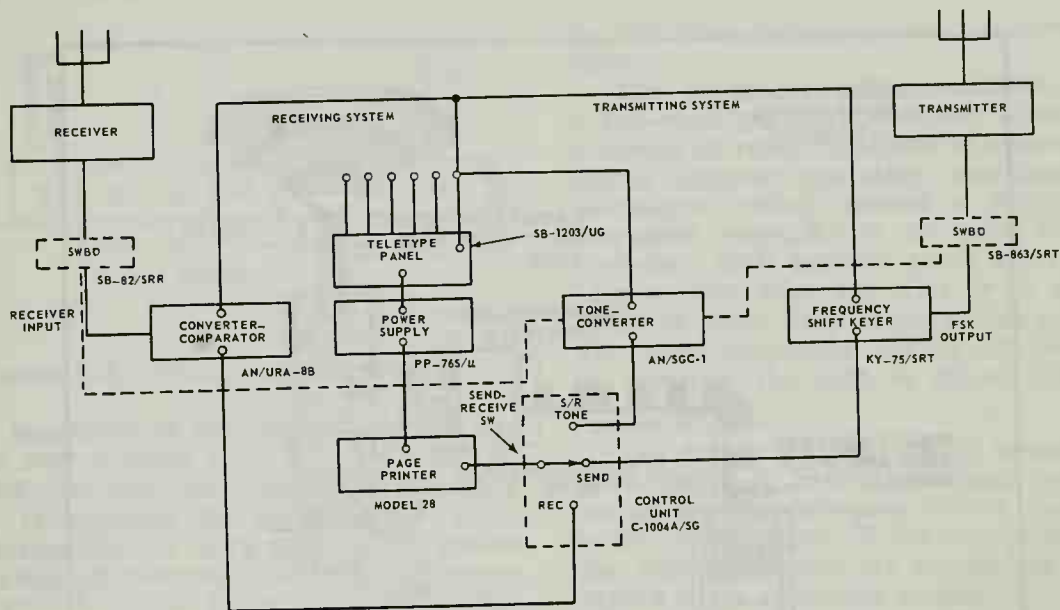
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Figure 8-1.—Block diagram of a representative radio teletype system.

by shifting a constant amplitude carrier between two fixed frequencies. The two frequencies represent marking and spacing conditions of the teletypewriter code.

An identical keyer, intended for shore station use, is the KY-58/GRT.

For radio teletypewriter operation, the keyer functions so that current impulses from the teletypewriter keyboard produce a marking signal. This action causes the transmitter to emit a frequency above its mean frequency. Conversely, opening of contacts on the keyboard produces a spacing signal. The spacing signal, or open keyboard, causes the transmitter to emit a frequency below its normal assigned frequency.

The primary purpose of the frequency shift keyer is to replace the conventional exciter of the CW transmitter with a source of RF excitation that can be shifted (in frequency) a small amount upward and downward to produce RF telegraph code. The unit responds to both neutral (DC) and polar input signals. Neutral signals cause a current flow in the line during the marking condition, and no current flow during the spacing condition. Polar signals allow current in one direction through the line for mark, and in the opposite direction for space. The keyer is used principally for comparatively long-distance communications in the high-frequency range.

The keyer is composed of two subunits, which are the modulator-power supply and the amplifier-oscillator. The modulator-power supply (lower panel) incorporates the power supply and all modulator circuits up to but not including the reactance tube (discussed later). The amplifier-oscillator section (upper panel) comprises the crystal oven, reactance tube, 200 KC oscillator, and all RF circuits and controls. All frequently used operating controls are located on the front panel. Additional controls are mounted on a subpanel behind the front panel.

The input radiofrequency to the frequency shift keyer may be derived from an external oscillator, such as the master oscillator of the transmitter. Or from the self-contained crystal oscillator (fig. 8-5). The overall output frequency range of the keyer is 1.- to 6.7 MC. Because the RF input is mixed with a 200 KC signal from the frequency-modulated oscillator to obtain the sum frequency, the input frequency must range from 0.8 to 6.5 MC.

The frequency of the 200 KC oscillator is frequency-modulated by the balanced reactance modulator. The reactance modulator varies the amount of reactance across the oscillator tuned circuit, causing oscillator frequency to increase or decrease a small amount in response to mark and space signals.
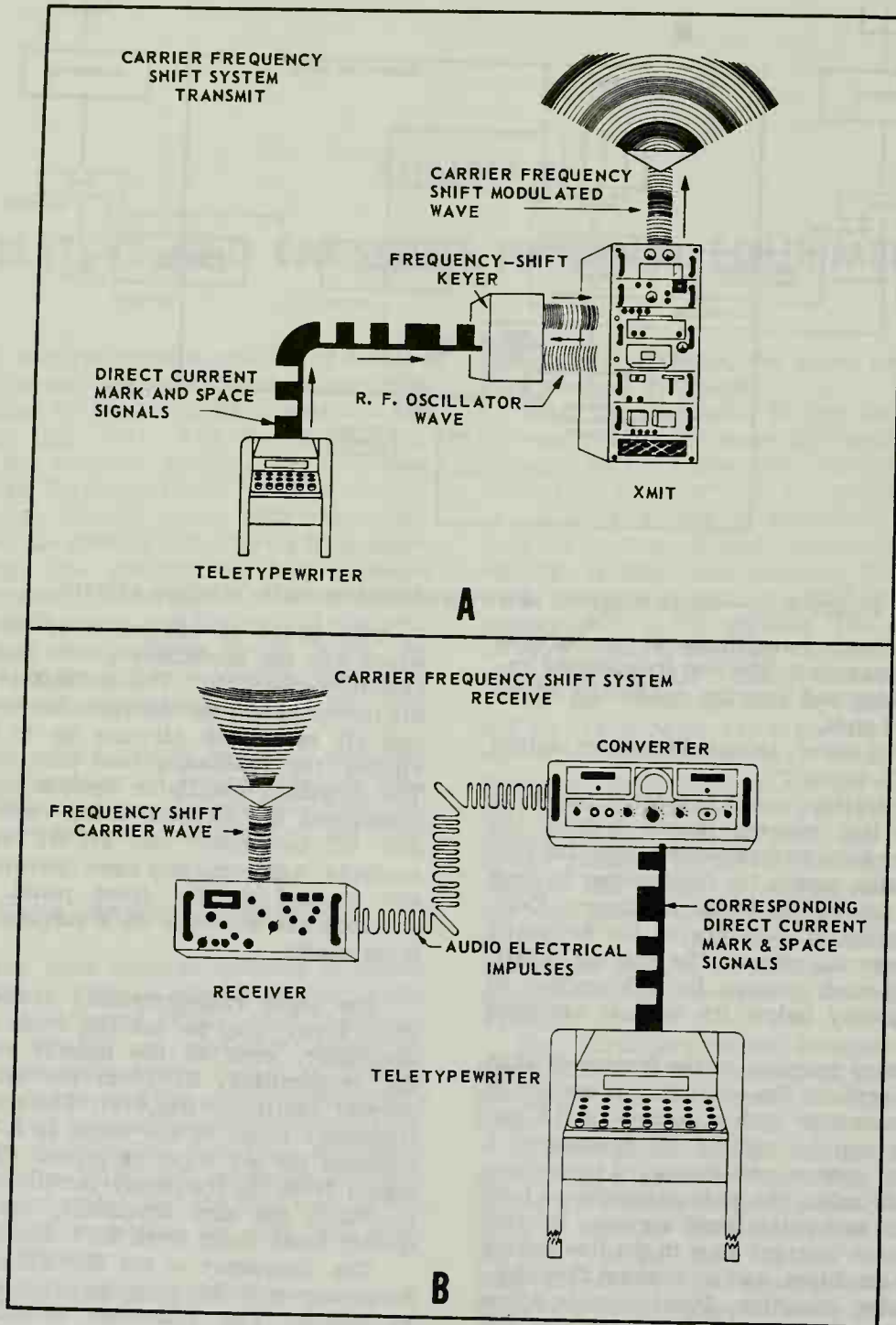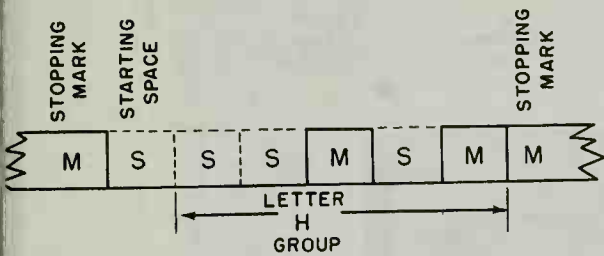
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Figure 8-2.—Frequency-shift system.

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Figure 8-3.—Teletype code group.

The magnitude of the frequency shift is adjustable over a range from $0/$ to ± 500 CPS or 1,000 CPS total shift. The transmitter frequency usually is adjusted for an 850-cycle shift, which means that the mark signal is 425 cycles above assigned frequency, whereas the space signal is 425 cycles below the carrier. The equipment is capable of being keyed up to 240 dot cycles (transitions from mark to space and vice versa) per second.

The radiofrequency output of the cystal oscillator and the frequency-modulated output of the 200 KC oscillator are combined in the balance mixer circuit. The frequency of the crystal oscilator is balanced out. Therefore, only the sum and difference frequencies resulting from mixing the two oscillator frequencies are present in the output of the balanced mixers. The circuits of the balanced mixers are tuned to the sum frequency.

The buffer amplifier serves to isolate or premit light loading at the balanced mixer output circuit, providing greater discrimination against unwanted modulation components. A ganged tuning control simultaneously tunes the output of the crystal oscillator, balanced mixers, and buffer amplifier. The power amplifier supplies an output of 6 watts into a 75-ohm noninductive resistive load.

The test-operate and calibrate circuit comprises a five-position switch utilized to select the circuit arrangement required for frequency shift or facsimile operation. It also selects the arrangement required to perform alignment adjustments for carrier, mark, and space.

Two diode pulse limiters are employed to furnish a modulating wave effectively free from amplitude and wavefront variations of the incoming telegraph signal. The output of the second limiter is coupled to a cathode follower, which acts as an impedance transforming device to match the low-impedance pulsing circuits to relatively high-impedance calibration circuits.

The phase modulation oscillator supplies a 200-cycle phase shift signal, which provides a means of reducing effects of selective fading during adverse operating conditions. Phase modulation, which causes a slight shift in frequency, normally is not used if the total frequency shift between mark and space signals is less than 600 CPS or if the dotting speed is faster than 35 dot cycles per second. The phase-modulation control provides a means of adjusting the shift in phase from 0 to 1 radian $(57.3°)$.
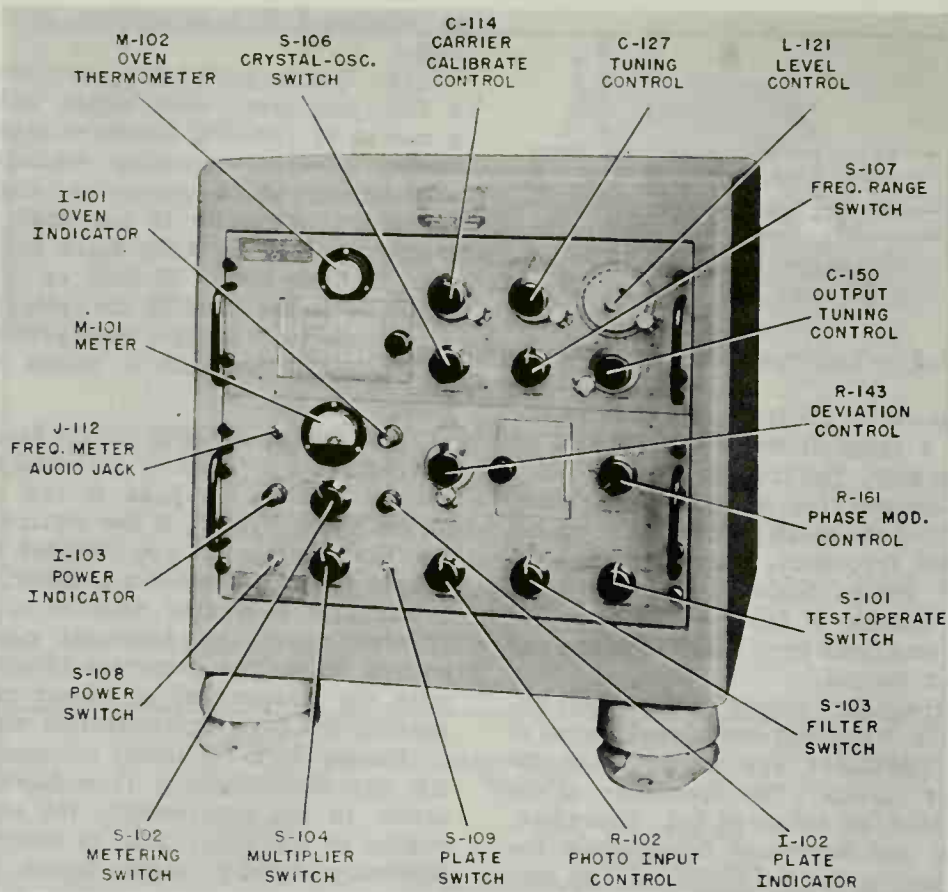
The frequency shift signal transitions are wave-shaped to restrict sideband radiation by means of four low-pass filters incorporated in the pulse shaper. If the square waves from the limiting circuit are rounded before being passed to the reactance modulator, fewer sidebands are generated because of the partial elimination of the harmonic content of the square wave. The reduction of sidebands permits the assignment of closer radio channel spacing for telegraph operation with the keyer.

During high-frequency transmission, when RF carrier frequency is multiplied by some factor in the transmitter, the amount of devaition of frequency shift is multiplied simultaneously by the same factor. It therefore becomes necessary to reduce the amount of deviation by a factor equal to the multiplication factor of the transmitter. This action is accomplished by the deviation multiplier, which provides a means of dividing frequency deviation by a factor of 1, 2, 3, 4, 5, 8, 9, or 12. Thus, the transmitter output is held to a certain preset deviation regardless of the frequency multiplication employed in the transmitter.

The deviation control circuits act to keep wave-shaping termination at a fixed value. The calibrated dial of the deviation control reads the actual amount of deviation realized at the output of the transmitter.

When the test-operate switch is placed in PHOTO position, the equipment is in readiness for facsimile operation. Photo facsimile operation is similar to frequency shift key (FSK) operation, except that during photo operation, limiting and wave-shaping circuits are not utilized. The photo control circuit provides a means of adjusting photo input voltages.

The power supply contained in the modulator-power supply is designed to operate from a 115- or 230-volt. AC, single-phase, 60 cycle

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M-102 OVEN THERMOMETER

S-106 CRYSTAL-OSC. SWITCH

C-114 CARRIER CALIBRATE CONTROL

C-127 TUNING CONTROL

L-121 LEVEL CONTROL

I-101 OVEN INDICATOR

M-101 METER

J-112 FREQ. METER AUDIO JACK

I-103 POWER INDICATOR

S-108 POWER SWITCH

S-107 FREQ. RANGE SWITCH

C-150 OUTPUT TUNING CONTROL

R-143 DEVIATION CONTROL

R-161 PHASE MOD. CONTROL

S-101 TEST-OPERATE SWITCH

S-103 FILTER SWITCH

S-102 METERING SWITCH

S-104 MULTIPLIER SWITCH

S-109 PLATE SWITCH

R-102 PHOTO INPUT CONTROL

I-102 PLATE INDICATOR

8-4.—Frequency-shift keyer KY-75/SRT, front panel compartment identification.

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input source. The supply furnishes 270 volts DC at 220 ma, 4.35 amp at 6.3 volts AC, 0.6 amp at 12.0 volts AC and 3 amp at 5.0 volts AC. A 5.0-volt winding furnishes filament voltage for the rectifier tube.

TONE-MODULATED SYSTEM

The tone-modulated system is illustrated in figure 8-6. The teleprinter (TTY) sends out a signal consisting of direct-current, on-and-off pulses.

To transmit messages by the tone-modulated system (fig 8-6, (1)), a teleprinter, a tone terminal, and a transmitter are needed. The teleprinter sends out a direct-current signal of marks and spaces, and the signal is changed to either of two audio tones in the tone terminal. The tones may be 500 cycles for a space and 700 cycles for a mark. The transmitter impresses the audio tones on the carrier and

sends out an amplitude or tone-modulated carrier wave.

To receive messages with the tone-modulated system (fig 8-6, (2)), a radio receiver, a tone converter, and a teleprinter are needed. The tone-modulated carrier wave enters the receiver, which extracts the signal intelligence and sends the audio tones to the tone converter. The converter changes the audio tones into direct-current mark and space pluses for the teleprinter.

In practice, the same tone terminal is used for both the sending and the receiving circuits because it contains both a transmit "keyer" unit and a receiver "converter" unit.

Tone Converter AN/SGC-1A

Shipboard communications employ tone modulation for short range (UHF and VHF) transmission. Tone modulation employs a tone
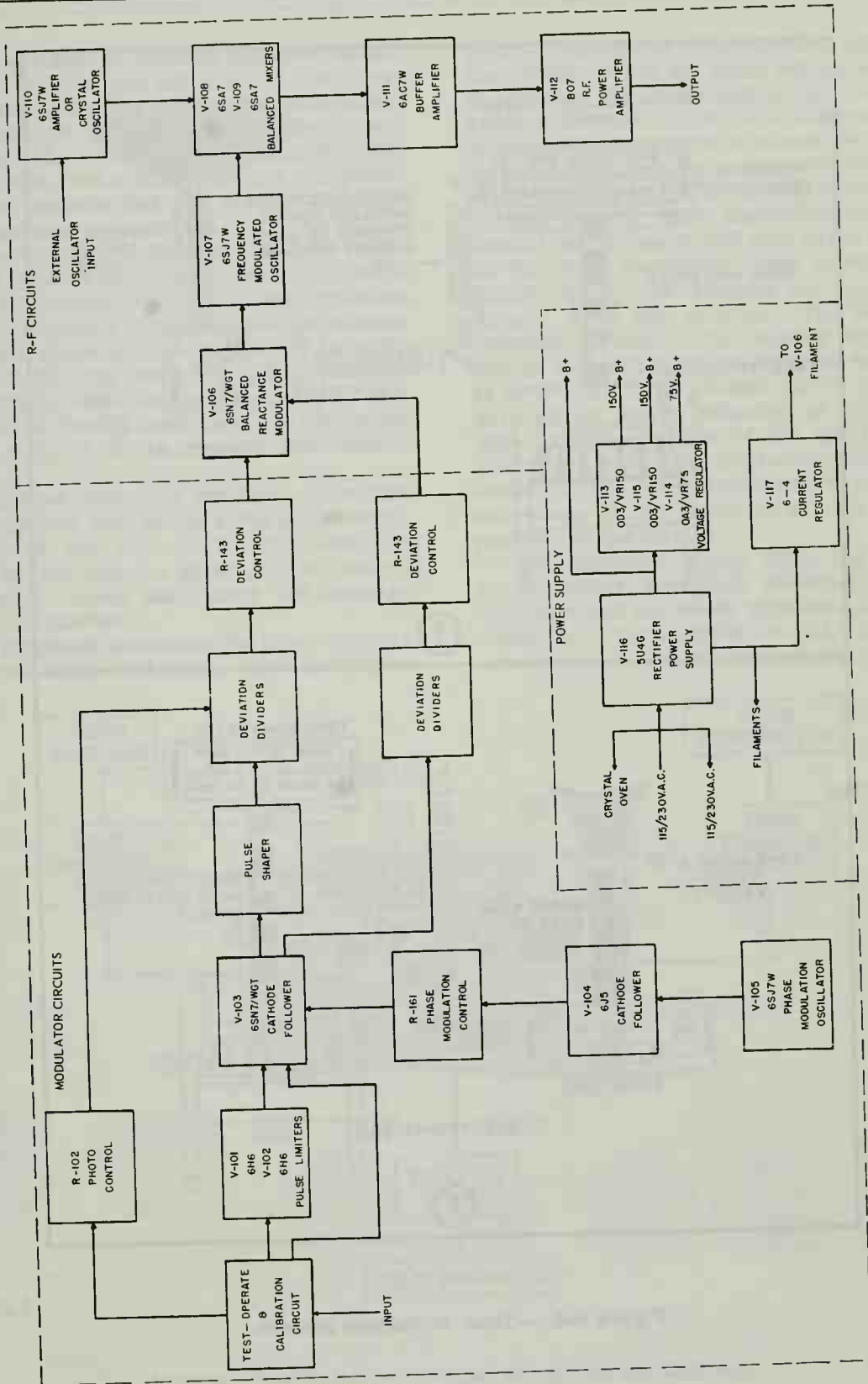
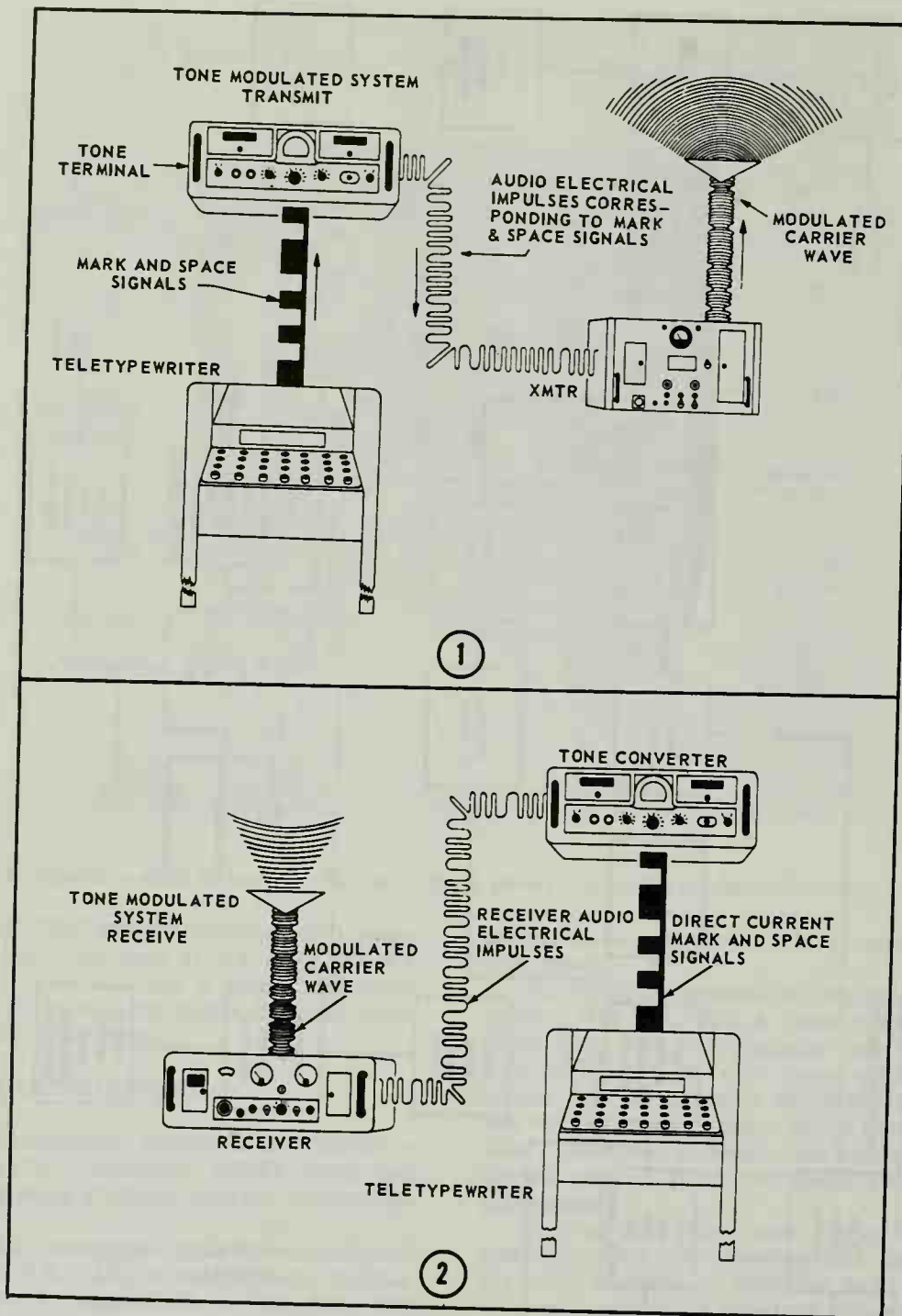Fig. 8-5.—Block diagram of frequency-shift keyer KY-75/SRT, front panel compartment identification.

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TONE MODULATED SYSTEM
TRANSMIT

TONE
TERMINAL

AUDIO ELECTRICAL
IMPULSES CORRES-
PONDING TO MARK
& SPACE SIGNALS

MODULATED
CARRIER
WAVE

MARK AND SPACE
SIGNALS

TELETYPEWRITER

XMTR

①

TONE CONVERTER

TONE MODULATED
SYSTEM
RECEIVE

MODULATED
CARRIER
WAVE

RECEIVER AUDIO
ELECTRICAL
IMPULSES

DIRECT CURRENT
MARK AND SPACE
SIGNALS

RECEIVER

TELETYPEWRITER

②

Figure 8-6.—Tone-modulated system.

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converter, one type of which is included in Radio Teletype Terminal Set AN/SGC-1A, as illustrated at the center of figure 8-7. Other associated equipment includes a radio receiver and transfer panel, a radiophone transmitter and transfer panel, and a teletypewriter and panel. The blocks indicate that any suitable standard Navy components may be used.

In tone modulation transmission, the teletypewriter pulses are converted into corresponding audio tones, which amplitude modulate the voice-frequency transmitter. Conversion of the audio tones is accomplished by an audio oscillator in the tone converter, which operates at 700 cycles when the teletype loop is in a closed-circuit (mark) condition and at 500 cycles when the loop is in an open-circuit (space) condition.

An internal relay in the tone converter closes a control line to the radio transmitter, which places the unit on the air when the operator begins typing a message. The control line remains closed until after the message has been transmitted.

When receiving messages, the tone converter accepts the mark and space tones coming in from an associated radio receiver and converts the intelligence of the tones into signals suitable to operate the make and break contacts of a relay connected in the local teletypewriter DC loop circuit. This action causes the local teletypewriter to print in unison with the mark and space signals from the distant teletypewriter.

The receive level calibrated attenuator is located at the upper left side of the front panel on the tone converter. This control permits adjustment of the level of the incoming tone signals from the receiver. The loop current rheostat is next to the receive level control, and is adjusted to 60 ma when the teletype loop is in the mark, or closed circuit condition. A zero-center meter indicator is located at the upper middle portion of the front panel. An associated switch is located directly below the meter. It has several positions to permit measurements to be taken in all the necessary portions of the circuit.

Two indicator lights flank the upper part of the meter. One light indicates the receive condition and the other indicates the transmit condition. Both lights are off when the tone converter is in the standby condition.

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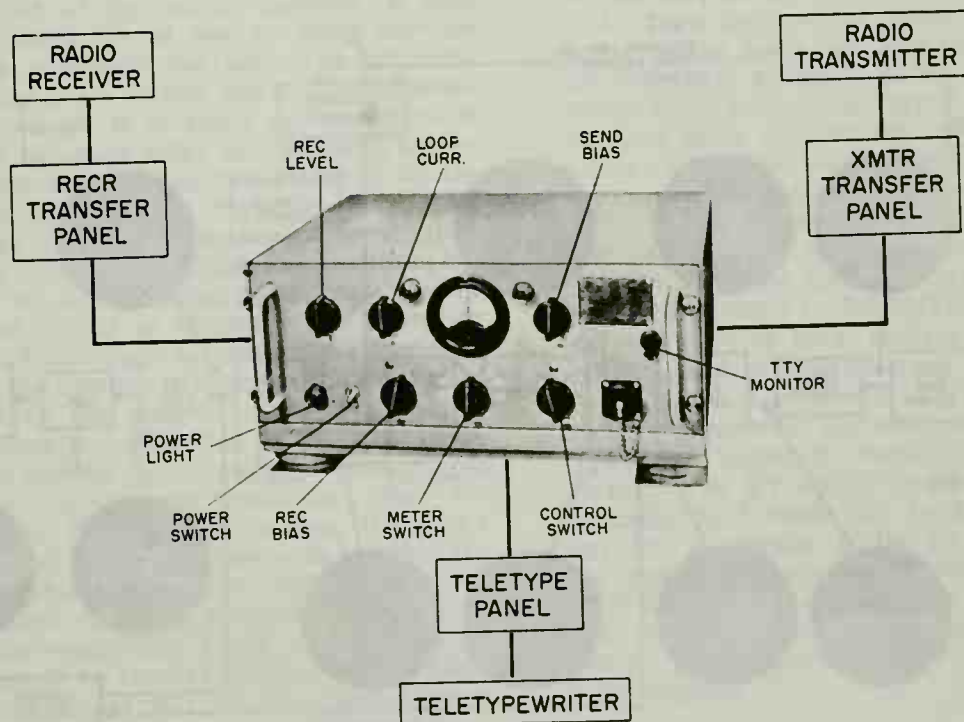
Figure 8-7.—Teletype terminal equipment AN/SGC-1A.

The send bias rheostat is located at the right of the meter. This control permits correction of any teletype distortion (for example, unequal length of mark and space signals) in the local teletypewriter loop when sending a teletype message.

At the far right is a jack marked, TTY monitor. A test or monitoring teletypewriter may be patched into this jack, thereby placing it in series with all other equipments in the loop. NOTE: headphones must not be plugged into this jack.

The power indicator light is located at the lower left side of the front panel of the tone converter. The power ON-OFF switch is located next to it.

The receive bias potentiometer is located at the right of the power switch. This control enables correction of distortion (unequal length of mark and space tones) in the receiving tone circuit.

The control switch is located at the right of the meter switch. The position of the control switch determines the function of the tone converter to either receive or to transmit teletype signals.

A 115-volt, 60-cycle convenience outlet is provided at the lower right of the front panel. It bypasses the power switch.

A block diagram of the tone converter is illustrated in figure 8-8. Typical oscilloscope patterns are included. The attenuator is located at the input of the receive circuit to permit adjustment of the level of the incoming 2-tone signal. The bandpass filter passes all frequencies in the band from 400 to 800 cycles and rejects all other frequencies. The amplifier limiter stages have a constant-output level. The frequency discriminator filter selects the fundamental frequencies of 500 cycles and 700 cycles and routes them via separate germanium rectifiers to corresponding d-c amplifiers (in the DC amplifier block) and associated relay coils in the receiver relay block. A 700-cycle signal causes the receive relay to close the teletype DC loop, a 500-cycle signal causes the relay to open the loop. Thus, corresponding mark and space signals are developed in the teletype loop circuit.

In sending, the mark and space signals in the DC loop cause the send relay to apply either a DC voltage or no DC voltage
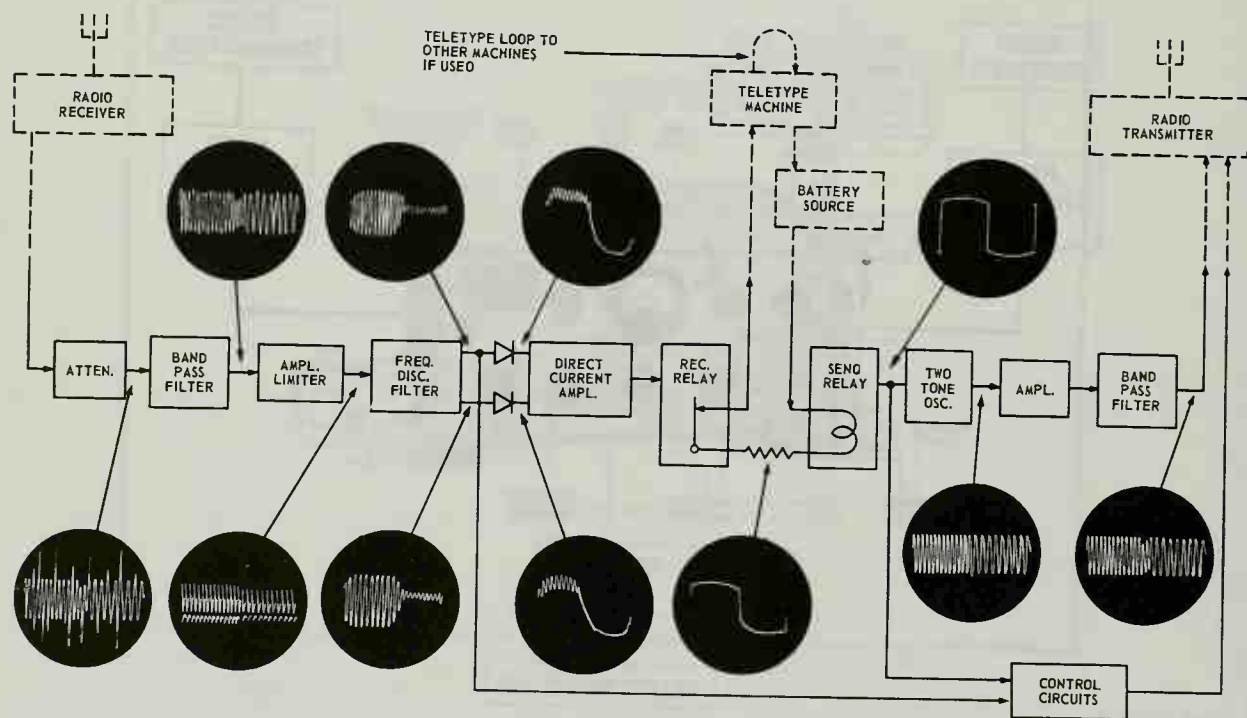
Figure 8-8.—Block diagram of tone converter.

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respectively to the two-tone oscillator. A mark signal causes the oscillator to operate at 700 cycles; a space signal causes it to operate at 500 cycles.

The output of the oscillator passes through a level-controlling potentiometer (not shown in the diagram) to the amplifier stage. The output of the amplifier passes through a bandpass filter to the ship's radio transmitter.

The control switch is used to change from one operating condition to another or to permit the accomplishment of specific operating or maintenance functions. If it is desired to prevent the sending of a message by the teletypewriter, the control switch may be turned to REC/STDY. Thus, the equipment cannot change to the transmit condition even though the teletypewriter is operated, but it can receive messages or remain in the standby condition.

In carrying on communications, the equipment should not be operated with the control switch turned to TRS because the equipment is then locked in the transmit condition and cannot receive any message until released by turning the control switch to one of the other two operating positions.

Usually, for half-duplex communication, this switch is set at the AUTO position. In auto condition, operators at two or more stations having this equipment can engage in full communication. After a station has completed sending its message, it is ready for reception of any return message after an automatic three-second time delay. When the switch is in the AUTO position, the tone converter may be in one of three conditions: receiving, transmitting, or standby. When in the standby condition, the reception of an incoming mark tone causes the control circuit to change to the receiving condition. Following the end of the incoming message, the internal circuits of the equipment shift back to the standby condition. When in this condition the operation of the local teletypewriter causes the circuits to change from standby to the transmit condition. After the last letter of the message is sent, there is a time delay of about three seconds and then the internal equipment circuits shift back to the standby condition. The interlocking functions prevent the equipment from shifting directly from transmit to receive, or vice versa. Thus, an incoming signal will not interrupt an output signal, nor will keying the local teletypewriter, when receiving, cause the circuit to shift to the transmit condition. The control circuits also,

when shifted to transmit condition, cause the control contacts of the transmit control relay to the radio transmitter to close, thereby placing the transmitter carrier on the air.

After a station has sent its message, it is ready for reception of any return message following an automatic three-second time delay. Because of the small time delay inherent in the operation of the control circuits of the local and distant terminals the first character transmitted is usually lost. Therefore, the first character typed should be the "letters" key.

When placing the tone converter in operation, proceed as follows:

1. Turn the power switch (fig. 8-7) to ON.
2. Turn the control switch to TRS (transmit position). If the transmitter power has been turned on, the transmitter will send out a continuous tone; however, this does not matter for a few moments. If desired, the transmitter power may be left off until certain other adjustments have been performed.
3. Turn the meter switch to LOOP CURRENT. Adjust the control marked, LOOP CURR, until the meter reads 60 on the upper scale. If the meter reads zero, the source of loop current may not be energized.
4. Turn the control switch to AUTO. Then hold down the space bar on the teletypewriter and turn the meter switch to SEND BIAS. The meter should read zero on the upper scale. If it does not read zero, correct the reading to zero by means of the send bias control. Turn the meter switch to OFF before releasing the space bar.
5. The radio transmitter may be turned on if it was left off. When a teletypewriter message is received from a distant station, turn the meter switch to RECEIVE LEVEL and adjust the receive level control until the meter indicates 0 dbm (lower scale).
6. The last adjustment is the receive bias control adjustment for which an incoming teletypewriter signal is required from a distant station. The ET should request that a distant operator hold down his teletypewriter space bar for a minute. While he is holding down the space bar, turn the meter switch to RECEIVE BIAS and adjust the receive bias control until the meter reads zero on the upper scale. Then return the meter switch to the OFF position.

The equipment is now adjusted for operation with its associated teletypewriter receiver and

transmitter for communication with other stations similarly equipped.

The AN/UGC-9 tone shift keyer converter (not illustrated) is the transistorized version of the AN/SGC-1. It serves the same purpose and performs in the same manner. Because transistors are much smaller than conventional electron tubes, the AN/UGC-9 is smaller in size than the AN/SGC-1. Thus, it is better suited for shipboard installation.

Basic Radio Teletype
System

When the carrier frequency-shift system (long range) is combined with the tone-modulated system (short range), several more pieces of equipment are needed-a TELETYPE PANEL, a POWER SUPPLY, a SWITCHING CONTROL, a TRANSMITTER SWITCHBOARD, and a RECEIVER SWITCHBOARD, as illustrated in figure 8-9.

The teletype panel is capable of handling six channels, or "loops." The power supply furnishes the direct "looping" current for all teletype direct-current signals. Located at

the teleprinter is the switching control, which is used to select the desired system. The transmitter and receiver switchboards are used to join the radio teletype systems with other communication systems on board ships.

As has been stated, the tone-modulated system and the carrier-frequency-shift system are combined to form one teletype system in shipboard communications. The tone-modulated system is used only for short-range, or "line-of-sight," communications in the UHF band. Manmade and atmospheric static and signal fading are not major problems in this band, and no special equipment to counteract these effects is needed.

The frequency-shift system, used in the LF to HF bands, is the best way to send the rapidly keyed signals of the teletypewriter over long distances. Fading and interference are sometimes major problems in these bands.

Because a single carrier usually does not fade simultaneously in areas separated by more than one wavelength, and because fading of carriers of different frequencies usually
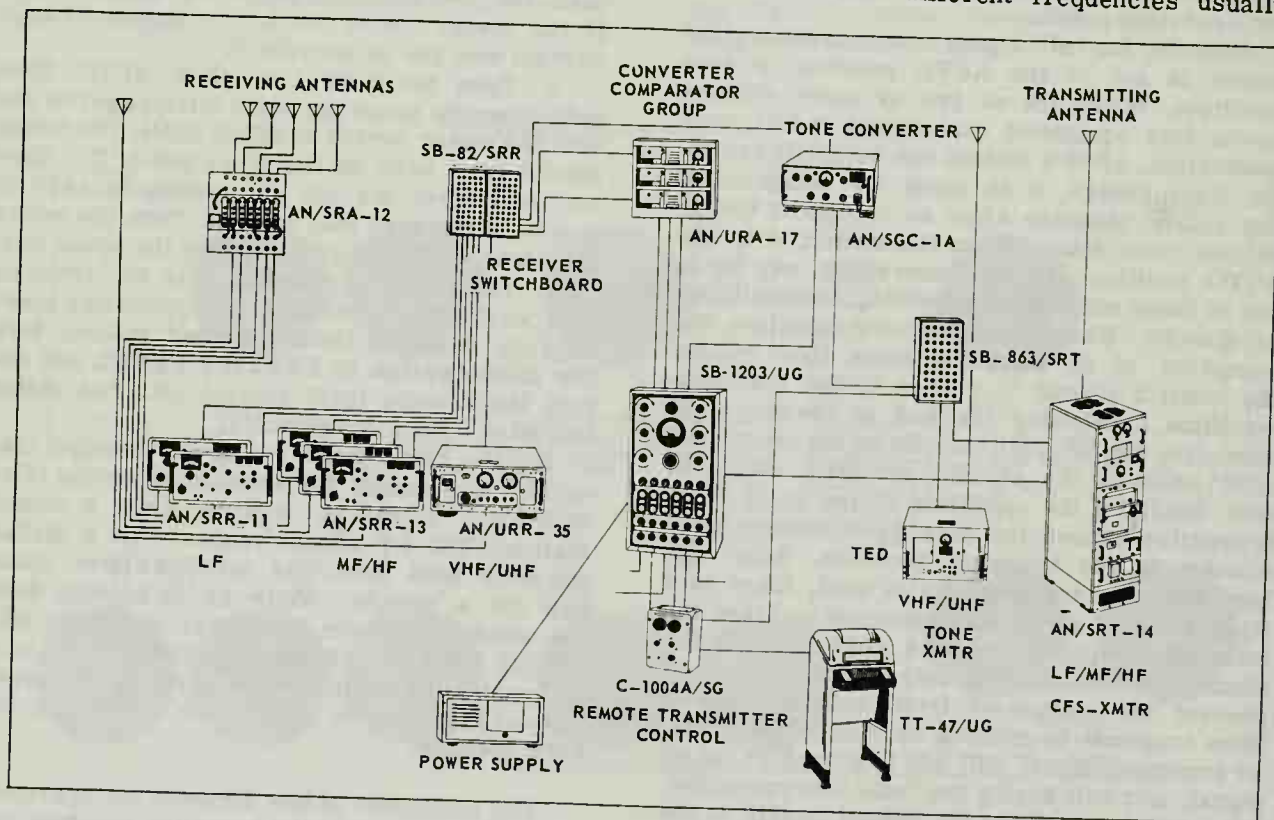
Figure 8-9.—Teletype system (basic).

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oes not occur simultaneously at the same
oint, the Navy has taken advantage of this
ituation by the use of two methods of DIVER-
ITY RECEPTION (fig. 8-10).

In SPACE DIVERSITY reception (fig. 8-10, A),
ne signal is transmitted, and this signal is
eceived by two receivers. Antennas for these
eceivers are separated by a distance greater
han one wavelength. The outputs of the re-
eivers are fed into two frequency-shift con-
erters and then into a COMPARATOR, which
elects the best signal for the teletypewriters.

In FREQUENCY DIVERSITY reception (fig.
-10,B), two or more identical signals are
ransmitted on different frequencies. Two re-
eivers, two converters, and a COMPARATOR
re used, as in space diversity. The receiving
ntennas are not separated.

For the tone-modulation system (higher
requencies), the transmitter may be the TED
and the receiver may be the AN/URR-35; for
he carrier frequency-shift system (lower fre-
quencies), the transmitter may be the AN/
URC-32, and the receiver may be the AN/SSR-
11 and AN/WRR-2, system.
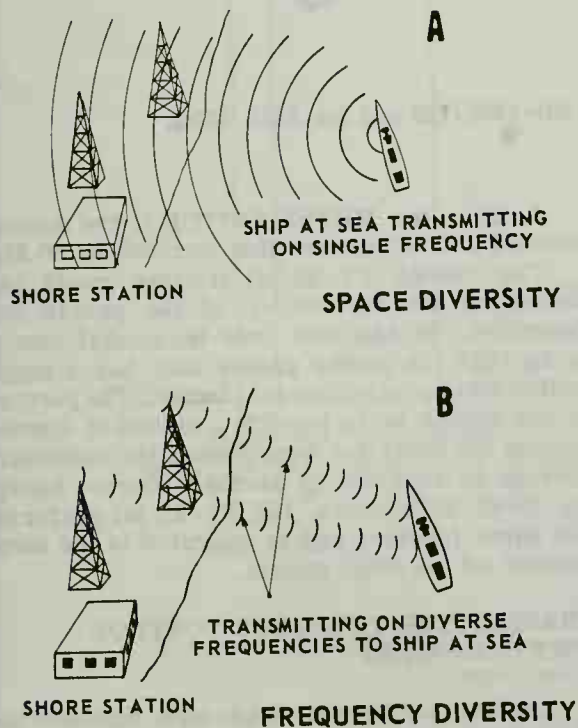
A basic teletype system employing diversity
reception is illustrated in figure 8-9.

COMMUNICATION PATCHING PANEL SB-1203/UG and SB-1210/UGQ

Teletype panels and communication patching
panels are used to interconnect the page printers
and terminal equipments. Communication patch-
ing panels SB-1203/UG and SB-1210/UGQ (fig.
8-11) contains 6 channels. The SB-1210/UGQ
is intended for use with cryptographic devices,
whereas the SB-1203/UG is a general-purpose
panel. Since the panels are almost identical
only the SB-1203/UG will be discussed.

The permanent and patching connections
provide many circuit arrangements. Each chan-
nel comprises a circuit of 3 looping jacks
(LPG), 1 SET jack, 1 MISC. jack, and a rheostat
for adjusting line current. The LPG and MISC.
jacks are identical, and are standard type
phone jacks. The SET jacks incorporate the
features of a double-pole double-throw switch,
as will be seen later. The 6 line current
rheostats provide individual channel current
adjustment.

The CURRENT METER is a DC milliam-
meter. The METER SWITCH is a two-pole,
seven-position, rotary selector switch. When
the METER SWITCH is turned to any one of
the 6 channels, line current in the selected
channel will be indicated on the CURRENT
METER.

Figure 8-12 shows a simplified schematic
of a single channel. The other five channels
are identical. Terminal equipment is connected
to terminals 1 and 2 of terminal board TB-101,
and the teletypewriter is connected to terminals
1 and 2 of TB-102. When line current is not
supplied by the remote station loop, provisions
are made to connect a local source of 115
volts d-c across terminals 1 and 2 of TB-104.
These connections are paralleled across the
corresponding terminals of each loop, and
local current can be connected in or out of
each loop by removable straps.

Resistor R119 limits the current in each
loop to a maximum of 100 ma. Line current
is adjusted by the 2500-ohm rheostat R108.
One terminal of the rheostat is connected to
the first looping jack J101. The SET jack
J119 is connected to terminals 1 and 2 of TB-
102 completing the circuit between the ter-
minal equipment and the teletypewriter.

SHIP AT SEA TRANSMITTING
ON SINGLE FREQUENCY

SHORE STATION

SPACE DIVERSITY

TRANSMITTING ON DIVERSE
FREQUENCIES TO SHIP AT SEA

SHORE STATION

FREQUENCY DIVERSITY
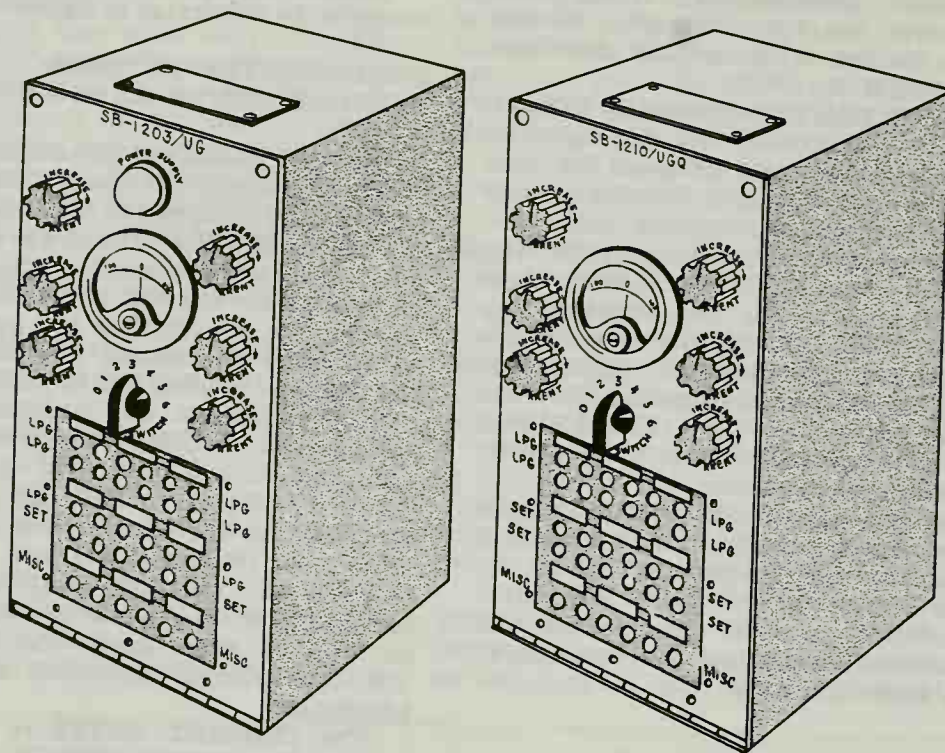
70.10

Figure 8-10.—Diversity reception.

Figure 8-11.—Teletype patch panels SB-1203/UG and SB-1210/UGQ.

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The teletypewriter may be transferred to any other channel by patching it from the SET jack J119 to one of the looping jacks in the channel desired. If it is desirable for the teletypewriter in this channel to be inoperative, a dummy plug is inserted in SET jack J119. An additional teletypewriter may be connected to the MISC. jack and patched into any channel.

Operating Adjustments

1. Turn all line current rheostats counterclockwise to allow minimum current.
2. Turn on local or remote power.
3. If the teletypewriter to be used is wired in the same looping channel as the terminal equipment to be used, no patching is required. If the teletypewriter is not wired in the same channel as the terminal equipment, insert one end of a patchcord in the proper SET jack and the other end in either of the looping jacks in the desired channel.

4. Turn the METER SWITCH to the desired channel and adjust the line current to 60 MA.

The model TT-23/SG teletype panel (not shown) is the forerunner of the panels just described. It has one less horizontal row of jacks than the newer panels and, has a toggle switch incorporated in each channel. The purpose of the switch is to provide a means of disconnecting the local d-c supply when the necessary current is supplied by another source. Except for these differences, the TT-23/SG performs the same function and is operated in the same manner as the other panels.

TRANSMITTER TELETYPE CONTROL
UNIT C-1004A/SG

Another equipment used with teletype installations aboard ship is a control unit, for example, Navy Model C-1004A/SG (fig. 8-13). This unit permits control of a teletypewriter radio circuit from a remote position. It provides

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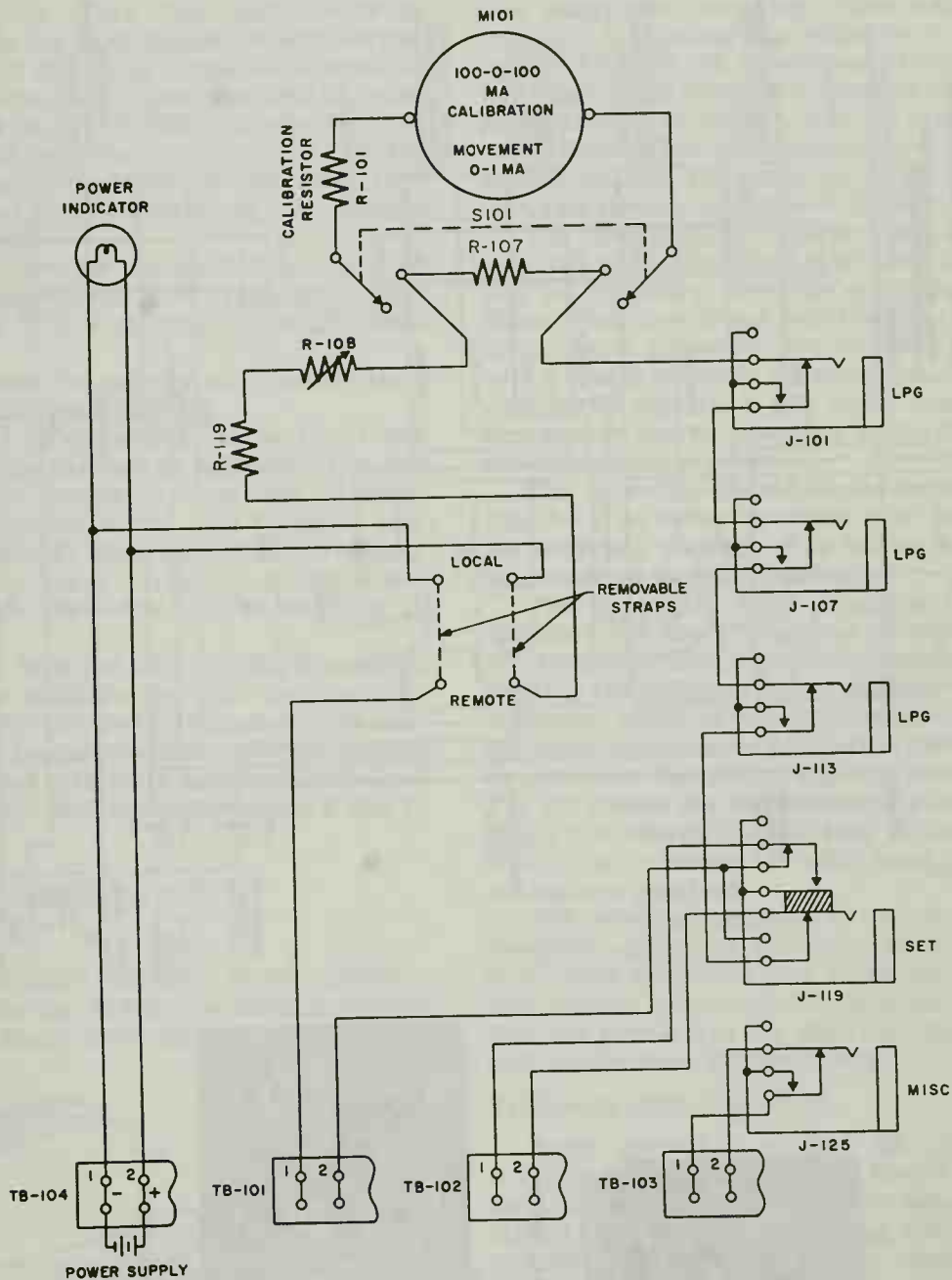
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Figure 8-12.—Communication patching panel SB-1203/UG, single channel simplified schematic diagram.

a transmitter power ON-OFF switch, a power-on indicator lamp, a carrier-on indicator lamp, and a three-position rotary selector switch. The rotary selector switch, S103, provides the following functions:

1. Connects a send-receive teletypewriter to (A) a frequency-shift keyer circuit (CFS send), (B) a frequency-shift converter or comparator circuit (CFS receive) or, (C) a tone terminal on a send receive basis (tone S/R).
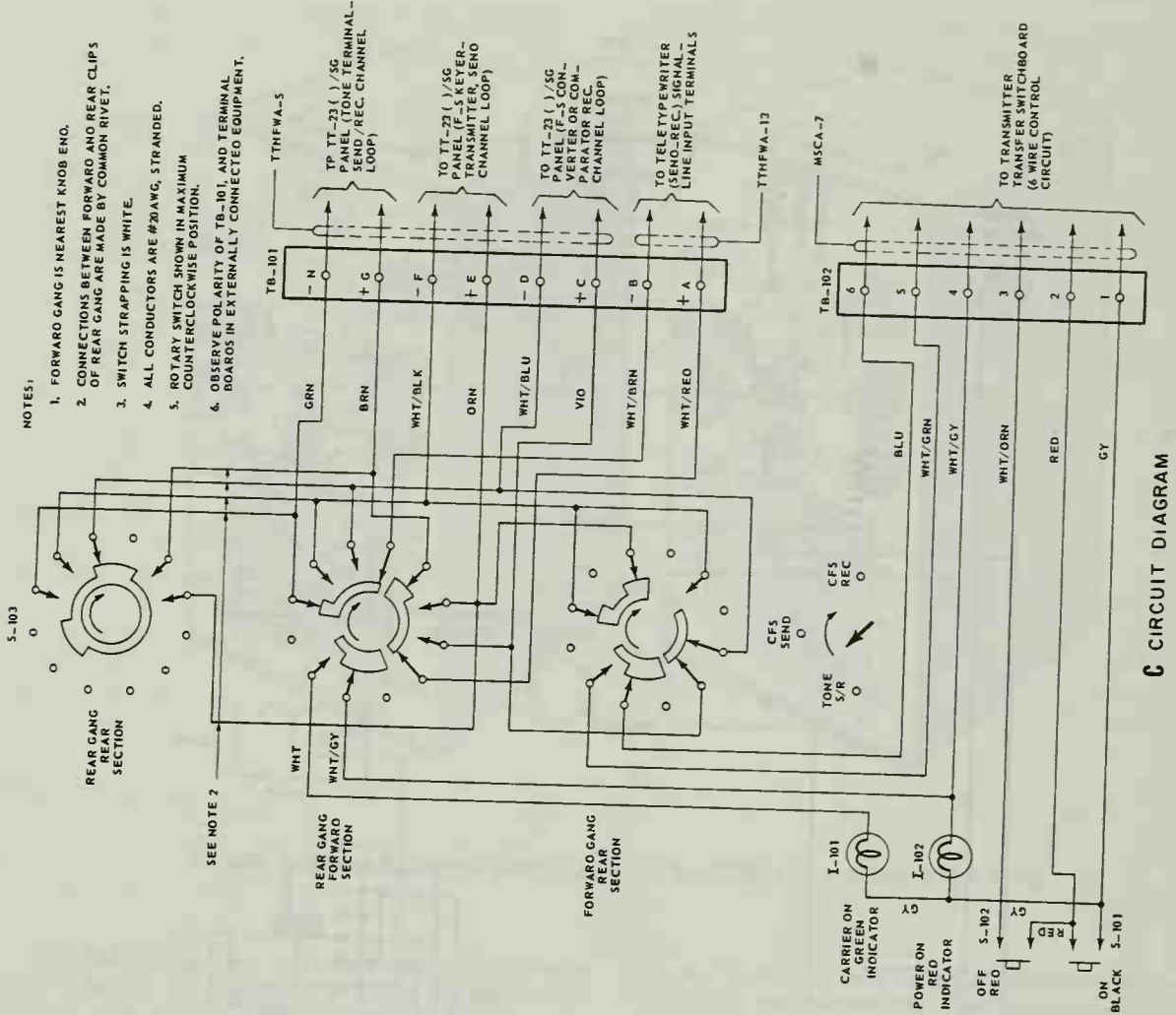
269
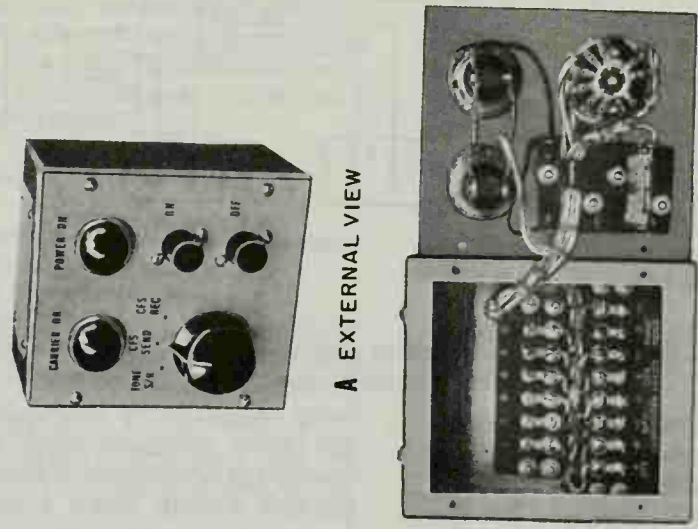
Figure 8-13.—Transmitter teletype control unit C1004A/SG.

2. Shorts the other two unused sets of rminals when the send-receive teletypewriter connected to the set of terminals associated th a particular switch position; that is, when e switch is in the TONE S/R position, the equency-shift keyer terminals, E and F (CFS nd) and the frequency-shift converter ter- inals, C and D (CFS receive) are shorted lose circuited).

3. Turns on the transmitter carrier by osing a circuit in the radio transmitter (ter- inals 5 and 6 are shorted in the CFS send sition only).

4. Energizes the carrier-on indicator lamp the CFS SEND position only.

When S103 (Fig. 8-13C) is in the TONE R position, the carrier-on indicator lamp and e transmitter carrier are off; the teletype- riter is connected to the tone terminal loop erminals G and H) while the unused terminals, equency-shift keyer terminals E and F and equency-shift converter terminals C and D, e shorted.

When S103 is in the CFS RECEIVE position, e carrier-on indicator light and the transmit- r carrier are off; the teletypewriter is con- ected to the frequency-shift converter circuit hile the unused terminals, tone terminals G and and frequency-shift keyer terminals E and F, e shorted.

ONVERTER-COMPARATOR ROUP AN/URA-17

The AN/URA-17 consists of two identical nverter units (fig. 8-14), one of which is shown closeup in figure 8-15. Each converter has its

FREQUENCY SHIFT CONVERTERS
CV-483/URA-17
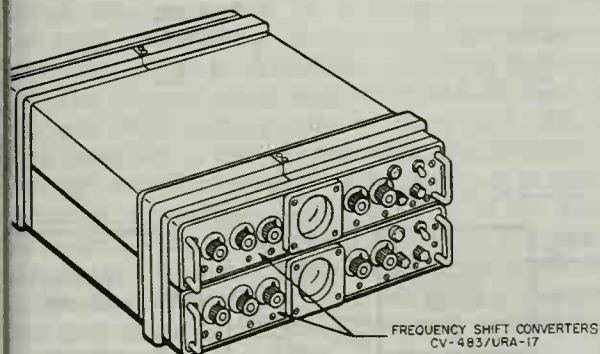
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Figure 8-14.—Converter-comparator group
AN/URA-17.

own comparator circuitry. This design feature achieves a considerable reduction in size from model AN/URA-8B (discussed later), wherein the comparator occupies a separate chassis. The physical size of the AN/URA-17 is further re- duced through use of semiconductors and printed circuit boards. The complete equipment is less than half the size of the AN/URA-8B.

The comparator-converter can be operated with two radio receivers in either space diver- sity or frequency diversity receiving systems. When conditions do not require diversity oper- ation, each converter can be used separately with a single receiver for reception of frequency shift RATT signals. In this latter usage, the two converters can be operated in two independent communication circuits.

For diversity operation, the function switch (fig. 8-15) on both converters must be placed in the diversity position. The teletypewriter may be connected to either converter.

The simplified block diagram (fig. 8-16), indicates the basic functions of converting the RF frequency shift signal into a signal for con- trolling the DC loop of the teletypewriter. The frequency shifts of the audiofrequency output of the radio receiver are converted into DC pulses by action of the audiofrequency discriminator. The DC pulses are fed into the loop keyer, which opens and closes the DC loop of the teletype- writer in accordance with mark and space characters received.

The principal functions of the circuits of the complete equipment are represented in figure 8-17. Two receivers and a teletypewriter are also shown, connected for diversity operation. The two converters are identical. One is shown as a single block for simplicity.

Frequency-Shift Converter

Each converter of the AN/URA-17 (fig. 8-17) is a single unit whose filters and trans- formers are mounted around the sides of printed circuit boards. In the following discussion, the converter is covered as three functional sec- tions: (1) signal processing circuits, (2) keyer circuits, and (3) power supplies.

The input signal from the receiver is applied to the audio input connector at the rear of the converter. The input transformer matches the 8000-ohm impedance of the bandpass filter to the 600-ohm line from the receiver. A center tap is provided for accommodation of balanced inputs. The input transformer is encased with the bandpass filter.
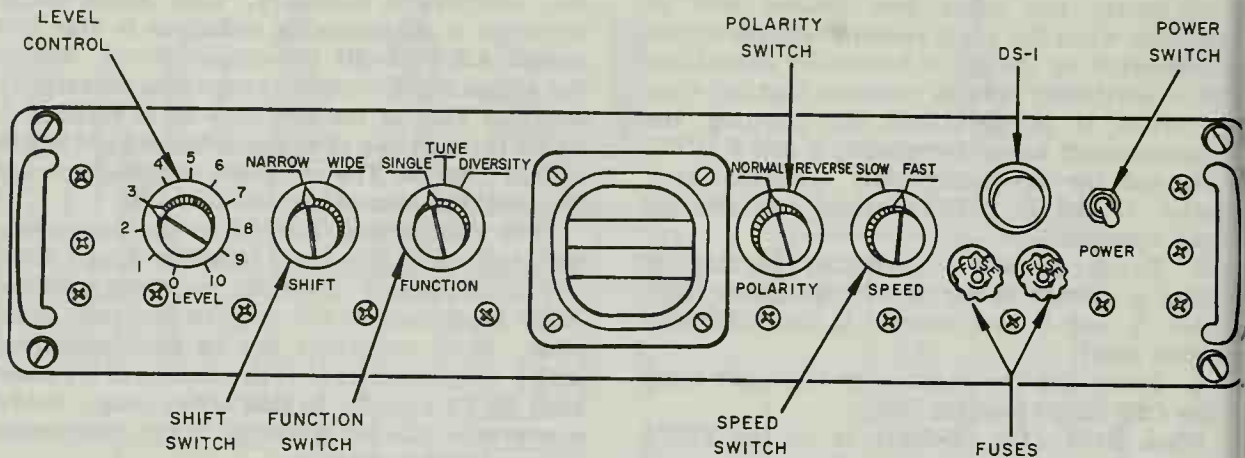
50.7

Figure 8-15.—Frequency-shift converter CV-483/URA-17, front panel controls.

High-frequency noise pulses are attenuated by bandpass filters, and both extremes of the shifted audio signal are passed to the limiter. The shift switch selects the correct filter for the shift width of the input signal. The narrow filter is used when the center frequency of the input signal is 1000 cps, with shifts of 5 to 100 cps each side of center. The wide filter is used when the center frequency of the input signal is 2550 cps, with shifts of 100 to 500 cps each side of center.

The detector output level is held to within 2 db by the limiter, with input signals of from 60 microwatts to 60 milliwatts. Both diodes have a high forward resistance to signals below ap-

proximately 0.6 volt in amplitude. Their resistance to signals of greater amplitude is low. Strong noise pulses are removed from the input signal, and the signal to the discriminator is held at a constant level with fading input signals. The signal level is maintained at approximately 0.6 volt. The output from the limiter is amplified by a common emitter amplifier that uses the level control as the collector load. The level control is used for adjusting the signal level to the discriminators. During reception of a narrow-shift signal, the level must be higher than when receiving a wide-shift signal. The amplified signal is applied to the discriminator through the shift switch.
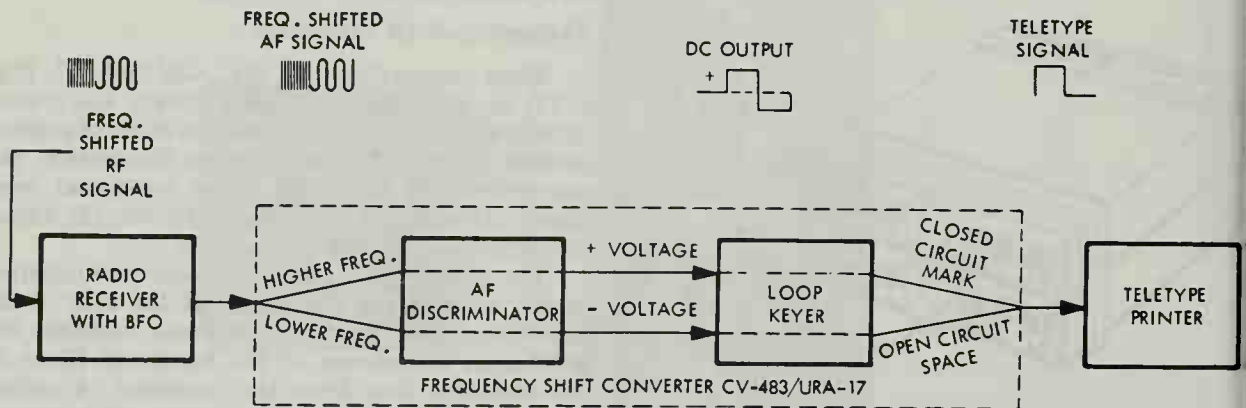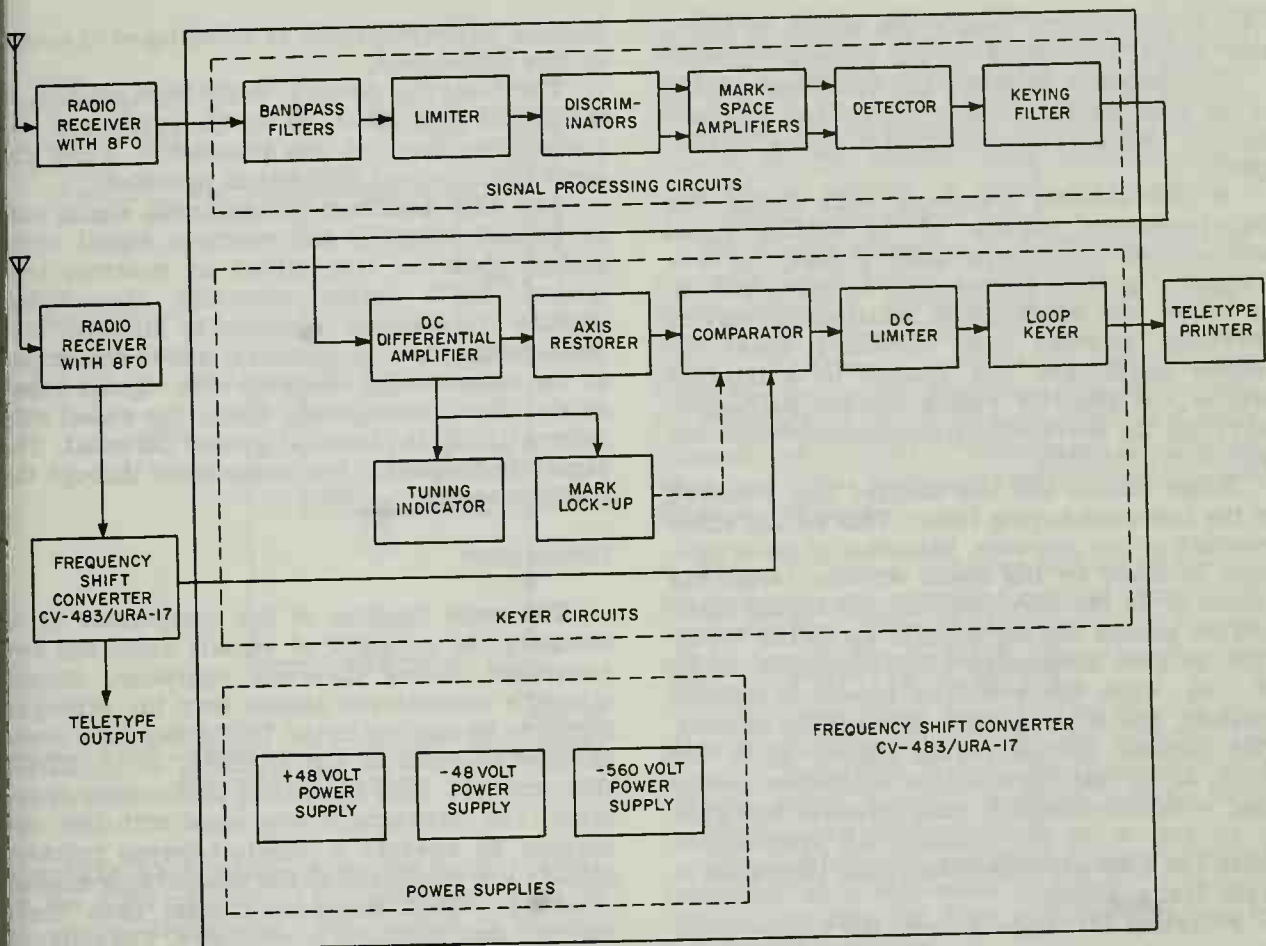

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Figure 8-16.—Frequency-shift receiving system, simplified block diagram.

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Figure 8-17.—Converter-comparator group AN/URA-17, functional block diagram.

Discriminators determine the frequency versus amplitude slope of mark and space signals. Each discriminator is a frequency-selective network, consisting of two resonant networks with overlapping frequency response patterns.

The narrow-shift discriminator is used for signals with shift widths of 10 to 200 CPS. At one terminal of the narrow-shift discriminator, voltage output increases with frequency to maximum at approximately 1200 CPS. At the other terminal, voltage output increases as frequency decreases to a maximum at approximately 800 CPS. The crossover point, at which voltages from the two output terminals are equal, is 1000 CPS ±15 CPS.

The wide-shift discriminator is used for input signals, with shift widths between 200 and 1000 CPS. Crossover frequency of the two resonant networks is 2550 CPS ± 40 CPS. One of the outputs increases with frequency to maximum at about 3400 CPS. The other output increases as frequency decreases to a maximum at approximately 1700 CPS.

The output from one terminal of the discriminator is applied to the base of the first mark amplifier (mark-space amplifier block). Output from the other terminal is applied to the base of the first space amplifier.

The first mark amplifier is a common emitter with fixed base biasing. The mark gain control in the collector circuit provides adjustment of signal gain by controlling degeneration in the emitter circuit. The mark signal is capacitor coupled to the base of the second mark amplifier. Space amplifiers are identical to mark amplifiers. The space gain control is in the collector circuit of the first space amplifier.

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Two transformers couple the output of mark and space amplifiers to the detector circuit.

The detector rectifies and combines outputs of the discriminator transformers into a pulsating DC signal containing mark-space intelligence.

In conventional frequency shift keying, the high-frequency portion of the shifted signal corresponds to teletype mark pulses; the low-frequency portion corresponds to space pulses. At times, the mark-space relationship may be reversed. Garbled copy results if these reversed characters and applied to a teletype printer. A polarity switch for the purpose of inverting the mark-space relationship when required is available.

Noise pulses and the carrier are removed by the low-pass keying filter. This keying filter consists of two sections. Selection of these sections is made by the speed switch. When the switch is in the slow position, the keying filter section passes keying signals up to 100 WPM. This section attenuates all frequencies above 45 CPS. When the switch is moved to the fast position, the other section of the filter is used. This section passes keying signals up to 400 WPM. In the fast position, the equipment can be used with four-channel, time division multiplex up to 100 WPM per channel. All frequencies above 180 CPS are attenuated when the switch is in the fast position.

Pulsating DC signals from the keying filter are converted by keyer circuits into on-off pulses for operating the teletype printer relay.

The DC differential amplifier amplifies the mark-space signals before they are applied to the DC limiter. This differential amplifier contains five transistors, four of which form two emitter-coupled amplifiers. The fifth transistor is the output and feedback transistor.

The input signal from the keying filter is applied to the base of one of the transistors in the first emitter-coupled amplifier. Input to the base of the other transistor in this amplifier is from the collector of the fifth transistor in the output circuit of the DC differential amplifier. The feedback signal from the output stabilizes the gain of the DC differential amplifier.

Both the input signal and feedback signal are amplified by the other emitter-coupled amplifier. This amplifier also mixes these two signals and applies them to the output transistor.

The vertical center control, in the emitter of the first transistor, adjusts the vertical centering of the tuning indicator display. Zero

balance of the amplifier is established by means of this adjustment.

The linearity control, in the base circuit, of the output transistor, is for the purpose of adjusting the bias of this transistor. Thus, the most linear signal possible is provided.

The axis restorer maintains the signal axis at ground potential and restores signal symmetry when the transmitter or receiver frequency drifts during operation. Mark-space signals are clamped to ground by different circuits within the axis restorer, and are combined at the output. By clamping both signals separately, then recombining them, the signal axis automatically is placed at ground potential. The signal is coupled to the comparator through the function switch.
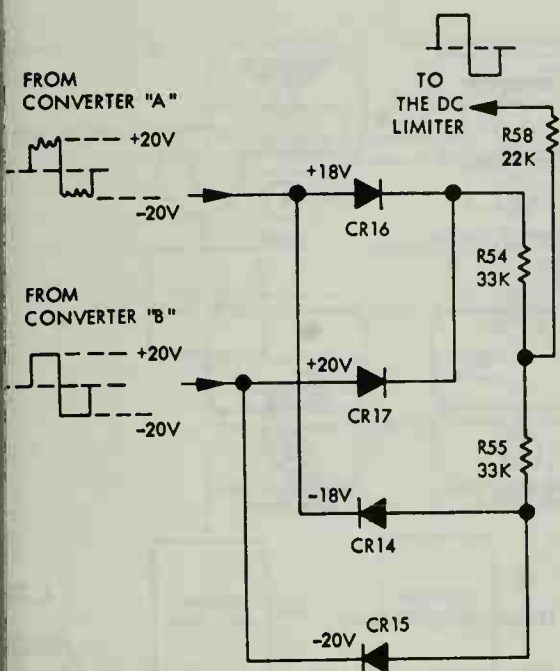
Comparator

The main function of the comparator is to compare the strength of signals from the two receivers during diversity operation. Signal strength comparison allows only the stronger signal to be applied to the DC limiter. The comparator consists of CR14, CR15, CR16, CR17, R54, and R55. (See fig. 8-18.) In diversity operation, two converters are used with two receivers to operate a single teletype printer. Signals are compared at the comparator in each converter. Both converters must have their function switch placed at diversity. The teletype printer may be connected to either converter.

In figure 8-18, peak amplitudes of both input signals are equal, but the signal from converter A contains noise. The signal from converter B, which has a constant peak value of 20 volts, will develop a greater voltage across R54 and R55. This action places a reverse bias of 2 volts on CR14 and CR16, preventing converter A from actuating the DC limiter.

Besides diversity setting, the function switch has two other positions—single and tune. In single, positive mark pulses cause CR16 to conduct; negative space pulses cause CR14 to conduct. In the tune position, the signal input to the DC limiter is removed while the receiver is being tuned.

The DC limiter is a class B push-pull circuit. This circuit supplies about 20 db of post-detection limiting, and aids in proper operation during reception of signals, containing strong noise pulses.

The signal from the comparator is applied simultaneously to the base of two transistors.

162.24

Figure 8-18.—Comparator circuit, simplified schematic diagram.

positive mark signal causes the mark transistor (an NPN transistor) to conduct, but cuts off the space transistor. The signal is phase-shifted 180° and is fed to a PNP transistor, causing it to give a strong positive signal at the output. A negative space signal causes the space transistor (a PNP transistor) to conduct and at the same time cuts off the mark transistor. This signal is shifted 180° and delivered to an NPN transistor, causing a strong negative signal at the output. Mark and space signals then are combined and fed to the loop keyer. Switching action of the loop keyer is controlled by the DC limiter.

The function of the loop keyer is to operate the teletype printer loop relay. Two transistors are contained in the loop keyer circuit. Depending on the signal applied, the first transistor causes the second to conduct or not conduct. When the second transistor conducts, the teletype printer relay closes, causing a mark signal to be transmitted. When the transistor doesn't conduct, the relay opens, causing a space signal to be transmitted. A positive signal from the DC limiter activates the loop keyer circuit; a negative signal cuts off the loop keyer circuit.

During traffic interruptions, the mark lockup circuit provides a strong, artificial mark signal to the DC limiter. When a deep fade in the signal occurs, or if the signal-to-noise ratio is low, noise pulses could cancel the small positive bias on the DC limiter input, allowing garbled copy to be printed. A steady mark signal is transmitted between messages.

The tuning indicator is a cathode-ray tube. Controls for horizontal adjustment, centering, focus, intensity, vertical adjustment, vertical center, and linearity are provided on the converter chassis as screwdriver adjustments. When the associated receiver is tuned properly, the CRT pattern (fig. 8-19) is centered vertically. Correct level control adjustment is achieved when horizontal lines of the pattern coincide with those on the bezel.

Three DC power supplies furnish all operating voltages and operating current required by the converter. Tapped primaries on the two power transformers allow operating one-line voltages of 105, 115, or 125 volts.

Both the positive and negative 48-volt supplies consist of a full-wave bridge rectifier and a series regulator circuit. Each circuit has a voltage adjust control, enabling voltage to be maintained at a constant level.

The negative 500-volt supply uses two diodes connected as a half-wave rectifier. A voltage divider provides the high voltages required by the cathode ray-tube.

AN/URA-8B

The AN/URA-8B (fig. 8-20) is the tube version of the AN/URA-17. The units of the AN/URA-8B are much larger and require an extra unit for the comparator in a diversity set-up with two converters. The same functions are carried out as with the AN/URA-17.

FACSIMILE EQUIPMENT

Facsimile equipment is used to transmit still images over an electrical communications system. The images, called pictures or copy in facsimile terminology, may be weather maps, photographs, sketches, typewritten or printed text, or handwriting. The still image serving as the facsimile copy or picture cannot be transmitted instantly in its entirety. Three distinct operations are performed. These are (1) scanning, (2) transmitting, and (3) recording.
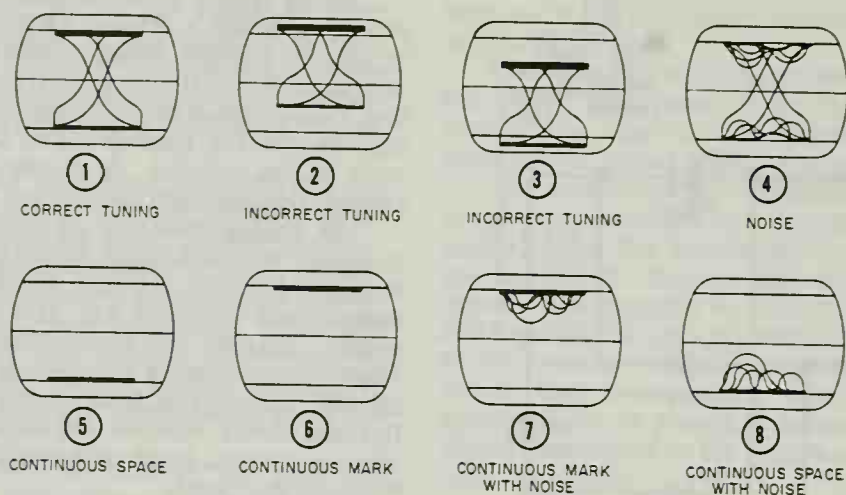
Figure 8-19.—Monitor oscilloscope patterns.

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The SCANNING operation is that of subdividing the picture in an orderly manner into a large number of elemental segments. This process is accomplished in the facsimile transmitter (fig. 8-21) by a scanning drum and photocell arrangement.

The picture to be transmitted is mounted on a cylindrical scanning drum, which revolves at a speed of 1 revolution per second and travels along a lead screw at the rate of 12.5 inches in 20 minutes. (The lead screw has 96 threads per inch.) Light from an exciter lamp illuminates a
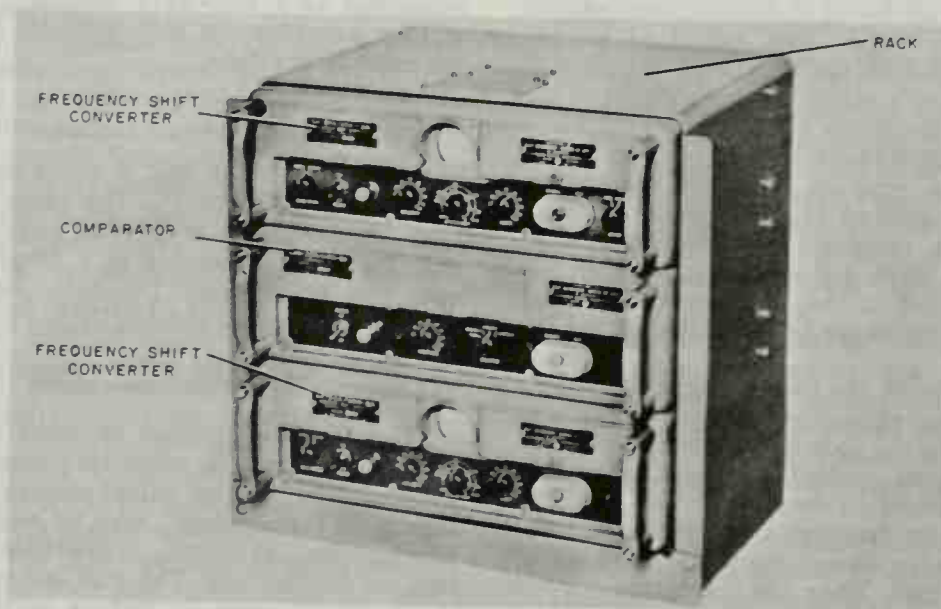
Figure 8-20.—Frequency shift converter-comparator group AN/URA-8B.

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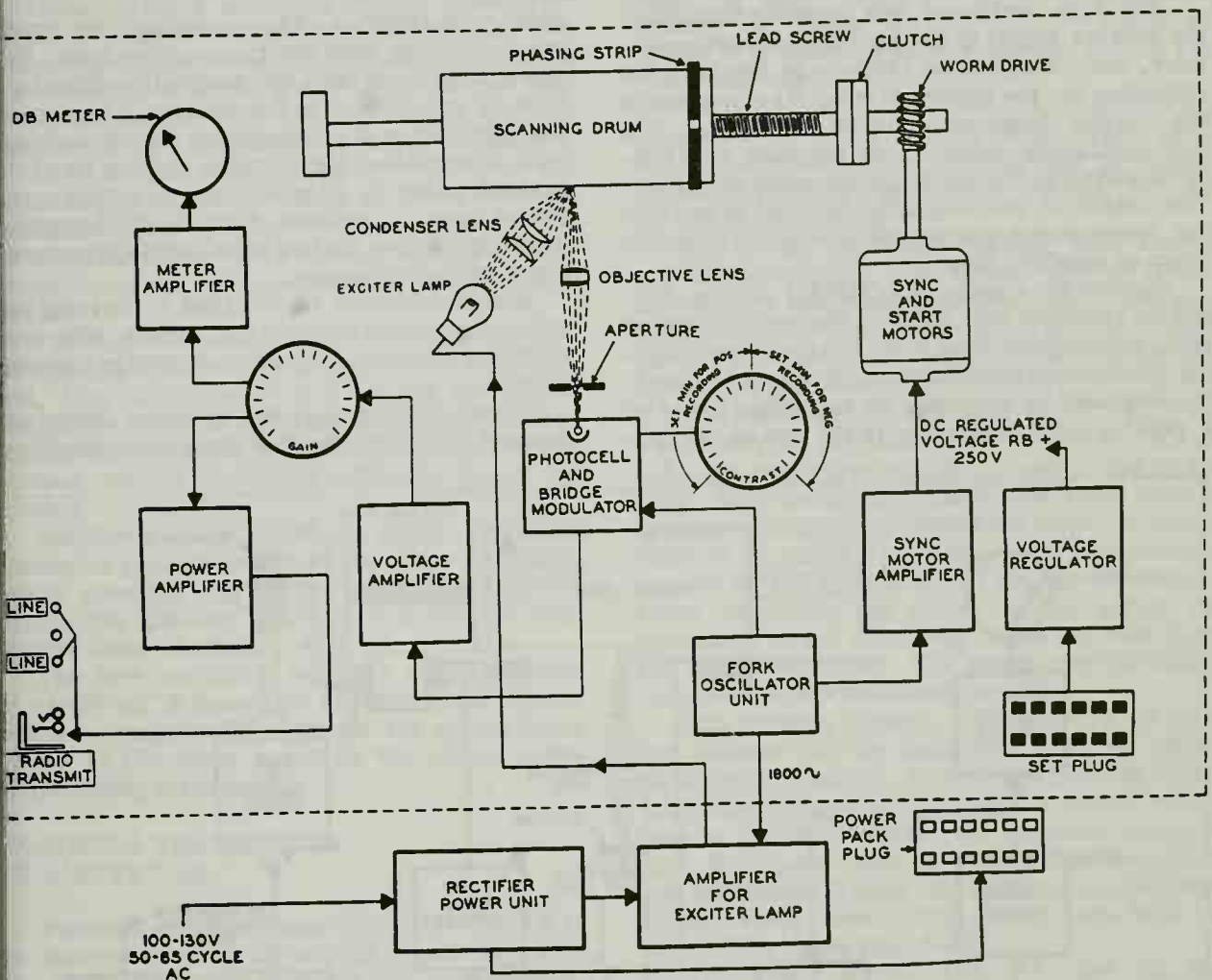
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Figure 8-21.—Facsimile transceiver-transmitting block diagram.

small segment of the moving picture and is reflected by the picture through an aperture to a photocell. During the transmission of a complete picture, the light traverses every segment of the picture as the drum slowly spirals past the fixed lighted area.

At any instant, the amount of light reflected back to the photocell is a measure of the lightness or darkness of the tiny segment of the picture that is being scanned. The photocell transforms the varying amounts of light into varying electrical signals.

The fork oscillator unit develops an output voltage (MODULATION VOLTAGE) that is applied across the bridge modulator. The frequency of this voltage is 1800 cycles. When the bridge

is balanced (photocell dark), the output voltage is zero. When the amount of light falling on the photocell varies, the resistance of the cell varies. This action unbalances the bridge and produces an output voltage that varies in amplitude with the variations in light. Thus, the 1800-cycle voltage is amplitude modulated in the bridge modulator.

The modulated signal is amplified in the voltage amplifier, the proper level is established in the gain control, and the signal is further amplified in the power amplifier before going to the radio transmit output jack. The modulated signal is fed to external equipment (discussed later) and then to a radio transmitter for transmission as desired.

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The fork oscillator unit, besides furnishing the carrier signal to the photocell bridge modulator, also supplies an 1800-cycle signal to the amplifier for the exciter lamp. This output keeps the exciter lamp at constant brilliancy. The fork oscillator output also supplies an 1800-cycle output to the synchronous motor amplifier. The output of the amplifier is used to operate the synchronous motor that drives the scanning drum at constant speed.

Electrical signals RECEIVED by the facsimile receiver (fig. 8-22) at the radio receiver jack, are amplified and serve to actuate a recording mechanism that makes a permanent recording (segment by segment) on recording paper on a receiver drum similar to the one in the fac-simile transmitter. The receiver drum rotates synchronously with the transmitter drum. This action continues until the complete original picture is reproduced in its entirety. The recording mechanism may reproduce photographically with a modulated light source shining on photographic paper or film; or, the recording mechanism may reproduce directly by burning a white protective coating from specially prepared black recording paper.

Synchronization is obtained by driving both receiving and transmitting drums with synchronous motors operating at exactly the same speed.

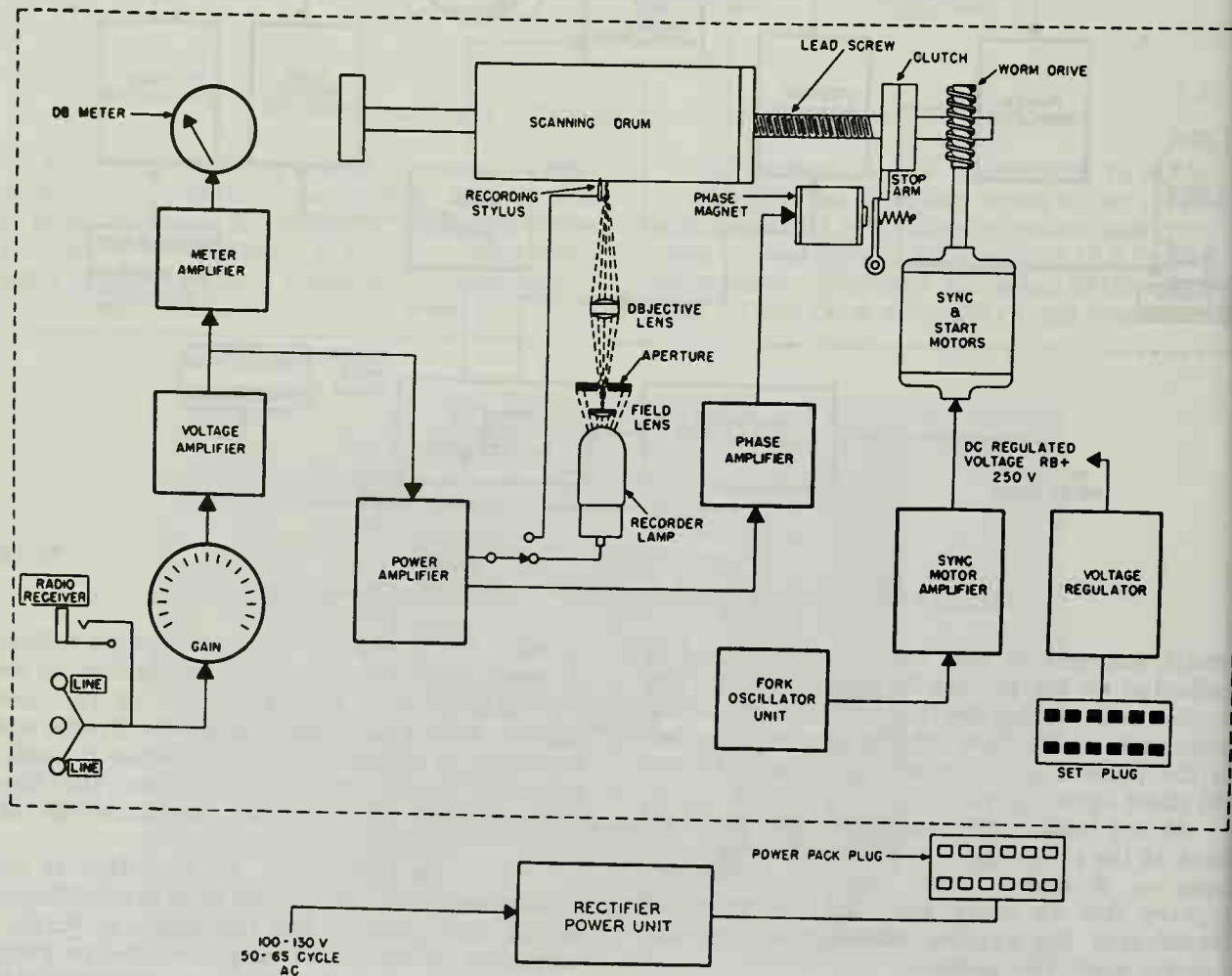Framing (orienting) the receiver drum with respect to the transmitter drum is accomplished

Figure 8-22.—Facsimile transceiver-receiving block diagram.

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by transmitting a series of phasing pulses just before a picture transmission is to begin. The pulses operate a clutch mechanism that starts the scanning drum in the receiver so that it is properly phased with respect to the starting position of the scanning drum in the transmitter.

A start motor mechanically coupled to the synchronous motor serves to increase the synchronous motor speed above synchronism during the starting period after which it coasts down to synchronous speed when operating on 1800-cycle power.

The facsimile signal from the radio receiver circuit (fig. 8-22) is attenuated at the gain control, then amplified in the voltage and power amplifiers. The power amplifier output drives either the recording lamp for photographic recording, or the recording stylus for direct recording.

Another circuit from the power amplifier transmits phasing pulses to the phase amplifier, which operates the phase magnet and clutch during the phasing process just before each picture transmission.

The fork oscillator serves a single purpose of receiving. It generates an 1800-cycle signal, which is amplified to operate the synchronous motor at the same speed as the motor in the transmitting transceiver.

FACSIMILE TRANSCEIVER TT-41B/TXC-1B

Facsimile Transceiver TT-41B/TXC-1B is an electro-mechanical optical facsimile set of the revolving-drum type for the transmission and reception of page copy. It is used for transmission of maps, photographs, sketches and printed or handwritten text over regular voice communications channels, either wire or radio, between two or more stations. Colored copy may be transmitted, but all reproduction is in black, white and intermediate shades of gray. Received copy is recorded either directly on chemically treated paper, or photographically in either negative or positive form. The equipment will transmit or receive a page of copy 12 by 18 inches in 20 minutes at regular speed or in 40 minutes with half-speed operation. A block diagram of the overall equipment is illustrated in figure 8-23.

The complete electrical circuit of the transceiver and power unit may be subdivided into the following eight principal components: (1) fork oscillator unit, (2) photo cell bridge modulator, (3) signal amplifier circuit, (4) phasing circuit, (5) motor circuit, (6) B+ regulator circuit, (7) rectifier power supply, and (8) exciter lamp supply.

The fork oscillator unit (V1, V2, and V3) generates a highly stable 1800-cycle audio signal for distribution to the bridge modulator circuit, the motor amplifier circuit, and the exciter lamp circuit.

The bridge modulator circuit, built around a 5652 photo tube, V24, is used only on transmission. It amplitude-modulates the 1800-cycle signal in accordance with the variations in light intensity of the small portion of the transmitted copy being scanned at any instant.

The signal amplifier circuit (V10, V11, V12, V13, and V14) amplifies the 1800-cycle AM signal. On transmission the circuit is fed by the photo tube bridge modulator; the output goes to the line terminal and thence via auxiliary equipment to the radio transmitter. On reception the circuit is fed by the signal on the line from a radio receiver; the output drives either the stylus for direct recording or the recorder lamp for photo recording. The signal amplifiers are also used in the talk-back circuit.

The phasing circuit, V15 and V16, is used for phasing of the facsimile receiver before each transmission. A series of phasing pulses from the transmitter actuates a clutch mechanism, which positions the receiver drum so that the clamp bars of both drums (receiver) and transmitter) pass the scanning mechanisms at the same time. The clamp bars hold the copy to the drum (fig. 8-24).

The motor circuit (V4, V5, and V6, fig. 8-23) amplifies the 1800-cycle signal from the fork oscillator unit and drives the synchronous motor at constant speed. A multivibrator, V26, is used to provide half-speed operation of the synchronous motor.

The B+ regulator circuit (V7, V8, V9, and V25) provides a closely regulated voltage of 250 volts from the rectifier power supply of 250 volts. Regulated B+ voltage is used on critical transceiver circuits.

The rectifier power supply (V22 in the power unit) provides an unregulated 450-volt B+ output. The power unit also provides 6.3 volts a-c for filament operation and 115 volts a-c for starting the synchronous motor (not shown in the figure).

The exciter lamp supply (V17, V18, V19, V20, V21, and V23 in the power unit) amplifies and regulates the 1800-cycle signal from the fork oscillator unit to provide a constant amplitude,
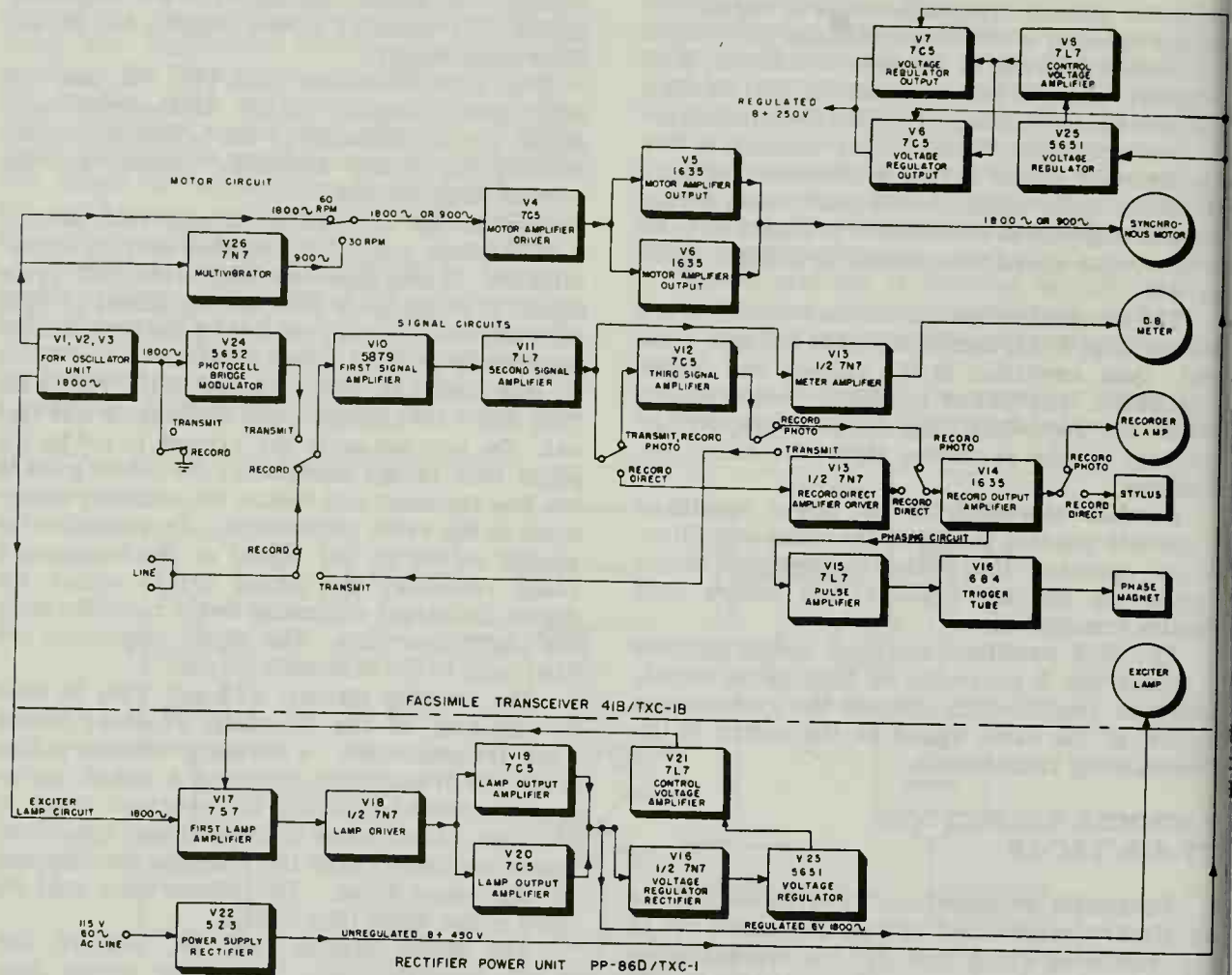
1.243

Figure 8-23.—Block diagram of facsimile transceiver TT-41B/TXC-1B and rectifier
power unit PP-86-D/TXC.

constant frequency signal for the transceiver exciter lamp, which illuminates the transmitter copy with a constant brilliancy.

All electrical operating controls of the facsimile transceiver, except the motor speed-control switch are located on the sloping front panel of the transceiver (fig. 8-25). The motor speed-control switch is located on the left end of the base of the transceiver. Two mechanical controls, the drum engaging lever and the clamp bar, are located on the drum. Input and output connections are located on the right-hand end of the transceiver. The power unit has no operating controls.

The power ON-OFF switch is located on the right-hand side of the control panel. It makes and breaks one side of the 11-volt, 60-cycle AC line circuit feeding the primary of the main power transformer located in the rectifier power unit. In the OFF position power is removed from all parts of the facsimile set. In the ON position plate and filament voltages are applied to all tubes, and power is available for starting and operating the synchroneous motor.

The selector switch located to the left of center on the front panel is the principal control for determining the function of the transceiver. It
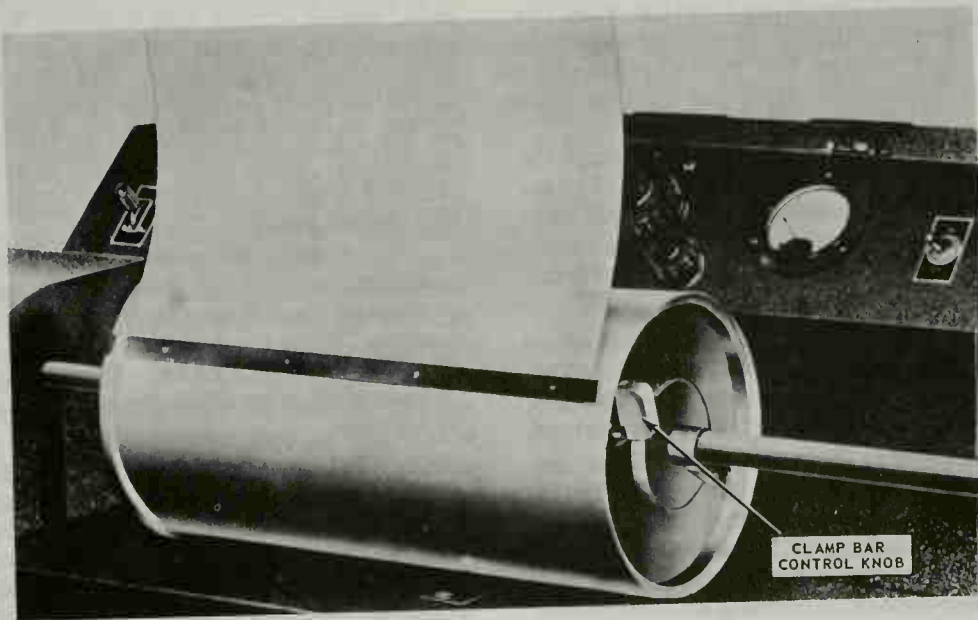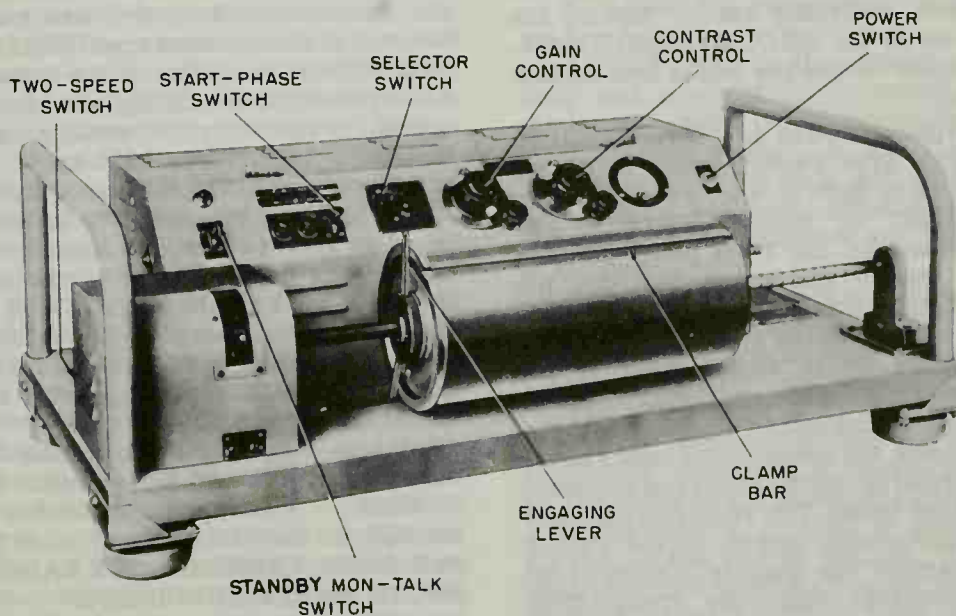
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Figure 8-24.—Placing copy on drum.

as five positions, which are labeled (from top to bottom) TRANSMIT, SET RANGE, STANDBY, RECORD PHOTO, and RECORD DIRECT.

In the TRANSMIT position the facsimile transmitting circuit is established, and impulses representing elements of the facsimile

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Figure 8-25.—Facsimile transceiver controls.

copy are sent out over the facsimile circuit. This is the normal operating position when the transceiver is transmitting copy. In this position the switch also closes the radio transmitter relay circuit associated with the carbon mike jack.

In the SET RANGE position the transmitting circuit is established as in the TRANSMIT position, except that impulses are not sent out over the line. This position is used when the operator is engaged in setting the contact range for transmitting a specific piece of copy.

In the STANDBY position the electron tubes are kept ready for operation of the transceiver. This is the normal setting during periods when copy is not being transmitted or received.

In the RECORD PHOTO position the receiving circuit is established to permit use of the recorder lamp for recording the received copy on photographic paper. This switch position is used to set the gain and to phase the receiving transceiver before recording either "photo" or "direct." A raised stud on the selector switch nameplate enables the operator to set the switch at record PHOTO by "feel" in the dark room.

In the RECORD DIRECT position the receiving circuit is established to permit use of the stylus instead of the recording lamp. The stylus causes recording of received copy on electrosensitive recording paper. Moving the selector switch to the RECORD DIRECT position mechanically moves the stylus into contact with the paper.

The start-phase switch is a three-position, nonlocking lever type switch, which is spring loaded to return to neutral when released from either the START or PHASE positions. The switch is located on the front panel between the selector switch and spare fuse for the motor. The transmitting and receiving operators throw the start-phase switch to the START position momentarily to apply power to the start motor and to bring the synchronous motor above the normal operating speed. When the switch is released, the synchronous motor slows down to its normal operating speed and continues to operate at that speed. The receiving operator throws the start-phase switch to the PHASE position to energize the phasing circuit of the receiving transceiver while the transmitting operator is sending phasing pulses. When the relatively short-phasing operation is completed, the machines remain synchronized.

The motor speed control switch is a three-pole, double-throw toggle switch located on the left end of the base of the transceiver to the rear of the motor cover (fig. 8-25). It is used in the 60 RPM position for normal operation or in the 30 RPM position to provide half-speed operation. When transmitting over long radio circuits under adverse conditions, the 30 RPM half-speed position can be used.

The standby-mon-talk switch is a three-position (two locking, one nonlocking) key switch located at the left on the front panel. The three position, from the top to the bottom, are labeled STANDBY, MON, TALK. The talk-back circuit is operated only when the selector switch is on STANDBY.

In the STANDBY position, with a speaker plugged into the speaker jack, the speaker will be connected in the circuit and the operator can monitor the communication channel. If the standby mon-talk switch is in the STANDBY position, and the selector switch is also in the STANDBY position, the synchronous motor will not operate. This position is used to receive voice communications when it is not necessary to have the motor running.

In the MON position, with a speaker plugged into the speaker jack, the speaker is also connected and the operator can monitor the communication channel. In this position, however, the motor receives power and can be started and run at synchronous speed. This is the normal position for receiving voice communications, when phasing transmitting and receiving equipments, preparatory to sending and receiving facsimile copy. Some 1800-cycle interference will be noticed in the talk-back circuit.

In the TALK position, with a speaker plugged into the speaker jack, the speaker becomes the microphone of a voice intercom system. The switch must be held in the TALK position; if pressure is released, it will snap back to the MON position. The TALK position is used when talking to the facsimile operator on the other end of the circuit.

It should be noted that the talk-back circuit only operates when the selector switch is at STANDBY. Voice communications cannot be carried on over this circuit when the selector switch is at TRANSMIT, SET RANGE, RECORD PHOTO, or RECORD DIRECT.

It should also be noted that the talk-back circuit will not function over a radio circuit that uses auxiliary radio equipment unless the

auxiliary equipment is bypassed for voice communications.

The contrast control is a calibrated potentiometer located to the left of the db meter on the front panel of the transceiver; it adjusts the contrast range (difference between minimum and maximum signal strength) of the transmitted signal by adjusting the magnitudes of the currents in the photo-tube bridge circuit. This control is operated only by the transmitting operator. The dial on this control is numbered from 1 to 100, with the applicable ranges for use when transmitting either positive or negative indicated on the dial.

The contrast control may be used to balance the bridge in the photo-cell bridge modulator (V24, fig. 8-23) either on maximum light (white) or on minimum light (black). In either case when the bridge is balanced, the output signal has a minimum amplitude. Thus, if the contrast control is adjusted so that the bridge is balanced when the light is a maximum, the output signal will have minimum amplitude on white and maximum amplitude on black. This type of transmission is called positive transmission. When the bridge is balanced on minimum light, the output signal will have minimum amplitude on black and maximum amplitude on white. This type of transmission is called negative transmission.

The gain control, located between the contrast control and the selector switch, is a dual potentiometer, which serves to control the level of the signal handled by the transceiver both in transmitting and receiving. It does not change the contrast range.

The front panel db meter, which serves as a guide in setting gain and contrast controls, is calibrated in db with reference to a zero-power level of 6 milliwatts in 600 ohms.

To turn the transceiver on, throw the power on-off switch to ON and turn the selector switch to SET RANGE. Normally a 5-minute warmup period should be allowed before making adjustments or actually transmitting.

To place copy on the drum, raise rear edge of clamp bar (fig. 9-24) by turning the clamp-bar control knob to its clockwise position. Place the lower edge of the copy (face up) under the rear edge of the clamp bar. Turn the knob to its counterclockwise position, thus closing the clamp on the leading edge of the copy and opening the other clamp. Revolve the drum forward, wrapping the copy around the drum. Insert the loose edge of the copy under the clamp; pull the copy tight around the drum with a wiping motion of the hand, and close the clamp by turning the clamp-bar control knob back to its center position. Examine the copy after it is placed on the drum to make sure there are no bulges.

The following adjustments (fig. 8-25) should be made for transmitting positive for direct recording:

1. Turn selector switch to SET RANGE.

2. As a starting point, set the gain control at 65 and the contrast control near zero.

3. Turn the drum so that the whitest portion of the copy is illuminated by the spot of light. Carefully turn the contrast control to a higher setting until the meter reads the required minimum. Check the meter reading and move the drum slightly backward and forward while examining a different white portion of the copy to be sure that the whitest spot is used. For radio operation, the required minimum will depend on the type of auxiliary equipment being used with the radio transmitter. The wire line minimum (1/16 inch below the -10 calibration mark on the db meter) is recommended.

4. Move the drum so that the blackest portion of the copy is illuminated by the spot of light. Be sure that the blackest portion actually is used by again rotating the drum slightly backward and forward while checking the meter.

5. Adjust the gain control so that the meter reads + 2 DB.

6. Shift back to the whitest portion of the copy and readjust the contrast control for a minimum reading of -10 DB on the meter.

7. Repeat steps 4, 5, and 6 until the final adjustments of the gain and contrast controls give meter readings, differing by at least 12 DB between the blackest and whitest portions of the copy.

Adjustments for transmitting negative are not given here in the interest of brevity but may be obtained by referring to the appropriate technical manual.

The transmission of facsimile signals by amplitude modulation on the radio channel ordinarily results in a change in the loudness or levelness of the received signals with fading. If fading occurs, streaks or level changes will appear in the record picture. Therefore it is desirable to transmit a type of signal over radio circuits whose amplitude when demodulated, at the receiving station will be independent of changes in the volume or level of the radio signal. This action may be accomplished by a

frequency shift of the signal in which different picture values are represented by different frequencies. Because the facsimile transceiver generates an amplitude-modulated signal in transmitting and the recorder is designed to operate on an amplitude-modulated signal, picture signal conversion equipment is necessary between the facsimile transmitter and the radio transmitter, and between the radio receiver and the facsimile recorder.

The TT-66( )/TXC facsimile transceiver, not shown, is a newer transceiver that performs the same functions as the TT-41B/TXC-1B. Because they have different operating speeds, however, the two transceivers are not compatible. The newer set has selectable operating speeds of either 90 RPM or 45 RPM, whereas the older equipment operates at either 60 RPM or 30 RPM.

The components of two radio facsimile transmission systems (fig. 8-26, B) will be described in the following portion of this training course. The first system employs audio frequency=shift (AFS) keying (normally short range) and includes one auxiliary unit between

the facsimile transmitter and the radio transmitter. This unit is the modulator, radio MD-168/UX. The other system is carrier frequency-shift (CFS) keying (long range), and includes two auxiliary units between the facsimile transmitter and the radio transmitter. These units include Keyer Adapter KY-44A/FX and Frequency-Shift Keyer KY-75/SRT. (Only the Keyer Adapter will be required if the Frequency Shift Keyer is built into the radio transmitter.)

Modulator, Radio MD-168/UX

Modulator, Radio MD-168/UX (fig. 8-27) is designed to convert amplitude modulated facsimile signals from a facsimile transmitter (for example, Facsimile Transceiver TT-41B/TXC-1B acting as a transmitter) to audio frequency-shift facsimile signals of 1500 to 2300 cycles suitable for modulating a radiophone transmitter.

A block diagram of the modulator is illustrated in figure 8-28. Basically, the unit consists of a preamplifier; a keyer (detector); a variable-frequency, phase-shift oscillator; a
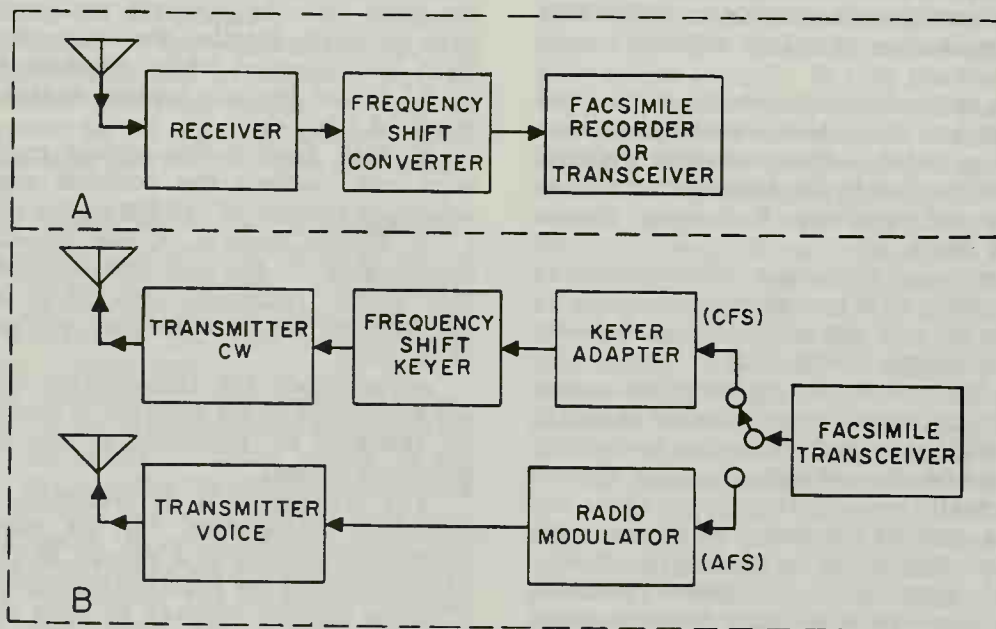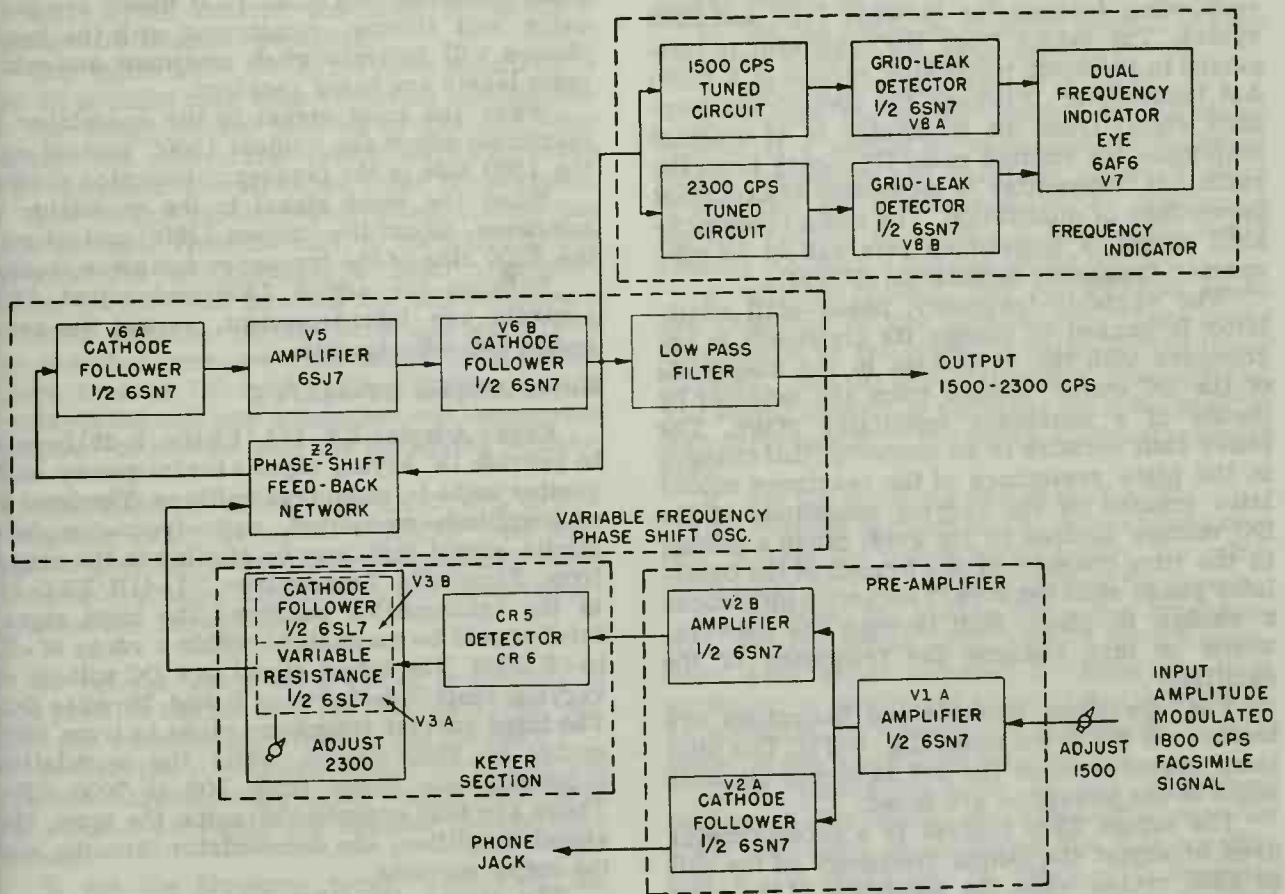

Figure 8-26.—Radio facsimile systems.

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Figure 8-27.—Modulator radio MD-168/UX.

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Figure 8-28.—Block diagram of modulator radio MD-168/UX.

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frequency indicator; and a power supply (not shown). The amplifier increases the received facsimile signal to the proper level for operating the phase-shift oscillator and a pair of earphones. The input signal to the modulator may be monitored by using the phone jack. The frequency limits of the output signal may be monitored by means of a dual-type, tuning-eye indicator.

The input signal to the modulator has a frequency of 1800 cycles and an amplitude that varies in accordance with the light and dark segments of the picture being scanned at the facsimile transmitter.

The output signal from the modulator is an audio signal in which 1500 cycles represent the maximum signal input (in amplitude) and 2300 cycles represent the minimum signal input (in amplitude) to the modulator from the facsimile transmitter. Amplitudes between maximum and minimum signals are changed to corresponding frequencies between 1500 and 2300 cycles. The output from the modulator is connected to the audio modulator section of a radio AM transmitter. Because the audio frequency-shift signal from the modulator is of constant amplitude the emitted radio frequency from the radio AM transmitter is modulated at a constant percentage of modulation. The audio frequency-shift signal is sometimes referred to as subcarrier-frequency modulation (SCFM).

The variable-frequency, phase-shift oscillator is caused to change its frequency in accordance with the variations in the magnitude of the DC output voltage from the detector by means of a reactance modulator stage. The phase shift network is so connected that changes in the plate resistance of the reactance modulator (caused by the varying magnitude of the DC voltage applied to its grid) cause a change in the time constant of one branch of the oscillator phase shift network. This action introduces a change in phase shift through the network, which in turn changes the frequency of the oscillator.

The operating controls and indicators are located on the front panel (fig. 8-27). The pilot lamp operates when the set is turned on. Both sides of the powerline are fused.

The adjust 2300 control is a potentiometer used to adjust the output frequency of the unit to 2300 cycles when the amplitude of the input facsimile signal is a minimum.

The adjust 1500 control is a potentiometer used to adjust the output frequency of the unit to 1500 cycles when the amplitude of the input facsimile signal is a maximum.

The 2300-1500 frequency indicator is used to indicate the proper frequency limits of the output audio frequency-shift signal. The left half of the indicator closes when the output signal has a frequency of 2300 cycles and the right half closes when the output signal has a frequency of 1500 cycles.

The gain of the amplifier is such that when the adjust 1500 control is set at the proper position, the audible level at the earphones will be a comfortable level.

When the power ON-OFF switch is in the OFF position, power is removed from the entire unit.

When operating the unit, throw the power switch on and allow a 5-minute warmup period. Turn the adjust 1500 control to the extreme clockwise position. When an incoming signal is being received, the 2300-1500 tuning eye indicator will flicker; monitoring with the headphones will indicate when maximum and minimum levels are being received.

When the input signal to the modulator is maximum adjust the "adjust 1500" control until the 1500 side of the frequency indicator closes.

When the input signal to the modulator is minimum, adjust the "adjust 2300" control until the 2300 side of the frequency indicator closes.

Because the adjust 2300 and adjust 1500 controls are interdependent, repeat the preceding two adjustments.

Keyer Adapter KY-44A/FX

Keyer Adapter KY-44A/FX (fig. 8-29) is used to provide DC keying signals for frequency-shift exciter units in radio transmitters. The input is an amplitude-modulated, audio-frequency, facsimile signal that may be similar to the output from Facsimile Transceiver TT-41B/TXC-1B in the transmitting position. The input signal level should be maintained within a range of -20 to +6 DBM. The output signal is a DC voltage of varying amplitude between 0 and 20 volts DC. The input carrier frequency range is from 1500 cycles to 7000 cycles, while the modulation sidebands can range from 100 to 7000 CPS. There are four essential circuits: the input, the signal amplifier, the demodulator circuits, and the output circuits.

The power ON-OFF switch turns the set completely on or off. A preliminary warmup period of a few minutes should be allowed before using the set.
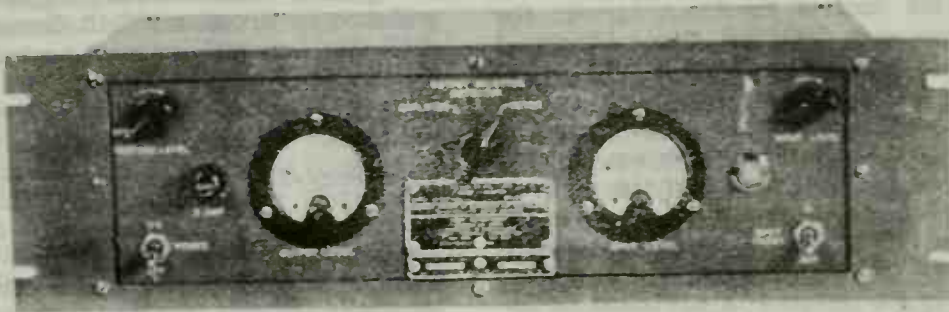
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Figure 8-29.—Keyer adapter KY-44A/FX.

The test osc. level control is used for alignment of the unit, and to set the frequency shift keyer. Setting the level control to 12V will shift the FS keyer to mark, a zero setting will present a space signal to the keyer (such as KY-75/SRT). This control should be in the OFF position for normal operation.

The input-filter IN-OUT switch should be in the IN position only when receiving frequencies in the range of 900 to 2500 cycles and when signals are noisy. The switch should be in the OUT position for all other input frequencies.

The output selector switch determines the type of output from the unit. In the AMPLIFIER position the unit becomes a linear amplifier. In the DETECTOR position the unit becomes a detector with the carrier still unfiltered from the detected envelope. The output level meter reads average DC values. The KEYER position converts the unit into a detector with the carrier removed to provide only the DC keying signals.

FREQUENCY-SHIFT KEYER KY-75/SRT.— Frequency-shift keyer KY-75/SRT was discussed previously with teletype equipment. Some common operating adjustments for photo transmission are listed below.

1. Set the test operate switch at PHOTO. In this position the limiting (V-101 and V-102, fig. 8-5) and waveshaping circuits are not used.

2. Set the crystal-oscillator switch at the position corresponding to the socket position of the crystal providing the desired channel frequency.

3. Set the freqency range switch to the desired output frequency of the keyer. This frequency is the sum of the crystal frequency and the 200 KC frequency from V-107.

4. Set the input filter switch at PHOTO.

5. Set the multiplier switch at the position corresponding to the frequency multiplication factor employed in the transmitter. For example, if the multiplcation factor is 8, the switch should be set at "X8."

6. Set the phase-modulation control at OFF (extreme counterclockwise position).

7. Set the metering switch at GRID.

8. Unlock the tuning control. Set the tuning control at a setting corresponding to the keyer output frequency and carefully adjust it about this setting for a maximum meter reading. A normal reading is about 1.5 MA (actual meter reading 0.5 MA). Lock the tuning control.

It will be noticed that two current peaks are observed on the panel meter. These peaks correspond to the resonant peaks for the lower sideband, and the upper sideband, respectively. The tuning control is normally set at the position that corresponds to the upper sideband resonant peak.

9. Set the metering switch at PLATE.

10. Release the lock on the output tuning control. Adjust the output tuning control for minimum plate current, as indicated on the meter.

11. Release the lock on the output level control. Set the output level control for the maximum grid drive required to drive the first amplifier or multiplier stage of the transmitter, as indicated by a maximum reading on the grid meter of the associated transmitter.

Care should be taken in this adjustment because if the tuning range is located near the lower markings on the tuning dial it is possible that a dip may also be obtained in the plate current near the higher markings of the dial due to the second harmonic of the keyer frequency.

12. Set the deviation control at the desired deviation. This control functions to vary the amount of frequency deviation. The control dial has a multiplication factor of 100 for FSK operation and 200 for photo operation.

In normal operation the deviation control is adjusted to obtain the desired deviation, as read directly on the calibrated dial of the control. After the multiplication factor employed in the transmitter is determined, the multiplier switch is set at a position corresponding to this factor. In this manner the amount of deviation is determined and held constant despite any ensuing multiplication in the transmitter. Fine adjustment of the shift may be checked on a receiver test setup, by varying the KY44A/FX test. oscillator level control from 12 V to OV and reading the total shift of the keyer (900 cycles for photo).

FACSIMILE RECEIVING EQUIPMENT

To receive radio facsimile signals of either radio carrier frequency-shift transmission or audio frequency-shift transmission, conventional superheterodyne receivers may be used. With either system, the signal output of the radio receiver is an audio frequency-shift signal in which 1500 cycles represent the maximum signal and 2300 cycles represent the minimum signal output from the facsimile transmitter at the sending terminal.

In order to convert the audio frequency-shifted signal output of the radio receiver into an amplitude-modulated signal suitable for operation of a facsimile recorder or transceiver, an additional unit must be interposed between the

receiver and the recorder (fig. 8-26, A). This unit may be Frequency Shift Converter CV-172-A/U. The facsimile recorder may be similar to Facsimile Recorder RD-92A/UX or the previously discussed transceiver TR-41B/TXC-1B.

Frequency-Shift Converter CV-172A/U

Frequency-Shift Converter CV-172A/U (fig. 8-30) is used to convert 1500- to 2300-cycle facsimile signals received from a radio circuit to AM signals suitable for operating a facsimile recorder (for example, Facsimile Recorder RD-92A/UX). The unit contains provision for audible monitoring of the incoming signal and for visual checking of the frequency limits.

Another method that may provide for even more reliable tuning is to turn off the auxiliary oscillator and tune slowly through the frequency-shift signals. As the signals are approached from one side a thumping sound will be heard. Continuing to rotate the tuning dial slowly in the same direction will cause this sound to disappear almost completely and then to reappear again before it disappears on the other side of the signal. The proper tuning position corresponds to the quiet zone between the two thumping sounds.

Satisfactory frequency-shift reception may be obtained with high stability radio receivers without the need for an auxiliary oscillator provided the receivers are properly tuned. If the converter and BFO have been sufficiently stabilized, the BFO can be used as the heterodyne oscillator to provide the audio frequency
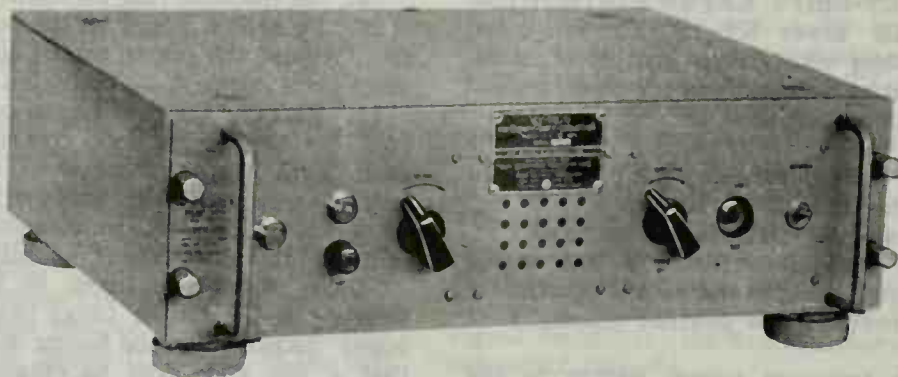

Figure 8-30.—Frequency-shift converter CV-172A/U.

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utput. Turn off the BFO before making the uning adjustments. After the proper tuning adjustments are made, as previously described, urn on the BFO and vary its control for the proper audio frequency output (1500 to 2300 ycles) from the radio receiver. Do not change he receiver tuning to vary the audio frequency.

The ET who is familiar with tuning c-w radio signals has observed that a-f or tone output from the radio receiver changes as either the main tuning dial or the BFO control is moved. Thus it is possible to have the right sense of signals (no inversion) and still not be tuned properly for optimum results. Improper tuning results in an inferior signal-to-noise ratio and may cause broadening of lines or multiple images somewhat similar to those caused by multipath transmission of the radio signals.

The facsimile signal obtained from the radio receiver is fed through an amplifier and filter to a limiter and then through a frequency discriminator (fig. 8-31). The output of the amplifier is controlled by an input level control potentiometer to adjust the signal level to the limiter. A loudspeaker with a separate volume control connects to the output of the line amplifier. This output also feeds two tuned circuits resonated at 1500 and 2300 cycles respectively to operate a tuning eye frequency indicator.

The input signal from the radio receiver is an audio frequency-shift signal in which 1500 cycles represent black and 2300 cycles represent white for the RD-92A/UX recorder. The discriminator is a low-pass filter having a cut-off frequency of 1500 cycles. The output from the filter at 2300 cycles can be adjusted to be from 10 to 20 DB lower than the output at 1500 cycles. Thus, the input signal to the coverter is changed from an a-f-s signal of substantially constant amplitude to an output signal of varying amplitude in which the maximum amplitude (corresponding to 1500 cycles) is 10 to 100 times the minimum amplitude (corresponding to 2300 cycles).

To operate the frequency-shift converter, turn the input level control clockwise (fig. 8-30). This action applies power to the unit. Allow a few minutes warmup time.

Adjust the input level and volume controls fully clockwise until a signal is received.

When a signal is received, adjust the input level control until the speaker does not distort. (Distortion exists when there are overtones.) This adjustment provides sufficient level to operate the limiter.

Adjust the volume control for a convenient speaker level.

When a 1500-cycle signal is being received, the "1500" tuning eye will close and the output will be the maximum signal level for the facsimile recorder. Also, when a 2300-cycle signal is being received, the "2300" tuning eye will close. The difference in output levels will depend upon the trimmer control (not shown), which provides a variable-resistance bypass around the discriminator circuit.

When the radio receiving equipment is properly adjusted, the tuning eye indicators will show that 1500- and 2300-cycle signals are being fed to the input of the frequency-shift converter. When a copy that has mostly background is being received, the high stability radio receiver output may be adjusted by changing the BFO frequency to close the tuning eye corresponding most closely to the background signal frequencies.

The CV-172 A/U is not the only frequency shift converter used by the Navy in facsimile installations, but it is the one most commonly found aboard ship. Others you may encounter are models CV-97/UX and the CV-1066/UX. They all perform the same function.

