THESIS

GÜRŞT UNCA

PREFACE

This thesis was prepared in accordance with the requirements of the Engineering Graduate School program of Robert College. The subject of the thesis is Single Sideband Communication Systems. This is almost a new field in communications and gaining wide interest because of its advantages over the other communication systems.

This thesis consists of nine chapters; first two are about the general characteristics of the SSB system, third is about SSB generation, the succeeding two about SSB generators and receivers, and the following two about frequency control techniques of the SSB systems. Chapter 8 is devoted to the consideration of Linear Power Amplifiers in SSB systems. Since it is very important to maintain linearity in SSB systems, much space has been saved for this purpose. In the last chapter, two special linear power amplifiers are given in detail.

I want to express my deep gratitude to Prof. Dr. Mustafa Santur for his valuable suggestions and urgings in preparation of this thesis, Mr. M. Hananel and Mr. D. Dall for their great help in practical field and supplying me with the necessary documents. I am also indebted to Miss Ayşe Dinç who read the thesis through in its elementary form and corrected the grammatical mistakes and helped in printing.

June 15, 1965 GÜR TUNCA
Bakırköy, Istanbul

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CHAPTER 1

MODULATION SYSTEMS IN COMMUNICATIONS

Before passing directly into Single Sideband (SSB) theory, I thought it worthwhile to spend a little time on different type of modulation systems in communications, comparison of their performances, advantages and disadvantages, although quite a large space is saved for the comparison of single sideband with other modulation systems in the next chapter.

Therefore, this chapter is a summary of what we have seen in our Electronics courses.

1.1. Modulation Spectra and Spectral Bandwidth

If a carrier wave which can be expressed as a rotating vector by

$$i = Ie^{j\theta_o} = Ie^{jw_o t}$$

is used in the transmission of intelligence, then the intelligence may be represented by the modulation of the amplitude, phase, or frequency of the rotating vector. Carrier modulation is used as a means of transmitting intelligence because it utilizes the propagation properties of h-f and u-h-f electromagnetic waves and because the intelligence may be recovered from the modulated wave.

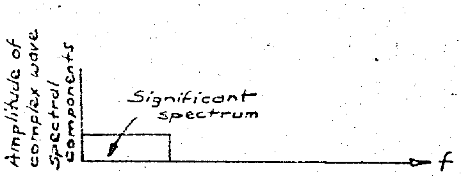
Although the primary requirement of the modulation process is frequency conversion, the additional purposes for modulating can be stated as follows:

1. The signal frequencies are changed to be transmitted to a preassigned band, and this change of the signal frequencies can be repeated as often as necessary by the process of remodulating.
2. We know that the superposition of a carrier wave onto an audio signal causes the transfer of the audio signal band into the radio signal band with the sideband products around the carrier frequ-

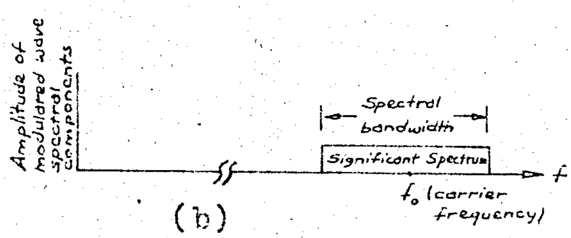
ency. At these frequencies the design of reasonably small and efficient radiators, as well as the design of high-power amplifiers, is simplified.

3. The resulting bandwidth may be increased or decreased over that of the original signal to be transmitted, depending on the type of modulation process used. It is also known that the signal-to-noise ratio at the receiver is a function of the bandwidth of the modulated signal (*). Therefore the modulation process provides an additional improvement in received signal-to-noise ratio.

As I have indicated in the second statement above, when a carrier is modulated in any known fashion, spectral components which are called side frequencies appear in the vicinity of the carrier. It is important to note here that side frequencies are often referred to as sidebands. But it is preferable to use the term sideband only when just one pair of side frequencies is present. The group of side frequencies above and below the carrier frequency are called the "upper sideband" and "lower sideband" respectively. The frequencies in these two bands may or may not undergo a change in amplitude and phase. They are propagated with the carrier to the receiver which is a device in which demodulation process takes place. This process is just the reverse of the modulation process, and means to separate the audio signal from the carrier. The general form of modulated spectra is given in Fig.1(a) and (b).



(a)



(b)

FIG.1

(*) Chapter 5, An Introduction to the Principles of Communication Theory, John C. Hancock, McGraw-Hill Book Company, Inc., 1961.

