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# DIRECTED STUDY

# SINGLE-SIDEBAND COMMUNICATIONS

FEDERAL AVIATION AGENCY ★ UNITED STATES OF AMERICA ★

## FEDERAL AVIATION AGENCY

### OFFICE OF PERSONNEL AND TRAINING

TRAINING DIVISION

# SINGLE-SIDEBAND COMMUNICATIONS

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Approved for use in training FAA electronic engineers and technicians in single-sideband communications.

This publication is not intended to replace, substitute for or supersede official regulations or directives which should be consulted for final authority.

O C Lott

O. C. Lott
Chief, Training Division
Office of Personnel and Training
Federal Aviation Agency

Date: March 26, 1963

# PREFACE

The increased use of Single-Sideband requires training in the basic theory and techniques of SSB communications. This course is designed for technicians working with SSB equipment. To fully understand the operation of the equipment, the technician must acquire certain basic SSB concepts. It is assumed that the student will be generally familiar with the concepts of conventional AM communications. However, a brief review of AM concepts is included for completeness.

In order to make the course as practicable as possible, a transmitter and a receiver used in FAA installations are considered from a functional point of view. A discussion of terminal equipment common to SSB systems is included in the last chapter.

This second edition of the text incorporates changes which we feel will improve the course.

This course was prepared by Lawrence (Larry) M. Wood. Valuable assistance was given by Bill Holland.

W. R. Hudgins
Chief, Communications Section

REFERENCE LIST

SINGLE-SIDEBAND COMMUNICATIONS

DFC-25

The following references will be helpful to students who wish to pursue the subject of single sideband further. If you know of additional references, please list them below and return to us. This will enable us to be of additional services to other students.

Proceedings of the IRE, Single Sideband Issue,
December 1956

Single Sideband Communications Handbook,
by Harry D. Hooten, Howard Sams Publishing Co.

Single Sideband Communications,
by Philco Corp., Philadelphia, Pennsylvania

Fundamentals of Single Sideband,
Collins Radio Co., Cedar Rapids, Iowa

Report on Single Sideband Techniques & Design Requirements,
W.B. Bruene, Engineering & Research Div.,
Collins Radio Co., Cedar Rapids, Iowa

Single Sideband for the Radio Amateur,
Published by the American Radio Relay League

Sideband Handbook, by Don Stoner,
CQ Technical Series, Cowan Publishing Corp.

Please list additional references below and mail to:

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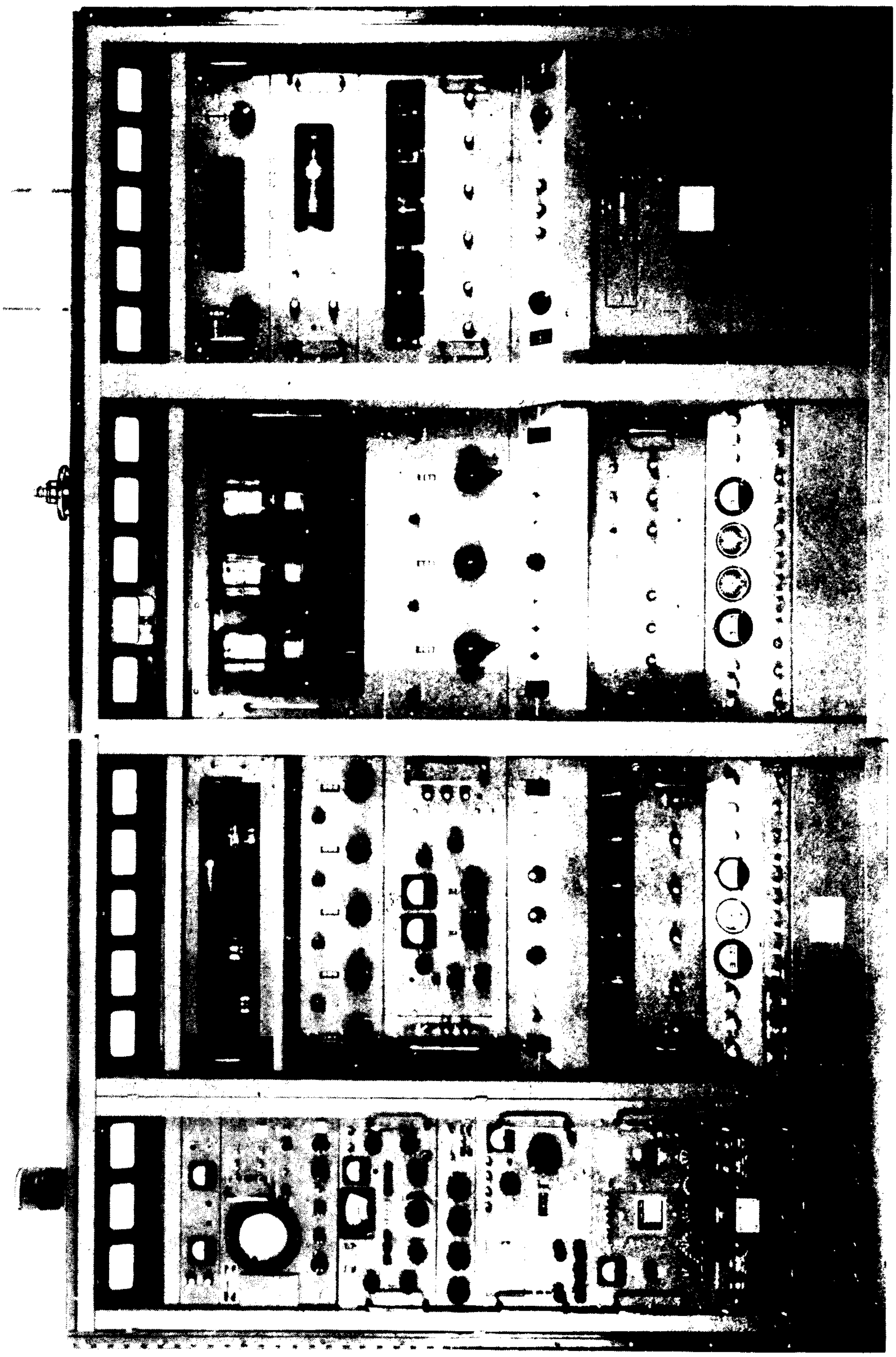
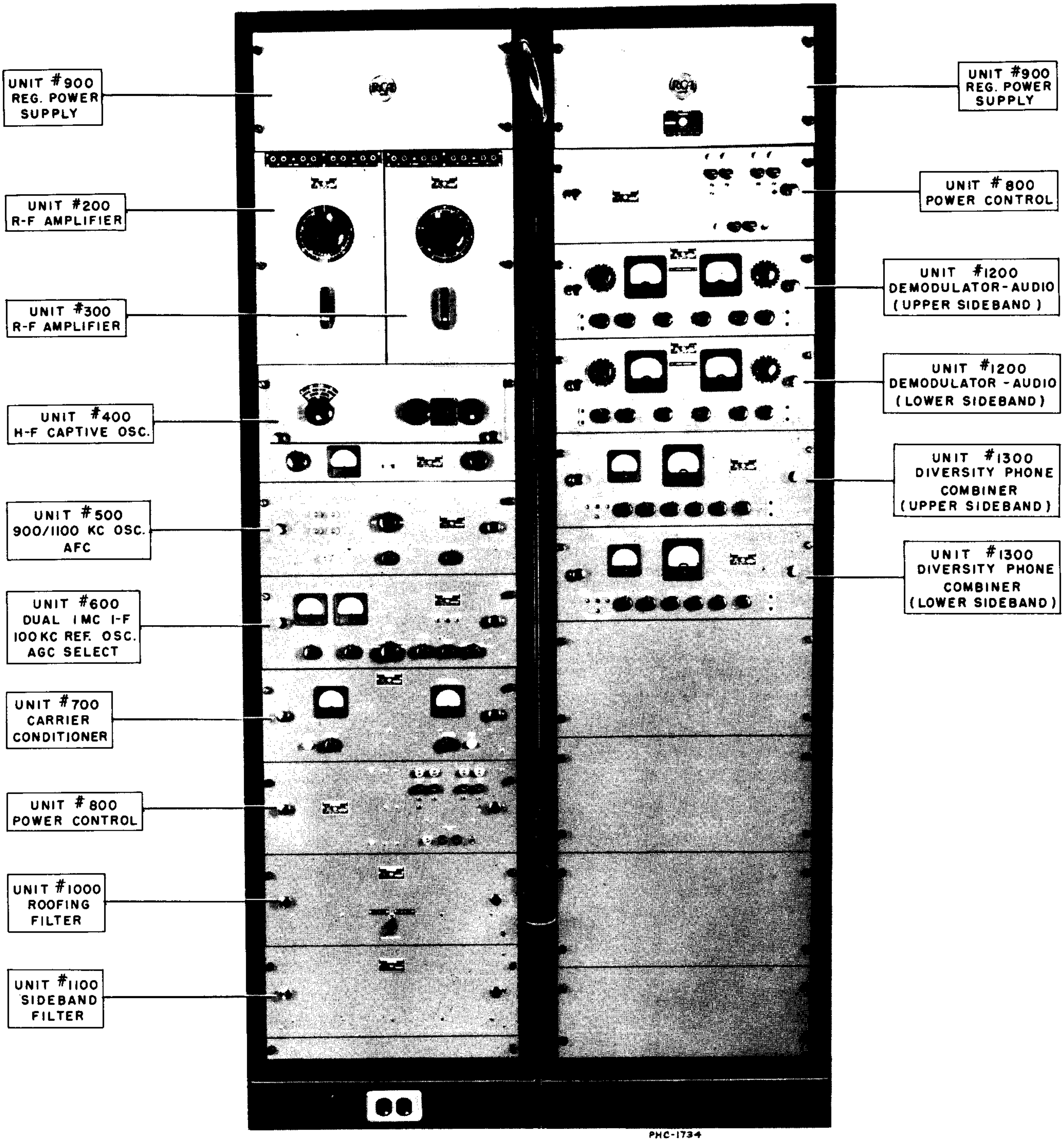


(Courtesy of TMC)

AN/FRT-40 sideband transmitter manufactured by TMC.

(Courtesty of Radio Corporation of America, Radiomarine Products)




SSB-R3 sideband receiver manufactured by RCA.

# CHAPTER 1

# INTRODUCTION

## 1.0 DEFINITION

The term "single sideband" applies to the type of modulation utilized in radio communications in which one sideband or one sideband and the carrier are suppressed from the composite AM wave before transmission. The presence or absence of the carrier is not a factor in the definition of single-sideband transmissions. Single sideband applies equally well when two sidebands are transmitted and the lower sideband contains different modulation than the upper sideband (independent sideband). To clarify, transmissions that contain the following components of the AM wave are single sideband: (1) one sideband and a full-level carrier, (2) one sideband and a low-level carrier, (3) one sideband and no carrier, (4) two sidebands and a full-level carrier with the two sidebands containing different modulation (independent sideband), (5) two sidebands and a low-level carrier with the two sidebands containing different modulation (independent sideband), (6) two sidebands and no carrier with the two sidebands containing different modulation (independent sideband).

There are several types of single-sideband transmission, and in the process of producing each type the composite AM wave is altered by suppression of some of its frequency components. This concept is discussed in detail in Chap. 2. First, however, a brief historical account of SSB is given to break the ground for a more analytical treatment of SSB theory.

## 2.0 HISTORY OF SINGLE SIDEBAND

In 1914, it was established mathematically that an amplitude-modulated wave consists of a carrier and two sidebands. For more than a decade thereafter, the physical reality of sidebands continued to be argued. It was alleged that sidebands were only a mathematical fiction.

As a result of experiments conducted at the U.S. Navy Radio Station at Arlington, Virginia, in 1915, the physical reality of sidebands was established. At the suggestion of H.D. Arnold, the antenna of a radio telephone transmitter was tuned to one side of the carrier frequency so that one sideband was passed and the other was attenuated. The original speech was reproduced from the radiated single-sideband signal. Here was recognition that the combining of the audio and carrier in a non-linear element to produce sum and difference frequencies was more than a mathematical fiction. Because satisfactory communication was possible when only one sideband was transmitted, it was also clear that one sideband contained all of the intelligence of the original speech.

Mr. John R. Carson of the American Telephone and Telegraph Company did a considerable amount of research on SSB and concluded that in the process of transmitting conventional AM the greatest amount of energy radiated from an antenna was contained in the carrier. Carson's experiments further proved that the carrier contained no useful information but served only as a reference for the sidebands; therefore, it could be eliminated without sacrificing any of the intelligibility of the transmitted signal. Because the greatest amount of energy is contained in the carrier, the elimination of the carrier permits a considerable saving of input power for the same effective power output, thus providing a more efficient method of communication than conventional AM.

An early single-sideband application was the first commercial wire carrier telephone system. Single-sideband permitted twice as many channels, thus increasing the limited frequency spectrum available for wire transmission. SSB soon became standard in most systems of wire transmission because of its early success in this application.

In 1922 a reliable radio speech transmission employing SSB was made from Long Island to New South Gate, England. This led to the development of a New York-London circuit in 1927. By 1936 Bell Telephone Laboratories had completed extensive tests on SSB equipment, and the design for production models was initiated. By 1946 more than 50 commercial SSB radio circuits were in operation throughout the world.

Shortly after amateur radio communication was restored, following World War II, the increase in the number of amateurs created a very congested condition on the amateur bands. Experimentally-minded amateurs deplored the difficult communications that resulted from this congestion and began to explore the narrow-band, high-reliability capabilities of single-sideband transmission. The merits of SSB were discussed quite thoroughly by amateurs during the late forties and the inevitable path of progress led to adaptation of many existing AM transmitters and receivers to the specific requirements of SSB. The success of these amateurs caused nearly every manufacturer of amateur equipment to produce SSB transmitters and receivers. As a result of this interest in SSB, literally hundreds of articles covering various phases of SSB theory and practice have been published.

Thus far, commerical and amateur experience with SSB has brought about only two basic methods of generating single-sideband signals. In one method (filter method) a carrier is modulated with the intelligence signal in a balanced modulator which suppresses the carrier; then one sideband is eliminated by means of a sideband filter. In the second method (phasing method) the carrier and one sideband are eliminated by a system of balanced modulators in which the carrier and modulating signal applied to one modulator are shifted 90° with respect to those signals applied to another modulator. The carrier is canceled in the balanced modulators, and the undesired sideband is balanced out in the common load of the system.

## 3.0 TYPES OF SINGLE- AND DOUBLE-SIDEBAND COMMUNICATIONS

During the thirty years that single sideband has been used commercially, several types of sideband communications have been developed. Certain types of communications have advantages over others; the type used depends on the circumstances. A brief description of a few of the common types of single- and double-sideband systems is given in the following paragraphs.

(a) In what is perhaps the most common type of SSB communication, one sideband or two independent sidebands are transmitted and the carrier is suppressed as much as is practicable. In the reception of such a transmission, it is necessary to insert a locally generated signal close to the proper frequency. The frequency of the locally generated signal must be maintained within close tolerances.

(b) In another type of single sideband, a pilot low-level carrier is transmitted along with one sideband or with two independent sidebands. This low-level carrier is amplified separately in the receiver. The carrier can then be injected into the demodulator as a reference for detection or it can be used to synchronize a local oscillator in the receiver to give automatic frequency control.

(c) In still another type, one sideband or two independent sidebands and a carrier which varies inversely with the signal level is transmitted. This allows a sufficient average carrier level for automatic frequency control without reducing the sideband power below the full transmitter rating.

(d) In the double-sideband system, two sidebands containing the same modulation are transmitted and the carrier is suppressed. For reception, the double-sideband signal requires a locally generated carrier of proper frequency and <u>phase</u> for demodulation. Although the double-sideband system requires as much spectrum as conventional amplitude modulation, all of the energy transmitted is useful sideband energy. This allows a great saving in power requirements. Also, the problem of two strong carriers beating together and covering the RF spectrum with undesired signals is partially eliminated.

In the past, many systems have used automatic frequency control of the reinsertion frequency at the receiver because the frequency accuracy requirements for single-sideband communications are very precise.

Automatic frequency control is not always necessary in present equipment because the frequency stability requirements of the equipment can be met with present oscillator design techniques. However, automatic frequency control will be used in FAA equipment because it is necessary to communicate with foreign governments whose equipment cannot always meet frequency stability requirements necessary for reliable communications.

Single sideband is used extensively in long-range point-to-point communications. Several long-range single-sideband circuits are now used by the FAA; additional circuits are contemplated for the future.

The circuits in use now and those proposed for the future are as follows:

(a) SSB circuit 365 operating between Honolulu and Wake Island uses Collins Type 310-F6A SSB exciters and 205J-1 linear amplifiers with RCA Type SSB-R3 receivers.

(b) Circuit 350 operating between Honolulu and San Francisco uses Western Electric Type LD-T2 transmitters with Western Electric Type LD-R1 and RCA SSB-R3 receivers.

(c) New York/San Juan, Circuit 210, uses AN/FRT-39 transmitters and SSB-R3 receivers.

(d) New York/Santa Maria, Circuit 205, uses AN/FRT-40 transmitters and SSB-R3 receivers.

(e) All of these stations utilize AN/FRT-39 transmitters and SSB-R3 receivers: Miami/San Juan, Balboa/Bogota, Balboa/Guayaquil, Guam/Manila, Guam/Wake Island, and Honolulu/Midway.

It must be remembered, of course, that sideband is widely used in FAA systems other than those mentioned above.

## 4.0 ADVANTAGES OF SSB

Single-sideband systems have many advantages over other methods of communications. The advantages of SSB over AM will be stated briefly in the following paragraphs. A more detailed analysis will be included in the theory portions of the course.

(a) SSB has a reduced frequency spectrum, permitting nearly twice as many channels for a given band of frequencies. Doubling the number of channels in the 2-megacycle to 30-megacycle range is especially important.

(b) There is, as has been stated, an increase in effective power because all of the power goes into the single sideband, which carries the useful voice intelligence. The power gain in the SSB system is from 3 to 12 decibels over the equivalent AM system.

(c) SSB is less subjected to certain interference and fading conditions than is AM. This interference in AM occurs when the phase relationship of the wave components are not maintained during transmission.

(d) In AM communications systems the carrier remains on the air during periods when modulation is absent. If one station transmits while another (having nearly the same carrier frequency) is on, there is a heterodyning of the two carriers, resulting in interference. In SSB, with voice break-in, this interference is reduced because signal radiation ceases when the operator stops speaking into the microphone. A station may enter the network as soon as the "talk power" leaves the air. Even when two stations transmit at the same time, a receiving station can read through the interfering station the same way you are able to choose one conversation from several going on around a conference table.

## 5.0 DISADVANTAGES OF SSB

Very briefly, some of the disadvantages of SSB are as follows:

(a) SSB has extreme stability requirements which increase the complexity and cost of the equipment. Because of these stability requirements, proper maintenance is more difficult and requires more expensive test equipment. The tuning of SSB transmitters and receivers is more complicated than the tuning of AM equipment.

(b) It is often difficult or impossible to convert existing AM systems to SSB systems. This would not be a disadvantage, however, if SSB had been used before AM systems were developed.

The reasons for some of the advantages and disadvantages may not be apparent now, but they will be more apparent after the appropriate theory has been covered.

ATTENTION

Examination 1 is to be worked at this time.
This exam covers the material in Chap. 1.
It is included with the material.

CHAPTER 2

# MODULATION AND SIDEBAND THEORY

## 1.0 INTRODUCTION

In discussing single-sideband transmission theory we shall be concerned first with the development of the AM wave and then with the alterations necessary to obtain a single-sideband signal. The initial discussion will recall to mind the various components of the AM wave, and will thus aid in a better understanding of the techniques used in the transmission and reception of SSB. The brief discussion of the AM wave includes a qualitative discussion of the components of the AM wave followed by the corresponding mathematical analysis.

In the process of amplitude modulation, the radio-frequency carrier and the low-frequency audio signal are impressed simultaneously upon a final amplifier stage to produce an amplitude modulated wave. The entire AM wave is referred to as the composite wave. The composite wave and the separate frequency components contained in the composite wave are the subjects of this chapter.

It is sometimes convenient to represent the components of the AM wave or the entire wave by vectors. The vector discussion included in this chapter will strengthen your understanding of amplitude modulation.

## 2.0 BASIC CONCEPTS

An amplitude-versus-time graph is often used in discussing amplitude modulation. The graph of the amplitude-modulated wave appears as a wave of single frequency but with varying amplitude. For simplicity, a carrier modulated by only one frequency of sinusoidal form is often used to illustrate concepts of the AM wave, as indicated in Fig. 2-1(a). The area encompassed by smooth curves joining successive positive and successive negative peaks is called the envelope. The envelope is not a waveform itself but merely a sketch of the waveform, and is generally used to represent the entire AM wave. See Fig. 2-1(b).

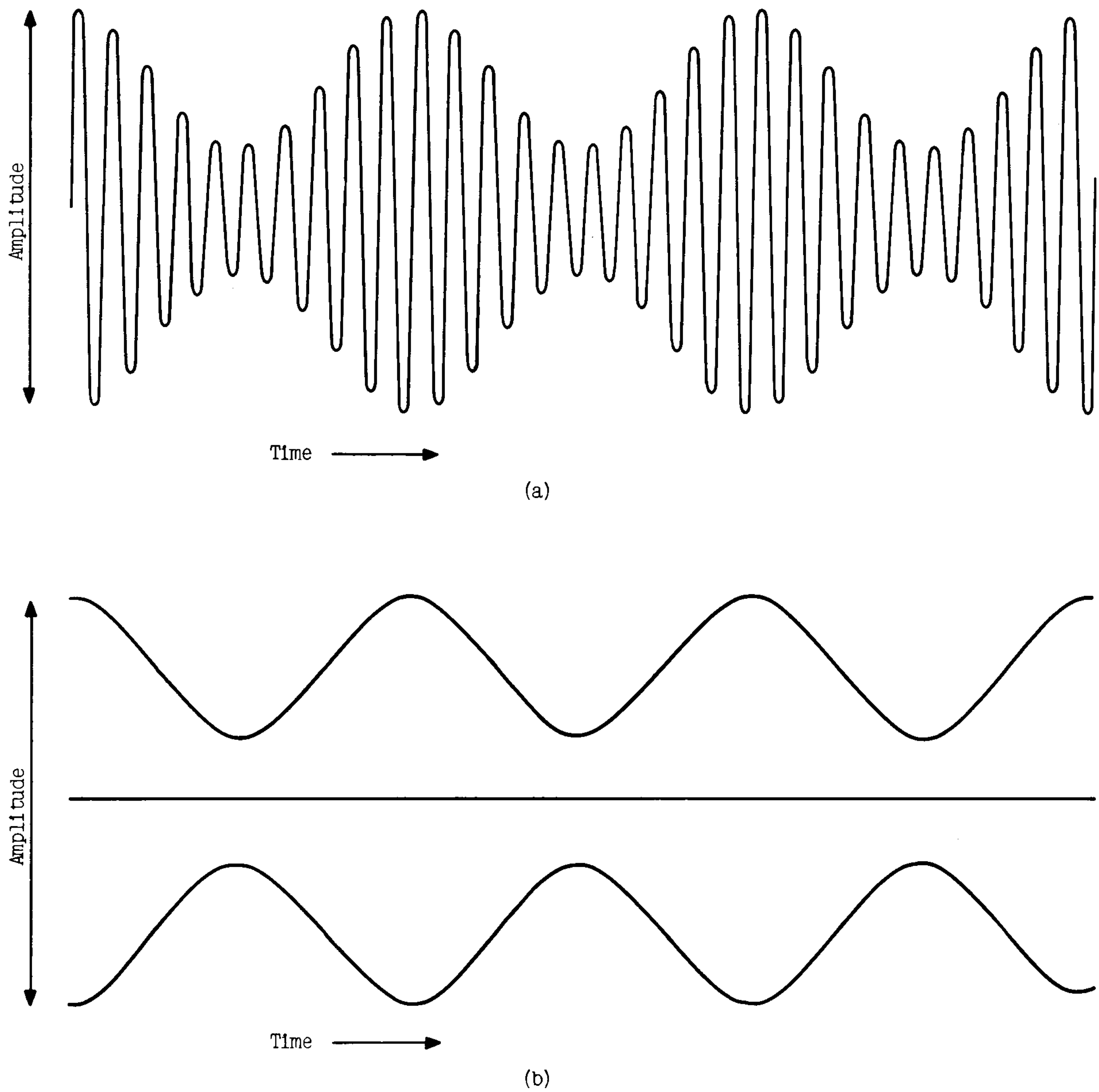



Fig. 2-1. Waveform of an amplitude-modulated wave.

The waveform shown in Fig. 2-1(a) is produced by modulating the carrier of an AM transmitter with a single tone of sinusoidal form. Basically the modulation takes place as follows.

An oscillator generates a carrier frequency and the carrier is subsequently amplified to a level suitable for transmission. The modulation is usually applied to the Class C final stage and in series with the dc supply to that stage, as indicated in Fig. 2-2.

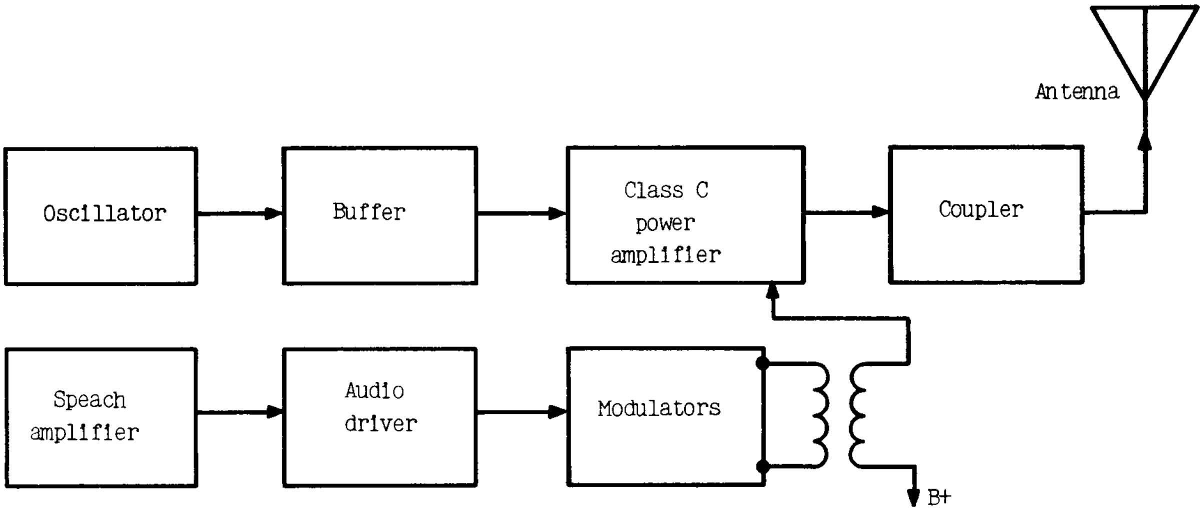



Fig. 2-2. Basic AM transmitter.

The audio signal adds to and subtracts from the plate voltage of the final stage; thus the amplitude of rf carrier oscillations in that stage varies at the audio rate as shown by Fig. 2-1(a). The wave appears as a single rf frequency varying in amplitude; however, the sum and difference frequencies and original carrier frequency can be shown to exist in the composite wave by the following method.

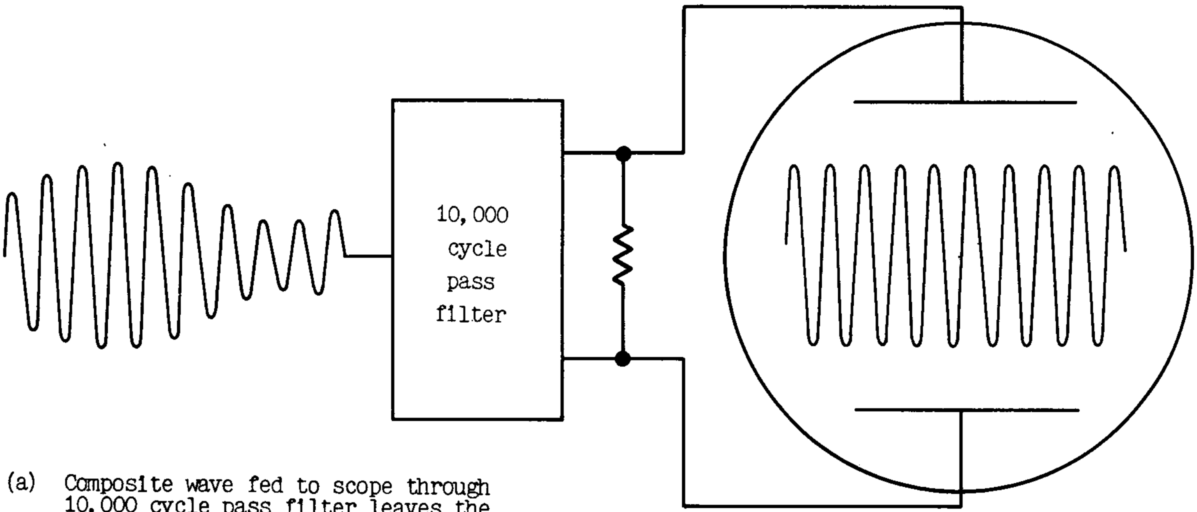
An AM wave developed from a 10,000-cycle carrier frequency and a 1,000-cycle modulating frequency (Fig. 2-3(a)) is fed to the vertical deflection plates of an oscilloscope through a filter passing 10,000 cycles and rejecting 9,000 and 11,000 cycles. The carrier of constant amplitude is presented on the scope. Likewise, when the AM wave is fed to the scope through a filter rejecting 10,000 cycles, but passing

9,000 and 11,000 cycles, a distinct waveform is presented on the scope (Fig. 2-3(b)). This waveform will be recognized later as being sideband energy only (both upper and lower). If the carrier and upper sideband, 10,000 and 11,000 cycles, are filtered out then the scope presentation is that of a waveform having a longer time base per cycle than the carrier shown on the scope in Fig. 2-3(a). That is, we now see the lower sideband (Fig. 2-3(c)).

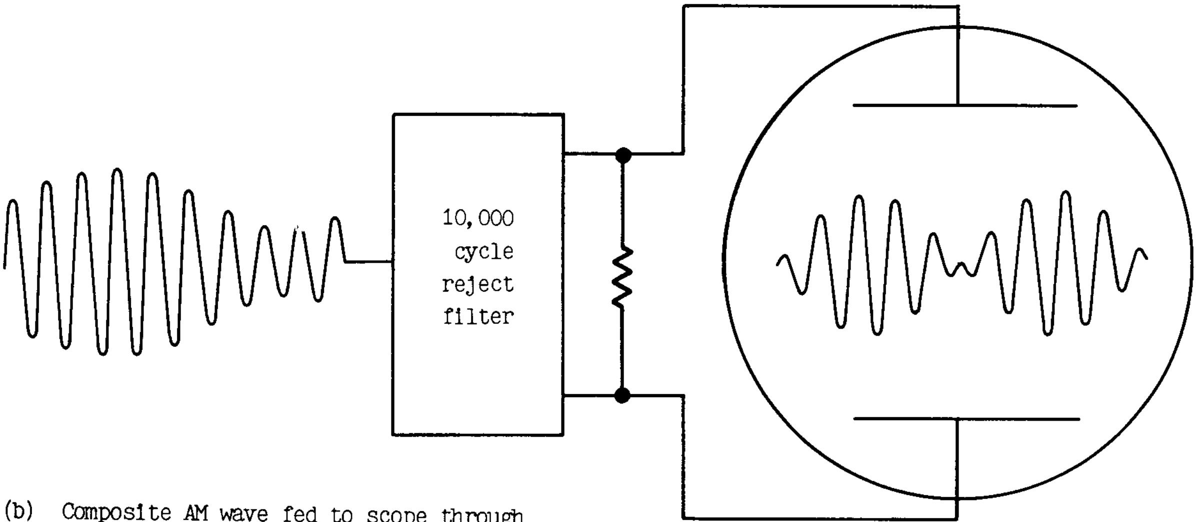
If the 10,000- and 9,000-cycle signals were filtered out, a wave of shorter time base per cycle than the carrier would appear on the scope. This would represent the upper sideband; however it is not included in Fig. 2-3.

Thus, a carrier modulated by a single sine wave tone (such as 1,000 cycles) forms a composite AM wave that is the resultant of three frequency components. The amplitudes of these frequency components are constant, and it is only the resultant of these components that make the wave appear to vary in amplitude. This concept should be kept clearly in mind because the waveform can actually be obtained by varying the amplitude of a single frequency.

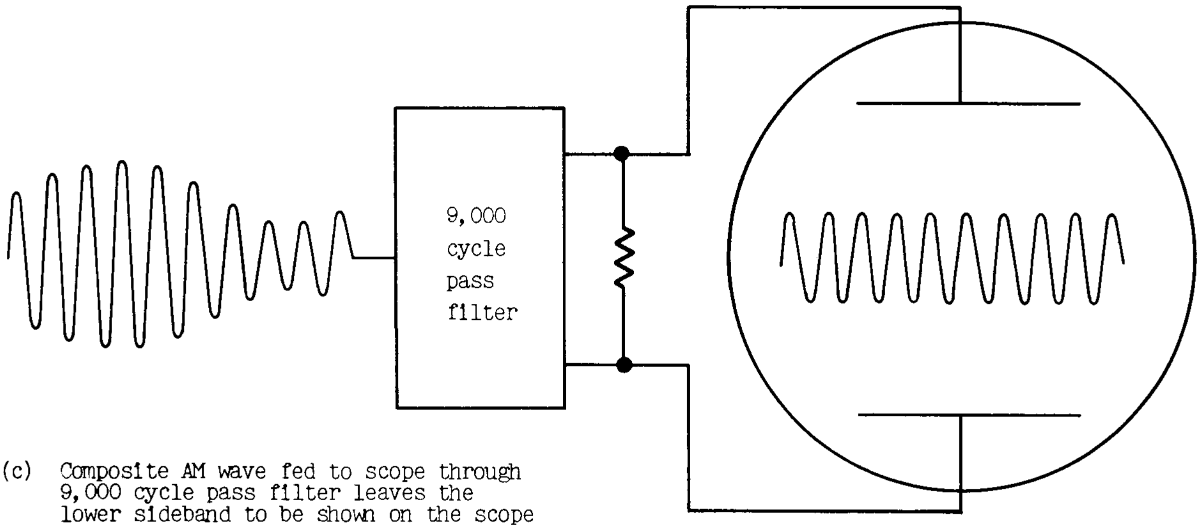
During the process of amplitude modulation the audio signal is moved up in frequency to form the sidebands. The frequency relationship of the sidebands and the carrier in the AM wave can be shown graphically by means of an amplitude-versus-frequency graph. The frequency relationship shown in Fig. 2-4(a) represents modulation by a single frequency.




(a) Composite wave fed to scope through 10,000 cycle pass filter leaves the carrier to be presented on the scope




(b) Composite AM wave fed to scope through 10,000 cycle reject filter leaves sidebands only




(c) Composite AM wave fed to scope through 9,000 cycle pass filter leaves the lower sideband to be shown on the scope

Fig. 2-3.

Voice communication requires modulation by a band of voice frequencies lying between approximately 300 and 3,000 cycles. A carrier modulated by these frequencies is represented in Fig. 2-4(b). Note that the sidebands occupy a frequency spectrum from 3,000 cycles below the carrier frequency to 3,000 cycles above for a total bandwidth of 6,000 cycles.

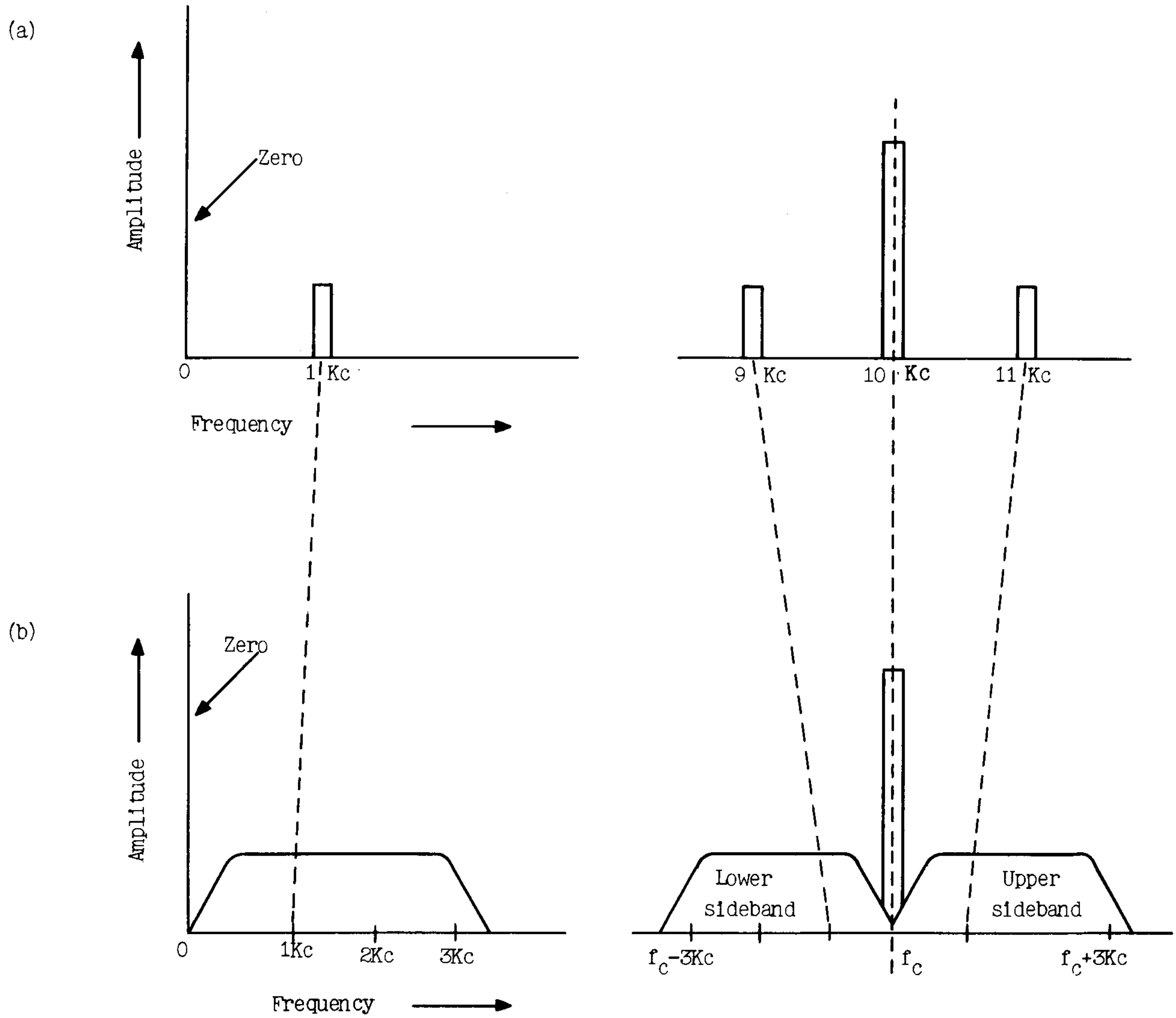



Fig. 2-4. Amplitude versus frequency graph showing frequency relationships.

So far in this chapter, we have reviewed the familiar method of obtaining an AM wave in a transmitter and the manner in which such waves are represented graphically. It has been brought out that although an AM wave appears to be a carrier of a single frequency, the wave is actually a combination of the sidebands and the carrier forming a composite AM envelope. The quantitative considerations of the composite wave, such as the power and voltage relationships of the sidebands and carrier, will be included in the following mathematical analysis.

## 3.0 ANALYSIS OF THE AMPLITUDE-MODULATED WAVE

### 3.1 Definition of Symbols

It will be necessary to define various symbols used in amplitude modulation. The more important terms are

$m$ = modulation factor

$E_{cm}$ = maximum carrier voltage

$E_{sm}$ = maximum sideband voltage

$E_c$ = rms value of carrier voltage

$e$ = instantaneous value of voltage

$P_c$ = power in the carrier component

$P_s$ = power in both sidebands

$P_u$ or $P_L$ = power in upper or lower sideband

$\omega_a$ = angular velocity of modulating frequency*

$\omega_c$ = angular velocity of rf carrier

$\omega_c + \omega_a$ = angular velocity of upper sideband component

$\omega_c - \omega_a$ = angular velocity of lower sideband component

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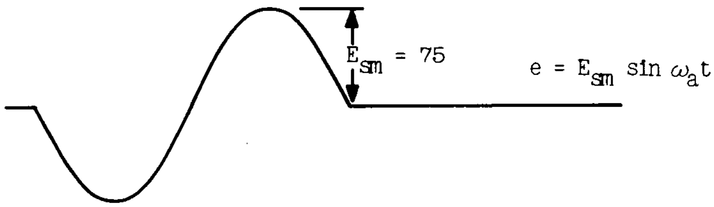
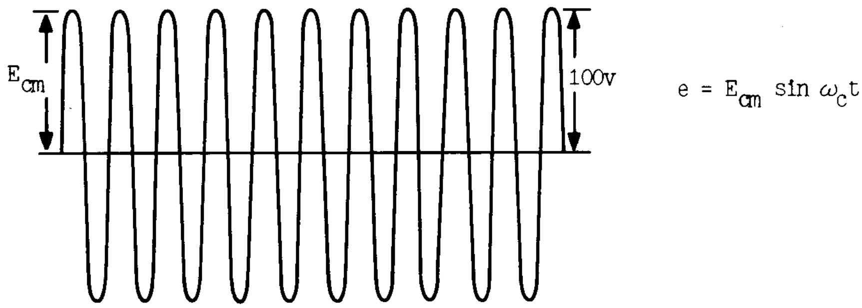
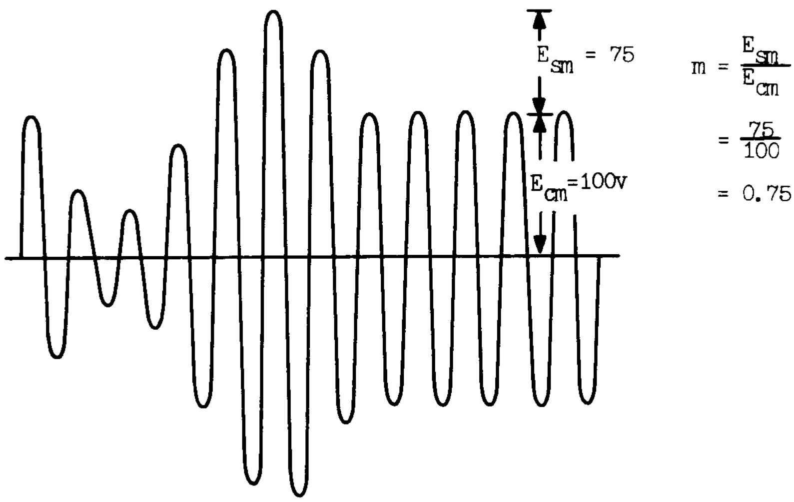
* $\omega$ being equal to $2\pi f$.




(b) Signal to be impressed on carrier




(a) Carrier of maximum amplitude $E_{cm}$




(c) Modulation envelope

**Fig. 2-5.**

Using these symbols, we shall develop those equations describing the AM wave that will contribute most to an understanding of both AM and single-sideband theory.

### 3.2 Modulation Factor

In the process of amplitude modulation, a carrier of a given amplitude is added to and subtracted from to a certain degree, depending on the amplitude of the modulating signal. The strength of the modulating signal determines the amplitude of the sidebands.

A term that indicates the relative amplitudes of the carrier and sidebands in the AM wave is the modulation factor. By definition, the modulation factor, m, is simply the ratio of the sideband voltage to the carrier voltage. Symbolically, $m = E_{sm}/E_{cm}$, where $E_{cm}$ is the maximum amplitude of the carrier wave before modulation and $E_{sm}$ is the amount that the sideband voltage adds to, or subtracts from, the carrier. As an example, the carrier shown in Fig. 2-5 is 100 volts in amplitude. When this carrier is modulated with the signal voltage shown in part (b), having a maximum voltage of 75 volts, the AM wave of part (c) results. The modulation factor of this wave is equal to $E_{sm}/E_{cm} = 75/100 = 0.75$, or the wave is said to be modulated 75 percent.

## 4.0 EQUATIONS OF THE AM WAVE

From the relationship already established it takes only a few steps to write an equation that will express the form and content of an amplitude-modulated wave more exactly.

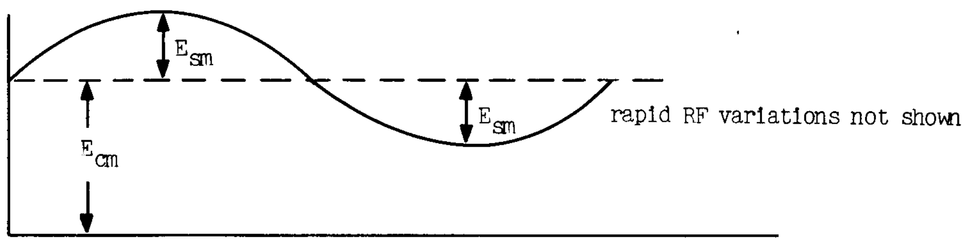



Fig. 2-6.

Assume a constant-amplitude carrier of amplitude $E_{cm}$ (Fig. 2-6). Now if a sinusoidal modulating audio voltage of maximum value $E_{sm}$, expressed by $E_{sm}\sin\omega_a t$, is applied to it, the modulating component will add to and subtract from the constant-amplitude component. The envelope is then expressed as:

$$\text{Envelope} = E_{cm} + E_{sm}\sin\omega_a t.$$

However, the envelope shows only the effect of the modulating component and does not show the rapid rf variation of the total wave. The influence of the carrier component must be introduced to the expression because the slowly varying audio wave is modulating a rapidly varying carrier wave. Therefore, applying the rf sine wave function to the expression gives the instantaneous voltage (e) of the amplitude-modulated wave,

$$e = (E_{cm} + E_{sm}\sin\omega_a t)\sin\omega_c t \tag{2-1}$$

Expand Eq. (2-1) and two terms result:

$$e = E_{cm}\sin\omega_c t + E_{sm}\sin\omega_a t\sin\omega_c t \tag{2-2}$$

Remembering that $m = E_{sm}/E_{cm}$; then $E_{sm} = mE_{cm}$, and Eq. (2-2) in general form becomes

$$e = E_{cm}\sin\omega_c t + mE_{cm}\sin\omega_a t\sin\omega_c t \tag{2-3}$$

The first term of Eq. (2-3) describes the carrier component and the second the total sideband component. The two terms plotted separately, become as shown in Fig. 2-7 and 2-8. Note in Fig. 2-8 that two lobes are shown and each lobe represents both upper and lower sidebands.

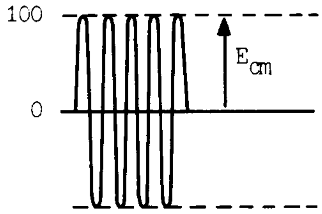



Fig. 2-7. Carrier component $E_{cm}\sin\omega_c t$.

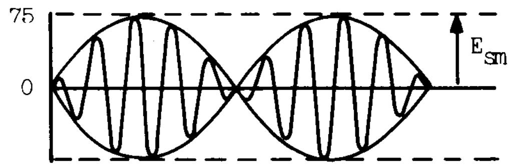



Fig. 2-8. Total sideband component $m\,E_{cm}\sin\omega_a t\sin\omega_c t$.

It should be kept in mind that the two waves in Figs. 2-7 and 2-8 represent two distinct parts of the complete wave. They may be radiated in combined form from a single antenna (plate modulation) or they may be radiated as discrete waves from separate antennas and then combined upon reception (space modulation) to make up a complete modulated wave.

If the total sideband component of Fig. 2-8 were radiated by itself it would be a double-sideband transmission. The wave (and each lobe of the wave) represents both the upper and the lower sidebands. However, the general equation for an amplitude-modulated wave may be put in a different form to express separately the upper and lower sideband terms as well as the carrier.

The sideband term (the second term) of the general Eq. (2-3) contains the product of the sines of two angles. That is

$$e = E_{cm} \sin \omega_c t + mE_{cm} (\sin \omega_c t)(\sin \omega_a t) \qquad (2\text{-}3)$$

The trigonometric identity for the product of the sines of two angles is $\sin\phi \sin\theta = \frac{-1}{2} \cos(\phi + \theta) + \frac{1}{2} \cos(\phi - \theta)$.

Substituting the identity yields

$$\underbrace{e = E_{cm} \sin \omega_c t}_{\text{Carrier Term}} \underbrace{- \frac{mE_{cm}}{2} \cos(\omega_c + \omega_a)t}_{\text{Upper Sideband Term}} \underbrace{+ \frac{mE_{cm}}{2} \cos(\omega_c - \omega_a)t}_{\text{Lower Sideband Term}} \qquad (2\text{-}4)$$

The plus and minus signs within the parentheses indicate the respective sidebands. That is, $\omega_c + \omega_a$ indicates that one sideband (upper) is higher in frequency than the carrier, and $\omega_c - \omega_a$ indicates that the other sideband (lower) is lower in frequency than the carrier.

If $\omega_a$ (the angular velocity of the audio) is due to a single audio frequency then the two sideband terms can be plotted separately as single-frequency sinusoidal waves. See Fig. 2-9(a) and 2-9(b). The graphical addition of the upper and lower sidebands results in the total sideband component, which was shown in Fig. 2-8 and is shown again in Fig. 2-9(c) to show the graphical addition.

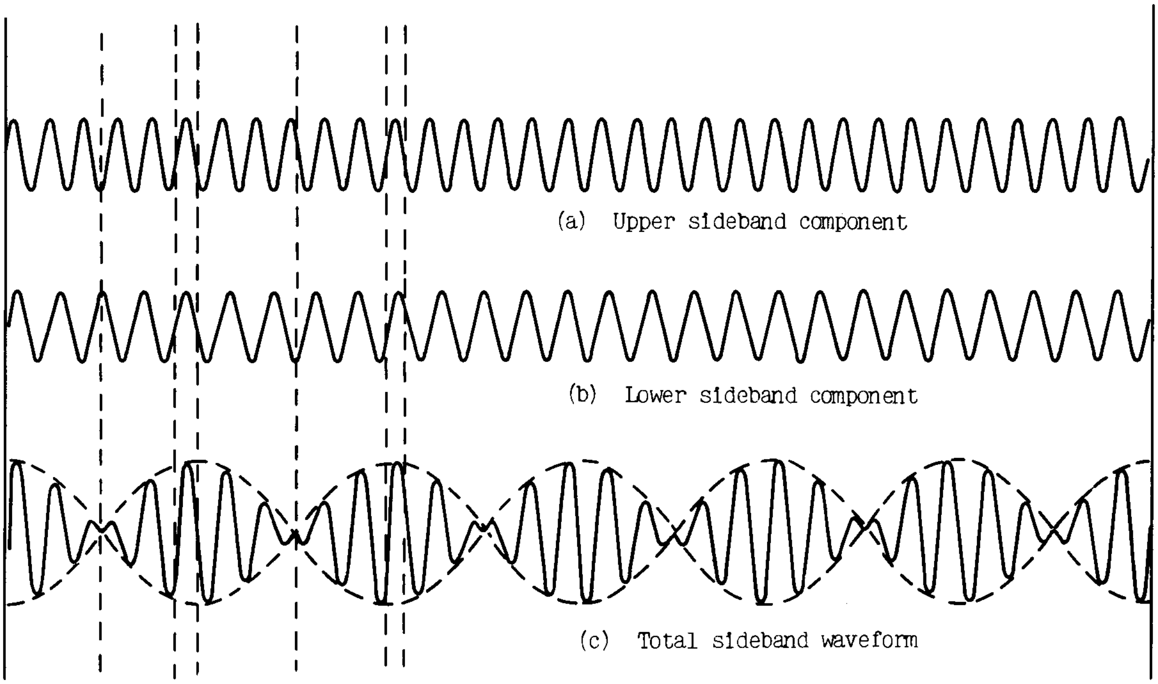



Fig. 2-9. Adding of upper and lower sidebands resulting in total sideband component.

Figure 2-9(a), (b), and (c) is shown again as Fig. 2-10(a), (b), and (c). By placing all the separate components and their combined waveforms on the same page one may more readily see how the separate sideband terms combine to form the total sideband term; likewise, one can more readily see how the total sideband term combines with the carrier term to form the amplitude-modulated wave.

Part (a) of Fig. 2-10 represents the upper sideband term of Eq. (2-4), and Part (b) represents the lower sideband term. If the two terms are added graphically, as shown in Part (c), the total sideband wave is formed. Part (c) represents the total sideband term of Eq. (2-3) which we later divided into two separate parts in obtaining Eq. (2-4).

The carrier term of Eq. (2-4) is represented by Part (d). When the carrier term and the total sideband term are added graphically, the composite amplitude-modulated wave, as represented by Part (e), results.

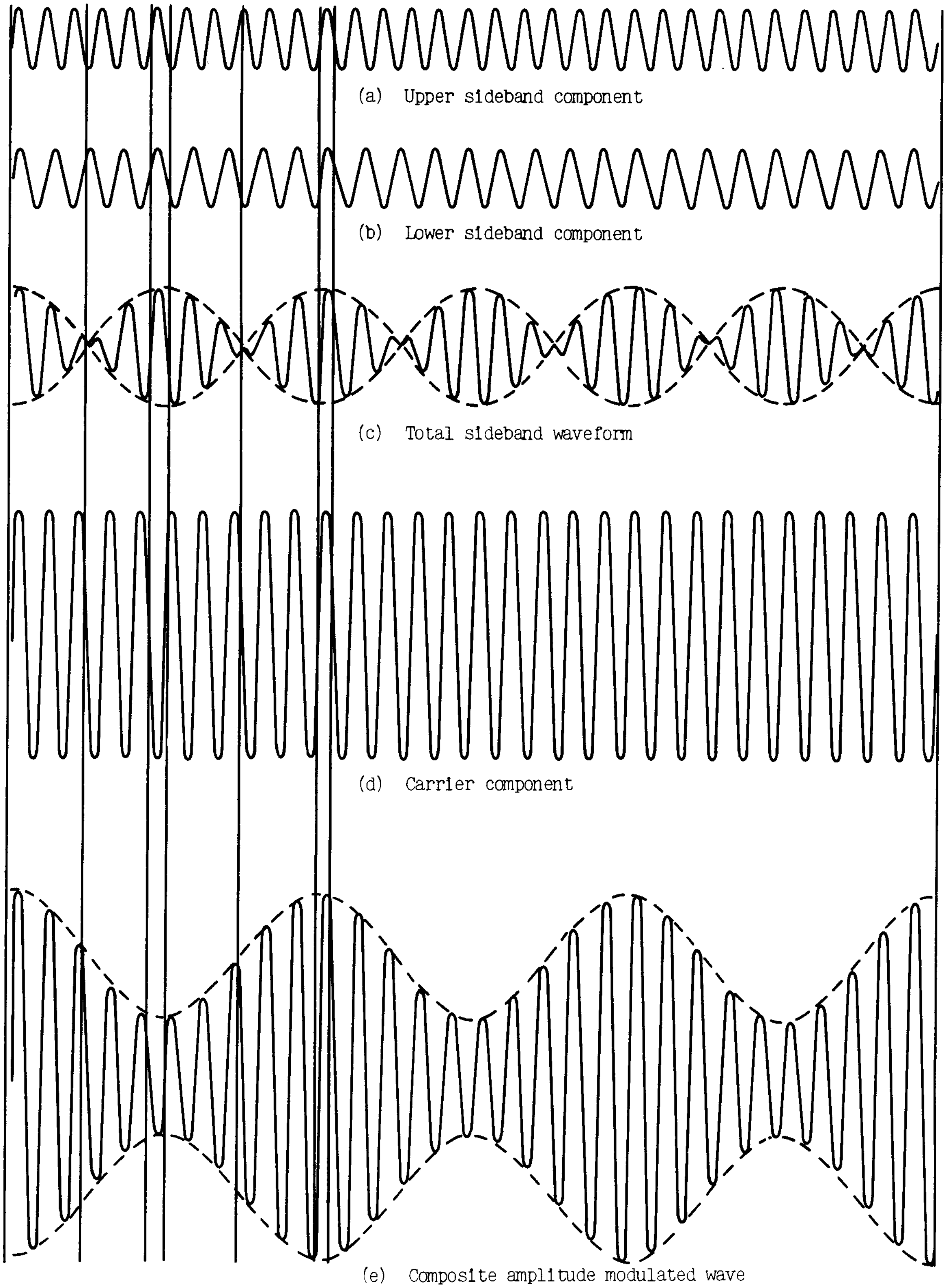



Fig. 2-10. Adding of separate components to form composite AM wave.

In the discussion of Fig. 2-10 we have combined the separate components to show how they will form the composite AM wave. Conversely, we might say that if the separate components will combine to form a composite wave, then the composite wave must contain the separate components. This is actually the case.

## 5.0 POWER IN THE AMPLITUDE-MODULATED WAVE

It is desired to obtain expressions for power when the voltage applied is represented by the modulated wave Eq. (2-4).

$$e = \underbrace{E_{cm}\sin\omega_c t}_{\text{Carrier Component}} - \underbrace{\frac{1}{2} mE_{cm}\cos(\omega_c + \omega_a)t}_{\text{Upper Sideband Component}} + \underbrace{\frac{1}{2} mE_{cm}\cos(\omega_c - \omega_a)t}_{\text{Lower Sideband Component}} \tag{2-4}$$

Voltage e is the instantaneous sum of the right-hand expressions at any time (t). Therefore, each of the expressions on the right must be instantaneous values of the carrier, upper sideband, and lower sideband components, respectively.

The total power in a composite wave must equal the sum of the powers in each individual component. It is then necessary to evaluate the power in the three individual right-hand members of Eq. (2-4).

### 5.1 Carrier Power

The first term ($E_{cm}\sin\omega_c t$) represents the instantaneous value of the carrier voltage. Then $0.707E_{cm}$ is the rms value, $E_c$. As seen in Fig. 2-10, the carrier component is a sine wave of constant maximum amplitude and the power produced by a sine wave of voltage is the rms voltage squared divided by the resistance to which it is applied; that is,

$$P_c = \frac{E_c^{\,2}}{R} \tag{2-4}$$

### 5.2 Sideband Power

The second and third terms of the right-hand member of Eq. (2-4) represent the upper and lower sideband voltage components respectively; and the portion, $mE_{cm}/2$, represents the maximum value of either the upper or the lower sideband voltage.

Figure 2-10 shows that each sideband is a sine wave. Therefore, the power for either sideband dissipated in a load resistance is

$$P_u = P_L = \left(\frac{mE_c}{2}\right)^2 \frac{1}{R} = \frac{m^2 E_c{}^2}{4R} \qquad (2\text{-}6)$$

The upper and lower sidebands contain the same amount of power and can be combined to indicate total sideband power:

$$P_s = P_u + P_L$$

$$P_s = \frac{m^2 E_c{}^2}{4R} + \frac{m^2 E_c{}^2}{4R} = \frac{m^2 E_c{}^2}{2R} \qquad (2\text{-}7)$$

Noting that

$$P_c = \frac{E_c{}^2}{R} \quad \text{then } E_c{}^2 = P_c R$$

and substituting $P_c R$ for $E_c{}^2$ in Eq. (2-7) yields

$$P_s = \frac{1}{2}\, m^2\, P_c \qquad (2\text{-}8)$$

Thus we have the expression relating modulation factor (m) with carrier and total sideband power.

5.3 Power in a 100% Modulated Wave

When the total sideband component voltage, $E_{sm}$ is equal in magnitude to that of the carrier component, the modulation factor is unity (100%) and the maximum instantaneous voltage of the modulated wave is twice that of the unmodulated carrier. Also, the minimum instantaneous voltage equals zero. The maximum instantaneous power in the 100% modulated wave is 4 times the carrier power because the voltage is doubled. The minimum instantaneous power is zero.

As stated previously, the total power is equal to the sum of the powers of the individual components:

From Eq. (2-8), $P_t = P_c + P_s$; (2-8)

$$P_s = \frac{m^2 P_c}{2};$$

then, $$P_t = P_c + \frac{m^2 P_c}{2};$$

after factoring, $$P_t = P_c(1 + \frac{m^2}{2}). \qquad (2\text{-}9)$$

In the special case of 100% modulation, m = 1; and substituting m = 1 in Eq. (2-9) gives

$$P_t = P_c(1 + \frac{1}{2})$$

$$P_t = \frac{3}{2} P_c$$

Therefore, $P_c = \frac{2}{3} P_t$, which indicates that (for 100% modulation) two-thirds of the total power of the complete wave is in the carrier and one-third is in the sidebands. For example, a 100% modulated wave containing 150 watts will have two-thirds of 150 or 100 watts in the carrier and one-third of 150 or 50 watts in the sidebands, 25 watts being contained in each sideband.

One hundred percent modulation is normally assumed when making power comparisons of AM and SSB systems. Such a comparison will be made after a brief review of vector notation.

## 6.0 VECTORIAL ANALYSIS OF AMPLITUDE MODULATION

### 6.1 Plotting of Vectors Representing Trigonometric Functions

The phase difference between two waves can be represented by the angle between their representative vectors. Angles of lag are represented by angles measured clockwise from the reference, and it is customary to draw the vector to be taken as a reference along the positive X-axis. To illustrate, the three voltages

$$e_1 = E_{1m} \sin(\omega t + 30°)$$

$$e_2 = E_{2m} \sin(\omega t + 60°)$$

$$e_3 = E_{3m} \sin(\omega t - 30°),$$

are represented by the vectors $E_1$, $E_2$, $E_3$, of Fig. 2-11(a), drawn with the X-axis as a reference. If $E_1$ is taken as a reference, the three representative vectors may be drawn as in Fig. 2-11(b). In either case $E_2$ is seen to lead $E_1$ by 30 degrees, and $E_3$ lags $E_1$ by 60 degrees.

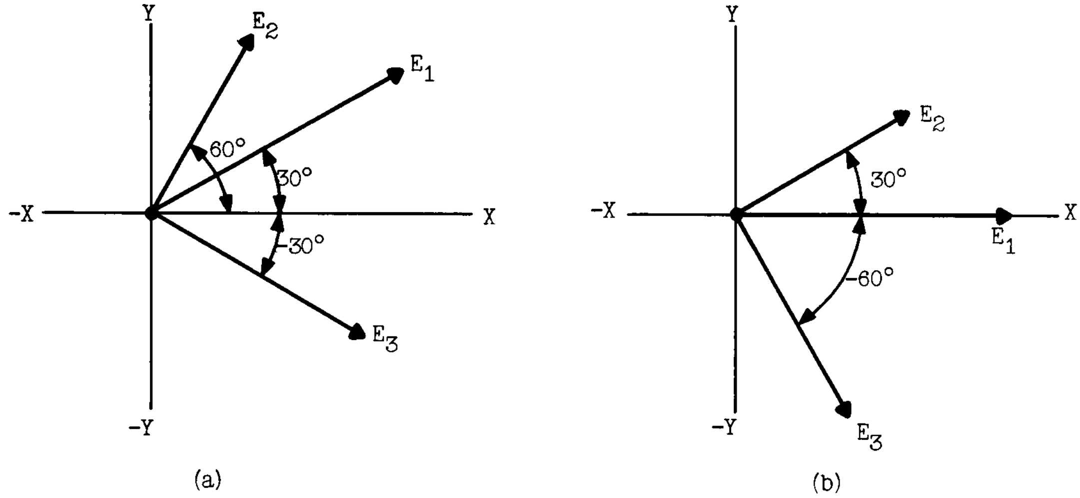



Fig. 2-11.

Sine and cosine waves may also be shown vectorially in proper phase relationship. Consider the four voltages:

$$e_1 = E_{1m} \sin\omega t$$

$$e_2 = E_{2m} \sin(\omega t + 60°)$$

$$e_3 = E_{3m} \cos\omega t$$

$$e_4 = E_{4m} \cos(\omega t + 30°)$$

In Fig. 2-12, vectors $E_1$ and $E_2$ are first plotted, using $E_1$ as reference along the X-axis. Next, the phase relationship of $E_3$ is determined as follows. Because a cosine wave is simply a sine wave advanced by 90 degrees, $e_3 = E_{3m} \sin(\omega t + 90°)$; and the vector, $E_3$, is plotted leading $E_1$ by 90 degrees. By the same reasoning, $e_4 = E_{4m} \sin(\omega t + 120°)$; this can be plotted as vector $E_4$ in Fig. 2-12.

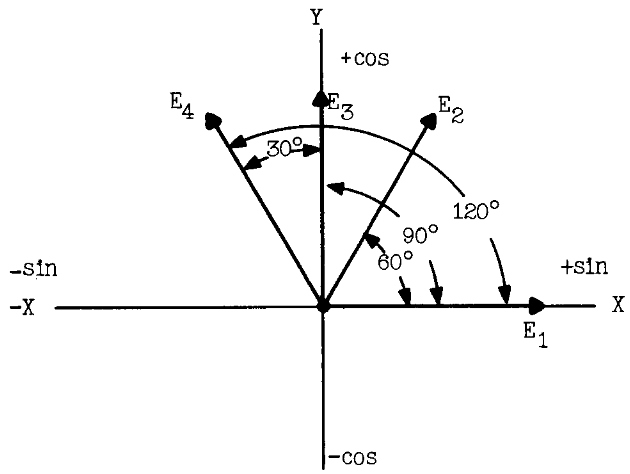



Fig. 2-12. Plotting of sine and cosine terms.