Facsimile Recorder RD-92A/UX

The output of Frequency-Shift Converter CV-172A/U may be fed to Facsimile Recorder RD-92A/UX (fig. 8-32) to make recordings of copy transmitted by Facsimile Transceiver TT-41B/TXC-1B (previously described) or equipment having the same transmission characteristics. The facsimile recorder is a self-contained unit comprising an amplifier-power supply chassis and four plug-in assemblies. The assemblies (fig. 8-33) are: (1) the audio frequency oscillator, (2) the amplifier detector, (3) the amplifier modulator, and (4) the recorder subassembly. A circuit test switch mounted on the front panel provides means for quickly testing all important circuits. Neon light indicators across individual fuses and tube heaters instantly indicate a blown fuse or burned out tube heater filament.

Facsimile Recorder RD-92A/UX performs its functions of recording pictures, drawings, or messages by rotating a drum at a constant speed, while feeding a stylus needle along the drum, one scanning line for each revolution until the complete drum has been covered. This function is performed by means of three
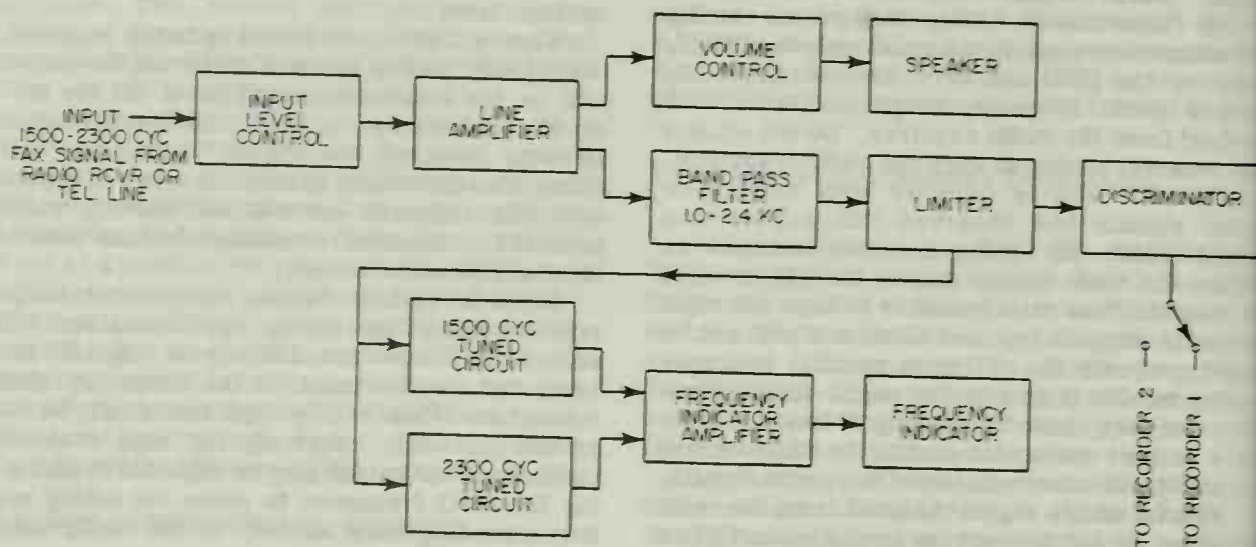
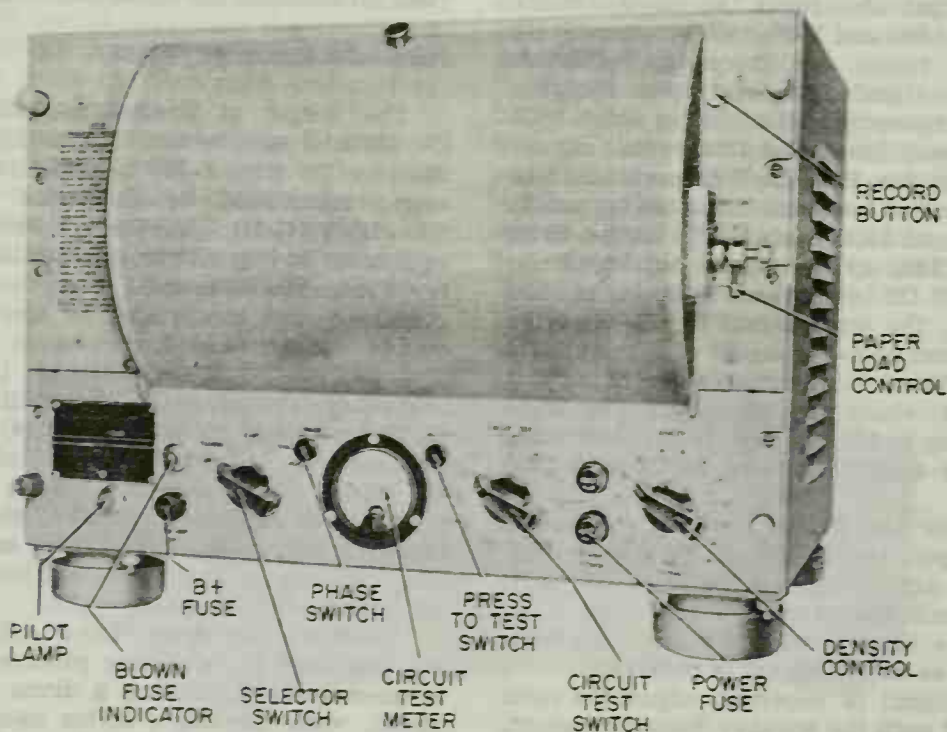Figure 8-31.—Block diagram of frequency-shift converter CV-172A/U.

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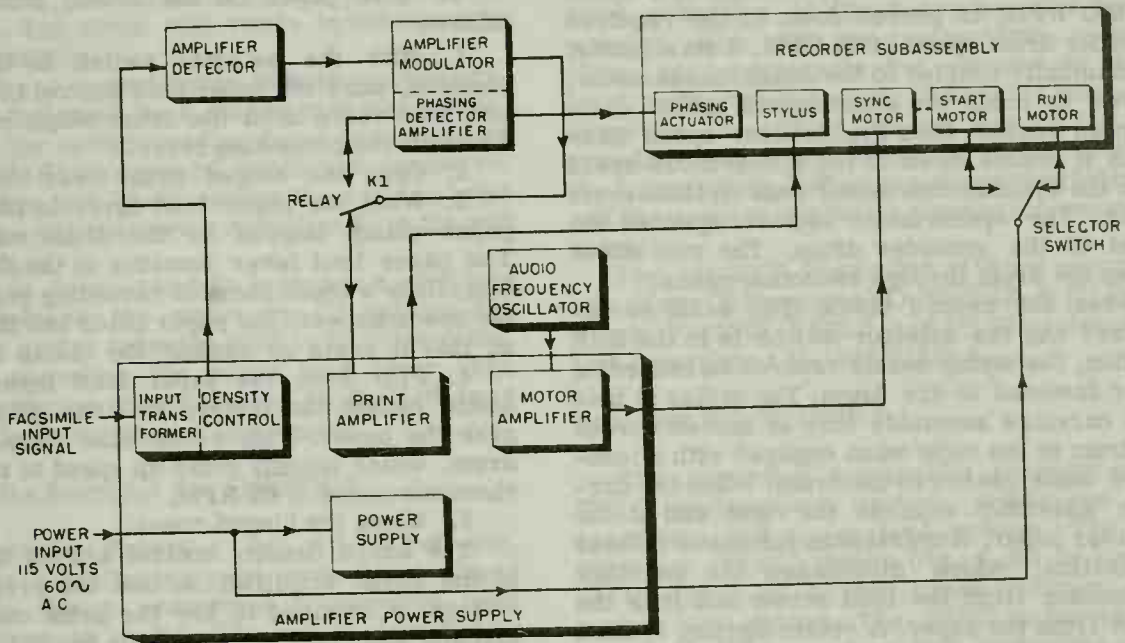Figure 8-32.—Facsimile recorder RD-92A/UX.

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Figure 8-33.—Simplified block diagram of facsimile recorder RD-92A/UX.

motors and suitable gears and mechanical linkages contained within the recorder subassembly. The power, signal, and phasing currents for proper operation of the recording mechanism are obtained from the outputs of the electronic subassemblies.

The amplifier-detector unit (fig. 8-33) receives the input facsimile signal. This signal, consisting of phasing pulses and facsimile intelligence between the frequency limits of 900 and 2700 cycles, is amplified by a class A amplifier and demodulated through the action of a full-wave rectifier and low-pass filter to form a varying DC facsimile signal.

The varying DC facsimile signal output of the amplifier-detector unit is coupled to the amplifier modulator subassembly where it combines with a signal from a 15 KC oscillator in a modulator stage. The 15 KC output of the modulator varies with the amplitude of the d-c facsimile signal. The resulting modulated signal connects through the normally closed contact of relay K1 to the print driver and amplifier in the amplifier-power supply assembly.

The signal from the amplifier modulator unit is amplified by the print driver and amplifier to sufficient intensity, so that, when connected to the stylus needle in the recording mechanism

assembly, it will record on the recording paper points representing varying shades of dark elements of the transmitted copy.

For phasing pulses, relay K1 is operated to transfer the signal from the amplifier modulator to the phasing detector and amplifier, located in the amplifier-modulator subassembly. The phasing pulses, which are transmitted at the beginning of each copy, are amplified to operate the phasing actuator in the recorder subassembly.

The phasing actuator, on receiving the phasing pulses, releases the stop bar on the synchronous-drive mechanism in the proper position to frame the recorder drum with the drum of the facsimile transmitter.

The AF oscillator generates an 1800-cycle signal which is amplified by the buffer amplifier (not shown) and motor amplifier in the amplifier-power supply assembly and the coupled to the recording mechanism assembly to operate the synchronous motor. The 1800-cycle signal also connects to the test signal position on the density control (fig. 8-33), where it is available for connection to the input of the signal amplifier for testing the various circuits with the circuit test switch.

The synchronous motor, rotating at a speed of 1800 RPM, is geared down to the required recorder drum speed of 60 RPM. A start motor, mechanically coupled to the synchronous motor, serves to bring the synchronous motor up to a speed higher than synchronous speed after which it coasts down to the synchronous speed when the synchronous motor runs on 1800-cycle power. The synchronous motor regulates the speed of the recorder drum. The run motor drives the drum through reduction gears.

When the record button (fig. 8-32) is depressed and the selector switch is in the RUN position, the stylus needle records on recording paper fastened to the drum. The stylus is held in a carriage assembly that is moved across the drum to the right when engaged with a lead-screw shaft geared to the drum. When the carriage assembly reaches the right end of the recorder paper, it operates an automatic release mechanism, which disengages the carriage mechanism from the lead screw and lifts the stylus from the paper. A return spring, located in the left side gear box, then pulls the carriage back to the left side of the drum so that it will be ready for the next copy.

OPERATING ADJUSTMENTS—To operate Facsimile Recorder RD-92A/UX (fig. 8-32) proceed as follows:

1. Turn the selector switch to the STANDBY position. Note that the pilot light lights up, indicating that power is being applied to the recorder. Wait about 1 minute for the tubes to warm up. The recorder may be left at STANDBY While waiting for a transmission so that the set is ready for immediate operation.

2. Turn the selector switch to the START position. Wait about 5 seconds for the start motor to bring the synchronous motor above synchronous speed.

3. Turn the selector switch to the SYNC position. Wait until the synchronous motor coasts down, then locks in synchronous speed. This is distinguished by a distinctive high-pitch tone.

4. If the motor does not lock in but falls below the synchronous speed, switch back to the START position and repeat steps 2 and 2. If the motor does not come down to synchronous speed, turn to STANDBY and allow the motor to stop. Omit step 2 and switch directly to the SYNC position.

5. Turn the selector switch to the RUN position. It is necessary for the drum to rotate into the proper position for loading the paper.

To load paper on the drum, proceed as follows:

1. With the selector switch in the RUN position, push the paper load control to the left and hold there until the drum stops rotating. Then lift the projecting lever.

2. Open the hinged cover over the drum.

3. When the paper load lever is lifted, the paper clamp fingers on the drum will open. The paper load lever remains in the UP position. Drop a fresh sheet of recording paper into the space between the paper guide and the drum so that it rests up against the clamp fingers.

4. Flip down the paper load lever. This action causes the fingers to close quickly and grab the paper. This action also releases the drum, which quickly picks up speed to the synchronous speed of 60 RPM.

5. Close the hinged cover.

The adjust density control affects the gain of the signal amplifier so that the proper d-c voltage is obtained to key the print oscillator circuit. Incorrect setting of the density control may result in faulty recording. Proceed as follows:

1. Set the density control when facsimile signals are being received. When steady signals of maximum signal level are received, preferably on phasing signal, start near zero and advance the control to the lowest point that gives maximum reading on the meter on the front panel. This reading is normally about 100.

2. In some types of copy it is desirable to advance the dial setting of the density control slightly beyond the point that gives a maximum meter reading. Try this procedure if the copy is too light.

3. Leave the density control at the setting that gives the best recording.

The phasing operation is performed with the recorder drum stationary. Phasing pulses may be identified by a downward dip of the meter pointer occurring once a second. To phase the recorder proceed as follows:

1. Switch to the SYNC position and wait for the drum to stop.

2. Turn the phase button to LOCAL.

3. When phasing pulses are received, depress the phase button and hold depressed for five pulses. While phasing, two clicks per second are usually heard; one when the phasing actuator trips, and another when the stop bar passes the drum drive coupler. Press and hold for 5 meter pulses.

4. Turn the selector switch to the RUN position. The drum will rotate in the properly phased position.

5. Press record button when the copy starts. This is indicated by a change in meter pulses. Usually the meter reading drops down to about zero and flicks upward instead of downward. The stylus now feeds across the drum to print the copy, and releases automatically at the end of travel.

If it is desired to take less than a complete copy, release the stylus by (1) turning the selector switch from the RUN to SYNC position, or (2) operating the paper load lever. The stylus returns automatically to the left-hand end of the drum when it is released.

Facsimile Recorder Set AN/UXH-2

Facsimile Recorder Set AN/UXH-2 (fig. 8-34) is a continuous page recorder designed to make direct recordings transmitted over land wires or radio. The set is designed to operate at 60, 90, or 120 scans per minute. When receiving from a transmitter with the proper signals, the unit will automatically phase start recording at the beginning of a transmission, stop when the transmission is complete, and compensate for changes in signal level during the recording. When this automatic operation is utilized the set may be left unattended. Manual operation may be utilized when the transmitting station is not equipped to transmit the necessary control signals for automatic operation.

The AN/UXH-2 consists of three major assemblies: (1) Recorder RO-76/UXH-2, (2) Power Supply PP-1901/UXH-2, and (3) Electrical Control Amplifier AM-1845/UXH-2.

Recorder RO-76/UXH-2

The recording mechanism consists of the SYNC system, run system, stylus and base assembly, and paper feed system. The synchronous (SYNC) motor (fig. 8-35) is brought up to speed by a split-phase induction start motor (not shown). The SYNC motor is then supplied with power by a fork-controlled oscillator circuit, and operates at a synchronous
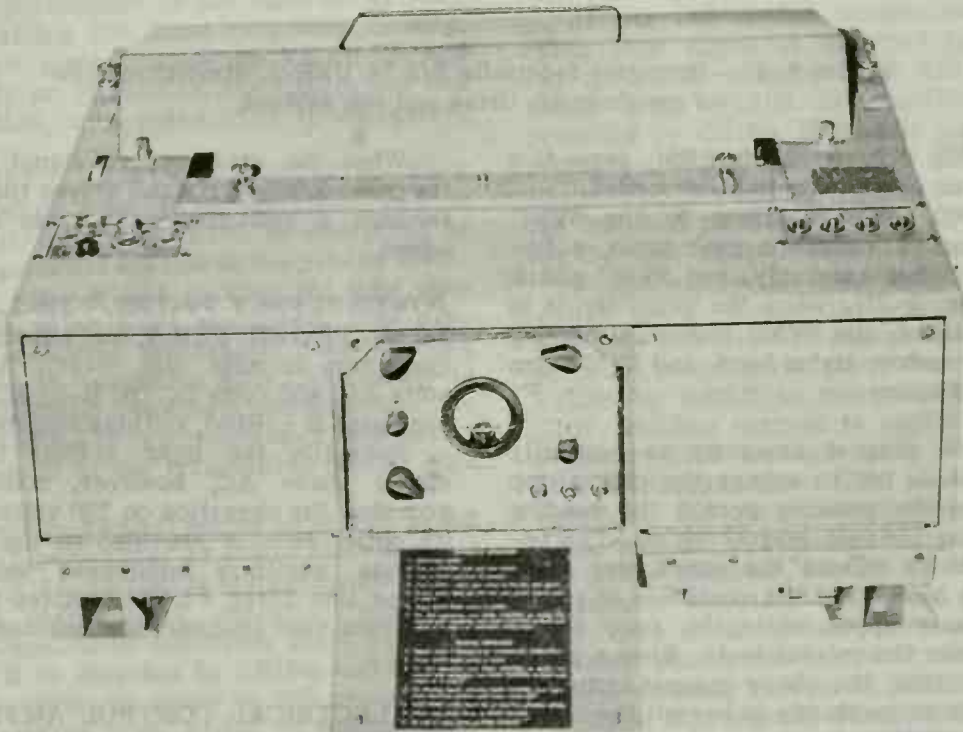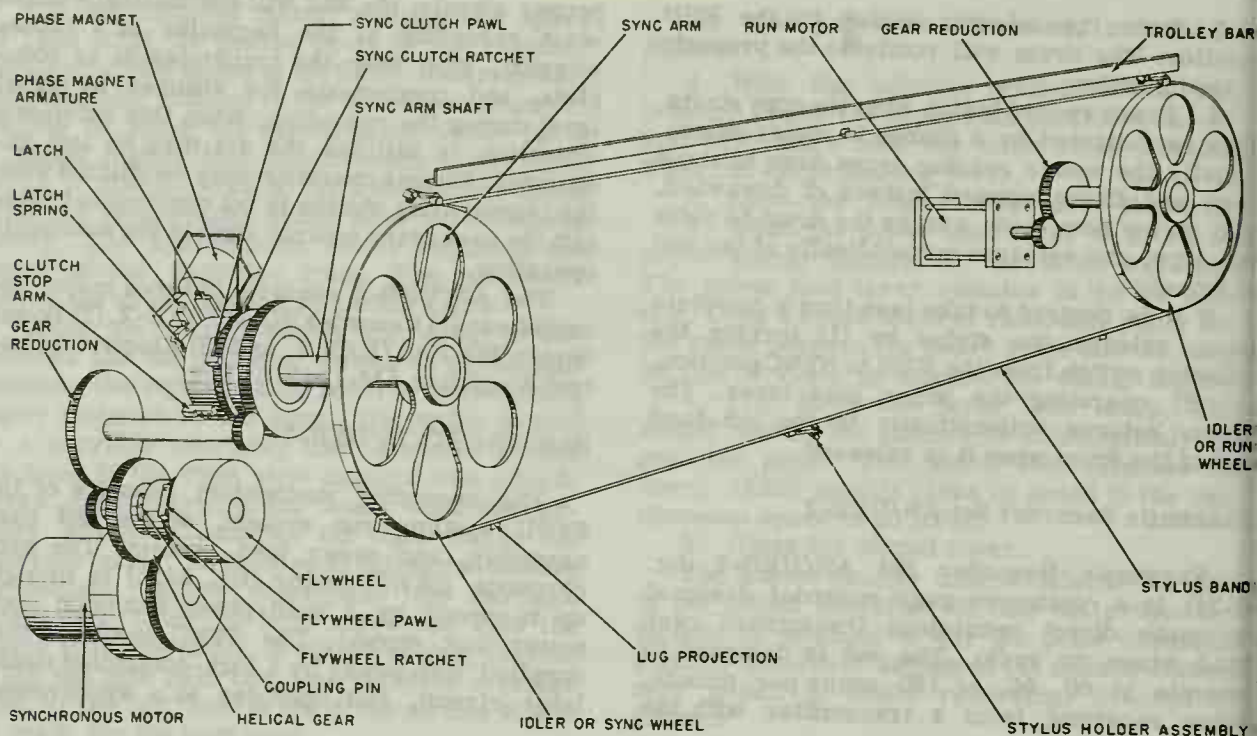

70.88

Figure 8-34.—Facsimile recorder set AN/UXH-2.

70.89

Figure 8-35.—Recorder facsimile RO-76/UXH-2, simplified view
of synchronous drive and run system.

speed of 600, 900, or 1200 RPM, depending upon the scans per minute to be recorded.

The SYNC shaft is driven by the SYNC motor through a reduction gear train, a flywheel and ratchet assembly, and SYNC clutch assembly (fig. 8-35). When the recorder is in standby condition, the SYNC motor is running and the run motor, stylus band, and SYNC arm shaft are stationary.

The phase magnet armature is normally held back so as not to engage the clutch stop arm. During the phasing period the holding current to the phasing magnet is off. The released armature blocks the clutch stop arm, but the sync clutch ratchet continues to rotate at synchronous speed while the sync clutch pawl slips over the ratchet teeth. When a phase pulse is received, the phase magnet armature is pulled back momentarily to permit the clutch to resume rotating in phase at synchronous speed. Subsequent phase pulses will pull the phase magnet armature back each time the stop arm passes it.

When the start-record signal is received, the run motor starts and drives the stylus band through a reduction gear train and the run wheel.

POWER SUPPLY PP-1901/UXH-2.—The Power Supply PP-1901/UXH-2, and associated voltage regulator circuits (fig. 8-36), furnishes 6.3 volts AC, 400 volts DC (HI B+), 225 volts DC regulated B+ (RB+), and bias supplies for the set.

Normally the input voltage is 115 volts single phase AC, however, connections are provided for operation on 220 volts AC. A utility outlet J601 is provided on the input line to operate auxiliary equipment when desired. Power line filter FL 601 isolates the recorder set from other equipments connected to the power line.

ELECTRICAL CONTROL AMPLIFIER AM-1845/UXH-2.—Electrical Control Amplifier AM-1845/UXH-2 consists of signal amplifier circuits, automatic level control and print amplifier circuits, automatic control circuits, fork

oscillator and SYNC motor drive circuits, and speaker amplifier and test signal circuits.

SIGNAL, AMPLIFIER, ALC, AND PRINT CIRCUITS.—The input facsimile signal from the secondary of T301 (fig. 8-36), is fed through gain control R301 to ALC amplifiers V1A and V1B. ALC lockout switch S306 is a single-pole double-throw switch and is part of gain control R301. If the gain control is set at the automatic position, the ALC feedback network controls the bias on V1A and B, automatically compensating for variations in the level of the facsimile signal. Setting the control at the manual position locks out the feedback network, and bias voltage 3 derived from the regulated power supply is applied to the ALC amplifiers. Resistor R301 is then used to adjust the magnitude of the signal input to V1A.

The output signal voltage from V1B is fed to print driver amplifier V4A, then to print amplifiers V5 and V6. Contrast potentiometer R27B is used to control the amount of bias on the grids of V5 and V6. Potentiometer R27A is ganged to R27B so as to vary the signal input to V4A to correspond to the contrast setting, thus maintaining the same recording density. There is no plate voltage on V5 or V6, and consequently no voltage on the trolley bar and stylus needles, until relay K102 is energized.

AUTOMATIC CONTROL CIRCUITS.—When selector switch S301 is turned from OFF to SYNC, start/stop relay K101 is energized by a positive potential applied to the grid of start/stop relay tube V102B to keep the tube conducting and relay K101 energized. This prevents the recording mechanism from starting before S301 is turned to RUN. When S301 is turned to RUN and awaiting a transmission, the positive potential is removed from V102B so that it will respond to the transmitted control signals.

The amplified signal output from V4B is fed through high-pass filter FL103 to control signal detector CR104, and out to the automatic start/stop and start-record circuits. When the amplitude-modulated 300-cycle start signal is received, it is detected by CR104, and filtered by FL102 which is tuned to 300 cycles. The 300-cycle output from FL102 is fed to start/stop signal amplifier V101A. The output of V101A is rectified by CR102 to cut off start/stop relay amplifier V102B, and deenergize relay K101. Deenergizing K101 operates contacts to release the armature of phase actuator E501, energize phasing timer K1, and tune FL102 to the 450-cycle stop signal frequency.
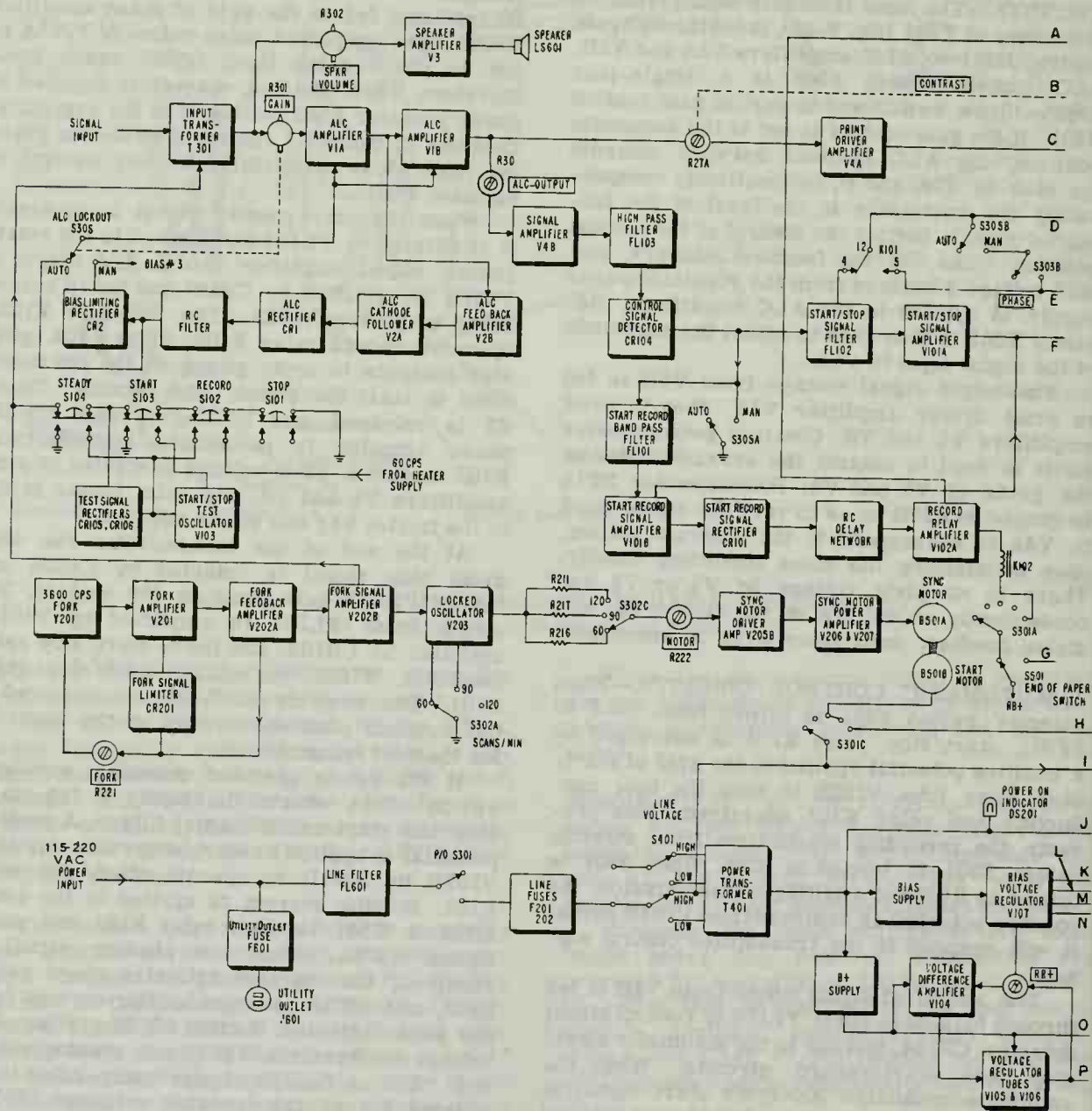
After the start signal, phasing pulses are transmitted for 15 seconds. The phase pulse is filtered and fed to the grid of phase amplifier V205A. The amplified pulse output of V205A is fed to phase delay lamp DS202 which fires thyratron V204. Holding current is supplied to phase actuator E501 pulling back the armature. Contacts of timer K1 remain open during phasing, then close maintaining holding current to actuator E501.

When the start record signal is received, it is filtered by FL101 and applied to the start-record signal amplifier V101B. The output of V101B is rectified by CR101 and fed to record relay amplifier V102A. The output of V102A operates record relay K102. Relay K102 operates contacts to apply power to the run motor B502 to start the stylus band rotating. Timer K1 is released, and holding current for the phase actuator is maintained through relay K102 contacts. Plate voltage is applied to print amplifiers V5 and V6, and print power is fed to the trolley bar and stylus needles.

At the end of the transmission the 450-cycle stop signal is detected by CR104 and fed to FL102 (now tuned to 450 cycles). The output from FL102 is amplified by V101A, rectified by CR103, and fed to start/stop relay amplifier V102B energizing start/stop relay K101. The contacts of K101 are restored to the original positions setting up the recorder for the next transmission.

If the set is operated manually, auto/man switch S305A shorts the inputs to the start/stop and start-record signal filters. A positive potential is applied to start/stop relay amplifier V102B by S301D to operate start/stop relay K101. Holding current is applied to the phase actuator E501 through relay K101 and phase switch S303A. When the phasing signal is received, the operator activates phase switch S303, and S303A removes holding current from the phase actuator. Section S303B applies plate voltage to thyratron V204, and phasing pulses fire V204 as described previously. After three phasing pulses the operator releases S303 to reapply holding current to E501.

When copy signals start, the operator throws the start/record switch (S304B) momentarily to start-record. This places a positive potential on the grid of record relay amplifier V102A causing it to conduct, energizing relay K102.
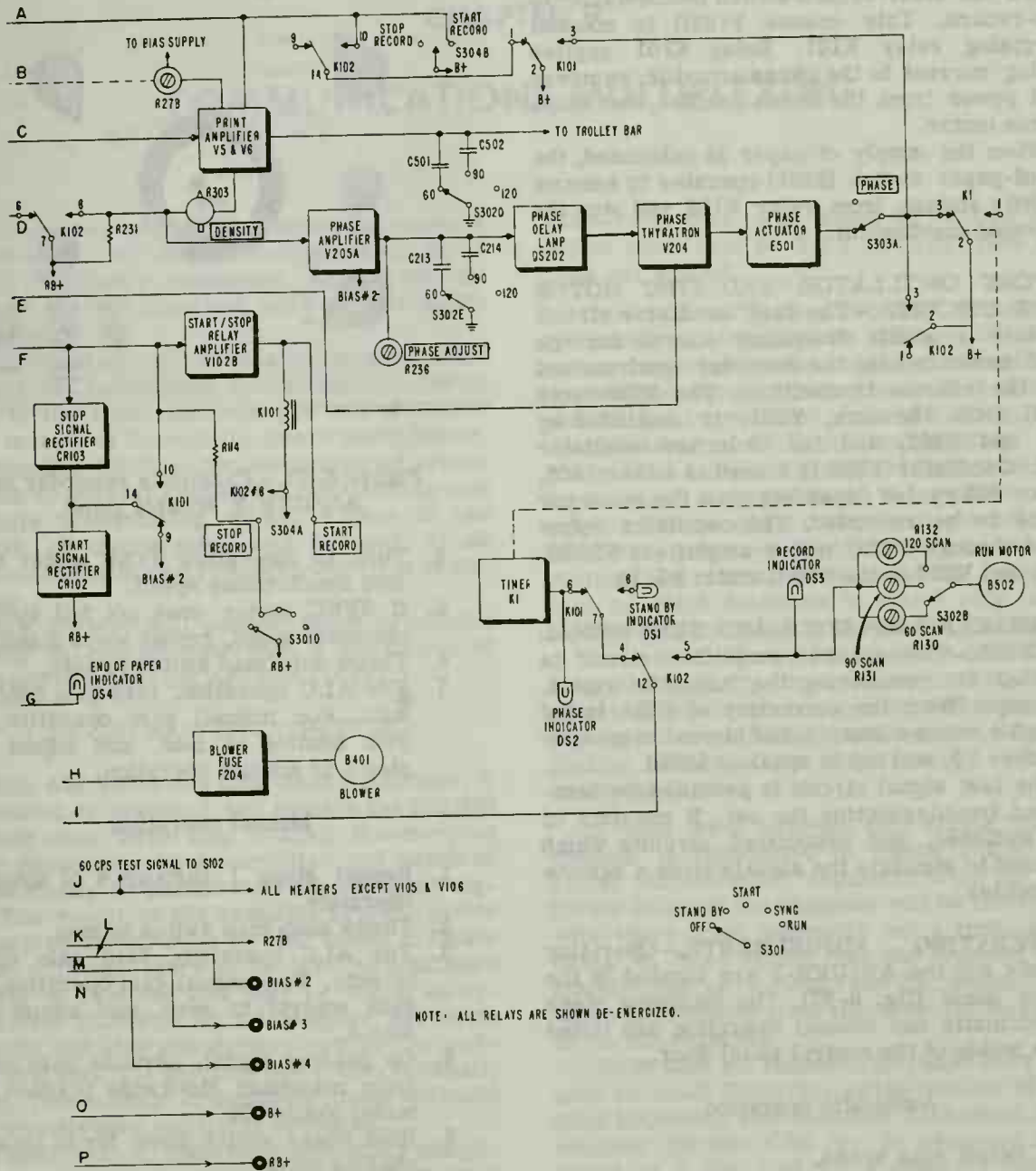
R302

SPKR VOLUME

SPEAKER AMPLIFIER V3

SPEAKER LS601

A

CONTRAST

B

R301 GAIN

SIGNAL INPUT

INPUT TRANSFORMER T301

ALC AMPLIFIER V1A

ALC AMPLIFIER V1B

R27A

PRINT DRIVER AMPLIFIER V4A

C

R30

ALC-OUTPUT

ALC LOCKOUT S305

AUTO   MAN   BIAS #3

12 K101

4   5

AUTO   MAN S305B

S303B

D

PHASE

E

SIGNAL AMPLIFIER V4B

HIGH PASS FILTER FL103

BIAS LIMITING RECTIFIER CR2

RC FILTER

ALC RECTIFIER CR1

ALC CATHODE FOLLOWER V2A

ALC FEED BACK AMPLIFIER V2B

CONTROL SIGNAL DETECTOR CR104

START/STOP SIGNAL FILTER FL102

START/STOP SIGNAL AMPLIFIER V101A

F

STEADY S104   START S103   RECORD S102   STOP S101

60 CPS FROM HEATER SUPPLY

START RECORD BAND PASS FILTER FL101

AUTO   MAN

S305A

TEST SIGNAL RECTIFIERS CR105, CR106

START/STOP TEST OSCILLATOR V103

START RECORD SIGNAL AMPLIFIER V101B

START RECORD SIGNAL RECTIFIER CR101

RC DELAY NETWORK

RECORD RELAY AMPLIFIER V102A

K102

SYNC MOTOR

3600 CPS FORK Y201

FORK AMPLIFIER V201

FORK FEEDBACK AMPLIFIER V202A

FORK SIGNAL AMPLIFIER V202B

LOCKED OSCILLATOR V203

R211

R217   120   S302C

R216   60

90

MOTOR R222

SYNC MOTOR DRIVER AMP V205B

SYNC MOTOR POWER AMPLIFIER V206 & V207

B501A

S301A

FORK SIGNAL LIMITER CR201

90

120

60

S302A

SCANS/MIN

B501B   START MOTOR

BB+

S501 END OF PAPER SWITCH

G

FORK R221

S301C

H

I

115-220 VAC POWER INPUT

LINE FILTER FL601

P/O S301

LINE FUSES F201 202

S401   HIGH

LOW

HIGH

LOW

POWER TRANSFORMER T401

BIAS SUPPLY

POWER ON INDICATOR DS201

J

BIAS VOLTAGE REGULATOR V107

L

K

M

N

UTILITY OUTLET FUSE FE01

LINE VOLTAGE

UTILITY OUTLET J601

B+ SUPPLY

VOLTAGE DIFFERENCE AMPLIFIER V104

RR+

VOLTAGE REGULATOR TUBES V105 & V106

O

P

296

70.90

Figure 8-36.—Facsimile recorder set AN/UXH-2, overall block diagram.

Relay K102 applies power to the run motor and sets up the set to record.

When the transmission ends, the operator throws the start-record switch momentarily to stop-record. This causes V102B to conduct energizing relay K101. Relay K101 applies holding current to the phase actuator, removes print power from the stylus needles, and stops the run motor.

When the supply of paper is exhausted, the end-of-paper switch (S501) operates to remove the RB+ voltage from relay K102 and stop the recording mechanism.

FORK OSCILLATOR AND SYNC MOTOR DRIVE CIRCUITS.—The fork oscillator circuit provides a stable frequency source for the SYNC motor to keep the recorder synchronized with the remote transmitter. The 3600-cycle signal from the fork, Y201, is amplified by V201 and V202, and fed to locked oscillator V203. Oscillator V203 is locked at either 1200, 900, or 600 cycles depending upon the scans per minute to be recorded. The oscillator output is fed through SYNC motor amplifiers V205B, V206, and V207 to the SYNC motor B501A.

SPEAKER AMPLIFIER AND TEST SIGNAL CIRCUITS.—The speaker amplifier circuit is provided for monitoring the facsimile signal. The output from the secondary of T301 is fed through a volume control (not shown) to speaker amplifier V3, and out to speaker LS601.

The test signal circuit is provided for testing and troubleshooting the set. It consists of four switches, and associated circuits which are used to simulate the signals from a remote transmitter.

OPERATING ADJUSTMENTS.—Operating controls for the AN/UXH-2 are located on the control panel (fig. 8-37). The following steps for automatic and manual operation are listed on the inside of the control panel door.

Automatic operation

1. Select scan speed.
2. Turn to standby and wait one minute.
3. Turn to SYNC and wait 15 seconds.
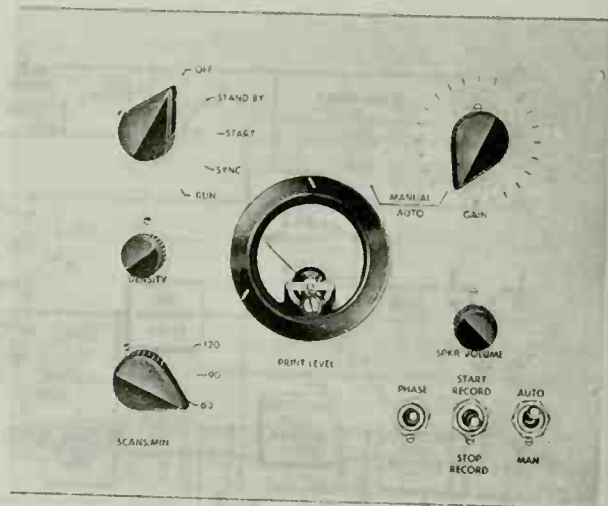
70.91

Figure 8-37.—Facsimile recorder set AN/UXH-2, control panel.

4. Turn to run—allow SYNC motor to fall into synchronous speed.
5. If SYNC motor does not fall into synchronous speed, repeat steps 3 and 4.
6. Throw auto/man switch to auto.
7. For ALC operation, turn gain control to auto. For manual gain operation, turn gain control to man. and adjust as in step 4 of manual operation.

Manual operation

1. Repeat steps 1 through 4 of automatic operation.
2. Throw auto/man switch to man.
3. For ALC operation, turn gain control to auto. For manual gain operation, turn gain control to man. and adjust as in step 4.
4. On phasing signal, advance gain control from maximum clockwise position until meter reads 100.
5. Hold phase switch down for at least two phasing pulses.
6. When copy starts, throw to start-record.
7. At end of copy, throw to stop-record.

CHAPTER 9

COMMUNICATIONS MULTIPLEXING

The number of communications networks in operation per unit of time throughout any given area is constantly increasing. As a result, all areas of the RF spectrum have become highly congested.

To a great extent, the maximum permissible number of intelligible transmissions taking place in the radio spectrum per unit of time is being increased through the use of multiplexing. Multiplexing involves the simultaneous transmission of several intelligible signals using only a single transmitting signal. Either of two methods of multiplexing may be used. These are time-division and frequency-division multiplexing.

This chapter will present basic principles relating to multiplexing with a discussion of telegraph terminal set AN/UCC-1C(V).

TIME-DIVISION MULTIPLEXING

With AM voice and tone communications, it is desired to transmit and receive the full 360 of each sine wave (fig. 9-1,A). However, an audio signal may be transmitted and received satisfactorily by periodically sampling the signal. The result of the sampling process yields a received signal such as that shown in figure 9-1,B. Although there is no limit to maximum number of samples that may be made, an approximate minimum of 2.5 samples per cycle of audio will give satisfactory results. In practical systems, 2.4 samples per cycle are usually taken. This concept of sampling forms the basis for time-division multiplex operation.

Figure 9-2, A, illustrates in a highly simplified form the basic principle of time-division multiplexing. Assume that a 3,000-cycle tone is applied to each of the six channels in the transmitter. Assume also that the rotating switch turns fast enough to sample, in turn, each of the six channels 2.4 times during each cycle of the 3,000-cycle tone. The speed of rotation of the switch must then be 2.4 x 3,000 or 7,200 rotations per second.

If the transmitter and receiver switches are synchronized, the signals will be fed in the proper sequence to the receiver channels. The transmitted samples from transmitter channel 1 will be fed to receiver channel 1. Thus, in the time-division method of multiplexing, many channels of audio are combined (with time spacing between components of the separate channels) to form a single output (multiplexed) chain. The chain is transmitted (via wire or radio facilities) to distant demultiplexing receivers, each of which functions to select only the information pertaining to its particular channel.

A mechanical system is shown here for the sake of simplicity. In actual operation, no mechanical switching arrangement will provide the high speed of switching used in practical multiplexing systems. For this reason, electronic switching is used.

A sine wave sampled four times for each channel is illustrated in figure 9-2,B. In an actual transmission, segments from the waveforms in all of the channels will be interspaced with these four segments on a time-sharing basis. In other words, only one segment can be transmitted at a time; and the segments are taken in sequence from the wave-forms existing in the six channels.

More than six channels (perhaps 24 or more) may be used. However, as the number of channels increases the sampling rate must be increased (28,800 CPS for 24 channels as opposed to 7,200 CPS for 6 channels using the above example). In addition the width of each segment must be proportionately reduced. Decreasing the pulse width will decrease the minimum required rise time and increase the required bandwidth even more because of the
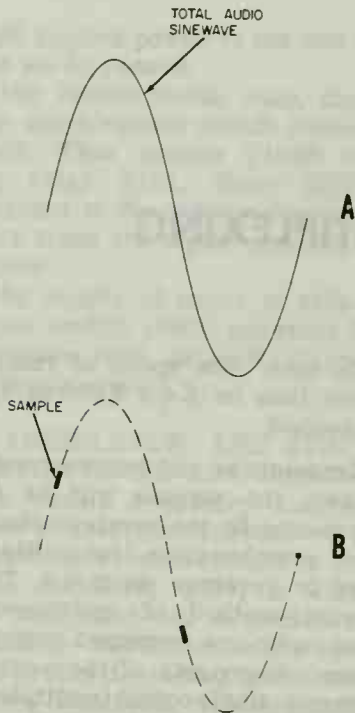
299

TOTAL AUDIO
SINE WAVE

A

SAMPLE

B

1.189

Figure 9-1.— Components of a sine wave

increase in the number of harmonics contained in the sharper leading edge of the pulse.

The bandwidth is also affected by the shape of the sampling pulse and the method of varying the pulse to carry the modulation. Commonly used methods of time division multiplexing include pulse amplitude modulation (PAM), pulse width or duration modulation (PWM or PDM), pulse position modulation (PPM), and pulse code modulation (PCM). Pulse amplitude modulation is the type illustrated in figure 9-2.

In time-division multiplexing, to use time in the most efficient manner, the bandwidth necessary to transmit n channels would be only slightly wider than n times that necessary for one channel. However to prevent crosstalk, etc., the minimum bandwidth is not generally utilized.

FREQUENCY-DIVISION MULTIPLEXING

Frequency-division multiplexing (fig. 9-3) is the older of the two methods of multiplexing. In this system, different subcarrier frequencies are modulated by the signals of different chan-

nels and then transmitted simultaneously over the same cable or on the same radio frequency carrier. On the receiving end the channels are separated by filters and are demodulated individually. The total bandwidth required for a frequency-division multiplexing system is the sum of the bandwidths of the individual channels, plus the sum of the necessary guard-band frequencies between channels.

The resultant signal in each channel will be approximately the same as that which would be produced by the use of a single independent channel, except that nonlinear elements in the system may cause mixing of the various carrier frequencies and crosstalk between channels. Frequency-division multiplexing is a satisfactory multiplexing method for transmission by cable (land line) because equipment is available which gives good fidelity, good signal-to-noise ratio, nearly constant gain, and very low nonlinear distortion.

A greatly simplified block diagram of a frequency-division multiplexing system is illustrated in figure 9-3. For simplicity it may be assumed that the signals are transmitted over a cable; nevertheless a radio carrier could be used.

Channels 2 through 6 each employ an oscillator whose frequency is heterodyned with the audio input to the respective channels in a balance modulator. The frequency produced in the channel as a result of the heterodyning represents an audio modulated subcarrier. A sharp cutoff filter is used in each channel and allows only the passage of the upper sideband. Channel 1 and the 5 upper sidebands are used to modulate the carrier output (transmitter end). There is no need for channel 1 to employ an oscillator; it feeds its audio input (between 300 and 3,000 cycles) directly to the transmitter, and fills the low end of the modulating spectrum.

The bandwidth and position (with respect to frequency) of each of the six channels are indicated at the bottom of the figure. The total spectrum that may be occupied in this particular example is 19.5KC.

Only the upper sideband (sum frequencies generated in the balanced modulator) is utilized in each channel. The suppressed carrier frequency is indicated by a dashed line between the various channels. This is the frequency used to generate the sideband indicated to the right-hand side of the line. For example, 3.3 KC is the oscillator frequency for channel 2.
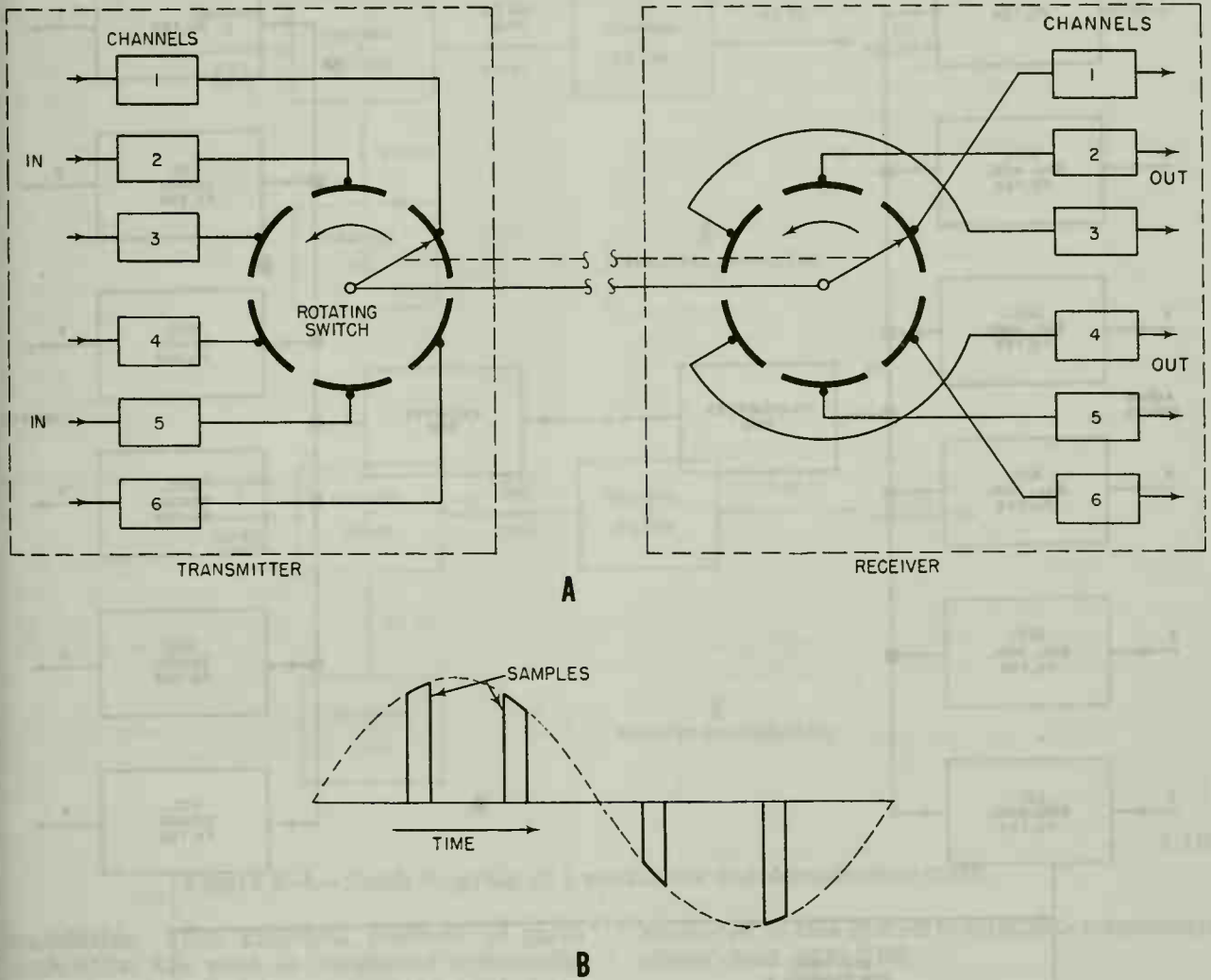
1.190

Figure 9-2.—Basic principle of time division multiplexing.

Although the upper sidebands are transmitted in this example, the lower sidebands could have been transmitted just as well.

The channel 2 circuits at the transmitter and the receiver are used as examples in the block diagram of figure 9-4. For simplicity, a 1-KC tone is assumed to be the channel input.

At the transmitter, the 1-KC signal is fed to the balanced modulator; a 3.3-KC signal is also fed to the balanced modulator from the channel 2 oscillator. The balanced modulator suppresses the carrier and generates upper (4.3 KC) and lower (2.3 KC) sidebands. Only the upper side-band is passed by the channel 2 filter.

The channel 2 demodulator in the receiver accepts the 4.3-KC signal and heterodynes this input with the channel 2 oscillator signal (3.3 KC). The carrier is suppressed, and the two sidebands (7.6KC and 1-KC are fed to the channel 2 demodulator filter. The filter passes the 1-KC signal to the reproducer.

TIME-DIVISION MULTIPLEX SYSTEMS (GENERAL)

Either the amplitude or the frequency of the transmitter signals (samples) from the respective channels may be varied to affect

301
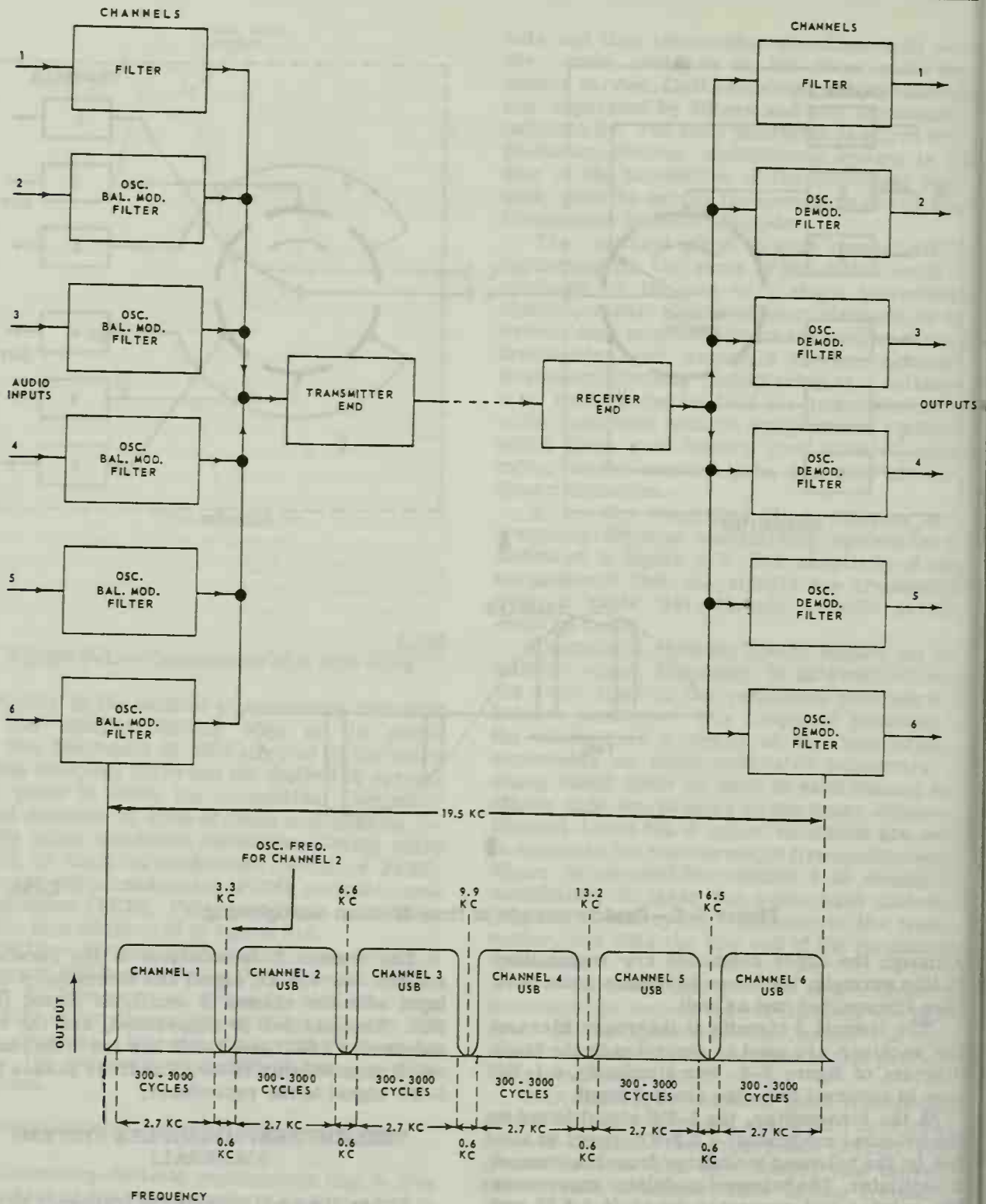
1.191

Figure 9-3.—Block diagram of a frequency division multiplexing system.
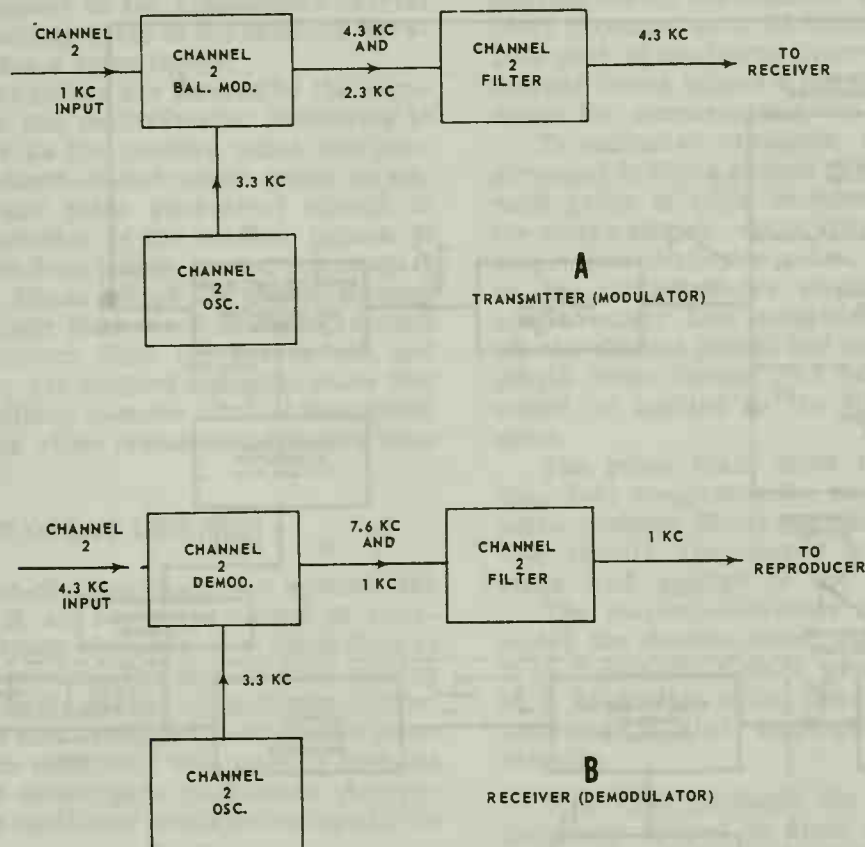
1.192

Figure 9-4.—Block diagram of a modulator and demodulator units.

modulation. Four practical methods of pulse modulation are used as mentioned previously. In each case, a modulating pluse is made to vary with respect to the amplitude of the audio signal.

The sampling process allows modulating pulses at a fixed rate to be introduced to the RF transmitter carrier, or to the wire cable equipments. No less than 2.4 of these samples must be transmitted for each cycle of the audio in order to produce an intelligible signal at the reproducer.

If the rate of sampling throughout the multiplex system is adjusted to 9.3 KC, then the maximum allowable audio frequency $= \dfrac{9300}{2.4} = 3875$ CPS. This frequency represents the highest audio frequency that may be used. In practice, for better fidelity translation it is desirable to provide more than the minimum number of samples per second. For this reason, filters would be employed in this system to eliminate frequencies above about 3000 CPS.

BLOCK DIAGRAM

Sampling the audio signal of several channels through the use of electronic switching can be accomplished in several ways. A block diagram analysis of one method used in time-division multiplex is shown in figure 9-5. In this system, it is first necessary to develop a voltage that varies linearly with time to establish the basis for time-division multiplex. A bootstrap type sawtooth (time base) generator, which uses a linearity stage to improve both voltage output and voltage linearity is used for this purpose.

The sawtooth generator output voltage (through the rise time) is applied to the grid of a normally cutoff sawtooth selector stage (V4). Conduction of this stage occurs when the
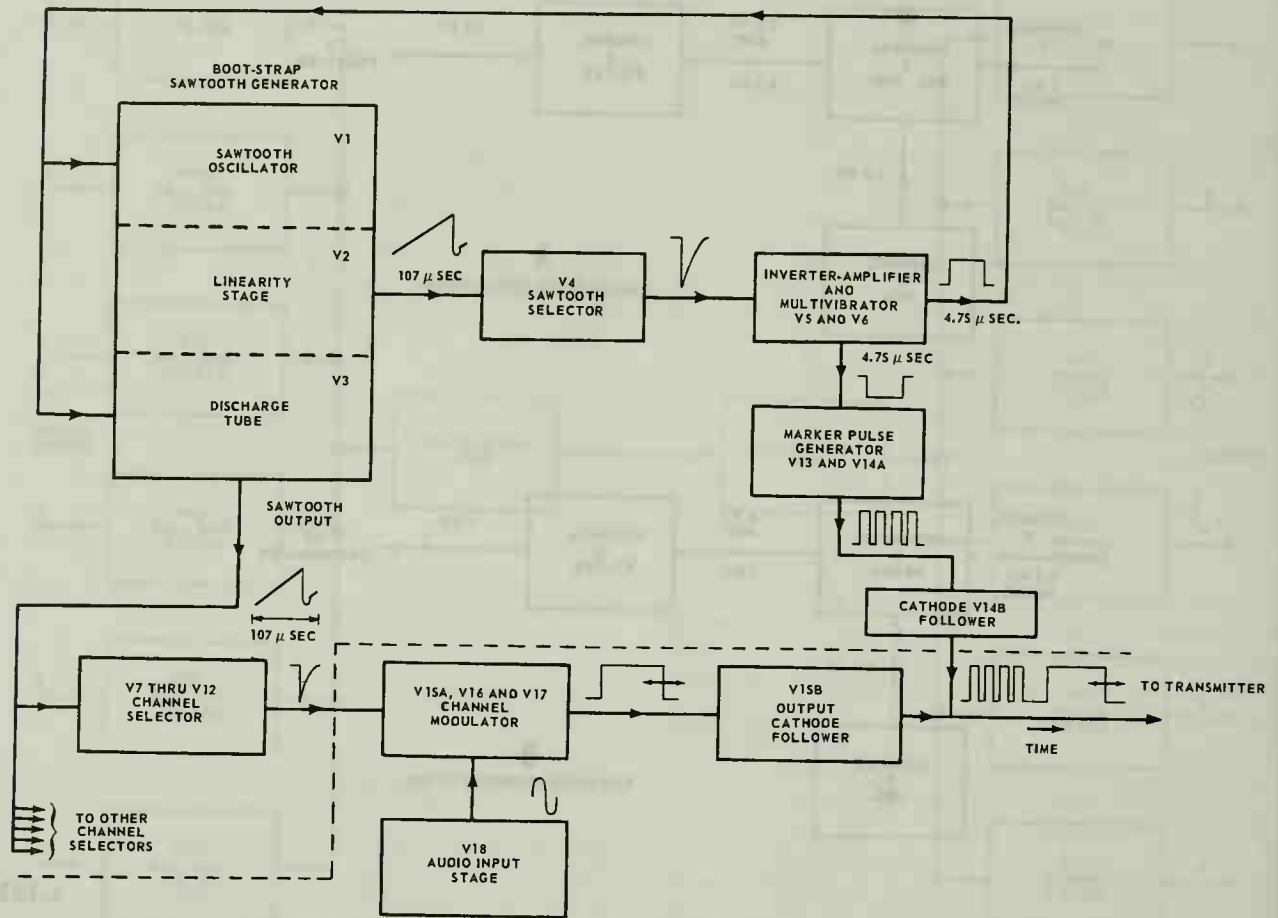
1.193

Figure 9-5.—Time-division multiplex block diagram.

sawtooth amplitude reaches a potential slightly less than the bias voltage on the selector. When the sawtooth voltage becomes sufficiently high to cause conduction of the selector, a trigger pulse is produced at the grid of a one-shot multivibrator (V5 and V6). The multivibrator output is returned along dual paths to the sawtooth generator osciallator and discharge tube to cause the sawtooth flyback. The amplitude of the sawtooth may be varied by manually varying the bias on the selector.

The sawtooth generator output waveform is fed along paralleled paths to the channel selectors (V7 through V12) one of which is contained in each channel of the multiplex system. Each individual selector is biased at a different level so that each is energized at a different time by the sawtooth voltage.

As each channel selector is energized, it feeds a pulse to its respective channel modulator, causing this circuit to produce a channel pulse. The channel pulse thus produced (for this channel alone) is then modulated in accordance with the type of pulse modulation used by the incoming audio.

Included in the transmitter output is a marker pulse(s) (sometimes called a sync group). The primary purpose of the marker pulses is to act as a synchronizing voltage between the receiving and transmitting multiplex terminals. These pulses may be of any form but are usually of a nature that makes them easily distinguishable from the modulation pulses at the receiving terminal of the multiplex system. The marker pulse (or pulses), along with the total modulation pulses from each channel

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(in time) presented to the transmitter carrier during one complete cycle of the sawtooth waveform, constitutes a video frame.

The marker pulses are formed by the negative gate from the multivibrator (occurring at the same time as the positive pulse that produces the sawtooth decay) which gates an oscillator (marker pulse generator) circuit to cause the generation of the marker pulses. In figure 9-5, four 5-μs pulses spaced 0.8 μs apart are produced. These pulses are passed through a cathode follower (impedance matching) circuit to the transmitter. Both marker pulses and channel pulses are coupled along the same line to the transmitter, because each of the pulses is added to the video frame at a different time interval.

FUNCTION OF DELAY LINE

In any time-division multiplex system, the development of an accurate method of energizing the various channels at a given time is of utmost importance. The system must employ some type of highly stabilized master oscillator. In the system just discussed, a triggered bootstrap sawtooth oscillator was used in both the multiplex and demultiplex equipments. A crystal-controlled oscillator would serve equally as well.

A block diagram of a time-division multiplex system using a crystal-controlled master oscillator and delay line is shown in figure 9-6. The oscillator generates an 8-KC sine wave, which is clipped by the pulse generator circuit to produce a square wave output. The oscillator frequency (8 KC) is the frame or base frequency of the multiplex system. This output is coupled along dual paths to a marker pulse generator, and to an artificial delay line.

The artificial delay line comprises 24 sections of pi-connected inductors and capacitors, each of which introduces a delay (5.2-μs) to the passage of the pulse along the line. A connection is made to a modulator channel from each section of the delay line. Thus, 23 channel modulators can be activated at time intervals of 5.2 μs (fig. 9-7,A) by the trigger pulses from the line. The presence of the controlling pulse enables the modulator to generate one channel pulse per frame (8,000 per second). The frequency of the audio signal applied to the modulator varies the position of the pulse (as indicated by the double-headed arrow through the waveform) and this comprises the signal

intelligence for the channel. Thus the 24 section delay produced up to 23 channels, and during the 24th part of the frame interval a double-pulsed marker (three pulses in latest equipment) is produced for synchronizing the multiplex system.

To minimize crosstalk, the modulators are arranged in three groups of eight. The output of each group of eight modulators is combined in the mixer shaper, which shapes each modulator output into a narrow pulse. The output of each of the mixer-shaper stages is applied to a single mixer line amplifier, which combines all the channel pulses and marker pulses into a single video frame. The mixer line amplifier output is applied to the RF transmitter circuits.

The pulse train input to the demodulator (fig. 9-8) comprises the marker, and up to 23-pulse position (time) modulated channel pulses. The signals are shaped in a clipper-shaper stage and applied to the marker-separator.

The marker-separator circuit extracts (detects) the double-pulsed marker and converts it to a suitable square waveform, which is fed to a 24-section delay line. This delay line is identical to that employed in the multiplex circuits.

The line produces the controlling pulses (properly spaced in time sequence) for activating the individual channel demodulators. The first output pulse is delayed 5.2 μs from the input pulse. Each of the remaining pulses is delayed successively 5.2 μs from the preceding pulse.

The first 23 pulses from the delay line form the control or deblocking pulses for the 23 demodulators. Each demodulator employs a coincidence stage (not shown), which requires the simultaneous application of the channel pulse and the delay line deblocking pulse to cause conduction. The entire video frame is applied to each demodulator. Since the deblocking pulses from the delay line are applied at specific intervals, only one demodulator will be activated for a single interval. The demodulated signals are filtered and amplified and reproduced at the output.

The time division multiplex system just discussed is widely used at Navy communication stations for radio line systems to remotely key transmitters or to modulate transmitters. Each channel could carry a set (3 KC) of telegraph terminal tones (discussed later) or 1 (3 KC) voice channel.
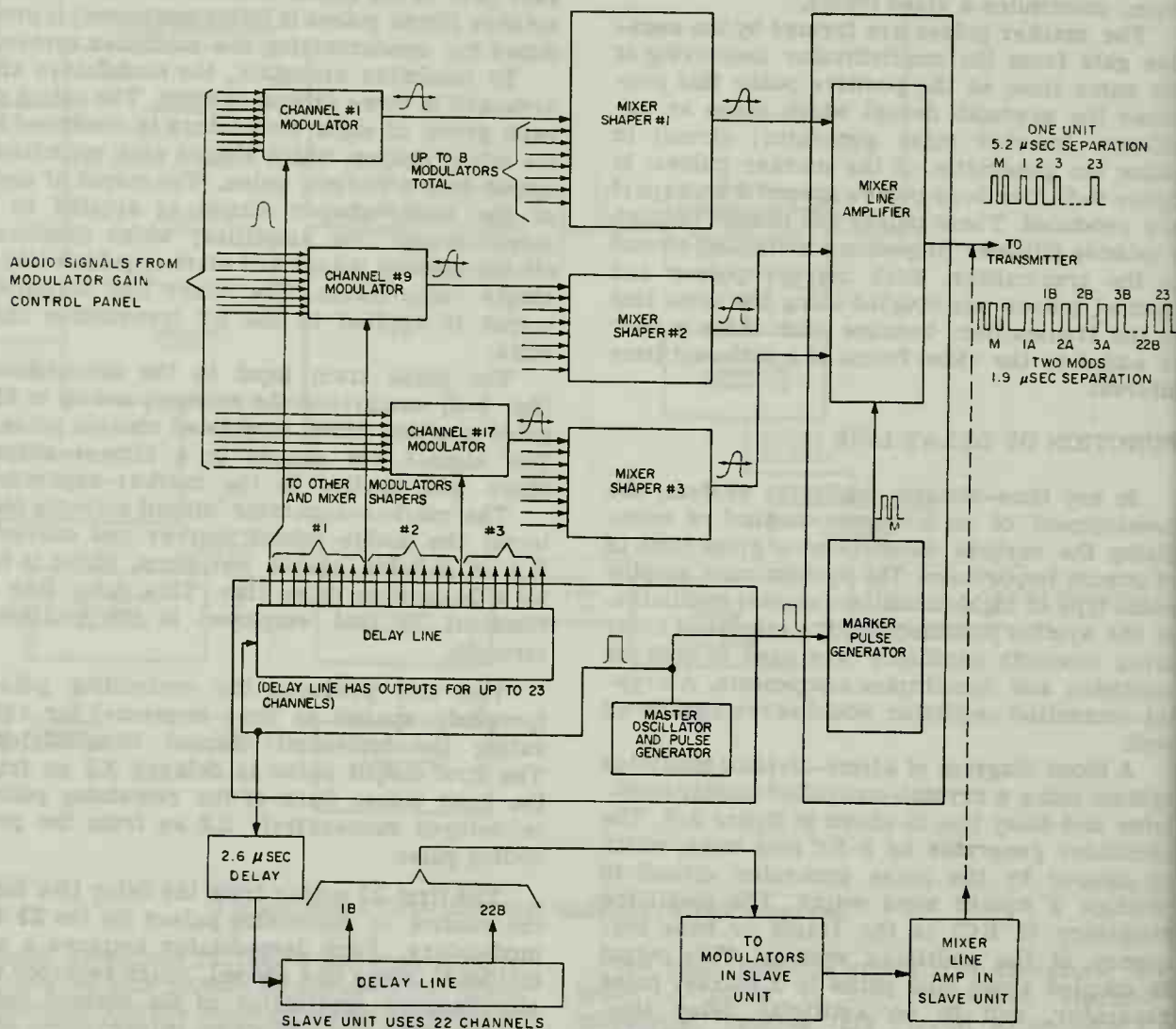
CHANNEL #1 MODULATOR

UP TO 8 MODULATORS TOTAL

AUDIO SIGNALS FROM MODULATOR GAIN CONTROL PANEL

CHANNEL #9 MODULATOR

CHANNEL #17 MODULATOR

TO OTHER MODULATORS AND MIXER SHAPERS

#1   #2   #3

MIXER SHAPER #1

MIXER SHAPER #2

MIXER SHAPER #3

MIXER LINE AMPLIFIER

ONE UNIT 5.2 μSEC SEPARATION

M 1 2 3 23

TO TRANSMITTER

1B 2B 3B 23

M 1A 2A 3A 22B

TWO MODS 1.9 μSEC SEPARATION

DELAY LINE

(DELAY LINE HAS OUTPUTS FOR UP TO 23 CHANNELS)

MASTER OSCILLATOR AND PULSE GENERATOR

MARKER PULSE GENERATOR

M

2.6 μSEC DELAY

1B   22B

DELAY LINE

SLAVE UNIT USES 22 CHANNELS

TO MODULATORS IN SLAVE UNIT

MIXER LINE AMP IN SLAVE UNIT

109.32

Figure 9-6.—Block diagram of multiplex modulation circuits.

If the circuits are needed, another modulator (demodulator) at the receiving end) may be slaved to the first modulator to give a 45 channel output. This may be 45 channels (300-3600 CPS) or 22 channels (300 - 7200 CPS). Figure 9-6 shows a simplified hookup of the slave unit modulator.

The pulse from the master oscillator (fig. 9-6) is delayed for 2.6 usec and applied to the slave delay line. This gives a combined pulse train spaced 2.6 usec apart (fig. 9-7,B). The marker channel and channel 23 are not used

so the output train contains 3 markers and 45 information pulses.

Figure 9-8 shows the demodulator system which requires a demultiplexer unit to separate the master and slave pulses and 2.6 delay line to synchronize the slave delay line gates.

TELEGRAPH TERMINAL AN/UCC-1C(V)

The AN/UCC-1C(V) (fig. 9-9) is a frequency division multiplex carrier-telegraph terminal
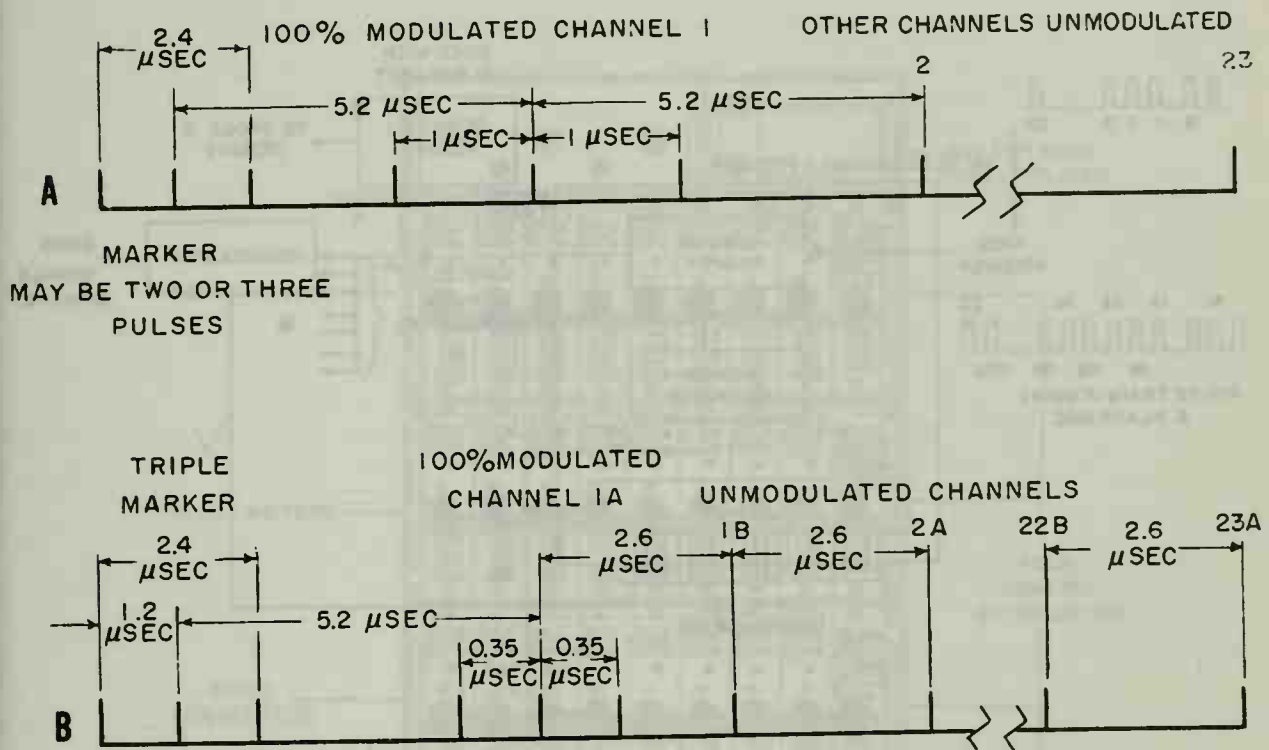
Figure 9-7.—Pulse relationship, 23 channel and 45 channel.

162.25

equipment for use with single sideband or double sideband radio circuits, voice frequency wire lines or microwave equipments. The equipment is of modular construction comprised of solid state components.

FUNCTIONAL DESCRIPTION

The terminal provides up to 16 different narrow-band voice-frequency tone channels, each passing a different band of frequencies. Associated with each channel are two oscillators and a gate circuit. One oscillator constantly generates the mark frequency; the other constantly generates the space frequency. The gate circuit is keyed by a telegraph loop causing one frequency or the other to pass (frequency-shift keying). The two frequencies are symmetrically disposed with respect to the center of the channel pass-band. The output from any set of different channels (fig. 9-10) can be combined on a single line for transmission over a single 3-kilocycle bandwidth communication link. In place of 16 narrow-band channels, the Telegraph Terminal can provide eight narrow-band and four wide-band channels. The wide-band channels use the band portion of the upper band narrow-band channels plus the band extending from 3025 to 3815 CPS. However, extension of the wide-band channels above 3265 (CPS) prevents the use of the multiplexing function.

The Telegraph Terminal is modular in construction so that the number of channels can be varied by increasing or decreasing the total number of modules.

Two 3-kilocycle bandwidth voice-frequency signals can be accommodated by the Telegraph Terminal by a multiplexing unit for transmission over a single 6-kilocycle communication link. The two 3-kilocycle signals may contain 16 telegraph tone channels in each 3-kilocycle band or 16 telegraph tone channels in one 3-kilocycle signal and speech in the other 3-kilocycle signal.

Diversity-combination switching is provided in the telegraph terminal at both the transmitting and receiving stations; thus, one telegraph signal can be used to key two different tone channels (Keyers) at the transmitting

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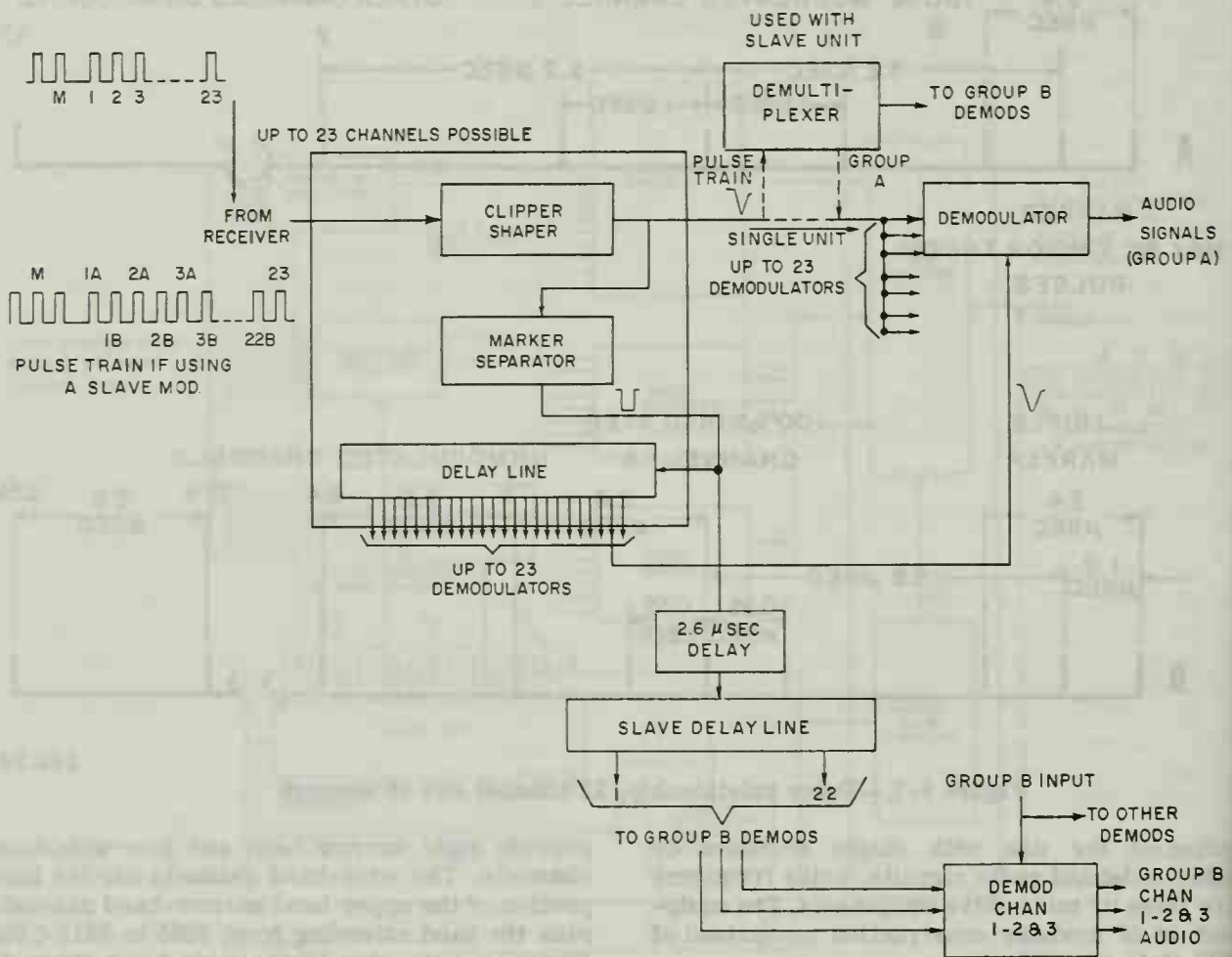
109.33

Figure 9-8.—Block diagram of multiplex demodulation circuits.

station. In this case, the corresponding tone channels (Converters) at the receiving station both develop keying signals. However, the best signal of the diversity pair operates the receiving telegraph loop, thereby preserving the telegraph signals when fading occurs in one of the two channels in the frequency-diversity pair. Diversity switching at the receiving station also permits space-diversity or RF diversity operation, in which tone signals obtained from two different radio receivers are used to operate two identical sets of tone channels. For this type of operation, each pair of identical tone channels (Converters) provides a single keying signal. Frequency-diversity operation can be combined with either space diversity or RF

diversity operation. In this case, one set of four tone channels produces a single keying signal.

Integrated circuits of two types, a differential amplifier and a flip-flop, are used throughout Telegraph Terminal AN/UCC-1C(V). Although each integrated circuit is housed in a transistor enclosure, it performs a much more complex function than a single transistor. External electronic components such as resistors and capacitors are used with integrated circuits as they are with transistors, when the application of the device requires their use.

The integrated circuit consists of a number of transistors or diodes and other circuit elements formed on a single chip of N-type
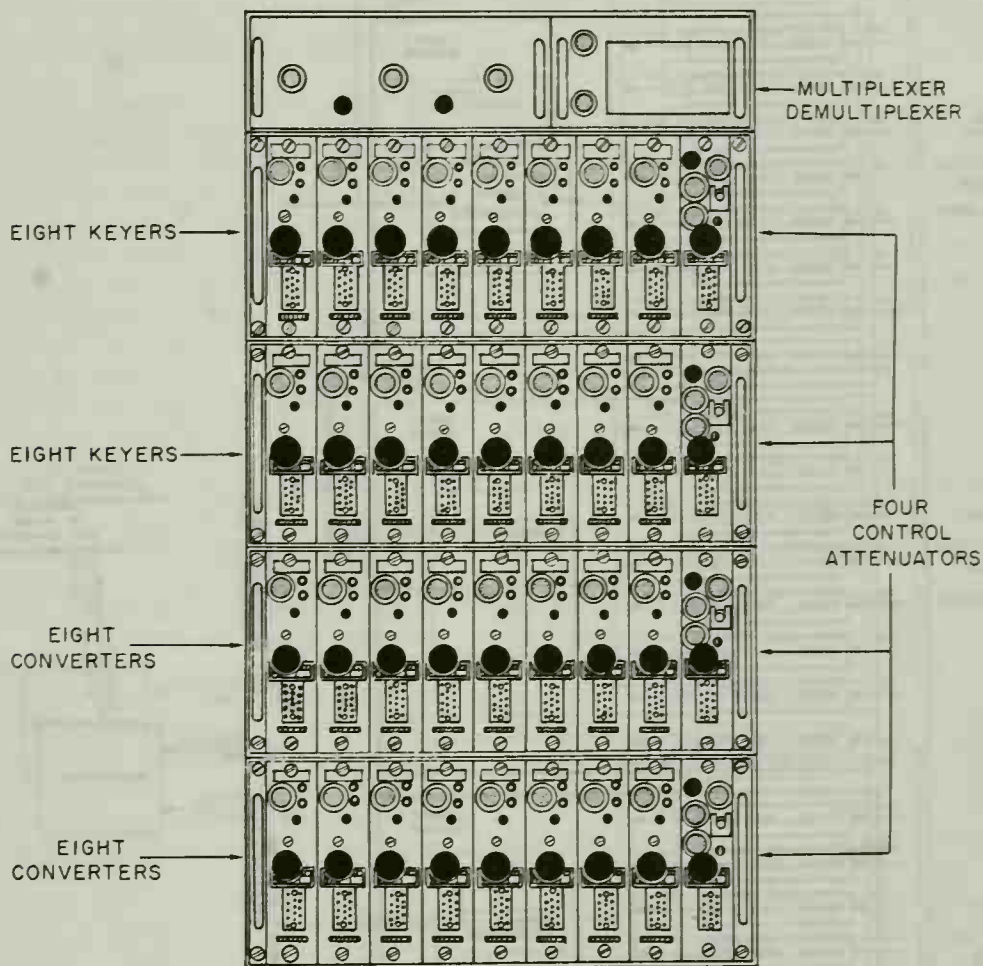
162.26

Figure 9-9.—Telegraph terminal AN/UCC-1C(V)

doped silicon crystal. The circuit elements are formed by semiconductor construction methods including masking, oxide photolithographic etching, diffusion, evaporation, epitaxial growth, and deposition. Then, lead wires are connected to the integrated circuit by thermal compression bonding and it is sealed in the enclosure or envelope.

In addition to the basic application as a differential amplifier, the integrated differential amplifier is used in a Schmitt trigger, a crystal-controlled oscillator, and a free running multivibrator. The adaption of this integrated circuit to these applications was undertaken in the design of Telegraph Terminal AN/UCC-1C(V) to improve reliability and to reduce the

number of spare parts required. External electronic components are connected into the circuit to adapt it to these applications.

Standard schematic symbols for integrated circuits have not been adopted as yet. The symbols used to represent the differential amplifier and the flip-flop have been approved for use in this application by the Naval Ship Systems Command, Technical Publications Section. These symbols were developed by placing the standard logic symbol for the circuit within a standard envelope or enclosure. All signal paths have been drawn through the envelope to the logic symbol. Other connections to the circuit such as voltage supplies, common connections, gain control, and operating bias are drawn only to the envelope.

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CABINET (MASTER)

CONVERTER (425 CPS)
CONVERTER (595 CPS)
CONVERTER (765 CPS)
CONVERTER (935 CPS)
CONVERTER (1105 CPS)
CONVERTER (1275 CPS)
CONVERTER (1445 CPS)
CONVERTER (1615 CPS)

CONTROL-ATTENUATOR

ELECTRONIC KEYING SIGNALS TO 8 DC TELEGRAPH LOOPS

CABINET (SLAVE)

CONVERTER (1785 CPS)
CONVERTER (1955 CPS)
CONVERTER (2125 CPS)
CONVERTER (2295 CPS)
CONVERTER (2465 CPS)
CONVERTER (2635 CPS)
CONVERTER (2805 CPS)
CONVERTER (2975 CPS)

CONTROL-ATTENUATOR

ELECTRONIC KEYING SIGNALS TO 8 DC TELEGRAPH LOOPS

SIGNAL INTERCONNECT

COMPOSITE VF SIGNAL (6 KC BANDWIDTH) FROM COMMUNICATION LINK

COMPOSITE SIGNAL

COMPOSITE VF SIGNAL (3 KC BANDWIDTH)

DEMULTIPLEXER

DIRECT PATH

TRANSLATED PATH

COMPOSITE VF SIGNAL (3 KC BANDWIDTH)

CABINET (MASTER)

CONVERTER (425 CPS)
CONVERTER (595 CPS)
CONVERTER (765 CPS)
CONVERTER (935 CPS)
CONVERTER (1105 CPS)
CONVERTER (1275 CPS)
CONVERTER (1445 CPS)
CONVERTER (1615 CPS)

CONTROL-ATTENUATOR

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CONVERTER (2125 CPS)
CONVERTER (2295 CPS)
CONVERTER (2465 CPS)
CONVERTER (2635 CPS)
CONVERTER (2805 CPS)
CONVERTER (2975 CPS)

CONTROL-ATTENUATOR

SIGNAL INTERCONNECT

ELECTRONIC KEYING SIGNALS TO 8 DC TELEGRAPH LOOPS

109.34

Figure 9-10.—Telegraph terminal AN/UCC-1C(V), overall functional block diagram.

Special test facilities required to align the Telegraph Terminal are incorporated in the portable Test Set.

KEYER

The DC telegraph loop is connected to the Schmitt trigger circuit (fig. 9-11) through the 20-MA/60-MA switching circuit and the adjustable trigger-level bias control circuit. The adjustable bias sets the trigger level at one-half the loop current value for netural operation and at zero current for polar operation. The input level applied to the Schmitt trigger is the sum of the trigger-level bias and the voltage drop produced by loop current through resistors (not shown), of the 20-MA/60-MA switching circuit. The change of input resistance for 20-MA or 60-MA loop operation assures that the voltage amplitude of the signals applied to the Schmitt trigger circuit is the same for both types of operation.

In netural operation, the input transistor section of the Schmitt trigger is turned on (conducting) when loop current flows and is turned off (not conducting) in the absence of loop current. In polar operation, the input transistor section is turned on during positive loop current and is turned off during negative loop current.

The output signal of the Schmitt trigger circuit and the output frequencies of the mark and space crystal-controlled oscillators are used to drive the diode gates. When the SIGNAL SENSE switch is in the NORMAL position, a loop current (mark input) causes the gate circuits to pass the output of the mark (higher frequency) oscillator. When the switch is in the REVERSE position, a loop current causes the gate circuits to pass the output of the space (lower frequency) oscillator.

The mark and space oscillators operate at frequencies which are 128 times the mark and space frequencies of the channel. They are controlled by crystals to maintain the frequency within a close tolerance. Both oscillators operate at all times but the output of each is passed through its gate only when the gate is enabled by the Schmitt trigger. The oscillator frequnecy passed through the gate is divided by 128 in the divider circuit. The divider is a seven-stage digital count down circuit, each stage is a flip-flop integrated circuit device. The output of the divider circuit (channel mark or space frequency) is applied through the

output amplifier to the output bandpass filter.

Keyer Diversity Combinations

The cabinet is wired in such a way that the Keyers in specific cabinet stations can be connected in frequency-diversity pairs. These pairs of cabinet stations are: A1 and A2, A3 and A4, A5 and A6, and A7 and A8.

With the DIVERSITY switches of the Keyers in the ONE position, each Keyer is connected to a separate input telegraph loop. Turning the DIVERSITY switches of two Keyers in the stations which are wired to form a diversity pair, to the TWO position connects them to the same input telegraph loop. Both channels process the same input telegraph signals. Turning the DIVERSITY switches of the four Keyers in cabinet stations A1 through A4 or in A5 through A8 to FOUR, connects the four Keyers to the same input telegraph loop. The DIVERSITY switches of all Keyers in a diversity combination must be in the proper position.

The signals from the output bandpass filter are applied to the TONE OUTPUT switch. When the switch is set to PARALLEL, it applies the channel output through the Control-Attenuator to the composite tone output line. When the switch is set to INDIV, it applies the channel output directly for individual tone output.
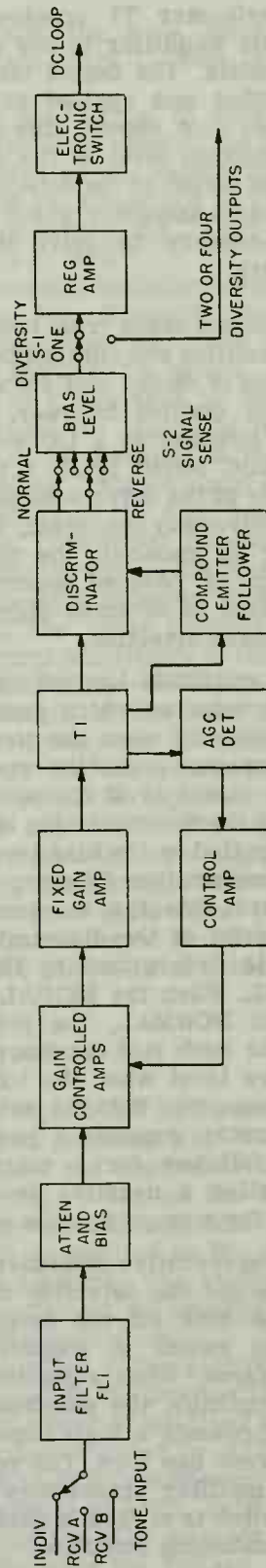
CONVERTER

Either a composite tone signal or an individual tone signal is fed to the input filter, (fig. 9-12), a bandpass filter which blocks all channel frequencies except those of the Converter's channel. The filter also contains the adjustable delay network that provides the means for equalizing delays for diversity operation. The output of the filter is applied to the attenuator and bias network. This network provides the AGC attenuation and also biases the input gain-controlled amplifier. The output of the gain-controlled amplifier is coupled to the input of the second gain-controlled amplifier generator in the integrated circuit device. The output of the second gain-controlled amplifier is applied to the input of a fixed gain amplifier whose output is coupled to the limiter stage which clips the signal removing any amplitude-modulated components from the waveform applied to the discriminator.
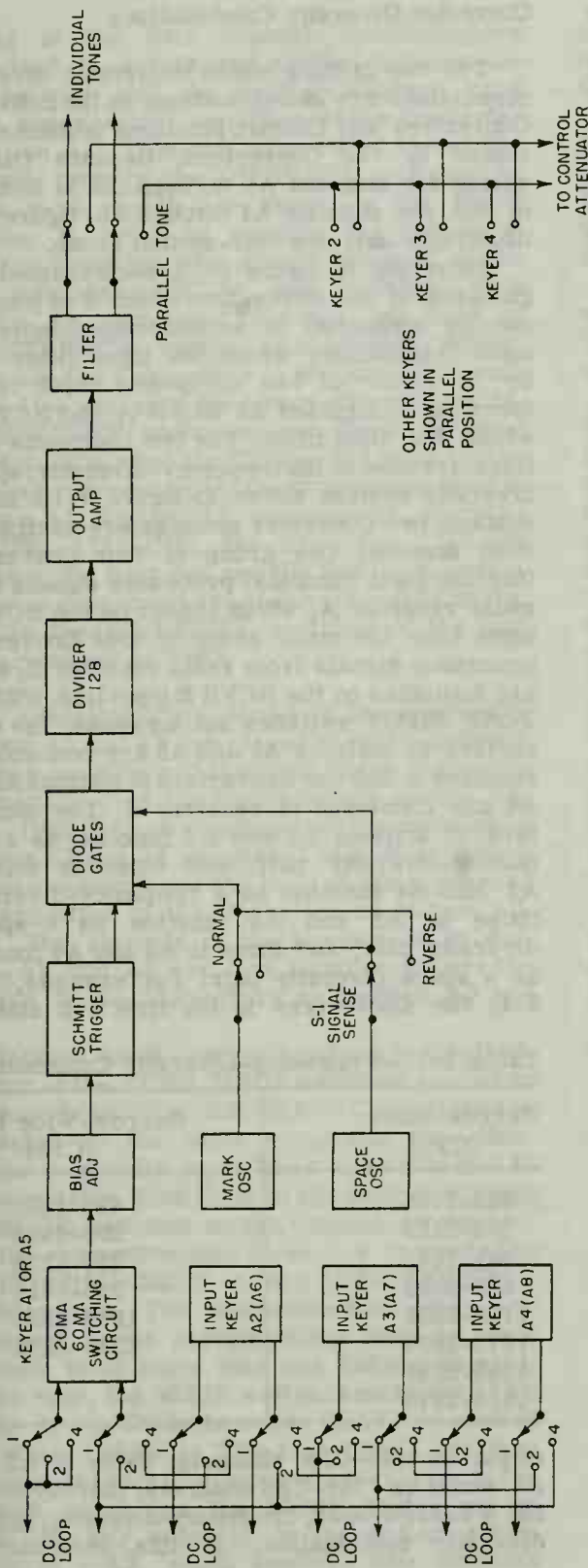
162.2

Figure 9-11.—Block diagram, frequency shift keyer.

162.28

Figure 9-12.—Block diagram, frequency shift converter.

Transformer T1 couples the output of the fixed gain amplifier to the gain and level control circuits. The output taken at one terminal is rectified and applied to the AGC detector transistor (not shown), this voltage establishes the conduction level of the AGC detector. The amplified output of the detector is applied to the AGC control amplifier which provides the power gain necessary to drive the gain-controlled amplifiers.

The output taken from two other terminals of T1 is rectified and filtered by a rectifier-filter consisting of diodes (not shown), and applied to a compound emitter follower. The output of the emitter follower is a DC voltage proportional to the AGC output level. It is used to limit the amplitude of the drive voltage to the discriminator. In diversity operation, when the gain of the receiver is reduced, the discriminator drive waveform is reduced assuring that a linear combination of detected signals results in the mark space detection.

The amplitude-limited output is applied to the discriminator which generates a DC signal of one polarity when the tone signal is at the mark frequency and of the opposite polarity when the tone signal is at the space frequency. The output of the discriminator is added to the bias level supplied by the bias level network and fed through an emitter follower (not shown) to the output level detector, a regenerative amplifier. The polarity of the discriminator output connection is determined by the SIGNAL SENSE switch S2. When the SIGNAL SENSE switch S2 is set to NORMAL, the polarity of the connection is such that the discriminator supplies a negative level when the tone signal is at the mark frequency. With S2 set to REVERSE, the discriminator supplies a positive level to the emitter follower for a mark frequency tone and supplies a negative level to the emitter follower for a space frequency tone.

The regenerative amplifier provides a stable reference for the detection threshold. A negative input cuts off the amplifier holding the electronic switch in conduction so that loop current flows. When a positive input is applied to the amplifier, the electronic switch is held off and presents a high impedance so that no loop current can flow. The regenerative action of the amplifier transistors assures that the output switch is always in either a nonconducting or full conducting condition.

Converter Diversity Combinations

The two factors which determine diversity connections are switch settings on the individual Converters and Cabinet positions (stations) occupied by the Converters. Because Cabinet wiring for stations A1 through A4 is identical to that for stations A4 through A8, figure 9-13 illustrates only one four-station group.

Referring to figure 9-13, each channel, regardless of the station into which it is plugged, can be connected to an associated individual input transformer within the input filter FL1, or to either of two composite input transformers (designated as RCVR A and RCVR B) within the input filter. The two composite input lines are used in the frequency-diversity/space-diversity system shown in figure 9-13. In this system, two Converter modules are assigned to each channel. One group of four Converters (for the eight channels) processes signals from radio receiver A, which appear on the RCVR A input line: the other group of four Converters processes signals from radio receiver B, which are furnished on the RCVR B input line. With the TONE INPUT switches set as shown, the Converters in stations A1 and A3 are connected to receiver A and the Converters in stations A2 and A4 are connected to receiver B: The converters in stations A1 and A3 function as a frequency-diversity pair, and those in stations A2 and A4 function as a frequency-diversity; those in A1 and A2 function as a space-diversity pair, and those in A3 and A4 function as a space diversity pair. For example, (table 9-1) the Converters in the first two stations

Table 9-1.—Frequency-Diversity Combinations

Narrow-Band (CPS)	Narrow-Wide Band (CPS)
425-1785	425-1105
595-2125	595-1445
765-1955	765-1275
935-2295	935-1615
1105-2465	1955-2805
1275-2805	2380-3230
1445-2635	
1615-2975	

might be 425-CPS channels; those in A3 and A4 would be 1785-CPS channels, thereby forming a narrow-band, frequency-diversity/space-diversity combination. In this combination,

utputs of the four channel discriminators re added in series and applied between the ias network and the emitter-follower input) the regenerative amplifier of the Converter 1 station A1. This can be verified by tracing he path through the DIVERSITY switches on gure 9-13, which are shown in the FOUR osition. In order to combine the outputs of ll four channels, the circuit grounds of the our channels must be tied together. This unction is performed through one bank of he DIVERSITY switch and cabinet wiring.

To use the Converters in stations A1 through 4 as two separate frequency-diversity pairs, he DIVERSITY switch on each Converter must e set to TWO and the Converters must be re-rranged. For example, the Converters used in tations A1 through A4, respectively, might be 25-CPS, 1785-CPS, 595-CPS, and 2125-cps hannels, the first two and the last two function-ng as narrow-band frequency-diversity pairs.

With the DIVERSITY switches of the four Converters set to TWO:

1. Discriminator outputs of the modules in tations A1 and A2 are placed in series between he bias network and the input to the regenera-ive amplifier of the Converter in station A1.

2. Discriminator outputs of the modules in tations A3 and A4 are placed in series between he bias network and the input to the regenera-ive amplifier of the Converter in station A3.

3. The grounds of the Converters in sta-ions A1 and A2 are tied together as are the ;rounds of the Converters in stations A3 and 4.

Since no space-diversity pairs are involved, ll four of the TONE INPUT switches are set to he same position so that all four Converters are ;onnected to the same composite tone line.

The composite input lines are connected to he composite tone lines in the cabinet through he MODE switches on the Control-Attenuator, nd the composite-tone lines are connected to he INTERCONNECT A and INTERCONNECT 3 connectors. The composite-tone lines are ;onnected to the corresponding lines in other Cabinets when more than one Cabinet is used; n this case, the MODE switch is set to MASTER n one of the Cabinets and to SLAVE in each of he other Cabinets. The tone inputs are then received through the input transformers in the Cabinet which has the MODE switch set to MASTER. For example, in the system shown in figure 9-13, which employs four Cabinets

containing Converters, the input lines of the top Cabinet and of the third Cabinet from the top are used.

Each converter can be set to operate in-dividually as a receiving channel by turning the DIVERSITY switch on the module to ONE. Each converter can be connected to an indi-vidual input line by setting the TONE INPUT switch on the module to INDIV.

Figure 9-13 illustrates the intermediate AGC circuit connections for diversity operation. Notice that these connections are made through another bank of the DIVERSITY switch.

When a group of Converters is used in a diversity combination, the inputs to the AGC control amplifiers of all modules in the group are connected together. This common connec-tion is at the collector of the AGC detector of each module. The Converter receiving the strongest input signal produces the most nega-tive output from the AGC detector. This signal overrides those of the other Converters and drives all AGC control amplifiers in the group. The AGC control amplifiers of all Converters in the group are driven by the AGC signal of the Converter receiving the strongest input signal. This causes the discriminator drive signals of all Converters except the one with the strongest input signal to be attenuated with respect to the discriminator drive signal of the Converter receiving the strongest input signal. Thus, only the strongest input signal produces sufficient discriminator output to drive the re-generative amplifier (output level detector) and the output switch.

CONTROL-ATTENUATOR

When the Keyer TONE OUTPUT switches are in the PARALLEL position (fig. 9-11), the signals are combined and applied to the input transformer, T1 (fig. 9-14) of the Control-Attenuator through the MODE switch. The out-put signal from T1 is applied to the input at-tenuation and bias network. The attenuation of this network is sufficient to provide the op-erating range for the amplifiers.

The output is applied to the gain controlled amplifier and coupled to the input of the fixed-gain amplifier. The output of the fixed gain am-plifier is transformer coupled by A1T1 to the output amplifier. It is also transformer coupled, through other terminals of A1T1, rectified, and applied to the base of the AGC detector transis-tor (not shown). Because the voltage required
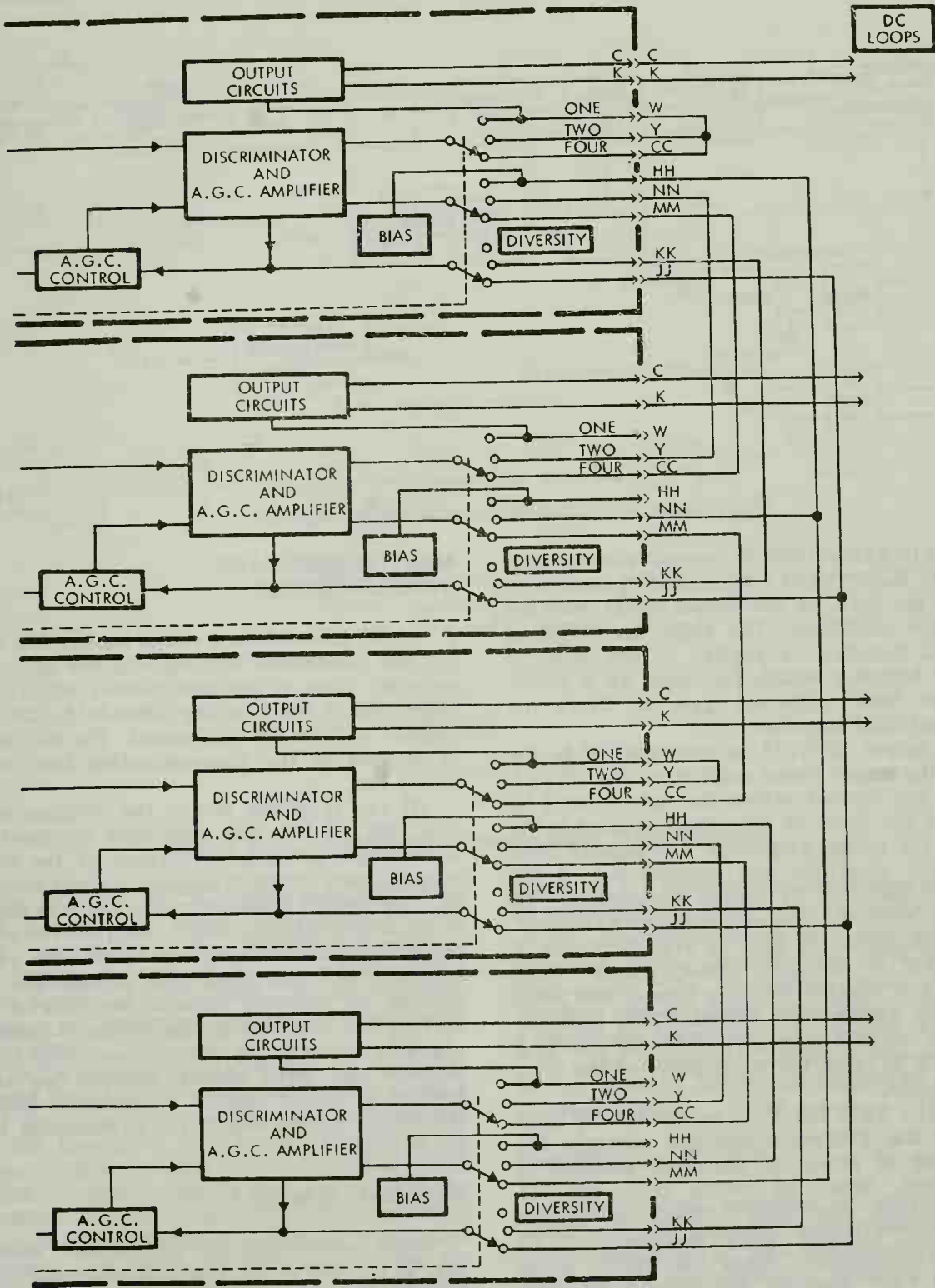
162.29
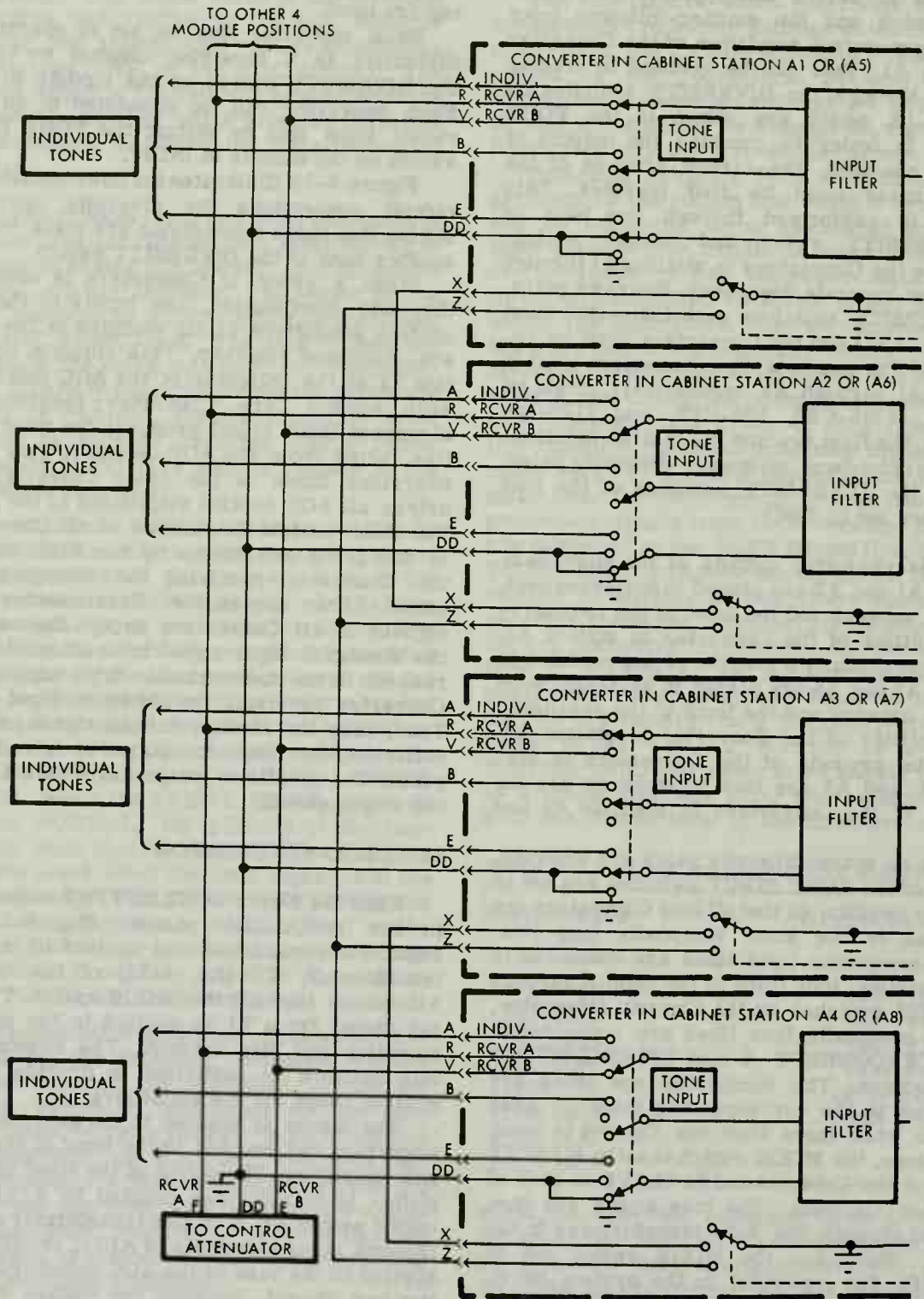
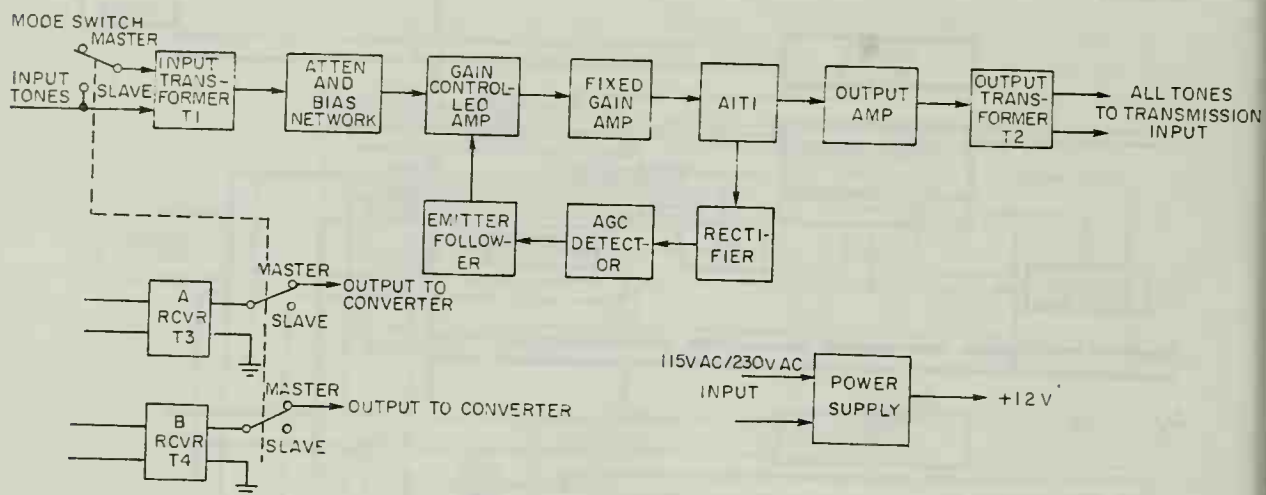Figure 9-13.—Converter diversity combination, block diagram.

Figure 9-14.—Control-attenuator block diagram.

162.30

to cause this transistor to conduct is essentially constant, it serves as a reference to determine (detect) the level of the output signal from the fixed-gain amplifier. The amplifier output of the AGC detector is applied to the base of emitter follower which functions as a power amplifier with sufficient gain to drive the gain-controlled amplifier.

The output of A1T1 is also applied to the base of the output power amplifier. The LEVEL control (not shown) varies the output level by adjusting the level of the signal applied to the base of the output amplifier. The signals from the power amplifier are coupled out of the Control-Attenuator by output transformer T2 which is located in the rear chassis assembly.

Transformers T3 and T4 are the RCVR A and RCVR B isolation transformers. When signals are applied to their inputs, and when the MODE switch is in the MASTER position, the input signals are applied to the RCVR A and RCVR B lines for application to the Converter inputs.

The 115 VAC/230 VAC switch on the front panel of the Control-Attenautor controls the application of power to the other modules in the Cabinet. When the switch is in the 115 VAC position, the input voltage is applied across the 115-volt input terminals of all power transformers; when the switch is in the 230 VAC position, the input voltage is applied across the full primary windings of all power transformers.

MULTIPLEXER AND DEMULTIPLEXER

At the transmission end of the link (fig. 9-10) the two composite 3KC signals are applied on separate lines to the multiplexer, which translates one of them to the 3265-5915-CPS band signals to form the 6KC signal. The 6KC signal is applied to the communication link on the single line.

At the receiving end of the communication link (fig. 9-10) the telegraph terminal reverses the process performed at the transmission end; voice-frequency signals from the 6KC bandwidth communication link are applied to the demultiplexer, which reverses the multiplexer frequency-translation action. The resultant two 3KC bandwidth tone signals are then applied on separate lines to two sets of receiving tone channels having the same bandpass characteristics as the channels used at the transmission end. Each channel accepts one voice-frequency signal which it converts into an electronic keying signal for a receiving telegraph loop.

Multiplexer Function

When performing the multiplexer function (fig. 9-15) one of the two composite-tone input signals to be multiplexed is applied to the DIRECT PATH jack, and the other to the TRANSLATED PATH jack (Mux-Demux inputs).

A MULTIPLEXING FUNCTION

B DE-MUTIPLEXING FUNCTION
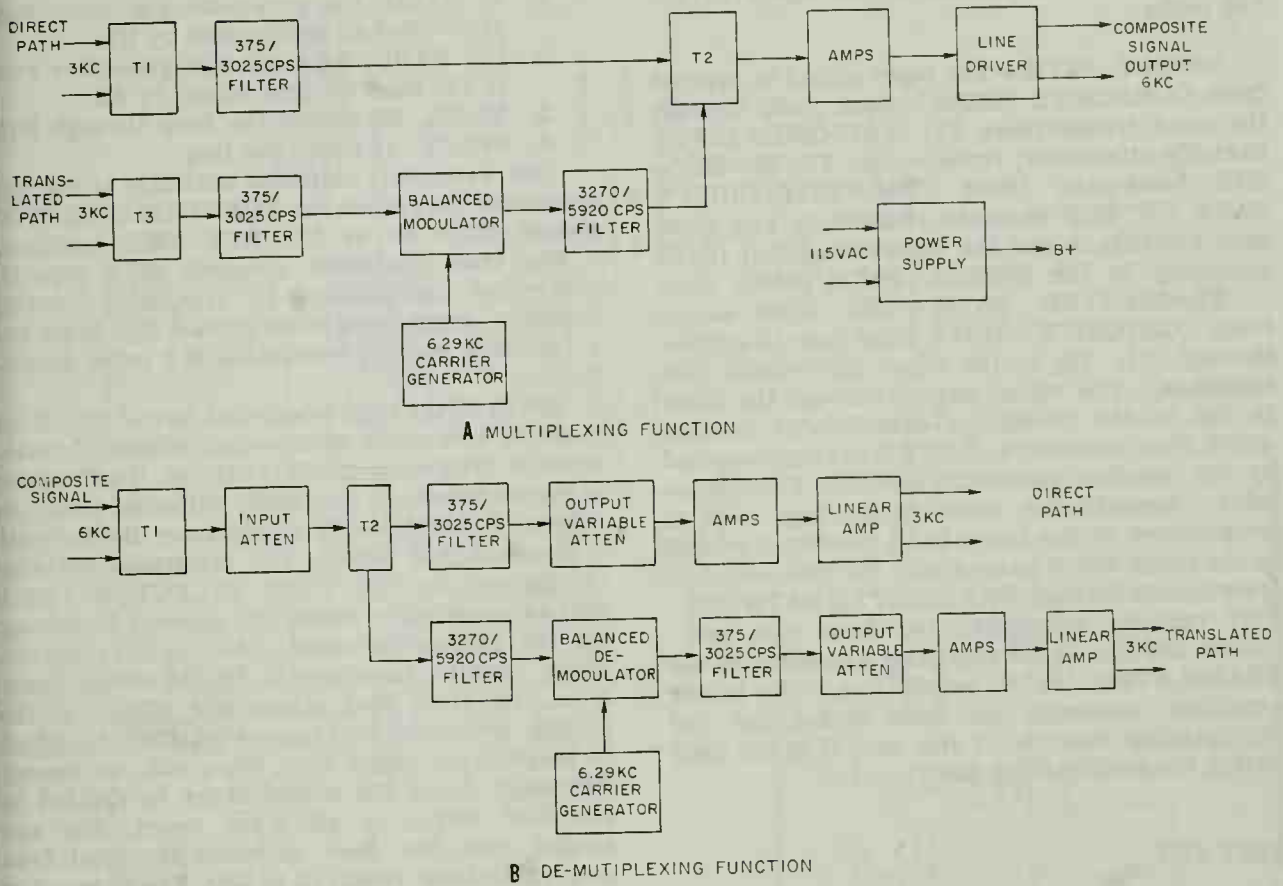
162.31

Figure 9-15.—Multiplexer demultiplexer block diagram.

The DIRECT PATH input passes through an input transformer which isolates the floating input circuit from the Mux-Demux ground. The output is applied to a bandpass filter which passes the 375-3025-CPS band of frequencies for summing with the translated path output signal.

The TRANSLATED PATH input is applied to an input transformer which isolates the floating input from the Mux-Demux ground and matches the low impedance of the input circuit to the higher input impedance of the bandpass filter. A filter removes signals that are outside the 375-3025-CPS pass-band, the output is fed to the modulator-demodulator section which also receives a 6.29KC carrier supplied by the carrier-generator assembly. The modulator-demodulator, a balanced lattice type modulator, suppresses both the input and the carrier frequencies, and passes the upper and lower side-bands. Of the modulator-demodulator output fed to the bandpass filter, only the lower sideband of the modulated signal is passed (the upper sideband falling outside the 3270-5920-CPS pass-band). The translated path signal is summed with the direct path signal in T2 to form a composite signal with a pass-band of 375-5920 CPS. The composite signal is then amplified to the desired level and applied to a transmission input.

Demultiplexer Function

Demultiplexer operation is essentially the reverse of the multiplexer operation described above. Because two separate outputs are furnished in the demultiplexing function, two variable attenuators and linear amplifiers are used; a third variable attenuator establishes the proper composite input signal level for

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application to the filters at the inputs of the two paths.

DIRECT PATH.—The input signal is applied from COMPOSITE SIGNAL input jack, through the input transformer T1, COMPOSITE INPUT variable attenuator, transformer T2, 375-3025-CPS band-pass filter COMPOSITE/DIRECT PATH OUTPUT variable attenuator, two fixed gain amplifiers and the linear amplifier (D/C) assembly to the DIRECT PATH output line.

TRANSLATED PATH.—THE input signal from COMPOSITE SIGNAL input jack is applied through T1, T2, to the filter (3270-5920-CPS bandpass). The signal passed through the filter is fed to the modulator-demodulator section which also receives a 6.29 KC carrier supplied by the carrier generator section. The modulator demodulator converts the high band of frequencies to the lower band, feeding its input to the filter which passes only the lower band of frequencies through TRANSLATED PATH OUTPUT variable attenuator, two fixed gain amplifiers and the linear amplifier assembly to the TRANSLATED PATH output line. The linear amplifier assembly is used during the demultiplexing function of the unit; it is not used during the multiplexing function.

TEST SET

The Test Set (fig. 9-16) consists of a reversals generator, a tone generator, an audio amplifier and loudspeaker, a meter, switches to control operation, and a power supply circuit. THE FUNCTION SELECTOR switch is the primary control of Test Set operation. It determines the connection and routing of input and output signals. Because, in various positions, it functions with the various circuits in the Test Set, these positions will be discussed with the circuits they affect.

The reversals generator (fig. 9-17) consists of a free running multivibrator driving a transistor switch output circuit. The active element of the multivibrator is an integrated circuit differential amplifier. The MODE SELECTOR switch S1 determines the mode of operation of the reversals generator, controlling the timing circuitry and providing loop closure to pins A (output) and P (ground) of the test connector P1. The positions of S1 and the functions of the reversals generator are as follows:

1. 75 BAUD, the reversals generator runs at 75 baud as determined by R2.
2. 150 BAUD, the reversals generator runs at 150 baud as determined by R1.
3. MARK, S1 closes the loop through R16.
4. SPACE, S1 opens the loop.

The reversals output is available at the test connector only when the FUNCTION SELECTOR switch (S2) is in the REV. OUT position.

The tone generator consists of a crystal-controlled oscillator, 12 frequency control crystals, eight integrated circuit flip-flops and a feedback circuit consisting of a pulse amplifier.

Seven of the eight integrated circuit flip-flops make up a variable seven-stage counter (countdown or frequency divider) circuit. The function of the eighth stage is slightly different from the others; it functions as a frequency divider with a symmetrical output. The frequency division is changed by the TONE SELECTOR switch (S4) as required to obtain the correct frequency for the channel selected. The variable seven-stage counter functions in the following manner. The first four stages are always in the circuit. Switching determines whether the output is taken from stage four, five, six, or seven. Feedback from the output stage is applied to previous stages to alter the count. For example, take the case in which the total frequency division required is 240. The symmetry
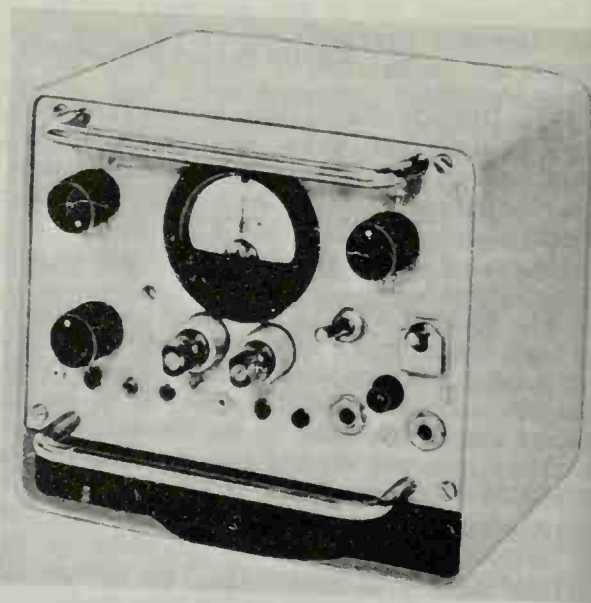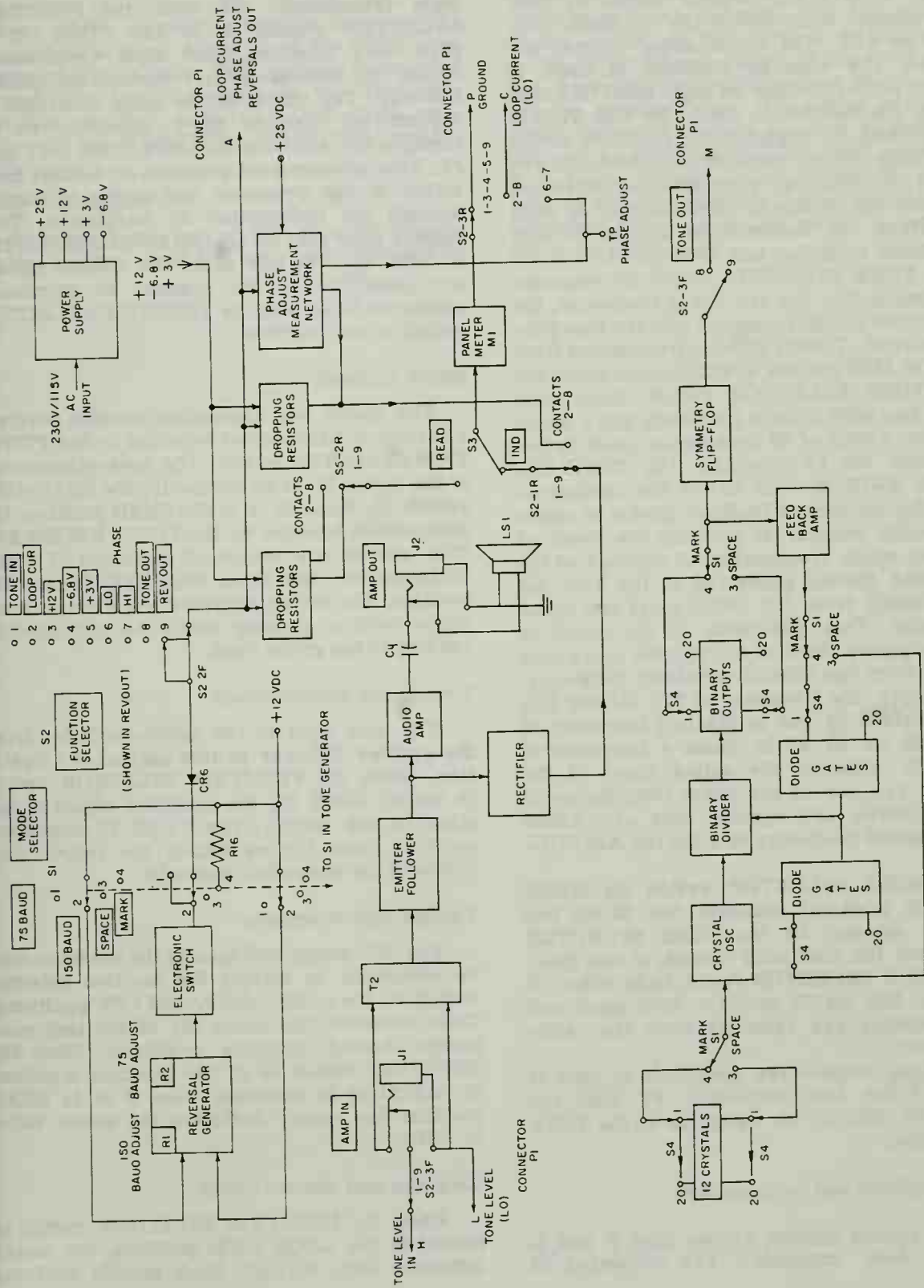

Figure 9-16.—Telegraph test set.

162.32

162.33

Figure 9-17.—Telegraph test set block diagram.

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flip-flop divides the counter output by two, so the counter must divide the oscillator frequency by 120. The seven-stage counter divides by 128 when no feedback is used. A feedback numeral count of eight must be added to cause the counter to divide by 120. This is accomplished by applying the feedback to the fourth stage. Other counting combinations are obtained by switching, which changes the output stage and the feedback combination. In each combination, the feedback is applied through the feedback amplifier and diode gates.

The TONE SELECTOR switch S4 connects two crystals (one for the space frequency, the other for the mark frequency) into the tone generator circuit. Twenty channel frequencies from 425 cps to 3230 cps are available upon command of the TONE SELECTOR switch. Since each channel has both a mark frequency and a space frequency, a total of 40 frequencies must be derived from the 12 crystals. The TONE SELECTOR SWITCH also alters the configuration of the network of feedback diodes to cause the flip-flop counter to produce the required mark and space frequencies. A number of the frequencies (tones) generated by the Test Set differ slightly from the actual mark and space frequencies. This difference is the result of using frequency division to generate more than one tone from one crystal oscillator frequency. For example, the frequency of 2Y2 (91,840 KC) can be divided by 144 to obtain a frequency of 637.8 CPS or by 80 to obtain a frequency of 1148 CPS. Although the output tones of the Test Set are not at the exact frequencies of channel marks and spaces, they are within the tolerances necessary to align the AN/UCC-1C (V).

The MODE SELECTOR switch (in MARK or SPACE position) connects one of the two crystals, selected by the TONE SELECTOR switch, into the oscillator circuit, at one time. If the MODE SELECTOR switch is in either 75 BAUD or 150 BAUD position both mark and space crystals are removed from the oscillator circuit.

The tone outputs are connected to pins M and P of the test connector P1 when the FUNCTION SELECTOR switch is in the TONE OUT position.

Audio Amplifier and Loudspeaker

Tone signals applied across pins N and L of the test connector are connected to

input transformer T2 when the FUNCTION SELECTOR switch is in the TONE IN or REV OUT position. The input transformer drives the audio amplifier through the emitter follower. The output of the audio amplifier is capacitively coupled by a capacitor to the loudspeaker and also to AMPLIFIER OUT jack J2. This arrangement provides an audible indication of the presence and keying of signals through the loudspeaker or earphones. Tone signals also may be applied to the AMPLIFIER In jack J1. In this case, the audible signals are available at the speaker or earphones connected to J2 with the FUNCTION SELECTOR switch in any position.

Meter Circuits

The meter and associated circuits perform a variety of functions as selected by the FUNCTION SELECTOR switch. The mode of operation of the meter M1 is determined by the READ-IND. switch S3. When S3 is in the READ position, the parameters selected by the FUNCTION SELECTOR switch are measured in terms of volts, milliamperes, and dbm. When S3 is in the IND. position, the meter provides a go/no-go indication with a passing range as indicated by reading in the green band.

Tone Level Measurement

The tone input to the audio-amplifier from the emitter follower is also applied to a rectifier. When the FUNCTION SELECTOR switch is set to TONE IN, the rectified signal is applied to the meter. The TONE IN measurement is used for equalizing the output tone levels of the individual channels.

Voltage Measurements

The DC supply voltages in the modules may be measured by turning the function selector switch to the +12V, -6.8V, and +3V positions. This connects the meter M1 of the test connector through dropping resistors. When the READ/IND switch is in IND position a go/no-go indication is obtained; when it is in READ position the meter indicates the actual value in volts.

Loop Current Measurement

When the FUNCTION SELECTOR switch is turned to the LOOP CUR. position, the meter indicates loop current. Each module contains

a resistor in series with the DC loop. The voltage which is proportional to loop current is applied to pins A and C of the TEST connector of the module. This voltage is coupled through the test connector of the Test Set, dropping resistor, and the LOOP CUR, contacts of the FUNCTION SELECTOR switch, to the meter.

Delay Adjustment

The phase angles of the channel signals applied to pin A of the test connector and to the PHASE ADJ test point are compared by a delay comparison network. The DC component of the output of this network is proportional to the difference in phase of the input signals and dependent in polarity upon which channel has the greater delay. The output of the delay comparison network is applied to meter M1 through the FUNCTION SELECTOR switch when it is set to PHASE LO or PHASE HI. The delays of the two channels may be equalized by adjusting the modules to obtain a zero indication on the meter.

CHAPTER 10

ELECTRONIC AIDS TO NAVIGATION, PART I

Electronic aids to navigation include electronic equipments such as radio direction finders, radio compass equipment, loran equipment, and tacan equipment. This chapter will discuss radio direction finder and loran equipments using the AN/URD-4 and AN/UPN-15 as representative equipments. An overal block diagram of the equipments will be presented along with circuit analyses of special circuits.

RADIO DIRECTION FINDER AN/URD-4

Essentially, the radio direction finder (RDF) is a sensitive receiver to which a directional antenna is connected. Radio direction finding is used primarily in air navigation. It is also used in the location of personnel afloat in liferafts or lifeboats, provided these craft are equipped with a radio transmitter, and that the RDF equipment is capable of receiving the transmitted frequencies.

Radio Direction Finder, AN/URD-4 (fig. 10-1) produces a visual indication of the bearing of a received signal. Fundamentally, this is accomplished by means of an indicator unit, which presents an oscilloscopic display of the output of a receiver.

Direction finding information is presented in the form of a propeller-shaped pattern on the face of the cathode-ray tube of the indicator. The pattern can be made to shift position to form a distorted figure 8 "sense bend" the apex of which indicates the relative bearing of the received signal. The method of obtaining the propeller-shaped pattern and sense bend will be treated later. Directional information is displayed with a bearing accuracy of plus or minus five degrees.

Provision is made on the indicator cabinet for headphones, which are used for aural reception simultaneously with D/F (direction finding) operation. Aural receiption is possible on AM or FM signals only. Direction finding information is displayed on the cathode-ray tube for any received signal whether modulated or unmodulated. The equipment is designed for either shipboard or shore installation. A functional block diagram of the shipboard system is illustrated in figure 10-2. Signal data converter CV-538/URD-4 is not required for shore installations.

ANTENNA

The antenna of the AN/URD-4 direction finding system is a two-dipole vertically polarized, Adcock array, which is rotated about the vertical axis by a motor at a constant speed. The antenna is most sensitive to signals in the vertical plane and least sensitive to signals in the horizontal plane. Accurate direction finding information is provided, however, from signals polarized as much as 45° from vertical.

Each time a dipole passes in a plane with a given signal, maximum voltage is generated in the antenna. Because there are two dipoles in the antenna array, the separate dipoles pass through the maximum signal path at different times to introduce a modulated component on the receiver RF signal. The modulated RF voltage is fed to the receiving portion of the equipment.

The rotating portion of the antenna is coupled to the stationary portion (fig. 10-2) through a rotating capacitance joint. The energy transfer from the antenna to the receiver is accomplished through this capacitance connection.