1.2. Modulation Systems

The five most widely used modulation systems in electrical communications are given below:

1. Amplitude modulation (double sideband)
   a. Amplitude modulation-double sideband/suppressed carrier
   b. Amplitude modulation-double sideband
2. Amplitude modulation (single sideband) (*)
3. Frequency modulation
4. Pulse modulation
5. Multiplex systems

Phase modulation is not listed above since it has very limited application in communications.

In commercial sound and television transmission, the following maximum spectral bandwidths are assigned:

A. Sound broadcasting
   1. Amplitude modulation--10 kc
   2. Frequency modulation--150 kc for commercial frequency modulation, 50 kc for television

B. Television broadcasting
   Video--single-sideband amplitude modulation--6 Mc
   Sound--frequency modulation--6 Mc

1.3. Amplitude Modulation (AM)

The main process in an amplitude-modulating system is the multiplication of the basic information source by a carrier wave, that is, $A(t)\cos w_c t$. Therefore, the purpose of the transmitter is to achieve this frequency translation, and the purpose of the receiver is

(*) Also termed "vestigial-sideband amplitude modulation" since complete elimination of one sideband could not be achieved up to recent times.

to perform the inverse operation so as to recover the information source. Frequency domain representation of an amplitude modulation-double sideband-suppressed carrier system is shown in Fig.2, where F(f) represents the assumed Fourier transform of the information or modulation source, X(f) represents the translated spectrum, that is, the spectrum of $e_m(t)\cos w_c t$. The conversion of X(f) in the receiver is achieved by multiplying $e_m(t)\cos w_c t$ by $\cos w_c t$. The resulting spectrum is denoted by Y(f), and the recovery of $e_m(t)$ may be accomplished by filtering. The block diagrams of the

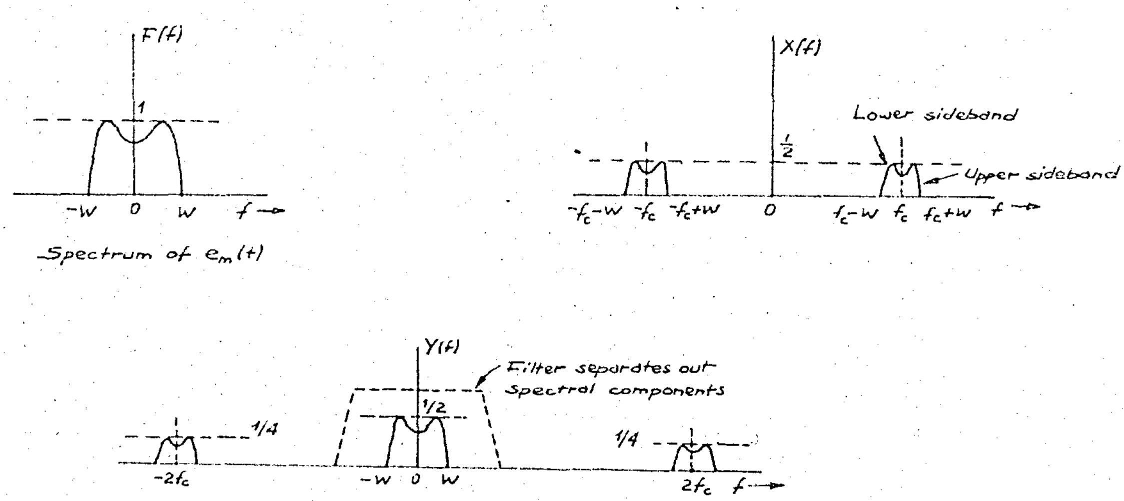


FIG.2

actual system is given in Fig.3. This system is termed "amplitude-modulated, double-sideband, suppressed carrier" (AM-DSB/SC).

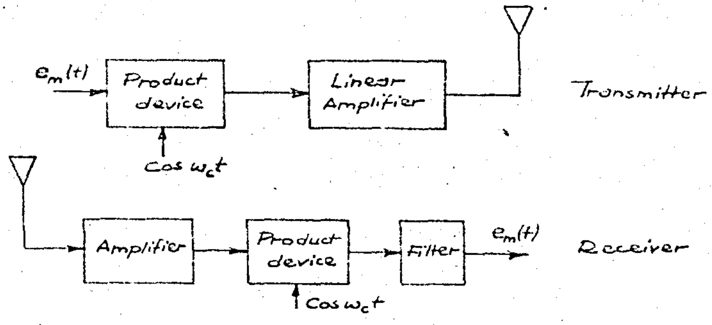


FIG.3

The main disadvantage of this system, as with the case of single sideband, suppressed carrier system, is the necessity of knowing the carrier frequency $\cos w_c t$ at the receiver. For the detection purpose, a carrier wave is used in the receiver which is in synchronism with the one that is used in the transmitter. For the synchronization purpose, a pilot carrier is transmitted.