It should be noted in Fig. 2-12, that $e_1$ is a positive sine wave and is plotted along the positive X-axis. Also $e_3$ is a positive cosine wave and is plotted along the positive Y-axis. It follows that $-\sin\omega t = \sin(\omega t + 180°)$ and must therefore be plotted along the minus X-axis. Then $-\cos\omega t = \sin(\omega t - 90°)$ and is likewise plotted along the minus Y-axis. This provides a basis for labeling a set of coordinate axes as shown in Fig. 2-13.

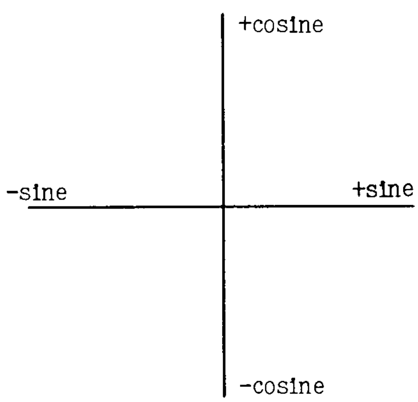



Fig. 2-13.

By the use of this set of coordinate axes the following voltages are plotted vectorially in Fig. 2-14. Note especially the vector representing $e_2$. It is a minus sine term with a phase of +20, and this places it 20 degrees advanced counterclockwise from the minus sine axis.

$$e_1 = E_{1m} \sin(\omega t + 20°)$$

$$e_2 = -E_{2m} \sin(\omega t + 20°)$$

$$e_3 = E_{3m} \cos(\omega t - 45°)$$

$$e_4 = -E_{4m} \cos(\omega t - 45°)$$

$$e_5 = -E_{5m} \cos(\omega t + 30°)$$

$$e_6 = E_{6m} \cos(\omega t + 30°)$$

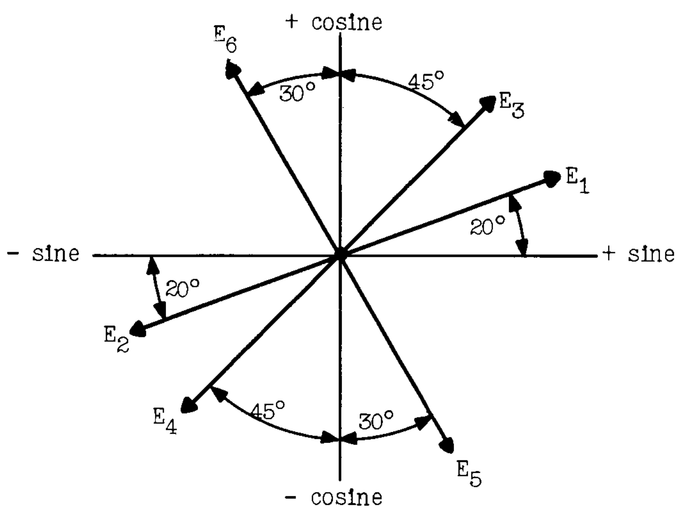



Fig. 2-14

## 6.2 Carrier and Sideband Vectors

The general equation for the composite amplitude modulated wave is plotted vectorially as follows:

$$e = E_{cm} \sin\omega_c t - \frac{mE_{cm}}{2} \cos(\omega_c + \omega_a)t + \frac{mE_{cm}}{2} \cos(\omega_c - \omega_a)t \quad (2\text{-}4)$$

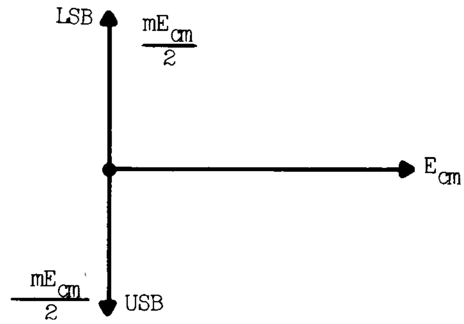



Fig. 2-15. The carrier, upper sideband, and lower sideband as represented by plus sine, minus cosine, and plus cosine functions, respectively.

There are several concepts to keep in mind when representing AM by vectors. The carrier vector is usually considered to have unit length with respect to the sideband voltage. Because each sideband contains one-half the total sideband voltage($mE_{cm}/2$), each vector is drawn to this proportional scale.

The carrier vector rotates with an angular velocity, $\omega_c$. The rate of rotation of the upper sideband vector is $\omega_c + \omega_a$, which signifies a faster rotation rate than the carrier vector. The lower sideband rotation rate is $\omega_c - \omega_a$, which is a slower rotation rate than the carrier.

All three vectors rotate in a counterclockwise direction. Due to the various angular velocities, the upper sideband vector appears to rotate counterclockwise and the lower sideband vector appears to rotate clockwise when the carrier is stopped for a reference.

Although a vector diagram represents a dynamic condition, any one vector diagram holds true only for a given instant of time, and it is therefore necessary to draw a diagram for each instant of time that the vector relationship is desired. For convenience, the vector diagrams for AM are normally drawn for three different relationships:

1. When the sidebands are not adding to or subtracting from the carrier. At this instant the amplitude of the envelope is the same as the maximum amplitude of the carrier component, and the vector diagram shows that the two sideband vectors cancel [see part (a) of Fig. 2-16].

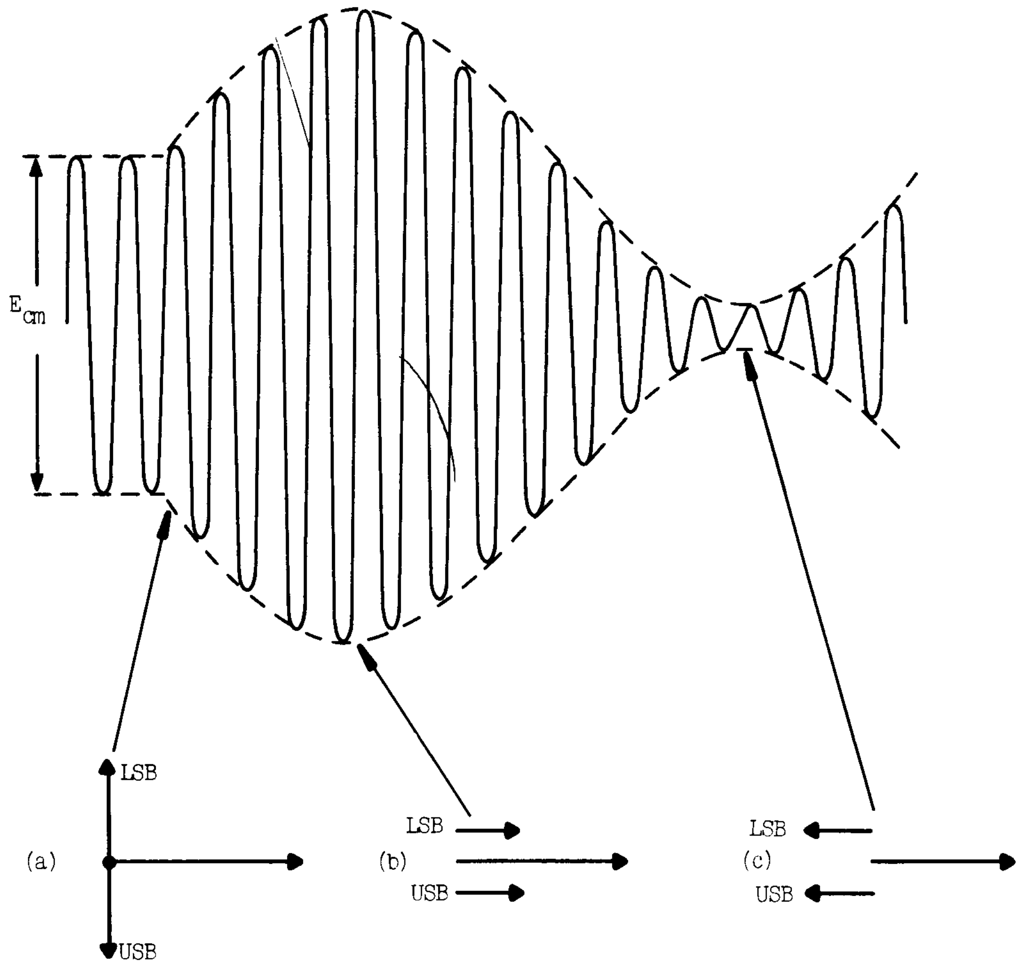



Fig. 2-16. Vector diagrams of AM wave.

2. When the sideband vectors combine and add to the carrier. At this instant the instantaneous value of the composite wave is maximum. See Part (b).

3. When the sideband vectors combine and subtract from the carrier, as illustrated in Part (c) Fig. 2-16. At this time the instantaneous value of the composite wave is minimum.

## 7.0 POWER CONSIDERATIONS

Several attempts have been made to evaluate SSB systems and to compare them with the more familiar AM systems. However, there is no single method of comparison that will be satisfactory to all.

Single-sideband systems have gains over AM systems ranging from 3 to 12 db, depending on the reference. Some of this gain is realized because of the narrower bandwidth of the receiver, but in this discussion we will be concerned with the relative sideband power available for detection at a receiver whose bandwidth is sufficient to receive both sidebands.

Suppose that a transmitter does not radiate the carrier, but only the two sidebands. There is no difference in the available sideband power at the receiver, with or without the carrier. One definite conclusion follows. If a 100-watt carrier is modulated 100% by a sine-wave tone, the composite AM wave will contain 150 watts and there will be 50 watts in the sidebands. Fifty watts in a double-sideband suppressed-carrier signal having the same capabilities as 150 watts of AM gives the suppressed carrier signal a 4.77 db gain advantage.

$$\text{db} = 10 \log \frac{150}{50} = 4.77$$

The power available at the receiver from a transmitted 50 watt single-sideband signal is 3 db below a double sideband signal containing 25 watts in each sideband (50 watts total). This is explained as follows.

In the detection process the double-sideband signal produces two sideband voltages that combine. They may be represented by two sideband vectors,

$E_L$ and $E_u$; and for convenience, the amplitude of each is represented as 5 volts (power being proportional to $E^2$, $5^2 = 25$). The vectors appear to rotate in opposite directions at a rate equal to the modulating frequency. When the two vectors come together in one direction they combine to produce 10 volts (positive). They produce 10 volts negative when combining in the opposite direction. This is a peak-to-peak voltage of 20 volts. See Fig. 2-17(a).

The voltage of the received SSB signal of 50 watts is represented vectorially in Part (b) as having an amplitude of 7.07 volts ($7.07^2 = 50$). During rotation, this sideband vector goes from 7.07 volts in one direction to 7.07 volts in the other direction for a peak-to-peak voltage of 14.14. Then the gain of double-sideband over single-sideband is:

$$\text{db} = 10 \log \frac{\frac{20^2}{R}}{\frac{14.14^2}{R}} = 10 \log \frac{20^2}{14.14^2} = 3 \text{ db}$$

Thus, we can conclude that SSB has a 1.77-db advantage over AM because double-sideband suppressed carrier has a 4.77 db advantage over AM and a 3 db advantage over SSB.

In the preceding discussion a comparison was made using average power as the reference. In making a comparison using peak power as the reference, one must remember (page 16) that the peak instantaneous power of a 150 watt composite AM wave is 400 watts. Then if we increase the SSB signal to a peak power of 400 watts, 200 watts average, an additional 1.24 db gain will be realized as db = 10 log 200/150 = 1.24 db. Then using peak powers as the reference gives single-sideband

1.77 + 1.24 = 3 db gain over AM.

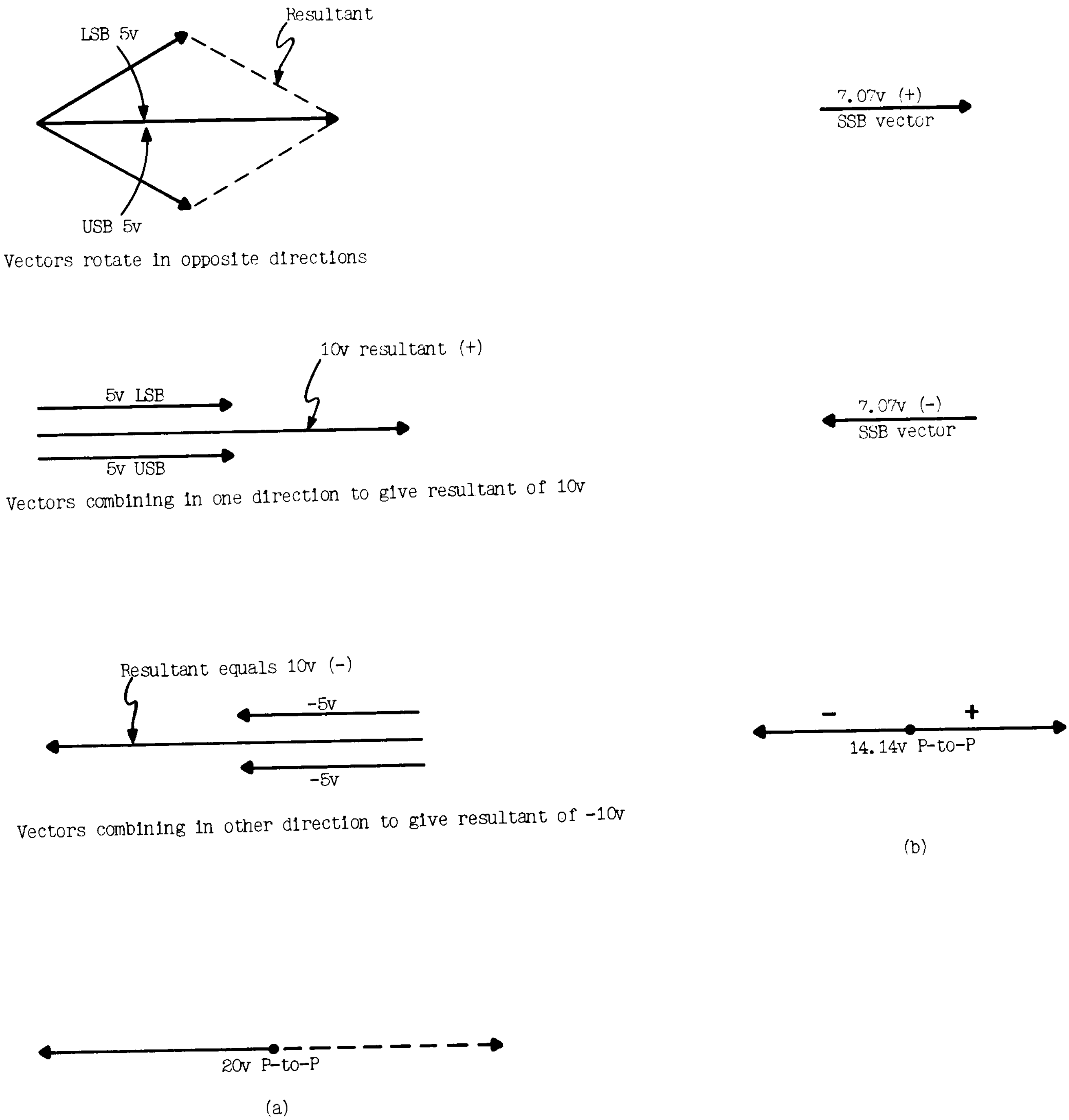



Fig. 2-17. Vector diagrams showing peak-to-peak voltages resulting from detected DSB and SSB signals.

In this discussion, only the radiated signals have been considered. It will be shown later that narrowing the bandwidth of a receiver for SSB gives an additional 3 db advantage as far as random noise is concerned.

## 8.0 SUMMARY

In this review of AM it was shown that an AM wave is developed by varying a carrier in amplitude at the modulating frequency. This process of modulation produces sum and difference frequencies, called sidebands. The composite wave is not composed of only one frequency, but is made up of a constant-amplitude carrier that is added to and subtracted from by the sidebands.

Modulation factor was defined as $E_{sm}/E_{cm}$ Using this relationship, graphical analysis, and trigonometric identities, the general equation for the instantaneous voltage (e) for a composite wave was developed.

The general equation consisted of three terms:

1. One term having an angular velocity $\omega_c$, represented the carrier.
2. Another term, having an angular velocity equal to $\omega_c + \omega_a$, represented the upper sideband.
3. The remaining term, whose angular velocity was equal to $\omega_c - \omega_a$, represented the lower sideband.

It should be remembered that this review has not included modulation by several frequencies (such as voice). Such a treatment is beyond the scope of this course at the present time.

Figure 2-10 is probably the most important figure in this chapter because it shows the separate components of an AM wave and how they combine to form the composite wave. If Fig. 2-10 is not understood at this time, the figure and the corresponding reading material should be reviewed. On Pages 14 and 15, equations were developed to show the relationship of the power in the carrier and sidebands for any degree of modulation (m).

However, the power relationships were developed for the 100% modulated wave; and to simplify the discussion, 100% modulation was assumed in the power comparison that followed. The power comparison concluded that single-sideband has a 1.77 db advantage over AM when average power is used as a reference, and a 3 db advantage when peak power is used as the reference.

A review of plotting trigonometric functions lead to the analysis of AM by means of vectors. Vectors were used in the power comparison and should be very helpful throughout the course.

PRACTICE PROBLEMS 2-1 (Answers at back of book.)

Choose the correct answer in the first four questions.

1. A 3.5-megacycle carrier is modulated by a 300-cycle tone. The resulting AM wave contains frequency components of

   A. 3.5 Mc, 3.2 Mc, and 3.8 Mc.

   B. -300 cycles, +300 cycles, and 3.5 Mc.

   C. 3500 Kc, 3499.7 Kc, and 3500.3 Kc.

   D. 3500 Kc, 3497 Kc, and 3503 Kc.

2. A transmitter having an operating frequency of 10 megacycles is transmitting an upper-sideband signal containing 3000-cycle modulation. What is the frequency of the transmitted signal?

   A. 10, 000 Kc.

   B. 10, 003 Kc.

   C. 10, 000.3 Kc.

   D. 9, 997 Kc.

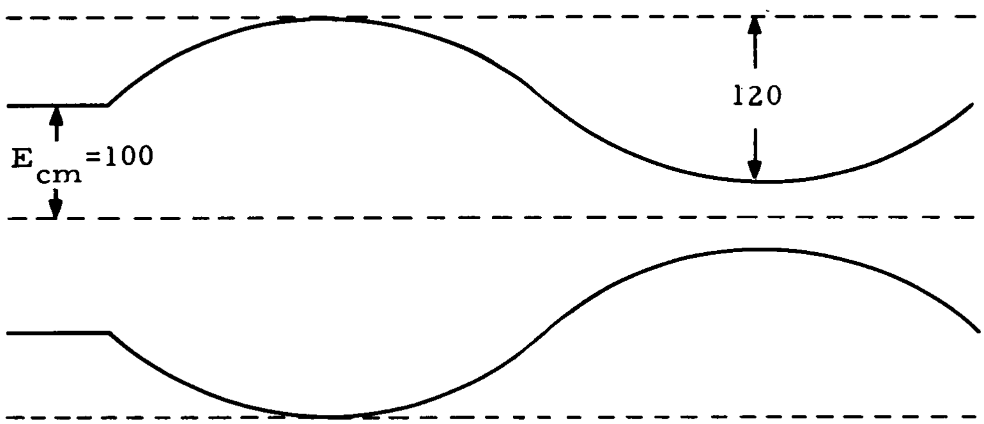
3. What is the modulation factor, m, of the envelope shown in the figure?

A. 0.600.

B. 0.833.

C. 0.500.

D. 0.750.




4. An RF carrier of 2000 volts rms is modulated to a depth of 15%. What is the maximum total sideband voltage, $E_{sm}$?

A. 300 volts.

B. 150 volts.

C. 133 volts.

D. 424 volts.

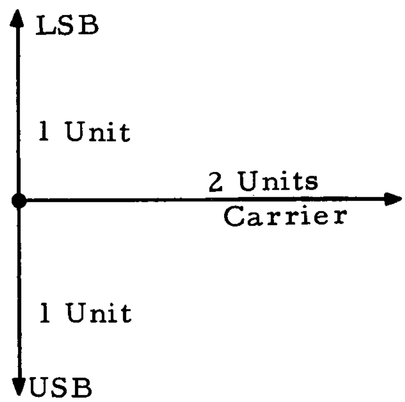
5. What audio power is required to produce 100% modulation of a 1-kilowatt carrier?

6. A 1000-volt, 746-kilocycle carrier is modulated with a 400-volt, 800-cycle audio signal. What is the frequency and amplitude of the lower sideband voltage? Voltages given are rms value.

7. The carrier voltage for a 35% modulated AM wave is 1000 volts rms. What is the rms value of the lower sideband voltage?

8. What is the ratio of the total sideband power to the carrier power in the previous problem?

9. What is the modulation factor for an AM wave represented by the vector diagram below?



10. A 750-watt carrier is modulated 50%. What is the sideband power, $P_s$, in the resulting AM wave?

11. A 100-volt carrier is modulated 50%. Before expansion, the expression for the AM wave is $(100 + 50 \sin \omega_a t)(\sin \omega_c t)$. What is the expression when it is expanded to three terms showing the carrier, lower sideband, and upper sideband components?

12. An AM wave is applied to a 1000-ohm resistor. What is the percentage of modulation if the carrier component is 400 volts rms and there is a power of 20 watts in the total sideband component?

13. Define peak-envelope power, PEP.

ATTENTION

Examination 2 is to be worked at this time. This exam covers the material in Chap. 2. It is included with the material.

## CHAPTER 3

## BALANCED MODULATORS

### 1.0 INTRODUCTION

In an AM transmitter, the modulation is normally applied at the final class C amplifier in series with the B+ supply voltage to that stage. This is high level modulation which means that the modulation is applied after the carrier has been amplified to full transmitting power level. In contrast, the modulation process in an SSB transmitter is performed at a low level. Low level modulation has the disadvantage of requiring linear amplifiers after the modulating stage, but it is necessary to perform the modulation at relatively low frequency and power levels to effect proper elimination of the undesired sideband.

The modulation of an SSB transmitter is normally performed in a balanced modulator where the intelligence signal is heterodyned with a carrier to raise the frequency of the audio intelligence signal. This heterodyning process produces in the output of the balanced modulator two frequencies not present in the input. They are the sum and difference frequencies (upper and lower sidebands). The carrier (one of the components of the AM wave) is not present in the output of the balanced modulator. This is the chief reason for using a balanced modulator.

To obtain a single-sideband signal, the double-sideband suppressed carrier output of the balanced modulator is fed to a suitable filter or a phasing system which eliminates the unwanted sideband. Thus, a single-sideband signal is made available for linear amplification and subsequent transmission.

The block diagram of a basic SSB transmitter in Fig. 3-1 shows the order in which the functions are performed in the transmitter. This block diagram represents a transmitter utilizing the filter method of SSB generation. Although we are primarily concerned with balanced modulators in this chapter, a brief description of the block diagram is given now to help visualize the other functions performed in an SSB transmitter.

The audio that modulates the transmitter is fed to a balanced modulator where it is raised in frequency by mixing with the 250-kilocycle signal from the carrier oscillator. The mixing process produces sum and difference frequencies; the carrier is eliminated by the symmetry of the balanced modulator. The two sidebands in the output of the balanced modulator are fed to an upper sideband filter where the lower sideband is eliminated. The upper sideband is fed to a mixer where it is heterodyned with an oscillator signal to raise it to the operating frequency and is amplified in linear amplifiers before it is radiated from the antenna.

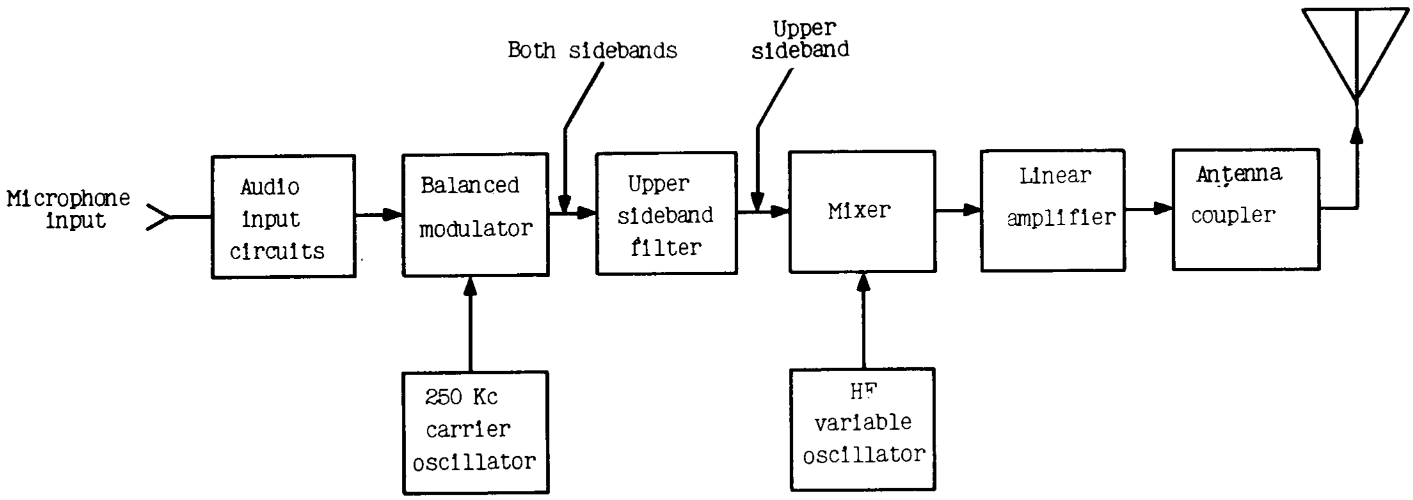



Fig. 3-1. Block diagram of basic single-sideband transmitter utilizing the filter method of single-sideband generation.

A second method of SSB generation, the phasing method, utilizes two balanced modulators. Because the filter and phasing methods both require balanced modulators, it is appropriate to discuss them in this chapter. The two methods of SSB generation are discussed in the next chapter.

## 2.0 VACUUM-TUBE BALANCED MODULATORS

Crystal diode rectifiers or vacuum tubes may be used in balanced modulator circuits. Vacuum-tube modulators are normally easier to understand; therefore, they will be discussed first.

In discussing balanced modulators, we shall be more concerned with phase relationships of the applied signals than with their amplitudes; therefore, amplitudes are considered to be unity.

In the balanced modulator of Fig. 3-2, the audio modulating signal is applied through a transformer to the grids of V1 and V2 in phase opposition. The RF carrier is applied in phase to the grids of the two tubes. Before

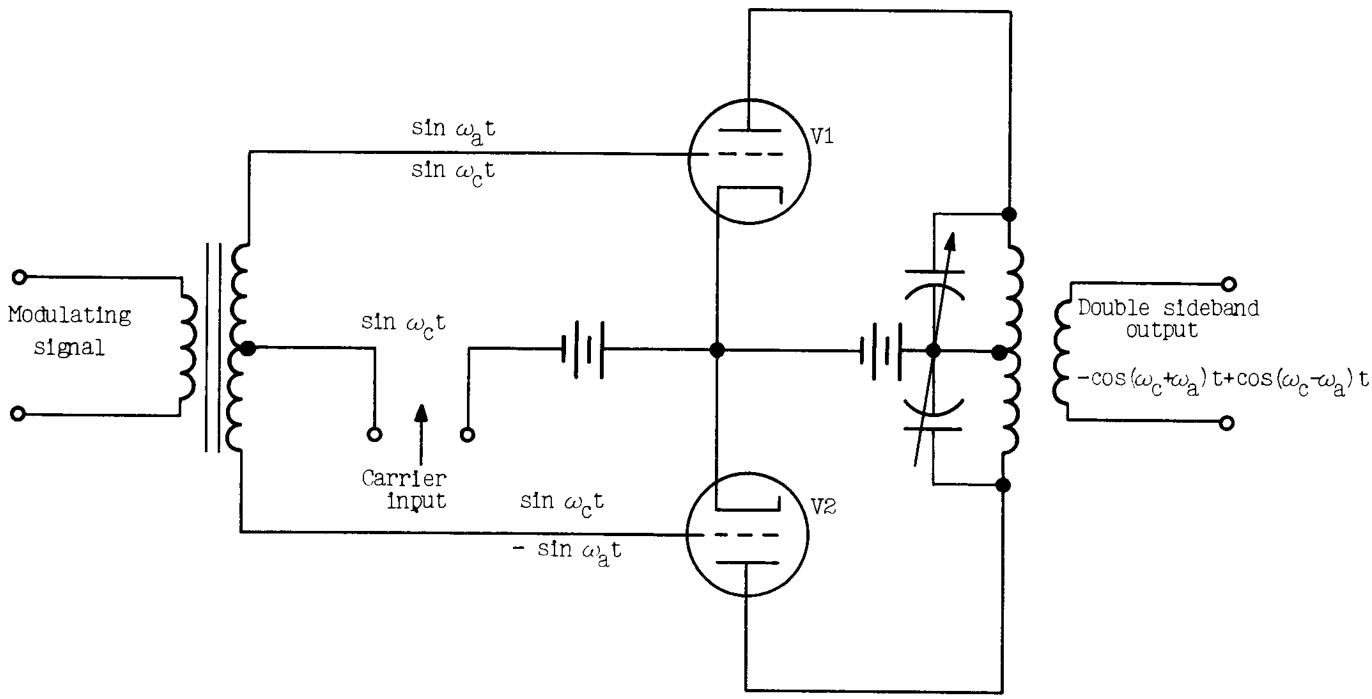



Fig. 3-2. Basic vacuum-tube balanced modulator in which the audio is applied to the grids in phase opposition and the carrier is applied to the grids in phase.

going into the mathematical explanation of the generation of a double-sideband signal in this circuit, let us digress for a moment to review the expression for the plate current transfer function of a vacuum-tube circuit.

The current in each vacuum tube is some function of the input grid voltage, $i = f(e)$. For example, if a tube curve is linear, $i = ke$, where k is a constant. Normally a tube curve is not linear, and the plate-current grid-voltage transfer curve is similar to Fig. 3-3. This curve can be expressed accurately by

$$i = k_o + k_1 e + k_2 e^2 + k_3 e^3 + \ldots$$

This curve is very closely approximated by the expression

$$i = k_o + k_1 e + k_2 e^2$$

which we will use in this discussion.

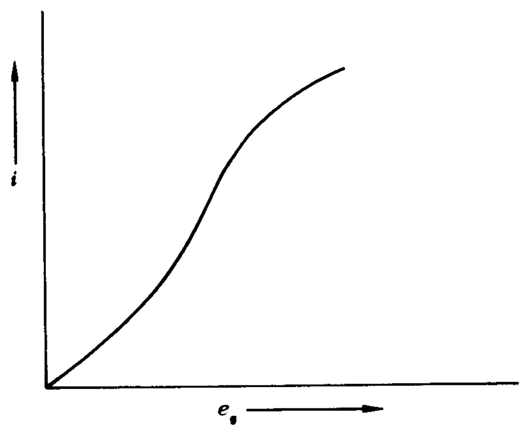



Fig. 3-3. Typical plate-current grid-voltage vacuum tube characteristic curve.

The audio input to the grid of V1 is given as $\sin \omega_a t$. The audio at the grid of V2 is 180° out of phase with the audio at V1 and is $\sin(\omega_a t + 180°) = -\sin \omega_a t$. The carrier is given as $\sin \omega_c t$ on the grid of both tubes. The instantaneous signal ($e_1$) appearing on the grid of V1 is then $\sin \omega_c t + \sin \omega_a t$; on the grid of V2, the instantaneous voltage ($e_2$) is:

$$e_2 = \sin \omega_c t - \sin \omega_a t$$

Substituting the instantaneous grid voltages in the general expression for i of the transfer curve gives the plate current of V1 as:

$$i_1 = k_o + k_1(\sin \omega_c t + \sin \omega_a t) + k_2(\sin \omega_c t + \sin \omega_a t)^2$$

$$= k_o + k_1 \sin \omega_c t + k_1 \sin \omega_a t + k_2 \sin^2 \omega_c t + 2k_2 \sin \omega_c t \sin \omega_a t + k_2 \sin^2 \omega_a t$$

and after using the two trigonometric identities:

$$\sin^2 A = \tfrac{1}{2}(1 - \cos 2A)$$

$$\sin A \sin B = -\tfrac{1}{2}\cos(A + B) + \tfrac{1}{2}\cos(A - B)$$

$$i_1 = k_o + k_1 \sin \omega_c t + k_1 \sin \omega_a t + k_2/2(1 - \cos 2\omega_c t) + k_2\left[-\cos(\omega_c + \omega_a)t + \cos(\omega_c - \omega_a)t\right] + k_2/2(1 - \cos 2\omega_a t)$$

and similarly the plate current for V2 is developed:

$$i_2 = k_o + k_1(\sin \omega_c t - \sin \omega_a t) + k_2(\sin \omega_c t - \sin \omega_a t)^2$$

$$= k_o + k_1 \sin \omega_c t - k_1 \sin \omega_a t + k_2 \sin^2 \omega_c t - 2k_2 \sin \omega_c t \sin \omega_a t + k_2 \sin^2 \omega_a t$$

$$= k_o + k_1 \sin \omega_c t - k_1 \sin \omega_a t + k_2/2(1 - \cos 2\omega_c t) - k_2\left[-\cos(\omega_c + \omega_a)t + \cos(\omega_c - \omega_a)t\right] + k_2/2(1 - \cos 2\omega_a t)$$

Thus, it is seen that the plate current of the balanced modulator tubes contain dc, audio, carrier, upper and lower sideband, and second harmonic components.

The plate circuits are linked magnetically to the output circuit, and since the two plate currents are in opposite directions in the primary of the output transformer, the effect of the two currents is represented by taking the difference of $i_1$ and $i_2$. With a perfectly balanced circuit the similar terms which are positive in both expressions will disappear leaving

$$i_1 - i_2 = 2k_1 \sin \omega_a t - k_2 \cos(\omega_c + \omega_a)t + k_2 \cos(\omega_c - \omega_a)t$$

The first term is audio and will not be passed by the output circuit of the balanced modulator. The other two terms are the upper and lower sidebands.

There are many variations of vacuum-tube balanced-modulator circuits. A balanced modulator similar to the one used in the localizer of the Instrument Landing System is shown in Fig. 3-4. The same manner of analysis as used for the circuit of Fig. 3-2 can also be used for this circuit. However, we have chosen to discuss this circuit in a qualitative manner for those who prefer the intuitive discussion over the mathematical.

In Fig. 3-4 the RF excitation voltage is applied to the two grids in phase because they are connected in parallel; the plates are fed with audio voltages in phase opposition. The tubes are biased for class C operation and, therefore, there is no output from a tube except when the RF voltage on the grid is positive, and then only when the audio on the plate is also positive. It should be noted that no dc voltage is supplied to the plates; they are supplied with alternating voltages. Note also that when the plate of one tube is positive with respect to its cathode, the plate of the other tube is negative with respect to its cathode. Therefore, only one tube can conduct at a time. At any instant of time, the amplitude of the RF output of a tube is nearly proportional to the value of instantaneous plate voltage applied to that tube. As a result, the amplitude of the RF output varies at the frequency of the plate voltage.

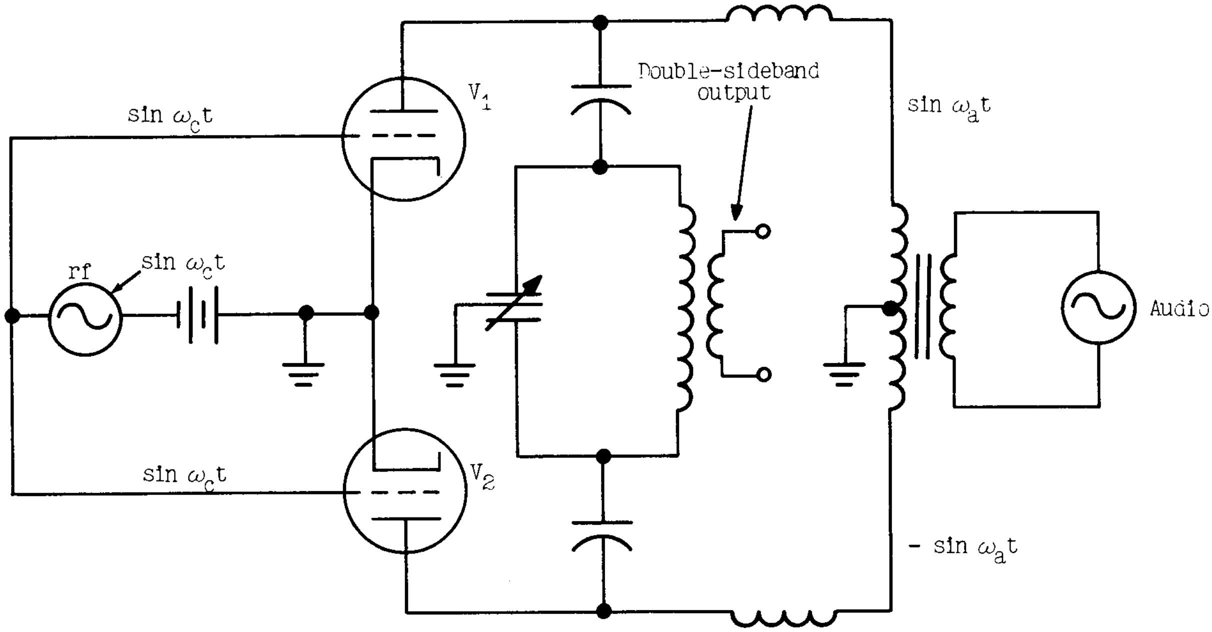



Fig. 3-4. Basic vacuum tube balanced modulator circuit employing ac on the plates.

During the time in which the plate of V1 is positive, the RF output voltage will appear as shown in Fig. 3-5(a).

When the audio plate voltage is reduced to zero, the RF in the plate tank circuit is also zero because both tubes are cut off.

The RF output of V2 is similar except that it conducts when the plate of V1 is negative and the ensuing plate current of V2 causes a circulating tank current in the plate circuit that is 180° out of phase with that produced when V1 conducts. Thus, the output of V2 would be as shown in Fig. 3-5(b). The combined output of the two tubes would appear as shown in Fig. 3-5(c) in which the RF voltage reverses phase at each time the plate voltages drop to zero. The output consists of both sidebands and is identical to the total sideband component mentioned in Chap. 2, while discussing the AM wave. See Fig. 2-10(c).

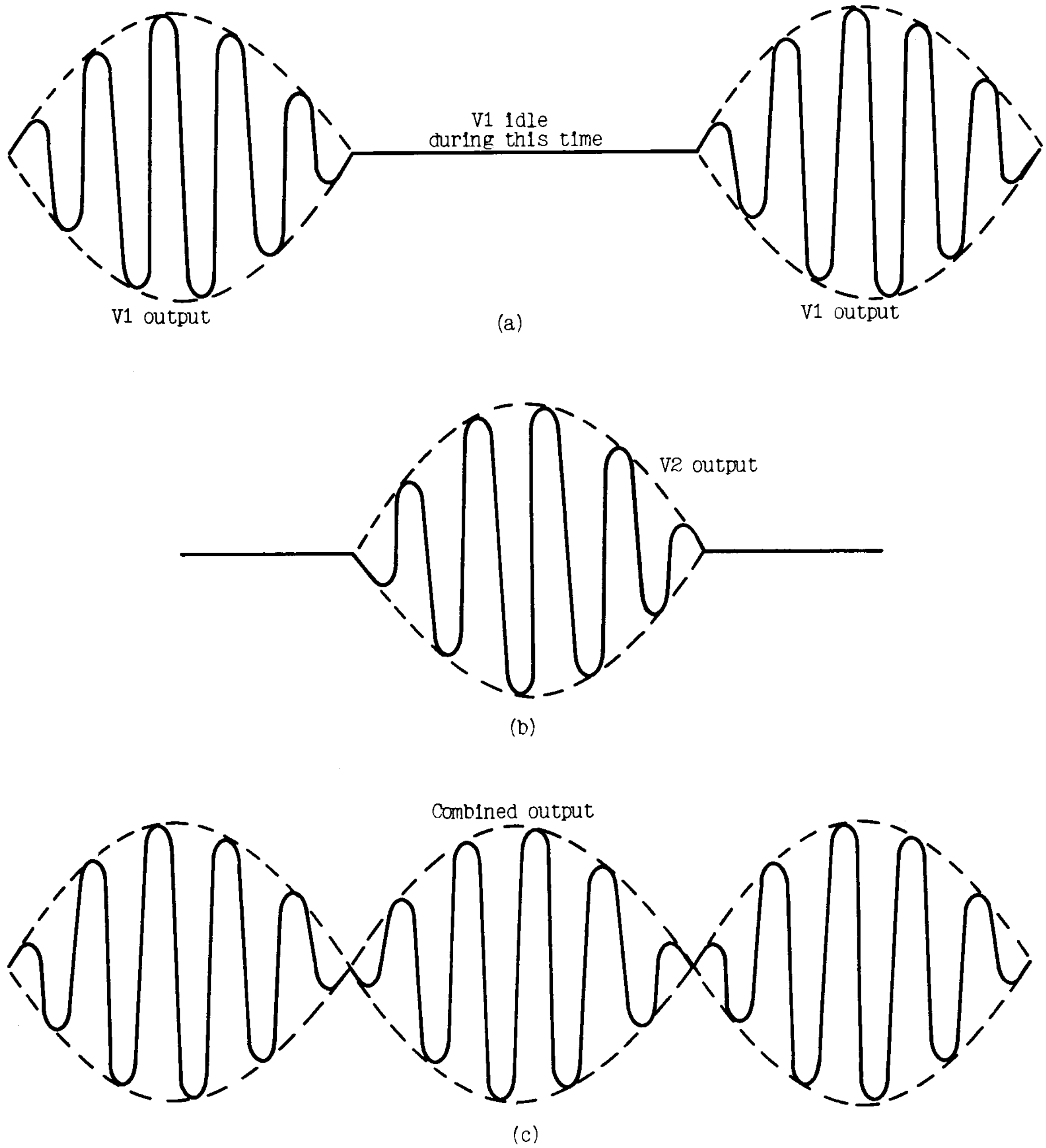



**Fig. 3-5. Output of sideband generator.**

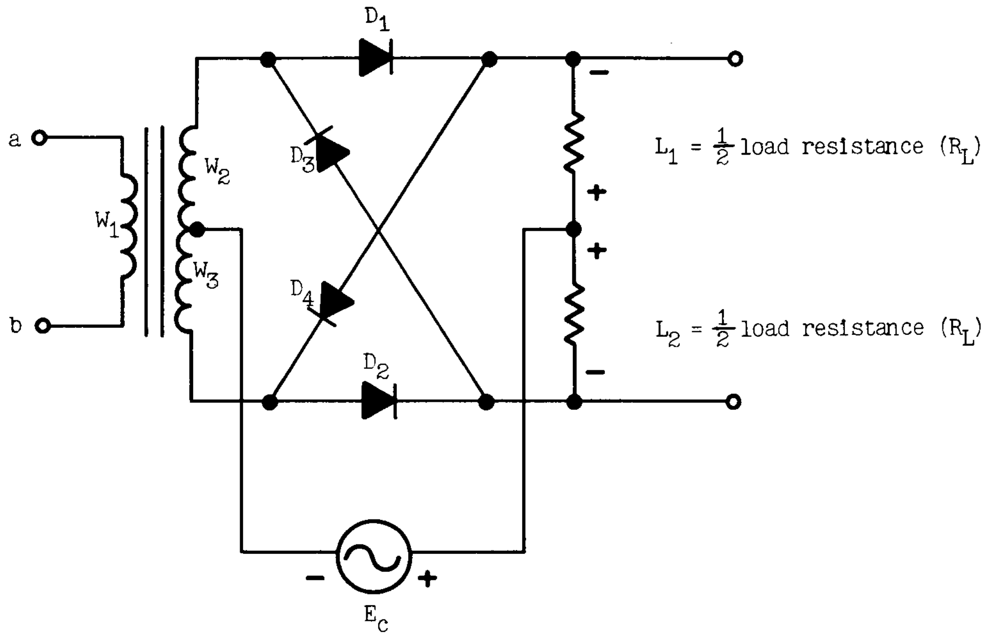
## 3.0 DIODE BALANCED MODULATORS

Diode balanced modulators have several advantages over tube modulators, which make their use in sideband application very desirable. Some of the characteristics of diodes are: they are stable, compact, require no filament power, and have longer life than tubes thus requiring less maintenance.

Balanced-modulator circuits employing rectifiers are many and varied. Three of the basic circuits are called ring, series, and shunt. The name in each case refers to the manner in which the diodes are connected in the circuit. The ring modulator is probably the most common of the three basic types. Because the theory of operation is nearly the same for all three, the ring modulator will be the only diode modulator discussed.

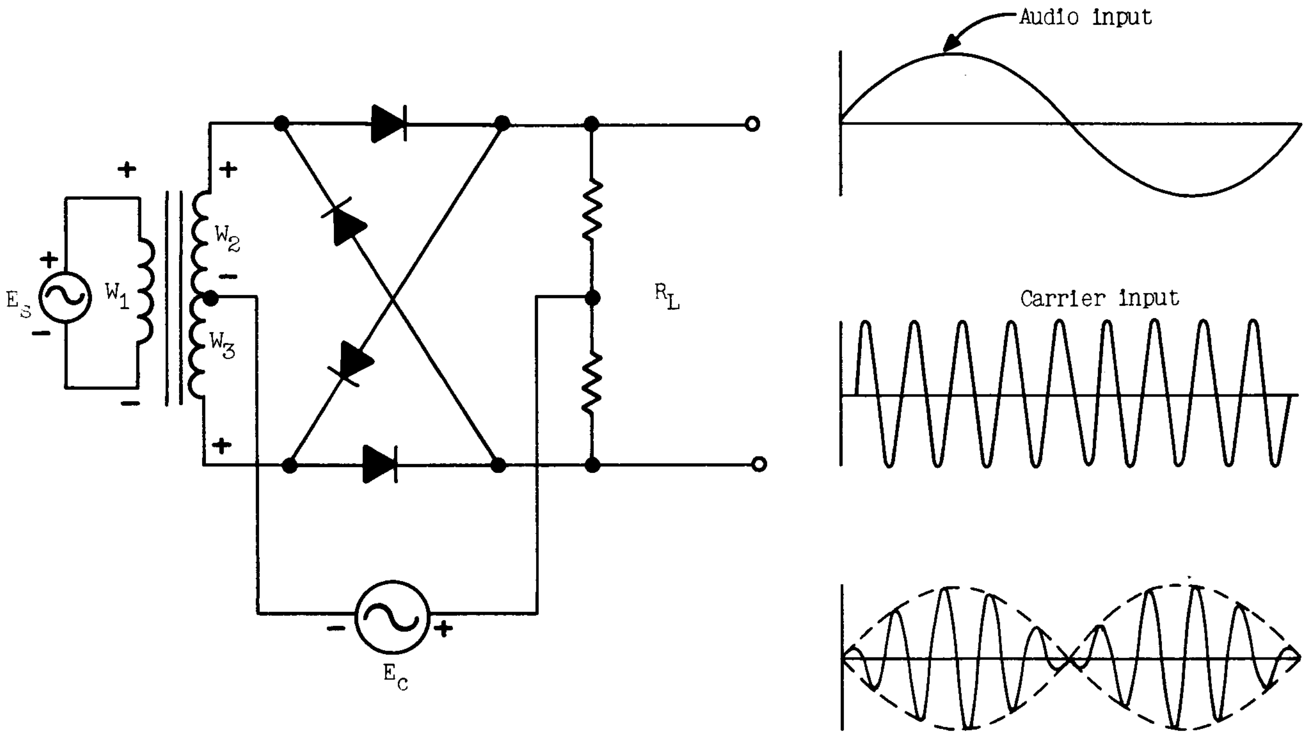
The purpose of the ring modulator is to modulate the carrier by means of a signal voltage and to produce in the output upper and lower sidebands without the carrier. The basic principle of the ring modulator, as is true of other balanced modulators, is to introduce the RF carrier in such a manner that it will not appear in the output of the stage. There will be an output signal, however, when both the audio modulation and RF carrier signals are present simultaneously at the modulator input. This output signal will consist of only the upper and lower sideband frequencies generated in the ring modulator by the mixing of the two input signals.

First, the operation of the modulator circuit will be considered when only the carrier signal is applied to the circuit; that is, in the absence of any audio modulating signal. In Fig. 3-6(a), with the carrier voltage, $e_c$, of the instantaneous polarity shown, the electron flow is from the left side of the carrier generator to the junction of $W_2$ and $W_3$. From this junction, electrons take two paths. One path is from the junction through $W_2$, $D_3$, and $L_2$, developing a voltage across $L_2$ of the polarity shown. The current in the other path (through $W_3$, $D_4$, and $L_1$) develops a voltage in $L_1$



D1
a
b
W1
W2
W3
D3
D4
D2
L1 = 1/2 load resistance (RL)
L2 = 1/2 load resistance (RL)
Ec


(a)



ES
W1
W2
W3
RL
Ec
Audio input
Carrier input
Double-sideband output


(b)

Fig. 3-6.

that opposes the voltage in $L_2$. With the carrier current flowing in opposite directions through the two halves of the load, the resultant carrier voltage across the total load ($L_1$ and $L_2$) is zero. When the polarity of the carrier input voltage, $e_c$, reverses, the current in each half of the load will reverse and there will be no output as before.

We have observed that there is no carrier output when there is no audio modulating signal input (Fig. 3-6(a)). Also, there is no audio output when there is no simultaneous carrier input because the diodes present a short circuit to the audio signal.

It is worth noting here that when either one of the input signals (audio or carrier) is not present at the modulator there is no output of the modulator, This is the reason there is no output from an SSB transmitter when no audio is entering the speech input circuits.

Let us now consider the signal voltages that will be developed across the load when $e_s$ and $e_c$ are applied simultaneously. Experience will tell us that the sum and difference frequencies in the output result from the mixing that takes place because of the nonlinear action of the diodes in the circuit. In an attempt to describe the operation when both signals are applied, we will show that the output will have the same waveform as shown in Fig. 3-6(b). This waveform represents a double-sideband signal which was illustrated in Fig. 2-8 and discussed in the accompanying text. The expression for this waveform can be separated into two terms, representing the upper and lower sidebands separately. Because the output of the ring modulator can be shown to have the same waveform as in Fig. 3-6(b) and Fig. 2-8, it follows that the output contains the upper and lower sideband frequencies.

To continue with the operation when both signals are applied simultaneously, the audio signal, $e_s$, having the instantaneous polarity indicated, aids the carrier current in $W_3$ and opposes the carrier current in $W_2$.

The currents are no longer equal in the two halves of $R_L$ and a voltage is therefore developed at the output. During the next half cycle of the carrier input, the audio signal maintains the same polarity because its frequency is much lower than the carrier frequency. Again there is an unbalance in the output and a resultant voltage is developed between the output terminals.

We see then, that when the positive half cycle of the audio is being applied, its effect is to unbalance the circuit, permitting the RF pulses to appear across the output. There is also an output on the negative half cycle of the audio as the circuit is unbalanced in the opposite direction. The instantaneous amplitude of the audio signal determines the amount of unbalance and, in turn, the instantaneous amount of unbalance determines the maximum amplitude of the RF output. It follows that the output is in the form of lobes and is a double-sideband signal as shown in Fig.3-6.

## 4.0 AN APPLICATION OF A DIODE BALANCED MODULATOR

Fig.3-7 shows the 250-kilocycle oscillator and balanced modulator section of the sideband exciter unit, which is found in the AN/FRT-39 single-sideband transmitter. The method of sideband generation employed in this unit is not to be confused with the phasing method of sideband generation because of its two balanced modulators. This unit consists of two channels which employ the filter method of SSB generation. One channel utilizes the upper sideband and the other channel utilizes the lower sideband; each channel is modulated with different intelligence. This method is called independent sideband operation.

The audio inputs at T103 and T104 contain different intelligence. The RF carrier from the 250-kilocycle oscillator, V105, is fed between the center of R265 and ground for the upper modulator, CR115, and between the center of R266 and ground for the lower modulator, CR116. The 250-kilocycle carrier is canceled in each modulator and each modulator output (at T125 and T126) contains upper and lower sidebands.

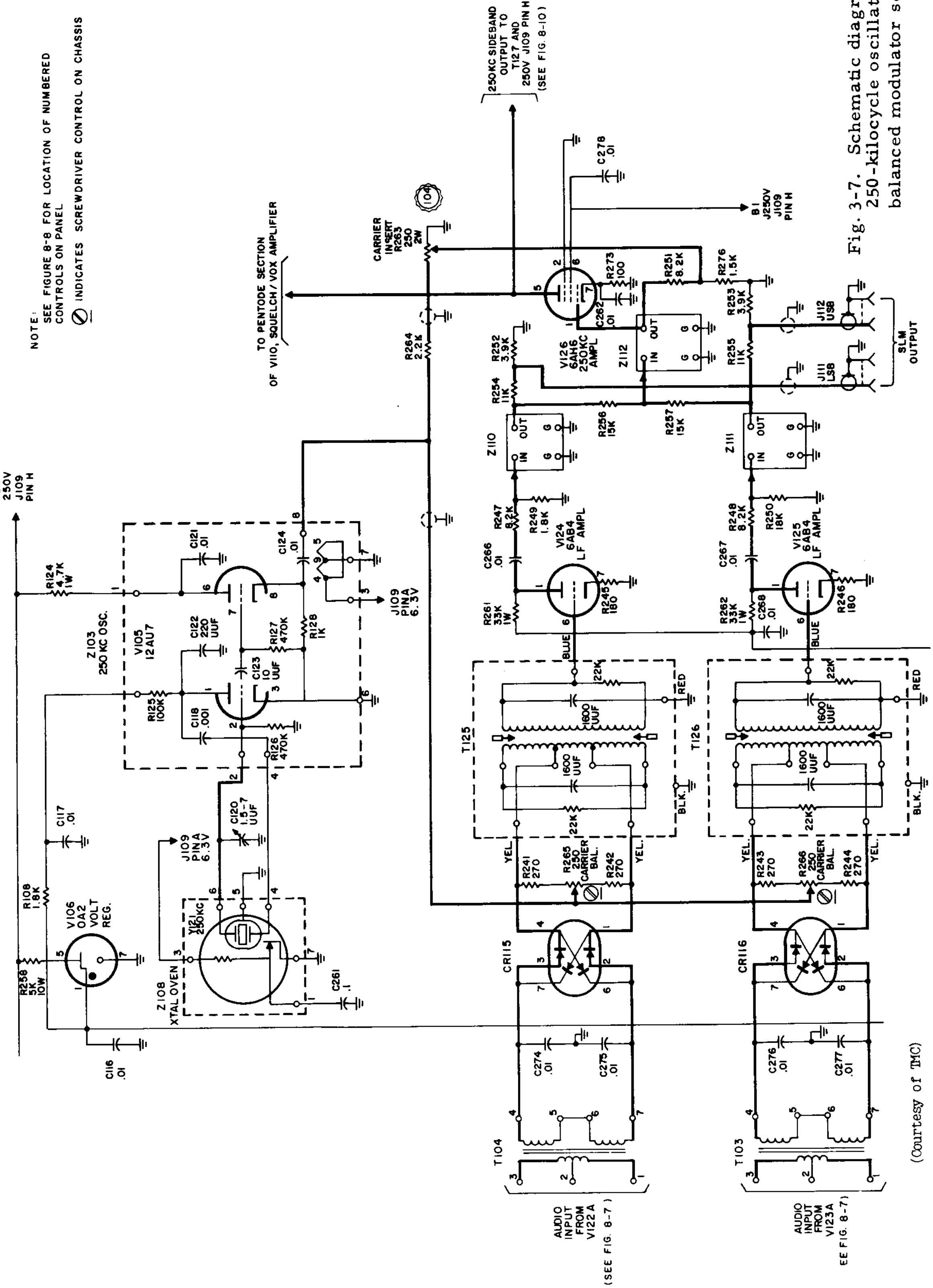



Fig. 3-7. Schematic diagram SBE-3, 250-kilocycle oscillator and balanced modulator section.

The output of the upper modulator is passed on to the grid of V124 where it is amplified and fed to sideband filter Z110 which rejects the upper sideband. The output of Z110 consisting of the lower sideband is fed through Z112 where the 250-kilocycle carrier, that remains due to imperfect balancing in the modulator, is filtered out.

Both sidebands from the lower modulator CR116 are fed to the grid of V125 to be amplified and passed on to Z111 where the lower sideband is eliminated. The upper sideband is passed on through Z112 and amplified by V126. The output of V126 contains LSB and USB signals, each sideband containing different information. In practice, one sideband could be used as a teletype channel and the other used as a voice channel.

## 5.0 SUMMARY

Balanced modulators are used to perform the modulation process in SSB transmitters. The process raises the modulating signals to higher RF frequencies. The balanced modulator has audio and RF inputs which combine in the modulator to form sum and difference frequencies. The carrier is balanced out (or reduced to a low level) because of the symmetry of the circuit.

Diodes are used more often than tubes in the modulators of sideband applications. They are more reliable and their stability is high in comparison with vacuum tubes.

Balanced modulators are used in both the filter method and phasing method of single-sideband generation. The filter method requires one modulator for each channel while the phasing method requires two modulators. In the filter method, the undesired sideband is eliminated by a very precise sideband filter. The characteristics required of sideband filters and the filter method of single-sideband generation will be discussed in Chap. 4.

PRACTICE PROBLEMS 3-1

1. The balanced modulator circuit of an SSB transmitter

   A. eliminates the undesired sideband.

   B. raises the audio in frequency.

   C. modulates the carrier.

   D. eliminates both sidebands.

2. In a filter method of SSB generation containing one channel,

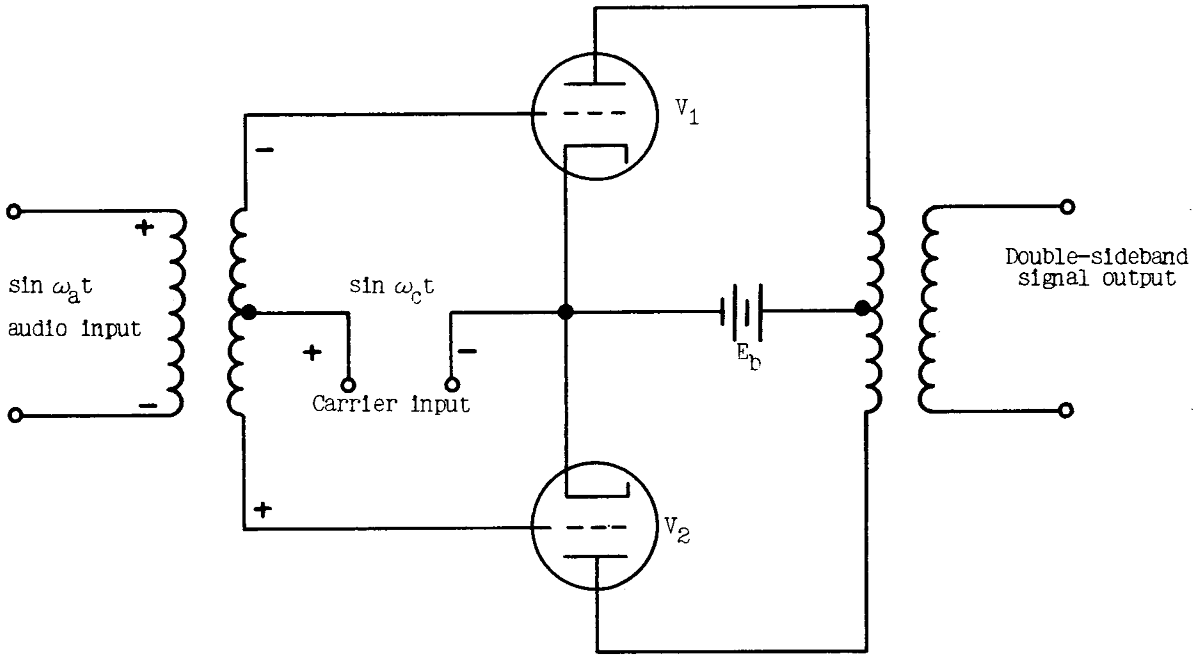
   A. a double-sideband signal is generated in a system of two balanced modulators and then passed on to a filter where one sideband is eliminated.

   B. a single-sideband signal is generated in a system of two balanced modulators and passed on to a filter which eliminates the carrier.

   C. a single-sideband signal is generated in a balanced modulator and fed to a filter which removes the residue carrier.

   D. a double-sideband signal is generated in a balanced modulator and fed to a filter which eliminates one sideband.

3. With the instantaneous input voltages at the points shown in the following figure, what is the voltage on the grid of V2?




A. $(\sin \omega_c t - \sin \omega_a t)$

B. $(\sin \omega_c t + \sin \omega_a t)$

C. $(\sin \omega_c t)(\sin \omega_a t)$

D. $\frac{1}{2} \cos(\omega_c - \omega_a)t - \frac{1}{2} \cos(\omega_c + \omega_a)t$

4. Referring to the figure in prob. 3, and assuming that the expression $i_p = k_o + k_1 e_g + k_2 e_g^2$ is the equation for the plate current of both V1 and V2, what is the expression for the plate current of V1 with the inputs shown. (Leave your answer in a form that will show the sideband components.)

ATTENTION

Examination 3 is to be worked at this time.
This exam covers the material in Chap. 3.
It is included with the material.

# CHAPTER 4

# FILTER AND PHASING METHODS OF SINGLE-SIDEBAND GENERATION

## 1.0 INTRODUCTION

It was mentioned earlier that there are two common methods of generating a single-sideband (SSB) signal. The most familiar of these two is the filter method in which a double-sideband signal is generated in a balanced modulator and then passed through a selective filter which rejects one of the sidebands. This method requires the use of special filters; these filters are one of the topics of this chapter.

The phasing method of SSB generation removes the undesired sideband by the use of a balancing process. This method uses two balanced modulators (discussed in Chap. 3), an audio 90-degree phase-difference network, and an rf 90-degree phase-difference network. Phasing theory and phase-shift networks are discussed briefly because this method is not as important in FAA applications as the filter method.

Heterodyning is necessary to place the sideband signal in the proper location in the rf spectrum. Heterodyning techniques are discussed briefly in this chapter.

## 2.0 FILTER METHOD

The transmission and reception of SSB demands the use of highly selective filters. They are used to eliminate undesired frequency components while passing the desired signal with little or no attenuation. For example, a double-sideband rf signal containing modulation frequencies between 300 and 3,000 cycles has a 600-cycle spread between the lower and upper sidebands; see Fig. 4-1(a). The function of the sideband filter is to eliminate almost completely those frequencies that are only 600 cycles removed from those that need to be passed with little or no attenuation.