A two-phase alternator generates horizontal and vertical sweep voltages for application to the horizontal and vertical deflection plates of the cathode-ray tube. The phase of the two voltages are correct to form a circular pattern on the indicator. This pattern is present when no signal is being received.
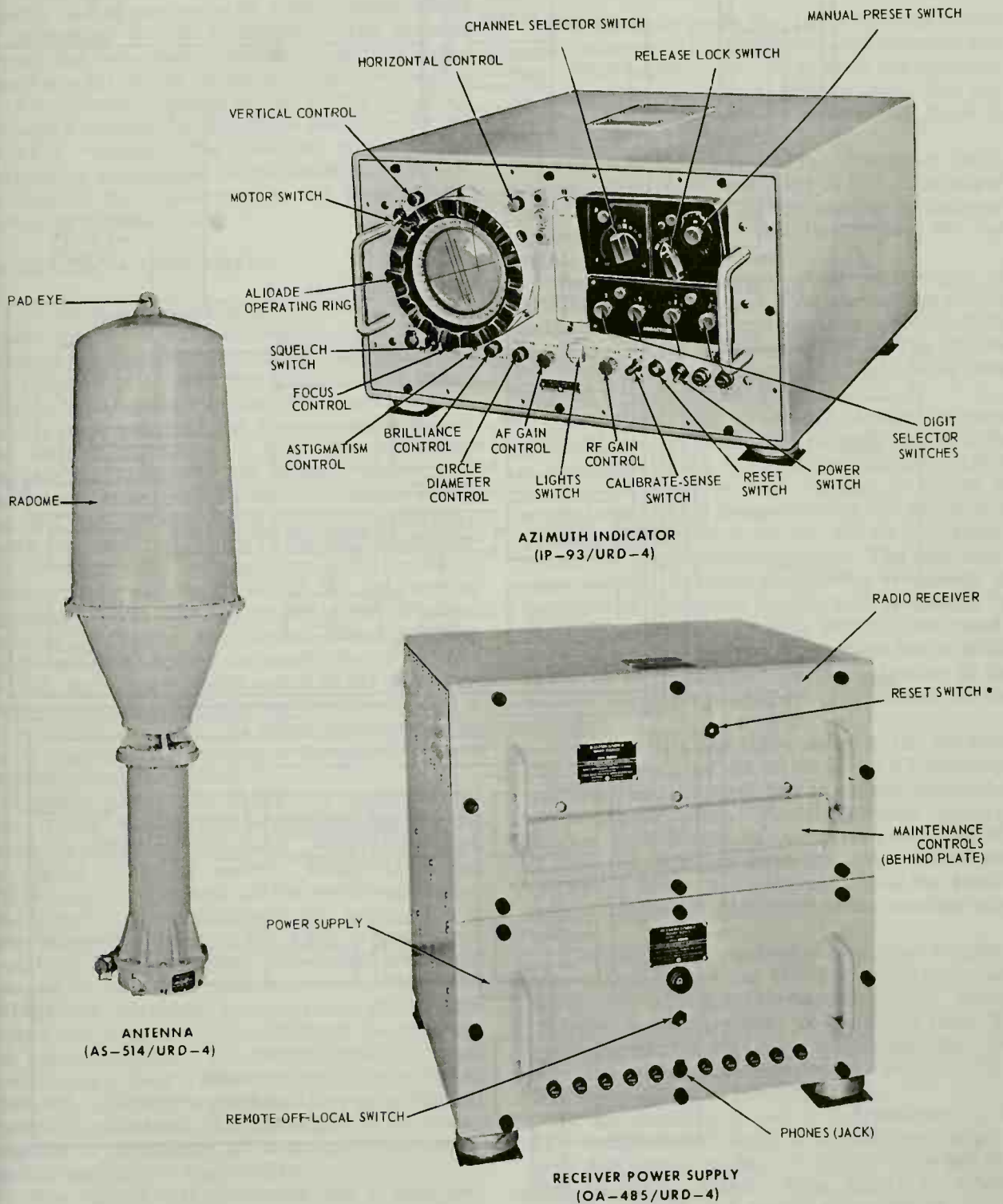
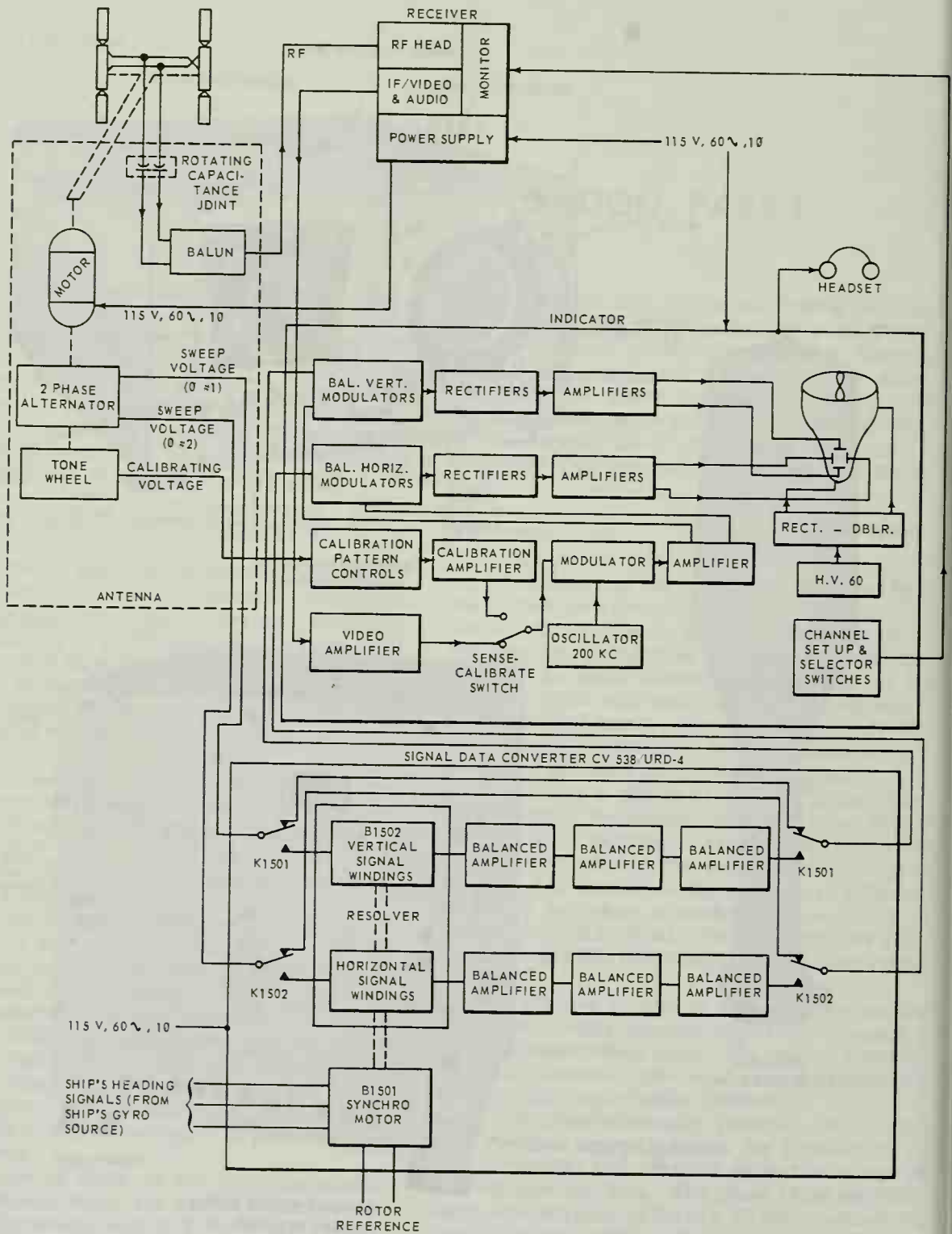Figure 10-1.—Direction finder set AN/URD-4, major components.

74.130.1

74.130.2

Figure 10-2.—Direction finder set AN/URD-4A, overall functional block diagram.

Calibration voltage for calibration of the indicator unit is generated by a tone-wheel, which is connected to the alternator. The antenna array, the two-phase alternator, and the tone-wheel are all driven by the antenna drive motor.

RF antenna voltage is fed to the receiver through a balance-to-unbalance device referred to as a "balun." The balanced output of the antenna is unbalanced by the balun to allow RF voltage to be transferred to the receiver by means of coaxial cable.

SIGNAL DATA CONVERTER

In shipboard installations of the direction finder set, the signal data converter is connected between the receiver and the azimuth indicator. Its purpose is to shift the phase of the sweep signals with respect to the video signal to produce a true geographical bearing indication at the azimuth indicator. If the video and sweep signals are applied to the azimuth indicator in the same phase relationship that exists at the antenna, the signal bearing indicated by the azimuth indicator is relative to the ship's bearing. To determine the true bearing of the received signal, it is necessary to add the ship's true heading. When the equipment is set for true bearing operation the signal data converter performs this addition automatically. Resolver B1502, positions synchro motor B1501 which receives ship's heading information from the ship's gyro circuits. When the signal data converter is in use, the true bearing of the signal is read directly from the indicator.

Relays K1501 and K1502 are controlled by the TRUE-STBY-REL switch located on the azimuth indicator front panel. When this switch is placed in its REL position, relays K1501 and K1502 are deenergized and the sweep signals are fed directly to the indicator for application to the deflection plates of the CRT to produce a relative signal bearing. When this switch is placed in its TRUE position, relays K1501 and K1502 are energized and the sweep signal currents are caused to pass through the windings of resolver B1502. The sweep signals are then fed through their respective amplifiers to the azimuth indicator to produce a true geographical bearing indication. Amplification of the sweep signals is necessary to compensate for losses in the windings of the resolver.

The signal data converter has a self-contained power supply for operation of the signal amplifier stages.

RECEIVER

The receiver group (fig. 10-2) includes the r-f head, monitor, IF video and audio chassis, and the power supply. The RF head of the receiver is essentially a tuned amplifier unit. The receiving frequency passed by the RF head is determined by the monitor circuit.

The monitor controls the frequency of a variable frequency oscillator (VFO). The output of the VFO is multiplied and used as the heterodyning frequency (in the first mixer) for the signal input to the RF head.

The monitor consists of several stages of heterodyning. It provides its controlling action to the VFO by operating a tuning capacitor (which is controlled by a tuning motor driven by the monitor) and by controlling a reactance tube shunted across the VFO.

Essentially, the frequency control system performs in the following manner: A sample of the output of the variable frequency oscillator in the RF head is beat against a series of crystal-controlled frequencies in the monitor to produce a frequency to be fed to a crystal-controlled phase discriminator. The discriminator output, in turn, causes the frequency of the VFO to approach the correct frequency by controlling either the tuning motor or the reactance tube. The frequency at which the monitor operates is determined by the selection of the various crystal-controlled frequencies in the monitor.

An isolation amplifier stage in the RF head (fig. 10-3) couples the output of the VFO through another isolation amplifier to a first mixer circuit in the monitor. The first mixer receives another input from a harmonic generator, which, in turn, is supplied from the first crystal oscillator. This oscillator produces the fundamental frequency generated in the monitor circuit.

The harmonic generator produces frequencies ranging from the 4th to the 13th harmonic of the oscillator fundamental frequency. Either of these harmonics may be selected from the generator circuit by tuning a transformer. The selected output is heterodyned with the VFO frequency output in the 1st mixer.

If the monitor is set to a frequency in the 225- to 299.9-MC range, a selector switch permits the output of the 1st mixer to be fed to a low bandpass filter. This signal is subsequently passed to a bandpass amplifier, and to a 2nd bandpass filter. When the monitor is

MONITOR
FIG 2-15
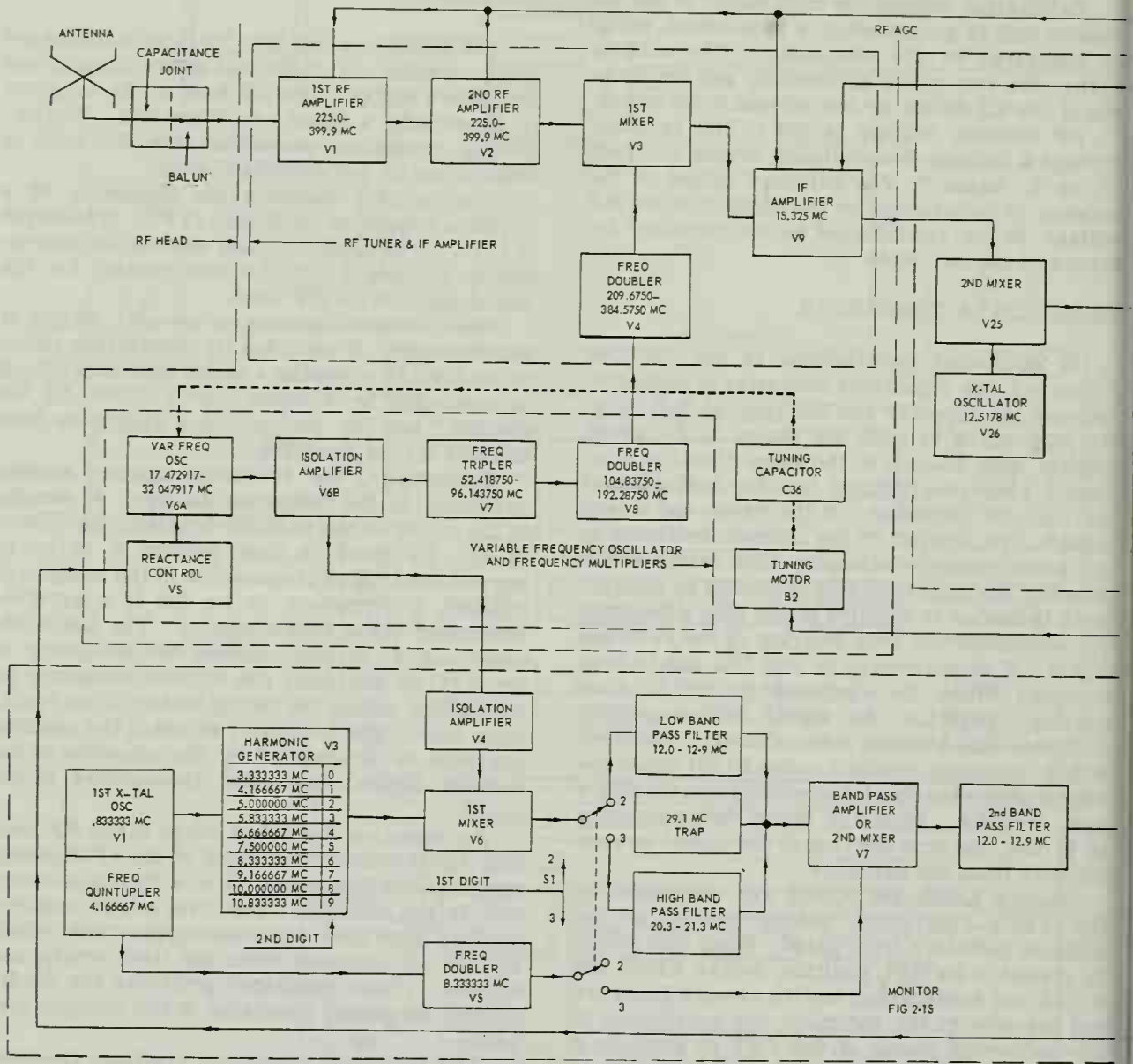
set to a frequency in the 300.0- to 399.9-MC range, the 1st mixer output is fed to a high bandpass filter and, in turn, is mixed with the tenth harmonic of the 1st crystal oscillator at the 2nd mixer. This action produces a beat frequency, which is fed to a 2nd bandpass filter. The bandpass amplifier or 2nd mixer stage functions as an amplifier for frequencies from 225 to 299.9 MC and as a mixer stage for frequencies from 300.0 to 399.9 MC. The tenth harmonic of the 1st crystal-oscillator fre-

quency is produced by the action of a frequency quintupler and frequency doubler, which are cascade connected between the oscillator and 2nd mixer circuits.

A 2nd crystal oscillator provides a choice of five frequencies. The 3rd mixer combines the output of the second bandpass filter with the selected output of the 2nd crystal oscillator to produce a beat frequency. This output is fed through the 3rd bandpass filter to the 4th mixer. The 3rd crystal oscillator provides a
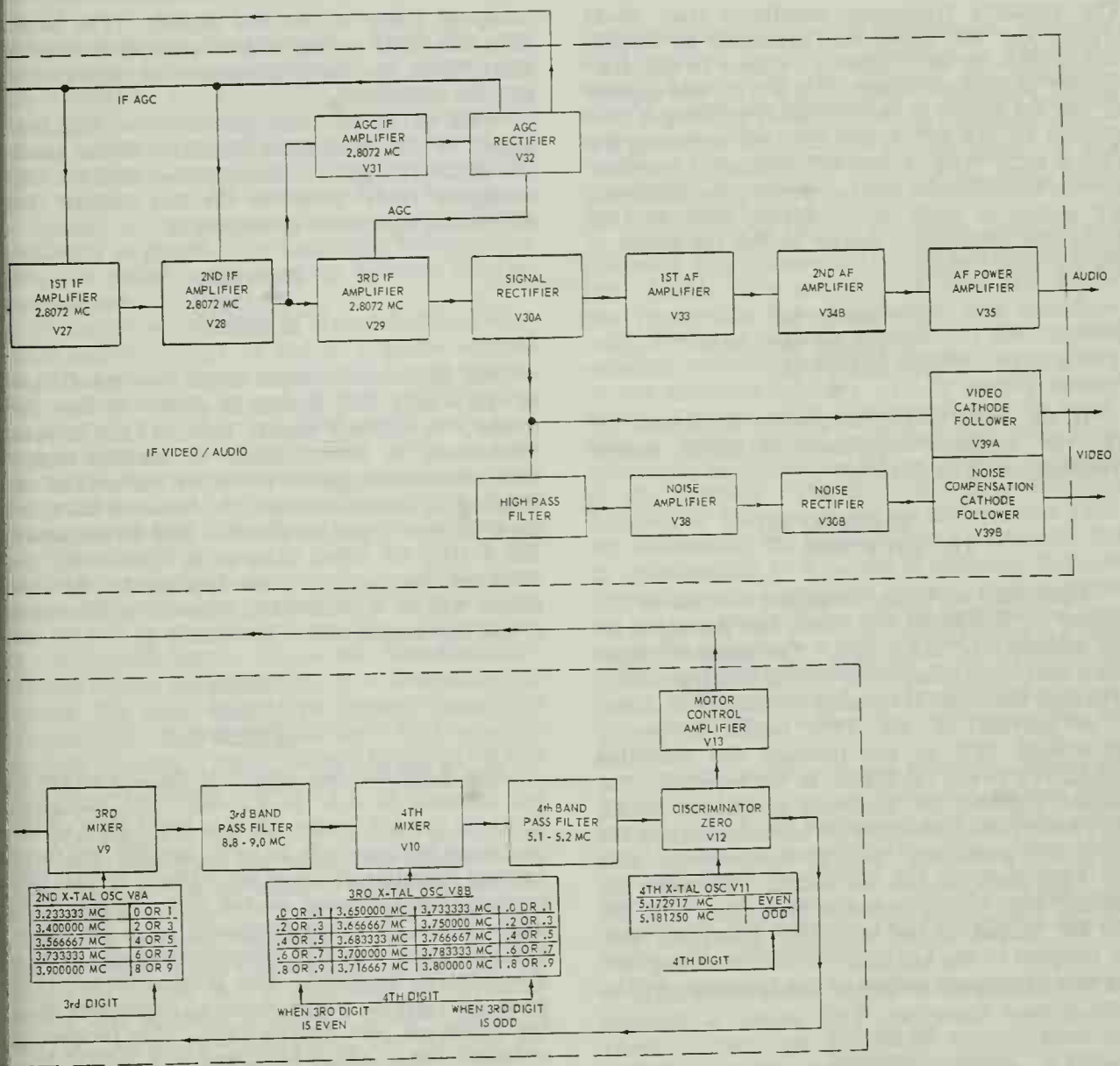
109.1

Figure 10-3A.—Block diagram of RF head, monitor, and video-audio chassis.

choice of 10 frequencies. Either of five of these frequencies may be selected when the chosen frequency has an even third digit, with either of the other five frequencies being selected when the third digit of the chosen frequency is an odd number. The selected crystal-controlled frequency is heterodyned in the 4th mixer stage with the input from the 3rd bandpass filter. The resulting output is passed through the 4th bandpass amplifier to the discriminator. The frequency applied to the discriminator will always be 5.172917 MC when the fourth digit is even, and 5.181250 MC when the fourth digit is an odd number.

In order to gain a better understanding of the monitor operation, consider the following example.

Assume that a frequency of 323.0 MC is selected by the channel selector switch (fig. 10-1) as the input frequency to the RF head.

The variable frequency oscillator (fig. 10-3) tracks the RF input, and produces an output, which will be multiplied 12 times in the frequency multiplier stages (3 x 2 x 2), and applied to the 1st mixer in the RF head, yielding a frequency 15.325 MC lower than the incoming RF (323.0 MC). The 15.325-MC difference frequency can therefore be subtracted from the incoming RF signal to yield the frequency being applied from the multiplier stages to the 1st mixer in the RF head. Thus,

$$\begin{array}{r} 323.000000 \text{ MC} \\ - 15.325000 \text{ MC} \\ \hline 307.675000 \text{ MC} \end{array}$$

If the 307.675-MC frequency is divided by 12, the variable frequency oscillator output frequency will be derived.

$$12\overline{)\,\dfrac{25.639583 \text{ MC}}{307.675000}}$$

Thus, the variable frequency oscillator operates at 25.639583 mc when the RF input (at the antenna) is 323.0 MC. The monitor output must therefore control the VFO at this frequency, provided the input frequency remains 323.0 mc.

A portion of the VFO output frequency (25.639583 MC) is fed through two isolation amplifiers to the 1st mixer in the monitor. The second digit of the incoming frequency (323.0 MC) determines which of the harmonics from the harmonic generator will be heterodyned with the VFO input in the 1st mixer. Because the second digit of the assumed frequency is 2, the 5.0 MC output of the harmonic generator will be coupled to the 1st mixer. The resulting difference frequency output of the 1st mixer will be

$$\begin{array}{r} 25.639583 \\ - 5.000000 \\ \hline 20.639583 \text{ MC} \end{array}$$

When the frequency selected by the channel selector (fig. 10-1) is in the range from 225.0 to 299.9 MC, the switch, S1, at the output of the 1st mixer in the monitor (fig. 10-3) will be in the position shown (up). For input frequencies from 300.0 to 399.9 MC, S1 is in the position opposite to that shown (down).

Since the chosen frequency is 323.0 MC, S1 at the 1st mixer output in the monitor is down (opposite to that shown), and the output frequency of the mixer is fed through the high bandpass filter to the 2nd mixer. (The bandpass amplifier or 2nd mixer operates as a mixer stage when the input frequency is above 300.0 MC, as discussed.)

With S1 in the down position, the 10th harmonic of the 1st crystal oscillator in the monitor (8.333333 MC) is heterodyned with the high bandpass filter input in the 2nd mixer. The difference frequency produced is

$$\begin{array}{r} 20.639583 \\ - 8.333333 \\ \hline 12.306250 \end{array}$$

The 2nd mixer output is fed through a 12.0- to 12.9-MC, 2nd bandpass filter to the 3rd mixer. A second input, from a 2nd crystal oscillator, is also applied to the 3rd mixer. The selected frequency from the 2nd crystal oscillator is dependent upon the value of the third digit in this input frequency. For a frequency, 323.0 MC, the third digit is 3. Likewise, the selected frequency to be applied to the 3rd mixer will be 3.400000 MC. Thus, the difference frequency output of the 3rd mixer is

$$\begin{array}{r} 12.306250 \\ - 3.400000 \\ \hline 8.906250 \text{ MC} \end{array}$$

The 8.906250-MC output of the 3rd mixer is fed through an 8.8- to 9.0-MC, 3rd bandpass filter to the 4th mixer. If this third digit of the incoming frequency is an even number, the third crystal oscillator provides either of the five frequencies indicated on the left. If the third digit of the incoming frequency is an odd number, either of the five frequencies indicated on the right is selected. The frequency selected from the left or right group is dependent upon the value of the fourth digit. The frequency selected for a 323.0-MC input is 3.733333 MC. Thus, the input frequency to the discriminator will be

$$\begin{array}{r} 8.906250 \\ - 3.733333 \\ \hline 5.172917 \text{ MC} \end{array}$$

Since the fourth digit of the input frequency (323.0 MC) is even, the 4th crystal oscillator also supplies an input of 5.172917 MC to the discriminator.

The discriminator is a phase-sensitive device, which develops an output voltage representing the resultant of any phase deviation

between the two frequencies. When the two frequencies are exactly 90° out of phase, the discriminator error voltage produced represents zero error. Any deviation in phase develops either a positive or negative discriminator voltage output, the polarity depending upon whether the 4th bandpass filter frequency is leading or lagging the 4th crystal-oscillator frequency in phase. The amplitude is dependent upon the phase angle existing between the two frequencies.

If the error voltage is large, the discriminator output voltage is fed to a motor control amplifier whose output controls a tuning motor. The tuning motor, in turn, drives a VFO tuning capacitor. The discriminator output and associated circuits cause the tuning motor to tune the VFO to that frequency which, when fed through the monitor unit, is approximately the same frequency as that of the 4th crystal oscillator.

When the frequency is close to that of the desired operating frequency, the reactance tube in the RF head takes control of the VFO to bring the circuit to the exact frequency. This action is described later. Since the discriminator voltage output tunes the VFO to a frequency that equals the sum of all the crystal-controlled frequencies, it follows that the VFO frequency may be changed a given amount by changing any one of the crystal-controlled frequencies the same amount.

The output of the VFO is coupled through an isolation amplifier and multiplier stages to the 1st mixer in the RF head. The 1st and 2nd RF amplifiers in the RF head amplify the input signal to a level sufficeint to provide proper heterodyning of the two signals. The 1st mixer output of 15.325 MC is amplified in the 1st IF stage and coupled through a coaxial cable to the IF, video-audo chassis. The RF and 1st IF amplifiers are AGC controlled.

A second frequency heterodyning process takes place at the input to the IF, video-audio chassis to produce a 2.8072-MC, IF frequency. The remaining circuits of the IF video-audio chassis are conventional receiving circuits.

INDICATOR

A block diagram of the indicator unit is shown in figure 10-4. The indicator provides the means by which the detected signal is amplified and mixed with the sweep voltages, which were originated at the antenna. The indicator also presents an oscilloscopic display of the

received direction finding information. Because the receiver must be installed a relatively close distance to the antenna, the indicator unit is normally situated at a remote location relative to the receiving and antenna equipments. The frequency of operation of the equipment is usually selected remotely from the front panel of the indicator.

Mounted just behind the front panel of the indicator is a 20-position rotary turret assembly (fig. 10-5). The memory cylinder fits inside the removable turret assembly and mechanically actuates various contacts of the channel selector switch. Arranged around the periphery of the memory cylinder are 20 evenly spaced detented bars, which are parallel to its longitudinal axis. Along each bar are four tabs, which can be positioned laterally along the length of the cylinder. The bars and their tabs are accessible through a slot in the cylinder case. By positioning the four tabs on a particular bar in relationship to a series of indexing digits on the cylinder, a particular channel frequency can be set up. Since there are 20 longitudinal bars along the cylinder, it follows that 20 of the total 1750 frequencies available within the tuning range of the equipment can be selected as channel operating frequencies by presetting the cylinder.

The alternator voltages from the antenna are fed to the indicator, as discussed. These voltages serve as the horizontal and vertical sweep input voltages, respectively. The phase of the vertical input is shifted 90° from that of the horizontal to produce a circular (Lissajous) pattern on the cathode-ray indicator in the absence of a video signal.

The input video signal is amplified and fed through a sense D/F calibrate switch (fig. 10-4) to the first modulator circuit when the switch is in the sense D/F position. A 200-kc carrier oscillator output is also fed to the first modulator, and is therefore modulated in this stage by the video signal. The resulting signal output is amplified and applied from a cathode follower circuit to both the vertical and horizontal modulator stages.

The input video signal is applied to the video differential amplifier. The video differential amplifier is a duo-triode tube, one section of which receives a d-c control voltage whose magnitude is proportional to the amplitude of the prevailing noise. The other section of the tube receives a negative-going video signal.
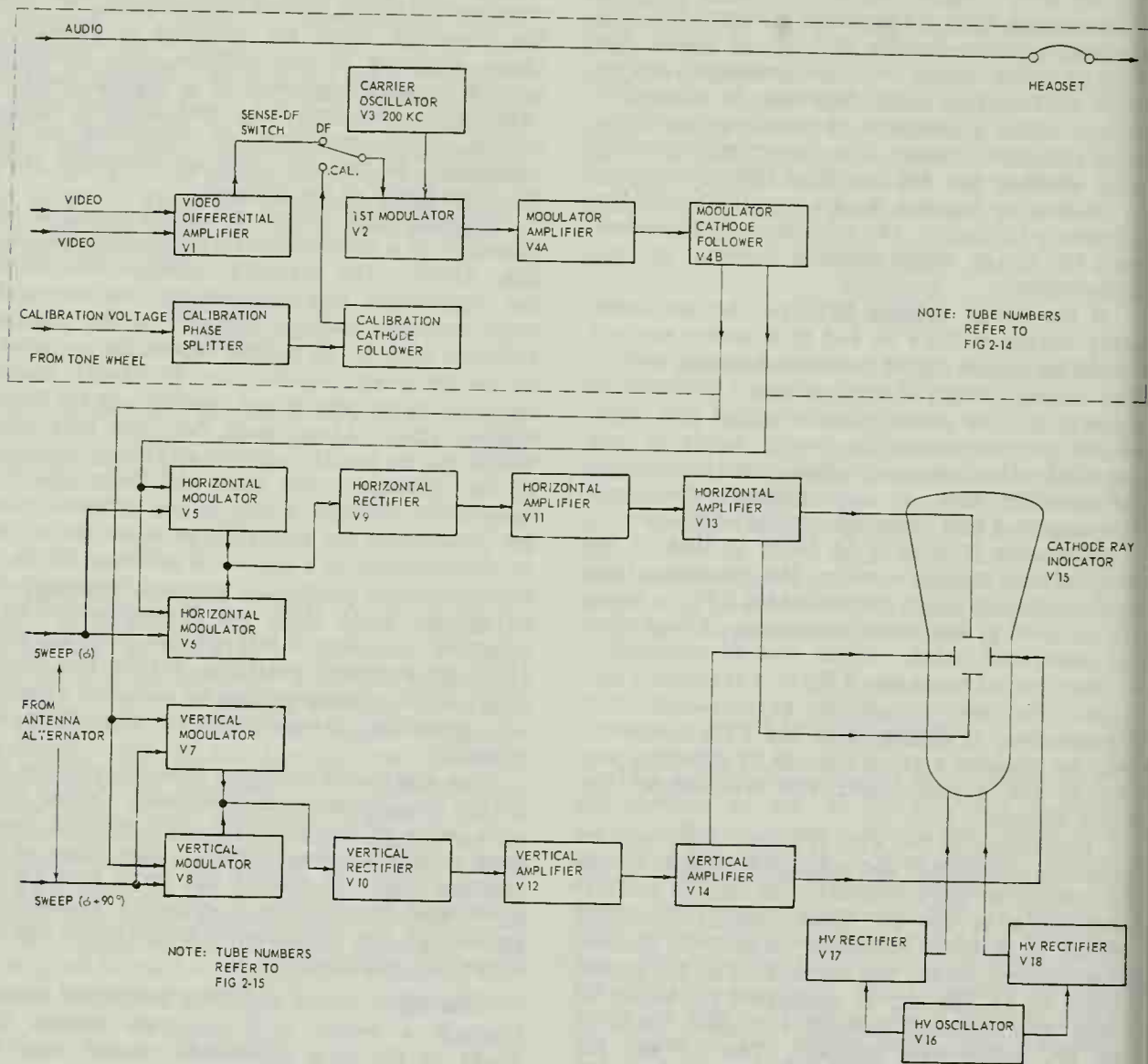
74.128.1

Figure 10-4.—Azimuth indicator, block diagram.

The video output of the video section causes the current in the noise section of the video differential amplifier to vary.

When the video input is maximum negative (representing the maximum input at the antenna) the current through the noise section of the tube is maximum, and is effective in supplying a negative-going grid voltage, which cuts off the first modulator tube. When the first modulator is cut off, no video signal can pass through the subsequent stages to the indicator, and a spot appears in the center of the indicator screen.

Conversely, when the video input is minimum negative, the conduction of the noise section of the video differential amplifier is minimum, and the first modulator stage is effective in supplying a video signal through the subsequent stages to the indicator screen. Thus, a video signal is fed to the indicator during the time of the nulls at the antenna, and the tips of the propeller-shaped pattern produced on the

FIRST DIGIT GROUP

FIRST TAB

SECOND DIGIT GROUP

SECOND TAB

THIRD TAB

THIRD DIGIT GROUP

FOURTH TAB

FOURTH DIGIT GROUP

DIAL LAMP

CHANNEL SELECTOR SWITCH

109.2
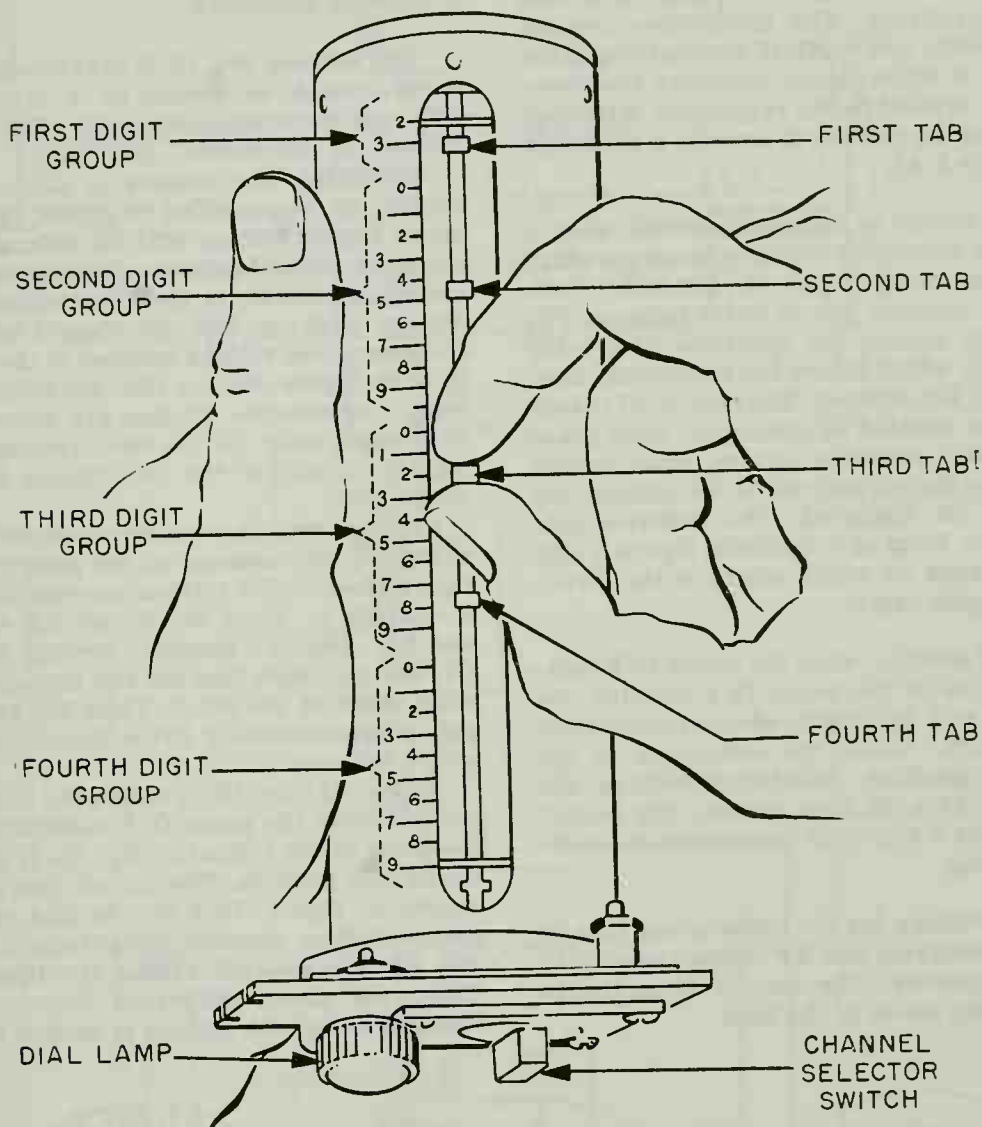
Figure 10-5.—Memory indicator.

indicator points along the nulls of the received signal.

If the sense D/F switch is placed in the CALIBRATE position, the calibration voltage (also originated at the antenna) is fed to the modulator. Thus, in this condition, the calibration voltage modulates the 200-kc oscillator carrier, and is subsequently applied to the horizontal and vertical modulators.

The calibration voltage input is derived from a 36-toothed tone wheel which generates 36 cycles of input voltage for each rotation of the

antenna shaft. The calibration input thus operates the first modulator 36 times as often as the video signal, and the pattern consisting of 36 propeller-shaped lobes is presented on the indicator for calibration purposes.

The chosen modulated voltage applied to the vertical and horizontal modulators is further modulated by the alternator voltage input (from the antenna) to each of the modulator stages. The resultant voltage from the separate horizontal and vertical modulators is rectified, amplified (in two stages) and subsequently

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applied to the vertical and horizontal deflection plates, respectively. The modulation process in the horizontal and vertical modulators, in the presence of a video signal, produces a voltage, which when applied to the respective deflection plates causes the pattern to assume a propeller shape (fig. 10-6 A).

A sense circuit is put into operation when it is desired to determine the direction from which the signal is being received. Normally, both sides of the antenna are in exact balance. The sense circuit causes the operation of a relay (shown later), which alters the electrical characteristics of the antenna. The electrical change of the antenna creates an unbalance, and a phase shift in the field response pattern, which causes the propeller-shaped pattern on the cathode-ray indicator to be distorted. The distorted pattern is in the form of a distorted figure 8 (fig. 10-6B), the apex of which points in the direction of the signal origin.

As stated earlier, when the sense D/F calibrate switch is in the sense D/F position, the video signal will be displayed as a propeller-shaped pattern. When the switch is in the CALIBRATE position, calibration voltage will be displayed as a 36-lobe pattern (fig 10-6C) which provides a means of calibrating the indicator.

The high voltage for the cathode-ray tube is obtained by rectifying and RF voltage generated by an RF oscillator. The voltage is applied to the accelerating anode of the tube.

ANTENNA CIRCUITS

The antenna (fig. 10-7) is a two-dipole Adcock array, which is rotated by a direct-coupled, 900-rpm induction motor, B1. The motor also drives an alternator, G1, and a wheel, E1.

Normally, the antenna is electrically connected, as represented in figure 10-8A. Maximum signal voltage will be induced into both antennas simultaneously when the signal is received, as shown in the illustration. Since the dipoles are one-half wavelength at the mean frequency, the voltage induced in the two halves of each dipole will be 180° out of phase. Likewise, the separate dipoles are spaced one-half wavelength apart (at the mean frequency), causing the voltage in the two dipoles to be out of phase 180°.

Assume that the received signal induces a voltage in the antenna of the polarity shown in figure 10-8, A. An antenna current is conducted for dipole 1, which is through L2, C2, T1, C1, and L1. Dipole 2 conducts through L4, C2, T1, C1, and L3. Note that the two signals reinforce each other at the balun. Thus, the receiver input is approximately twice that received by a single antenna.

Relay K1 (fig. 10-7) alters the circuit of the antenna when the sence D/F calibrate switch on the panel of the indicator (fig. 10-1) is moved to the SENSE position. The circuit then appears as shown in figure 10-8 B. In this condition, a phase-shifting network comprising L1, L2, R1, and R2 is connected across the dipole 1 input. Thus, the phase difference between the two voltages from both dipoles is shifted to alter the
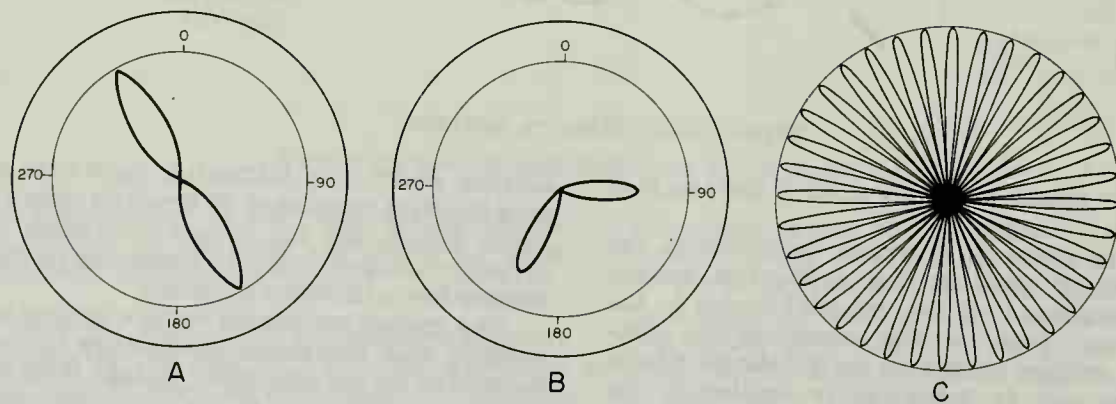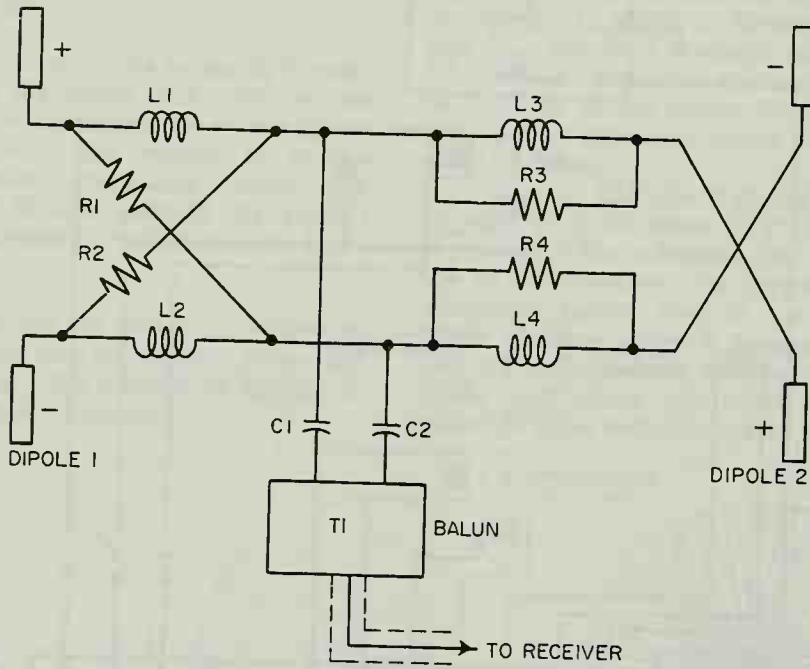

Figure 10-6.—Indicator patterns.

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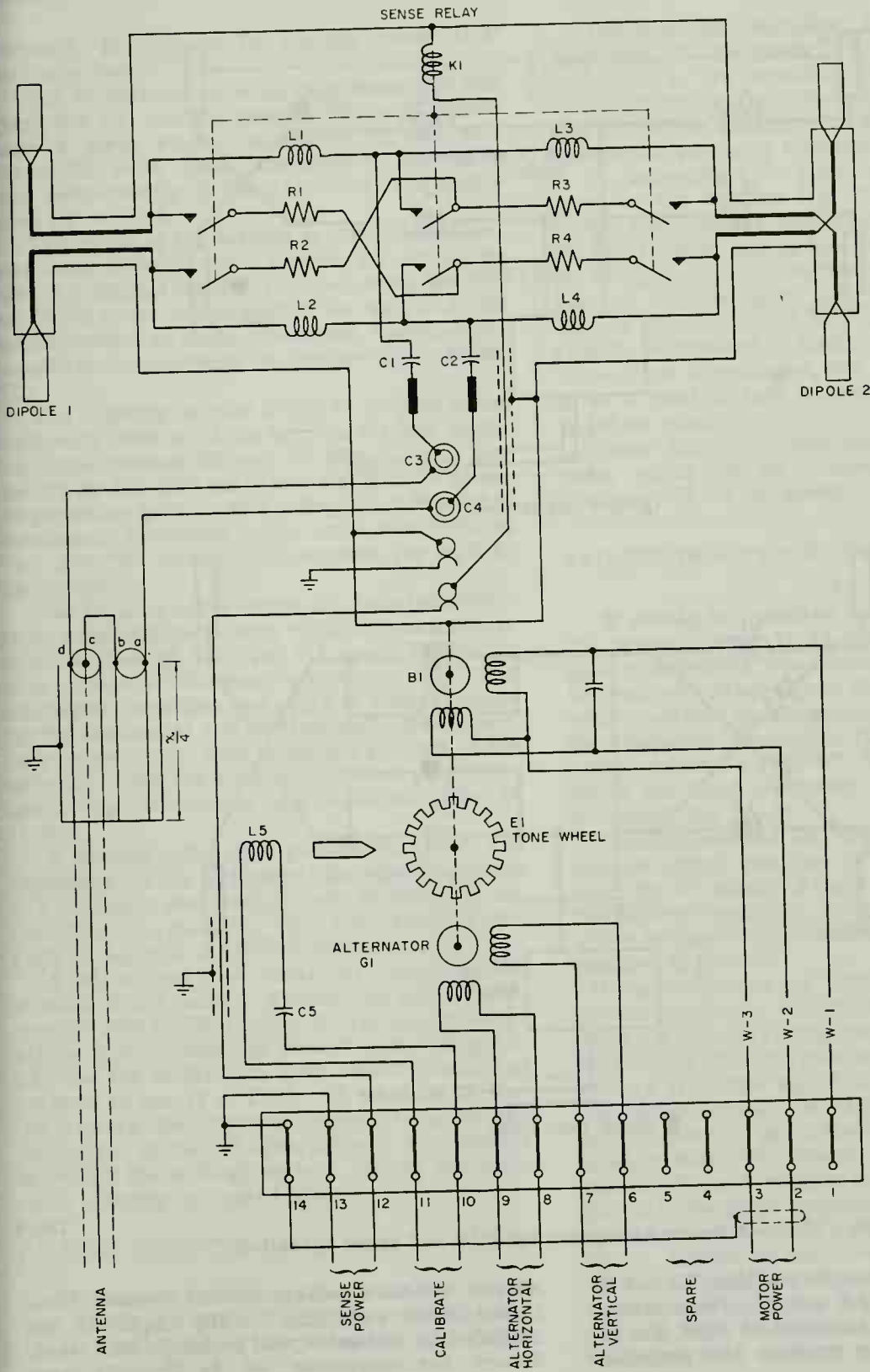
Figure 10-7.—Antenna schematic diagram.

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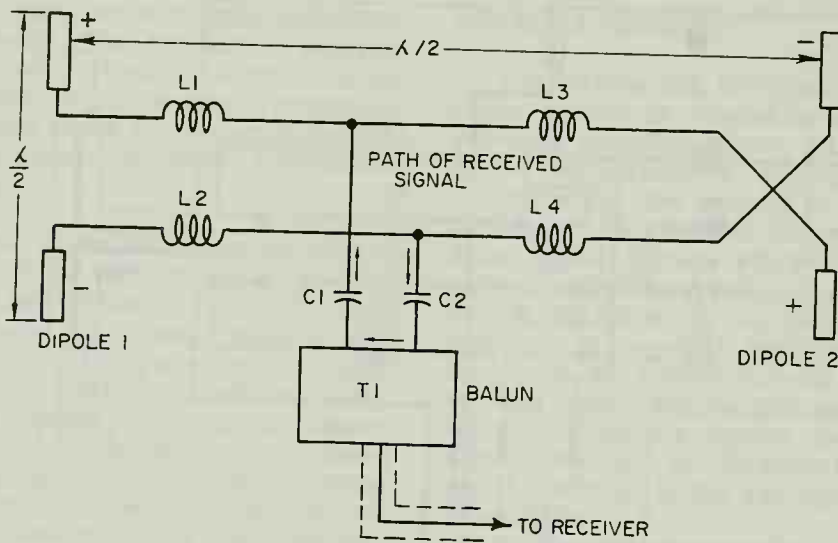
Figure 10-8.—Antenna schematic for D/F and sense operation.

109.4

pattern of the indicator. In addition, R3 and R4 are shunted across L3 and L4, respectively, causing attenuation of the dipole 2 input. The effect of the phases to produce the propeller shaped indicator pattern and the "sense" (distorted figure eight (fig. 10-6B) display on the cathode-ray indicator will be considered later. Power for operating K1 is obtained from

erminal 12 (fig. 10-7), via the sense D/F :alibrate switch.

Due to antenna rotation (approximately 900 :pm) the RF voltage induced into each dipole :rom a given source is caused to vary at a sinusoidal rate. Thus, the input to the balun :and subsequently to the receiver) is a modu-.ated RF voltage.

The output of the antenna array at the antenna :erminals is balanced to ground. To permit the use of a coaxial cable for transferring the output of the array to the input of the receiver, the antenna output is made unbalanced. The balance-to-unbalance condition is created by a balun, T1.

The rotating antenna array is coupled to the stationary portion of the antenna through rotating capacitances C3 and C4 (fig. 10-7). The energy is fed into the balun one-quarter wavelength above ground. The voltage at this point is maximum. A coaxial cable connected between "c" and "d" of the balun supplies the input to the receiver.

The antenna drive motor, B1, is an induction-type, single-phase motor, which drives a single antenna shaft at 900 rpm (15 rps). The motor also drives a two-phase alternator, G1. The alternator produces one cycle of sweep voltage for the horizontal and vertical deflection plates, respectively, for each complete rotation of the antenna. Thus the frequency of the sweep voltage to the cathode-ray tube deflection plates is 15 cps.

A 36-lobe calibration pattern (fig. 10-6C) is presented on the indicator tube when the sense D/F calibrate switch (fig. 10-1) is placed in the CALIBRATE position. One lobe appears for each 10 degrees of the 360-degree total.

A 36-toothed tone wheel, E1, driven by the antenna drive motor, affects the highly permeable core of L5 (fig. 10-7) to induce an alternating current into the circuit comprising L5, C5, and the calibration input circuit (connected between 10 and 11 of TB1). C5 tunes L5 so that the voltage developed is correct to cause the production of the 36-lobe pattern. The manner by which the induced voltage affects the indicator circuits is treated later.

RF HEAD CIRCUITS

The signal from the antenna is fed via C1 (fig. 10-7) to the RF head of the receiver (fig. 10-3). The input is a 15-cps modulated RF voltage in the 225.0- to 399.0-mc range.

The first RF amplifier, V1, the second RF amplifier, V2, the mixer, V3, and the frequency doubler, V4, are capacitance tuned by capacitors contained in their respective cavities. This type of tuning is utilized to obtain a high Q and to provide frequency stability.

To understand the operation of the cavities, consider the diagram (fig. 10-9A), which illustrates the physical construction of the cavity. Figure 10-9 B, shows the electrical equivalent of the circuit. The center post of the cavity acts as an inductance, L1. In conjunction with ganged capacitors C1, C2, and C3, L1 is electrically equivalent to a quarter-wavelength rod whose base is grounded, and whose opposite end is at a point of high impedance. The cavity is tuned to resonance by C1, C2, and C3. C4 is a trimmer capacitor, and slug Lv is a tuning vane, which can be utilized to adjust the inductance, L1, of the cavity.

VFO AND REACTANCE TUBE CIRCUITS

It should be recalled that the frequency of the mixer output is 15,325 mc. During frequency selection, reactance tube V5 receives a DC control voltage from a discriminator circuit, which controls the frequency output of the variable frequency oscillator. The VFO circuit comprises tubes V5, V6, V7, V8, and V4, and produces the input frequency, which is fed to the V3 mixer (fig. 10-3).

Remember that the control voltage polarity may be either positive or negative, depending upon the direction of the error frequency from the desired frequency. In actual operation, zero error voltage from the discriminator is slightly positive to ground.

The reactance of C41 in series with C38 (in the reactance tube circuit (fig. 10-10)) develops a capacitive reactance, which is equal to or greater than the resistance of R31. The r-f voltage from the tap of L18 to ground is also developed across C38, R31, and C41. The voltage across C41 is applied to the grid of the reactance tube, V5. Due to the capacitances involved, the r-f voltage at the grid of V5 lags the L18 (variable frequency oscillator) voltage by approximately 90°). The plate current of V5 is always in phase with the grid voltage. Thus, the plate r-f current lags the vfo oscillator r-f voltage by an angle that approaches 90 degrees. V5 is therefore an inductive reactance shunted across the oscillator circuit.
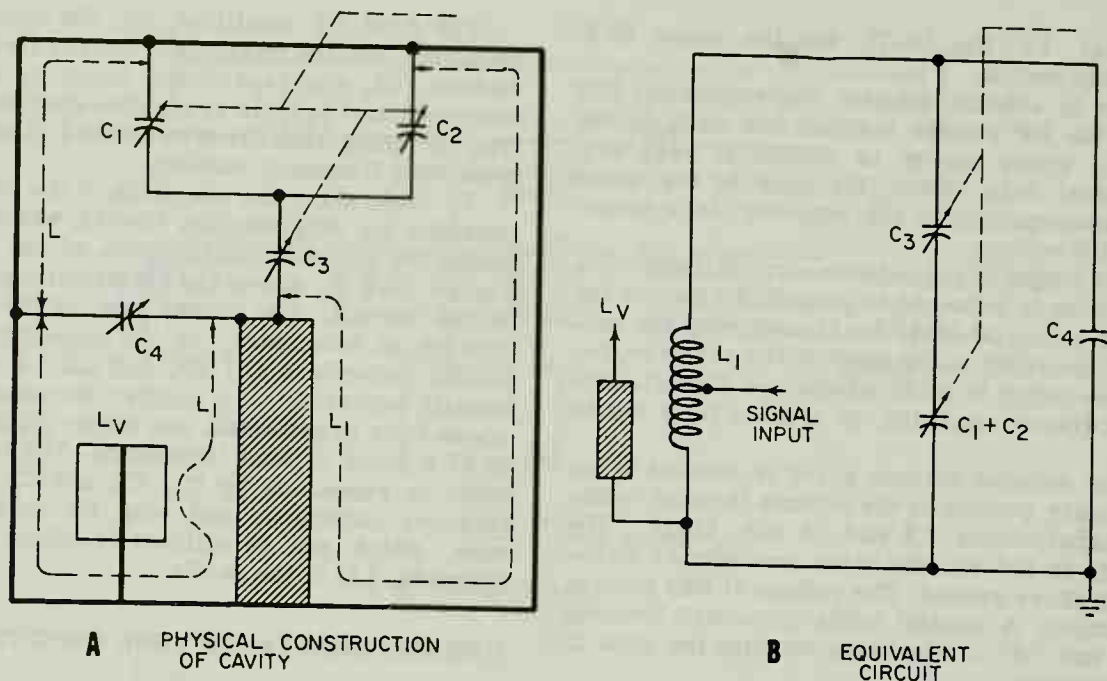
Figure 10-9.—RF head cavity, physical and electrical equivalent circuits.

The d-c control voltage is fed through a de-coupling network comprising C46 and R27 to the L17-C45 junction. Assume that this voltage goes more positive to ground. Tube V5 will conduct a heavier current, and therefore acts as a smaller inductive reactance across the oscillator to increase the oscillator frequency output. Conversely, if the control voltage goes in a negative direction, a larger lagging volt-age is applied to the oscillator, and the oscil-lator frequency decreases. Thus, by controlling the bias at the grid of V5 with the DC control voltage, the tube current and, in turn, the induc-tive reactance (as seen by the VFO) is also con-trolled. Cathode resistor R26 develops a bias of such a value that the control voltage up to approximately +5 volts DC is the equivalent of a zero error signal from the monitor.

The variable frequency oscillator circuit, V6A, is connected as a series-fed Hartley os-cillator whose feedback is developed between the lower and upper sections of L18. C36A pro-vides oscillator tuning, and C37 is the trimmer capacitor. The corrected oscillator output frequency is coupled by C33 to the control grid of a cathode follower. V6B, which serves as an isolation stage between the oscillator and the

frequency tripler stage, V7, and output doubler, V8.

V8 acts as a frequency doubler stage. The V8 output is again doubled by frequency doubler stage V4.

The frequency of this signal is 12 times that at the output of the VFO, V6A. With the VFO fre-quency corrected, the heterodyning process in the mixer stage, V3, produces a 15.325-MC sig-nal, which is fed through a filer circuit to the grid of the first FF amplifier, V9. This stage, like the first and second RF amplifiers, is AGC controlled. Amplifier V9 is a conventional IF amplifier stage whose output is fed to the IF video-audio chassis.

FREQUENCY SELECTION
CONTROL CIRCUITS

As stated earlier, the memory cylinder lo-cated behind the front panel of the indicator can be preset to select any one of 20 of the total 1750 available frequencies. Frequency selec-tion is initiated by the four lateral bars (fig. 10-5) which mechanically actuate various con-tacts of the channel selector switch.
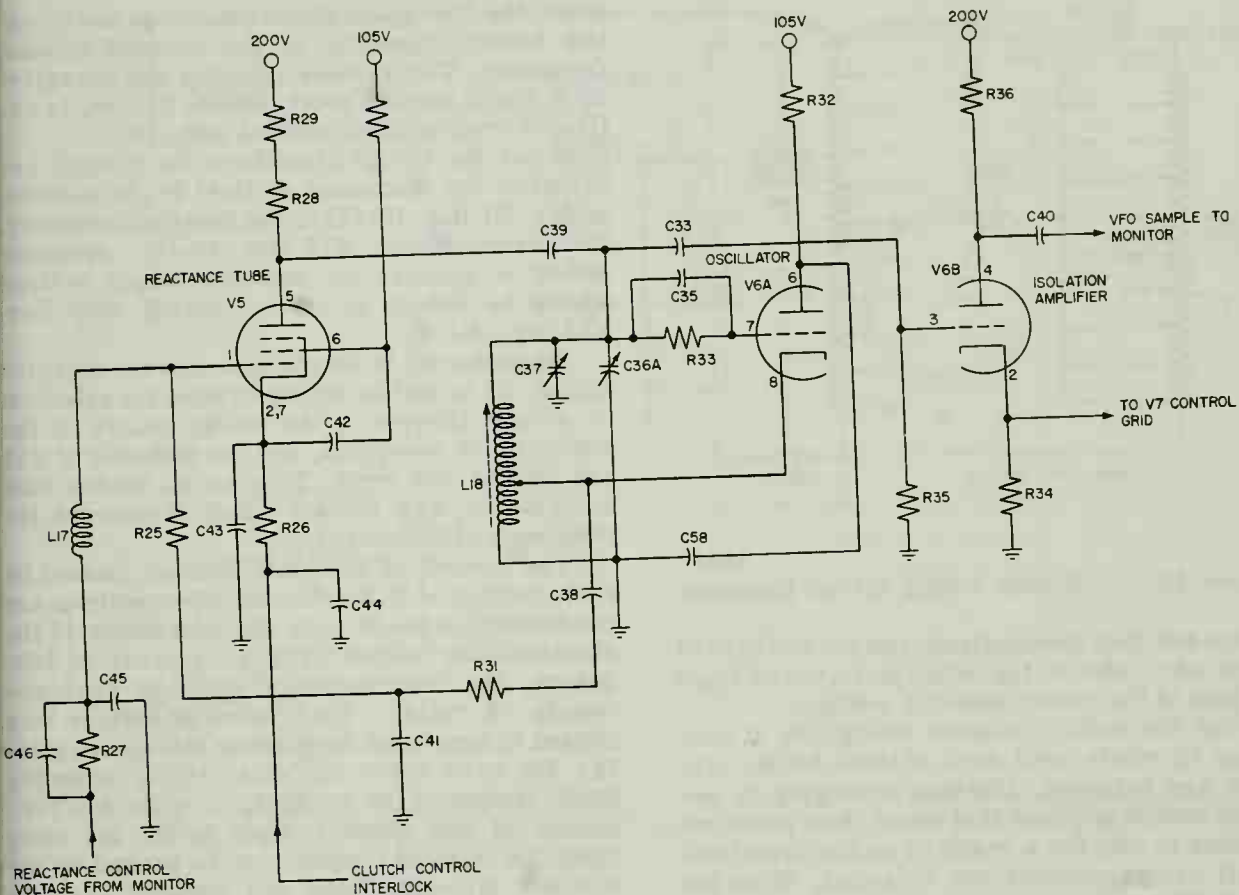
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Figure 10-10.—Reactance control circuit.

When one of the preset frequencies is to be selected, the channel selector switch is turned to the new position, which causes the unbalance of bridge circuits. These circuits are similar to that shown in figure 10-11. Three similar bridge circuits exist, corresponding to the second-third-, and fourth-digit controls of the frequency selection system.

To select the new frequency, the channel selector switch chooses a tap on its associated bridge resistor, which corresponds to the position of the actuating bar on the memory cylinder. It should be recalled from the block diagram discussion that the monitor circuits produce a discriminator DC control voltage, which controls the frequency output of the vfo in the RF head. Thus, a change in the input frequency to the RF head necessitates a change in the monitor frequencies so that the vfo will be locked in at the correct frequency.

The channel selecting bridge circuits (fig. 10-11) are established between the indicator and monitor. If that section of the channel selector switch (for a particular bridge) and associated rotary selector switch in the monitor are not at the same tap on their respective resistors, a difference in potential will exist at the ends of the bridge. The voltage thus established is placed across a bridge relay, K4, and the relay is energized.

With K4 energized, its contacts close to connect a clutch relay, K1 (fig. 10-12), between a +27.5-volt-supply and ground. The 3-4 contacts of K1 close to apply motor power to the selector motor, B1. The clutch relay also releases its associated clutch face by pulling a pawl, P, out of a notch on a detent plate, CP1. The released clutch face allows the selector motor to operate, driving the wiper on the rotary selector switch S1F, until it reaches a tap, which again balances the bridge. The bridge and clutch

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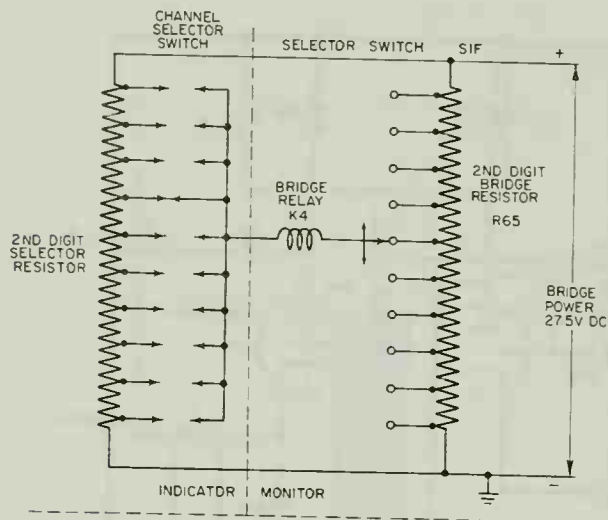
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Figure 10-11.—Bridge tuning circuit diagram.

relays are then deenergized, and the clutch pawl drops into a slot on the detent plate, which stops the arm of the rotary selector switch.

When the motor becomes energized, it continues to rotate until each of three bridge circuits are balanced. Clutches belonging to selector switch sections that reach their positions continue to slip (as a result of motor drive) until all bridge circuits are balanced. When the last bridge relay deenergizes, the clutch relay also deenergizes, and the motor stops.

Sections of the selector switch are ganged to their respective clutch plates (not shown). The clutch plates (when released) drive their respective switch sections to select the proper combination of components in the harmonic generator and mixer, and the second, third and fourth crystal oscillators of the monitor.

TUNING MOTOR CONTROL CIRCUITS

The tuning motor controls the tuning capacitor of the VFO by three clutches designated the high-, medium-, and low-speed clutches, respectively. The total frequency range of the vfo is divided into six equal ranges. The high-speed clutch drives the tuning mechanism at high speed until the segment is reached in which the selected frequency lies. The medium-speed clutch then takes control to drive the tuning capacitor at medium speed until the approximate range of the half power points on the response curve of the 4th mixer is reached. At this

point, the low-speed clutch takes over and drives the tuning capacitor to the desired channel frequency. The various clutches are energized by a clutch control relay, which, in turn, is activated from a clutch control amplifier.

After the bridge circuits in the monitor are balanced (as discussed earlier) by the selector motor, B1 (fig. 10-12) to the desired frequency, the discriminator, V12 (fig. 10-13) produces either a positive or negative output voltage which is fed to a motor control amplifier, V13 (fig. 10-13).

When the vfo is off channel, and the selector motor, B1 is setting up the system for selection of a new frequency, the bridge relays of the monitor are energized, and the cathodes of V13 are tied to +30 volts. This action biases both sections of V13 beyond cutoff throughout the frequency selection period.

The amount of the plate current passed by each section of V13 (when the tube sections are conducting) depends upon the magnitude of the discriminator output. For zero error in frequency, the discriminator output is approximately +5 volts. This value of voltage was chosen to represent zero error since zero voltage for zero error can exist either when the input frequency is correct, or when the frequency at the monitor input is too far away from the desired frequency to be passed by the monitor circuits. When the monitor circuits are on frequency, the discriminator output is +5 volts. The grid (pin 7) voltage is +8 to ground, while the cathode is approximately 10.2 volts. The V13B grid (pin 3) is about 5.3 volts to ground. Thus, V13B receives about 5 volts bias and does not conduct a sufficient current to energize its plate relay, K7.

When the input frequency is too far above the chosen frequency to be passed through the monitor (as may be the case when a new channel is selected), the discriminator output voltage is zero. The V13A grid voltage decrease from +5 volts to zero causes a smaller current through K8 (in the V13A plate circuit), and the K8 contacts are released (position shown). Likewise, the smaller current through R53 reduces the V13B bias. The increased V13B current through the solenoid of K7 energizes this relay to move the contacts down (opposite to that shown). This action causes the tuning motor, B2, to turn by completing the motor circuit from ground at the 2-4 contacts of K7 (K7 energized), through S4A (in the position shown), through the motor filter, and through
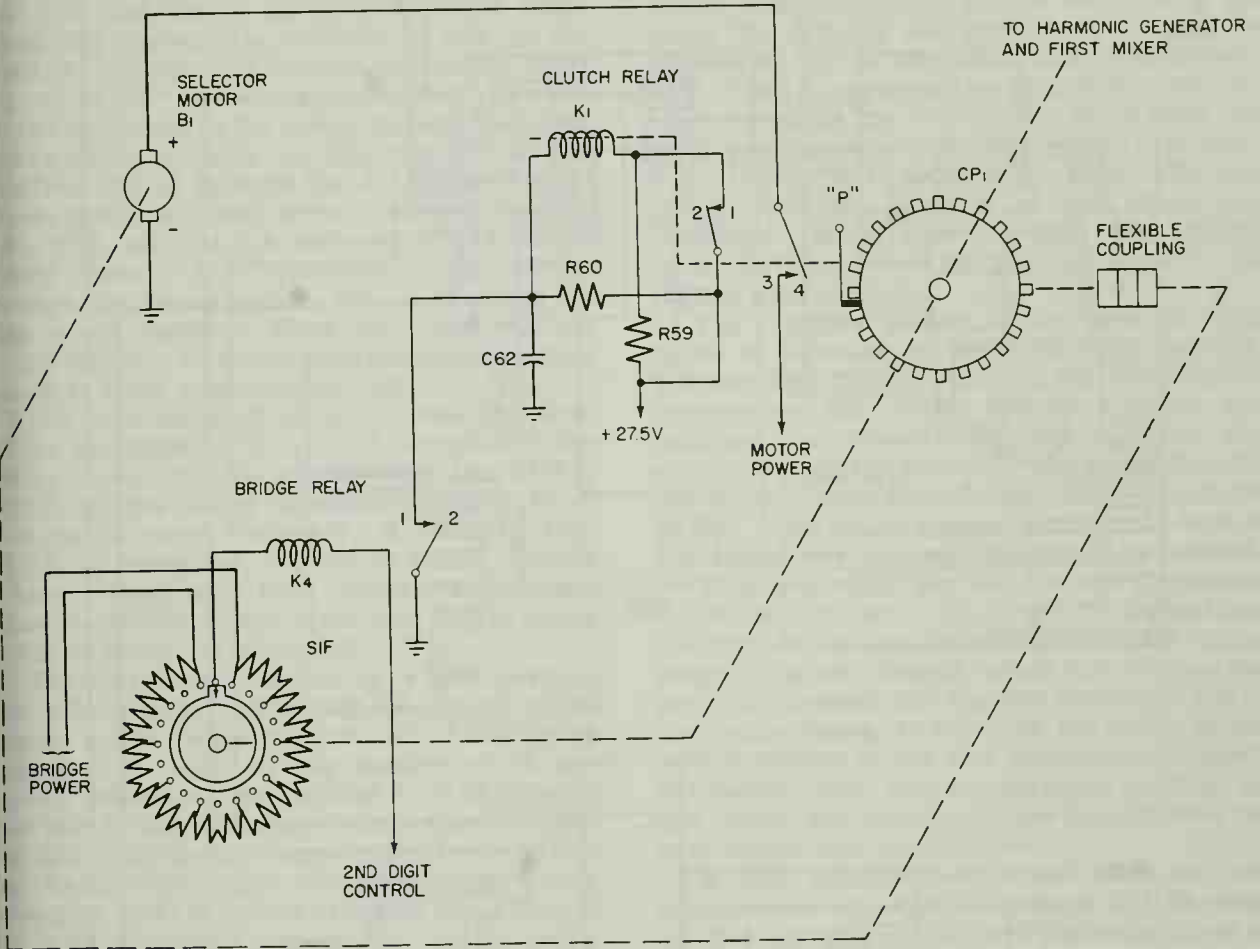
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Figure 10-12.—Frequency selection control circuits.

the motor from a to b, through another motor filter, and the contacts of S4B (in the position shown), through the 8-7 contacts of K8 (K8 deenergized), through R56, the 3-2 terminals of K14, through the 3-4 contacts of K13, and to the +30-volt unfiltered motor power. The polarity of the voltage applied to the motor is such that the vfo tuning capacitor is driven towards the selection of a lower frequency.

When the 4th mixer, V10, output (fig. 10-10) is near the chosen frequency, the discriminator, V12, circuit is functioning, and the reactance tube, V5, (fig. 10-10) begins to operate. The discriminator, V12, output rises to approximately five volts positive. The conduction of V13A (fig. 10-13) increases, and its plate relay, K8, becomes energized. V13B continues to conduct, and both relays are now energized.

The 8-7 contacts of K8 are opened when K8 is energized, while the 4-2 contacts of K7 (when the relay is energized) are closed. With S4A in the position shown this action removes drive power from the motor.

If the frequency at the discriminator input tends to fall below the zero error frequency (vfo low in frequency) the magnitude of the discriminator, V12, output voltage will be increased in the positive direction. V13A conducts heavily, and K8 is energized.

The bias increase on V13B causes the V13B plate current to decrease, and K7 is deenergized. Power is now applied to the motor from a ground on the contacts of a high speed clutch control relay, K12 (discussed later). The motor circuit is connected from a ground on S5, through the 7-8 contacts of K9, through
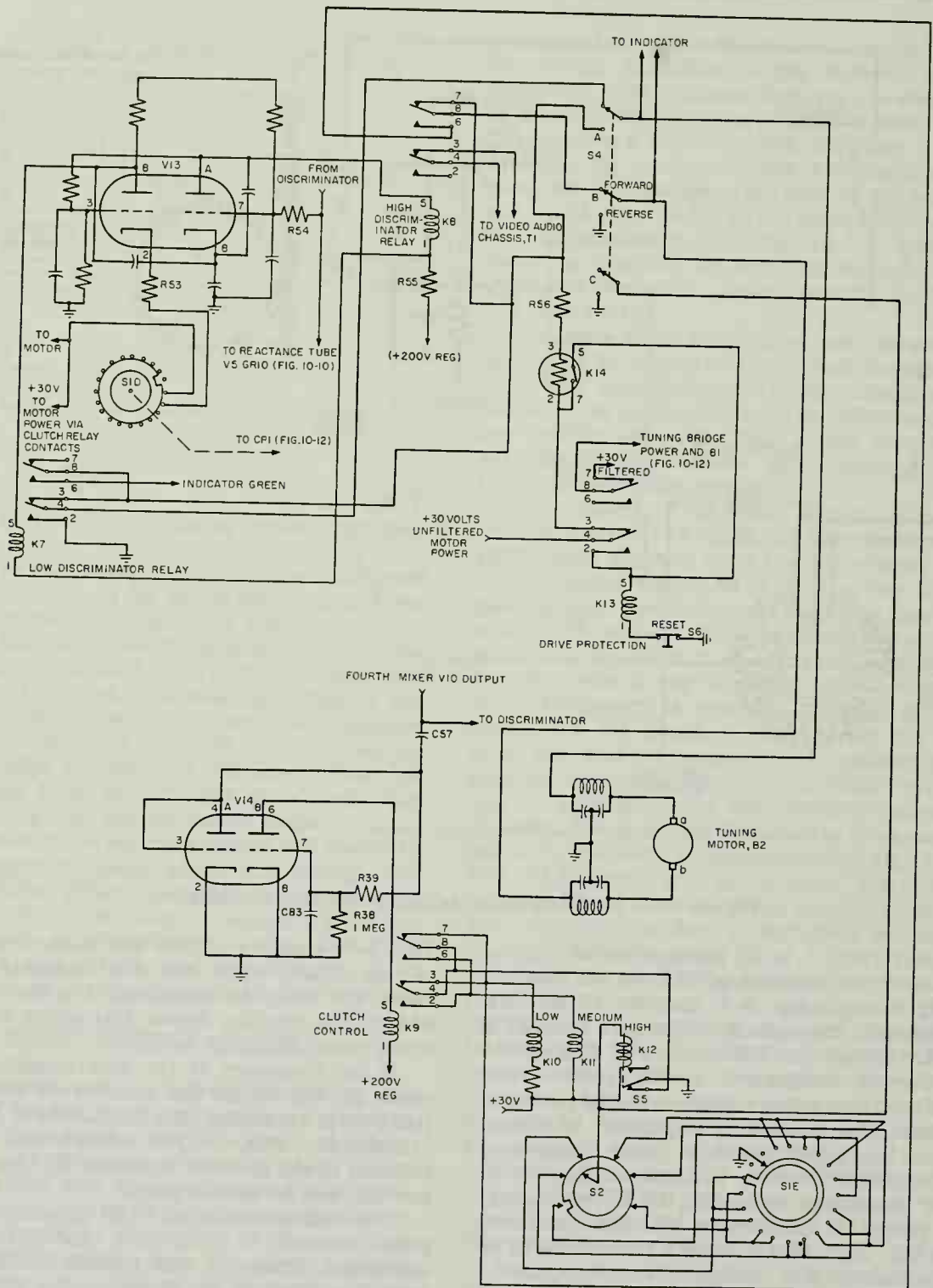
Figure 10-13.—Tuning motor control circuits.

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the 6-8 contacts of K8 (closed in this error condition) through the contacts of S4B (in the position shown), and through the motor filter circuit to "B" on the motor, B2. The +30-volts power is applied to the motor through the upper motor filter, through the S4A contacts (in the position shown) through the 3-4 contacts of K7 (deenergized for this error condition) through R56, K14, and the 3-4 contacts of K13, to the motor power. The direction of the current through the tuning motor, B2, is reversed for this error condition. Thus, the motor will run in the opposite direction until the discriminator output is again approximately +5.

From this action, it should be seen that when K7 is energized, a DC voltage is applied to the tuning motor of such polarity that the VFO if driven (by its tuning capacitor C36) (fig. 10-3) towards a lower frequency. If relay K8 (fig. 10-13) is energized, the vfo is driven towards a higher frequency. Both relays are energized when the VFO frequency is correct, and the tuning motor is braked at this point.

Forward-reverse switch S4 is cam operated, and effective in controlling the speed of the tuning motor. When the low end of the tuning capacitor is reached, the contacts of S4 are moved downward, and section C of S4 grounds one end of the high-speed clutch thus energizing K12. When the high-speed clutch is energized, the tuning motor (and consequently the tuning capacitor C36) is driven at a high speed toward the high frequency end of the tuning range.

The action of S1E is to divide the entire frequency range into six frequency segments. When the high-speed clutch, K12, is energized, the tuning motor continues to operate at high speed until the sector in which the preset frequency lies is reached. At this time S1E (which is driven by CP1) operates to remove the ground from the high-speed clutch, K12. When the tuning system is far off frequency, K9 is energized with its 4-2 and 8-6 contacts closed. This action will be considered later. The deenergizing of K12 permits S5 to supply a ground for the medium-speed clutch, K11. The K11 circuit is completed from ground, through the lower contacts of S5, through the 4-2 and 8-6 contacts of K9 in parallel through the K11 solenoid, to the +30-volt motor power. With K11 energized, the tuning motor is driven at medium speed until the selected frequency beings to pass through the monitor.

When the output of the 4th mixer (V10, figure 10-3) is such that it can pass through the

fourth band-pass filter circuit comprising the tuned T5 primary and secondary, a signal is coupled by C57 to the clutch control amplifier, V14. V14A is connected as a diode (fig. 10-13), which rectifies the V10 output. Each time the input signal goes positive with respect to ground, V14A conducts to effectively short R39 and R38. Negative portions of the signal are developed across R38 and R39, and applied from the junction of the two resistors to the control grid of V14B.

The negative pulses at the grid of V14B cause a decrease in the V14B plate current. The smaller current through the K9 solenoid deenergizes this relay, and the contacts are released (as shown). With K9 released, the ground is removed from K11 and supplied to the low-speed clutch, K10, through the 3-4 contacts of K9. This action causes the motor to turn at low speed until the exact frequency is reached.

When searching for the desired frequency, the tuning capacitor, C36, travel is always from the high to the low frequency end of the tuning range. If a new channel, which lies between the previous channel and the low frequency end of the tuning range, is selected, the tuning motor will be driven to the new channel and stopped. The tuning drive may start at high, medium, or low speed, depending upon how far removed the new channel lies from the old.

If K12 remains deenergized when the new frequency is selected, S5 grounds K11 (through the 2-4 contacts of K9), and the motor turns at medium speed. If the new frequency is such that the monitor continues to receive an input, K9 will be deenergized as a result of negative pulses at the V14B grid (which reduces the V14B plate current), and the ground from S5 is supplied through contacts 3-4 of K9 to the low-speed clutch. Thus, the motor will turn at low speed to the new frequency.

If a new channel frequency, which lies between the previous frequency and the high frequency end of the tuning range is selected, the tuning capacitor, C36, will first be driven to the high frequency end. Cam operated switch S4 then reverses the direction of rotation of the motor, causing the motor to start back again to search for the new channel frequency.

In the event that the vfo is not set to the new channel frequency after traveling completely across the tuning range, it is returned to its high frequency end at high speed by the reversal of S4. In the reverse position, S4A and S4B connect the motor directly in series with R56,

K14, and the 3-4 contacts of K13, across the +30-volt power bypassing relays K7 and K8 and thus remove the motor entirely from the control of the monitor.

When the VFO reaches the high frequency end of its range, S4 returns to the position shown. The monitor circuit discriminator relays, K7 and K8, again take control, and the tuning capacitor, C36, is driven back towards its low frequency end in search of the new channel frequency.

VIDEO DIFFERENTIAL
AMPLIFIER CIRCUITS

The video differential amplifier, V1 (fig. 10-14) receives the video input signal from V39A at its B section grid. At the same time, a DC cathode potential proportional to the level of the noise is applied to the A section grid of V1 from V39B. Both sections of V1 are connected between the -150- and +87-volt supplies.

The bias applied to each section of V1 is approximately 1 volt, and both sections of V1 are normally conducting. If the noise component of the signal to V39B (fig. 10-3) increases, the cathode-to-ground potential of V39B (across R1, fig. 10-14. decreases. This action lowers the conduction of V1A. Note that the cathode-to-ground potential will decrease for both sections of V1 since the two cathodes are tied together by R2 and R2A. R2 can be adjusted to balance the currents of V1A and V1B in the absence of an input signal.

The video signals to the V1B grid are negative pulses from V30A (fig. 10-3). For zero video input to V39A (or when the received video pulse amplitude reduces to zero) V39A conducts heavily, and the positive voltage at the V1B grid (fig. 10-14) is increased. V1B conducts heavily, and the cathode potential of both sections of V1 increases. Due to the increased common cathode potential, V1A can only conduct a small current through R6. The less positive potential at the top of R6 is applied through R7 and S1B (in the position shown) to the control grid of the first modulator, V2. Since the V1A conduction is small, the corresponding negative-going voltage to the V2 grid is not sufficient to cut off V2.

As the video signal increases in negative amplitude, V1B conducts less. The resulting decrease in cathode potential of both sections of V1 allows V1A to conduct a heavier current through R6. This action decreases the V1A plate voltage and the V2 grid voltage. The decrease control grid potential at V2 cuts off

the first modulator stage. Thus, for strong portions of the video (corresponding to maximum received signal at the antenna) the first modulator, V2, is held beyond cutoff to produce zero output. For zero video input, which corresponds to the nulls of the video, the first modulator stage, V2, is above cutoff, and thereby produces an output (as discussed later).

FIRST MODULATOR AND 200-KC
OSCILLATOR CIRCUITS

The first modulator stage, V2 (fig. 10-14) provides a method of combining the video signal with an internally generated r-f signal to produce a modulated video output. The modulation process is necessary so that the proper mixing of the video and alternator voltages from the antenna can be accomplished in subsequent modulator stages.

The DC potential at the grid of V2 is determined by the setting of the arm of circle diameter control, R9. The V2 grid potential, in turn, determines the operating level of the modulator.

In the SENSE-D/F position of S1, the signal from the video differential amplifier, V1, is applied across R10 to the V2 control grid. When the video input signal at V1 is zero, V2 is conducting. A 200-KC carrier oscillator (V3 of fig. 10-14) is connected as a transitron circuit. The transitron oscillator utilizes the negative transconductance of the tube to establish oscillations and develop an output frequency. Note that the screen potential of V3 is higher than that of the plate.

Screen grid signal current (which varies at the frequency of the oscillator) causes a negative voltage to be developed at the suppressor grid of V3. The cathode potential of V3 is about +4.6 volts. This further increases the suppressor-cathode negative potential.

Under these conditions (with the screen effectively controlling the tube current) the plate current is greatly reduced because the plate voltage has less influences on the electrons leaving the cathode than the screen. Many of the electrons, which do succeed in passing through the screen grid (and would normally add to the plate current) are repelled by the negative suppressor back to the screen grid. By selection of the proper relationship of plate and screen voltages, the respective currents will be approximately equal.

The negative suppressor potential produces a negative resistance effect in the tube. This is true because above a certain critical screen
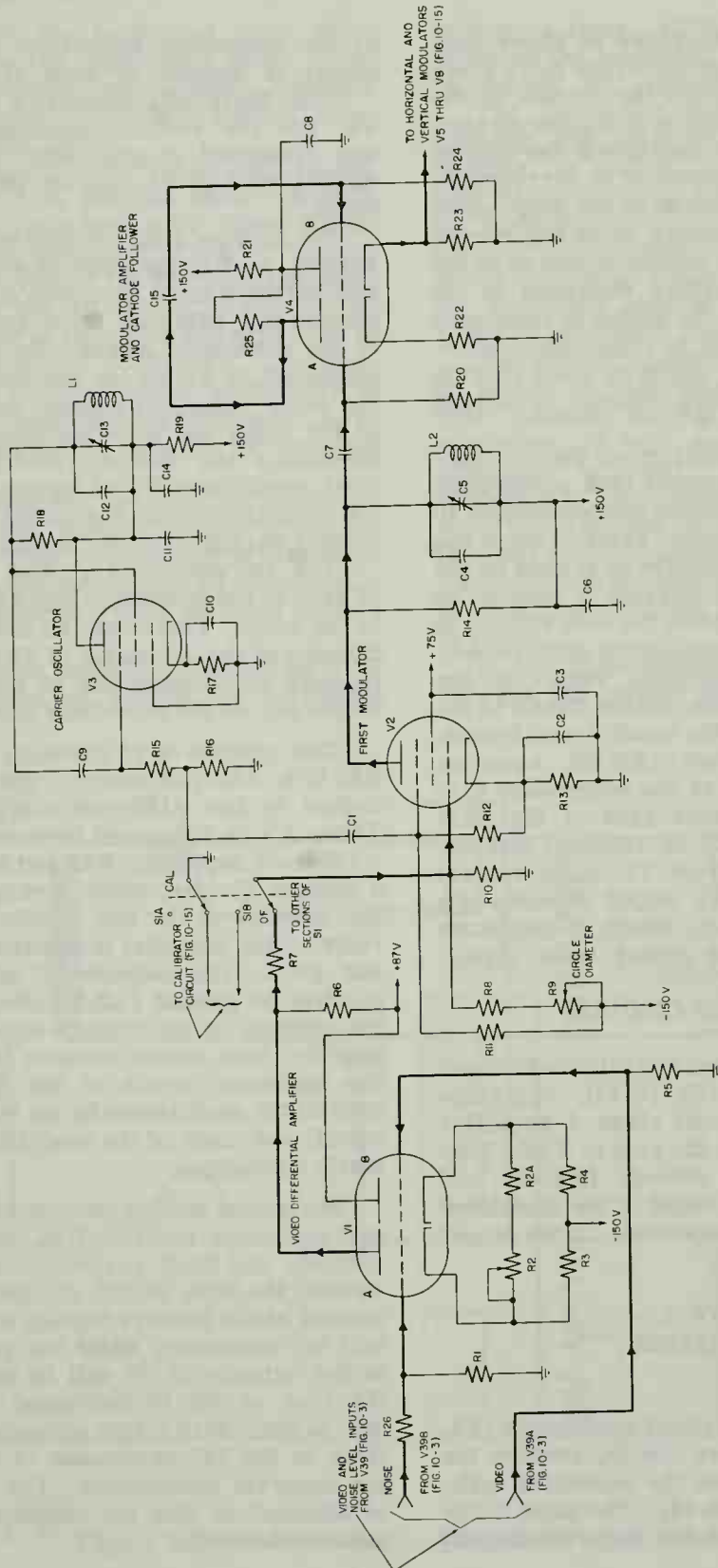
Figure 10-14.—Azimuth indicator (video amplifier and first modulator circuits).

109.10

potential any further increases in suppressor voltage in a positive direction allow more electrons to be passed through the screen to the plate, resulting in a decrease in screen current. Since the screen current decreases, the amount of the suppressor bias produced is correspondingly reduced. The reduction in the suppressor bias is equivalent to a positive-going suppressor voltage, which allows a further increase in the plate current and a further decrease in the screen current. Thus, the action is regenerative, and a small change in suppressor potential produces a greater effect on plate current than will a corresponding change in screen potential.

When the circuit is oscillating, the alternating component of the screen tank comprising L1, C12, and C13 is applied simultaneously to the screen and suppressor grids. When the voltage is made more positive to ground by the tank oscillations, screen current is reduced as discussed. Conversely, when the tank voltage is going less positive, screen circuit is increased. The screen current therefore varies at the oscillator frequency. The screen output to the oscillator tank supplies the tuned circuit losses.

The oscillator frequency is 200 KC. A portion of the signal developed at the suppressor grid is fed to the suppressor grid of the first modulator, V2. It should be recalled that the video input is applied from V1 to the control grid of V2. Thus, the V2 output consists of a 200-kc carrier frequency, which is amplitude modulated by the video pulsed input signal.

MODULATOR AMPLIFIER CIRCUITS

The modulator amplifier and cathode follower are combined in tube V4 (fig. 10-14). This stage functions as a conventional class A amplifier and supplies its output to the grid of V4B. This section functions as a cathode follower and matches the impedance output of the modulator amplifier to the input impedance of the stages that follow.

HORIZONTAL AND VERTICAL MODULATOR AND AMPLIFIER CIRCUITS

The horizontal and vertical modulators (V5, V6, V7 and V8 of figure 10-15) receive the modulated video input from the modulator cathode follower, V4B (fig. 10-14). The input to the modulators is coupled to the suppressor grid of its respective modulator. Thus, the same signal is applied to each of the modulators.

The horizontal modulator stages, V5 and V6, and the vertical modulators, V7 and V8 are connected in push-pull. Horizontal modulators V5 and V6 operate into separate tuned tanks.

To produce a single pattern on the indicator screen, each of the modulator tubes (horizontal and vertical) must be driven so that their video outputs are balanced from saturation to cutoff of the modulator stages. By supplying the DC potentials to all of the modulator stages from the same regulated source, and by the control of the video and no video input condition at the horizontal and vertical modulators from the first modulator (V2 of figure 10-14), each of the modulator stages supplies video to the indicator during the same period.

The DC potentials applied to the elements of each of the modulator tubes are approximately the same. Therefore, an analysis of the relationship of the potentials of V5 will be sufficient to show the relationship of the potentials that exists for all the modulator tubes.

The cathode of V5 connects through R44 and R43 to a -150-volt source. The V5 plate is connected to the +150-volt supply as discussed. Thus, V5 is connected between the -150- and +150-volt supplies. R43 and R44 are a part of a voltage divider, which develops a potential at the junction of the two resistors of about +0.7 volts. This potential is applied to the suppressor grid. The cathode DC potential is more positive to ground (+3.5 volts) as a result of the addition of the voltage across R44. Thus, a negative bias exists between the cathodes and the suppressor grids of the modulators. The bias on the modulators is not sufficient to cause cutoff, and each of the modulator tubes is normally conducting.

Horizontal modulator control, R31, and vertical modulator control, R38, serve as balance controls for their respective modulators. By moving the arm of R31 clockwise (upward) the amount of the positive voltage at the V5 cathode will be decreased, while the positive potential at the cathode of V6 will be increased. Thus, the bias of V6 is increased and its gain is decreased. This action compensates for variations in the DC resistance of the two tubes of the separate modulators. The vertical control is adjusted so that the output of both tubes is equal (balanced).

In addition to the modulated video input which is fed to the suppressor grids of the vertical and horizontal modulators), a second input at the control grids is applied from alternator, G1, (fig. 10-7).

As the alternator rotor turns (fig. 10-16), voltages are induced into the stator windings "a" and "b"), which bear a 90° phase difference. The voltage developed by the "a" winding is applied to the grids of the horizontal modulators, V5 and V6 (fig. 10-15). R105 (fig. 10-16) provides a method of balancing the grid input to the separate modulator amplifiers. Likewise, voltage applied to the vertical modulators is fed from the "b" winding of the alternator. R116 acts as the grid balancing resistor for the input to V7 and V8.

The phase difference between the horizontal and vertical input signals at the alternator should be 90°. This relationship is necessary to produce the circular pattern on the cathode-ray tube of the indicator when no video input is being received.

A phase shifting network, comprising R102 and C58, is connected across the horizontal

modulator input. R102 is variable so that the phase of the horizontal input can be adjusted to be exactly 90° relative to the vertical input.

The horizontal and vertical gain balance controls, R96 and R107 (fig. 10-15) respectively, function to control instantaneous changes in tube resistance (and tube conduction) of each of the modulator tubes. The action is necessary to prevent the introduction of distortion, which would otherwise result in nonsymmetrical output amplitude signals from the horizontal (or vertical) modulator tubes.

The simplified diagram of figure 10-17 should be helpful in understanding the circuit operation. Assume that R31 has been set so that the conduction of both sections of the horizontal modulator is equal. The corresponding fixed resistors in each section of the bridge are of equal value so that the current through path abc is equal to the current through path adc, if the V5 and V6 grid-cathode resistances are equal. In this condition, the voltage bd (across R96) is zero.

If the grid-cathode resistance of V6 decreases, the current through path adc increases.

109.12
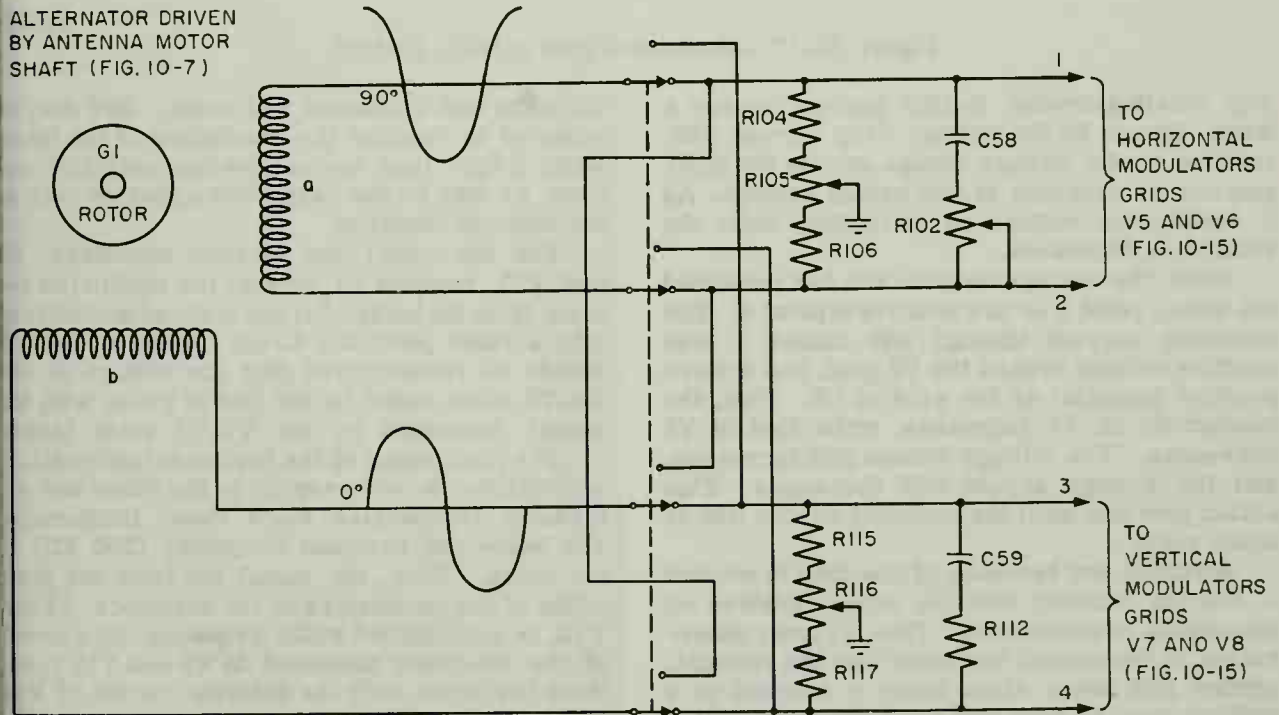
Figure 10-16.—Alternator and modulator input circuit.

347
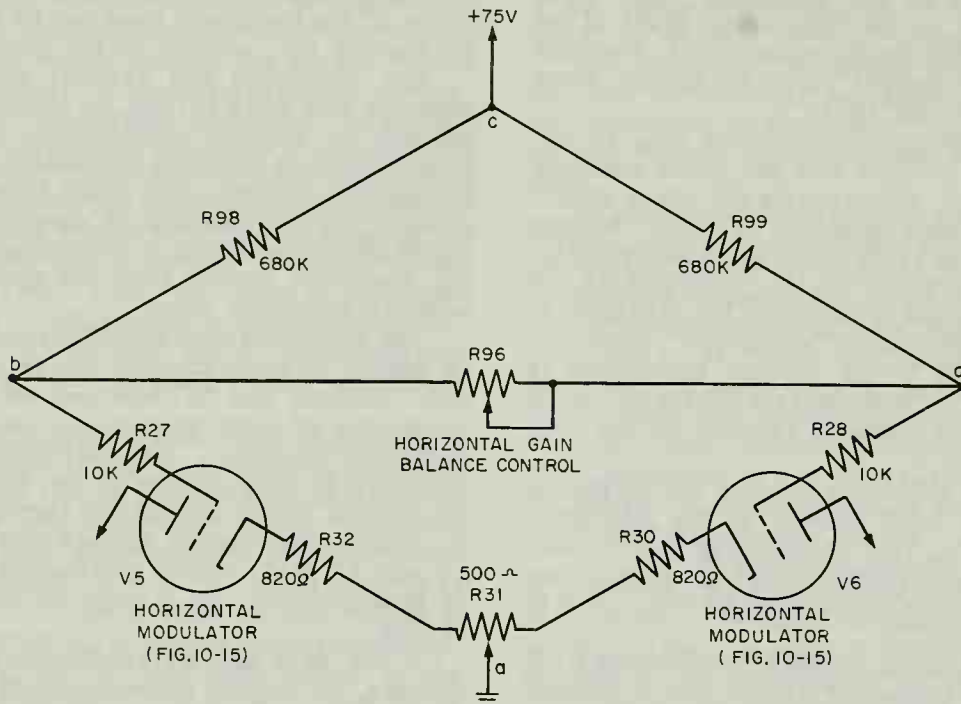
Figure 10-17.—Horizontal gain balance control.

109.13

Any small increase in this current causes a large change in the voltage drop across R99, relative to the voltage change across the other smaller resistances in this bridge section. As a result, the voltage cd increases, while the voltage da decreases.

Since the voltage division abc has remained the same, point b is now positive to point d. The resulting current through R96 causes a less positive voltage toward the V6 grid, and a more positive potential at the grid of V5. Thus, the conduction of V5 increases, while that of V6 decreases. The voltage across R98 increases, and the voltage across R99 decreases. This action prevails until the potential across R96 is again zero.

Actually, any tendency of one tube to conduct a heavier current than the other receives an immediate counteraction. Thus, a closer observation of the circuit will show that the potential across R96 never rises above a fraction of a volt so long as both tubes are operating within their conduction ranges.

If section abc of the bridge attempts to produce a predominating current, the reverse

action to that discussed will occur. R96 may be adjusted to balance the conduction of the tubes when larger than normal current variation occurs, as may be the case when either or both of the tubes is replaced.

The horizontal and vertical rectifiers, V9 and V10, function to convert the modulated r-f input from the horizontal and vertical modulators into a video pulsating direct current output. It should be remembered that the voltage at the V5-V6 plate tanks is 90° out of phase with the signal developed by the V7-V8 plate tanks.

The plate tanks of the horizontal and vertical modulators do not respond to the video and alternator frequencies since these frequencies are below the resonant frequency (200 KC) of the tanks. Thus, the signal fed from the plate tanks of the modulators to the defectors, V9 and V10, is a modulated radio frequency. Since each of the detectors combined in V9 and V10 functions the same, only the detector circuit of V9A will be discussed.

Note that the alternator 15-CPS input to the horizontal modulators, V5 and V6, from terminals 1 and 2 (fig. 10-16) will be negative toward

CATHODE-RAY
TUBE INDICATOR

+275 VOLTS

V15

POS. HV SUPPLY

CAP

10

7   8

11

NEG. HV SUPPLY

+275 V

NEG. HV SUPPLY   NEG. HV SUPPLY

+275 VOLTS

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109.11

Figure 10-15.—Azimuth indicator (horizontal and vertical modulators and amplifiers).

one of the grids of the horizontal modulators (fig. 10-15) while positive toward the other grid. In this manner, the grids of V5 and V6 are always fed 180° out of phase as a result of the alternator input.

It should be recalled that the 200-KC AM-signal from V4B (fig. 10-14) is fed in phase (at the suppressor grids) to each of the modulator stages, V5 through V8 (fig. 10-15). Thus, the modulator receives both the video and alternator signals, as discussed. (Video as used in this discussion refers to the component of the signal voltage, which is received at the RDF antenna, and not to the signal frequency.)

Assume an instantaneous input polarity from the alternator to the V5 and V6 grids, as shown. Due to the more positive V5 grid, V5 will produce a greater driving signal to the V5 plate tank than will be produced for the V6 plate tank. As a result of this action, a greater magnitude signal voltage will be applied to the V9A horizontal rectifier through C45 than will be fed to the V9B rectifier through C47.

If an unmodulated 200-KC signal (containing no video or alternator voltage) were applied to V9A and V9B, both diode sections would produce equal voltages negative to ground across C50 and C53, respectively. Since V9A and V9B will produce equal amplitude outputs negative to ground for an unmodulated 200-KC input, the algebraic sum of the voltages will be zero.

When the 200-KC modulated signals from V5 and V6 are developed across the plate output tanks, one output will be of greater magnitude than the other due to the positive alternator input at one control grid (the grid of V5 in this case), and a more negative input to the V6 control grid. Thus, the negative voltage to ground across C50 will be of greater magnitude than the negative voltage to ground across C53. The detector output voltage at any instant is the algebraic sum of the C50 and C53 voltages.

Note that the negative voltage to the control grid of horizontal amplifier, V11B becomes more negative, while the voltage to the V11A control grid becomes less negative. Thus, the two grids of the horizontal amplifiers, V11B and V11A, are fed 180° out of phase.

The video signal applied from the vertical detector, V10A, to the vertical amplifier, V12B (in the same manner as described for V9A), is 90° out of phase with the signal from V9A to V11B. The 90° phase difference in the vertical and horizontal signals must exist, and must be preserved, in order to produce the various indicator patterns that will be discussed later. The 90° phase difference in the vertical and horizontal signals was introduced as a result of the alternator voltage input at the horizontal and vertical modulators, V5 through V8, as discussed.

The horizontal and vertical amplifiers, V11 through V14 (fig. 10-15) increase the video signal amplitude, which is to be applied to the respective deflection plates of the cathode-ray tube in the indicator. Direct-coupled amplifiers are employed in the horizontal and vertical amplifiers, V11 through V14, to eliminate the need for coupling capacitors.

In addition to the phase-shifting action just discussed, it should be noted that if a capacitor were used to couple the video signal from one of the amplifier stages to the other, the charge time of the coupling capacitor would not only delay presentation of the pattern, but would also cause the indicator pattern to move either upward or downward on the screen as the capacitor is charged and discharged by the incoming video signal. Also, the size of the capacitor would have to be relatively large in order to offer low reactance to the various components of the video signal.

Horizontal amplifiers V11 and V13 are series connected between the -150- and 275-volt supplies. The cathode of V11B is connected to the -150-volt supply and maintained at -80 volts. The potential is derived by the conduction of V11B from the -150-volt source, to the +75-volt supply. The grid potential of V11B is held at -55 volts. This potential is obtained from a separate voltage divider (comprising R64 and R63), which is also connected from the 150-volt supply to ground. Thus, during quiescence, the grid potential (-55 volts) of V11B is positive with respect to the cathode (-80 volts), and grid current flows.

The positive-going video signal from V9A increases the current through V11B and causes the V11B plate voltage to decrease. Note that at any given time, the grid input signal to the V11A and V13A sections of the horizontal amplifiers is 180° out of phase with the signal applied to the V11B and V13B amplifiers. Thus, the V13A output to the pin 7 horizontal plate of V15 is less positive during the same period that V13B is supplying a more positive potential to the pin 8 horizontal plate of V15. The simultaneous application of the video signal to both horizontal deflection plates of V15 increases

the control of the horizontal plates on the electron stream.

The vertical amplifiers, V12 and V14, function the same as the horizontal amplifiers, V11 and V13. The output of V14 is a video signal to the vertical plates (pins 10 and 11) of V15, which is 90° out of phase with the video applied to the horizontal deflection plates (pins 7 and 8), as discussed. The action of the horizontal and vertical voltages at the cathode-ray tube, V15, to produce the various indicator patterns is discussed below.

CIRCULAR PATTERN

If no video input is applied to the first modulator (V2 of figure 10-14) from the video differential amplifier, V1, V2 will conduct continuously, and only the 200-kc oscillator output of V3 will be fed through V2 and V4 to the modulators (V5 through V8 of figure 10-15). In the modulator stages, V5 through V8, the 200-kc radio frequency generated by V3 (fig. 10-14) will be modulated by the vertical and horizontal 15-cycle alternator inputs (fig. 10-15), which are 90° out of phase. The modulators, V5 through V8, conduct continuously, and the resulting output to the vertical and horizontal deflection plates of V15 (after passing through the detector and amplifier stages) produces vertical and horizontal sine wave voltages, respectively, which are 90° out of phase. Thus, the circular pattern shown in figure 10-18 A, is produced, representing zero video signal input at the antenna.

PROPELLER-SHAPED PATTERN

Signals received at the antenna are 15 cps modulated by the antenna rotation. The modulation introduced on the incoming signal is effective in controlling the output of the video differential amplifier, V1 (fig. 10-14), as discussed earlier.

The video signal at the V1B grid is negative going. When the V1B input is maximum negative (corresponding to maximum received signal at the antenna) the effect is to increase the V1A plate current through R6, which cuts off the first modulator stage, V2, and no 200-kc carrier from V3 can pass to the modulators, V5 through V8 (fig. 10-15). When the V1B input is minimum (corresponding to the nulls of the received signal), the V1A plate current through R6 is decreased and the first modulator, V2, is

conducting. A video signal will be passed through the modulators, detector, and amplifiers of figure 10-15 only when the first modulator, V2, is conducting, or when the nulls of the received signals are being applied to the V1B grid.

The period of conduction of the first modulator, V2, is small as compared to the total duration of a single pulse of the video signal. Thus, the signal applied to the vertical and horizontal deflection plates of V15 is a pulse, which corresponds in time to the reception of the nulls of the received signal.

The phase relationship of the vertical and horizontal pulses is always 90°, as discussed. If the vertical and horizontal voltages were complete sine waves, a circular pattern would be produced for the 90° phase relationship. However, since the vertical and horizontal inputs are pulses, the pattern displayed on the indicator (V15) is in the form of a propeller (fig. 10-18E) the tips of which point in the direction of the nulls of the received video. An azimuth scale which encircles the cathode-ray tube screen permits the operator to read, in degrees, the position of the two nulls of the received signal.
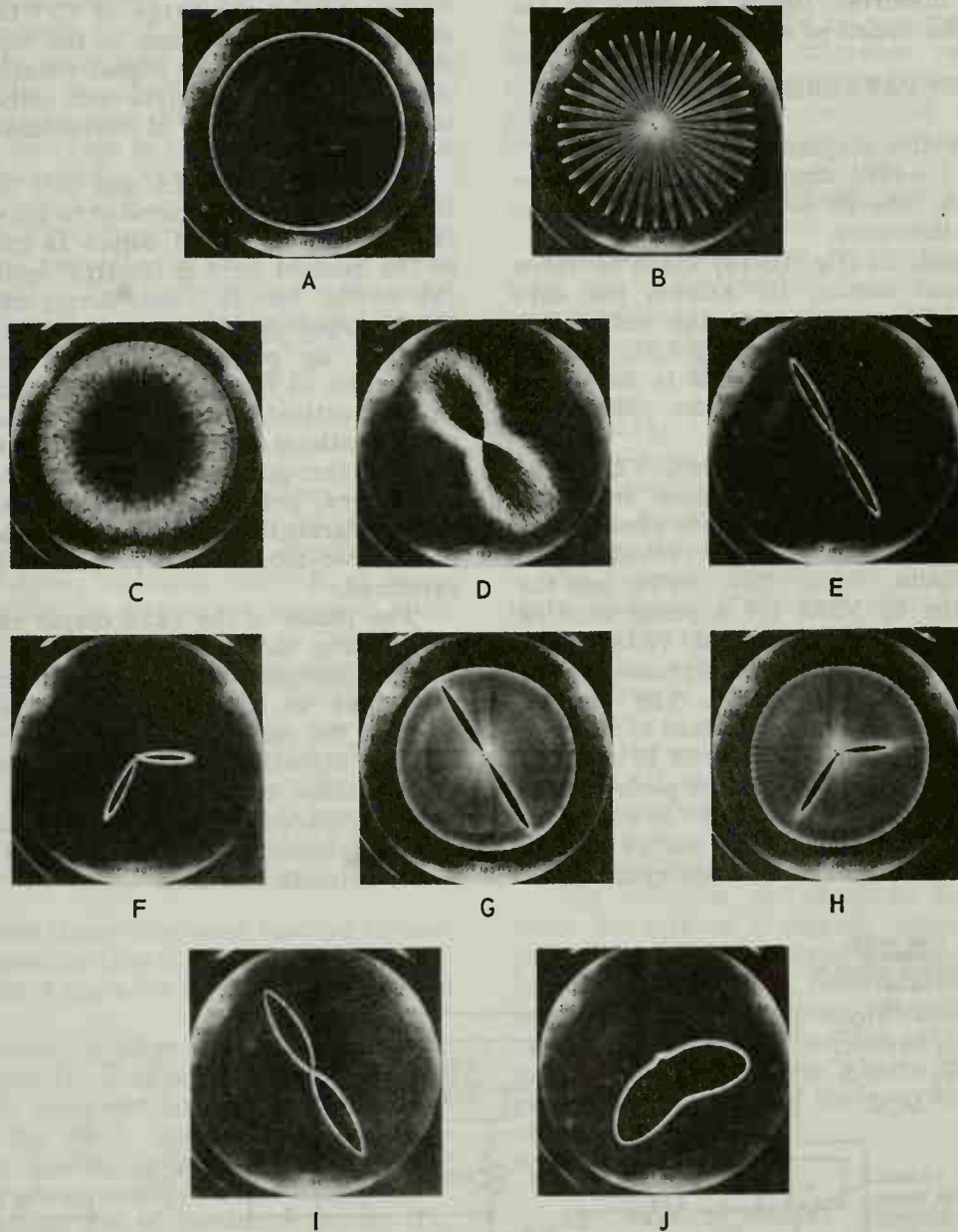
SENSE PATTERN

The circuit of the antenna in D/F operation is as shown in figure 10-8A. The two sections of the Adcock array are in exact balance during D/F operation, as discussed.

When the calibrate-sense switch on the front panel of the indicator is held in the SENSE position, a sense relay, K1 (fig. 10-7) energizes, allowing its contacts to alter the antenna electrical connections to that shown in figure 10-8B. The effect is to introduce a phase lag in one section of the array, while simultaneously introducing a phase lead in the other section. This action causes the nulls to appear at a different position in the rotation of the antenna, and at a different position on the indicator screen.

In addition to the action just discussed, when the sense-D/F switch is depressed, sections S1C, S1D, S1E, and S1F or S1 (fig. 10-15) interchange sweep voltages to the balance modulators, V5 through V8. This causes the pattern to be shifted 90° on the indicator screen.

The combined effects of the sense-D/F switch, S1, and the sense relay, K1 (fig. 10-18F) causes the rotation of the pattern by 90° and a shift in the position of the nulls, so that the distorted figure 8 (fig. 10-18F) is formed. The

A. CIRCULAR PATTERN
B. CALIBRATE PATTERN
C. NOISE PATTERN
D. BEARING PATTERN, NO MODULATION, WITH NOISE
E. BEARING PATTERN, NO MODULATION
F. SENSE PATTERN NO MODULATION
G. BEARING-100 PER CENT TONE MODULATION
H. SENSE-100 PER CENT TONE MODULATION
I. BEARING PATTERN - STRONG SIGNAL - r-f GAIN LOW
J. SENSE PATTERN - STRONG SIGNAL - r-t GAIN LOW

109.14

Figure 10-18.—Typical operation patterns.

apex of the distorted figure 8 points in the direction of the received signal.

CALIBRATION PATTERN

The calibration display is a 36-lobe "daisy" pattern (fig. 10-18B) displayed on the cathode-ray tube, V15. The 36-lobe pattern is used to calibrate the indicator.

A tone wheel, E1 (fig. 10-19) which is driven by the antenna motor, B1 Affects the core material of L5A to generate the calibration voltage in the circuit comprising L5A, C5, and the primary of T1. This voltage is developed at the T1 secondary across the calibration gain control, R173.

The calibration phase splitter, V23A, receives the calibration signal input from the arm of R173. The plate and cathode resistances of V23A are of equal value to provide equal amplitude outputs. R175, C77, R174, and the tube resistance of V23A for a phase shifting network for the plate output of V23A. The amount of the phase shift is determined by the position of the arm of R174. The cathode potential of V23A (which is applied to the arm of R174) is degenerative, since it is approximately 180° out of phase with the plate output. The degenerative action is reduced by a smaller regenerative action due to the charge of C77 through R177 on the negative half cycle of the input, and the discharge of C77 through V23A during positive portions of the input. The resulting phase of the signal obtained from the combination of the plate and cathode voltages is applied to the grid of conventional amplifier, V23B.

When sections S1A and S1B of the sense-D/F cal switch are operated to the CAL position (as shown), the V23B output is coupled by C79 to the control grid of the first modulator stage (V2 of fig. 10-14). Thus, during calibration the V23B output causes the first modulator, V2, to produce an output pulse 36 times for each revolution of the antenna. The resulting signal to the vertical and horizontal deflection plates of the cathode-ray tube, V15 (fig. 10-15), after passing through the modulators, detector, and amplifiers, produces 36 lobes on the indicator screen during the same period normally required for the two-lobe (single-propeller) pattern to be produced.

The phase of the V23B output can be shifted by rotating the cal-phase control, R174. This action simultanously changes the position of all the lobes of the calibrate pattern. In this manner, the calibration necessary for determining the direction of a received signal can be accomplished by moving the tips of the 36-lobe pattern (spaced every 10 degrees) until 4 of the tips correspond with 0, 90, 180, and 270 degrees on the azimuth indicator.
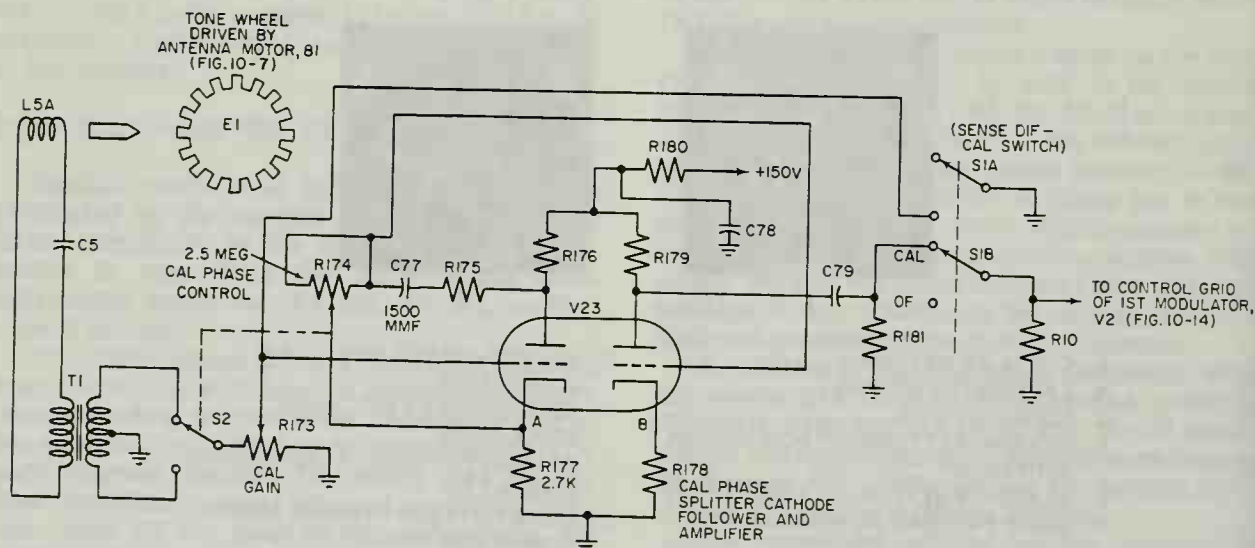

Figure 10-19.—Calibration voltage input.

109.15

The cal-gain control, R173 is used to set
e output voltage level at V23B. This action
ll adjust the amplitude of the signal pulses
the deflection plates of V15. The control is
operly adjusted when the spokes of the 36-
be pattern converge in a spot near the center
the cathode-ray tube screen.

When the sense-D/F-cal switch, S1 is placed
the D/F position, the grid of V23A is directly
rounded. This prevents the amplification of the
librate signal in V23, and eliminates the pos-
bility of allowing an amplified calibrate signal
be capacitively coupled to the CRT deflection
ates through the open contacts of S1.

LORAN

The principle of loran navigation involves
stablishing a fix at the intersection of two or
ore "lines of position" as discussed in
asic Electronics, NavPers 10087-B. A loran
ne of position is a line of constant time
ifference between the arrival of two pulse
odulated RF signals transmitted from a pair
f loran transmitting stations (master and
lave transmitters). The stations are usually
cated several hundred miles apart. Because
F energy travels at a constant velocity, the
me difference in reception of the two signals
a measure of the difference in the distance
om the transmitting stations to the point of
bservation.

An obtained time-difference reading is used
locate a specific line of position on a special
hart. The chart contains various loran lines of
osition.

If a second time difference is obtained from
second loran transmitting pair, the new time
ifference is used to locate a second line of
osition on the chart. The intersection of the
wo lines of position obtained from the time
ifference readings establishes a "fix."

The receiving set of the loran system (fig.
0-20) is basically an instrument for measuring
he very small periods of time that elapse
etween the arrival of signals from the two
tations of the loran transmitting pair. To
ccomplish this, a variable timed interval is
enerated within the receiving set and displayed
n the screen of a cathode-ray tube. When the
eceived signals are positioned at the beginning
nd end of the timed period, the time difference
etween arrival of the signals at the location
f the receiving set may be read from a
evolution type time-difference counter.

A second reading obtained in the same
manner from a different transmitting pair pro-
duces a second time-difference reading on the
counter. The first and second time-difference
readings are used to determine geographical
location from loran charts and tables (not shown).

LORAN RECEIVING SET AN/UPN-15

Originally designed for loran A operation
only, many AN/UPN-12( ) receiving sets have
been modified to accommodate both loran A and
loran C signals. Modification is accomplished
by adding a small receiver-control unit and
associated components to the existing
AN/UPN-12( ) set (fig. 10-20). When so mod-
ified, the nomenclature of the receiving set is
changed from AN/UPN-12( ) to AN/UPN-15( ).

When functioning as a loran C receiver-
indicator, the set utilizes the signals received
by the receiver-control unit mounted atop the
main chassis. This unit contains a 100-KC
radio receiver of the tuned radio frequency
type. The controls that affect its operation
as a loran C receiver are on the front panel
of the unit.

With the equipment set for loran A operation,
the 100-KC receiver is isolated from the set
and the four-channel superheterodyne receiver
in the main chassis is used to receive the
loran signals.

The indicator unit of the set displays either
loran A or loran C signals. When the received
pulses are aligned as prescribed for the particu-
lar mode of operation, time difference readings
are taken from a counter. By taking a second
reading from a different set of loran stations
and referring to loran charts and tables the
geographic position of the ship is determined.

ANTENNA COUPLER

The antenna coupler (fig. 10-21) permits
matching of the antenna impedance to the input
impedance of the receiver. The antenna coupler
consists of a 17-position rotary switch, a
tapped inductor, a neon lamp, and a spark
gap.

When the rotary switch, is adjusted for
maximum signals on one of the four channels,
the antenna and antenna lead-in reactance, and
the section of the inductor in use, form a
tuned circuit, which is broadly resonant over
the band of frequencies that the equipment is
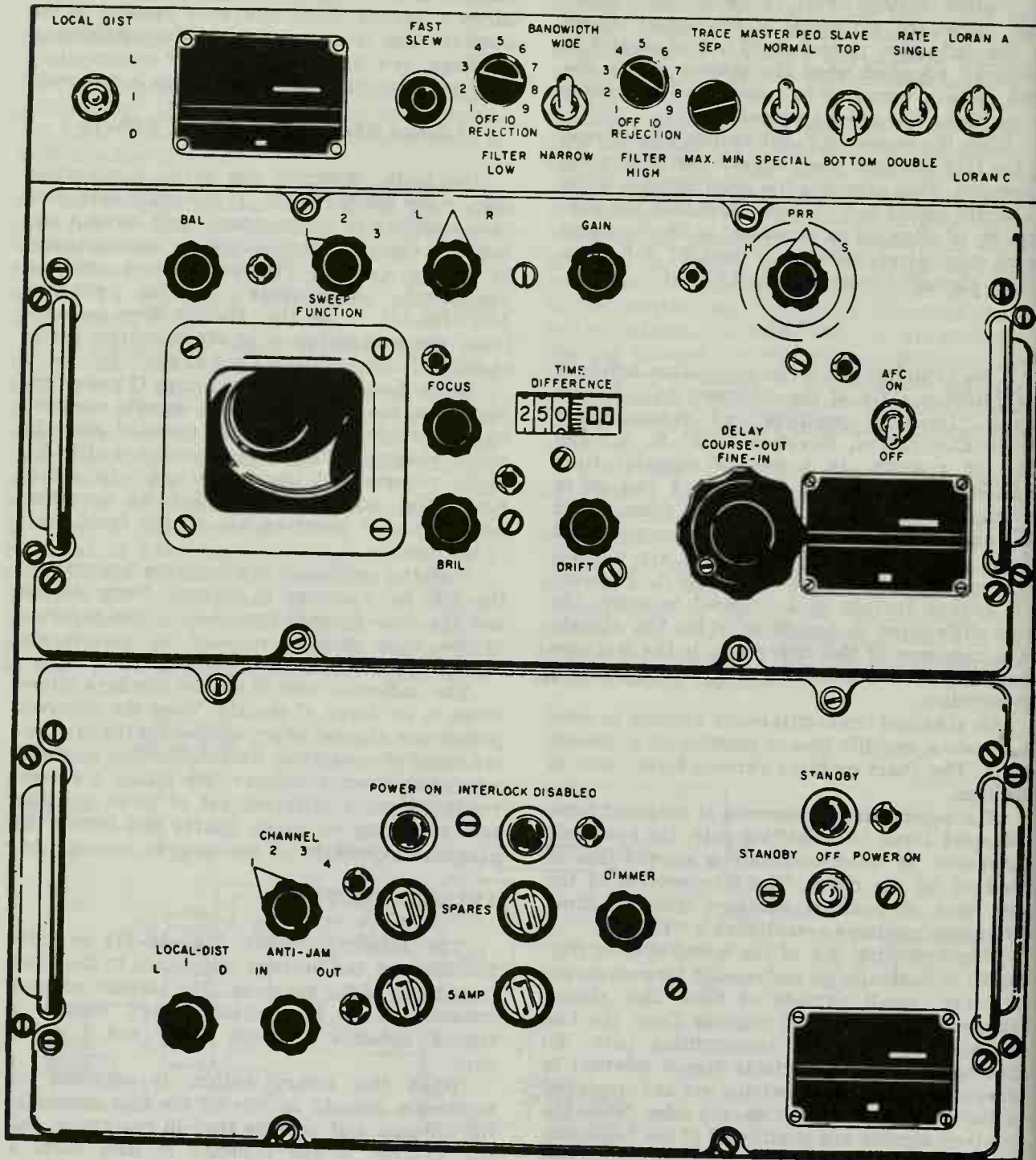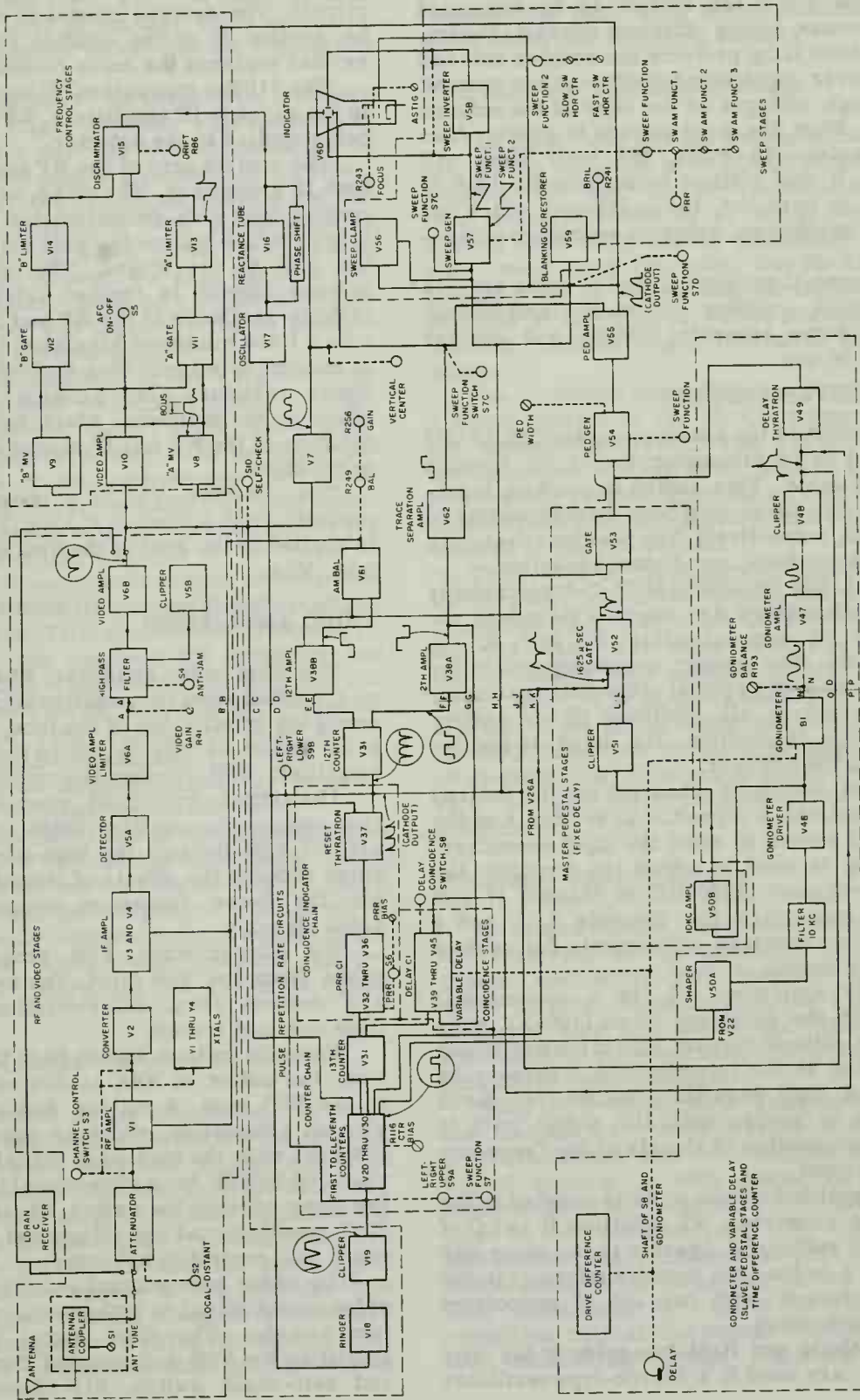designed to receive. The multiple taps on the

162.34

Figure 10-20.—Loran receiving set, AN/UPN-15.

70.97

Figure 10-21.—Loran receiving set AN/UPN-15, block diagram.

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inductance allow the circuit to be resonated with antennas having different characteristics.

The neon lamp protects the coupler coil and the receiver input components against burnout due to high voltages at the input of the antenna coupler. When the input voltage is greater than approximately 60 volts, the neon lamp will conduct. If the voltage is sufficiently high to cause it to burn out, the spark gap, which has a higher breakdown, provides secondary protection.

The loran C receiver uses the same antenna and antenna coupler on most installations although some separate antenna and coupler units are in use.

R-F RECEIVER CIRCUITS

The output of the antenna coupler (fig. 10-21) is applied to an attenuator switch, S2, through a coaxial cable. This switch is marked, local-dist, and has three positions corresponding to L, I, and D, respectively. The positions designate local, intermediate, and distant reception.

In the L and I positions of S2, T-type resistor network attenuators are used to introduce attenuations of approximately 20 and 40 db, respectively. The proper impedance match for the interconnecting coaxial cable and the primary is maintained in all positions of the switch, S2. In the D position, the input is applied directly to the primary.

Channel switch S3 when set for the desired channel, selects a capacitor in section A of the switch (not shown) to shunt the input secondary. Slug tuning is used to adjust the circuit to the highest frequency (1950 KC of channel 1), and trimmer capacitors for channels 2, 3, and 4 provide the necessary adjustments for each of the other channels.

The RF amplifier, V1, is a conventional circuit with the exception of an 1100-KC wave trap in the cathode circuit. This circuit provides degeneration at the intermediate frequency by inserting a high parallel resonant impedance effectively in series with the signal path to increase attenuation of signals at the frequency in the V1 output.

The amplified output of V1 is coupled to the grid of the converter, V2. Sections B and C of the channel switch, S3, operate in the same way as section A to tune the input transformer to the selected channel with the aid of associated trimmer capacitors.

The cathode and first two grids of the converter, V2, are used in a Pierce-type oscillator circuit. The desired crystal is connected to V2 by section D of the channel switch, S3. The crystal replaces the more common tank circuit.

The 1100kc converter (V2) output is amplified in the cascade IF amplifiers, V3 and V4, before being applied to the detector.

The control grid of the RF amplifier, V1, and both IF amplifiers, V3 and V4, are returned to a control bias gain potentiometer in the balance circuit which varies the grid bias to determine the gain of the stages. The operation of the balance circuit is treated later. Unbypassed cathode resistors in the RF and IF stages provide a fixed minimum bias when the gain control is maximum (no grid bias applied to the stages). These resistors also produce a degenerative effect at the cathodes, which tends to prevent detuning of the RF and IF stages due to changes in grid bias.

The detector, V5A is a conventional diode rectifier with a shunt R C load. The V5A output is applied to the grid of the first video amplifier tube, V6A.

VIDEO AMPLIFIERS

The first video amplifier, V6A, also serves as a limiter when the negative input voltages exceeds the cutoff value of the tube. The positive-going output of V6A is fed to the second video amplifier, V6B, through the antijam switch, S4.

The antijam switch, S4, when in the IN position couples the positive output of V6A to the grid of V6B through the high-pass filter. This filter reduces the effects of jamming by removing the lower frequency components of the video signal.

The clipper stage, V5B, removes negative peaks from the V6B input. The negative pulses are derived from differentiation in a short time constant circuit.

When the antijam switch is in the OUT position, the output of V6A is coupled through S4 to the V6B grid. A voltage divider inserted to give an attenuation, which is equal to the loss obtained when the high-pass filter is used. Thus, the input to V6B for a given signal level remains the same whether the switch, S4, is in the IN or OUT position, and no adjustment of the front panel gain control is necessary.

The video gain control sets the level of the video output signal in both positions of the anti-jam switch. The video signal from V6A is amplified by V6B and coupled through a relay and self-check switch, S10 (discussed later),

o the video amplifier, V7. The video is also applied to the grid of another video amplifier (V10) for a purpose which will be considered later.

Video amplifier V7 is essentially a clipper and amplifier stage. Detector V5B operates to remove negative portions of the video output of V6A. Some of the negative portions of the video, which are generated by differentiation at the V6B grid will be amplified in V6B. These negative voltages at the V6B plate, cause the V7 grid to draw a current through the grid limiting resistor. This action develops a negative voltage at the V7 grid, and therefore limits positive alternations of the grid input.

Negative alternations at the V7 grid (from V6B) are amplified by V7, and appear as positive-going pulses at the V7 plate. These positive video pulses are coupled to the upper deflection plate of the cathode-ray tube, V60.

When loran C is selected, a relay connects the common antenna and coupler to the loran C receiver (fig. 10-21).

The loran C receiver is simple in design and operation, it differs from conventional receivers in that it is a TRF (tuned radio frequency). Consequently, the received RF frequency and the intermediate frequency (100 KC) are the same, and undetected intermediate frequency is applied as video through a relay to V7 and V10 the indicating circuits.

The relay to select loran A or loran C also disables the receiver not in use.

The circuits described in the following pages pertain to loran A counting and timing. Loran C is the same but some of the circuits as shown have been modified to present the loran C signals correctly. Several units of amplification and dividers along with multiple ganged switches and relays are not shown on figure 10-21.

TIMER AND INDICATOR

To this point in the discussion, the loran signal input has been selected (by the antenna, RF amplifiers and converter stages), amplified (in the IF stages), detected (by video detector), and fed through stages of video amplification and clipping to the upper vertical deflection plate of the CRT. To hold the loran signal stationary on the screen of the indicator tube while readings are being taken, it is necessary to employ frequency control circuits to synchronize the repetition rate of the loran

receiving set with the loran signals from the transmitting pair. This action is accomplished in the timer and indicator chassis.

Oscillator and Reactance Circuits

The time base for the synchronization system in the receiver is an 80-KC sine wave oscillator, V17 (fig. 10-22). The oscillator, V17, is crystal controlled by Y5 (at the normal frequency of 80 KC), and is essentially a tuned-plate, tuned-grid oscillator. The 80-KC crystal is connected in series with parasitic suppressor R94 at the grid of V17 to prevent oscillations at frequencies not controlled by the oscillator.

The reactance tube, V16 receives at its grid a small portion of the 80-KC voltage appearing across the crystal. The V16 grid is held negative with respect to the cathode by a negative voltage received from the AFC circuits (discussed later) supplemented by cathode bias developed at the V16 cathode.

The value of the apparent capacity across the crystal, Y5, depends upon the gain of the reactance tube, V16, which in turn depends upon the V16 bias. The value of bias is controlled (to an extent) by a negative voltage from the AFC circuit. The range of frequency control as a result of the action of the reactance tube, V16, is small (less than ± 30 cps), but is sufficient to compensate for differences in pulse repetition rate between the transmitting and receiving systems.

Frequency Control Circuits

As stated earlier, the basic timing system of the receiving set is controlled by the 80-KC crystal oscillator, V17. However, to hold the received pulse stationary on the screen of the indicator tube, the timing base must be synchronized to an exact multiple of the reception rate of the received signals.

Both AFC (Automatic Frequency Control) and manual drift circuits are available for use in synchronizing the receiving set with the received signals. In both cases, the control is accomplished by variation of the grid bias applied to the reactance tube, V16. The AFC circuit automatically changes the grid bias of V16, which changes the constants of the oscillator, V17, sufficiently to lock in the receiving set repetition rate at the proper relation to that of the transmitting group. Adjusting the drift
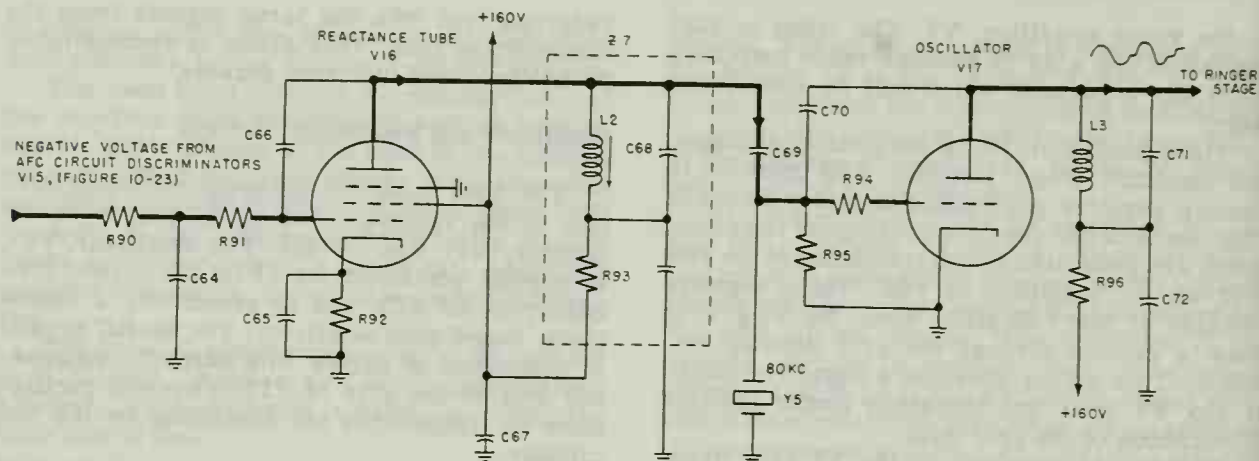
Figure 10-22.—Frequency control circuits (oscillator and reactance tube).

109.17

control on the receiver front panel will manually set the grid bias to control the reactance tube, V16.

The automatic frequency control circuits are shown in figure 10-23. V8 and V9 are one-shot multivibrators referred to as the A and B multivibrators, respectively. The one-shot cycle is described below. In the quiescent condition, V8B conducts, and V8A is cut off. The positive pulse on the V8A grid causes V8A to conduct. The resulting decrease in V8A plate voltage lowers the V8B grid below cutoff. The V8B plate voltage rises sharply as V8B plate current is cut off. The V8B grid rises above cutoff in approximately 80 μs, at which time V8B again conducts. The accompanying sudden increase in voltage cuts off V8A, and the one-shot cycle is completed.