Second system of amplitude modulation utilizes only one of the sidebands around the carrier frequency in transmission instead of transmitting the total spectrum about the carrier frequency f_c. Since this is the subject of this thesis, detailed description of the system will follow in the next chapter, except a brief discussion given here. The system is illustrated in Fig.4. The

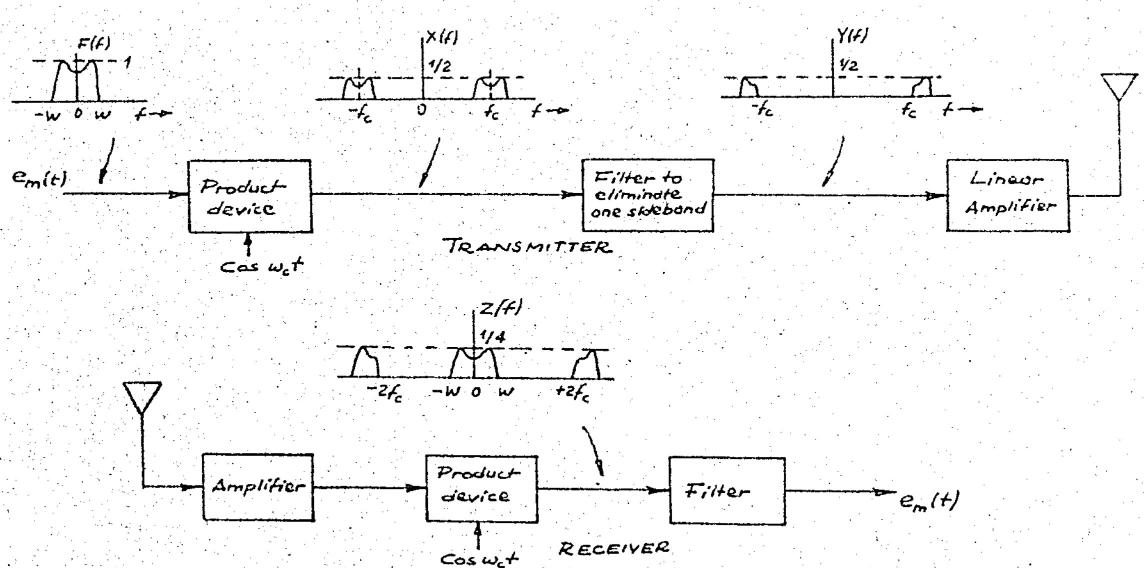


FIG.4

advantage of the system over the AM-DSB/SC is that the required channel bandwidth is reduced by one-half. But the system has the same disadvantage of coherent detection.

Since coherent detection is the main problem in the two modulating systems above, another system which didnot necessitate the coherent detection would be convenient to devise. As a starting point, it must be stated that an AM-DSB/SC wave has information contained in both the amplitude and phase. This is shown in Fig.5, where $e_m(t)$ is assumed sinusoidal. Therefore, coherent detection is necessary to extract the information in phase and amplitude. The sign change in $e_m(t)$ causes the carrier to change phase. Thus, the first way to be thought of is to modulate the carrier by a source that does not change sign. This can be achieved by preventing $e_m(t)$ from going negative by adding a bias component. The process is indicated in Fig.6. This type of modulation is termed amplitude modulation, double sideband (AM-DSB). Now all the information is in the amplitude and an amplitude-sensitive device may be used to recover $e_m(t)$. Therefore, the advantage of this type of modulation over the other two is that the ease with which the reception occurs. But the disadvantage is that, the carrier component consumes a large amount of power. A system block diagram is given in Fig.7.