The design of filters with characteristics sharp enough to select the desired sideband and reject the other sideband is difficult at relatively low radio frequencies (100 kc); it is almost impossible at the higher rf frequencies (say 40 mc). It is more difficult to separate the sidebands at the higher carrier frequencies because the ratio of sideband separation

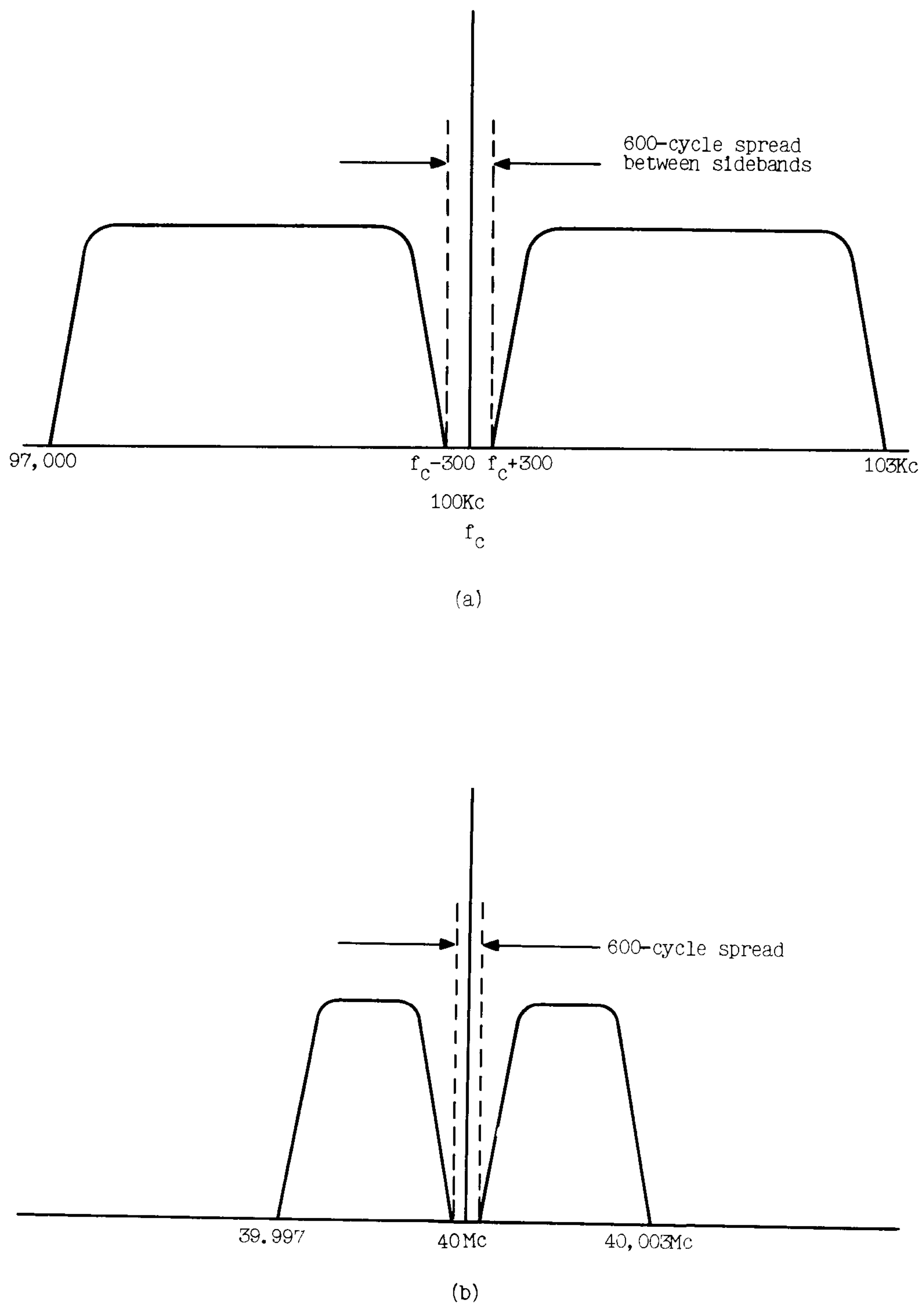



Fig. 4-1. 600-cycle sideband separation at carrier frequency of 100 kilocycles and 40 megacycles.

to carrier frequency is much smaller at the higher carrier frequencies. For example, Fig. 4-1(a) represents a 100-kc carrier with sidebands existing 300 cycles on both sides of this carrier. The ratio of sideband separation to carrier frequency is

$$\frac{600}{100 \times 10^3} = 0.006.$$

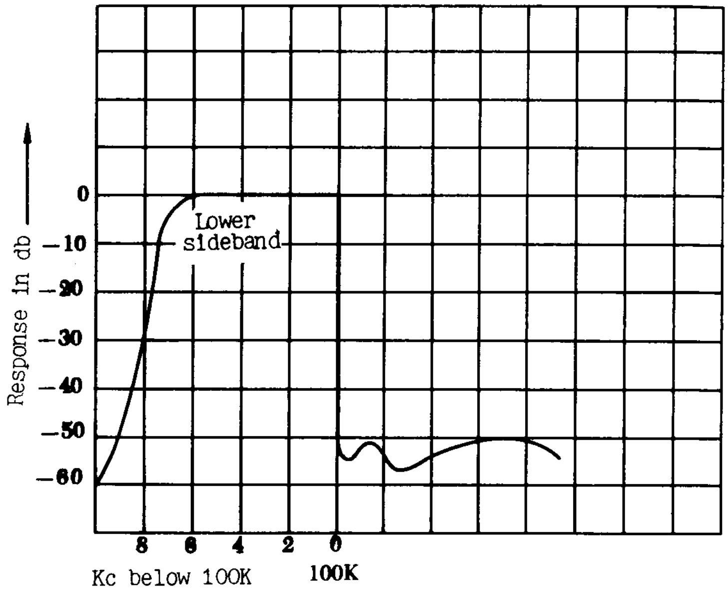
Figure 4-1(b) shows the same sideband separation of 600 cycles and a 40-mc carrier. The sideband separation to carrier ratio in this case is

$$\frac{600}{40 \times 10^6} = 0.000015.$$

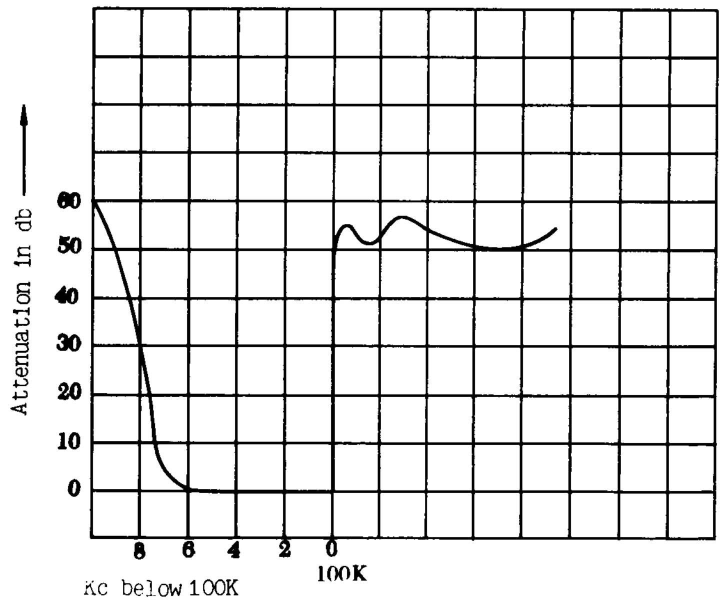
The smaller the sideband separation to carrier frequency ratio, the more difficult it is to reject one sideband without significantly attenuating the other. For several reasons, including the one mentioned above, it is not feasible to perform the filtering process at the final operating frequency. Another reason is that a separate filtering unit would be needed for each operating frequency. Therefore, the undesired sideband is filtered at a lower fixed frequency, and a variable frequency oscillator is used to heterodyne the single-sideband signal to any desired operating frequency in the transmitter tuning range. If the sideband signal is generated at too low a frequency, multiple frequency conversion must be used in order to reach a relatively high operating frequency. Multiple conversion brings about a considerable increase in circuit complexity with an accompanying loss in economy and reliability. Because of these factors, it is very desirable to generate the single sideband at a frequency which requires a minimum number of frequency conversions to reach the operating frequency. This requires generation of the sidebands high enough in frequency to reduce the number of conversions required and at the same time low enough in frequency to eliminate effectively (by means of a sideband filter) the undesired sideband.

## 2.1 Filter Characteristics

Filter characteristics enable one to better understand the rejection qualities of filters. A few of the more important characteristics are given here. However, our discussion of filters is limited because most sideband filters are hermetically sealed and require no field maintenance.




(a) Response curve of sideband filter




(b) Attenuation curve of sideband filter

Fig. 4-2.

2.1.1 Characteristic curve. The output of a filter plotted with respect to frequency is called its characteristic or response curve. The shape of a characteristic curve depends on the design and construction of the filter. Many of the filter characteristics may be determined from the curve in a manner similar to the way characteristics of vacuum tubes are taken from tube curves. The characteristic or response curve of a sideband filter is shown in Fig. 4-2(a). Another form of the characteristic curve known as an attenuation curve is shown in Fig. 4-2(b). Both curves are for the same filter. It can be noted that the filter having this curve is designed to eliminate the upper sideband from the output of a modulator stage having a carrier injection frequency of 100 kilocycles. This curve is for the lower sideband filter of the SSB-R3 receiver. The broad curve (7000 cycles) allows for the reception of two channels on each side of a suppressed carrier.

2.1.2 Insertion loss. It is desirable to have a filter that will pass the desired signal with no attenuation. Almost all filters have losses and the amount of attenuation a filter presents to desired frequencies is termed insertion loss, and, if the input impedance is equal to the output impedance, it is expressed in decibels as: $db_{loss} = 20 \log (E_{in}/E_{out})$.

2.1.3 Bandpass. The bandpass of a filter is equal to the bandwidth of the widest signal that the filter can pass without exceeding specified amounts of attenuation. The specified attenuation is generally 3 decibels down from the flat top portion of the characteristic curve. For example, a filter whose characteristics are represented by Fig. 4-2(a) has a bandpass of approximately 7000 cycles.

2.1.4 Shape factor. The term defining the merit of a filter (for sideband use) is called shape factor, and is the ratio of the bandpasses at the 60-decibel and 6-decibel points. The shape factor gives some indication of the steepness of the skirts while taking bandpass into consideration. As an example of shape factor, a filter 6 kilocycles wide at the 60-decibel points and 3 kilocycles wide at the 6-decibel points has a shape factor of 2.

The shape factor of the curve in Fig. 4-2 is approximately 10,500/7500 = 1.4 which is exceptionally good. The closer the shape factor is to one, the more effective the filter is in eliminating the undesired sideband.

## 2.2 Mechanical Filters

Mechanical filters can be made with rejection characteristics needed for single-sideband applications. Other features of mechanical filters (thermal stability, compactness, light weight, and ruggedness) make them well-suited for sideband use.

Two types of electro-mechanical filters are illustrated in Fig. 4-3. In part (a), a twisting or torsional motion is imparted to the resonator elements by the action of the transducer when a signal is fed to the input; in part (b), a longitudinal motion is imparted.

In each of the two types of filters, the basic principle of operation is the same. Electrical (RF) energy is supplied to the input, which is tuned to the center frequency of the pass-band. In the transducer, the electrical energy is converted to mechanical energy; and this energy passed through the resonant filter elements to the output where it is reconverted to electrical energy. The bandpass of the system depends largely on the geometry of the mechanical elements.

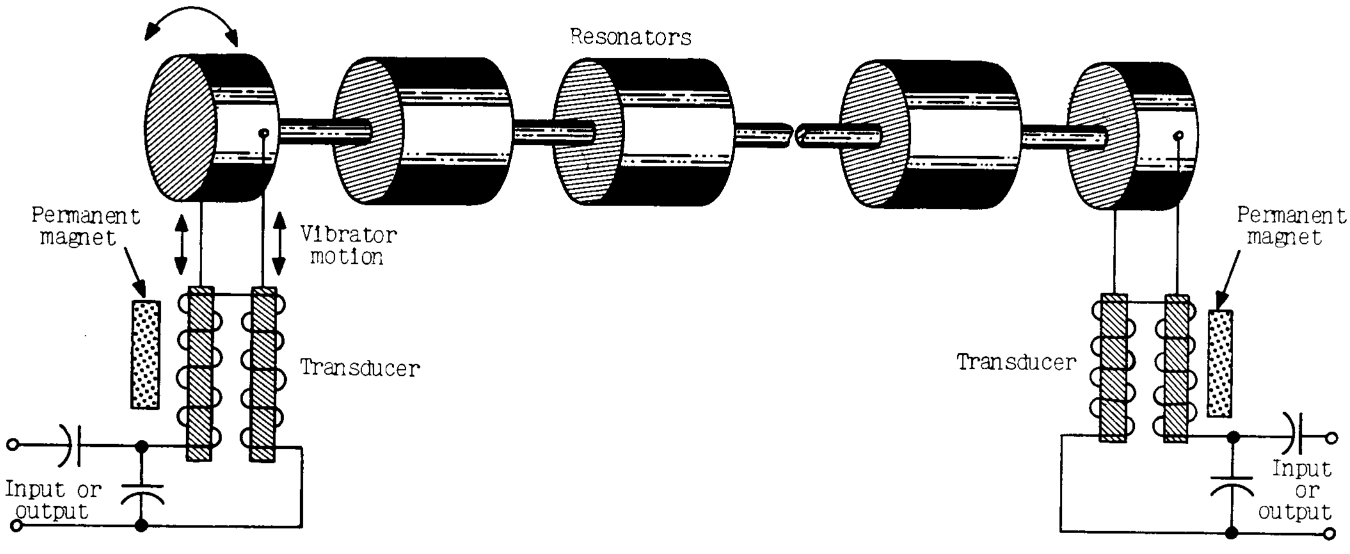
Within the coils at the input and output of the filter, are the magnetostrictive elements. These elements are made of a substance that lengthens and shortens under the influence of an alternating magnetic field. This change in length provides the motive force for setting the resonators in motion. For proper operation of the transducers, a biasing magnetic field is needed; this field is supplied by small permanent magnets.

There is, of course, a certain amount of insertion loss due to the filter. It is expressed as

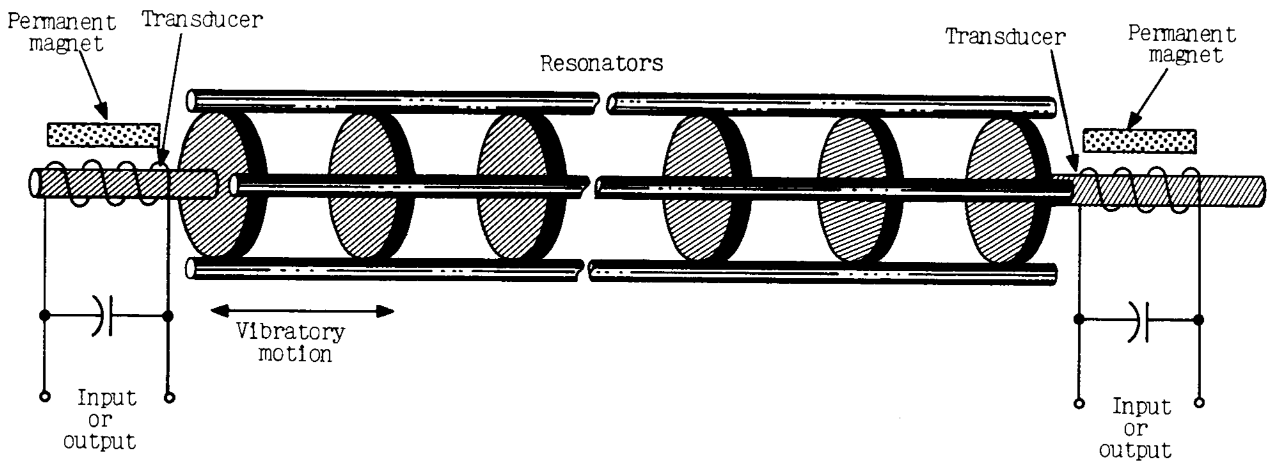
$$db_{loss} = 20 \log \frac{E_{in}}{E_{out}}$$

This loss is kept low by the use of high efficiency ferrite transducers — that is, by the use of transducers having low eddy-current losses, low mechanical losses, and high magnetostrictive properties.

The selectivity of a mechanical filter increases as the number of resonators increase;however, there is a practical limit to the number that may be used.




(a) Torsional filter




(b) Longitudinal filter

Fig. 4-3. Mechanical filters.

Mechanical filters are generally placed in the equipment circuitry between two amplifier stages. The filter couples the signal from one stage to the next and the impedances to which the filter is coupled should be high in order to avoid undesired damping. Response curves for typical mechanical filters are illustrated in Fig. 4-4.

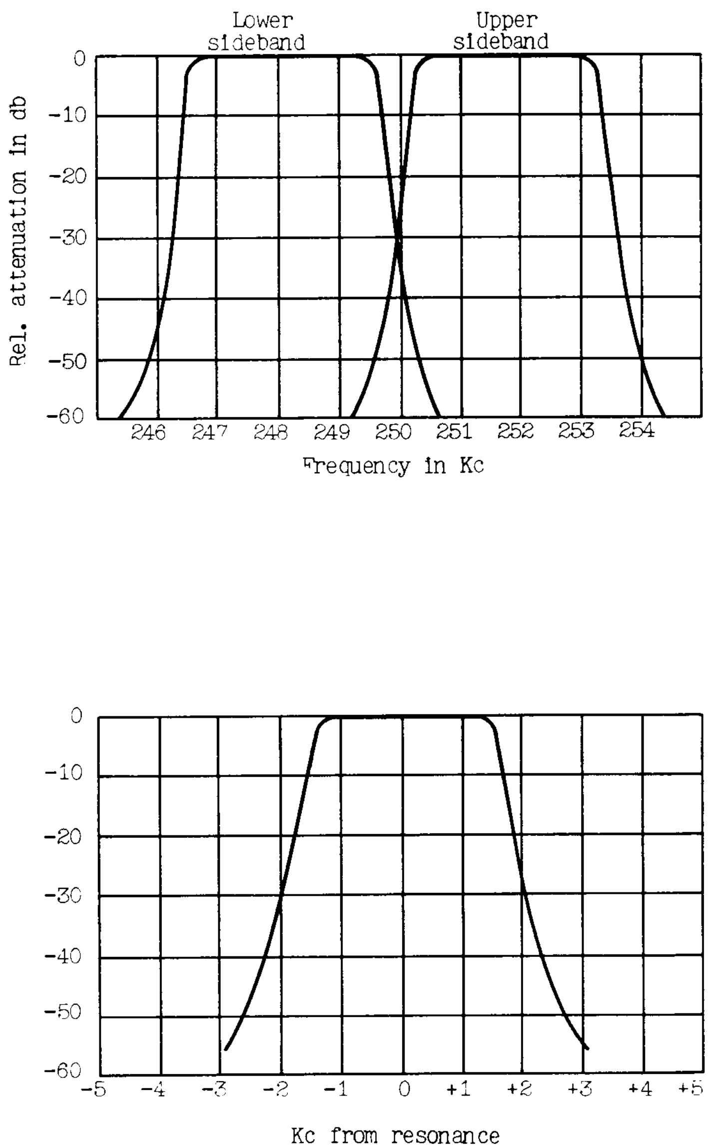



Fig. 4-4. Characteristic curves of mechanical filters.

## 2.3 Crystal Lattice Filters

Crystal lattice filters consist of combinations of piezoelectric crystals. Crystal filters achieve high attenuation outside the pass-band because of the high "Q's" of the crystals themselves. Response characteristics of crystal filters make them equally as suitable for sideband use as mechanical filters.

2.3.1 Piezoelectric crystals. An individual piezoelectric crystal, with its holder, has both a series-resonant and a parallel-resonant frequency. These two resonant frequencies are best explained by using the equivalent circuit of the crystal and its case. The crystal alone has an equivalent circuit of an inductor and capacitor in series. The capacity of the crystal holder adds an equivalent capacitor in parallel with the series combination; see Fig. 4-5. The series resonant point* occurs at the frequency at which L and C are series resonant. The parallel-resonant point occurs when L, C, and $C_H$ are parallel resonant. The series resonance is called a "zero" of impedance and the parallel resonance is called a "pole" of impedance.

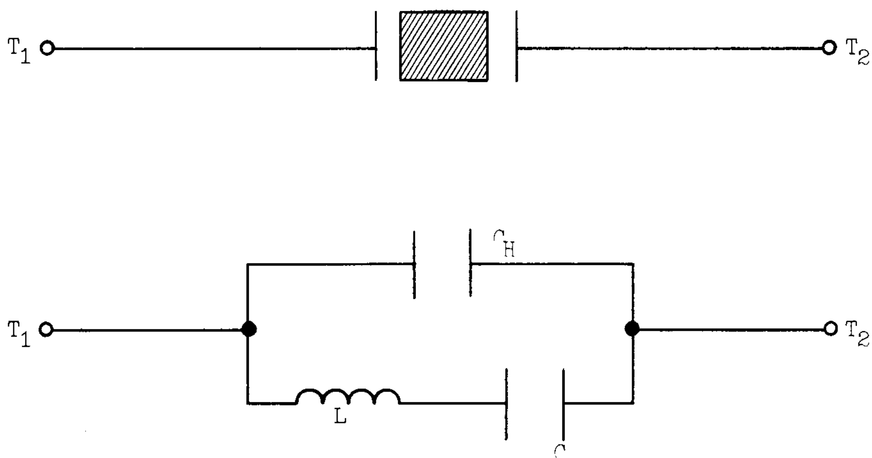



Fig. 4-5. Crystal and equivalent electrical circuit.

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* The term resonant point is used for convenience in referring to the resonant frequencies, both series and parallel, of a crystal.

The resonances appear very close together on a reactance versus frequency plot such as that given in Fig. 4-6.

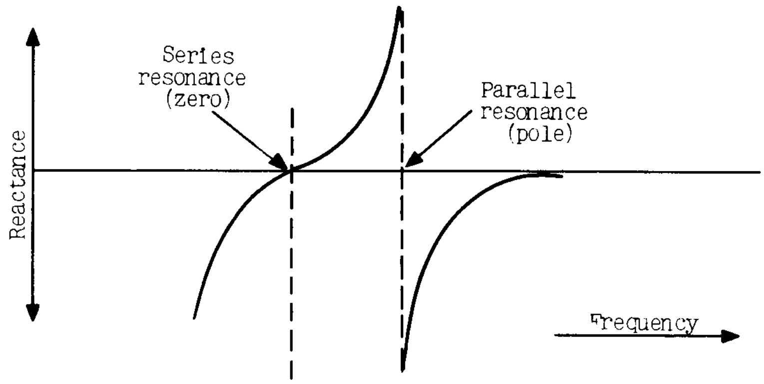



Fig. 4-6. Reactance curve of a single crystal showing zeroes and poles.

2.3.2 Analysis of a basic filter. The zeroes and poles are utilized in crystal networks to obtain a filter with desirable response characteristics. For example, a simple network (half-lattice circuit) containing two crystals is shown in Fig. 4-7.

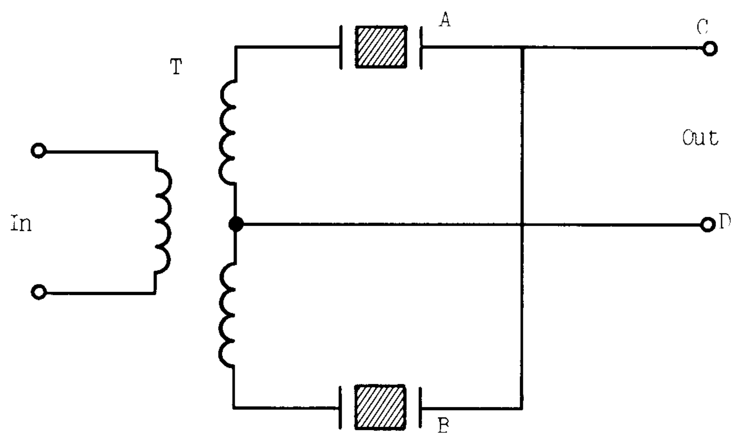



Fig. 4-7.

When the impedances of crystals A and B are equal, and an input is applied, it may be seen that the output voltage between points C and D is zero. To obtain a bandpass the crystals must be ground to different frequencies. Then at the zero of crystal A, the impedance balance of A and B is upset and a voltage appears between points C and D.

The balance is also upset at the pole of crystal A. The unbalance occurs for the zero and pole frequencies of crystal B also, with the unbalance in the opposite direction. The pass-band of the crystal filter extends from the zero of crystal A to the pole of crystal B as indicated in Fig. 4-8. The

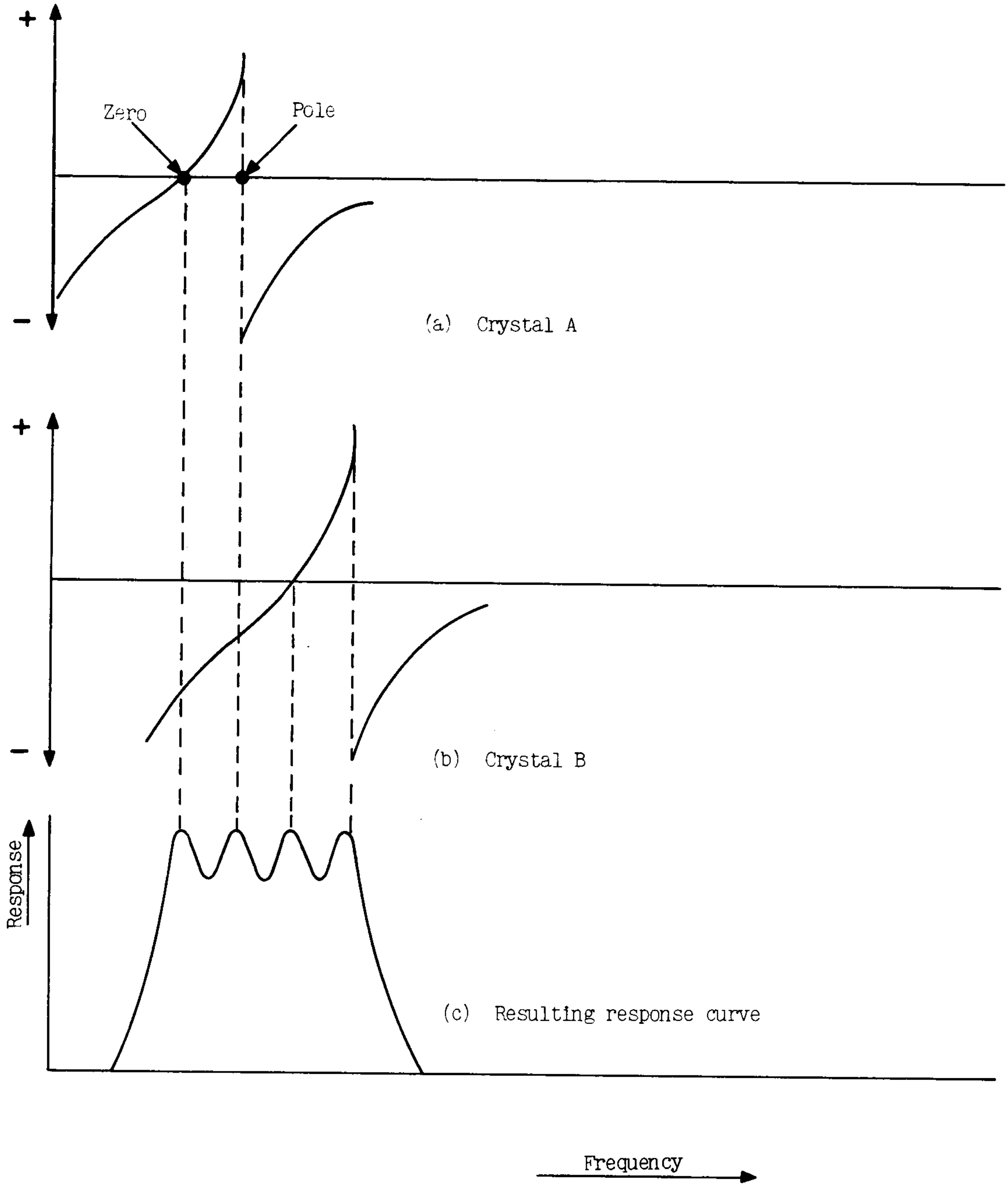



Fig. 4-8. Reactance curves of different frequency crystals and the response curve that results when used in a half-lattice circuit.

response curve of a basic filter may be uneven across the top portion, especially if the frequencies of A and B are far apart. The uneveness of the response curve can be reduced if the poles and zeros of the crystals are chosen carefully.

Figure 4-9 represents a response curve for a filter where the series-resonant frequency of crystal B is arranged to coincide with the parallel-resonant frequency of crystal A. This will give an essentially flat pass-band from the zero of crystal A to the pole of crystal B. Note that the steep sides of the response curve are due to the high Q's of the crystals and the broad top is due to the spreading effect brought about by choosing A and B to be slightly different in frequency.

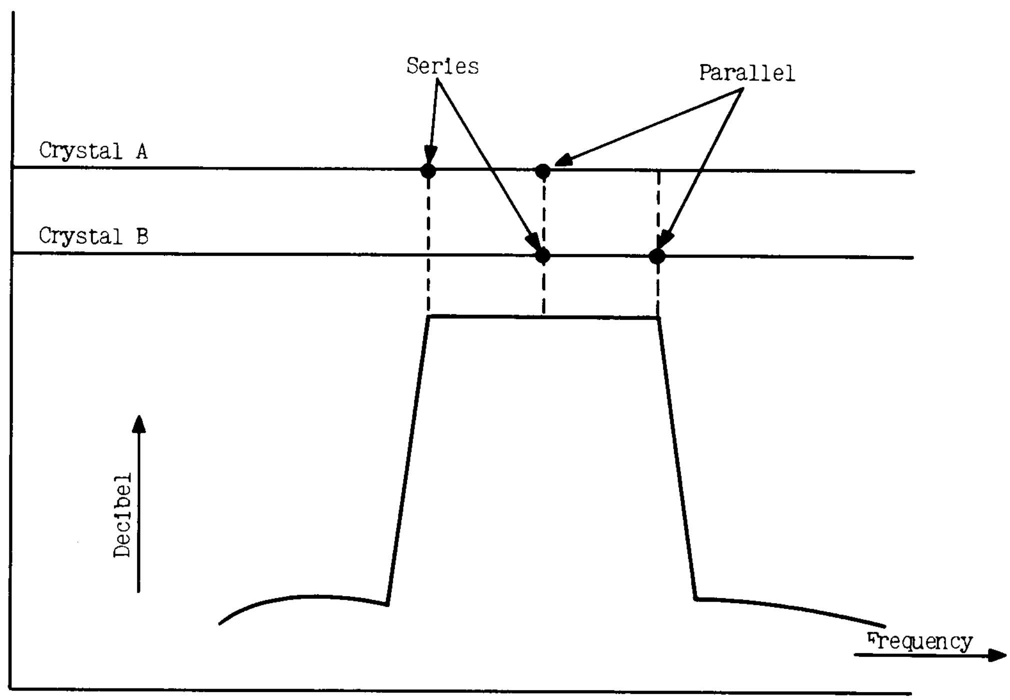



Fig. 4-9. Response curve for paired crystals.

2.3.3 External coupling. Until now we have not considered the effects of external coupling circuitry on the pole-zero spacing of a single crystal. In Fig. 4-6, it was seen that the zero and pole points occur close together. The placing of an inductor in parallel with the crystal has no effect on the series resonant point; however, the parallel resonant point is moved to a higher frequency. The added inductor also creates a new parallel-resonant frequency at a point lower in frequency than the zero point. Figure 4-10 is a reactance curve for a crystal shunted by an inductor. The input and output transformers of a filter section are usually used to shunt the crystals to obtain the spreading effect.

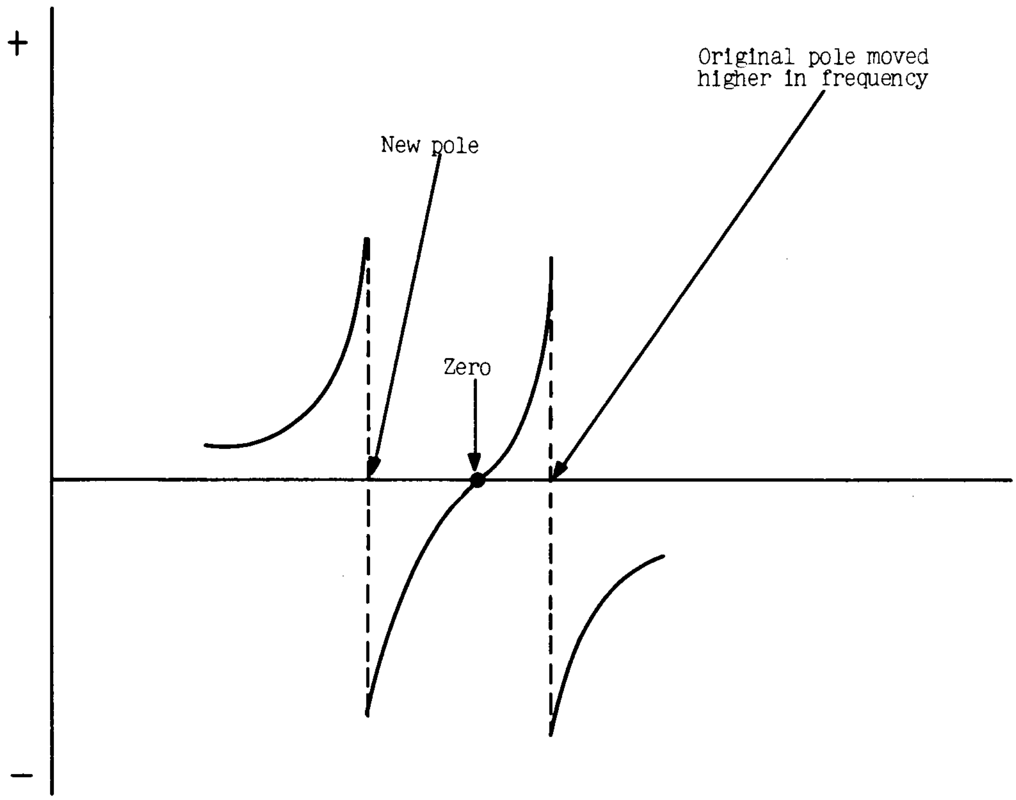



Fig. 4-10. Reactance curve for crystal shunted by an inductance.

In practice, a crystal filter similar to that shown in Fig. 4-11 is used.

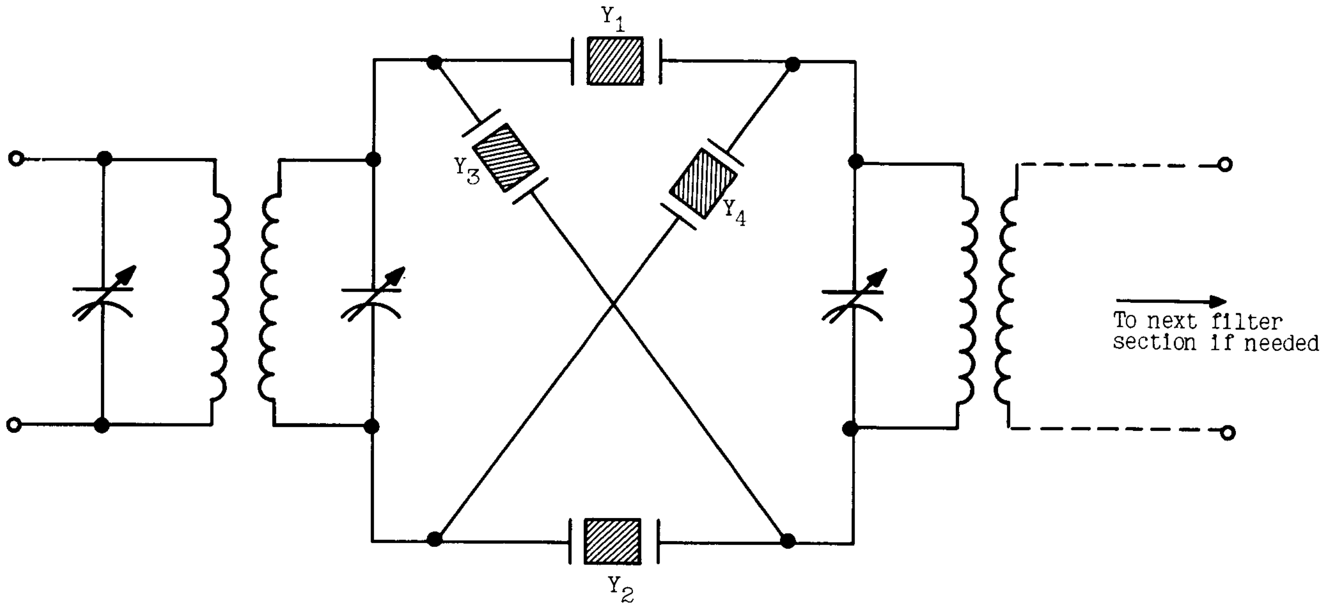



Fig. 4-11. Diagram of a useable crystal filter.

$Y_1$ and $Y_2$ are paired in frequency the same as $Y_3$ and $Y_4$. While the input and output windings are not directly across the crystals, the effect is the same as if they were. $Y_3$ and $Y_4$ should be about 2 or 3 kilocycles higher in frequency than $Y_1$ and $Y_2$. Any overspreading of the resonant frequency points can be compensated for by the trimmer capacitors in parallel with each transformer.

The input L-C network is tuned to the center of the pass-band, and the band-pass of the filter is determined by the frequency separation of the two identical pairs. This circuit is normally found in sideband applications. It is often necessary to use two of these sections in series to provide proper sideband rejection.

2.4 Heterodyning

In the filter method of SSB generation, the first modulation process takes place at a frequency that is determined by the frequency pass-band of the sideband filter. Considerations of filter design and cost make 250 kilocycles a popular choice of frequency for the modulation process.

When the sideband is generated at 250 kilocycles, and the transmitter operating range is 2 - 32 megacycles, two conversions are normally used to reach the operating frequency.

For example, if the transmitter is to operate on a suppressed carrier frequency of 18 megacycles, the 250-kilocycle sideband signal is heterodyned to the 2 - 4-megacycle range by a heterodyning process. A second heterodyning process raises the 2-4-megacycle signal to 18 megacycles.

After one sideband has been eliminated by means of a precise sideband filter, we perform two additional heterodyning processes which generate both sidebands again. However, these heterodyning processes which generate additional sidebands cause little difficulty.

It has been shown that it is necessary to eliminate the undesired sideband at frequencies near 250 kilocycles. When this 250-kilocycle sideband signal is mixed with the oscillator frequency in the heterodyning stage, the sidebands resulting from this mixing are 250 kilocycles on each side of the oscillator frequency. The sideband separation to carrier ratio is high enough in this case to permit the elimination of the undesired sideband in conventional tuned amplifiers or IF stages.

As an example, a 1000-cycle audio tone produces frequencies of 251 kilocycles and 249 kilocycles in the balanced modulator. An upper sideband filter (passes upper sideband) will reject the 249-kilocycle sideband. Here, the lower sideband which we want to reject completely is only 2 kilocycles removed from the upper sideband which we want to pass with no attenuation.

In the first heterodying process a carrier insertion frequency of 3 megacycles produces sum and difference frequencies of 3,251,000 cycles and 2,749,000 cycles. The desired sideband will be removed 502 kilocycles from the undesired sideband. This rejection can be performed easily by tuned circuits, because the sideband separation to carrier ratio is high. The sideband separation to carrier ratio is even higher after the second heterodyning stage; therefore, it will not present sideband rejection problems either.

### 2.5 An Application of the Filter Method

A block diagram, representing one channel of the SBE unit in the AN/FRT-39 is shown in Fig. 4-12. This equipment uses the filter method of sideband generation. The SBE unit receives the audio signal, transfers it to sidebands in a balanced modulator, rejects one sideband, and heterodynes the desired sideband to the operating frequency. The unit performing all of these functions is commonly called an exciter, or sideband exciter.

The SBE unit generates the sidebands at 250 kilocycles, and eliminates the undesired sideband by use of sideband filter Z110. The 250-kilocycle sideband is raised to the 2-4.25-megacycle range in the first heterodyning stage. The first heterodyning, or mixing stage, serves as the final heterodyning stage when the operating frequency is less than 4.25 megacycles. For output frequencies above 4.25 megacycles, the output of the first mixer is raised in frequency by the second mixer. The second mixer stage has injection frequencies from 8 to 34 megacycles in the 2-megacycle steps.

## 3.0 PHASING METHOD OF SINGLE-SIDEBAND GENERATION

The phasing, or phase-shift method of single-sideband generation produces without filters an output containing only the desired sideband; that is, the carrier and the undesired sideband are both eliminated without the use of filters.

We will discuss the theory of the phasing system briefly because it is

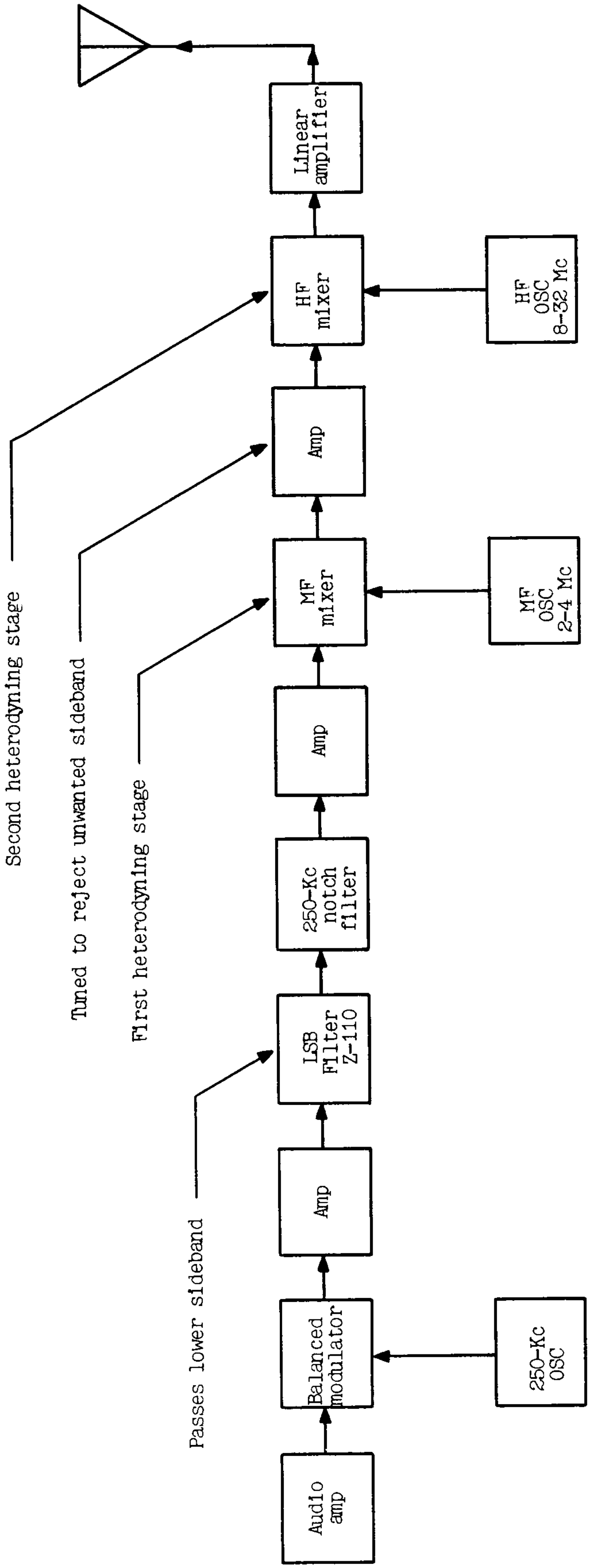



Fig. 4-12. Block diagram of one channel of AN/FRT-39 transmitter.

used in many applications of SSB communications. However, the phasing method is not as stable as the filter method, and this results in the greater use of filter systems in the FAA equipment.

### 3.1 Audio and RF Phase-Shift Networks

A block diagram of a phase-shift system is illustrated in Fig. 4-13. Two important units of the system are the audio and RF phase-shift networks. From a single input signal these networks provide an output voltage for two sections. The voltages of both sections must be equal in amplitude and separated in phase by 90°.

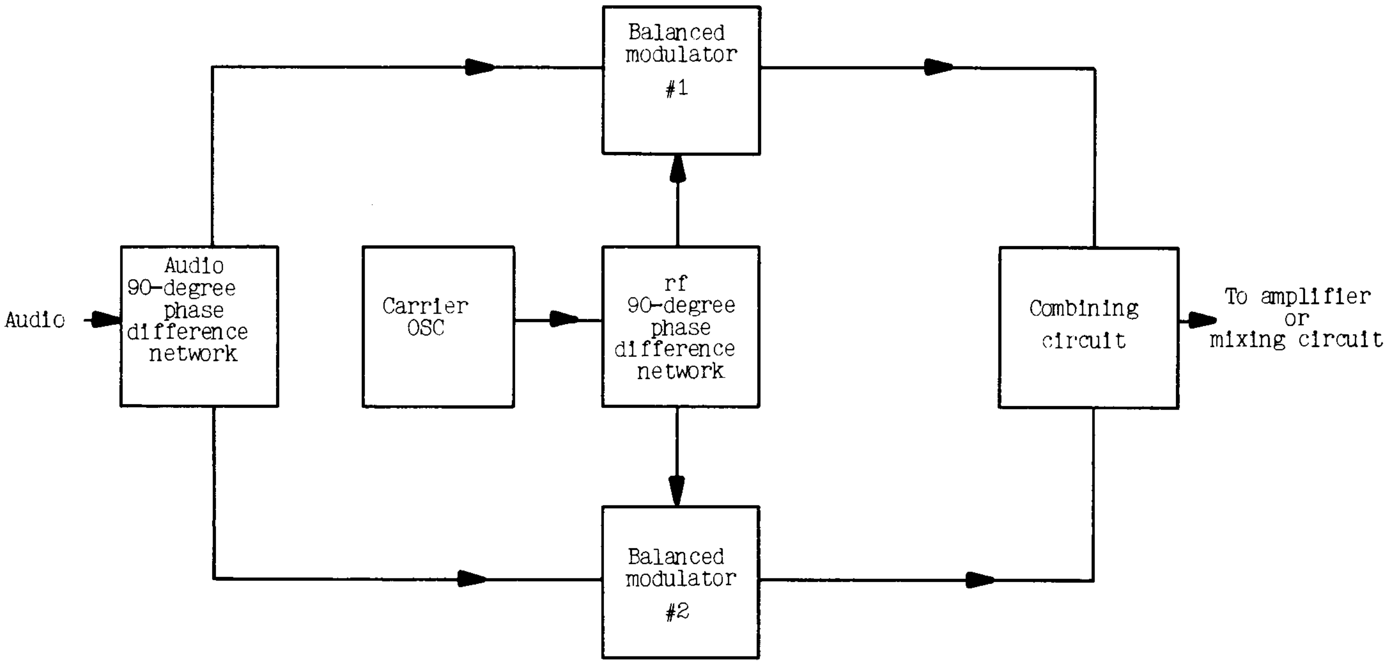
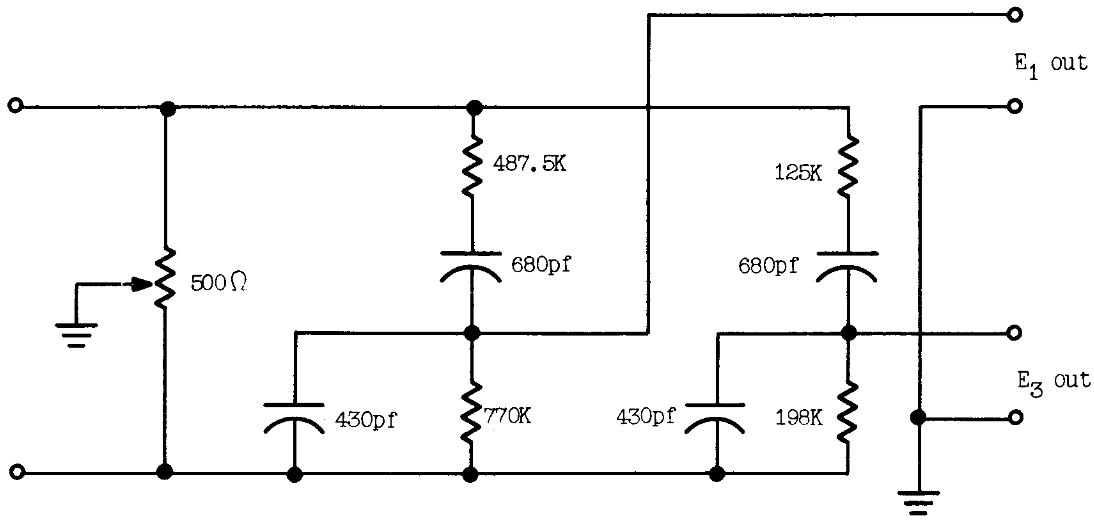



Fig. 4-13. Block diagram of the phasing method.

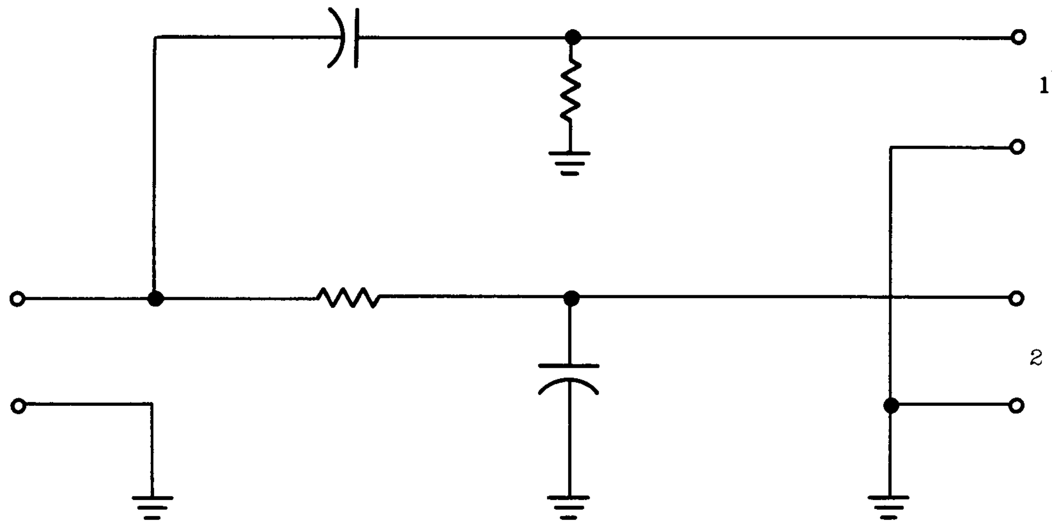
For all input frequencies within the speech channel of 300 to 3000 cps, the audio phase-shift network should provide two output voltages having the same amplitudes but differing in phase by 90 degrees. The phase and amplitude accuracy of the voltages directly affect the sideband suppression in the phasing system.

The design of audio phase-shift networks is more complex than that of rf phase-shift networks. The rf phase-shift network is designed to operate at one carrier frequency. The audio network must provide the same phase shift to all frequencies in the audio band.

Vacuum tubes are sometimes used in both networks, but the proper amplitude and phase relationships can be maintained by using resistors and capacitors only. Basic audio and rf phase-shift networks are shown in Fig. 4-14(a) and (b), respectively.




(a) Audio, 90°-phase difference, network




(b) RF, 90°-phase difference, network

Fig. 4-14.

Since the audio network is rather complex it is difficult to see that the two outputs differ in phase by 90 degrees. A mathematical analysis of the network for a 1000-cycle input frequency is given in the appendix.

In the rf network, the input is fed to two parallel branches. The No. 1 branch has its output voltage taken across a resistance; the No. 2 output is taken across a capacitor. At the operating frequency the resistance of each branch equals the reactance of that branch. The two branch currents are in phase thus making the voltage across the resistance at the No. 1 output 90 degrees out-of-phase with the voltage across the capacitance at the No. 2 output.

## 3.2 Theory of Sideband Cancellation

The trigonometry involved in the analysis of the phase-shift method may be more meaningful if the development of the equations for the amplitude modulated wave, which was covered in Chap. 2, is reviewed at this time. For convenience, Eq. (2-4) is given below. The expression for the instantaneous value of voltage (e) of an AM wave is:

$$e = \underbrace{E_{cm} \sin \omega_c t}_{\text{Carrier Component}} - \underbrace{\frac{mE_{cm}}{2} \cos (\omega_c + \omega_a)t}_{\text{USB}} + \underbrace{\frac{mE_{cm}}{2} \cos (\omega_c - \omega_a)t}_{\text{LSB}} \qquad (2\text{-}4)$$

For our analysis, $E_{cm}$ and the modulation factor, m, are disregarded because it is desired here to show that the proper phase relationships are produced to cause cancellation of the undesired sideband when proper amplitudes are assumed. In practice, however, it is as difficult to maintain the proper amplitude for adequate cancellation as it is to maintain the proper phase relationships.

As indicated in Fig. 4-15, one audio output from the phase-difference network is fed to balanced modulator No. 1; the other output from the phase-difference network is fed to balanced modulator No. 2. The output from the carrier oscillator is fed to the rf phase-difference network where two outputs (which differ in phase by 90 degrees) are fed to balanced modulators No. 1 and No. 2.

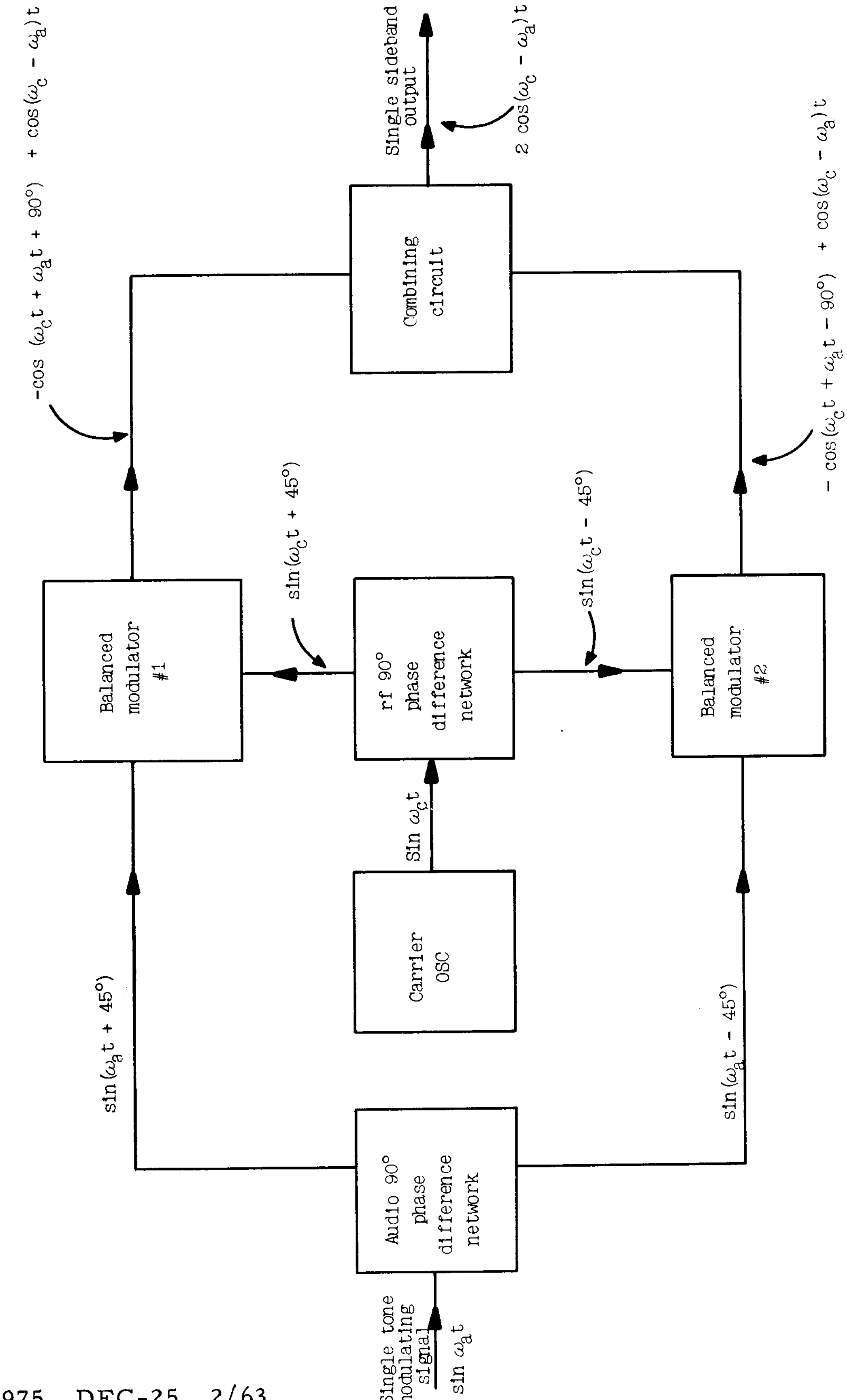



**Fig. 4-15. Phase-shift method of generating a single sideband.**

If properly adjusted and operating, the balanced modulators eliminate the carrier; however, the output of each contains upper and lower side-bands. If the signal contains many frequencies, as in speech, the upper and lower sidebands contain many frequency components. For the purpose of this discussion the audio signal is assumed to be a single frequency; each sideband then consists of only one frequency.

The audio tone entering the phase-shift network could be expressed in terms of the maximum voltage times sin $\omega_a t$, however, amplitude is disregarded and the signal going to the 90° phase-shift network is expressed as sin $\omega_a t$. The audio signals from the 90° phase-shift network is then expressed as $\sin(\omega_a t + 45°)$ and $\sin(\omega_a t - 45°)$. The output of the carrier oscillator is given as sin $\omega_c t$, and when passed through the phase-shift network will produce two signals which are $\sin(\omega_c t + 45°)$ and $\sin(\omega_c t - 45°)$. The two inputs to balanced modulator No. 1 produces an output expressed by the equation for an AM wave except that in the balanced modulator, the carrier component is balanced out. Thus, the output of modulator No. 1 takes the form of the last two terms of the AM wave equation. This form is obtained by multiplying $\sin(\omega_c t + 45°)$ and $\sin(\omega_a t + 45°)$, as shown:

From the identity,

$$\sin A \sin B = -\frac{1}{2}\cos(A + B) + \frac{1}{2}\cos(A - B).$$

By dropping the 1/2 amplitudes,

$$\begin{aligned}
&\sin(\omega_c t + 45°)\sin(\omega_a t + 45°) \\
&\quad = -\cos\left[\omega_c t + 45° + (\omega_a t + 45°)\right] + \cos\left[\omega_c t + 45° - (\omega_a t + 45°)\right] \\
&\quad = -\cos(\omega_c t + \omega_a t + 90°) + \cos(\omega_c t - \omega_a t) \\
&\quad = -\cos\left[(\omega_c + \omega_a)t + 90°\right] + \cos(\omega_c - \omega_a)t,
\end{aligned}$$

which gives the upper and lower sideband output of balanced modulator No. 1.

Treating the inputs to modulator No. 2 in a like manner,

$$\sin(\omega_c t - 45°)\,\sin(\omega_a t - 45°).$$

$$= -\cos\left[\omega_c t - 45° + (\omega_a t - 45°)\right] + \cos\left[\omega_c t - 45° - (\omega_a t - 45°)\right]$$

$$= -\cos(\omega_c t + \omega_a t - 90°) + \cos(\omega_c t - \omega_a t)$$

$$= -\cos\left[(\omega_c + \omega_a)t - 90°\right] + \cos(\omega_c - \omega_a)t$$

The two outputs show that the upper sidebands are 180° out-of-phase and will cancel in the combining circuit. The lower sidebands are in phase and will add in the combiner, resulting in a single-sideband output from the system.

This concludes the discussion of the methods used to generate a single-sideband signal. The next chapter deals with the amplification of the sideband signal in linear amplifiers before it is radiated from the antenna.

PRACTICE PROBLEMS 4-1

1. A transmitter equipped with a lower sideband filter is conveying 3000-cycle modulation. The oscillator injection frequency in the balanced modulator is 250 kilocycles. The carrier oscillator injection of the only heterodyning stage is at 4 megacycles. What is the frequency of the transmitted sideband signal?

   A. 3,747,000 cycles

   B. 247,000 cycles

   C. 3,753,000 cycles

   D. 3,750,300 cycles

2. In the previous problem the actual transmitted signal frequency was found. What is the suppressed-carrier frequency (operating frequency)?

3. A double-sideband suppressed-carrier signal contains voice modulation between 300 and 3000 cycles. If the suppressed-carrier frequency is 100 kilocycles, what is the sideband separation to carrier frequency ratio?

   A. 0.0006

   B. 0.006

   C. 0.003

   D. 0.0003

4. In the discussion of piezoelectric crystals,

   A. the series-resonant point is called a pole of impedance and the parallel-resonant point is called a zero of impedance.

   B. the crystal alone has an equivalent circuit of an inductor and capacitor in parallel.

   C. the capacity of the crystal holder adds an equivalent capacitor in series with the equivalent circuit of the crystal alone.

   D. the parallel-resonant point is called a pole of impedance and the series-resonant point is called a zero of impedance.

5. A transmitter equipped with a lower sideband filter utilizes two heterodyning stages. The oscillator injection frequency in the balanced modulator is 250 kilocycles. The injection frequencies in the first and second heterodyning stages are 4.5 megacycles and 24 megacycles, respectively. The difference frequencies of the two heterodyning stages are utilized. What is the frequency of the transmitted signal when the transmitter is modulated with 1500 cycles?

   A. 4250 Kc

   B. 4251.5 Kc

   C. 19748.5 Kc

   D. 19254.5 Kc

6. Is the transmission of the previous problem an upper or a lower sideband signal?

7. If the transmitter in prob. 5 were equipped with an upper-sideband filter, what would be the frequency of the transmitted sideband with the same 1500-cycle modulation?

ATTENTION

Examination 4 is to be worked at this time.
This exam covers the material in Chap. 4.
It is included with the material.

CHAPTER 5

# AMPLIFICATION

## 1.0 INTRODUCTION

Chapter 4 dealt with the generation of single-sideband signals with either a filter or a phasing system. The single-sideband signals are normally generated at low frequency and power levels. After their generation, it is necessary to heterodyne the sideband signals to the operating frequency and to amplify them to the desired transmitter output power level.

The sideband signals must be raised to the output power level in linear amplifiers. Linears are necessary because the sideband signals, in practical transmissions, contain several frequency components. If a sideband transmitter were to transmit a single tone only, the signal could be amplified in class C amplifiers in the same manner that the single-frequency carrier is amplified in an AM transmitter. When seveal frequencies are modulating the sideband transmitter, which is the case with voice or multiplexed teletype, the sideband signal contains several frequencies. If we attempted to amplify such a signal in class C amplifiers, harmonic and intermodulation distortion would result. This distortion is in the form of undesired frequencies in the assigned channel and frequencies in adjacent channels (splatter).
A truly linear amplifier is one that produces an output that is directly proportional to the input. Such an amplifier requires a tube which has a perfectly straight $i_p$ versus $e_g$ characteristic curve. This ideal tube has not yet been produced and therefore the nonlinear characteristic curve is one cause of distortion which we must live with.

Other causes of distortion are improper grid bias, poor regulation of the driving signal and supply voltages, and self-oscillations. To obtain an acceptable degree of linearity, amplifier stages are designed to operate over the more linear portions of the tube curves. Also, negative feedback, grid swamping resistors, neutralization, and various other means are sometimes utilized. In this chapter, the cause and effect of distortion in amplifiers and the techniques used to reduce distortion to a tolerable level are demonstrated mathematically and qualitatively.

## 2.0 CLASSIFICATION ACCORDING TO OPERATING CONDITIONS

In addition to good linearity, it is desirable that SSB power amplifiers have high power gain in order to obtain the output power level with few amplifier stages. An ordinary class A amplifier satisfies the linearity requirement but it has low efficiency and therefore its use in high-power stages is ruled out. Class C amplifiers have high efficiency but are not suited for SSB use because they are not linear. Their use in SSB transmitters would cause excessive distortion of the modulation components contained in the SSB signal. The amplifiers used for SSB are a compromise between good linearity and high efficiency. They are usually class A in the low power stages and class AB or class B in the high power stages. A description of the various classes of amplifiers follows.

The different operating conditions are best demonstrated on the grid voltage plate current transfer curve of Fig. 5-1. The figure is divided into three regions. The A region represents the linear portion of the curve. The B region is in the vicinity of plate-current cutoff and the C region lies beyond plate current cutoff. The points of operation for the different classes are labeled on the figure. This discussion will review the familiar characteristics of the different classes of operation.

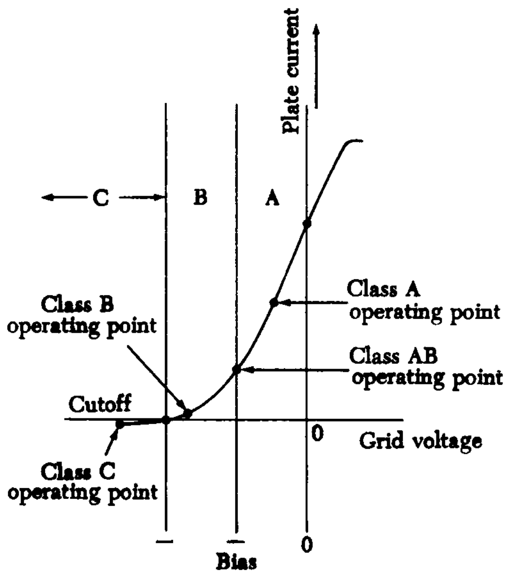



Fig. 5-1. Typical tube transfer curve showing class A, AB, B, and C operating points.

## 2.1 Class A

In class A operation the control grid-bias voltage is adjusted so that the tube operates over the most linear portion of its characteristic curve. This condition usually prevails when the bias voltage is less than half the cutoff value for the tube. In this type of operation, the curved region of the tube characteristic near cutoff is avoided and the grid voltage is so selected that the grid remains negative during the complete cycle of the input signal voltage. As a result, plate current is in the tube for 360° of the applied input cycle, and the output waveform in the plate circuit is a faithful reproduction of the input waveform. Voltage amplification in class A amplifiers is high, but the power output is relatively small because both current and voltage are restricted to small variations.

Classes of amplifiers are often compared with each other in terms of plate efficiency, which is defined as the ratio of the ac power output developed across the load to the dc power supplied the plate circuit. The efficiency of a class A amplifier is very poor, and in practice is in the range of from 2 to 20 per cent. Class A amplifiers are used whenever linearity and high voltage amplification with low power output is desired. For example, the first amplifier stages in AM and SSB receivers are class A RF amplifiers.

## 2.2 Class B

In a class B amplifier, the grid-bias voltage is adjusted at or near the cutoff value. Current is in the plate circuit for approximately one-half or 180° of the input voltage cycle. The output waveform approaches in shape and resembles the positive half of the waveform applied to the input of the tube. A tank, in the output circuit, is often used to replace the other half of the input signal.

A single tube operating as a class B amplifier has relatively high distortion, which makes it somewhat undesirable for use where fidelity of reproduction is a factor. However, two tubes are often used in push-pull operation to produce about twice the output of a single tube, and in addition contains a minimum of distortion due to the inherent cancellation of even order harmonic components. In push-pull operation, the tubes are operated 180° out of phase and when operated class B push-pull, one tube amplifies the positive half cycle of the input signal, and the other tube amplifies the negative half cycle.

The magnitude of the signal required to drive an amplifier operating class B is greater than for class A, and the voltage amplification possible is reduced when compared with class A. A single tube operating in class B delivers a greater amount of power than a single tube operating class A. This is partly due to the greater current swing that is permissible in class B operation. The plate efficiency of class B operation is approximately 60 per cent.

## 2.3 Class AB

Class AB operation will now be described in terms of class A and class B operation. The class AB amplifier operates between class A and class B. The performance of this amplifier also falls between class A and class B with the advantages of more power output than class A and less distortion than class B. Compared with class A, class AB operation produces more distortion, more power output, and has a greater plate efficiency. Compared with class B, class AB produces less distortion, less power output, and has a lower plate efficiency.