The A multivibrator, V8, receives a positive rectangular trigger pulse from the pedestal amplifier. (The origin and purpose of the pedestal pulses are treated later.) The control grid of V8B is returned to the +160-volt supply, and is therefore the normally conducting section of V8. The positive rectangular output of the pedestal circuit is coupled through isolating resistor R46 and a differentiator circuit. The positive pulse resulting from the differentiation of the input triggers V8A into conduction to initiate the A multivibrator cycle.

The A multivibrator, V8, furnishes a positive rectangular pulse from V8B, to the suppressor grid of the A gated amplifier tube, V11, and a negative rectangular pulse from V8A, to the B multivibrator, V9. The duration of the output

of each section of V8 is approximately 80 μs The pulse furnished to the grid of the B multivibrator, V9, is differentiated to form a sharp negative pulse. This pulse occurs at the same time as the leading edge of the input pulse to V8A. The operation of the B multivibrator, V9, is identical to that for the A multivibrator, V8.

Since a positive pulse from V8A is required to trigger V9A, the V9A conduction occurs 80 μs later (in time) than V8A when the V8A plate output (shown in dotted lines) is positive going. The positive rectangular output of V9B is applied (80 μs later than the pulse from V8B) to the suppressor grid of the B gated amplifier, V12, and is also approximately 80 μs in duration.

The A and B gated amplifiers, V11 and V12, respectively, are normally held beyond cut off by the negative bias (from the -180-volt supply) applied to the suppressor grids of V11 and V12. The suppressor grid of V11 is connected to the -108-volt supply.

Before the A multivibrator, V8, is triggered (at V8A) the plate voltage at V8B is approximately 45 volts, and the voltage divider comprising R50 and R63 applies about -36 volts to the suppressor grid of V11.

When V8A receives its positive trigger pulse, this tube section conducts; and the accompanying drop in the V8A plate potential is fed through C52 to the grid of V8B. V8B is cut off as a result of the negative-going grid input from V8A, and the V8B plate voltage rises toward the +160-volt supply potential. Likewise, the suppressor grid potential at the V11 grid (from the
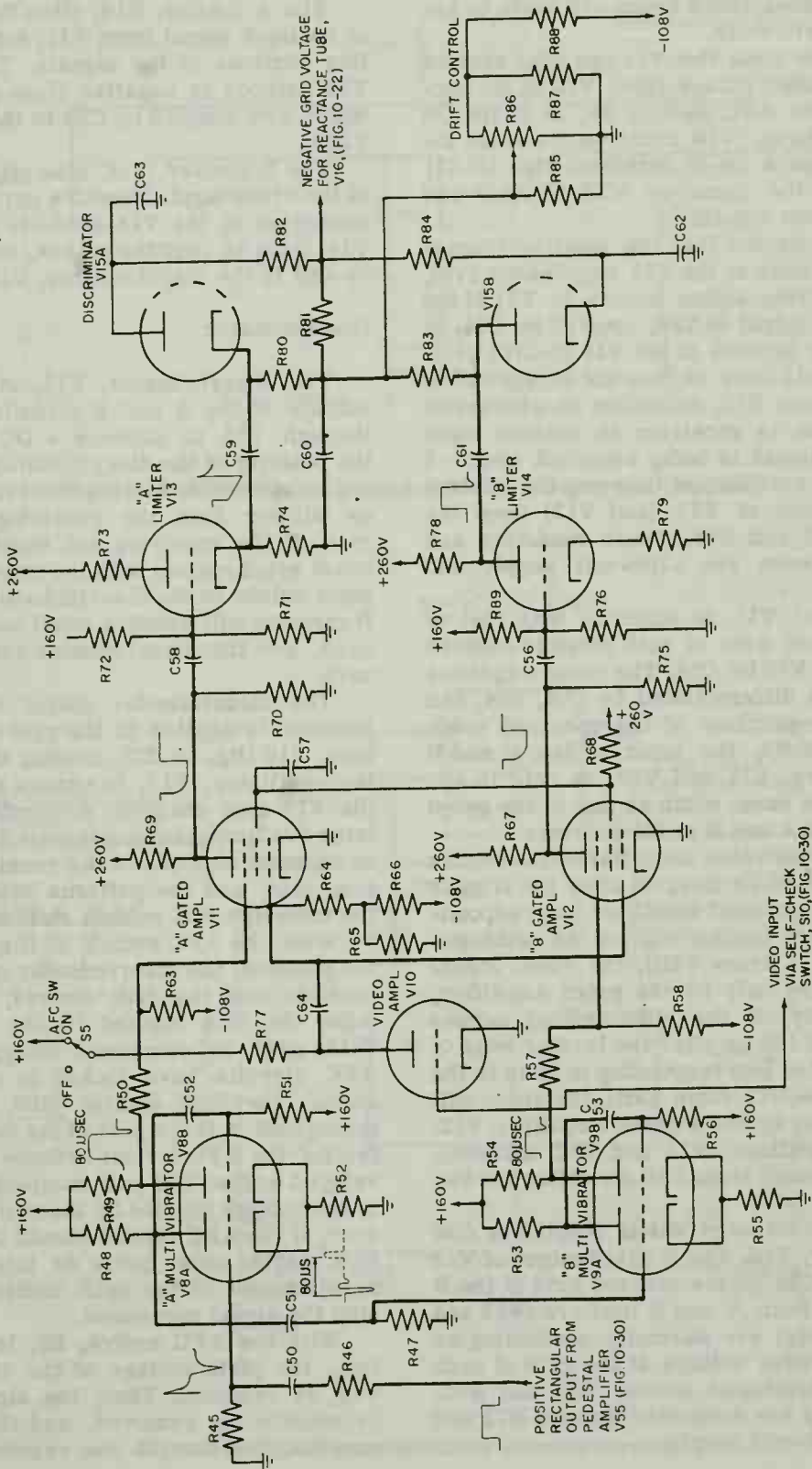
358

Figure 10-23.—Frequency control circuits (A and B gating circuits and discriminator).

109.18

359

R50-R63 junction) rises from -36 volts to approximately zero volts.

Note at this time that V11 can also receive the positive video pulses from V10 at its control grid if the AFC switch, S5, is in the ON position, as shown. V10 receives its video input from Loran A or C receiver (fig. 10-21) and functions the same as V7B to limit and amplify the video signals.

During the period that the positive trigger from V8B prevails at the V11 suppressor grid, a coincident firing action occurs in V11 if the positive video output of V10, coupled by C64, is simultaneously applied at the V11 control grid.

Video signals may or may not be applied to the V11 grid from V10, depending on whether or not the system is receiving an antenna input signal. If no signal is being received, about -7 volts bias is established between the control grid and cathode of V11 (and V12) from the junction of R65 and R66. These resistors are connected between the -108-volt supply and ground.

Operation of V11 is identical with that of V12. The control grid of both tubes is coupled to the plate of V10 by C64. The video amplifier (V10) output is differentiated by C64, R64, and R65 so that, regardless of the received video signal pulse width, the input of the A and B gated amplifiers, V11 and V12, is held to approximately the same width as that of the gated pulses from the A and B multivibrators.

When the (V10) video amplifier output occurs during the first 80-μs interval after the trigger pulse from the pedestal amplifier (corresponding in time to the leading edge of the rectangular pulse to V11 from V8B), the video signal will be amplified only by the gated amplifier, V11. Conversely, if the video signal occurs between 80 and 160 μs after the leading edge of the pedestal pulse (corresponding in time to the trigger pulse to V12 from V9B), the video will be amplified only by the B gated amplifier, V12. Both gated amplifiers, V11 and V12, will conduct when the video from V10 straddles the two gated periods.

The A gated output of V11 is coupled by C58 to the A limiter, V13. The B gated output of V12 is coupled by C56 to the control grid of the B limiter, V14. Both A and B limiters (V13 and V14, respectively) are normally conducting as a result of positive voltage at the grid of each of the tubes developed across R71 and R76. These resistors are connected through R72 and R89, to the +160-volt supply.

The A limiter, V13, clips the positive peaks of its input signal from V11, but amplifies negative portions of the signals. The output at the V13 cathode is negative (less positive) pulses, which are coupled by C59 to the discriminator, V15A.

The B limiter, V14, also clips positive peaks of the video input. Negative portions of the video appearing at the V14 grid are developed at the V14 plate as positive pulses, which are coupled by C61 to the discriminator, V15B.

Discriminator

The discriminator, V15, utilizes the gated outputs of the A and B circuits comprising V8 through V14 to produce a DC output voltage, the polarity of the discriminator output depending upon whether the received signals are faster or slower than the receiving set repetition rate. If the receiving set repetition rate is in exact synchronism with the received signal, the input pulses to the discriminator from the A and B circuits will produce equal and opposite voltages, and the discriminator resultant output is zero.

The discriminator output at the R81-R82 junction is applied to the grid of the reactance tube, V16 (fig. 10-22), causing the V16 output to the oscillator, V17, to change as the voltage of the V17 grid changes. Accordingly, the oscillator (V17) output changes until it is synchronized to an exact multiple of the receiving set repetition rate, and the patterns being observed on the indicator will remain stationary.

When the AFC switch, S5 (fig. 10-23) is in the ON position, the discriminator operates in conjunction with the drift control, R86, to determine the bias applied to the reactance tube (V16) grid, as discussed. Normally, when the AFC circuits have locked in on the received signal, the drift control, R86, may be varied throughout a large part of its range without affecting the AFC circuit because the voltage developed across R81 will compensate for changes in the voltage applied by the drift control. However, if the AFC circuit should lose control due to a sudden noise pulse or interfering signal, readjustment of the drift control setting would stop the signal movement.

With the AFC switch, S5, in the OFF position, the plate voltage of the video amplifier, V10, is removed. Thus, the signal to the discriminator is removed, and the drift control governs, the bias on the reactance tube, V16.
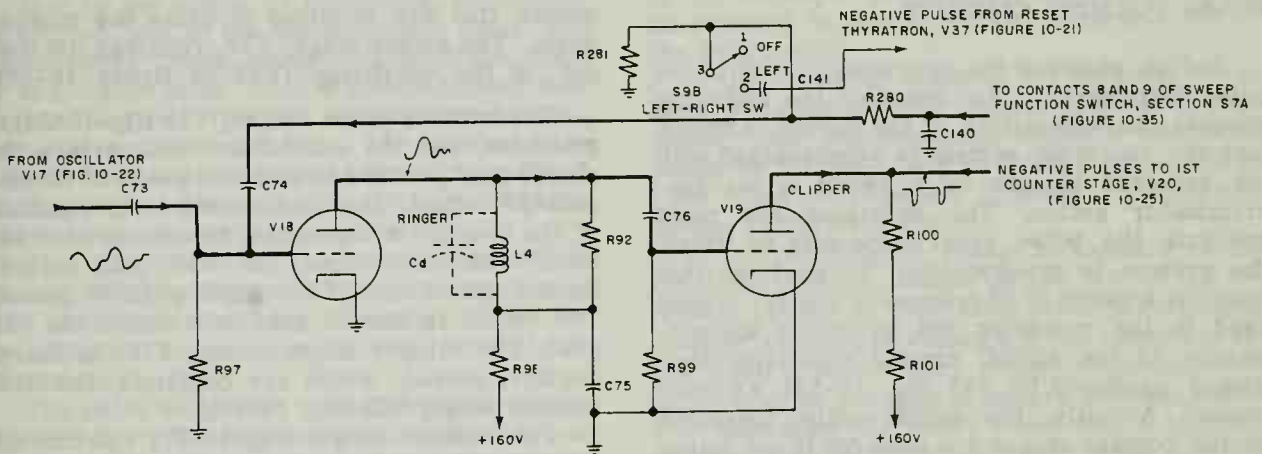
109.19

Figure 10-24.—Pulse-forming circuits (ringer and clipper).
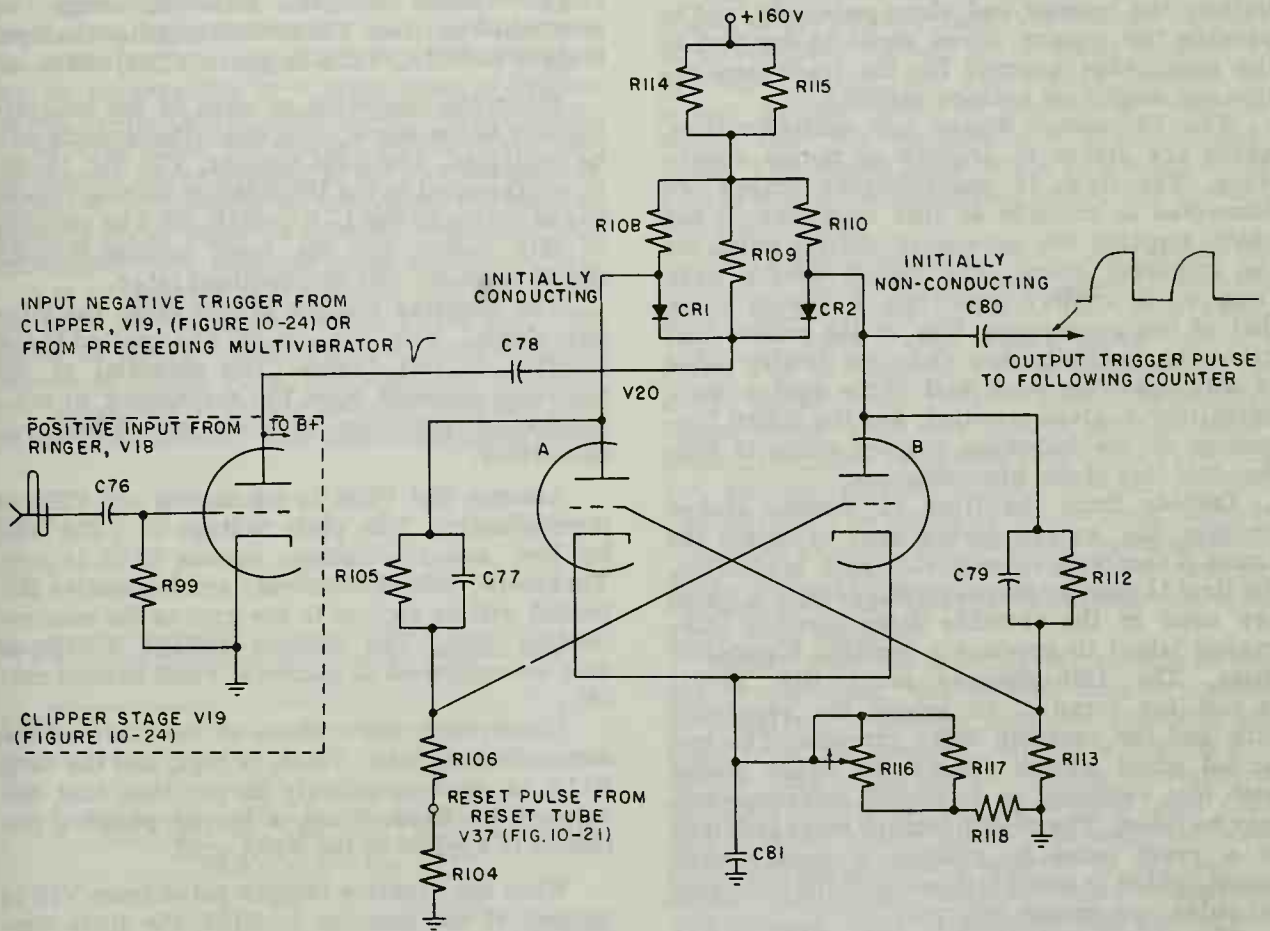
109.20

Figure 10-25.—Representative counter circuit.

PULSE FORMING CIRCUITS

Before studying the operating details of the pulse forming circuits, consider the following general characteristics of the circuit. Assume that the receiving system is synchronized with the received signals as a result of the discriminator action. The receiving set must generate the pulse recurrence rate to which the system is synchronized. To perform this function a group of 13 frequency counter stages used in the receiving set to obtain subharmonics of the 80-KC crystal oscillator frequency produced by V17 (fig. 10-22), as discussed. Actually, the subharmonics produced by the counter stages are used for three major purposes: to obtain the receiving set pulse repetition rate as selected by the PRR (Pulse Repetition Rate) switch on the receiver front panel, to produce the trigger pulses used to initiate the master and slave pedestals, and to provide two square waves equal in duration to the recurrence interval for the trace separation and amplitude balance circuits.

The 13 counter stages are multivibrators, which are driven to produce an output square wave. The first 11 multivibrator stages are connected in cascade so that each stage in the chain supplies the necessary driving pulse for the following stage. The output square wave from each multivibrator has a period twice that of the recurrence rate of the pulses that drive the multivibrator. Only one driving pulse is obtained from each half of the square-wave cycle (for a given counter), and the output frequency of the following counter stage is thus one-half that of the preceding one.

Outputs from the first 11 counter stages produce the square waves used to obtain the Loran A receiving set repetition rate. In addition, the first 11 counter stages produce outputs, which are used in the variable delay circuits (discussed later) to produce a variable B pedestal pulse. The 13th counter stage (fig. 10-21) is use for loran C to extend the repetition rate and the variable delay circuits. The extended count switch affects the count period such that readings up to 50,000 microseconds may be taken. The twelfth counter stage is driven by a reset pulse to produce a square wave output, which is equal in duration to the receiving set pulse recurrence interval.

The ringer and clipper stages, V18 and V19 (fig. 10-24) are included in the pulse-forming circuits to produce the sharp, negative driving pulses that are required to drive the counter chain. The ringer stage, V18, receives the output of the oscillator (V17 of figure 10-22).

The cathode of the ringer, V18 (fig. 10-24) is grounded, and the oscillator output drives the ringer grid positive on positive peaks of the oscillator output. Due to the long time constant of the coupling components, as compared to the 80-kc oscillator input, the V18 grid is held beyond cutoff except on input positive peaks. The action is that of grid-leak bias at the V18 grid. The clipper stage output, V19, is sharp, negative pulses, which are coupled to the first counter stage, V20 (fig. 10-25).

The counter stages (fig. 10-21) V20 through V31 (V951 for Loran C) are flip-flop (Eclles-Jordan) multivibrators. The cascade connection of the 13 multivibrators allows the output pulse of one counter stage to serve as the input trigger pulse for the following stage, as mentioned earlier. Thus, V20B supplies the input trigger to V21A. V21B triggers V22A, and so on.

Since the operation of each of the multivibrators is the same, only one of the circuits will be explained. The first counter, V20 (fig. 10-25) is represented in the illustration without regard to the action of the L-R switch, S9. The purpose of this switch and the input applied through the S9 contacts will be explained later.

The negative trigger pulses from the clipper stage, V19 (fig. 10-24) are applied to the junction of two diodes. The potential at the junctions depends upon the conducting or nonconducting condition of V20A and V20B, respectively.

Assume that V20A is conducting and V20B is nonconducting. The plate voltage of V30A will be low, and the voltage across R106 is low. Therefore, only a relatively small positive potential will be applied to the grid or the nonconducting tube. The cathode voltage developed will be sufficient to maintain V20B beyond cutoff.

Conversely, the voltage at the plate of the nonconducting tube, V20B, is high, and the drop R113 is correspondingly larger than that developed by R106. Thus, a larger positive potential is applied to the V20A grid.

When the negative trigger pulse from V19 is applied at the junction of R109, the diode connected to the "off" plate will have the larger voltage difference placed across its terminals and will therefore conduct a heavy current.

Due to the action of the diodes, the trigger is applied only to the nonconducting stage. The regenerative action originating from the effect of the input pulse on the nonconducting tube causes the generation of a square wave having a fast-rise and fast-fall time. The counter stage output (V20 in this case) is taken from the B section of the tube, and applied through 080 to the following counter stage.

The cathodes of all the counter stages (V20 through V30) are returned through the counter bias network to ground. Resistor, R116, can be adjusted to set the counter bias at the proper level.

The pulse repetition rate coincidence stages, V32 through V36A (fig. 10-21), utilize the various outputs of the counter chain to drive a particular pulse repetition rate for the receiving system chosen by the PRR switch on the receiving set front panel. Because each counter stage must be triggered twice by a negative pulse before it produces a trigger pulse for the following stage, the point (in time) at which any tube switches can be represented by a given count of driving pulses applied to the input of the counter chain.

By combining the voltage values of the output square waves present at the various counter plates, it is possible to determine the point at which a given count is reached. The count will correspond to a selected pulse repetition rate.

Let us analyze this concept with the aid of the plate waveforms of V28A, V29A, and V30A shown in figure 10-26. Each of the three plate outputs of V28A, V29A, and V30A will be negative going only at counts between 1792 and 2048 on the scale shown. This coincident occurrence
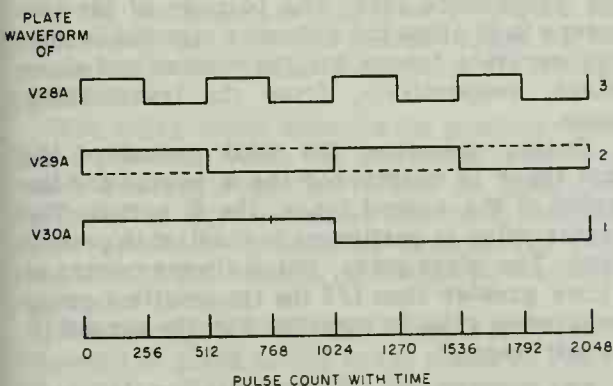
of all negative voltages at the chosen plates correspond in time to a given count selected by the PRR switch.

If the B section output of V29 shown on dotted lines were substituted in place of the A section output, the simultaneous, negative indication would be obtained only at counts from 1270 to 1536. This corresponds in time to a different setting of the PRR switch.

It is possible to obtain a negative output from the counter stages during any of the intervals plotted on the bottom scale (if the proper counter plates are chosen). If the outputs of all the binary plates were properly combined, it would also be possible to secure a unique negative indication for every count from 0 to 2048. Since only 24 different repetition rates are used by the various transmitting groups it is not necessary to use all of the outputs of the first 11 (13 for Loran C) counter stages.

The plates of the third through the tenth stages of the counter chain (V22 through V29) are direct-coupled through an isolating resistor to the PRR switch, S6. The switch selects the proper counter plates to produce the negative output that corresponds to a predetermined count. The chosen plate output is coupled to the grid of one of the amplifying tubes, V32 through V35 (fig. 10-21).

The pulse count (fig. 10-26) represents the number of pulses in time required to produce the effect indicated for either of the tubes V28A, V29A, or V30A. Only counts above 1024 are used. The A plate of V30 is negative for the entire period from 1024 to 2048. The negative pulses from V30A are always used in determining the pulse repetition rate of the receiving set, and are applied to the V36A control grid (fig. 10-21). The plates of V32 through V36A are connected together and returned through a common load resistor to the +160-volt supply. Cathode PRR bias control permits the bias to the tubes to be adjusted.

Recall that the input (grid) voltage of any one of the tubes, V32 to V35, may be positive or negative, depending upon the potential applied from the counter plate to which the grid of the tube is connected through a section of the PRR switch, S6. The V36A input from V30A may also be either positive or negative. When the input pulse from a counter to any section of the coincidence indicator chain comprising V32 through V36A is positive, that section will draw saturation current. When the counter input to a particular section of any one of the tubes

PLATE
WAVEFORM
OF

V28A   3

V29A   2

V30A   1

0   256   512   768   1024 1270 1536 1792 2048

PULSE COUNT WITH TIME

109.21

Figure 10-26.—Negative coincidence indication from combined plate waveforms.

(V32 through V36A) is negative, that section will be held beyond cutoff for the duration of the counter input. V32 through V36A are paralleled, with a common plate load resistor connected to the +160-volt B supply.

Since an output is produced from the coincidence indicator chain during the time that all sections of V32 through V36A are cut off, the width of the output cannot be greater than the width of the negative half of the smallest square wave applied to the chain. The first counter output pulse to the coincidence indicator chain is applied from V22A (or V22B) through S6 to one of the grids of V32. One of the V22 outputs will always be applied to V32, depending on the position of S6. The output frequency of the third counter V22, is 10 kc, and the width of the negative pulse is 50 μs (one-half the total pulse duration). This represents the maximum time that all of the tubes in the coincidence indicator chain (V32 through V36A) can be held cut off, and likewise sets the output pulse width of the coincidence indicator.

Changes in the plate to ground voltage of V32 through V36A are coupled to the control grid of the clipper amplifier stage, V36B. The V36B grid is normally held at about -34 volts negative to the cathode, and V36B is nonconducting. Positive inputs at the V36B grid cause an output at the V36B plate. Negative input voltages are not amplified by V36B because negative inputs drive the V36B grid further into cutoff.

The negative V36B output is differentiated in the short, time-constant circuit. The positive pulse resulting from the differentiation is applied to the grid of a reset thyratron, V37.

The reset thyratron, V37, is normally held beyond cutoff by a fixed bias voltage from the -108-volt supply. The positive differentiated output of V36B (when the coincidence chain comes back into conduction) triggers V37 into conduction. The resulting positive output at the V37 cathode is a very sharp positive pulse. The point (in time) at which the reset pulse occurs corresponds to 50 μ sec after the time during which the simultaneous negative voltages from the counter plates arrive at the grids of V32 through V36A. Thus, the recurrence rate of the reset pulse at the V37 cathode is determined by the selection of the counter plates by the PRR switch, S6, to supply the simultaneous negative pulses to the grids of all the coincidence indicator chain tubes, V32 through V36A. The output reset pulse from V37 is fed from the V37 cathode to the control grids of the B sections of

V21, V22, and V24 through V30 (through their respective grid resistors). The reset pulse is also applied to the first counter, V20, and to the fourth counter, V24, if the L-R switch, S9, is in the OFF position.

The reset pulse causes a switching of all the conducting tube sections in the counter chain, and thereby resets the chain. When the point (in time) is again reached where all of the counter outputs applied to the coincidence indicator chain are negative, another reset pulse will be produced 50 μ sec later and will likewise again reset the counter chain. Thus, the pulse repetition rate (frequency) at which this action occurs is determined by the time required for the counter chain to reach the condition of all negative outputs to the coincidence indicator plus 50 μ sec. This, in turn, corresponds to a given setting of the PRR switch, S6.

The positive reset pulse from the V37 cathode will appear at the B grids of the second, third, and fifth through eleventh counter stages, as a result of a permanent connection. With the L-R switch, S9A, in the OFF position, the positive reset pulse from the cathode of V37 is also applied to the B grid of the first counter stage, V20, and the B grid of the fourth counter, V23. Thus, all counter stages are reset on the same side at a rate determined by the PRR switch, S6.

In the receiving system, two sweeps are produced during the time of one recurrence interval of the signals received from the transmitting groups (fig. 10-27). One reset pulse from V37 is required to produce each of the two sweeps. Thus, the output frequency of the reset thyratron, V37 must be twice the receiving system recurrence rate. The purpose of the two sweeps is to allow the indicator circuits to display separate traces for the master and slave pulses, respectively, from the transmitting group.

During operation, the total period of the first trace is considered the A period and the period of the second trace, the B period. The master pulse is positioned in the first (A period) trace. The slave pulse, which always occurs at a time greater than 1/2 the transmitting group recurrence rate, is positioned on the second (B period) trace.

The twelfth counter stage, V31 (fig. 10-21) is active in producing the A period and B period traces. The operation of the twelfth counter stage, V31, is the same as that discussed earlier with the use of figure 10-25.
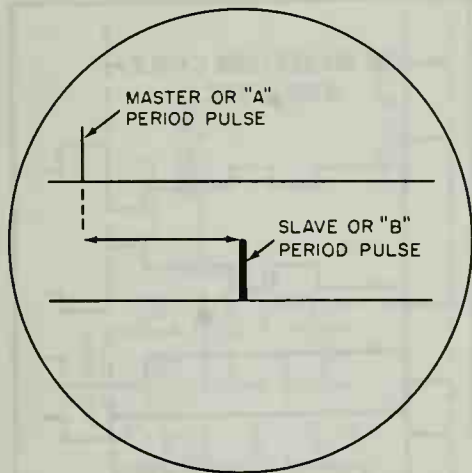
109.22

Figure 10-27.—Cathode-ray tube showing A and
B period traces.

VARIABLE DELAY CIRCUITS

The variable delay coincidence stages, V39
through V45 (fig. 10-21), produce a pulse that is
continuously variable in delay over the entire B
period. The actual delay is read on the time-
difference dial on the receiver front panel.

The coincidence action of selected counter
outputs of any of the tubes (V20 through V30) with
the output of the goniometer circuits (V47 and
V48) in the delay coincidence indicator stages
(V39 through V45) produces the variable delay.
The V39 to V44 output is combined with the
output of the twelfth amplifier to produce the
variable B pedestal.

It is sufficient at this time to regard the
goniometer as a device whose output varies
continuously in phase throughout its 100-μs
period.

The delay crank controls the position of the
lower pedestal within the B period. The delay
crank accomplishes this action by operating
switch S8 in such a manner that the B pedestal
pulse occurs at an interval of time that cor-
responds to the reading indicated on the time-
difference counter.

S8A through S8D are the sections of S8 which
connect the grids of V39 to the plates of V22 in
the counter chain when the delay crank is ro-
tated. The common cathode voltage at the V39
grid maintains each section of V39 cutoff when
its input signals are momentarily disconnected
through the rotating contacts of S8.

V39A receives the grid input from the A
plate of the third counter, V22, through S8A.
The V39B grid input is coupled from the B plate
of the third counter, V22. The duration of the
third counter output is 100 μs, having positive
and negative alternations of 50 μs each. When
either section of the tube is conducting, the
voltage drop across the common plate load re-
sistor produces a negative rectangular output
pulse, which is 50 μs in width. This pulse is
direct-coupled from the V39 plate to the grid of
V40.

As the delay crank is turned, S8, which is
geared to the crankshaft also turns, and the
positive alternation of the input wave form from
the third counter is constantly being applied to
one of the V39 grids.

The time-difference counter changes at a
rate of 100 units per revolution of the goniome-
ter (S8) shaft. One revolution of the goniometer
corresponds to 100 μs. A gear reduction of 4 to
1 between the S8 shaft and certain switch sec-
tions of S8 cause these switch sections to turn
90° for every revolution of the goniometer. Since
the goniometer shaft directly drives the me-
chanical counter, the counter will change
400 units as the output of V39 moves through
a 400-μs cycle (one revolution of the S8) and
returns to the starting point. Figure 10-28
shows the waveform at the plate of V39 for every
22.5° interval between 0° and 360° points of
the S8 drive shaft rotation.

GONIOMETER AND CLIPPER CIRCUITS

The output of the third counter stage, V22,
is coupled to the control grid of a shaper stage,
V50A (fig. 10-29).

The grid of V50A begins to draw a current
because of the connection of the V50A grid to
the +160-volt supply. With the square wave
applied, negative grid-to-cathode (bias) voltage
is established at the V50A grid.

Shaper V50A amplifies the square-wave in-
put and applies its output to a 10-kc filter. The
filter output is a 10-kc since wave, shich is
applied to the grid of the goniometer driver,
V46.

Goniometer driver V46 is a cathode follower
with a positive grid bias. The cathode potential
of V46, as a result of the V46 tube conduction is
about +42 volts. These potentials hold V46 at
the proper operating point.

The output of V46 is a 10-KC sine wave,
which is coupled to the goniometer, B1. Thus,
the period of the goniometer input is 100μs, the
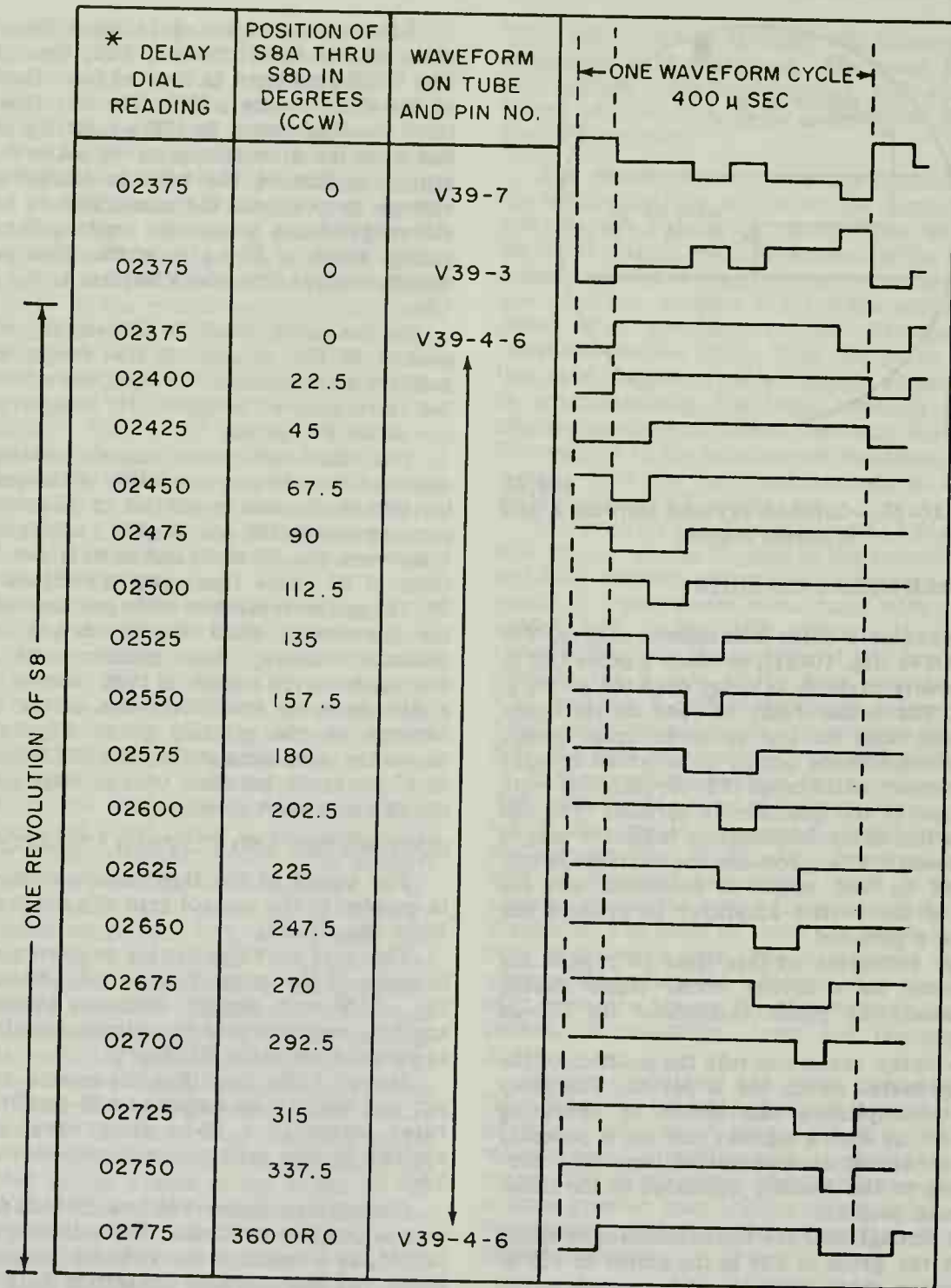
Figure 10-28.—Gating waveforms from delay coincidence indicator.

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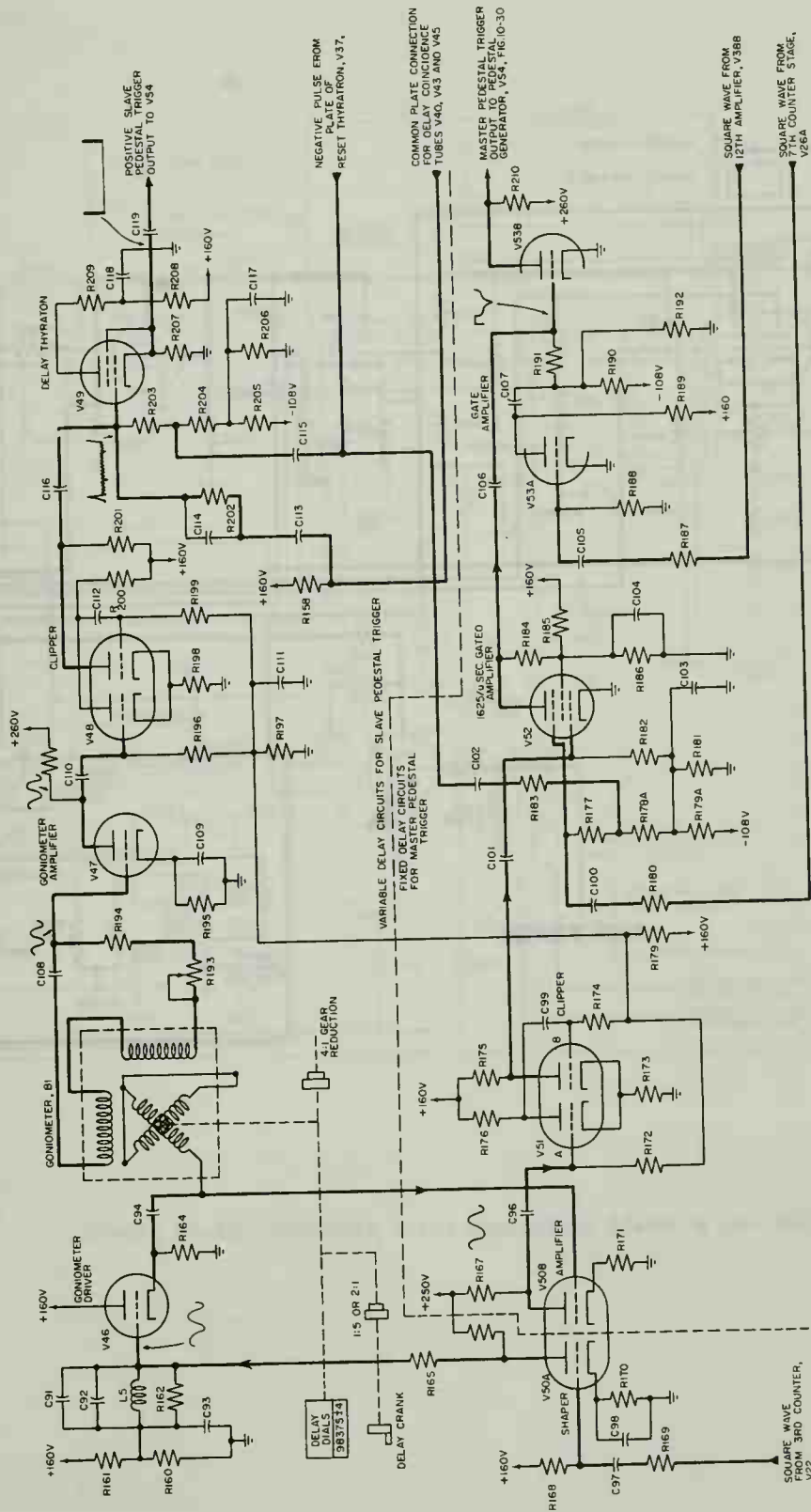
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Figure 10-29.—Goniometer, clipper, and fixed delay circuits.

109.25
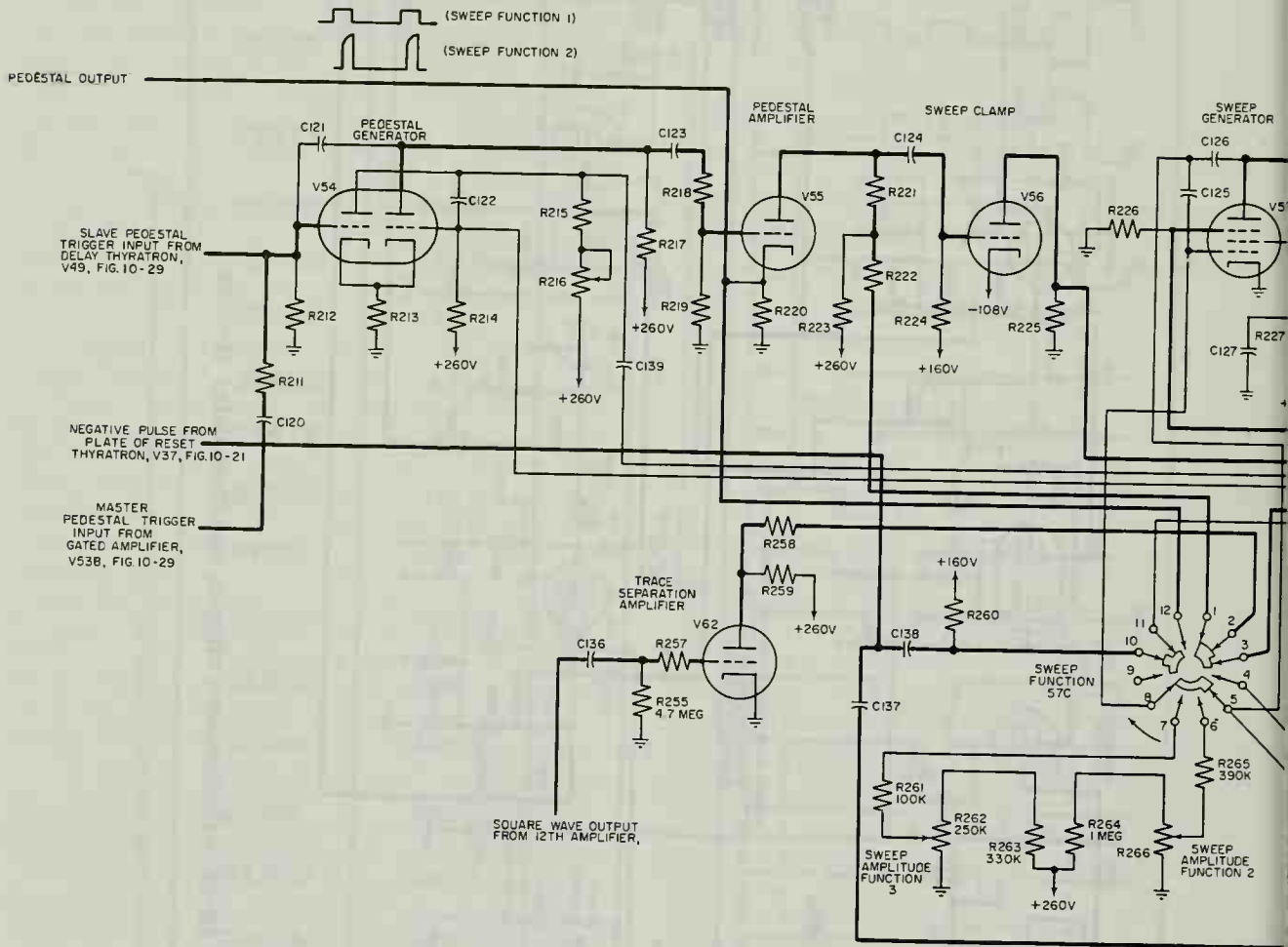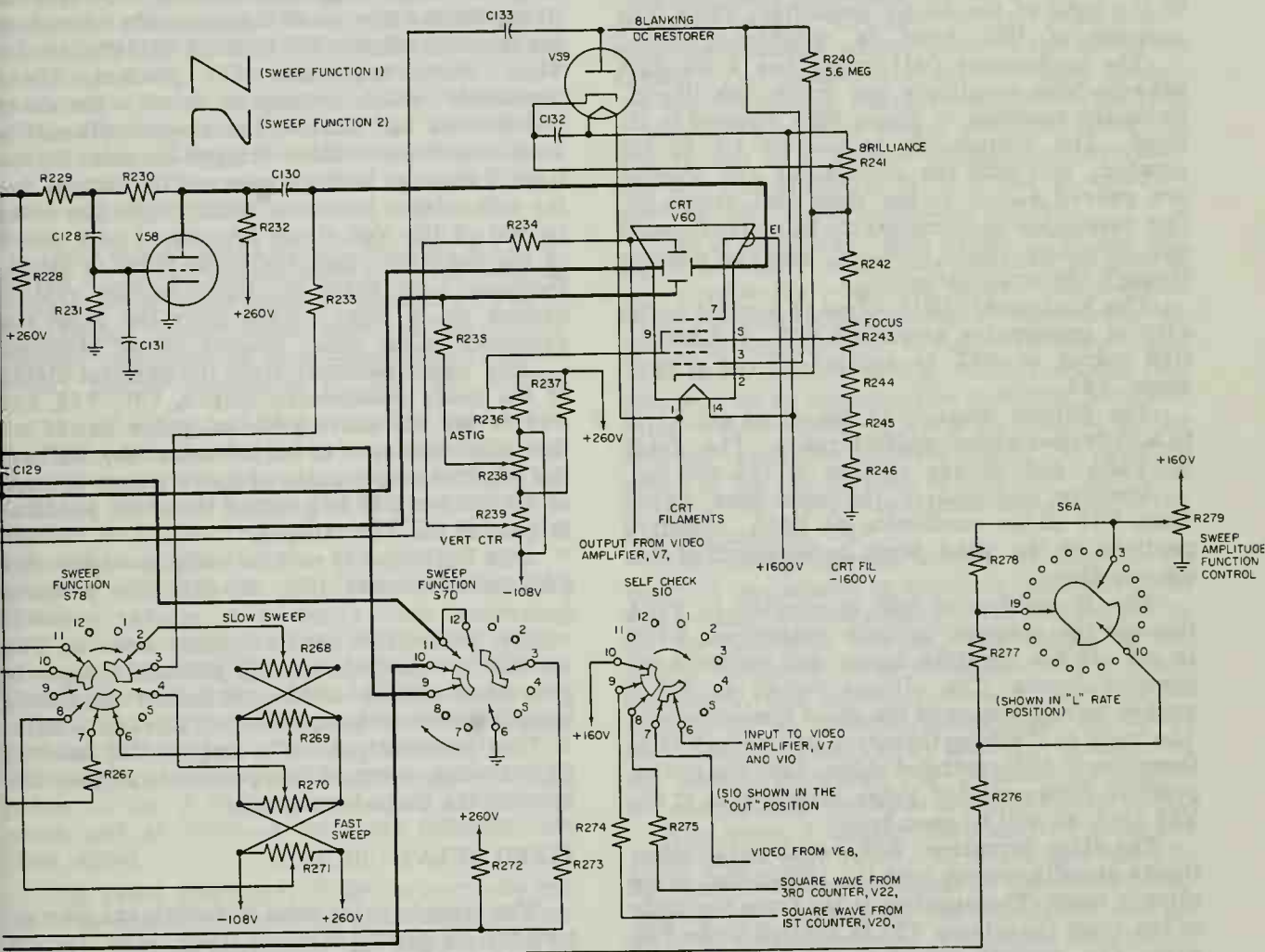
Figure 10-30.—Pedestal, trace separation, blanking and self-check circuits.

same as that of the widest pedestal pulse from the delay coincidence indicator, V45.

The goniometer driver output is also coupled to the grid of the 10-KC amplifier, V50B. The purpose of this input is considered later.

The goniometer (B1) output has a constant peak-to-peak amplitude per cycle, but is continuously variable in phase with respect to its input. The goniometer is geared 1:1 to the counter, and both the goniometer and counter are geared 4-to-1 to the shaft that drives S8. One revolution of B1 moves the B1 output through one complete cycle (100 μs), and the counter through 100 scale divisions.

The goniometer (B1) output is applied to the grid of goniometer amplifier V47. The amplified output of V47 is applied to the clipper stage, V48.

The clipper stage, V48, functions similarly to a cathode-coupled multivibrator. The grids of V48A and B are tied to a +17-volt tap.

Positive portions of the input sine waves from V47 cause conduction of V48A. Negative portions of the V48A input cause cutoff of this tube section.

The B section of V48 is coupled to V48A through the common cathode resistance. V48B is cut off for negative inputs and conducts for positive inputs. The clipper output is not a square wave because of the short time constant. The input to the delay thyratron V49, from V48 is therefore a differentiated pulse. Only the sharp, positive differentiated pulse is effective at the V49 grid, as will be seen later.

The delay thyratron, V49, receives two other inputs simultaneously with the application of the clipper input. The negative pulse from the plate of the reset thyratron, V37 is applied to the V49 grid. The positive output of the delay coincidence tubes, V40, V43, and V45 (fig. 10-21) is fed to the grid of V49.

The position (in time) of the positive output of the clipper stage, to the delay thyratron grid varies continuously as the goniometer shaft is turned. The positive trigger, which is applied to the V49 grid from V48 in the presence of the positive delay coincidence input to V49 causes the delay thyratron to produce a positive-going slave pedestal trigger pulse. Because the delay coincidence input pulse from V45 is only available during the B period, and because the input pulse from the goniometer is continuously moving through a 100-μs period as the goniometer shaft is turned, the time of the conduction of V49 can vary throughout the 100-μs interval.

The exponential slope of the input from the reset thyratron, V37, is sufficiently negative to keep the delay thyratron, V49 (fig. 10-29) from firing during the short interval after reset of the counter stages (V20 through V31) (fig. 10-21) when many transients are present. These transients could be coupled through the delay coincidence and goniometer stages with sufficient amplitude to false-trigger the delay thyratron, V49, even in the absence of the normal delay coincidence indicator chain output. The delay period of the V49 firing introduced as a result of the negative reset thyratron pulse is small. Because zero delay for the receiving system occurs an A delay period after the reset the accuracy of the delay reading is not affected.

The delay pedestal from the parallel plates of the delay coincidence stages, V40, V43, and V45 is not the slave pedestal pulse and is not the pulse displayed on the indicator tube. Rather, the positive output pulse of these stages is used at the grid of V49 to produce the slave pedestal trigger at the V49 cathode.

The V49 output is coupled to the slave pedestal generator circuits (fig. 10-30). The pedestal generator also receives a master pedestal trigger pulse from the fixed delay circuits. The method by which the master pedestal trigger is generated will be considered before discussing the generation of the master and slave pedestals.

The pedestal generator stages, V54 and V55 (fig. 10-30), produce the pedestals as they are seen on the indicator screen.

FIXED DELAY CIRCUITS

The trigger pulse used to initiate the master pedestal is produced in the fixed delay circuits (fig. 10-29) by combining a pulse obtained from the leading edge of the sine-wave output of the 10-kc amplifier (after passing through clipper stage, V51) with square waves from the seventh counter stage, V26, and the twelfth amplifier, V38. These inputs produce the master pedestal trigger pulse an A delay period after reset in the A trace period. Thus, the time of the master pedestal trigger corresponds to the leading edge of the master pulse, which begins 1625 μs after the beginning of the A period.

In operation, the time-difference counter is adjusted to give a zero reading 1625 μs after the beginning of the A period. Thus, any change in the time-difference counter reading from zero in the B period trace (when the difference is being determined) represents the actual difference

s between the master and slave pulses less
ne-half the recurrence interval.

The 10-KC amplifier, V50B (fig. 10-29) re-
eives the goniometer driver output and couples
n amplified 10-KC sine-wave to the grid of clip-
er stage V51A. The clipping action in V51 is
roduced in the same manner as the V48 clipper
n the variable delay circuits. The 10-KC
quare-wave output of V51B is differentiated
nd applied to the control grid of the 1625-μs
ated amplifier, V52.

The 1625-μs gated amplifier, V52, also re-
eives a 1600-μs square wave from the seventh
ounter stage. This input is applied to V52 (a
oincidence stage) which requires the simul-
aneous application of input pulses (from V51 and
726A) before conduction of V52 will occur.

The time constant of the coupling network to
he V52 suppressor grid is long enough to hold
he suppressor at approximately zero volts for
25-μs interval before the clipper output is
pplied to the control grid of V52. When the
751 output arrives at the V52 control grid, a
egative output pulse is produced at the V52
late.

The negative input pulse, coupled to the sup-
pressor grid, from the reset thyratron, V37 (fig.
10-20) prevents firing of the 1625-μs gated am-
plifier, V52 (fig. 10-29) due to transient voltages
present immediately after the reset period. The
1625-μs gated amplifier, V52 produces an output
pulse on the A delay period after each reset
pulse and at 1600-μs intervals following the
first output.

The gated amplifier, V53B, is normally cut
off. The output pulse of the 1625-μs gated am-
plifier, V52, is applied to the control grid of
V53B. This input pulse alone is not sufficient
to cause the conduction of V53B.

The control grid of V53A receives and ampli-
fies the square-wave output of the twelfth ampli-
fier, V38B (fig. 10-20) which is negative during
the A period and positive during the B period.
Since in V53A (fig. 10-29) the input pulse is in-
verted at the V53A plate, the V53A output is
positive going during the A period. The time
constant (in the V53A coupling network) main-
tains the V53B grid above cutoff just long
enough for the first input pulse from the 1625-μs
gated amplifiers to the V53B grid to be ampli-
fied in this stage. The V53B output represents
the master pedestal trigger output pulse to the
pedestal generator circuits (fig. 10-30).

PEDESTAL CIRCUITS

The trigger pulses produced in the fixed and
variable delay circuits are used in the pedestal
generator stages, V54 and V55 (fig. 10-30) to ini-
tiate the pedestal pulses seen on the indicator
screen. The master and slave trigger pulses
are applied (during separate periods) to the grid
of pedestal generator stage, V54. The circuit of
V54 is that of a cathode-coupled multivibrator
and operates similarly to the A and B multivi-
brators (fig. 10-23) discussed earlier.

The pedestal generator stage, V54B (fig. 10-
30) is normally conducting because its grid is
connected to the +260-volt supply. The pedestal
trigger pulses (either master or slave from V49
or V53B, respectively, of figure 10-29) are ap-
plied to the V54A grid (fig. 10-30), causing this
tube section to conduct. The resulting positive-
going output pulse of V54B is coupled to the
pedestal amplifier control grid. The positive
output at the V55 cathode is coupled to the AFC
circuits (fig. 10-23) to initiate the action of the
A multivibrator, V8.

The sweep function switch, S7 (fig. 10-30)
controls the manner in which the pedestal pulses
will be used and the nature of the pattern shown
on the indicator screen. The modes of the
switch, referred to as sweep function 1, sweep
function 2, and sweep function 3, describe the
manner of the indicator display and the use of
the pedestal pulses rather than the position of
the switch contacts. The cathode-ray display
for each of the S7 modes is shown in figure 10-31.

The loran A display has a master pulse and
slave pulse (fig. 10-31, A) which are tuned by
the receiver and adjusted by the controls to
display the master on the top trace and the
slave on the bottom trace. The pulses are brought
to the leading edge of the pedestal by the delay
(counter) control (fig. 10-20) and then sweep
function switch, S7 is placed in step 2 to give
display in figure 10-31, B. The pulses are then
lined up with the lower pulse adjusted under the
upper pulse by the counter control. The function
switch, S7 is then placed in step 3 (fig. 10-30), C
and the counter adjusted to superimpose the
leading edge of the two pulses. The counter
readings are plotted on a loran chart using loran
A tables, and usually checked against one or two
other station fixes.

The loran C display (figure 10-31, C) shows
that the loran C signal is a group pulsed technique
instead of the loran A single pulse. The master
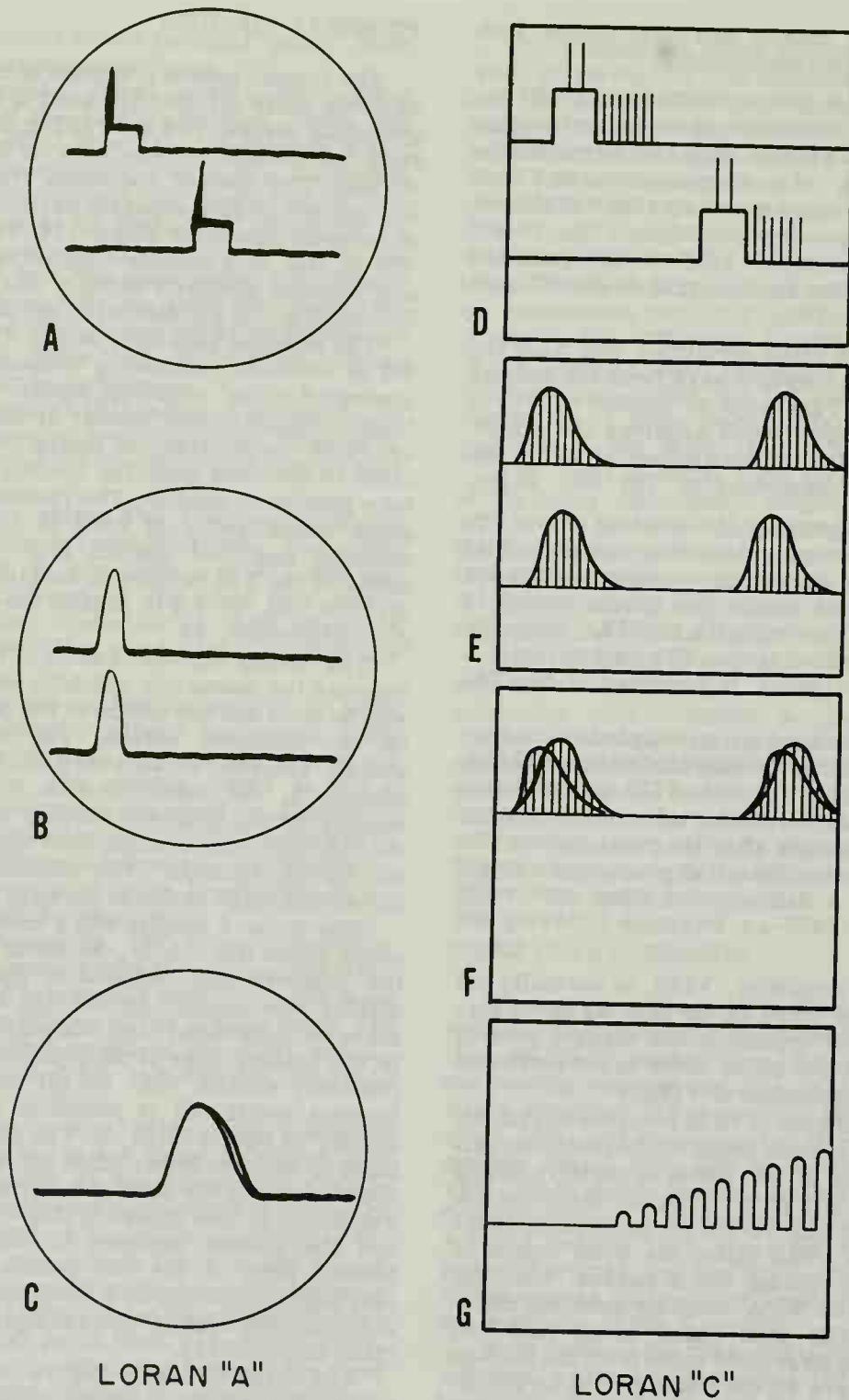group transmission is comprised of nine

LORAN "A"

LORAN "C"

Figure 10-31.—Cathode-ray tube presentations for sweep functions 1, 2, and 3.

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hase-coded pulses while the slave group comrises eight phase-coded pulses. To take a eading of the counters for a fix, adjust the elay counter crank to line up the two pedestals nd pulses. Set the sweep switch (S7) to position (fig. 10-31, E) and again align the two bottom ulses with the top pulses. Adjust the TRACE EP control to superimpose the two sweeps fig. 10-31, F). Set the sweep function switch 7, to position 3 (fig. 10-31, G) and adjust the elay counter to align the two RF pulses. Record he counter reading and follow the same procedure to check the same master pulse against second set of slave pulses. The counter readngs are plotted on a loran chart using loran C tables.

TRACE SEPARATION

The output of the twelfth amplifier stage, V38A (fig. 10-21) is fed into the trace separation amplifier, V62 (fig. 10-30). The purpose of the trace separation amplifier, V62, is to produce a potential (at the V60 vertical deflection plates) which causes the electron beam to be presented in the upper section of the indicator (V60) tube screen during the A period, and towards the lower section of the V60 screen during the B period.

The output of V38A is positive during the A period and negative during the B period. The V38A output is coupled to the grid of V62.

When the equipment is initially set into operation, positive portions of the V38A output cause the V62 grid to draw current, and bias is established. V62 amplifies and inverts the input square wave from V38A.

In sweep function 1 the V62 output which is negative going during the A period and positive during the B period, is coupled to the lower deflection plate of cathode-ray indicator tube, V60. Since this output is applied to the lower vertical deflection plate of V60, the V60 electron beam is repelled by the negative-going V62 output during the A periods, and the A period trace is therefore presented on the upper section of the V60 screen. During the B period, the V62 output is positive going, and the B period trace will be attracted downward toward the now more positive lower deflection plate.

Note that during sweep function 1, the pedestal amplifier (V55) plate output (which is also negative going as discussed) is applied through the

1-3 contacts of S7C to be superimposed on the trace separation output of V62. Because the pedestal pulses from the V55 plate are also negative-going voltages, the potential on the lower deflection plate of V60 is made more negative for the duration of the pedestal pulses in the A and B period traces, and the pedestal pulses are presented to the V60 observer (fig. 10-31A).

SWEEP CIRCUITS

In sweep function 1, the negative-going plate output of the reset thyratron, V37 (fig. 10-21) is coupled to the suppressor grid of the sweep generator stage V57 (fig. 10-30) The V37 plate output goes negative at reset, and rises exponentially when V37 again cuts off.

The negative drop in potential that is applied to the V57 suppressor grid, holds the suppressor sufficiently negative to cut off the V57 plate current.

Since the time constant of the charge circuit of C126 is short, C126 will become completely charged by the time that the V57 suppressor input rises sufficiently to allow V57 to again conduct. The suppressor potential goes slightly positive at the end of the negative reset pulse, as a result of the plate potential rise of V37.

When the V57 suppressor grid is driven positive, plate current will again flow. The resulting drop in the V57 plate voltage and rise of plate current cannot occur at a faster rate than that required for C126 to lose its accumulated charge. The negative potential thus developed at the V57 grid to ground is applied to the V57 control grid, and opposes the rise in the V57 plate current.

The high grid-to-ground capacity of V57 permits the grid voltage to change only at a very slow rate. Consequently, the effective time constant for the discharge of C126 is sufficiently large to hold the negative-going V57 plate output linear over the duration of the positive suppressor input.

The resulting V57 plate output is a linear-rising sawtooth waveform. The sawtooth sweep duration during sweep function 1 is approximately 1500 to 25,000 μs, depending upon the PRR switch, S6.

The sweep amplitude function 1 control, R279 (fig. 10-30) adjusts the magnitude of the

positive-voltage applied to the V57 control grid, thereby locating the operating point of the tube and determining the amplitude (and width) of the V57 plate sawtooth output. For L rates, R277 is shorted by contacts 10 and 19 of S6A (as shown). For S rates (fewer cycles per second than for L rates), both resistors, R277 and R278, are in the C126 discharge path, and the discharge time is increased. Thus, fewer sawtooth voltage wave-forms per second will be produced by V57. For H rates contacts 10 and 3 short R277 and R278, and the sawtooth output frequency is increased.

In sweep function 2 each of the switch sections (S7B, and S7D) is rotated 1 contact to the left from the position shown. For sweep function 3, the switch sections are rotated 2 contacts to the left.

During sweep function 2 and sweep function 3, the pedestal pulses are not displayed on the indicator screen. Instead, the pedestals are coupled through C124 to the grid of the sweep clamp stage, V56. The negative-going grid input is amplified in this stage.

Because of the larger negative voltages at the V56 cathode (-108 volts) V56 will conduct and amplify strong negative signals. The V56 grid is negative to the cathode due to the discharge of C124 through R224.

During sweep function 2 the V56 plate pedestal output is coupled through contacts 11 and 9 of S7D of the suppressor grid of the sweep generator, V57. The sweep amplitude function 2 control, R266, is selected by contacts 8 and 6 of S7C, and is used to adjust the sawtooth amplitude for the sweep function 2 mode of operation. The sweep duration for sweep function 2 is approximately 1250 μs.

For sweep function 3 operation, the sweep clamp (V56) plate pedestal output is coupled through contacts 12 and 9 of S7D to the sweep generator (V57) suppressor grid. The sweep amplitude function 3 control, R262, which is used to adjust the V57 plate sawtooth output amplitude, is selected by contacts 8 and 7 of S7D. The duration of the sweep for sweep function 3 operation is about 160 μs.

As stated earlier, R279 functions to adjust the sweep amplitude (or sweep duration) for sweep function 1. The sweep duration is different for each position of the sweep function switch, S7. To obtain adequate sweep amplitude range for each of the sweep functions (1,

2, and 3) the time constant of the C126 discharge must be changed.

For sweep function 2 (S7 one position to the left) C125 is placed in series with C126.

The resistors in the discharge path during sweep function 3 selected by the 8 and 7 contacts of S7C included R261, R262 (upper section), and R263. Thus, R266 and R262 adjust the sweep amplitude for sweep function 2 and sweep function 3 modes of operation, respectively.

The output of the sweep generator, V57, is coupled through C129 to the left horizontal deflection plate of the cathode-ray tube, V60. The V57 output is also coupled through R229 and C128 to the control grid of the sweep inverter, V58.

The sweep inverter (V58) grid is grounded through C131. This capacitor bypasses the grid resistor, R231, to reduce differentiation of the input, which would otherwise occur across R231, thereby altering the shape (rise time) of the sawtooth input. The V58 input sawtooth is inverted in the stage, and coupled to the right horizontal deflection plate. Inverse feedback through R230 from plate to grid of V58 is used to hold the amplitude of the V58 output equal to the output of V57, thus supplying the same magnitude voltage to the left and right horizontal deflection plate.

BLANKING CIRCUITS

The blanking d-c restorer, V59 (fig. 10-30) prevents the appearance of the sweep voltage retrace on the V60 screen. V59 is connected across the potential difference existing between the V59 plate and cathode. These resistors form a voltage divider from a -1600-volt potential to ground. The plate of V59 is positive to the cathode, and the tube conducts. C133, at the plate of V59, charges from the plate potential (approximately -1600 volts) through contacts 11 and 10 of S7C toward the 160-volt supply.

It will be recalled that the negative pulses arriving from the plate of the reset thyratron, V37 terminate the sawtooth wave-form at the V57 plate. At the same time, the reset thyratron negative plate output is applied to C133. Thus, during the period of the negative trigger, C133 discharges through R240, developing a large negative potential at the control grid of the cathode-ray tube, V60. This action blanks the indicator tube during the periods of the sawtooth waveform flyback.

ATHODE-RAY TUBE CIRCUITS

The accelerating potentials required for he operation of the cathode-ray tube, V60 ig. 10-30) are obtained from separate positi ve and negative supplies (not shown). Both upplies develop a 1600-volt output. The ositive supply is connected to the third anode, 1, of V60.

A voltage divider is connected between the 1600-volt source and ground. Potentials otained from this divider provide the voltages required at the cathode (pin 2) control grid (pin) and second anode (pin 5) of V60. The control rid potential is determined by the setting of the rm of the brilliance control, R241, and the reulting conduction of V59. The second anode ; connected to the focus control, R243, nd the cathode to the junction of R241 and 242.

Since the cathode of V60 is highly negative ith respect to ground, the heater (between pins and 14 of V60) is returned to the -1600-volt upply to keep a low potential difference between he heater and cathode of V60. The heater of V59 s also returned to the -1600-volt supply for he same reason.

The voltage at the first anode (pin 9) of 760 is controlled by the astigmatism adjustment, R236. This control is connected between the -260- and -108-volt supplies. Vertical center potentiometers, R238 and R239, limit the voltage range of the controls. The connection of the vertical controls between the -108- and +260-volt supplies provided push-pull vertical centering. The astigmatism control, R236 functions in conjunction with the focus control, R243, to determine the defintion of the presentation.

The DC potentials at the vertical deflection plates of V60 are determined by the settings of R238 and R239. The DC potentials on the horizontal deflection plates depend on the position of the sweep function switch section, S7B. For sweep function 1 (S7B as shown) the settings of R268 and R269 (slow and fast sweep controls, respectively) determine the potential on their respective horizontal plate. For sweep functions 2 and 3, the horizontal deflection plate potentials are obtained from R270 and R271, respectively. Since the sweep period during sweep function 1

is considerably longer (15,000 to 25,000 μs) than the sweep used for sweep functions 2 and 3 (1250 and 160 μs, respectively), the use of two sweep controls (fast and slow) eliminate timely readjustment of the horizontal plate potentials when switching from one sweep function to another.

The self-check switch, S10 permits markers to be produced on the indicator tube so that the sweep length and the alignment of the goniometer and time-difference counter can be checked. When S10 is in the OUT position the video output from V6B (fig. 10-21 is connected through contacts 7 and 6 of S10 (fig. 10-30) to video amplifiers, V7 and V10.

When S10 is rotated one position to the left, 100-μs marker pulses are coupled from the output of the third counter stage, V22 to the video amplifier, V7. This input is subsequently applied to the upper vertical deflection plate of the cathode-ray tube.

When S10 is rotated two positions to the left 25-μs markers are obtained from the first counter stage, V20. These pulses are similarly displayed on the cathode-ray tube.

LEFT-RIGHT SWITCH

In operation, received signals at the selected pulse repetition rate are locked in on the indicator screen by use of either the manual or automatic frequency control circuits.

The L-R (left-right) switch, S9 (fig. 10-24) enables the operator of the receiving set to position the signals at any desired location on the trace without having to disturb any of the other controls.

When S9 is in its normal (OFF) position, the switch is not effective in controlling the position of the indicator pattern.

The reset pulse from the cathode of V37 is applied through the contacts of S9A to the B grid of the first counter stage, V20. The V37 cathode reset pulse is also fed to the B grid of the fourth counter, V23. Since the B grid of all the other counter stages are permanently controlled by the positive reset pulse, all of the counter stages are then reset on the same side, and the counters

V20 through V31, produce the proper output count selected by the PRR switch. This represents the normal action of the reset and counter stages.

When the L-R switch is operated to the RIGHT position in sweep function 1, the positive reset pulse from the V37 cathode is coupled to the A grid of the first counter stage, V20. The reset pulse is also coupled to the A grid of the fourth counter stage, V23. Thus, the first and fourth counters (V20 and V23, respectively) are reset to the side opposite from normal.

Each counter stage normally produces one output pulse for each two input triggers. However, when the first and fourth counters are reset at the opposite side, the first input trigger pulse from the preceding stage (or from the clipper stage, V19 in figure 10-24) after the reset pulse will cause the first and fourth counters to produce an output. After the first output pulse of V20 and V23 is produced these stages will produce an output on each odd pulse input.

Because the first and fourth counters are driven by trigger pulses with repetition rates of 12.5 and 100 μs, respectively, the count selected by the PRR switch, S6, will be reached 112.6 μs earlier than normal, and the receiving set repetition rate is higher than that of the received signals. The signals will appear at a later time on successive traces, and drift across the screen to the right.

The sweep time for sweep functions 2 and 3 is much less than in sweep function 1, and a slower drift is necessary to permit the operator to control the received signals.

The path for the reset pulse input to the fourth counter, V23, is open. As a result, only the first counter, V22, is reset to the opposite side, and the selected count is reached only 12.5 μs earlier than normal. This produces a much slower drift of the received signal across the indicator screen.

When the L-R switch, S9B (fig. 10-24) is operated to the LEFT position in sweep function 1 the negative plate output of the reset thyratron, V37 is coupled to the control grid of the ringer tube, V18. The input pulse is differentiated. This negative input holds the ringer tube, V18, beyond cutoff for a period of nine driving pulses (112.5 μs). The receiving set pulse repetition rate is therefore made lower than that of the received signals. This action causes the received sig-

nals to appear at an earlier time on successive sweep traces, and drift across the screen to the left.

C140 normally charges negative to ground during the period of the negative input pulse.

The C140 discharge maintains the negative cutoff potential at the grid of V18 for the required period.

For sweep functions 2 and 3, the cutoff time of V18 is reduced as necessary to produce a slower drift of the received signals. During sweep functions 2 and 3, C140 is shorted to ground. The cutoff period of V18 is reduced to that equal to 1 driving pulse (12.5 μs). Thus, the drift toward the left end of the cathode-ray tube is made slower.

AMPLITUDE BALANCE

The amplitude balance stage, V61 (fig. 10-32) provides a means of equalizing the amplitude of the received master and slave pulses during the A and B periods, respectively. The cathodes of V61 are tied together and returned to the -108-volt supply. The plates of V61 are connected through separate sections of R249, each section of which is in series with R256 to ground.

Both sections of V61 are normally conducting, and the plate potential of the A and B sections of V61 is negative ground. The negative voltage at the terminal 2 arm of R249 is applied to the control grid of the r-f and i-f stages, V1, V3, and V4 (fig. 10-21).

The twelfth amplifier (V38B) cathode output is applied to the A grid of V61 (fig. 10-32) and is negative during the A period, and positive during the B period. The V38B plate output is fed to the grid of V61B, and is positive during the A period and negative during the B period. The width of the twelfth amplifier output is equal to the width of the recurrence interval of the received signals with the positive and negative alternations equal to one-half the total recurrence period.

The potential at terminal 2 arm of R249 for the separate halves of the twelfth amplifier input depends upon the setting of the terminal 2 arm. If the terminal 2 arm is moved to the right or
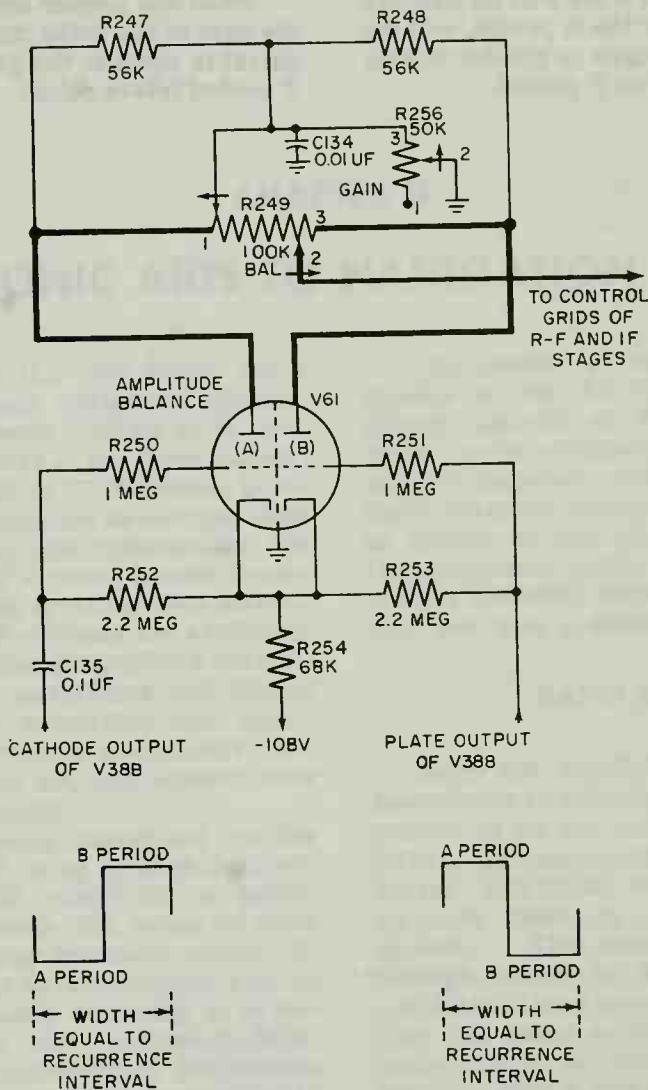
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Figure 10-32.—Amplitude balance circuit.

left, the negative potential fed to the r-f and i-f stages will be larger for one-half cycle of the input (one-half recurrence period) than for the other.

Assume that the terminal 2 arm of R249 is moved to the extreme right (as shown). Since the V61B grid is positive during the A period and negative during the B period, the V61B conduction through R249 and R256 during the A period is greater than the conduction during the B period. Thus, the voltage at the terminal 2 arm of R249 will be more negative during the A period than during the B period. This increased negative potential is fed to the grids of r-f and i-f stages (V1, V3, and V4 of figure 10-31), and the gain of these stages during the A period is less than the gain during the B period. Likewise, the amplitude of the pulse in the B period is increased, and the A period pulse amplitude is decreased.

By positioning the terminal 2 arm of R249 (fig. 10-32) to the extreme left position, the reverse effect to that just discussed can be obtained. Since the input of V61A is negative during the A period and positive during the B period, the conduction of V61A is smaller for the A period than for the B period. Thus, the negative voltages

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at the terminal 2 arm of R249 will be smaller for the A period than for the B period, and the gain of the r-f and i-f stages is greater during the A period than during the B period.

When the proper setting of R249 is reached the gain of the video during the A period (master pulse) is equal to the gain of the video during the B period (slave pulse).

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CHAPTER 11

ELECTRONIC AIDS TO NAVIGATION, PART II

The U.S. Navy and U.S. Air Force use a radio beacon as an air navigation system, called TACAN. The word TACAN is formed from the underlined letters of the term Tactical Air Navigation. TACAN is an electronic polar coordinate system that enables an aircraft pilot to read—instantaneously and continuously—the distance and bearing of a radio beacon transmitter installed on a ship or at a ground station.

Two types of TACAN systems are available, one for shore installation (AN/GRN-6 series), and one for shipboard installation (AN/SRN-6 series). Each system comprises four major groups (fig. 11-1), receiver-transmitter, antenna, power supply-test set, and monitor (may include remote control unit).

A new TACAN is being introduced into the fleet. The AN/URN-20 is of solid state (transistorized) construction except for the power output tube. Two cabinets will house all units except antenna, including automatic switch off. The AN/URN-20(V) will be two complete TACAN equipments with automatic switching to an active unit if the ''on air'' equipment fails.

To supply correct information, the antenna must be maintained in a position perpendicular to the earth's surface. Thus the antenna radiation pattern maintains a constant relationship to the azimuth plane. In addition, the antenna radiation pattern must be constantly referenced to magnetic north. The magnetic variation (discussed later) for each area, whether shipboard or shore based, must be set on the equipment.

The shore based antenna, after initial installation, will not vary its position, but the shipboard installation must compensate for magnetic variations, azimuth bearing, roll, and pitch. To correct these variables, the magnetic variation is reset as the ship moves from one magnetic zone to another. The shipboard antennas are constructed with an azimuth, roll, and pitch servosystem to stabilize the antenna in its proper attitude.

This chapter presents a block diagram discussion of the AN/SRN-6 Tacan System, and circuit analysis of the receiver special circuits, coder-indicator, amplifier-modulator, control-duplexer, and antenna system. In the block diagram discussion, particular emphasis is placed on the purpose and relationship of the component parts. In the discussion of the special circuits, emphasis is placed on operation, and how it affects the remaining circuits.

RADIO SET AN/SRN-6

Radio Set AN/SRN-6 is used for shipboard installation to provide the radio beacon or ground portion of an air navigation system. The AN/SRN-6, its associated antenna group and accessories, and Radio Set AN/ARN-21 (located in aircraft) make up a complete air navigation system. The beacon supplies information through which an aircraft equipped with an AN/ARN-21 can accurately determine its position. As many as 100 aircraft may simultaneously obtain navigational information in conjunction with a single installation of the AN/SRN-6. To provide for a large number of transmitting stations, the system is capable of receiving on any one of 126 frequencies (channels) in the range of 1025 to 1150 mc. Transmission of information also takes place on 126 channel frequencies in the ranges of 962 to 1213 mc.

Two types of antenna are available for use, low band and high band. Each antenna operates on 63 channels, corresponding to low-band frequencies and high-band frequencies, respectively. Low-band installations transmit at frequencies between 962 and 1024 mc inclusive, and receive at frequencies between 1025 and 1087 mc. High-band installations transmit in the range of 1151 to 1213 mc, and receive in the range of 1088 to 1150 mc.
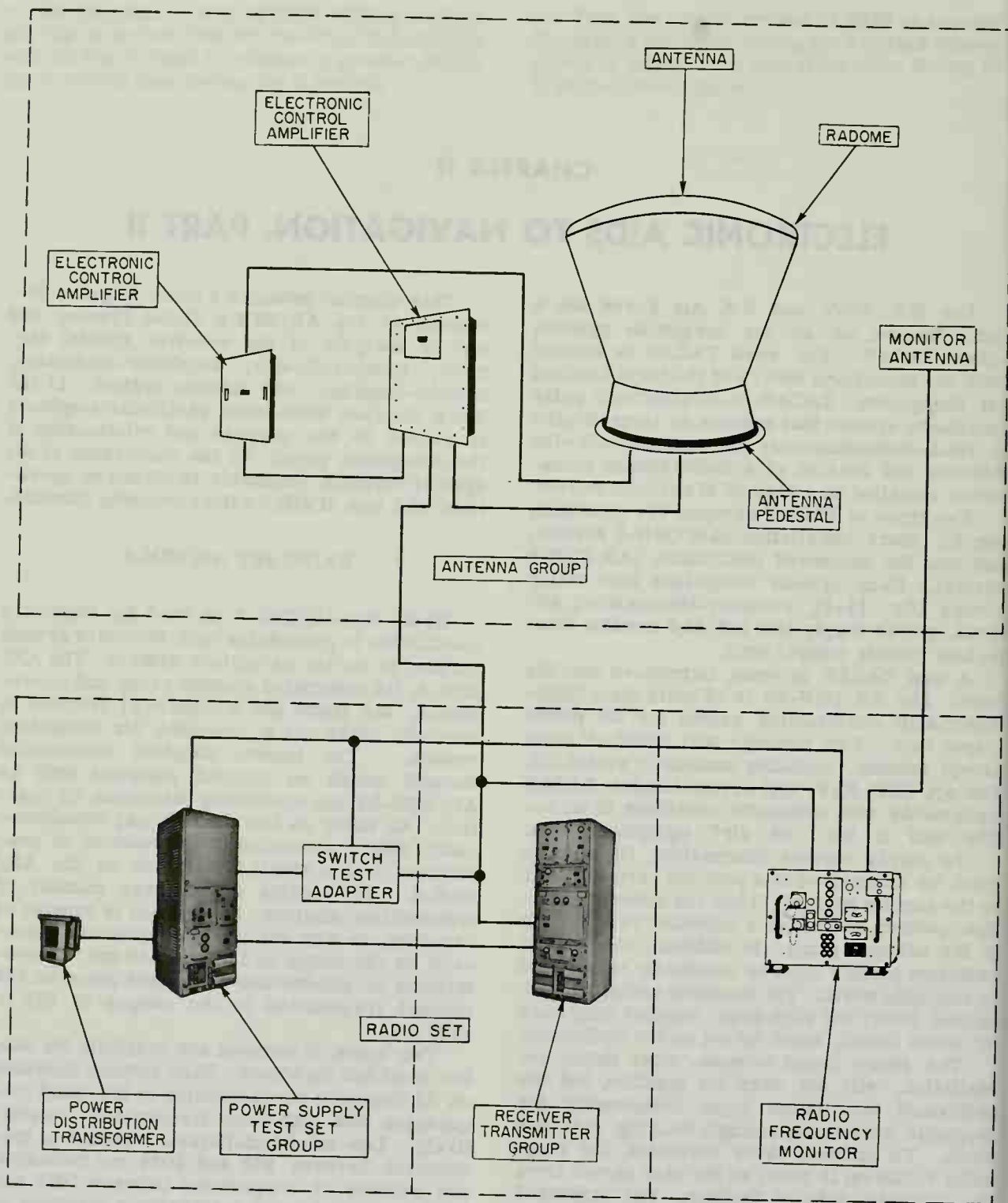
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Figure 11-1.—Major components of radio beacon AN/SRN-6.

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Two frequencies are used in each channel: one for receiving, and one for transmitting. The frequency used for receiving in low-band installations is 63 mc above the frequency used for transmitting in the same channel. In high-band installations, the receiving frequency is 63 mc below the transmitting frequency.

No two stations within interference distance of each other are assigned the same channel. The pilot can switch channels to select any tacan transmitter within range. In aircraft equipped with tacan receiving equipment, an azimuth indicator shows the position of the transmitting source in degrees of magnetic bearing from the aircraft. Also, the distance in nautical miles to the same reference point is registered as a numerical indication, similar to that of an automobile odometer. In the illustration (fig. 11-2) the aircraft is 106 miles from the carrier, and the ship is on a magnetic bearing of approximately 230° from the aircraft.

Two radio frequencies are employed (fig. 11-2), one frequency (Y) is used for transmissions to the aircraft; another frequency (X) is used for transmissions from the aircraft. The surface-to-air frequency carries bearing and range intelligence as well as station identification information. The transmission from the aircraft-to-surface unit is required to trigger the distance-measuring system.

When the pilot closes the proper switch on his set control, his aircraft radio set transmits
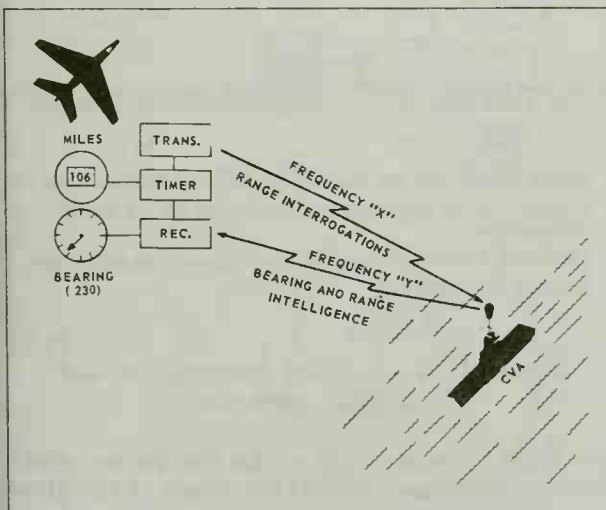

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Figure 11-2.—Dual-frequency transmission.

distance interrogation pulses only (frequency X), spaced 12 μs part. The interrogation pulses are detected by any ship or station operating on the same channel, and in reply, transmits back to the aircraft on the assigned channel pulse-pairs on frequency (Y).

The time delay between the distance interrogation pulse-pair (transmitted from the aircraft) and the corresponding distance pulse-pair reply (transmitted from the radio beacon) is adjusted at the radio beacon to exactly 50 μs. The aircraft radio set deducts 50 μs from the total time elapsed between the interrogation and reply. The distance between the aircraft and the radio beacon is determined by measuring the total time elapsed between initial transmission of the distance interrogation pulse-pair and the reception of the corresponding radio beacon reply. Other circuits convert the time difference to equivalent dial indication in miles. Bearing information is radiated continuously on frequency Y. Under normal circumstances, the range over which aircraft can obtain reliable distance information from the beacon is about 200 miles.

The Radio Beacon AN/SRN-6 output consists of the bearing information signals, the beacon identification code, distance information signals, and random pulses used to make up a constant duty cycle. To aid the pilot in identifying a particular transmitter, the radio beacon periodically transmits its identifying call in International Morse Code, thus enabling the interrogating aircraft to determine the radio beacon with which it is in contact. The characters of the code consist of a train of pulse-pairs generated at a fixed rate of 2700 pulse-pairs per second. During the transmission of the code, the 2700 pulse-pairs are grouped so that the reproduced tone will be at a frequency of 1350 cps. A motor-driven mechanical keyer in the coder-indicator accomplished the coding. The aircraft receiver detects these regularly occurring pulse-pairs and reproduces the code as a keyed 1350-cps audio tone.

The bearing information reference burst has priority over the identification call. The identification call has priority over the distance information signals. The relative duration of these signals is such that there is, effectively, no interruption of distance information or identification call.

When the radio beacon is in operation, reference burst information will be transmitted at regulated intervals. The beacon identification

code signals will be transmitted every 37.5 seconds. Any airborne Radio Set AN/ARN-21, which is tuned to the beacon radio frequency, will receive the reference burst and identification signals.

The pulse-pairs sent by the AN/ARN-21 (aircraft equipment) are transmitted on a carrier frequency which is the same as that of the radio beacon. The pulse-pairs are transmitted at a random rate peculiar to the AN/ARN-21 aboard a specific aircraft. When receiving, the AN/ARN-21 selects its own particular distance-measuring reply from all other signals by comparing the repetition rate of the reply pulses with the repetition rate of those originally transmitted. The receiver locks in on the reply pulses whose repetition rate is the same as the rate of its interrogation pulses.

The radio beacon operates on an assigned crystal-controlled carrier frequency. As stated earlier, as many as 100 aircraft can simultaneously interrogate the beacon on its single channel frequency. It is quite possible that two aircraft may interrogate at the same time and with the same pulse rate. If this does occur, each of the aircraft may receive conflicting information. However, the chances of such an occurrence are highly improbable.
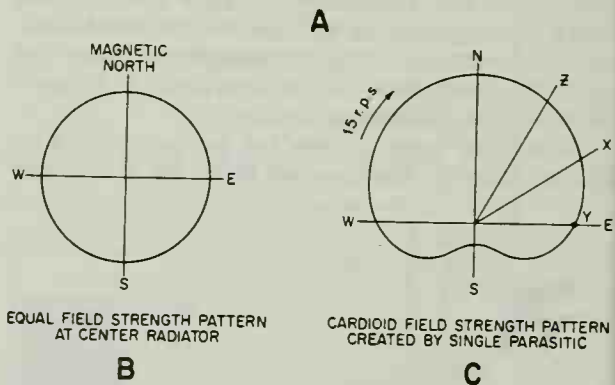
ANTENNA CONSTRUCTION

Bearing information originates from the antenna in the radio beacon. A single stationary vertical antenna (fig. 11-3, A) provides an omnidirectional pattern, such as shown in figure 11-3, B. A 15-cps parasitic element is properly spaced from the antenna so as to deflect the r-f energy to produce the cardioid pattern of figure 11-3, C. The parasitic element is motor driven so that it encircles the antenna at a rate of 15 revolutions per second. Thus, the parasitic element of the antenna modulates the total radio beacon output with a subaudio (15 cps) frequency component.

Mounted concentrically with the 15-cps antenna element is an arrangement of nine vertical reflectors embedded in an outer fiberglass cylinder, which rotates simultaneously with the inner cylinder at the 15-rps rate. Mechanically linked to each of the rotating cylinders, and mounted horizontally below the central array is an assembly referred to as the pulser plate (fig. 11-4, A). On the upper edge of the pulser plate (fig. 11-4, B) there is one soft iron slug. On the lower edge of the circular plate there

ANTENNA CUTAWAY

A
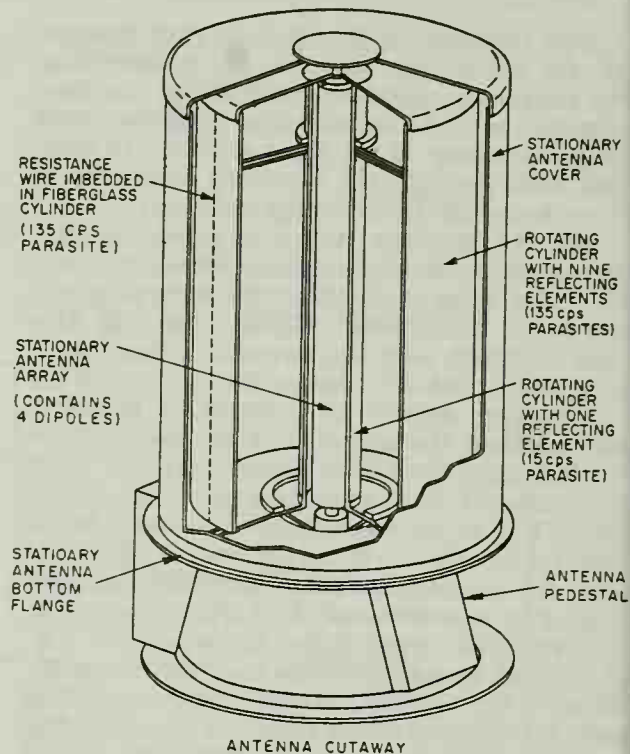
EQUAL FIELD STRENGTH PATTERN AT CENTER RADIATOR

B

CARDIOID FIELD STRENGTH PATTERN CREATED BY SINGLE PARASITIC

C

A. CONSTRUCTION OF ANTENNA AND ANTENNA PARTS.
B. EQUAL FIELD STRENGTH PATTERN OF CENTER RADIATOR.
C. CARDIOD FIELD STRENGTH PATTERN CREATED BY SINGLE PARASITIC.

32.74

Figure 11-3.—Antenna construction and antenna patterns.

are eight similar slugs. As the pulser plate rotates, the upper and lower slugs of the plate pass through air gaps of pickup coils. As the slug passes through the air gap of the pickup coil, the net reluctance of the magnetic circuit changes, and a pulse is induced in the coil
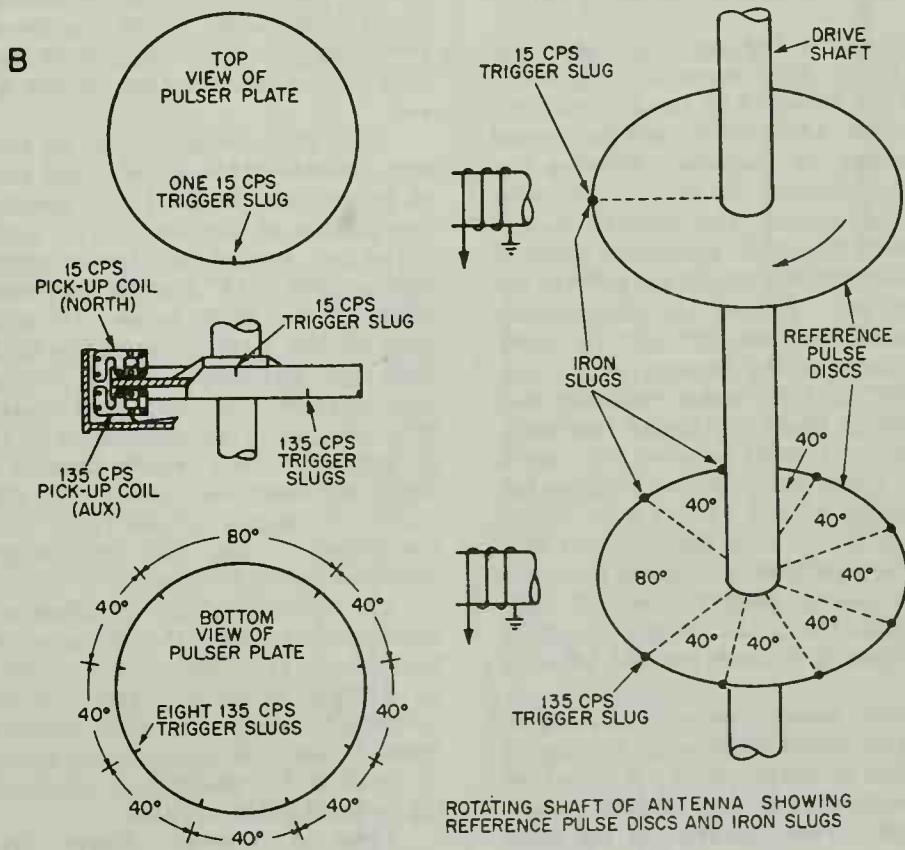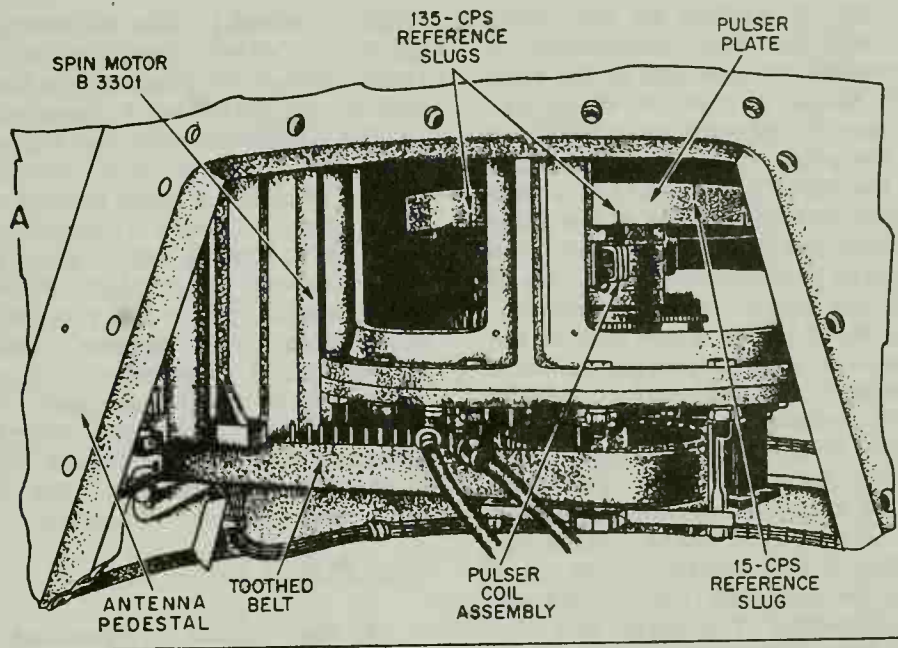
Figure 11-4.—Reference burst pulser assembly.

32.75

winding. One coil is excited by the upper slug (15 cps) once for each revolution. A second coil is excited by each one of the eight lower (auxiliary) slugs, in turn, at 40° intervals, except for one 80° interval when the space is occupied by the pulse from the upper slug. The pulse from the upper slug occurs at a time between the eighth and first pulse of the outer cylinder. The eight auxiliary slugs generate a total of eight cycles per revolution at a rate of 360°/40° x 15 cycles or 135 cycles per second.

The pulse produced by the upper slug on the pulser plate is used to establish magnetic north as a reference point in time. The pulse triggers the generation of a coded burst, each pulse of which triggers the transmitter r-f output from the beacon antenna. As stated above, the 15-cps pulse is generated once during each revolution of the reflector. The pulser plate causes the generation of the 15-cps trigger pulse each time the peak of the radiation lobe points due east (an arbitrary point). The pulse thus obtained causes the generation of the north reference burst.

The time difference between the reference point on the 15-cps sine wave and the north reference burst is detected by the aircraft receiver of Radio Set AN/ARN-21 and converted to a meter reading in degrees, showing the bearing from the aircraft to the beacon with respect to magnetic north. The aircraft measures bearing from the north reference point to the next positive-going half amplitude point of the modulation envelope. Figure 11-5 illustrates the signal received by four aircraft (in their respective positions as the beacon signal rotates through 360°) and the meter readings that would correspond to their particular bearings. Note that all the aircraft receive the north reference burst at the same instant of time but at a different phase of the 15-cps a-m wave.

The 135-cps signal alters the cardioid pattern to form a new pattern, as shown in figure 11-6. Thus, the antenna pattern basically consists of a large lobe (from the 15-cps director) and 9 smaller lobes (produced by the 135-cps directors).

The total radio beacon output consists of the internally generated reference-frequency modulated or keyed by pulse-pairs. The pulse-pairs are transmitted at a rate of 3600 pulse-pairs per second. Nine hundred of the total pulse-pairs per second are the 15-cps and the 135-cps reference bursts. (The reference bursts are initiated by the 15-cps and 135-cps

trigger pulses.) The additional 2700 pulse-pairs are either identification code distance reply pulses or random squitter pulses generated to maintain a constant duty cycle.

The composite radio beacon pattern is shown in figure 11-7. In 1/15 second (1 cycle of 15-cps signal), 12 north reference burst pulse-pairs occur. Over a 1-second period, 12 x 15 pulse-pairs occur, which total 180 north reference bursts per second. Seven hundred and twenty auxiliary (135 cps) reference burst pulse-pairs occur each second. These pairs are divided into 6 pulse-pairs occurring 120 times per second (8 times per cycle). Random pulse-pairs (totaling 2700 pulse-pairs) are inserted between the reference bursts. Identity pulse-pairs are substituted for the random pulse-pairs once every 37.5 seconds.
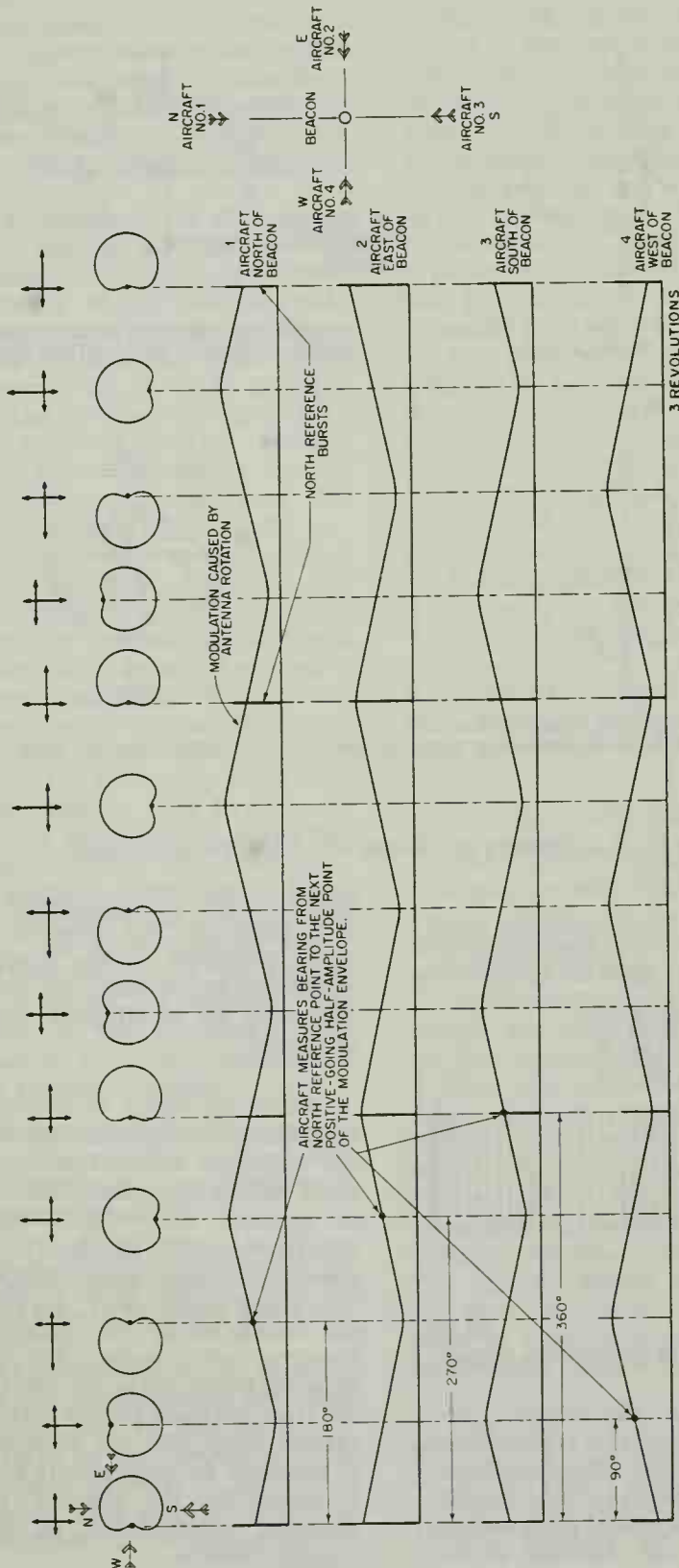
Transmission Priorities

Of the signals transmitted by the radio beacon, all except distance-measuring information are initiated within the beacon. Distance reply signals are supplied by the radio beacon only when interrogated by the airborne radio set.

All signals transmitted by the radio beacon are characterized by the fact that they consist of pulse-pairs, with 12-μs spacing between the two pulses of the pair. The number of pulse-pairs per second and the spacing between the pulse-pairs (the spacing between the leading edge of the first pulse-pair and the leading edge of the first pulse of the next pair) depend upon the particular signal element and is a characteristic of that particular signal only. However, it is the spacing of 12 μs between the pulses of a pair which provides the aircraft radio set with the means of distinguishing between the signal pulses from the radio beacon and other pulses that may be present on the received radio frequency.

The three elements of information (bearing, identification call, and distance-measuring information) are transmitted by the radio beacon as a train of pulse-pairs. To prevent interference between the three signal elements, the radio beacon is equipped to assign a sequence of priority by which each element enters into the overall pulse train.

First in order of priority are the bearing reference bursts. Because they occur at a fixed rate, they are of relatively short duration and convey the most important portion of

Figure 11-5.—Meter readings for various aircraft positions.

385

HORIZONTAL PATTERN OF 15 CYCLE
FUNDAMENTAL MODULATION

HORIZONTAL PATTERN OF 135 CYCLE
NINTH HARMONIC MODULATION ONLY

HORIZONTAL PATTERN COMBINATION OF
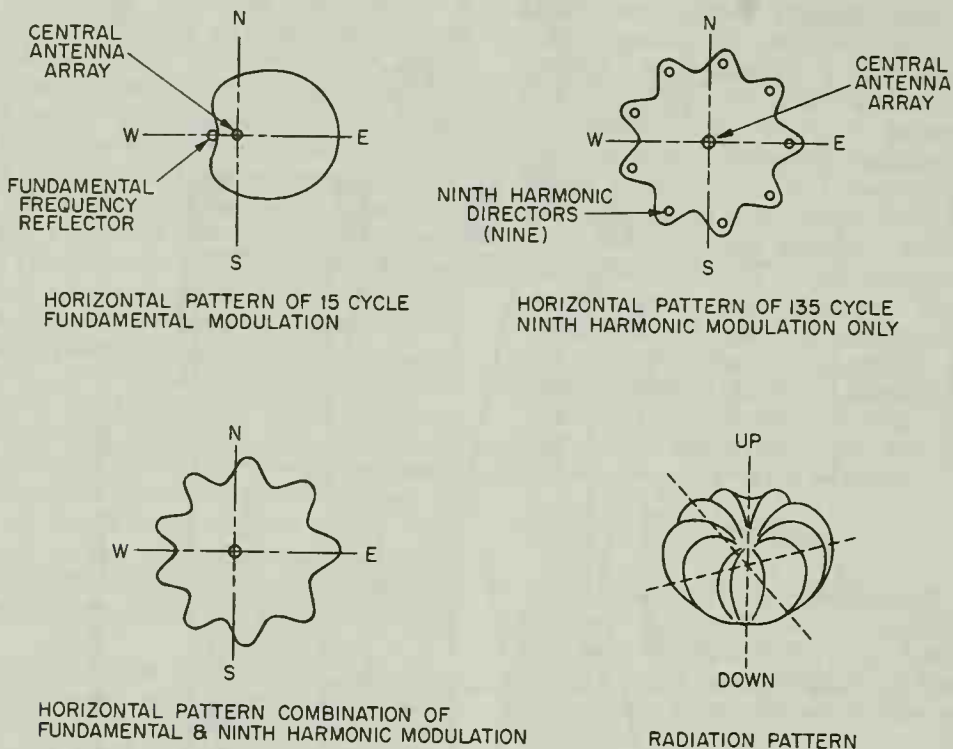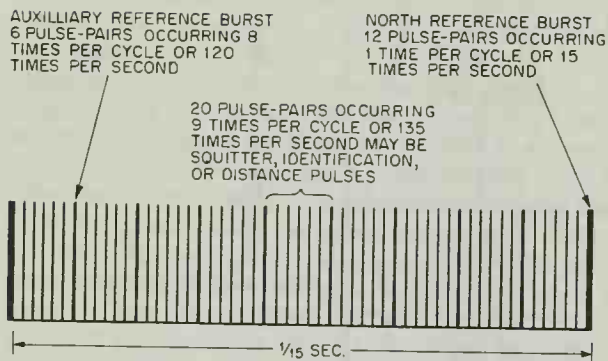FUNDAMENTAL & NINTH HARMONIC MODULATION

RADIATION PATTERN

32.78

Figure 11-6.—Effects of 15-cps and 135-cps modulation.

32.80

Figure 11-7.—Composite radio beacon signal.

the information supplied by the beacon. Accordingly, reference bursts will be transmitted regardless of the number of interrogations.

Second in order of priority is the identity call signal. The identity call signal also occurs at a fixed rate. However, the duration of the identity call is sufficiently long, so that the utilization of the portion of the total duty cycle required for the insertion of the bearing reference burst does not materially affect the reception of intelligible identity code by the aircraft.

Third in order of priority is the distance information signal. The interrogation pulses are not initiated within the beacon, and their arrival within the pulse train must be controlled to prevent their interfering with the other signal elements of the radio beacon. This is done by allowing the interrogation pulses to enter the pulse train only during a time interval not occupied by the reference burst or the identity call signals. The period of the reference burst pulse is small in comparison to the time between pulses of the bursts. A memory circuit built into the aircraft receiver makes it possible to compensate for the time interval occupied by the identity pulses so that no apparent interruption of distance-measuring signals results.

Last in order of priority are "squitter" pulses consisting of random noise pulses, which are used as fill-in signals when the maximum number of aircraft are not interrogating the beacon. This is necessary to maintain the modulation pattern and to permit a constant duty cycle.

The functional arrangement of the major units of the radio beacon is illustrated in figure 11-8. Pertinent waveforms are shown at the input and putput of the stages to aid in understanding the overall system operation.

A single antenna is used for both transmission and reception. This is made possible through the use of the control-duplexer. Distance interrogation pulse-pairs are received by the antenna and passed through a coaxial switch (fig. 11-8A) along a coaxial line, and through the control-duplexer to the radio receiver. In the receiver, the received signal passes through a low-pass filter (fig. 11-8B) to the hybrid mixer stage of the receiver. The low-pass filter, rejects frequencies between 1,650 and 10,000 mc to prevent interference from other pulse-producing equipments operating in or near the beacon frequency range.

The local oscillator signal for the mixer stage is derived from the frequency-multiplier section of the radio beacon. The basic local oscillator frequency is generated in a crystal oscillator and doubler stage. The oscillator-doubler output is further multiplied in three stages of frequency multiplication which include the second doubler, the third doubler, and a tripler stage. The tripler output is applied simultaneously to a keyed r-f amplifer and to the mixer stage in the receiver. The heterodyning process in the mixer produces a 63-mc intermediate frequency signal.

The mixer output is amplified in three stages of preamplification. The output is subsequently amplified in the i-f stages to a level sufficient to drive the Ferris discriminator.

The Ferris discriminator detects the interrogation and random noise pulses from the 63-mc i-f signal and effects adjacent and near-adjacent channel rejection.

The video output is applied to the blanking gate state. During the transmission of output pulses by the transmitter of the beacon, a 10 μs pulse generated in the frequency multiplier-oscillator video chassis is applied to the blanking gate stage. This pulse disables the receiver for the duration of the transmission.

The output of the blanking gate stage is amplified in the final video amplifier, and fed to a coincidence decoder circuit. The decorder produces a single pulse for each input pulse-pair which has the pulses of the pair separated by 12 microseconds. The decoding of each pulse-pair is followed by a 40 μs blanking condition in the receiver to ensure that the 2700 pulses at the receiver output will be distributed over a period of one second. The delay line ensures the generation of one pulse for each input pulse-pair.

The pulse output of the coincidence decoder triggers a one-shot multivibrator, which produces three outputs. One initiates the blanking gate pulse in the blanking-gate generator. This action is described later. A second is applied to a pulse amplifier. The third is fed to the receiver output cathode follower and is applied to the coder-indicator chassis (fig. 11-8, C).

Pulse amplifier V408, pulse counter V409, d-c regulator V407A, and d-c clamp stage V409A (fig. 11-8B) are included in a circuit referred to as a "squitter control circuit." This circuit produces a negative (squitter) voltage which is fed back to the i-f amplifiers, V301 and V302.

CONTROL-DUPLEXER

The control-duplexer consists of a passive duplexer network that permits the transmitter output signal and the receiver input signal to be fed through the same transmission line and antenna (fig. 11-8A) via the switch test adapter. Control circuits located in the control-duplexer permit the radio set to be operated by switches on the front panel of the control-duplexer.

The control-duplexer output signals are routed through a Switch Test Adapter, SA-420/URN-3. The switch test adapter consists of a heavy-duty coaxial switch, S6151, and pickoff diodes, CR6151 and CR6152. The switch, S6151, transfers the radio beacon transmitter output from the antenna to the dummy load as desired. The pickoff diodes supply detected signals to the test equipment. A biasing voltage is applied to CR6152 to permit power measurement by the slide-back method.

The pulsed r-f transmitter output from the Klystron r-f amplifier, V1304, (fig. 11-8B) is fed through matching slug Z1303 to a transmission line r-f filter (Z1156 or Z1157, fig. 11-8A). Each of the filters consists of a pair
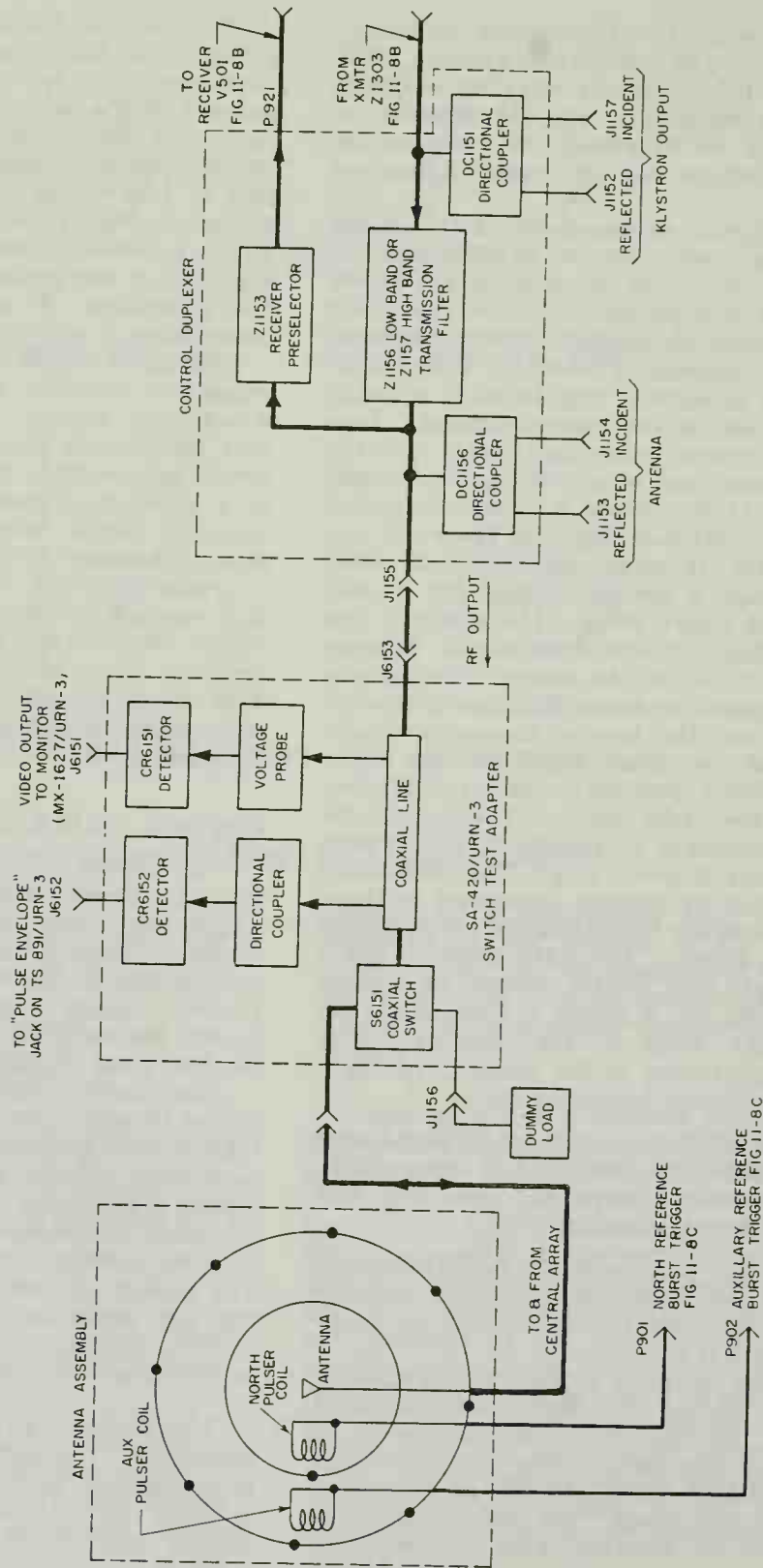
32.81

Figure 11-8.—Radio beacon, block diagram. A. Antenna assembly, switch test adapter and control duplexer.

32.81

Figure 11-8.—Radio beacon, block diagram. B. Transmitter-receiver.

Figure 11-8.—Radio beacon, block diagram. C. Coder-indicator.

32.81

of tunable resonant cavities. These filters are inserted in series with the transmission line to limit the overall spectrum of the transmitter output.

The transmission line filter is located in the line between the transmitter and the antenna, with the receiver input connection branching off on the antenna side of the filter. The receiver input path also contains a filter, Z1153. The transmission line filters (Z1156 and Z1157) are tuned 63 megacycles, one above, and the other below the receiver frequency, so that Z1153 (which is tuned to the receiver frequency) appears as an open circuit at the transmitter frequency. Thus, the transmitter output is fed directly to the antenna.

Samples of the klystron-incident and klystron-reflected fields and of the antenna-incident and antenna-reflected fields are made available (40db down), by directional couplers, DC1151 and DC1156, respectively. During tuning, the klystron output may be obtained at J1152 and J1157, while the antenna output can be obtained at J1153 and J1154.

The control-duplexer contains a desiccator (drying agent) that controls the humidity of the air in the filter cavities (Z1156 or Z1157 and Z1153). Because of temperature changes, the filter cavities tend to breathe (draw in and expel air). This air exchange takes place through the desiccator, with the desiccant absorbing moisture from the air, thereby maintaining a relatively dry atmosphere in the cavities.

RADIO RECEIVER

The radio receiver (fig. 11-8B) with the exception of the special circuits (Ferris discriminator, blanking gate and squitter control), is a conventional superheterodyne type.

Ferris Discriminator

The Ferris discriminator circuit, V306 (fig. 11-9), receives its input signal from the final (fifth) i-f amplifier, V305. The function of the Ferris discriminator is to detect the signal and provide adjacent and near-adjacent channel rejection.

The input signal to the Ferris discriminator is coupled to two tank circuits. Low impedance points on the tank circuits are used to feed the discriminator diodes, V306A and V306B, respectively. In this way, the tube capacitance is shunted across a smaller portion of the total tank impedance. Thus, the discriminator is relatively unaffected by variations in tube capacitance. In addition, the circuits are temperature compensated to maintain frequency stability.

Both of the tank circuits feeding V306 are tuned to 63 mc. Because of the loading difference imposed on the tanks, the "Q" of the tank feeding V306B is reduced to approximately one-third the "Q" of the tank feeding V306A. Each half of V306 had a different frequency output characteristic because of the differences in the signal developed by each tank circuit.

Since the "Q" of the tank feeding V306A is higher than that of V306B, the negative voltage developed by V306A (fig. 11-10) is greater than the positive voltage developed by V306B (fig. 11-10) at the same frequency (63 mc). Therefore, only the negative-going narrow band of frequencies centered at 63 mc are passed.

The algebraic sum of the V306 output voltage is represented in figure 11-10. The normal V306 output is negative interrogation signals, within the 63-mc i-f pass band. Note the positive voltages that appear at the output of V306 for adjacent channel frequencies.

Clamping diodes V308A and V308B (fig. 11-8, B) effectively clamp the grids of V307A and V307B respectively, to ground and thereby eliminate the adjacent channel voltages.

The output of the Ferris discriminator and video amplifier consists of interrogation pulse-pairs spaced 12 μs apart.

The video signal (interrogation pulse-pairs) and noise pulses from V307B are routed to the first video amplifier, V402A (fig. 11-8,B). A portion of the input signal is fed to V405A, which serves as an output stage to provide a portion of the video signal derived from the Ferris discriminator as it appears before decoding.

The video signal is amplified in V401A and coupled to the blanking gate stage, V401. The gating action of this stage will be considered later.

The output of the blanking gate stage is coupled to the final video amplifier, V402B. The output is applied directly to the suppressor grid and through a 12 μs delay line to the control grid of the decoder tube, V403. This tube is biased at cutoff, which requires the simultaneous application of positive voltages on both the control and suppressor grids for conduction. Since the decoder is biased at cutoff, the pulse input must be of sufficient amplitude to cause conduction. This means
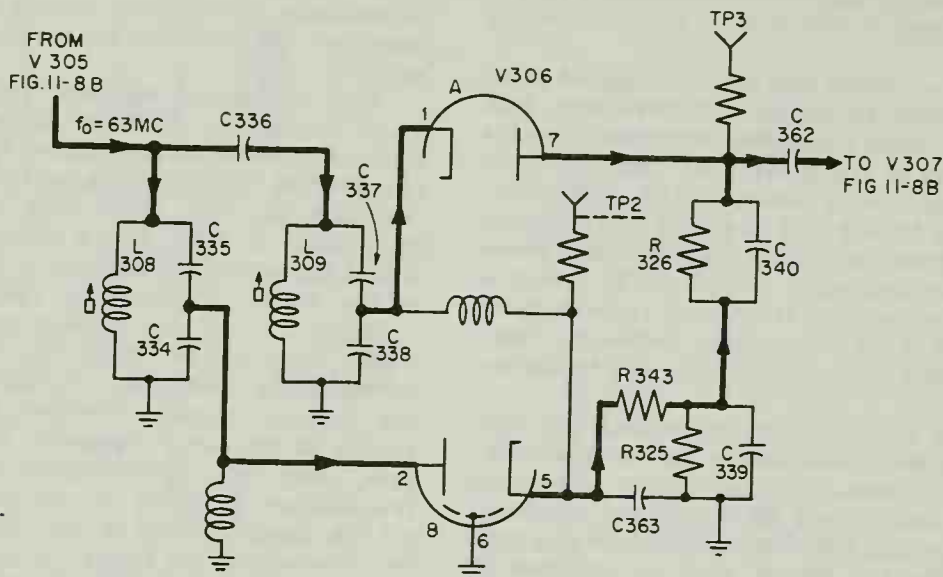
Figure 11-9.— Ferris discriminator.

32.89

that not all decodable noise or interrogation pulse-pairs will be decoded, but only those above a minimum value. The receiver is designed to maintain the number of decoded pulse-pairs at 2700 ± 90 pulses per second.

The decoder output, a single negative-going pulse for each pulse-pair, triggers a one-shot multivibrator, V404. Triode section V404A normally conducts and the negative input from V403 cuts it off, initiating the multivibrator action.

Tube V404 provides three output signals; one feeds the radio receiver blanking one-shot multivibrator, V406, the second is coupled to a pulse amplifier, V408, in the radio receiver squitter-control circuit, and the third feeds cathode follower stage V405.

The output of V405B is a video pulse with an amplitude of about +30 volts and a duration of between 3 and 4μs. This is the output of the receiver which is applied to output jack J504 for connection by inter-unit cabling to the keyer circuits of the coder-indicator. The output is also available at jack J407 for test purposes.

Blanking Gate Circuits

The blanking circuit (fig. 11-11) performs a dual purpose. It controls the decoder input circuit for 40μs after a pulse-pair has entered

the circuit, and eliminates undesirable feedback from the klystron during transmission.

The B section of the blanking multivibrator, V406, is normally held beyond cutoff. The A section normally conducts, since cathode bias alone controls the current through this tube section.

The negative-going output from the plate of V404B (fig. 11-8, B) cuts off V406A (fig. 11-11), initiating multivibrator action. The positive output at the cathode of V406B is coupled to the blanking phase inverter and isolation amplifier, V407B. Because V407B receives a high value of bias (-105 volt supply), conduction takes place only on strong positive pulses from V406B. The negative output pulse is fed to the suppressor grid of the blanking gate tube V401.

It should be recalled that the second pulse of the input pulse-pair caused the triggering action at the V404 multivibrator. Likewise, the trigger to the blanking multivibrator, V406A, corresponds in time with the second pulse of the input pair. The blanking pulse to V401 ensures the cutoff of this stage for 40μs from the leading edge of the second pulse of the pair.

An additional blanking time in V401 (fig. 11-8, B) is caused by a positive 10μs pulse received from the frequency multiplier-oscillator. This pulse ensures blanking of the

A
V306A OUTPUT CHARACTERISTIC

B
V306B OUTPUT CHARACTERISTIC

ADJACENT CHANNEL
FREQUENCIES

C
V306A AND V306B COMBINED
OUTPUT TO V307A
FIG.II-8B

63 MC IF BAND

63 MEGACYCLES

32.90

Figure 11-10.—Ferris discriminator, c-w
frequency response characteristics.

receiver during transmission of a pulse from
the beacon transmitter.

Since each pulse is approximately $3 \mu s$
wide, and the blanking gate input corresponds
in time to the second pulse of the pair, the
elimination of the first pulse in the pair (by
the decoder circuit) adds $3 \mu s$ to the total time
between the arrival of the leading edge of the
first pulse to V401 and the leading edge of the
next pulse. Thus, the total time between pulses
at the output of the receiver, is equal to the sum
of the $40 \mu s$ delay in V401, the $10 \mu s$ delay
caused by the pulse from the frequency multi-
plier - oscillator, the original $12 \mu s$ between
pulses of a pair, and the $3 \mu s$ created by the
omission of the first pulse - pair, or
$65 \mu s$.

Squitter Control Circuit

The output of V404B multivibrator cathode
(fig. 11-8,B) which is also applied to the pulse
amplifier, V408 (fig. 11-11), is used to derive
the squitter-control voltage. Squitter control
involves the action of a circuit in its use of
noise pulses to maintain the total pulse-pairs
in the receiver at 2700 + 90 pulses per second.

Grid limiting is permitted at the input to
V408, which renders the output of this stage
substantially independent of the amplitude of the
input pulse. The output of V408 is coupled to the
pulse stretcher and counter diode, V409B. As
V409B conducts only on negative input alter-
nations, C415 charges negative toward the plate
of the diode. Capacitor C415 holds most of its
charge during positive alternations at the V409
cathode. As a result of this action, the voltage
on the grid of V407A is essentially a negative
d-c voltage.

The potential at the cathode of V407A is ap-
plied to the first and second i-f amplifiers
(V301 and V302) as squitter (gain) control
voltage. When the bias voltage on the grid of
V407A is correct to effect a squitter rate of
2700 ± 90 pps, the voltage of the cathode of
V407A is approximately -5 volts with respect
to ground.

Since without the application of the bias
voltage to the grid of V407A (when the equip-
ment is initially turned on), the voltage drop
may result in a high positive voltage (in ex-
cess of $+20$ volts d-c) applied to the i-f
amplifiers, a d-c clamp diode, V409A, op-
erates to prevent the voltage rising above
$+5$ volts while the receiver is warming up.
Without the clamping action of V409A, the
$+20$ volts applied to the grids of the amplifiers
will completely block the receiver, and may
cause damage to the receiver tubes.

Power Supplies

The radio receiver power supplies pro-
vide regulated -105 volts d-c for the video
amplifier circuits, and regulated +150 volts
d-c for the complete receiver.

Metering Circuit

A meter and a 6-position meter selector
switch, located on the receiver front panel are
used to permit the operator to observe the op-
erating condition of the mixer crystal diodes,
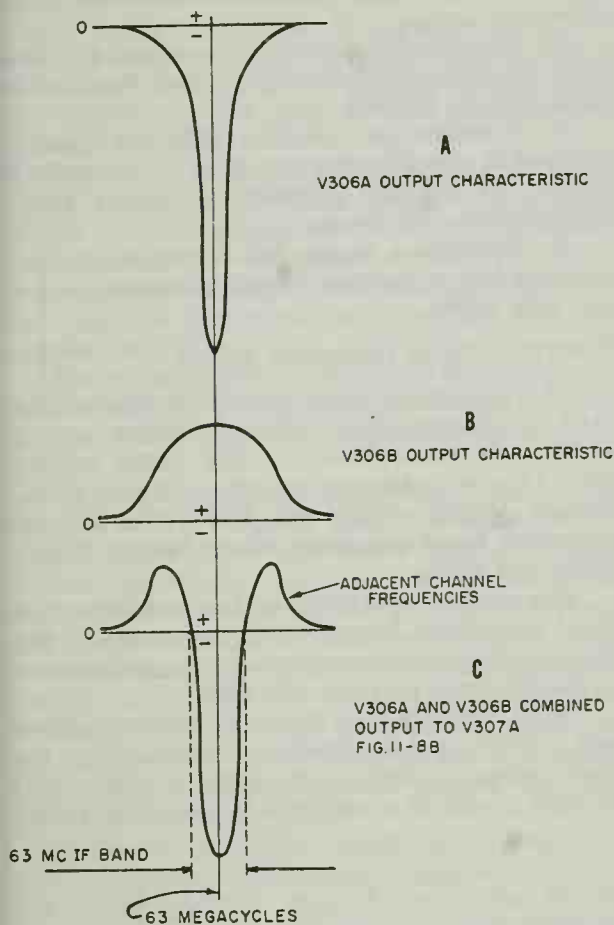the B (+200 volt) and C (-200 volt) master

power supplies, and the squitter-control output.

CODER-INDICATOR

The coder-indicator of the Radio Beacon, AN/SRN-6 performs the following functions:

1. Generates the 15-cps reference burst each time a 15-cps trigger pulse is generated in the antenna.

2. Generates the 135-cps reference burst each time a 135-cps trigger pulse is generated in the antenna.

3. Generates a radio beacon identification call of 1350 cps in International Morse Code keying for transmission at specific intervals.

4. Processes the distance interrogation pulses appearing at the receiver output, and in combination with time delays in the receiver and transmitter, adjusts their net transit time through the beacon to exactly 50 μs.

5. Assigns priorities of transmission to the components of the signal, which consists of the bearing reference bursts, the radio beacon identification call, and the replies to distance interrogations and random noise pulses. Generation of the reference burst component has priority over the generation of all other output components.

6. Combines the various components of the signal.

7. Encodes the transmission signals to give them the characteristics identifying them as the radio beacon signals.

8. Provides an accurate 1350 cps signal to be used in checking the accuracy of the speed at which the antenna parasitic elements rotate about the central array.

9. Provides a means for adjustment to compensate for variations between magnetic north and true north.

Reference Burst Generator Stages

The reference burst generator stages (fig. 11-8,C) produce the reference burst signals which are transmitted by the radio beacon. The 15-cps reference burst generator stages include V601B, V602, and V603. The 135-cps reference burst generator stages include V601A, V609, and V610.

The circuit operation of the two reference burst generator stages is the same except that the size of the frequency generating components in the separate circuits will differ.

The trigger pulses from the antenna contain both positive and negative excursions. When the input pulse to V601A is positive-going, grid current produces a net voltage drop at the V601A
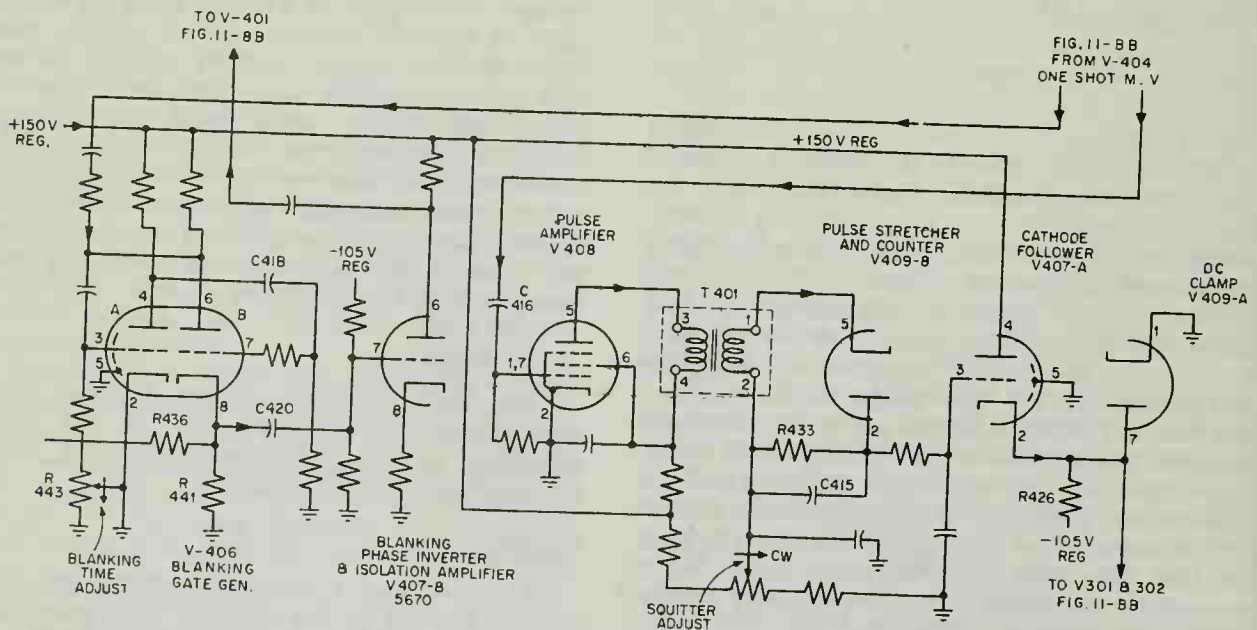

Figure 11-11.—Blanking generator and squitter control (pulse counter).

32.89

grid which is held very near to zero volts. When the trigger pulse is negative-going, the grid is negative, and the resulting output of V601A is a sharp positive-going pulse, which is coupled to the grid of the 135-cps gate generator, V609.

Tube V609 is a one-shot keying multivibrator which is used to provide a keying or gate pulse to the pulsed oscillator V610A which permits the generation of six pulses of 41.7 kc. These pulses, after amplifying and shaping, represent the 135-cps reference burst pulses and are coupled to V604A.

The pulse output of V609 plate is also coupled to the mixing network with the 15-cps gate and to the control grid of the priority gate state, V611.

The 33.3 kc pulsed oscillator, V603, in the 15-cps reference burst circuit also applies its output to the same reference burst amplifier, V604A. A single amplifier can be used for both the north and auxiliary reference burst since these signals do not occur simultaneously.

Burst amplifier V604A operates at approximately zero bias and is therefore normally conducting. This stage inverts and amplifies the oscillator (V610 or V602) output to an amplitude which is sufficient to trigger the one-shot multivibrator, V605.

Blocking oscillator V606 is employed to produce sharp voltage spikes capable of driving the encoding delay line, DL 601. The V606B section provides two outputs; one of 6 positive-going spikes which are applied to delay line DL 601, and a second of 6 negative-going spikes for each reference burst trigger (at V610A), as a synchronization control for the 135-cps gate generator stage, V609.

The 15-cps reference burst generator circuit, consisting of V601B, V602, and V603, is designed to produce 12 pulses spaced $30 \mu s$ apart. The pulsed oscillator, V603 operates at 33.3 kilocycles and is gated by the cathode output of V602B. Diodes in grid of V604A isolate the 15-cps and 135-cps pulsed oscillators. The feedback circuit from V606B to the V609A gird drives V609A further into cutoff during the period of the 12 input pulses developed by V605 to ensure that the 135-cps gate generator does not operate during this time.

Encoding Delay Line and Output Amplifiers

The combined action of the 135-cps reference burst circuits is to produce a group of bearing reference pulses, which consist of 6 pulse-pairs with 12 μs spacing between pulses of a pair, and $24 \mu s$ between pulse-pairs. These pulses recur 120 times per second at a 135-cps rate, as determined by the position of the soft iron (auxiliary) slugs on the pulser-plate assembly. The 15-cps reference burst circuits must produce 12 pulse-pairs with a $12 \mu s$ spacing between pulses of a pair, and $30 \mu s$ between pulse-pairs. The pulses recur 15 times per second as determined by the north slug on the pulser-plate.