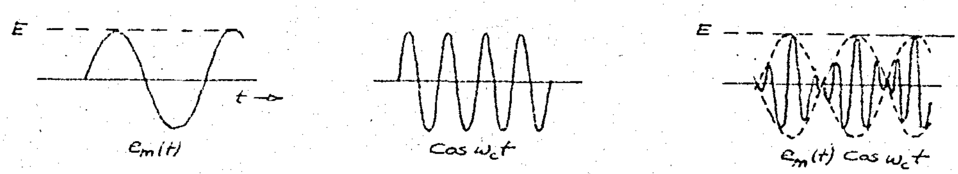


FIG.5

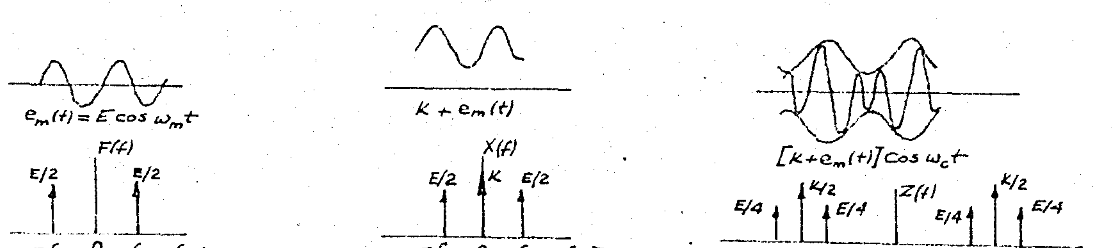


FIG.6

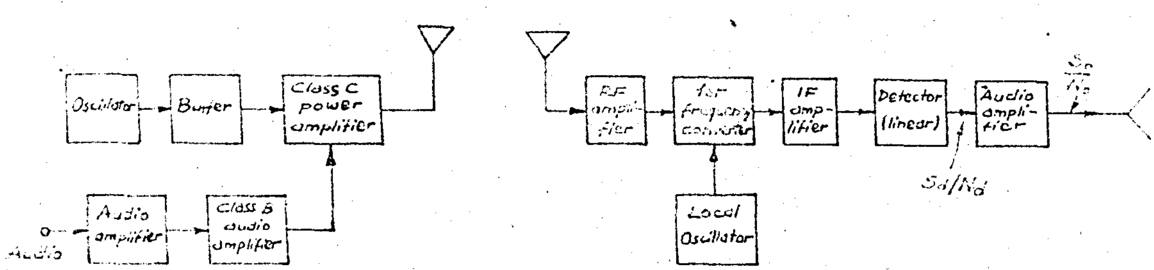


FIG.7

1.4. Comparison of AM Systems

A concise form of comparison is given below only by formulae:

1. AM-DSB

$$e(t) = E_c(1 + m_a \cos w_m t) \cos w_c t$$

$$P_t = (E_c^2/2) + (E_m^2/4) = P_c\left(1 + \frac{m_a^2}{2}\right)$$

$$P_{SB} = E_m^2/4$$

$$S_o/N_o = (E_m^2/2)/4\eta f_m = 2P_{SB}/N_{in} = 2S_{in}/N_{in}$$

2. AM-DSB/SC

$$e(t) = E_m \cos(w_c - w_m)t + E_m \cos(w_c + w_m)t$$

$$P_t = E_m^2 = P_{SB}$$

$$S_o/N_o = (E_m^2/2)/\eta f_m = 2P_{SB}/N_{in} = 2P_t/N_{in} = 2S_{in}/N_{in}$$

3. AM-SSB

$$e(t) = E_m \cos(w_c - w_m)t \quad \text{(lower sideband)}$$

$$P_t = E_m^2/2 = P_{SB}$$

$$S_o/N_o = (E_m^2/2)/2\eta f_m = P_{SB}/N_{in} = S_{in}/N_{in}$$

where P_t = total transmitted power,
P_{SB} = sideband power,
S_o/N_o = post-detection signal-to-noise ratio,
S_{in}/N_{in} = predection signal-to-noise ratio
η = height of the noise power spectrum.

1.5. Frequency Modulation (FM)

It is convenient to consider the "direct" FM transmitter, although there are several systems for producing FM signals. The block diagram of the transmitter is given in Fig.8. The frequency of the master oscillator or "primary oscillator" is controlled by a reactance tube. The oscillator output frequency is multiplied nine times in two stages by means of two frequency triplers; the resulting wave is impressed on the input of a class C power amplifier which feeds into the antenna. The main problem here is one of center-frequency stabilization (*), and the problem of monitoring is solved by beating a sample of the output signal against a reference signal from a piezo fixed-frequency standard; any center-frequency drift (**) is impressed at the input of a frequency discriminator. The discriminator yields a control voltage, which is a function of the center-frequency drift. This voltage is impressed with suitable polarity into the modulating reactance-tube circuit to compensate for this drift.