## 2.4 Class C

Although we are not concerned with class C operation in linear amplifiers, class C operation is included here for completeness. In class C operation, the grid-bias is usually adjusted for two or more times cut-off value. The input signal voltage required for class C is quite large, and plate current exists for less than 180° of the input voltage cycle. The output-plate current waveform represents a small part of the positive peaks of the input grid signal. A class C amplifier, due to the larger possible grid signal swing, produces a greater power output as compared to classes A and B. The plate efficiency for class C operation is high, being on the order of from 70 to 85 per cent. In general, class C amplifiers are used where a large power output and high plate circuit efficiency are desired. The output plate-current waveform for class C operation contains a large amount of distortion which, within limits, is overcome by the characteristics of the tuned circuits usually used in the plate circuit of the tube.

## 2.5 Subscript Numbers

A subscript number is used with the amplifier class designator to indicate whether or not the tube is operated in the positive grid region during part of the cycle. A subscript "1" attached to a particular class of operation means that no grid current exists at any time in the cycle of operation. The subscript "2" attached indicates that grid current does exist at some time during the cycle of operation. As examples, class $AB_1$ indicates that the grid never goes positive so that no grid current is drawn. Class $AB_2$ indicates that the grid does go positive so that grid current is drawn.

## 3.0 DISTORTION

### 3.1 Amplifier With Second Order Curvature

It is interesting to consider the spurious frequency components generated in amplifiers because of the nonlinear characteristic curves by utilizing the expression for plate current of a typical amplifier tube. Recall that the transfer function was used in Chap. 3 to demonstrate mixing which generated sidebands in the balanced modulator.

The expression for the plate current of a typical tube whose characteristic curve is represented by Fig. 5-1 is approximately

$$i_p = k_o + k_1 e_g + k_2 e_g^2$$

Disregarding quiescent current $k_o$, the expression for $i_p$ is

$$k_1 e_g + k_2 e_g^2$$

Then, with two signals $E_1 \sin\omega_1 t$ and $E_2 \sin\omega_2 t$ applied at the grid simultaneously,

$$e_g = E_1 \sin\omega_1 t + E_2 \sin\omega_2 t$$

and

$$i_p = k_1(E_1 \sin\omega_1 t + E_2 \sin\omega_2 t) + k_2(E_1 \sin\omega_1 t + E_2 \sin\omega_2 t)^2$$

and upon squaring the last term

$$i_p = k_1(E_1 \sin\omega_1 t + E_2 \sin\omega_2 t) + k_2(E_1^2 \sin^2\omega_1 t + 2E_2 E_1 \sin\omega_2 t \sin\omega_1 t + E_2^2 \sin^2\omega_2 t)$$

and after using the trigonometric identities

$$\sin A \sin B = -\tfrac{1}{2}\cos(A+B) + \tfrac{1}{2}\cos(A-B)$$

and

$$\sin^2 A = \tfrac{1}{2}(1-\cos 2A)$$

$$i = k_1E_1 \sin\omega_1 t + k_1E_2 \sin\omega_2 t + \frac{k_2E_1^2}{2} - \frac{k_2E_1^2}{2}\cos 2\omega_1 t - k_2E_2E_1\cos(\omega_2+\omega_1)t + k_2E_2E_1\cos(\omega_2-\omega_1)t + \frac{k_2E_2^2}{2} - \frac{k_2E_2^2}{2}\cos 2\omega_2 t$$

Here we can see that the tube plate current will contain four kinds of frequencies: (1) zero frequency (direct current) represented by $k_2E_1^2/2$; (2) the two input frequencies; (3) second harmonics of the two input frequencies; (4) sum and difference frequencies of the two input signals. These frequencies are located in the spectrum as demonstrated in Fig. 5-2.

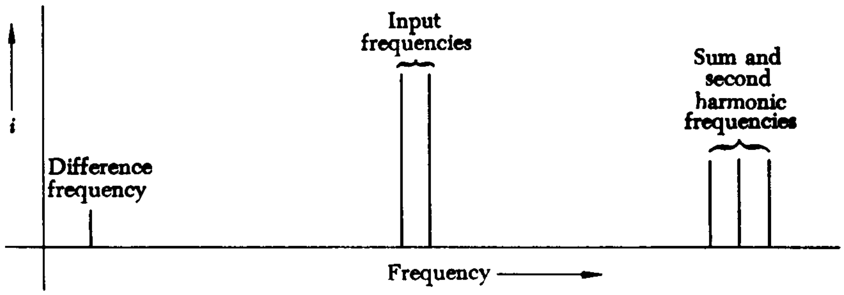



Fig. 5-2. Spectrum caused by mixing and second harmonic distortion.

Note that the undesired products of distortion are displaced from the input frequencies by several cycles. Amplifiers designed for SSB will be frequency selective (3-kilocycle to 6-kilocycle bandpass depending on number of channels) and the first, third, and fourth kinds of frequencies described above will almost completely be eliminated. So it is seen that the distortion caused by the second order curvature in tube curve is eliminated because of frequency selective circuits. This is not the case when third order curvature which exists in the more accurate expression for plate current is considered. In addition to the third order curvature term of the transfer function, the second and third harmonics are intensified when a stage is operated over nonlinear portions such as in class B or C and even some in class AB.

3.2 Amplifiers With Third Order Curvature

A more accurate expression for $i_p$ is given by

$$i_p = k_1 e_g + k_2 e_g{}^2 + k_3 e_g{}^3$$

With two input frequencies as before, the expression for $i_p$ is

$$i_p = k_1(E_1\sin\omega_1 t + E_2\sin\omega_2 t) + k_2(E_1\sin\omega_1 t + E_2\sin\omega_2 t)^2 + k_3(E_1\sin\omega_1 t + E_2\sin\omega_2 t)^3$$

The first two terms gives us the same components as in the previous example; i.e.,

$$k_1E_1\sin\omega_1 t + k_1E_2\sin\omega_2 t + \frac{k_2E_1{}^2}{2} - \frac{k_2E_1^2}{2}\cos 2\omega_1 t - E_2E_1\cos(\omega_2 + \omega_1)t$$

$$+ E_2E_1\cos(\omega_2 - \omega_1)t + \frac{k_2E_1^2}{2} - \frac{k_2E_2}{2}\cos\omega_2 t$$

In addition to these terms in the output, the last term,

$$k_3(E_1\sin\omega_1 t + E_2\sin\omega_2 t)^3 = k_3(E_1^3\sin^3\omega_1 t + 3E_1^2\sin^2\omega_1 t E_2\sin\omega_2 t$$

$$+ 3E_1\sin\omega_1 t\, E_2^2\sin^2\omega_2 t + E_2^3\sin^3\omega_2 t)$$

is present.

By the trigonometric identities,

$$\sin^3 A = \frac{3\sin A - \sin 3A}{4}$$

$$\sin^2 A\sin B = \frac{\sin B}{2} - \frac{\sin(2A + B)}{4} + \frac{\sin(2A - B)}{4}$$

it is seen that the last term caused by third order curvature will cause the following frequencies in addition to those caused by first and second order terms.

1. Same frequency as input signal.
2. Frequencies equal to three times those of the input signals (third harmonics).
3. Frequencies of the type $2\omega_1 t + \omega_2 t$, and $2\omega_2 t + \omega_1 t$ (sum of second harmonics of one input and the other original input frequency).
4. Frequencies of the type $2\omega_1 t - \omega_2 t$ and $2\omega_2 t - \omega_1 t$ (difference of second harmonics of one input and the other original input frequency).

Again, a frequency selective amplifier will eliminate many of these undesired frequency combinations; however, mixing between second harmonic components gives frequency components which fall within the bandpass of the amplifier and these cannot be eliminated in any manner. This can be demonstrated rigorously with trigonometry, but we have chosen to use a simplified example to demonstrate the frequency combinations which cause us trouble.

Consider the amplification of a USB signal in a nonlinear amplifier. The suppressed carrier frequency is 5 megacycles. The frequencies in the signal are 5001 kilocycles and 5002 kilocycles which corresponds to 1000- and 2000-cycle modulation. The input frequencies and their second and third harmonics are listed as follows:

| | | | |
|---|---|---|---|
| Input frequencies | 5001 Kc | and | 5002 Kc |
| Second harmonic of input frequencies | 10002 Kc | and | 10004 Kc |
| Third harmonic of input frequencies | 15003 Kc | and | 15006 Kc |

The mixing of harmonics and original input frequencies result in many spurious frequencies. Four possible combinations of difference frequencies are given:

10004 Kc – 5001 Kc = 5003 Kc
10002 Kc – 5002 Kc = 5000Kc
15003 Kc – 10004 Kc = 4999 Kc
15006 Kc – 10002 Kc = 5004 Kc

Sum frequencies of harmonics and input signals also result, but they are attenuated by a tuned plate circuit.

In the first example, the mixing product of 5003 kilocycles results from mixing the original 5001 kilocycles with the second harmonic of 5002 kilocycles. The 5003-kilocycle signal appears as an undesirable modulating component.

The 5000-kilocycle signal in the second example is the same frequency as the suppressed carrier, and its appearance at this point is highly undesirable.

The frequency of 4999 kilocycles appears as a lower sideband, a frequency that was so painfully removed by the sideband filter during the generation of the single-sideband signal.

5004 kilocycles appears as another undesirable modulation component in the upper sideband.

Additional harmful frequencies are generated, but the four examples of mixing products just given demonstrate the undesirable effects of nonlinear amplification when the second and third order curvature products are both considered.

## 4.0 AMPLIFIER TANK CIRCUITS

A tuned parallel capacitor and inductor in the plate circuit of an RF power amplifier is referred to as a tank circuit. The impedance of a tank circuit at resonance is high and falls off rapidly on either side of resonance. Therefore, the tank circuit provides a low impedance path from plate-to-cathode for harmonic components and other spurious frequencies which are considerably removed from the desired bandpass.

At resonance the impedance of a tank circuit is purely resistive and therefore contains no inductive or capacitive reactance. The high impedance at resonance limits dc current to a minimum, gives maximum circulating current in the tank, and the maximum output of the amplifier circuit.

During the interval that plate current flows most of the available plate voltage is dropped across the tank circuit supplying energy to the tank. This energy stored in the tank continues to supply the output circuit when the tube is cut off. This has the effect of eliminating some of the distortion which would otherwise be generated because of discontinuities in plate current when the stage goes into cutoff.

The amount of energy that can be successfully stored in a tank circuit is determined by the "Q" of the circuit and is expressed as $X_L/R$. Q may also be expressed by the ratio of the energy stored each cycle to the energy lost each cycle times $2\pi$.

$$Q = \frac{\text{energy stored each cycle}}{\text{energy lost each cycle}} \times 2\pi$$

A circuit with a very low Q will permit harmonics in the output; i.e., the bandpass will be broad, while a very high Q is objectionable because some of the higher modulating frequencies are suppressed because of the narrow bandpass. It has been found that a Q of approximately 10 provides the desired bandpass.

Thus, the tank circuit helps to eliminate or reduce distortion by not passing harmonics and mixing products, and by replacing in the plate circuit the portion of the input cycle which is clipped when the tube goes into cutoff. In this manner the tank circuit permits us to go into class AB or B operation for efficiency and still maintain an acceptable degree of linearity.

## 5.0 AMPLIFIER CIRCUITS CLASSIFIED BY EXTERNAL CIRCUIT CONNECTIONS

Linear RF power amplifier circuits are quite conventional, being either grid-driven or cathode-driven amplifiers with extremely stringent design considerations to produce maximum linearity. The tube operating point must be well chosen and accurately maintained.

Normally, class A pentode amplifiers are used in low-level power amplifiers to produce enough power to drive the higher level stages. Until recently, triodes were the only transmitting type tubes available in the medium and high power sizes. Now tetrodes are also used in high power stages. The tetrode has a screen grid between the control grid and plate which provides an additional accelerating potential to the electron stream and adds an electrostatic shield between the plate and the control grid. The electrostatic shield reduces the plate-to-grid capacitance of the tube.

Amplifiers may be classified by their external circuit connections. When classified in this manner, there are three basic types: grounded cathode, grounded grid, and grounded plate. The grounded plate is not useful in SSB power applications, and therefore is not discussed here.

### 5.1 Grounded-Cathode Triode Power Amplifier

Figure 5-3 is a schematic of a basic grounded-cathode (grid driven) triode amplifier. Triodes will always require neutralization to prevent oscillation when used in this circuit because of its large plate-to-grid interelectrode capacitance. The tubes are operated class $AB_1$ or class

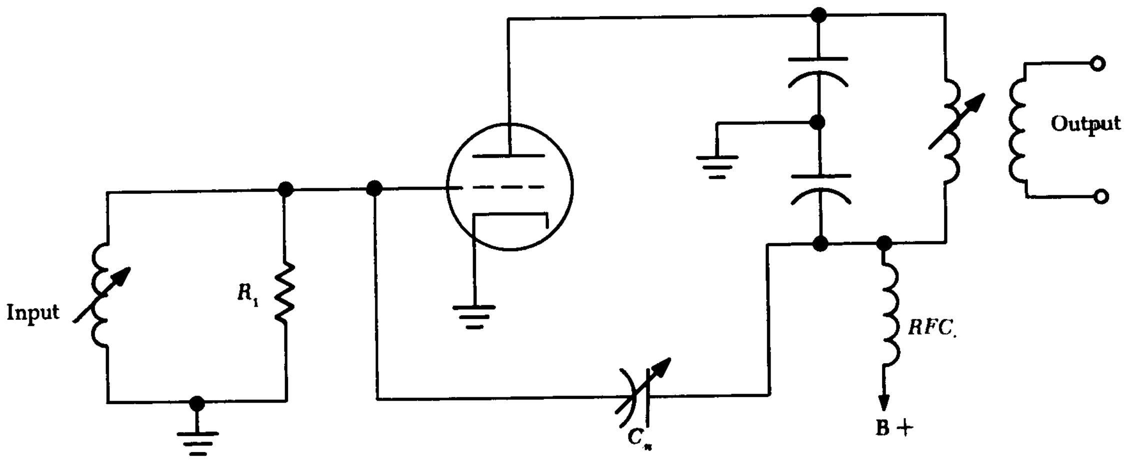



Fig. 5-3. Basic grounded cathode power amplifier circuit.

$AB_2$. The only types of triodes capable of class $AB_1$ operation have low amplification factors. Tubes suitable for class $AB_2$ operation have amplification factors up to 20.

The swamping resistor in the grid circuit aids in maintaining a constant input impedance. In $AB_2$ operation, the grid circuit appears as a varying load to the previous stage when grid current is drawn. When no grid current is drawn, the grid-to-cathode resistance is high (appears as an open); during periods of grid current it is rather low (about 500 ohms). With no swamping resistor, the grid circuit impedance varies from nearly infinity down to 500 ohms. With the resistor in the circuit, the impedance varies from 1000 ohms (with no grid current) to slightly less than 500 ohms (with grid current). This variation in impedance represents a much smaller change than before.

$C_n$ is part of the neutralizing circuit. More will be said of neutralization later.

## 5.2 Grounded-Grid Triode Power Amplifiers

Figure 5-4 is a simplified schematic of a grounded-grid triode power amplifier. This circuit is commonly known as a cathode-driven circuit. The grid is at RF ground and the input signal is fed to the cathode. In this circuit the control grid becomes an effective screen between the plate and the cathode, thus reducing the plate-to-cathode capacity.

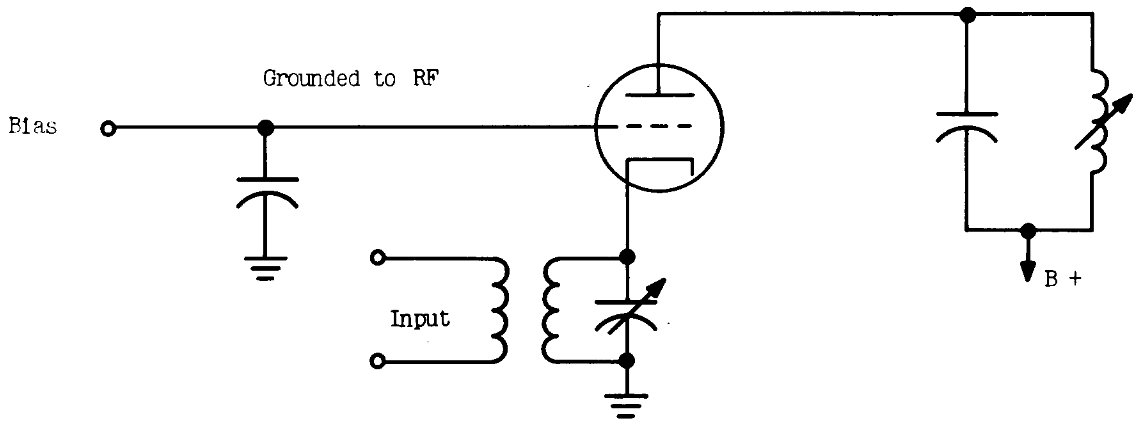



Fig. 5-4. Basic grounded grid power amplifier circuit.

The reduction in capacity decreases the amount of neutralization required or it may eliminate the need for neutralization. Therefore, triodes with high amplification factors can be used.

The relatively large driving power required by the grounded grid amplifier is recovered in its plate circuit; this results from the plate current through the cathode (input) circuit. The tube current in series with the input causes the input impedance to be low and eliminates the need for a swamping resistor.

### 5.3 Grid-Driven Tetrode Power Amplifier

Tetrodes are used in grid-driven circuits because of low grid-to-plate capacities. However, some neutralization is generally required if the tube is to operate at relatively high frequencies.

The tetrode, a high-gain tube, requires relatively little driving power and relatively small grid swing when used in amplifier circuits. This permits the paralleling of tubes and a resultant reduction in the number of stages which simplifies tuning.

## 6.0 NEUTRALIZATION

The grid-to-plate capacitance of the tube, and any external coupling between the input and output circuits may provide a path for regenerative feedback in the amplifier. When the phase and amplitude of the feedback voltage are sufficient to replace the grid circuit losses, the amplifier will break into self-oscillation. Self-oscillation in an amplifier is undesirable in that it produces distortion in the output. In order to analyze the conditions that produce amplifier instability, the input admittance of a vacuum tube must be considered.

### 6.1 Input Admittance

The input admittance of a vacuum tube is a measure of the ability of the input circuit of the tube to admit a signal and is represented by the symbol $Y_g$. The $Y_g$ of a tube is the reciprocal of the input impedance $Z_g$, or the ratio of the input current to the input voltage.

$$Y_g = \frac{I_g}{E_g}$$

Only the input admittance of a grounded cathode type amplifier will be discussed here, because this type will be most frequently encountered. The input admittance equation for a grounded cathode amplifier is given:

$$\underbrace{Y_{in} = -(\omega C_{gp} A \sin\theta)}_{\text{Conductance}} + \underbrace{j\omega \left[ C_{gk} + C_{gp} \; (1 + A\cos\theta) \right]}_{\text{Susceptance}} \qquad (5\text{-}1)$$

Where:

$\omega = 2\pi f$ = Angular velocity of input signal.

A = voltage gain of the amplifier.

$C_{gp}$ = Interelectrode capacity, grid-to-plate.

$C_{gk}$ = Interelectrode capacity, grid-to-cathode.

$\theta$ = Impedance angle of plate load.

$j = \sqrt{-1}$

The first term of Eq. 5-1 is the real portion of the input admittance, or conductance, represented by the symbol G. The second term, preceded by the j operator, is capacitive susceptance, represented by the symbol $B_c$. To determine the input capacity of the amplifier, the second term of the equation can be divided by $j\omega$ to obtain

$$C_{in} = C_{gk} + C_{gp} \, (1 + A \operatorname{Cos} \theta) \qquad (5\text{-}2)$$

It can be seen from Eq. 5-2 that the grid-to-plate capacity is increased by a factor A Cos $\theta$, thus making the effective input capacity of the tube greater than the interelectrode capacities. This phenomenon is called the "Miller Effect".

An inspection of Eq. 5-1 shows that the factors that affect input admittance are: frequency of the input signal, voltage gain of the amplifier, the interelectrode capacities, and the impedance angle, $\theta$, of the plate load. For purposes of simplicity, the input frequency and the voltage gain of the amplifier will be considered as constants. The interelectrode capacities of a grounded-cathode amplifier are represented in Fig. 5-5.

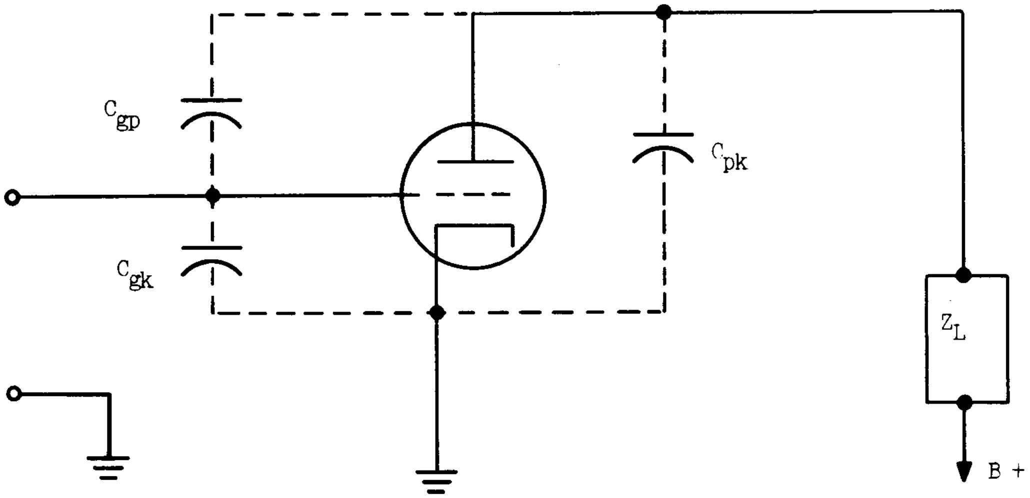



**Fig. 5-5. Grounded cathode amplifier showing interelectrode capacities.**

The condition necessary for an amplifier to oscillate is that the input admittance of the amplifier has a conductance component that is negative. Thus, when the real portion of Eq. 5-1 is negative, the amplifier tends to oscillate. The only term in the real portion of Eq. 5-1 that can be either positive or negative is $\sin\theta$. The angle $\theta$ is determined by the plate load impedance, $Z_L$. When $Z_L$ is inductive, $\theta$ will be positive or between $0^\circ$ and $90^\circ$. The sine of an angle in the first quadrant is positive, therefore, the first term of Eq. 5-1 will be negative. A negative conductance component shows that energy is being transferred from the output circuit to the input circuit through $C_{gp}$. If this feedback exceeds the input circuit losses, the amplifier will oscillate.

When $Z_L$ is capacitive, $\theta$ is negative, or between $0^\circ$ and $-90^\circ$. The sine of an angle in the fourth quadrant is negative, which will make the first term of Eq. 5-1 positive. This indicates a positive conductance component of the input admittance and that the feedback through $C_{gp}$ is degenerative. Under these conditions the amplifier will be stable.

When the load impedance is a pure resistance, the angle $\theta$ is zero. The first term of Eq. 5-1 then goes to zero and the input admittance is composed only of capacitive susceptance. Because no energy is fed back to the input circuit, the amplifier is stable with a resistive load. The following summary chart shows input circuit conditions with different types of plate load impedance.

| Plate Load | Input Circuit |
|---|---|
| Resistive | Capacitive |
| Inductive | Negative resistance and capacitance |
| Capacitive | Resistance and capacitance |

Amplifiers utilizing triodes and operating at radio frequencies must have special circuitry to counteract the tendency for oscillation when the load impedance is inductive. Such counteraction of the tendency to oscillate is called "neutralization". Neutralization is the process of preventing the feedback of voltage through the grid-to-plate capacitance, or as feeding back into the input circuit, a voltage equal in amplitude to, and $180^\circ$ out-of-phase with the voltage fed back through the grid-to-plate capacitance.

## 6.2 Neutralizing Circuits

Several circuit variations for neutralizing an amplifier are in use. Any circuit may be used provided it fulfills the requirement that the feedback voltage through $C_{gp}$ is eliminated, or that the feedback voltage through the neutralizing circuit be equal in amplitude and 180° out of phase with the feedback voltage through $C_{gp}$. Three types of neutralizing circuits are discussed in the following.

6.2.1 Grid neutralization. The circuit for "grid" neutralization is shown in Fig. 5-6. The neutralization is taken directly from the plate of the amplifier and coupled through the neutralizing capacitor $C_n$ to the bottom of the grid tank circuit. The grid tank provides the 180° phase shift necessary for cancellation of the feedback through the grid to plate capacity $C_{gp}$. Note that for this type of neutralizing circuit, the bottom of the grid tank cannot be at RF ground potential.

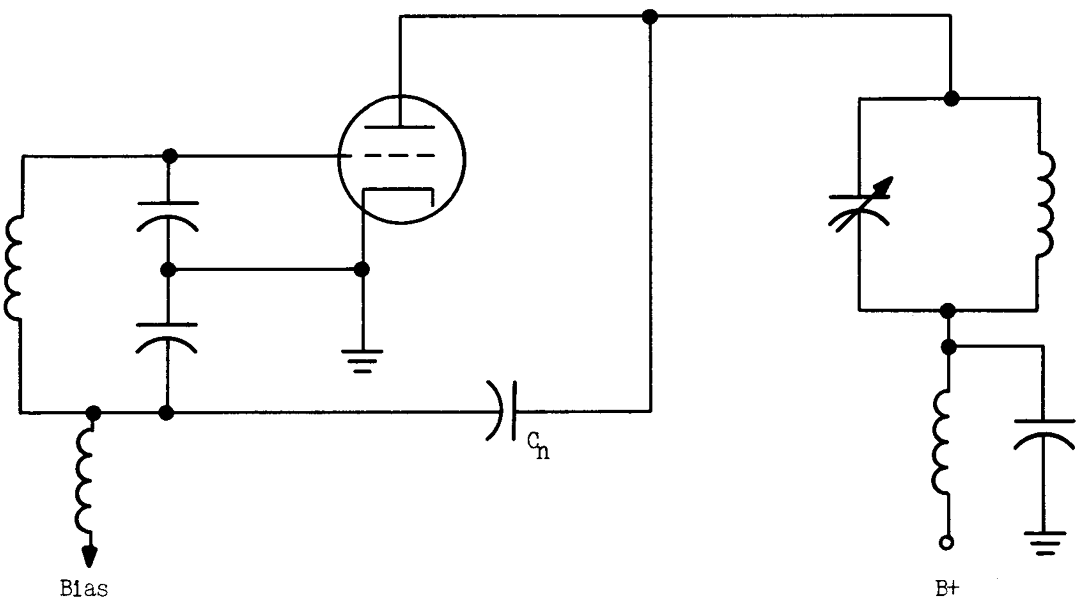



Fig. 5-6. Grid neutralizing circuit.

6.2.2 <u>Plate neutralization.</u> The circuit for "plate" neutralization is is shown in Fig. 5-7. In this type of circuit, the neutralizing feedback is taken from the bottom of the plate tank and coupled through the neutralizing capacitor $C_n$, directly to the grid. The plate tank provides the necessary 180° phase shift. In this circuit, the bottom of the plate tank cannot be at RF ground potential

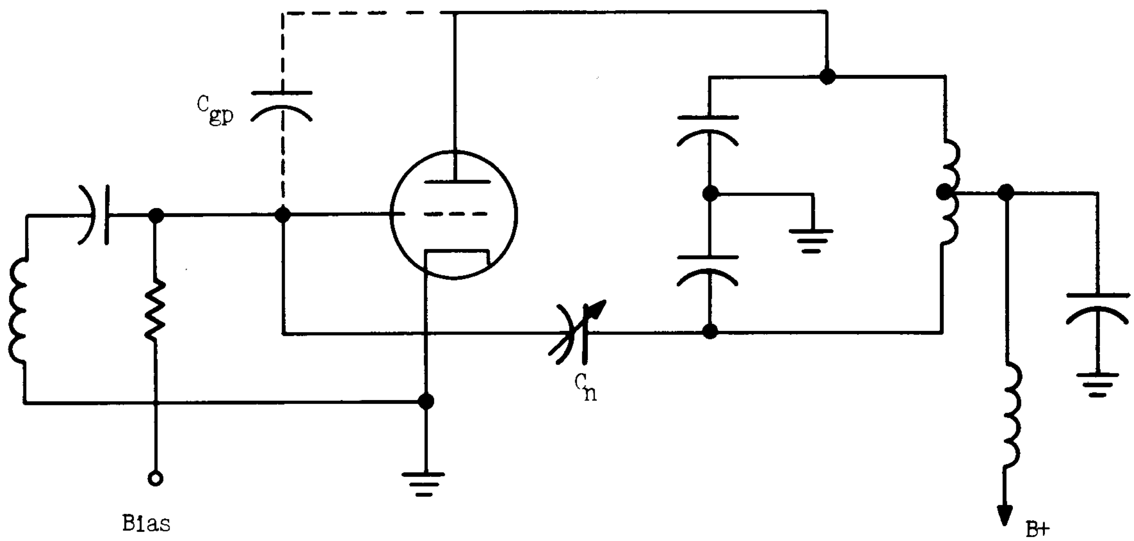



**Fig. 5-7. RF amplifier utilizing plate neutralization.**

6.2.3 <u>Push-pull neutralization.</u> For push-pull neutralization, the neutralizing feedback is obtained by connecting the plate of each tube through the neutralizing capacitor $C_n$, to the grid of the opposite tube. Cancellation of feedback through the grid-to-plate capacity $C_{gp}$ results, due to the fact that the plates of a push-pull stage are 180° out of phase. The circuit for push-pull neutralization is shown in Fig. 5-8.

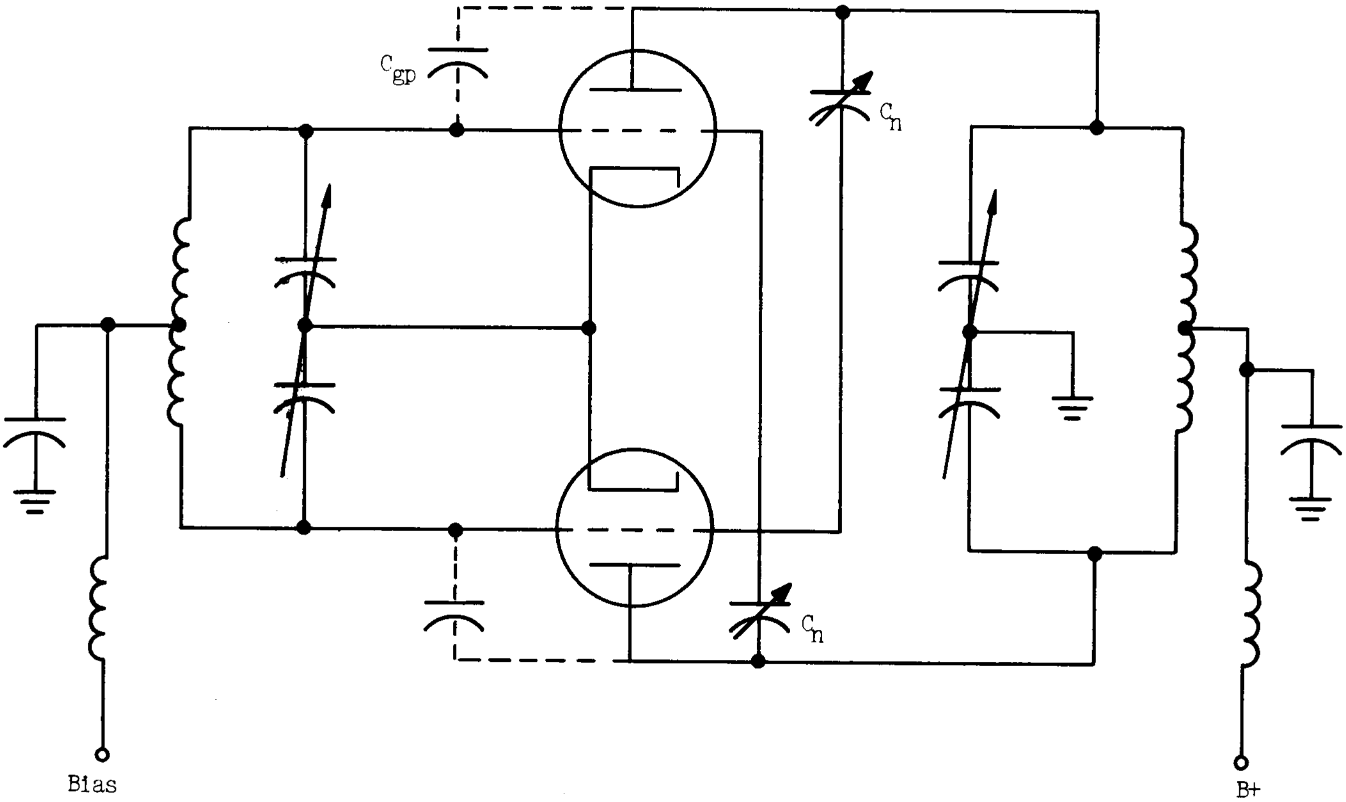



Fig. 5-8. Push-pull neutralizing circuit.

ATTENTION

Examination 5 is to be worked at this time.
This exam covers the material in Chap. 5.
It is included with the material.

CHAPTER 6

# RECEIVING SINGLE SIDEBAND

## 1.0 GENERAL CONSIDERATIONS

### 1.1 Basic

Most of the previous chapters have discussed single-sideband(SSB) transmission theory. This chapter explains the very interesting concepts of SSB reception. Much of the theory of SSB transmission covered thus far is also applicable to SSB reception. For example, in the explanation of amplitude modulation it was shown that the composite AM wave consists of separate and distinct frequency components. In the discussion of SSB generation it was shown how all the frequency components (except the desired sideband) are eliminated. In this chapter we will show how the carrier is reinserted to heterodyne with the received sideband signal to enable the demodulation of the audio intelligence.

Many of the factors influencing SSB receiver design also influence the design of AM receivers. Although we have taken much of the AM communications theory for granted thus far, this chapter discusses design considerations that are common to both AM and SSB communication receivers. AM and SSB communication receivers are invariably of the double-conversion superheterodyne type which provides high image rejection, sensitivity, and selectivity. The reason double-conversion receivers provide these three desirable features is given under the section on theoretical circuit considerations.

The reception of an SSB signal is essentially the reverse of transmission. The RF SSB signal is picked up at the antenna, amplified in Class A amplifiers, converted down to one or more intermediate frequencies, and reduced from IF to audio frequencies in the demodulator. Except for demodulating the audio, SSB reception is very similar to AM reception.

There are two common types of SSB transmissions. In one, the carrier is suppressed completely; this type is called single-sideband suppressed carrier (SSBSC). * In the other type the carrier is transmitted at reduced level, down 20 or 30 db; thus, the term single-sideband reduced carrier (SSBRC) is applied. Note that it is not common practice to transmit a full-level carrier in SSB systems.

The absence of a normal carrier in the received SSB signal accounts for the principal difference between single-sideband and AM receivers. To recover the audio intelligence from an SSB signal when the carrier is completely suppressed at the transmitter, a reference carrier must be reinserted at the receiver. The locally generated carrier is normally inserted at the receiver demodulator. However, it can be inserted a stage or two preceding the demodulator under certain conditions. The reinserted carrier must have nearly the same frequency relationship with the sideband signal as the previously suppressed carrier. Therefore, the oscillator in the transmitter and the one that produces the reinserted carrier in the receiver must have extremely good frequency stability.

There are two methods of utilizing the low-level carrier of an SSBRC transmission to aid in its reception. The low level carrier can be separated from the sidebands, amplified, and fed directly to the demodulator to aid in demodulation. In the other method, the low-level carrier is utilized in an automatic frequency control circuit to control an oscillator injection frequency.

The SSB-R3 receiver discussed in the next chapter is capable of receiving SSBSC and SSBRC signals. For demodulating SSBSC signals it has a highly stable carrier oscillator. For demodulating SSBRC signals by the first method mentioned above, the receiver has a carrier conditioner for separating the carrier from the modulation and amplifying it to the proper level for the demodulator circuit. For the second method of reception, it has an automatic frequency control circuit which controls the frequency of the second

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*When the carrier is below the sideband signal by 50 or 60 db it is considered to be suppressed completely because the level of the carrier is not sufficient to aid in signal reception.

oscillator. In this manner the second IF signals always have the proper frequency relationship with the carrier oscillator signal at the demodulator.

The reception of a CW signal is similar to the reception of an SSBSC signal with single-tone modulation. For this reason, in explaining SSBSC reception, we shall compare it with the reception of CW.

## 1.2 Noise Considerations

In the reception of all types of RF signals, amplified noise competes with desired signals in the output of receivers. Noise can be divided into two classes: noise from external sources and noise generated within the receiver.

Noise from external sources is composed of undesired signals from adjacent-channel systems, atmospherics, automobile ignitions, and other electrical interference. Adjacent-channel interference can be reduced by proper channel separation and proper receiver design. Automobile ignition and other electrical interference can be reduced by proper location of the receiving site. The degrading effects of all external noise voltages are proportional to the bandwidth of the receiver being considered.

Thermal agitation noise generated within the receiver is due to the random motion of electrons. This noise is present at the terminals of any conductor. Being random in nature, thermal agitation noise voltages occupy an infinite frequency band. Noise voltages in a given conductor are proportional to the bandwidth, the resistance of the conductor, and its absolute temperature. In equation form, thermal noise voltage is expressed as:

$$E_n = \sqrt{4\,K\,T\,\Delta f\,R}$$

where $E_n$ = rms noise voltage

$K$ = Boltzmann's Constant = $1.38 \times 10^{-23}$ joules/°K

$T$ = absolute temperature

$\Delta f$ = bandwidth in cps

$R$ = resistance in ohms

Because power is equal to the voltage squared divided by the resistance, noise power is stated as:

$$P_n = 4 K T \Delta f$$

Therefore, the degrading effects of internal noise is directly proportional to the bandwidth of the receiver being considered.

Because the bandwidth required for SSB reception is one-half that required for AM, one-half as much noise power appears in the output of SSB receivers. It was explained in Chap. 2 that a double-sideband suppressed-carrier system had a 3-db advantage over an SSB system if the receiver has a bandwidth sufficiently wide to receive both sidebands. However, when a receiver is specifically designed for SSB, proper filters will narrow the bandwidth and give SSB the same system gain that is available from double-sideband suppressed-carrier systems.

## 1.3 Fading Considerations

Long-range high-frequency communication is carried out by means of sky waves which are reflected or refracted by the ionosphere. Fading occurs when the strength of sky waves varies because of fluctuations in the reflecting and refracting properties of the ionosphere. This fading is dependent on frequency, being more rapid at higher frequencies.

Because fading depends on frequency, the different frequency components of a transmitted signal are affected differently. Distortion in a signal, when various frequencies are affected differently, is called selective fading.

Proper reception and detection of an AM signal requires that the phase and amplitude relationships between carrier and sidebands remain undisturbed. Selective fading changes these relationships and causes incoherent sideband addition; this, in turn, causes distortion and less than full utilization of transmitted sideband power. If selective fading causes the carrier to be partially or completely canceled, harmonic and intermodulation distortion results upon detection.

The loss of one of the two sidebands of an AM transmission results only in a reduction of signal voltage out of the demodulator. This is not entirely detrimental to the signal because the other sideband contains the desired intelligence. However, the bandwidth of the AM receiver is sufficient to receive both sidebands and the noise level remains constant. This results in a 3-db deterioration in the signal-to-noise ratio of the received signal. The complete loss of one sideband would probably be unusual. Even so, the signal-to-noise ratio available from a receiver suffers from any reduction in the level of either one or both sidebands.

Selective fading does not impair SSBSC and SSBRC systems as much as it does AM. The phase relationships of the different frequencies in an SSB signal are not necessary for detection even though some distortion results when these relationships are disturbed. Selective fading does cause amplitude-versus-frequency distortion in SSB signals which is responsible for the strange sounds that characterize overseas news broadcasts. Intelligibility studies have been made of comparable AM and SSB systems that have the same radiated sideband power. These studies indicated that SSB provides a signal that is from 0 to 9 db above that of an AM system under various conditions of propagation, and during severe selective fading SSB reception was established where AM was completely out of service.

It is seen that selective fading (which is very detrimental to AM) is not a serious problem with SSB. Periods exist, however, when the entire SSB transmission fades completely or is reduced in amplitude for short intervals. It is shown in the next chapter how the effect of this short-interval fading is reduced.

## 2.0 BLOCK DIAGRAM ANALYSIS

### 2.1 Block Diagram Analysis of an AM Receiver

The block diagram of an SSB receiver can seem complicated to a beginner in SSB. For this reason we shall discuss the familiar AM receiver and then proceed to the CW and SSB receivers.

Figure 6-1 is a block diagram representing a double-conversion communications receiver similar to those used by the FAA. For the sake of discussion, assume that the AM wave being received has a carrier frequency of 5 mc and that the wave contains 1000-cycle modulation.

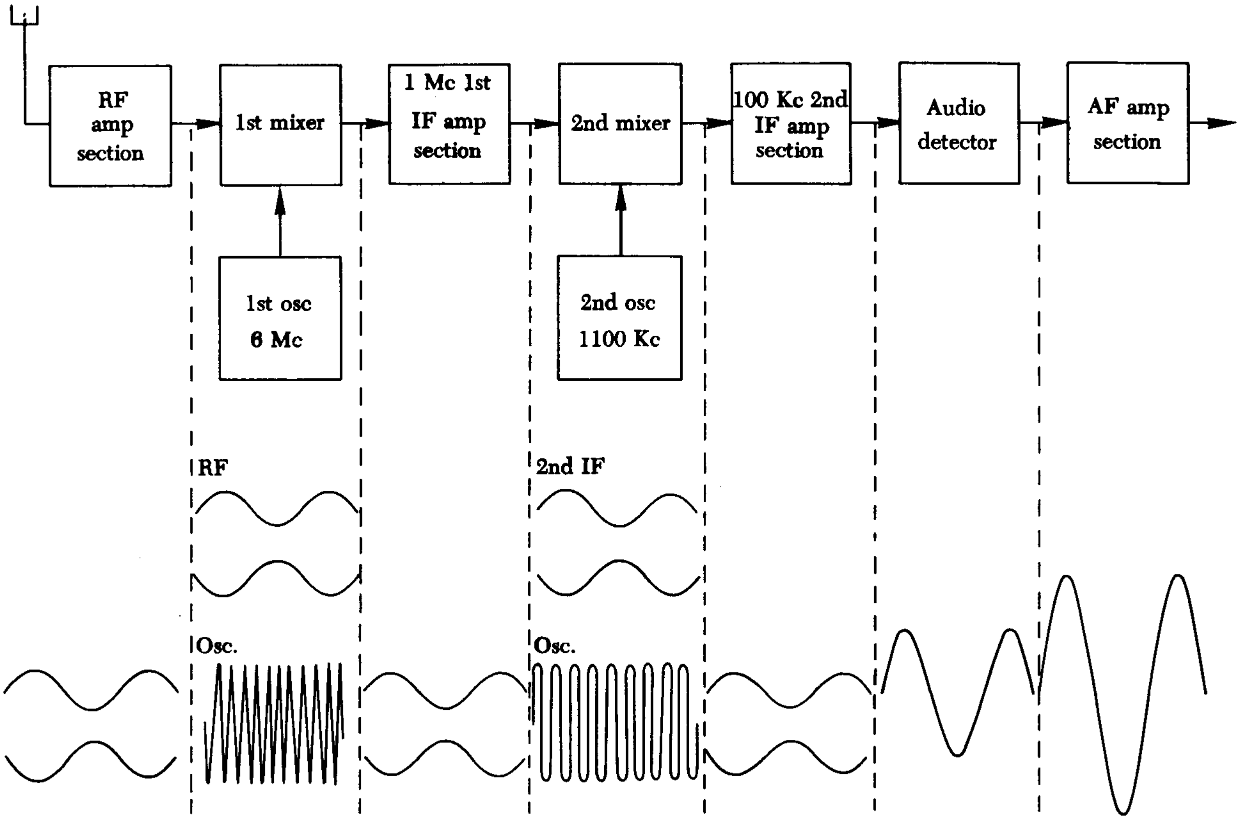


**Figure 6-1. Block diagram of an AM receiver.**

The received signal is amplified in the RF amplifier section. The output of the RF amplifier section is combined with the output of the first oscillator to produce the first intermediate frequency. The waveforms on the block diagram show the signal at various locations in the receiver.

The envelope of the IF signal is the same as that of the received RF signal. To explain the waveform of the IF signal, assume the first intermediate frequency is 1 mc. The oscillator frequency will be 1 mc above the carrier input frequency (5 mc + 1 mc = 6 mc). Recall that the received AM wave contains 1000-cycle modulation; therefore, the three received signals are 4999 kc, 5000 kc, and 5001 kc. These three frequencies form sum and difference frequencies with the 6-mc oscillator signal to form 6 new frequencies.* The difference frequencies are:

6000 kc - 4999 kc = 1001 kc
6000 kc - 5000 kc = 1000 kc (frequency of carrier in 1st IF)
6000 kc - 5001 kc = 999 kc

The sum frequencies are:

6000 kc + 4999 kc = 10999 kc
6000 kc + 5000 kc = 11000 kc
6000 kc + 5001 kc = 11001 kc

The sum frequencies are rejected by the IF section; the difference frequencies (amplified by the first IF) comprise the carrier, upper, and lower sideband of a new AM wave which has the same envelope as the received AM wave.

The first IF signal mixes with the 2nd oscillator signal to again produce sum and difference frequencies. For a second IF of 100 kc, the second oscillator frequency is 1100 kc and the difference frequencies from the mixing are:

1100 kc - 1001 kc = 99 kc
1100 kc - 1000 kc = 100 kc (frequency of carrier in 2nd IF)
1100 kc - 999 kc = 101 kc

These three frequencies form a new AM wave which has the same envelope as the received RF and first IF. Sum frequencies are generated in the second mixer, and again they are rejected.

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*Only 6 new frequencies, if the mixing of the harmonics due to nonlinearities are neglected.

The second IF signal is fed to a detector where the sidebands are mixed with the carrier to recover the difference frequency as audio. The 1-kc audio is obtained as follows:

100 kc - 99 kc = 1 kc
101 kc - 100 kc = 1 kc

The recovered 1-kc audio is amplified in the audio amplifier section, and finally a 1-kc tone is heard at the speaker. Note that we get a 1-kc signal from each sideband. These two 1-kc signals combine coherently as explained in the power comparison of Chap. 2.

This detailed discussion of the waveforms and frequencies in the different sections of the AM receiver has been included to help with the explanation of SSB reception. Note that the upper sideband of the received signal is converted to a lower sideband in the first mixer and converted back to an upper sideband in the second mixer. The lower sideband of the received signal is likewise inverted twice in the two mixers. The frequency of the carrier in the 1-mc IF was 1 mc, and in the 100-kc IF the carrier was 100 kc. The sidebands were displaced from the carrier frequency in the two IF sections by an amount equal to the modulation frequencies that the AM wave contained. It will be seen that the sideband modulation in the IF's of SSB receivers is displaced from the suppressed carrier frequency by an amount equal to the modulation frequency.

## 2.2 Block Diagram of a CW Receiver

It was stated earlier that the reception of a CW signal is similar to that of an SSBSC signal having single-tone modulation. This statement will be clarified at this time.

In CW transmission, a single RF frequency is radiated. The frequency of the transmitted signal is the same as the assigned operating frequency. This radiated carrier signal is made to convey intelligence by turning it on and off. The intelligence is recovered at the receiver in the form of tones of different duration. To simplify discussion we will assume that the key is held down at the CW transmitter. Thus the only intelligence being conveyed is that of a steady tone.

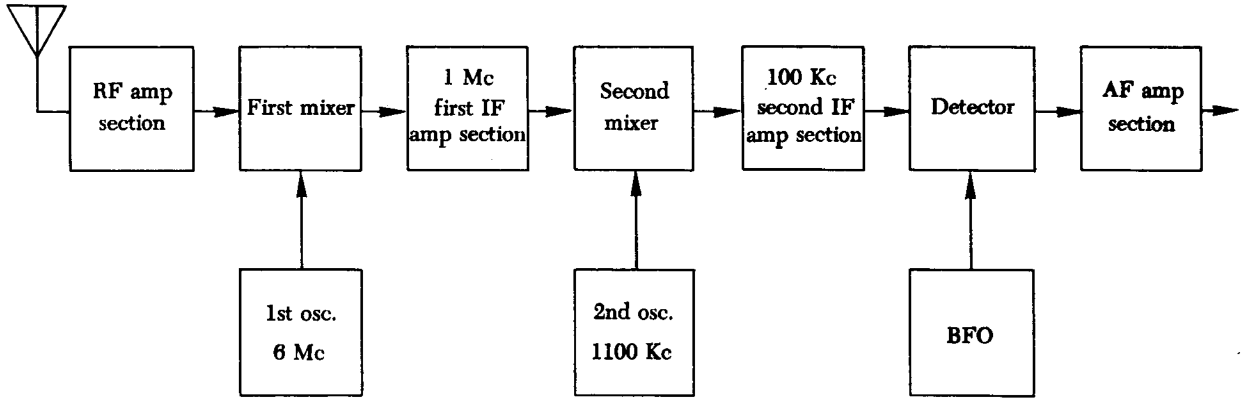


FIGURE 6-2. Block diagram of a double-conversion CW receiver.

The reception of the radiated CW signal is explained by using the block diagram of the double-conversion CW receiver shown in Fig. 6-2. The operating frequency and the first and second oscillator frequencies are considered to be the same as they were in the previous discussion of the AM receiver. The 5-mc steady carrier at the antenna is amplified in the rf amplifier section and fed to the first mixer. The 6-mc oscillator injection frequency and the received signal produce a 1-mc intermediate frequency. The 1-mc IF signal and the 1100-kc second oscillator injection frequency produce a second intermediate frequency of 100 kc. The 100-kc second IF signal is fed to the detector where it mixes with the signal from the beat frequency oscillator (BFO).

The BFO is tunable between 97 and 103 kc. This is a range of 3 kc below to 3 kc above the second intermediate frequency to allow for a variation in pitch of the recovered audio. If the BFO frequency is either 101 kc or 99 kc, mixing with the 100-kc second IF signal gives a 1-kc difference frequency to be amplified by the audio amplifier section. Varying the BFO from 100 kc to lower or higher frequencies gives an increasingly higher pitched tone out of the detector. A 100-kc BFO frequency gives a zero beat and no output from the detector.

It was shown in Chap. 2 that an SSB signal with single-tone modulation is a single RF frequency. This frequency is displaced from the operating frequency by an amount equal to the modulation frequency. It is not possible at the receiving end to distinguish between the radiation from a CW transmitter and a single-tone modulated SSB signal with the carrier suppressed completely. Both are single-frequency RF transmissions. Therefore, the reception of a single-frequency SSB signal is similar to that of CW, and therefore SSB can be received on AM receivers that are equipped with a BFO. It is often necessary to modify such receivers before they will receive SSB satisfactorily, but most BFO receivers can receive SSB well enough to recover the intelligence from voice transmissions.

## 2.3 Block Diagrams of SSB Receivers

2.3.1 Basic SSBSC receiver. Figure 6-3 is the block diagram of a basic SSBSC receiver. Note that it is very similar to the CW receiver. A sideband filter has been added. The filter is not necessary for the reception of SSB but it is necessary to obtain the signal-to-noise ratio advantage available from the narrower bandwidth of SSB signals. It will be seen later that sideband filters are necessary for the separation of sidebands when receiving independent sideband transmissions.

Lets assume that a USB signal with a 5-mc operating frequency is being received. With one kilocycle modulation, the frequency of the received signal is 5001 kc. The first oscillator frequency is the same as that used in the two previous systems. The output of the first mixer is 6000 kc minus 5001 kc or 999 kc. Note that the received USB signal is converted to a lower sideband signal and is now 1 kc below 1 mc in the first IF. The output frequency of the second mixer is 1100 kc minus 999 kc or 101 kc. This is a USB signal because it is 1 kc above the 100-kc second IF.

The carrier-oscillator signal and the 101-kc second IF signal mix in the demodulator, and the 1-kc difference frequency is recovered and amplified in the audio amplifier section. The carrier oscillator is not variable as is

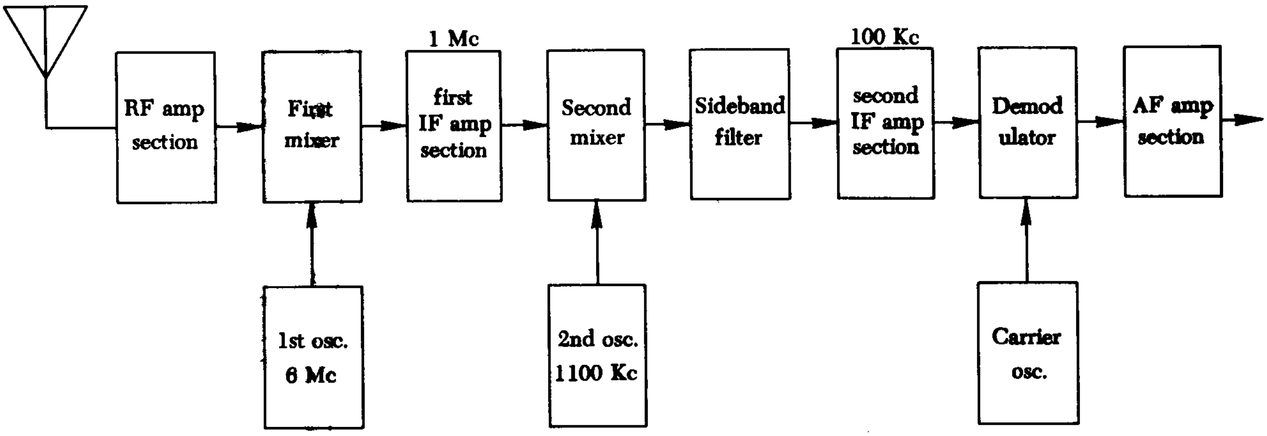


**Figure 6-3. Block diagram of basic SSBSC receiver.**

the BFO, but instead is very stable so that the reinserted carrier is located in the spectrum at the same place as the carrier would be if it had been transmitted. The stable carrier oscillator ensures that the frequency of the demodulated audio is the same as the audio that originally modulated the transmission. It has been determined that the reinserted carrier must not vary from the optimum frequency by more than 50 cycles for dependable communications. Frequency errors up to 200 cycles in the reinserted carrier may still give intelligible speech under ideal conditions, but weak signals and noisy circuits do not permit errors above about 50 cycles. Also,the proper operation of teletype circuits requires a reinserted carrier equally as stable.

In the discussion thus far, we have seen that the CW transmitter radiates a signal at the operating frequency. The SSB suppressed carrier transmission is displaced from the operating frequency by an amount equal to the modulation it contains. This difference in the two radiated signals causes a slight difference in their reception.

The CW signal was 1 mc in the first IF and 100 kc in the second IF. The

BFO was varied from 100 kc by an amount which gave the desired tone. The SSB signal was displaced from 1 mc in the first IF by an amount equal to the modulation (1 kc). In the second IF it was displaced 1 kc from 100 kc. The carrier oscillator injected at the demodulator was 100 kc. This gave a difference of 1 kc which was the transmitted intelligence.

2.3.2 SSB receiver for independent-sideband reception. Figure 6-4 represents a receiver for independent-sideband reception. Up to the second mixer, the diagram is the same as that of Fig. 6-3. After the second mixer, the two independent sidebands are separated by sideband filters and fed to separate IF amplifiers, demodulators, and audio amplifier sections.

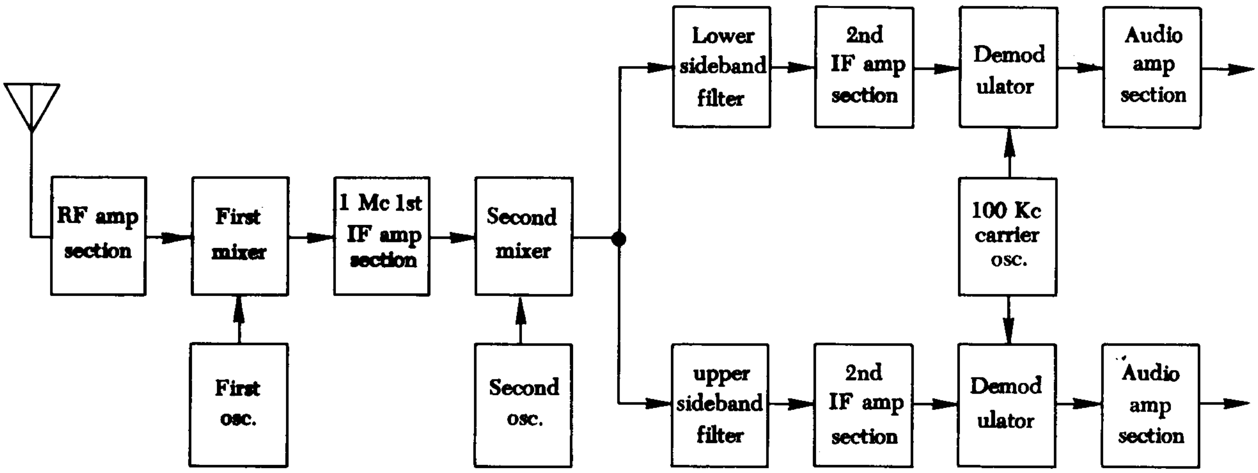


FIGURE 6-4. SSB receiver for independent-sideband reception.

## 3.0 THEORETICAL CIRCUIT CONSIDERATIONS

Figure 6-5 is a block diagram of an SSB receiver; it is similar to the diagram of Fig. 6-3. Blocks representing automatic frequency and gain control circuits have been added to Fig. 6-5. The dashed lines divide the diagram into four sections corresponding to the manner in which receiver theory is covered in this discussion.

The RF section consists of the RF amplifier stages, the high-frequency oscillator, and the first mixer. The IF section consists of the first and second

IF amplifier sections, filters, and the second oscillator. The carrier oscillator, demodulator, and audio amplifiers are contained in the audio section. The automatic gain and frequency control circuits are in the AGC and AFC section.

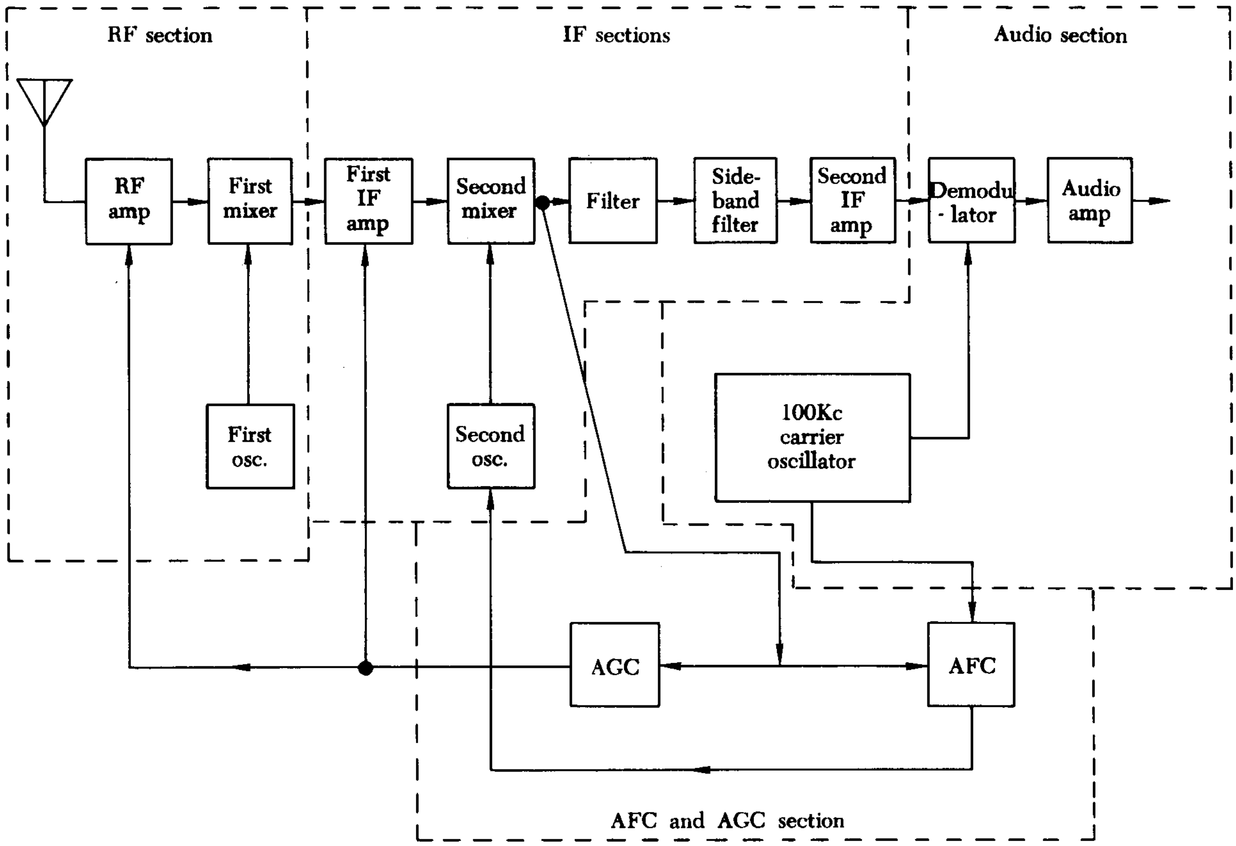


FIGURE 6-5. Block diagram of SSB receiver divided into the four main sections.

## 3.1 RF Section

Communication receivers invariably have one or more RF amplifier stages. The use of RF amplifier stages before the first mixer provides increased sensitivity, image rejection, and signal-to-noise ratio. RF amplifier stages must use low-noise amplifier tubes because the noise introduced in one of the early stages is amplified by all succeeding amplifier stages. The SSB-R3 receiver (discussed in the next chapter) has three RF stages utilizing type 6BA6 tubes. The 6BA6 is a remote-cutoff pentode. Although pentodes generate more noise than triodes, pentodes have the advantage of low grid-to-plate capacitance which permits high gain without neutralization.

For good sensitivity, high amplification ahead of the first mixer is desirable. This provides a strong signal level at the mixer that overrides noise generated in the mixer. However, in strong signal areas, high amplification adversely affects the selectivity of the receiver because adjacent-channel signals are highly amplified and cause additional spurious mixing products upon reaching the mixer. As a compromise, the gain of RF amplifiers is limited so that the level of strong adjacent-channel signals will not overload the RF stages.

The first frequency conversion is accomplished in the first mixer. The SSB signal, after being amplified in the RF amplifier section, mixes with the first oscillator signal in the first mixer to produce the difference frequency for the first IF.

A 6BE6 tube is used in the first mixer stage of the SSB-R3. The 6BE6 type tube is a pentagrid converter. Pentagrid mixers have the advantages of excellent isolation between signal and oscillator circuits and high conversion gain.

### 3.2 First and Second IF Sections

To translate an RF signal to an intermediate frequency, an oscillator injection frequency higher than the RF signal is normally used. To translate an RF signal of 15 mc to an intermediate frequency of 500 kc, an injection frequency of 15.5 mc can be used. The IF signal is the difference between the RF and injection frequencies. An undesired signal frequency of 16 mc reaching the mixer will also result in a difference frequency of 500 kc and will be amplified by the 500-kc IF amplifier. This second frequency (16 mc) is called an image frequency and is a source of trouble when the selectivity of the RF amplifier is not sharp enough to reject it. In the previous example, with a 500-kc IF, the image frequency is displaced from the desired signal by 1 mc. By choosing a higher first intermediate frequency, say 1 mc, the image frequency is removed from the desired signal by 2 mc. For example, when a 15-mc signal is received, the injection frequency is 16 mc and the image frequency of 17 mc is displaced from the desired signal by 2 mc. It may therefore be concluded that the higher the choice of first intermediate frequency, the further the image is displaced from the desired frequency, and the easier it is for the RF amplifier to reject the image frequency. Therefore, the first intermediate frequency is relatively high.

Actually, the first IF is a compromise between good image rejection and the elimination of spurious responses. The lowest receiver operating frequency determines how much compromise is necessary, as can be seen by utilizing the following example. In Fig. 6-6 a 1001-kc signal is received at the input. Due to nonlinearity the second harmonic (2002 kc) of the input signal is generated in the mixer. The second harmonic mixes with the 1500-kc oscillator signal to give a difference frequency of 502 kc. This difference frequency is amplified by the IF amplifier, and a 2-kc signal is demodulated in the audio section and appears as a spurious response in the output. Spurious responses result because the carrier-to-intermediate frequency ratio is too low. To eliminate spurious responses of this nature, the intermediate frequency should be lower than one-half the operating frequency minus the highest audio frequency to be received. This is expressed as $1/2\ f_0 - f_{a(high)}$. When the intermediate frequency is as low as that determined by the preceding equation, however, optimum image rejection is not obtained.