The output at the blocking oscillator V606B cathode (fig. 11-8C) is about 166 volts peak-to-peak, and is fed to the primary of autotransformer T603 (not shown). This transformer matches the impedance to the encoding delay line, DL601, and has a step-up ratio of 1:3. Thus, the amplitude of the input signal to the delay line is approximately 500 volts peak-to-peak. The encoding delay line introduces the major portion of the standard zero distance delay for distance reply pulses.

For each pulse input to the encoding delay line, the delay characteristics of the line produces 2 positive pulses spaced 12 μs apart. These pulses are delivered to the pulse amplifier, V607A. These pulses are tapped from the 5 and 3 terminals of DL601, and have an amplitude of approximately 15 volts peak-to-peak. When an input pulse is applied to the line, the initial pulse from the line is delayed $32 \mu s$, while the second pulse is delayed $44 \mu s$.

The negative-going pulses appearing at the plate of V607A are coupled to the grid of the A section of one-shot multivibrator V615.

The V615B cathode output consists of the shaped reference burst pulses (either north or auxiliary) which are coupled to cathode follower V607B. The action of V615 ensures uniformity of the amplitude and width of the reference pulses.

Cathode follower V607B functions to isolate V615 from the output, and to match the output circuit to the 50-ohm line which feeds the signal to the transmitter. A portion of the output of V607B is delivered to test jack J607 on the front panel of the coder-indicator.

Identification Code Circuits

The circuits comprising V612, V613A, V614, V613B, and CR603 (fig. 11-8C) make up the 1350-cps tone generator, the tone of which is keyed to produce the identification code

characters for the radio beacon. Identification call amplifier stage, V613B, actually produces 2 pulses every 740μs, or at a rate of 1350 cps.

The identification tone output of V613B is keyed to V604B and V611, amplified and shaped in V605, V606, DL601, V607A, V615, and V607B before it is fed to the frequency multiplier-oscillator of the radio beacon. You will remember that the encoding delay line, DL601, produces 2 pulses spaced 12μs apart for each pulse applied at the input to the line. Thus, during a 1-second period of the identification call, 1350 pairs of 100μs-spaced pulses appear at the input to the encoding delay line, with 2700 pulse-pairs at the DL601 output. The resulting beacon pulse output is maintained at the required rate.

The operation of the identification call circuits is as follows: The oscillator tone (1350cps) is generated in V612. A 135-cps synchronization signal obtained from the B section plate of the 135-cps gate generator stage (V609B) is applied to the control grid of V612A. The signal is amplified and applied as a positive (regenerative) synchronization input to the oscillator tank circuit. The tank circuit actually operates at the tenth harmonic of the synchronization input. The 1350-cps output of V612A is the basic identification tone signal. The circuit of V612B is that of a cathode follower. The 1350-cps output of the oscillator is coupled to amplifier, V613A. Negative alternations of the 1350-cps signals at the grid of V613A are amplified in this stage, and coupled to the A section grid of multivibrator, V614.

Twin triode V614 is connected as a one-shot multivibrator. The purpose of this stage is to produce 2 output pluses for each pulse applied at its input. Thus, the output pulse rate of the identification stages during the transmission of the identification tone is 2700 pulses per second, which are grouped at a 1350-cps rate. These pulses are encoded (double-pulsed) and amplified in the shaping stages (V605, V606, and DL601) so that the output pulse rate from the transmitter during the transmission of identification call is maintained at 2700 pulse-pairs per second.

Diode CR603 (fig. 11-8C) functions as a clamping diode which presents either a low or high impedance path to ground for the identification call signals depending on the position of the switches (S607 open, and S604 closed) in the identification call mechanical keyer assembly, A602. When the microswitches (S607

and S604) are closed, ground is applied to the plate side ("P") of CR603. When "P" is grounded, CR603 becomes a low impedance path to ground for the negative pulses which are fed to the grid of V613B. Thus, the tone signal is grounded out through the diode, and does not appear at the V613B grid. When the ground is removed from the diode, by the action of the keyer wheel assembly, a negative bias voltage of approximately -50 volts is applied to plate "P" of the diode. With -50 volts on the plate of CR603, the diode acts as a high impedance, and the tone signals are developed at the V613B grid. Under this condition, V613B passes the 1350-cps pulse-pairs to the priority gate stage, V611.

The keyer assembly, A602, comprises a motor-driven code keying wheel, a code start timing cam, and two microswitches, S607 keying switch and S604 timing switch. The keyer assembly in conjunction with the gate blocked distance reply amplifier, V604B, controls the selection of either identification tone distance reply pulses as input signals to the priority gate stage, V611. The keying wheel and start code timing cam initiate a beacon identification coded call every 37.5 seconds.

The code keying wheel revolves at a speed of 8 rpm. The periphery of the wheel is divided into segments. The segments can be moved outward from the center of the wheel. By positioning the segments, code characters of long or short duration (dots and dashes) can be set on the wheel.

A 1350-cps antenna synchronization oscillator, V608, produces an output voltage, which in conjunction with a signal voltage from the antenna speed tachometer (not shown) can be used to check the speed of rotation of the antenna.

The tachometer produces a 675 cps output. The two voltages will cause a 2 to 1 Lissajous pattern to be formed on an oscilloscope screen when the antenna speed is correct.

TRANSMITTER

Actually, the transmitter can be considered to contain a frequency multiplier-oscillator chassis and an amplifier-modulator chassis. It should not be forgotten however, that the local oscillator signal for the beacon receiver is also obtained from the frequency multiplier-oscillator chassis.

Frequency Multiplier-Oscillator

The frequency multiplier-oscillator chassis of the radio beacon comprises a carrier generating (r-f) chain and a video chassis. The carrier generating chain produces the beacon radio frequency carrier. The video chassis produces two shaped pulses, one pulse, called the klystron gate pulse, is $10\mu s$ wide and is amplified in succeeding stages of the transmitter to a level sufficient to key the high voltage applied to the klystron. The second pulse, called the $3.5\mu s$ shaped pulse, is a specially shaped video pulse which is used to intensity-modulate the klystron beam current. This pulse has a relatively long rise and decay time in order that the final r-f output pulse will occupy a minimum of frequency spectrum.

SHAPED PULSE GENERATOR.—The positive $1.5\mu s$ pulses from the coder-indicator chassis are applied via J1401 to the oneshot multivibrator, V1402, of the shaped-pulse generator circuit, in the frequency multiplier-oscillator chassis (fig. 11-8B). Multivibrator V1402 ensures uniformity of shape of the input pulses. To prevent false triggering action, CR1402 is inserted in series with the signal path of the trigger pulse to block any negative alternations.

Power amplifiers, V1403, V1404, and V1405 are operated in parallel in order to handle the heavy currents required in producing the high-powered pulse necessary to drive the subsequent shaping circuits. The primary of a step-up transformer (not shown) acts as the common plate load of the power amplifiers. Fixed bias cuts off the power amplifiers in the absence of a signal.

When a positive pulse from V1402B is applied to the power amplifiers, plate current flows and a negative-going rectangular pulse approximately 800 volts in amplitude is developed across the primary of the transformer. A positive pulse, approximately 2300 volts in amplitude is induced in the secondary of the transformer, and is applied to the shaping network through the series diode, V1406.

The shaping network comprises capacitors and coils in a low-pass filter circuit, which produces a $3.5\mu s$ pulse. The delay time of the pulse throughout the shaping network (fig. 11-8, B) is approximately $2.8\mu s$.

Positive output pulses from the shaping network are coupled to the paralleled grids of the cathode followers, V140 and V1411. The cathode followers are operated in parallel to lower the output impedance of the $3.5\mu s$ pulse shaping network.

The output of V1407 and V1411 is fed to the amplifier-modulator chassis (fig. 11-8, B). The signal is applied to the klystron amplifier, V1304. You will note that in the frequency multiplier - oscilator, each pulse from the coder-indicator is shaped before it is applied to the klystron amplifier. The space between pulses is maintained at $12\mu s$

In addition to applying the $3.5\mu s$ pulse to the klystron, the r-f carrier must also be applied at the same time to produce the required beacon output. The r-f carrier frequency is produced in the carrier generating chain by a Butler type crystal-oscillator, V151. The oscillator is followed by a group of frequency multiplier stages. These stages multiply the basic oscillator frequency 24 times.

The V1504 output (fig. 11-8, B) which represents the oscillator frequency multiplied 24 times is applied simultaneously to the hybrid mixer in the receiver, and constantly to the first keyed r-f amplifier, V1505. Tubes V1505 and V1506 are keyed r-f amplifiers operated in cascade and employing lighthouse tubes in conjuction with a broadband coaxial-type resonant cavity. The r-f output of V1505 and V1506 is keyed by the $10\mu s$ pulse produced in the frequency multiplier-oscillator and supplied via V1409. This negative pulse, coupled to the cathodes, allows the passage of the r-f carrier.

The $1.5\mu s$ pulses at J1401 are also fed through delay lines DL1401 and DL1402 in series to the one-shot multivibrator stage consisting of V1408 and V1401. The total input delay through DL1401 and DL1402 is about $2.8\mu s$. This corresponds with the delay of the shaped $3.5\mu s$ pulse through the shaper network to V1304. Positive input pulses from DL1402 trigger V1408 to start a $10\mu sec$ multivibrator action. Diode CR1401 blocks negative pulses coming from DL1402 which might cause false triggering of the multivibrator during the time that V1408 is conducting. Pulses from V1401 are coupled to pulse amplifier, V1409.

The positive $10\mu s$ pulse from V1401 permits V1409 to conduct, thereby producing a plate output across the primary of a transformer (not shown), which couples the $10\mu s$ output pulse from the plate of V1409 to the

cathodes of r-f amplifiers, V1505 and V1506. The pulses reduce the existing grid-cathode bias on V1505 and V1506 (fig. 11-8, B) to a point which allows these tubes to conduct. The r-f output of V1506 which consists of 10μs pulses of r-f energy at the carrier frequency is fed to the klystron, V1304 (fig. 11-8, B), in the amplifier-modulator. A second output from V1409 is developed across the unbypassed cathode and fed to the receiver blanking gate, V401, to serve as the 10μs receiver blanking pulse. This pulse blanks the receiver during the period that the transmitter is operative.

Amplifier-Modulator

The amplifier-modulator consists of the klystron r-f amplifier, V1304, a regulated bias power supply, and the associated control circuits. A sectional diagram of the klystron tube is shown in figure 11-12. The klystron operates as a three-cavity amplifier, and depends on the changes introduced in the velocity of the stream of electrons in the first and second drift spaces (velocity-modulation). The changes in the r-f potential across the klystron resonator cavities alternately slow some of the electrons and accelerate others. This action causes bunching of the electrons during one period, and creates a space between bunches during the next period. The transit time between the two electron groups (or bunches) produces an alternating r-f current through the klystron.
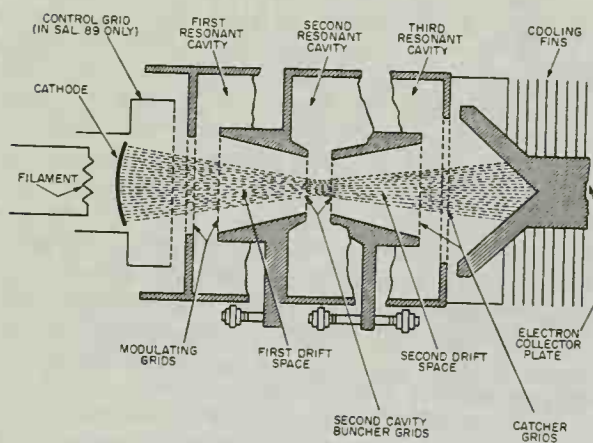
32.93
Figure 11-12.—Sectional diagram of klystron tube.

If the area of the klystron cavity is either increased or decreased, the rate of bunching in the klystron will be changed. Cavity tuning is accomplished by mechanically changing the area of the cavity by adjustable screws.

The combined effect of grid-modulation and velocity-modulation of the klystron beam current results in high-powered r-f output pulses from V1304 (fig. 11-8B). The shape of the pulse is controlled by the 3.5 μs pulse to ensure that the radiated r-f power will occupy a minimum of frequency spectrum about the assigned carrier frequency as discussed.

In the quiescent state a constant negative potential of -12 kilovolts from the high voltage power supply (not shown) is applied to the V1304 cathode. However, the klystron beam current is cut off by a -120 volt potential applied between the control grid and cathode. This potential is obtained from a series regulated bias supply which is ungrounded.

The 10 μs r-f pulse envelope from V1506 is applied to the first cavity of V1304. The 3-5 μs-shaped pulse is applied from V1407 and V1411 via T1372 to V1304 between the cathode and grid so that the peak amplitude of the pulse occurs at about the center of the 10 μs r-f pulse envelope The shaped pulse modulates the r-f to form the output pulse. Thus, the action of the klystron amplifier is like that of a coincidence stage.

The output pulses from the klystron amplifier have a minimum peak power of 7.5 kw, and a repetition rate identical to the repetition rate of the pulse train at the output of the coder-indicator. The output is coupled to a double-slug tuner, 21303, (fig. 11-8,B) which provides a means of matching the output impedance of the klystron with the input circuit impedance of the control-duplexer.

ANTENNA CONTROL SYSTEMS

The shipboard antenna system consists of the antenna proper, antenna pedestal (fig. 11-3), a speed control servosystem, a bearing servosystem, and a roll and pitch stabilization servosystem. The basic functions of the antenna were considered earlier in this chapter. The following discussion presents a block diagram analysis of the various antenna control systems.

Antenna Speed Control System

As discussed previously, the speed of the rotating cylinders carrying the parasitic reflectors must be maintained constant at 900

rpm, since the speed determines the modulating frequencies. To control this speed, a speed control servosystem is employed. A tachometer that is an integral part of the motor which rotates the antenna (spin motor) generates a 675-cps signal when the antenna cylinders are spinning at the required 900 rpm.

The speed control circuit (fig. 11-13) consists of a preamplifier and a power amplifier. The preamplifier comprises a phase-sensitive detector, an equalizing network, a voltage amplifier, and an oven control amplifier with its associated circuits. The power amplifier consists of three saturable reactors with their associated feedback rectifiers and power factor correction transformers.

The spin motor, B3202, which is geared to the tachometer, G3202, causes the tachometer to generate the 675-cps signal when the antenna cylinders are being rotated at the proper speed (900 rpm). Any deviation from this speed causes a corresponding change in the frequency of the signal from the tachometer. This frequency is fed to the input of the antenna speed control preamplifier circuit.

The phase-sensitive detector in the preamplifier contains a tuned circuit which is resonant at 675 cps. If the tachometer frequency is the same as the resonant frequency of the phase detector, no error voltage will be developed. However, if the tachometer frequency varies from the required 675 cps. the phase detector will produce an output voltage which is proportional to the magnitude of the frequency error.

The error voltage is produced in the rectifier unit which is a part of the phase detector. The output of the rectifier unit is fed to the speed error indicator located on the front panel of the coder-indicator, and through an equalizing network to the voltage amplifier.

The heat-controlled oven contributes to the frequency stability of the tuned circuit in the phase-sensitive detector. A separate amplifier used for controlling the oven temperature senses any change in the temperature and operates a relay that controls the current to the heating elements in the oven. Direct current for the heat control elements is supplied by a bridge rectifier, which also supplies d-c bias for the voltage amplifier.

The voltage (magnetic) amplifier receives the speed error voltage developed in the phase-sensitive detector circuit, amplifies it, and converts the error voltage to its equivalent d-c voltage. The equalizing network stabilizes the operation of the amplifier by removing ripples from the error signal, thereby preventing the amplifier from saturating because of noise voltages.
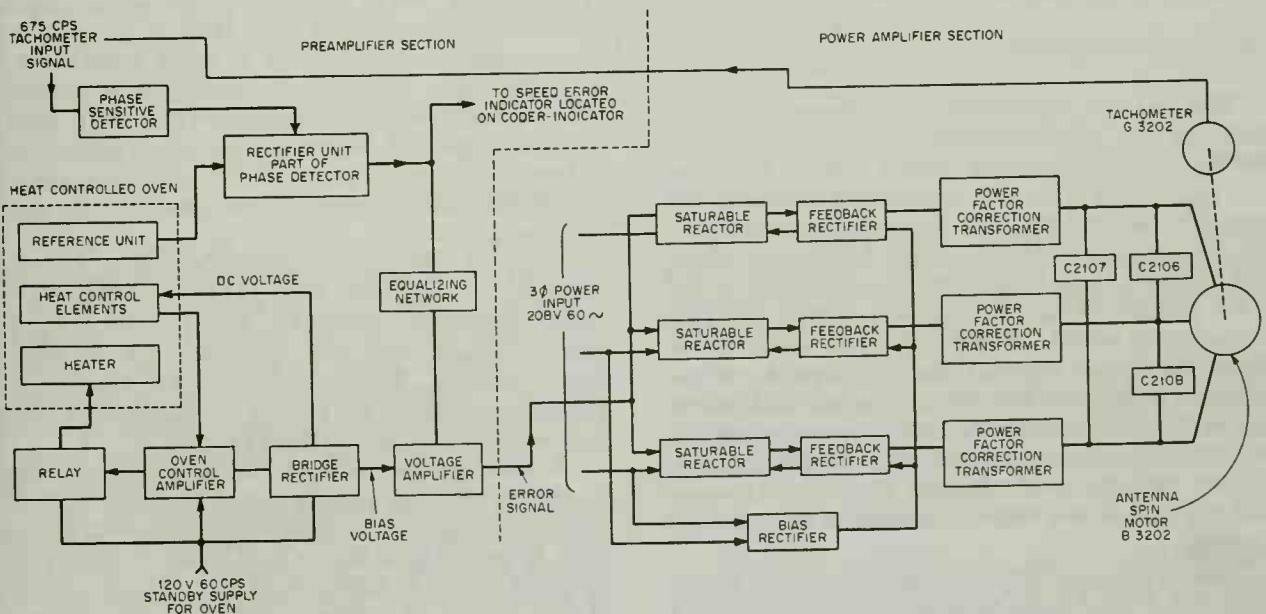

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Figure 11-13.—Antenna speed control circuit, block diagram.

The output of the voltage amplifier is applied to the control winding of three saturable reactors. The load winding of each of the reactors is in series with one phase of the 3-phase power line, which supplies power to the spin motor. Thus, the line current supplied to the speed control motor is directly controlled by the effective impedance of the series reactors.

For low values of d-c control current to the spin motor, the saturation is low and the impedance is high. The motor voltage is decreased and the motor will slow down. Conversely, for large values of control current the motor voltage will be increased, due to the lowered reactor impedance, and the motor will speed up.

A bias rectifier is placed across one phase of the input power to supply the three separate bias windings of the saturable reactor. Each bias winding is shunted by a single dry-disk rectifier (not shown). The rectifier damps oscillatory currents and supresses harmonics of the line frequency that are introduced in the windings of the reactor.

In the AN/SRN-6, the power factor correction capacitors C2106, C2107, and C2108 are not used. The output from each of the saturable reactors is fed through its associated auto-transformer to separate control phases of the 3-phase motor. The transformers step up the line voltage for each phase of the input power to compensate for losses introduced by the power windings of the saturable reactors.

Bearing Servosystem

The bearing servosystem of the radio beacon provides an electromechanical means of adjusting the pulser-plate (fig. 11-4), of the beacon to ensure that any calibrated AN/ARN-21 receiver will receive the proper magnetic bearing. The pulser-plate positions the 15-cps pulser coil so that the 15-cps reference burst of the radio beacon is in its correct position relative to magnetic north. The 15-cps reference burst signal is pulsed so that the tenth pulse of the train of pulses of the 15-cps burst will occur precisely at the positive zero crossover of the 15-cps fundamental and 135-cps ninth harmonic.

The bearing servosystem (fig. 11-14), used with the stabilized shipboard antenna operates as follows:

The angular position signals for true north from the ship's gyro compass are continuously transmitted through the ship's bus to the radio beacon. The magnetic variation unit located in the coder-indicator corrects the ship's gyro compass information for the difference between magnetic north and true north. This correct signal maintains the radio beacon 15-cps reference pulse coil subassembly fixed with respect to correct (magnetic) north.

The magnetic correction signals are manually set into the 1- and 36-speed differential transmitters, CDX601 and CDX602, of the magnetic variation subassembly. The sum of the gyro compass and magnetic variation signals is forwarded to the bearing 1- and 36-speed control transformers, T3201 and T3202, in the antenna base.

As long as the rotors of the control transformers are at the same effective angular position as those of the differential transmitters after summing up the signals of the ship's gyro compass and magnetic variation signals, no control transformer output signal is sent to the bearing servo amplifier. If the equilibrium should be upset because of displacement of the rotors of CDX601 and CDX602 to new positions by changing the setting of the magnetic variation unit, or if the ship's course were to be changed, an error voltage would be introduced into the differential transmitters and transmitted from the control transformers to the servoamplifier.

The magnitude of the error voltage fed to the balanced bearing magnetic amplifier is a function of the rate of change of the control transformer shaft position. If the error signal from the differential transmitters exceeds 2.5 volts (the equivalent of a shaft position 2.5 degrees away from null), the one speed transformer voltage takes control. For shaft position less than 2.5 degrees away from null (equivalent to voltages below 2.5 volts), the output of the 36 speed control transformer is the controlling signal.

Bearing Magnetic Amplifier—The bearing magnetic amplifier (fig. 11-15), utilizes two phase-sensitive detectors; one for the 1-speed error voltage, and the other for the 36-speed error voltage. The output signals of the phase-sensitive detectors are fed to respective control windings of saturable reactor, L2111.

A deemphasis network, CR2146, is connected in series with the 1-speed control winding. This network uses two rectifiers connected back-to-back in order to obtain a nonlinear bilateral resistance. The resistance of this circuit is high at low levels of voltage, owing to the characteristics of the diodes (called Zener effect), and the resistance is decreased considerably
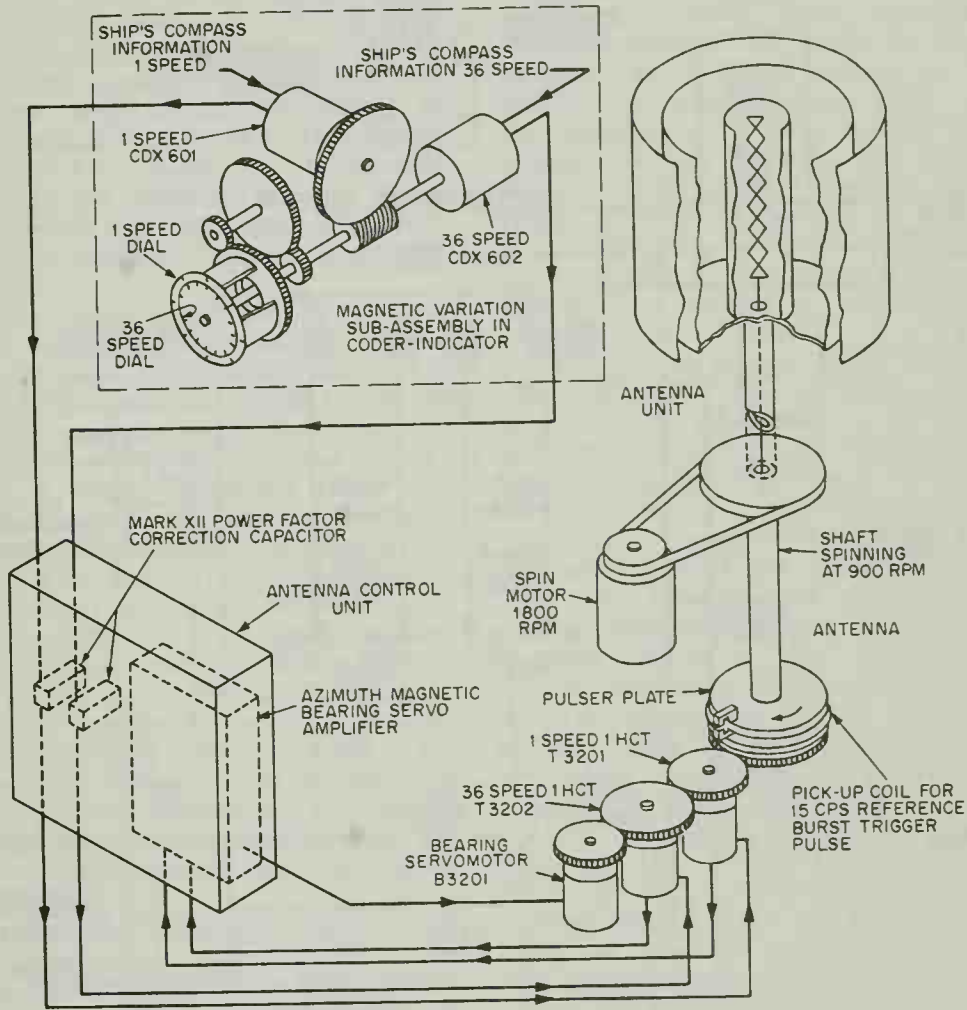
32.84

Figure 11-14.—Shipboard bearing servosystem.

for high levels of voltage. Thus, when the 1-speed transformer error voltage is low, the gain of the 1-speed circuit is decreased. Conversely, when the error voltage is large, the gain is increased and the 1-speed error voltage has the greater control. Under this condition, the lowered resistance of the CR2146 diode pair permits a much higher current to enter the 1-speed control winding.

The series rectifiers, one in each leg of the balanced amplifier circuit, prevent the load current in the leg from reversing. The circuit is said to be balanced, since the center-tap of the output transformer, T2109, is at the electrical center between each branch of the circuit.

The two branches of the amplifier circuit are designated the A and B branches, respectively. The flux of branch A is in opposition to the flux of branch B.

The 120-volt 60-cycle input at terminals 1-2 of TB2107 cause a quiescent current alternately through the two legs of the A and B branches of the amplifier. The resulting voltage developed across the output transformer, T2109, for the quiescent condition is always zero.

Parallel resistances R2148, R2155 and R2149, R2156, connected across the load windings, are used for balancing the a-c output across the 5-4 and 4-3 terminals of output
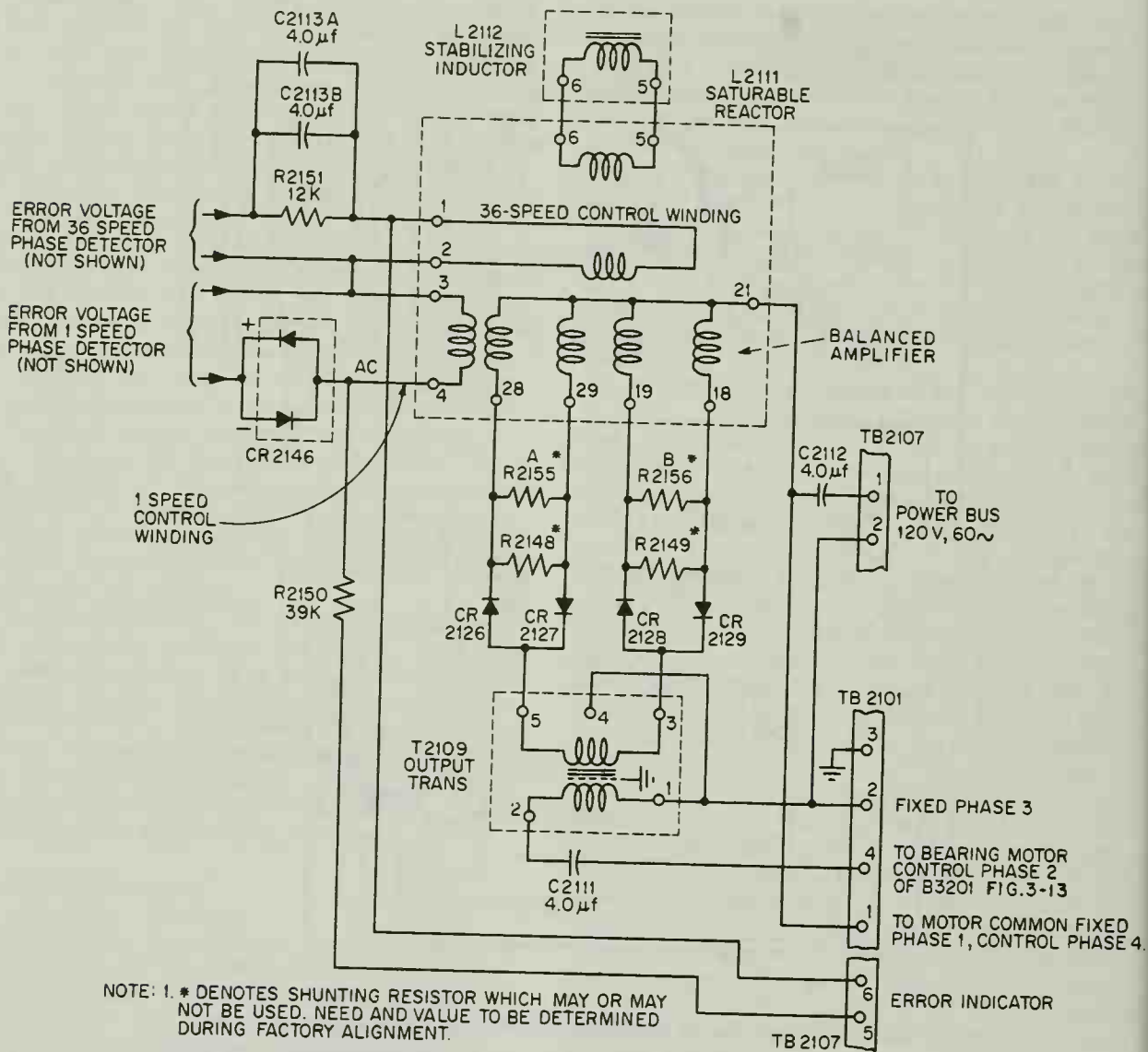
401

32.105

Figure 11-15.—Balance bearing magnetic amplifier, schematic diagram.

transformer, T2109. The value of the resistances is chosen to give zero a-c voltage output for zero d-c current in the control windings. These resistors also set the bias current in the load windings.

The operation of the amplifier is as follows: The mmf of branch A is in opposition to the mmf of branch B. Assume that there is an error voltage which tends to saturate the core of branch A, and desaturate the core of branch B. This condition unbalances the circuit, resulting in a heavy current flow in the load winding of the saturated branch. If the polarity of the error voltage is reversed, branch B becomes saturated, while branch A becomes desaturated.

The branch conducting the greater current will produce a predominant alternating flux across one-half of the primary of the output transformer, T2109. This output is coupled from

402

he secondary of T2109 to the control phase winding of the bearing servomotor, B3201.

Because of the balanced arrangement of the two branches of the amplifier with respect to the center-tap, the mmf in the 5-4 section of T2109 will be predominant for one error and the mmf of the 3-4 section of T2109 will predominate for the opposite error. Thus, the current through the motor (and the direction of motor rotation) reverses with a reversal of error voltage.

Antistickoff Voltage.—As stated earlier, signals from the magnetic variation subassembly of the coder-indicator are introduced into the 1- and 36-speed differential transmitters in order to cause the pulser plate assembly in the antenna pedestal to be corrected to magnetic north. A single calibrated from 0 to 360, and fastened to the coarse (1-speed) differential transmitter and control transformer (CDX601 and T3201, fig. 11-14) yields only a fair degree of accuracy in proving information to the servomotor, B3201. A second dial, fastened to the fine (36-speed) differential transmitter and control transformer, CDX602 and T3202, is calibrated so that one revolution of the fine control transformer, T3202, equals only 10° of the full coverage of the coarse control transformer, T3201. This provides a method of reading the position of the control shaft with 36 times the accuracy of a system when only the coarse control is provided.

The voltage which controls the bearing servomotor is shown in figure 11-16. The curves shown have the appearance of sine waves but actually represent an infinite number of rms voltage values for error signals throughout a 360° rotation of the rotor of the control transformer.

The amplified control transformer output (fig. 11-15) drives the bearing motor. The motor is a 2-phase induction motor having a continuously excited fixed phase and a variable control phase. The direction of the motor torque is determined by the instantaneous polarity of the control voltage from the control transformer with respect to that of the fixed phase.

When the error voltage is any value above the zero reference line, the direction of the motor rotation is assumed to be counterclockwise (ccw). Conversely, when the error voltage is below the axis the instantaneous polarity of the control phase of the motor is reversed with respect to the fixed phase and the motor rotates in the opposite direction (clockwise).

If the error voltage is less than 2.5 a deemphasis network which consists of Zener diodes connected as shown in figure 11-15, permits the 36-speed voltage to control the motor. For error signals greater than 2.5°, the deemphasis network breaks down and the 1-speed voltage controls the motor. Thus in figure 11-16A, the portion of the curve from 357.5°(corresponding to 2.5° on either side of 0°), the 36-speed voltage controls the servo system. For errors larger than 2.5° the 1-speed voltage is shown to be the controlling signal.

The use of an even gear ratio between the fine and coarse control transformers presents a problem in synchronizing the servo system. Note that at 180° the 36-speed voltage crosses the zero axis three times in the area from 175° to 185° to produce a negative voltage at 179° and a positive voltage at 181°. When the system is at 179° the error voltage drives the motor clockwise; at 181° the servomotor is driven counterclockwise. This causes the system to lock-in at 180°, so that the system synchronizes at a false null of 180°.

The elimination of the 180° false null is accomplished by the addition of an a-c voltage, called an antistickoff voltage (fig. 11-17) to the 1-speed voltage (fig. 11-16 B). The addition of the antistickoff voltage shifts the curve upward at 180° so that the crossover point occurs at 185°. The 180° false null condition cannot exist, since for an error condition of 185° or less, the error voltage is positive and the servomotor is driven counterclockwise away from the false null to null at 0°. If the error condition exists above 185° the servomotor will be driven clockwise away from the false null to null at 360° or zero degrees. Thus, the addition of the antistickoff voltage to the 1-speed control transformer voltage causes only the 0° null position to exist as a stable point in the servosystem.

To correct the 1-speed zero at the zero shaft angle position, the 1-speed control transformer is re-zeroed to a point 2.5° from zero. This action cancels the error introduced by the addition of the antistickoff voltage at the zero shaft position, and adds the 5° error to the 180° crossover point.

ROLL AND PITCH SERVOSYSTEMS

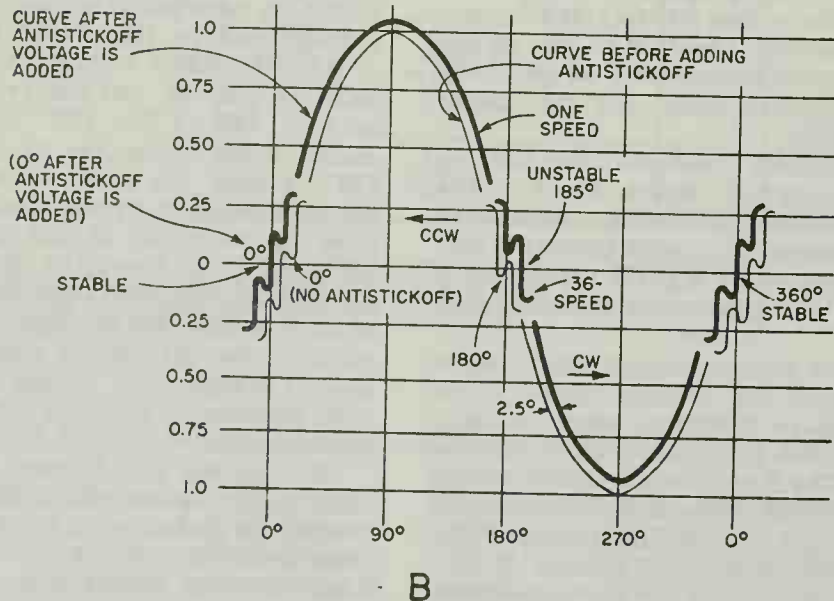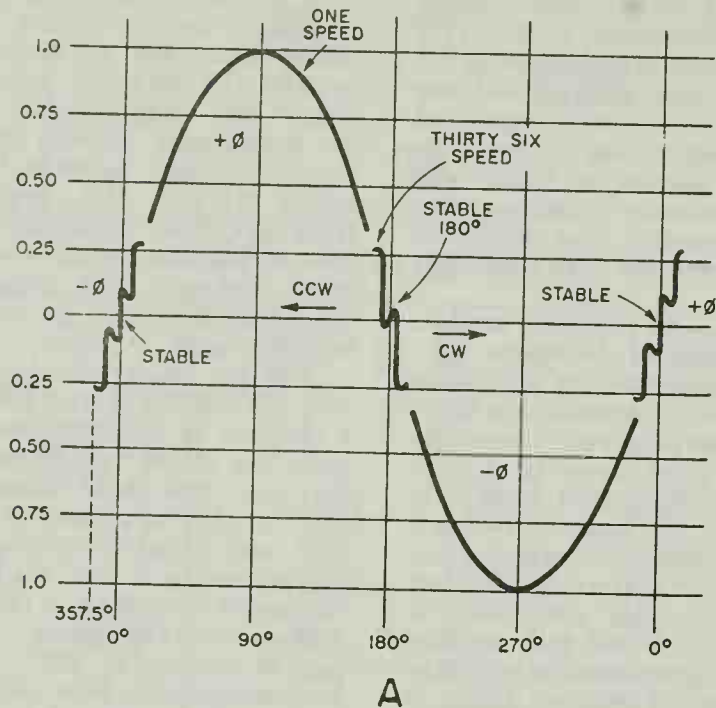The roll and pitch servosystems (fig. 11-18) are separate but similar circuits which
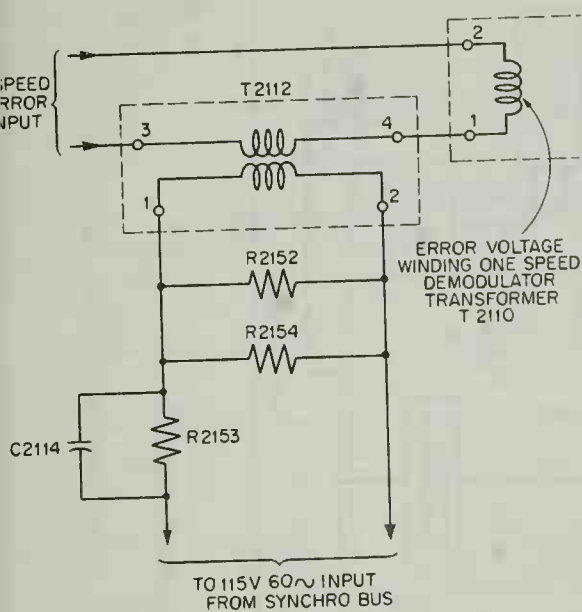
Figure 11-16.—Addition of antistickoff voltage.

32.106

maintain the antenna in the vertical position with respect to a horizontal plane despite roll and pitch motions of the ship.

The roll servosystem can correct up to 25° of roll, while the pitch servosystem can correct up to 6° of pitch. Roll and pitch information is obtained from the stable element of the ship

404

32.107

Figure 11-17.—Method of adding antistickoff voltage.

which is gyroscopically controlled to maintain a true horizontal position.

A 2-position (stow-stabilized) switch, S606, is located on the front panel of the coder-indicator. In the STOW position, the antenna is locked perpendicular to the deck of the ship. When S606 is in the STABILIZED position, (which is its normal position aboard ship) the antenna will be held in a fixed position by the servosystems so long as the roll or pitch of the ship does not exceed the specified limits as described above.

Roll and pitch information is fed from the bus of the ship through the stow-stabilized switch, S606, in the STABILIZED position, to the control transformer. The control transformer passes the error signals to the roll or pitch magnetic amplifier.

The roll and pitch magnetic amplifiers are almost identical units. The following discussion applies to both systems unless otherwise noted.

The synchro error voltage (11-18), from the roll or pitch control transformers is fed to a phase-sensitive detector. The reference voltage for the phase detector is obtained from the 115-volt, 60-cps roll and pitch reference bus.

The d-c output error voltage from the phase detector appears between the center-tap of the two secondaries of a transformer (not shown), The phase-sensitive detector load is connected between these points, and consists of the series, aiding control windings of the magnetic pre-amplifier.

The tachometer feedback current from either G3001 or G3002, is fed to one set of the control windings of the magnetic preamplifier. Interference between the phase detector and tacho-meter circuits is prevented by the isolation network (acceleration feedback network).

The isolation network impedance in the ta-chometer input circuit also serves the purpose of changing the tachometer voltage from rate feedback to acceleration feedback.

Without the use of a rate-to-acceleration feedback circuit, when the load is turning at a constant speed, a steady d-c voltage (propor-tional to speed) would be fed back to the pre-amplifier in opposition to the error signal. Therefore, a greater error signal would be re-quired to overcome the d-c voltage and produce a certain unbalance in the amplifier and a cer-tain speed of the servo motor than would be required if there were no feedback.

In order to produce a greater error signal there must be a greater error. Since the feed-back voltage is proportional to speed, the error is also proportional to speed. This is called a velocity error, or rate feedback.

To eliminate this velocity error, a group of capacitors (not shown) are connected in the feed-back circuit and block the d-c component of the voltage. The a-c component of the voltage at the frequency of hunting (several cycles per second) is fed back without much loss.

With this circuit arrangement, only changes is speed will produce a feedback signal through the capacitors. This action is called accelera-tion feedback, which further improves the stability of the servosystem.

The bridge rectifier consisting of four diodes (not shown), serves as a high impedance shunt across the feedback capacitors for low amplitude tachometer feedback voltages. For high am-plitude feedback voltages, a diode (not shown), with a low impedance, causes the rectifier to act as a low impedance shunt across the feed-back capacitors. This feature results in the tachometer feedback signal having a greater damping effect when the servosystem is cor-recting rapidly than when the system is cor-recting slowly. The low impedance diode is
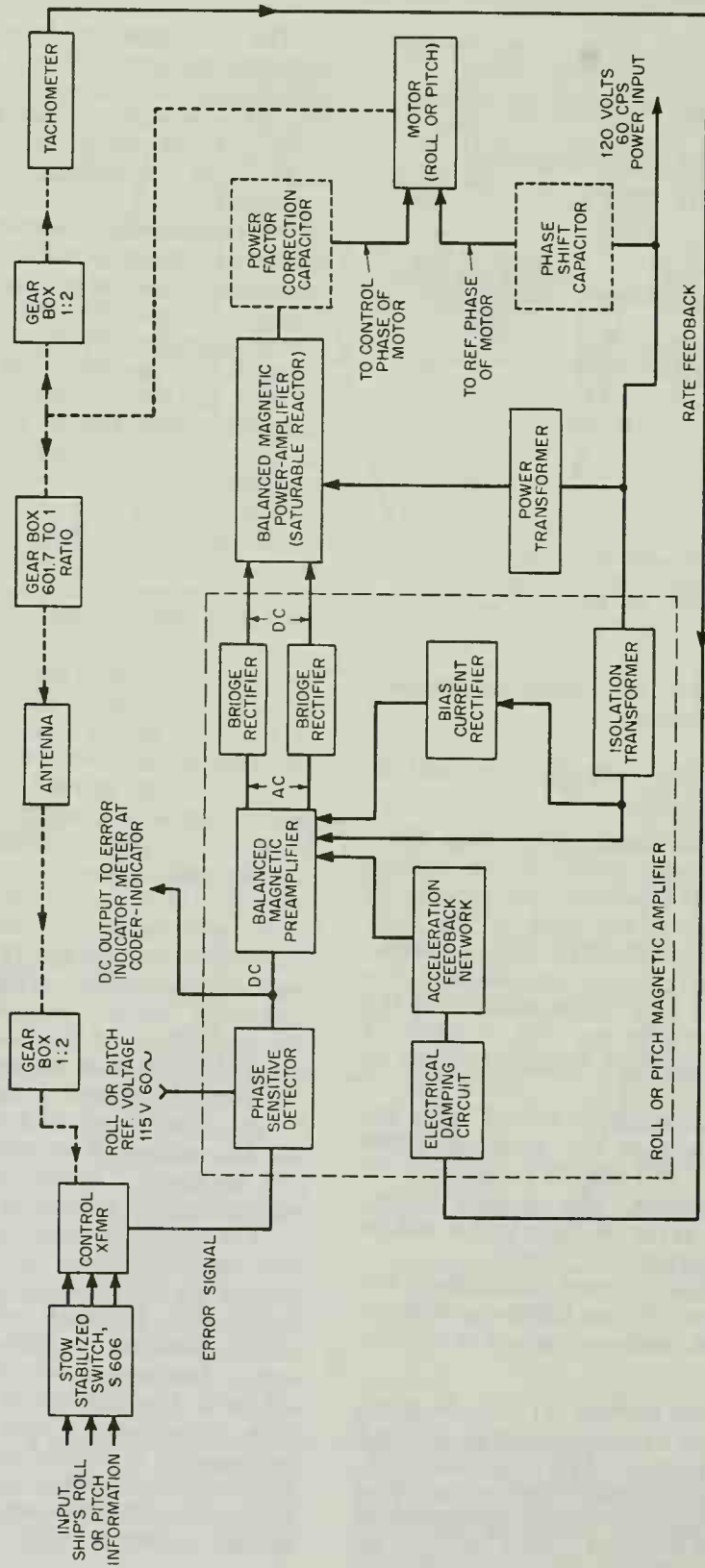
Figure 11-18. – Roll and pitch servosystem, block diagram.

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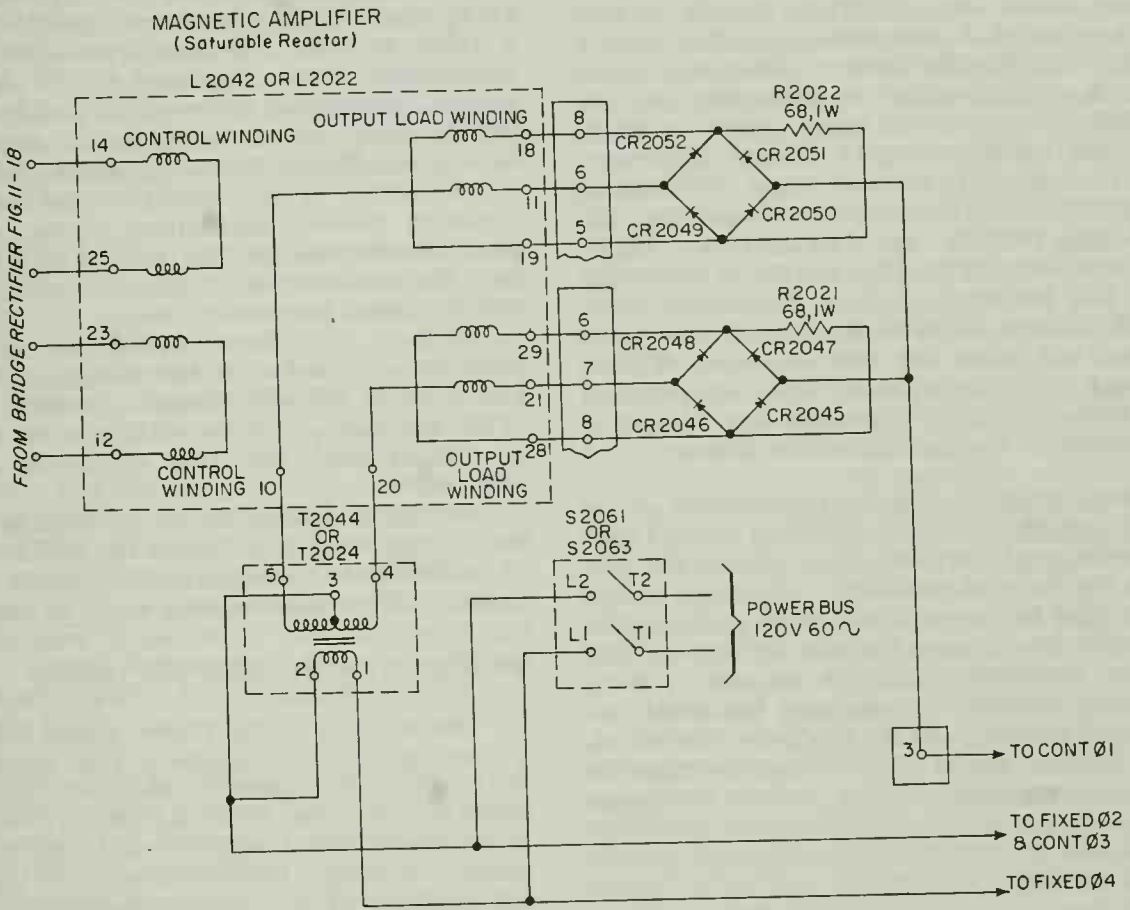
32.108

Figure 11-19.—Roll or pitch magnetic amplifier.

selected for its Zener characteristic, which means that the diode will have a very high front-to-back resistance ratio for low amplitude a-c signals, and a very low front-to-back resistance ratio for signals above a certain amplitude.

As stated earlier, the error signal from the phase-sensitive detector is developed across the three control windings (not shown), of the preamplifier which are connected in series aiding. A rectified bias current is conducted through the bias winding section of the control winding by the action of the bridge rectifier comprising of four diodes (not shown),

two resistors (not shown), in parallel for current limiting, and two balancing resistors (not shown), in the separate halves of the bias windings. A capacitor (not shown) filters the bias voltage to produce the required steady bias current.

An error signal in the control windings will cause an impedance increase in one of the load windings (either upper or lower), while simultaneously lowering the impedance in the other load winding. The control windings output, after being rectified in separate bridge circuits, causes a greater controlling signal

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to be fed to one of the control windings of the magnetic amplifier than to the other.

The two load windings of the magnetic amplifier, L2022 (fig. 11-19), in conjunction with the 5-4 and 4-3 terminals of T2024, form a bridge circuit. The control phase winding of the roll or pitch motor is connected from the transformer center-tap to the center-tap of the two load winding circuits. During quiescence (no-error signal), the a-c input at T2024 is rectified in the bridge circuits comprising CR-2045 thru CR2048, and CR2049 thru CR2052. The impedance of the 1819 and 29-28 load windings will be equal. Since the transformer, T2024, voltage is equal across both legs of the bridge, and since the impedance in each of the legs is the same, no difference in potential will exist across the control phase winding of the motor and the motor does not rotate.

When an error signal is introduced at the 14-25 and 23-12 control windings of the L2022 magnetic amplifier, the signal will tend to saturate the core of one-half of the output bridge circuit (for instance, the top half), while simultaneously desaturating the core of the other load winding (the bottom half). The impedance of the saturated winding is lower than that of the desaturated winding, and the bridge is unbalanced. Thus a difference in potential exists across the motor control phase winding, and the motor rotates in a given direction. The 5-4 section of T2024 will provide the source voltage for the motor winding.

When the bridge is unbalanced, the motor essentially is placed across one-half of the transformer (T2024) in series with one load winding of the bridge. The current through the motor will be alternating, changing as the transformer secondary voltage changes. If the input error signal from the magnetic preamplifier tends to saturate the core of the lower load winding, the impedance of this winding will be decreased, while the impedance of the upper load winding is simultaneously increased. In this condition, the source voltage for the motor is derived from the 4-3 section of T2024. Note that the instantaneous direction of current through the control phase of the motor will be reversed, since the instantaneous voltages across the separate halves of the transformer (T2024) are 180° out of phase. An instantaneous change in the control winding current with respect to the fixed phase winding will reverse the direction of rotation of the motor.

The servomotor, through its associated gear train, moves the antenna to its corrected position. At the same time, the gear train nulls the error signal at the control transformer through a 1:2 gear box. The gear train also turns a tachometer which produces a rate feedback voltage. This signal is converted into acceleration feedback and applied to the balanced magnetic preamplifier. The feedback is degenerative to the action of the amplifier and therefore tends to prevent oscillations in the roll or pitch servo-systems. The method used to convert the rate feedback to acceleration feedback was discussed previously.

Blowers (not shown) are connected to the pitch and roll motors to help dissipate the heat generated in the unit. Brake solenoids, a part of the apparatus, lock the position of the antenna in roll or pitch when the equipment is de-energized.

The bubble canopy (fig 11-1), which protects the antenna and pedestal from the weather, also decreases wind resistance. Without the bubble canopy, larger servomotors would be required.

POWER SUPPLY - TEST SET GROUP

The top half of the power supply - test set group (fig. 11-1), consists of four units which will completely check out the operation of the radio beacon. The lower half of the group contains three power supplies with overload and control circuits. The lowest shelf in the radio set contains blower units to maintain the units at the proper operating temperatures, and fuses for primary power.

Power Supply Group

The low voltage power supply consists of two d-c regulated outputs of + 250V and -375V from two conventional controlled series resistance, tube type rectifiers. These voltages are used throughout the equipment.

The medium voltage power supply consists of a d-c regulated output of + 1000V, and is used in the frequency multiplier-oscillator.

The high voltage power supply delivers an unregulated output of about -12 KV to the klystron in the amplifier-modulator.

Test Set Group

To premit a rapid check of important operating characteristics and maintenance of the radio

eacon, four test equipment units are mounted the power supply cabinet. These units are:) TS-891/URN-3 power meter-pulse counter,) OS-54/URN-3 oscilloscope, (3) TS-890/URN 3 pulse analyzer-signal generator, and (4) G-121A/URN-3 pulse generator. An additional it SA-420/URN-3 switch-test adapter, mount- d outside of the cabinet is used with the TS- 91/URN-3 power meter-pulse counter to com- lete the test equipment.

SWITCH TEST ADAPTER.—The switch test dapter SA-420/URN-3 (fig. 11-1), is mounted n a wall or bulkhead near the radio set. The nit (fig. 11-10,A) consists of a coaxial line ith a voltage probe and a crystal detector hich makes a detected sample of r-f avail- ble to the radio frequency monitor MX-1627/ RN-3 (discussed later), for power monitoring urposes. The coaxial line also contains a di- ectional coupler and a crystal detector which akes a detected sample of r-f available to the ower meter-pulse counter (TS-981/URN-3), or peak power measurements. The switch est adapter provides a manually operated oaxial switch for connecting the transmitter -f output to either the transmitting antenna or dummy load.

POWER METER-PULSE COUNTER.—The ower meter-pulse counter TS-891/URN-3 (fig. 1-20) measures the peak r-f power output of he AN/URN-3 transmitter in its frequency range of 960 to 1215 mc. The switch-test adapter must be used in conjunction with the ower meter-pulse counter for this function, since the power measuring circuit is incomplete without both units. Power is measured by means of a slide-back diode circuit and can then be read in kilowatts on the PEAK RF POWER meter. The R-F POWER SWITCH is used in conjunction with the PEAK PWR VOLTAGE control and the peak r-f power meter circuit, and when depressed, closes this circuit for peak r-f power measurements. The PEAK PWR VOLTAGE CONTROL is used to establish the slide-back voltage for the power meter cir- cuit during power measurements. It is operative only when the RF POWER switch is depressed. The detected r-f pulse envelope is sent through the FUNCTION SWITCH to the oscilloscope for visual observation during peak power measure- ments. Another circuit measures the average repetition rates of all pulse sources in the AN/ URN-3 and its test equipment over a range of

80 to 8000 pulses per second. The repetition rate is counted by means of an analog type circuit and is read in pulses per second on the PULSE COUNT METER on the front panel of the power meter-pulse counter. Suitable gating, phase reversal and pulse shaping circuits are provided in order to perform the necessary counting of both fixed and random rate pulses. The PULSE COUNT CAL adjustment is a locked screwdriver adjustment control which is used for calibrating the PULSE COUNT METER and should be made only by technical personnel. When the COUNTER SELECTOR SWITCH is in the "-" position, the pulse counter will count negative input pulses at the front panel COUNTER INPUT connector. With the switch in the "+" position, the pulse counter will count positive pulses at the same input. In the SQUITTER and REPLY RATE positions, the receiver squit- ter and reply rates are measured respectively. The RANGE SWITCH has two positions, X1 and X10. In the X1 position, the PULSE COUNT METER range is 6 to 800 pulses per second. In the X10 position, the meter range is 0 to 800 pulses per second and all readings on the PULSE COUNT METER dial should be multi- plied by 10. The FUNCTION SWITCH controls the functional operation of the unit. The switch is labeled the same as another FUNCTION SWITCH on Oscilloscope OS-54/URN-3. When these two switches are set to corresponding positions, all appropriate interconnections nec- essary for normal test of the AN/URN-3 are made. With the FUNCTION SWITCH in the GENERAL TESTING position, the power meter- pulse counter can be used as a general piece of test equipment for counting the repetition rate of pulses fed to a front panel receptacle COUNTER INPUT and whose characteristics fall within the equipment capabilities.

Table 11-1 lists the FUNCTION SWITCH positions and the tests that can be performed in those positions.

Table 11-1—Function Switch Positions.

Switch Position	Tests
1. OPERATING TEST	Peak r-f power Visual pulse shape Output pulse count Transmitter spectrum

Table 11-1—Function Switch Positions—
Continued

Switch Position	Tests
2. RECEIVER SENSITIVITY	Squitter count Reply count Reference pulse burst count Receiver sensitivity
3. SYSTEM DELAY	Overall zero-distance time delay Video unit zero-distance time delay
4. GENERAL TESTING	Each test equipment available for individual use

OSCILLOSCOPE.— The oscilloscope OS-54/URN-3 has the basic circuits found in a conventional oscilloscope and differs only in having manually selected preset conditions selected by a FUNCTION SWITCH on the front panel. The switch positions are labeled the same as the pulse meter-pulse counter, OPERATING TEST, RECEIVER SENSITIVITY, SYSTEM DELAY, and General Testing. In the first three positions, vertical, trigger, and sweep, inputs are applied from external equipment through BNC connectors at the rear of the oscilloscope. Circuit parameters are switched in appropriate sections of the oscilloscope to furnish the preset horizontal sweep speed and vertical sensitivity required for the selected function test. Front panel controls capable of affecting fixed conditions are inoperative. In the GENERAL TESTING function, the preset conditions are switched out and vertical, trigger, and sweep inputs can be applied to the front panel receptacles to use the oscilloscope in the conventional manner.

PULSE ANALYZER-SIGNAL GENERATOR. —The pulse analyzer-signal generator TS-890 URN-3 (fig. 11-21) may be functionally divided into two sections, the signal generator portion and the pulse analyzer (receiver) portion. They are related only in that the signal generator output is used in a mixer to beat with the receiver input signal and produce a 63 mc receiver input frequency.

The signal generator produces a 1025- to 1150-mc output, either pulsed or cw, for testing the beacon receiver, and also as a heterodyne signal for turning the pulse analyzer section to the output of the beacon transmitter. The output frequency can be continuously varied over the entire range and can be checked against a single, internal, crystal-controlled frequency. Alternatively, the output can be derived from the internal crystal. Output amplitude is variable over a range of 10 microvolts to 0.5 volts. This output may be set to a standard reference level, as indicated on an OUTPUT LEVEL INDICATOR, by means of a POWER SET control. The standard reference may be attenuated over a wide range by means of an accurately claibrated R.F. OUTPUT (DBM) attenuator. Thus, a signal of known frequency and amplitude is delivered to R.F. OUTPUT jack for application to the beacon receiver. This signal may be either pulse or cw. A suitable modulation pulse is obtained from a companion test unit pulse-sweep generator SG-121A/URN-3 (discussed later).

The pulse analyzer is, in a sense, a radio receiver. The local oscillator for this receiver is the signal generator. The signal fed to the pulse analyzer is taken from the output of the beacon transmitter. It is fed, through an attenuator and filter-mixer circuit in the rf unit, to the receiver i.f. stage as a 63-mc signal. The receiver employs a POWER COMPARISON INDICATOR in an arrangement that permits measuring side-band energy with great ease.

The filter-mixer circuit heterodynes the beacon transmitter output with the signal generator output. A feature of the tacan system is that the receiver and transmitter frequencies of all channels are exactly 63 mc apart. Thus the output of the preselector is always a 63-mc signal.

The receiver can be tuned to any frequency in the two bands from 962 to 1024 mc and from 1151 to 1213 mc. The received signal is applied to a metering circuit through a very narrow band amplifier. The receiver can be tuned to the exact center frequency of the signal by switching crystals (BAND SHIFT knob) in the local oscillator. This gives a narrow band of 63 mc i-f bandwidth at 0.8 mc or 1.0 mc either side of center. This permits metering the spectral distribution characteristic (side-band energy) of the output pulse transmitted by the output pulse transmitted by the beacon. The step attenuator is part of the crystal
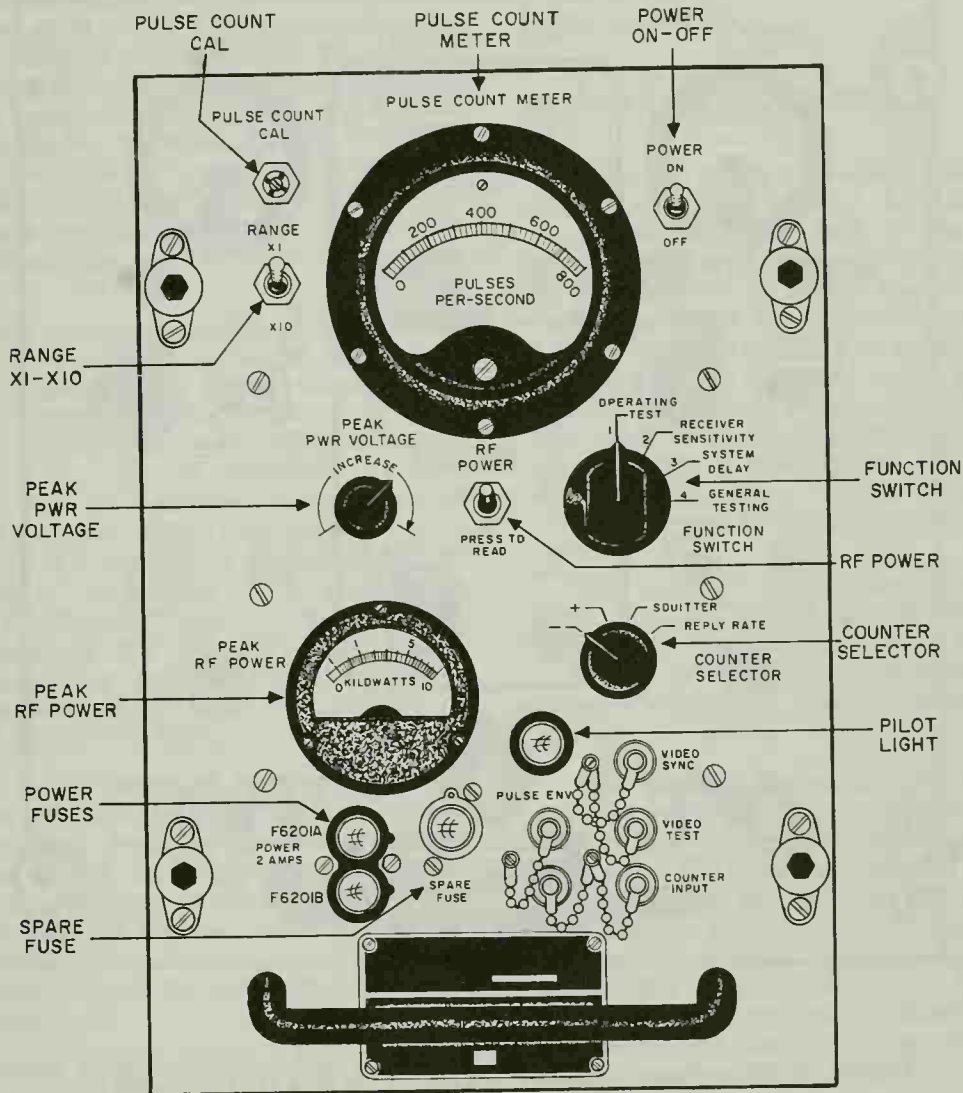
162.35

Figure 11-20.—Power Meter-Pulse Counter TS-891/URN-3.

turret-step attenuator (BAND SHIFT knob) as-sembly and is arranged to insert a discrete value of attenuation for each crystal frequency. For the on-channel frequency, attenuation is 40 db. For the frequencies 0.8 mc off channel, attenuation is 5 db. For the 1.0-mc off-channel frequencies, attenuation is zero db. These values are used so that if the beacon output pulse has normal spectral distribution, attenuator output will be uniform under all conditions. That is, as the side-band energy falls off with change of receiver frequency, the attenuation is reduced by an amount calculated to keep the output uniform. The output section of the receiver is essentially a metering amplifier. The output is fed to the POWER COMPARISON INDICATOR to provide a relative indication of the signal level at the output of the step attenuator. The narrow bandwidth of the receiver is obtained by means of sharply tuned circuits in this unit.

Thus, to use the pulse analyzer, it is only necessary to set the signal generator to the desired channel on the digital type CHANNEL
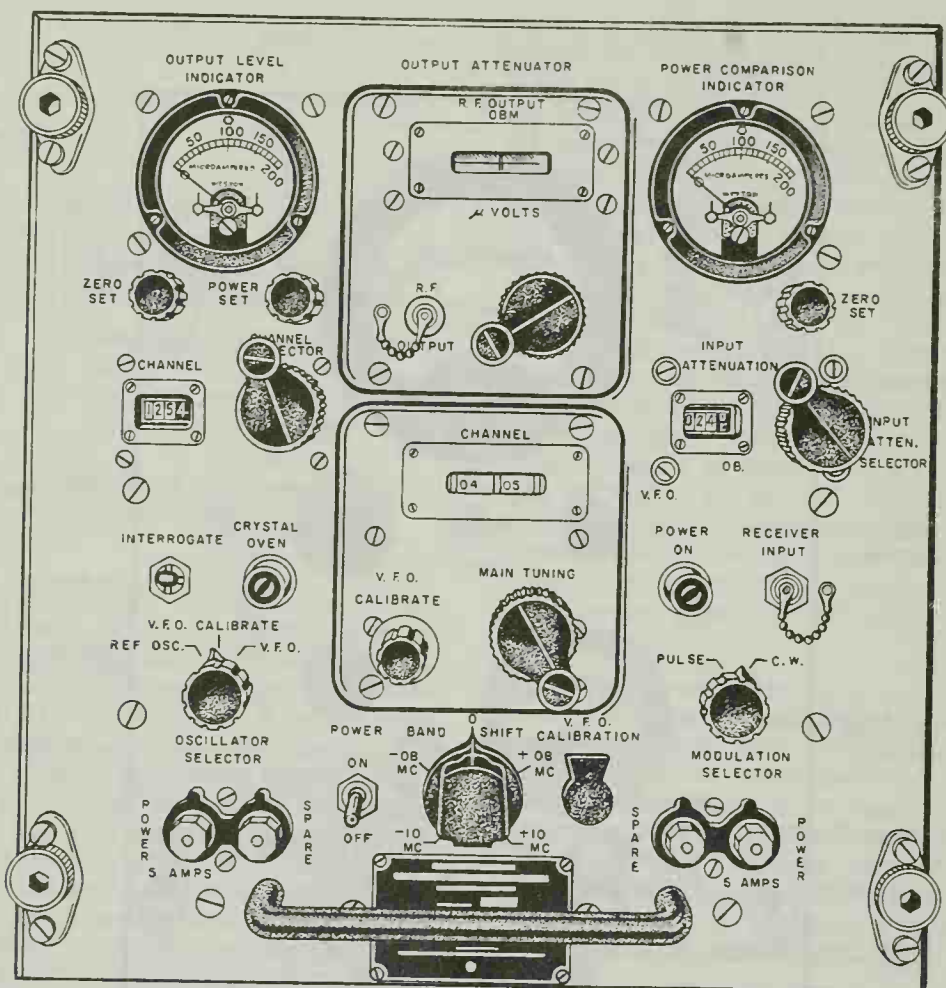
162. 36

Figure 11-21.—Pulse Analyser-Signal Generator TS-890/URN-3

indicator, set the input attenuator to make the POWER COMPARISON INDICATOR read midscale, set the signal generator for cw operation, and then read relative output at the upper and lower side-band frequencies. Any deviations from the required spectral distribution characteristic will be shown on the meter as deviations from the midscale reading.

PULSE SWEEP GENERATOR.—Pulse Sweep Generator SG-121A URN-3 is an accurately calibrated test instrument, which provides a quick means of checking the salient features of the AN/SRN-6 beacon per-

formance. Essentially, the pulse sweep generator serves as the pulse modulator for Pulse Analyzer—Signal Generator TS-890/ URN-3, and provides the timing circuitry for measurement of zero-distance time delay of Radio Set AN/SRN-6.

The pulse-generating chain of the instrument supplies video pulse-pairs, shaped to correspond to the pulse waveform of Radio Set AN/ARN-21, the airborne interrogator for the AN/SRN-6 equipment. These pulse-pairs are available from the pulse sweep generator at a repetition rate of 40 to 5000 pulse-pairs per second. The spacing between the members within

a pair of pulses may be adjusted in 0.5 microsecond steps, with a ± 0.2 microsecond tolerance, from 11 to 13 microseconds. The amplitude of this paired-pulse signal is continuously variable, by means of a front panel control from zero to 200 volts when the PULSE OUTPUT jack is loaded with an impedance as low as 650 ohms, shunted by as much as 800 picofarads. Facilities are provided for either internal or external triggering of the pulse-generating chain.

The sweep-generating chain of the instrument generates a sawtooth waveform, and a carefully synchronized trigger for the pulse-generating chain. Both the sawtooth and the trigger are frequency-controlled by a crystal oscillator. The period of this crystal oscillator is a submultiple of the zero-distance time delay of Radio Set AN/SRN-6. Repetition frequencies of 39.840 kc, 40.160 kc, 45.454 kc, or 47.620 kc are selected by the setting of the front panel CRYSTAL SELECTOR control. The peak-to-peak amplitude of the sawtooth waveform is 20 volts, when the output terminal is loaded by an impedance of 100,000 ohms shunted by as much as 270 picofarads.

A video mixer circuit is provided in the instrument as a means of combining two of various types of outputs from Radio Set AN/SRN-6 into a video signal. This output signal is used to allow a presentation of the time difference comparison between the two outputs on Oscilloscope OS-54/SRN-6. By this means, the zero-distance time delay of the AN/SRN-6 beacon is to be measured.

Pulse Sweep Generator SG-121A/URN-3 is utilized in only a few of the test measurements controlled by the function switch, table 11-1. These measurements are:

1. Operating Test - The pulse sweep generator circuitry is not used.

2. Receiver Sensitivity - The pulse sweep generator circuitry is used in measuring squitter count and reply pulse count.

3. System Delay - The pulse sweep generator circuitry is used in measuring overall zero-distance time delay and video unit zero-distance time delay.

4. General Test - The paired-pulse output signals and trigger output signal from the pulse sweep generator are available for other tests, as desired.

RADIO FREQUENCY MONITOR

Radio Frequency Monitor MX-1627/URN-3 (fig. 11-22) and Antenna AT-592/URN-3 (fig. 11-1), which form part of the special test equipment, are a receiver-transmitter and antenna employed in either ship or shore installations to check the performance of Radio Set AN/SRN-6. The monitor interrogates Radio Set AN/SRN-6 with a pulsed signal simulating that of airborne Radio Set AN/ARN-21 and receives the reply of Radio Set AN/SRN-6. The monitor provides a visual (and has provision for an external) alarm signal when Radio Set AN/SRN-6 deviates from its proper performance.

The monitor is fed input signals directly from the AN/SRN-6 transmitter output r-f line J6151 (fig. 11-8A) and radiated signals from the AN/SRN-6 transmitter antenna, via Radio Frequency Monitor Antenna AT-592/URN-3.

The monitor antenna should be mounted in a vertical position. Its center line should line up with a line extended from the N (north) mark line of the AN/SRN-6 beacon antenna which is located on the base of the beacon antenna and also on the radome for shipboard installations. The exact position of the antenna along this line should be 2 feet (measured horizontally) from the bottom flange of the stationary antenna cover and 3 feet above the bottom flange (fig. 11-3).

Basically the monitor (fig. 11-23) consists of a crystal controlled receiver-transmitter. The monitor transmitter emits 30 coded r-f pulse-pairs per second on the AN/SRN-6 receiver channel frequency. These pulses are fed to the monitor antenna which is in the field of the AN/SRN-6 antenna. If the AN/SRN-6 system is functioning properly, a correctly coded pulse-pair is transmitted by the AN/SRN-6 50 micro-seconds after challenge by the monitor. The monitor receives these reply pulses, reference bursts and squitter pulses. The monitor will then give a visual alarm indication whenever any of the monitored properties of the AN/SRN-6 transponder fall outside of tolerance specifications. There are
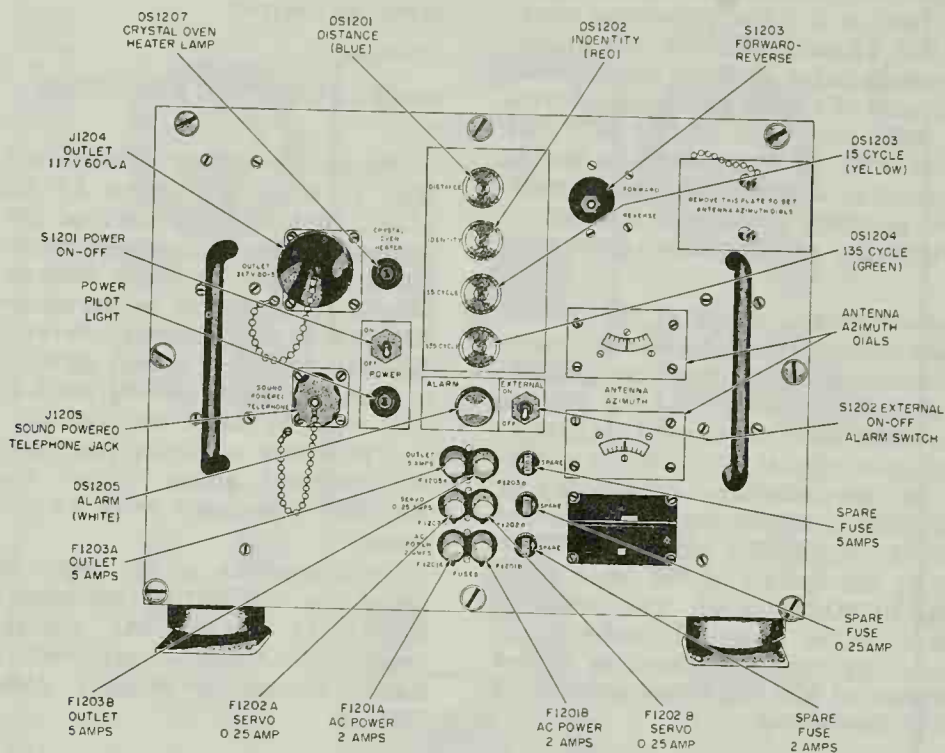
Figure 11-22. — Radio Frequency Monitor MX-1627/URN-3.

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four operating lights plus an alarm light. The faulty replies cause one or more of the alarm lights to go out, giving a rough indication of the nature of the faulty operation. If the fault persists, a visual alarm (on the front panel) is actuated. The monitor also has provision for an external alarm, if desired.

The monitor, therefore, has a vital function providing complete supervision of the functioning of Radio Set AN/ SRN-6 in a form convenient for local or remote indication.

The following visual indicators (fig. 11-22), will indicate when performance requirements of the Radio Set AN/SRN-6 are not within tolerance.

(1) DISTANCE — blue light — normally on (off for five seconds at end of identity keying cycle).

(a) Transponder time delay — light stays on if transponder nominal delay of $50 \mu sec$ changes less than $\pm 2 \mu sec$. Light goes off if delay changes more than $\pm 5 \mu sec$.

(b) Transmitter frequency — light stays on if change in transponder frequency from nominal is less than 0.2 mc. Light goes off if change is greater than 0.5 mc.

(c) Receiver sensitivity — light goes off if AN/SRN-6 receiver sensitivity decreases by 12 db. Light stays on if decrease is less than 6 db.

(d) Transmitter code — light goes off if pulse separation of transponder transmitted code varies by more than $3 \mu sec$ from its nominal value of 12 microseconds.

(e) Reciver decoder — light goes off if transponder receiver is not decoding properly.

(f) Transmitter peak power — light stays on if transponder peak power output decreases up to 4 db. Light goes out if decrease amounts to 8 db.

(g) Transponder count down — light goes out if less than 50 percent of monitor interrogations receive replies. Light stays on if more than 75 percent of interrogations are received.

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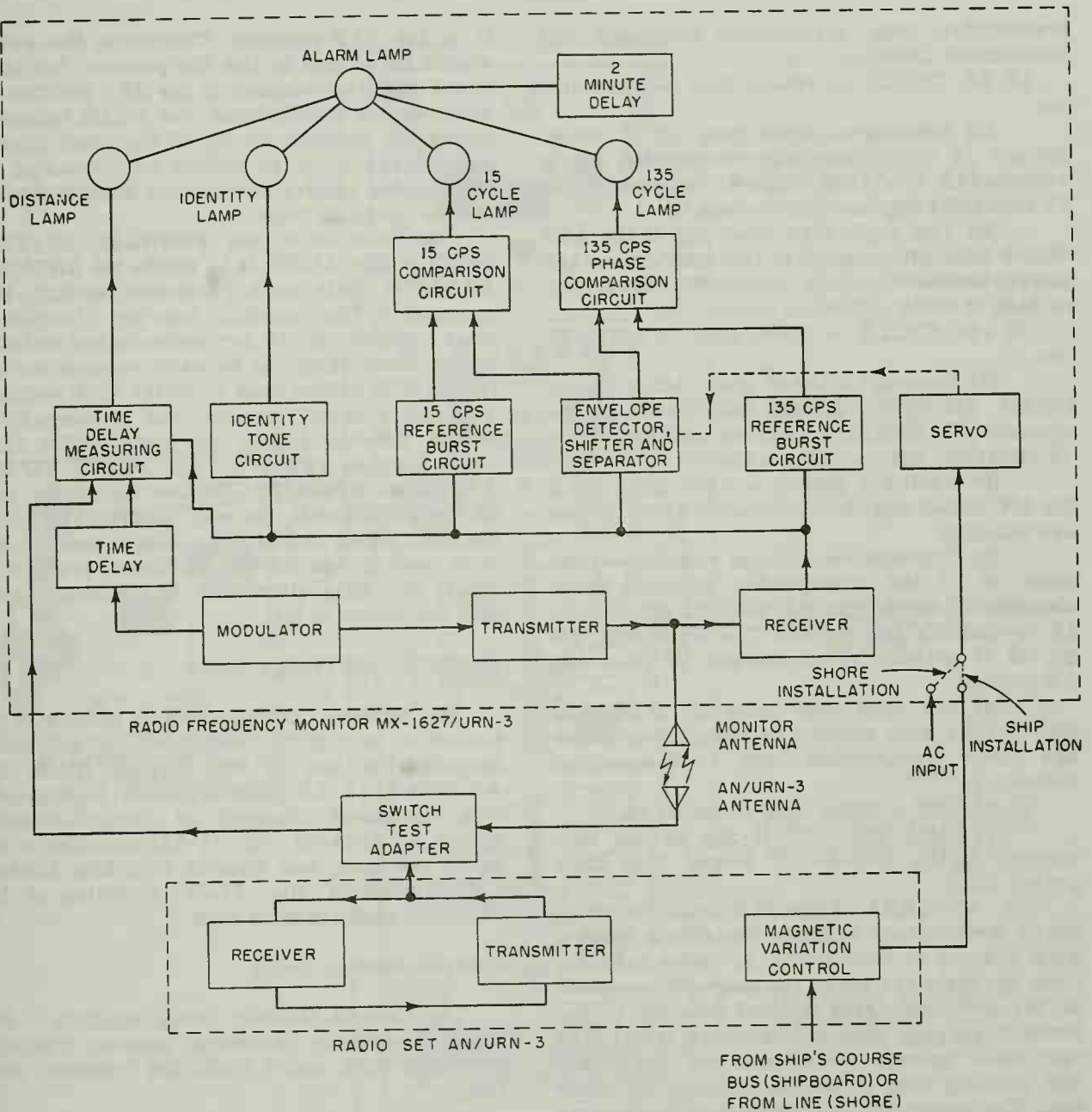
162.38

Figure 11-23.—Overall system, simplified block diagram.

(h) Receiver frequency — light goes off if transponder receiver frequency changes.

(2) IDENTITY — red light.

(a) Normal operation — light flashes on and off following identity keying at 1 1/4 minute intervals. Towards the end of the identity keying cycle, the DISTANCE lamp will go off for a short period of time (approximately 5 seconds) and then light again. This is normal and should not be considered as a possible malfunctioning of the AN/SRN-6.

(b) The light goes off if the AN/SRN-6 fails with respect to transmitter average power,

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transmitter code, transmitter frequency and transmitter identity.

(3) 15 CYCLE — yellow light — normally on.

(a) Bearing — light goes off if transponder 15 cycle bearing information error exceeds±12 electrical degrees (shipboard) or ±7 electrical degrees (shore-based).

(b) The light also goes off if the AN/SRN-6 fails with respect to transmitter average power, transmitter code, transmitter frequency or lack of north reference group.

(4) 135 CYCLE — green light — normally on.

(a) Bearing — light goes off if transponder 135 cycle bearing information error exceeds ±20 electrical degrees (shipboard) or ±7 electrical degrees (Shore based).

(b) Auxiliary pulses — light goes off if the AN/SRN-6 auxiliary reference group pulses are missing.

(c) Transponder antenna rotation — light goes of if the transponder antenna speed changes by more than 0.8 percent of nominal, 15 revolutions per second. The light does not go off if antenna speed changes by less than 0.2 percent.

(d) The light also goes off if the AN/SRN-6 fails with respect to transmitter average power, transmitter code, or transmitter frequency.

(5) ALARM — white light normally on.

(a) Light goes off if any of the four colored lights remain off longer than their preset delay.

The ANTENNA AZIMUTH dials in shipboard installations indicate the ship's heading with respect to true north. In shore installations the dials are set to the magnetic variation of the particular geographical location by the forward-reverse switch (discussed later). The two dials operate simultaneously, the lower dial rotating nine times the speed of the upper one. The upper dial is graduated in increments of 10 degrees from 0 to 360. The lower dial is graduated in increments of one degree from 0 to 9 and contains four identical scales. The arrangement is therefore a simple vernier in which the reading of the lower dial is added to the reading of the upper dial for all values above 10 degrees. For values below 10 degrees the reading is taken directly from the lower dial.

A SERVO switch (not shown) enables the servosystem in the ON position and disables it in the OFF position. Therefore, the switch should be placed in the ON position for ship-based installations and in the OFF position for shore-based installations. The switch is located inside the monitor on top of the video chassis and access to it is gained by releasing the front panel captive screws and withdrawing the unit on its guide rails.

The function of the FORWARD-REVERSE SWITCH (fig. 11-22) is to rotate the ANTENNA AZIMUTH dials to a particular setting, thus correcting the monitor bearing circuits to read magnetic north for shore-based installations. When using the forward-reverse switch, the SERVO switch must be in the OFF position. For shore-based beacons this momentary-on switch (located behind the plate on the front panel marked REMOVE THIS PLATE TO SET ANTENNA AZIMUTH DIALS) should be held in the FORWARD (raised) position to rotate the ANTENNA AZIMUTH dials counterclockwise and in the REVERSE (down) position to rotate the dials clockwise. In the center position the switch is off.

REMOTE CONTROL GROUP

The Remote Control Group is primarily intended to permit the unattended and automatic operation of one or two TACAN (Radio Set AN/SRN-6 or AN/GRN-9) shore equipments. The equipment consists of Control-Monitor Group AN/GRA-34 (fig. 11-24) installed at the radio set site, and Remote Switching Control C-2234/GRA-34 (fig. 11-24) installed at the desired remote control area.

Control-Monitor Group

The Control-Monitor Group consists of two Radio Frequency Monitors, Antenna Transfer Switching Unit, and a Radio Set Transfer Control.

RADIO FREQUENCY MONITORS.—The switching of the two TACAN radio sets is accomplished by using two identical Radio Frequency Monitors MX-2229/GRA-34 (fig. 11-25), which provide an alarm and voltage source to initiate operation of relays. The MX-2229/GRA-34 monitors are MX-1627/URN-3 monitors (discussed previously), with Field Change 1-AN/GRA-34 incorporated.

The monitors (fig. 11-25) are normally used in a system which contains two TACAN shore
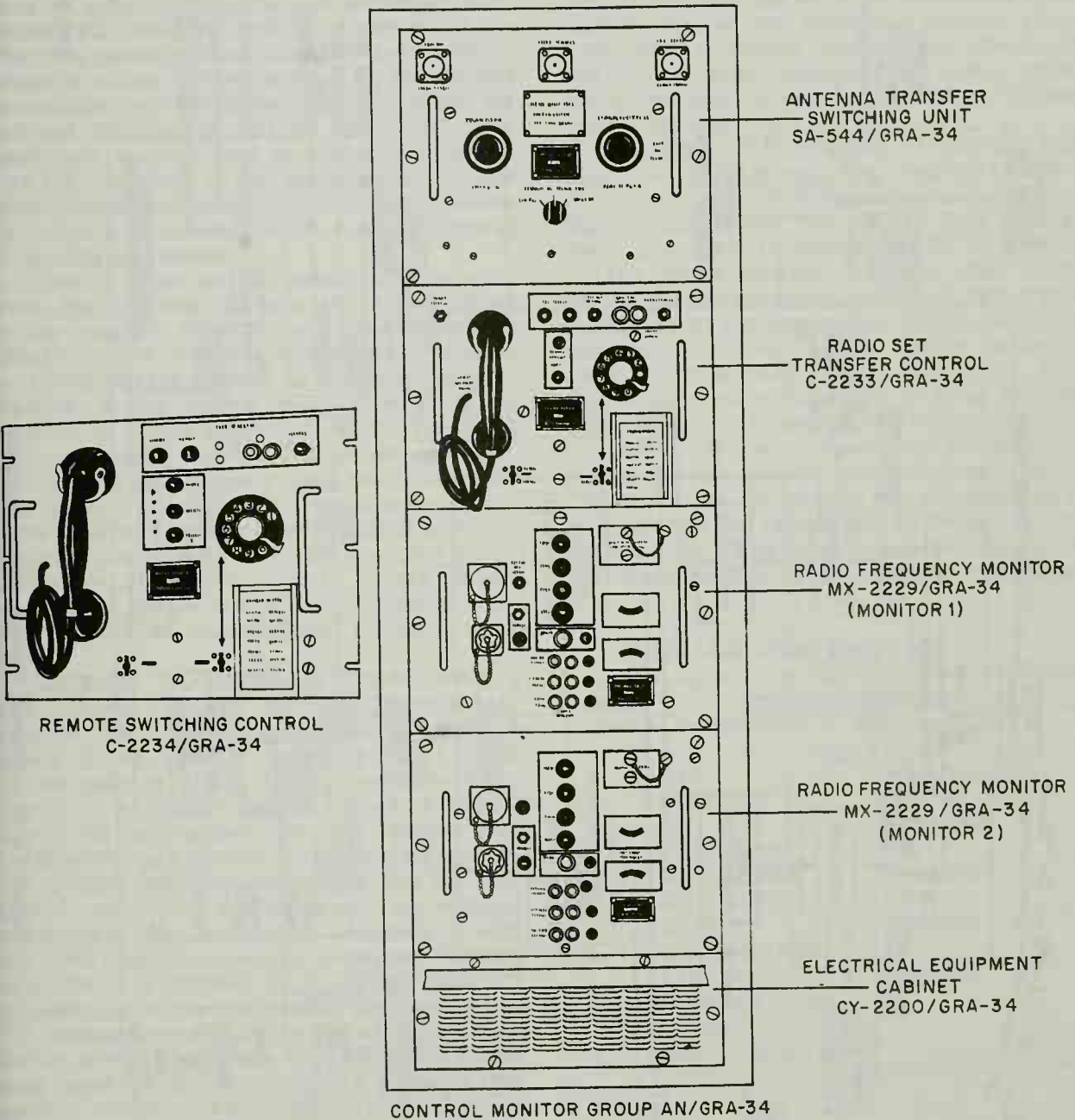
162.39

Figure 11-24.—Control Monitor Group and Remote Switching Control.

equipment transmitters, one in operation and one on standby. Although both monitors will monitor the performance of the operating transmitter, the alarm output of only one monitor at a time will be effective. This monitor will initiate a transfer action when a malfunction somewhere in the system causes an alarm. This is referred to as alarm number 1.

As a result of alarm number 1, the output of the alternate monitor will become effective and the alarm output of the original monitor will be ineffective. If the alternate monitor
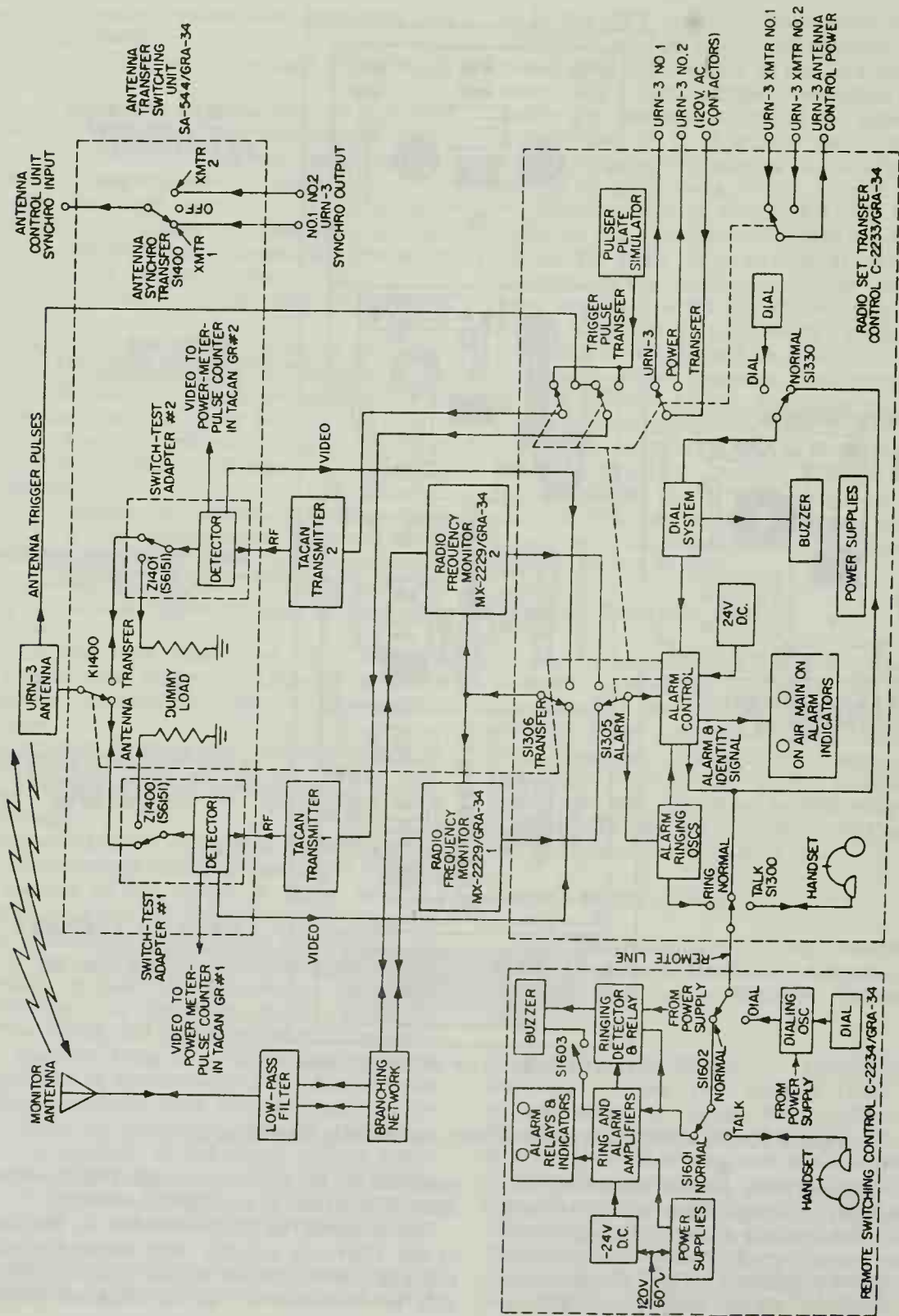
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162.40

Figure 11-25.—Overall System, block diagram.

es not indicate an alarm, the operating trans-
mitter will continue working. However, if the
alternate monitor also indicates that an alarm
situation exists (alarm number 2), the standby
transmitter will be placed in operation and the
original transmitter will be turned off. The
equipment may be used in a system with one
TACAN radio set. In this instance, the TACAN
transmitting equipment will cease to function
as a result of alarm number 2, since there is
no standby transmitter.

After a delay for the monitors to recover
from the previous alarms, the alarm output
of the original monitor once more becomes
effective. The original monitor will initiate
an alarm (alarm number 3) if there is a mal-
function in the system being monitored. As
a result of alarm number 3, the alarm output
of the alternate monitor will become effective.
If the alternate monitor does not indicate an
alarm, the transmitter will remain in operation.
If the alternate monitor also indicates an alarm
situation (alarm number 4), the operating trans-
mitter and the antenna will be turned off and
the TACAN station will be completely shut down.
The AN/GRA-34 equipment will remain in oper-
ation.

ANTENNA TRANSFER SWITCHING UNIT.—
The Antenna Transfer Switching Unit SA-544/
GRA-34 (fig. 11-24) automatically switches
either of the TACAN transmitters r-f output
to one antenna group input in response to an
alarm or a proper dialing signal (discussed
later). To accomplish this, the chassis (fig.
11-25) contains a relay controlled coaxial switch,
two switch-test adapters (dismantled SA-420/
URN-3 units), two dummy loads, and suitable
connectors. The switch-test adapters (Z 1400
and 1401) can be manually positioned to con-
nect the r-f outputs to a dummy load for
maintenance or to the coaxial switch (K 1400)
for automatic switching to the antenna group.
During normal operation, the switch-test adap-
ters must be in the ANTENNA position. The
antenna synchro transfer switch (S 1400) is
used in the initial installation of the equipment
to set the area magnetic variation. The switch
is then kept in the OFF position.

RADIO SET TRANSFER CONTROL.—The
Radio Set Transfer Control C-2233/GRA-34

(fig. 11-24) contains all of the equipment perti-
nent to manual and automatic control, alarm
indication, telephone system, and simulation
testing. It also contains the power supply
which provides power for the entire transfer
control chassis.

The MAIN ON light (fig. 11-25) will light
when the TACAN radio set master switch is
in the remote position and the transmitter
high voltage is on. The ON AIR light will light
when a monitor is operating with no alarms.

The alarm ringing oscillator (fig. 11-25),
during normal operation, will send 1900 cycles
modulated by the transmitter keying tone (from
the active monitor) to the remote handset to
identify which radio set is on the air. When
the S1300 switch is placed in Ring, the 1900
cycle oscillator is switched to 800 cycles which
is sent over the remote line to sound the
buzzer on the remote unit (discussed later).
The buzzer in the control unit is energized for
ringing from the remote unit by the dial system,
dialing "O." The buzzer may be silenced by
the Push to Stop Buzzer on the front panel.

Table 11-2.—Dial Digit Code.

Dial Digit	Program or Function	
	Transmitter 1	Transmitter 2
1	OFF	NO CHANGE
2	STANDBY	NO CHANGE
3	ON	STANDBY
4	STANDBY	ON
5	NO CHANGE	STANDBY
6	NO CHANGE	OFF
7	RESET*	RESET*
8	OBSTRUCTION LIGHTS ON	not used
9	OBSTRUCTION LIGHTS OFF	
0	RING LOCAL SITE	

* Resets automatic operation with last
  "on" transmitter as main.

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Ten different manual control functions may be performed in response to pulses from the dial system as listed in table 11-2.

The Pulser Plate Simulator switch on front of the Radio Set Transfer Control unit controls a unit which provides trigger pulses similar to those normally supplied by the TACAN antenna pulser plate. The purpose of the simulator is to permit testing of a transmitter without disturbing the operating transmitter.

The Monitor servoswitch is used to set the area magnetic variation on initial installation.

REMOTE SWITCHING CONTROL

The Remote Switching Control unit (fig. 11-24) performs the ten manual control functions in response to a dial system in parallel with the control transfer unit. The remote unit (fig. 11-25) also provide visual and aural alarm indications through lights and a buzzer using circuits similar to the control transfer unit. The remote handset monitors the operating TACAN keying tones from the control transfer unit to indicate equipment Main On. The remote handset is also used for voice communications to or from the local unit or control transfer unit.

CHAPTER 12

SEARCH RADAR

This chapter presents a block diagram analysis of Radar Set AN/SPS-10. The directional coupler and echo box used to obtain systems performance checks are considered. Height finding radar and the moving target indicator (MTI) are also discussed.

RADAR SET AN/SPS-10

Radar Set AN/SPS-10 (fig. 12-1) is used primarily in the detection, ranging, and tracking of surface targets. It can be used to a limited extent to provide the same information regarding low flying air targets.

Target range and bearing intelligence is displayed on a standard Navy Plan Position Indicator (PPI). The AN/SPS-10 can be operated as a beacon or as an IFF (identification, friend or foe) system.

GENERAL DESCRIPTION

Radar Set AN/SPS-10 operates in a frequency range of 5450 to 5825 MC. The pulsed output frequency is generated by keying a magnetron oscillator in the transmitter. The peak power is between 190 and 285 KW.

During radar operation the output pulse duration is either 0.25 or 1.3 μs ($\pm 10\%$). The pulse repetition rate (PRR) can be adjusted between 625 and 650 pulses per second. Approximately 0.08 μs of each magnetron keying pulse is required to place the magnetron in operation. Thus, the widths of the input trigger pulses are 0.33 and 1.38 μs, respectively.

For beacon operation, the pulse duration is increased to 2.25 μs, and the PRR is decreased to a range of from 312 to 325 pulses per second. The total duty cycle of the system 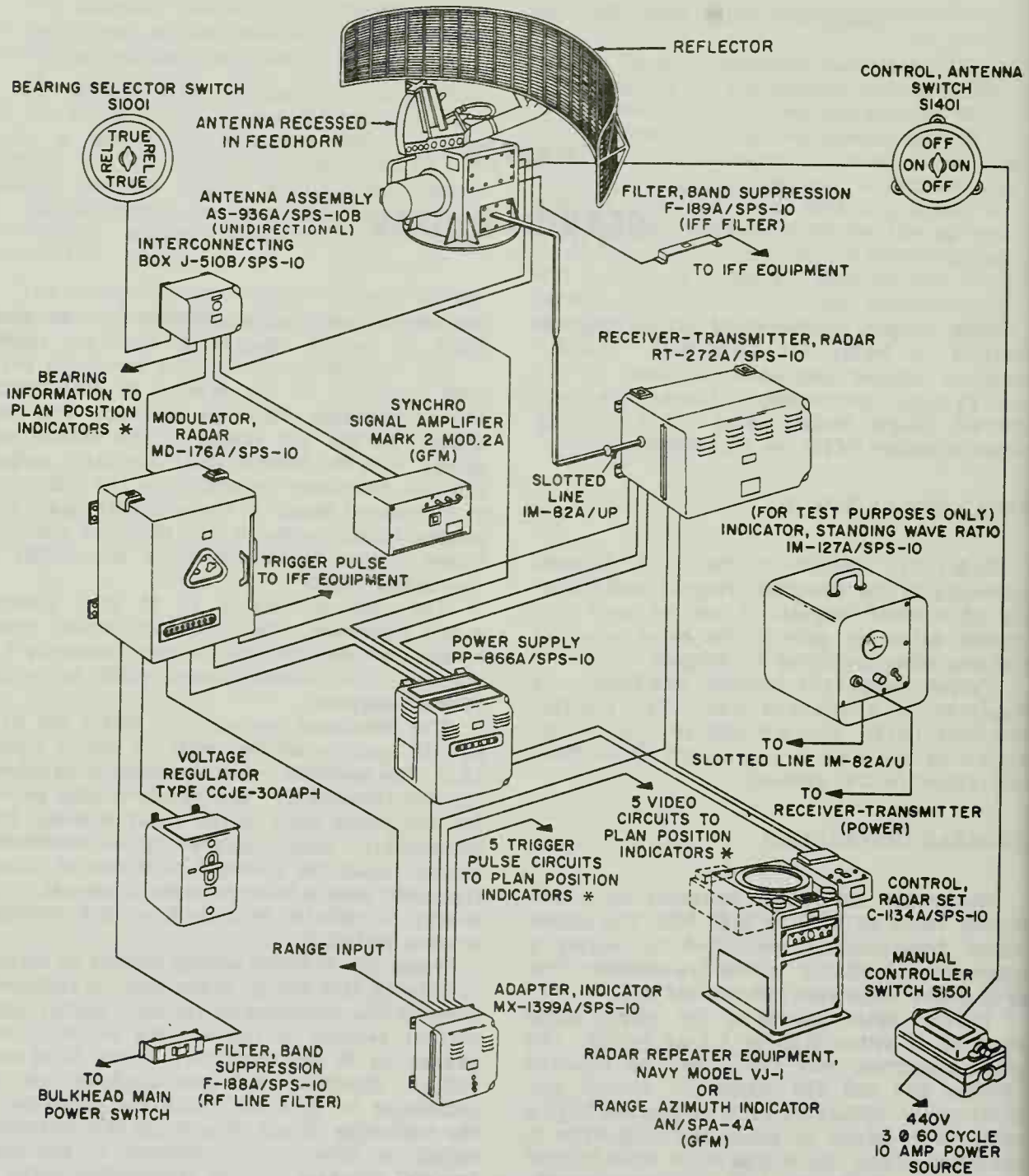(ratio of pulse duration time to pulse repetition time) must be maintained at less than 0.001 to permit magnetron recovery time.

The beacon function of the radar set permits operation of the system in conjunction with responding ships or aircraft. The response signals are mixed in the beacon receiver with the beacon local oscillator output and are ultimately presented on the indicator. The received signal is therefore stronger, and causes bright spots in the indicator pattern. These bright spots are used to identify a responding object.

The antenna rotates at 16 rpm. During radar operation, the horizontal beam width of the radiated pattern is approximately 1.5 degrees. The vertical beam width is within 12 to 16 degrees.

The functional operation of Radar Set AN/SPS-10 can be studied with the aid of figure 12-2. The modulator provides modulating pulses for the transmitter, and synchronizing pulses for the other units of the radar system. The transmitter, when triggered by the modulator output, transmits a series of bursts of radio-frequency pulses (electromagnetic waves). This energy is radiated by a unidirectional rotating antenna reflector.

When the radiated energy strikes an object, or objects, this energy is scattered or reflected. Some of this scattered or reflected energy (echo pulses) returns to the antenna reflector and passes on to the receiver, where it is amplified, detected, and then displayed on the indicators to give the range and direction of the reflecting object. The range of a reflecting object is directly proportional to the time interval required for the transmitted pulse to travel to the reflecting object and return to the receiver. The direction of the reflecting object can be translated into true or relative bearing intelligence by means of the rotating antenna reflector and a synchro system.

BEARING SELECTOR SWITCH
S1001

TRUE
REL
REL
TRUE
TRUE

ANTENNA RECESSED
IN FEEDHORN

REFLECTOR

CONTROL, ANTENNA
SWITCH
S1401

OFF
ON ON
OFF

ANTENNA ASSEMBLY
AS-936A/SPS-10B
(UNIDIRECTIONAL)

INTERCONNECTING
BOX J-510B/SPS-10

FILTER, BAND SUPPRESSION
F-189A/SPS-10
(IFF FILTER)

TO IFF EQUIPMENT

BEARING
INFORMATION TO
PLAN POSITION
INDICATORS ✱

MODULATOR,
RADAR
MD-176A/SPS-10

SYNCHRO
SIGNAL AMPLIFIER
MARK 2 MOD.2A
(GFM)

RECEIVER-TRANSMITTER, RADAR
RT-272A/SPS-10

SLOTTED
LINE
IM-82A/UP

(FOR TEST PURPOSES ONLY)
INDICATOR, STANDING WAVE RATIO
IM-127A/SPS-10

TRIGGER PULSE
TO IFF EQUIPMENT

POWER SUPPLY
PP-866A/SPS-10

VOLTAGE
REGULATOR
TYPE CCJE-30AAP-1

TO
SLOTTED LINE IM-82A/U

TO
RECEIVER-TRANSMITTER
(POWER)

5 VIDEO
CIRCUITS TO
PLAN POSITION
INDICATORS ✱

5 TRIGGER
PULSE CIRCUITS
TO PLAN POSITION
INDICATORS ✱

CONTROL,
RADAR SET
C-1134A/SPS-10

MANUAL
CONTROLLER
SWITCH S1501

RANGE INPUT

ADAPTER, INDICATOR
MX-1399A/SPS-10

TO
BULKHEAD MAIN
POWER SWITCH

FILTER, BAND
SUPPRESSION
F-188A/SPS-10
(RF LINE FILTER)

RADAR REPEATER EQUIPMENT,
NAVY MODEL VJ-1
OR
RANGE AZIMUTH INDICATOR
AN/SPA-4A
(GFM)

440V
3 ∅ 60 CYCLE
10 AMP POWER
SOURCE

✱ NAVY STANDARD PLAN POSITION INDICATORS
EITHER RANGE AZIMUTH INDICATOR AN/SPA-4A (GFM)
OR RADAR REPEATER EQUIPMENT NAVY MODEL VJ-1

32.172

Figure 12-1.—Radar Set, AN/SPS-10.
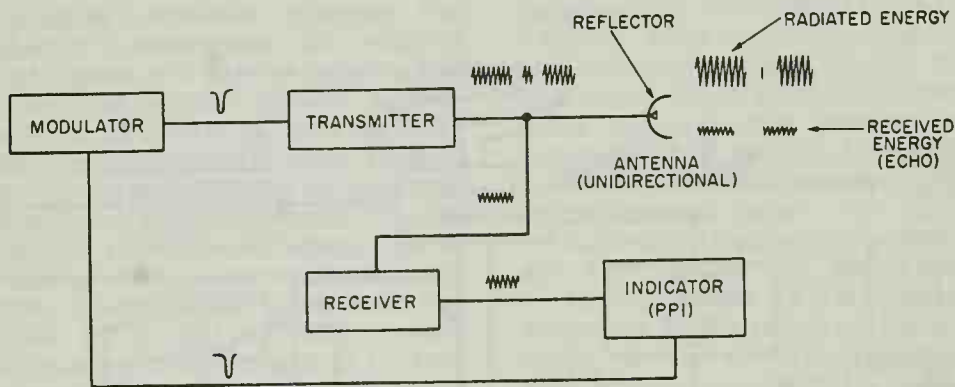
32.173

Figure 12-2.—Radar Set, AN/SPS-10, functional diagram.

RADAR SET AN/SPS-10 DETAILED BLOCK DIAGRAM

A detailed block diagram (fig. 12-3) shows all of the major units of Radar Set AN/SPS-10. The main system flow is indicated by heavy lines, and supplementary circuits by light lines.

Radar Set AN/SPS-10 is designed to operate from a primary power source of 115 volts, ±10%, 60 cycles, single phase, 90% power factor, and requires approximately 3.5 KVA. The antenna requires a power source of 440 volts, 3 phases, 60 cycle, 10 amperes. Primary power for system distribution is fed to the modulator through a line filter (Filter, Band Suppression F-188A/SPS-10). This filter serves to eliminate the transfer of RF noise from the AC power source to the radar set and from the radar set to the power source.

Primary power is delivered to the modulator and divided into regulated and nonregulated sources for distribution to the radar set. An electronically controlled voltage regulator maintains the regulated primary voltage at 115 volts, AC ± 2 percent. A power supply provides the necessary unregulated and regulated DC voltages required for proper operation of the various circuits.

In the modulator the 115 volts AC, is stepped up to a high voltage by a transformer, rectified and filtered. It is then fed through a resonant charging choke and charging diode to one of three pulse-forming networks, (WLP, a long-pulse network: WSP, a short-pulse network: or BEACON, a beacon-pulse network). Any one pulse-forming network may be selected by a motor-actuated high-voltage switch. A charge restorer diode discharges the residual mismatch voltage of the pulse-forming network after each pulse.

A thyratron keyer tube, together with one of the selected pulse-forming networks, converts the high-potential DC voltage into negative, rectangular, high-voltage pulses. The length of the pulses is sufficient to compensate for the delay in the starting of the transmitter. The transmitter then develops RF pulses having the following lengths: 0.25 (WSP), 1.30 (WLP), or 2.25 (BEACON) microseconds, depending on the pulse-forming network selected. The pulse repetition rate is determined by an RC coupled phase-shift oscillator whose frequency is variable from 625 to 650 PPS. Within these limits (625 to 650 PPS) the repetition rate is adjustable by means of a pushbutton switch, which permits the capacitance of the RC network to be varied. The pulse repetition rate for beacon operation is one-half the radar repetition rate.

The negative pulses (main pulses) generated in the modulator are conducted to a pulse transformer in the transmitter, where the amplitude is increased through the pulse transformer and fed to a magnetron oscillator, with no change in polarity or waveform. Application of the pulses to the magnetron places the magnetron in instantaneous operation, and causes it to generate microwave energy for a time interval equal to the length of each pulse.

The microwave output energy of the magnetron oscillator is conducted to the antenna assembly through a duplexer, the slotted line, and rectangular waveguide. This energy is then

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32.174

Figure 12-3.—Radar Set, AN/SPS-10, detailed block diagram.

directed by a feedhorn (foghorn type) to the surface of a slotted parabolic reflector. The antenna-reflector concentrates and radiates the energy into space in a narrow beam having a cosecant-squared vertical pattern. A cosecant-squared vertical pattern permits the radar set to receive returned signals from aircraft targets that otherwise might not be detected. The antenna reflector forms a cosecant-squared pattern by phase addition and subtraction of microwave energy components radiated to its surface by the feedhorn. As the antenna reflector rotates (16 rpm), the beam sweeps over the surrounding areas to search for both surface and aircraft targets within the limits of the height of the antenna reflector and the parameters of the radar set.

When the radiated energy strikes a target, a small portion of this energy returns to the antenna reflector. The reflector directs this reflected energy to the receiver through the waveguide, slotted line, duplexer, and TR cavity. The TR cavity acts as an automatic electronic shorting switch, unblocking the receiver to the waveguide while the transmitter is not radiating, and blocking the receiver from high-energy pulses of the transmitter while the transmitter is radiating. Two ATR (anti-TR) tubes, located near the nagnetron oscillator, present a high impedance to the returned signal energy, thus preventing dissipation of the returned signal at the magnetron oscillator during the receiving interval.

The received signal is directed through the TR cavity to a mixer. In the mixer the returned signal frequency is mixed with the frequency of the radar or beacon local oscillator. The radar local oscillator operates at a frequency 30 MC higher than the returned signal frequency. The beacon local oscillator operates at 5420 MC (30 MC below the received beacon signal). The difference-frequency component of the mixer output (intermediate frequency) is selected and amplified by the IF amplifier in the receiver.

Both the radar local oscillator and the beacon local oscillator are coupled to the receiver mixer. This method of coupling permits the use of a common IF amplifier for either beacon or radar reception. Either the beacon or radar oscillator is turned on independently for the desired type of operation.

After eight stages of IF amplification, the signal is detected to convert it to video pulses, which are then fed through a video limiter and

amplifier. The amplified video signal from the receiver is then delivered to the indicator adapter. After two more stages of video amplification, the signal is fed to five cathode followers. These cathode followers couple the video signal to plan position indicators. The low-impedance outputs of the cathode followers minimize distortion and loss of the video signal through coaxial cables.

Automatic-frequency-control (AFC) circuits are provided for the radar and beacon local oscillators. Radar AFC maintains a constant IF frequency in the receiver, regardless of minor drifts in transmitter frequency. The radar AFC circuit, incorporating a 30-MC amplifier and discriminator, compares the magnetron frequency with the radar local oscillator frequency, and thereby provides a voltage to correct the radar oscillator frequency for the proper IF frequency.

The beacon AFC circuit serves to maintain the proper local oscillator frequency during beacon operation. It utilizes a beacon reference cavity pulse which is pretuned to 5420 MC.

The sensitivity-time-control (STC) circuit reduces the gain of the receiver instantaneously in coincidence with each transmitted pulse. The gain is reduced for a short period of the sweep time; it then rises exponentially to normal gain. The STC circuit is of value in reducing sea-return saturation at close ranges so that nearby targets can be distinguished. It does not affect the presentation on the plan position indicator of targets which are beyond the STC range. The time interval needed to fully restore the receiver gain limits the range and determines the maximum effective STC range.

The fast-time-constant (FTC) circuit makes it possible to distinguish more clearly individual targets in a cluttered area, which otherwise would appear as a large mass on the plan position indicator. This feature is useful when there are many strong close-in targets. Blotting out of nearby target areas on the plan position indicator is thus greatly reduced.

The instantaneous-automatic-gain-control (IAGC) circuit prevents intense signals from saturating the IF amplifier. The IAGC circuit reduces the gain of the IF amplifier for the duration of the received signal.

The video signal that is delivered to the plan position indicator is again amplified and applied to the signal grid of the cathode-ray tube in the plan position indicator, causing intensity modulation of the electron beam. The target signal

is thus converted into visible intelligence on the screen of the plan position indicator, where it appears as a bright spot.

Circuits are provided so that the range and bearing of the target can be determined at the plan position indicator. Target range is determined from the target signal. A synchro signal amplifier, which is connected to the radar set through the interconnecting box, delivers "true bearing" (own ship's course) information to the plan position indicator. A bearing selector switch, located adjacent to the radar set control, directs either true or relative bearing information to the plan position indicator.

An IFF coaxial filter (Filter, Band Suppression F-189A/SPS-10), connected to the Antenna Assembly, minimizes interference from the radar set to the IFF equipment that is used on conjunction with Radar Set AN/SPS-10.

MODULATOR

The modulator of the AN/SPS-10 provides four pulses:

1. A negative 5 KV modulating pulse (50-ohm termination) of variable width for transmitter (magnetron) operation. (One of three pulse widths may be selected.)

2. A 25- to 50-volt positive trigger pulse (50-ohm termination for synchronization of IFF equipment.

3. A 25- to 50-volt positive trigger pulse (75-ohm termination for synchronization of the indicator adapter.

4. A 150-volt negative pulse (75-ohm termination) for synchronization of other external equipment used in conjunction with the radar set (such as test equipment).

Modulator Block Diagram

The modulator (fig.12-4) comprises two major sections, namely, the trigger pulse generator and the modulator pulse generator. The trigger pulse generator supplies a +150-volt trigger pulse to the modulator pulse generator causing the latter to produce the output modulating pulse to the transmitter and auxiliary equipment.

Synchronization of all units of the radar set is controlled by a low-power repetition rate (master) oscillator, V109. This oscillator supplies a sine wave to limiter, V108A,

at a selected repetition rate between 625 and 650 cps. An RC network, Z103, in conjunction with a PRR control determines the oscillator output frequency and the resulting pulse repetition rate.

Limiter, V108A, clips the sine wave and shaper, V108B, converts the clipped sine wave input into a squared wave output. The square wave is differentiated by R131 and C112 into sharp voltage spikes. Positive portions of the spikes are reproduced at the V106A cathode and fed through T105 to the blocking oscillator V106B input. These pulses trigger the blocking oscillator, V106 B, to produce a negative-going output and are applied to the primary of T105. The T105 secondary output pulse is positive-going to cathode follower, V105. This positive-going, + 150-volt signal is fed through T106 and serves as a trigger pulse for keyer thyratron, V104. The pulse is 4 microseconds in duration and has a sharp leading edge.

The modulator pulse generator circuit produces a high voltage main pulse (and the low-amplitude synchronizing pulses for external and auxiliary equipments) at a rate determined by the trigger pulse to the keyer thyratron, V104. Thus, the V104 trigger pulse rate is the same as the repetition rate of the oscillator, V109. Thyratron, V104 keys a high voltage circuit by periodically short-circuiting the high voltage output at the V104 trigger pulse rate.

The high voltage rectifier of the modulator pulse generator comprises T101, V101, and V102. This supply charges pulse forming network, Z101, through charging diode, V103, to approximately 14 KV. The charging diode presents an easy charge path (during the time that rectified pulsating DC voltage is rising), but at a high discharge resistance when the high voltage pulse tries to decrease. Thus, the pulse forming network cannot lose its charge until V104 fires.

Thyratron, V104, operates as a coincidence stage, and requires the simultaneous application of the high voltage at its plate and the input trigger pulse from T106 at its control grid. The conduction of V104 effectively reduces the V104 plate voltage to ground, and a low resistance path exists in parallel with the charged Z101 network. Thus, Z101 discharges through the conducting resistance of V104. The output pulse from Z101 to ground (at J103) is -5 KV, and is fed to the magnetron in the transmitter.
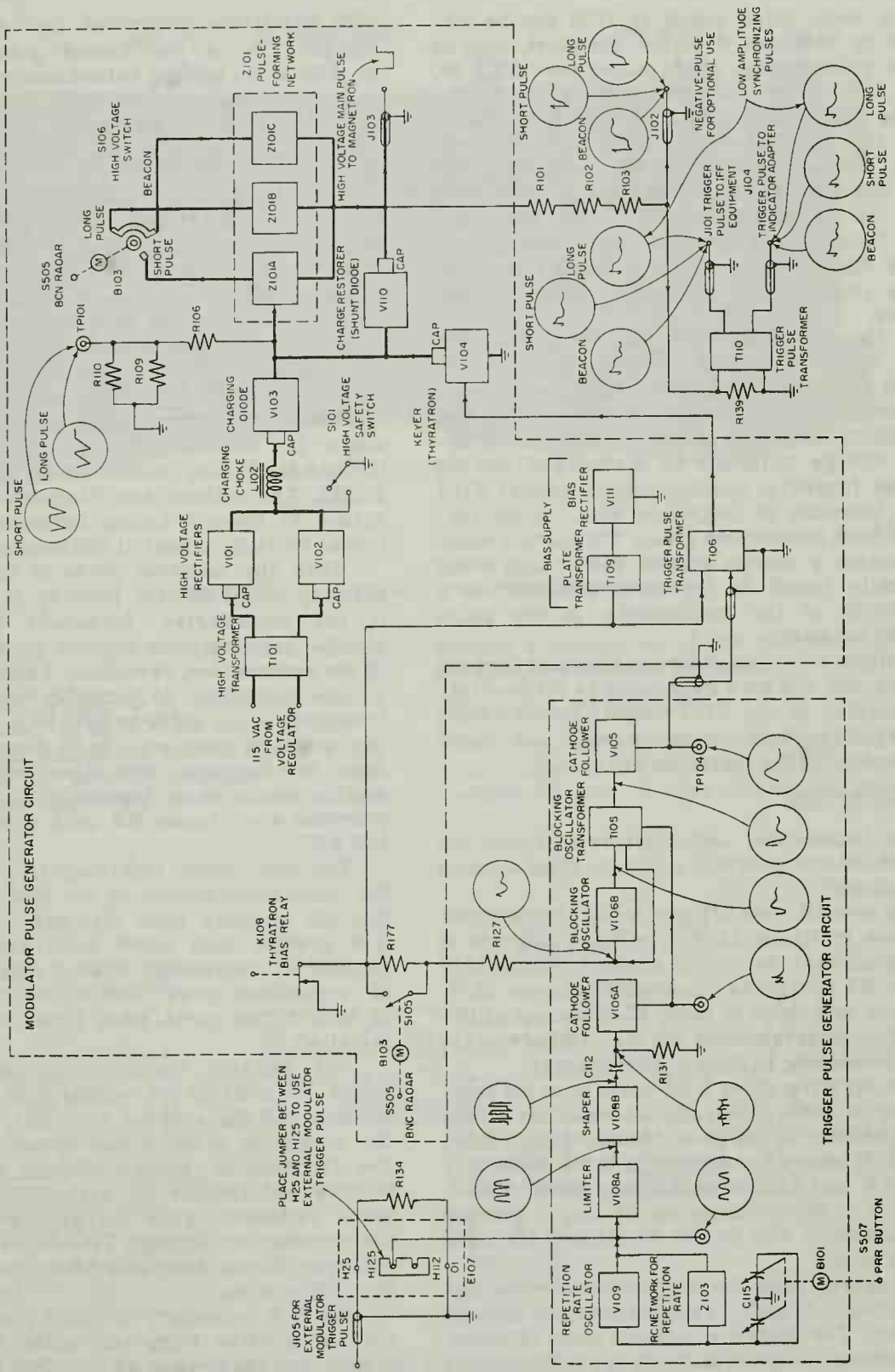
Figure 12-4.—Modulator, block diagram.

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The main pulse output at J103 can be adjusted by S505 and S106 for the short, long, or beacon pulse widths. These pulses are 0.33, 1.38, and 2.33 microseconds respecitvely.

The main pulse output at J103 is negative-going and a positive-going overshoot pulse will occur immediately after the negative-going pulse has expired. This overshoot pulse tries to make the output terminal of Z101 (at the Z101-R101 junction) positive to ground. Charge restorer diode, V110, conducts during the period of the overshoot to eliminate positive output at J103.

At the end of the V110 conduction, charging diode, V103, again conducts to charge Z101. The charge restores diode, V110 which remains nonconductive until the end of the next pulse.

Without the action of V110, the positive overshoot voltage could not be discharged, and the reverse (positive) residual charge across Z101 would increase in magnitude with the application of each subsequent pulse. This action would accumulate a charge across V104 which would eventually result in thyratron arc-over or a breakdown of the components in the pulse forming network.

A slight mismatch between the pulse forming network and the load impedance is deliberately incorporated at the Z101 output. The resulting positive overshoot pulse is used for rapid deionization of the thyratron keyer.

TRANSMITTER

The transmitter and receiver of Radar Set AN/SPS-10 are operated using the same antenna and duplexer.

The overall view of the receiver-transmitter is shown in figure 12-5. The block diagram of the transmitter section is shown in figure 12-6 and the RF system is illustrated in figure 12-7. Frequent reference to these illustrations will be helpful in understanding the material presented in the remaining portion of this chapter.

The negative-going 5 KV pulse from the pulse forming network, Z101, (fig. 12-4) is fed via J103 to the primary of the pulse transformer, T3002, (fig. 12-6) where it is stepped up in the secondary to 20 KV and applied to the magnetron, V3001. The magnetron produces an RF output between 5450 and 5825 MC for the duration of the input trigger pulse.

RF energy from the magnetron is fed through the duplexer, W3010, to the antenna and antenna reflector. The duplexer contains two ATR tubes, a directional coupler, and a radar AFC coupler

with adjustable attenuator (not shown). The duplexer is of the branch guide type, that is, with arms leading to the broadband and do not require tuning. A choke joint (air gap) couples the duplexer to the magnetron.

Transmitter Circuit Analysis

The -5 KV pulse from J103 of the modulator (fig. 12-4) is fed through E3018 (fig. 12-8) to the primary of pulse transformer, T3002. The input pulse duration is 0.33, 1.38, or 2.33 microseconds, depending upon the combination of the sections in the pulse forming network as discussed. The T3002 secondary voltages are approximately 20 KV. The parallel windings equalize the voltage on both sides of the magnetron filament. Filament transformer, T3001, provides 11 volts for the magnetron filaments through the bifilar (noninductive) secondaries of T3002. The cathode and filaments are tied together to prevent arcing between these electrodes by large potential differences.

When the negative pulse is applied to the primary of T3002, the polarity at the left end of the secondaries, terminals 5 and 6, is always negative with respect to the right end of the secondaries, terminals 3 and 4.

The magnetron is manually tunable over its frequency range (5450 to 5825 MC). This enables the operating frequency to be changed to minimize interference from other radar sets operating at the same frequency. The magnetron provides a minimum RF peak power output of 160 KW.

The high-power high-frequency pulse from the magnetron passes to the duplexer, W3010, through a choke joint (air gap). The air gap (not greater than 0.060 inch) allows for expansion of components without appreciable loss of transmitter power and avoids the necessity of hand-fitting parts when a new magnetron is installed.

The duplexer conducts the magnetron RF output to suitable waveguide and then to the antenna and the antenna reflector, from which the energy is radiated into space. A portion of the transmitted energy strikes a target or targets and returns (by reflection) to the antenna reflector. This energy travels through the duplexer and through TR tube, V3002, to the receiver during intervals when the magnetron is not operating.

The 7-8 secondary of T3002 produces a low voltage positive triggering pulse for the STC circuit and the beacon AFC circuit in the radar
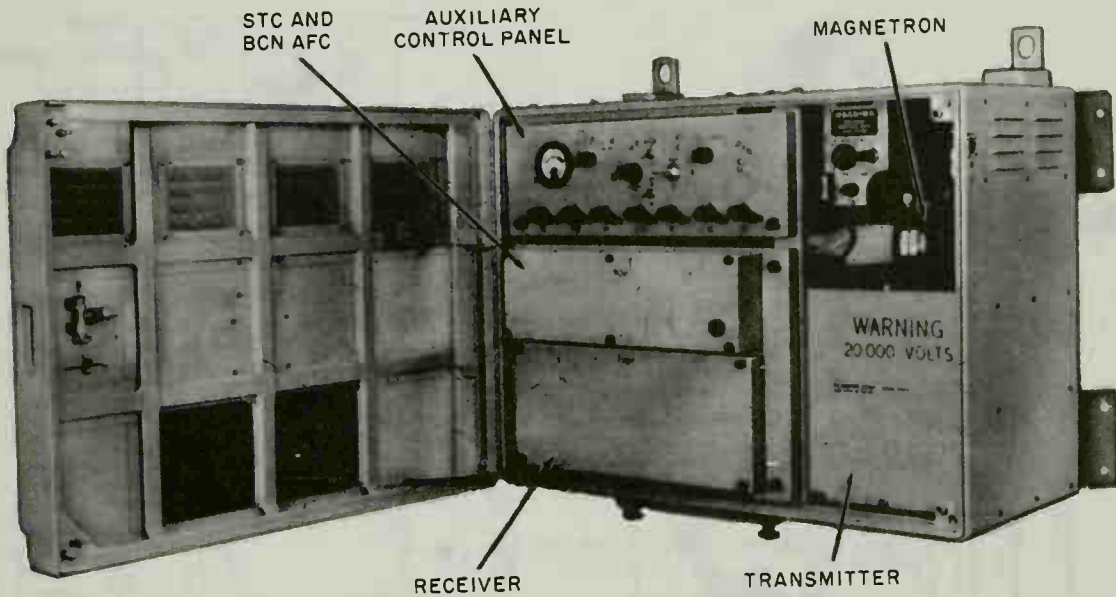
32.179

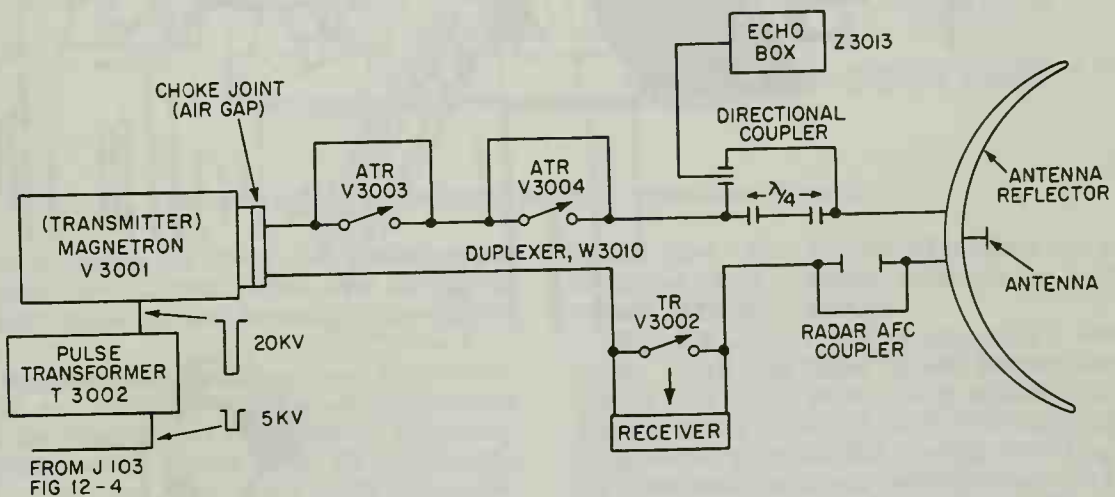Figure 12-5.—AN/SPS-10, receiver-transmitter.

receiver. These circuits will be discussed later.

Spark gap, E3006, is a protective device which arcs-over at 2200 volts peak. In case of an open in the magnetron ground lead, the gap will arc-over to prevent excessive transient voltages from appearing in the circuit and causing a breakdown of the capacitors (C3033 and C3004) dielectric material.
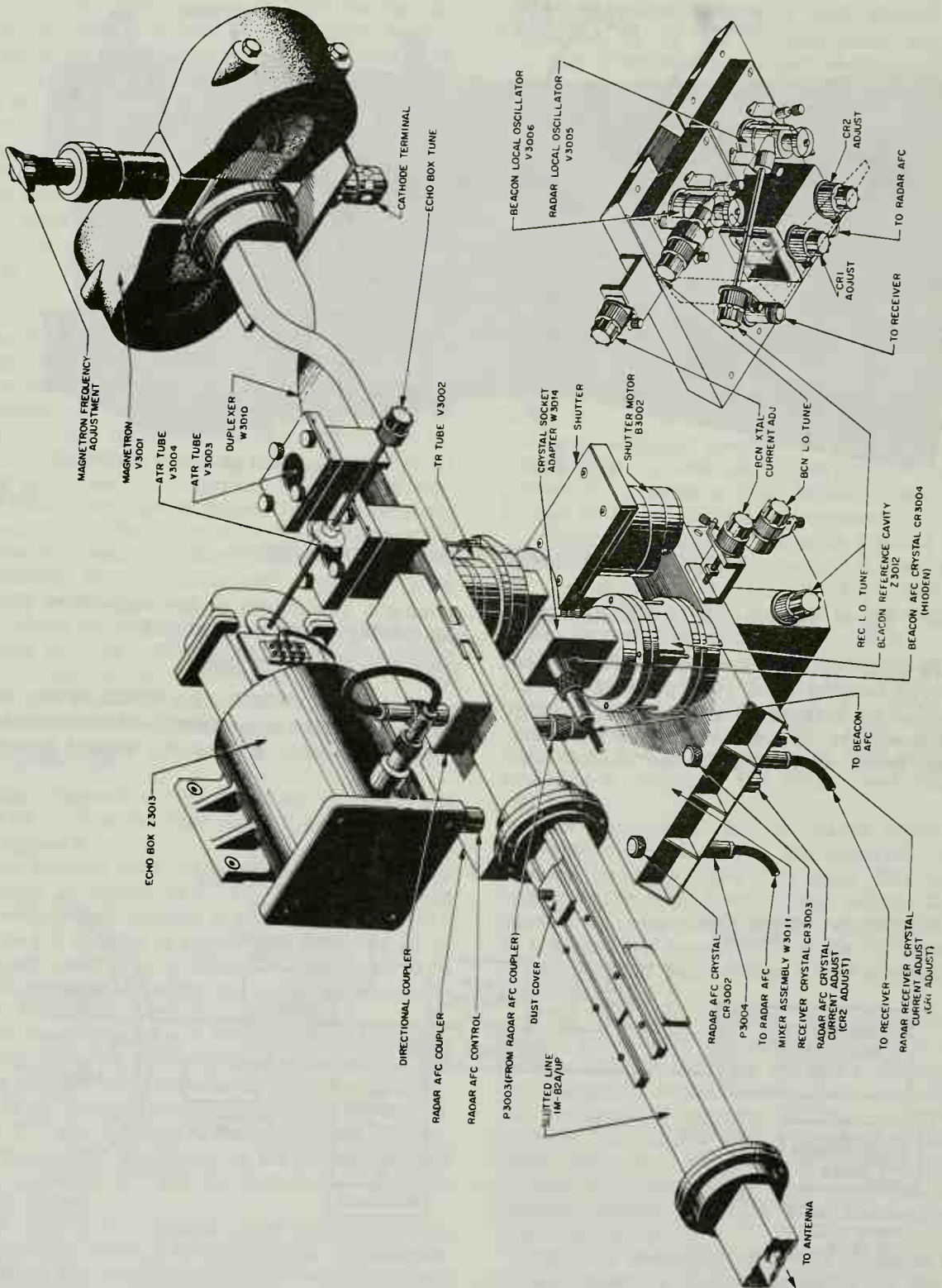
MICROWAVE SWITCHING

Th duplexer W3010, of AN/SPS-10 (fig. 12-8) makes possible transmission and reception from a single antenna. It does this without damaging

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Figure 12-6.—AN/SPS-10, transmitter, block diagram.

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Figure 12-7. — R-F section of receiver-transmitter in AN/SPS-10.

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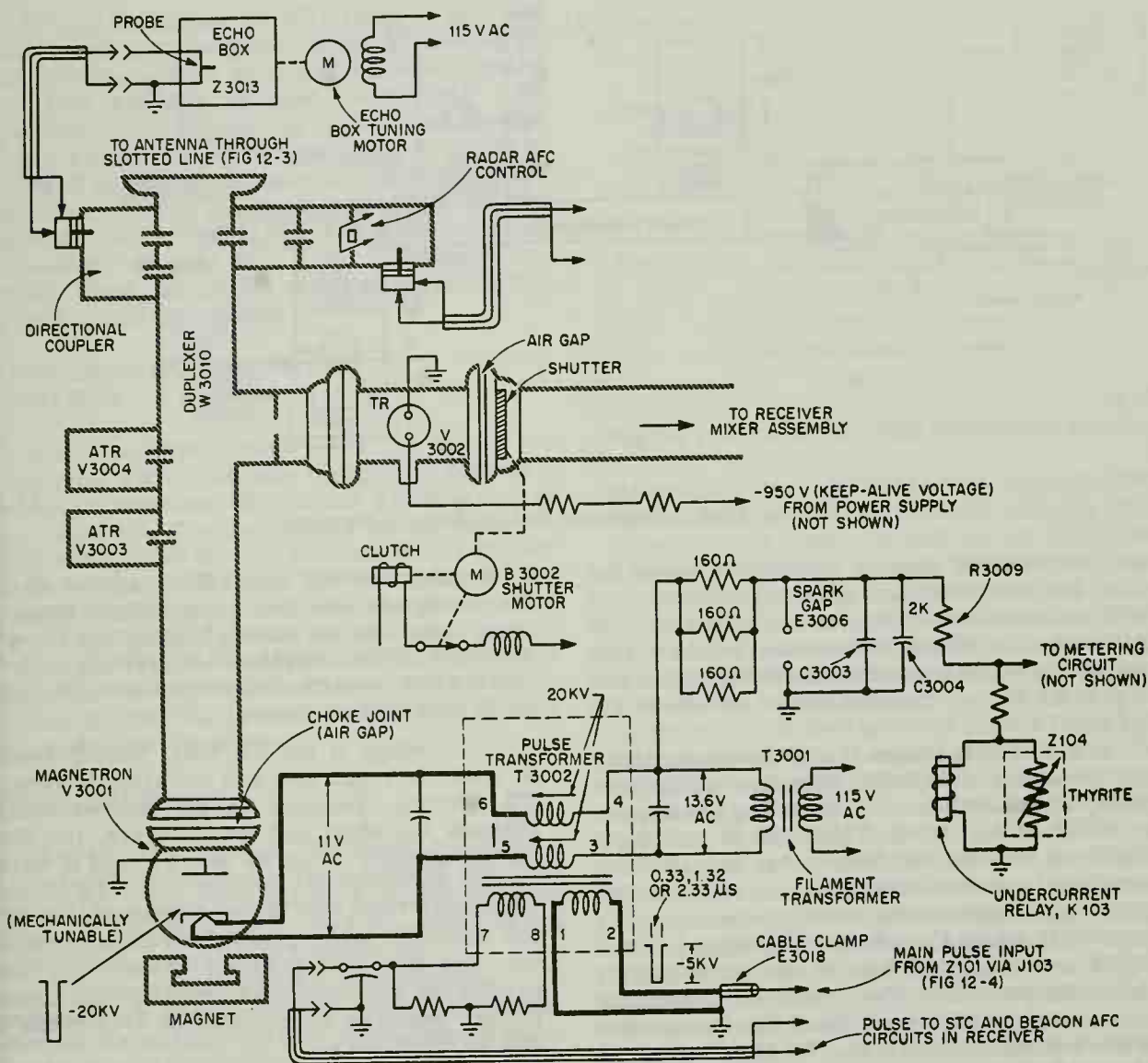
32.182

Figure 12-8.—Transmitter section, schematic diagram.

the receiver during the transmission of the high-power pulse and without undue loss of the re-ceived echo signal in the transmitter during re-ception.