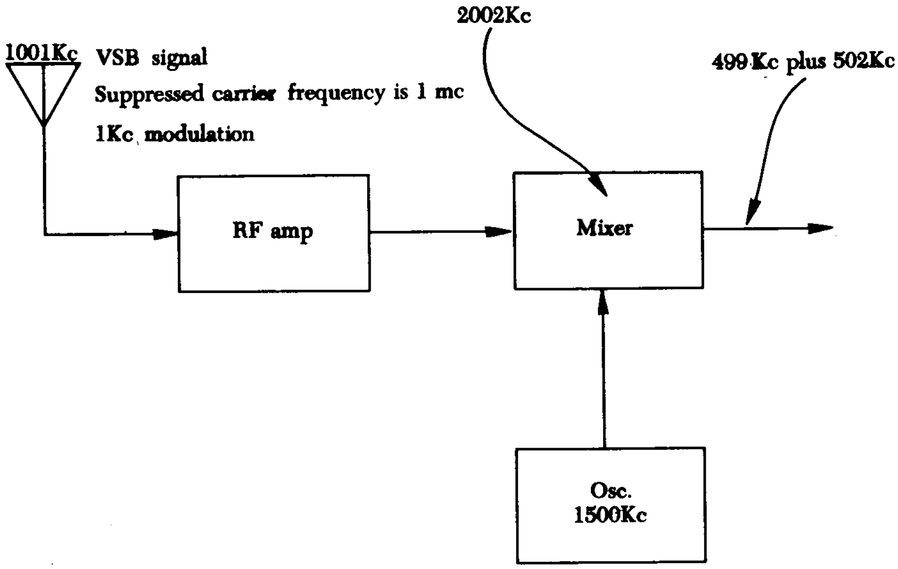


FIGURE 6-6. Illustration of spurious response because of low carrier frequency to IF frequency ratio.

A compromise must be made between good image rejection and satisfactory elimination of spurious responses. Careful selection of a mixer tube and its operating point is necessary to obtain an acceptable degree of suppression of these responses.

Because of its relatively low frequency, the second IF provides the main selectivity for a communications receiver. The frequency-selective filters are located in the second IF because sharper response characteristics can be obtained at that frequency. Optimum receiver selectivity is obtained when the bandwidth is wide enough to pass the required intelligence and at the same time narrow enough to reject signals in the adjacent channel. The shape factor of a filter is a relative measure of a filter's ability to perform this function. For the reception of one channel of SSB, the 3-db points of a filter characteristic curve should be separated by approximately 3 kc. Then the shape factor of the filter will determine how much the adjacent-channel signals are attenuated.

In receivers designed for independent sideband reception, there is another filter in addition to the sideband filters in the second IF. (See Fig. 6-7.) The total signal (both independent sidebands) first encounters a filter with a bandpass of 6 kc. This filter has steep outside skirts to eliminate unwanted frequencies outside of the two independent sidebands. From this filter the independent sidebands are fed to sideband filters which separate the sidebands. Note that the two independent sidebands are separated by approximately 400 to 600 cycles (depending on how much the lows are suppressed at the transmitter) and the carrier side of the sideband filters are of necessity very steep. To obtain this steep side requires that the steepness of the outside skirt be compromised; this is the reason for the first (6-kc bandpass) filter. This 6-kc filter in the SSB-R3 receiver is called a roofing filter. Roofing filters with a 12-kc bandpass are utilized in the

SSB-R3 when it is necessary to receive 4-channel independent sideband.

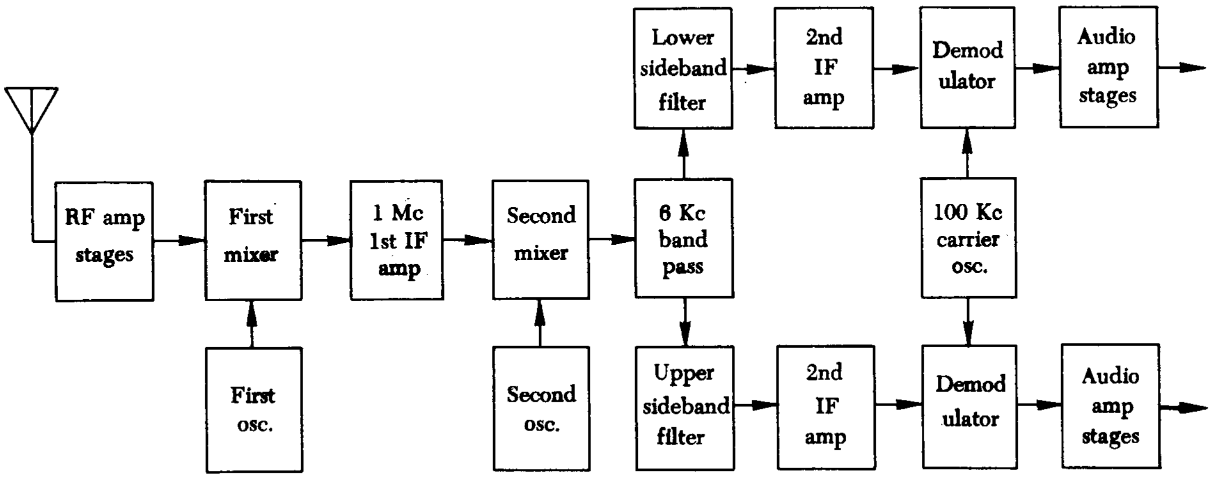


FIGURE 6-7. Block diagram of independent sideband receiver showing added band pass filter to improve outside skirts selectivity.

The attenuation curve of a 6-kc roofing filter for the SSB-R3 receiver is superimposed on the curves of its sideband filters as shown in Fig. 6-8. The curve for the lower-sideband filter is a dashed black line, the curve for the upper-sideband filter is in solid black, and the curve for the roofing filter is in dots. It is seen that the use of the roofing filter in conjunction with the sideband filters gives very selective filtering for each channel.

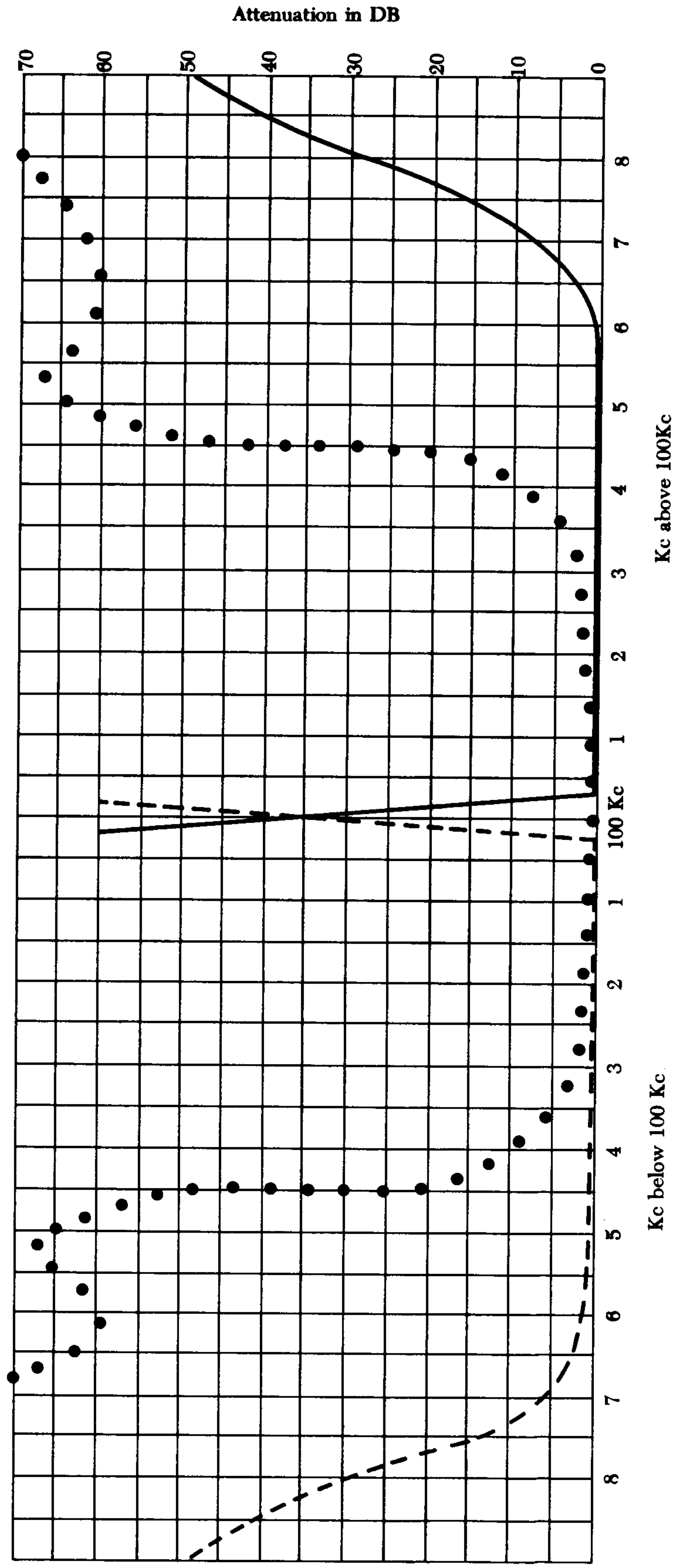


FIGURE 6-8. Attenuation curves of sideband filters and roofing filters superimposed.

### 3.3 Audio Section

The audio section of the basic SSB receiver of Fig. 6-5 consists of the carrier oscillator, demodulator, and audio amplifier circuits. Our main interest in this section is the operation of the demodulator. The process of demodulation consists of the mixing of sideband and carrier oscillator signals and recovery of the difference frequency as the desired audio.

Demodulation of single-sideband signals may be accomplished by means of envelope detectors, as in ordinary AM; but this method is not common in the reception of suppressed or reduced carrier transmissions. The most common method of demodulating an SSB transmission is by the use of some type of product detector. A product detector is a demodulator that performs multiplication of the incoming sideband signal and the carrier insertion signal.

One type of a product detector is the balanced demodulator shown in Fig. 6-9. The balanced demodulator is similar to the balanced modulator circuits discussed in Chap. 3 except that the input and output signals are reversed. Note that Fig. 6-9 shows the audio at the output instead of at the input. Also, the sideband signal which is the output of the balanced modulator is one of the inputs to the demodulator. The carrier frequency is an input in both circuits. You will recall that the signals at various points in the balanced modulator circuits were found by multiplying the two input signals. No difficulty should be encountered in seeing that this circuit is a multiplying circuit and therefore is a type of product demodulator.

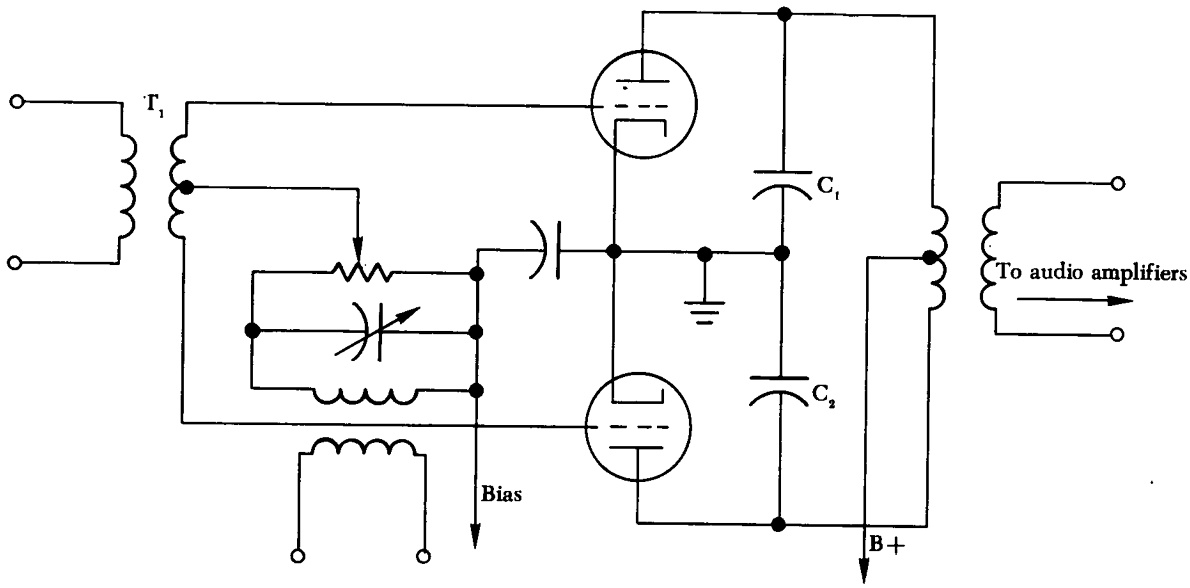


FIGURE 6-9. Balanced demodulator.

Another type of product detector circuit shown in Fig. 6-10, utilizes a pentagrid converter type in a conventional heterodyning or mixing circuit. The IF sideband signal is applied to one grid and the carrier injection to another grid. The sum frequency as well as the difference audio appear from the mixing but they are separated so widely in frequency that there is no problem in eliminating the sum frequencies while passing the audio on to amplifiers.

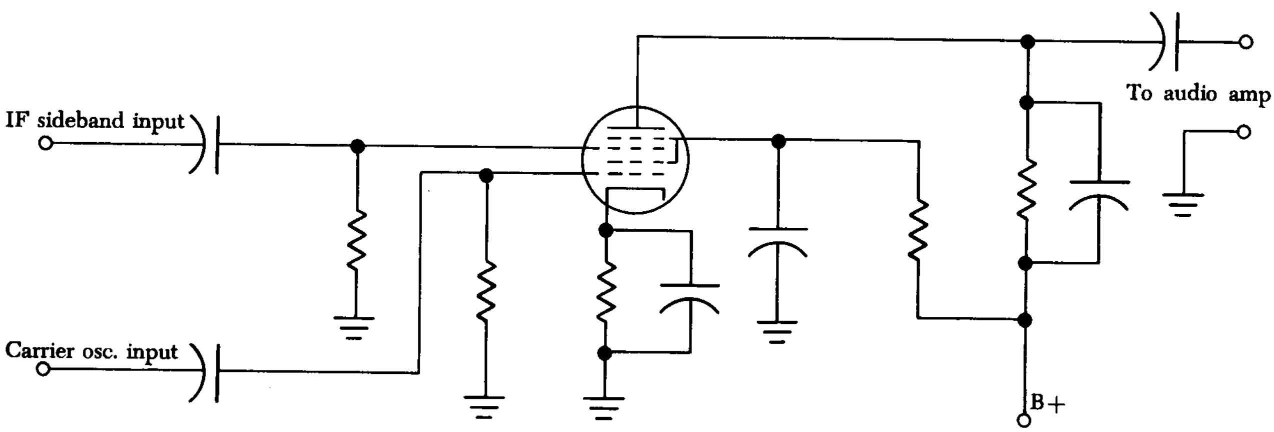


FIGURE 6-10. Product detector utilizing a pentagrid converter.

The product demodulator circuit used in the SSB-R3 receiver is shown in Fig. 6-11. The sideband and carrier signals are applied to the cathode of V2B via cathode followers. This provides good isolation of the two input signals. The grid of V2B is grounded and application of the two inputs at the cathode cause its plate current to contain the sum and difference frequencies as well as the original input frequencies. All frequencies except the audio are rejected by the filter in the plate circuit of V2B.

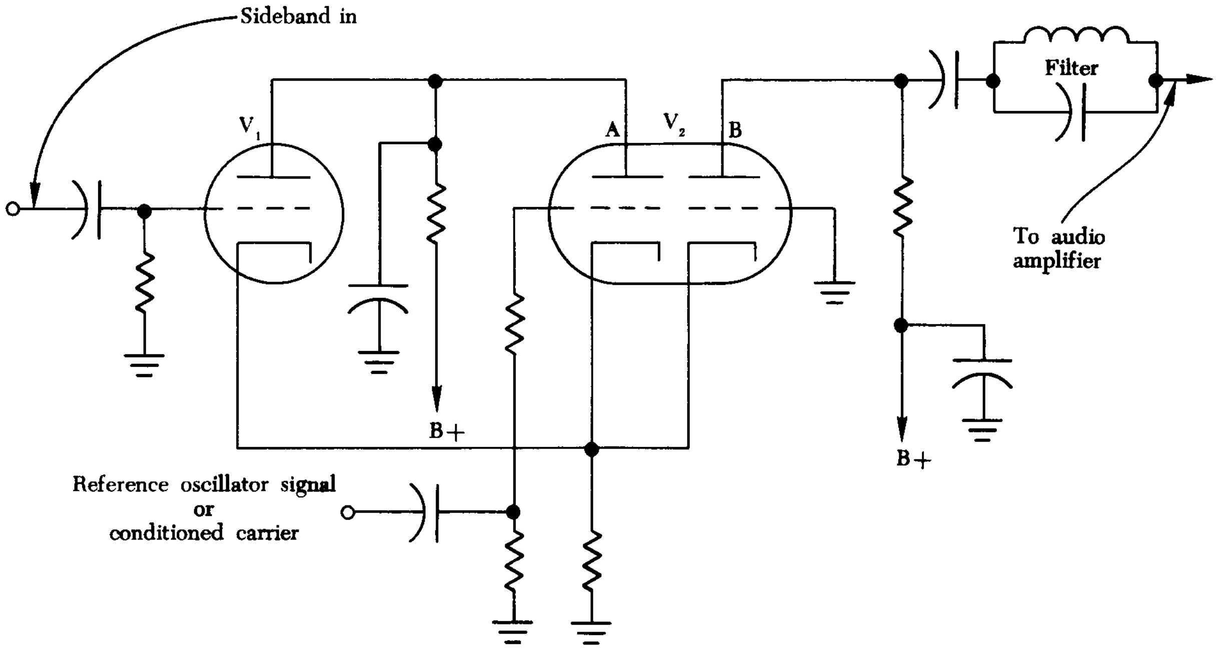


FIGURE 6-11. Product detector of SSB-R3 receiver.

The demodulated audio frequencies consist of voice or teletype signals which are fed from the receiver site to the control station by lines or radio link.

## 3.4 Frequency and Gain Control Circuits

### 3.4.1 Automatic frequency control.

Automatic frequency control (AFC) is utilized to ensure that the receiver remains tuned to the transmitted frequency. AFC compensates for instabilities in the transmitter and in the receiver and thus permits more reliable communications. There are so many different techniques for automatic frequency control that no attempt will be made to cover AFC generally. Instead, the method of AFC utilized in the SSB-R3 is explained briefly here. In the next chapter the theory of the AFC circuitry will be explained in detail.

AFC cannot be used in SSB unless a pilot or reduced carrier is transmitted. The SSB-R3 receiver amplifies the pilot carrier along with the single-sideband signal. At the second intermediate frequency the carrier is separated by filters and conditioned to the proper level before being fed to the AFC circuit. The frequency of the conditioned carrier is compared with that of a highly stable oscillator in a comparison circuit. The output of the comparison circuit is amplified and fed to a motor that drives a tuning capacitor in the second oscillator. Tuning the oscillator in accordance with changes in signal frequencies ensures that the carrier frequency in the second IF remains at 100 kc. Thus, when the 100 kc reference signal is inserted for demodulation, the audio signals occupy the same position in the audio spectrum that they occupied when the transmitter was modulated.

3.4.2 Automatic gain control. The function of an automatic gain control (AGC) system is to keep the signal output of a receiver relatively constant for changing input signals.

The AGC voltage in an SSB receiver is taken from the modulation envelope. This envelope varies with modulation which makes it necessary to utilize a long-time-constant circuit to smooth out the variation. The signal for the AGC circuit is taken from the second IF or from the demodulated audio. This AGC information is rectified, amplified, and fed back to RF and IF amplifiers.

Some sideband receivers provide for several types of AGC which can be switched in by the operator. The SSB-R3 receiver has carrier-only AGC, carrier-with sideband AGC, and sideband aggregate AGC. These different methods of AGC operation will be explained during the detailed circuit analysis of the SSB-R3 receiver in the next chapter.

ATTENTION

Examination 6 is to be worked at this time. This exam covers the material in Chap. 6. It is included with the material.

CHAPTER 7

# SSB-R3 SINGLE-SIDEBAND RECEIVER

## 1.0 INTRODUCTION

A brief analysis of the SSB-R3 receiver is given in this chapter. Chapter 8 will include a discussion of the AN/FRT-39 transmitter. These equipments are chosen for discussion because both are found at most International Flight Service Stations. It was decided to discuss the receiver while receiver concepts are fresh in mind and to leave the transmitter until last. It is realized that this order is reversed from the order in which transmitter and receiver theory was covered.

Because the SSB-R3 is designed for space-diversity reception, it is necessary to discuss the advantages of space-diversity reception before considering receiver circuitry.

Single sideband is used in the FAA for long-range point-to-point communications in the high-frequency band. High-frequency systems depend on signals reflected from the ionosphere. The paths of reflected signals are subject to frequent changes that result in short-interval fading. Experience has shown that short-interval fading seldom occurs simultaneously at two locations that are separated by several wavelengths. Therefore, the effect of fading can be reduced by using a system of two antennas and two receivers. The antennas are separated by several wavelengths and each is connected to a separate receiver input. By electronically selecting the stronger signal and squelching the weaker signal, an output consistently as good as the stronger signal is made available. In this manner, the effect of short-interval fading is reduced and communications become more dependable.

Another type of fading, called long-interval fading, results from a day-to-night shift in the ionosphere. The shift from day to night causes the angle of reflection of the transmitted wave to change. Because different frequencies have different reflecting characteristics, the effect of long-interval fading is reduced by lowering the operating frequency at night. The frequency range of the SSB-R3 receiver is more than sufficient to allow the necessary frequency shift from day-to-night operation. The frequency range is from 2.8 to 28 mc.

The various units of the receiver system are built for rack mounting. The units are interconnected by cables terminated with coaxial connectors or Jones plugs. Antenna connections and all operating controls are located on the front panel. A picture, included at the beginning of the text, shows the physical location of the several units.

A block diagram of the SSB-R3 is shown in Fig. 7-1. The diagram represents two receivers with oscillators and other circuits common to both. One receiver alone is very similar to the block diagram considered in Chap. 6.

It is convenient to refer to the units of the receiver in the same manner as the manufacturer has numbered them. It should be understood that every unit of the receiver cannot be represented by one block on the block diagram. In some instances, it takes more than one block to represent the functions of a unit and, in some cases, a block may represent a function that is started in one unit and completed in another. Therefore, for purposes of study, it is not convenient to divide the receiver the way it is divided by the block diagram of Fig. 7-1. Instead we will consider the receiver in an order corresponding to the order that the circuits are encountered by signal flow through the receiver. This order leaves the discussion of afc and agc circuits until last. In tracing the signal through the receiver, momentary digressions from the signal path will be made to discuss the signal sources for frequency conversions and demodulation. This order has been decided upon because it enables one to more readily understand the operation of the receiver.

The schematic diagrams used in circuit discussions are included in foldouts at the end of this chapter.

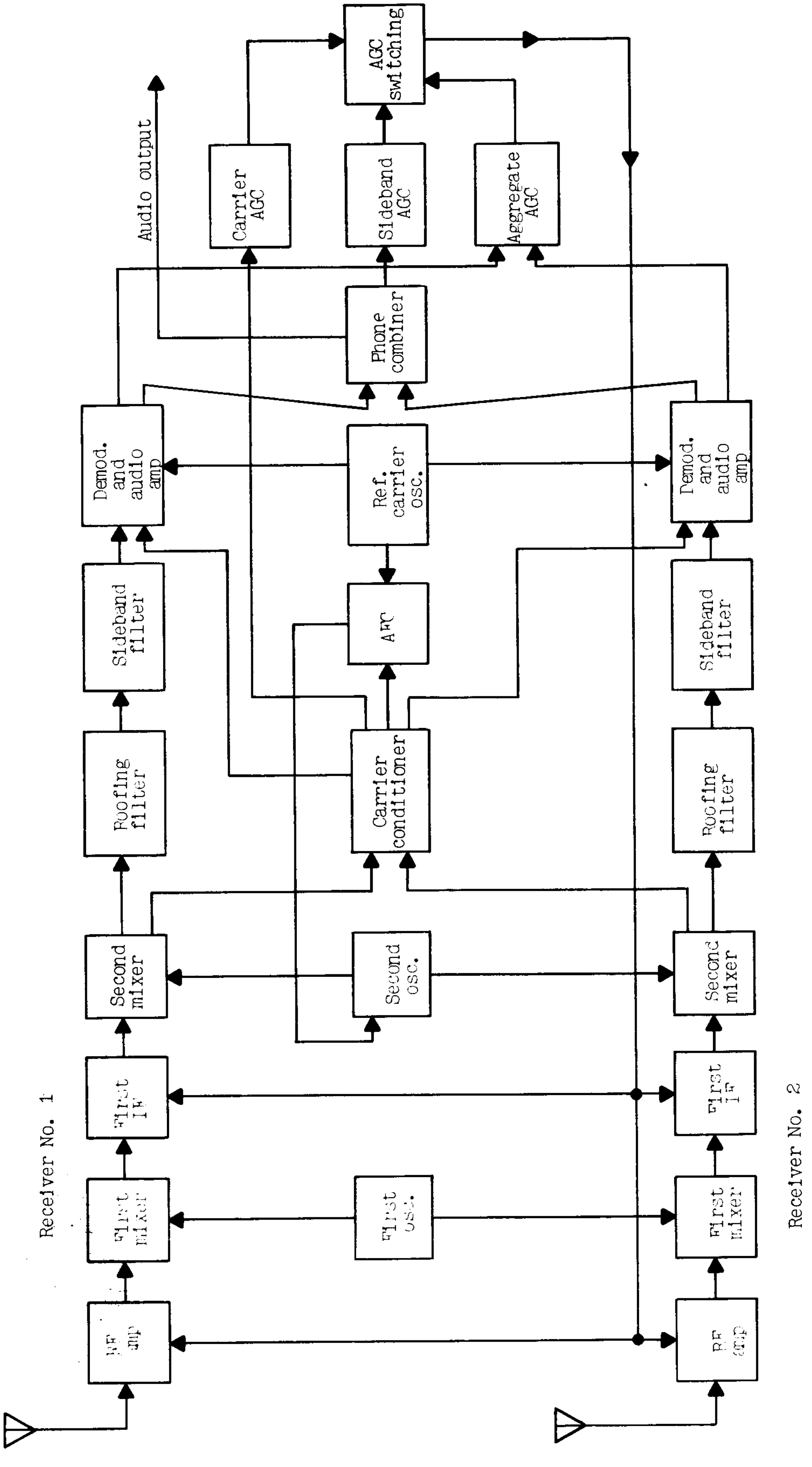


Fig. 7-1. Block diagram of SSB-R3 dual diversity receiver.

## 2.0 RF AMPLIFIER AND FIRST CONVERTER

The picture of the SSB-R3 receiving system (near the front of the text) shows two rf amplifier units. Two rf amplifier units are necessary to cover the frequency range of 2.8 to 28 mc. One unit covers the range of 2.8 to 9 mc and the other covers the range from 9 to 28 mc. Within each amplifier unit separate amplifier sections are necessary to cover that units frequency range. Unit #200 utilizes four separate rf amplifier sections. Two sections, operating at the same frequency, cover the range from 2.8 to 5 mc. The other two sections, operating at the same frequency, cover the range of 5 to 9 mc. Two rf sections operating at the same frequency in each case provide for dual diversity reception. In unit #300, two rf sections cover the frequency range of 9 to 16 mc and the remaining two sections cover the range of 16 to 28 mc.

A look at Figs. 7-2 and 7-3 shows that each rf amplifier unit contains four amplifier sections for a total of 8 sections in both units. Only one unit is operated at any one time and only two amplifier sections of the operating unit are utilized at one time.

Both rf amplifier units are identical except for frequency coverage. Therefore, only unit #300 will be discussed. The schematic of unit #300 is shown in Fig. 7-2. The upper half of the schematic contains amplifiers for receiver #1 and the lower half contains amplifiers for receiver #2. The two amplifiers of receiver #1 or receiver #2 will never be used at the same time. Instead, an amplifier from receiver #1 is used simultaneously with an amplifier from receiver #2.

Each rf amplifier section has three stages utilizing 6BA6 tubes. Each operating amplifier feeds to the first converter, which uses a 6BE6. The high-frequency oscillator signal is fed to the converter through J304-1 (for receiver #1) from the high-frequency oscillator unit. The 1-mc output of the converter is fed to an IF transformer which is used for both bands. A 12AU7 is used in a cathode-follower output stage. The 1-mc output is fed from J303-1 to the 1-mc IF amplifier located in unit #600.

It should be remembered that unit #200 contains four separate rf amplifier sections also and that the unit provides for operation between 2.8 and 9 mc. Figure 7-3 is a schematic diagram of that unit.

## 3.0 FIRST OSCILLATOR

The oscillator signal fed to the first converter, in the rf amplifier unit at J304-1, comes from unit #400, the high-frequency oscillator unit. shown in Fig. 7-4. In addition to providing the oscillator sig- the 6BE6 first converter tubes in the rf amplifier units, the high-frequency oscillator unit provides bandswitching for the rf amplifier unit. A schematic of the oscillator unit is shown in Fig. 7-4.

The oscillator frequency must be tuneable from 3.8 to 29 mc to obtain a difference frequency of 1 mc with received signals between 2.8 and 28 mc. It is difficult to design a variable-frequency oscillator that is stable because continuous crystal control is not possible. The complicated appearance of Fig. 7-4 is due to the captive-type circuit that controls the output frequency of the oscillator. The frequency of the oscillator is phase-locked to a crystal-controlled oscillator and to a low frequency stabilized variable-frequency oscillator. In this manner, the oscillator is made stable and variable.

The frequency of the oscillator is selected by a four-position switch, which selects the frequency band, and a tuning dial that adjusts the variable frequency. The band switch in the oscillator also switches heater voltages to the rf amplifiers for the corresponding band of operation. The oscillator frequency is 1 mc higher than the incoming signal frequency. However, the tuning dial is calibrated to indicate the signal frequency.

The oscillator signal fed to the first converter originates in the stage containing V2. Most of the remaining circuitry controls the frequency of V2 oscillations. The rf output of V2 is fed in parallel to three isolation amplifier stages, V15, V16, and V17, having separate outputs.

Two outputs from the isolation amplifiers are used for dual diversity operation; the third output is used when three-receiver diversity operation is employed. The operation of this circuit is not to be confused with automatic-frequency control. This circuit maintains the oscillator frequency constant. An automatic-frequency control circuit forces the receiver to stay tuned to a carrier signal whether it is drifting or not. A brief description of the oscillator control circuits follows.

A 1-mc signal generated by crystal oscillator V10 is fed to a series of harmonic amplifiers, V11, V12, and V13. The harmonic amplifiers provide the desired harmonic from 6 to 31-mc at the proper output level. The proper harmonic is selected by ganged switch S4. The correct output level from the harmonic circuit is maintained by an agc voltage fed back to the grids of V12 and V13. The agc voltage is developed by amplifying the dc output of CR4 and CR5 in the dc amplifier, V14B.

The output from cathode follower stage V14A is fed to mixer V4. A sample of the output from the high-frequency oscillator, V2, is also applied to the mixer through an isolation amplifier, V3. The output from the mixer is applied to the phase detector diodes, CR1 and CR2, through bandpass circuits which include T1, V5, and T2. Also, an output from the stabilized low-frequency variable oscillator is coupled in push pull, by the phase splitter V8-B, to the phase detectors, CR1 and CR2.

Positive and negative voltages from the phase detector are applied to the grid of reactance tube V1. The voltages on the grid of the reactance tube change the impedance presented to the oscillator frequency-determining network thereby controlling the frequency of the oscillator. The rf output of the oscillator is fed to the isolation amplifiers mentioned earlier and in turn to the rf amplifier unit.

## 4.0 FIRST IF AND SECOND CONVERTER

Having discussed the source of oscillator signal for the first frequency conversion, we are ready to continue with the 1-mc output of the rf amplifier unit. This signal is fed to unit #600 consisting of the 1-mc dual IF, the 100-kc oscillator, and the agc selector circuit. See Fig. 7-5. This unit provides two identical signal channels, each comprising two 1-mc amplifier stages, one 100-kc converter, and a cathode-follower output stage. This unit also contains the highly stable crystal-controlled oscillator, that supplies the 100-kc reference signal, and a dual-channel sideband aggregate agc system with an agc switching circuit. We will discuss the IF and converter stages now and return to the others at a more appropriate time.

The two identical 1-mc signals from the 12AU7 coupler, in the rf amplifier unit, are fed to their corresponding IF amplifiers in unit #600. The output of receiver #1 enters this unit at J601; the output of receiver #2 enters at J603.

The relatively high frequency for the first IF was chosen to provide good image rejection in agreement with design considerations discussed in Chap. 6.

After being amplified by two amplifier stages, the 1-mc IF signals are fed to the 6BE6 second converter stage. The oscillator feeds the first grid of each converter through J619. Provision is made to inject the second oscillator signal into the second converter at either 900 or 1100 kc. We will digress for a moment to consider how switching the oscillator frequency enables the reception of either the lower or upper sideband when only the upper-sideband filters are installed.

### 4.1 Sideband Switching Considerations

Normally, a sideband receiver is capable of receiving only one sideband signal at a time. However, all receivers are designed to switch from lower to upper sideband reception. When both upper and lower sideband filters are installed, the SSB-R3 system is capable of receiving both sidebands simultaneously, provided separate demodulator stages are available. FAA installations provide for the reception of both sidebands in independent sideband operation. There is then little need for switching the oscillator frequency. However, if an SSB-R3 installation has only upper sideband filters, it can only receive one sideband at a time and the 900- or 1100-kc choice of second oscillator frequency permits switching from one sideband to another (from lower to upper or vice versa).

To begin our study of sideband switching, assume that only upper sideband filters are installed and that an upper sideband signal conveying 1000-cycle tone modulation and having a suppressed carrier frequency of 10 mc is being received. The rf signal is actually 10,001 kc. This 10,001-kc signal mixes with 11 mc in the first converter to produce a difference frequency of 999 kc for the first IF. This 999-kc sideband signal mixes with 1100 kc in the second converter, to produce a difference frequency of 101 kc for the second IF. This 101-kc signal appears as an upper sideband in the 100-kc IF and is passed by an upper sideband filter designed to pass frequencies from 100 to 103 kc. A 1000-cycle tone will eventually be demodulated.

Next, assume that a lower sideband signal, conveying 1000-cycle tone modulation and having a 10-mc suppressed carrier frequency, is being received. The rf signal is 9,999 kc, and when mixed with a 11-mc oscillator signal, produces a first IF frequency of 1,001 kc. When mixed with 1100 kc in the second mixer, a 99-kc signal appears in the second IF. This signal will not be passed by the upper sideband filter and nothing appears in the output. However, if an oscillator signal of 900 kc is injected in the second converter, the difference frequency, 101 kc, is passed by the upper sideband filter. Thus, sideband switching is affected by changing the second oscillator signal.

Both the 900- and the 1100-kc second oscillator frequencies will produce a 100-kc* IF signal. The 100-kc IF signal is fed to a 12AU7 output stage before being applied to two separate units, the roofing filter (unit 1000) and the carrier conditioner (unit #700).

At this time, we will trace the signal through the roofing filter; later we will consider the 100-kc signal going to the carrier conditioner.

## 5.0 ROOFING FILTER

The roofing filter (Fig. 7-6) contains two sets of filters and two pairs of cathode-follower output stages. One set of filters and one pair of output stages are used for each receiver. The roofing filters narrow the bandwidth of the 100-kc IF signal. A 7-kc (3.5-kc on both sides of center) roofing filter is used for independent sideband operation with one complete voice channel on each side of the carrier frequency. A 12-kc roofing filter is used for independent sideband operation having two complete voice channels on each side of the carrier frequency. The inputs to the roofing filter come from unit #600, entering at J1001 and J1002 for receivers #1 and #2, respectively. Power is fed to the roofing filter unit through P1001. The outputs of the unit (at reduced bandwidth) are fed to the sideband filter unit, #1100, Fig. 7-7.

## 6.0 SIDEBAND FILTERS

The two receiver outputs of the roofing filter enter the sideband filter unit at J1101 and J1102. The upper and lower sideband filters are contained in this unit and provide for independent sideband operation. The lower sideband filters are present but are used only when both sidebands are received. FL1, FL2, FL3 and FL4 of this unit are

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* Actually the carrier, if present, would be 100 kc in the second IF. The sideband signal will be displaced from 100 kc by an amount equal to the modulating frequency.

crystal lattice filters similar to the ones discussed in Chap. 4 of this text. The cutoff frequency of each filter on the carrier side is sharp to prevent crossfeed from the other sideband. Each filter is designed to pass from very low audio frequencies up to 6 kc from the carrier frequency. The 6-kc limit provides for two voice channels in each sideband.

Considering only receiver #1, the signal entering J1101 (Fig. 7-7) takes two paths and is amplified by both V1A and V1B. CR1 and CR2 form a diode clamping circuit to prevent excessive input voltage from damaging the filters. The output of V1A goes to the upper sideband filter (passes upper sideband) and the V1B output goes to the lower sideband filter (passes lower sideband). The separated sidebands from the filters are fed to cathode follower output stages. Four separate output signals are available from the sideband filter unit. The output from V3A and V4A consist of upper sideband signals; they are fed to the same demodulator unit where they are demodulated separately. The output from V3B and V4B consist of lower sideband signals; they are fed to the same demodulator unit which is identical to the one to which the upper sidebands are fed.

## 7.0 SECOND IF AMPLIFIER, DEMODULATOR, AND AUDIO AMPLIFIER

Unit #1200, Fig. 7-8, contains the demodulators and audio amplifiers for two receiver channels. The output of the sideband filter unit enters the demodulator at J1201. The 100-kc IF sideband signal is amplified by V1 and passed through IF transformer T1 to the demodulator circuit. The reference signal, necessary for demodulating the sideband signal, is applied to the grid of the second tube of the demodulator stage. The reference carrier entering this unit comes from one of two sources. One source is the 100-kc reference oscillator signal coming from unit #600. The other source of reference signal is the carrier conditioner unit, #700. At the present, we will assume that a carrier reference signal is present for demodulation. The sideband signal and the reference carrier signal mix in the demodulator. The sum frequency resulting from the mixing is filtered out by tuned circuits following the demodulator stage. The difference sideband signal is passed by the tuned circuits and fed to a phase inverter to drive push-pull audio amplifiers. The audio output from this unit consists of either voice communication or frequency-shift teletype signals.

## 8.0 SIGNAL SOURCES FOR DEMODULATION

Before we continue to the next unit with the demodulated audio signal, the two sources of carrier reference signals for demodulation will be discussed. Refer to the block diagram in Fig. 7-1. In addition to the second IF signal entering the demodulator unit, there is an input from the 100-kc reference oscillator, and another input from the carrier conditioner. Only one of the two sources will be utilized at a time.

### 8.1 Reference Carrier Oscillator

The reference carrier oscillator is used when no pilot carrier (carrier at reduced level) is transmitted. It provides a stable signal for the demodulator unit. This signal bears the same frequency relationship to the sideband as that of the carrier before suppression at the transmitter.

The 100-kc reference oscillator is contained in unit #600, Fig. 7-5. The oscillator circuit consists of the crystal-controlled oscillator, V10; a stage of amplification, V11A; and a cathode-follower output stage, V11B. Feedback is applied to the oscillator (from the output stage) to keep the output level constant.

### 8.2 Carrier Conditioner

When a low-level carrier is transmitted along with the sideband information, the reception process is somewhat simplified. The received carrier maintains the proper frequency relationship with the transmitted sideband signal even though there is a slight drift in transmitter frequency. However, the amplitude of the low-level carrier is not sufficient for demodulation. Thus, the purpose of the carrier conditioner is to ensure that the pilot carrier has the desired level for demodulation. The block diagram shows that a portion of the signal from the low-frequency conversion is fed to the carrier conditioner. The output of the carrier conditioner is fed to the demodulator.

A schematic diagram of the carrier conditioner, unit #700, is shown in Fig. 7-9. This unit provides two functions in addition to supplying a reference carrier for the demodulator. It provides a carrier automatic gain-control voltage and a voltage for squelching automatic-frequency control in unit #500 when the carrier becomes too weak for proper afc action. The two latter functions will be discussed later.

The carrier signals to be conditioned are taken from the second IF, jacks J606 and J609 of Fig. 7-5. A carrier is taken from the IF of each receiver and both signals are conditioned in similar circuits. Only the conditioner for receiver #1 is discussed.

The signal coming from the second IF and entering the carrier conditioner at J701 (Fig. 7-9) has not passed through the roofing or sideband filters. This signal contains the sideband as well as the carrier. The composite signal is fed to the grid of amplifier V1A from R1, a 100,000-ohm level control. V1B (the other triode section of the 12AT7) is a cathode-follower which couples to the carrier filter (passes the carrier). The sideband is filtered out and the carrier is amplified in V2, limited in V3A, and fed to cathode-follower output stage, V3B. The output, coupled from the cathode of V3B through C11 and T1 to a diode gating system, CR5 and CR6, will be considered at this time. The gating system selects the strongest carrier from the two receivers. The selected carrier is further amplified and limited by stages V7A, V7B, V8A, V8B, and V9 and passed to cathode-follower output stage V10A where the conditioned carrier for the demodulator is taken from J703. Another output of V10A, going to carrier rectifier diodes, will not be considered at this time.

## 9.0 RECEIVER OUTPUT

We digressed from a discussion of the demodulator and audio amplifier unit to discuss the two sources of carrier reinsertion signals for the demodulator. In Fig. 7-8, it may be seen that the audio output from push-pull output stage V6 is coupled through T3 to J1211 or J1212 (for receiver #1).

The diversity audio leaving the demodulator and audio amplifier of each receiver can be combined in various ways by the different equipments that are available with the SSB-R3 system. For example, in the diversity phone combiner, the two receiver signals are amplified separately and the stronger receiver signal is passed to the output. The output from the receiver having the weaker signal is suppressed.

The two audio diversity signals are combined in a manner best suited for the installation. Voice communications can be applied to a communications link to be relayed from the receiver site to the communications control center. Diversity teletype signals can be converted in additional equipment to dc pulses which, in turn, key tones that modulate the communications link. These tones are converted to keying pulses at control and fed to teletype loops.

## 10.0 AUTOMATIC FREQUENCY AND GAIN CONTROL

We have followed the signal from the input of the receiver to the audio output with as few digressions as possible. In doing so, we have neglected the afc and agc circuits of the receiver. Starting with the afc circuit operation, we will go back and secure these loose ends.

### 10.1 Automatic Frequency Control

Automatic frequency control is used when a carrier is received. AFC is accomplished by controlling the frequency of the second oscillator in such a manner that the difference between the carrier, present in the first IF, and the second oscillator frequency is always 100 kc.

The carrier utilized in afc operation is taken from unit #700, the carrier conditioner, and applied through J501 to the afc circuit in the second oscillator and afc unit, Fig. 7-10. The conditioned carrier is amplified by V1A and coupled to cathode follower stage, V1B. The output from V1B is applied to T1. The secondary terminals, 4 and 5, of T1 applies the conditioned carrier to the grids of V5B and V4B. The grids of V5B and V4B are connected to opposite ends of the secondary winding; thus, the carrier signals on the grids are 180 degrees out-of-phase with each other. The two outputs of the phase-shift networks are 180 degrees out-of-phase with each other. The output of the phase-shift network going to V5A is 90 degrees advanced in phase with respect to the carrier at terminal 5 of T1. The output of the phase-shift network going to V4A is 90 degrees delayed in phase with respect to the carrier at terminal 5 of T1. When the carrier at terminal 5 is used as a reference, the conditioned carrier on the grids of the driver tubes can be represented vectorially as shown in Fig. 7-11.

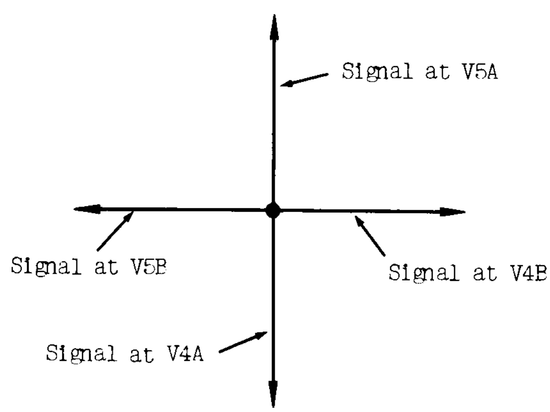



Fig. 7-11. Vector diagram representing relative phase of conditioned carrier on the grids of the driver tubes.

There is another source of signal for afc operation. The 100-kc reference signal from the oscillator in unit #600 is applied to an amplifier, a cathode follower output stage, and to T2, in that order. From the secondary of T2, the reference signal is fed to the center tap of T1, which applies the reference signal in phase to all four driver tubes. The output of the driver tubes are fed to the grids of motor control tubes. The plate load of each control tube is one of the windings of the afc motor. In addition to assuming that the conditioned carrier and the 100-kc reference signal are both present at T1 at exactly the same frequency and amplitude, assume that the carrier and 100-kc reference are in phase at the grid of V4B. The two signals will be 90 degrees out-of-phase at V4A, 180 degrees out-of-phase at V5B, and 270 degrees out-of-phase at V5A. The two signal voltages add vectorially on the grids of the tubes in the manner shown in Fig. 7-12.

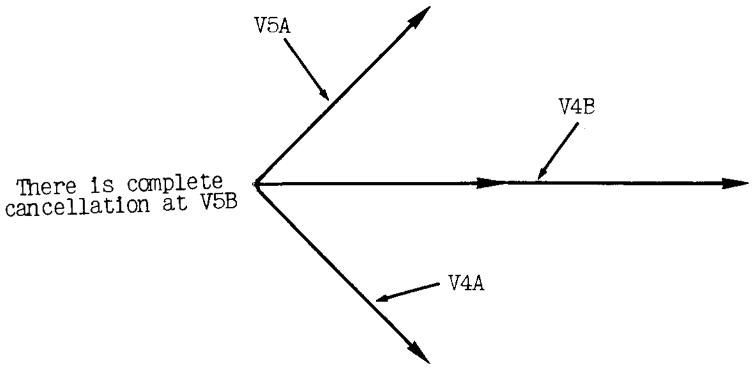



Fig. 7-12. Vectorial addition of conditioned carrier and reference carrier on the grids of the driver tubes.

As long as the two signals remain at the same frequency and phase relationship as assumed above, pulses in the motor-control tube-plate circuits feeding the motor windings will be proportional to the resultant voltage driving the corresponding driver tube. With the phase relationship assumed above, the current in winding #1 is greatest. The currents in windings 2 and 3 are equal but less than in winding #1. Winding 4 has no current. The same relative amount of current exists in the same windings cycle after cycle if the frequencies remain the same and the same assumed phase relationship continues to exist. The afc motor will not rotate because there is no rotating field under these conditions.

To see how afc operation is obtained, assume that the conditioned carrier is slightly higher in frequency than the 100-kc reference signal. Beginning at the instant when the maximum values of both waves appear simultaneously on the grid of V4B, we will analyze afc operation. In representing the voltages on the grids of the driver tubes, we will stop the 100-kc reference vector. Then the carrier vector will appear to rotate counterclockwise at a rate equal to the difference frequency. See Fig. 7-13.

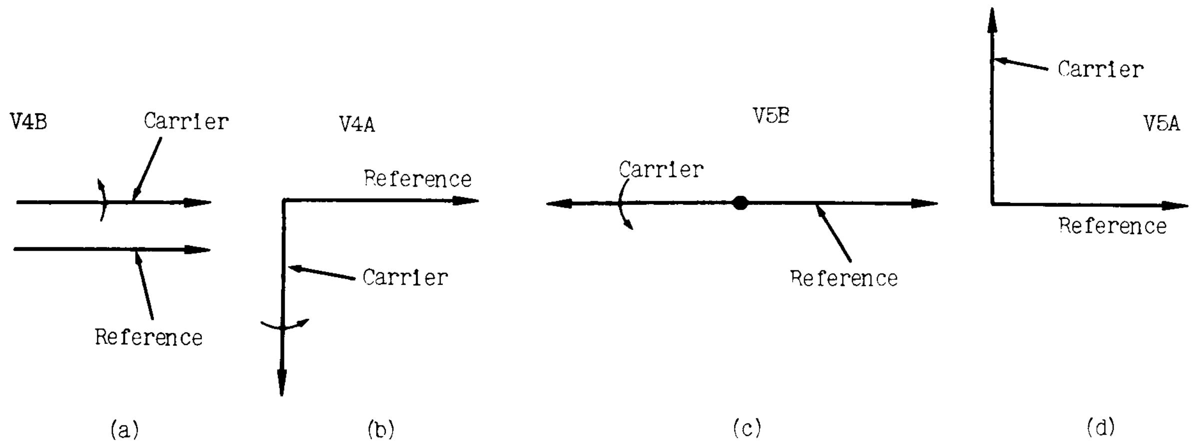



Fig. 7-13. Vectorial addition of voltages on grids of the driver tubes at the instant the two voltages are in phase at V4B. Note that in (b) (c), and (d) the voltages will come in-phase on the grids of V4A, V5B, and V5A in that order.

From the vector diagram, it can be seen that the carrier vector is in phase with the reference carrier on the grid of V4B at the instant chosen to begin the discussion. This corresponds to the time that the voltage on the grid of V4B is its maximum possible value, and when a maximum value of current exists in motor winding #1. As time goes on, the carrier vector, rotating counterclockwise, goes out of phase with the reference at V4B. Next, the carrier goes in phase with the reference on the grid of V4A and, as a result, the current in winding #2 then will be maximum. The maximum values of current, in turn, appear in windings #3 and #4, in that order as the carrier and reference vectors, representing grid voltages, come in phase at the grids of V5B and V5A. The result is a rotating field in the motor which causes rotation of the armature. The motor armature, in turn, rotates the afc range capacitor in a direction that increases the second oscillator frequency.

Rotation of the field and motor armature is in the opposite direction when the carrier frequency is lower than the 100-kc reference and the afc motor then turns the afc range capacitor to lower the second oscillator frequency. Thus, the afc range capacitor is designed so that when the carrier frequency raises or lowers, the second oscillator frequency raises or lowers a corresponding amount to maintain a difference frequency of 100 kc for the second IF.

10.1.1 AFC indication. A system of neon bulbs on the front panel of the receiver gives an indication of afc operation. A neon bulb connected in the plate circuit of each motor control tube lights once for each time maximum current exists in that tube. The four neon bulbs are located in a circle on the front panel of the unit in the clockwise order of V9, V8, V7, and V6. When the carrier frequency is above the reference, the bulbs light in a manner that gives the appearance of clockwise rotation; and when the carrier is below the reference, the rotation appears to be in a counterclockwise direction. Of course, the faster the appearance of rotation of the lights, the greater the deviation of the carrier from the 100-kc reference.

AFC SQUELCH operation is squelched when the level of the received carrier is too low for proper afc operation. In Fig. 7-9, we have already considered one output of V10A which went to the demodulator via J703. Another output from V10A goes to the carrier rectifier consisting of CR7 and CR8. The RC circuit following CR7 and CR8 filters out dc ripple at the carrier frequency. However, when the carrier is weak and noisy, the lower frequency noise variations will reach V11 where the rectified voltage is passed to J706. From that jack, the squelching voltage is fed to unit #500, Fig. 7-10, where it is amplified by V3A and then utilized in a voltage-divider circuit. When the carrier becomes noisy, the voltage-divider circuit operates to bias V1A to cutoff and the carrier coming in at J501 therefore is not passed to the afc circuit. Thus, afc operation is squelched.

### 10.2 Automatic Gain Control

The agc selector switch, S2, in unit #600, Fig. 7-5, provides for 5 types of gain control which includes sideband-aggregate agc, carrier-only agc, carrier and sideband agc, external agc, or manual gain control. Sideband-aggregate agc is discussed first.

10.2.1 Sideband-aggregate agc. A signal from the second IF (J606 in Fig. 7-5 for receiver #1) is applied to J1201 in Fig. 7-8. A portion of the signal near the input jack is applied to V15A, a cathode follower. The signal for receiver #2 enters J1202 of Fig. 7-8 and feeds to cathode follower V15B whose output goes to J1209. These takeoff points, being near the demodulator stages contain the same sideband signal that reaches the demodulator. These two receiver sideband signals are combined upon rectification; thus, this method of agc is called sideband aggregate agc. The outputs of J1207 and J1209 go to J615 and J617, respectively in Fig. 7-5. The sideband signal of each receiver is amplified in its corresponding amplifiers (V12, V13 for rec. #1; V14 and V15 for rec. #2) and fed to cathode followers. The output diodes of each receiver section are connected together and the combined agc output is fed to the agc selector switch, S2. The diversity balance of each receivers agc signal is provided by R75 and R95 in the grid circuits of V12 and V14. The balance is indicated when meters M1 and M2 have the same reading.

We have just considered the combined source of sideband-aggregate agc from diodes CR1, CR2, CR3, and CR4 which is fed to #3 contact on S2. From that contact, the sideband-aggregate agc voltage is fed to agc diode V9 which contains the time-constant switch, S5, in its plate circuit. Three choices of time constants provide for slow, medium, and fast discharge times, which are 57, 7.2, and .72 seconds, respectively. From the time-constant switch, S5, the sideband-aggregate agc voltage is fed to the top wafer of S2 and then to S4. Switch, S4, permits agc voltage to be applied to either one or to both receivers. From S4, agc voltage goes to the same stages in the receivers regardless of the type of agc selected.

From S4, agc voltage is fed to the IF amplifier of each receiver in this unit and to J620. From J620, agc voltage is fed to the grids of rf amplifiers by way of P402 in the High-Frequency Captive Oscillator Unit, Fig. 7-5. In Fig. 7-5, it can be seen that agc for both receivers enter at P402 and leaves for unit #300 at J409 and for unit #200 at J408. In units #200 and #300, agc voltage is fed to all rf amplifiers.

10.2.2 Carrier-only agc. In Fig. 7-9, the carrier signal is amplified in several stages and then rectified by CR1 and CR2 before going to J707. The rectified carrier signal is then fed to switch, S2, in unit #600 through P602. This source of agc voltage can be chosen by S2 and its agc voltage fed to the agc diode V9 and to S4. AGC can be chosen for either one or both receivers by S4 and then fed to the rf and first IF amplifiers, as explained before.

10.2.3 Carrier-and-sideband agc. When S2 is placed in the carrier-and-sideband position, the carrier-only agc from P602 is combined with a sideband-only agc which originates in the diversity phone combiner and enters unit #600 through P603. We have not discussed the diversity phone combiner in detail; however, as previously explained this unit selects the autio output of the receiver having the stronger signal. A portion of this selected signal is fed to two diode-rectifier circuits before entering unit #600. These two agc voltages, carrier and sideband, are both applied to the agc diode V9.

10.2.4 External agc. The source of signal for external agc enters unit #600 at P604. This external source may be a test signal for adjusting the agc system.

10.2.5 Manual gain control. Manual gain control R119 in unit #600 permits bias adjustments for maintenance purposes.

This ends our discussion of the SSB-R3 receiver. We have attempted to give a general explanation of its operation. Only a few of the circuits have been discussed in detail. At this time, it is difficult to know which circuits need additional explanation. If additional explanation of circuit theory is desired please let us know. Your comments and questions will influence the contents of future printings.

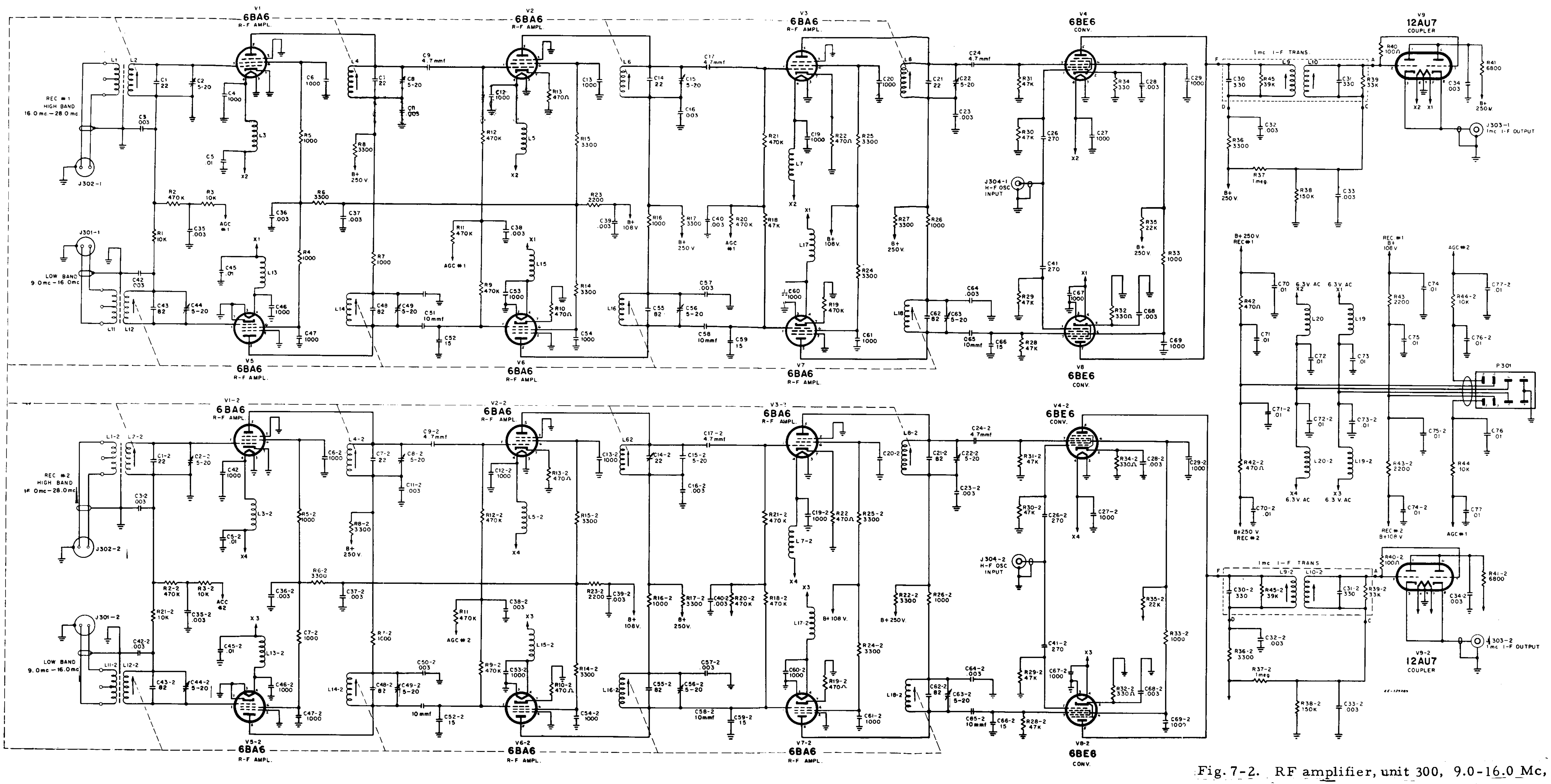
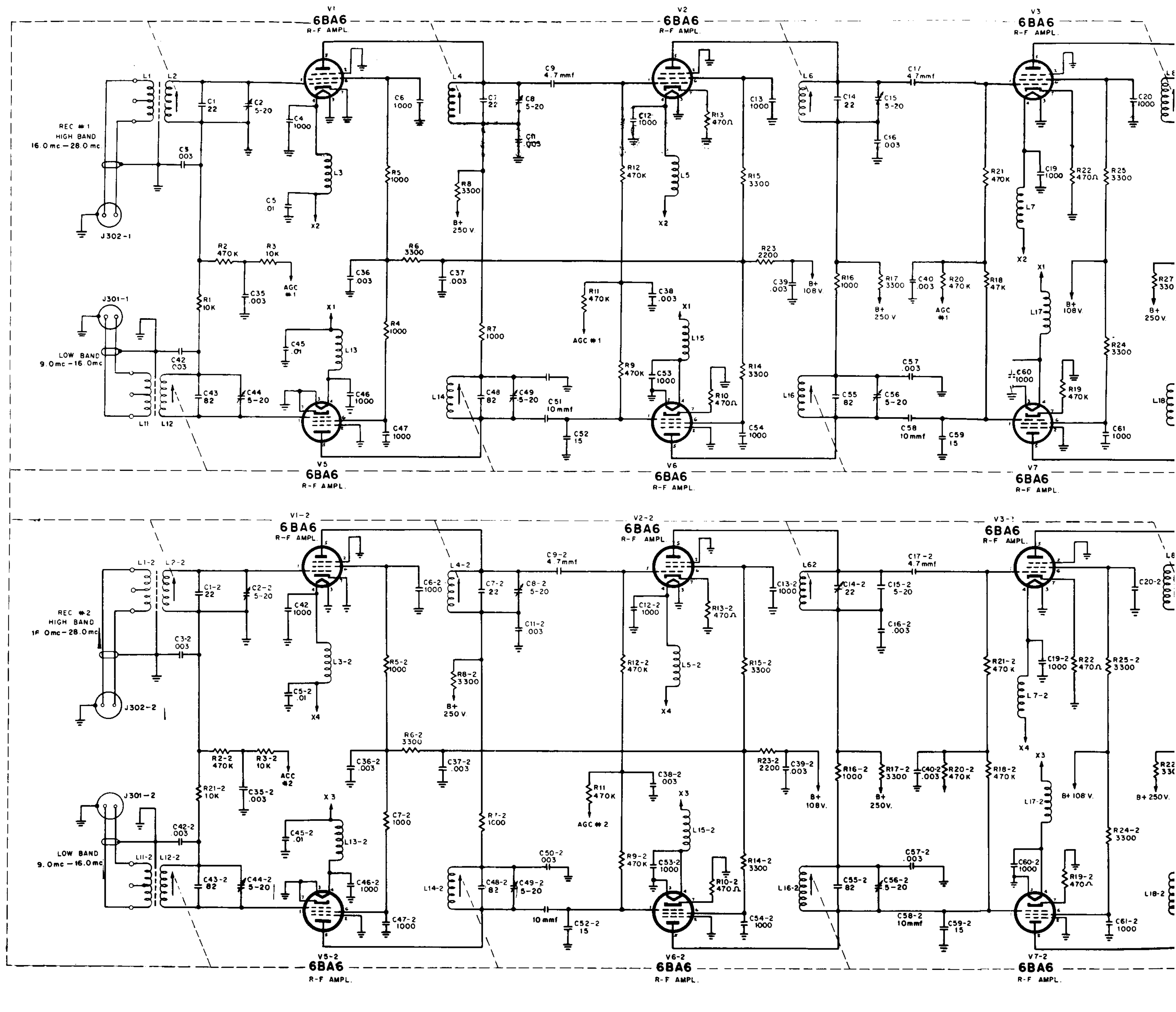



Fig. 7-2. RF amplifier, unit 300, 9.0-16.0 Mc, 16.0-28.0 Mc, schematic diagram.

(Courtesy of Radio Corporation of America, Radiomarine Products.)

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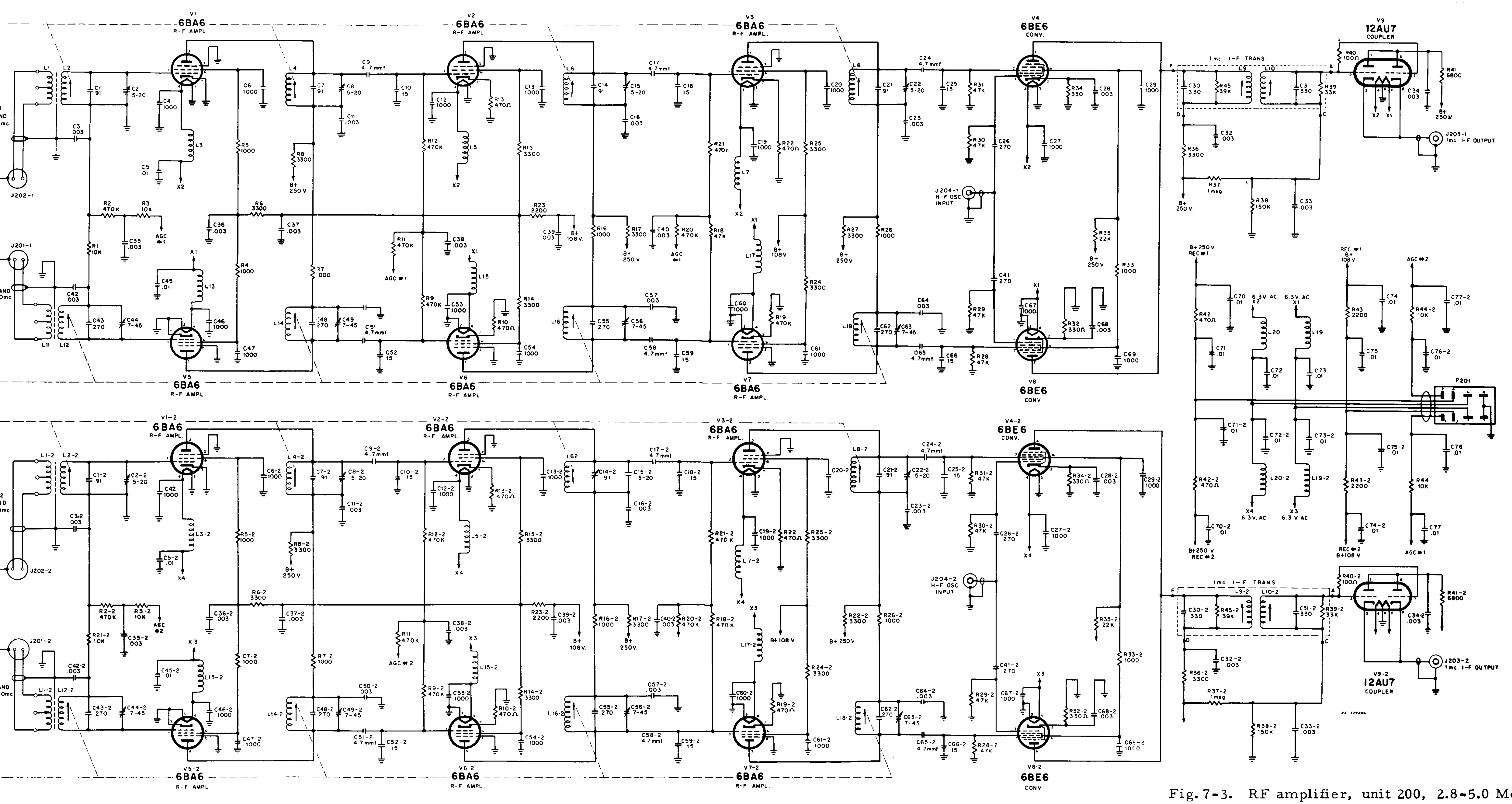


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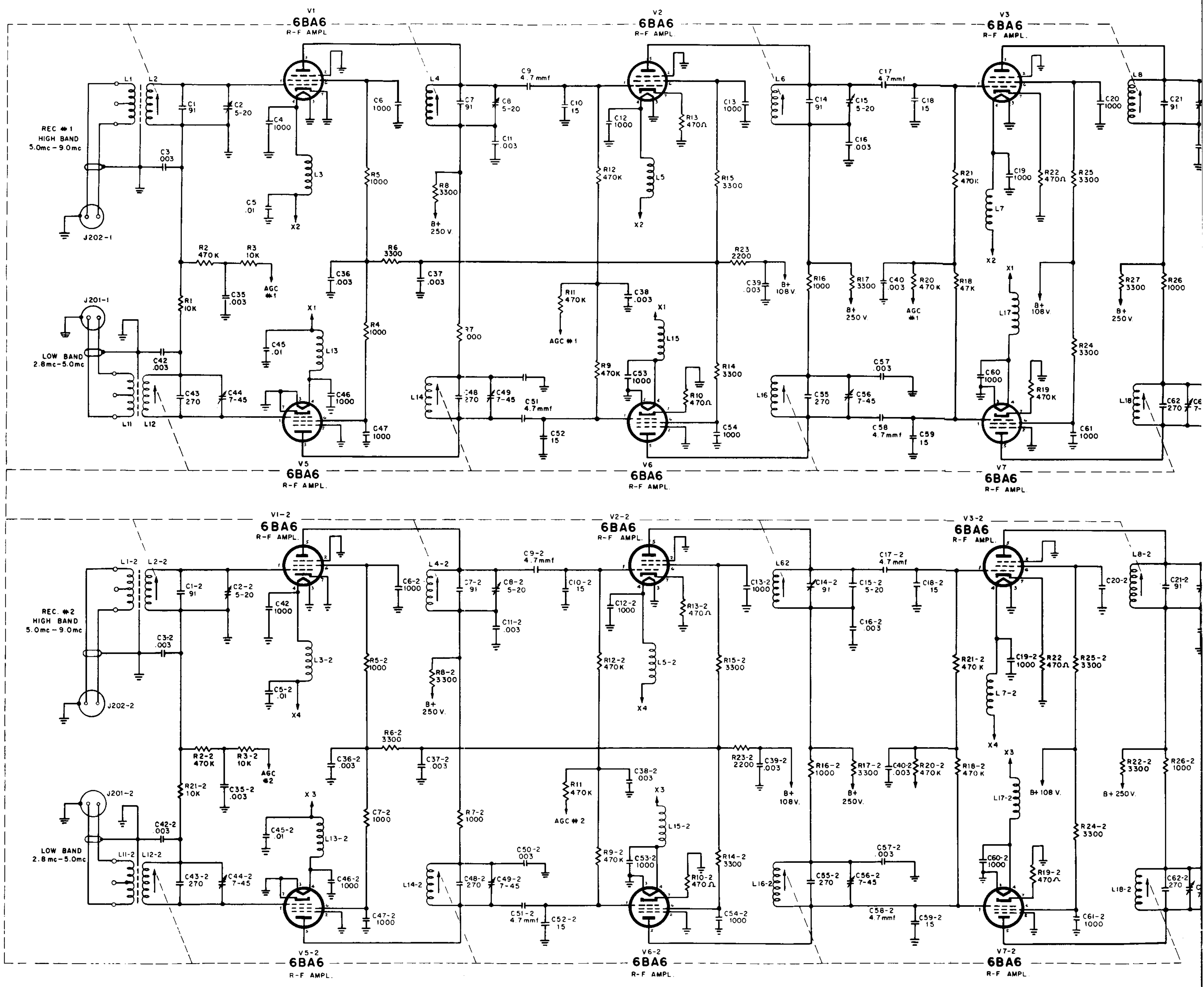
Fig. 7-3. RF amplifier, unit 200, 2.8-5.0 Mc, 5.0-9.0 Mc, schematic diagram.

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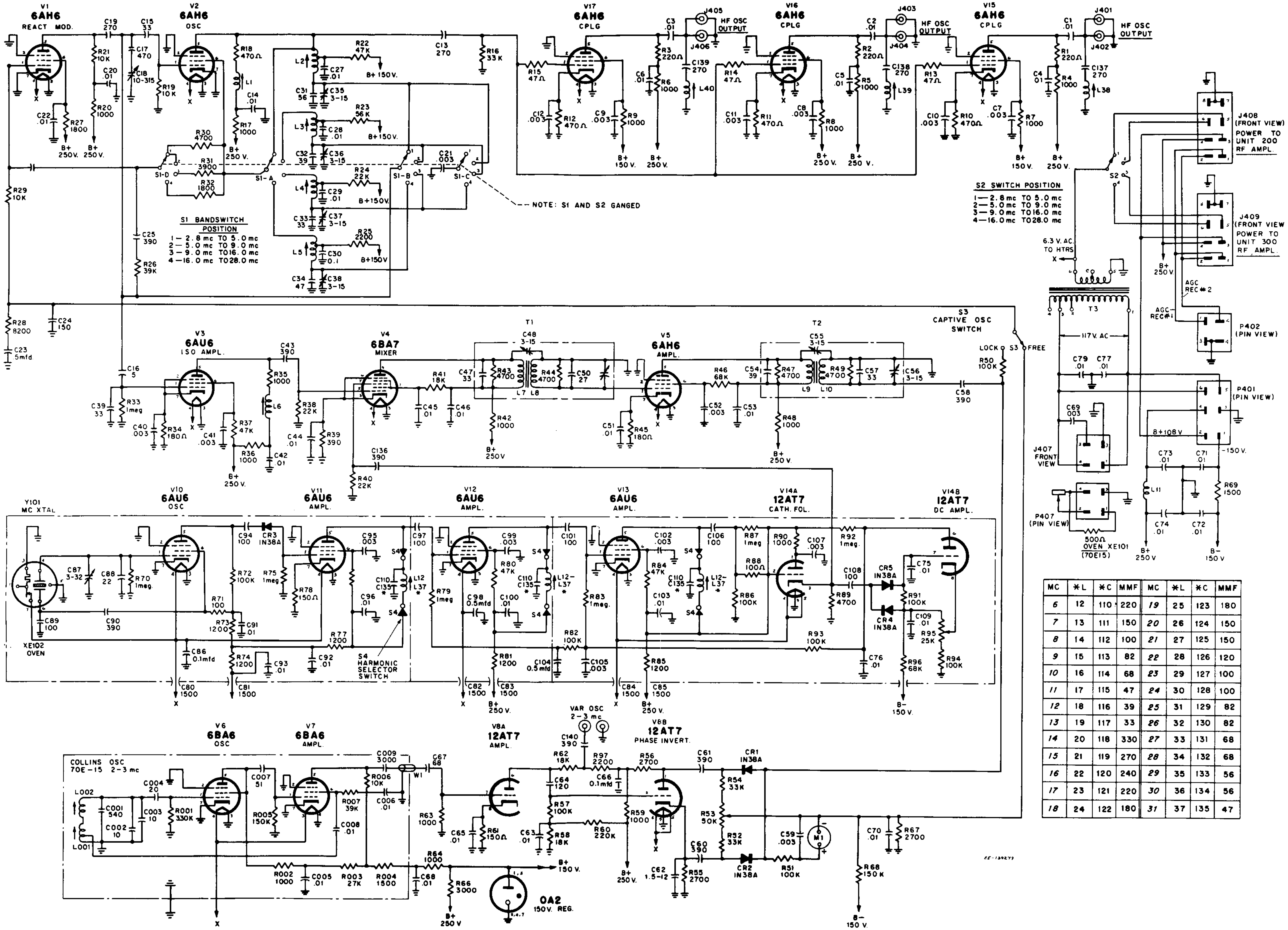
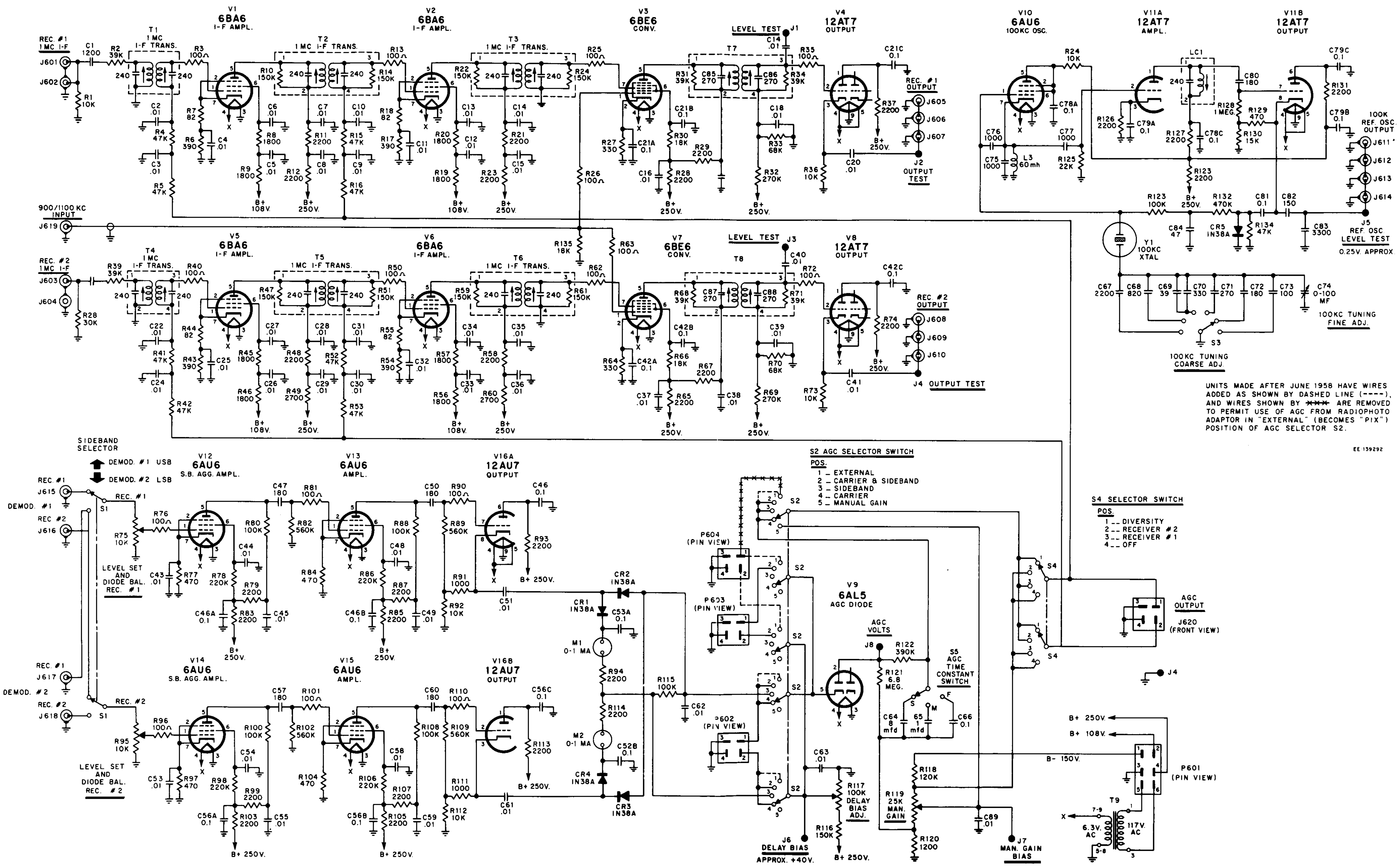



| MC | *L | *C | MMF | MC | *L | *C | MMF |
|---|---|---|---|---|---|---|---|
| 6 | 12 | 110 | 220 | 19 | 25 | 123 | 180 |
| 7 | 13 | 111 | 150 | 20 | 26 | 124 | 150 |
| 8 | 14 | 112 | 100 | 21 | 27 | 125 | 150 |
| 9 | 15 | 113 | 82 | 22 | 28 | 126 | 120 |
| 10 | 16 | 114 | 68 | 23 | 29 | 127 | 100 |
| 11 | 17 | 115 | 47 | 24 | 30 | 128 | 100 |
| 12 | 18 | 116 | 39 | 25 | 31 | 129 | 82 |
| 13 | 19 | 117 | 33 | 26 | 32 | 130 | 82 |
| 14 | 20 | 118 | 330 | 27 | 33 | 131 | 68 |
| 15 | 21 | 119 | 270 | 28 | 34 | 132 | 68 |
| 16 | 22 | 120 | 240 | 29 | 35 | 133 | 56 |
| 17 | 23 | 121 | 220 | 30 | 36 | 134 | 56 |
| 18 | 24 | 122 | 180 | 31 | 37 | 135 | 47 |

(Courtesy of Radio Corporation of America, Radiomarine Products.)

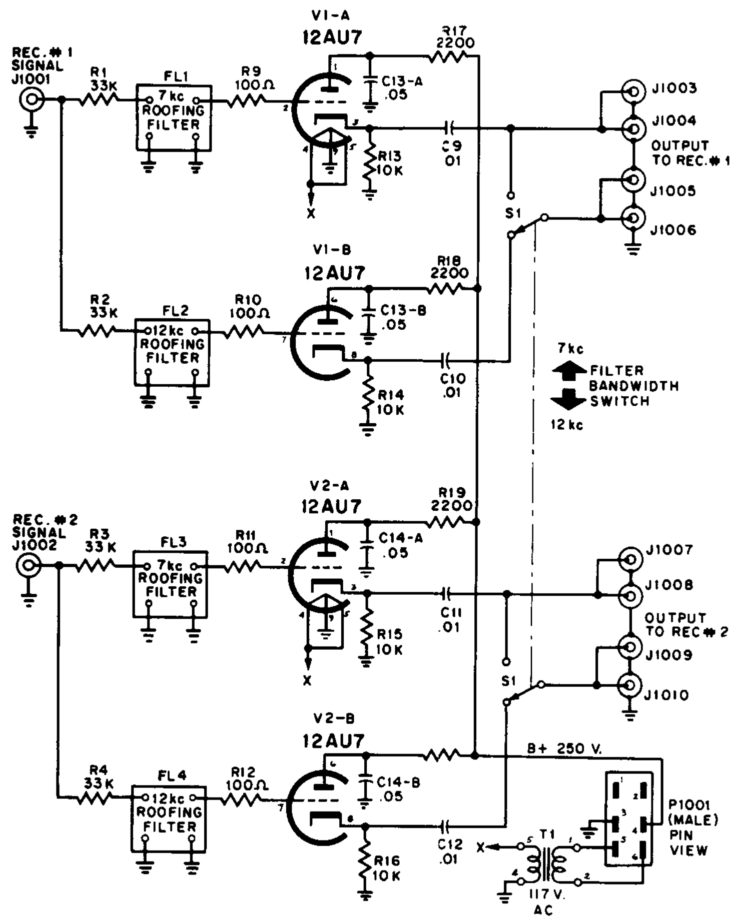
Fig. 7-4. HF captive oscillator, schematic diagram, unit 400.




(Courtesy of Radio Corporation of America, Radiomarine Products.)

Fig. 7-5. 1 Mc dual IF, 100 Kc oscillator, AGC selector, unit 600, schematic diagram.

(Courtesy of Radio Corporation of America, Radiomarine Products.)




*NOTE:* UNIT 1000A (MI22707) CIRCUIT IDENTICAL EXCEPT FOR FILTERS. FL1 AND FL3 SUPPLIED FOR 3 kc. BW. FL2 AND FL4 OPTIONAL FOR OTHER BANDWIDTHS.

Fig. 7-6. Roofing filter, unit 1000, telegraph filter, unit 1000A, schematic diagram.

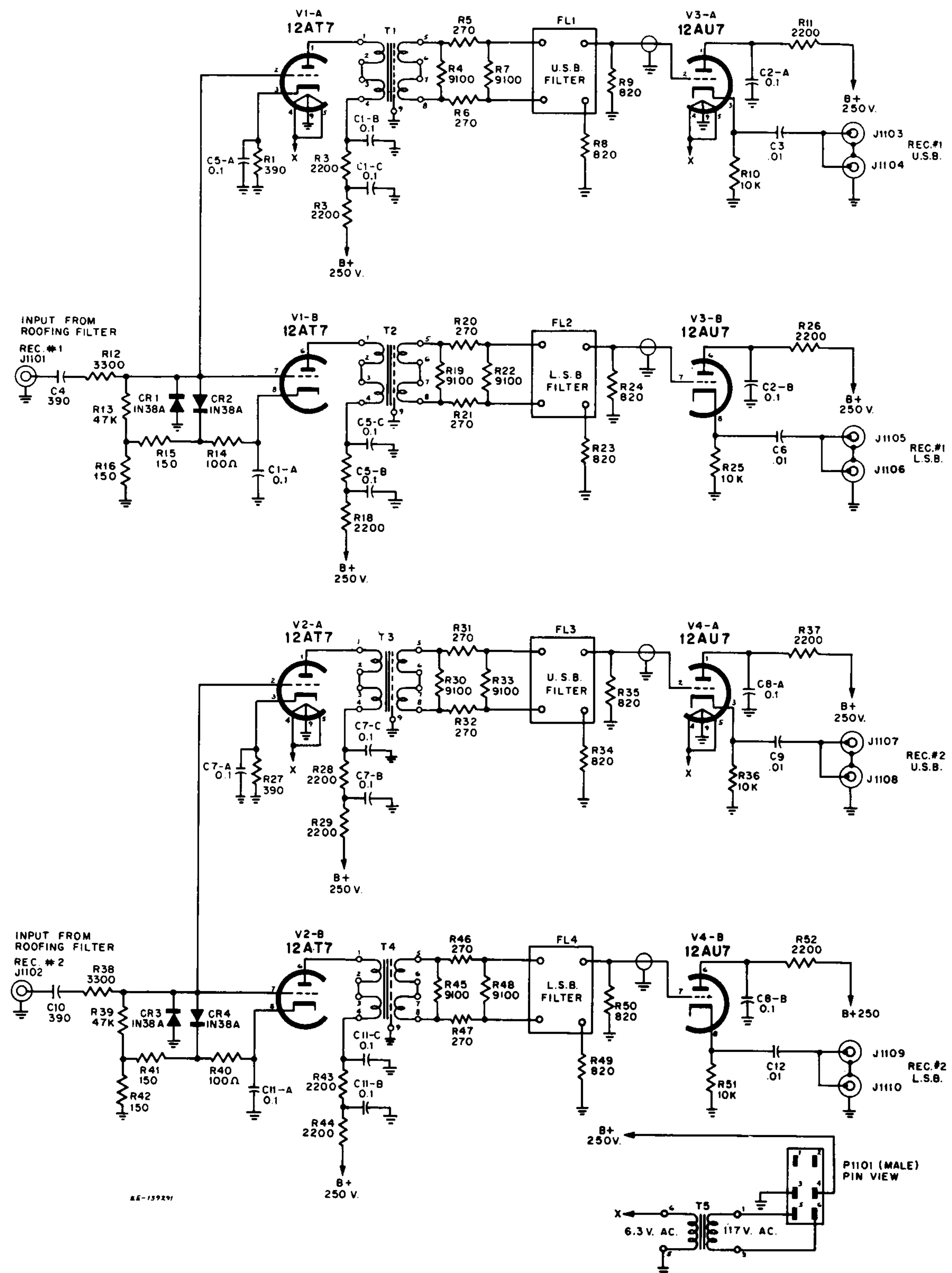


Fig. 7-7. Sideband filter, unit 1100, schematic diagram.

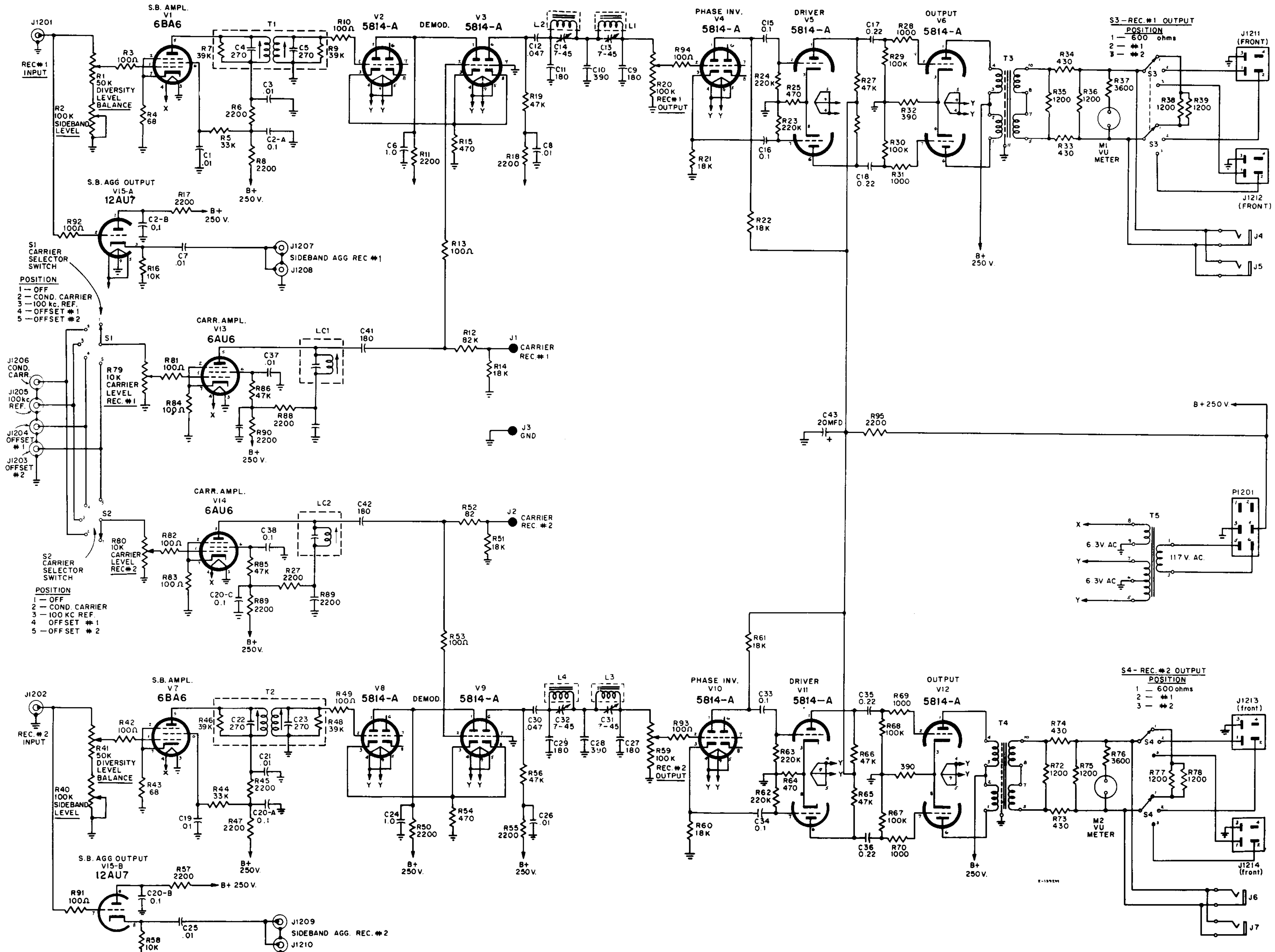
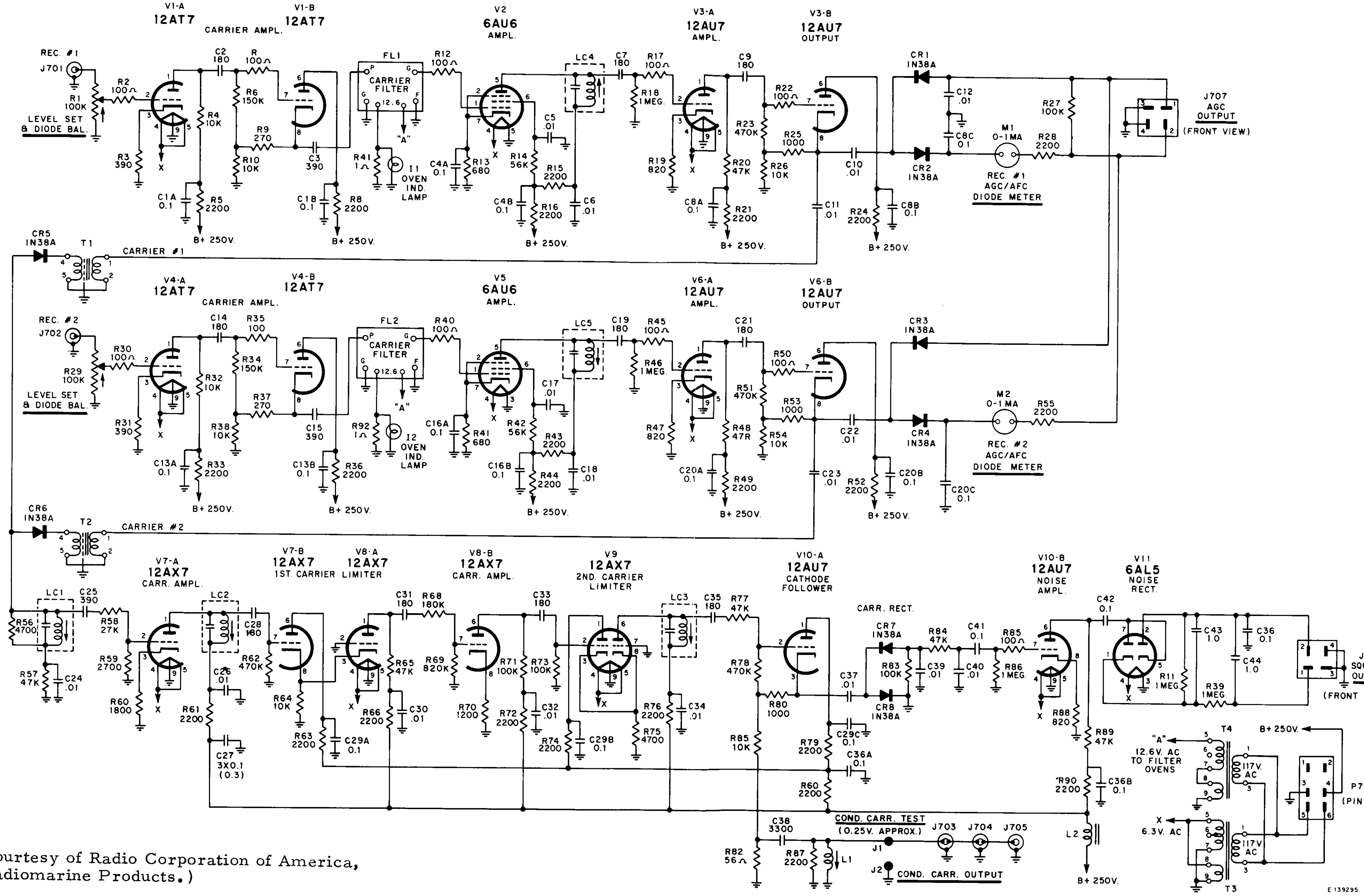


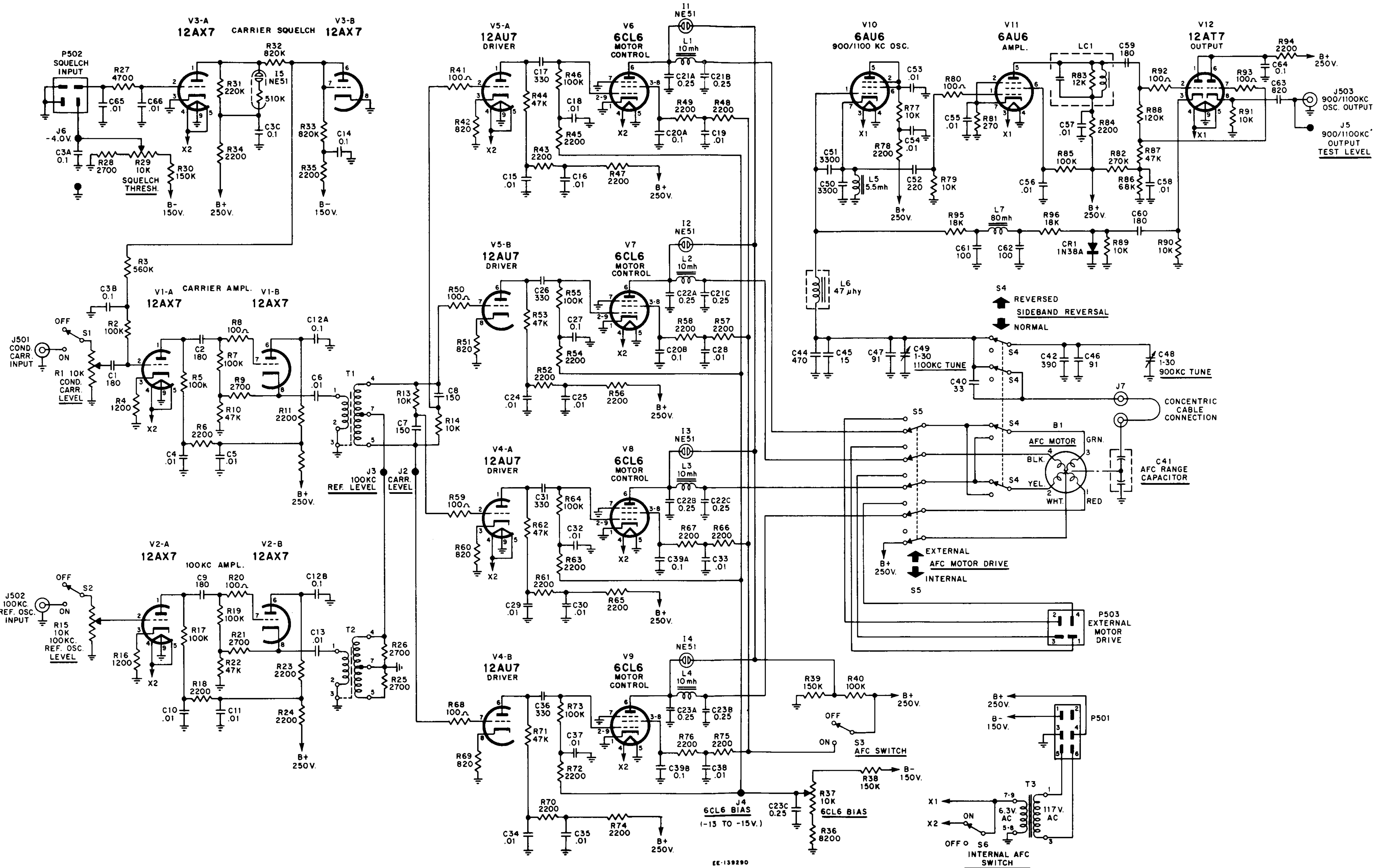

Fig. 7-8. Demodulator, audio amplifier, unit 1200, schematic diagram.

(Courtesy of Radio Corporation of America, Radiomarine Products.)



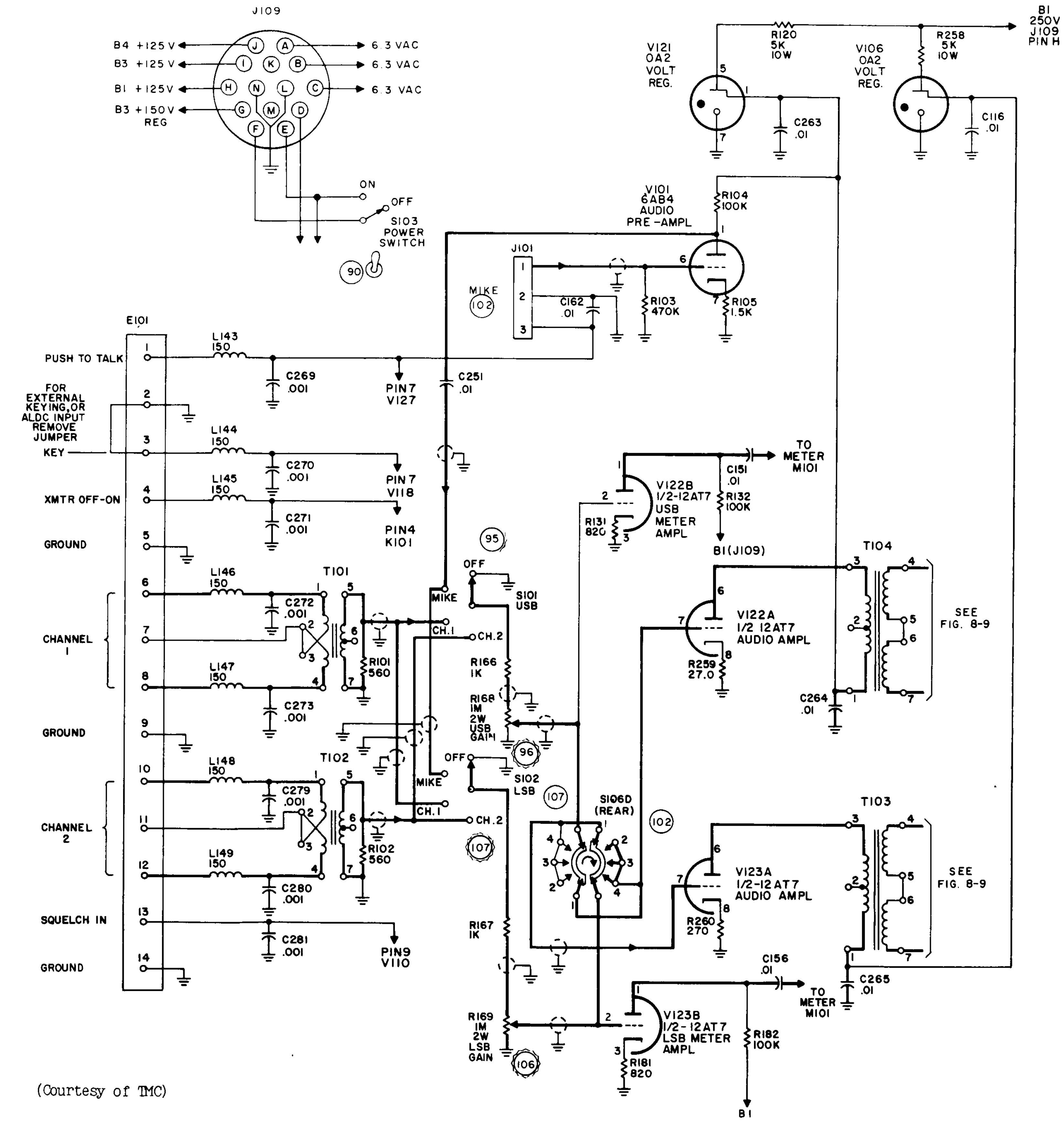

(Courtesy of Radio Corporation of America, Radiomarine Products.)

Fig. 7-9. Carrier conditioner, unit 700, schematic diagram.



(Courtesy of Radio Corporation of America, Radiomarine Products.)

Fig. 7-10. 900/1100 Kc oscillator, AFC, unit 500, schematic diagram.




(Courtesy of TMC)

Fig. 8-7. Schematic diagram, SBE-3, audio input section.

ATTENTION

Examination 7 is to be worked at this time. This exam covers the material in Chap. 7. It is included with the material.

CHAPTER 8

# THE AN/FRT-39 SINGLE-SIDEBAND TRANSMITTER

## 1.0 INTRODUCTION

The AN/FRT-39 transmitter provides a 10,000-watt peak envelope power (PEP) output throughout the 2-28 mc range. On long-range circuits, where additional power is needed, the AN/FRT-39 is used as a driver for a 40-kw PEP final. This arrangement is designated as the AN/FRT-40. Unless otherwise specified, our discussion will pertain to the AN/FRT-39. Both arrangements, the AN/FRT-39 and -40, are capable of the following types of transmission:

a. CW (keyed carrier)

b. Frequency-Shift Telegraphy

c. Single-Sideband Suppressed Carrier

d. Double-Sideband Suppressed Carrier

e. Independent Sideband With Reduced Carrier

f. Single or Double Sideband With Carrier

We will be concerned primarily with independent sideband reduced carrier operation because this is utilized by the FAA.

The AN/FRT-39 transmitter utilizes the filter method of SSB generation. As a review, we will discuss briefly the functions performed in a basic transmitter utilizing the filter method.

Refer to the block diagram of Fig. 8-1. The audio intelligence to be transmitted is fed through audio input circuits and then applied to the balanced modulator. The carrier oscillator signal is also fed to the balanced modulator. The output of the balanced modulator consists of the sum and difference frequencies of its two inputs. This output is fed to an upper-sideband filter where the lower sideband is eliminated.

The upper-sideband signal from the filter is fed to a mixer stage where it is raised to the operating frequency. If only one heterodyning stage is used (as shown in the figure), the upper-sideband signal from the filter is converted to a lower-sideband signal as the result of taking the difference of the mixing process, and a lower-sideband signal is transmitted. A second heterodyning stage would convert the lower-sideband output of the first heterodyning stage to an upper-sideband signal and this upper-sideband signal would be transmitted.

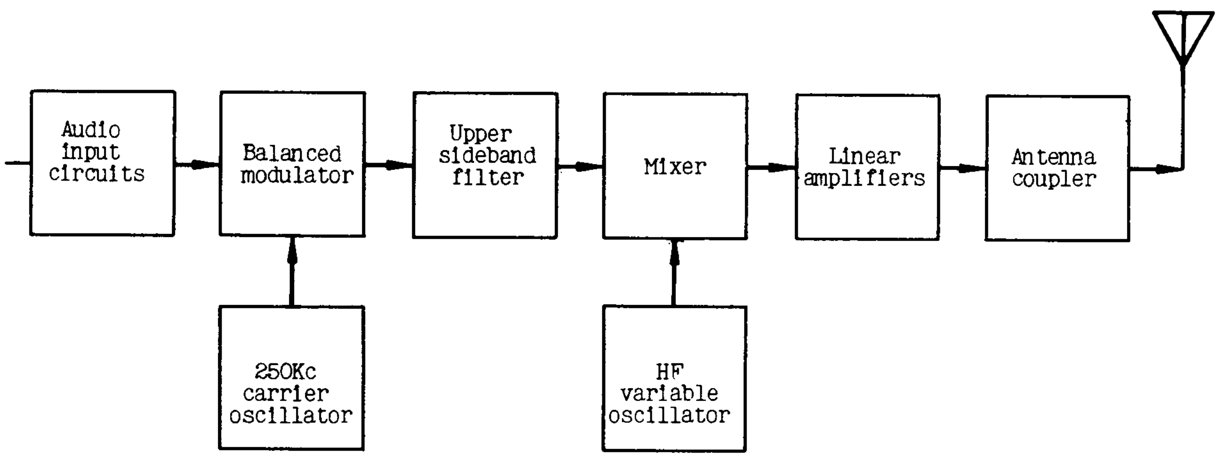



Fig. 8-1. Block diagram of a basic single-sideband transmitter utilizing the filter method of single-sideband generation.

The heterodyning stages in transmitters are actually modulation stages that utilize balanced modulators. Both sidebands are generated in each modulation stage as they are in the first balanced modulator when the audio is raised to the 250-kc region. It is necessary to utilize a precise sideband filter after the first balanced modulator because the sideband separation to carrier frequency ratio is low in that stage. The sideband separation is much greater in the second and third balanced modulators, therefore, tuned circuits easily eliminate the undesired sidebands. This concept was explained in detail in Chap. 4.

A unit block diagram of the AN/FRT-39 is shown in Fig. 8-2. A comparison of Figs. 8-1 and 8-2 indicates that the sideband exciter unit contains the audio input circuits, the balanced modulator, the carrier oscillator, the sideband filter, and the oscillator and mixer stages for heterodyning the sideband signal to the operating frequency.

Actually, there are two audio input channels; two balanced modulator circuits, and two sideband filters in the sideband exciter unit to generate two independent sidebands. After the sidebands are generated separately they are combined and then heterodyned to the operating frequency. Thus, the function of the sideband exciter is to amplify two audio channels, to

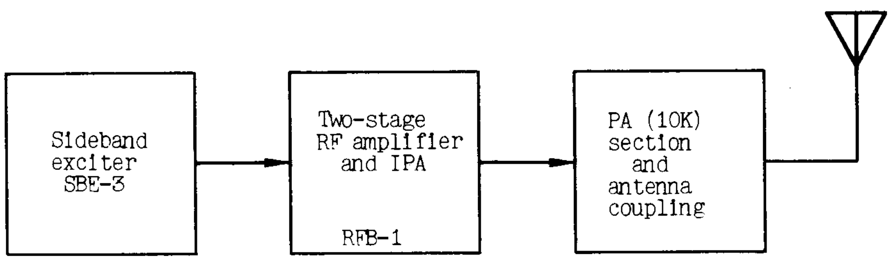



Fig. 8-2. Block diagram representing the main units of the AN/FRT-39 single-sideband transmitter.

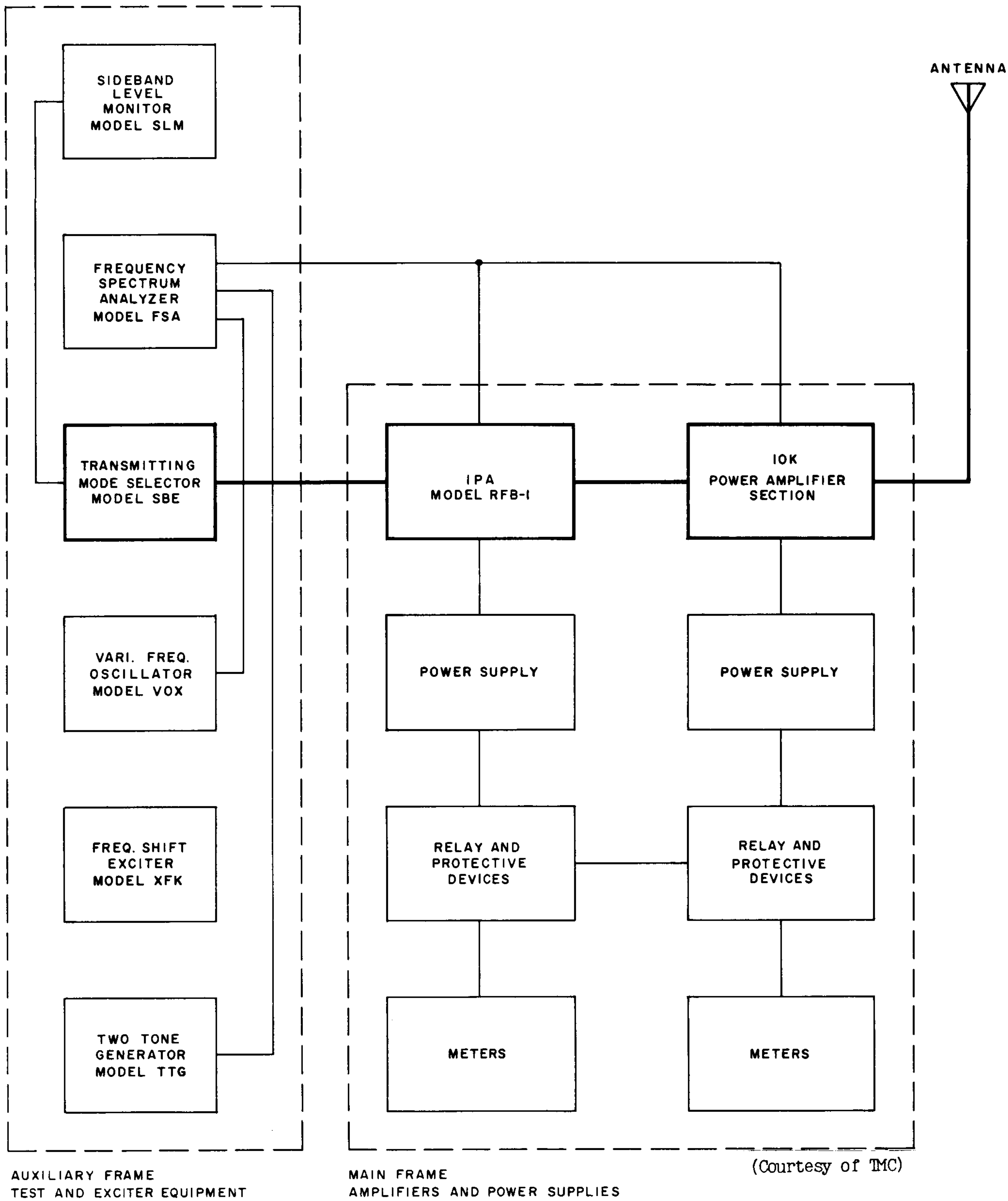
convert the separate audio into low-level rf single-sideband signals, and to apply them to linear amplifiers.

The next block in the figure represents two rf amplifier stages and the intermediate power amplifier stage. The IPA drives the 10-kw final represented by the last block.

Figure 8-2 represents the portion of the transmitter that generates and amplifies the SSB signal. It does not include blocks for the test equipment, power supplies, protective devices, and meter circuits. Figure 8-3 is a more complete diagram of the transmitter showing these additional blocks. The blocks are enclosed in dashed lines that represent the cabinets containing the units of the transmitter system. The sideband exciter and auxiliary equipment are in the left frame of the AN/FRT-39 as shown in Fig. 8-4. The frame on the right contains the two stages of linear amplifiers; the IPA, 10-kw final, and their power, metering, and protective circuits.

This chapter deals primarily with the sideband exciter unit. The linear amplifiers are discussed only briefly. The auxiliary equipment and power supply circuits are not discussed.

When the 40-kw final is used, two additional frames are required for the amplifier and its associated power, metering, and protective circuits. These two frames are located on the right in the picture of the AN/FRT-40 system shown in Fig. 8-5.




(Courtesy of TMC)

Fig. 8-3. Block diagram of AN/FRT-39 transmitter.

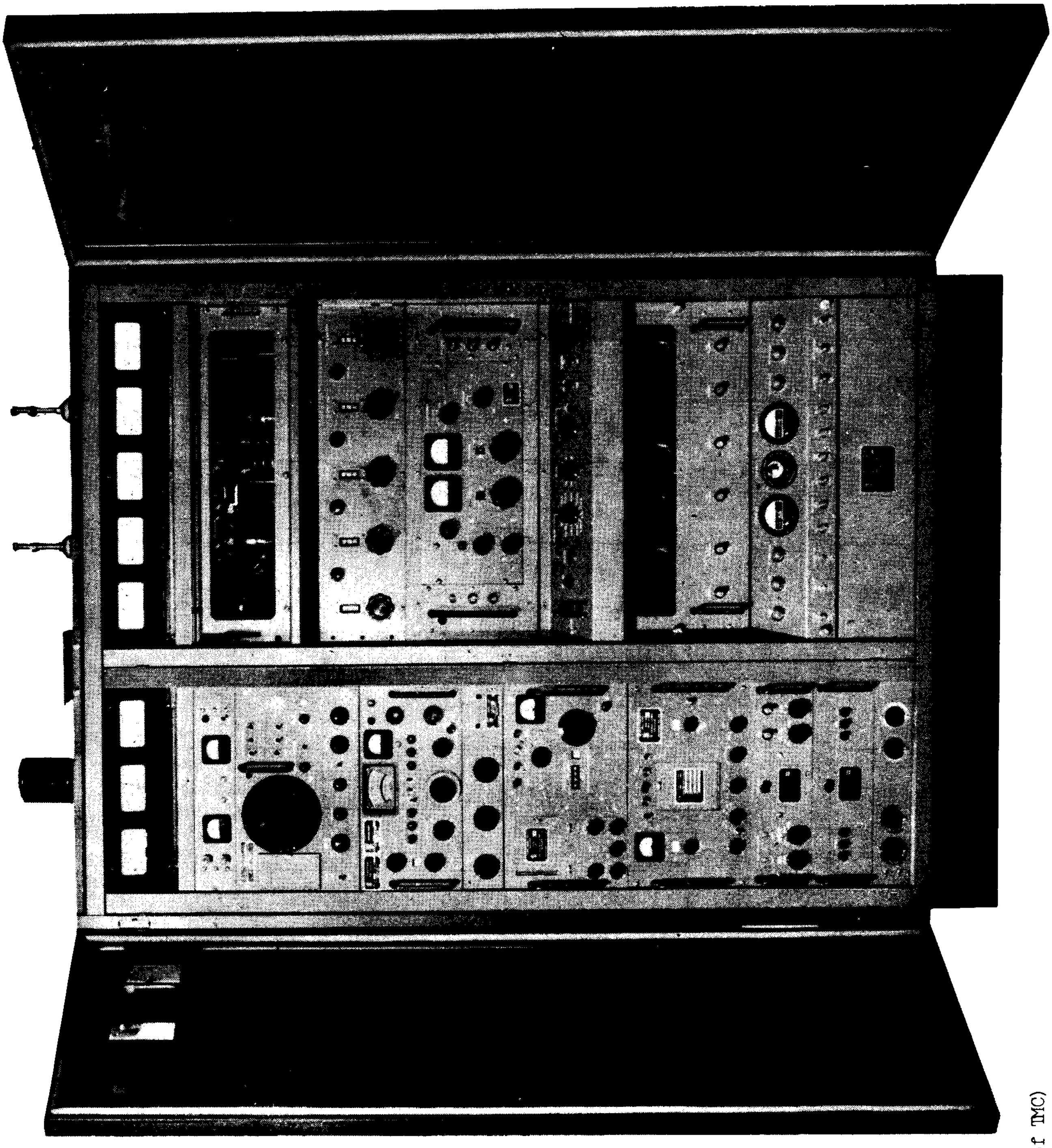
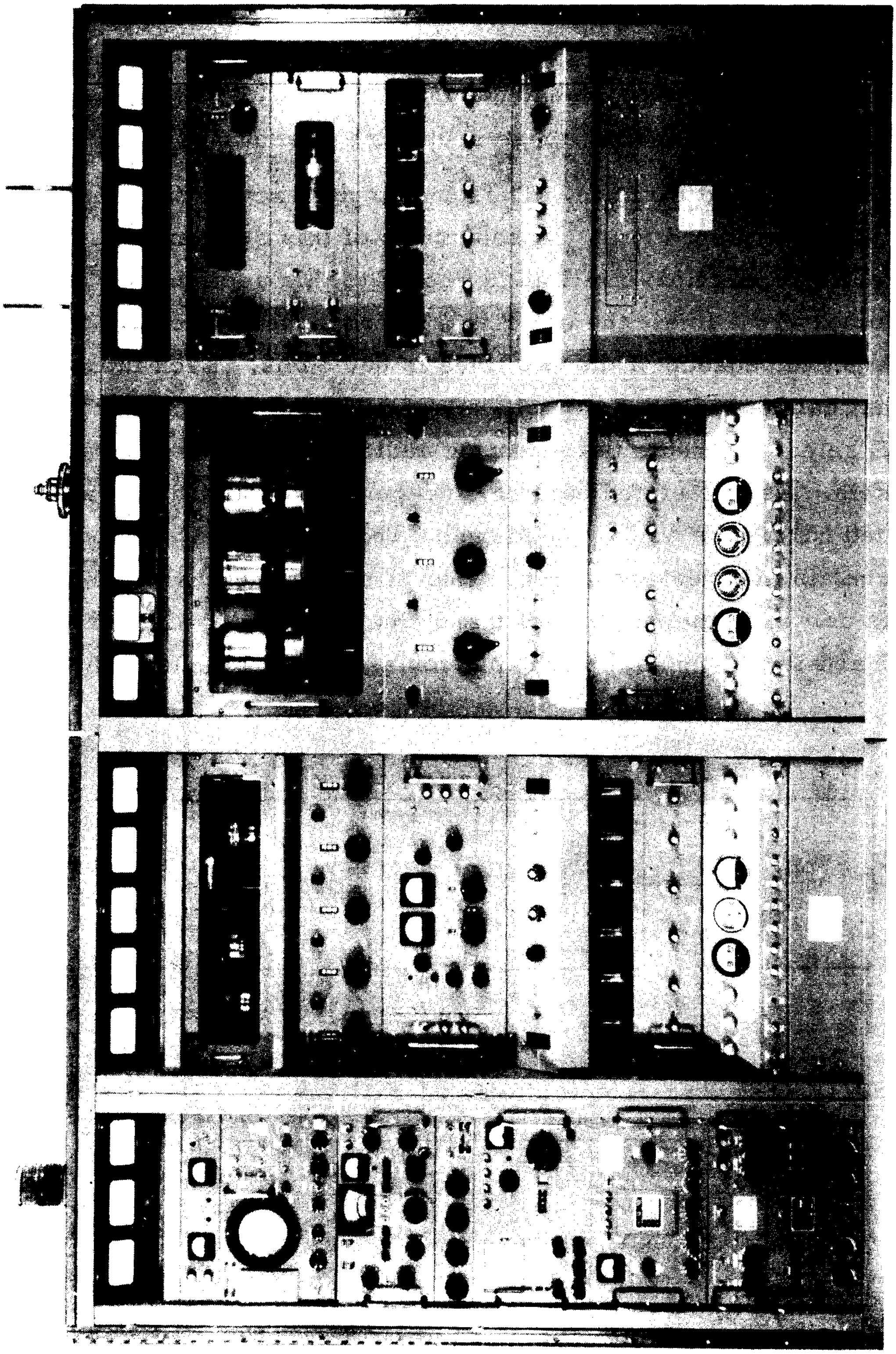


Fig. 8-4. Front view, Model AN/FRT-39, 10 kilowatts.

(Courtesy of TMC)



(Courtesy of TMC)

Fig. 8-5. Front view, Model AN/FRT-40, 40 kilowatts.

All schematic diagrams are included at the end of this chapter.

## 2.0 CIRCUIT CONSIDERATIONS

### 2.1 Sideband Exciter

The sideband exciter (designated SBE-3) has five principle sections of interest in our discussion; these sections are shown in Fig. 8-6.

2.1.1 Audio input circuits. The audio input circuits of the sideband exciter are considered first. A schematic of this section is shown in Fig. 8-7. There are connections for 3 audio inputs to this section. Two connections are for external lines; the other is a microphone jack on the front panel. The input impedance of the two lines is 600 ohms. Each input has a center tap that can be grounded for a balanced input. Switches S101, S102, and S106D select and route the incoming audio to the upper or lower sideband channels. As was stated earlier, independent sideband operation with reduced carrier is being considered. This means that the lower sideband will contain different intelligence than the upper sideband. The line inputs to this unit are labeled channel 1 and channel 2 and will be referred to as line 1 and line 2 inputs. Each line contains separate audio that is intended

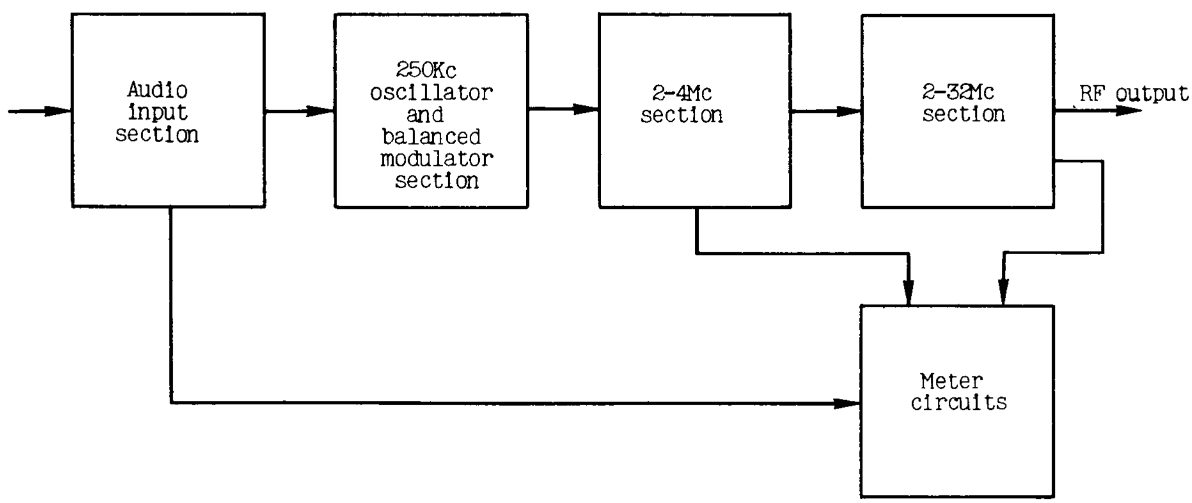



Fig. 8-6. Block diagram representing the five principal sections of the sideband exciter unit (SBE-3).

for separate sidebands. The SBE-3 unit amplifies the two audio signals separately; two separate balanced modulator circuits and two sideband filters (upper and lower) are utilized to obtain independent sideband operation.

Figure 8-8 shows the front panel of the SBE-3 unit with all components numbered. This figure will help to relate the front panel controls to their corresponding schematic components.

The function of lower- and upper-sideband switches S101 and S102 can be seen on the schematic of Fig. 8-7. For the upper sideband, S101 can select inputs from line 1, from line 2, or from the microphone. The same is true of S102 for the lower sideband.

R168 and R169 are the sideband gain controls. The center arms are connected to S106D for inversion of the upper and lower sideband inputs when operating in the 2 - 4.25 mc range (to be explained later). Audio is also taken from the center arms to feed meter amplifiers V122B and V123B. The outputs of these amplifiers are rectified and fed to a VTVM circuit. The metering circuit indicates the level of incoming signals so that overloading of the following balanced modulator section can be avoided.

It should be noted that R168 always controls the gain of the audio that eventually is transmitted as the upper sideband and R169 controls the gain of the audio for the lower sideband. The voltage at the center arm of R168 goes to either V122A or V123A, depending on the setting of S106D. The setting of S106D depends on whether the frequency of operation is below or above 4.25 mc. The frequency of operation determines the number of conversions required as will be explained in later sections.

2.1.2 <u>250-kc oscillator and balanced modulator section.</u> This section was discussed briefly in Chap. 3 as an application of diode balanced modulators. Figure 8-9 is a schematic of the section. Some of the components shown in the output of Fig. 8-7 are repeated at the left of Fig. 8-9 to provide a smooth transition from one schematic to the next. The same procedure is followed in the other schematic diagrams of the SBE unit.

The balanced modulator containing CR115 has two inputs; the audio from V122A and the carrier from the 250-kc crystal oscillator shown at the top of the schematic. The output of T125 consists of both sidebands centered around 250 kc. Both sidebands from T125 are amplified in V124 and applied to lower-sideband filter Z110 where the upper sideband is eliminated.

The circuitry for the other balanced modulator circuit is essentially the same. Both sidebands, centered around 250 kc are present at T126. They are amplified by V125. Z111 eliminates the lower sideband.

The lower sideband, which contains modulation from one line input, and the upper sideband which contains modulation from the other line input, are both available at Z112. Z112 has high attenuation at 250 kc to reduce carrier residue from the balanced modulators. The sidebands are attenuated only slightly by filter Z112. V126 amplifies the independent sidebands and applies the sideband signals to the 2 - 4 mc section. The residue carrier from both balanced modulators was eliminated by Z112 so that a controlled amount of carrier may be reinserted and transmitted along with the sidebands. R263 (carrier insert) selects any degree of carrier insertion from -55 db to full output of the exciter unit.

2.1.3 2-4 mc section. V126 and R263 are shown again in Fig. 8-10 as the input to the 2-4 mc section. The function of this section is to raise the sidebands to the 2-4.25 mc range by a heterodyning process. When the operating frequency is between 2 and 4.25 mc, this section serves as the final heterodyning stage. When the operating frequency is above 4.25 mc, the 2-32 mc section provides the second frequency conversion to raise the sidebands to the operating frequency.

The sidebands from V126 are applied to the vacuum-tube balanced-modulator circuit containing T127 and V113. The source of injection signal is either a variable master oscillator (contained in a separate unit) or the crystal controlled oscillator, V115B. When the VMO is used, its signal enters at J104. When only a few different operating frequencies are utilized there is little need for a continuous tuning VMO, and the proper choice of crystals in oscillator V115 provides for the necessary frequency shift from day to night operation, etc.

The injection frequency is 250 kc higher than the desired output of this section. For example, when a 3-mc output is desired, the injection frequency is 3, 250 kc and the difference between the two signals provides the 3-mc output. The frequency of the injection crystal oscillator signal is determined by MF XTAL switch S107. The MF tuning control, designated C167 on Fig. 8-8, is ganged to C167A and C167B to tune the circuit containing V114 to the output frequency of this section. The MF tuning control is also ganged to one scale of the indicator which is designated 2-Position Indicator on the figure. Although the MF tuning control tunes C167A and C167B to the output frequency, the dial that it is ganged to, indicates the oscillator injection frequency.

2.1.4 <u>2-32 mc section.</u> This section of the sideband exciter provides for output frequencies from 4.25 mc to 32.25 mc. A schematic is shown in Fig. 8-11. The amplified output of the previous section is fed to diode balanced-modulator circuit Z107. The high-frequency oscillator signal is fed to the balanced modulator through J110 from oscillator amplifier V116. The difference frequency output of the balanced modulator is at the final operating frequency and is amplified by the cascaded amplifier stages V118, V119, and V120 which raises the single-sideband signals to the rated one-watt PEP output of the sideband exciter. The tuning capacitors of the three amplifier stages are ganged together. Band switching is accomplished in the amplifier circuits by S106A, S106B, and S106C to continuously cover the operating range of 2 to 32 mc. It may be remembered that the D section of S106 inverts the upper and lower sideband inputs in the audio input section when the SBE-3 is operating below 4.25 mc.

Crystal-controlled oscillator V117 supplies injection frequencies for balanced modulator Z107 from 8 to 34 mc in 2-mc steps. Oscillator frequencies are selected by band MCS switches S108A and S108B. The choosing of crystals to obtain the desired operating frequencies will be explained in a later section.

A small portion of the output voltage from the sideband exciter is applied across voltage divider R210 and R211 where it is fed to CR114 via C176. CR114 rectifies the signal and produces a dc voltage proportional to the envelope peaks. The output is indicated on M101 when meter switch S109 is in the rf out position.

2.1.5 Metering section. The block diagram of Fig. 8-6 shows inputs to the metering section from 3 other sections of the SBE unit. A schematic of the metering section is shown in Fig. 8-12. Inputs from the audio input section are taken from the center taps of resistors R168 and R169 and amplified by V122B or V123B. R168 is the upper-sideband gain control, and the signal at its center tap is amplified by V122B and applied to the USB terminal of meter switch S109. The signal at the center tap of R169 is amplified by V123B and applied to the LSB terminal of S109. S109 can select, for the meter, the audio of either sideband. The separate audio signals can be compared by switching S109 between the USB and LSB positions.

Samples of the signals from the 2 - 4 mc and 2 - 32 mc sections are fed to the meter circuit for tuning purposes. The signal from the 2 - 32 mc section enables the meter circuit to give an indication of the output of the SBE unit.

2.1.6 Crystal frequencies. A few examples of choosing crystals for desired operating frequencies is given now to aid in a general understanding of the sideband exciter unit.

The first consideration is the first modulator stage. The audio entering the transmitter is raised to the 250-kc region in the first modulation process. Because the sideband filters must be precise, they cannot be tunable and are therefore designed to operate at a set frequency of 250 kc. Therefore, the carrier injection frequency in the first modulator is 250 kc regardless of the operating frequency.

When the operating frequency ($f_0$) is below 4.25 mc, the crystal frequency ($f_x$) of the oscillator in the 2-4 mc section is determined by the formula:

$$f_x = f_0 + 0.25 \text{ mc}$$

For example, for an $f_0$ of 2.00 mc, $f_x$ = 2.25 mc. For these low operating frequencies no frequency conversion takes place in the 2-32 mc section. The oscillator in that section injects an 18-mc signal to provide bias for the diodes of the balanced modulator to prevent intermodulation distortion.

For an $f_0$ above 4.25 mc, the output of the second modulator is modulated by the 3rd. modulator; thus, the operating frequency is expressed by the formula:

$$f_0 = f_{x3} - (f_{x2} - 0.25 \text{ mc})$$

where $f_{x2}$ and $f_{x3}$ are the crystal frequencies of the 2nd. and 3rd. modulator stages respectively.

Frequencies of crystals in the 2nd. modulation stage are between 2 and 4.5 mc, and the crystal frequencies in the 3rd. modulator range from 8 to 34 mc in 2-mc steps.

A table of a few operating frequencies with corresponding crystal frequencies in the 2nd. and 3rd. modulators is given below (all frequencies are in megacycles):

| $f_0$ | $f_{x2}$ | $f_{x3}$ |
|---|---|---|
| 2.25 | 2.50 | 18 |
| 4.00 | 4.25 | 18 |
| 8.00 | 2.25 | 10 |
| 15.00 | 3.25 | 18 |
| 21.50 | 2.75 | 24 |
| 27.25 | 3.00 | 30 |
| 32.25 | 2.50 | 34 |

The frequency conversions for an $f_0$ of 2.25 mc are explained as follows. The sidebands are centered around 250 kc in the balanced-modulator section. The crystal frequency of 2.50 mc in the 2-4 mc section mixes with the sidebands centered around 0.250 mc to give a difference frequency of 2.25 mc, which passes to the amplifiers in the 2-32 mc section without further modulation.