The duplexer uses an arrangement of micro-wave (TR and ATR) switches which distinguish between the transmitted and received pulses by the relative power levels. One TR (transmit-receive) tube and two ATR (anti-transmit-receive) tubes are used. These tubes are posi-tioned electrically on the duplexer as shown in

figure 12-9. The TR and ATR tubes are essen-tially the same in action, although the ATR tube is less critical in design.

The TR tube V3002 is basically a resonant cavity spark-gap tuned to the transmitter fre-quency and filled to a low pressure with gas. The switching action of the tube is accomplished automatically by the RF energy, resulting from the difference in power level between the trans-mitted pulse on and off condition. The trans-mitted pulse is sufficiently strong to strike an
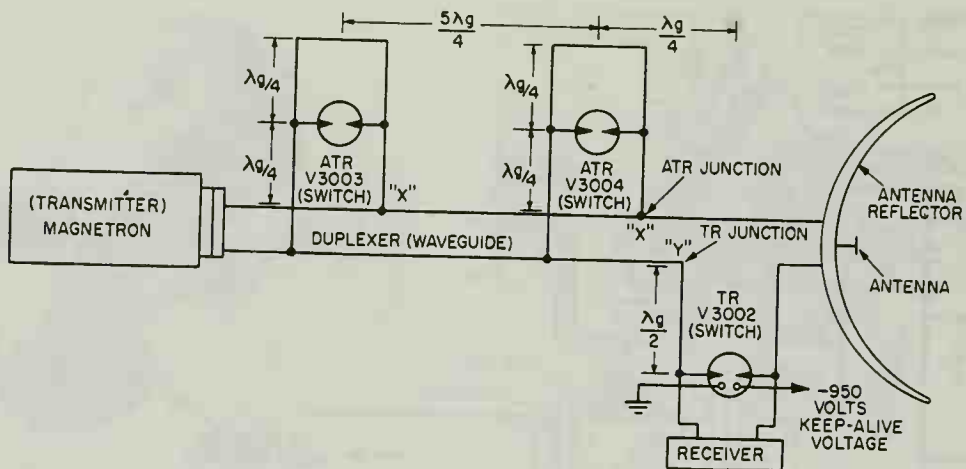
Figure 12-9.—Duplexer (microwave switching).

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arc (for a short circuit) across the gap of the tube. The combination of spark gap and resonant cavity increases the instantaneous voltage across the gap before firing. This causes the tube to fire nearer the beginning of the transmitted pulse, and to present a more complete short circuit during the arc.

In order to increase the TR tube sensitivity and to ensure complete ionization during each pulse, a -950 volt keep-alive voltage is applied to V3002 at all times. This voltage maintains V3002 at a point just below that required for complete ionization. The high-power transmitter pulse will immediately raise the voltage to a point well above the critical ionization point of V3002 and cause the tube to arc more rapidly and more smoothly. The target echoes contain considerably less power than the transmitted pulse and will not fire the TR tube during the interval between transmitted pulses.

The duplexer has two ATR tubes spaced one and one quarter wavelengths apart (mid-band). Two tubes are used in order to ensure effective firing over the full magnetron frequency band. The actual spacing of the tubes is in terms of fractions of the duplexer (waveguide) wavelength which is longer than the free-space wavelength. Because of fringing effects at the waveguide junctions, the actual spacing of these components is determined experimentally.

When the magnetron, V3001, oscillates, the high-power microwave burst passes into the duplexer and energizes the TR and ATR tubes by striking an arc in them. The shorts in the

ATR tubes (V3003 and V3004) appear across the waveguide one-fourth wavelength from the open ends. The impedance looking into the ATR junctions from points X is infinite and the transmitter energy continues down the waveguide toward the antenna.

The arcing of the TR tube, V3002, results in a short-circuit one-half wavelength from the TR junction. Because of the half-wavelength between the short and the duplexer, the short circuit appears to the RF energy as if it were a closed switch at the opening to the TR junction (point Y). Thus, the RF energy continues down the duplexer to the antenna and antenna reflector. The only loss in power encountered in the passing of the magnetron output to the antenna is that required to maintain the arcs in the TR and ATR switches.

The input path to the receiver is short-circuited during the arc-over in V3002 so very little of the transmitted energy enters the receiver during magnetron operation. The TR tube arc-over does not represent zero resistance but will be of the order of a few ohms. The small input voltage developed across the input terminals of the receiver is well below a value which will damage the receiver.

When the magnetron, V3001, is off (receive period) the TR tube, V3002, is effectively open, and the waveguide is not short-circuited at the TR junction. The received echo pulse will then pass with ease along the matched input path to the receiver.

During reception, the ATR tubes (V3003 and 3004) are not energized (open). In this condition, the impedance at points X looking toward the ATR switches is zero, since both ATR branches now appear as half-wave shorted stubs. The line is effectively short-circuited at points X and the impedance at Y looking toward the transmitter is that seen at the open end of a shorted quarter-wave stub and a five-quarter wavelength shorted stub (both impedances infinite). Thus, all of the received energy (echo) is routed into the receiver.

DIRECTIONAL COUPLER AND ECHO BOX

The directional coupler (fig. 12-6) and the echo box, Z3013, function together to provide a means of checking the overall performance of the radar set. Transmitted RF energy from the waveguide is fed through the directional coupler to the echo box. The echo box stores the energy during transmission and releases it to the receiver immediately after the transmitting period. This received energy from the echo box is displayed on the PPI-scope and can be interpreted by the observer to give an indication of the system performance.

Directional Coupler

As the name implies, the directional coupler is so connected in the transmission line that either the transmitted energy (traveling down the line) or the reflected energy (traveling in the opposite direction) may be sampled.

The one-way directional coupler (fig. 12-10) samples RF energy traveling from the magnetron to the antenna and rejects energy traveling in the opposite direction.

The unit is coupled to the main waveguide through two small holes located one-quarter wavelength apart. One end of the directional coupler is terminated in its characteristic impedance, and absorbs most of the energy reflected to that end of the coupler. The output from the coupler is taken from a probe at the other end, and fed through a coaxial cable to the echo box probe.

A small portion of the magnetron RF energy (incident wave) is fed into the two holes on the directional coupler. The voltage at the probe in the directional coupler consists of two components (E_{ac} and E_{abc}). The paths traveled by the two components from the magnetron to the
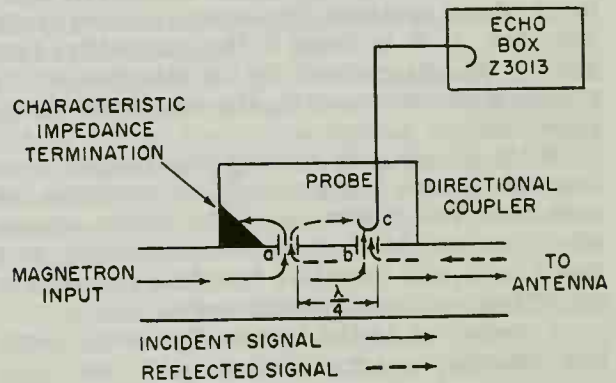
32.184

Figure 12-10.—Directional coupler and echo box.

directional coupler probe are approximately equal and the voltages are only slightly out of phase. These voltages add at the directional coupler probe, and a large incident voltage is produced at the echo box probe.

There are also two paths for reflected energy which exist in the guide because of slight mismatches in impedance between the duplexer and antenna. Along one path the reflected signal (indicated by dashed arrows) passes from b to c in the directional coupler, while in the other path the reflected signal travels one-quarter wavelength farther to a and an additional one-quarter wavelength back to c. The latter path is one-half wavelength longer than the original path. The two reflected voltage components therefore arrive at the directional coupler probe 180 degrees out of phase and cancellation occurs. Thus, the input RF energy from the magnetron to the echo box probe is essentially free of reflected RF signal in the waveguide.

The ratio of power traveling up the waveguide (from magnetron to antenna) to power traveling down the guide (from antenna to magnetron) is greater than 15 db. The standing-wave ratio resulting from the insertion of the directional coupler is negligible.

Echo Box

The echo box, Z3013 (fig. 12-10) is equivalent to a high-Q tuned circuit. The energy stored in the echo box during the transmission interval decays exponentially during the receive interval. The time required for the energy to drop to the noise level of the receiver is known as the "echo box ringtime," and depends on two factors: (1) the amount of the transmitted energy

stored by the echo box; and (2) the overall noise level of the receiver. The energy stored by the echo box, if it is tuned to the transmitter frequency, is determined by the duration of the transmitted pulse and by the magnitude of the pulse energy.

If the power output of the transmitter were decreased, less energy would be stored in the echo box. The discharge of the stored energy after the transmitted pulse would begin at a lower level, and therefore take less time to decay. This would yield a shorter ring time.

A reduction of the transmitter power (echo box charging energy) to one-half (3 db down) will reduce the ringtime approximately 1 microsecond. The noise level of the receiver sets the level at which the ringing is no longer perceptible on the plan position indicator. An increase in the receiver noise level by 3 db will also reduce the ringtime approximately 1 microsecond.

Because the magnetron frequency is tunable, the echo box must be tunable over the same frequency range. An echo box tune knob is provided on the receiver-transmitter (fig. 12-7) to manually tune the echo box cavity to the magnetron frequency.

When the equipment is operated from the radar set control (fig. 12-1), the echo box tuning is accomplished by a motor (not shown). The motor is operated by a pushbutton switch on the radar set control panel marked ECHO BOX. The motor drives the echo box tuning shaft so that the tuning sweeps over the entire range of transmitted frequencies. Ringing will occur whenever the resonant frequency of the echo box is the same as the output frequency of the magnetron.

The echo box ringtime indications appear on the plan position indicator as lobes (fig. 12-11, A). These lobes should extend outward on the indicator a minimum of 4000 yards. The echo box lobes will appear at a rate of 10 times per minute. When the echo box switch is off, the lobes will not appear in the PPI pattern (fig. 12-11, B).

ANTENNA ASSEMBLY AND ANTENNA CONTROL CIRCUITS

The antenna assembly (fig. 12-12) radiates the pulses of microwave energy from the magnetron, and directs the echo signals to the receiver through the waveguide as discussed. The antenna assembly consists of a truncated parabolic reflector, a pedestal, a feedhorn assembly, and a synchro (located in the antenna pedestal) which provides bearing information.

The antenna pedestal also contains a unidirectional 440-volt 60-cycle 3-phase 1-horsepower antenna drive motor, B1301 (fig. 12-13). Power for the motor is applied through manual
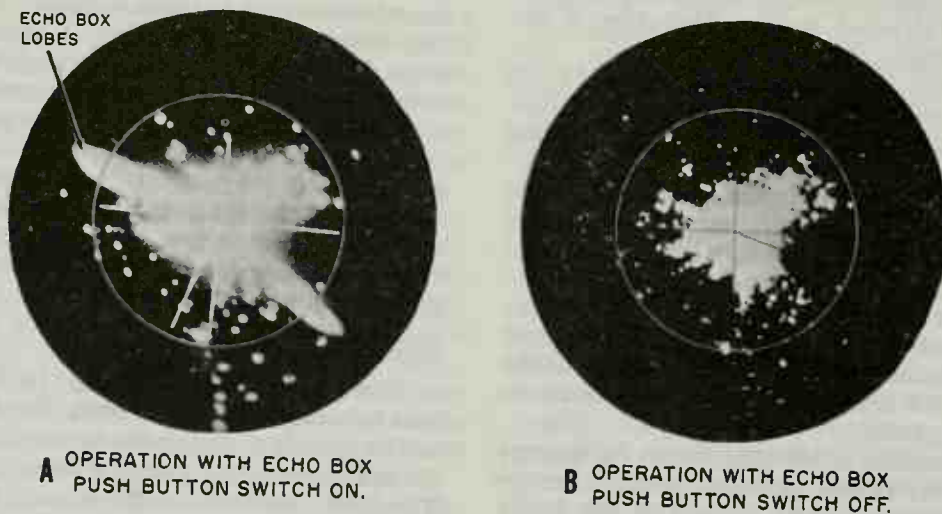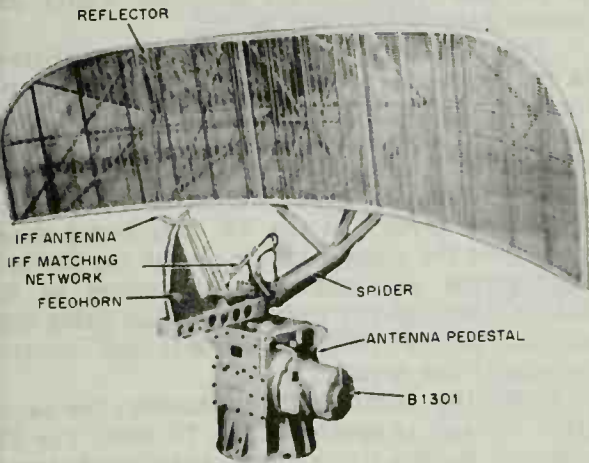

A OPERATION WITH ECHO BOX
   PUSH BUTTON SWITCH ON.

B OPERATION WITH ECHO BOX
   PUSH BUTTON SWITCH OFF.

Figure 12-11.—Echo box patterns.

32.185

32.186

Figure 12-12.—Antenna assembly.

controller switch S1501, and control antenna switch S1401. The motor turns at 1725 rpm.

An oil heater, HR1301 (in the antenna pedestal) provides a means of maintaining the ambient temperature of the motor lubricant above 25° C. If the temperature drops below this value, a heater switch (not shown) closes to allow a heater warmup current to flow through the heater element.

Drive motor B1301 is mechanically coupled to the antenna through a reduction gear train and a rotary joint. The turning of the rotary joint causes rotation of a main shaft (which drives the antenna), a 5G synchro, and two cam-operated microswitches. The gear reduction to the main shaft causes the antenna to turn clockwise at 16 ± 1 RPM.

The main shaft is sufficiently locked by the action of the gear train during rotation to prevent error introduction in the antenna speed by normal forces acting on the antenna reflector. When S1501 or S1401 is open, B1301 does not rotate. The locking feature provided by the gear train when the antenna is not rotating prevents the reflector from drifting during high winds, and acts as a safety during maintenance.

Synchro B1302 is driven at the same speed as the antenna reflector, and provides an output voltage to the synchro signal amplifier which is proportional to the position of the antenna reflector. During alignment, the antenna reflector is rotated (by operating S1501) until it points directly toward the ship's bow. The synchro, B1302, is then rotated to produce an electrical
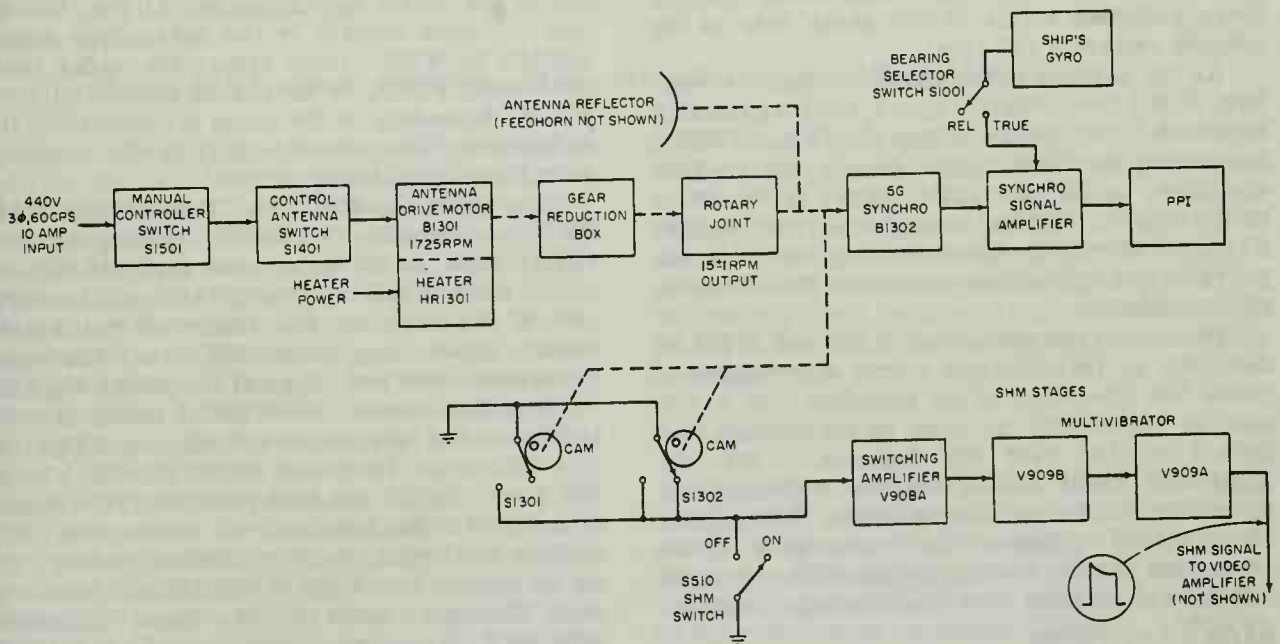
32.187

Figure 12-13.—Antenna pedestal and antenna control circuit, block diagram.

435

zero. In operation, the antenna reflector and B1302 maintain correspondence. The B1302 output is fed through the synchro signal amplifier to the PPI-scope. This voltage drives the sweep on the indicator at the same rate and through the same angle as the antenna reflector, and bearing information relative to the ship's bow can be read from the indicator.

The bearing selector switch, S1001, provides a means of selecting either true- or relative-bearing information for presentation on the plan position indicators. When S1001 is in the TRUE position (opposite to the position shown), gyro compass information and relative bearing information (from B1302) are combined in the synchro signal amplifier. With these inputs, the synchro signal amplifier computes true-bearing information for the plan position indicators.

The antenna assembly also contains two cam-operated microswitches, S1301 and S1302. These switches produce the ship's heading marker (SHM) on the indicator each time the antenna reflector crosses the ship's bow. The marker appears as a group of dashed lines which extend outward on the ship's trace. An SHM switch, S510, provides on-off control of the marker signals.

When S510 is in the ON position, the operation of the SHM stages (V908 and V909) is controlled by S1301 and S1302. The cams which operate these switches rotate at the same rate as the antenna reflector (15 rpm).

As the antenna rotates past 000 degrees relative, S1301 momentarily opens, and the ground is removed from the switching amplifier, V908A, permitting the SHM stages to operate. In this condition, a 100-KC SHM signal is fed from multivibrator, V909A, through the video amplifier (not shown) to the indicator, where it appears on the indicator sweep as a 100-KC intensity modulation.

The cam arrangement of S1301 and S1302 is designed so that there is a time difference between the operation of the switches. The S1302 cam is set so that as soon as the antenna reflector rotates past 000 degrees, S1302 will close and shunt S1301, thereby replacing the ground on V908A and cutting off the SHM output. Thus, S1302 prevents the occurrence of an erroneous shm on the display due to the rise and fall time of a single cam-operated microswitch.

RADAR RECEIVER

A simplified block diagram of the radar receiver in the AN/SPS-10 is shown in figure 12-14.

The microwave portion of the receiver consists of TR tube V3002, mixer assembly W3011, and beacon reference cavity W3012.

The input echo signal to the radar receiver is fed from the antenna and antenna assembly through a slotted line, a TR tube, V3002, and a shutter to the mixer assembly, W3011. The shutter is operated by a shutter motor (not shown). When the radar set is shut down, the keep-alive circuit in the TR tube is not energized, and the shutter motor closes the shutter. This action prevents a large amount of RF energy from nearby radar sets from being dissipated in the receiver crystal and causing damage to the crystal.

When the radar set is in standby, the keep-alive voltage is applied to the TR tube. The TR tube will readily fire and present a short circuit across the receiver input terminals when strong r-f signals are received. The shutter opens slowly when the radar set is energized, and is not fully open until the TR tube is partially ionized by the keep-alive voltage. When power is removed from the shutter motor, a motor clutch closes the shutter.
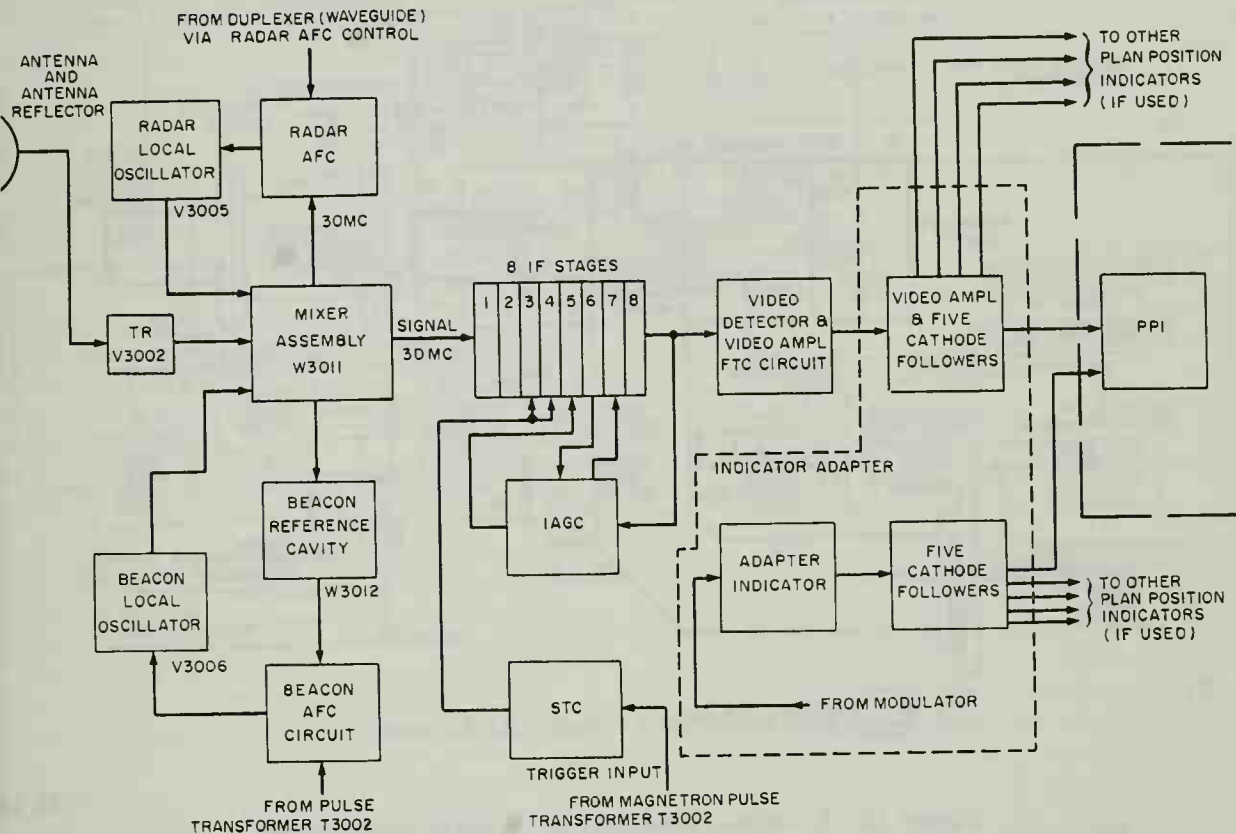
Mixer Assembly

As in conventional mixers, two signals are fed to the mixer assembly, W3011 (fig. 12-15). One of these signals is the heterodyne signal applied to W3011 from either the radar local oscillator, V3005, or the beacon local oscillator, V3006, depending on the mode of operation of the radar set. The second signal is the received echo from the antenna.

The mixing in W3011 by CR3003 is similar to the more familiar action in a pentagrid converter tube. In the latter case separate signals on the control and injector grids simultaneously control the plate current. The result yields plate output signal components which are the sum, difference, and two original frequency signals.

In a like manner, the CR3003 output is controlled by the heterodyne and echo input signals. The difference frequency output of W3011 is 30 MC for radar or beacon operation. This signal is selected in the tuned circuit comprising C301 and the T301 primary. The T301 secondary signal is applied to the 1st IF amplifier. There are eight IF stages which are conventional amplifiers with STC (sensitivity-time control) and IAGC (Instantaneous Automatic Gain Control).

The radar local oscillator, V3005, and beacon local oscillator, V3006, are type 6115 reflex

32.190

Figure 12-14.—Radar receiver and adapter indicator, simplified block diagram.

klystrons. These oscillators employ resonant cavities as the frequency controlling device, which supplies microwave energy to the crystal mixer assembly, W3011. The klystrons (V3005 and V3006) are velocity modulated, i.e., the RF potential on the klystron cavity alternately causes acceleration and deceleration of the electron stream. This action produces a grouping or bunching of the electrons in the klystron. The frequency at which the potential changes in the klystron is determined by the resonant frequency of the klystron cavity.

Reflex klystrons V3005 and V3006 each contain a built-in cavity which is tunable by changing the vacity dimensions. A screw attached to the strut on the V3005 klystron (fig. 12-15) is operated by the receive local oscillator tune control. As the control is turned, a stress is exerted on the strut causing it to increase or decrease the cavity dimensions. This permits the radar local oscillator to be tuned 30 MC

higher than the incoming frequency, or over a range from approximately 5480 to 5855 MC.

Because beacon reception takes place on 5450 MC only, the beacon oscillator must operate at a single frequency. Klystron V3006 is tuned 30 MC below the incoming frequency by the beacon local oscillator tune control to 5420 MC. The beacon signal is fed through the same IF stages as in radar operation.

Radar AFC Circuit

The radar AFC circuit (fig. 12-16) maintains a constant IF signal in the receiver, regardless of minor drifts in the transmitter (magnetron) frequency. The method used to obtain automatic frequency control is to vary the repeller voltage of the radar local oscillator klystron, V3005, by means of a sawtooth voltage. The AFC circuit locks-in at a frequency which is 30 MC above the magnetron output.
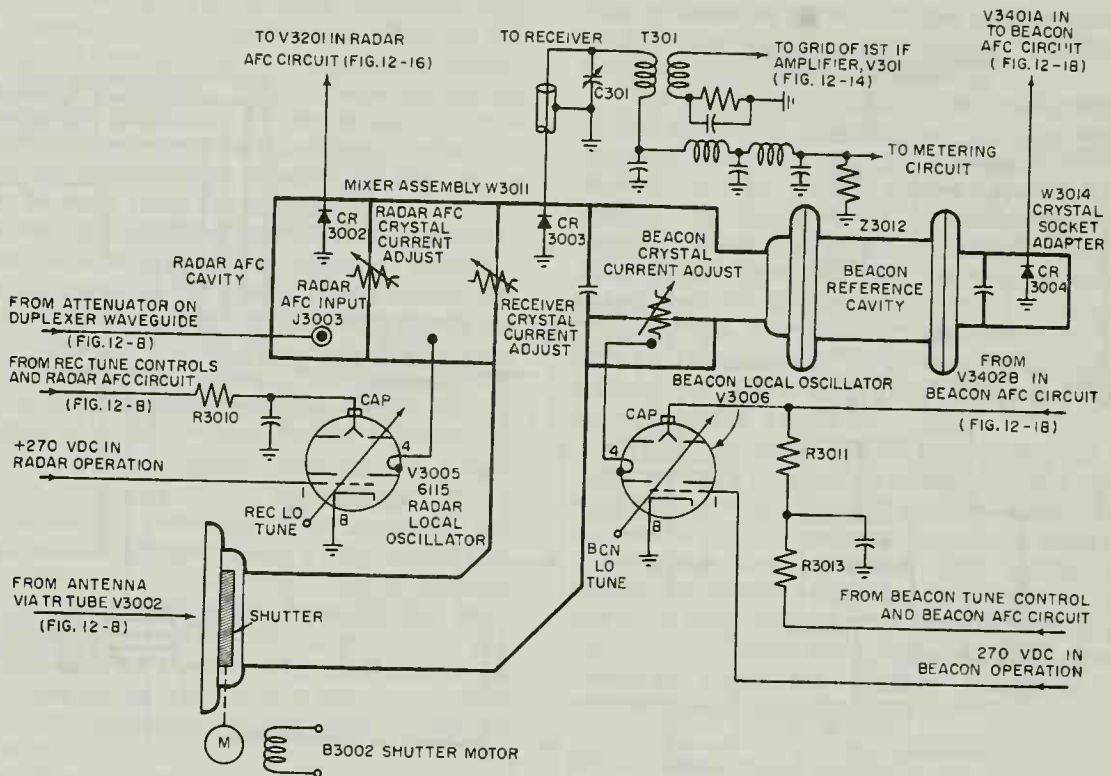
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32.192

Figure 12-15.—Mixer assembly of receiver-transmitter.

The radar AFC coupler, connected to the waveguide, couples a small portion of the transmitted energy to CR3002 in the radar AFC mixer. A radar AFC control permits adjustment of the amount of the transmitted signal fed to the mixer.

An input from the radar local oscillator, V3005, is also applied to CR3002 via a radar AFC crystal current adjust. The two mixer input signals differ in frequency by 30 MC ± the transmitter magnetron error frequency.

The output of the AFC mixer crystal, CR3002, consists of pulses of RF energy which recur at the same rate as the transmitted output. This signal is amplified in a two-stage IF amplifier strip comprising V3201 and V3202. The bandwidth characteristics of the IF amplifiers allow passage of the 30 MC ± magnetron error signal but rejects all other frequency components. The output of IF amplifier V3202 is fed to an S/B AFC discriminator, V3203.

A schematic diagram of the radar AFC circuit is shown in figure 12-17. The discriminator (V3203A and B) comprises L3201 and L3202 which are tuned to resonate with distributed capacitance at 32.5 and 27.5 MC, respectively (2.5 MC above and 2.5 MC below the 30 MC desired intermediate frequency).

The radar AFC circuit is actually a closed loop, with the radar local oscillator, V3005, acting as the signal source in the loop. The V3005 output (5480 to 5855 MC) is heterodyned in the radar AFC mixer cavity with the AFC signal from the waveguide (5450 to 5825 MC) as discussed. If the transmitter remains on-frequency, the difference frequency will be constant at 30 MC. However, if the magnetron frequency drifts, it is necessary to produce a comparable drift in the radar local oscillator frequency to maintain the required difference frequency of 30 MC.

The radar AFC circuit performs this function by amplifying (in V3201 and V3202) the difference frequency obtained in the radar AFC cavity and analyzing the IF output in the tuned circuits of discriminator, V3203, to determine the direction and the amount the IF signal differs from 30 MC.

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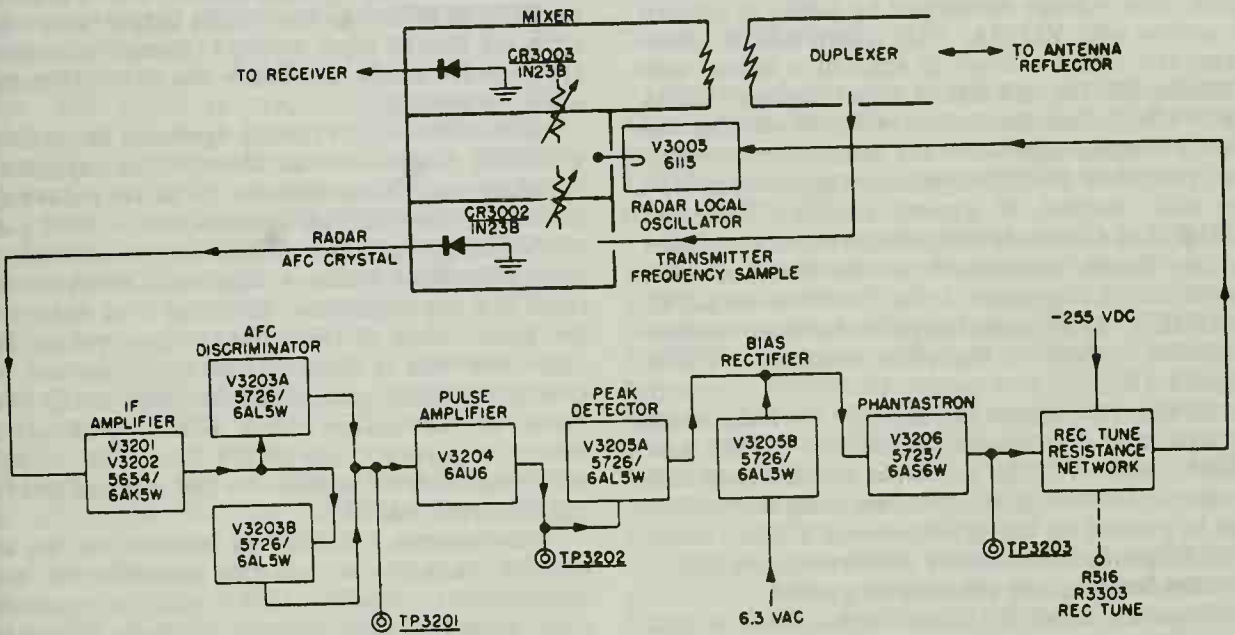
162.41

Figure 12-16.—Radar AFC, block diagram.
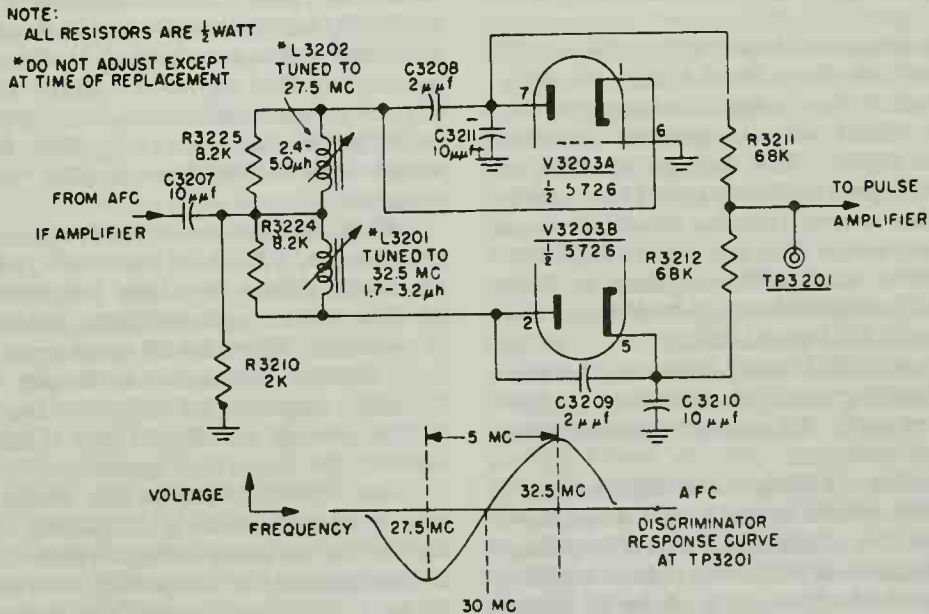
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Figure 12-17.—Radar AFC discriminator, simplified schematic diagram.

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The IF input signal from V3202 is applied to the junction of the two discriminator tuned circuits. The voltage developed by L3202 is applied in series with V3203A, C3211, and R3210. Likewise, the L3201 voltage is applied in series with V3203B, C3210, and R3210. Twin diodes, V3203A and V3203B, are connected in such a way that the AFC IF signal applied to the discriminator circuit produces rectified pulses of opposite polarity with respect to ground across C3211 and C3210. The algebraic sum of the average voltages across these capacitors is the discriminator output which appears at the junction of R3211 and R3212. This output is applied between control grid and cathode of the pulse amplifier, V3204 (fig. 12-16).

If the IF signal is exactly 30 MC, equal voltage will be developed across L3202 and L3201. Tube V3203A conducts on alternate half cycles of the IF signal to produce a negative voltage to ground on the upper side of C3211, while V3203B produces an equal voltage across C3210 with the lower plate positive to ground.

When an input IF signal higher than 30 MC is applied to the discriminator, the L3201 tuned circuit will produce a larger resonant step-up voltage than will L3202. Likewise, a larger voltage positive to ground will be developed at the lower side of C3210 than will negative voltage at the upper side of C3211. The resulting positive voltage is the discriminator output to V3204.

Conversely, when an input signal below 30 MC is received at the L3201-L3202 junction L3202 will produce the larger voltage and the voltage across C3211 will be greater than the voltage across C3210. The voltage applied to the V3204 control grid will be negative to ground.

The response curve for the discriminator circuit is shown below the schematic (fig. 12-17). The negative and positive peaks at 27.5 MC and 32.5 MC respectively, result from the stagger tuning of L3201 and L3202.

Adjustment of L3201 and L3202 may be required after changing a major component in the discriminator circuit. Subsequent adjustments should not be necessary.

Pulse amplifier, V3204, is a conventional pentode amplifier which inverts and amplifies the signal from the discriminator. The pulse appearing at the plate of V3204 varies in amplitude, depending on the frequency of the IF input to the discriminator. As the frequency rises higher than 30 MC, the plate output of V3204 decreases in amplitude. Conversely, as the frequency decreases below 30 MC, the V3204 output increases in amplitude. The waveform at TP3202 indicates the V3204 output which occurs as the IF input sweeps through a range of frequencies above and below the 30 MC IF during a 10 μs period.

Peak detector, V3205A, operates like a diode clamping circuit. Minus 250 volts are applied to the plate and to the cathode. Negative pulses appearing at the V3205A cathode from V3204 permit V3205A to conduct. A capacitor in conjunction with a resistor forms a relatively short charge time for the capacitor, enabling it to charge to the peak value of the V3205A input pulse. The capacitor has a long discharge time and the charge remains essentially constant at the peak value of the pulses from V3204. This action converts pulses from V3204 into a DC control voltage which is applied to the control grid of phantastron, V3206.

Phantastron, V3206, is controlled by the negative charge across the capacitor as discussed. To make the AFC circuit more sensitive, an initial bias of about 2.5 volts is applied between the control grid and cathode of V3206 by the action of the bias rectifier, V3205B. This enables V3206 to follow a lower amplitude negative signal voltage from V3205A.

Phantastron, V3206, generates a sawtooth voltage at its plate which is applied to the repeller of radar local oscillator, V3005. The klystron, V3005, output frequency changes as the repeller voltage changes, increasing in frequency as the repeller voltage is made more negative, and decreasing in frequency as the repeller is made less negative. R3303 sets the range over which the repeller voltage can be changed.

When the intermediate frequency to the discriminator, V3203, is incorrect, the phantastron sawtooth output provides a corrective voltage for the radar local oscillator until the proper IF is reached. When the IF is correct, peak detector, V3205A, produces an output which is sufficiently negative to maintain the phantastron, V3206, beyond cutoff, thereby locking the oscillator to the correct frequency.

The V3005 output to the radar AFC mixer cavity should be at a frequency 30 MC above that of the magnetron input from the waveguide. If the magnetron frequency decreases, the difference frequency output from the AFC cavity will increase.

If the magnetron frequency increases, the difference will decrease.

Local oscillator changes in the opposite direction produce the same effects. The sweeping action (searching) occurs only for large intermediate frequency errors, i.e., when the IF error is outside the 6 MC bandpass of the IF stages. For small errors, V3206 acts as a DC amplifier to produce a corrective voltage for the repeller of the local oscillator, V3005.

Assume an output intermediate frequency from V3202 which is slightly higher than 30 mc. The discriminator, V3203, output to the V3204 control grid will be positive-going. The negative-going plate output of V3204 is fed through V3205A (without inversion), and applied as a negative input between grid and cathode of phantastron, V3206. The V3206 plate voltage increases toward ground potential, resulting in amplification of the DC grid input voltage.

Because the intermediate frequency is too high (above 30 MC) the V3206 positive-going plate output to V3005 reduces the negative voltage at the klystron repeller, and thus produces a decrease in the local oscillator frequency. This action produces a local oscillator injection frequency to the radar AFC mixer cavity which nulls the intermediate frequency error.

If the intermediate frequency from V3203 is slightly lower than 30 MC, no input negative voltage will be applied to the grid of V3206, and the V3206 grid will automatically become less negative. The normal discharge action produces a negative-going plate output voltage from V3206. This negative voltage is the corrective voltage which is of the proper polarity to counteract the low-frequency local oscillator error.

Tube V3206 acts like a DC amplifier (as described above) only when there is sufficient bias applied to the control grid (approximately 6 volts). If the control grid bias should drop below this value, the tube would no longer act as a DC amplifier, but would, in effect, become a sweep generator, performing in much the same manner as a free-running multivibrator.

For proper operation of the AFC circuit, the AFC discriminator must be precisely aligned with the signal IF mid-frequency. If the alignment is inaccurate, the AFC circuit will lock in at a local-oscillator frequency which is higher or lower than the signal IF channel midfrequency, and received signals will be poorly represented on the PPI or completely absent from the presentation at the indicator.

The AFC circuit is particularly sensitive to the magnetron frequency spectrum, and will give erratic results if the magnetron doubles its frequency (double-modes) or jumps in frequency beyond the range of the AFC operation.

Beacon AFC Circuit

The general theory of operation of the beacon AFC circuit (fig. 12-18) is essentially the same as given for radar operation. Hence, the beacon AFC circuit will be discussed only as it differs from radar AFC.

During beacon operation the AN/SPS-10 works in conjunction with other beacons (generally shore-based) or landmarks to obtain navigational information. In beacon operation the AN/SPS-10 magnetron sends out a signal on any frequency within its range and triggers other land-based or shipboard beacon transponders which reply on 5450 MC. The reply is mixed with the output of local oscillator V3006 which always operates at 5420 MC (30 MC below the transponder reply frequency).

The function of the beacon AFC circuit is to maintain the frequency of the beacon local oscillator, V3006, at 30 MC below the received beacon signal. The transmitter of the AN/SPS-10, during beacon operation, may operate on any frequency in its range, while beacon reception is limited to a single frequency (5450 MC). Thus, the beacon local oscillator frequency must be held at 5420 MC (30 MC below the received signal) regardless of the frequency output of the transmitter. This action is accomplished through the use of the beacon reference cavity.

The beacon AFC circuit consists of a beacon AFC crystal, CR3004, three pulse amplifiers, V3401A, V3401B, and V3402A, which amplify the pulse from the beacon crystal, a peak detector, V3403A, a phantastron, V3404, a beacon tune resistance network, and a pulse amplifier, V3402B, which amplifies a positive pulse from pulse transformer, T3002.

A portion of the energy from the beacon local oscillator, V3006, is fed through the beacon crystal current adjust in the beacon reference cavity to beacon AFC crystal, CR3004. The beacon reference cavity appears as a high-Q series tuned circuit, and the crystal current through CR3004 is maximum when the frequency of the beacon local oscillator is the same as the resonant frequency of the beacon reference cavity. Thus, the rectified CR3004 current, plotted versus frequency, has a conventional tuned circuit response characteristic, i.e., maximum at resonance and diminishing on either side of resonance.
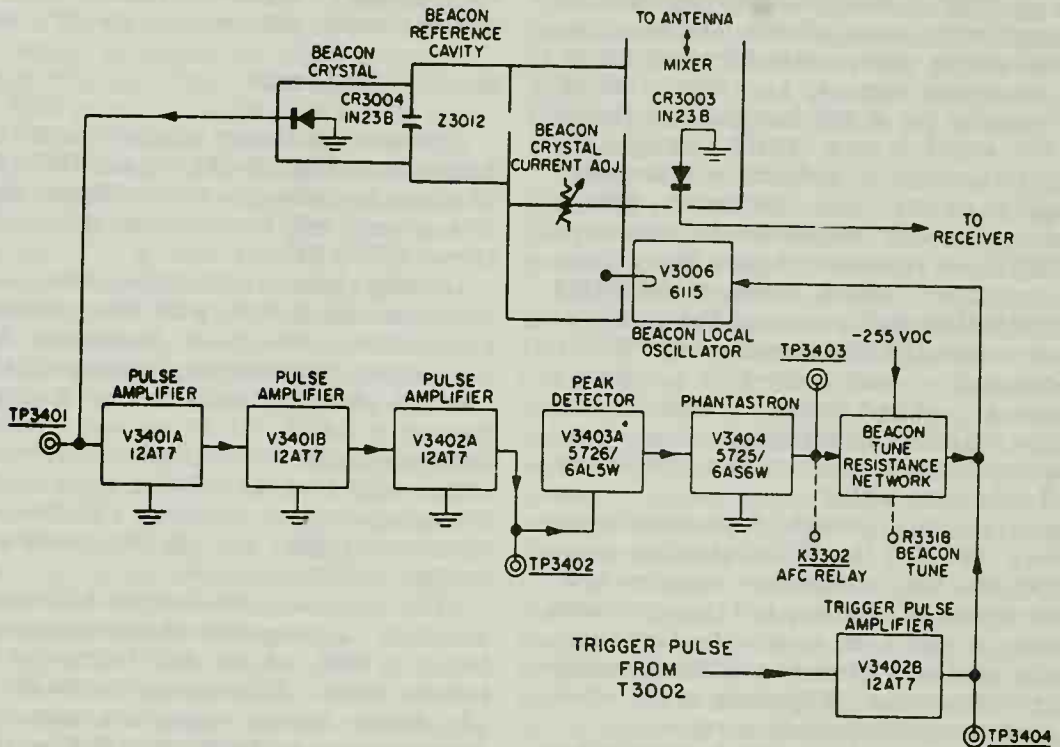
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Figure 12-18.—Beacon AFC circuit, block diagram.

The high-Q characteristics of the beacon reference cavity makes the circuit rigidly tuned to 5420 MC. This rigidity at resonance makes possible the use of the reference cavity as a reference frequency source for the beacon local oscillator. Thus, the primary purpose of the beacon AFC circuit is to maintain the beacon local oscillator at the frequency of the beacon reference cavity, or 5420 MC.

To perform this function, some type of discriminator action must be provided which will provide zero control on V3006 when the oscillator frequency is correct, and a corrective voltage to V3006 when the oscillator frequency is incorrect. A schematic of the beacon AFC circuit is shown in figure 12-19. The method by which discriminator action takes place is as follows:

A positive trigger pulse from T3002 is applied to the grid of pulse amplifier, V3402B. During beacon operation, the recurrence frequency of the pulse is 325 cps. Voltage divider, R3428 and R3429 maintains the cathode potential of V3402B approximately 60 volts above ground.

The V3402B grid is at ground potential, and V3402B is cut off. The arrival of the positive pulse from the 7-8 winding of T3002 on the grid of V3402B causes a saturation plate current through R3430. The negative-going plate output of V3402B is applied through C3410 and R3425 to the junction of limiting diodes CR3401 and CR3402. Because V3402B is driven immediately from cutoff to saturation and back to cutoff, the pulse output of V3402B is clipped and will have a much steeper rise and decay time than that of the waveform at its grid.

Diode CR3402 short-circuits positive overshoot in the V3402B negative output pulse. A negative potential is applied to the anode of CR3401 to prevent this diode from conducting until the cathode of CR3401 assumes a more negative potential than the plate. When this occurs, CR3401 conducts and the voltage from the CR3401-CR3402 junction to ground is limited to the voltage developed across R3424. The arm of R3424 can be adjusted to permit a variable negative pulse of approximately zero to 2 volts to be obtained.
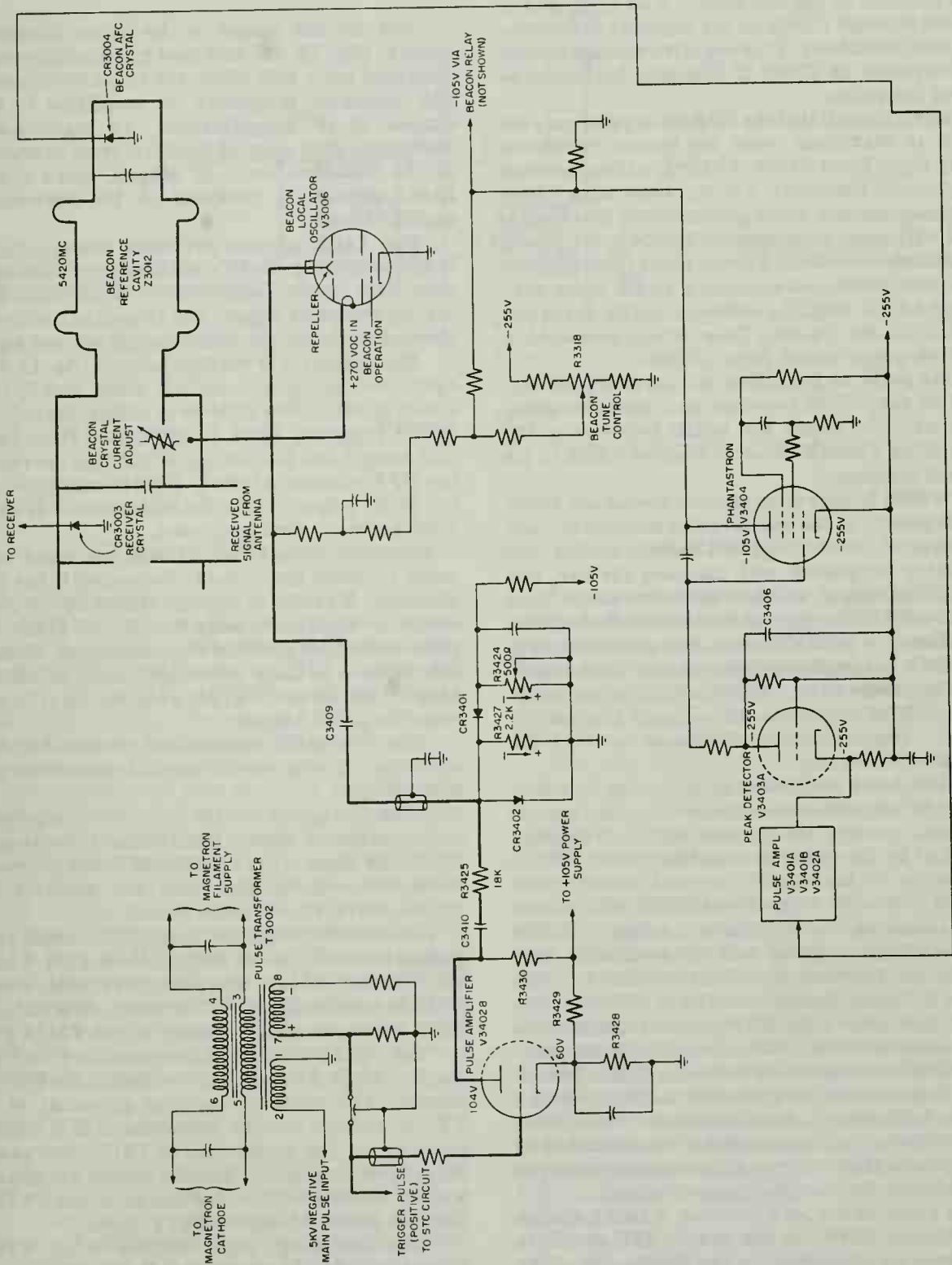
32.195

Figure 12-19. — Beacon AFC circuit, simplified schematic diagram.

The negative output pulse at the CR3401-CR-3402 junction is approximately 2 μs long, and is applied through C3409 to the repeller of the beacon local oscillator. This negative voltage causes the frequency of V3006 to increase for the duration of the pulse.

You will recall that the CR3004 crystal current output is maximum when the beacon reference cavity input from V3006 is exactly at the resonant frequency of the cavity. The increase in the V3006 frequency during the 2 μs period of the repeller pulse will cause the output of CR3004 to drop from its maximum value to a lower value (provided the local oscillator was operating at the same frequency as the beacon reference cavity at the beginning of the pulse). This action produces a negative-going output from CR3004.

This pulse is amplified and inverted and applied to the V3006 repeller as a positive-going pulse which opposes the initial negative-going pulse from V3402B so as to restore V3006 to its correct frequency.

If V3006 is operating at some frequency above the frequency of the beacon reference cavity, and the negative pulse is applied to the repeller, the oscillator frequency will increase further, and the CR3004 output will decrease even more. This negative CR3004 output is fed through three pulse amplifiers, a peak detector, and phantastron to produce a corrective voltage at the V3006 repeller. The phantastron, (V3404) output to the repeller of V3006 will be positive-going and the oscillator frequency will decrease to null the oscillator error.

If the local oscillator is operating at a frequency below the resonant frequency of the beacon reference cavity, the upward shift in frequency produced by the negative repeller pulse at V3006 will cause an increase in crystal current from CR3004. Thus, the pulse from CR3004 will be positive. Likewise, the phantastron output to V3006 will be negative-going and the oscillator frequency will increase to null the oscillator error.

The action of the local oscillator and reference cavity to produce a negative pulse for frequencies above the desired oscillator frequency and a positive pulse for frequencies below the desired oscillator frequency is very similar to the action of a standard frequency discriminator circuit. This discriminator action corrects the frequency of the local oscillator for a difference frequency of 30 mc below the received beacon signal.

The action of the peak detector, V3403A and the phantastron, V3404, in the beacon AFC circuit is the same as described in the discussion of the radar AFC.

Receiver IF Amplifiers

The 30 MC output of the mixer assembly W3011, (fig. 12-14) obtained by heterodyning the received echo with either the radar or beacon local oscillator frequency, is amplified in eight stages of IF amplification. An instantaneous automatic gain control (IAGC) circuit is provided at the fifth and sixth IF stages, and a similar IAGC circuit is provided at the seventh and eighth stages.

The IAGC circuit prevents intense signals from saturating the IF amplifiers by reducing the gain of the controlled stages for the duration of the strong received signal. The IF gain is restored to normal immediately following the intense signal.

The operation of the IAGC circuit (fig. 12-20) is controlled by antijam switch, S504, and by IAGC relay, K303. When S504 is in either the IAGC or BOTH position, K303 is energized from the 24 volt supply and the two IAGC circuits operate. In the OFF position of S504, K303 deenergizes, and the IAGC outputs from the cathodes of V311A and V312A are shorted to ground.

A portion of the IF signal from the output of the sixth IF stage (fig. 12-14), is coupled to the plate detector, V311B. A similar output of the eighth stage is applied to plate detector, V312B. The plate detectors provide negative-going controlling signals to their associated cathode follower stages (V311A and V312A, respectively) to produce the IAGC output.

The two IAGC circuits are identical and the operation of only one of the circuits will be considered.

Bias voltage of V311B (fig. 12-20) places the tube just above cutoff. Positive half cycles of the 30-MC IF input at the V311B grid cause increased conduction through the tube, and negative half cycles drive V311B below cutoff.

The rectified negative-going plate output video pulses from V311B are coupled to the grid of cathode follower, V311A. The pulses are integrated by C336 to remove the 30-MC IF component and produce a more constant voltage on the V311A grid.

The V311A cathode is returned to a -105 volt supply, while its plate is connected to the 270 volt supply. The normal operating potential of the V311A cathode (during quiescence) is 0 volts to ground. As the conduction of V311A increases, its cathode attempts to become positive to ground, while a decrease in conduction produces a V311A cathode potential negative to ground.

Negative-going input signals from V311B during strong IF signals cause the V311A conduction to decrease. This action produces a
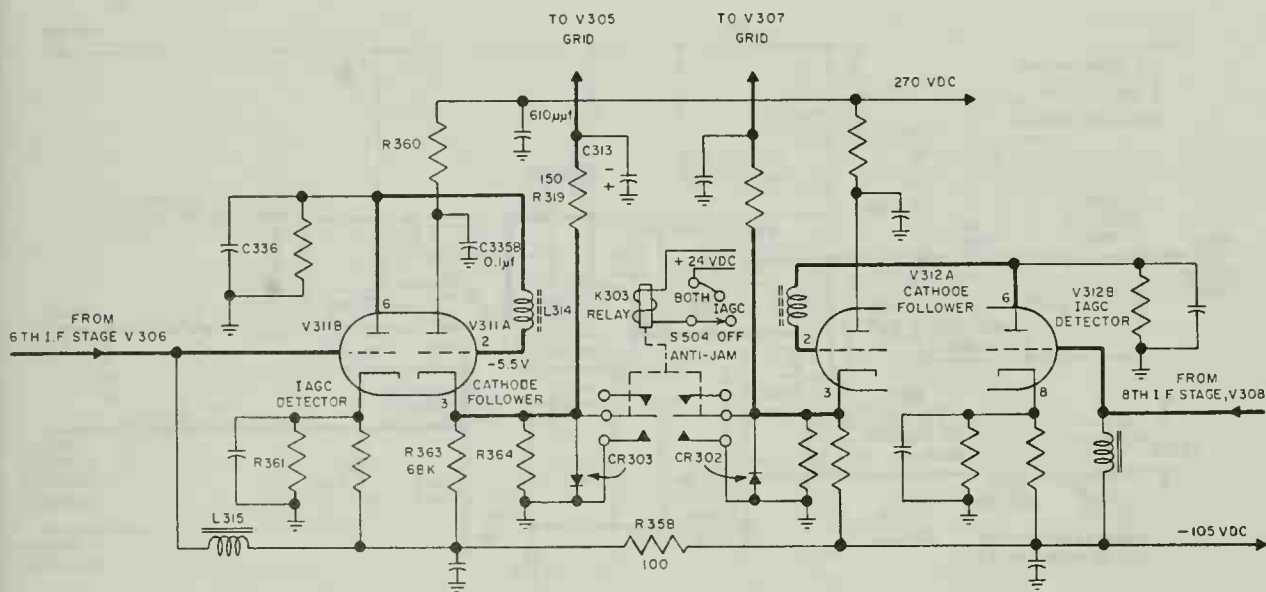
32.196

Figure 12-20.—Instantaneous automatic gain control (IAGC) circuit.

negative voltage to ground at the V311A cathode, and C313 charges to this voltage through the -105 volt supply, R358, R363, and R319. The negative voltage at the top of C313 is applied to the fifth IF amplifier (V305) grid (fig. 12-14). This action reduces the gain of V305 and therefore reduces the input signal amplitude to all of the subsequent IF amplifiers.

The reduction in gain in the IF amplifiers is not instantaneous because the voltage on the grid of C305 can become more negative only as fast as C313 can charge through the path described above. This provides a slight delay in IAGC action. At the end of the video pulse to V311A, C313 can discharge very rapidly through R319, V311A, and C335B, thus permitting an almost immediate return of the control grid bias of V305 to its normal operating level. Rectifier, CR303, prevents the cathode of V311A (and the V305 grid) from going more than a fraction of a volt positive to ground.

Video Circuit

The receiver video circuit (fig. 12-14), consists of an FTC (fast-time constant) circuit, a video detector, a video amplifier, and a cathode follower. The video circuit detects and amplifies the video signal and feeds its output through the indicator adapter to the plan position indicator.

The output of the eight IF amplifier stages is coupled to germanium crystal diode, CR301 (fig. 12-21), which is used as the video detector. The negative video output pulses of CR301 are filtered by C326 and L312 to remove the 30 MC IF signal. The filtered output is fed through the FTC circuit to video amplifier, V309.

The FTC circuit aids in presenting relatively short-duration target signals which occur in the midst of long-duration signals. The FTC circuit is controlled by the anti-jam switch, S504, which is located on the panel of the radar set control.

The FTC circuit changes the time constant of the coupling circuit between the video detector, CR301, and video amplifier, V309. When S504 is in the OFF position (as shown) FTC relay, K301, is in the deenergized position, and its contacts short C327. The coupling circuit consists of C328 and R336, and the input signal is therefore differentiated at the V309 grid across R336. This arrangement represents a relatively long time constant coupling circuit, and, for long-duration signals, the video pulse to V309 may be long enough to blot out neighboring short-duration signals at the video detector output.

When K301 is energized, by placing S504 in either the FTC or BOTH position, its contacts move down (opposite to the position shown). The
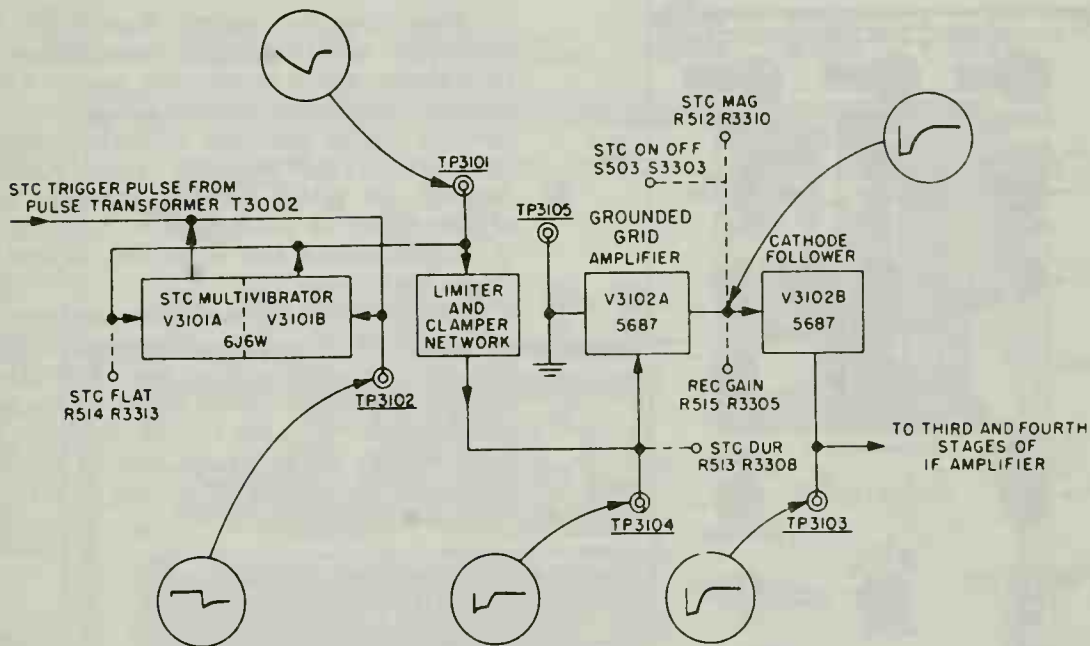
445

Figure 12-21.—Video circuit.

32.198

detected output video signal is now coupled through C327, which, in conjunction with R335 forms a fast-time constant coupling circuit. Differentiation across R335 produces a shorter duration video pulse than would be obtained from C328 and R336 for the same video pulse input, and the duration of the V309 input for all signals is considerably reduced.

The differentiation of the video signal reduces clusters of target signals on the plan position indicator to make individual targets easily distinguishable. FTC is useful in preventing target area blotting of the plan position indicator to achieve a clearer presentation.

The FTC action differs from IAGC in that FTC affects the time duration of the video pulses, whereas IAGC action affects the amplitude of high-intensity video pulses.

Video amplifier, V309, inverts and amplifies the video signal from CR301. Inductor L313 is connected in the plate circuit of V309 to produce the proper high-frequency response in the leading and trailing edges of the V309 video output pulse. The V309 output is fed to parallel-connected cathode followers a V310A and V310B. Parasitic oscillations are suppressed by R345 and R346.

The V310 cathode output is fed to the indicator adapter and to the detector meter circuit. The value of R342 is selected so that the total output impedance is 75 ohms.

Sensitivity-Time Control Circuit

The sensitivity-time control (STC) circuit (fig. 12-22), provides a means of reducing the receiver sensitivity for close-in targets, thus minimizing the sea return echo on the plan position indicators so that nearby targets can be distinguished. The circuit is actuated by a positive trigger pulse from pulse transformer T3002.

The circuit is controlled by STC switch S3303, STC magnitude R3310, STC duration R3308, and STC flat R3313. The output of the STC circuit is coupled to the 3rd and 4th IF amplifiers.

Many of the controls mentioned above are located on the radar set control (fig. 12-23). All of these controls are located on the auxiliary control panel (fig. 12-24). These will be discussed later.

The STC circuit (fig. 12-22) consists of multivibrator, V3101, a limiter and clamper network, a grounded grid amplifier, V3102A, and a cathode follower, V3102B.

Stage, V3101, is a monostable multivibrator. In the absence of an input trigger pulse from T3002, the stable state of the multivibrator is V3101A conducting and V3101B cutoff. When a positive pulse enters the STC circuit from T3002, V3101B is driven into conduction and V3101A is cut off. During the period before

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162.44

Figure 12-22.—STC circuit, block diagram.

V3101A again conducts, a negative-going pulse appears at the plate of V3101B. This pulse is the STC multivibrator output pulse which has a width determined by the position of the STC flat control, R3313.

The negative pulse from the V3101B plate is fed through the limiter clamper network to the cathode of ground grid amplifier V3102A. The limiters prevent any positive overshoot portions of the pulse while the clamper references the positive peak of the pulse.

Ground-grid amplifier, V3102A, conducts during the negative input pulse. The time of conduction of V3102A is determined by STC duration control R3308. By the combined actions of the STC flat and STC duration controls, the STC pulse now has a variable flat section and variable decay.

The grounded-grid amplifier output pulse has the same shape as the cathode input. The gain in V3102A depends on the plate voltage, which can be adjusted by STC magnitude control R3310. The resulting output waveform from V3102 A, adjusted for proper flatness, duration, and magnitude, is applied to the V3102B grid.

Cathode follower, V3102B, couples the negative-going STC output pulse to the 3rd and 4th IF amplifier control grids. As the STC input trigger pulse from T3002 occurs at the instant the transmitter magnetron output is keyed-on, the STC circuit output will provide its negative-going input to the 3rd and 4th IF amplifier during the time strong sea return echoes are received.

The cathode follower is connected between a -105 volt and a +105-volt supply. The DC potential at the cathode of V3102B, normally about -14 volts, can be varied by adjusting the receiver gain control, R3305. The range of this adjustment is from 0 to -15 volts. Because the V3102B cathode potential is applied along a DC path to the 3rd and 4th IF amplifier, it follows that the gain of the IF stages and of the receiver can be varied by positioning the arm of R3305. The STC voltage is superimposed on the V1302B cathode potential, and is independent of the receive gain setting.

The action of the STC circuit varies from that of IAGC. The STC circuit attenuates close-range video echo pulses by reducing the IF gain during the first portion of the receive period. This reduces blurred bright spots on the PPI caused by close-in targets. The IAGC circuit is

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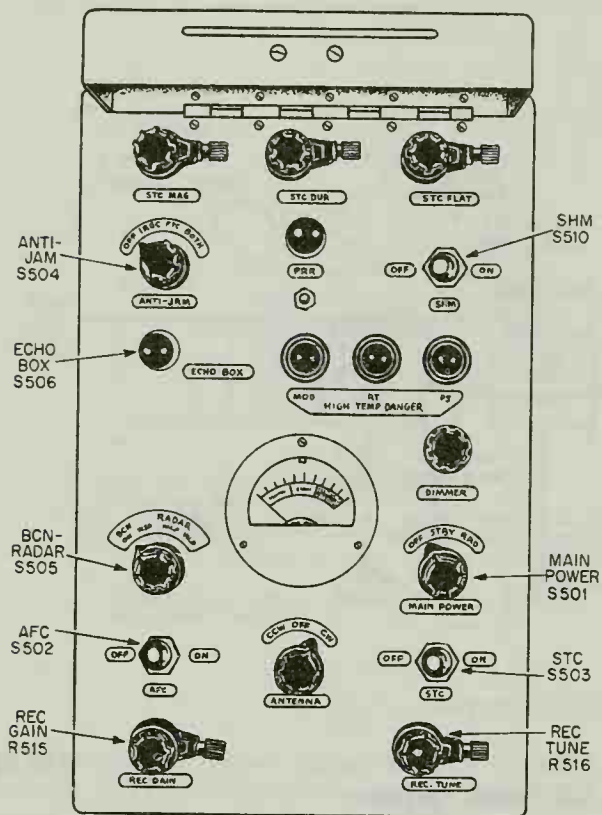
32.193
Figure 12-23.—Radar set control, front panel.

active on strong echo signals regardless of target range.

Indicator Adapter

The primary function of the indicator adapter (fig. 12-14) is to supply trigger pulses and video signals at the proper output levels and at the proper time to the PPI. Provisions are made in this unit for the introduction of range markers and ship's heading marker (SHM) signals.

The indicator adapter consists of a power supply; a trigger pulse delay circuit; a trigger pulse shaper; a video amplifier, an SHM multivibrator circuit, and a range mark input circuit. The trigger pulse delay circuit receives the triggering pulse from the modulator and after a selected time delay, this circuit feeds the trigger pulse to the five plan position indicators. The video amplifier receives the video signal from the receiver and feeds the amplified signal to the five plan position indicators. The SHM and range mark signals are mixed with the video signal in the video amplifier.

Radar Set Control

The radar set control (fig. 12-23), contains operating controls, high-temperature danger indicators, and a monitoring meter. The meter and an associated meter selector switch are used to obtain readings of magnetron current, receiver crystal current, radar AFC crystal current, or beacon crystal current. The scale of the meter is calibrated to indicate current in milliamperes and in POOR, NORMAL, and DANGER areas.

The controls of the radar set control that are duplicated on the auxiliary control panel of the receiver-transmitter are ineffective when LOCAL-REMOTE switch S3301 on the auxiliary control panel is in the LOCAL position. The adjustments on the radar set control are effective when LOCAL-REMOTE switch S3301, is in the REMOTE position.

Auxiliary Control Panel

The auxiliary control panel (fig. 12-24), contains the receiver adjustments and switches necessary for tests, adjustments, and operation of the radar set.

Many of the controls on the auxiliary control panel are duplicated on the radar set control. LOCAL-REMOTE switch S3301 determines which set of controls is activated. With S3301 in the LOCAL position, adjustments must be made at the auxiliary control panel. With S3301 in the REMOTE position, adjustments must be made at the radar set control. The dual arrangement of receiver adjustments can be seen by comparing the controls found in figures 12-23 and 12-24.

HEIGHT FINDING RADAR

The AN/SPS-10 Radar Set has only range and bearing information to consider for presentation of target information on a PPI Height Finding Radar must have a stabilized antenna system and furnish pitch and roll information. Circuits used to properly present this information are similar to those discussed in chapter 11, Electronic Aids to Navigation, Part II.

In Height Finding Radar certain other criteria must be met. This includes the determination of ''slant range'' or height of the target. This is accomplished by converting the slant range of the target into its X and Y components (actual height and range components). The
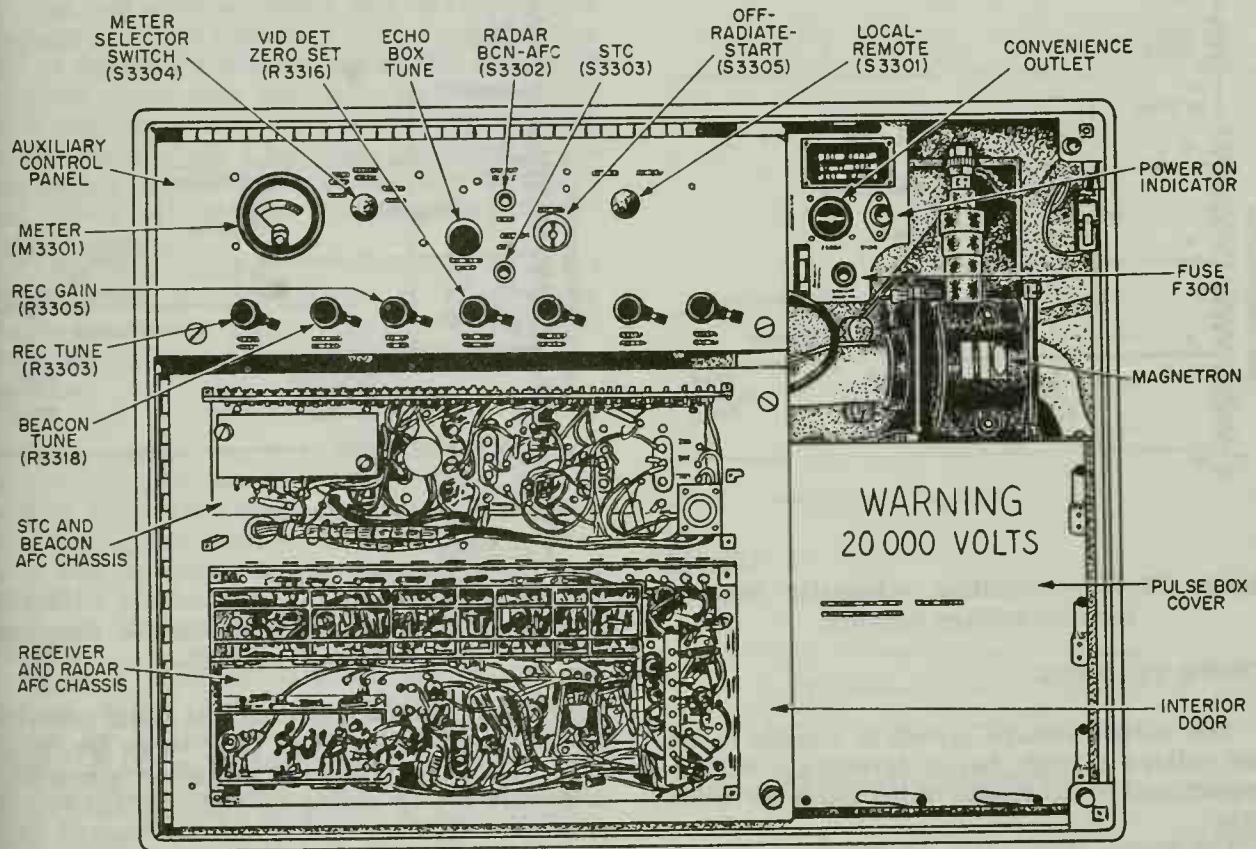
32.189

Figure 12-24.—Receiver-transmitter chassis showing auxiliary control panel.

circuits that perform this function are called sine and cosine circuits.

SINE VOLTAGE

The sine voltage circuit develops a DC voltage proportional to the sine of the beam elevation angle, which is the sum of the reflector elevation angle and the angular position of the antenna scanner. Several methods are available for use in developing the sine voltage. The method that will be described provides maximum accuracy with a minimum of adjustments.

A specially wound precision potentiometer (P1) that includes a fixed 15.474-ohm dropping resistor (R1) and sine commutator are used (fig. 12-25). When the output of a regulated 300-volt supply is connected to the end terminals, 67.6 volts DC will appear across the 4516-ohm section of the potentiometer.

Three moving arms (3, 4, and 5 on the diagram) are used on the potentiometer, and they are mounted on a carriage instead of the usual shaft. These arms pick off voltages that represent the reflector elevation angle. These voltages are applied to the sine commutator, R3, which is a high-speed potentiometer having a resistance taper designed to match the sine potentiometer taper between 10° and 21° rotation. This taper introduces an error in the beam elevation angle voltage for most reflector positions, but the use of arm 4 reduces the error to a negligible amount. Arm 4 also reduces the output impedance of the circuit. The commutator arm, 2, is turned to the scanner angle. Because the angles involved are small, the sum of the sine of two angles is very nearly equal to the sine of the sum of two angles and the arm picks off a voltage that is proportional to the beam elevation angle.

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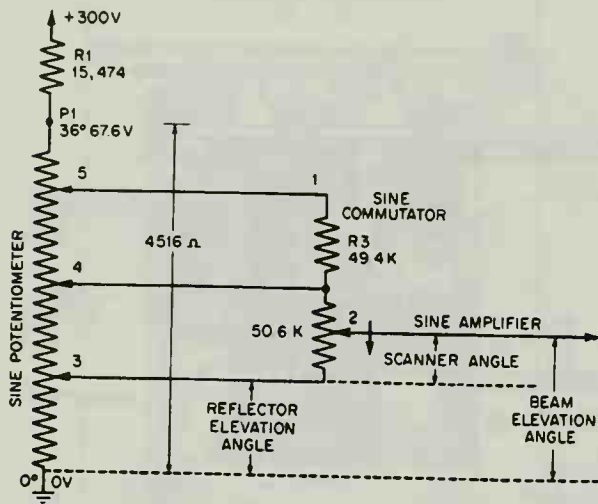
109.102

Figure 12-25.—Simplified schematic diagram for sine voltage network.

COSINE VOLTAGE

The cosine voltage circuit is similar to the sine voltage circuit, but it develops a voltage proportional to the cosine of the beam elevation angle.

The cosine circuit (fig. 12-26) uses a specially wound precision potentiometer that includes a fixed 12,634-ohm dropping section, R1, on the +300-volt (regulated) side and a fixed 6,348-ohm dropping section, R3, on the ground side, of the power supply and a sine commutator, R4. A potential of 115 volts DC corresponds to 0° elevation, and a potential of 93 volts DC corresponds to 36° elevation. The potentiometer arms and cosine commutator are similar in speed, taper, error, and output impedance to those used in the sine voltage network.

The scanner covers a vertical 11° sector. The position of the scanned area depends upon the reflector elevation. Mechanical coupling is used between the scanner and the sine and cosine commutators. The direction of coverage is such that the beam scans the sector from the sky toward the water. Thus, the sine and cosine voltage viewed on an oscilloscope would appear as shown in figure 12-27.

In this system, the reflector elevation can be set anywhere from 4° above the horizon to 29° above the horizon. The 11° sector scan extends from 7° above the reflector elevation to 4° below
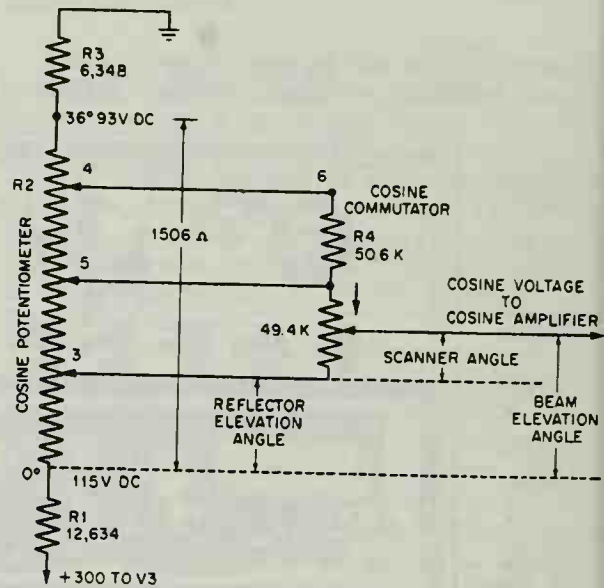
109.103

Figure 12-26.—Simplified schematic diagram cosine voltage network.

the reflector elevation. Thus, the beam could be made to scan as far down as 0° or as far up as 36°. The lower limit of reflector elevation, nominally 4°, is called the reflector tilt (T). It is accurately determined at the factory and stamped inside the elevation data box for each antenna.
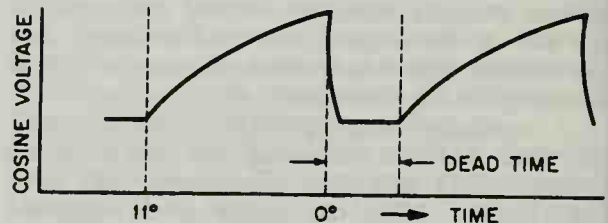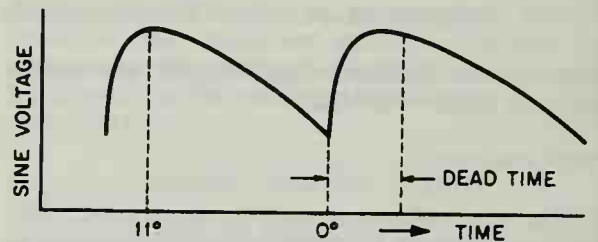
109.104

Figure 12-27.—Sine and cosine waveforms.

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The reflector elevation data is transmitted through an 8-to-1 step-up gear train, so that 25° of reflector elevation causes 200 degrees of rotation of the sine and cosine potentiometers. This step-up ratio means that one-degree error at the start of the gear train produces only a one-eighth degree error at the potentiometers and thus provides more accurate data from the potentiometers.

Additional information concerning synchros is contained in Introduction to Electronics, NavPers 10084, Basic Electricity, NavPers 10086-A, and Ordnance pamphlet-1303.

MTI RADARS

Pulse radars, such as the AN/SPS-10, can measure range but they cannot reject clutter echoes from fixed targets. A CW doppler radar can distinguish between fixed and moving targets and measure velocity, but cannot measure range. A combination of the two can be used to measure the range and velocity of a moving target.

Such a combination radar set may be designed either as a pulse-modulated doppler radar or as a moving target indicator (MTI) radar. If the pulse length is more than 10 percent of the wavelength, the radar set is classified as a CW pulse-modulated Doppler system; if the pulse length is less than 10 percent of the wavelength, the set is classified as an MTI system. The shorter pulse length of MTI radars gives them greater resolution and permits them to handle more targets simultaneously. However, the longer pulse length of pulse-modulated Doppler radar permits more efficient recovery of Doppler frequencies and better rejection of clutter echoes from stationary targets.

Although pulse-modulated Doppler has good rejection of ground clutter, the emphasis in development for naval applications is on MTI. This is due to the better resolution of MTI radar.

Although MTI radars are pulse radars, the principles of MTI and pulse-modulated Doppler radars are so similar that the basic principles of MTI will be discussed at this time.

ECHOES

If the echoes from fixed and moving targets, as illustrated in figure 12-28A, are presented on an A scope, they appear as shown. The echoes from a moving target vary in amplitude and give the signal a fluttering appearance; the echoes

A. FIXED AND MOVING TARGETS
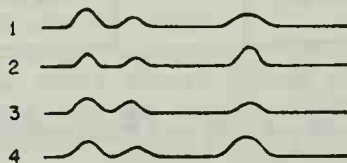
B. INDIVIDUAL TRACES
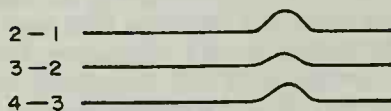
C. CANCELLATION OF ECHOES

162.45

Figure 12-28.—Echoes from fixed and moving targets.

from a stationary target have steady amplitude. The pulse-modulated Doppler system uses filters to reject the pulses of constant amplitude, whereas MTI systems use pulse-to-pulse cancellation of echoes. The four curves in figure 12-28B are the four individual traces shown superimposed in figure 12-28A. If the signals of sweep 1 are delayed for an interval equal to the repetition period and then subtracted from the signals of sweep 1, all signals of constant amplitude are canceled and only variations between successive pulses remain. The three traces in figure 12-28C show the canceled signals.

DELAY NETWORKS

One method of delaying the pulses is to use a supersonic delay line. The subtraction process can be carried out continuously by the arrangement shown in figure 12-29. The video signals from the receiver are sent through two channels,
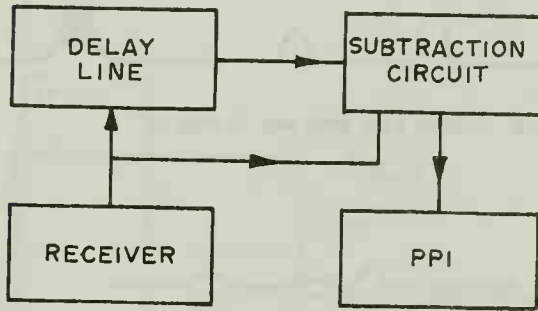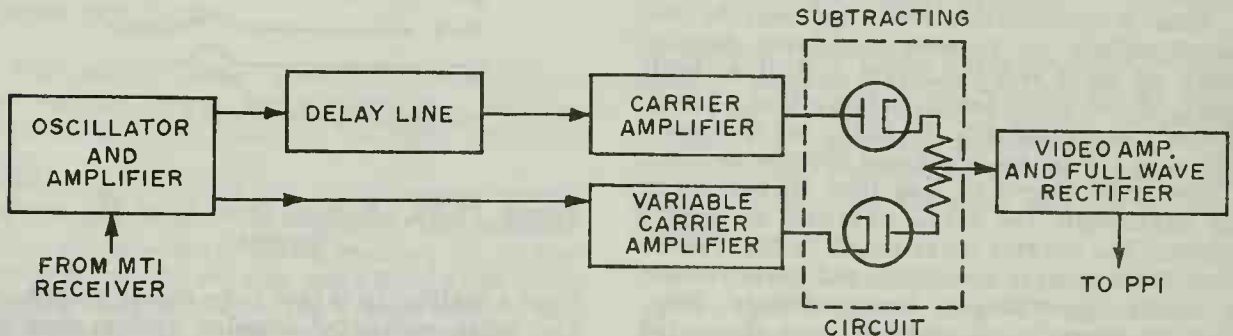
162.46

Figure 12-29.—Delay and subtraction of echoes.

one of which contains the delay line. The undelayed echoes and the delayed echoes are then compared in a subtraction circuit, after which the difference signals are presented on an A or PPI indicator.

One type of supersonic delay line consists of two electro-mechanical (piezoelectric) transducers separated by a medium in which the pulse travels at reduced speed. The video signal is introduced between mercury in the end-cell and that in the delay tube. The voltage across the piezoelectric crystal deforms the crystal. The vibrations are communicated to the mercury in the delay line as a supersonic wave, which travels in the mercury at the velocity of 1,700-feet per second. The length of the delay line and the temperature of the mercury determines the amount of delay. At the output end, the wave causes the crystal to vibrate and produce an alternating voltage. The output voltage is taken from another end cell. The end cells are used as electrodes to introduce and recover the video signals and also to terminate the delay line with the proper impedance.

The response of the crystals does not permit undistorted transmission of the video pulses at

162.47

Figure 12-30.—Compensating for distorted video pulses.
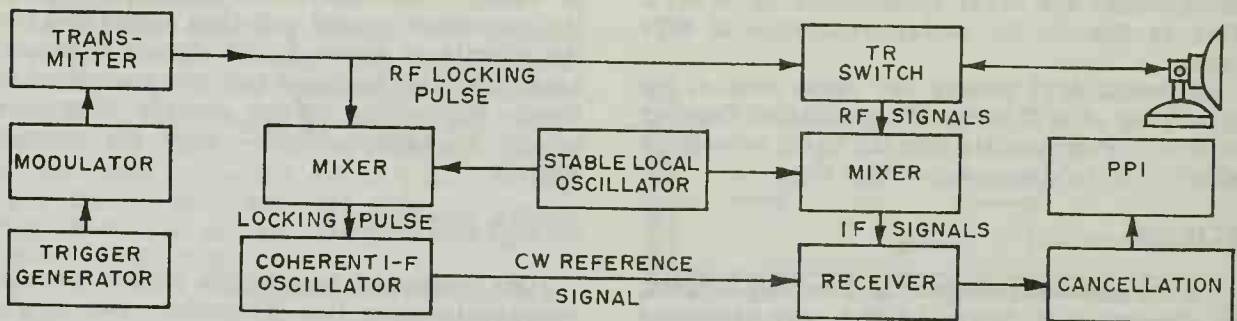
162.48

Figure 12-31.—MTI system.

the input and output. Also the delaying action of the mercury column causes additional distortion. To overcome this distortion, the delayed video pulses are used to modulate a carrier, the frequency of which is the resonant frequency of the crystal.

The video signals from the MTI receiver amplitude modulate a 15-MC oscillator (fig. 12-30). The amplitude-modulated carrier is sent through the delay line and then subtracted from the undelayed carrier in the subtraction circuit. The subtraction circuit consists merely of two diodes arranged to give opposite polarities at the output, so the signals of equal amplitude produce zero output.

In addition to the supersonic delay line, another device for delaying the echo signals in an MTI system is a storage tube, which is similar to the iconoscope used in television. However, mercury or quartz delay lines are used more often in MTI systems.

MTI SYSTEMS

A complete MTI system is shown in figure 12-31. The echoes are translated to zero frequency by heterodyning in the receiver at the intermediate frequency. The CW reference signal that is used in the heterodyning process is obtained from the coherent IF oscillator.

When pulsed, the transmitter starts with random phase from pulse to pulse. The phase of the reference oscillator must be matched, or locked to that of the transmitter at each instant of pulse transmission, since the phase change beating oscillator and returning echo is the basis for the recovery of the Doppler-shift pulses.

An RF locking pulse from the transmitter is used to phase the oscillator. The locking pulse is said to make the oscillator inherent with the transmitter, and the oscillator is called a coherent oscillator or "coho".

Moving-target indication circuits can be added to a radar aboard planes or ships by compensating for the velocity of the station. The compensation can be done by a phase-shift unit. The phase-shift unit changes the phase of the reference signal at the same rate that the phase of echoes from fixed targets is being changed by the motion of the station. Accurate velocity information could be obtained from an Inertial Navigation system, such as discussed in chapter 14 of this training course.

Various MTI systems are under development for naval purposes. The systems differ in the methods employed for interlocking the transmitter and coho at RF and IF frequencies, and in the type and location of the delay and subtraction circuits employed.

CHAPTER 13

RADAR REPEATERS

Modern combatant ships are equipped with two or more radar sets. Data provided by these sets is required at various stations, such as CIC, bridge, flag plot, and pilot house. The problems involved with getting this data to these stations are solved with the installation of radar repeaters.

A radar repeater can be compared to extra loudspeakers on a radio set. With radar, as many repeaters as needed can be connected to one radar set, which then becomes the signal source for all repeaters. A Guided Missile Cruiser for example, may have as many as 30 repeaters.

Each repeater has its own power supply, sweep circuits, and CRT. The repeater receives synchronizing triggers, video, and antenna position data from the master radar through connecting cables. An operator at a remote repeater can switch from one radar set to another, thereby obtaining data from several master locations without leaving his station.

Radar repeaters have several different types of displays, depending upon their intended use. We will list three types of displays available (standard, off-center, and range-height) and then give a brief description of each before discussing a representative repeater.

The STANDARD DISPLAY radar presentation is a conventional plan position indicator (PPI). The sweep path is circular around a point located in the center of the CRT screen. This central point represents the location of the radar antenna. The distance to an object being picked up by the radar may then be determined by using the range rings and/or range strobe.

The OFFCENTER DISPLAY is a modified standard display used to enlarge a sector being covered by the sweep. When a target is barely visible on the face of the CRT at some given range, this enlarging property can be utilized by moving (off-centering) the ship's position

(trace origin) to some other part or off the screen and centering the target indication at a lower sweep range (sweep speed), thereby increasing the relative size of the target. It is possible to use the off-centering cursor to determine the range and bearing between two targets on the display. The operator may discontinue the normal cursor operation and use the offset hand-crank (or joystick in the AN/SPA-8C) to operate the cursor. By doing this, he can place the cursor over the blip from a friendly plane or ship and determine the range and bearing between them and any target.

The RANGE-HEIGHT DISPLAY is, as the name implies, a radar representation, which makes it possible to determine both the range to an aircraft and the height at which the aircraft is flying.

This representation is so arranged that slant range is measured along the horizontal axis, and height along the vertical axis. Targets appear as vertical pips that are bisected with the height line for height measurement.

BLOCK DIAGRAM

A block diagram of a representative radar repeater (AN/SPA-8 series) and the waveforms at various points (fig. 13-1) show how the various sections are related to the operation of a complete unit.

All of the sections represented by individual blocks in the block diagram are not used for certain types of operation. On the other hand, one section may be connected to several other sections at the same time. In our preliminary discussion we will assume that the panel switch that controls the function of the repeater is in the CURSOR ON position because most of the sections, but not all, will be used for that function. Keep in mind, while reading the next several pages, that a detailed discussion of the
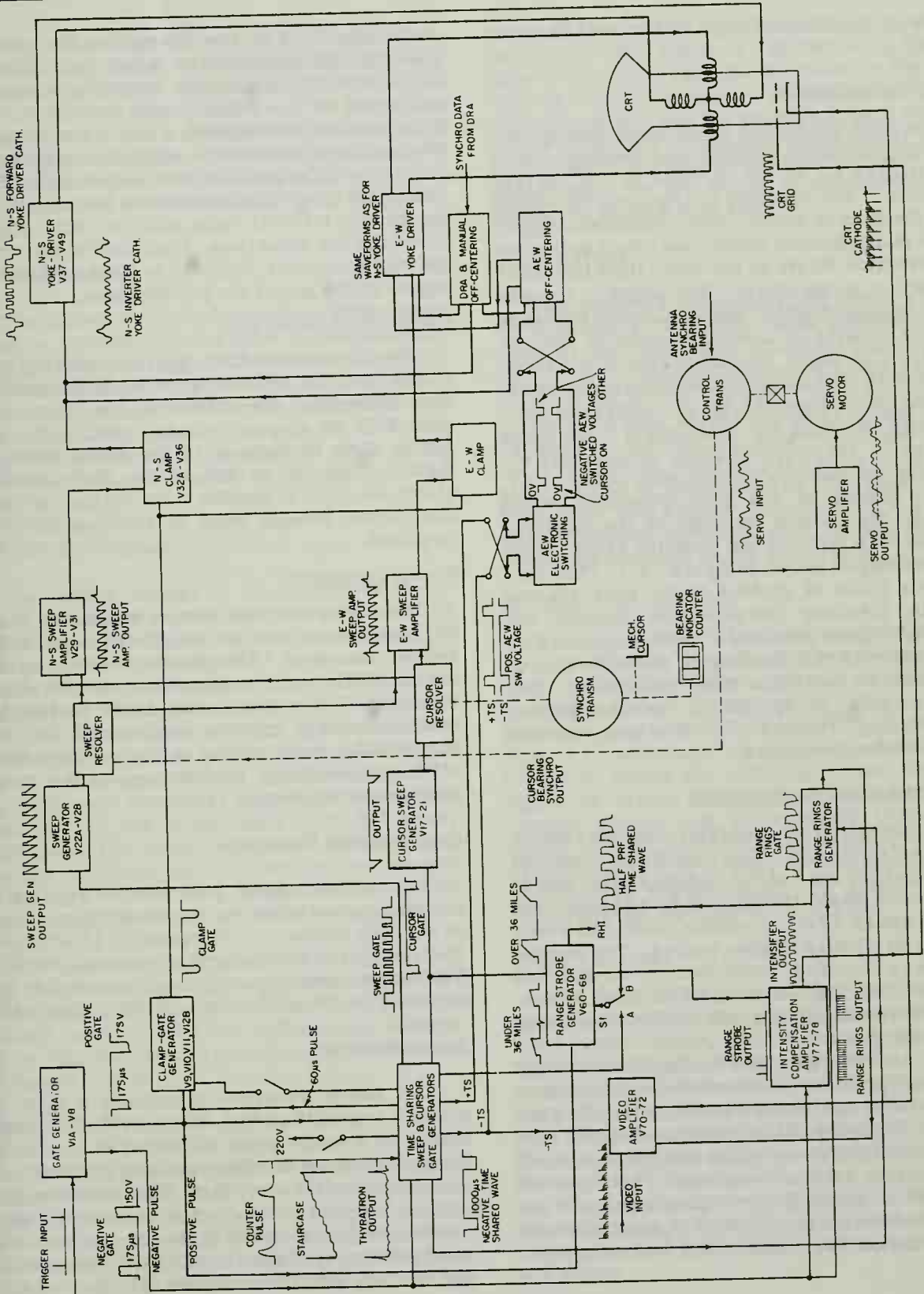
109.47

Figure 13-1.—Block diagram of a representative radar repeater.

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theory of operation of each section will be given later.

GATE GENERATOR

The gate generator is the basic timing circuit and furnishes positive and negative gates that trigger all of the circuits in the correct sequence. The repeater is triggered by pulses from the master radar. The input pulses trigger a gate multivibrator (discussed later) so that the multivibrator starts at the same time the radar transmitter pulse starts. The length of the gate from the multivibrator can be varied by the range selector control in another section of the repeater. The pulse repetition rate of the radar transmitter also controls the length of the multivibrator gate. However, the gate generator automatically shortens the gate length if the range selected by the operator is not satisfactory for use with the pulse repetition rate of the radar.

The output of the gate generator which actuates the display circuits of the indicator contains both positive and negative gates having the waveshapes shown in figure 13-1. The negative gate pulse is applied to the time-sharing circuits, intensity compensation circuits, and the range rings generator. The negative gate pulse controls the starting of a shock-excited oscillator in the range rings generator. The positive gate is applied to the clamp-gate, time-sharing, sweep and cursor gate, and the range-strobe generators.

TIME-SHARING GENERATOR

The time-sharing generator contains a staircase counter circuit. This counter is triggered by the trailing edge of the negative pulse and is used as the basic timing unit to determine the type of sweep (PPI or cursor) that will occur during a particular trigger period. The counter actuates a flip-flop multivibrator circuit that furnishes two time-sharing waves used to control coincidence mixers that produce the sweep and cursor gates.

The positive gate fed to the time-sharing generator provides a second coincidence gate that operates the sweep- and cursor-gate generators. Because of the staircase counter circuit action, nine sweep gates are generated and then a cursor gate is generated. This nine-to-one ratio of sweep gates to cursor gates is not maintained for all functions. It is, however, used in the "Cursor On" function that we are describing.

Switch S1 is in the (B) position as shown, when the function control is set for "Cursor ON" operation, and the time-sharing wave output is fed to the range strobe generator. The time-sharing wave provides one of the two coincidence gates required to enable the range strobe generator. The positive gate provides the other coincident gate. The gate pulses coincide during the cursor interval only, and the range strobe occurs at the same time. The output of the time-sharing generator includes a 60-μs pulse that begins at the end of the positive gate.

Clamp-Gate Generator

The 60-μs pulse from the time-sharing generator and the positive gate from the positive gate generator are mixed in the clamp-gate generator to form a clamp gate. After the clamp gate is formed to the shape shown in figure 13-1, it is fed to the N-S and E-W clamp sections to disable them during the video and cursor sweeps when no tracking strobe is required.

Sweep Generator

The sweep gate that occurs on nine out of ten triggers (when time sharing is on) is fed to the sweep generator. The sweep generator uses a sweep oscillator to produce a negative sawtooth waveshape. Provisions are made so that the sawtooth voltage may be divided and switched at a definite range setting so that a voltage wave of the proper slope is delivered to the sweep resolver on all ranges.

Cursor Sweep Generator

The cursor sweep generator is similar to the sweep generator but its sweep occurs only on the 10th trigger out of a group of 10 triggers. The circuit is idle during the other nine sweeps. The cursor sweep generator is actuated by the cursor gate from the time-sharing generator.

Sweep Resolver

The sweep resolver is actually a form of synchro transmitter that has two rotor windings wound on a single rotor displaced 90 electrical degrees from each other, and one primary and one auxiliary stator winding. The rotor is mechanically coupled to the shaft of the control transformer, which responds to the antenna rotation and position. A single, linearly increasing voltage is fed into the resolver primary stator

winding and produces a current flow in the winding. Because of the current flow in this winding, voltages are induced in the auxiliary stator winding and the two rotor windings. The envelopes of the induced rotor voltages are made to vary as the sine and cosine, respectively, of the shaft angle of the radar antenna. The auxiliary stator winding is used for feedback purposes discussed later.

Cursor Resolver

The cursor resolver is identical to the sweep resolver, except that its rotor is mechanically coupled to the bearing control counter and to the mechanical cursor. By using the mechanical cursor adjustment (handwheel), the operator can place the cursor line anywhere on the CRT display. The electronic cursor will move along with the mechanical cursor and remain parallel to it.

N-S and E-W Sweep Amplifiers

The output voltage of the resolvers is too small for effective use; therefore it must be amplified. Eventually, as we shall show, the cursor voltages in amplified form are applied to the CRT yoke windings. These windings are designated N-S (north-south) and E-W (east-west) to denote their position in the circuitry.

The output voltage from one secondary winding of the sweep resolver and the output voltage from one secondary winding of the cursor resolver are fed to the input of the N-S sweep amplifier. The output voltages of the secondary winding remaining on each resolver are coupled to the input of the E-W sweep amplifier. Both sweep amplifiers are straight voltage amplifiers.

The amplifier input circuitry is arranged so that the sum of the voltages from each resolver is applied to the input of the amplifier. Straight voltage amplification then builds these voltages up to values suitable for clamping and mixing in the yoke drivers.

N-S and E-W Clamps

The clamping circuits are required to prevent shifting of the start time of the sweep voltages as the resolver output varies. To prevent time shifting, the output of the sweep amplifiers is clamped to near zero volts during the off time of the sweep. Feedback type clamp circuits are

used. The period of time during which the clamping action takes place is determined by the off time of the clamp gate. As mentioned in the discussion of the clamp gate generator, a 60-μs pulse is mixed with the positive gate to form a clamp gate. The clamp gate is purposely made 60 μs longer than the sweep gate to insure sufficient time for the sweep to recover control before the clamps take hold.

Yoke Drivers

Two separate amplifiers are used to drive the N-S and E-W yokes. Both amplifiers are supplied with sweep and off-centering voltages, and produce deflection-yoke currents proportional to the sum of the input voltages. The rate at which the yoke driver amplifier output varies the current through the yoke coils determines the sweep range in miles per radius. The yoke driver output is determined by the range selector control (not shown). This control is a four-stack, nonlinear potentiometer that controls the amount of feedback in the amplifiers.

N-S and E-W Limiters

The N-S and E-W limiters are used to prevent overloading the yoke drivers, which overload easily. These stages can only supply enough current to deflect the spot on the CRT one radius from the center before they overload. Therefore, it is necessary to keep the amplifier from trying to follow the incoming signal when the signal is strong enough to exceed the limit. Limiter stages are incorporated in the yoke driver circuit, and these stages cause the yoke driver gain to drop almost to zero when the spot is deflected approximately 10 percent of the radius beyond the edge of the screen.

Video Amplifier

The three voltages fed to the video amplifier are the negative time-sharing pulse, the video input signal voltage, and the signal voltage from the range-rings generator. The video amplifier combines these three voltages into a single voltage called the video output voltage. The amplifier output is coupled to the cathode of the CRT through a cathode follower stage that provides the necessary impedance match.

The negative time-sharing signal blanks out the video signal during the cursor time and produces an output pedestal that intensifies the cursor sweep.

Provisions are made to change the point or points to which the range-rings oscillator voltage is fed. This is done by switching, which provides a means by which range rings centered about the origin of the sweep or range marks on the electronic cursor, or both, are produced.

Intensity Compensation Amplifier

A negative gate from the gate generator and a pulse from the range strobe generator are fed to the intensity compensation amplifier. As the name implies, this amplifier compensates for changes that would cause variations in spot intensity. To do this, a blanking voltage is applied to the CRT grid, except during the interval of the negative gate. The amplifier receives the range strobe pulse and adds it and the cursor origin intensification pulse to the cursor sweep in the CRT display. Compensating circuits in the amplifier help to hold the spot brightness constant for different repetition rates and ranges.

DRA and Manual Off Centering

The DRA and manual off-centering section is used to cancel the motion of the ship so that land masses are stationary on the display. Information from the ship's dead reckoning analyzer (DRA) is required when this type of display is used. The DRA operation uses information from a synchro, but provisions have been made so that the operator can manually control the DRA feature. When manual off-centering is used, the operator can manually control the DRA feature. Also when manual off-centering is used, the operator can center the sweep on some point other than his own ship's position. Distances and angles to a third point can then be determined.

AEW Off-Centering

For airborne early warning applications, equipment in the block labeled, AEW OFF-CENTERING, is substituted for that used for DRA and MANUAL OFF-CENTERING. For plane tracking purposes, the radar operator off centers his scope display so that it is centered on the plane being tracked, and the tracking strobe then follows his own ship. This kind of operation is time-shared off-centering. That is, time-shared off-centering voltages are produced in the AEW electronic switching circuits. These voltages,

by electronic switching, determine the time during which the AEW off-centering potentiometers will be effective.

AEW Electronic Switch

The AEW electronic switch provides a DC off-centering pedestal voltage that is applied to the input of the yoke driver through AEW potentiometers. The potentiometer output is used to off-center either the video or cursor sweeps, depending upon the type of operation being used— that is, tracking, final, or general purpose off-centered cursor operation. In the tracking and general purpose operation of the indicator a roving cursor is provided. The name comes from the fact that the operator can move this cursor about over the display. The DC pedestal voltage is applied only during the interval of the cursor sweep.

When the indicator is being used in tracking operations, the roving cursor may be replaced by a tracking strobe. When this is done, the DC pedestal voltage is applied during the tracking strobe time (cursor interval).

To displace the video sweep for fixed operation, the DC pedestal voltage is applied during the video sweep time.

Range Strobe Generator

The positive gate pulse from the gate generator is applied to the input of the range strobe generator and used as one of two coincidence gates for enabling the range strobe generator. If the function control switch S1 is in position A, the positive time-sharing wave from the time-sharing generator is used as the second coincidence gate.

When this connection is used, a range strobe appears on the cursor.

When S1 is in the position B, a half-frequency gate from the range-rings generator is used for the second coincidence gate, and a range strobe ring appears on the display. The position of the range strobe or the radius of the range strobe ring is determined by a control that is geared to the wipers of the potentiometers in the range strobe generator. The control is also geared to counters that indicate the distance of the range strobe from the origin of the cursor, or the distance of the range strobe ring from the origin of the video sweep.

The indications of distance along the cursor are in the form of bright spots. The increased spot intensity at the origin of the cursor sweep is derived by coupling the cursor gate pulse into the circuits of the range strobe generator.

The range strobe generator, marked RHI (Range Height Indicator) furnishes range information to a remote point in the form of an output of a thyratron tube in the range strobe generator.

The thyratron is triggered by the range strobe at the correct time for RHI indications.

Range-Rings Generator

A negative gate from the gate generator is used to enable the range-rings generator. After being triggered by the negative pulse, the range-rings generator delivers uniformly spaced pulses to the video amplifier. The pulses cause spots of increased intensity to appear on the sweep. As the sweep revolves, these bright spots along with the screen persistence of the CRT, cause rings to appear on the display. The spacing between rings is determined by the tuned circuit in use in the range rings generator. By switching in different tuned circuits, the range ring spacing can be set to 1, 2, 5, 10, 20, or 50 miles. This action enables the operator to determine within close limits the range (distance) to a target.

CIRCUIT ANALYSIS

In order to recognize faulty operation and to determine possible causes of trouble, the ET must know how each section (fig. 13-1) in the complete unit functions. In addition, he should know the waveforms that should be present at various test points. Where possible, we will show the kind of waveform that should be expected at the different test points along with the circuit analysis.

GATE GENERATOR

The gate generator section which is regulated by the radar triggers, gate trigger, and gate multivibrator circuits, is the basic timing unit of the indicator and includes the fundamental timing circuits for the whole set. If operating improperly, it will affect the operation of the entire repeater. Therefore, any defect in the gate generator section may cause distortion of waveforms at other circuits.

Gate Generation and Shaping

A schematic diagram for gate generation and wave shaping is shown in figure 13-2. The gate generator is basically a one-shot multivibrator consisting of V2A and V2B, triggered by the amplifier, V1A. Tube V3 is a cathode follower and is the driver for V2B. Other tubes in the generator circuit provide a means for sharpening the leading edge of the gate, and for obtaining continuous variable control of the gate length.

The gate trigger amplifier, V1A, is normally biased below cutoff. The gate action is triggered by the leading edge of a positive 50-μs pulse that is coupled to the grid of V1A by C1. The plates of V1A and V2A are directly coupled to the grids of V3.

The application of a positive pulse (test point 1 waveform) from the trigger multivibrator to the grid of V1A results in a negative pulse at the plate and at the grids of V3. The negative pulse at the grids of V3 is reproduced across cathode resistor R5 and is coupled to the grid of V2B through the RC network composed of C2 and R6. Normally V2A is cut off and V2B is conducting. When the leading edge of the negative pulse from V2A reaches the grid of V2B (V2B normally conducting) the plate current of V2B is reduced. The rise in plate voltage of V2B causes V2A to conduct. This action causes a further decrease in the voltage at the plate of V2A, and thus drives the grid of V2B below cutoff. All of the action that has been described occurs on the leading edge of the negative pulse and completes the multivibrator switching.

When the switching action has been completed, capacitor C2 discharges, and the voltage at the grid of V2B rises exponentially toward the voltage established by the voltage divider R7, R8, and R9. The time constant is determined primarily by C2 and R6. When the grid of V2B rises above cutoff, the tube conducts and its plate voltage drops. This action drives the grid of V2A toward cutoff. The voltage rise at the plate of V2A causes an increase in the voltage drop across R5 which causes the grid potential of V2B to rise sharply to a higher level. The plate voltage of V2B then drops sharply and cuts off V2A. The rapid charging of C2 through the grid circuit of V2B and the cathode of V3 completes the multivibrator cycle, and the circuit is then ready for the next trigger.

The output at the plate of V2B is a positive-going, square-wave gate. Its leading edge starts with the trigger pulse, and its trailing edge occurs when V2B again conducts. The length of the gate is determined by the length of time that V2B is cut off. The cut-off time of V2B is determined by the magnitude of the initial negative
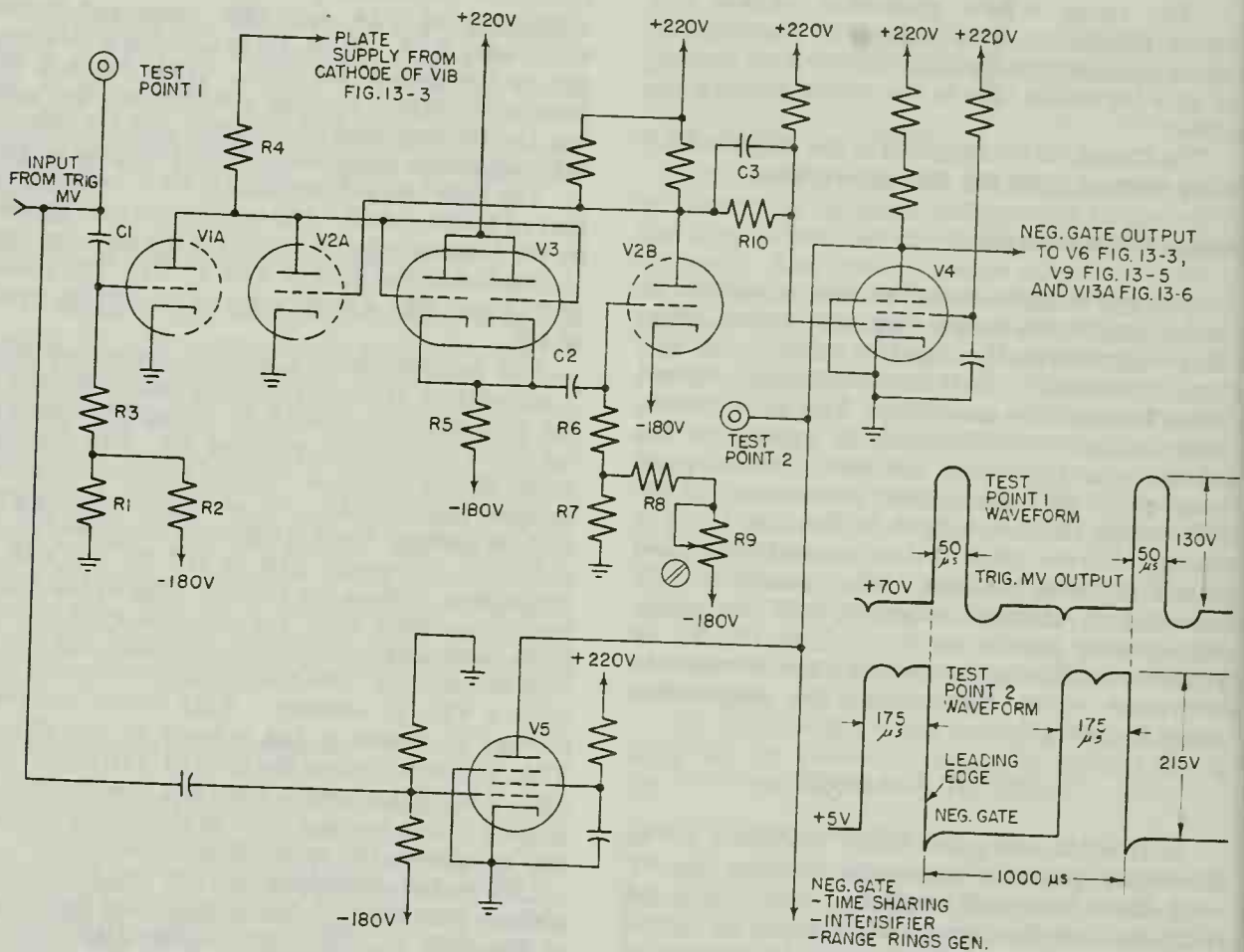
Figure 13-2.—Gate generation and shaping.

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swing at the plates of V1A and V2A. The more negative the grid of V2B is driven, the longer it takes C2 to discharge enough to raise the grid above cutoff level.

Another factor in the cutoff time of V2B is the time constant of the RC network composed of C2 and R6. The longer the time constant, the longer the gate.

The voltage supplied by the voltage divider formed by R7, R8, and R9 also affects the length of the gate. The gate length may be adjusted by setting potentiometer R9. The greater the voltage at the junction of R7 and R8, the faster the grid will rise toward the cutoff level and the shorter will be the gate.

The maximum gate time is preset for a gate length of approximately 3660 μs. This is equivalent to 300 nautical miles, based on one nautical mile for 12.2 μs.

The gate waveform from the plate of V2B is coupled to the control grid of the negative-gate generator, V4, and appears in an inverted form at its plate, as shown in the waveform for test point 2.

The output from the plate of V4 is the negative gate, which provides synchronous enabling of the following circuits in the indicator.

The gate shaper tube, V5, is used to speed up the leading edge of the negative gate. The positive 50-μs trigger pulse input to the gate generator is coupled to the normally cutoff control grid of V5. Because the plate load resistor for V5 is also the plate load resistor for the negative-gate generator, V4, the sharp

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drop in plate voltage is also present at the plate of V4.

The leading edge of the negative shaper pulse occurs at the start of the trigger input pulse. It is slightly ahead of the start of the gate waveform at the grid of V4, and as a result, the negative shaper pulse speeds the start of the leading edge of the negative gate.

Although the pulse at the grid of V5 is of short duration, the wave form at its plate has the full preset gate length duration because the plate voltage of V5 is held down by the output of V4.

A special circuit automatically shortens the length of the gate, when the equipment is connected to a radar having a repetition rate so high that succeeding trigger pulses would ordinarily fall within the gate period.

The waveforms of figure 13-2 indicate that a high repetition rate could cause alternate multivibrator triggers to occur during the negative gate interval and be counted down. The length of the negative gate is automatically shortened when the repetition rate of the radar is so high that there would be less than 175 μs between the end of one negative gate and the start of the next.

Gate Shortening and Length Control

The schematic for an automatic gate shortening and gate length control circuit is shown in figure 13-3. The automatic gate shortening circuit includes tubes V6, V7, and V1B.

The negative gate from V4 is coupled to the grid of V6, and produces a positive gate as shown in figure 13-4 A, at the plate of V6. The positive gate is applied to the grid of V7 through the integrating network, R16, R17, and C6, which partially integrates the pulse so that the rise time of the leading edge and the fall time of the trailing edge are increased. As a result the waveshape is changed to the waveshape shown in figure 13-4, B.

The trigger pulse and the leading edge of each positive gate are mixed at the grid of V7. This means that waveforms A and B are combined with the trigger pulses, and the resultant waveform is shown in figure 13-4, C.

When the PRF is low, the integrated gate voltage has time to return to the base line between gates. The DC voltage at the grid of V7 is adjusted by the recovery time potentiometer, R18, so that at PRFs, which allow about 175 μs or more recovery time, the top of the trigger pulse will be below the cutoff level of V7. Therefore, V7 never conducts when the PRF is low.

At high PRFs when the recovery time is less than 175 μs, the integrated trailing edge of one gate does not have time to return to the base line before the next gate starts. As a result, the trigger pulse, which is added at the start of the next gate, starts at a higher level and reaches a value that is high enough to make V7 conduct. The result is a waveform like the one in figure 13-4, D. Therefore, the voltage output of V7 is a series of pulses which are applied to the grid of V1B. These pulses are reproduced on the cathode of V1B, which supplies the plate voltage for V1A and V2A in the multivibrator.

The amount of plate voltage supplied by V1B controls the magnitude of the initial negative voltage change at the plate of V2A, which in turn controls the operation (cuts off) of V2B. The lower the plate voltage, the smaller the negative change at the plate, and the shorter the cutoff time for V2B. Therefore, the lower the supply voltage, the shorter the gate.

Shortening the gate allows more recovery time, which permits the integrated trailing edge of the gate to reach a lower level so that V7 will conduct less and produce the waveform as shown in figure 13-4, D. The shortening process continues until a balance is reached, and the conduction time of V7 (fig. 13-3) is just sufficient to give the amount of gate shortening desired. The exact condition for this balance is determined by the setting of R18. The final balance condition is effected by the position of S1. If S1 is closed the DC level at the cathode of V7 is raised. However, the operation of the automatic gate shortening circuit remains the same, except that the minimum recovery time has changed to 125 μs.

The gate length control circuit is designed so that there are at least six cathode-ray tube radii or 125 miles of sweep, whichever is shorter, on range selector settings below 34 miles. On settings above 34 miles, there are six radii or up to 300 miles of sweep available, whichever is less. When the range selector is set at 300 miles, the wiper of R20 is near the positive end. In this condition, diodes V8A and V8B cannot conduct because the cathode of V8B is positive with respect to its plate, and the diode acts as an open circuit. If the trigger input PRF is low and there is no automatic gate shortening from V7, the voltage at the grid of V1B is constant at approximately 150 volts, and the gate length is fixed at 300 miles.

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Figure 13-3.—Automatic gate shortening and gate length control.

When the wiper of R20 is turned toward the negative end of the resistance strip, a lower potential is tapped off the divider. When the potential reaches approximately 150 volts (at about the 50-mile range), the cathode of V8B becomes negative with respect to its plate and the diode conducts, which effectively ties the grid of V1B to the wiper of R20. After V8B conducts, the grid of V1B is no longer held at a constant level, and the grid voltage decreases smoothly as the range selector is turned toward a lower range. Accordingly, the voltage at the grid of V1B decreases and shortens the gate length smoothly.

Starting at the time V8B conducts, the gate length becomes equal to six times the range selector setting. The gate length continues to shorten smoothly until a switchover point from slow to fast range is reached at the 34-mi

setting of the range selector. Just before the switchover, the voltage at the grid of V1B is still sufficiently high to provide a gate length of six times the range selector setting, or about 204 miles. As the range selector dial is turned through the switchover point, a cam geared to the shaft of the range selector releases a sensitive switch that energizes relay K1. When the relay pulls in, the cathode circuit of V8A is completed to ground, and the diode, V8A, conducts. This lowers the voltage at the grid of V1B to a value that decreases the gate length from 204 miles to 125 miles. At the time the grid voltage of V1B decreases, V8B will stop conducting. The voltage from the wiper arm of R20 will then no longer affect the gate length.

The reason for switching in the 125-mile limit voltage as the range selector dial is moved through 34 miles is that, on the fast (short)

A. POSITIVE GATE AT PLATE OF V6 WITH LOW PRF

B. INTEGRATED GATE AT GRID OF V7 WITH LOW PRF

CUTOFF LEVEL AT GRID OF V7
50 μs TRIGGER PULSE

C. INTEGRATED GATE MIXED WITH TRIGGER PULSE
AT GRID WITH LOW PRF

CUTOFF LEVEL AT GRID OF V7

SWITCH POINT

V7
CONDUCTS   RECOVERY TIME   RECOVERY TIME EQUAL TO
            LESS THAN   ABOUT 175 μs (OR 125 μs)
            175 μs (OR
            125 μs)   ◄— ADJUSTED SHORTER GATE —►

D. EFFECT OF SWITCHING TO HIGH PRF

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Figure 13-4.—Automatic gate shortening wave-
forms.

ranges, the resolver drivers can tolerate no
more voltage than that which will provide 125
miles of sweep. This is because the sweep slope
for the fast range sweep is four times as steep
as for the slow (long) range sweep. Resistors
R21 and R22 are the maximum and minimum
gate length voltage adjustments.

Positive Gate and Clamp Gate Generators

Figure 13-5 shows the schematic for a
positive-gate and clamp-disabling gate genera-
tor. The positive-gate generator is V9, and the
input to this stage is the negative gate from V4.
The normal amplifying action of V9 reverses
the polarity of the negative input gate and pro-
vides a positive gate at the plate. This positive
gate is used to drive the number 2 grid of V10A,

which is a clamp gate generator. The signal is
also used to drive other circuits described later.

The clamp gate is used to disable the sweep
clamp circuits for a period of time longer (by
60 μs) than the sweep waveform. This delay
gives the sweeps enough time to recover before
the clamps operate to hold the sweep baseline
near zero volts between sweeps.

The clamp gate generator, V10, receives
signals from two different sources and operates
as a mixer. The positive gate is applied to the
number 2 grid. The number 7 grid receives a
60-μs, positive, square-wave pulse from the
time-sharing counter (fig. 13-6). This 60-μs
pulse arrives at the end of the positive gate
pulse and lengthens it so that the output at the
plates of V10 is a negative gate 60-μs longer
than the positive gate.

Tube V11 is an amplifier-inverter and ca-
thode follower. The output of the cathode fol-
lower is a positive clamp gate, which starts
at the same time as the normal gate but lasts
60 μs longer, and is applied to the sweep-clamp
circuits. The clamp gate is used to disable the
clamps, and clamping occurs only between clamp
gates. When the function selector switch is in
the AEW TRACK STROBE position, however, the
clamp gate is disabled via V12B during the
tracking strobe time, which is the normal time
occupied by the cursor sweep.

Time Sharing

Although the video and cursor sweeps ap-
pear to be present on the CRT screen at the
same time, they are actually present on a nine-
to-one shared basis. That is, a sequence of
nine video sweeps occur and then the 10th
video sweep is disabled, and a cursor sweep
occurs.

To obtain the nine-out-of-ten (video) and
one-out-of-ten (cursor) sweeps, the negative
gate from the gate generator is differentiated
and fed to a one-shot, cathode-coupled multi-
vibrator. The pulse output of one-half of the
multivibrator is applied to a staircase counter
circuit, which delivers one output pulse for
every 10 input pulses. It is the counted-down
output of the staircase counter that is used to
produce simultaneous disabling of the video
sweep gate generator and enabling of the cursor
sweep gate generator. The pulse output of the
second half of the multivibrator is used to re-
set a time-sharing, flip-flop multivibrator cir-
cuit after it has been flipped over by the output
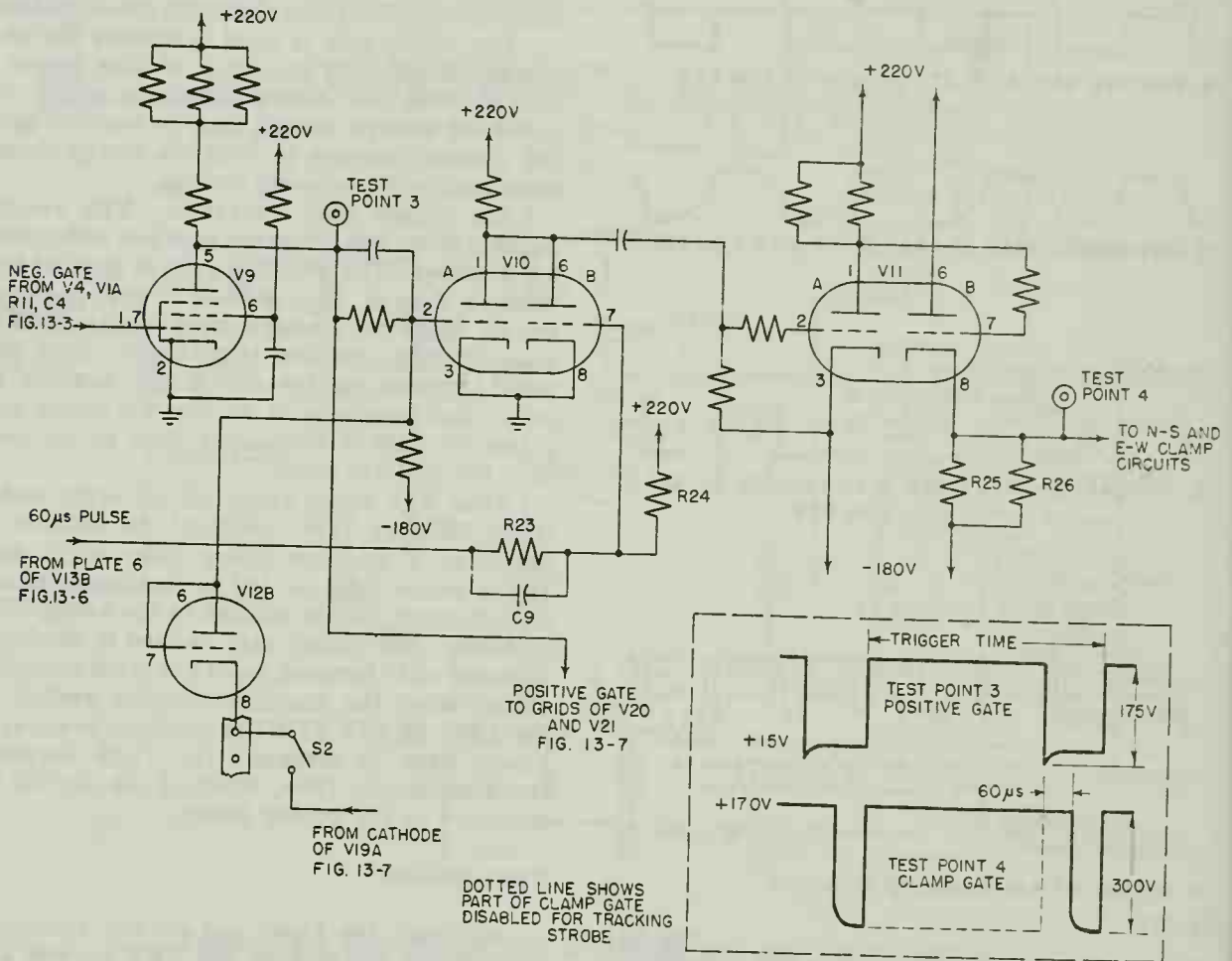
Figure 13-5.—Positive gate and clamp gate generator.

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of the staircase counter section. (The second output of the multivibrator is also coupled to the clamp gate generator.)

The outputs of the flip flop are fed to cathode followers, and the outputs of these stages are two rectangular waveforms of opposite polarity, which form the positive and the negative time-sharing waves. The positive time-shared wave is maintained at approximately +45 volts during the nine video sweep intervals, and swings more positive to +165 volts during the cursor interval. This wave is used to enable the cursor-gate generator. The negative time-sharing wave is maintained at approximately +160 volts during the nine video sweeps, and swings in a negative direction to +30 volts during the cursor inter-val. This waveform is used to enable the video sweep gate generator.

Figure 13-6 shows the schematic diagram and wave shapes for the time-sharing counter. The negative gate from the plate of V4 is coupled to the number 2 grid of V13 through a differentiating circuit C10 and R27. The leading edge of the gate wave-form produces a negative pulse, and the trailing edge, a positive pulse. Tube V13 is a dual triode, one-shot, cathode-coupled multivibrator, with grid 2 held below cutoff. When V13A conducts, multivibrator action is initiated. This action cuts off grid 7, which reduces the bias on grid 2, driving grid 7 well past cutoff. Section V13B will re-main cut off until C11 discharges enough through
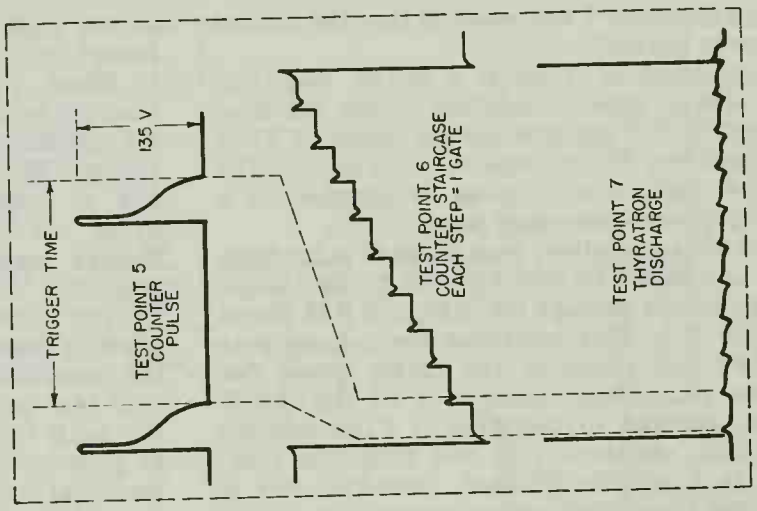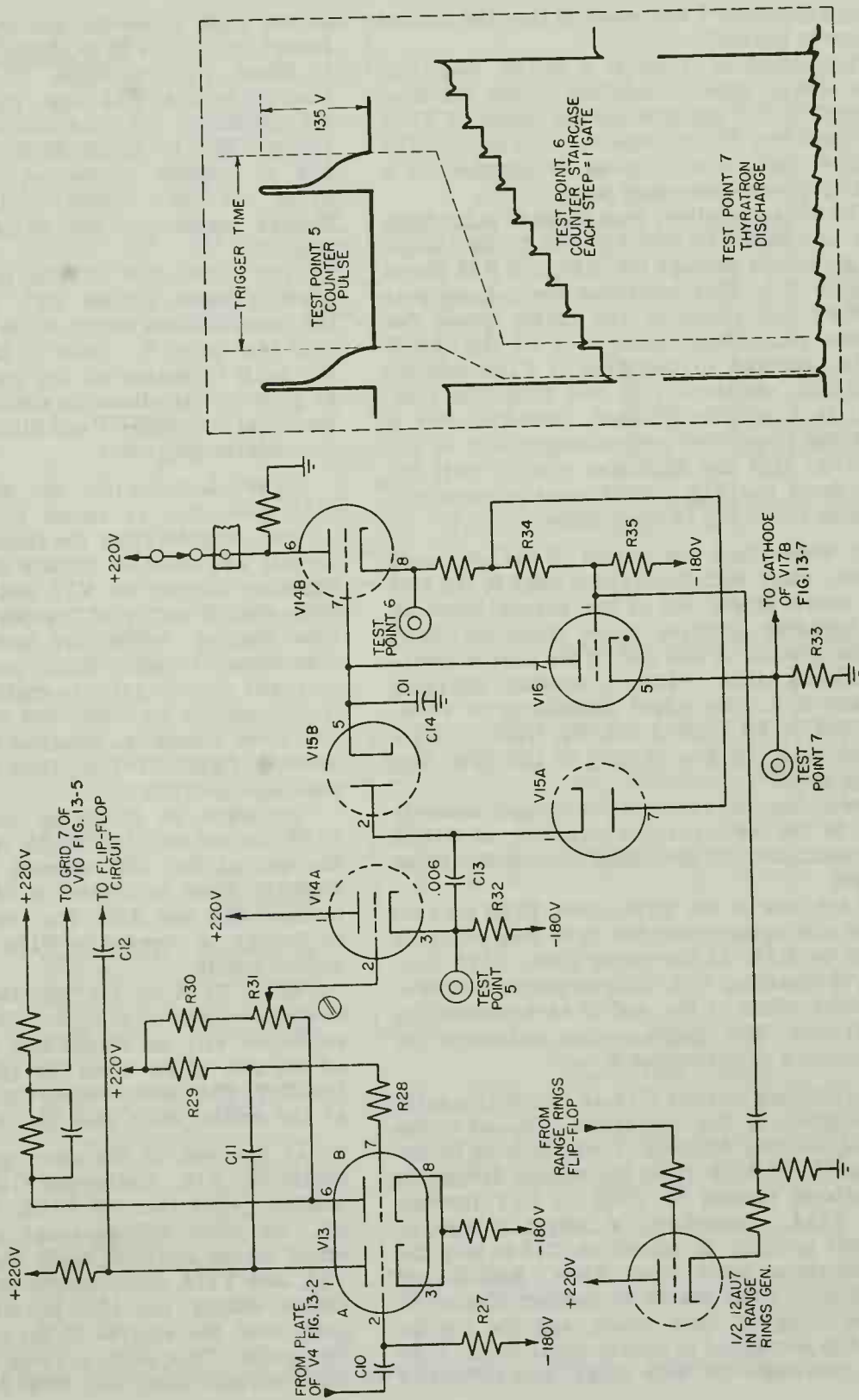
Figure 13-6.—Time-sharing (staircase) counter.

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R29 to raise grid 7 above cutoff then the circuit returns to normal.

The output of V13A is a 60-μs, negative, square-wave pulse controlled by the RC time constant of C11 and R29 and the output of V13B is a positive, 60-μs, square-wave pulse. The negative square-wave pulse is coupled to a flip-flop circuit described later.

The 60-μs positive, square-wave pulse from V13B is coupled to grid 7 of V10 in the clamp-gate generator through C9, R23, and R24 shown in figure 9-5. This lengthens the normal gate by 60μs and produces the clamp gate. The positive pulse from plate 6 of V3 (fig. 13-6) is also coupled to the grid of V14A through plate load resistors R30 and R31. The tube, V14A, is a cathode follower. Potentiometer R 31 is the countdown potentiometer and is adjusted so that the staircase counter will deliver, from the V16 cathode, one positive output pulse for every 10 input pulses.

To understand the action of the staircase counter, start with the trailing edge of the 10th gate, which is the end of the cursor interval. The adjusted positive pulse from the V14A cathode occurs at the end of the cursor interval causing diode V15B to conduct, charging C13 and C14. The output impedance of V14A, C13, and V15B form a low RC time constant so that C14 charges rapidly to the first-step voltage level.

The voltage increase on C14 is approximately equal to the ratio of the capacity of C13 to C14, multiplied by the adjusted positive pulse voltage.

At the end of the first pulse, V15B cuts off and the action maintains the first-step charging voltage on C14. At the same time, V15A conducts, discharging C13. This sequence of operation takes place at the end of each succeeding normal gate. The counter-pulse waveform can be examined at test point 5.

The voltage across C14 is coupled directly to the plate of the thyratron V16 and to the grid of cathode follower V14B. A part of the voltage is fed back from the voltage divider in the cathode circuit of V14B to C13 through diode V15A. Therefore, a slight charge of reversed polarity is placed on C13 to help the next charging pulse from V14A. Because of this action, V15B starts to conduct almost at the beginning of each pulse, and the charge on C14 is increased in nearly equal steps. This action continues for nine steps, and at the end

of the ninth pulse the charge on C14 is great enough to cause V16 to ionize.