For an $f_0$ of 8 mc, the crystal frequency in the 2-4 mc section is 2.25 mc, and this mixes with the 0.25-mc signal from the balanced modulator section to give a difference frequency of 2 mc. This 2-mc signal mixes with the 10-mc injection signal in the 2-32 mc section to give an operating frequency of 8 mc. Similar reasoning can be applied to arrive at the other operating frequencies.

2.1.7 Sideband inversion. Under the heading of "Sideband Inversion" it is desired to explain two concepts: the first is the inversion of the signal from lower to upper sideband, and vice versa, as a result of using the difference frequency from the heterodyning process. With this concept clearly in mind, it will be easy to explain the second concept; that is, the function of S106D in providing proper sideband inversion.

To explain the sideband inversions that occur, they will be considered under two conditions: when the operating frequency is below 4.25 mc and when it is above that frequency.

Below 4.25 mc, only one heterodyning stage is used to convert the sideband signal to the operating frequency and the inversion is explained as follows: Assume that the lower sideband from the balanced modulator section contains 1000-cycle modulation and the upper sideband from that section contains 2000-cycle modulation. The sidebands are 250 kc - 1000 cycles = 249 kc, and 250 kc + 2000 cycles = 252 kc. For an operating frequency of 3 mc, the injection frequency of the oscillator in the 2-4 mc section is 3250 kc. The difference frequencies of the sidebands and the 3250 kc injection signal are the desired output from the balanced modulator and are 3250 kc - 249 kc = 3001, and 3250 kc - 252 kc = 2998 kc. The sidebands are centered around 3 mc.

The 2-kc modulation is now contained in the lower sideband and the 1-kc modulation is in the upper sideband. No further heterodyning takes place in the 2-32 mc section.

For an operating frequency of 15 mc, the injection frequency in the 2-4 mc section is 3.250 mc as before and the sidebands are inverted in that section the same as they were in the previous example. An 18-mc injection frequency in the 2-32 mc section produces 18000 kc - 3001 kc = 14999 kc, and 18000 kc - 2998 kc = 15002 kc. Because the operating frequency is 15 mc, the 2-kc modulation is in the upper sideband and the 1-kc is in the lower sideband.

The function of S106D in Fig. 8-7 follows logically from the previous discussion. Remember that the audio at R168 is amplified by V122B and fed to the USB terminal of the meter switch. Because this same audio is metered as the USB signal, it must eventually be radiated as an upper sideband. S106D is wired to switch the audio signals at R168 and R169 to their proper amplifier stages. To explain, we will assume that S106D switches the audio at R168 to the wrong amplifier tube and we will observe the results. Next we will assume that S106D switches the audio properly.

The audio at R168 is always metered as the USB signal. Assume that S106D applies this audio to V122A. V122A feeds the balanced-modulator channel that is followed by a lower-sideband filter, and an LSB signal is generated. For an operating frequency above 4.25 mc, two sideband inversions take place, and the LSB signal changes from LSB to USB and back to LSB before being radiated. However, the audio in the radiated LSB was metered in the USB position.

It can be seen that if the operating frequency is below 4.25 mc, an LSB signal in the balanced modulator section is inverted to a USB in the one heterodyning process and is radiated as a USB. This condition is as it should be because the audio was metered in the USB position of the meter.

In conclusion, for operating frequencies below 4.25 mc, S106D must switch the audio at R168 to V122A and the audio at R169 to V123A. For operating frequencies above 4.25 mc, S106D must switch the audio at R168 to V123A and the audio at R169 to V122A.

### 2.2 RF Amplifier and IPA Unit

In the block diagram of Fig. 8-2 it is seen that the unit following the sideband exciter is the rf amplifier and IPA unit, designated RFB-1. This unit contains two Class-A amplifier stages and an intermediate power amplifier stage operated Class $AB_1$. A schematic of this unit is shown in Fig. 8-13. The RFB-1 is located on the main frame and receives its input from the sideband exciter unit which is located on the auxiliary frame.

The input signal is applied to the grid of the first Class-A amplifier stage (Fig. 8-13) via J201. The output from the plate of V201 is fed to the tuning elements selected by driver band switch S201A before going to the grid of V202. The tuning elements in the plate circuit of V201 consists of C203, C202, and L202, L209, L210, or L211. S201A selects one of the four inductors in band switching. Each inductor is slug tuned and is preset at the factory for its band of operation.

A portion of the rf signal from the 10-kw PA or from V203 is rectified and fed back to the grid of V201 via E201 for automatic load control. This control voltage limits the rf output of V201 on voltage peaks. This allows the final power amplifier to operate near its maximum power capability without being overdriven on signal peaks. The rectified control voltage is obtained from the plate circuit of V203 when it is the final amplifier. When the 10-kw amplifier is the final, the control voltage originates in its plate circuit.

The second Class-A amplifier stage is similar to the first. S201B selects the coils for band switching. Tuning in each band is accomplished by C232 and trimmer C231.

The output signal of the 2nd. amplifier is fed to the grid of V203 in the IPA circuit. The output of the IPA is fed to E203. A pi network consisting of L245, L246, C254, C269, C272, and C274 in the plate circuit of V203 is utilized for tuning and loading. S202 accomplishes band switching in this stage.

C255 is the neutralizing capacitor which is connected back to the bottom end of the grid coil via S201B. When the IPA is the final power amplifier, an automatic load control (ALC) voltage from the plate circuit of V203 is fed to E201 and then jumpered to the grid of V201. When the 10-kw final is used, the ALC voltage enters this unit at pin b of P201, and the jumper of E201 is connected as shown in Fig. 8-13.

2.3 Power Amplifier Unit

Next to be considered is the 10-kw power amplifier unit (Fig. 8-14). The PA is located in the top portion of the right frame (Fig. 8-4).

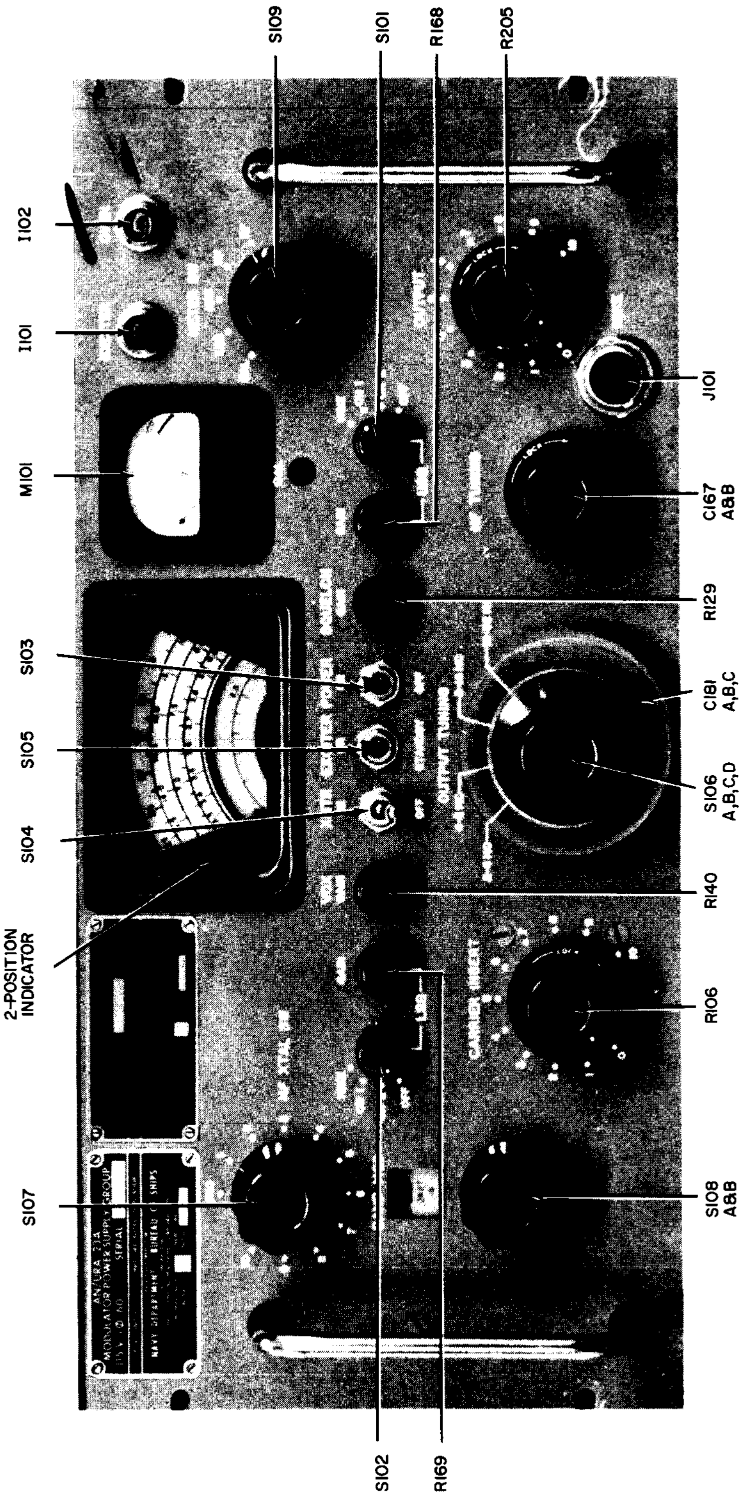
A 4CX5000A tube is used in the final Class-$AB_1$, grounded-grid amplifier stage. The input from V203 is applied to the cathode via J901, and the output of the final is applied to the antenna by E905 and E906 for a balanced output, and by J903 for an unbalanced output. Band switching is accomplished by band switch S900, which progressively shorts out sections of L903 and L902. Tuning is accomplished by C927 which makes up a pi network with L903, L902, and C928. C928 is the PA load capacitor. C916 accomplishes output balance, and L912 and L913 are output loading inductors.

When using an unbalanced output, thermocouple TC900 provides the load current reading on PA output meter M1004. When a balanced output is used, an rf meter is required for each line.

This ends our discussion of the transmitter. Auxiliary equipment and power supplies have not been discussed because they would add little to an understanding of SSB. Student comments may indicate that more detailed explanation of some circuits in the transmitter may be desirable.

If so, in future printings every effort will be made to expand the more important sections of the transmitter.

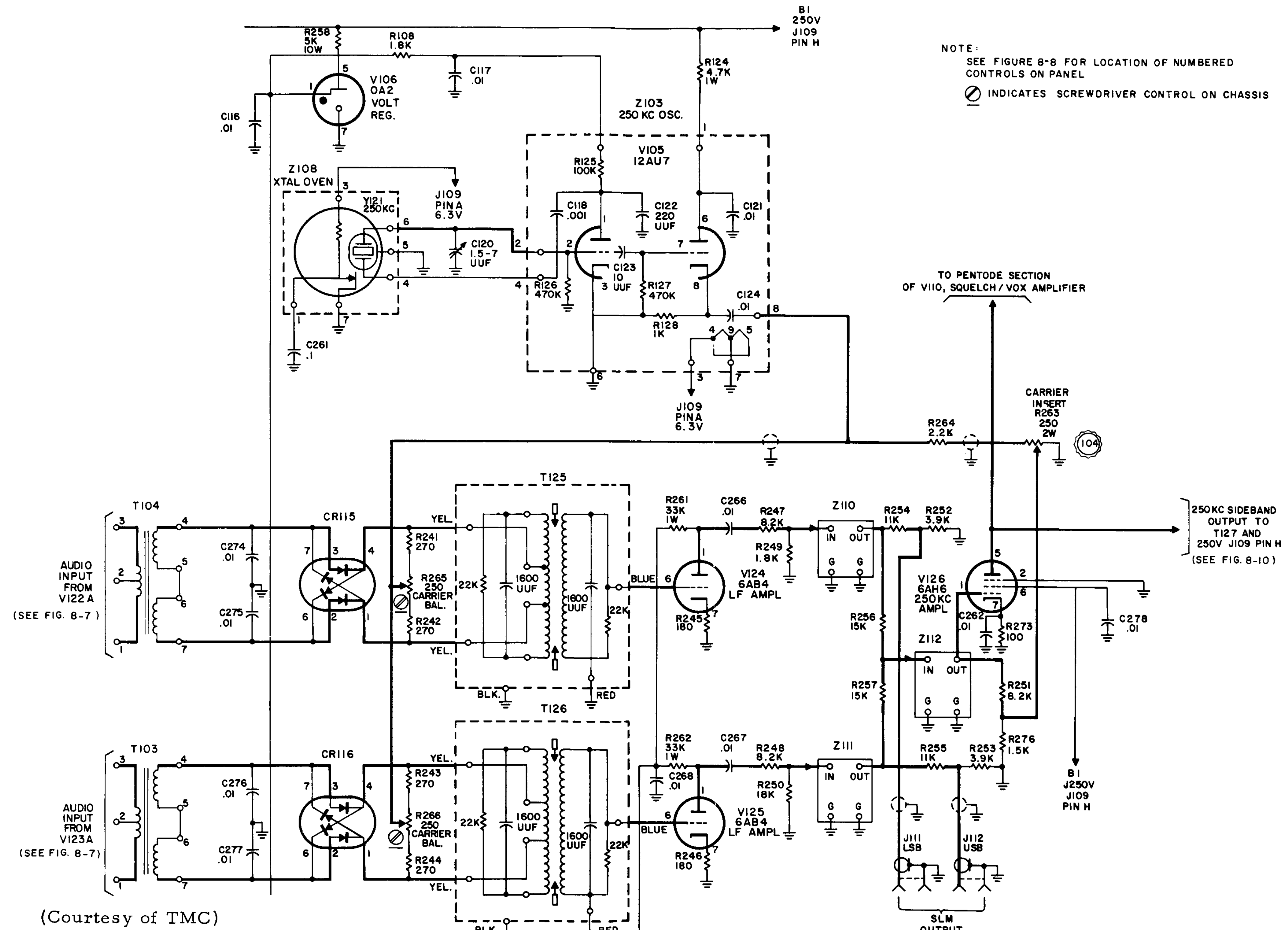
The purpose of considering equipment in this course is to strengthen an understanding of SSB theory. All variations in the utilization of such equipment cannot be considered. This makes it necessary to be very general in our discussion. When questions arise concerning modifications, cable connections, tuning or operating procedures, etc., MANOPS or the equipment instruction book should be consulted.




(Courtesy of TMC)

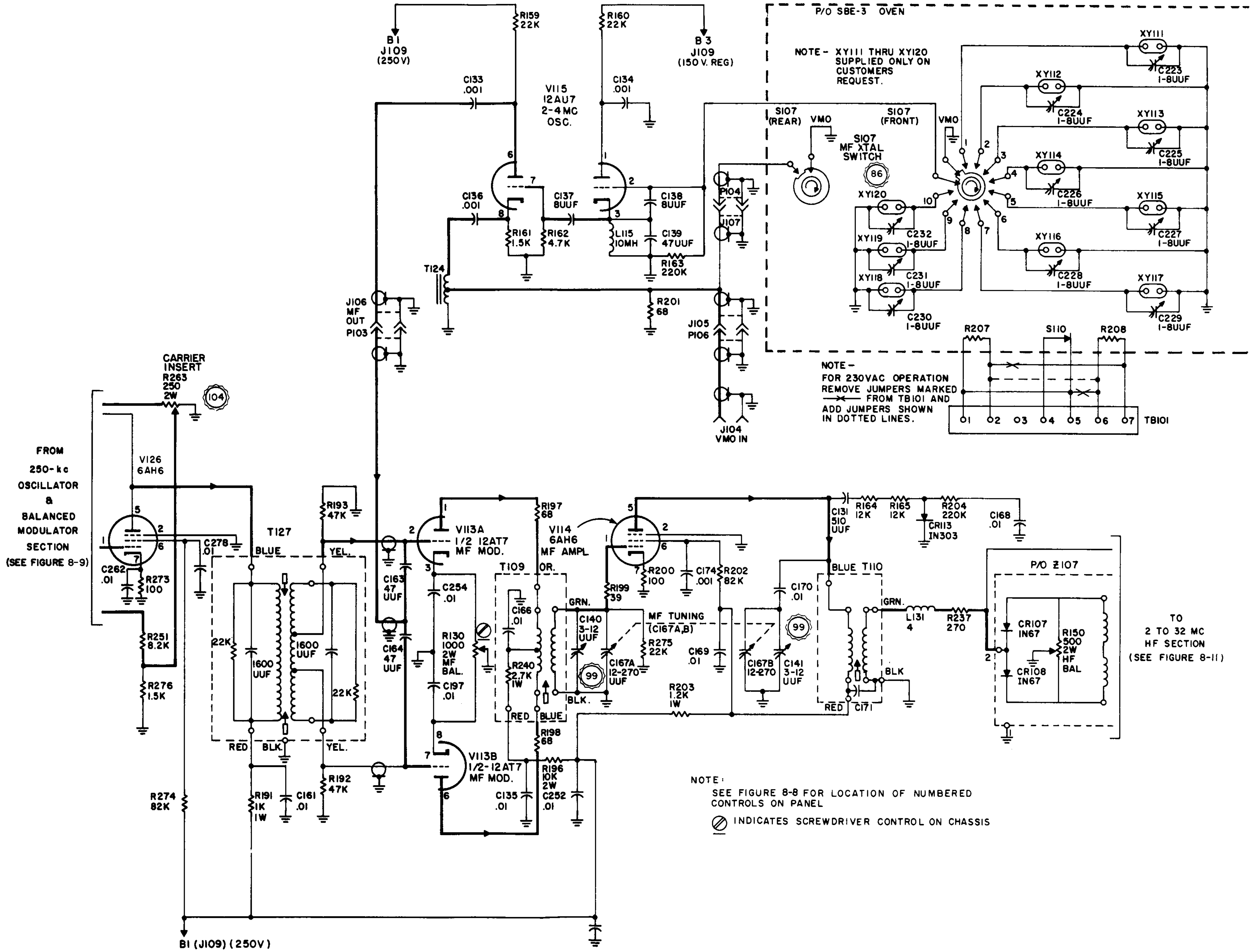
Fig. 8-8. Auxiliary frame operating controls.

NOTE:
SEE FIGURE 8-8 FOR LOCATION OF NUMBERED CONTROLS ON PANEL
⊘ INDICATES SCREWDRIVER CONTROL ON CHASSIS




(Courtesy of TMC)

Fig. 8-9. Schematic diagram SBE-3, 250-kilocycle oscillator and balanced modulator section.




(Courtesy of TMC)

Fig. 8-10. Schematic diagram, SBE-3, 2- to 4-Mc MF section.

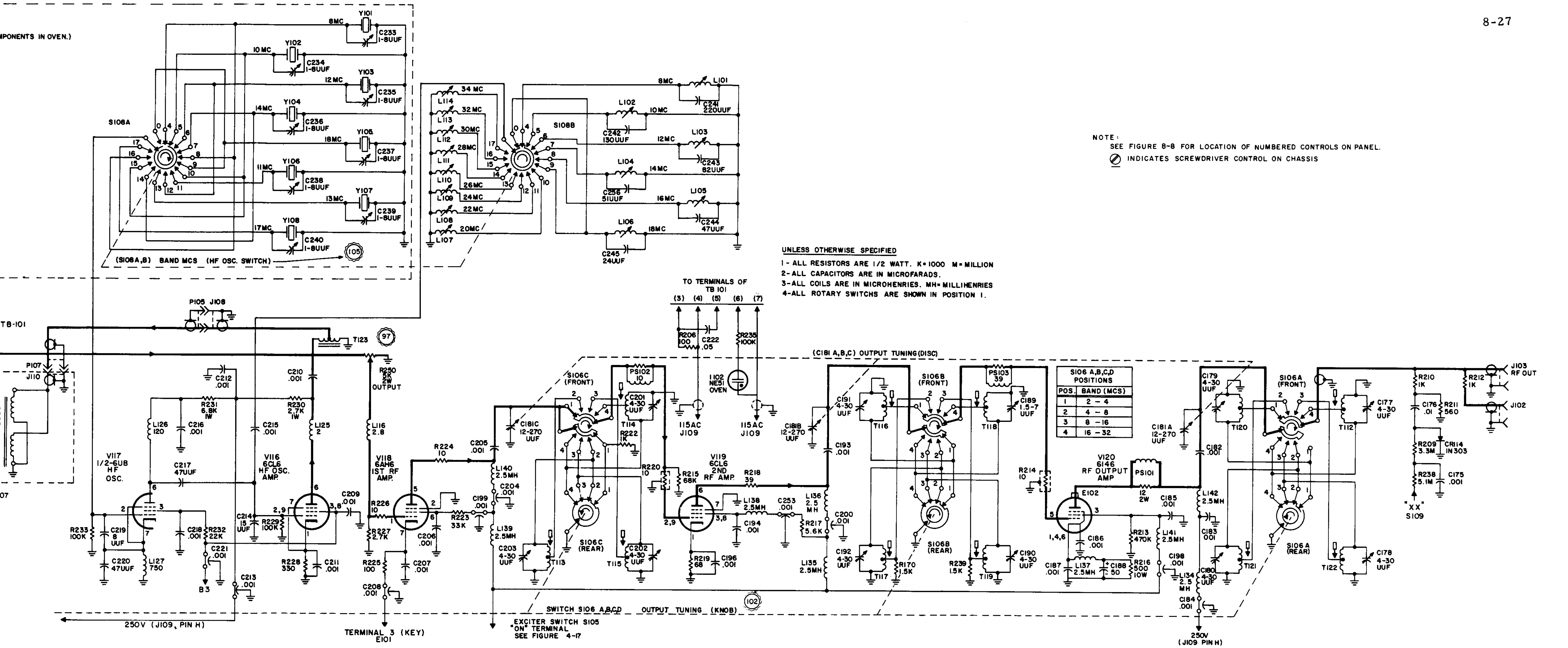
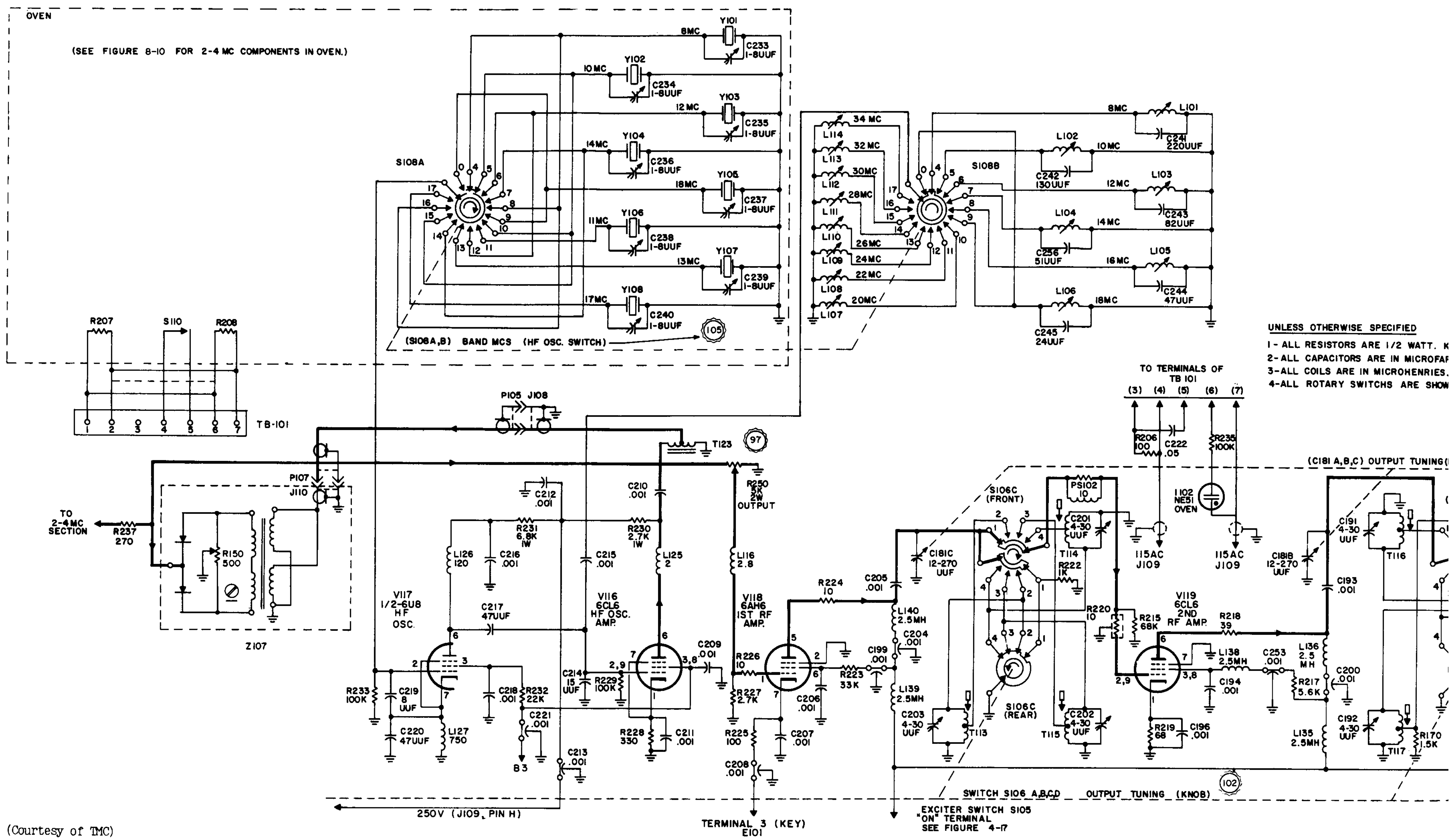
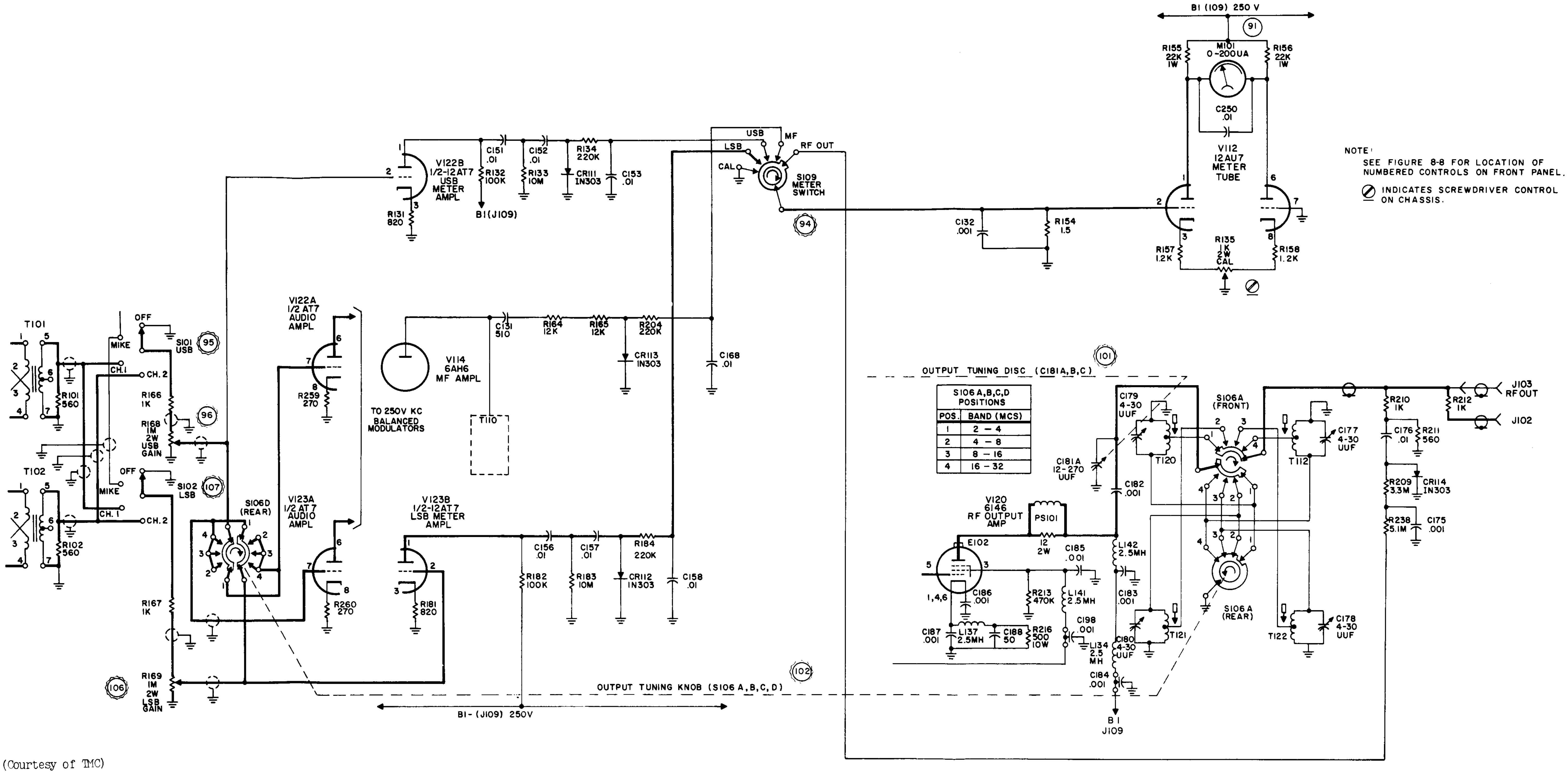



Fig. 8-11. Schematic diagram, SBE-3, 2- to 32-Mc HF section.



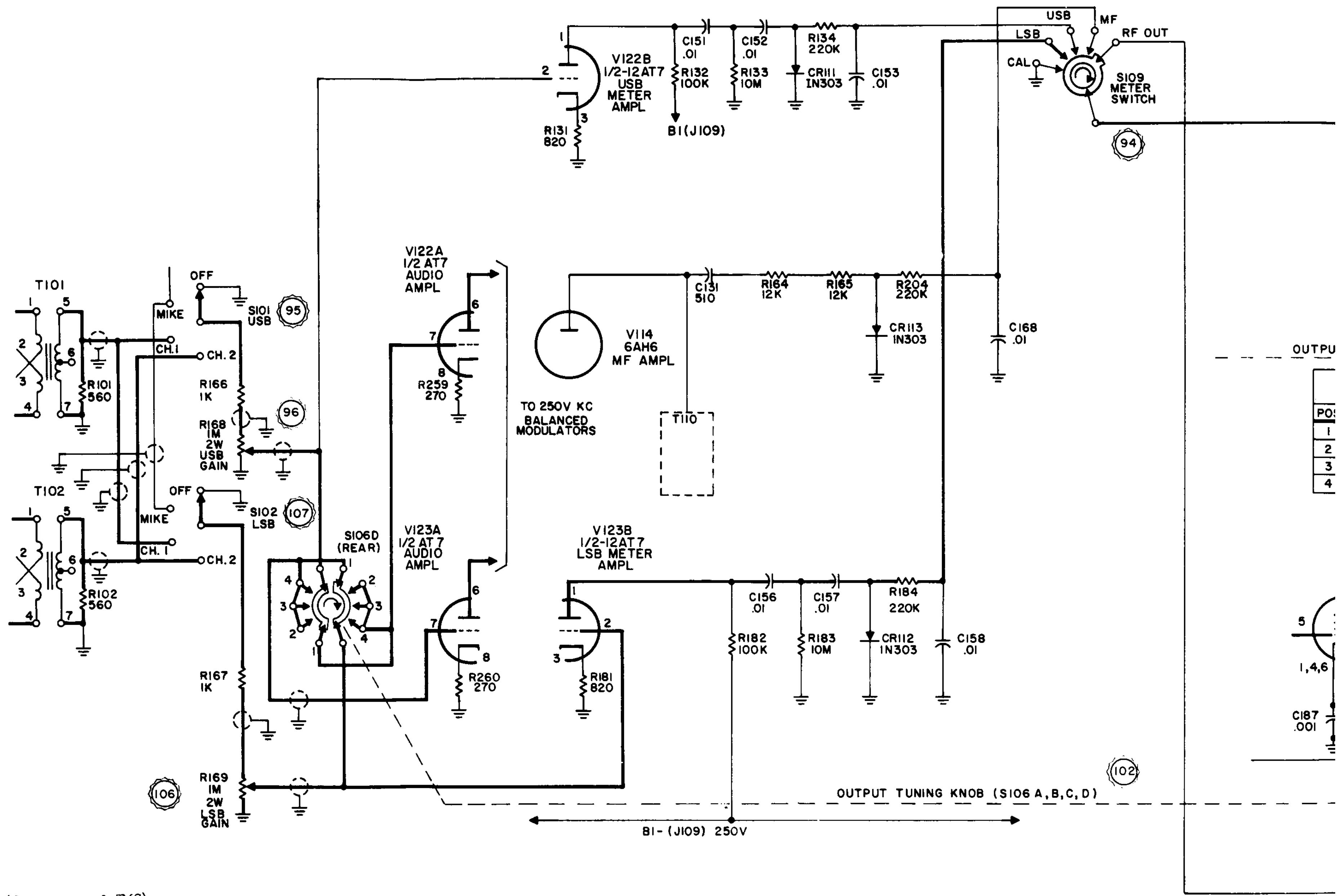

(Courtesy of TMC)



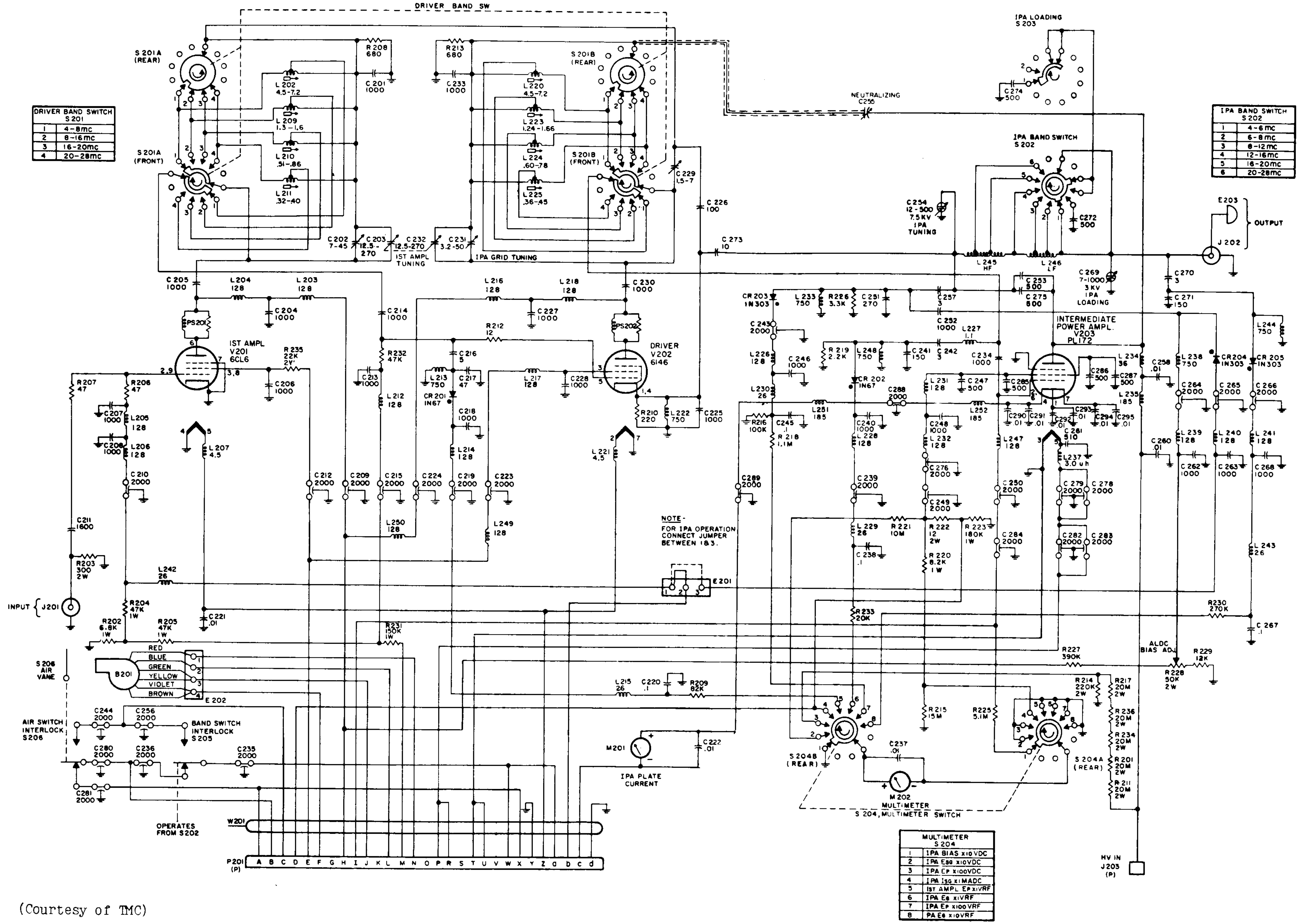

NOTE:
SEE FIGURE 8-8 FOR LOCATION OF NUMBERED CONTROLS ON FRONT PANEL.
⊘ INDICATES SCREWDRIVER CONTROL ON CHASSIS.

(Courtesy of TMC)

Fig. 8-12. Schematic diagram, SBE-3, M101 meter circuits.




(Courtesy of TMC)



(Courtesy of TMC)

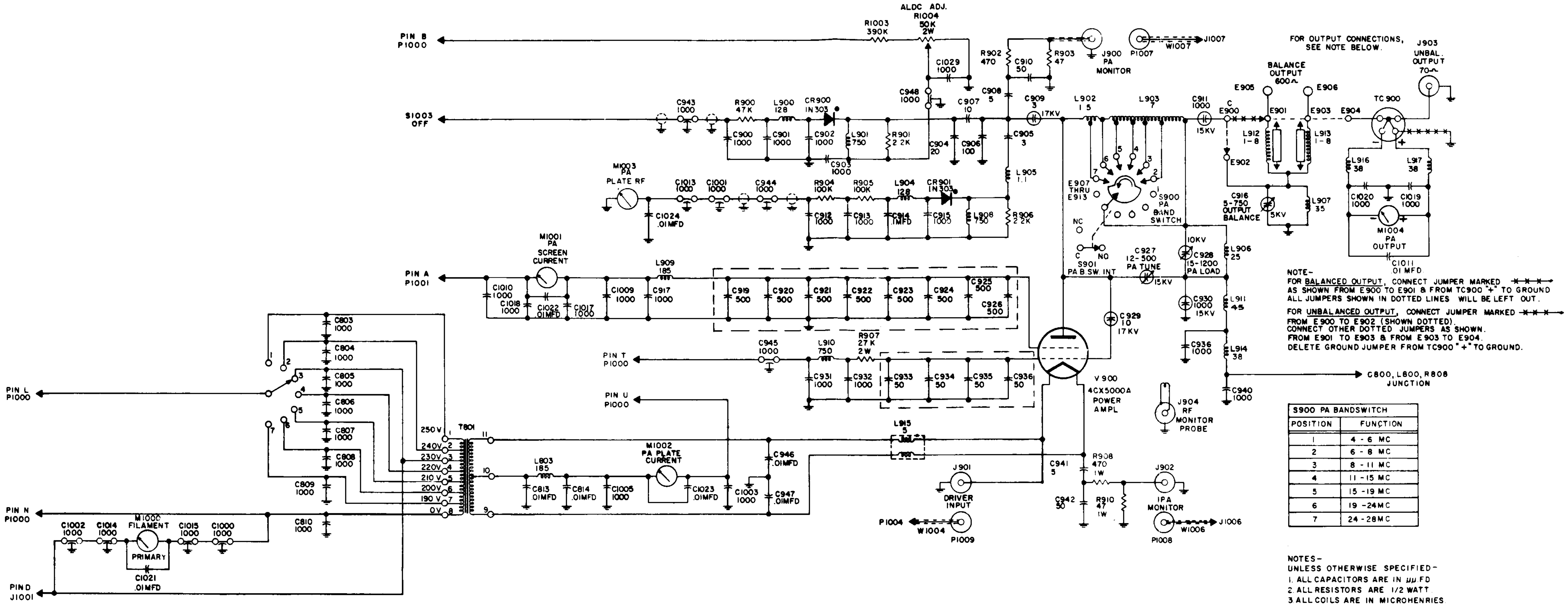
NOTES-

1-ALL CAPACITORS SPECIFIED IN DECIMALS ARE IN uf. ALL OTHERS ARE IN uuf.

2-ALL RESISTORS ARE 1/2 WATT UNLESS OTHERWISE SPECIFIED.

| LAST SYMBOLS | MISSING SYMBOLS |
|---|---|
| B201 | |
| C295 | C259, C277 |
| CR205 | |
| E203 | |
| J203 | |
| L252 | L201, L208, L219, L236 |
| M202 | |
| P201 | |
| PS202 | |
| R236 | R224 |
| S206 | |
| V203 | |
| W201 | |

Fig. 8-13. Schematic diagram, IPA and power supply.



NOTE-
FOR BALANCED OUTPUT, CONNECT JUMPER MARKED -×-×-× AS SHOWN FROM E900 TO E901 & FROM TC900 "+" TO GROUND. ALL JUMPERS SHOWN IN DOTTED LINES WILL BE LEFT OUT.
FOR UNBALANCED OUTPUT, CONNECT JUMPER MARKED -×-×-× FROM E900 TO E902 (SHOWN DOTTED). CONNECT OTHER DOTTED JUMPERS AS SHOWN. FROM E901 TO E903 & FROM E903 TO E904. DELETE GROUND JUMPER FROM TC900 "+" TO GROUND.

S900 PA BANDSWITCH

| POSITION | FUNCTION |
|---|---|
| 1 | 4 - 6 MC |
| 2 | 6 - 8 MC |
| 3 | 8 - 11 MC |
| 4 | 11 - 15 MC |
| 5 | 15 - 19 MC |
| 6 | 19 - 24 MC |
| 7 | 24 - 28 MC |

NOTES-
UNLESS OTHERWISE SPECIFIED-
1. ALL CAPACITORS ARE IN $\mu\mu$FD
2. ALL RESISTORS ARE 1/2 WATT
3. ALL COILS ARE IN MICROHENRIES
4. S900 SHOWN IN POSITION 1
5. S901 IS P/O INTERLOCK SYSTEM SEE MASTER SCHEM (CK-403) FOR CONNECTIONS
6. THE 800 & 1000 SYMBOL SERIES SHOWN ARE NOT P/O PA

| LAST SYMBOLS | MISSING SYMBOLS |
|---|---|
| C948 | C937, 938, 918 |
| CR901 | |
| *E913 | |
| J904 | |
| L917 | |
| R910 | R909 |
| S901 | |
| TC900 | |
| V900 | |

* NOTE
E907 THRU E913 ARE S900 CONTACTS

(Courtesy of TMC)



Fig. 8-14. Schematic diagram, PA section.

# CHAPTER 9

# TERMINAL EQUIPMENT

## 1.0 GENERAL

This chapter has been added since the first printing of the text. Some of its contents will be integrated into appropriate sections of the text if the need to do so becomes apparent later.

In this chapter, we shall tie the operation of some of the terminal equipment in with the SSB communication equipment to enable a better understanding of the overall communications circuit. This will broaden one's understanding of SSB practices in general and show how SSB circuits can be used advantageously.

Most of the equipment discussed here comes under the broad heading of "phase shift and FSK narrow-band data-handling equipment". Under this general heading is included most of the data handling terminal equipment found at International Flight Service Stations (IFSS), but we are concerned here with only the terminal equipment used in conjunction with SSB circuits.

The data handled by SSB communication circuits includes voice, teletype, and facsimile. Teletype data is by far the most common transmitted over SSB circuits; and therefore, in our discussion of terminal equipment, we will assume that teletype signals are being transmitted. The system theory is readily applicable to other types of data because all are handled in the form of audio frequencies or dc pulses in the terminal equipment.

Teletype signals going to a transmitter site are of such a nature that they must be processed in some manner before being transmitted over an intervening medium. An understanding of the equipment used to convey the data over the intervening medium has been the purpose of the preceding eight chapters. The function of the terminal equipment discussed in this chapter is to process the information for transmission and convert it after demodulation to operate the teletype printer. Before going into the detailed discussion of terminal equipment, it is first necessary to discuss the nature of the signals to be handled and then the general purpose and layout of communication facilities. This will give an idea of the function of the equipment and its physical location in the overall communication circuit.

### 1.1 Teletype Code

On most of our SSB circuits, at least one 3-kilocycle channel is used for voice communications. Voice is sometimes desirable because of the speed with which ideas can be transmitted and responses given back. However, teletype communication is often more desirable than voice because it gives a written record of the transmitted information. Also, several teletype circuits can be multiplexed into the 3-kilocycle bandwidth required for one voice channel.

Teletype uses the mark-space five-element teletype code. This code is composed of five information elements, plus a start and a stop pulse for a total of seven pulses. Each of the five information elements is represented by one of two possible conditions. The two conditions may be a dc current for a mark and no current for a space, or a given frequency for a mark and a lower one for a space. The stop pulse is made 1.42 times as long as the other equal pulses so that the teletype sending and receiving equipment will both have time to stop. This permits the two machines to start each new letter or character simultaneously, thus compensating for slight differences in running speeds of sending and receiving equipment. Because of the greater length of the stop pulse, teletype code is referred to as a 7.42 code. Some terminal equipment converts the assynchronous 7.42 code into equal-bit lengths, and the resulting synchronous code is called a 7.0 code.

## 1.2 IFSS Communication Facilities

Before discussing an overall communication circuit, it may be helpful to discuss a typical IFSS Facility for those who are not familiar with them. An IFSS Facility generally is made up of three geographically separated facilities: the IFSS Control Station, the IFSS Transmitting Station, and the IFSS Receiving Station.

1.2.1 IFSS Control Station. The IFSS Control Station is usually located in the immediate vicinity of an international airport. At the Control Station are the personnel who operate the teletypewriting equipment, communicate with the distant IFSS Stations, and maintain air/ground contact with the aircraft along the overseas or international air routes. At the Control Station, the dc teletypewriter signals are converted to audio tones and transmitted over VHF radio links or leased-wire services to the IFSS Transmitter Station. Similar audio tones are also received at IFSS Control from the IFSS Receiving Station and are converted to the proper type of dc signals required to operate the telecommunications equipment at the IFSS Control. The IFSS Control Facility is connected with other FAA stations and centers throughout the FAA's nationwide telecommunications system.

1.2.2 IFSS Transmitter Station. The Transmitter Station is located several miles from the IFSS Control Station in a sparsely populated area. Several acres of directional high-frequency antenna arrays are required. The teletype tones and voice communications from Control are applied as modulation to the high-frequency transmitters.

1.2.3 IFSS Receiver Station. The Receiver Site is located several miles from both the Control Station and the Transmitter Site, in an area as free as possible from industrial and other types of man-made electrical and electromagnetic interference to radio reception. The Receiver Site has several acres of directive receiving antenna arrays. The radio frequency telecommunications data and other intelligence from overseas IFSS Transmitting Sites are received and converted to audio signals or voice before being transmitted by VHF links or land lines to the IFSS Control Station.

### 1.3 Definition of Circuits and Channels

For this discussion it is necessary to define what is meant by the terms communication circuit, communication channel, teletype circuit, and subchannel.

A communication circuit is the communications link between two distant (overseas) IFSS Facilities. Such a communications circuit, using one transmitter, may handle only one teletype circuit, or it may handle several teletype circuits by the use of multiplexing.

A communication channel has a bandwidth of 3000 cycles. Channels are assigned 3000-cycle bandwidths because 3 kilocycles is sufficient for voice communications. The data-handling terminal equipment is normally designed for 3-kilocycle bandwidths to make it compatible with the 3-kilocycle channel allocations standard in most communication systems. We have seen that the utilization of SSB enables four 3-kilocycle communication channels on a communication circuit having a bandwidth of 12 kilocycles.

A teletype circuit consists of the teletype printing and sending equipment, terminal data-handling equipment, and the intervening circuit. Several teletype circuits are handled simultaneously on one communication channel by means of multiplexing.

A subchannel is the narrow band of frequencies allocated to each teletype or similar data circuit. That is, each teletype circuit of a multiplex communication system occupies a subchannel.

Thus, it follows that an SSB communication circuit is divided into 3-kilocycle communication channels; each communication channel is further divided into subchannels. Normally, a subchannel handles only one teletype circuit, but in some communication systems each subchannel contains two teletype circuits (Kineplex,for example). We will be consistent in the use of these terms in order to avoid confusion in terminology.

## 1.4 Multiplexing

Multiplexing is defined as the method of increasing the number of teletype or other information circuits handled by a communications circuit. Direct frequency shift communication circuits (where the carrier is shifted from 425 cycles above to 425 cycles below an assigned center frequency) have provided adequate and dependable communications, but such a communications circuit is limited to only one teletype circuit. When traffic between two IFSS Facilities exceeds that which can be handled by a single direct-frequency-shift circuit (which is normally the case), then additional direct-frequency-shift circuits must be added or different systems utilizing some form of multiplexing must be used. Because traffic is so great in the high-frequency spectrum multiplexing systems are usually preferred.

1.4.1 Time-Division Multiplexing. Time-division multiplexing systems used by the FAA permit four teletype circuits to be handled on a single communications circuit. The information on each teletype circuit is sampled in a sequential order and the transmitted signal is shifted to the mark or space frequency to correspond to the information on the circuit being sampled. The sending and receiving equipment must be synchronized so that the receiving equipment will divert the incoming signals to the proper decoders. Time-division multiplexing systems are not as economical spectrum-wise as some of the more advanced frequency-division multiplexing systems.

1.4.2 Frequency-Division Multiplexing. A frequency-division multiplexing system is any system of combining a number of subchannels into one communications channel that utilizes a frequency difference to maintain division among the subchannels.

Tone Keyed. Probably the oldest and the simplest-to-explain method of frequency-division multiplexing is the tone-keyed system. Here, the teletype pulses simply turn on and turn off audio "subcarriers", all of which lie within the bandpass of one communication channel. A typical example of this system is the 1621 Tone Channeling Equipment where five audio subcarriers, operating at 595 cps, 935 cps, 1235 cps, 1615 cps, and 1955 cps, convey five subchannels of information over one communication channel.

Frequency-Shift Keyed. In frequency-shift keyed frequency-division multiplexing systems, the mark and space impulses of a teletype circuit cause an audio oscillator to alternately shift between two frequencies. One frequency corresponds to a mark while the other corresponds to a space. By using a frequency shift of ±42.5 cps from a center frequency for each subchannel and a 170-cps separation between center frequencies, a total of 16 channels can be placed in a 3-kilocycle communication channel (425 cps for the lowest oscillator and 2975 cps for the highest). Equipment that accomplishes this, the Dual-Frequency Shift Tone Keyer and the Frequency-Shift Diversity Converter, will be discussed later under Equipment.

Twinplex. Twinplex is a specialized method of frequency-division multiplexing that allows the transmission of two teletype circuits over one frequency-shift communication circuit. Although it is not as conservative of the spectrum as many of the other multiplexing methods listed, it does double the traffic handling capabilities of a direct frequency-shift circuit while retaining many of its advantages (simplicity, reliability, etc.).

In Twinplex, the frequency is shifted among four different frequencies rather than between the conventional two. To explain, let the four frequencies be $f_1$, $f_2$, $f_3$, and $f_4$. Let M1 represent a mark on teletype circuit No. 1, M2 a mark on teletype circuit No. 2, S1 a space on circuit No. 1, and S2 a space on circuit No. 2. With two teletype circuits, there are four possible combinations of mark and space. Both circuits can be mark, both can be space, or one circuit can be a space and the other a mark. A glance at the following table shows how any one of the four possible combinations can be transmitted by shifting the carrier to one of four possible frequencies.

| Mark-Space Combination To Be Transmitted | Radiated Frequency |
|---|---|
| M1 M2 | $f_1$ |
| M1 S2 | $f_2$ |
| S1 M2 | $f_3$ |
| S1 S2 | $f_4$ |

Kineplex. The Kineplex system applies the principle of Twinplex to phase-modulated frequency-division multiplexing. Here, twenty subcarriers, spaced 110 cps apart from 605 cps to 2695 cps, are each modulated with two channels of information (teletype circuits) for a grand total of forty information channels on one communication channel.

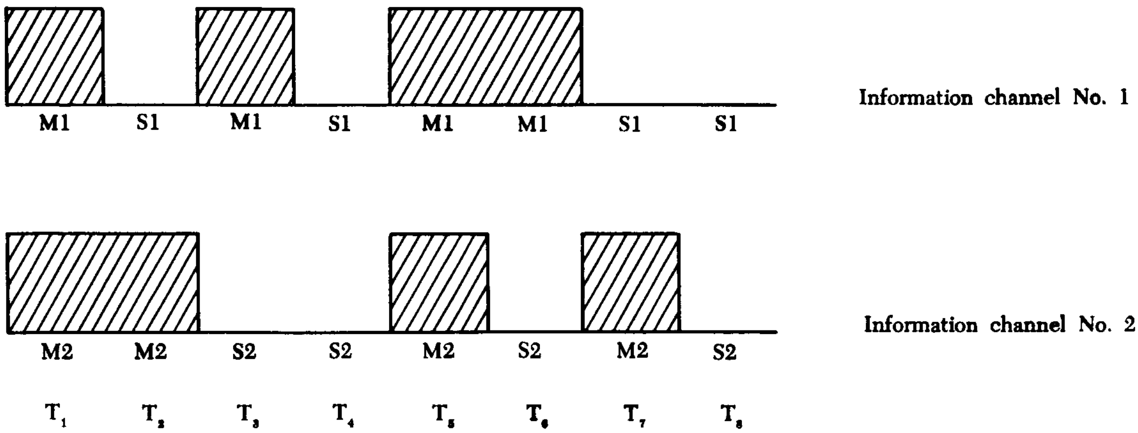
The method used to combine two teletype channels on one subcarrier oscillator is very similar to the method used in Twinplex, except that in Kineplex each one of the four possible combinations is represented by a phase angle of an audio subcarrier with respect to a particular reference ($\phi_1$, $\phi_2$, $\phi_3$, or $\phi_4$) rather than by a unique frequency ($f_1$, $f_2$, $f_3$, or $f_4$). This is shown in the following table.

| Mark-Space Combination | Radiated Phase | Standard Vector |
|---|---|---|
| M1 M2 | $\omega t + 45$ |  |
| S1 M2 | $\omega t + 135$ |  |
| S1 S2 | $\omega t + 225$ |  |
| M1 S2 | $\omega t + 315$ |  |

This phase modulation of the subcarriers has certain bandpass advantages over a frequency-shift system; however, it has the disadvantage of requiring a phase reference for demodulation. This disadvantage is circumvented by using the phase of the previous mark-space combination as a reference for the succeeding combination. This is demonstrated in Fig. 9-1.

It is apparent that assynchronous teletype code does not meet the inherent timing requirements of this method of combining two information channels. Also, each bit (mark or space) from information channel No. 1 must occur simultaneously with its traveling companion from information channel No. 2. It is therefore necessary for a code converter to retime the nonsynchronous teletype information so that each bit occupies the same time period (7.0 code) and so that all bits from information channel No. 1 coincide with bits from channel No. 2.

Another requirement of the Kineplex system is that the transmitter (coder) must be synchronized with the receiver (decoder). This results from the fact that if the receiver is going to compare the phase of each bit of information with the phase of the previous bit, it must know precisely where one bit ends and another begins. This synchronization is accomplished by sending a synchronizing tone at 2915 cps, well above the highest information subcarrier frequency of 2695 cps yet within the allocated 3-kilocycle communication channel.




| Time | M/S combination | Phase | Standard vector | Previous combination vector | Transmitted vector |
|---|---|---|---|---|---|
| $T_1$ | M1 M2 | 45° | | 0° | 45° |
| $T_2$ | S1 M2 | 135° | | 45° | 180° |
| $T_3$ | M1 S2 | 315° | | 180° | 135° |
| $T_4$ | S1 S2 | 225° | | 135° | 0° |
| $T_5$ | M1 M2 | 45° | | 0° | 45° |
| $T_6$ | M1 S2 | 315° | | 45° | 0° |
| $T_7$ | S1 M2 | 135° | | 0° | 135° |
| $T_8$ | S1 S2 | 225° | | 135° | 0° |

FIGURE 9-1.

## 1.5 Block Diagram of a Communication System

A simple radio teletype system is shown in Fig. 9-2. Its operation is as follows. First, imagine that a message is initiated at IFSS No. 1 and it is desired to transmit this message to IFSS No. 2. The message is first converted to teletype coded dc pulses on TTY (teletype) circuit No. 1-1 by a teleprinter or a tape reader. These dc pulses are fed into a multiplex coding system where they are converted to audio and combined with from 1 to 39 other TTY signals. The combined output of the multiplex system (communication channel No. 1) is fed as one audio channel out of a possible four to an SSB transmitter via a channel shifter. The four audio channels are independent-sideband reduced-carrier transmissions and are transmitted as RF to a receiver of IFSS No. 2. The receiver detects the audio, separates the communication channels with the help of a channel shifter and diverts communication channel No. 1 to a multiplex decoding system. The multiplex decoding system separates the subchannels, converts the audio back to dc pulses and directs the message to teleprinter No. 1-2. Here, the message is read and considered.

Now, imagine that IFSS No. 2 has an answer to the message from IFSS No. 1. The answer is converted to dc pulses, again on TTY channel No. 1, fed to a multiplexing system at IFSS No. 2 and modulated on their transmitter. It is received by the receiver of IFSS No. 1, separated, decoded, converted, and diverted back to teleprinter No. 1-1 at IFSS No. 1.

A little arithmetic will show that for a system using the Kineplex multiplexing system and transmitting independent sideband with two 3-kilocycle communication channels on each side of the carrier, it is possible to accommodate 160 60-wpm teletype circuits. It should be noted here that this is the maximum utilization of spectrum space that present technology will allow, and the actual number of TTY circuits utilized depends upon the need and equipment used.

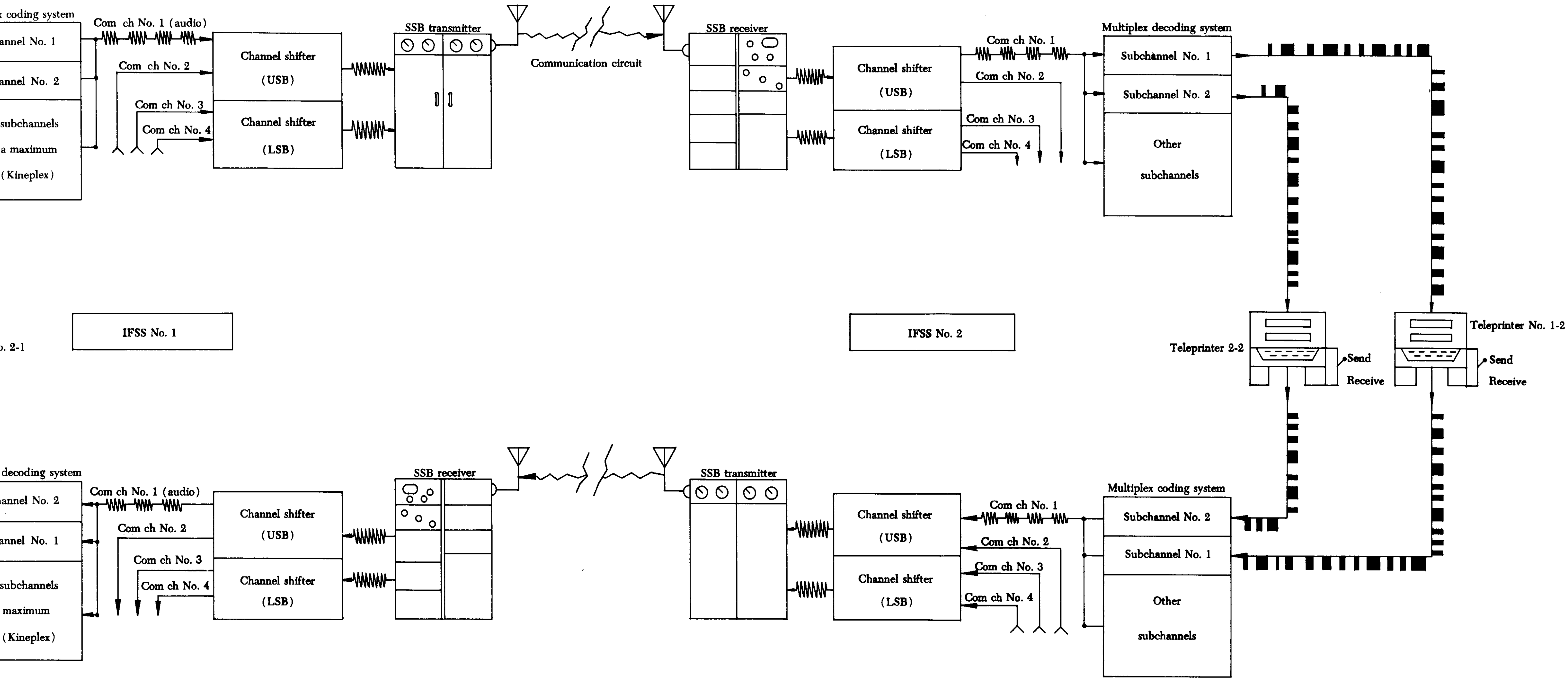
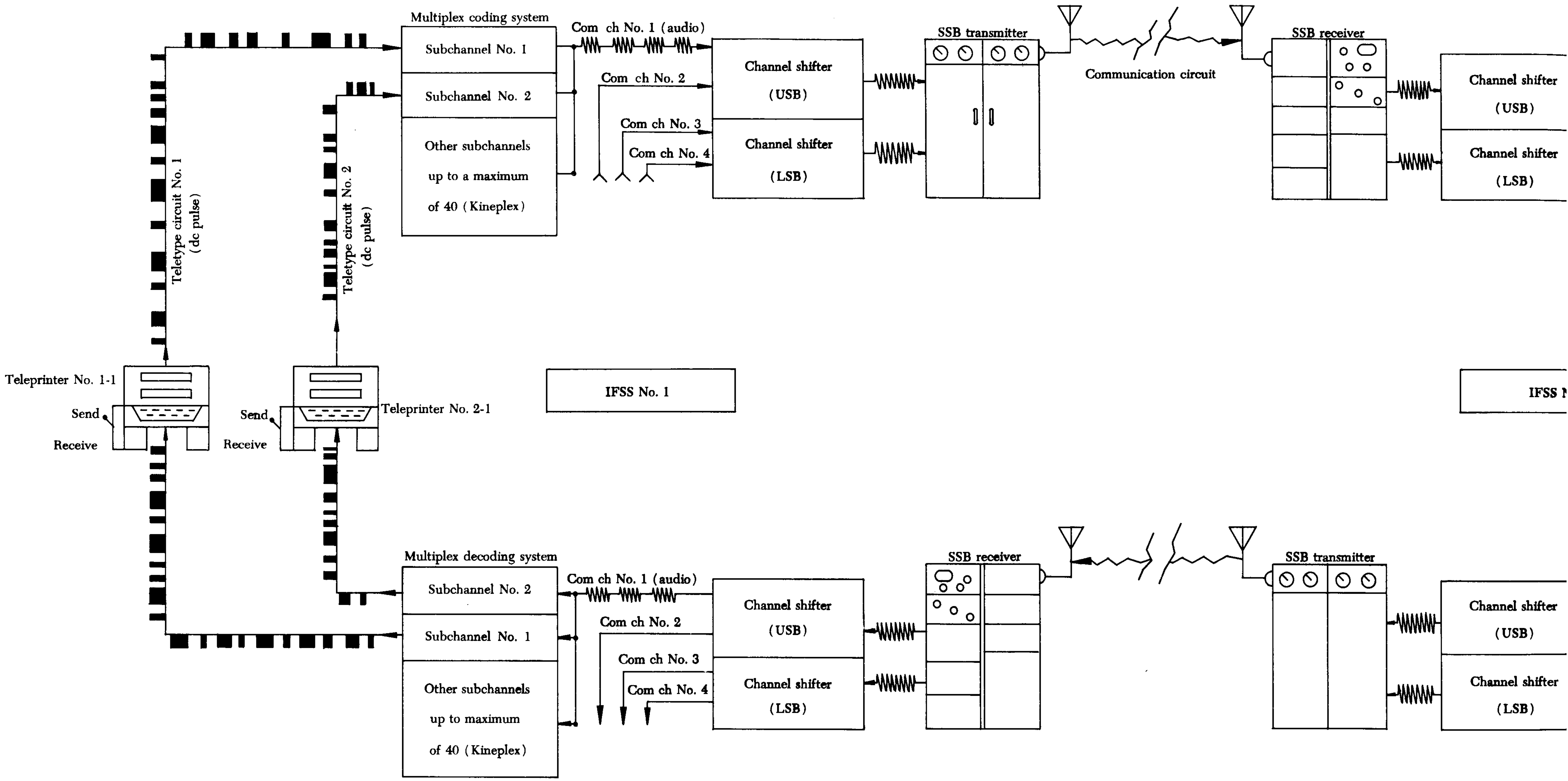



FIGURE 9-2. Simplified radio teletype communications system.