When V16 conducts, it acts as a short circuit across C14 and rapidly discharges the capacitor. As C14 discharges, the plate voltage of V16 drops to a value that is too low to sustain ionization, and the counting cycle can then repeat itself. The staircase voltage waveform can be examined at test point 6.

The conduction of V16 produces a sharp positive pulse across R33, which constitutes the counted-down output of the staircase counter (see test point 7.) Thus we have one pulse out for each 10 pulses in. The output pulse is used to produce simultaneous enabling of the cursor sweep-gate generator and disabling of the video sweep-gate generator.

The schematic for the sweep and cursor gate generator is shown in figure 13-7, A. Two outputs from the time-sharing counter circuit are used to actuate the time-sharing, flip-flop circuit of V17 which provides two time-shared waves of opposite polarity. These time-sharing waves are used to enable the coincidence mixers, which generate the sweep gate and cursor gate; to enable the AEW electronic switch circuits; and to control cursor and video intensity. The time relationships are shown in figure 13-7, B. Tube V17 is a bi-stable flip-flop multivibrator.

To trace an operating cycle, assume that V17B is cut off and V17A is conducting. At the end of the 10th normal gate, the 60-μs negative pulse is applied to the plate of V17B through C12 and R36. The corresponding drop in voltage is coupled to V17A driving the grid beyond cutoff.

When V17A is cut off, the rise in plate voltage causes V17B to conduct. This new condition will be maintained during 9 out of 10 normal gates when the video sweep is on the CRT. The pulse coming from V13 (fig. 13-6) at the end of each gate has no further effect.

At the end of the ninth gate, the counter thyratron, V16, discharges C14, and the sharp positive pulse cuts off V17B. When V17B cuts off, its plate voltage rises and cancels the effect of the negative pulse that arrives from V13, and V17A again conducts. This condition exists during the 10th normal gate (cursor gate) until the arrival of the next 60-μs negative pulse. This pulse arrives at the end of the 10th normal gate and cuts off V17A, which

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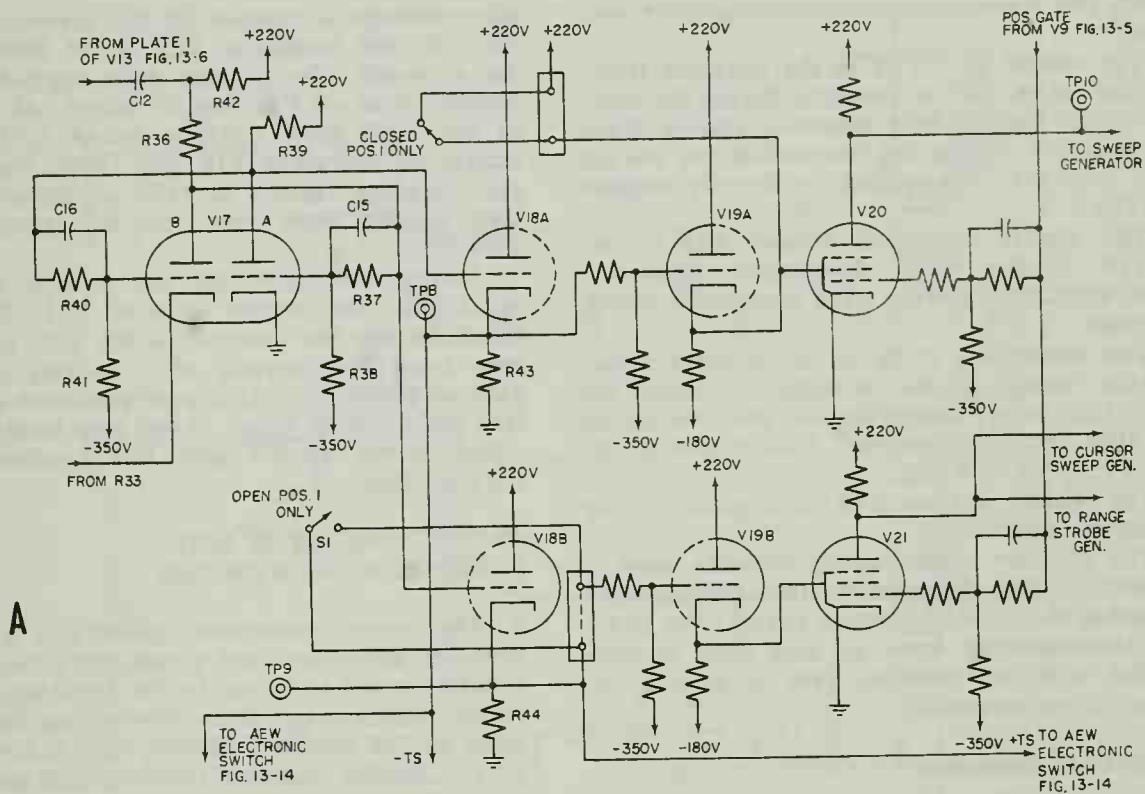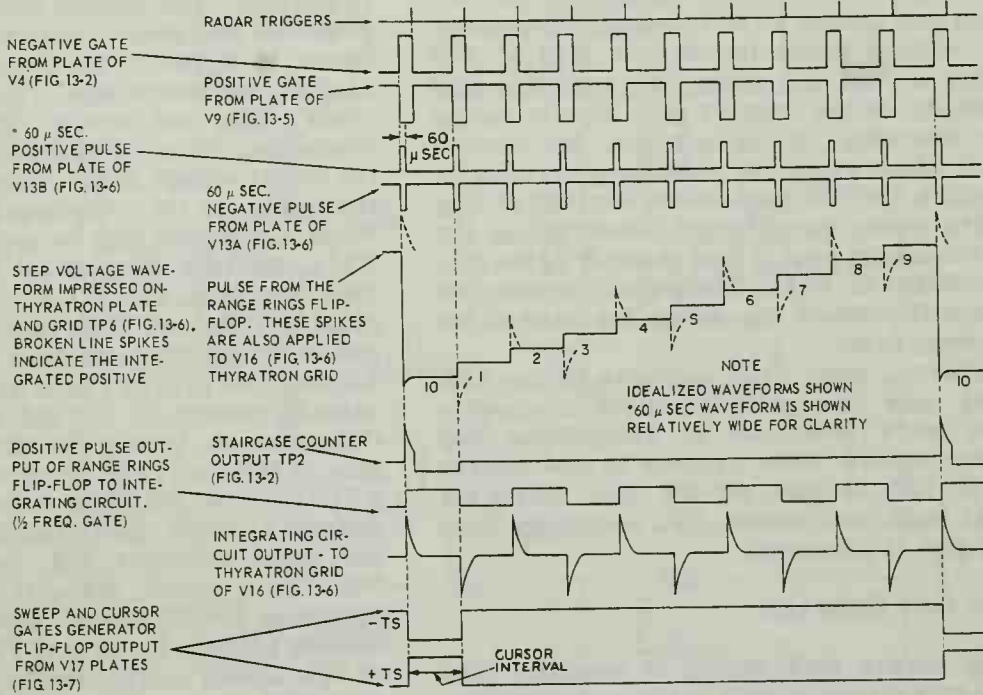
Figure 13-7.—Sweep and cursor gate generators and time relationships.

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starts the nine-to-one time-sharing cycle over again.

The output of V17A is the negative time-sharing wave, and is positive during the nine-out-of-ten normal gate intervals (sweep time) and negative during the one-out-of-ten cursor time interval. The output is directly coupled to V18A.

The signal developed across R43 is delivered to the video sweep-gate generator video amplifier, and the AEW electronic switch circuits.

The output of V17B is the positive time-sharing wave, which is negative during the nine video sweep intervals and positive during the 10th (cursor) interval. This output is directly coupled to V18B.

The signal across R44 is applied to the cursor gate generator.

The positive time-sharing wave is used in conjunction with the negative time-sharing wave to enable the AEW electronic switch. The positive time-sharing wave is also used in connection with the positive gate to enable the range strobe generator.

Sweep-Gate Generator

During the nine-out-of-ten normal gate intervals that section V17A is cut off, its positive plate voltage keeps the screen grid of V20 positive so that this sweep-gate switching tube is able to follow control grid signals during these intervals. At these times, the control grid of V20 is receiving positive-gate voltages. Therefore, the V20 plate output consists of nine negative sweep gates, which occur during the time the screen grid is held positive by the cut-off condition of V17A. The sweep circuits that follow will produce one sweep for each of the nine sweep gates.

However, when V17A conducts on the 10th normal gate, its plate produces a negative square wave (examined at TP8). When this negative square wave arrives at the screen grid of V20, it cuts off the tube during the normal tenth gate period. As a result, no video sweep gate is generated.

Cursor Gate Generator

The cursor gate circuit is similar to the sweep-gate circuit in both design and operation.

During the nine-out-of-ten normal gate intervals when V17B is conducting, its low level

plate voltage is coupled to the screen grid of V21. In this condition, V21 is not conducting. As a result, the positve gate applied to the control grid of V21 has no effect. At the end of the ninth gate, V17B is cut off by the discharge of thyratron V16 (fig. 13-6). The sharp rise in plate voltage at V17B produces a positive square wave that can be examined at TP9.

During the time that the square wave is applied to the screen grid of V21, the tube conducts for the interval of the 10th gate. At this time the arrival of a normal positive gate at the V21 control grid produces a negative plate output pulse of one gate length. This pulse is the cursor gate that produces one cursor sweep.

SWEEP GENERATOR AND
SWEEP RESOLVER DRIVER

The sweep generator consists of a linear sawtooth generator and a resolver driver. The resolver rotor is driven by the repeater servomotor, which responds to antenna bearing input signals. The resolver driver, V28, is a current source for the magnetic resolver that produces the E-W and N-S sweep inputs to the sweep amplifiers. A schematic diagram for the sweep generator and sweep resolver driver is shown in figure 13-8. Before the sweep gate from V20 is applied, the switch tube, V22A, the cathode follower, V25, and diodes V23A and V23B are conducting. When V23A and V23B are conducting, the initial voltage at the cathode of V25 is nearly the same as the control-grid voltage of V24. When a negative gate is applied to the grid of V22A, the tube cuts off for the duration of the gate. As V22A cuts off, a positive pulse is applied to V23A and V23B cathodes which cuts off the diodes for the pulse time interval. When this happens, the grid of V24 is no longer fixed at the cathode voltage of V25 but increases because C16 charges. During the time of the negative gate, V23A and V23B along with R50, complete a DC feedback path which determines the voltage difference which must exist between the grid of the sweep amplifier V24, and the cathodes of V25. As a result, there is only one possible operating condition that gives a voltage difference between the grid of V24 and the cathodes of V25. Direct current feedback is used to make sure that the sweep will start from the same voltage for each gate and that its average value will not change or drift.
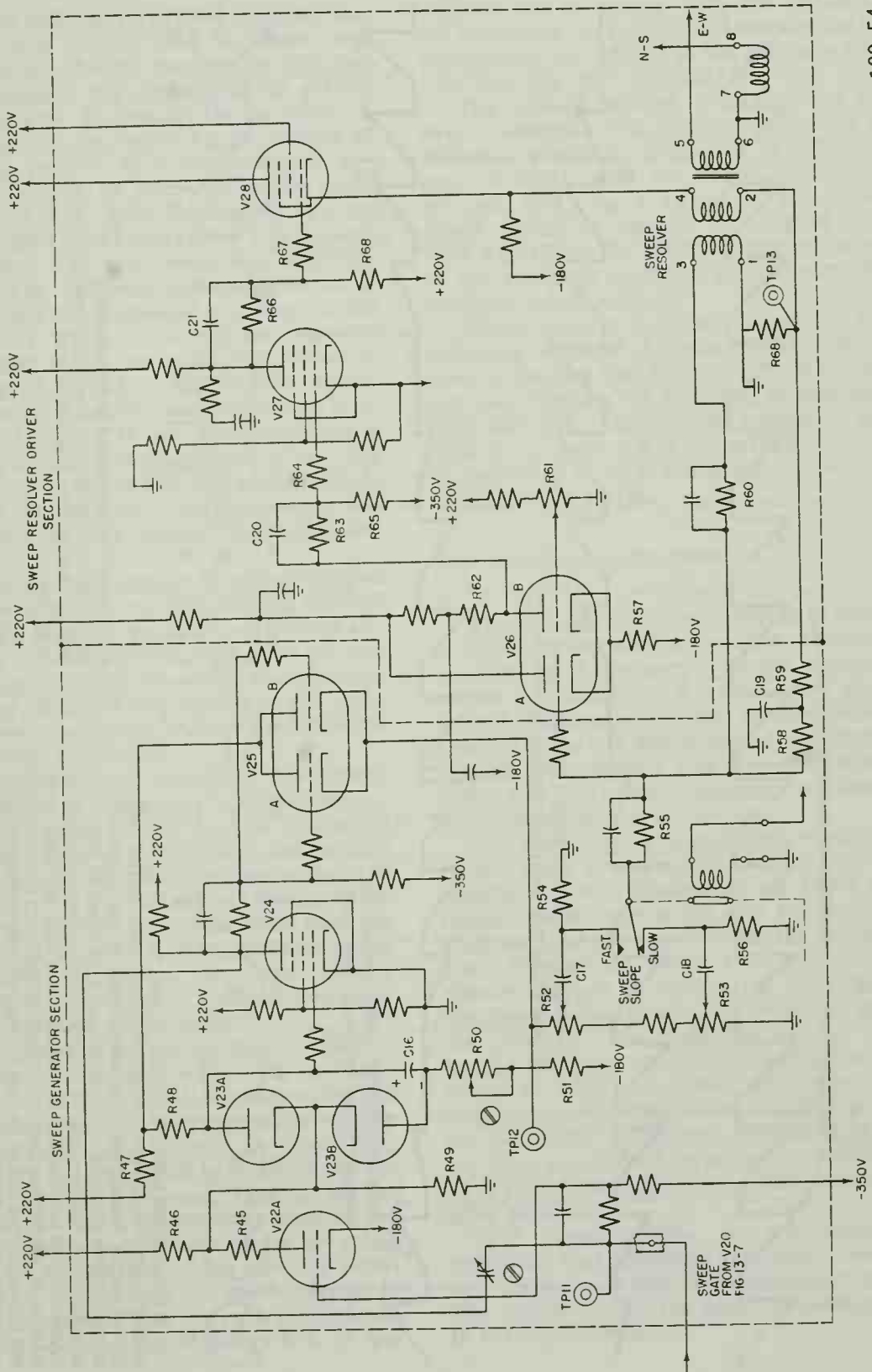
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Figure 13-8.—Sweep generator and sweep resolver driver.

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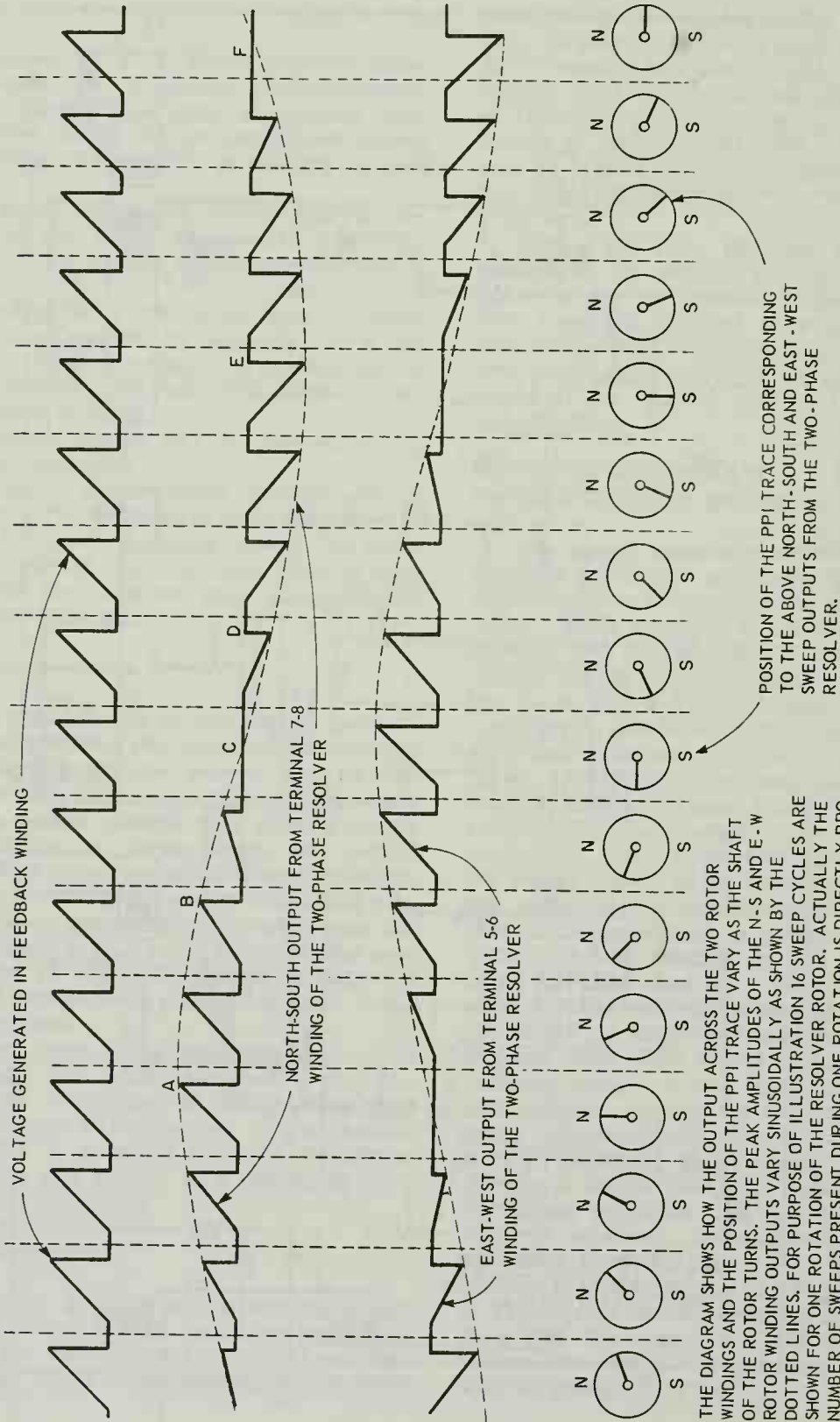
THE DIAGRAM SHOWS HOW THE OUTPUT ACROSS THE TWO ROTOR
WINDINGS AND THE POSITION OF THE PPI TRACE VARY AS THE SHAFT
OF THE ROTOR TURNS. THE PEAK AMPLITUDES OF THE N-S AND E-W
ROTOR WINDING OUTPUTS VARY SINUSOIDALLY AS SHOWN BY THE
DOTTED LINES. FOR PURPOSE OF ILLUSTRATION 16 SWEEP CYCLES ARE
SHOWN FOR ONE ROTATION OF THE RESOLVER ROTOR. ACTUALLY THE
NUMBER OF SWEEPS PRESENT DURING ONE ROTATION IS DIRECTLY PRO-
PORTIONAL TO THE RADAR PRF (60 TO 2000 PPS) AND INVERSELY PRO-
PORTIONAL TO THE ROTATIONAL SPEED OF THE INPUT ANTENNA SYNCHRO
DATA. THIS GIVES A VISUALLY SMOOTH ROTATION OF THE PPI TRACE.

VOLTAGE GENERATED IN FEEDBACK WINDING

NORTH-SOUTH OUTPUT FROM TERMINAL 7-8
WINDING OF THE TWO-PHASE RESOLVER

EAST-WEST OUTPUT FROM TERMINAL 5-6
WINDING OF THE TWO-PHASE RESOLVER

POSITION OF THE PPI TRACE CORRESPONDING
TO THE ABOVE NORTH-SOUTH AND EAST-WEST
SWEEP OUTPUTS FROM THE TWO-PHASE
RESOLVER.

Figure 13-9. — Output of the two-phase resolver after clamping.

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As soon as V23A and V23B stop conducting, the voltage drop across R48 becomes considerably smaller. As the junction of R48 and C16 is connected to the grid of V24 the grid of V24 tends to rise to nearly the B+ value. A small increase in the control grid voltage of V24 however, results in a much larger decrease in voltage at its plate and at the grids and cathodes of V25. This large drop in plate voltage is coupled back to the grid of V24 through sweep capacitor C16 and sweep step potentiometer R50. The net result is cancellation of most of the original increase in voltage at the control grid of V24. Therefore, the negative feedback action keeps the voltage increase at the control grid of V24 to very small values. There is a small constant current through R48, which flows through C16 and R50 producing a linearly increasing voltage across C16. The voltage at the grid end of C16 increases slightly while at the same time a large linearly decreasing sweep voltage appears at the R50 end of C16.

Because very little voltage drop is developed by the constant-sweep current flowing through R50, the large negative-going sweep voltage appears at the cathodes of V25, test point 12, and the top of the series-connected sweep slope potentiometer, R52 and R53. Resistor, R51, provides the proper operating current for V25 without loading down the output.

The negative sweep output of the sweep generator is fed to the sweep resolver driver, V26, through capacitor C17 or C18. Resistors R54 and R56 are used to discharge the capacitor not in use. By discharging the capacitors in this manner, the DC voltage at the output side of the couping capacitor is maintained near the value it would have if it were coupled to the resolver driver. Operating the circuit in this manner reduces transients caused by switching between the fast and slow ranges.

The first stage of the sweep resolver driver, V26, is a cathode follower (V26A), the second is an amplifier (V26B). A negative sweep voltage is applied to the grid of V26A decreasing the voltage drop across common cathode resistor R57. Therefore, the bias on V26B is reduced and the V26B plate current increases. This causes the plate voltage of V26B to decrease making the grid of V27 more negative and the plate more positive. The application of the positive signal to the grid of V28 causes its plate current to increase and draw more current through the cathode return circuit. Winding 2-4 of the

resolver is in series with the cathode, and the increased current through the winding causes voltages to be induced in the feedback winding, voltages to be induced in the feedback winding, 1-3, and the two secondaries, 5-6 and 7-8.

The voltage induced in the feedback winding has a sawtooth characteristic applied to the feedback resistor, R60, and is mixed at the grid of V26A with the sweep voltage coming through R55. As a result of the mixing, a small signal is applied to the grid of V26A. This signal has the effect of making the voltage on the feedback winding a good reproduction of the input sweep voltage.

Direct current feedback is used to hold the operating voltages at the desired levels. This is done by feeding the resolver current through R68 and then filtering the resulting voltage with R59, C19, and R58. The filtered voltage is then applied to the grid of V26A. The DC operating level is controlled by R61, which determines the bias on the grid of V26 B.

Two-Phase Sweep Resolver

The two-phase sweep resolver is essentially a transformer with a rotating secondary. It consists of a primary (stator) winding connected to terminals 2 and 4, two secondary (rotor) windings connected to terminals 5 and 6 and terminals 7 and 8, and an auxiliary stator winding (terminals 1 and 3) used to sample the stator flux and provide degenerative feedback to the input of the driver circuit.

The axes of the two secondary windings are at right angles to each other causing their mutual inductances with respect to the fixed primary inductance to vary as the sine and cosine of the shaft angle of the rotor. Thus, the turning of the rotor through the stator field provides a mechanical means of modulating the amplitude of the sweep generator output, which is applied to the primary. The resulting sweep voltages across the rotor secondary windings are 90° out of phase with each other, and each varies in amplitude according to the angular position of the particular secondary winding in relation to the fixed position of the primary winding. Figure 13-9 shows the output of the two-phase resolver after clamping.

For simplification, we will describe the operation that produces the output sweep voltage across only one of the secondary windings. The operation for the other secondary winding is exactly the same.

The positive-going sweep, sawtooth voltage at the grid of the sweep resolver driver, V28, controls the current flowing in the primary stator winding of the sweep resolver. This changing current sets up a magnetic field around the primary, which varies in intensity with the current waveform. The waveform may be examined at test point TP13.

Assume that the rotor has turned to bring the axis of the secondary (5-6) winding into a position parallel to that of the primary (2-4) winding, as shown at point A in figure 13-9. At this position, the magnetic field of the primary winding cuts the turns of the secondary (5-6) winding at right angles and induces maximum voltage across the secondary. As it is moved from its parallel position in relation to the primary, the secondary is cut by fewer electromagnetic lines of flux, and the induced voltage across it decreases correspondingly as shown at point B.

When the rotating (5-6) secondary has revolved through an angle of 90°, making its axis perpendicular to the axis of the primary (2-4) winding, its turns are parallel to the magnetic field of the primary and there is a minimum of flux linkage. Therefore, negligible voltage is induced as shown at point C.

As the rotor turns past 90°, the secondary winding is now reversed with respect to the primary winding, and the voltage induced across the secondary will be of opposite polarity as shown at point D. However, this induced voltage will increase in amplitude because of greater flux linkage as rotation continues, rising from 0 at the 90° position to maximum voltage at 180°, as shown at point E.

At 180° the secondary output voltage is equal in amplitude, but opposite in polarity to the output voltage at 0°, which is the start of the cycle. As rotation continues, the secondary output decreases until it reaches 0 voltage again at 270° (point F). At 270° the polarity of the output voltage is again reversed, and rises to a maximum at 360° (point A). The secondary output sweep voltage is modulated, therefore, by an envelope that varies sinusoidally as the rotor turns through 360.

A sweep output voltage is being similarly induced and modulated across the other rotor winding. However, since the axes of the two rotor secondary windings are at right angles to each other, the voltage across the second winding is being developed one quarter of a rotation earlier than the voltage across the

first. Therefore, the voltage peaks, nulls, and reversals of the second winding lead those of the first by 90°.

By means of a servo followup system consisting of the synchro bearing input, the servo amplifier, the drive motor, and the control transformer, the shaft angle of the sweep resolver (as shown in the small circle Fig. 13-9) accurately reproduces the bearing angle of the radar antenna. Therefore, the sweep on the PPI display points in the same direction as the radar antenna.

N-S Sweep Amplifier

The purpose of the N-S sweep amplifier, shown schematically in figure 13-10, is. to supply a low-impedance source voltage that is approximately 25 times the sum of the output voltages of the N-S windings of the sweep and cursor resolvers. This action provides a voltage level suitable for mixing with the off-centering voltages in the input circuits of the N-S yoke driver.

The voltages from the N-S sweep and cursor resolvers are mixed at the grid of V29. The sweep and cursor input networks are identical, but only the N-S chain may be individually adjusted. The potentiometers, R69 and R70, are in series with the resolver windings and are adjusted to compensate for normal difference in the N-S and E-W sweep signals.

For DC and low frequency signals, the input impedance for either sweep or cursor signals is 20,000 ohms. This impedance is adjusted in the N-S chain only. The d-c feedback impedance is R74 (500,000 ohms). The closed-loop gain to each amplifier input is determined by the ratio of the feedback impedance to the input impedance and, in this case, is 25 to 1.

To operate properly, the input DC voltage to the amplifier must be approximately -3 volts, and the output voltage at the plate of V31(TP14) must be approximately 20 volts under no-signal conditions. Because the input signal from the resolvers are at DC ground under no-signal conditions, it is necessary to bias the feedback network in order to keep the desired relationship between input and output voltages. In adjusting the amplifier, R72 is used to set the bias so that the output voltage at TP14 is + 20 volts under no-signal conditions.

The error signal is amplified and inverted in the plate circuit of V29.
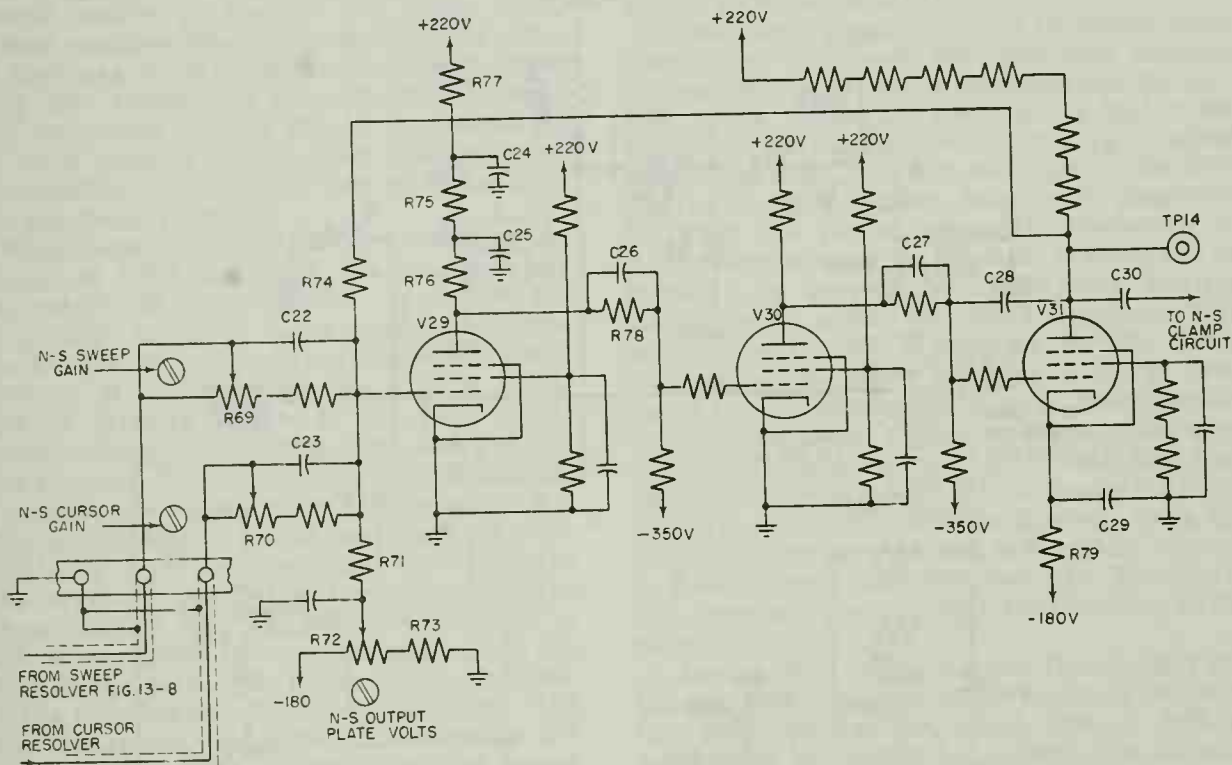
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Figure 13-10.—Schematic diagram of N-S sweep amplifier.

The signal from the plate of V29 is DC coupled to the grid of V30. Capacitor C26 provides an AC path for the high-frequency components in the signal. Tube V30 is a straight amplifier with its output DC coupled to V31. The path for high-frequency signals is through C27. The thrid stage, V31, has a feedback loop through C28 to provide an output characteristic that is free from the limitations that would result from plate bypassing. Another feature of this feedback system is that it permits the amplifier to have low impedance.

The output waveshape of the N-S sweep amplifier can be examined at TP 14. It should be a sawtooth wave with nine sweep slopes going negative or positive, depending on the direction setting of the sweep resolver. The tenth or cursor slope varies in polarity according to the direction setting of the cursor resolver. North is negative and south positive. Capacitor C30 provides a means of coupling the plate of V31 to the clamp circuit。

N-S Clamp

The N-S sweep voltage from the plate of V31 is fed to the clamp circuit shown in figure 13-11.

The purpose of the clamp is to prevent the start of the sweep from shifting with changes of sweep or cursor direction, repetition rate, sweep length, or other causes. To obtain this stability, the base line of the sweep must be restored to a fixed potential (near zero volts) before the start of the next sweep.

The positive-going clamp gate from the cathode follower, V11B, (test point 4 in figure 13-5) controls the turning on and off of the clamp. This gate is a disabling voltage to the clamp, and clamping can take place only on the negative swing of the trailing edge and during the interval between clamp gates. A clamp gate starts simultaneously with the start of each sweep, but is approximately 60 μs longer than the sweep gate. As a result, clamping is off

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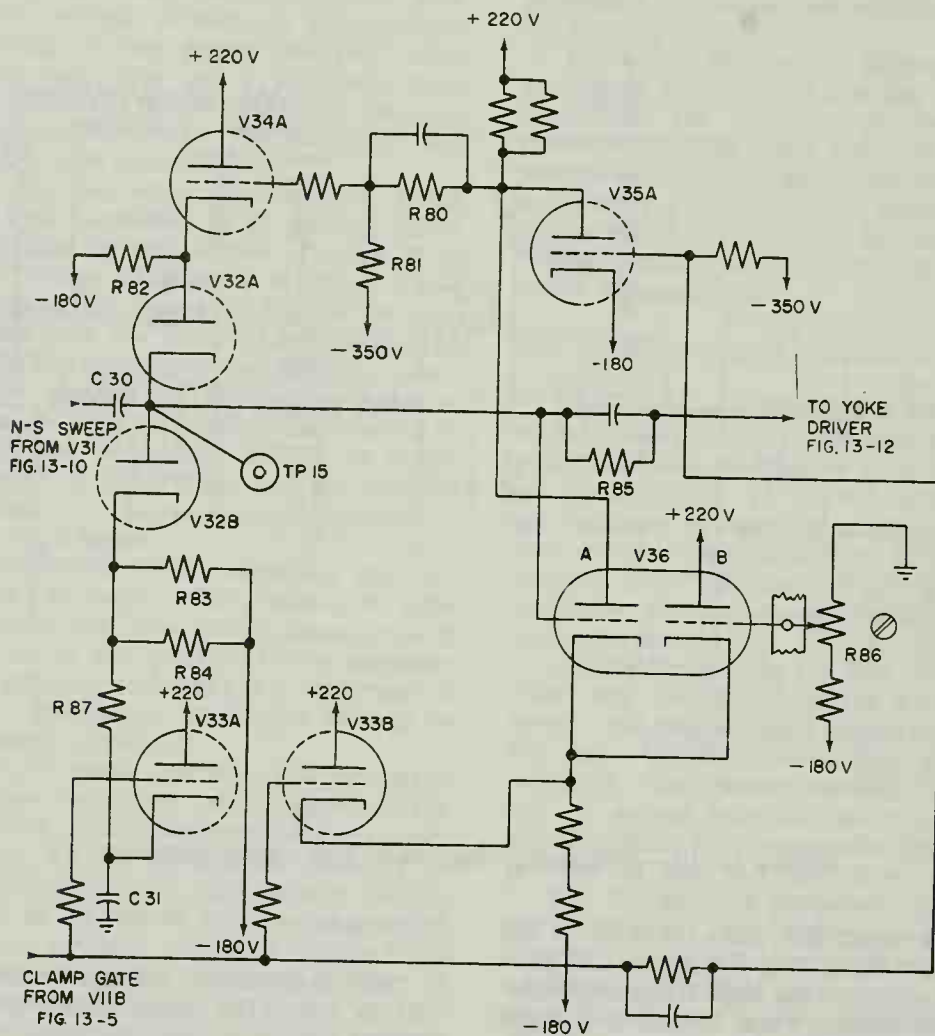
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Figure 13-11.—N-S clamp circuit.

during the sweep time, and for about 60μs thereafter. The clamp operates at all other times, even when there is no radar trigger.

The clamp gate must be longer than the sweep gate to give the 60μs additional time to recover toward the base line, thus preventing the circuit from starting to clamp to zero a signal voltage that should not be zero. If the last condition should happen, the baseline shift would increase, and clamping would be more difficult to achieve during the limited available time.

The clamp circuit is designed so that the clamp isolating diodes, V32A and V32B, conduct during the interval between clamp gates. During this time, the plate of V32A is approximately 1.5 volts positive and the cathode of V32B is approximately 1.4 volts negative.

Unclamping takes place when the unclamp switch tubes, V33A, V33B, and V35A conduct. These tubes conduct when the leading edge of the positive clamp gate is applied to their grids. The tubes conduct for the duration of the gate and are cut off between gates.

During the positive clamp gate, sweep and cursor voltages as high as ±100 volts are present at TP15, and these voltages must not be interfered with by the clamp. Each of the three

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unclamp switch tubes serves to isolate one possible source of load on the clamp point.

When unclamp tube V33A conducts, its cathode becomes slightly higher than the positive clamp gate voltage. As a result, the cathode of V32B is raised through R87 to approximately 120 volts positive. Tube V32B then becomes nonconductive and isolates this source of load from the clamp point.

When unclamp tube V33B is conducting, the cathodes of V36A and V36B are raised to approximately 120 volts positive. Therefore, the clamp point is never more positive than the cathode of V36A even with the +100-volts sweep signal on it. Consequently, the grid of V36A, which is directly coupled to the clamp point, can never draw grid current. This action eliminates a second source of load on the clamp point.

Another source of loading is eliminated when V35A is conducting. In this condition, the plate of V36A is made negative. The grid of V34A is made over 100 volts negative by the action of voltage divider R80 and R81. The cathode of V34A is energized through R82 and follows the action of the grid. This action makes the plate of V32A negative by a voltage greater than -150 volts. Therefore, no matter how negative the sweep becomes, V32A can never conduct during the positive clamp gate period.

One additional load source is connected to the clamp point (TP15). This is the input to the yoke driver section, which is a necessary load, and the clamp circuit is used to reduce the effect of this load in producing baseline shift.

During the time the clamp gate input is negative, V33A, V33B, and V35A are not conducting. Therefore, the clamp is operating and provides a means of automatically bringing TP15 to a voltage near zero. The action for this is as follows: assume that TP15 tends to go positive. Tube V36A will then conduct, and its plate will go more negative, which makes the grid of V34A go more negative. This action then causes the plate of V32A to go more negative. Less current will then flow through V32A, and some unbalanced current will flow through V32B. The lack of current balance between the two diodes acts to restore TP15 to zero since the unbalance is in a direction that will pull TP15 negative. As soon as TP15 has been restored to the proper voltage, the action stops. A reversed action takes place when TP15 becomes too negative.

Potentiometer R86 controls the voltage on the grid of V36B, which in effect, controls the grid bias voltage of V36A and can change the voltage at which the clamp is in balance.

Resistors R87 and C31 have important functions. When the clamp gate has gone negative and enabled the clamp, it is undesirable for current to be flowing in the negative half of the clamp circuit (V32B). Some time is required before the plate of V36B can rise to the point where feedback action effectively maintains TP15 at the proper potential. Therefore, if V32B were allowed to conduct, it would not be balanced by current from +220 volts, and the clamp point would be pulled negative. This action would make it difficult for the clamp to operate effectively because a charge produced by the unbalanced current would appear on C30. Therefore, C31 is used to delay the drop in the cathode potential of V32B until the other half of the clamp circuit is able to control TP15 properly. When the clamp is disabled, V33A can conduct more current than is normally allowed to flow through R87, R83, and R84. Therefore, C31 does not greatly delay the disabling to the clamp.

During each sweep, current flows from TP15 through R85. Because of the feedback action in the yoke driver, the output side of R85 is essentially connected to ground. The current flowing through R85 during each sweep charges C30. The capacitance of C30 is made as low as practical ($0.1\,\mu f$) because it must be discharged rapidly during clamp time. The low capacitance produces some distortion of the sawtooth voltage, which is coupled through it to the grid of the N-S yoke driver. The distortion causes the sawtooth to droop back slightly toward its starting potential. However, the sweep generator was designed to produce a negative sawtooth that curved downward. Thus the distortion produced by C30 and R85 is overcome.

NORTH-SOUTH YOKE DRIVER

The N-S yoke driver produces N-S deflection yoke currents that are proportional to the sum of the mixed N-S sweep voltage, manual DRA off-centering voltage, and AEW off-centering voltage. The circuitry used to perform this function is composed of two feedback-amplifier sections, a forward section with variable gain that is controlled by the range selector, and an inverter section whose amplifier gain is fixed. The circuitry also contains positive and negative limiters and a neutralizing system in order to minimize trouble from large input voltages.

Because electron tube characteristics vary considerably, a special feedback system is used to maintain an exact relationship of inputs and output. This feedback system is composed of adjustable capacitors and precision resistors. These parts are used in gain-determining networks, which mix the inputs with the desired output, and then deliver close to zero voltage whenever the output is in the desired relation with the inputs.

The desired outputs are supplied by two high-gain DC amplifiers, V40-V41, the forward section and V48-V49 the inverter section (fig. 13-12). Those amplifiers, because of their high gain, require very small input signals in order to provide the desired output. The input to each amplifier is supplied by the output of a corresponding feedback network. Part of the amplifier output is fed back to the input (negative feedback) and tends to cancel the voltage fed into the input resistor. The small voltage output from the mixing network represents error in the relation of output to input voltages. Basically, the signal at the amplifier input grid is practically zero, and the amplifier output adjusts itself. Therefore, the stable mixing network controls the input-output relationship, and variations in vacuum tube characteristics are unimportant.

The desired yoke driver output is a current of either polarity. In order to drive current through the deflection - yoke in either direction, the deflection-yoke coil is fed by the halves, L1 and L2, each half producing an opposite effect. When both coil halves carry equal currents, deflection is zero.

Deflection is produced by changing both currents in opposite direction, and by nearly the same amount.

One-half of the deflection coil is fed by the forward section of the deflection yoke driver; the other by the inverter section. The inverter section is driven by the output voltage of the forward section, and produces nearly equal but opposite changes in its output current.

The driver-amplifiers produce output voltages that are proportional to the output current. Therefore, the yoke driver system can be designed as a voltage amplifier, and since these output voltages are not appreciably affected by the yoke coil impedance, the output current is independent of the yoke impedance. In other words, the yoke drivers have constant-current outputs.

Forward Yoke Driver Feedback System

The relay contacts (fig. 13-12) are shown in the position used for sweep speeds from 34 to 300 miles per radius.

The three input networks along with one feedback network are connected to the grid of V37B. The feedback network has no polarity-inverting elements, and must produce near-zero voltage at the grid of V37B. The amplifier output voltage must be of opposite polarity to the inputs.

Changes in output voltage (and yoke current) should be proportional to the sum of the three input signals. In addition, a zero input voltage sum must always produce the same no-signal output current regardless of the gain control setting. Unless this requirement is met, the origin of the centered PPI display will vary as the range selector is varied. The no-signal output voltage is 18.17 volts, which corresponds to 92.7-ma cathode current for the output tube and to 87.0 ma through the forward coil of the deflecting yoke.

The voltage that is fed to the grid of V37B is proportional to the sum of the inputs, less a portion of the output that is fed back.

The sweep-speed selector control, R90, varies the fraction of output voltage fed back through R91 and C34 and controls the amount of feedback required to keep the voltage at the grid of V37B at zero. In order to obtain cancellation the grid voltage at the wiper of R90 must be the same for a given signal no matter what the position of the wiper. Therefore, the closer the wiper voltage is to ground, the more signal needed at the output cathode and the higher will be the input-output gain.

To keep the no-signal voltage (zero input) independent of the setting of the gain control, R90 is fed through a voltage divider, R89 and R92, which is in parallel with several other paths. The junction of these resistors is maintained at zero volts for the desired no-signal output voltage that will produce constant (zero) voltage on R90. Therefore, varying the wiper position on R90 with zero voltages will have no effect under no-signal conditions.

Resistors R93 and R94 form an adjustable load on the voltage divider output. Under no-signal conditions, the divider junction is at zero volts, the adjustable load varies the divider attenuation without changing the DC level.

For sweep speeds below 34 miles per radius, higher gain is required, and the feedback current to the grid of V37B is decreased by switching R95

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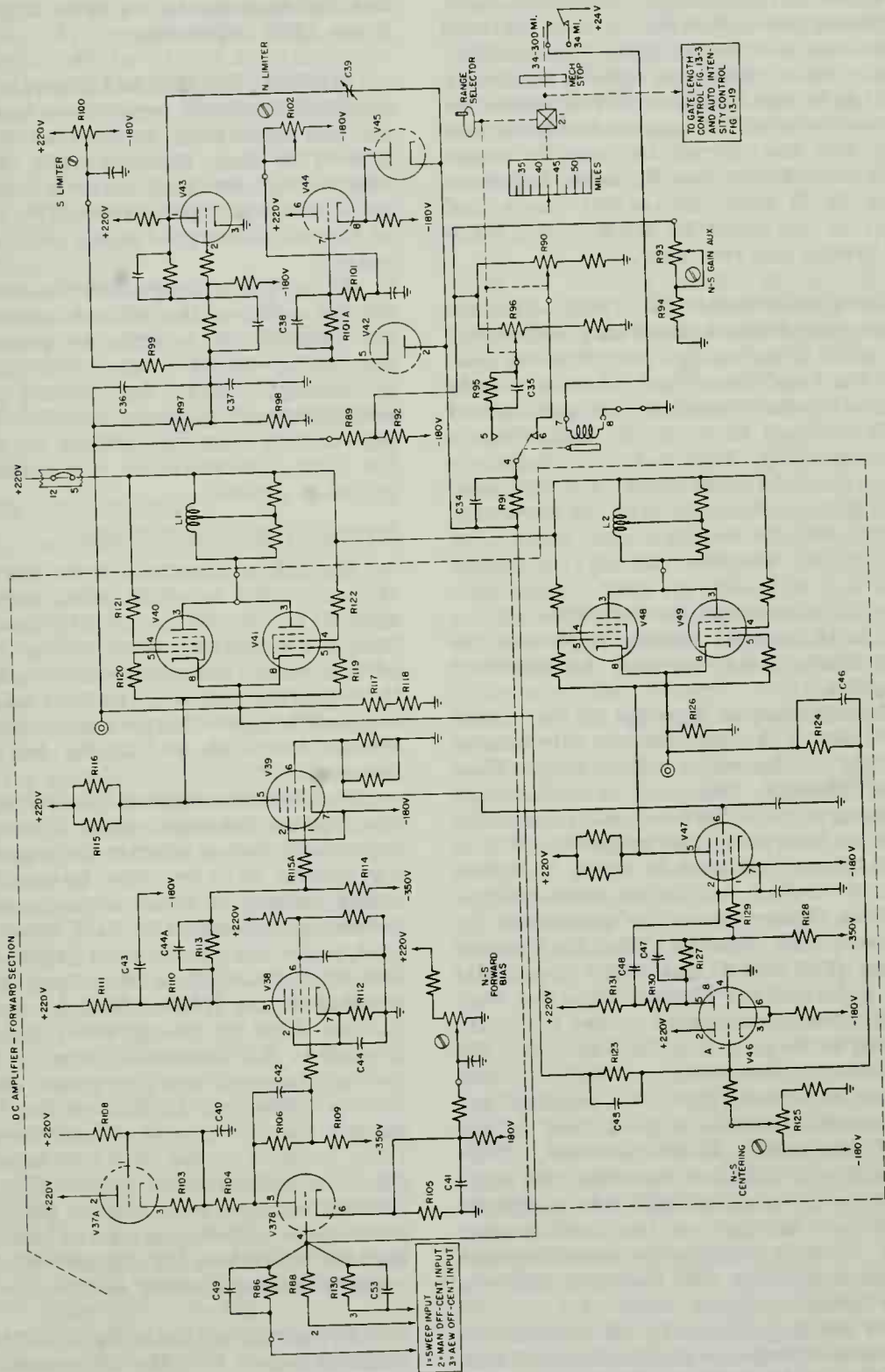
Figure 13-12.—N-S yoke driver.

in series with R91, and adding C35 to provide a high frequency path. Resistor R90 is replaced by R96 in order to have continuous rotation without a gap or sudden change in value as the sweep speed dial is turned. Resistor R96 is phased so that as the wiper of R90 approaches the end of its range, R96 has just started over its range. Thus, the switchover can be made anywhere within the 34-37 miles-per-radius range, and the result is the same as if the whole range were not broken into two parts.

LIMITERS FORWARD SECTION.—The inputs to the yoke driver system may correspond to many radii of deflection, and it is not possible for the amplifier output tubes to supply the required inputs. Therefore, the yoke driver system is designed to accept the larger inputs without overloading. High gain DC amplifiers using large amounts of feedback as in this case, contain large capacitors that will not permit the amplifier to recover rapidly from a large input signal. To offset the slow recovery, a limiter system is used to modify the input-output feedback network so that the forward yoke driver system gain is lowered enough to prevent the maximum inputs from reaching the amplifier overload point.

The limiter section consists of the biased diodes V42 and V45 which conduct only outside the useful part of the output voltage range. When these tubes conduct, they parallel the feedback path to the grid of V37B and greatly lower the impedance of this path. The output of V37B is reduced to about one-third in order to reduce the tendency toward oscillation by a voltage divider. This divider feeds the cathode of the south limiter, V42, directly. When the forward yoke driver (V40 and 41) cathode voltage falls between 1 and 3 volts, V42 conducts so that very slight increases in the output cancel increases in the inputs to the grid of V37B. Therefore, the system limits at this voltage without causing overload in the DC amplifier. The exact voltage where conduction starts is controlled by the setting of the south limiter control, R100.

For limiting in the north direction, the same voltage that feeds the south-limiter diode cathode is used to feed the plate of the north-limiter diode, V45. As soon as the output amplifier tube cathode reaches +36 to +38, the diode conducts and limits further changes.

Bias for keeping the diode cut off until the desired output voltage is reached is furnished by R102, the north-limiter adjustment. Tube

V44 matches the high divider impedance to the lower diode impedance.

LIMITER NEUTRALIZER.—The limiter diodes' interelectrode capacitance is always present and effectively in parallel with C34 and C35. This diode shunt capacity effect is undesirable because it is variable with the sweep-speed selector setting. Therefore, an adjustment of C34 and C35 is good at only one value of sweep speed.

To permit proper operation at all ranges, a voltage equal to the voltage across the diodes, but opposite in polarity is produced by the neutralizer-inverter tube, V43. This voltage is fed through C39, the neutralizing capacitor and is adjusted so that it feeds the grid with a current that is equal and opposite to the current from the diode interelectrode capacitance, and cancellation results.

Inverter Yoke Driver Section

The inverted voltage from the output stage is fed to the inverter grid, V46A. With the desired output voltage and current from V48 and V49, the inverted voltage at the grid of V46A will be nearly canceled by the feedback current through R124 and C46. The N-S centering potentiometer R125 provide a means of adjusting the bias voltage for V46A so that the desired output is obtained.

The Resistor R126 is the feedback sampling resistor for the output stage. The range selector adjustment has no effect on the inverter. Centering control R125 can also be used to provide a small amount of fixed off-centering. Any off-centering introduced by R125 however, must be held below six millimeters (approximately .25 inches) because the system may overlaod and produce severe trace-splitting.

The input to the inverter yoke driver is limited in the forward section so the inverter section does not require special overload protection. However, should the inverter be overloaded, the sweep may wander around with antenna rotation or be split into several parallel lines during timesharing. Therefore, the north-limiter voltage must not be too high and the coarse off-centering resistors (not shown) must be used to reduce the amount of off-centering required of the inverter section.

INVERTER SECTION DC AMPLIFIER.—The inverter signal to V46A is cathode coupled to V46B where it is amplified but not inverted in

he plate circuit. The plate output of V46B is
-c coupled to V47 through R127, C47, and R
28. The signal fed to V47 is amplified, in-
verted, and directly coupled to the grids of the
parallel-connected drivers, V48 and V49. These
ubes drive the inverter section of the N-S de-
lection coil, L2. Their cathode current is
closely proportional to the plate current and
lows through R126. The voltage across R126 is a
good sample of the output current and is used as
he output voltage.

The capacitor, C48 which couples R131 to
he suppressor grid of V47, is used to obtain the
desired frequency response. Capacitor C47 pro-
vides an AC path, between the plate of V46B and
the grid of V47, for high frequencies by par-
alleling R127 and thus preventing excess attenua-
tion and phase shifting because of the input
capacity of V47.

MANUAL OFF-CENTERING

The manual off-centering controls are geared
to the moving arm contacts of two 30-K poten-
tiometers, R132 and R133 (fig 13-13). These con-
tacts pick off DC voltages from the voltage
dividers that are connected across the + 220-and
the -180-volt power supplies. The arms of the
potentiometers feed the yoke drivers the normal
DC off-centering voltages of 0.8 volts-per-
mile in the fast (short) ranges, and 0.2 volts-
per-mile in the slow (long) ranges. The direction
of off-centering is the vector sum of the voltages
fed to the yoke drivers. Positive voltages off-
center the display to the east and south; negative
voltages to the west and north.

For the fast ranges the E-S and N-W potenti-
ometers, R134A-B and R-135A-B, are used to
preset the voltages at the midpoints of the posi-
tive and negative halves of R132 and R133. The
positive midpoint on both R132 and R133 is tap
4, and the negative midpoint is tap 2. The voltage
at the midpoints is measured with respect to
ground, and must be preset for -104 volts
from the negative and 104 volts from the
positive midpoints. The amount of off-centering
will then represent 130 miles.

For the slow ranges, R136 is used to set the
voltage between point 3 on both potentiometers
and ground to -52 volts, and R137 is used to set
the voltage between ground and point 5 on both
potentiometers to + 52 volts. The 52 volt values
represent 260 miles off-centering.

The circuit is switched from slow to fast
range condition automatically when the range
selector passes 34 miles (range decreasing).

When this happens, a circuit is closed to ener-
gize K2 and operate switch sections S3 and S4.

The mechanical action is set up so that one
revolution of the off-center control is equal to
approximately 20 miles off-centering for either
fast or slow ranges. A stop is geared to the
shaft to allow 51 turns from the zero off-
centering position in either direction.

AEW ELECTRONIC SWITCH

In the tracking function of the indicator, the
AEW electronic switch (fig 13-14) provides zero
volts across R138 and R139 during sweep time,
thus permitting no off-centering of the sweep by
this circuit. During cursor or tracking strobe
time, the circuit applies adjusted d-c voltages of
opposite polarity to both ends of the center
grounded R138 and R139. Thus during cursor
time, voltages of variable amplitude and polarity
are available from the arms of R138 and R139
to provide the required off-centering of the
cursor or tracking strobe in the direction
desired.

The AEW system uses radar carrying air-
craft to supply radar information to shipboard
installations. To reproduce this information, the
tracking indicator uses a tracking strobe, or the
origin of the electronic cursor, to follow the
position of own ship on its display. The AEW
electronic switch is necessary in order to cancel
the motion of the radar carrying plane. The
information is inverted for the final repeater
and then transmitted by synchros to the final
indicator where the display is centered on the
ship's position.

The inputs to the electronic switch are the
positive and negative time-shared waves from
V18A and V18B (fig. 13-7). The outputs are
positive and negative off-centering waves of
about 104 volts, which are applied to the opposite
ends of the potentiometer network. The switches,
S5A and S5B, are operated by a solenoid. When
the switches are in the position shown, the elec-
tronic switch applies the off-centering waves
during the nine video sweeps, and clamps at
near zero during the 10th (cursor) sweep. The
amplitude and polarity of the off-centering volt-
ages fed to the yoke drivers are controlled by
the potentiometer settings.

With switches S5A and S5B in the position
shown, the negative time-shared wave from
V18A (fig. 13-7) is applied to the grid of V50,
and the positive time-shared wave from V18B
is connected to the grid of V53A.
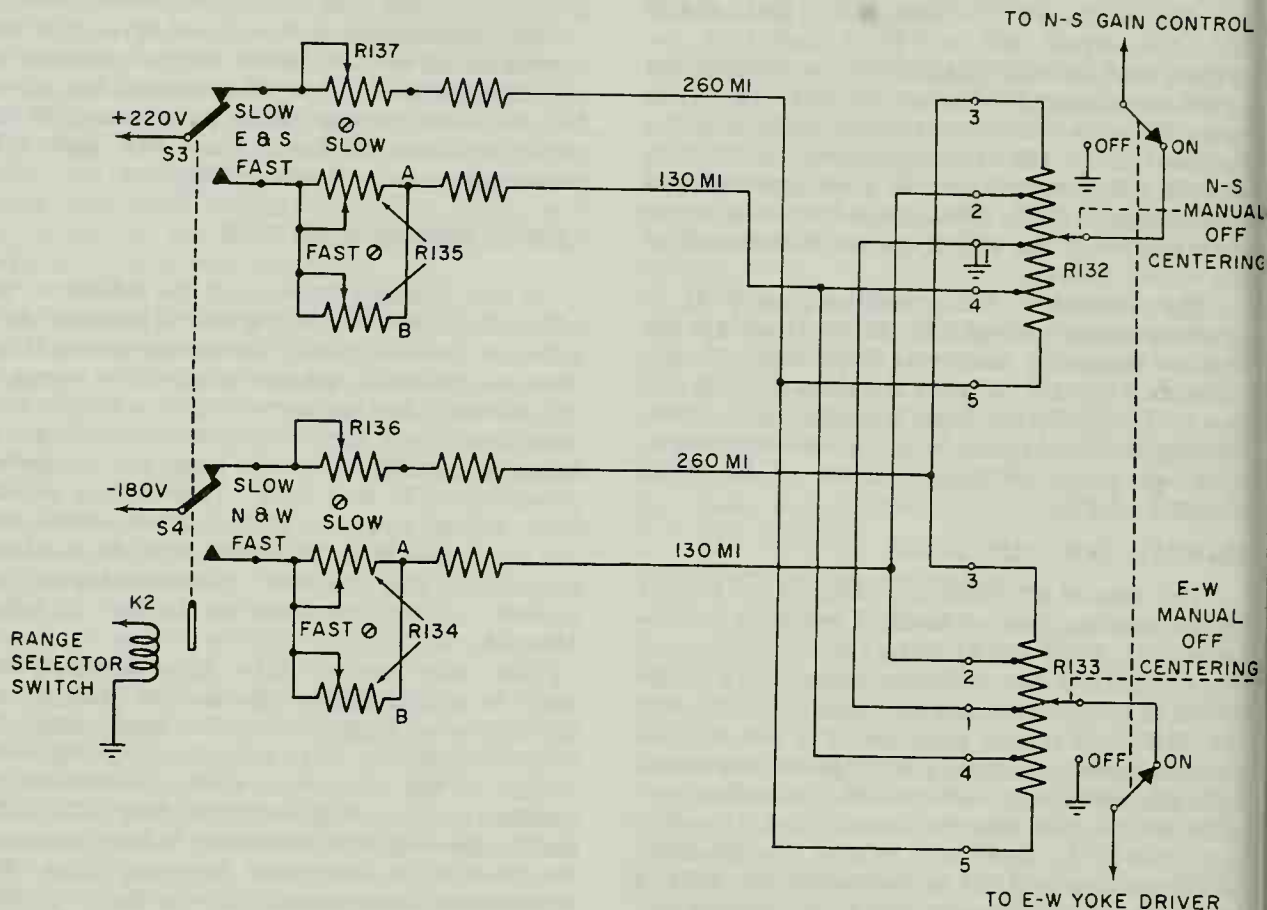
Figure 13-13.—Manual off-centering circuits.

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The negative time-sharing wave cuts V50 off during the cursor interval, and the tube conducts during the nine video sweeps, While V50 is cut off, the clamping diode V15A conducts. This action ties the grid of V52 to the voltage divider network.

The setting of R140 determines the amplitude of the positive driving voltage at the grid of V52 and the amplitude of the positive cathode output applied through S5B to one end of the off-centering network. The pulse may be observed at T P16. Normally R140 is set to maintain 87 volts positive during the cursor interval. Tube V50 conducts during the nine video sweep intervals because the negative time-shared wave is more positive for this interval. When V50 conducts, the grid voltage of V52 becomes less positive, and V51A stops conducting. When V52 stops conducting, the current through R143 drops, causing the voltage

at the cathode to drop toward a negative value at the start of the nine video sweep intervals.

When the cathode of V52 falls to zero, the clamping diode, V51B conducts. The conduction of V51B is controlled by the setting of R144. The cathode of V52 will be clamped to zero volts at the start and for the duration of the nine video sweep intervals. As soon as the next cursor interval occurs, the negative time-sharing wave cuts off V50, clamping the grid of V52 as described above. This method ensures rapid switching between the conducting and nonconducting states.

When a positive time-shared wave is fed to the grid of V53A, the tube conducts during the cursor interval and acts as an open circuit during the interval of the nine video sweeps. When V53A conducts, the positive-going output is coupled directly to the cathode of V54A and
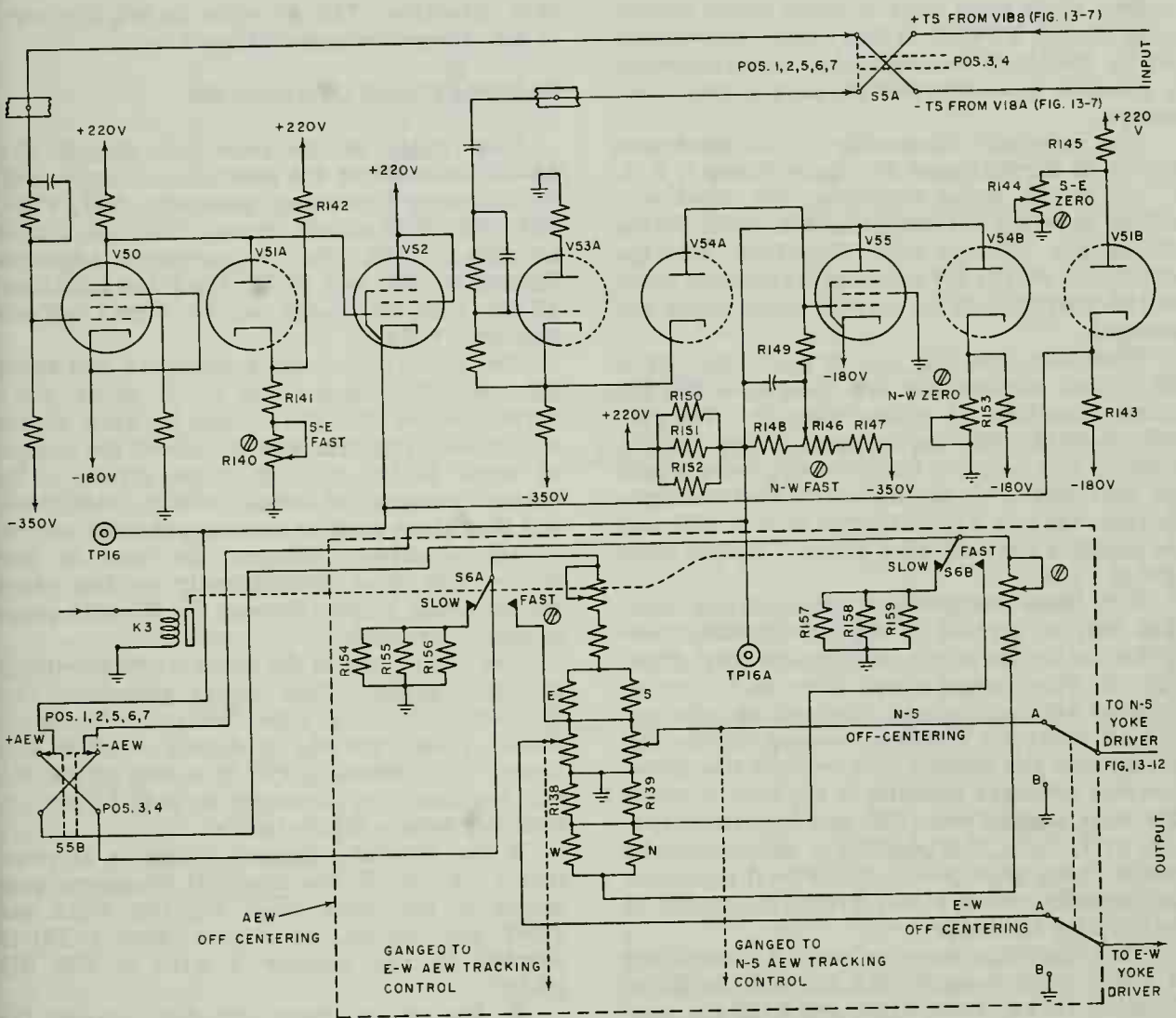
109.60

Figure 13-14.—AEW electronic swtich.

causes V54A to cut off. The plate voltage of V54A rises to the voltage determined by the setting of R146, the values of R147 and R148, and the plate voltage of V55. The plate of V55 will now be negative to ground because the tube is conducting. The voltage to which the plate of V55 drops is normally adjusted by R146 to 87 volts negative. This is the negative off-centering voltage fed to the AEW potentiometer network during the cursor interval. The small voltage fed back through R149 helps to reduce the output voltage drift at the plate of V55.

The positive, time-shared wave input to the grid of V53A cuts the tube off during the nine video sweeps because the positive time-shared wave is negative-going during this interval. Therefore the cathode of V3 goes negative at the start of the video sweeps, and because it is directly coupled to the cathode of V54A, V54A conducts. The resulting drop at the plate of V54A cuts off V55, and the plate of V55 rises. The clamping action of diode V54B keeps the plate of V55 from rising above zero volts because the potentiometer, R153, is set so that V54B will

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conduct if its plate tries to swing above ground potential. As a result of this action, the voltage fed to the AEW off-centering potentiometers is clamped to zero volts during the nine video sweeps.

The previous discussion of the electronic switch is for S5A and S5B in positions 1, 2, 5, 6, or 7. In these positions, the electronic switch provides off-centering potentials during the cursor interval only. Therefore, only the electronic cursor (or tracking strobe) will move as the positions of the potentiometer arms are changed.

When S5A and S5B are in positions 3 or 4, the circuit connections are changed so that the cursor functions as a fixed indicator. The circuit operation remains the same, except that the positive and negative time-sharing wave inputs are reversed. This means that the positive time-sharing wave is applied to the grid of V50, and the negative time-sharing wave is applied to the grid of V53A.

With these connections, off-centering voltages will be applied to the potentiometer networks during the nine video sweeps only. Then, only the video sweeps will move as the potentiometer arms are moved. The positive and negative off-centering voltages are switched by S5B so that they are applied to the ends of the potentiometer networks opposite to the ends to which they were applied when S5B was in position 1, 2, 5, 6, or 7. Thus, it is possible to obtain opposite motion of the video sweep at the final indicator, and opposite motion at the strobe, or cursor, at the tracking indicator.

When the range selector control is set above 34 miles (slow ranges), S6A and S6B, as shown in figure 13-14, R154, R155, and R156 are connected in the circuit of the E-W off-centering potentiometer, R138; and R157, R158, and R159 are connected in the circuit of the N-S off-centering potentiometer, R139. With the resistors connected in this manner, the current distribution in the off-centering circuits is such that one-half of 87 volts (43.5 volts) is available for off-centering in E-W and N-S directions.

When the range selector control is set below 34 miles (fast range), a cam-operated switch causes the coil of K3 to be energized, and the moving contacts cut the resistors, R154, R155, and R156 and R157, R158, and R159 out of the circuits of the E-W and N-S off-centering potentiometers. When K3 is energized in this manner, the full 87 volts output of the electronic switch is available for off-centering in the E-W and

N-S direction. The 87 volts is fed into tap 1 of the off-centering potentiometers.

RANGE STROBE GENERATOR

The range strobe generator circuit (fig. 13-15) consists of the coincidence mixer, V60; the bootstrap type sweep generator, V61, V62A, and V63; the pickoff diode, V64; the pickoff amplifier, V65, the regenerative amplifier formed by V66 and V67A; the pulse amplifier, V67B; and the range strobe output cathode follower, V68A.

The bootstrap circuit generates a fast sweep for the short ranges (up to 36 miles) and a slower sweep for the ranges between 36 and 300 miles. The fast sweep reduces the amount of error in the position of the strobe on the cursor because of small voltage variations, and thus gives greater accuracy at short ranges. Bootstrap circuit changes for slow or fast ranges are done automatically as the range strobe crank passes through the 36-mile range in either direction.

The timing unit of the range strobe generator is the bootstrap type sweep generator that includes the clamp tube, V61; the bootstrap diode, V62A; and the bootstrap cathode follower, V63. Although V63 is a dual triode, the two sections are connected in parallel and effectively form a single triode.

If the function selector switch is in position 1, 3, or 5, the one-half frequency gate output of the range rings flip-flop V59A and V59B (not shown), or cursor wave (+T8) is coupled to the number 1 grid of V60 (fig 13-15).

In the range rings flip-flop circuit the squarewave output of the flip-flop is one-half the PRF of the normal gate. Therefore, grids 1 (positive gate) and 3 of V60 (fig. 13-15) will be positive at the same time, only during the interval of every other gate. As a result, the PRF of the negative gate at the plate of V60 is one-half the normal gate, and a range strobe is generated, and appears on, every other video sweep.

When the function selector switch is in position 2, 4, 6, or 7, the cursor sweep appears on the display. For this type of operation, the positive time-sharing wave from V18B (fig 13-7A) is connected to grid 3 of V60, (fig 13-15), and the positive gate from V9 (fig 13-5) is connected to grid 1 of V60. The PRF of the negative range strobe generator
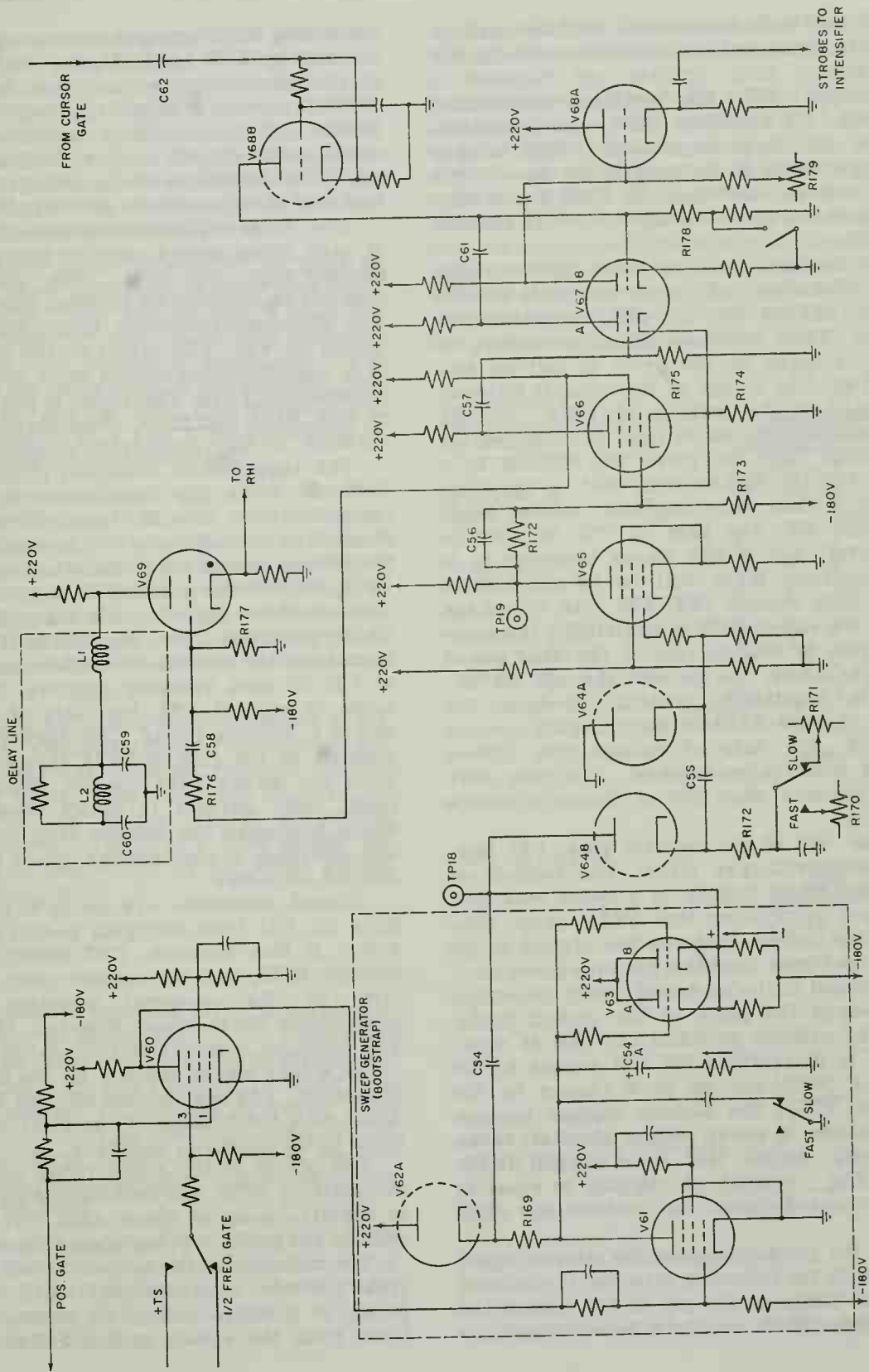
Figure 13-15. — Range strobe generator.

109. 61

gate is one-tenth the normal PRF and a range strobe is generated and appears on the cursor sweep only.

Normally, V61, the bootstrap clamp tube, and V62A, the bootstrap diode, are conducting. Because the diode impedance (V62A) is very low, the voltage at its cathode is close to the supply voltage, and because R169 has a high resistance the plate voltage of V61 is close to zero volts.

The leading edge of the negative range strobe generator gate from the plate of V60 cuts off V61 so that its plate voltage starts to rise. This increase in plate voltage is directly coupled to the grids of cathode follower V63. The output of the cathode follower is coupled to the cathode of V62A through C54. Only a very short time is required for the voltage rise to raise the cathode to a voltage that is higher than that on the plate of V62A. When this condition exists, diode V62A cuts off. The gain of V63 is close to one (unity), and if C54 (large capacity) is in good condition, there will be no appreciable voltage drop across C54. The rate of voltage rise at one end of R169 is practically the same as the rate of voltage rise at the other end of R169. Therefore, the current through the resistor is practically constant. Because the current through R169 is the charging current for C54A, the rate of voltage rise across C54A is practically constant. Charging current is constant when rate of change of voltage is constant.

At the end of the negative gate, V61 conducts and discharges C54A. The cathode of V62A then drops rapidly to a value less than the supply voltage so that V62A again conducts. This action restores the circuit to its original no-signal condition between gates.

The output of the bootstrap sweep generator is applied to the plate of the pickoff diode, V64B. The cathode potential of V64B is variable and is determined for fast ranges by the setting of R170 and on slow ranges by the setting of R171. The cathode current through R172 produces a sweep pattern identical to the plate sweep, except that it is clipped at the bottom. The amount of clipping is equal to the difference between the cathode and plate voltages.

When the leading edge of the clipped signal is applied to the following circuits, it produces the output range strobe pip at the instant the pickoff diode, V64B, conducts. Because the diode conducting time during the sweep is continuously variable by R170 and R171, the range strobe pip may be made to appear at any point on the sweep. Therefore, the radius of the range-strobe ring of the distance of the range strobe from the origin on the display tube are continuously variable. The waveform at the cathode of V64B for fast and slow sweeps is shown in figure 13-16.

The differentiated waveforms at the bottom of each group show a negative backswing at the trailing edge. The diode, V64A, eliminates the negative backswings by shunting them to ground. The differentiated output is amplified and inverted by V65. The signal at the plate of V65 is a negative-going square wave, which may be examined at TP19. The signal is peaked slightly by C56, R172, and R173, which forms a coupling network to the regenerative amplifier, V66.

The regenerative amplifier circuit includes V66 and V67A (fig 13-15), and is similar in operation to a one-shot multivibrator. In the normal, no-signal state, V66 conducts and holds the voltage across the common cathode resistor, R174, sufficiently positive to keep V67 cut off. This condition exists until the pickoff diode, V64B, conducts. Then, the differentiated waveform from the cathode of V64B causes the plate of V65 to start swinging negative. This action tends to cut off V66, the plate of which then starts a positive swing. The positive swing is coupled to the grid of V67A through C57 and R175 and makes V67A conduct. When this happens, more current is drawn through R175, which increases the voltage drop across R174 and provides a regenerative effect by making V66 cut off faster.

Circuit conditions are set up to prevent the plate of V65 from swinging positive immediately. If this happens, C57 would discharge through R175, and the multivibrator would return to its no-signal condition after approximately 60 to 70 μs. However, the plate of V65 remains comparatively negative until the range strobe sweep flattens on the fast range. Therefore, V66 remains cut off, and V67A conducts until the voltage at the plate of V65 returns to the no-signal condition.

The output of the regenerative amplifier at the plate of V67A is a rectangular wave, which is negative going at times when V67A is conducting and positive going when V67A is cut off.

The output from the cathode of V69 (a thyratron) provides range information to a remote point. A positive pulse, at strobe time, is taken from the screen grid of V66 and coupled

the grid of V69 through R176 and C58. Resistor R176 reduces the loading on the screen of V66. A differentiating network, formed C58 and R177 produces a sharp, positive like at the leading edge of the positive pulse, which causes V69 to fire. The thyratron, V69, then conducts, which causes its plate to go negative. This action causes a positive pulse appear at the cathode of V69.

The negative pulse initiated at the plate of 69 travels along the artificial delay line made of L1, C59, C60, and L2 and is reflected back to the plate of V69. This action quenches the thyratron, and conduction stops. The round trip time of the pulse through the artificial line is about 1 μs, and the range information at the thyratron cathode is a 1 μs pulse coincident in time with range strobe generation.

The negative and positive pips from the differentiator, C61 and R178, are fed to the grid of V67B. These pips are inverted by V67B, and its plate output is fed to the grid of the output cathode follower, V68A. The grid of V68A is below cutoff, and the negative pip is eliminated. The amount of voltage that holds the grid of V68A below cutoff is determined by the setting of R179.

An intensified spot is made to appear at the origin of the electronic cursor by coupling the negative cursor gate through C62 and V68B to the grid of V67B. Tube V67B then delivers the range strobe pip to the grid of the output cathode follower, V68A, at a time during the sweep that is determined by the range strobe control setting, R179.

VIDEO AMPLIFIER

The schematic diagram for a video amplifier is shown in figure 13-17. The video input line from the radar receiver may be terminated with a 75-ohm matching resistor, R180, by connecting a jumper. JU, between terminals 1 and 2. However, if several indicators are operated in a chain, R180 should be added in the last (most remote) indicator only, to avoid excessive loading of the line.

The video signal is fed to the grid of V70 through a long time constant network made up of C63 and R181. The video gain control, R182, controls the bias at the input grid of V70, and thereby adjusts the stage gain.

The amplifier that consists of V70, V71, and V72, plus a cathode follower in another section is designed for uniform gain over a wide frequency range. The coils, L3 and L4, are connected in series with resistors R183 and R184 to give high-frequency compensation. Otherwise, there would be reduced gain caused by a reduction in plate load impedance because of the shunting effect of the interelectrode capacitances and the distributed capacitance in the coil. The coils resonate with the interelectrode and distributed capacities to extend the high-frequency range of the amplifier, and have no effect on the low frequency response. Coils L5 and L6 form
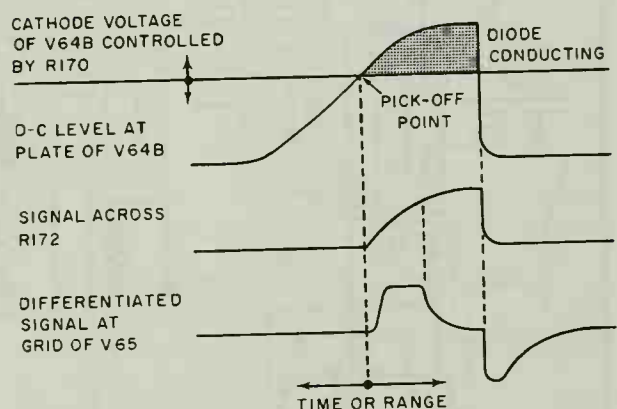
FAST SWEEP
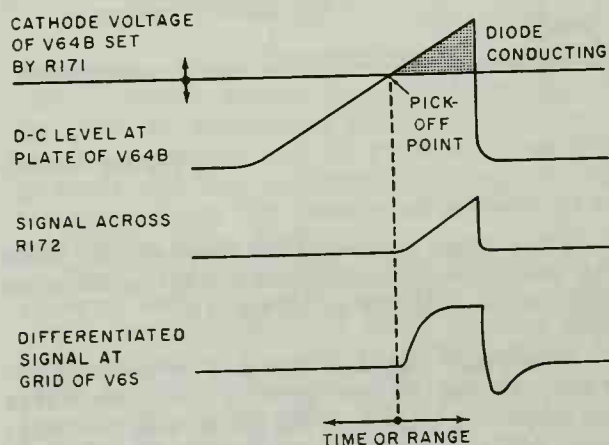
SLOW SWEEP

109.62

Figure 13-16.—Differentiated waveform.
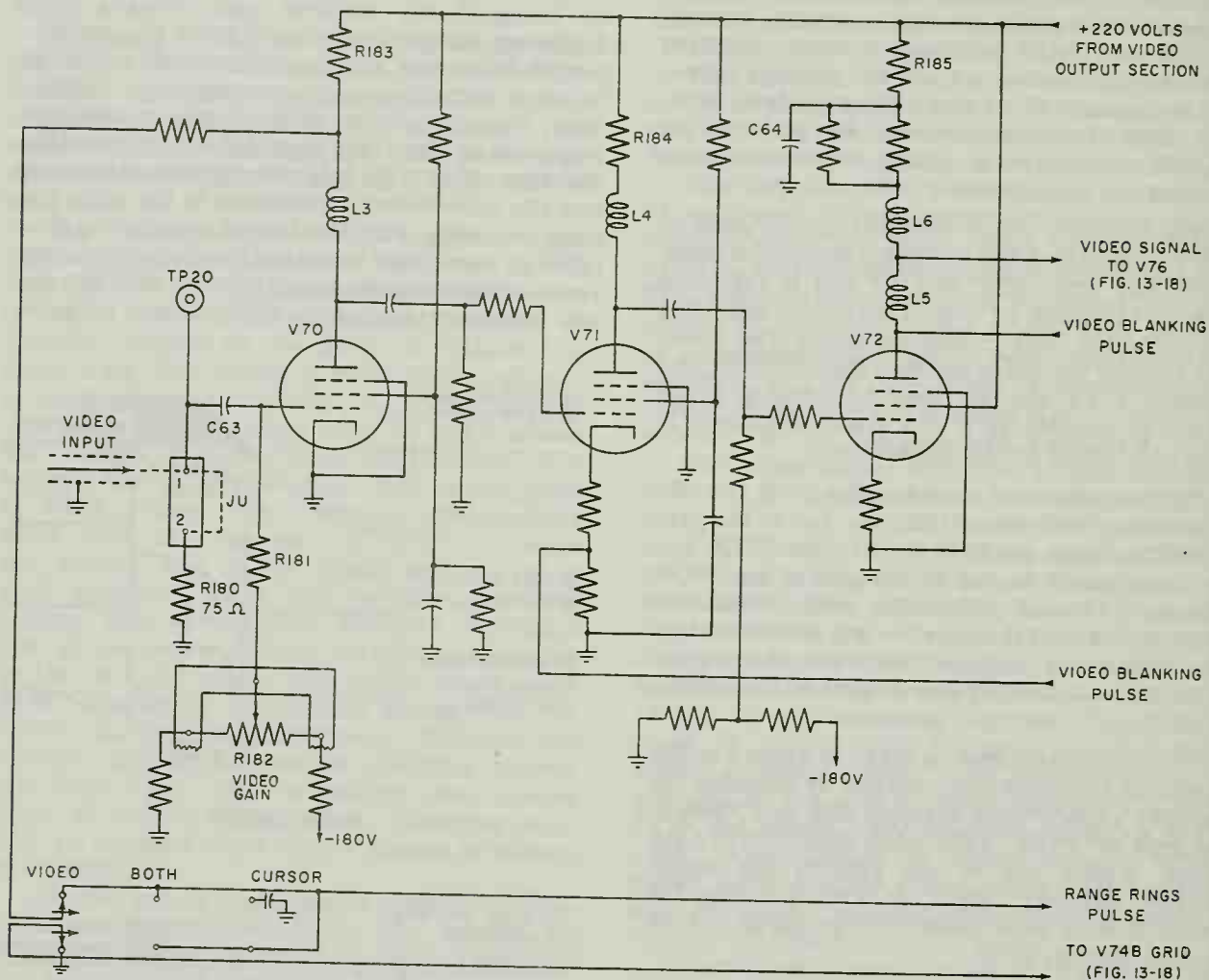
Figure 13-17.—Video amplifier.

109.63

a series-shunt compensating system. Coil L6 is the shunt compensation coil and gives the same high peaking effects as L3 and L4.

The video signal, which may be examined at TP20, is fed to the grid of V70 where it is inverted and mixed, at the plate, with the range ring pulse from another section. These combined signals are amplified and inverted again by V71. The amplification and inversion are again repeated in V72. A decoupling circuit, formed by R185 and C64, prevents feedback to V70 through the power supply.

VIDEO OUTPUT AND CURSOR UNBLANKING

The schematic diagram for the video output and cursor unblanking section is shown in figure 13-18.

The input stage, V73, is a DC restorer that clamps the DC level at the grid of V74A to the voltage set by the sweep intensity potentiometer, R186. If the voltage from R186 to the cathode of V73A is made more positive, the operating point of the grid of the output cathode follower, V74A, is raised. This action makes
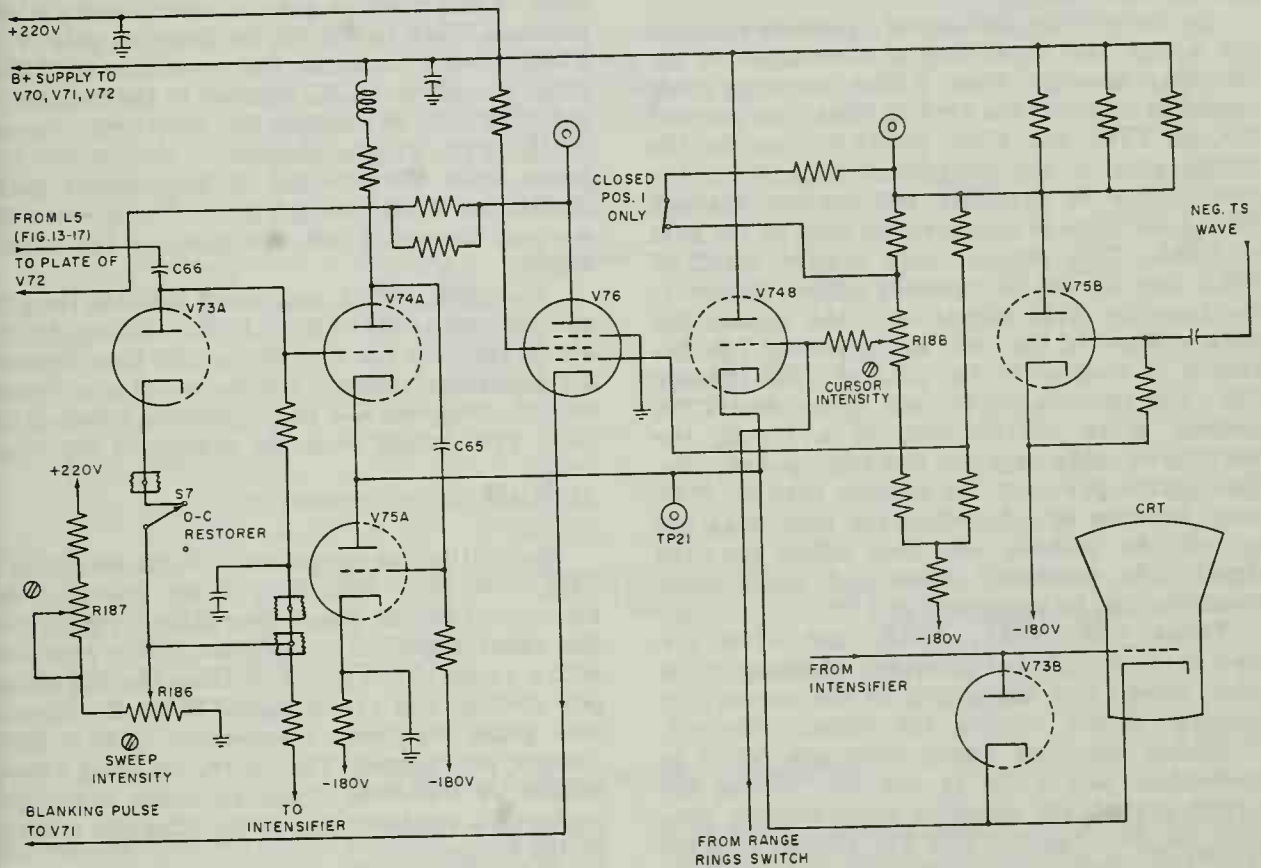
109.64

Figure 13-18.—Video output and cursor unblanking.

the signal to the CRT less negative and decreases the intensity of the sweep video display. The intensity of the sweep video display is increased by adjusting R186 so that the cathode of V73A is made more negative.

Under normal operating conditions, the DC restorer maintains the proper base line for video signals. Thus, it prevents the loss of weak signals because of a shift in base line when signals with large amplitude and high duty cycles are picked up. Under certain jamming conditions, it may be necessary to disable the DC restorer by opening S7. The potentiometer, R187, is used for automatic intensity compensation and will be discussed later.

The output cathode follower, V74A, feeds the negative video signal (examined at TP21) to the cathode of the CRT. Tube V75A is used as a cathode load for V74A to provide a low-output impedance. This provision assures that the cathode output follows the AC signal input to the grid. To explain the action of V75A, consider the cathode load of V74A as a resistor. In shunt with this resistance is a capacity that is equivalent to the distributed capacity of the cable going to the CRT.

When V74A conducts, the capacitance is charged with a positive polarity. The arrival of the leading edge of a large negative going video signal at the grid of V74A reduces the plate current of the tube. The cathode of V74A tries to follow the grid, but the capacity distributed in the cable to the CRT starts discharging through the cathode resistance. As a result, the leading edge of the cathode output signal lags the input signal in proportion to the rate of discharge of the capacity through the resistance. As the grid swings more negative,

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the cathode is prevented from following it by the action of the capacity, the bias increases, and the tube cuts off.

By using V75A instead of a cathode resistor the action just described is eliminated in the following manner. When a large negative going signal arrives at the grid of V74A, the current through V74A and V75A starts to drop, but the discharging to the distributed capacity in the cable tends to maintain the current through V75A, and thus to increase the bias on the grid of V74A. This action leads toward cutoff of V74A just as did the resistor action. However, the negative grid signal at V74A cuases the plate voltage of the tube to increase. The increase is coupled to the grid of V75A through C65. The positive pip through C65 makes V75A conduct more heavily and, as a result, the distributed cable capacity is discharged rapidly. This action prevents the cathode bias of V74A from building up, therefore the tube does not cut off. Its cathode can then follow the grid signal. The combined video and range rings waveform can be examined at TP21.

Tubes V76, V74A, V74B, and V75B are used in a circuit that provides blanking of the video sweep and unblanking of the cursor, or tracking strobe, during the cursor interval.

During the nine sweep intervals, V75B is conducting, and V74B is cut off. During the cursor period, the negative time-sharing gate from the cathode of V18B (fig 13-7A) cuts off V75B and produces a positive gate at its plate. This positive gate is also DC coupled to the grid of V74B through R188 the cursor intensity potentiometer.

The voltage drops at the plate of V76 because its grid has been made more positive by the positive gate and increases the current through the plate resistor. The negative output gate of V76 is DC coupled to the plate of V72 (fig. 13-17) and then through L5 and C66 (fig. 13-18) to the grid of V74A. This action drives the grid of V74A in a negative direction. To make sure V74A cuts off, the positive gate at the cathode of V76 is DC coupled to the cathode of V71 (fig. 13-17). The resulting positive gate at the plate of V71 is amplified and inverted by V72 and appears at the grid of V74A (fig 13-18) at the same time as the negative gate from the plate of V76 and thus assures that V74A will be cut off.

At the same time the above action occurs, a positive gate at the grid of V74B causes the tube to conduct. When this happens, a positive

gate is produced at the cathode of V74B and is coupled to the cathode output of V74A. The DC level of this positive gate is controlled by R188. Because V74A is cut off, the positive gate from V74B, alone, produces the waveform at TP21. This waveform is DC coupled to the cathode of the CRT and intensifies the electronic cursor on the PPI. The brightness of the cursor depends upon the strength of the positive pulse applied to V74B through R188. The greater the current through V74B, the dimmer the cursor display.

The diode, V73B, connected between the grid and cathode leads of the CRT is a safety device which prevents the grid of the CRT from becoming excessively positive in the event of a circuit failure. Otherwise, a large positive swing of the CRT grid might burn the screen of the tube.

AUTOMATIC INTENSIFIER

Normally, the birghtness of the sweep on a CRT will vary according to the speed of the sweep and to the pulse repetition frequency of the radar signal. The indicators are provided with a range of sweep speeds from 4 to 300 miles per radius, and are designed to handle signals with pulse repetition frequencies of 60 to 3000 sweeps per second. Therefore, changing sweep speeds or switching from one radar to another requires a readjustment of the intensity control by the operator.

To eliminate this condition, the automatic intensifier shown in figure 13-19 is used.

The automatic intensifier consists of V77A, V77B, V78A, and V78B and maintains the background brightness substantially constant for all sweep speeds and repetition rates without manual adjustment by the operator. The automatic intensity circuit does not affect the video gain.

In operation, a negative gate from the gate generator cuts off the normally conducting V77A, and a positive gate is produced at the plate of V77A, which is directly connected to the control grid of the CRT. Therefore, the positive gate turns on the display. The positive pulse from V77A is also coupled to the grid of V77B by the differentiating circuit made up of C67, R189, and R190. The positive pip of the differentiated waveform causes the normally cut off V77B to conduct. The negative output pip from the plate of V77B is integrated by C68 and R191. Any AC component in the pip is bypassed to ground by C68 and only a negative DC voltage is applied to the grid of V78A. The level of the negative
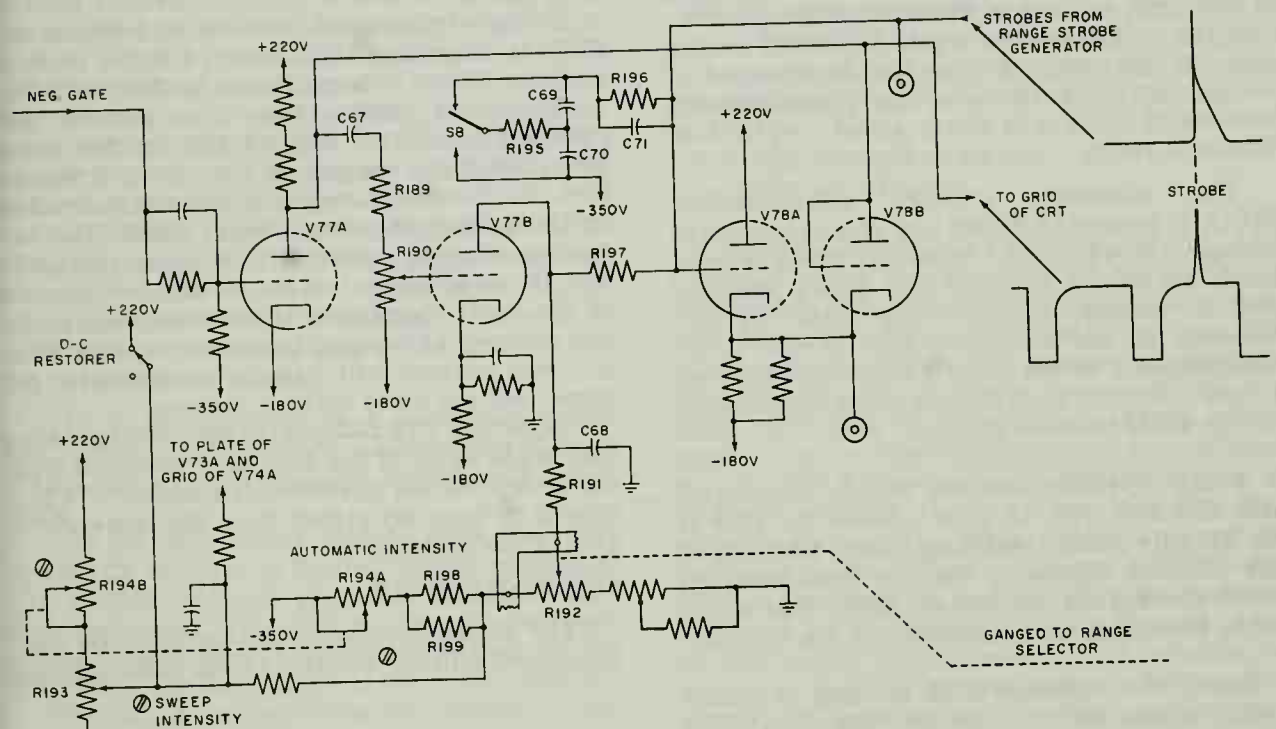
109.65

Figure 13-19.—Automatic intensifier.

DC voltage applied to the grid of V78A is determined by the pulse repetition frequency of the radar because, the higher the PRF, the greater the number of negative charges collecting on C68. Consequently, the greater the total negative bias on the grid of V78A.

The potentiometer, R190, sets the amplitude of the individual negative pulses that charge C68 and thereby determines the rate of change in DC level at the grid of V78A with the repetition rate. The negative DC level from the cathode of V78A is directly coupled to the cathode of V78B. Note that the grid and plate of V78B are tied together to form a diode. The voltage level at the cathode of V78B determines the plate voltage at which it will conduct. The amplitude of the positive gate at the plate of V77A is limited to the level at which V78B conducts.

Thus, an increase in the repetition rate of the positive gate that would normally increase the brightness of the CRT simultaneously produces a proportional increase in negative bias on the cathode of V78A. The increased bias acts to reduce the amplitude of the intensifier voltage and to prevent an increase in brightness. There-fore, the brightness level of the CRT is relatively unaffected by changes in the PRF of the radars.

The intensity of the video display must not vary as the range is changed. The fast sweep used for the shorter ranges would not normally appear as bright on the PPI as the slow sweep of the long ranges. To eliminate this condition, the range selector control on the panel is ganged to potentiometer R192, which controls the bias voltage for V78A as well as the plate voltage for V77B. Therefore, the range selector control moves the potentiometer arm to provide additional bias automatically as the range becomes longer or less bias as the range shortens.

The potentiometer, R193, provides a means of adjusting the CRT intensity voltage. The setting of this control determines, in part, the amount of negative bias available across R192. A more negative setting of the sweep intensity control, R193, lowers the operating point at the grid of the video output cathode follower, V74A, (fig. 13-18). As a result of this action, the signal to the cathode of the CRT is more negative, and the intensity of the video pattern on the CRT is increased. At the same time, the more negative

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setting of the sweep intensity control increases the range of the negative bias across R192. Thus, at different settings of the sweep intensity control the potentiometer arm of R192 will always pick off the correct bias that is required to compensate for variations of the sweep intensity because of changes in sweep speed over the 4-to 300-mile range.

The automatic intensity potentiometer, R194A is ganged to R194B and also controls the voltage across R192. However, these adjustments are preset and used only during installation or maintenance to compensate for differences in the amount of bias across R192. Readjustment of the controls is required when a new CRT is installed or when components change value because of age.

Relay operated switch, S8, in connection with C69 and C70, is used to blank the CRT at the 34-mile range switching point. Blanking at this point is necessary because transient currents develop (in the sweep chain) that would cause distortion and markings on the display.

Protective resistor R195 is used to reduce arcing at the switch contacts. When the relay operates S8 at 34 miles, C69 is connected to -350 volts. This sudden drop in voltage is coupled through R196 in parallel with C71 to the grid of V78A, and causes the cathode voltage of V78A to lower enough to cut off the CRT. Capacitor C69 then charges through R196, R197, R191, R198, R199, and R194 in approximately one second, so that V78A can conduct again and reestablish the normal bias on the CRT grid. The CRT is blanked during the one second that C69 is charging in order that transient currents, because of switching, do not appear on the display.

When the relay operates S8, C70 starts from zero charge and follows the same cycle as described for C69.

During the cursor interval, or the interval of every other sweep, according to the position of the switch in the range strobe generator, positive strobe pips are delivered from the range strobe generator to the grid of V78A. These positive pips produce sharp positive peaks on the positive gate from V77A, and also produce either the range strobe on the electronic cursor or the range strobe ring centered about the origin of the video sweep.

FOCUS

Proper focusing of the electron beam in the CRT is maintained by using a focus control circuit, shown schematically in figure 13-20.

The focus control tube is a pentode that provides a constant current through the focus coil. The plate current of the tube and, therefore, the current through the coil are controlled by the setting of potentiometer R200. The advantage of using a pentode in the focus circuit is that the tube current is essentially independent of the load. Therefore, if the resistance of the coil changes because of temperature variations, the coil current will remain substantially the same, and the focus will not change.

Capacitor C73 reduces the peak voltage swing that would occur if the tube were removed from its socket during operation. Capacitor C72 is a bypass to keep any pickup from the cable off the grid of the pentode.

SERVOAMPLIFIER

The sweep resolver rotation must be synchronized with the rotation of the radar antenna
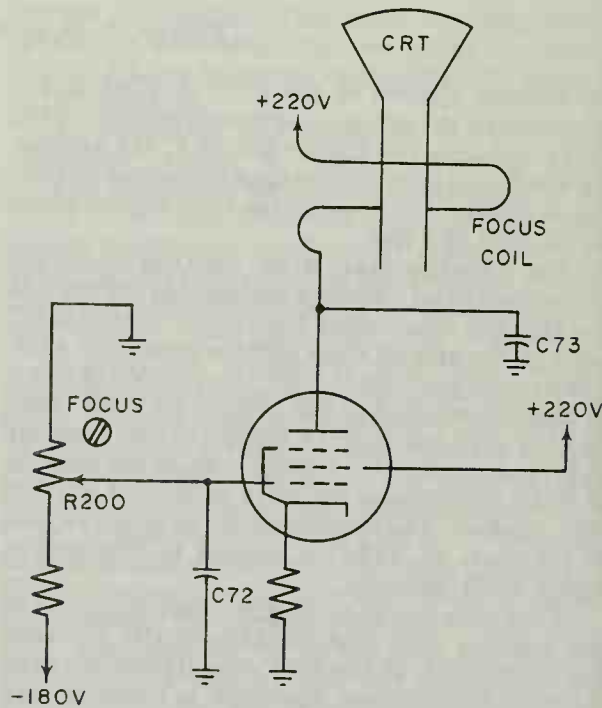
109.66

Figure 13-20.—Focus circuit.

being used. In one type of radar, a one-speed synchro followup system is used to synchronize the sweep and antenna rotation. The complete control (fig. 13-21) is called a servo loop, which consists of four major parts: (1) synchro bearing information from the radar, which is the input control signal, (2) a 1CT (control transformer), (3) a servoamplifier and (4) a drive motor and the necessary gears.

The bearing input may be north-stabilized by using correction voltages from a gyro compass. If no correction has been applied, bearing information relative to the heading of the ship or AEW plane, as the case may be, is obtained and fed to the servoamplifier.

The servoamplifier is a conventional unit and includes a bridged-T, antihunt network; a phase-shifting network; and a push-pull output stage.

The 60-cycle error voltage from the rotor of the 1CT synchro is applied across capacitor, C74, to filter out high-frequency transient voltages that result from movements of the synchro rotor. The bridged-T network consists of C75, C76, R201, and R202, which is an antihunt control.

The output voltage from the bridged-T network represents the sum of voltages proportional to the error signal and its derivative. The derivative voltage is proportional to the rate of change of the error signal voltage and acts to prevent hunting, thereby stabilizing the system.

The stabilizing action of the bridged-T network takes effect as the system is approaching a null after an error indication. The derivative signal starts slowing the motor down with dynamic braking before the null is reached. Braking in this manner reduces the overshoot so that the system will stabilize quickly. This type of damping is called error rate damping, and it permits the use of higher servo gain and greater freedom from oscillation than would otherwise be possible.

The bridged-T network output is fed to V79A. The output of this tube is fed to a phase-shift network made up of R203, R204, C77, and C78. This network produces approximately 90 degrees phase shaft in order to obtain maximum torque in the split phase drive motor.

The output of the phase-shift network is applied to the grid of V79B for amplification. The plate output of V79B is applied through C79 to R205, which is the servo gain control. A proportional part of the signal is applied to the amplifier and phase inverter V80A and V80B.

The output of V80B supplies the push-pull input signal for V81 and V82.

Negative feedback from plate to grid of each tube in the output stage, through R206 and R207, provides additional damping and keeps the motor from operating with no error signal.

JOYSTICK OFF-CENTERING CIRCUITS AN/SPA-8C

Joystick operation provided by the AN/SPA-8C, gives the operator of the indicator greater speed in off-centering the cursor origin and, in most cases, allows him to perform this function in a more natural manner. In addition, joystick operation requires the use of only one hand, permitting the other to be used for a strobe positioning or other operations. It is important to note the following three characteristics of CURSOR ORIGIN joystick operation that distinguish it from the other types of operation available with Indicator Group AN/SPA-8C: (1) it is available for general-purpose operation only, not AEW operation; (2) the amount of the cursor-origin displacement is not indicated on counters; and (3) the displacement information cannot be transmitted to other equipments. Time-shared voltage output of the electronic switch provides zero voltage across the potentiometers during the sweep period, thus permitting off-centering during the cursor period only. The shaft of R-1852 is geared indirectly thru electro mechanical storage clutches to the RANGE SELECTOR knob of the top panel. Depressing CURSOR ORIGIN joystick, de-energizes L-1851, allowing the wipers of R-1852 to select voltages which depend upon the range setting. The two voltages are equal in magnitude but opposite in polarity.

Motion of the CURSOR ORIGIN joystick controls potentiometers that apply the correct amount of cursor off-centering voltage to the E-W and N-S Yoke Drivers. There are two sets of these potentiometers, one set for coarse off-centering control and one set for fine off-centering control. When the joystick handle is depressed, the coarse potentiometers are activated and the joystick becomes a coarse control. In this condition, the cursor origin is capable of being displaced three radii in any direction from the sweep origin. When the handle is released the coarse potentiometers are held at their settings and the fine potentiometers are activated. Movement of the joystick in this condition results in fine control of off-centering
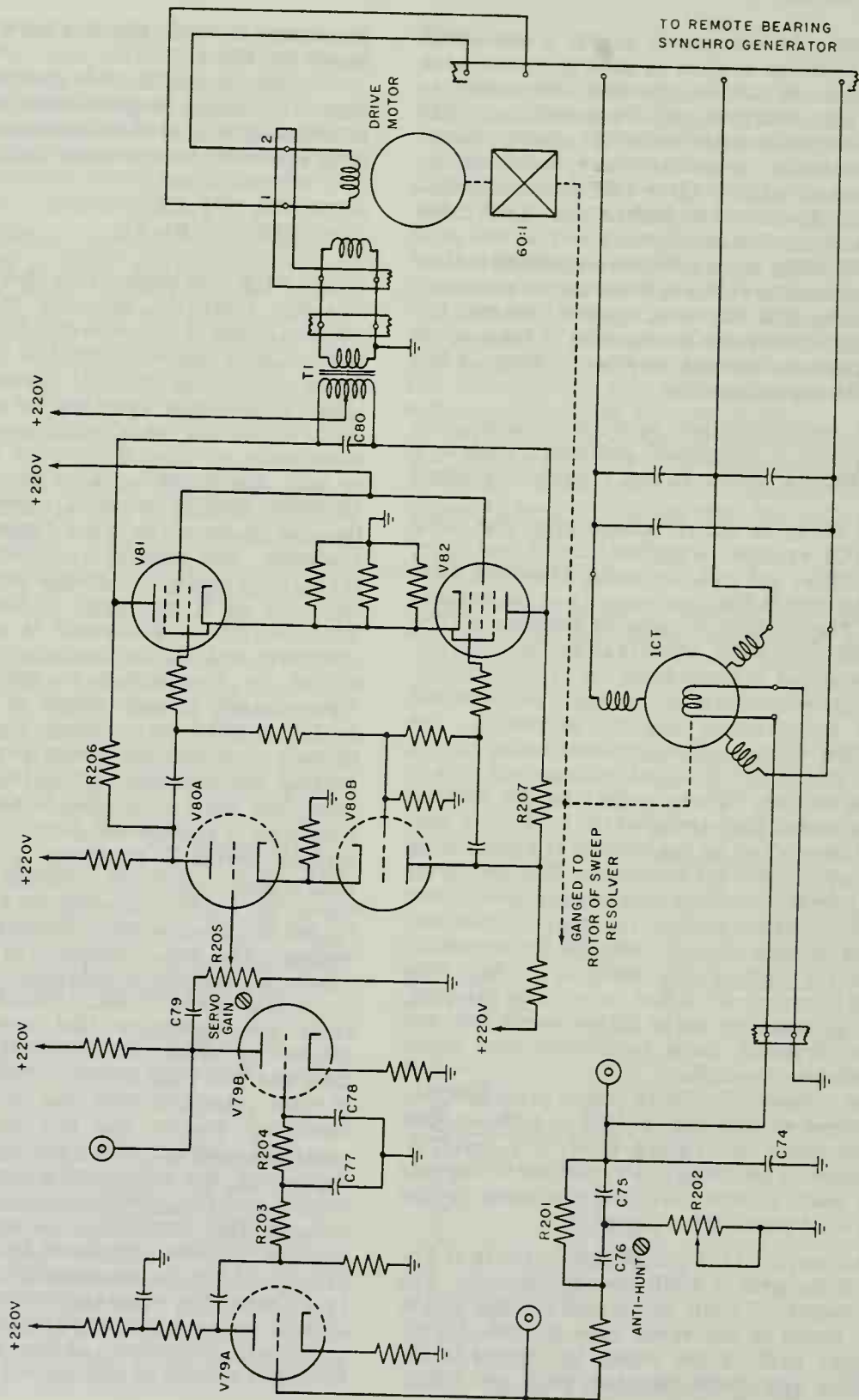
Figure 13-21. —Servoamplifier section.
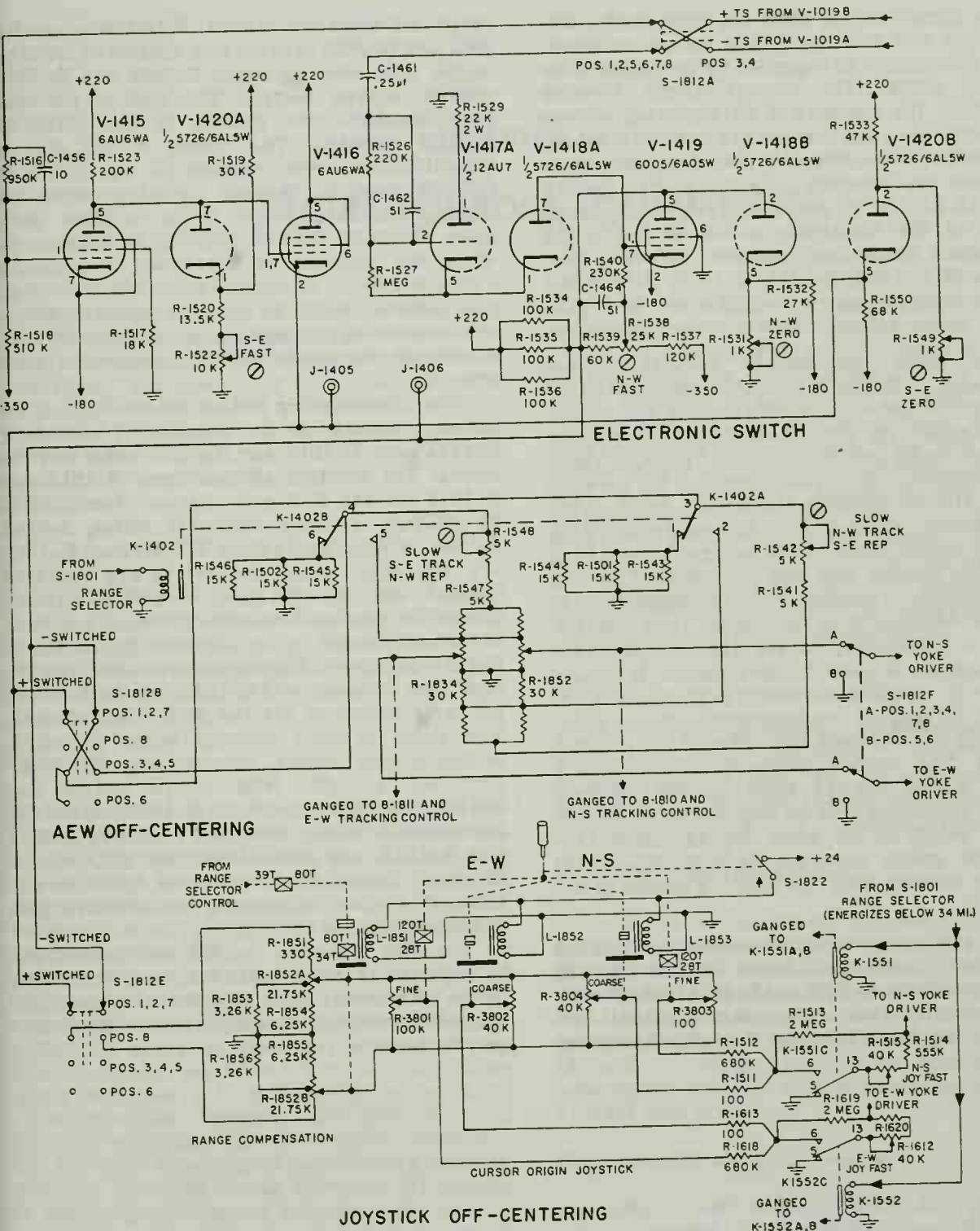
109.68

Figure 13-22.—Electronic switch, AEW, and joystick off-centering circuits, AN/SPA-8C.

of the cursor origin about the point on the display that was selected in the coarse condition. When the display is expanded by operation of the RANGE SELECTOR control during joystick operation, the amounts of off-centering voltage applied to the yoke drivers are reduced to maintain the above mentioned 3-radius displacement.

When the function selector switch S-1812 (Fig. 13-22), is in position 8, the time shared output of the electronic switch is fed to the non-linear range-compensation network, consisting of R-1853, R-1854, R-1855, R-1856 and tapped potentiometers R-1852A and R-1853B, in the range selector (sweep speed) assembly.

The range-compensation network applies decreasing voltages to the potentiometers as the display is expanded, thereby keeping the cursor displacement (in inches) substantially independent of range selector setting. Without range compensation, a given movement of the joystick would always displace the cursor through the same range (in miles) and, as a consequence the cursor origin might move several inches in an expanded display and only a fraction of an inch in a contracted display. Range compensation is applied because it is desirable, from the operator's standpoint, to see approximately the same amount of cursor displacement, in inches, relative to joystick movement regardless of the degree of display magnification. Range compensation affects only the action of the joystick control. The cursor origin will remain on a target during display magnification, because display expansion acts on both the target and the cursor origin to the same degree, and R-1852 does not rotate when the RANGE SELECTOR knob is rotated until the joystick handle is depressed.

The two range-compensated voltages are fed to opposite ends of CURSOR ORIGIN joystick potentiomenters R-3803 and R-3804 in the North-South cursor-origin off-centering circuit and R-3801 and R-3802 in the East-West cursor-origin off-centering circuit. Potentiometers R-3801 and R-3803 provide fine control of cursor-origin off-centering while R-3802 and R-3804 provide coarse control. The shaft of the fine potentiometers are geared to the CURSOR ORIGIN joystick. The shafts of the coarse potentiometers are coupled to the CURSOR ORIGIN joystick through electromechanical storage clutches which allow rotation only when CURSOR ORIGIN joystick is depressed. When the joystick is depressed it becomes a coarse control of cursor-origin off-centering. The wipers of all the joystick potentiometers are connected, through summing resistor networks, to the inputs of their respective yoke drivers.

The off-centering voltage pulses to the yoke drivers appear at the junction of resistors R-1511 and R-1512 for the N-S yoke drivers and at the junction of resistors R-1613 and R-1618 for the E-W yoke driver. For RANGE SELECTOR settings above 34 miles, switch S-1801 is open and relays K-1551 and K-1552 are energized since the relays are operated through the NO contacts of switch S-1801. Therefore, the input voltages to the yoke drivers are divided down by the networks consisting of R-1513 and the N-S input grid resistors for the N-S yoke drivers, and R-1619 and the E-W input grid resistors for the E-W yoke drivers. For range selector settings below 34 miles, switch S-1801 closes, deenergizing relays K-1551 and K-1552. With relays K-1551 and K-1552 energized, resistors R-1620 and R-1612 are shunted with R-1619, and resistors R-1514 and R-1515 are shunted with R-1513, so that the input impedance to the yoke driver is decreased, thereby increasing the effective gain of the yoke driver circuit.

Variable resistors R-1612 and R-1515 can be adjusted to compensate for small variations in the yoke driver input circuits so that the ratio of gains above and below the 34-mile switch point may be selected precisely.

CHAPTER 14

SPECIAL PURPOSE EQUIPMENTS

This chapter gives a brief overview of selected special purpose equipment used in the Navy. Equipment used in electronic warfare is described, and general information on some of the latest electronic aids to navigation is presented. The chapter closes with a discussion of the latest developments in satellite communications and implications for its use in the future.

ELECTRONIC WARFARE

Electronic Warfare is the science of intercepting the enemy and reducing the effectiveness of his electronic aids, by the utilization of electronic and nonelectronic equipment. It will be treated here under the two general headings: Passive and Active.

PASSIVE

As it pertains to Electronic Warfare, the term "passive" implies the state of being inactive but receptive. This means that a ship at sea which is electronically silent (no radio, radar, or sonar transmissions), with radio receivers on and listening, is engaged in passive Electronic Warfare.

For this purpose, any Navy radio receiver could be used, but in order for the countermeasures to be effective against an enemy radar, as many as possible of the following characteristics should be known about his enemy radar facilities: (1) the frequency, pulse width, pulse-repetition frequency, and peak power of the transmissions; (2) the receiver bandwidth and the time constants of the receiver coupling circuits; (3) antijamming features; (4) amount of shielding; (5) type of indicator; (6) antenna beamwidth; (7) types of scan; and (8) use of the radar. Some of this information is obtained by analyzing the enemy transmission, and some

may be obtained by examining captured equipment.

Special equipment has been developed for use in analyzing r-f transmissions. This equipment includes SEARCH RECEIVERS, which search the various frequency bands for the different types of emissions. This type of receiver should cover a wide band of frequencies. Also used with the search receiver is a PANORAMIC ADAPTER and a PULSE ANALYZER. The panoramic adapter measures the frequency, strength, and type of modulation of a transmission in a selected band of frequencies. The pulse analyzer measures the pulse rate and width of these transmissions.

ACTIVE

The act of reducing the effectiveness of the enemy's electronic aids is the "active" portion of Electronic Warfare. For the purpose of this discussion it will be broken down into electronic and nonelectronic countermeasures.

Electronic Countermeasures

In order to use countermeasures most effectively against enemy communication systems, the following information is needed; (1) the frequency of transmissions, (2) type of modulation, and (3) receiver bandwidth. If the foregoing information is known, sucessful jamming of enemy transmissions can be accomplished with special type transmitters designed specifically for this purpose.

Nonelectronic Countermeasures

Units called CORNER REFLECTORS are used to present strong echoes to enemy radars. When suitably placed, corner reflectors produce strong echoes that appear to the enemy radar

operator as a large naval force. Corner reflectors have other valuable uses. For example, they can be used on liferafts to assist friendly search radars to locate them.

ROPE is the name used for long streamers of aluminum foil. This foil, cut in lengths of about 400 ft, is dropped by aircraft within range of a enemy radar. The twisting and turning of the foil as it falls presents many different effective lengths to the enemy radar. Some of these lengths are highly resonant at the frequency of the radar, and therefore appear as strong target signals.

Window is the name used for short strips of aluminum foil dropped by aircraft to confuse the enemy. The foil is cut to slightly different lengths so that it causes strong reflected signals at the frequency of the enemy radar. The strips are packaged and dropped over enemy territory. While fluttering to the ground, the foil presents a multitude of targets to the enemy radars. His search and tracking radars follow the strong echoes presented by the window and find it difficult to track on the lesser echoes presented by the aircraft.

DECOYS consist of a wide variety of devices. Some of the most effective are balloons towing strips of aluminum foil. These strips, which vary in length, present strong reflections over a fairly wide band of frequencies. Aircraft may also tow streamers of foil as decoys. Corner reflectors might also be considered decoys.

Antijamming measures are used to reduce the effect of enemy jamming on our own equipment. In receivers some of the most important antijamming devices are special filters that pass only the most important parts of echo signals, thus rejecting as much of the jamming signal as possible. In the transmitters, a great many radar equipments have tunable magnetrons whose frequency may be varied at intervals to prevent an enemy jamming transmitter from locking on the radar signal.

FUNDAMENTALS OF INERTIAL NAVIGATION

Navigation is defined as the art or science of conducting a vehicle from one position on the earth to another position. There are two methods of navigation. One is position fixing by measuring position relative to some known object. A common example is celestial navigation. Loran is another example of navigation

by periodic position fixes. The other method, dead reckoning, measures speed and heading and, using these measurements, computes position change from an initial position fix. The most common example of this form of navigation is the ship navigator's position plot. One of the oldest automatic navigation systems is the dead reckoning analyzer, which takes its speed from the ship's log and its heading from the ship's gyrocompass to compute latitude and longitude.

An inertial navigation system is a self-contained, refined form of dead reckoning that continuously computes latitude and longitude by accurately sensing the accelerations of the ship with respect to the earth's surface.

Before learning how an inertial navigator operates, it is necessary to review a few fundamentals of physics and motion. The following paragrphs discuss the essential background material required for this purpose.

FUNDAMENTALS OF INTEGRATION

Since the process of integration is so frequently used in navigation systems, a simplified explanation is given. It is used here to explain the relationship of acceleration to velocity and distance traveled. In inertial navigation, distance traveled is calculated from measured acceleration by first integrating acceleration with respect to time to find velocity, and then integrating velocity with respect to time to find distance traveled.

Velocity is defined as a rate of change of distances with respect to time, $V = \frac{ds}{dt}$. The familiar unit of miles/hour is the dimension of velocity. Acceleration is a physical phenomenon equal to the rate of change of velocity with respect to time, $a = \frac{dv}{dt}$. The unit of acceleration curve is miles/hour^2. Because the area under the acceleration curve is proportional to velocity, if the mathematical process of integration is performed on the quantity acceleration, the result is velocity. Likewise, because the area under the velocity curve is proportional to distance if the mathematical process of integration is performed on the quantity velocity, the result is distance. Figure 14-1 illustrates the relationships of acceleration, velocity and distance.

In order to illustrate these relationships clearly, constant values of accelerations are chosen for each of the first four intervals (time 0 to time 4). Integrating the constant acceleration for time 0 to time 1 gives a
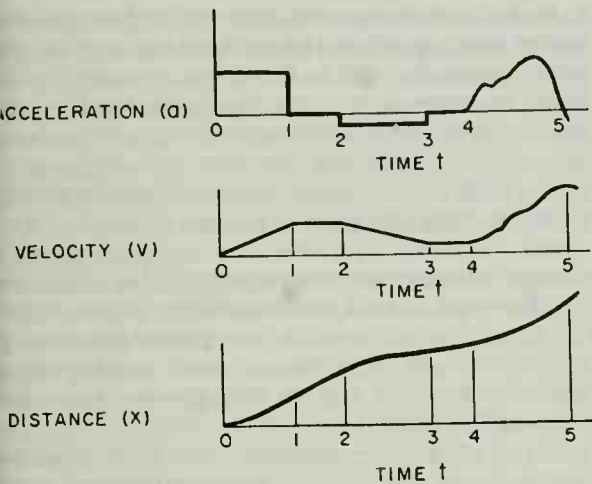
162.49

Figure 14-1.—Integration of acceleration to obtain velocity and distance.

velocity (V) which increases linearly. Velocity at the end of the first interval is actually computed by measuring the area under the acceleration curve (time 0 to time 1).

Recall that the fundamental formula for velocity is $V = at$; where $V = $ Velocity, $a = $ acceleration, and $t = $ time. The process of integration subdivides each interval shown in figure 14-1 into an infinite number of tiny intervals. Integration is, then, the process of adding, or summing, an infinite number of incremental areas under a curve to obtain the total area. The velocity curve, therefore, is a plot of subtotals of the incremental areas under the acceleration curve.

In the second period (time 1 to time 2), the acceleration is zero, and the velocity remains constant. In the third period, there is a constant negative acceleration which reduces the velocity subtotals at a constant rate. In the fourth period, there is no acceleration and no change in velocity. In the fifth period, a random acceleration and its approximate integral, the velocity, are shown.

The distance curve is a plot of subtotals of the incremental areas under the velocity curve.

An example of integration is a d-c motor driving a counter. If a voltage proportional to velocity is present, it could be used to drive the motor so that the speed of the motor would be proportional to velocity; and the number of revolutions or the counter reading would indicate distance traveled. Thus, velocity would be integrated to find distance.

NEWTON'S LAWS

In our definition of inertial navigation it states that the inertial navigator senses accelerations. The important thing to consider at this point is how to sense the quantity of acceleration. Newton's three laws of motion provide the answer to this problem.

Newton's first law states that a body remains in motion at a constant velocity unless acted upon by a resultant force. Constant velocity means both direction and magnitude, therefore velocity is a vector.

Newton's second law describes the magnitude of the acceleration as $F = M_s A_s$ where;

A_s = the acceleration of the ship

M_s = mass of the ship

F = vector sum of the forces

Two quantities are unknown, A_s and F. The quantity F cannot be directly determined because of the many varying parameters that enter into its solution.

Newton's third law states that for every reaction there must be an equal and opposite reaction.

For this application Newton's three laws result in the statement most familiar to us: a body at rest remains at rest unless acted upon by some external force and a body in motion remains in motion unless acted upon by some external force.

ACCELEROMETERS AND ACCELERATIONS

An accelerometer may be defined as a device which measures accelerations (changes in speed or direction). Its output is usually in the form of a voltage proportional to the acceleration to which it is subjected. Accelerometers are utilized in inertial navigation to provide information to the gyroscopes which in turn provide a stable platform.

Basically, an accelerometer consists of a mass suspended by constraining springs within a case. When the case and pendulum are traveling at a constant velocity, the mass remains in its center position as shown in figure 14-2, A. However, when the frame is accelerated (fig. 14-2, B), the mass will appear to move in the direction opposite to that of the accelerating force. Actually the mass lags behind since

497
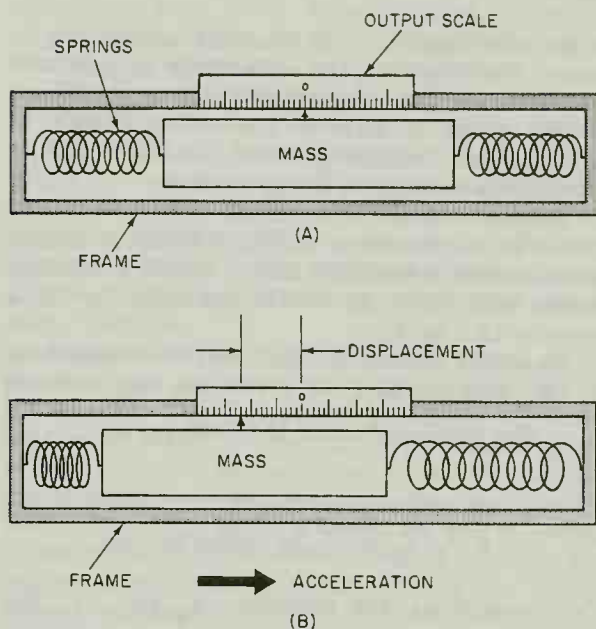
55.43(162)
Figure 14-2.—Basic Accelerometer.

the acceleration is applied to the frame, not the mass. The mass displaces enough for the constraining springs to apply a force proportional to the acceleration. The mass then moves with the frame maintaining its constant displacement. When the acceleration is removed from the frame, the constraining springs cause the mass to move (with respect to the frame or case) back to the neutral position. Thus, a body at rest or a body at constant velocity (zero acceleration) causes no displacement, providing the accelerometer is held horizontal.

In actual practice, the mass is held in its null or zero position with a calibrated spring or springs. To further dampen oscillations, the case is normally filled with a fluid. Since the deflection of the mass is proportional to the accelerating force, the amount of deflection (hence the accelerating force) is measured and through integration, velocity and distance can be determined.

The basic accelerometer demonstrates the principle of operation of inertial accelerometers. It must be kept in mind that these inertial accelerometers are sensitive to more than just the accelerations with respect to the earth. There are two types of forces that can cause a body to move: acceleration forces and field forces.

It is important to note that an accelerometer cannot distinguish between these two forces. The most common field force is gravity. This force is represented by the formula $(f = mg)$. The term (g) represents gravitational acceleration, 32.17 ft per sec per sec; the acceleration with which a body falls freely toward the earth's surface, neglecting air friction. If the accelerometer is tilted or placed on end, the force of gravity causes the mass to move in the same way as does an actual acceleration, even though the frame is at rest. A compensation must be made for these field forces, such as gravity, so that the quantity left is the acceleration with respect to the earth.

In order to compute the true distance traveled, we must sense the acceleration of the ship parallel to the reference plane of position. The accelerometer senses only components of acceleration parallel to its sensitive axis. Therefore, it must be maintained continuously parallel to this surface, independent of the pitching, roll, or yaw of the ship. This is accomplished by using more than one accelerometer and mounting them on gimbals so that the pivot axis of these gimbals are parallel to the motion caused by pitch, roll, or yaw. A device could be made that would sense the pitch, roll, and yaw of the ship and develop a signal that would be used to drive the associated gimbal with respect to the ship in a direction opposite to the disturbing motion, thereby maintaining the accelerometer's sensitive axis parallel to the reference plane of travel. In order to fully understand a device such as this, it will be necessary to learn a few fundamentals of gyroscopes and take a further look at inertial motions.

GYROSCOPE FUNDAMENTALS

A gyroscope, commonly called a gyro, is defined as a solid mass rotating about an axis contained within the mass. Usually the mass is symmetrically distributed about the axis which is called the spin axis. It is the rotation or spin of the mass which is the important feature of the gyroscope because it gives to the gyro characteristics or properties which distinguish it from other solid bodies. These characteristics are called rigidity and precession and are the result of fundamental laws of motion and inertia.

The principle of conservation of angular momentum (Newton's first law), states that a system will maintain its angular momentum if no external forces are applied to it. Thus the

tor (mass) of a gyro must maintain a constant
ngular momentum about its spin axis if no
ternal forces are applied. Both the amplitude
d direction of the angular momentum must
e conserved. The spin axis, therefore, tends
maintain the same direction in inertial
pace. This feature of a gyro is known as gyro-
cope rigidity and is a fundamental property
f gyroscopes.

Because of gyroscopic rigidity, moving the
uter supporting frame has no effect on the
irection in which the axle of a spinning gyro-
cope points. To change this direction, it is
ecessary to apply a force to the gyroscope
otor or its axle. A horizontal force (F) on one
nd of the axle (fig. 14-3) produces a turning
ffect which attempts to tilt the gyroscope about
e axis ab. If the rotor were not spinning,
e axle would tilt in response to the applied
orce. When the rotor is spinning, however, its
igidity causes it to resist any attempt to tilt
e axle about ab, and instead it turns about
e axis cd in the direction of arrow e. Con-
ersely, a force attempting to turn the gyro-
cope about the axis cd is similarly resisted
nd results in a tilt about the axis ab. If the
orce is applied in direction H, rotation will
ccur in the direction of arrow g. This rotation
f gyroscope about an axis at right angles to
he axis about which a force is exerted is

called precession. Precession takes place when-
ever any force tends to tilt the axle of a spinning
gyroscope rotor. The precession caused by the
applied force is always about an axis at right
angles to the axis about which the force is
applied.

As explained previously, a free-spinning
gyroscope can be moved in any direction with-
out altering the angle of its plane of rotation.
If this free-spinning gyroscope is placed on
the earth's surface at the equator with its
spinning axis horizontal and aligned east and
west an observer in space below the south
pole would note that the earth rotates clock-
wise from west to east and carries the gyro-
scope along. As the earth rotates, rigidity
of plane keeps the gyroscope wheel fixed in
space and rotating in the same plane at all
times. Figure 14-4A shows how this gyroscope
would appear to the observer in space. As-
sume that the gyroscope is set spinning at
0000 hours with its spinning axis aligned east
and west and parallel to the earth's surface.
At 0600, 6 hours after the gyroscope was
started, the scope is aligned with the original
starting position. At 1200 the earth has rotated
180° while the gyroscope retains its original
position. At 1800 the earth has rotated 270°
while the gyroscope retains its original position.
At 0000 the earth has rotated 360° and the
gyroscope is back in its original position.

This rigidity of plane appears quite dif-
ferent to an observer on the earth's surface.
As the earth rotates, the observer moves
with it and the gyroscope wheel appears to
rotate about its horizontal axis. Figure 14-4B
shows how this gyroscope appears to the ob-
server on the earth's surface beside the gyro-
scope. Assume that the gyroscope is set spinning
at 0000 hours with its spinning axis horizontal
and pointing west toward the observer. At
0600, 6 hours after the gyroscope was started,
the earth has rotated 90° and the gyroscope
axle apparently has tilted. To the observer, the
axle is horizontal again, but the axle points
away from the observer. At 1800, the gyroscope
axle is again vertical and points straight up.
At 0000, the earth has rotated 360° and the
gyroscope axle is back in its original position.

The rotation of the gyroscope axle as seen
by the observer on the earth's surface is known
as "apparent precession". If it is desired
to have the gyro maintain an orientation
referenced to earth, it must be precessed
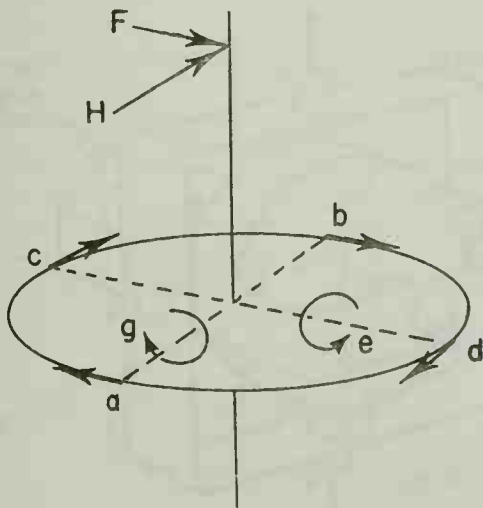
77.195

Figure 14-3.—Spinning rotor with a vertical
axis.
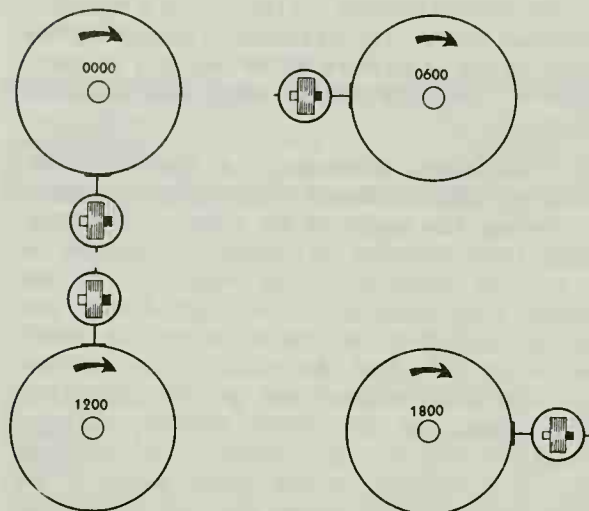
A
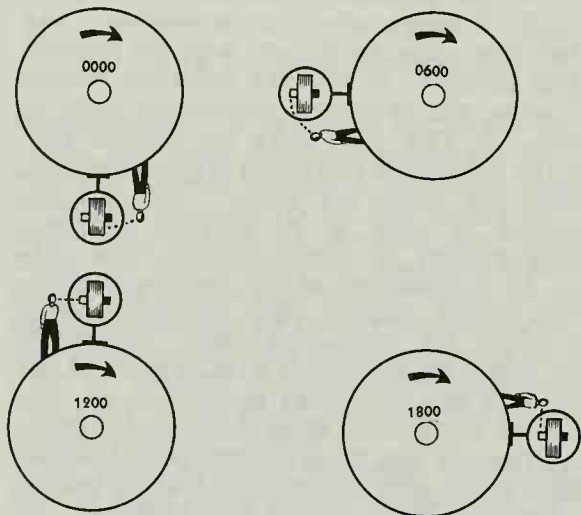
B

12.144

Figure 14-4.—Apparent precession.

(torqued) in such a manner as to remove the apparent precession.

Gyro Supports

The gyro described above is said to have two degrees of freedom. It consists of an axially symmetrical rotor supported in a case

in such a way that one axis is fixed relative to the case and the spin axis itself is free to assume any angular orientation relative to the case. Supporting the rotor within the case and providing the angular freedom is a gimbal system.

An example of such an arrangement is shown in figure 14-5.

The gyro rotor is free to precess about either of the two supporting gimbal axes (ab or cd), depending upon where the force is applied.

For application of the gyroscope in an inertial navigator, it is better to have a gyro which is able to precess about only one axis, called the output axis. Such a gyro will respond (precess) to rotation of the case about only one axis (known as the input axis). This allows the gyro to perform a control function about one axis and with three such gyros, mounted on a platform so that the input axis are mutually perpendicular, three-dimensional control of the platform is obtained.

By attaching the outer gimbal of the gyro (fig. 14-5) rigidly to the case, such a configuration is obtained. The gyro now has just one axis of freedom with respect to the case about which it can precess. This is called a single-degree-of freedom gyro. A gyro of this type with a few refinements is the one used for inertial navigation applications.
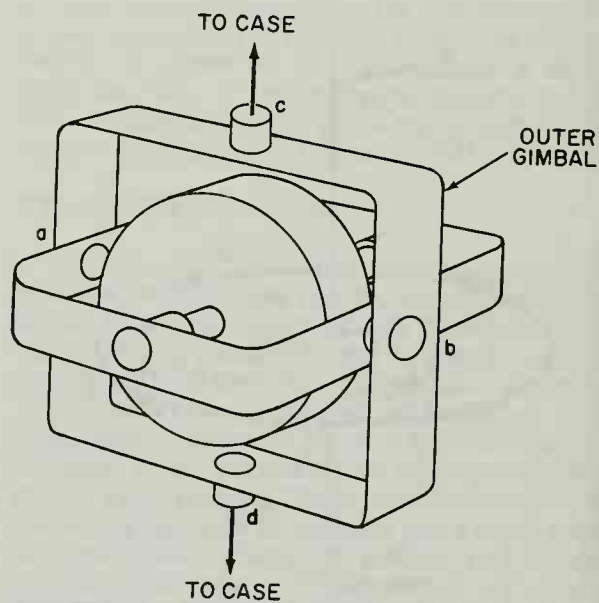
162.50

Figure 14-5.—Two degree of freedom gyro.

Inertial Navigation Gyroscopes

This refined gyro has as its rotor a 4-pole 3-phase synchronous hysteresis motor, which is supported on air bearings. The stator of the motor is part of the can surrounding the rotor. The motor receives a precision frequency signal so that the rotor speed and angular momentum remain constant.

As shown in figure 14-6, the rotating element and its cylindrical enclosure are enclosed inside another cylinder called the case. Between the two cylinders is a heavy viscous fluid contained by the outer case. The rotor and it's enclosure are adjusted for neutral buoyancy (no tendency to rise or sink) in the fluid and together are called the "float", and rotate about it's output axis on an almost frictionless bearing of oil. Gyro life depends on the number of starts and stops, rather than running time. On any power loss to a gyro, the rotor must be allowed to run-down to a complete stop before power is reapplied.

Mounted on the output axis are a pick-off and a torquer (not shown). The pick-off is a device used to measure any angular displacement of the float with respect to the case about the output axis. The signal representing this measurement from the pick-off is the

output of the gyro. The torquer is used to apply a rotational force to the float with respect to the case about the output axis causing an output from the gyro.

Gyro Drift

Gyro drift is any undesired precession of the rotor. In the case of the free gyro, drift will cause the spin axis to wander from its special orientation. For the single-degree-of-freedom gyro, since it can precess only about the output axis, drift causes the float to rotate about this axis. This produces an output since the case is not rotating about the output axis. This output appears exactly as if it had been caused by the gyro torquer or a rotation of the case about its input axis.

There are many causes of gyro drift. Some of them include bearing friction, unbalance, and change in environment. With each advance in gyro manufacturing, gyro drifts are reduced. Although gyro drifts have not been eliminated, gyros are built so that they will have a constant drift for long periods of time. When the drift remains constant it can be calculated and a compensation made.

In most inertial navigation systems, compensation does not eliminate the drift, but it can remove the effect of the drift. Since the effect is an output, the means to eliminate it is to apply a signal to the torquer. This signal, if proper, will counter the drift and the net result will be no gyro output. If the drift changes, then the signal to the torquer must also be changed. This signal to the torquer of a gyro which is designed to eliminate the output due to drift is called a gyro bias. An error in gyro bias has the same effect as an uncompensated gyro drift.

INERTIAL MOTIONS

As previously mentioned there are several motions in inertial space which are of concern in the study of inertial navigation. For the considerations of motions in this discussion the "fixed" stars are used for the inertial space reference. This is necessary because the inertial elements (gyros and accelerometers) which sense these motions operate using this reference. Some of these motions are desired to produce useful outputs while compensation for the others can only be accomplished if they are known.
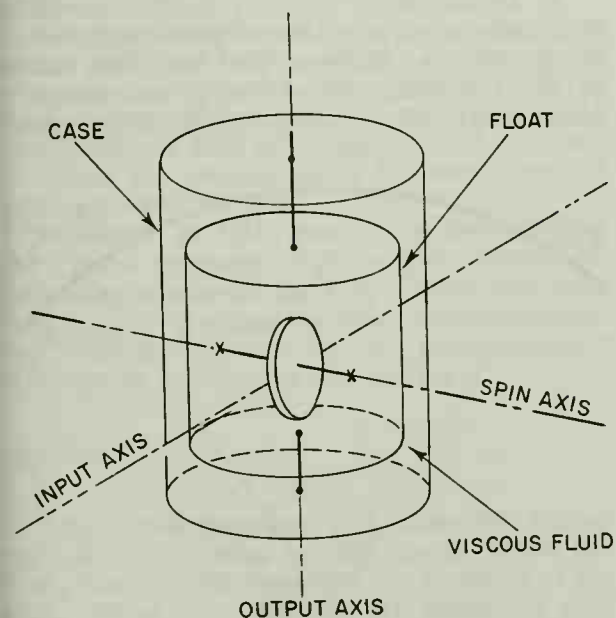
162.51

Figure 14-6.—Inertial navigation gyroscope.

For convenience these motions are divided into two categories; (1) rotations to which the gyro is sensitive and (2) accelerations to which the accelerometer is sensitive. Under rotations there are:

1. Ship's roll, pitch, and yaw
2. Earth's rotation
3. Changes in latitude and longitude

The accelerations of concern are:

1. Ship's accelerations with respect to the earth
2. Acceleration of gravity
3. Coriolis acceleration

Some of these are interrelated. For example, as will be seen, coriolis acceleration results because the earth rotates. The rotation of the earth also has an effect on gravity. There could be no latitude or longitude changes without an acceleration with respect to the earth.

There are a few inertial motions whose effects are negligible. That is, their effects are below the sensitivity level of the inertial elements in use today. These motions are the precession, nutation, and revolution of the earth, and the acceleration of the earth which results from the revolution of the earth about the sun.

Rotational Motions

When discussing rotational motions and their effects within an inertial navigator, it is most advantageous to consider the inertial navigator as maintaining a reference direction. The gyros will maintain a direction in inertial space, but since the inertial navigator is dealing with earth referenced values (position and velocity), the gyros must be controlled to maintain a reference with respect to the earth. The reference we will use is the local vertical (perpendicular to center of earth), and an orientation with respect to north.

The rotation of the earth causes the local vertical for a given position to change its direction in space. This change is not obvious to anyone on the earth because the local vertical maintains the same orientation with respect to the earth. The apparent precession of the gyros (previously discussed) would tend to keep the local vertical indication of the inertial navigator stationary in space. In order to prevent this, and keep the local vertical in respect to earth, gyro torquing signals are used. The value of these signals is determined so that the gyros, and therefore the inertial navigator, will rotate about the earth's spin axis at the same rate

that the earth does. As a result the inertial navigator which is controlled by the gyros will maintain the desired orientation with respect to the earth as the earth rotates in inertial space.

This gyro torquing signal which compensates for the rotational motion of earth in inertial space is called the earth torquing signal. It is applied to a particular single-degree-of-freedom gyro as a function of its orientation to the earth's spin axis. The vector sum of the individual signals to the gyros equals earth rate, the rotation rate of the earth in inertial space.

The earth rate torquing signal is only one part of the total torquing signal going to each gyro. The change in position of an inertial navigator on the earth's surface also causes the local vertical to change direction in space. This results from the fact that as the inertial navigator goes from one position to another it is changing from one local vertical to another. This is illustrated in figure 14-7. As the inertial navigator travels over the curved surface of the earth from left to right the "correct" orientation is shown at position 2 in the solid line figure. The dotted line figure at position 2 represents the inertial navigator after the change in position with no compensation for the change in local vertical.

It is necessary that our inertial navigator maintain the correct orientation at each position along the way. Knowing just the total change in local vertical orientation is not enough to have local vertical change exactly as necessary
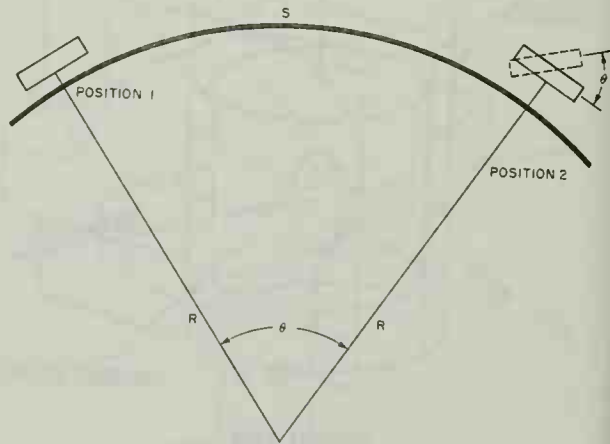
162.52
Figure 14-7.—Vertical change with movement over earth's surface.

ue to change in position. The amount that must change is the rate of change in respect to time. When R is assumed to be constant he time rate of change is:

$$\frac{d\theta}{dt} = \frac{1}{R}\frac{ds}{dt}$$

Where θ = function of the distance traveled
S = distance traveled
R = radius of the earth

Since the time rate of change of the distance, S, is velocity, V, the above equation becomes: $\frac{d\theta}{dt} = \frac{V}{R}$ and it becomes evident that the angular rate of change of the local vertical is simply the linear rate of change (velocity) converted into an angular rate by scaling to the earth's radius.

This is the rate at which the local vertical change direction in space and the inertial navigator must also change its direction at this rate to track the local vertical. The man who first described this motion and its compensation was a German physicist named Schuler. As a result, when the gyros receive a properly calibrated signal to compensate for the movement of the local vertical due to movement over the earth, the system is said to be Schuler tuned.

The remaining inertial rotation is that of the ship's rotations with respect to the earth. These are called roll, pitch, and yaw. These when converted into the inertial navigator's coordinates of roll, pitch, and heading are used to define the orientation of the local vertical with respect to the ship.

However, in this case it is desired that the gyro's property of precession be used to determine whenever the ship's rotational motions tend to displace the local vertical indication. Therefore, there is no "compensation" of the gyros for this inertial motion.

Accelerations

The accelerometers will sense the forces applied by the case on the mass within the case, and thereby sense the acceleration of the case. These accelerations are measured with respect to inertial space and therefore some of the accelerations which produce outputs do not result in movement of the inertial navigator

over the surface of the earth. The main inertial accelerations in this category are gravity and coriolis accelerations.

Gravity, which defines the local vertical, is the sum of two inertial accelerations; gravitational acceleration and centripetal acceleration. The gravitional acceleration is the result of the mass attraction between two bodies. The magnitude of this acceleration is a function of the mass of each body and the distance between the center of mass of each body; in this instance between the earth and the inertial navigator. Since the radius of the earth can be considered a constant and the mass of the earth is many times greater than the mass of the inertial navigator, or any other body on the earth's surface, the magnitude of the gravitational acceleration can be considered a constant. The direction of this acceleration is along the line between the mass centers of the two bodies.

Because the earth is rotating, all bodies on its surface are also subject to a centripetal acceleration in space. This acceleration is directed towards the rotation axis of the earth and is greatest at the equator. It decreases in magnitude as latitude is increased, due to a decrease in the effective radius.

Accelerometers cannot distinguish between the components of gravity. However, they will be affected by gravity if they are tilted. The output which would be produced would not be due to any acceleration of the inertial navigator over the surface of the earth so a compensation must be made. The easiest solution for this elimination of the effect of gravity is to place the accelerometer so that its sensitive axis is perpendicular to gravity. Since all accelerations which result in a change in position are perpendicular to the local vertical this orientation would allow the accelerometer to measure these accelerations without being affected by gravity. Two accelerometers, then, are placed perpendicular to the local vertical and perpendicular to each other so that together they can measure any acceleration which will result in a change of position on the earth's surface.

Another acceleration occurring in inertial space which is not an acceleration with respect to earth is coriolis acceleration. Since the inertially oriented accelerometers produce output due to coriolis acceleration, a compensation must be made to prevent errors.

A body experiences coriolis acceleration when it travels over a rotating surface in a direction perpendicular to the axis of rotation. A merry-go-round can be used to illustrate this coriolis acceleration.

Suppose we are on a merry-go-round, and we wish to travel from point A to point B, as shown in figure 14-8. Initially the platform is not turning. Naturally the shortest distance is the straight line AB, and this will also appear as a straight line to a person who is not on the platform. However, if the platform is turning in the direction indicated, then by the time we reach point B it will have moved to location B'. To the person who is not on the platform we will appear to have followed the curved line A'B', and according to Newton's First and Second Laws we must have experienced a force, and hence an acceleration, to the left.

Travel on the earth creates the same situation. In the northern hemisphere we will experience an acceleration, to the left of track, with respect to earth. In the southern hemisphere this acceleration is to the right. This acceleration which causes the path over the earth to curve is called coriolis and since an accelerometer makes its measurements with respect to inertial space, it will sense this acceleration, so we must correct for it.
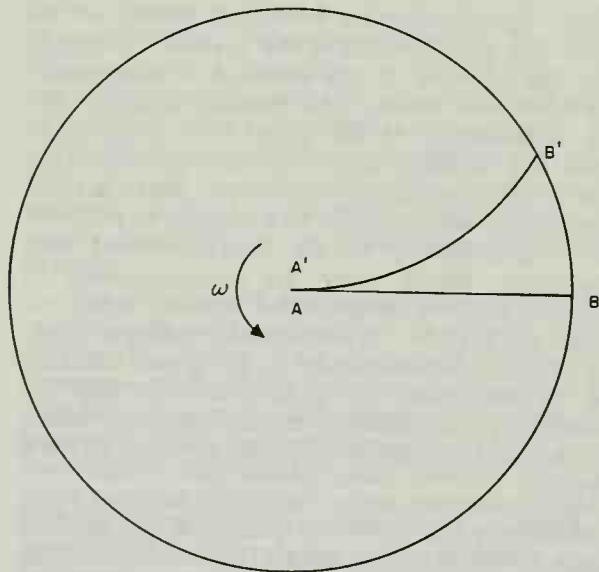
162.53
Figure 14-8.—Motion on a merry-go-round.

The coriolis acceleration is equal to twice the velocity with respect to the rotating base times the rate of rotation of the base, $(a = 2Vw)$ and is a function of latitude. A more careful analysis will reveal that the velocity is the component perpendicular to the axis of rotation and acceleration is directed perpendicular to both. Since the earth is rotating, any movement on its surface except in the north-south direction at the equator, has a coriolis acceleration associated with it. This will be sensed only by an inertially sensitive device.

With the effects of both gravity and coriolis accelerations eliminated from the velocity values, the accelerations remaining are those of the inertial navigator with respect to the earth. These accelerations result in position changes on the earth and will be measured by the accelerometers, since they are also accelerations in inertial space.

STABLE PLATFORM

The function of the accelerometers in the inertial navigator is to sense the accelerations of the ship with respect to the earth so that a computer can determine the ship's velocity and position. The function of the gyros is to maintain the accelerometers in the correct position to measure these accelerations. The function of the gimbals is to provide the physical support for the gyros and accelerometers and provide 3 degrees of freedom with respect to the ship. The gimbals, gyros and accelerometers together with associated electronics and gimbal torquer motors form a stable platform.

The purpose of the stable platform is to establish and maintain a reference system in which the measurements necessary to produce the inertial navigator's outputs are taken. The reference system most commonly used is based upon the local vertical and the direction of North. Both of these are referenced in respect to earth. As a result, in our application, "stable" in stable platform means fixed to or stable in respect to the earth. This means that any motion or force which might disturb the reference system must be accounted for and its effect must be eliminated. This would include motions in inertial space previously discussed as well as motions of the vehicle in which the inertial navigator is mounted.

Since the vehicle on which the stable platform is mounted operates in three dimensional

pace, it has three degrees of freedom with espect to earth. In order to maintain a reference xed to the earth, then, the stable platform just have three degrees of freedom with respect the vehicle. The stable platform contains ree gimbals, one for each degree of freedom ecessary for the stable platform. Each gimbal as rotational freedom about one axis with espect to its supporting element. The gimbal an be driven with respect to its support about the gimbal axis by a gimbal torquing motor.

Figure 14-9 is a cutaway view of a simple table platform. For this discussion, the gimbal hich is supported by the vehicle at points a nd b is called the pitch gimbal. Rotational 1ovement of this gimbal with respect to the ehicle about its axis is called pitch. The pitch imbal in turn supports the roll gimbal on an xis perpendicular to the pitch axis. Rotational 1ovement of the roll gimbal about its axis ith respect to the pitch gimbal is called roll. he roll gimbal supports the heading or azimuth

gimbal. The heading gimbal has one degree of freedom with respect to the roll gimbal, and rotational movement about the heading gimbal axis with respect to the roll gimbal is a change in heading. The heading gimbal through its own axis and the axes of the roll and pitch gimbals has three degree or complete rotational freedom with respect to the vehicle on which the platform is mounted. The heading gimbal is able to assume any orientation with respect to the vehicle within the limits of the gimbal physical restraints.

In order to understand how the stable platform functions to maintain the reference system, a simple sketch of a platform with gimbals, gyros, accelerometers, and servomotors is illustrated in figure 14-10.

Stabilization is accomplished by mounting the accelerometers and gyros on the platform such that the input axis of the gyros are perpendicular to the sensitive axis of the accelerometers. By referring to figure 14-10 it can be seen with the gyros on the stable element that: (1) the heading gimbal axis is always aligned to the Z gyro; (2) the roll gimbal axis is aligned to the X and Y gyros as a function of heading; (3) the pitch gimbal is aligned to the X and Y gyros as a function of heading and roll and to the Z gyro as a function of roll. Any motion about the input axis of the gyro (due to roll, pitch or yaw) causes the gyro to precess
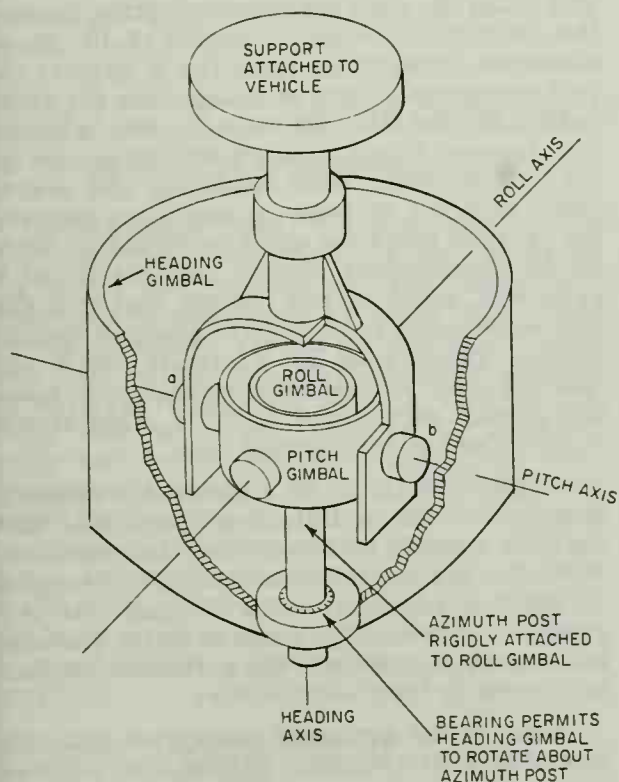
162.54

Figure 14-9.—Cut-away view of a stable platform.
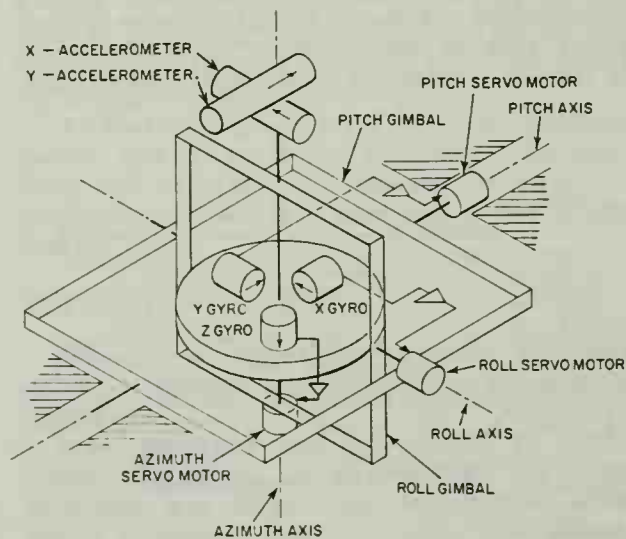
162.55

Figure 14-10.—Stable platform with inertial components.

about its output axis. The gyro pickoff measures this precession and sends an actuating signal to the gimbal servomotors which in turn drive the platform relative to the vehicle, in a direction opposite to the direction of the ship's motion. The platform is driven until the gyro pickoff signal is zero. The platform therefore moves with respect to the ship but remains stationary or aligned with respect to the reference plane of direction.

Any vector, such as acceleration, can be resolved into two right angle components. Since distance north-south, along a meridian of longitude, can be related to degree of latitude; and distance east-west, along a line of latitude, can be related to degrees of longitude, two accelerometers are mounted on the platform so that their sensitive axes are mutually perpendicular. One, the "X" accelerometer, is always aligned north and the other, the "Y" accelerometer, is always aligned east.

Figure 14-11 shows in block diagrm form how the system computes latitude and longitude. The X direction is defined as an axis along the north-south direction and tangent to the surface of the earth. The Y axis is along the east-west tangent to the earth. The X accelerometer senses accelerations in a north-south direction. Its output is used to compute velocity north and distance north which is related to latitude. The Y gyro senses any rotation of the accelerometer's sensitive axis from its aligned position. The Y accelerometer senses acceleration east-west. From this output is computed velocity and distance east-west which is related to longitude by the ratio 1/cos latitude,

(at the equator, one mile equals one minute of longitude. While at 45 degrees of latitude, one mile equals 1.4 minutes of longitude).

The X gyro senses rotation of the Y accelerometer from its aligned position and the Z gyro has its input axis aligned perpendicular to the plane containing the X and Y axes. The axis is parallel to the local gravity vector. It is precessed whenever the platform moves in azimuth from its northeast orientation.

Since the three gyro input axes are orthogonal (mutually perpendicular) no rotational motion of the platform from its aligned position will escape being sensed by one or more gyros.

Stabilization Loop

In figure 14-10 any ship's motion about its roll axis while headed north, will result in the X gyro sensing this rotation causing a signal to drive the platform with respect to the ship by way of the roll servomotor. Pitching motions will be sensed by the Y gyro which will cause the pitch servomotor to drive, keeping the platform leveled. Figure 14-10 shows electrical connections from the X gyro to the roll servomotor which is satisfactory for a ship headed 000 degrees, but when the ship is headed 090 degrees a roll position will be sensed by the Y gyro which must cause the roll servomotor to drive. However, as seen in the diagram, the Y gyro feeds the pitch servomotor. When on an intercardinal heading both the X and Y gyro will sense a roll motion that only the roll servomotor need drive to maintain stabilization. There must be a circuit which will combine the precessional signals and cause the proper servomotor to drive. This circuit is known as the stabilization loop.

Figure 14-12 is a simplified schematic diagram of the stabilization loop. Note that the gyro pickoffs feed resolvers, and that these resolvers are positioned according to the angles of heading and roll. The heading and roll resolvers are usually mounted on the gimbals, and are driven whenever the particular gimbals are driven by their servomotors.

Many of the problems associated with this type of inertial navigation system have not been discussed in this brief outline. Paramount among these are such things as the effects of Schuler Tuning, gyro drift, accelerometer error, platform misalignment and computer error.
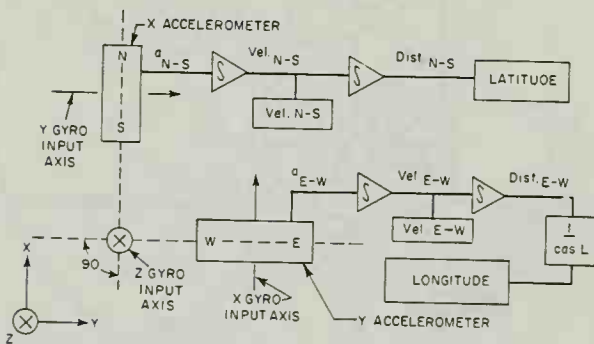
162.56
Figure 14-11.—Block diagram sequence for computing latitude and longitude.
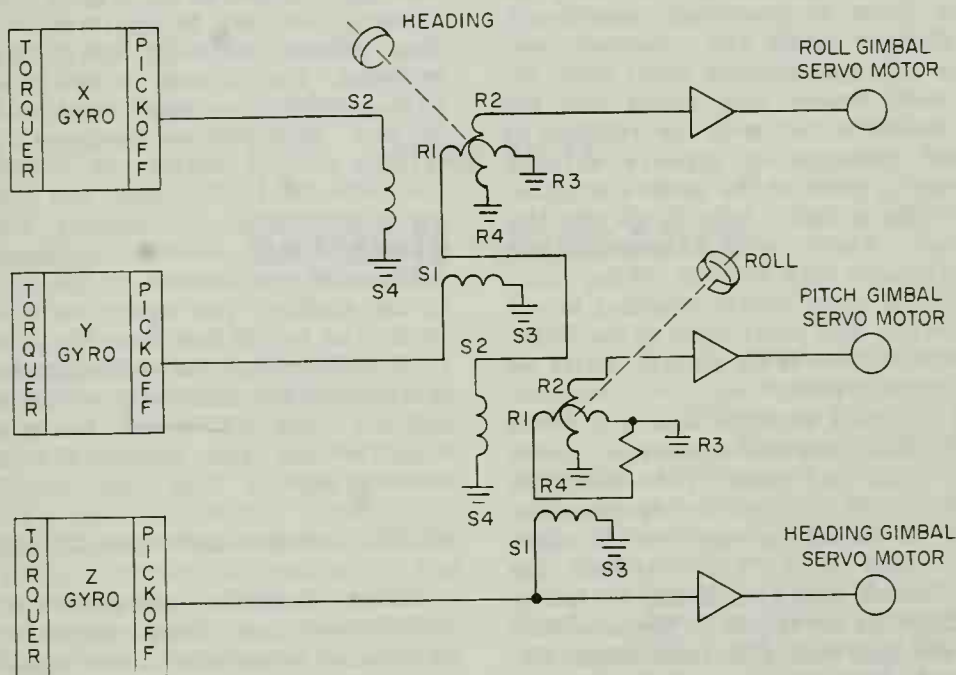
162.57

Figure 14-12.—Three-gimbal stabilization loop.

SATELLITE COMMUNICATIONS

In the year 1888, the first ship-to-ship "wireless" communication was made between a tug and a ship in the New York Navy Yard. This was accomplished by using large coils of wire wrapped around the hull of the ship. The first official U.S. Navy message was sent by wireless telegraph in 1899. In 1954 the U.S. Navy was using the moon as a "passive" relay between Hawaii and Maryland for communications. As early as 1962, the USNS KINGSPORT became the world's first satellite communications terminal afloat. In early 1965 the USS MIDWAY placed a call to the USS CANBERRA, 6,000 nautical miles away, and for the first time in history two combatant ships had established communication by means of satellite.

DEFENSE COMMUNICATIONS SATELLITE PROGRAM

Successful results from the test with satellite communication prompted the Defense Department to establish the Defense Communications Satellite Program (DCSP), under direction of the Defense Communication System. Within this system, satellite communication will form a part of the long-haul links between the switching centers. These links will employ the satellites in addition to other forms of existing communications media.

As the techniques of satellite communications evolve, and are perfected, satellite communications will provide the major part of all intercontinental links in the Defense Communications System.

TYPES OF SATELLITES

There are two basic types of satellites used as communication links, and two distinct orbit types.

The satellites are referred to as either active or passive. In this instance the active satellites are those that incorporate some device to either amplify or otherwise change, and re-broadcast the signal they receive.

Passive satellites do not include electronics equipment for the purpose of amplification of signals transmitted over the link. They provide

a reflective surface against which the radio waves are bounced. An example of this type would be the moon as previously mentioned.

The two distinct orbits are "random" and "synchronous". A synchronous orbit over the equator at such height and speed that the satellite is synchronized with the rotation of the earth and consequently appears to hang motionless over a point on the earth's surface.

Random orbits actually take in all satellite orbits achieved. These may vary from low altitude to extremely high altitude orbits, polar to equatorial, and from nearly circular to extremely eccentric. The polar orbit of the Navigation Satellite discussed in chapter 15 is an example of a random orbit.

In August of 1966 a multiple launch of seven communication satellites in an equatorial "near-synchronous" orbit was made. This means at an altitude below that required for synchronous orbit, so that the satellites separated in space would drift slowly in a "belt" around the equator. The launch was successful, but minor technical difficulties developed in the satellites resulting in less than desirable communication. Another multiple launch was made in January 1967 and it proven successful, the first continuous coverage satellite communication link around the world will be established.

There are presently several operational communication satellites in use, such as Telstar II, Relay I, Relay II, Syncom II, Syncom III, and Early Bird. Millions of Americans were able to witness the recovery operation of two Astronauts, on their home television sets, via way of Early Bird Satellite.

SATELLITE COMMUNICATIONS TERMINALS

The shipboard communication terminal designed for use as the shipboard link in the DCSP is designated AN/SSC-3. The first of this model was installed in USS PROVIDENCE (CLG-9) in October of 1966. Test results showed the terminal capable of communicating, while underway, voice and teletype through a satellite to another station.

The AN/SSC-3 utilizes a six-foot parabolic antenna; considered the largest communications antenna that can be operated on a combatant ship without seriously interfering with other systems. The antenna is set on a three-axis, gyro-stabilized pedestal which will operate in a 75-knot wind with an accuracy of 0.1 degree.

The control shelter is 7 feet high and 7 feet wide by 12 feet long, and contains its own air conditioning. All antenna acquisition and tracking control, baseband equipment, and status indicators are located on the control console in the shelter. The equipment shelter may be located up to 100 feet from the antenna.

The terminal is self-contained for easy installation and removal, so that when the ship returns from deployment, the terminal may be removed and then installed in another ship about to deploy.

FUTURE SATELLITE COMMUNICATION

When satellite communications are well established and Navy ships have need for permanent terminals, a new type will be integrated into the ship and be fully compatible with other electronic systems and equipment. When this happens, the number of conventional transmitters and receivers aboard ship can be reduced, thus eliminating portions of the maze of antennas which today interfere with each other and reduce the effectiveness of shipboard communications. This will result in better communications in all frequency bands, more capacity for reliable, long-range communications, and overall reduction in shipboard space and weight needed for communications equipment.

Through a rapidly advancing research and development program, the use of higher frequencies, and smaller antennas, relay of NTDS (Naval Tactical Data System) information, via satellite will become possible. Ships will eventually need the capability for tracking and exchanging information with many types of satellites other than communication satellites. Some examples of multiple satellite use might be intelligence, meteorology, navigation, and tactical data.

CHAPTER 15

SATELLITE NAVIGATION

Satellite Navigation was thought feasible after observation of Russia's first artificial earth satellite, Sputnik I. Scientists listened to the beep generated by Sputnik as it passed by and noted the Doppler-like shift in the received radio frequency signals. Doppler effect is an apparent change in a received frequency because of relative motion between the transmitter and receiver. If the distance between the transmitter and receiver is decreasing, the received frequency is higher than that which is actually transmitted; if the distance is increasing, the received frequency is lower than that transmitted. It was later demonstrated that accurate measurement of this Doppler shift pattern would permit the determination of a satellite orbit. From this successfully proved technique it was further reasoned that, working from a known satellite orbit, a listener could determine his position on the surface of the earth from an observed Doppler pattern. From this point followed the first successful satellite launch in April 1960, and the U.S. Navy Navigation Satellite System became an all-weather, highly accurate, fully operational navigation aid, that enables navigators to obtain accurate navigation fixes from the data collected during a single pass of an orbiting satellite.

NAVIGATION SYSTEM DESCRIPTION

The Navy Navigation Satellite System, shown pictorially in figure 15-1 is a worldwide, all-weather system of high accuracy, which enables navigators to obtain fixes approximately every two hours of the day or night. It consists of four earth-orbiting satellites, four tracking stations, two injection stations, the U.S. Naval Observatory, a computing center, and shipboard navigational equipment (Radio Navigation Set AN/SRN-9) of which there can be any number.

Each of the four satellites is completely transistorized and is placed in a circular polar orbit at an altitude of 500-700 (nominal 600) nautical miles. The orbital planes of the four satellites intersect at the earth's rotational axis and are spaced 45 degrees apart in longitude. Thus the orbital paths cross at the north and south poles as illustrated in figure 15-2. Although the orbital planes remain essentially fixed in space, the satellites appear to traverse the longitudinal meridians as the earth rotates beneath them. To an observer on the earth, the apparent effect is that the satellite orbits are moving westward.

The satellites are of octagonal shape with four windmill-like vanes to carry solar cells, and are stabilized in the earth's gravitational field so that directional antennas are pointed down at the earth rather than into space. The satellites contain an ultrastable oscillator, two coherent harmonically related transmitters, a digital clock, command receiver, digital memory, telemetry, and power supplies.

Each satellite orbits the earth in approximately 108 minutes. Throughout its useful life, each satellite transmits the following phase-modulated data every two minutes on two RF carriers: two synchronization signals, an identification code, a 400-hertz tone, fixed and variable parameters describing its own orbit, and fixed parameters describing the orbits of the other three satellites.

The fixed parameters describe the satellite's nominal orbits and are correct only for a 12 to 16 hour interval. The variable parameters describe the fine structure in the satellite's nominal orbit as a function of time and are correct only for the time interval they are transmitted by the satellite. Thus, the satellite's memory stores sufficient variable parameters to describe its orbit at two-minute intervals between subsequent injections of data
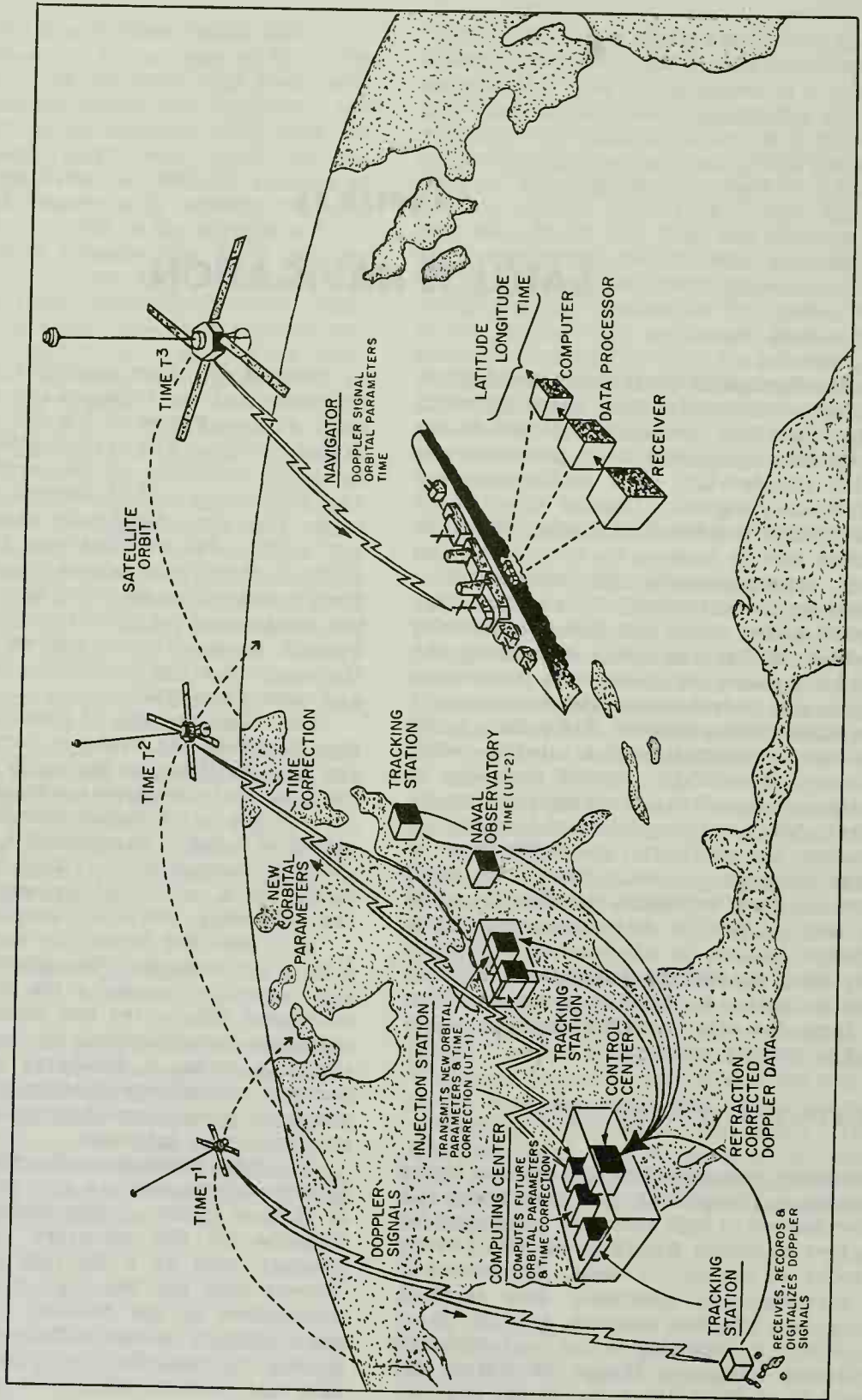
Figure 15-1. — Navy navigation satellite system.

162.58
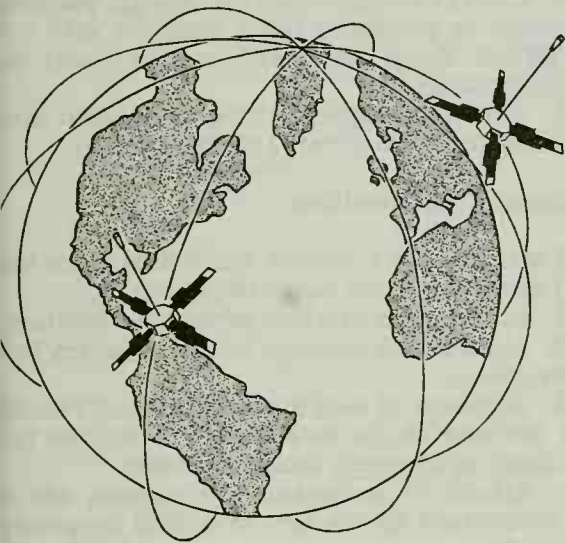
162.59

Figure 15-2.—Four polar orbits with 45° nodes.

into its memory. Each transmission is timed so that the end of the 78th bit, which is the last bit of the second synchronization signal, coincides with the even two minutes of Universal Time (UT-2). Thus, the satellites also serve as an accurate time reference for all navigators that are equipped with Radio Navigation Set AN/SRN-9 to receive and decode satellite transmissions.

Tracking Stations

In order to determine accurately its present and future orbit, each satellite is tracked as it passes within line-of-sight of each of the four fixed tracking stations. Each tracking station includes radio receiving and data processing equipment which functions to receive and decode the satellite transmissions. The received signal is digitized and sent with satellite time measurements, via the control center, to the central computing center. The tracking stations are located in the states of Hawaii, California, Minnesota, and Maine.

Naval Observatory

In addition to the Tracking Stations, the Naval Observatory receives signals during each pass. During a pass, the data processing equipment at the Naval Observatory compares the time of

reception of each 78th bit it receives with the even two minutes of Universal Time (UT-2). The times at which satellite time signals are received are then transmitted to the Control Center.

Control Center

The times at which satellite time signals are received are forwarded by the Control Center to the Central Computing Center. The Control Center monitors system operation and acts as a switching central for the routing of data to and from the Central Computing Center.

Central Computing Center

The Central Computing Center continuously accepts data inputs on the satellites from the four Tracking Stations and the Naval Observatory. Periodically, to obtain fixed orbital parameters for a satellite, the Central Computing Center computes an orbit for each satellite that best fits the doppler curves obtained from all Tracking Stations. Then using the computed orbital shape, the central computing center extrapolates the position of the satellite at each even two minutes in universal time for the next 12 to 16 hours, subsequent to data injection. These data together with data on the nominal shape of the orbits of the other three satellites, commands and time correction data for the satellite, and antenna-pointing orders for the Injection Station antennas are supplied to the Injection Stations via the Control Center.

Injection Stations

The Injection Stations, after receiving and verifying the incoming message from the Central Computing Center, store the message until it is needed for transmission to the satellite. As soon as the receiving equipment at the Injection Station receives and locks on the satellite's signals, the Injection Station reads the injection data and commands from storage and transmits them to the satellite. Transmission to the satellite is on a frequency different from those frequencies used by the satellite, and the bit rate is much higher; therefore, injection is completed in a matter of seconds. Once data injection is complete, the satellite continues to transmit at the normal two-minute intervals. Any time corrections for the satellite clock and any commands for the satellite are also

performed during the period of data injection. These precautions ensure that the navigational equipment, which depends on accurate satellite data for determining its positions, is provided the best possible data from each satellite. Any time that the satellite is within radio line of sight of the navigational equipment and has a maximum elevation angle at time of closest approach equal to or greater than 10 degrees and equal to or less than 70 degrees, the satellite's transmission can be used to compute the exact position on earth of the navigational equipment. Satellite passes suitable for use in obtaining a navigation fix will generally occur at least every two hours.

Shipboard Navigation System

The final link in the satellite navigation system is the shipboard navigation system (fig. 15-3). This equipment receives the satellite signals, measures the doppler shift, decodes the information content of the phase modulation, and performs the computations required to determine the ship's position.

The Radio Navigation Set AN/SRN-9 reduces the received satellite signals to a form which is suitable for navigational computations. One of the following three options may be used to perform the computation:

1. General purpose computer—Data must first be processed through a suitable buffer unit which converts the information into the computer format.
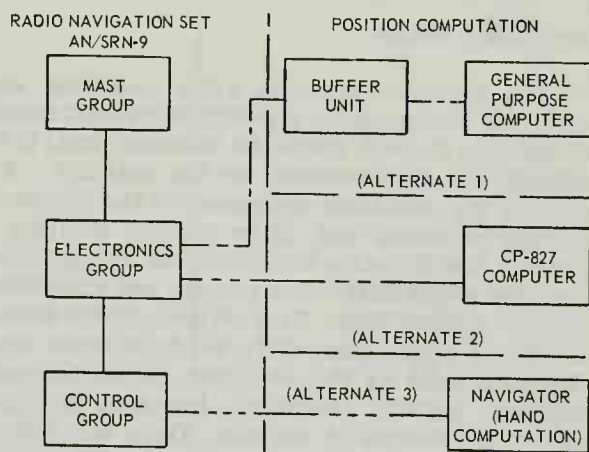
2. CP-827 Computer—This special purpose computer is presently being supplied with the AN/SRN-9 Radio Navigation Set and will be discussed later.

3. Hand computation—Using the printed data available from the Control Group printer.

Accuracy Considerations

There are five factors which determine the final accuracy of the navigational fix:
1. Accuracy of satellite orbit determination.
2. Accuracy of satellite orbit prediction for future times.
3. Accuracy of doppler measurements which is a function of the satellite oscillator and the receiving equipment's local oscillator.
4. Effects of ionospheric refraction, which are minimized by the use of a dual frequency system.
5. Accuracy of determination of ship's course and speed. (A 1-knot error in velocity North or South will contribute approximately a 0.2-nautical mile error to the navigational fix; errors in velocity East or West are less serious.)

INTEGRATED DOPPLER NAVIGATION

The navigational fix computed by Radio Navigation Set AN/SRN-9 is based on the shift in frequency (doppler frequency shift) that occurs whenever the relative distance between a transmitter and a receiver changes. Such a change occurs whenever a transmitting satellite passes within radio range of a receiver on earth and is due to the motion of the satellite in its orbit, the motion of the navigator on the surface of the earth, and the rotation of the earth about its axis.

As previously stated, the satellite message describes the orbital position of the satellite every two minutes on the even minute. To obtain a navigational fix, it is necessary only to determine the ship's location relative to the known satellite positions. The Radio Navigation Set AN/SRN-9 employs a so-called integrated doppler measurement for this purpose. In figure 15-4 positions of the satellite in its orbit are shown for times t_1 through t_4. The slant range from ship to satellite is given by S_1 through S_4. The positions of the navigator, P_1, through P_4, refer to the times at which the SRN-9 receiver recognizes the satellite synchronization signal, $t_1 + \Delta t_1$ through $t_4 + \Delta t_4$. Note that the times of reception are slightly

162.60
Figure 15-3.—Shipboard navigation system configuration.
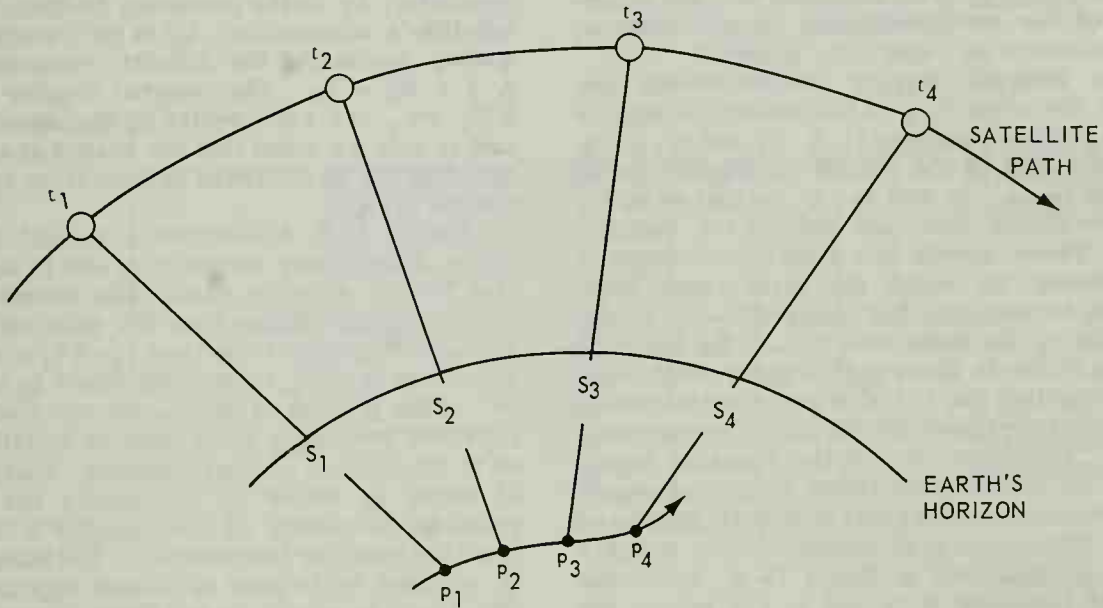
f_o = NOMINAL VALUE OF NAVIGATOR'S
REFERENCE FREQUENCY

c = SPEED OF LIGHT

Δ_f = DIFFERENCE BETWEEN NAVIGATOR'S
REFERENCE FREQUENCY AND
SATELLITE TRANSMISSION FREQUENCY

T = 2 MINUTES
(I.E. $t_2 - t_1$, $t_3 - t_2$, ETC.)

λ = LATITUDE

ω = LONGITUDE

t = TIME OF TRANSMISSION OF TIMING
MARK

$t + \Delta t$ = TIME OF RECEPTION OF TIMING MARK

N = DOPPLER COUNT

S = SLANT RANGE

$$N_{12} = f_{o/c} [S_2(\lambda, \omega) - S_1(\lambda, \omega)] + \Delta f.T$$

$$N_{23} = f_{o/c} [S_3(\lambda, \omega) - S_2(\lambda, \omega)] + \Delta f.T$$

$$N_{34} = f_{o/c} [S_4(\lambda, \omega) - S_3(\lambda, \omega)] + \Delta f.T$$

162.61

Figure 15-4.—Integrated doppler measurement.

later than the times of transmission due to the radio propagation time over the slant ranges S_1, S_2, S_3, and S_4. Also, since the propagation times Δt_1, Δt_2, Δt_3, and Δt_4, are not equal, the time intervals over which the SRN-9 makes its integral doppler measurements will differ slightly from the exact two-minute value, being somewhat smaller prior to satellite closest approach (while S is decreasing) and somewhat larger after closest approach (while S is increasing). This fact, however, does not affect the result of the measurement because the number of r-f cycles transmitted by the satellite between synchronization signals (exactly

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two minutes apart) must necessarily equal the number received by the navigator between receptions of the synchronization signals since no r-f cycles can be "lost" or "gained."

The integral doppler measurements are simply the count N_{12} of the number of doppler cycles received between $t_{1+\Delta}$ t_1 and $t_{2+\Delta}$ t_2, the count N_{23} of the number of doppler cycles between $t_{2+\Delta}$ t_2 and $t_{3+\Delta}$ t_3, and so on for all two-minute intervals during the satellite pass. These counts are a direct measure of the amount by which the slant range from satellite to navigator has changed (S_2-S_1, S_3-S_2, etc.) during the count intervals. This measure is quite accurate since each doppler count added (or subtracted) due to the relative motion means that S has decreased (or increased) by one wavelength. Therefore, one of the required inputs to the fix computation (slant range increment over two-minute intervals) is directly measured by the SRN-9 integral doppler count, suitably scaled as indicated in figure 15-4. Since the satellite positions at t_1 and t_2 are known, the receiver must be on some surface defined by the measured slant range difference between these two points. This surface is a hyperbola of revolution with foci at the known satellite positions. The ship must be located, therefore, somewhere along the curve defined by the intersection of this hyperboloid and the earth's surface. The next doppler count (between t_2 and t_3) will define a second curve, and the intersection of these curves gives the navigational fix.

In actual practice two factors complicate this simple explanation. First, the doppler signal which is counted consists of the doppler frequency plus a fixed, but not very accurately known, bias frequency (Δ f in fig. 15-4). Therefore, a third doppler count is required in order to solve for the three variables—latitude (λ), longitude (ω), and bias frequency (Δ f) as defined in figure 15-4. This means that integral doppler counts for at least three two-minute intervals must be used (and preferably more than three) in order to determine the three unknowns, λ, ω, and Δ f. The second complication is the motions of the ship during the satellite pass. To account for this, the best estimate of a ship's motion must be entered into the navigational computation along with the doppler counts and the satellite message.

The fact that the integral doppler counts N_{12} through N_{34} are directly affected by Δ f, is illustrated in figure 15-5, wherein f_G is the (constant) frequency of the navigator's reference oscillator, f_T is the (constant) frequency of the satellite's transmitter, f_R is the received frequency containing the doppler component, and Δ f = $f_G - f_T$. The integral doppler counts, N12, etc., are represented by the shaded area, and it will be noted that the shaded area is increasing as we progress in time from $t_{1+\Delta}$ t_1 toward $t_{3+\Delta}$ t_3.

Figure 15-5 illustrates a typical doppler curve of frequency versus time and is an actual plot from a satellite pass. The received frequency (f_R) is higher than the satellite transmission frequency from time $t_1 + \Delta t_1$ to slightly after time $t_3 + \Delta t_3$, to the point where f_R crosses f_T. This portion of the curve represents the received frequency from time of satellite lock on to the point of closest approach. That portion of curve f_R below f_T represents the actual received frequency as the satellite's distance from the receiver is increasing. The point where f_R crosses f_T is point of closest approach. At this time, f_R equals f_T and there is zero doppler.

There is only one Doppler frequency curve for any one position on earth that can be obtained for a given navigational fix by satellite. Figure 15-6 displays the characteristics of Doppler frequency shift for different locations about an arbitrary reference point on earth. Notice that in all cases the received frequency (f_R) is greater than the transmitted frequency (f_T) as the satellite approaches the navigator, in effect "compressing" the radio waves enroute. Notice also that the received frequency equals the transmitted frequency at the point of closest satellite approach, and that the received frequency decreases as the satellite recedes from the navigator, thereby "expanding" the radio waves between them. The four curves of figure 15-6 show the Doppler frequency variations received at selected positions, with respect to the satellite's subtrack. A subtrack is the path traced on the earth by a satellite passing directly overhead. When the navigator is located on the subtrack, the satellite passes directly overhead and the Doppler frequency variation is similar to that shown in figure 15-6, A. An observer north of the navigator would also be on the subtrack and would receive an identical Doppler frequency curve, but at an earlier time if the satellite were moving north to south, and at a later time if the satellite were travelling south to north. An observer to the west or east would record the variation at the same time as
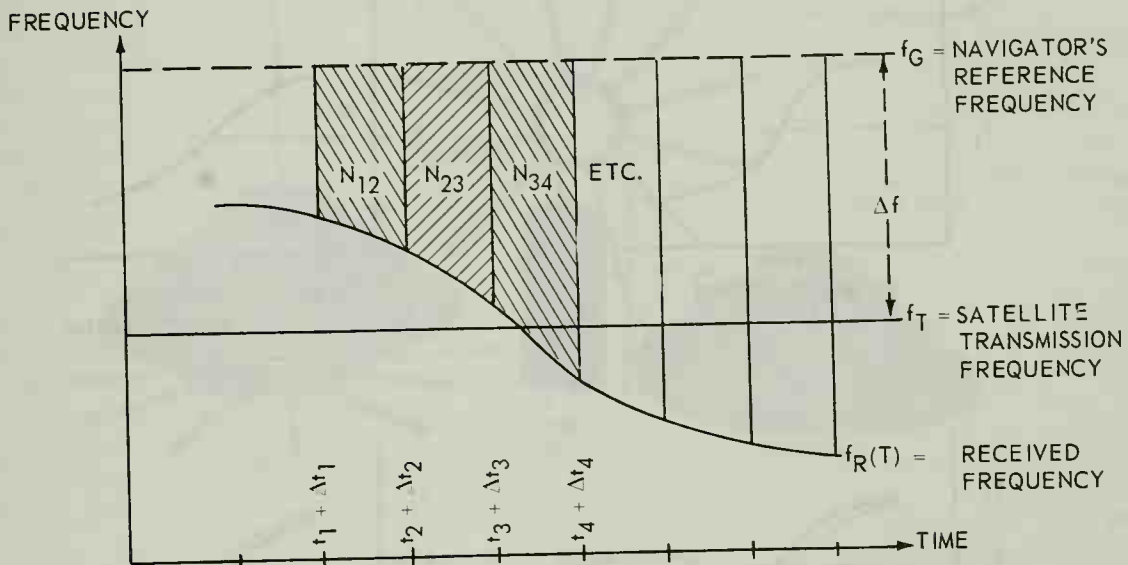
162.62

Figure 15-5.—Doppler frequency variation with time.

he navigator, but with altered shape as shown n figure 15-5,B,C, and D. Figure 15-5,B ex-hibits the more gradual rate of change of Dopp-ler frequency shift introduced by displacement to either east or west of the subtrack; while the effect of the earth's rotation is added to the curves of figure 15-5, C and D, showing how east and west displacements from the subtrack are distinguished from each other.

Thus, the shape of the received Doppler frequency curve and its time of reception depend on the receiver's position on earth, the satel-lite's location in space, and the rotation of the earth. If either the receiver's position or the satellite's position is known for the time of reception, the other can be determined from the received Doppler frequency curve.

AN/SRN-9 DESCRIPTION

This section provides a functional block dia-gram description of the Radio Navigation Set AN/SRN-9. The AN/SRN-9 Radio Navigation Set can be broken down to three major groups; Mast Group, Electronics Group, and Control Group which are treated separately in the following paragraphs.

Mast Group

The Mast Group (fig. 15-7), consists of a vertical whip antenna, upper and lower ground planes, and a housing assembly. The housing assembly contains all the electronic circuits associated with the Mast Group.

The Mast Group receives, separates, and amplifies the two signals from the satellite. The signal flow, in block diagram form, is illustrated in figure 15-8.

The two satellite signals are received by the dual frequency antenna. The antenna (fig. 15-7), is a simple rod of approximately 1/4 wavelength at 150 mc and 3/4 wavelength at 400 mc, mounted above a conical ground plane. A second ground plane consisting of 12 radial rods is located at the base of the Mast Group to isolate the Mast Group from any mounting structure and thus preserve the antenna pattern.

The diplexer (fig. 15-8) separates the com-bined signal from the antenna into its 150-mc and 400-mc components, each of which is con-ducted by coaxial cable to its respective pre-selector (preselective filter). If strong adjacent-channel signals are not anticipated, the output from the diplexer may be inserted directly into

DOTTED LINE = WITHOUT EARTH'S ROTATION
SOLID LINE = WITH EARTH'S ROTATION
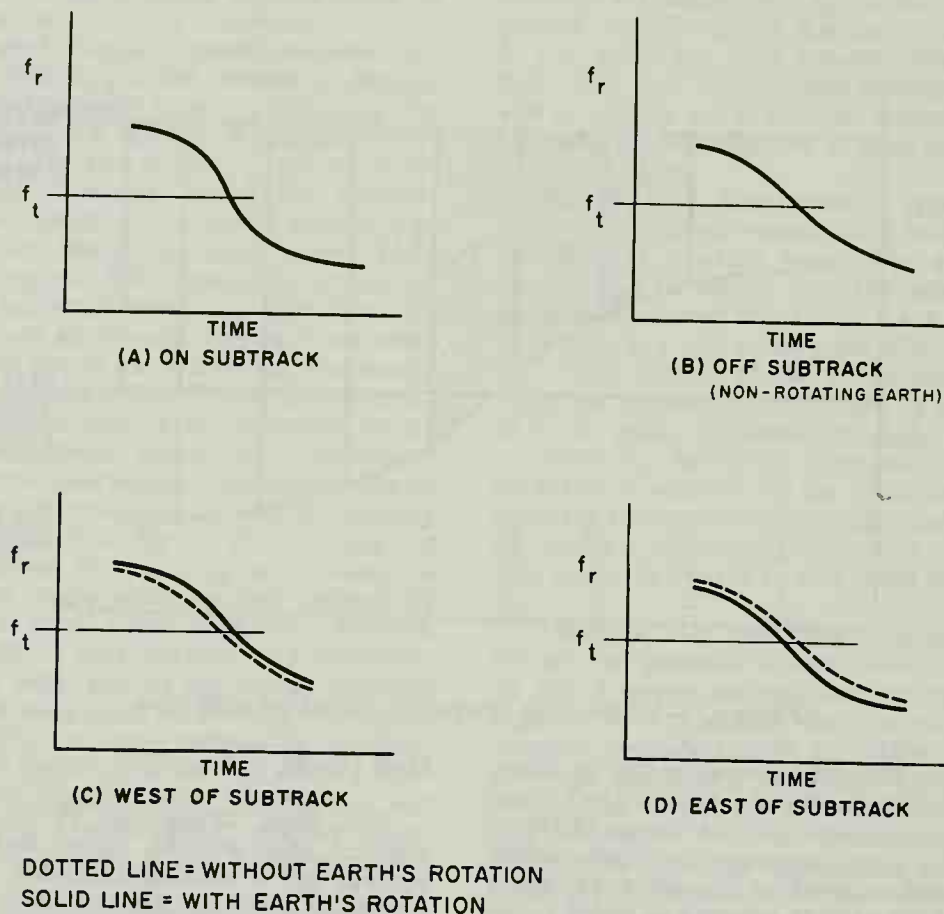
162.63

Figure 15-6.—Doppler frequency response to position changes.

the inputs of the amplifiers, resulting in an improvement in system sensitivity of approximately 3 db. When the preselectors are used, their outputs are connected to the input of the amplifiers. The 400-mc preselector provides discrimination against undesired off-frequency signals.

The 400-mc filter is connected between the second and third stages of the 400-mc amplifiers. Its function is to maintain a relatively constant center frequency and bandwidth characteristics of the amplifier with gain variations from Automatic Gain Control (AGC).

The 150-mc preselector and 150 mc filter are identical units. The preselector provides discrimination against undesired off-frequency signals, while the filter provides a constant bandwidth for the 150-mc amplifier regardless of the AGC voltage applied.

The 400-mc amplifier consists of eight r-f stages and three d-c amplifier stages, which provide AGC and temperature compensation. The amplifier exhibits a nominal gain of 70 db with a 50 db AGC range.

The 150-mc amplifier consists of five grounded-emitter, neutralized transistor stages and three d-c amplifier stages which provide AGC and temperature compensation. The amplifier exhibits a nominal gain of 70 db with a 55 db AGC range.

Electronics Group

The major elements of the Electronics Group (fig. 15-9) are the receiver unit, the data

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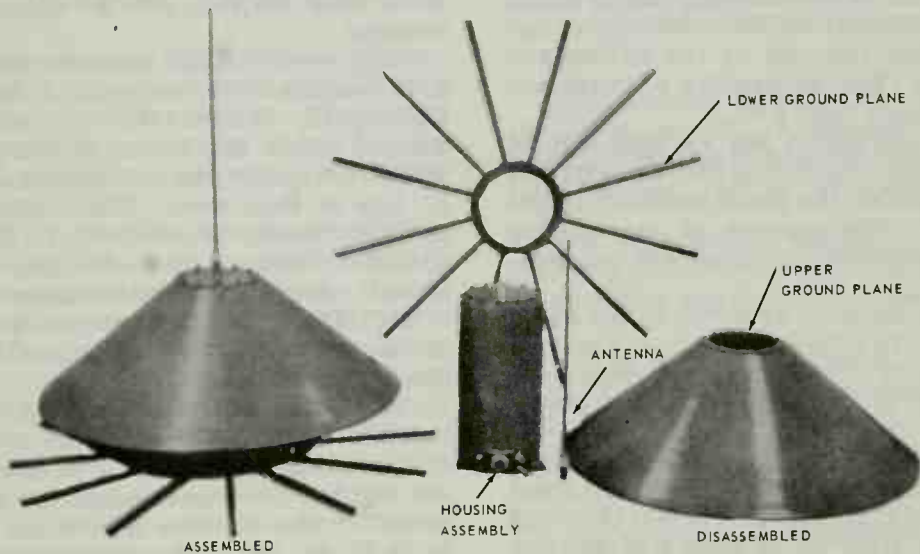
162.64

Figure 15-7.—Mast group.

processor unit, the power supply unit, and the oscillator. Before describing these units it will be necessary to become familiar with the received signal and message organization.

Data is received from the satellite in the form of phase modulation on the carrier
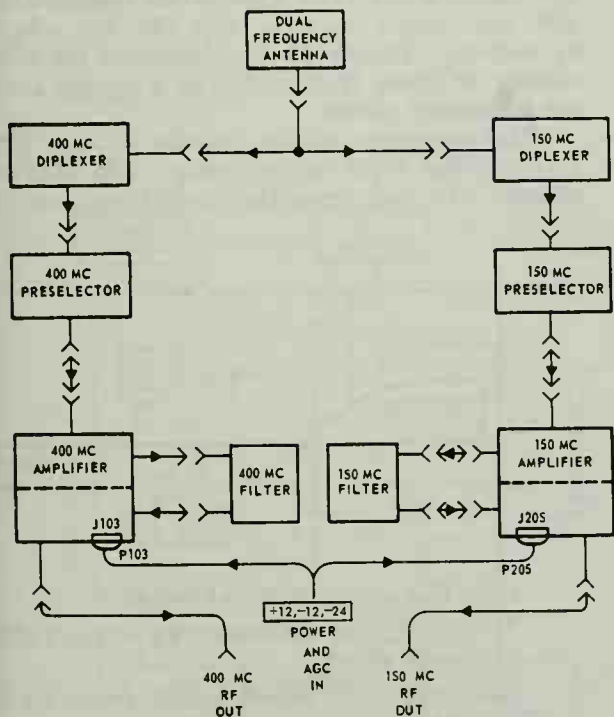
162.65

Figure 15-8.—Mast group, block diagram.
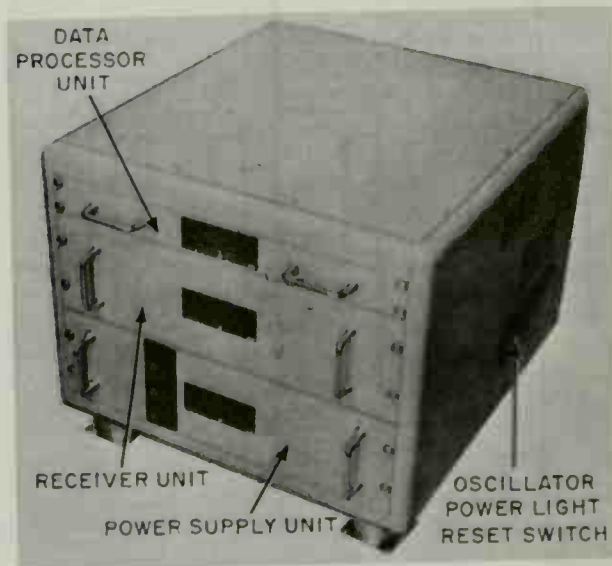
162.66

Figure 15-9.—Electronics group.

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frequency. The data is coded in binary form (fig. 15-10). The basic encoding unit is called a doublet and consists of two intervals of opposite phase shift followed by two intervals of no phase shift. Two successive doublets are required to define a binary bit.

The Radio Navigation Set AN/SRN-9 must make the correct association between adjacent doublets to translate the signal modulation into "logical bits." The process of determining this correct association is called "bit synchronization."

The satellite message received by the Radio Navigation Set AN/SRN-9 in a two-minute interval consists of 6103 binary bits organized into 156 words of 39 bits each plus a 19-bit word.

The last 25 bits of word 2 are a synchronizing pattern consisting of 1 zero, 23 ones, and another zero (0111111111111111111111110). The beginning of the first bit of word 3 is the two-minute time mark. Transmission of the message from the satellite is precisely controlled so that these two-minute time marks are normally transmitted within 1 millisecond of even integar two-minute times as determined by the Naval Observatory. Word 3 can be heard in the head set as a 400-cps tone. The beginning

of this so-called "beep" word may be used as a time mark to set the navigator's chronometer.

Only every sixth word starting with word 8 is required for navigation, thus words 8, 14, 20, 26, 32, etc. are all that will be required. Also the Radio Navigation Set AN/SRN-9 uses only the data contained in the first 32 bits of each word. These are transmitted as eight Binary Coded Decimal Excess Three (BCDXS3) digits. For example, the number 5 would normally appear in binary as 00000101. In BCDXS3 it would appear as 00001000, that is, each number is transmitted three higher than its actual value.

The Radio Navigation Set AN/SRN-9 recovers and records doppler data, satellite data, and timing information from the satellite signal. The higher frequency signal transmitted by the satellite is 400 mc - Δ_H, where $\Delta_H \cong 32$ kc since the frequency is offset 80 ppm. (The low channel likewise 150 mc - Δ_L, where $\Delta_L \cong 12$ kc.) This signal is shifted in frequency by d_H due to doppler effect and ξ_H due to ionosperhic refraction. The range of d_H is between 8 kc and that of ξ_H is between $\pm$ 3 cps. The signals as received are 400 mc $-\Delta_H + d_H + \xi_H$ and 150 mc $-\Delta_L + d_L$ and $+\Delta_L$. Figure 15-11 illustrates the association of these frequencies on a typical Doppler frequency curve.

The receiver is the middle drawer (fig. 15-9) of the Electronics Group. The received signals are fed from the Mast Group as two

162.67

Figure 15-10.—Modulation characteristics in binary coded form.

TCA (TIME-OF-CLOSEST APPROACH)

f_T – SATELLITE TRANSMITTED FREQUENCY = 400 MC – Δ_H

f_R – SATELLITE SIGNAL AS RECEIVED BY NAVIGATOR
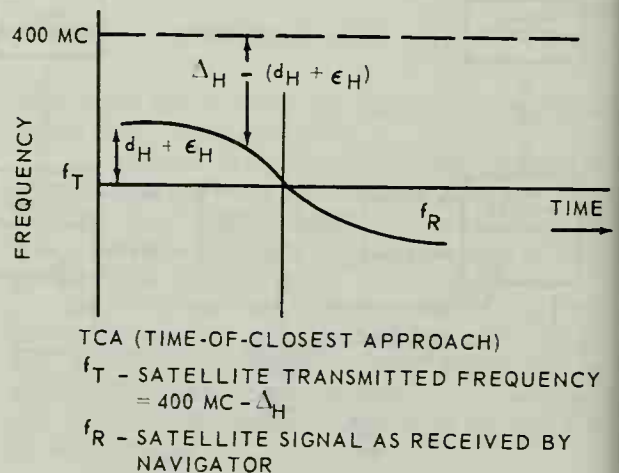
162.68

Figure 15-11.—Doppler frequency curve.

parate signals and consist of 400 mc - Δ H dH + ξ_H and 150 mc - Δ_L + d_L + ξ L, as shown fig. 15-12.

The 400-mc signed is fed to the 5-mc i-f am- ifier assembly, which consists of a 400-mc ixer and four amplifier stages. The mixer btracts the 400-mc incoming signal from a 05-mc local oscillator (LO) signal to produce i-f frequency of 5 mc.

The 5-mc i-f amplifier has an overall gain approximately 100 db and a bandwidth of 3 with AGC being applied to the first two stages.

The 5-mc phase comparator assembly com- ares the i-f amplifier output with frequency able quadrature reference signals. The phase omparator circuit is a fullwave, or balanced ype, which means there is no output of either e i-f input signal or reference signal but only e sum and difference frequencies of the two puts. The phase comparator has two outputs, e proportional to the sine and the other to e cosine of the phase difference between the -f signal and the reference signal.

The 5-mc buffer assembly consists of three dentical buffers and a tripler. The buffers upply 5-mc signals to the 5-mc phase com- arator, the 1.875-mc converter, and the trip- er.

The 16.8-mc voltage-controlled oscillator (VCO) generates the basic frequency from which the r-f mixer reference frequency (405 mc) is derived.

The output of the VCO (16.8 mc) is fed to the 16/101-mc multiplier assembly which synthe- sizes the signals to drive the mixer section of the i-f amplifier and the 101/404-mc multiplier.

The 101/404-mc multiplier provides ref- erence signals for the mixer section of the 5-mc i-f amplifier and the audio mixer.

PHASE-LOCK TRACKING.—Tracking of the satellite signal is accomplished by the phase- lock loop (fig. 15-13) which consists of the 5-mc mixer and i-f amplifier assembly, the 5-mc phase comparator, a phase loop control network, the VCO, the 16/101-mc multiplier, and the 101/404-mc multiplier. The loop op- erates to maintain the 101/404-mc multiplier output frequency at exactly 5 mc greater than the input signal from the satellite or 400 mc - Δ_H + d_H + ξ_H + 5 mc. Once lock-on of the sat- ellite signal is accomplished the phase-lock loop will automatically track the satellite as the output of the 101/404-mc multiplier will

162.69
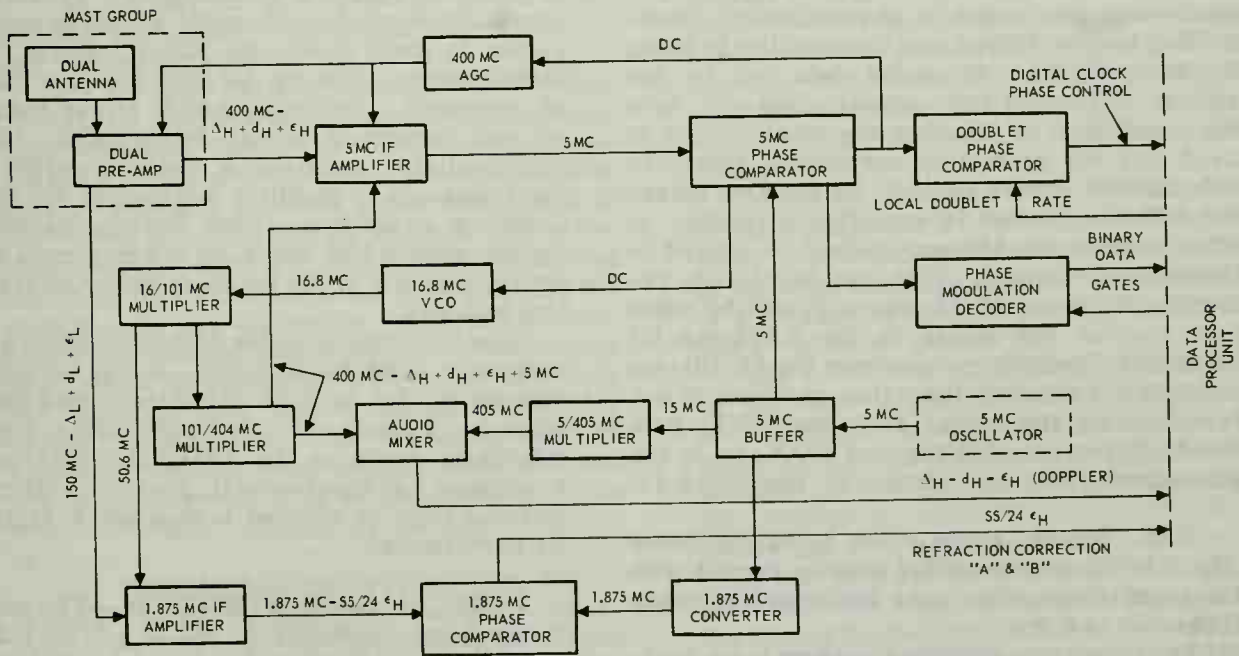
Figure 15-12.—Simplified block diagram of receiver unit.
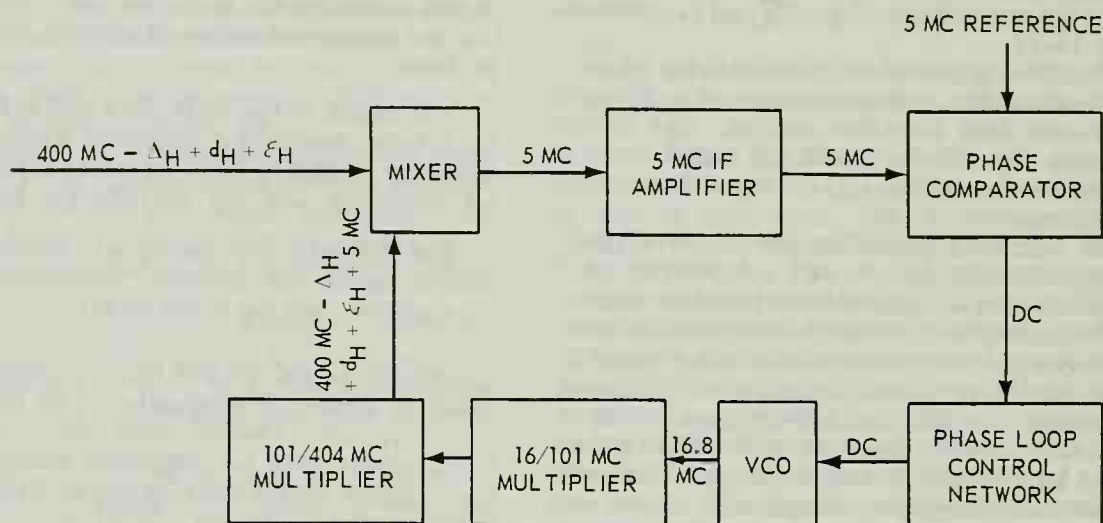
Figure 15-13.—Phase-locked loop block diagram.

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vary in frequency as the satellite signal varies in frequency.

This is possible since the VCO output varies according to the magnitude and polarity of its d-c input, which in turn depends upon the satellite signal and the local reference signal. The phase-lock loop must be in equilibrium, (lock-on has been achieved and the satellite is being tracked), before any useful data can be received. The 150-mc channel does not have the capability of tracking the satellite and is used only for refraction correction, therefore this channel cannot perform its function unless the 400-mc channel is operating properly. In other words the 150-mc channel is slaved to the 400-mc channel. This can be seen by referring to figure 15-12 where it will be noted that one of the inputs to the 1.875-mc i-f amplifier assembly comes from the 16/101-mc multiplier assembly the value of which is determined by the output from the VCO. Both the 16/101-mc multiplier and VCO are in the phase-lock loop as shown in figure 15-13.

5-MC OSCILLATOR.—The 5-mc oscillator (fig. 15-12) is a modified Pierce circuit with the crystal operating near its series resonant frequency of 5 mc.

The frequency stability for short term periods is 4 to 8 parts in 10^{11} and is to 6 parts in 10^{10} per day. The stability that we are interested in is the short term period; the

stability during the actual tracking of a satellite. This is apparent when it is recalled that the navigation fix is based upon the actual doppler count for two-minute intervals.

The oscillator must remain on at all times in order to maintain its stability. Provisions have been made for automatic switch to battery power in case of normal power failure. The nickel-cadmium battery is good for four hours of operation. Following initial power turn-on or any temporary power interruption, the 5 mc oscillator requires a warmup period for the frequency to stabilize sufficiently for high-accuracy navigation. This warmup period is approximately 10H hours, up to a maximum of 72 hours, where H is the length of power interruption in hours.

The importance of the stability of this 5 mc reference must be stressed. It may be well to remember that in case of failure of this oscillator and a ready spare is not available, a good substitute would be the AN/URQ-9 (described in chapter 3 of this manual), provided sufficient warmup time is allowed before any navigation fix is attempted.

DOPPLER RECONSTRUCTION.—The output of the 5-mc oscillator is fed to a 5-mc buffer stage (fig. 15-12) where two separate amplifiers are used to provide inputs to the 5-mc phase comparator and the 1.875-mc converter. Also included in the 5-mc buffer assembly is a

tripler where the 5-mc signal is multiplied to produce an output of 15 mc. The 15 mc-signal is fed to the 5/405-mc multiplier.

The output of the 5/405-mc multiplier is fed to the Audio Mixer which combines this input with the 404.9-mc signal from the phase-locked loop to obtain a nominal 32-kc signal containing the desired doppler frequency. The actual frequency present at the output of the audio mixer is approximately 32 kc and is made up of Δ H-dH-ξH as indicated in figure 15-12.

REFRACTION CORRECTION.—The purpose of refraction correction is to eliminate ξ H. This is accomplished because ξ is inversely proportional to transmitted frequency, whereas Δ and d are directly proportional to frequency. A definite relationship exists between these frequencies in that 150 mc is to 400 mc as 3 is to 8.

The 1.875-mc i-f amplifier (fig. 15-12) has as one of its inputs the low frequency signal (150 mc -ΔL + dL + ξL) with the other input being 50.6 mc derived from the phase-lock loop.

The 1.875-mc signal in the converter is obtained by multiplying the 5-mc signal from the 5-mc buffer by 3/8. This is done by dividing by 8 and then multiplying by 3. The 1.875-mc output is phase locked to the 5-mc input.

The outputs of the 1.875-mc phase comparator are called "Refraction Correction A" and "Refraction Correction B."

The Doublet Phase Comparator (DPC), (fig. 15-12), is a subsystem within the doublet-rate, phase-tracking loop. The DPC compares the phase of the local doublet rate with the satellite doublet rate and provides phase commands to the digital clock in the data processor when the phase error exceeds a certain threshold. The output of the DPC is the same wave shape as the doublet-rate flip-flop in the data processor except it is shifted 90 degrees in phase with respect to it. Phase synchronism is necessary to decode the incoming data and to define the doppler period's end points. When the output exceeds a certain positive value, than an advance or add command is given. If it is more negative than a certain value, a retard or subtract command is given.

The Phase Modulation Decoder takes phase-modulated doublet data from the doublet phase comparator, synchronously detects it, and places it in a binary format compatible with the digital section. Synchronous detection followed by an integration with end-of-bit sampling is used to afford immunity from noise errors. The digital section provides the properly timed gating signals for synchronous decoding.

DATA PROCESSOR UNIT.—The data processor unit (DPU), is the top drawer (fig. 15-9) of the Electronics Group. The DPU organizes the timing signals, satellite orbit parameters, and doppler counts derived from the satellite signals by the previously described sections of the Electronics Group and outputs this information to the computer and to the Control Group for printout. For discussion purposes of the DPU, a binary "one" is defined to be -9 ± 3 volts and binary "zero" to be 0 ± 0.2 volt. A pulse is defined to be a rapid change of state from "one" to "zero" (6 volts/nano sec slope).

Standard medium speed digital circuity is employed in the DPU with germanium transistors and diodes used throughout. All circuitry is standard and will be discussed in block diagram form, by referring to figure 15-14.

The 104-kc oscillator is a two-stage transistor amplifier with feedback through a quartz crystal mounted in an external oven at a temperature of 75 ± 2°C. The amplifier gain is sufficiently high so that the output transistor is alternately saturated and cut off, providing a squarewave output. The output of the 104-kc crystal oscillator is applied to FF01. Flip-flops 01-11 are binary dividers reducing the 104 kc to bit rate. FF10 is the doublet-rate flip-flop with an output of 101.7 cps or a period of about 10 msec. FF11 is the bit-rate flip-flop with an output of 50 cps or a period of about 20 msec.

One of the outputs of the bit-rate flip-flop (FF11) is applied to the digital clock which is made up of FF12-25 which performs the function of counting message bits and words. It consists of a "radix-39 counter", a "radix-6 counter", and a "radix-27 counter".

The radix of a numbering system is equal to the sum of the different characters which are necessary to indicate all the various magnitudes. As used here it simply means the number of bits that are to be counted before the counter is reset to zero state. In other words the "radix-39" counter will count 39 bits, the number of bits in each satellite word, and will then reset to zero to begin counting the bits of the next word. The "radix-6" counter will count the number of words so that only every sixth word is used. The "radix-27" counter counts the number of words that are
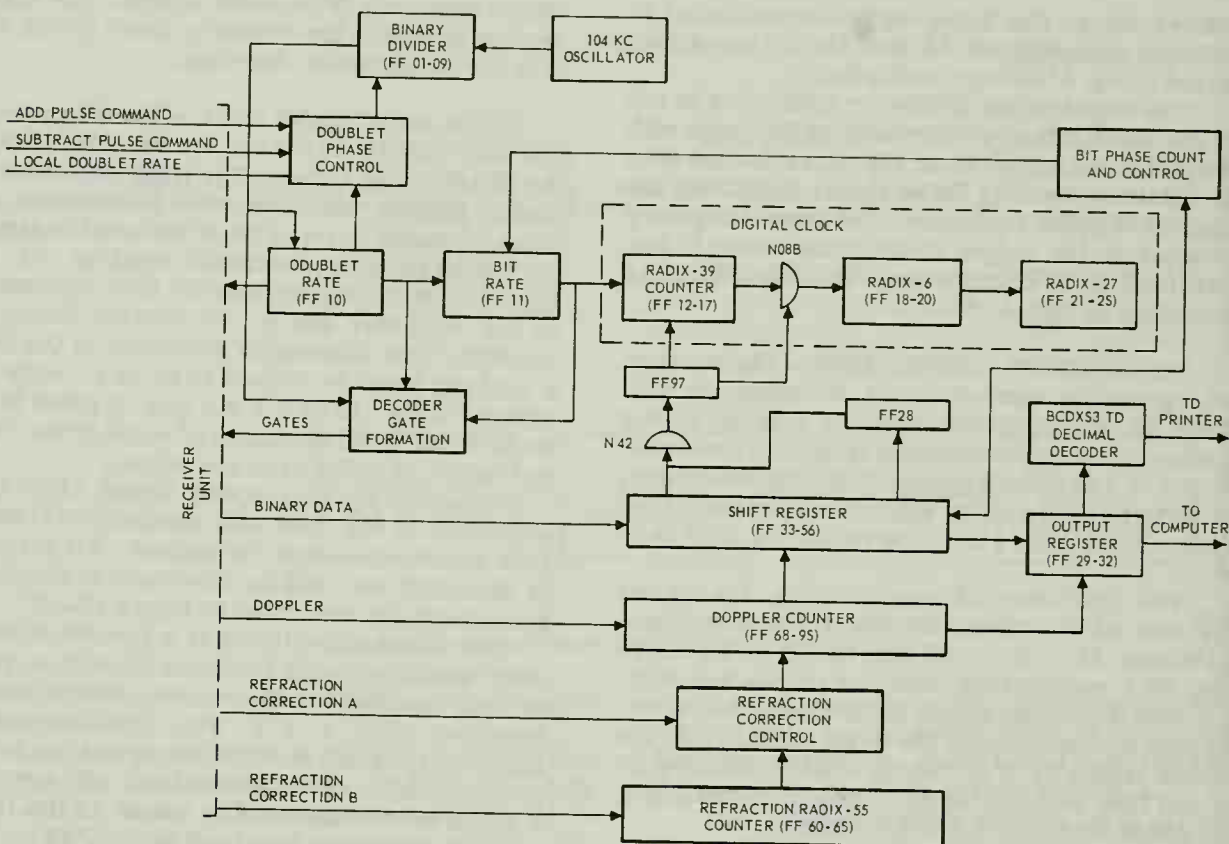
162.71

Figure 15-14.—Simplified block diagram of data processor unit.

used, in each two-minute message. As previously mentioned, only every sixth word is required for a navigational fix, beginning with word 8, with the sequence being 14, 20, 26, 32, etc. to word 152. The radix-27 counter then counts the number of words used, and the entire digital clock is reset to zero before the next two-minute message.

Flip-flops 12-17 form the radix-39 counter. They perform a straight binary count until the 39th state, at which time flip-flops 12, 15, and 16 are set to "one."

One count per word enters FF18 via NO8B. Flip-flops 18, 19, and 20 form the radix-6 feedback counter. When FF20 toggles from "zero" to "one," feedback sets FF19 to "one", thus six is as high as the radix-6 counter can count.

Flip-flops 21-25 form a radix-27 counter. Feedback occurs to set flip-flops 21 and 23 to "one" when FF25 toggles from "zero" to "one."

When the radix-27 counter reaches the 26th state, after six 39-bit words, FF21 is toggled from "zero" to "one" and flip-flops 18, 19, and 20 are immediately set to "one." This prepares the clock to count the last word of 19 bits of the satellite message. When the radix-39 counter reaches the 19th stage, all flip-flops of the radix-39 counter are set to "one". The beginning of the next bit counted by FF11 toggles flip-flops 12-25 to "zero," indicating the beginning of the next two-minute message.

Before any counting is to be done, doublet rate synchronization must be made. This is accomplished by the Doublet Phase Control Assembly. If the DPC in the receiver determines that the doublet-rate flip-flop (FF10) is lagging in phase, pulses are added to FF01. If FF10 is leading in phase, pulses are subtracted from FF01, until the "local doublet-rate" is in sync with the satellite doublet rate.

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Doublet demodulation by the phase modulation decoder in the receiver occurs after doublet rate synchronization. The phase modulation decoder output is applied to the shift register (FF33-56). The serial shift signal enters each doublet into FF33 and shifts the previously entered doublet into FF34. If the doublets in flip-flops 33 and 34 are different, a count pulse is sent to the add/subtract unit which is comprised of flip-flop 49-52 and used as an up-down counter. This count occurs when the satellite bit rate phase to which FF11 must be synchronized is in the "one" state. Once every 39 bits when FF17 toggles from "one" to "zero," flip-flops 49-52 are set to 0111 state (most significant bit given first). Whenever the flip-flops count from 0111 to 1111, a pulse is sent to assure that FF11 is in the "one" state. When FF11 is in the "one" state and FF49 switches from "one" to "zero," FF50 is toggled; and the counter is adding. When FF49 switches from "zero" to "one," FF50 is toggled and the flip-flops are subtracting.

If counts occur when FF11 is in the "one" state, they are added to the preset value of 0111. If counts occur during the "zero" state of FF11, they are subtracted from 0111. In either case eight counts in either direction will drive 0111 to 1111.

After bit synchronization, full bits are demodulated by the phase modulation decoder and shifted through the entire shift register and FF28.

Gate N42 is a 26 input NOR gate connected to flip-flops 33-56 and 28 so that when they contain the 25-bit synchronization pattern (0111111111111111111111110) the gate will be satisfied. The 26th input to N42 is the 26th bit input signal, which can be used to prevent the data processor unit (DPU) from obtaining message synchronization.

When the synchronization pattern is recognized by N42, FF97 is set to "one." When FF97 is set to "one," the following changes take place;

(1) Flip-flops 12-17 are set to the "one state
(2) Count pulses can enter FF18.
(3) No shift pulses can enter FF28.
(4) Message synchronization occurs.

When message synchronization occurs, the last bit of the synchronization pattern enters the shift register. Serial shift pulses occur about 5 milliseconds before each count pulse from FF11. Thus, the count pulse immediately following pattern recognition represents the satellite two-minute mark, or the beginning of word 3. After message synchronization the DPU begins to process data.

The 400-mc doppler signal from the receiver unit is converted to a square wave by a Schmitt Trigger and applied to the doppler counter. Flip-flops 68-95, which form the doppler counter, are divided into seven groups of four flip-flops each. Each group is a radix-10 feedback counter so that the doppler count is accumulated in decimal. The seven decades count in Binary Coded Decimal Excess Three (BCDXS3). Bits enter the decade serially from left to right so are in reverse order. As an example, when flip-flops 72-75 reach the 1100 (9) state the individual flip-flop states would appear as shown in figure 15-15, A.

Since this counter is a radix-10 counter the next pulse toggles FF72 and FF73 to "one" and resets FF74 to "zero." This toggles FF75 to "zero," sending a count pulse to the next decade. The state of the counter would then be as shown in figure 15-15. B. or in the "zero" state, ready for the next count. Figure 15-15,C shows a truth table for each of the 10 states of the radix-10 counter.

Every two minutes under control of the digital clock, the doppler counter transfers its contents to the shift register and resets to "zero." Preparation begins during states 26 and 27 of the digital clock radix-27 counter and it is significant that this reset operation is controlled by the incoming doppler frequency itself. In this

FF72 FF73 FF74 FF75

0	0	1	1

162.72

Figure 15-15A.—Radix-10 feedback counter in the "9" state (BCDXS3).

FF72 FF73 FF74 FF75

1	1	0	0

162.72

Figure 15-15B.—Radix-10 feedback counter in the "zero" state (BCDXS3).

	FF72	FF73	FF74	FF75
0	1	1	0	0
1	0	0	1	0
2	1	0	1	0
3	0	1	1	0
4	1	1	1	0
5	0	0	0	1
6	1	0	0	1
7	0	1	0	1
8	1	1	0	1
9	0	0	1	1

162.72

Figure 15-15C.—Radix-10 feedback counter
truth table (BCDXS3).

way no doppler count can be lost during the transfer and reset operation.

"Refraction Correction A" and "Refraction Correction B" signals are applied to two Schmitt Triggers which will be referred to as ST1 and ST2, respectively. These two signals come from the 1.875-mc phase comparator assembly in the receiver and are in phase quadrature, with a frequency 55/24 times the desired refraction correction rate.

The output of a Schmitt Trigger changes state rapidly each time its input crosses 0 volts. Because noise often accompanies the refraction correction signals as the signal voltage approaches zero, the signal plus noise voltage may cross zero several times. However, only one refraction correction count must occur.

Two outputs from each of the Schmitt Triggers are in phase opposition, so if the input of ST1 crosses zero volts in either direction it will set a flip-flop to "zero". Similarly, ST2 sets the same flip-flop to "one." Thus, the first count pulse from ST2 sets the flip-flop to "one," preventing a count until ST1 resets the flip-flop to "zero." In this way only one count pulse from the Refraction Correction B signal can enter the Refraction Counter until the Refraction Correction A signal again crosses through zero.

Flip-flops 60-55 (fig. 15-14) form a radix-55 feedback counter. The feedback sets flip-flops

60 and 63 to "one" when FF65 toggles from "zero" to "one." With a total of 55 pulses to FF60, the output of the radix-55 counter will be 24 pulses.

This output is taken from the set side of FF60. By utilizing the feedback, FF60 is in the "one" stage 24 times out of 55 input pulses. These pulses are applied to the refraction correction control and during this time are added to or subtracted from the doppler count in the doppler counter.

The primary data output from the DPU (Data Processor Unit) is from flip-flops 29 to 32 (output register). All doppler count and satellite message data is shifted, four bits at a time, into these output flip-flops.

In preparation for the doppler count transfer, the shift register and flip-flops 29 to 32 are filled with binary "zeros". Because the shift register is filled with "zeros," only the binary "ones" need to be transferred from the doppler counter. At time of transfer the doppler count in flip-flops 68 to 91 is put into the shift register while the count in flip-flops 92 to 95 is transferred directly to the output register. This transfer takes place just after the digital clock resets to all "zeros" at which time the two-minute mark timing signal is outputed. The shift register is delayed for four-bit times allowing time for the doppler count digits transferred directly to the output register to be decoded and entered into the printer.

After the four-bit delay the contents of flip-flops 33 to 56 are transferred into the output register and the serial shift pulses begin again. After each set of four bits has been shifted into flip-flops 33 to 56, it is then transferred out. When this process is completed satellite data is again allowed to enter the shift register.

Every bit received from the satellite is entered into the shift register except during the preparation and processing of doppler count data. The output transfers this data into the output register four bits at a time as previously discussed.

The BCDXS3 decoder module is a diode matrix which converts binary-coded inputs to decimal outputs. Certain other signals are provided as outputs of the DPU for use by a navigational computer, paper tape punch, or other device. These are end-of-word, digit rate, bit rate, bit sync signal, two-minute mark, and serial data.

Control Group

The Control Group (fig. 15-16) contains a number of interacting circuits and performs switching functions. The entire navigational fix can be monitored and controlled from the Control Group and associated computer.

Positions 1, 2, and 3 of the ten-position main control switch (S01) are used to perform a test of the power supply voltages; +12 volts, -12 volts, and -24 volts, respectively. If these voltages are correct, 50 microamps will flow through the 100-microamp meter, M01, and the needle will point to the green line at midscale.

Position 4 of switch S01 puts the navigational set into the search mode. Circuits involved in this operation are:

(1) Frequency search control
(2) Audio monitor
(3) Frequency monitor
(4) Doppler acquisition light

To tune the set, the operator uses two single-pole double-throw frequency search switches which are located under the FREQUENCY SEARCH bar. When the right side of the search bar is depressed, the set is tuned towards a higher doppler frequency by connecting -12 volts to the "manual tune" function. When the left side of the bar is depressed, 12 volts are connected to the "manual tune" function and the set is tuned towards a lower doppler frequency. When neither side of the bar is depressed, a small positive bias is applied to the manual tune" function. This small bias tunes the receiver to a doppler frequency of approximately 26 kc, which is near the frequency at which a satellite signal should first appear. The "manual tune" function circuitry is external to the control group and consists of resistors and capacitors which are used primarily to develop different time constants for the tuning bias which is then applied across the VCO voltage-variable capacitor.

A jack on the left side of the Control Group (not shown) provides an audio monitor signal to high-impedance headphones. This signal is the output of the 5-mc phase comparator monitored through a 100K ohm resistor. When the set is tuned near a satellite signal, the difference frequency may be heard in the headphones. The operator then tunes the frequency until a zero beat, or frequency null, is heard, indicating the receiver is in phase with the satellite signal.

Meter M01 has two scales as illustrated in figure 15-16. In the search mode the top scale of M01 displays the doppler frequency to which the navigation set is tuned. By this means the operator is guided to search for the satellite signal only in the frequency region where it is expected. The scale runs from 22 kc to 42 kc, which is wide enough to accomodate all possible satellite frequencies.

The Doppler ACQ (Acquisition) Light is another aid to the operator in locking on a satellite signal and is located in the upper right-hand corner of the Control Group panel. When the receiver is tuned to within approximately 30 cps of the satellite signal, the light begins to flash at that difference frequency. Thus, the light provides a visual tuning indication to supplement the audio monitor signal.

In the Pause Mode (Main Control switch S01 in position 5) the phase-lock loop in the DPU and Receiver is closed and the frequency search bar will have no effect. The signals reaching the Doppler ACQ light change from 30 cps to 1 cps. Because this signal is derived from the 5-mc phase comparator output, the light comes full on and stops flashing if the receiver is phase-locked to a satellite signal.

When the Main Control Switch is in position 5 (Pause Mode), the meter displays a measure of the phase error between the double rate flip-flop in the data processor unit and the doublet modulation rate received from the satellite. When the data processor unit achieves
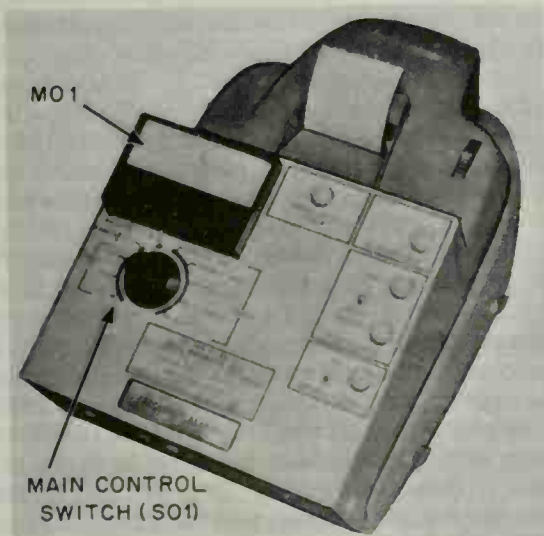
162.73

Figure 15-16.—Control group.

phase-lock, the meter needle will vary about the green center mark.

Positions 6-10 of the Main Control Switch (S01) comprise the track mode. In this mode the receiver tracks the satellite signal, and the data processor processes navigational data. These functions are monitored by the meter and by observing the output of the printer. While the receiver remains locked to the satellite signal, the only control functions are to select whether or not to use the refraction correction signals and, if necessary, to reject a data processor unit synchronization.

The Refraction Rate Light and Switch (fig. 15-16) gets its input from the 1.875-mc phase comparator and is the Refraction Correction B signal. It is a 3-volt peak-to-peak sinusoid with a frequency 55/24 times the refraction error frequency at 400 mc. Thus, the REFRACT CORR RATE light should flash at a rate seldom faster than 10 cps if the 150 mc satellite signal is being received properly.

The refraction control switch (fig. 15-16), allows the operator to select whether or not the data processor unit performs the refraction correction operation. If the REFRACT CORR RATE light is flashing in an erratic manner, it indicates that only noise is being received through the refraction channel, and the refraction control switch should be turned off.

When the Main Control Switch is advanced beyond position 5, the data processor unit is taken out of the high-speed bit track mode, and the "start bit sync" signal allows the data processor unit to achieve a bit synchronization. When this occurs, the DIGIT light (fig. 15-16), goes on. This whole process normally occurs within seconds of advancing S01 to the track mode.

Within two minutes after the DIGIT light comes on, the data processor unit will receive from the satellite a synchronization code. If the code is recognized, the data processor unit will achieve message synchronization and begin to process data. The MESSAGE light (fig. 15-16) will come on, and the printer will begin to print data.

In position 6 and 7 of S01, the two AGC voltages in the receiver are monitored. Position 6 is for the 400 mc AGC and position 7 for the 150-mc AGC. In each case the actual AGC voltages are monitored through 100K ohm resistors located in the receiver. Thus, full-scale deflection represents 10 volts of AGC, on the lower scale of meter MO1.

Another monitor function of the meter is the acquisition correlation function. This is the output of the 5-mc phase comparator monitored through a 28.7K ohm resistor. When this signal is at the correct value of +15 volts d-c, the meter will read center scale at the green line. The 400-mc AGC circuit is adjusted to maintain this condition.

DIGITAL COMPUTER CP-827/SRN-9

The navigation system measurements consist of three successive baselines. The satellite message is continuously transmitted during two-minute time periods, each message starting precisely on the even minute mark. A minimum of six full minutes, or three complete satellite messages, is required to calculate these baselines. Additional periods of received satellite transmission will increase the accuracy of these computations. After measuring the doppler shift and decoding the information content of the phase modulation, the computation may be performed in one of several methods.

Hand computation may be made to obtain a position fix, although the complexity of such calculations leads to hours of computational time and an almost certain probability of human error.

A general-purpose computer may be used, although a buffer unit is necessary to process and convert information into the appropriate computer format. The availability and the time sharing aspects of a general-purpose computer, along with the adaptive buffer unit requirement, make this a less desirable choice.

A special-purpose computer, designed specifically to operate with the AN/SRN-9 Navigation Set, can accept data on the format of the received signal. Such a computer must be able to store the results from a single satellite pass, and accept ship's information of estimated longitude, latitude, velocity north, and velocity east. It then must use the satellite and ship's position data to compute the ship's position fix in longitude, latitude, and coincident time. To assure the continuity of the satellite data, the special purpose computer should be able to sense any loss of signal in the lock circuits of the receiver and, when a loss of signal in these circuits is detected, reject information for the two-minute transmission interval as invalid.

The CP-827/SRN-9 is designed as a low cost computer, requiring a minimum of space, and permitting rapid calculation of a position fix to

meet the immediate needs of the Navy Navigation Satellite System.

In most cases, this computer will not render a "wrong" fix; when a malfunction is present, the computer will not progress through its sequence of operations, whether in response to taped program commands or to operator-controlled commands. Although the same basic circuits are used to execute these operations, an early identification of a source of trouble by isolating a malfunction to a distinct mode of operation permits a more rapid assessment of difficulty by the troubleshooter. To this end, a Test Device is furnished with each computer to give visual display of the binary content of all computer registers by means of a panel of lights in marked groupings.

The satellite message received during a satellite pass is cabled directly into the computer memory to a preprogrammed storage location. A taped program reduces the satellite data to a form which, when manual entry of ship's data has been added, will compute a position fix in a matter of seconds.

The final computations, as well as certain interim data, are printed out on the paper tape of the receiver control group printer. The fix is printed in decimal figures of longitude, latitude, time, and offset frequency.

Equipment Description

The CP-827 is a self-contained computer with three slide mounted drawers. Figure 15-17 shows the external view of the cabinet, its controls, switches, lights, and the tape feed and spools.

The computer consists of four major assemblies; Drawer I, Drawer II, Drawer III, and a Rear Door Assembly. The three drawers have front-mounted handles for their access and removal for maintenance.

Drawer I, the top drawer (fig. 15-18), has a blank front panel. Internally, the drawer is comprised of three rows of printed circuit board multipin connectors. Each row can accommodate up to 27 printed circuit logic cards.

Drawer II is the middle drawer of the cabinet. Most of the controls and indicators of the computer are mounted on the front panel of this drawer (fig. 15-19). An eight-position QUADRANT SELECT rotary switch allows selection of one of four quadrants in each of two planes of the core memory. A ten-position DIGIT SELECT rotary switch permits selection of any

digit from 0 through 9. Four incandescent lamps monitor the outputs of the manual entry flip-flops, (W01, W02, W03, and W04), to give visual indication of the sequence of operator's data entry into the computer. Five pushbutton switches control the STEP Y, TRANSFER, PRINT, RESET, and START functions. A single-pole double-throw READ-STORE switch allows either readout of memory or storage of information. The AC POWER ON micro-switch, when depressed, gives an indication that power is being applied to the computer. The ENTER DIGIT microswitch permits entry of the digit selected by the rotary DIGIT SELECT switch, and gives indication that the digit has been entered. The J7 ON-OFF is a two-position microswitch for turning on and off the J7 operational code, which controls whether instruction is to come from tape or from the stored program. The lighted portion of the switch indicates the on or off position. This switch should be left in the position where the ON or OFF condition may be controlled by computer operation.

Drawer II is almost identical to Drawer I in internal construction, with three rows of printed circuit board connectors, each row accumulating up to 27 cards.

Drawer III (fig. 15-20), the bottom drawer of the cabinet, contains the heavier computer components, such as the power supply and tape reader assembly. The tape reader is mounted on the front of the drawer. The stacked power supply inside the drawer furnishes regulated d-c voltages of + 12, +19, -6, -8, -12, -19, and an unregulated +18 volts d-c. Potentiometers are provided for each of the individual power supplies and are adjustable over $\pm10\%$ range. A zero to 25 VDC voltmeter and a rotary selector switch are mounted inside the drawer to monitor each of the power supply outputs. In the rear of this drawer is a 115 VAC, 50/60-cycle single-phase blower motor. Input line voltage to the computer is 115 ± 10 volts, single-phase, 60 cps power.

The Rear Door Assembly (fig. 15-21), has a four-part receptacle for the cabling harness of the test device. A horizontal multipin connector underneath this harness is for receiver-computer hookup. The blower exit vent is at the bottom of the assembly.

Printed Circuit Logic Cards

The cards used in the drawer of the computer are printed circuit boards with all circuit
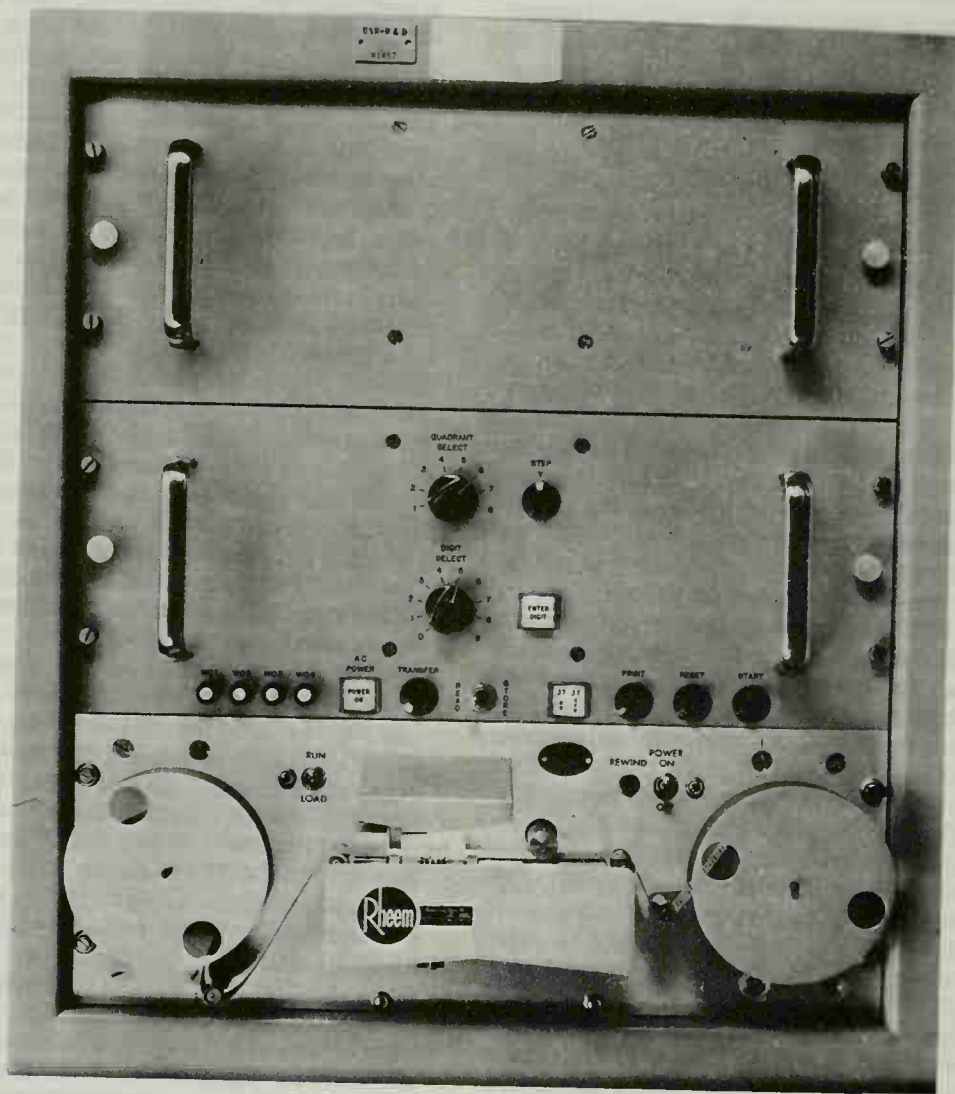
Figure 15-17.—Digital computer CP-827/SRN-9.

162.74

boards having the same card, or part number, being interchangeable. However, certain cards have potentiometer adjustments which may require realignment if they are interchanged.

Silicon semiconductor components are used on all cards. To achieve a minimal noise level on the d-c power distribution lines, the power lines on each card are bypassed with individual capacitors.

The thirteen types of cards in the computer comprise the modular circuitry to perform all of the operational commands by the three types of

input (receiver signals, taped instructions, or manual entries).

Tape Reader

The tape reader mounted on the front of Drawer III (fig. 15-17), is a Rheem Model RRS-102 Reader-Spooler combination. It uses an 8-channel punched tape with a reading speed of approximately 10 inches per second. Operating as the input medium, the function of the tape reader is to convert information from its

The light source is a horizontally mounted cartridge lamp with an average life of 4,000 hours. Care must be taken to properly focus the lamp and to ensure that no obstructions, such as dirt, incomplete tape perforations, etc., cover the photocell block apertures.

Receiver Data

The satellite data received during a satellite pass is fed directly into the computer memory to preprogrammed storage locations. The computer automatically interrogates the lock circuits of the receiver during this time and rejects as invalid any message interrupted by a loss of lock in the track. If a sequence of three, full two-minute satellite messages, is not received during a single satellite pass, the computer cannot compute a fix, and the printout will indicate (prior to manual entry of ship's data) that no fix is possible.

Satellite data is transmitted in BCDXS3 format. Any time that the printer of the control group is activated, the computer must operate through the decoder matrix of the receiver. Thus, all data injected into the computer via the receiver must also be in this format. As a 32 bit word is injected from the receiver, the computer is sequenced to accept each four bit series and shift its position to accept the next four bits. All computer data are in this serial binary form, with all numbers having a magnitude less than unity. The binary point is located between the 31st bit and the sign bit, which is in the 32nd position or the left-hand end of the printout. All numbers are stored as absolute value plus the sign bit.

162.75

Figure 15-18.—Drawer I, top view, showing logic cards and multipin connectors.

perforated form to electronic signals. As the tape passes under the reading lamp, the holes are recognized by means of photoelectric cells. Each cell is gated with certain logic circuits amplified, and sent to the output receptacle for external routing. The switches on the panel of the tape reader consist of two toggle switches, for power ON-OFF and tape RUN and LOAD conditions, and a pushbutton for the rewind condition of the tape spool.

162.76

Figure 15-19.—Drawer II, front view, showing control panel.
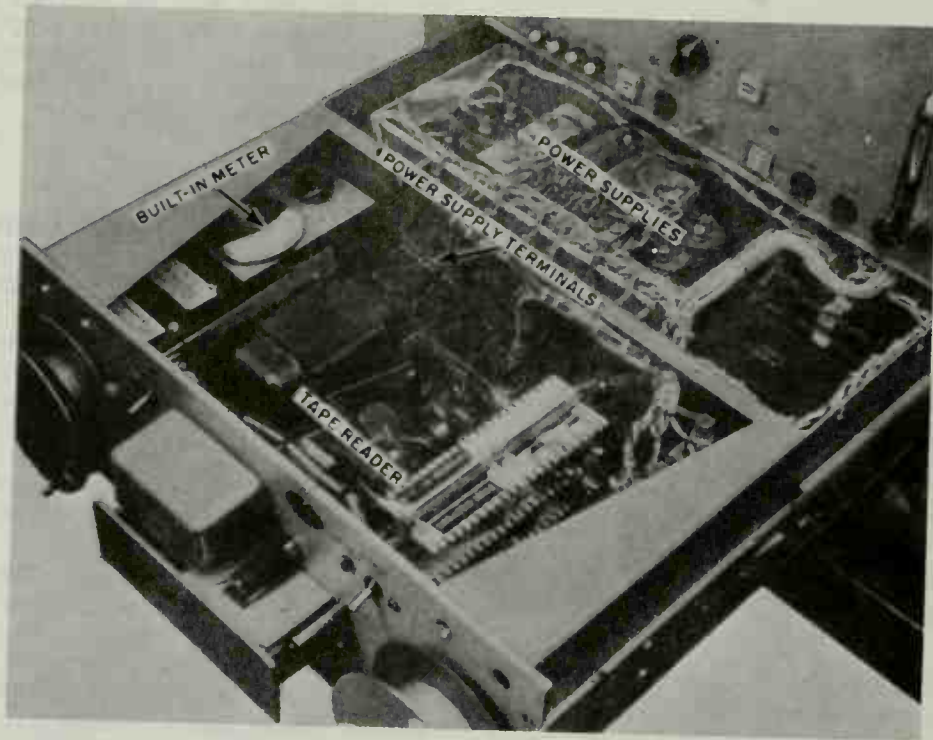
Figure 15-20.—Drawer III, open view, showing interior tape reader and power supply sections.

162.77

Satellite data are arranged with the least significant bit in position 1 of the printout, or the right-hand end of the word. The receiver lines feeding the computer are arranged to inject each series of data correctly into bit positions with the most significant bit in position four, and then these are end-around shifted for the computer to accept the next four bits.

The conversion from BCDXS3 is made in the computer prior to the arithmetic operations to simplify the logic. However, output must be reconverted for printing, since the printer must again operate through the receiver's decoder matrix. The Z register, through which all data must pass either entering or leaving the computer, will therefore display its content in XS3 format on the Tester panel lights prior to printing. Likewise, manual entries made in decimal form from the computer panel will be displayed in XS3 form on the (Z-register) Tester lights, although the printout will be in the same decimal form as used for entering the digits.

Core Memory

The computer utilizes two identical ferrite core memories with storage capacities of 128 words of 32 bits each. Each memory consists of the ferrite storage cores arranged in a matrix which is subdivided into four quadrants of 32 X 32 bits. The cores are threaded by three wires to give an X drive, a Y drive, and a bias-digit drive. Each quadrant also has its own sense line feeding a single sense amplifier. The 32 X and the 32 Y drive lines pass through each of the four quadrants of a memory plane. For any given memory cycle, one of the four quadrant lines will be true while the other three are false. During this time, the cores of the three quadrants not selected receive bias current, while the selected quadrant cores receive no bias current. After application of bias current, drive current is applied to one of the X drive lines, in opposition to the bias current. The Y drive current is then applied and only the core at the intersection of the X and Y drive lines in the

530
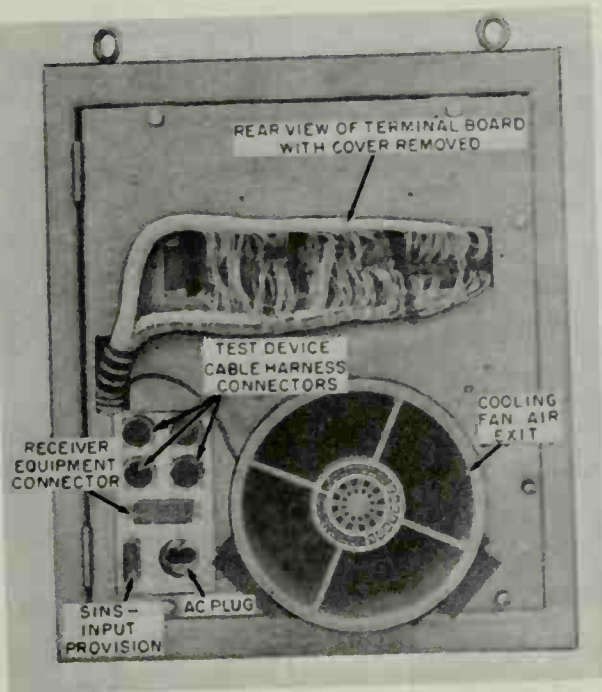
162.78

Figure 15-21.—Rear door assembly showing connectors, cooling fan, and terminal board (with cover removed).

selected quadrant receives full current and is switched.

The current steering switch techniques employed provide simplicity for X and Y drive selection and are designed to minimize dissipation in the current stabilization circuits and to provide greater timing tolerance. Sense signals of approximately 45 millivolts occur for each one bit output. A memory cycle consists of two beats, a read or clear and a restore or write beat. Each beat requires 4 microseconds, providing memory cycles of 8 microseconds each.

Whenever the content of a memory cell is scanned, the contents are read without changing the values within the cell.

Instruction Codes

Instruction codes to program the computer may be generated either from the punched tape or from the internally stored program of the memory. Operation from the control panel by the operator can drive words into memory storage, transfer words from memory and obtain a printout. Whether the taped instructions are

controlling the operation of the computer, or whether the manual controls are being used to enter data into the computer, the reset button of the computer must be pushed to generate a signal causing the proper sequence and logics to reset and wait for further instructions.

The start button then shifts the computer to a position where it awaits taped instruction. Four characters on the 8-channel tape define a 32 bit instruction. The first 8 bits of the instruction are used for the operation code. The second, third, and fourth groups of 8 bits each are used for the A, B, and C operands, since the CP-827 uses three address computer logic. The timing is based on the use of a stored program, and the variations of time with different instructions are due to the difference in time required for address decoding, according to the position in the memory which is being addressed. By placing the primary working storage in the small addresses of temporary storage registers, it is possible to keep the timing at a minimum for the bulk of a computing program. For this reason, eight special registers are set aside in the memory for use in performing the various arithmetic and logical operations, and to provide temporary storage of instructions.

Block Description

The organization of the computer is shown in block diagram form in fig. 15-22, and the front panel of the Test Device is shown in fig. 15-23.

Three types of input are shown in blocks A, B, and C, (fig. 15-22). These are the manual data entered at the control panel, the taped instructions at the tape reader, and the satellite data via the decoder matrix of the receiver. The output to the printer, Block E, is via the same decoder matrix of the receiver.

All information is time gated on passing to or from the Z register, Block 1. The Z register is the accumulator, the heart of the computer. All inputs and outputs pass through the Z register, which receives and sends signals from and to the adder.

The J register, Block 2, is the storage register for operational instructions. It controls the adder and the sequence of F states that are to be followed for each instruction.

After an operational code is sorted in the J register, the first two addresses are obtained, the operation is performed, and the results are stored. Then the third address is obtained and

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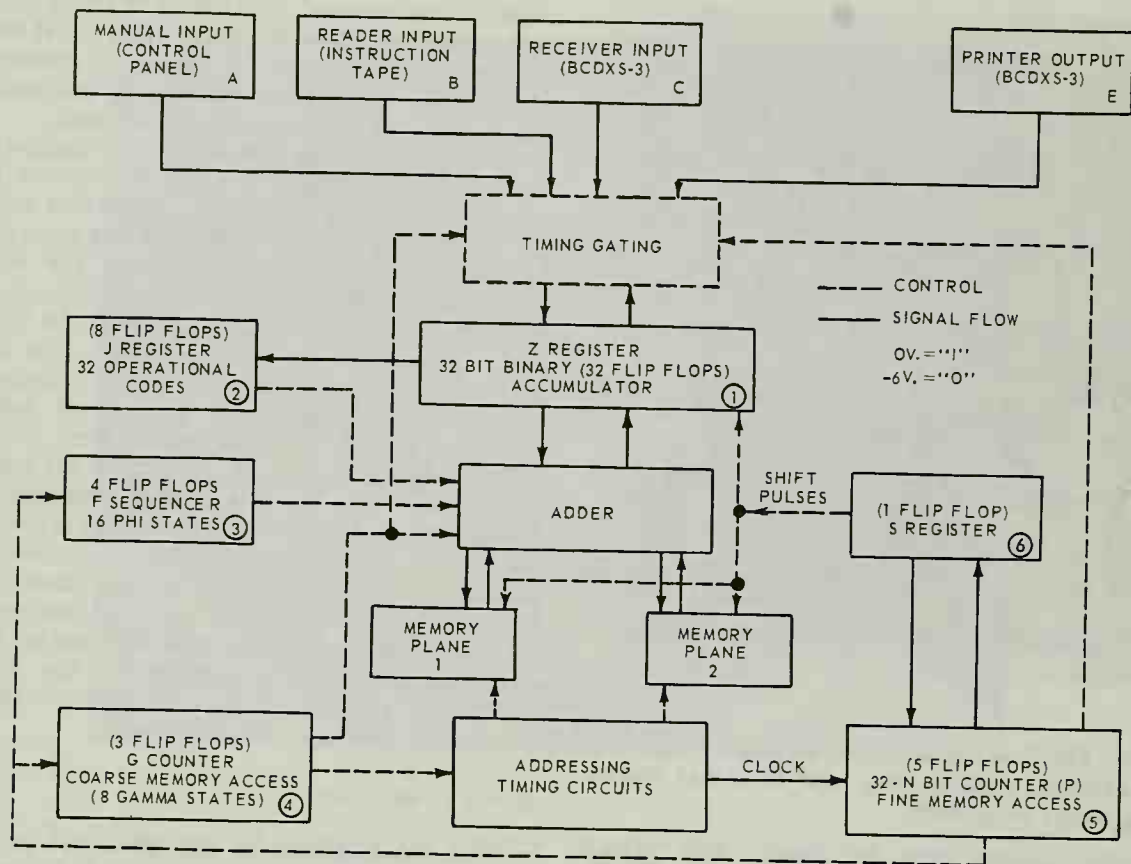
162.79

Figure 15-22.—Block diagram of computer organization.

decoded to obtain the cell into which the results are to be placed.

The F-register, Block 3, is called the "sequencer" and is actually a coarse control to sequence the computer to perform the desired operation while other registers in the computer derive the discrete timing and control necessary for the operations.

The G-register, Block 4, consists of three flip-flops and is used in 11 of the 16 F sequencer states. The G register, called the gamma counter, is used to synchronize the computer with unclocked input data and it controls finer timing for memory access. The G counter sequences, in a particular F sequencer state, to enable word timing signals for setting the memory addresses to select a word and to relocate this word as desired. In the F sequencer states where the G counter is used, the counter will not sequence to a new F state until the G counter

has sequenced properly. The eight states of the gamma counter are displayed on the Tester by the eight lights labeled γ.

Timing is provided in the computer by the clock system and the P counter. The basic clock pulse has a repetition rate of 125 kc and this rate is derived from a free running 1-mc oscillator. Elements of the P counter synchronize the clock pulse and also provide a means for inhibiting the clock. The five flip-flops of the P counter, Block 5, count a standard word length of 32 bits, as well as the 8-bit timing used in address manipulation and the 4-bit digit data.

Although the standard word is 32 bits, it may vary to a shorter word, as when tape data is to be entered or when four bit shifts are to occur. By setting the initial P counter state at the complement of the number to be counted, the P counter then acts as a down counter. For address manipulation, the complements of the address
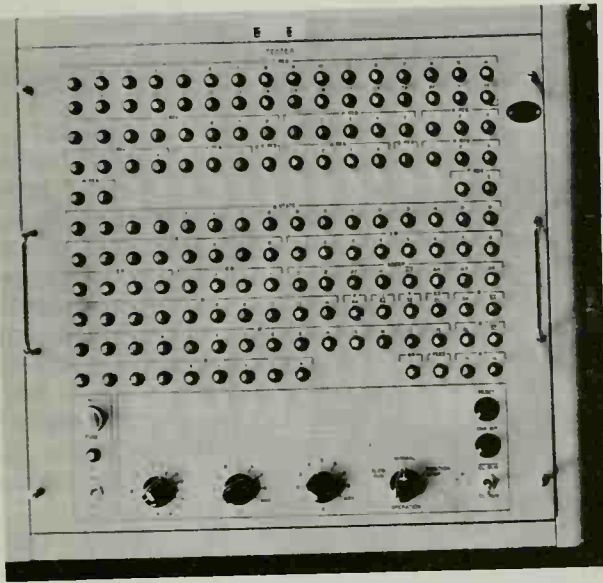
162.80

Figure 15-23.—CP-827/SRN-9 test device.

are injected. When this counter overflows, it signals the computer that the correct word length has been counted and turns off the S flip-flop to inhibit any further manipulation at this time. The P counter remains in the 32nd bit time for two full bit times during which logical functions are performed for correct computer sequencing.

The S-register, Block 6, is the main timing flip-flop which transmits shift pulses to the Z-register and the memory planes, as controlled by the P counter. The S counter remains off for either one or two pulses, during which time the memory drive circuits are reset and the F and G counters are reset to new states. When the Z-register is to be shifted, the P counter controls the S-register action to cause the shift. When the P counter overflows, this causes the S-register to turn off. Therefore, the S-register is on for the number of bit times conditioned by the state of the P counter. The F- and G-registers do not change their states until after the S-register has been turned off, allowing at least one clock pulse for sampling the end result of all states. This S flip-flop controls the memory by inhibiting or enabling the drive circuitry. Since all memory outputs must pass through the Z-register via the adder, the Z-register must see the same shift pulses as the memory. These are the pulses that drive the memory.

The adder controls all access to the memory and performs various arithmetic operations required by the operation instructions.

The W register (front of control panel in fig. 15-17) consists of four flip-flops, W01 through W04 which are used to drive indicator lights. Their primary purpose is to give a visual indication to the operator entering the ship's data or manual data. At this time, the lights are set and reset in such a sequence that by reading the binary number shown by the W01 through W04 indicator lights, the operator can determine what has been entered manually in the navigational program.

AN/SRN-9 RF TEST SET

The AN/SRN-9 RF Test Set is designed to provide fault isolation in the AN/SRN-9 Electronics Group and to provide a means of checking the AN/SRN-9 Mast Group. The AN/SRN-9 RF Test Set provides simulated satellite signals on 400 mc and 150 mc. The 400-mc output is phase-modulated to produce a test message.

Physical Description

The RF Test Set (fig. 15-24) operates on 115 VAC, 60-cps power. It also operates on battery power when it is required to the check the Mast Group. It is composed of two major sections: One side of the set contains the Modulation Switch, the fuse holder and voltage indicator lamps, and three pushbutton switches for BIT, MSG, and WORD synchronization. The remaining jacks and phase delay adjustments are for use with other test equipment and may be disregarded for purposes of this discussion. The major r-f test hookup points, selection dials, and attenuators are contained in an r-f section which can readily be removed for battery operation when it is necessary to climb the mast to check the Mast Group.

The selection dials are as follows:

1. POWER—3 positions, for OFF, INTernal, or BATtery.
2. DOPPLER—INTernal OSCillator selects 32 kc and EXTernal OSCillator selects 26 kc or 52 kc.
3. ATTENUATORS—10 db steps for both the 150-mc and 400-mc outputs. A jumper cable is furnished to hook either output into the center attenuation
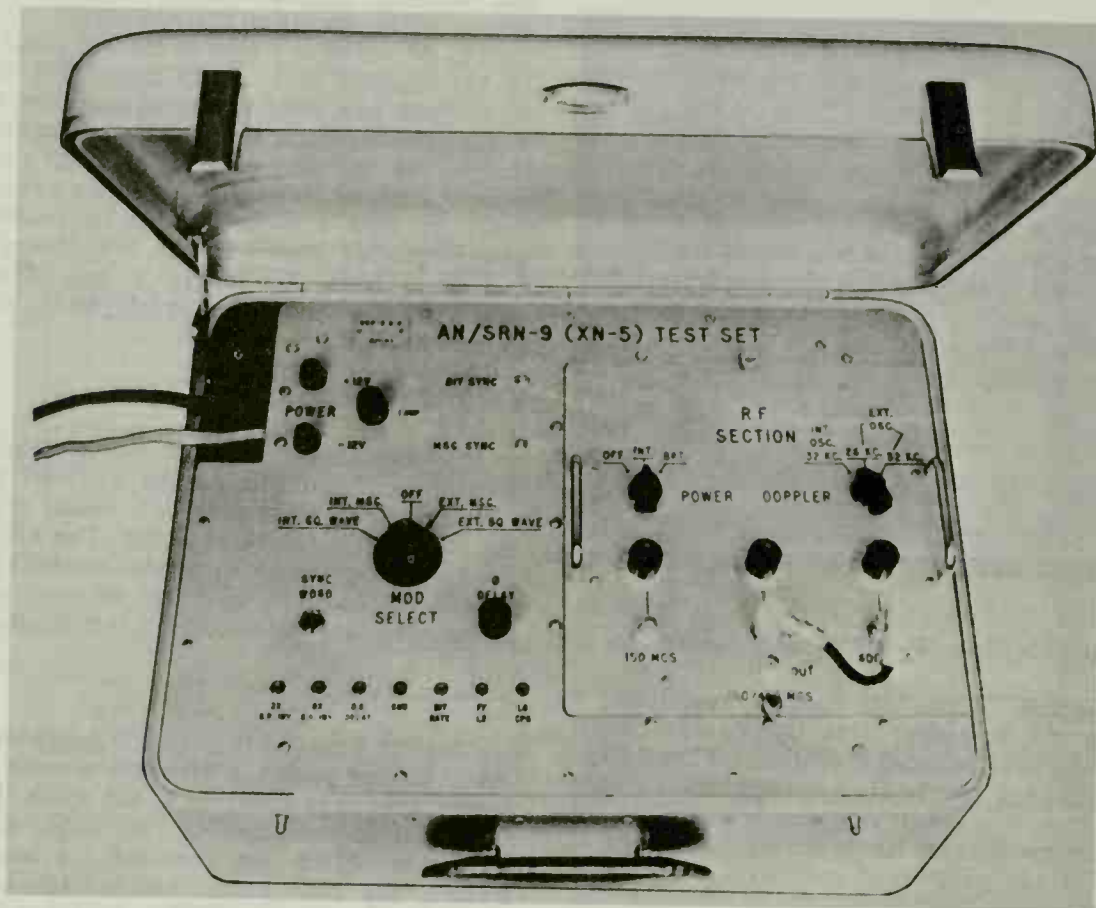
162.81

Figure 15-24.—SRN-9 RF test set panel controls and switches.

3. ATTENUATORS
Continued

selector which is in 1 db steps and thus permits finer selection of attenuation characteristics.

Four basic modes of test hookup can be used to isolate basic operational malfunctions. The sequence of these tests is as follows:

1. Readiness Check—prints a test message, variable doppler count.
2. Radiation Check—prints a test message, variable doppler count, varying AGC.
3. Mast Check—no test message, no doppler count.
4. Doppler Purity—prints test message and fixed doppler count.

Readiness Check

The purpose of this test is to check the ability of the Electronics Group to lock on a test signal and decode a test message. The RF Test Set will furnish a simulated satellite signal to the Navigation Set. Therefore, the procedure for acquisition and locking onto the signal is identical to that used for a satellite pass. The printout obtained from this check is a predetermined message of alternating words of zeros and nines, followed by a message of threes and sixes. Each two-minute message will consist of one of these combinations and will end with a doppler count which will not be stable for this check and may be disregarded. If, as a result of this check, all message printout is normal, it can be assumed that all sections of the

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Electronics Group are functioning properly and any loss of sensitivity will require a Mast or Radiation Check.

Radiation Check

The second check that can be performed is a radiation check. It renders a variable doppler count and varying AGC. This check isolates a completely malfunctioning Mast Group (Electronics Group not receiving any signal), without necessitating climbing the mast. The test set attenuation settings are at 0 db and the Modulation Section Switch on INT MSG. Using the normal AN/SRN-9 equipment setup with the Mast Group connected to the Electronics Group, if signal lock-on or proper message printout is not obtained, proceed to Check No. 3. If all indications are normal, proceed to Check No. 4.

Mast Check

If there is an improper message printout, or no test signal lock-on is obtained, climbing the mast is unavoidable. It is assumed that all connections of the Electronics Group, including a-c power, have been verified to be correct. For this test the RF section must be removed from the Test Set, and arrangements made for communications between the man on the mast and the operator monitoring the received signal at the Electronics Group of the Navigation Set. This test gives a sensitivity check of the Mast Group. A low sensitivity in the Mast Group would indicate the need for replacement.

Doppler Purity Test

The Doppler Purity Test provides a test message and a fixed doppler, although the message is of no consequence in determining the accuracy of results of a navigational fix. This test will reveal any power line noise which can cause doppler errors. If the Doppler Purity Check is not passed, the 115-VAC power must be checked for transients.

APPENDIX I
TRAINING FILM LIST

Certain training films that are directly related to the information presented in this training course are listed below under appropriate chapter numbers and titles. Unless otherwise specified, all films listed are black and white with sound, and are unclassified. For a description of these and other training films that may be of interest, see the United States Navy Film Catalog, NavWeps 10-1-777.

Chapter 1

ADVANCEMENT

MN—10402   The Electronics Technician (13 min.—1967).

Chapter 2

MAINTENANCE

MN—10043A   Planned Maintenance System—Introduction (20 min.—Color—1964).

MN—10043B   Planned Maintenance System—Maintenance Data Collection (15 min.—Color—1965).

MN—8639   Safety on the Job at Sea (16 min.—1957).

MC—4597   For Safety's Sake (13 min—1945).

MN—8990   115 Volts—Deadly Shipmate (19 min.—Color—1960).

MN—6754   Safety Precautions for Electronics Personnel (15 min.—1951).

Chapter 7

MULTICOUPLERS AND DISTRIBUTION SYSTEM

MN—6836   Shipboard Radio Communication—Remote Control Transfer Switchboards (11 min.—1951).

Chapter 8

TELETYPE AND FACSIMILE TERMINAL EQUIPMENT

MN—8099A   Radio Teletype Systems Afloat—General Principles of of Operation. (15 min.—1956).

MN—8099B   Radio Teletype Systems Afloat—Tone Modulated System (11 min.—1956).

Chapter 10

ELECTRONIC AIDS TO NAVIGATION, Part I

MN—5894A   Radio Direction Finders—Theory and Fundamentals
         (15 min.—1948).

MN—2731A   Loran—Basic Principles of Loran Operation
         (12 min.—1944).

Chapter 11

ELECTRONIC AIDS TO NAVIGATION, Part II

MN—6925C   Ground Aids to Air Navigation (Ship to Shore)
         (19 min.—1955).

Chapter 12

SEARCH RADAR

MA—6000A   Technical Principles of Radar—Part I—Introduction
         (22 min.—1945).

MA—9524   Theory of Moving Target Indicator Operation
         (20 min.—1960).

Chapter 14

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MN—6987A   Electronic Countermeasures—Introduction
         (13 min.—1953—Classified).

MN—9455B   The FBM Weapon System—The Navigation Subsystem
         (20 min.—Color—Classified—1963).

Chapter 15

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         (13 min.—1967).

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