Multiplex coding system
Subchannel No. 1
Subchannel No. 2
Other subchannels
up to a maximum
of 40 (Kineplex)
Com ch No. 1 (audio)
Com ch No. 2
Com ch No. 3
Com ch No. 4
Channel shifter
(USB)
Channel shifter
(LSB)
SSB transmitter
Communication circuit
SSB receiver
Channel shifter
(USB)
Channel shifter
(LSB)
Teletype circuit No. 1
(dc pulse)
Teletype circuit No. 2
(dc pulse)
Teleprinter No. 1-1
Send
Receive
Send
Receive
Teleprinter No. 2-1
IFSS No. 1
IFSS N
Multiplex decoding system
Subchannel No. 2
Subchannel No. 1
Other subchannels
up to maximum
of 40 (Kineplex)
Com ch No. 1 (audio)
Com ch No. 2
Com ch No. 3
Com ch No. 4
Channel shifter
(USB)
Channel shifter
(LSB)
SSB receiver
SSB transmitter
Channel shifter
(USB)
Channel shifter
(LSB)

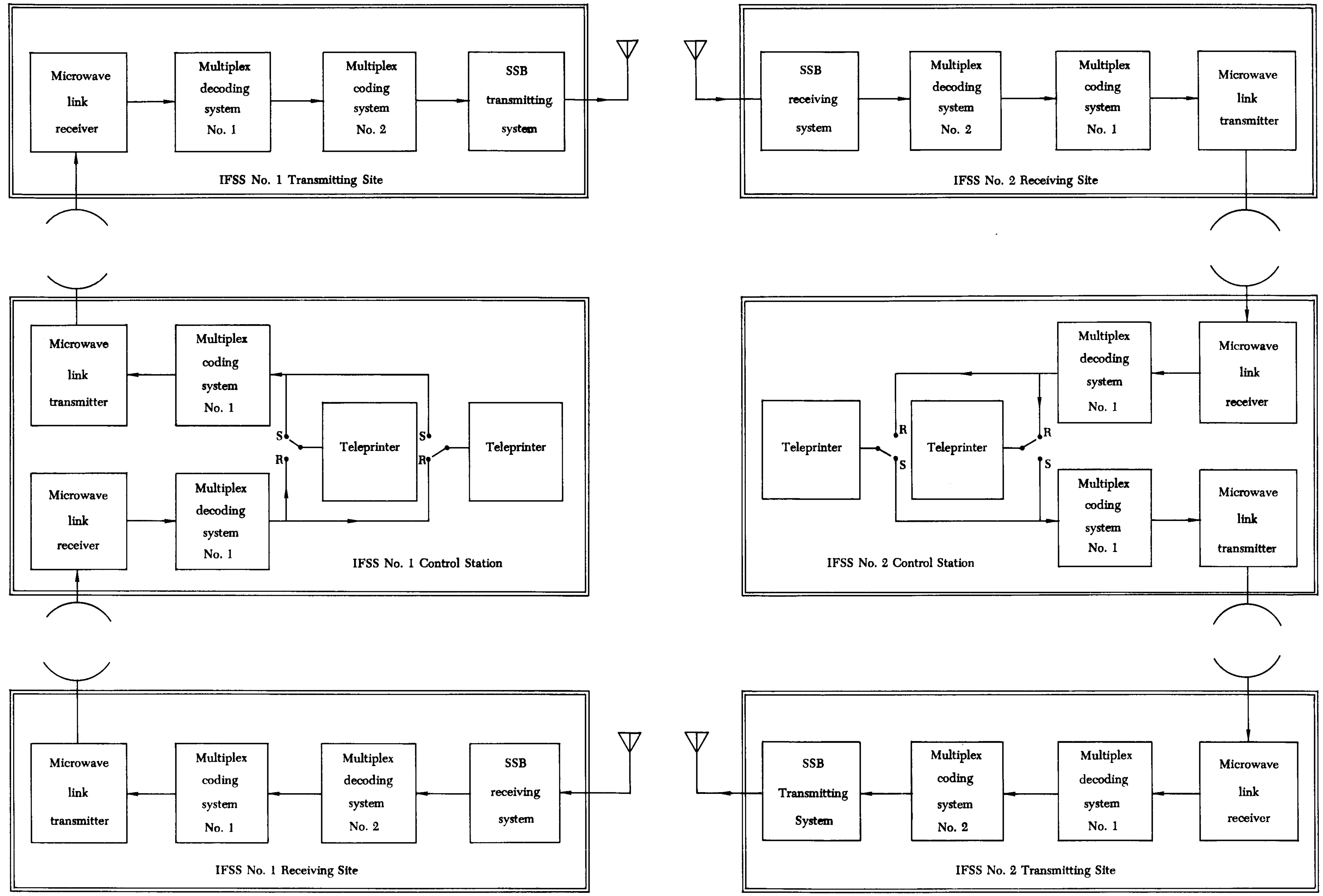



FIGURE 9-3. Typical radio teletype communications system.

For example, for those stations using the Dual-Frequency-Shift Tone Keyer (Type 153/3) and Frequency-Shift Diversity Converter (Type 174/1) as a multiplexing system, the maximum number of 60-wpm TTY circuits per communication channel drops to 16. If the 1621-tone channeling equipment is used, this number drops to four or five. Also, if an IFSS is not equipped with channel shifters, it will be limited to only two communication channels per communication circuit.

The system shown in Fig. 9-2 is a hypothetical one for it assumes that the transmitting, receiving, and control equipment are all located at the control site. As previously stated, this is not practicable and in almost all cases the equipment is distributed among three sites — transmitter, receiver, and control. The added problems of transporting the information between the sites further complicate the complete system, and there are many variations in the methods by which this is accomplished.

Figure 9-3 shows a typical system where microwave links are used for intersite connection and where, at both the transmitter and receiver site, the information is decoded to dc pulses and recoded by another multiplex system more compatible to the applicable linkage. Other possible intersite linkage systems include VHF link, UHF link, and land lines.

## 2.0 EQUIPMENT

### 2.1 Dual Frequency-Shift Tone Keyer, Type 153, Model 3

This frequency-shift tone keyer is used in multi-teletype-circuit communication channels to convert dc line pulses to frequency-shift tones for modulating the sideband transmitter. It is a double unit and therefore provides the frequency-shift tones for two data circuits. The two sections of this unit will operate on any of the standard-tone subchannels by changing a plug-in network. The oscillator frequency of each unit is shifted ±42.5 cps about the subchannel center frequency. The oscillator circuit is electronically switched from one feedback circuit to another to obtain the frequency shift. Figure 9-4 is a schematic of one section of the dual unit. A schematic of the other section would be identical.

When the space frequency is being generated, the right-hand section of V3 is made to operate as an amplifier and the feedback loop associated with this section causes oscillations at the space frequency. When the mark frequency is being generated, the left section of V3 is made to operate as an amplifier and its associated feedback loop causes oscillations at the mark frequency. The two feedback loops consist of R1/C1 for space and R2/C2 for mark. These feedback circuits are contained in the plug-in frequency-determining network enclosed in the dashed lines on the schematic.

To analyze the complete operation, we will first place the testing switch, S2, in the M position. This causes the unit to operate as though a mark pulse were received from the line input. The M contact of S2 is connected to B+ through R5. The grid voltage of V1 is now sufficiently positive to cause V1 to conduct. The two sections of V1 are connected in parallel. When V1 conducts, a negative pulse is received on the grid of V2A, cutting V2A off. V2B is caused to go into conduction by the positive pulse at the plate of V2A. When V2B conducts, the voltage drop across the cathode resistor of V3B (½ of R17) cuts V3B off. V3A is permitted to operate because V2A is cut off. Then V3A completes the feedback loop allowing oscillations at the mark frequency. The plates of V3 are connected in parallel. The operating section of V3 passes the feedback signal on to the left section of V4 (V4B) via C4.

The plate load for V4B consists of the tank circuit, L1/C3, which is tuned to the center frequency (½ way between space and mark frequencies). R5 is used on the lower frequency channels to reduce the Q of the tank circuit. C7 passes the output of V4B on to the grid of V4A. The cathode load of V4A is the parallel combination of R3 and R4 in the plug-in frequency determining network. R3 and R4 feed two RC phase-shifting networks connected to the grids of V3 which completes the feedback loop. The phase-shifting network R1/C1 feeds the grid of V3B, and R2/C2 feeds the grid of V3A. When V3A is operating (as is the case when a mark frequency is being generated), R2/C2 is the feedback network. Note that the output is taken across the resistor, and therefore, R2/C2 is a "leading" phase-shift network which accounts for the higher frequency of oscillation.

When V3B is operating (as is the case when a space frequency is being generated), R1/C1 is the frequency determining network which causes oscillations at a lower frequency because of the "lagging" feedback voltage. The following table shows the status of the input and oscillator tube sections for mark and space pulses. V4A, V4B and V5 are operating for both mark and space pulses.

| | V1 | V2A | V2B | V3A | V3B |
|---|---|---|---|---|---|
| Space | cutoff | conducting | cutoff | cutoff | conducting |
| Mark | conducting | cutoff | conducting | conducting | cutoff |

The plate circuit of V4A feeds V5 through C6. The output of V5 is transformer coupled to the channel filter BP1 located in the frequency-determining network. The output of BP1 between terminals 8 and 6 is fed to terminals 5 and 6 of E1. The output impedance characteristic of the channel filter is such that 18 filters can be paralleled. A typical system in the FAA communications circuit uses 8 dual frequency-shift tone keyers, which result in paralleling 16 filters.

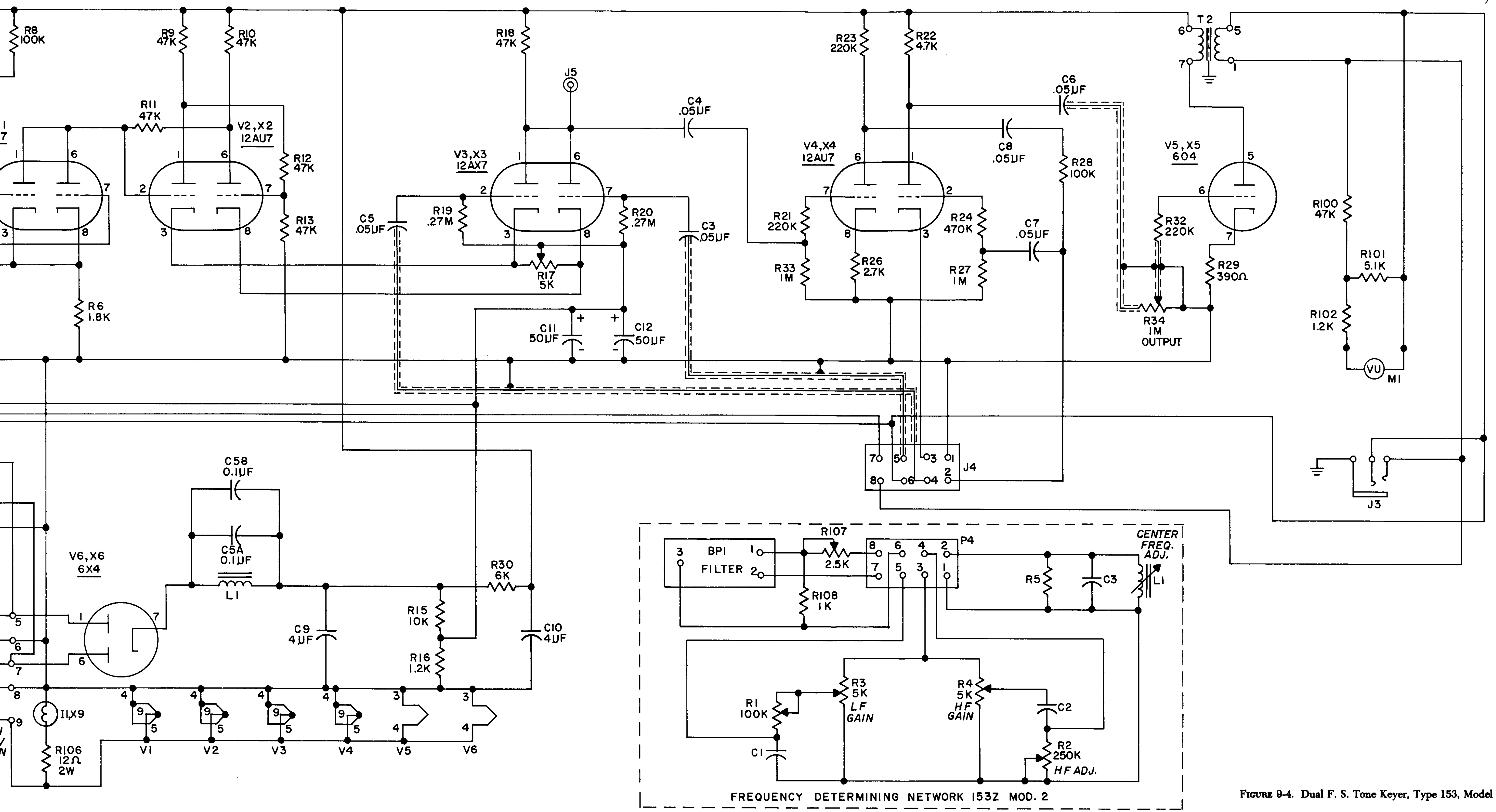
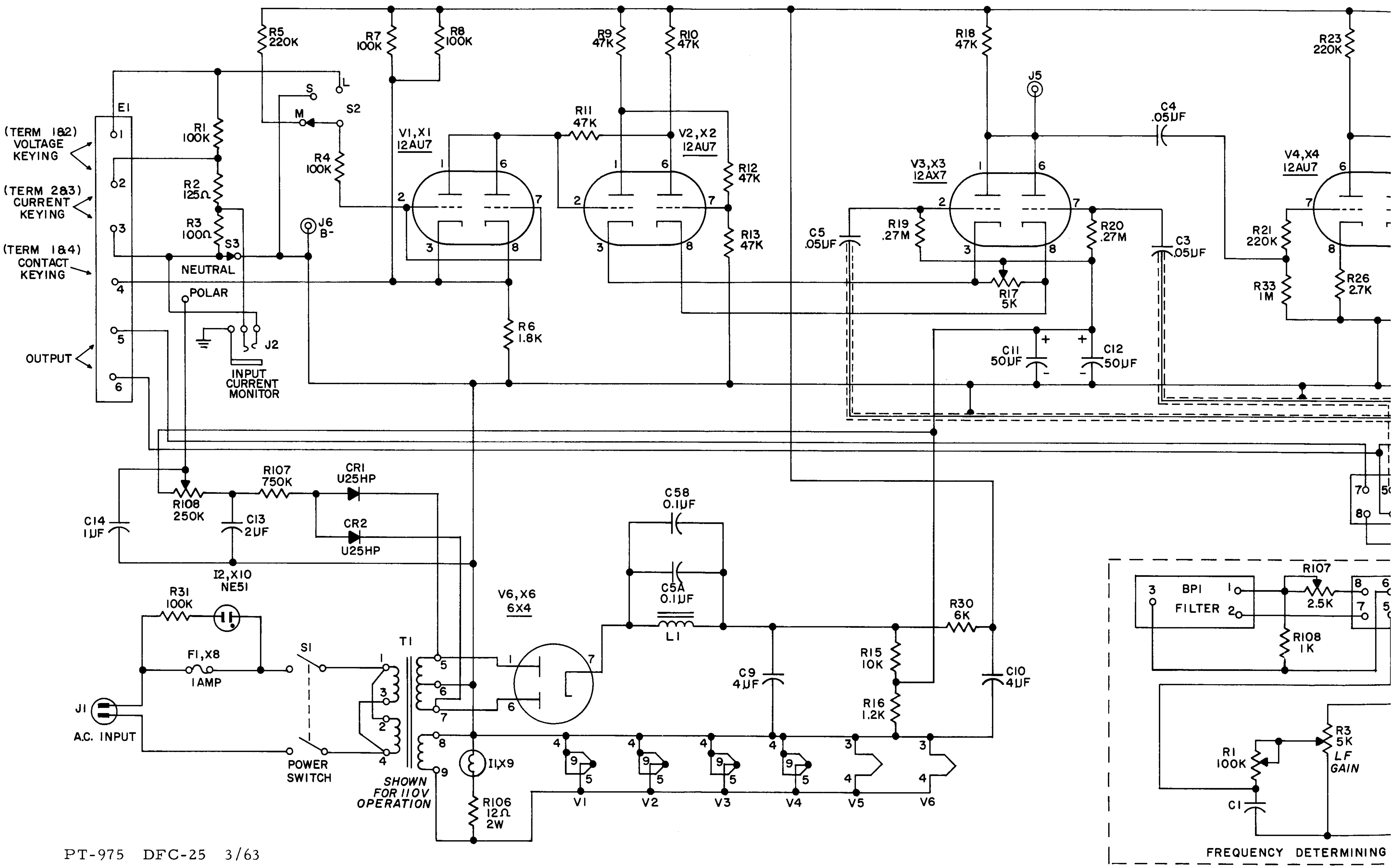



FIGURE 9-4. Dual F. S. Tone Keyer, Type 153, Model 3.

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## 2.2 Frequency-Shift Diversity Converter, Type 174, Model 1

The frequency-shift converter is used in multi-teletype circuit communications channels to convert frequency-shift audio tones into dc line pulses capable of operating teletype printing equipment. That is, it performs the inverse function of the Type 153 keyer unit and is used in conjunction with it.

The 174 is a dual unit designed for diversity operation. The output of one diversity receiver is fed to one section of the unit and the output of the other diversity receiver is fed to the second section. When 16 subchannels are contained within one communication channel, the communication channel from, say receiver No. 1, will be fed to one section of 16 converters while the same communication channel from receiver No. 2 (receiver No. 1 and No. 2 designate a pair of receivers operating space diversity) will be fed to the other section of the 16 converters.

Referring to the block diagram shown in Fig. 9-5, identical information from two different sources is fed into input No. 1 and No. 2. These signals first encounter a bandpass filter that passes only one subchannel of information and rejects the other 15. The bandpass filters for both sections of one converter are identical; however, each different converter has different filters corresponding to the particular subchannel it is going to convert. From the bandpass filter, the subchannel information is fed through an input transformer, through dc and ac amplifiers and is limited to remove any amplitude fluctuations that could degenerate the signal. The signal next encounters the discriminator circuit where tuned circuits (unique for each subchannel) separate the mark signals from the space signals. The mark and space signals are then rectified—the marks producing a positive dc voltage and the spaces producing a negative dc voltage.

Both sections of the converter are identical up to this point and each feeds its ±dc pulse output to the diversity combiner. The diversity combiner (with the help of the automatic diversity control) selects the stronger of the outputs and passes it through an impedance-matching device to a low-pass filter. The output of the low-pass filter is fed through ac and dc amplifiers to a dc keyer which controls an electronic relay. The electronic relay has its own separate power supply (see Fig. 9-7) and provides a floating output, isolated from the converter proper, capable of driving one or more teleprinters.

A tuning monitor, connected to the discriminator unit, provides a visual tuning indication by means of a CRT display.

The schematic of the Type 174, Model 1, converter is shown in Fig. 9-6. The input from receiver No. 1 enters terminals 6 and 7 of terminal board E3 and the input from receiver No. 2 enters terminals 1 and 2. They are then fed through a 100-cps bandpass filter (part of a plug-in unit that determines the particular subchannel to be converted). The audio tone outputs of the bandpass filters go through input transformers T1 and T2 that provide the out-of-phase voltages necessary to drive the balanced push-pull dc amplifiers V1, V2 and V7, V8. M1 and M2 monitor the relative input levels to the converters.

The signal is then capacity coupled to ac amplifier-limiters V3, V4 and V9, V10. Coupling diodes CR1, CR2, CR3, CR4 and CR5, CR6, CR7, CR8 assure symmetrical charging and discharging of the coupling capacitors, thus improving the recovery time of the limiters.

The output of the limiter is applied to the discriminator-tuned circuits contained in the same plug-in unit as the input filter. The "mark"-tuned circuit is tuned to approximately 45 cps above the applicable subcarrier frequency while the "space"-tuned circuit is tuned to 45 cps below.

The outputs of the mark tank circuits are full-wave rectified by V5 and V11 to produce positive pulses while the output of the space tank circuits are rectified by V6 and V12 to produce negative output pulses. These positive and negative pulses are fed to separate contacts of channel switch S2.

Channel switch S2 selects either the output of section one, the output from section two, or (in the diversity position) the stronger of the two outputs, to be fed to the combiner circuit.

With the channel switch in the diversity position, the mark output from section one is connected to the mark output of section two, and the space output of section one is connected to the space output of section two. Diversity combining takes place as follows. If section one mark output has a stronger mark signal than section two, the stronger positive output of section one will make the cathode of V11 more positive, cutting it off. This effectively disconnects the mark contribution of section two as long as section one has the stronger signal. The same action takes place for the space signals, and the resulting mark and space signals are combined across the combining network R38, R39, R40, C13, C14, and C15.

The diversity combining process is further enhanced by the action of the Automatic Diversity Control circuit associated with V21. In this circuit, an audio signal is taken from the plate of dc amplifier V2 in section one, rectified by CR9 and applied to the grid of V21A as a positive-going bias. V21A, in conjunction with a portion of R115, forms a bleeder network controlling the plate voltage of limiter amplifier V10 of section two. Therefore, a strong signal at the plate of V2 of section one will reduce the plate voltage and consequently the output of V10 in section two. In like manner, an audio signal from V8 in section two is used to control the output of the limiter V4 in section one via V21B. The overall result of this action is to amplify any amplitude differences that exist between the two channels. A section of S2 is used to disable this circuit when diversity operation is not desired.

Getting back now to the main signal path, the combined mark-space output is taken from the junction of R38 and R39 and fed to the grid of a cathode follower, which drives a low-pass filter selected by speed switch S3. For keying speeds of less than 60 wpm, S3 is placed in the "slow" position; for faster speeds the switch is placed in the "fast" position.

The output of the low-pass filter cathode-drives the ac amplifier V13B. V14 and V15 act as a dc restorer, assuring center slicing of the dc amplifier output. This is accomplished by charging C17 with positive signal elements through ½ of V14 while C18 is charged with negative signal elements through the other ½ of V14.

The average value of these positive and negative elements is obtained at the junction of R83 and R82. This average value, through V15B, determines the bias on V16A. Variation of the bias on the dc amplifier will compensate for long-term dc-level variations due to the ac coupling.

The output of V16A is directly coupled to the dc keyer V17. The dc keyer is an amplifier which is directly coupled to a buffer which in turn cathode-drives the amplifier. This regeneration results in a trigger action.

The sense switch, S4, selects either the output of the dc keyer or the output of the inverting amplifier to drive the Electronic Relay.

The Electronic Relay operates as follows: V18 is an electron-coupled oscillator, oscillating from screen to first grid with C23, L1, C24, C25 determining the frequency of operation. The frequency of oscillation is in the order of 400 kilocycles.

The plate output of V18 is gated on and off by the output of V16B or V17B as selected by the sense switch. T7 is tuned to the oscillator frequency (both primary and secondary). The output circuit is electrically isolated from the converter, since the only coupling between V19 and V20 and the converter is the magnetic coupling between primary and secondary of T7. The secondary is rectified and filtered in a voltage-doubler circuit consisting of CR13, CR14, C29, C30. The output of the voltage-doubler circuit is amplified by V19 which in turn drives V20. The current through V19 is 180° out of phase with current through V20. This combination of V19 and V20 therefore presents a balanced load on the output power supply, Fig. 9-7.

The output current, controlled by varying the screen voltage of V19 and V20 with R111, flows through output meter M3 and is available at either the rear terminal strip, E3, or the output jack, J4.

## 2.3 Channel Shifter

In our discussions of the SSB transmitter and receiver we spoke of two channels on each side of the suppressed carrier frequency. This is called 4-channel independent-sideband operation. We have seen that, through multiplexing, each one of the communication channels can handle several teletype or other data circuits.

What we want to explain here is how the information in the second channel of the upper and lower sidebands is placed 3 to 6 kilocycles away from the suppressed carrier frequency. Also, we want to explain how the information in the outside channels, after being demodulated at the receiver, is converted to frequencies between 0 and 3 kilocycles.

A device that performs these conversions is called a channel shifter. To explain the operation of the channel shifter, let's start with the two audio channels for the upper sideband entering the channel shifter on separate lines, as shown in Fig. 9-8. The audio signal from line 2 goes to an amplifier and low-pass filter; then into the combiner.

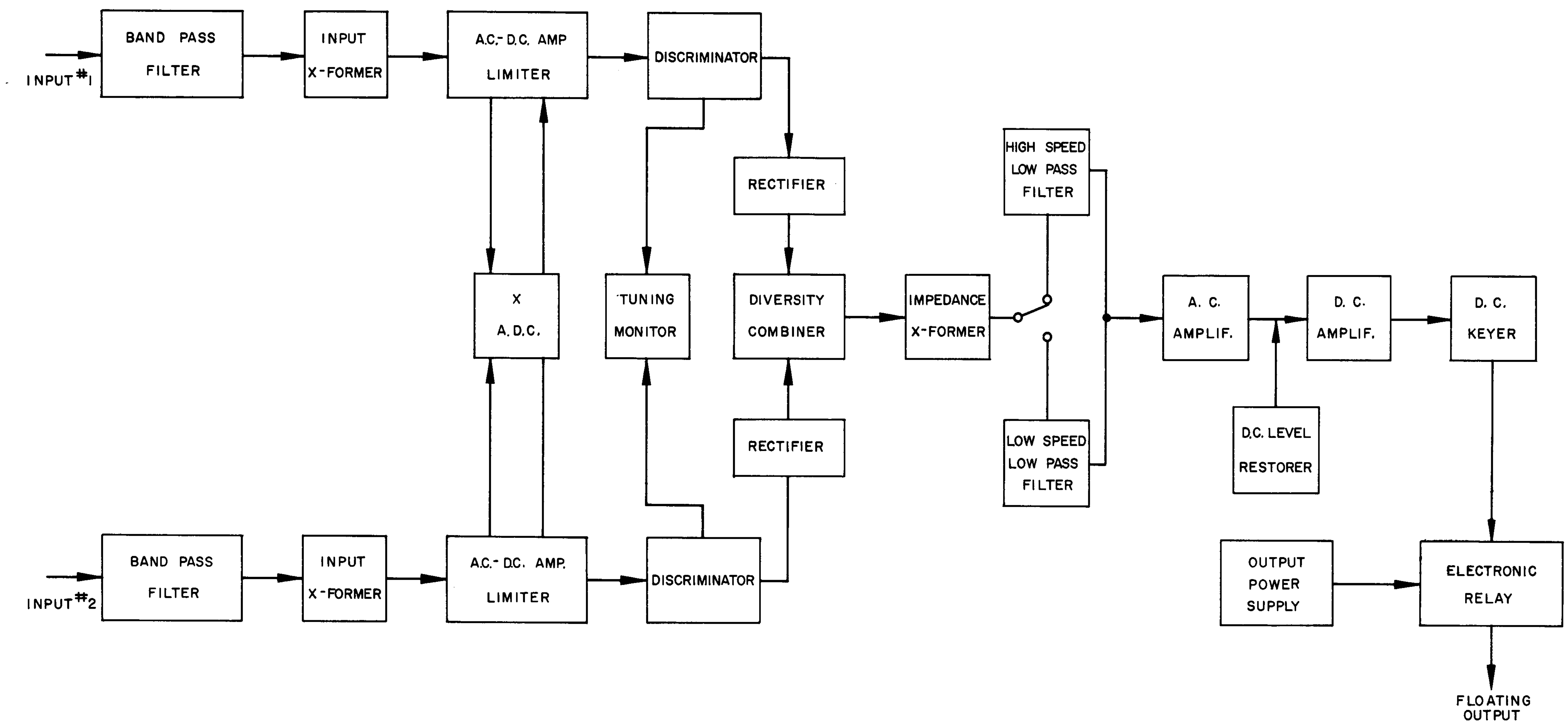



FIGURE 9-5. Block diagram F. S. Diversity Converter, Type 174, Model 1.

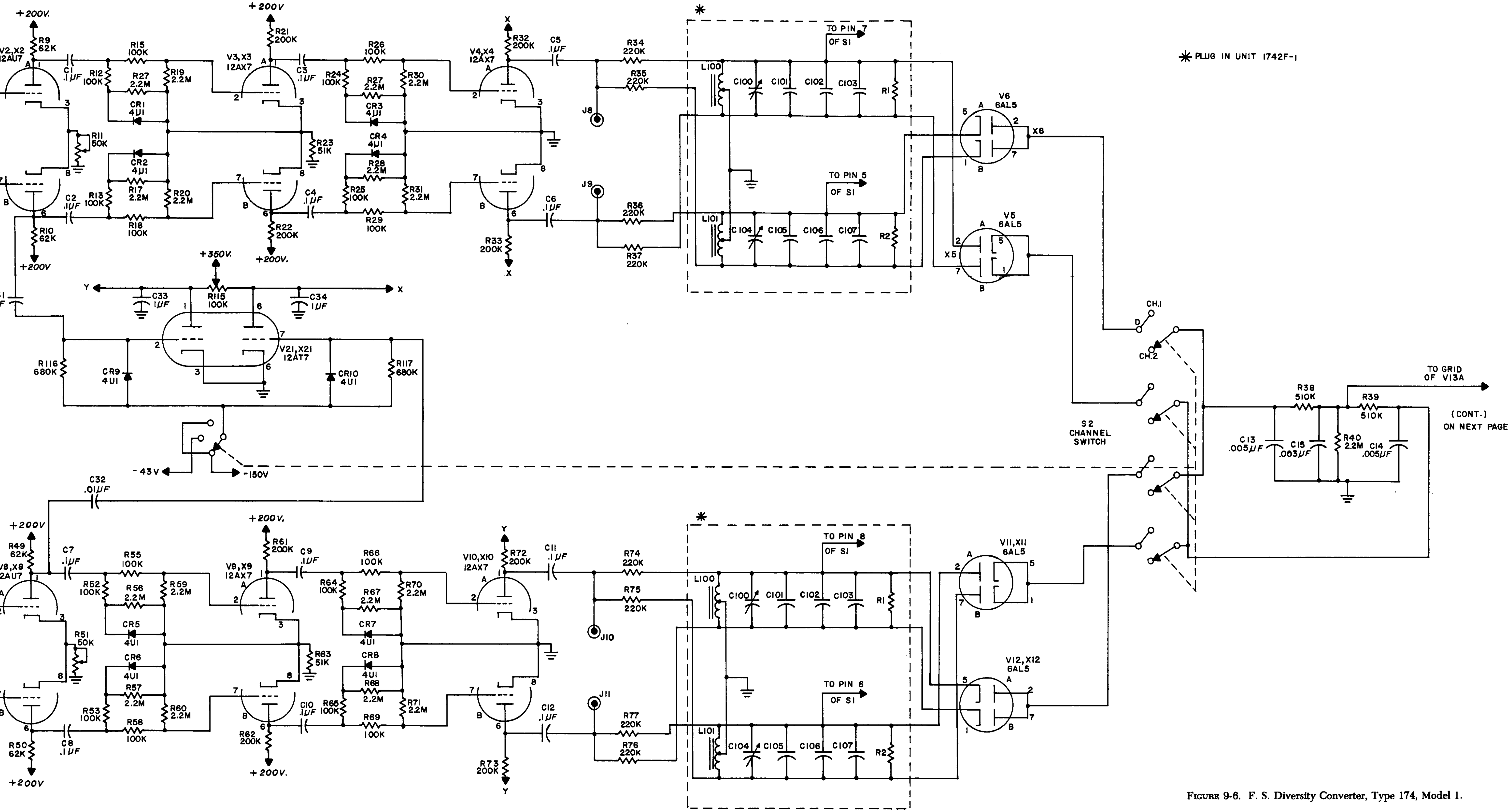
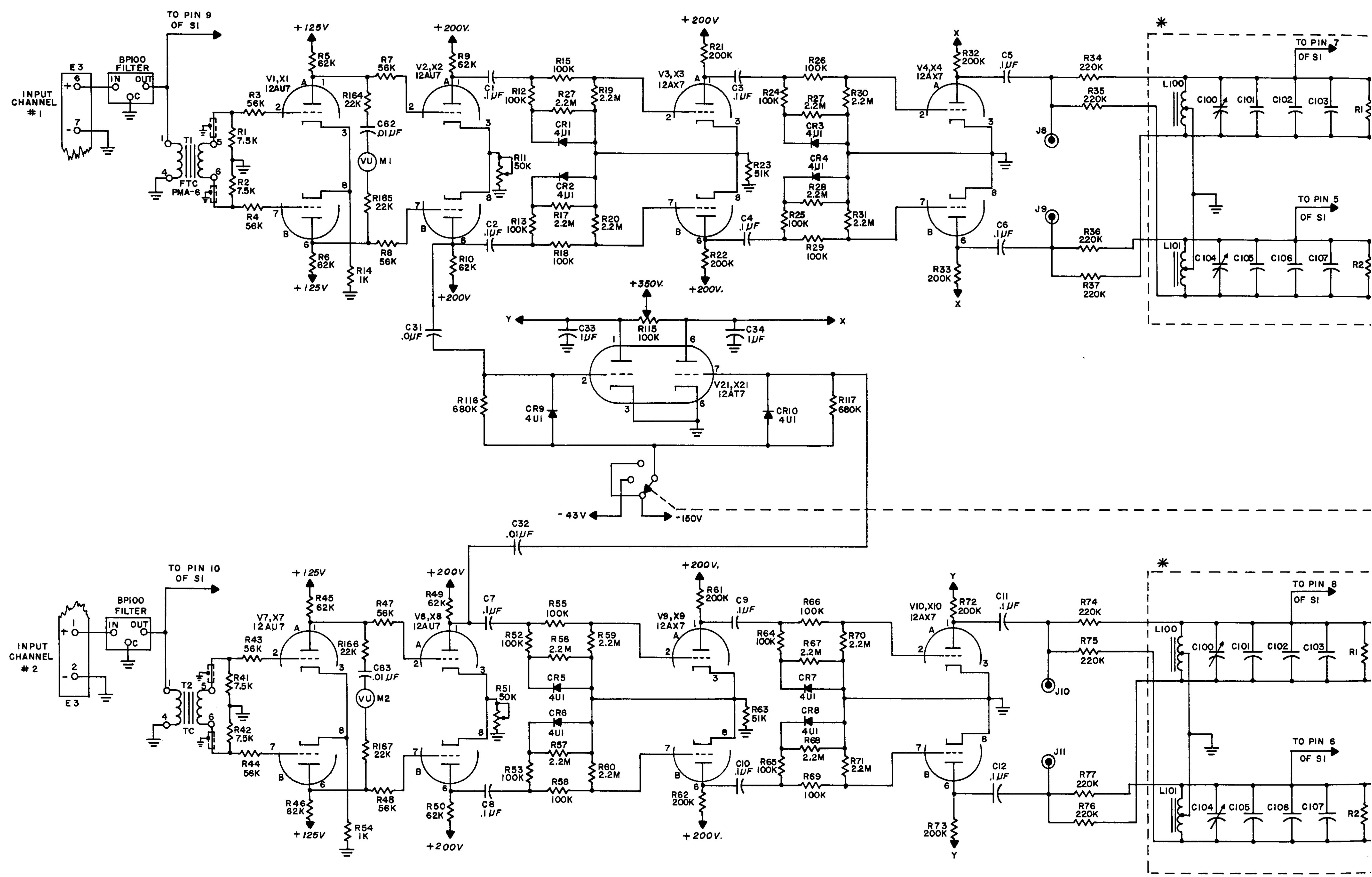



FIGURE 9-6. F. S. Diversity Converter, Type 174, Model 1.

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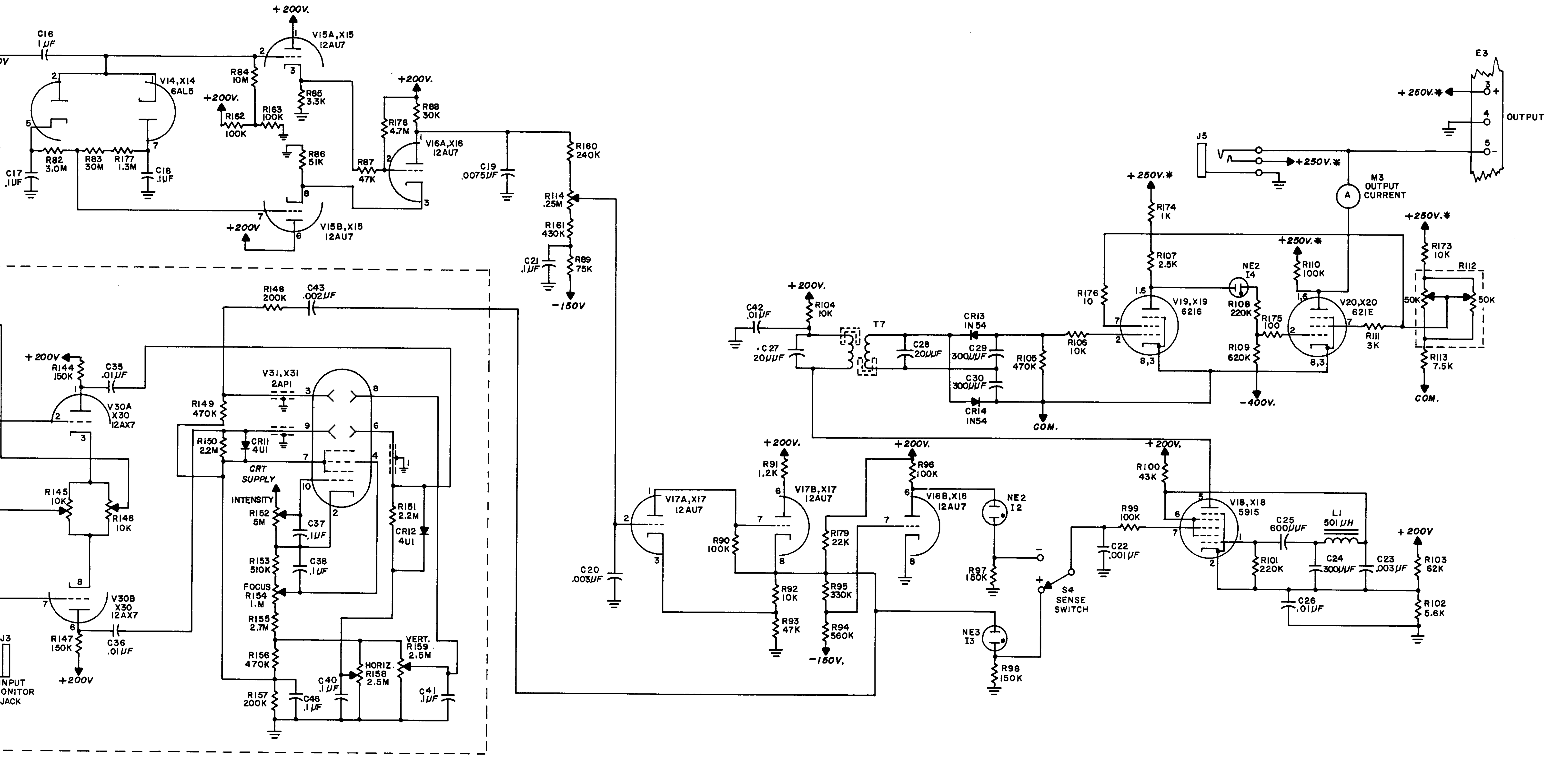
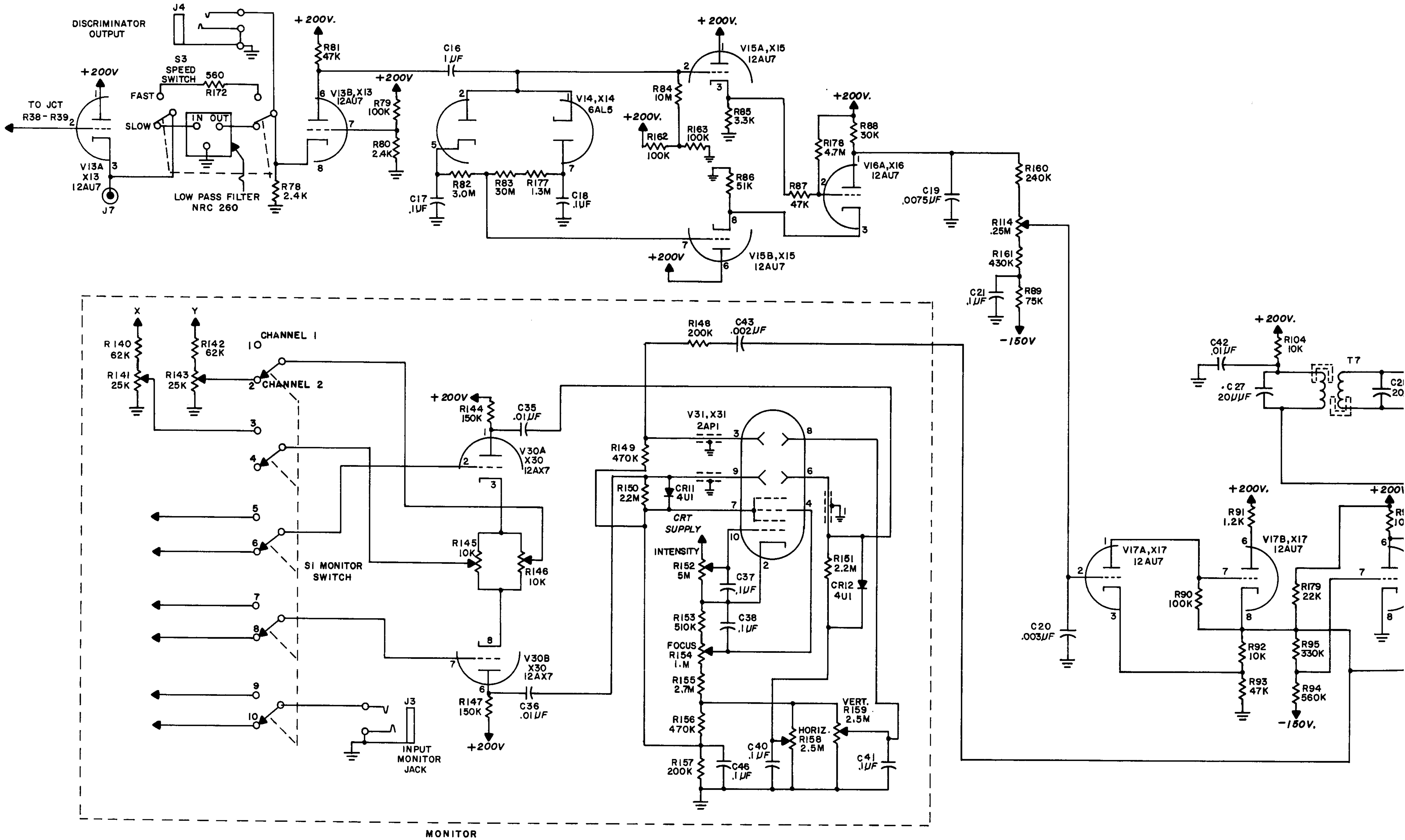



FIGURE 9-6 (Cont.) F. S. DIVERSITY CONVERTER

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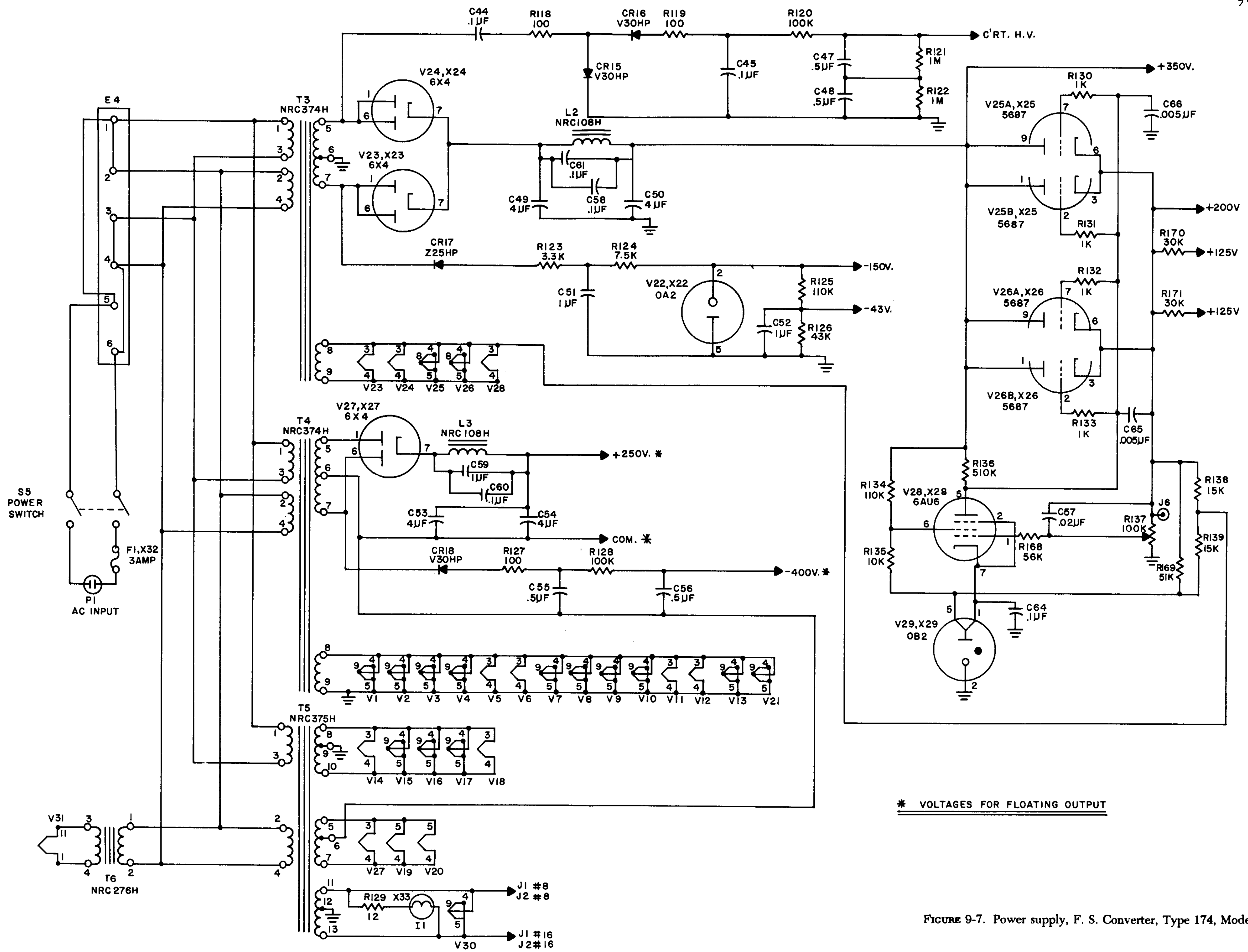



FIGURE 9-7. Power supply, F. S. Converter, Type 174, Model 1.

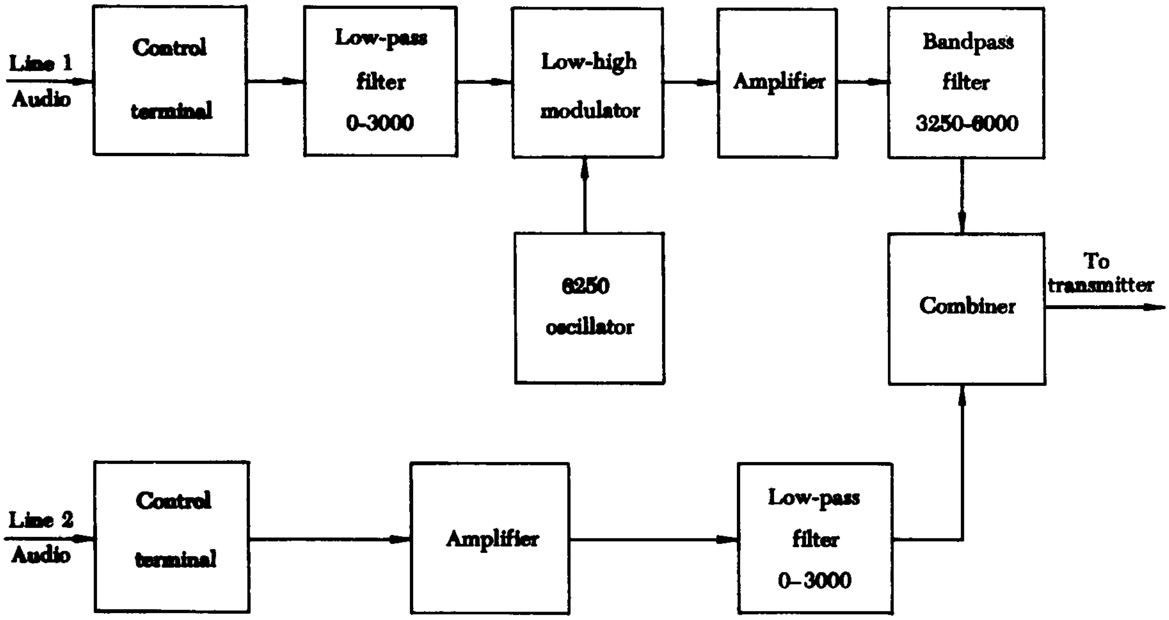



FIGURE 9-8. Simplified block diagram channel shifter.

The audio from line 1 is passed through a low-pass filter to remove all frequencies above 3000 cps. It is then heterodyned against the 6250-cps oscillator signal. A band of frequencies from 0 to 3000 cps will then produce a band of frequencies from 6250 to 3250 cps. As you can see, these are the difference frequencies produced by the heterodyning process. This inverted signal is then amplified and passed through a bandpass filter with a response of 3250 to 6000 cps to eliminate the original audio, the upper sideband, and the 6250-cps oscillator frequency. It, then, combines with the line 2 unshifted signal in the combiner to form the audio to be applied to the modulator and subsequently transmitted as the two upper-sideband channels.

The process used to restore the 3250- to 6250-cps channel to its proper 0- to 3000-cps perspective is just the inverse of the previous procedure. The audio output of one demodulator at the receiver site contains the two demodulated channels from the upper sideband. One channel contains frequencies between 0 and 3000 cps and the other contains frequencies between 6000 and 3250 cps (inverted). Let us now consider the conversion of the channel coming from the receiver site that has frequencies between 3 and 6 kilocycles to frequencies between 0 and 3 kilocycles. The block diagram of Fig. 9-9 indicates the main functions performed on the incoming signals.

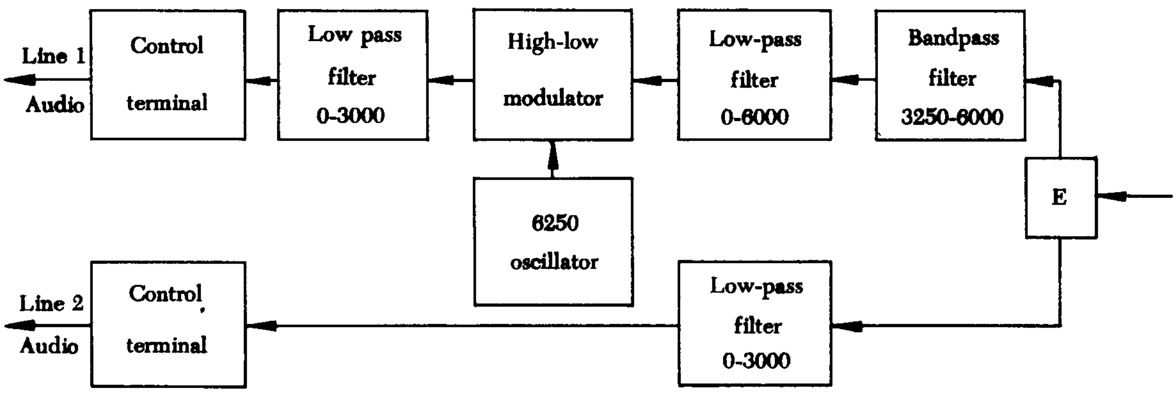



FIGURE 9-9.

The two channels are fed to a system similar to that used at the transmitter site. The line 2 audio is of the proper frequency and therefore is fed to the frequency-shift converter terminal equipment via link or lines. The other channel must be shifted from the 6000–3250-cps range to the 0–3000-cps range. It, then, progresses through a 3250–6000-cps band-pass filter to a mixer to be beaten against the 6250-cps injection frequency. The difference frequency is then passed through a low-pass filter, amplified, and sent over lines or link to the frequency-shift converters at the IFSS Control.

Note that the one set of equipments at the transmitter and the other set at the receiver site handle only the two channels in one sideband. A similar set of equipments is needed for the other sideband.

ATTENTION

Examination 8 is to be worked at this time. This exam covers the material in Chaps. 8 and 9. It is included with the material.

## ANSWERS TO PRACTICE PROBLEMS

### PRACTICE PROBLEMS 2-1

1. C.
2. B.
3. A.
4. D. To explain:

$$E_{cm} = 1.414 \times E_c$$

$$m = \frac{E_{sm}}{E_{cm}}$$

$$E_{sm} = mE_{cm}$$

$$= m(1.414 \times E_c)$$

$$= 0.15(1.414 \times 2000)$$

$$= 424 \text{ v}$$

5. 500 w.
6. 200 v rms, 745.2 Kc.
7. 175 v.
8. 0.06125.
9. 1.
10. 93.75.
11. $100 \sin\omega_c t - 25 \cos(\omega_c + \omega_a)t + 25 \cos(\omega_c - \omega_a)t$
12. 0.5.

13. Peak-envelope power, PEP, is defined as the rms power developed at the crest of the modulation envelope. Stated mathematically:

$$PEP = \frac{\left[0.707(E_{sm} + E_{cm})\right]^2}{R}$$

## PRACTICE PROBLEMS 3-1

1. B.
2. D.
3. B.
4. $e_g = (\sin\omega_c t - \sin\omega_a t)$ ; therefore,

$$i_p = k_o + k_1(\sin\omega_c t - \sin\omega_a t) + k_2(\sin\omega_c t - \sin\omega_a t)^2$$

$$= k_o + k_1(\sin\omega_c t - \sin\omega_a t) + k_2(\sin^2\omega_c t - 2\sin\omega_c t\sin\omega_a t + \sin^2\omega_a t)$$

$$= k_o + k_1(\sin\omega_c t - \sin\omega_a t) + k_2\sin^2\omega_c t - 2k_2\sin\omega_c t\sin\omega_a t$$
$$+ k_2\sin^2\omega_a t$$

Using the identities:

$$\sin^2 A = \tfrac{1}{2}(1 - \cos 2A)$$

and,

$$\sin A \sin B = \tfrac{1}{2}\left[\cos(A-B) - \cos(A+B)\right],$$

the equation becomes,

$$i_p = k_o + k_1(\sin\omega_c t - \sin\omega_a t) + \frac{k_2}{2}(1 - \cos 2\omega_c t) - \frac{2k_2}{2}\left[\cos(\omega_c - \omega_a)t\right.$$
$$\left. - \cos(\omega_c + \omega_a)t\right] + \frac{k_2}{2}(1 - \cos 2\omega_a t)$$

4. (continued)

$$= k_o + k_1(\sin\omega_c t - \sin\omega_a t) + \frac{k_2}{2} - \frac{k_2}{2}\cos 2\omega_c t - \frac{2k_2}{2}\cos(\omega_c - \omega_a)t$$

$$+ \frac{2k_2}{2}\cos(\omega_c + \omega_a)t + \frac{k_2}{2} - \frac{k_2}{2}\cos 2\omega_a t$$

Then, by grouping, this becomes

$$i_p = \underbrace{(k_o + \frac{k_2}{2} + \frac{k_2}{2})}_{\text{dc term}} + \underbrace{k_1(\sin\omega_c t - \sin\omega_a t)}_{\text{Linear term}} - \underbrace{\frac{k_2}{2}(\cos 2\omega_c t + \cos 2\omega_a t)}_{\text{Second harmonic term}}$$

$$- \underbrace{\frac{2k_2}{2}\left[\cos(\omega_c - \omega_a)t - \cos(\omega_c + \omega_a)t\right]}_{\text{Sideband term}}$$

## PRACTICE PROBLEMS 4-1

1. C.
2. 3,750,000 cycles.
3. B.
4. D.
5. C.
6. Lower sideband.
7. 19751.5 kilocycles.

The audio phase difference network given on page 4-19, Fig. 4-14 will be analyzed for the special case of a 1000-cycle input signal. A more general analysis (say for inputs between 300 and 3000 cycles) would be complicated considerably.

First, Fig. A-1 is divided into four impedance groups, $Z_1$, $Z_2$, $Z_3$, and $Z_4$, as shown. The values are given in Fig. A-2. $Z_1 + Z_2$ and $Z_3 + Z_4$ are listed in Fig. A-3. These impedance values are used throughout this analysis. Figures A-1, A-2, and A3 are on pages 8 and 9.

Two assumptions are made to further simplify the analysis; (1) The input signal is 1000 volts at a zero-degree phase angle. (2) The two output voltages, $E_1$ and $E_2$ feed into high-impedance loads. With these two assumptions, we can write two equations. (Remember, our final goal is to solve for amplitude and phase relationships of the two output voltages, $E_1$ and $E_2$.)

The value of the top portion of the potentiometer is designated as R; the lower portion will then be 500 - R. Then for the loop consisting of R, $Z_1$ and $E_1$ as shown by the arrow in Fig. A-2, we have the voltage equation

$$E_1 = \frac{(1000\underline{/0^\circ})R}{500} - E_{Z_1} = 2R\underline{/0^\circ} - E_{Z_1} \qquad \text{(A-1)}$$

The loop consisting of R, $Z_3$, and $E_2$, as shown by the arrow gives the equation

$$E_2 = \frac{(1000\underline{/0^\circ})R}{500} - E_{Z_3} = 2R\underline{/0^\circ} - E_{Z_3} \qquad \text{(A-2)}$$

Applying the voltage divider principle,

$$E_{Z_1} = \frac{1000\ Z_1}{Z_1 + Z_2} = 636 + j164.$$

Several steps in the last calculation were left out.

Also $E_{Z_3} = \dfrac{1000\ Z_3}{Z_3 + Z_4} = 608 - j147$

Substituting the values of $E_{Z_1}$ and $E_{Z_3}$ into Eqs. (A-1) and (A-2) gives

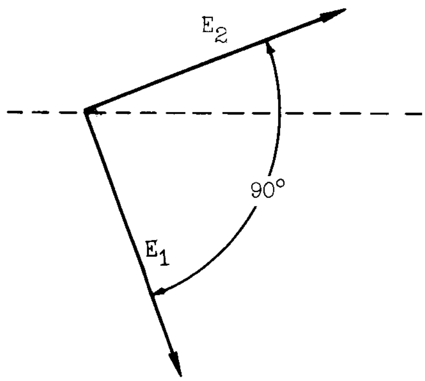
$$E_1 = 2R\underline{/0°} - (636 + j164)$$

$$E_2 = 2R\underline{/0°} - (608 - j147)$$

$$E_1 = 2R - 636 - j164 \qquad \text{(A-3)}$$

$$E_2 = 2R - 608 + j147 \qquad \text{(A-4)}$$

Our circuit is a 90° phase-difference network; thus we can draw a vector diagram for $E_1$ and $E_2$ as shown below.




Because there is a 90°—phase difference between the two voltage vectors, the tangent of the phase angle of one vector is equal to the cotangent of the phase angle of the other vector.

Or

$$\frac{2R - 636}{164} = \frac{147}{2R - 608}$$

This allows us to compute the value of R that will give a 90°—phase difference.

In solving for R we arrive at the quadratic equation

$$R^2 - 623R + 90,580 = 0$$

$$R = \frac{623 \pm \sqrt{389000 - 362,320}}{2}$$

$$= \frac{623 \pm 163}{2}$$

$$= 393$$

or $$R = 230$$

Each of these values will give a 90°—phase difference. We are also concerned with equal amplitudes. Substituting 393 into Eqs. (A-3) and (A-4) gives

$$E_1 = 786 - 636 - j164$$

$$= 150 - j164$$

$$= 222.4\underline{/-47.55°}$$

$$E_2 = 786 - 608 + j147$$

$$= 230.8\underline{/+ 39.55°}$$

Substituting 230 for R in Eqs. (A-3) and (A-4) gives

$$E_1 = 460 - 636 - j164$$

$$= 240.5\,\underline{/223°}$$

$$E_2 = 460 - 608 + j147$$

$$= 208.4\,\underline{/135.2°}$$

The value of 393 for the top portion of the potentiometer gives outputs of approximately equal amplitudes with a phase difference of 90°.

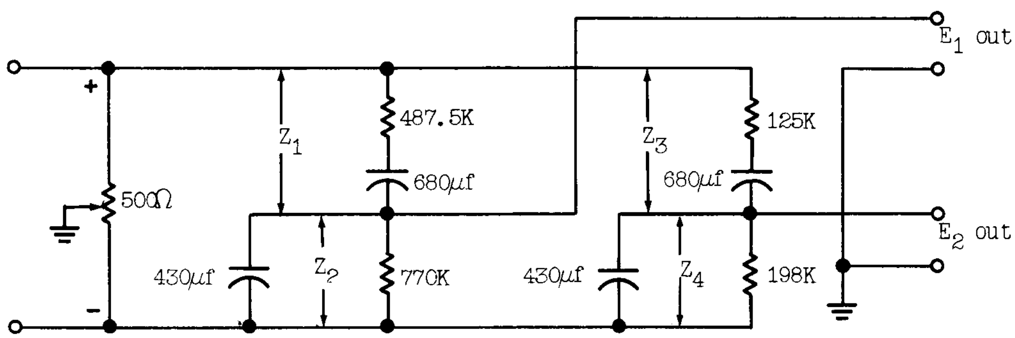



Fig. A-1.

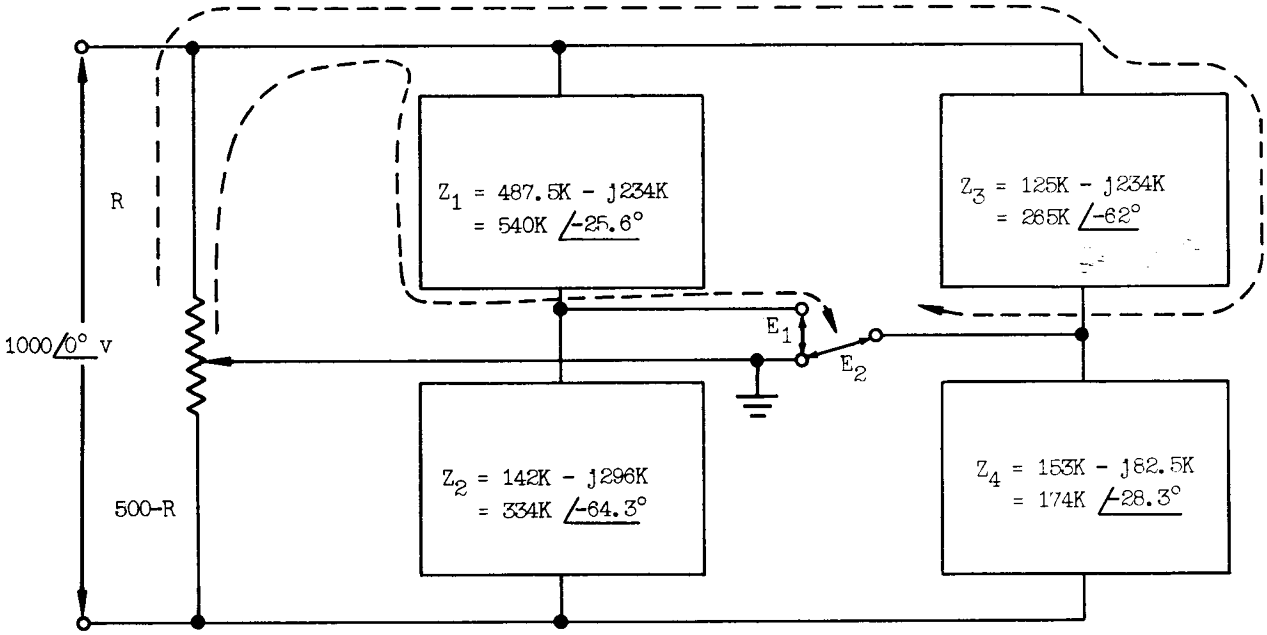

R
500-R
1000 /0° V
$Z_1$ = 487.5K - j234K
= 540K /-25.6°
$Z_3$ = 125K - j234K
= 265K /-62°
$E_1$
$E_2$
$Z_2$ = 142K - j296K
= 334K /-64.3°
$Z_4$ = 153K - j82.5K
= 174K /-28.3°

Fig. A-2.

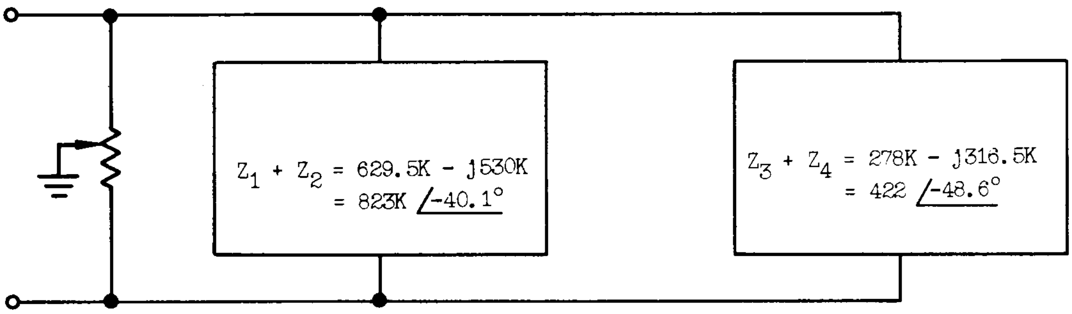

$Z_1$ + $Z_2$ = 629.5K - j530K
= 823K /-40.1°
$Z_3$ + $Z_4$ = 278K - j316.5K
= 422 /-48.6°

Fig. A-3.