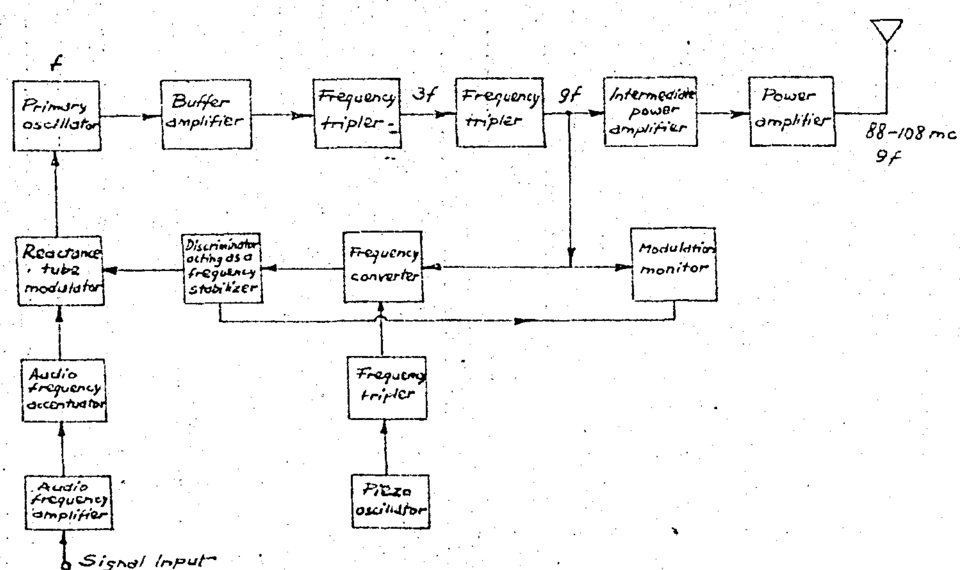


FIG.8

(*) For further details, E.S. Winlund, Drift Analysis of the Crosby Frequency Modulated Circuit, Proc.IRE, vol.29, July, 1941.

(**) For further details, Chapter 22, Harmonics, Sidebands, and Transients in Communication Eng'g., Louis Cuccia, McGraw-Hill.

In the following paragraph, frequency and phase modulations are discussed together under the topic of "Angle Modulation".

1.6. Angle Modulation--Frequency and Phase Modulation

It was seen in amplitude modulation how the magnitude of a sinusoidal wave

$$e(t) = A \sin (2\pi Ft + \emptyset)$$

can be varied, known as modulation, and can be made to carry information. It is also possible to keep the amplitude A constant, and change the argument of the sine function in accordance with the signal to be transmitted.

In "phase modulation", the phase $\emptyset$ in the expression above is varied in accordance with the signal. Thus, if the carrier is phase-modulated by a signal $\cos 2\pi\mu t$, the phase-modulated carrier is of the form

$$e(t) = A \sin[2\pi Ft + (\emptyset_o + \Delta\emptyset \cos 2\pi\mu t)]$$

The quantity $\Delta\emptyset \cos 2\pi\mu t$ in the above expression is called the phase deviation, and its instantaneous value may be expressed in radians. The degree of modulation is usually defined as the ratio of $\Delta\emptyset$ to the maximum phase deviation that the particular transmitting or receiving apparatus of interest at the moment is capable of handling. The degree of modulation in phase modulation is thus not a property of the signal alone, as it is in amplitude modulation, but is also defined in terms of the properties of the system in which it is used.

In "frequency modulation" (*), the instantaneous frequency of the first expression above is varied in accordance with the signal.

(*) For detailed work on Frequency Modulation; Frequency Analysis, Modulation and Noise, Stanford Goldman, McGraw-Hill Book Company, Inc., 1948.

Instantaneous frequency is defined as, when the carrier frequency is very high in comparison with the modulation frequency,

$$\text{instantaneous frequency} = \frac{1}{2\pi}\frac{d\theta}{dt}$$

when the frequency-modulated signal is expressed as

$$a = A \sin \theta$$

If $\theta = 2\pi Ft$, then

$$\frac{1}{2\pi}\frac{d\theta}{dt} = F$$

so that definition agrees with the usual one in case F is a constant.

Making use of the foregoing definition, we can find the form of a frequency-modulated signal, when the modulation is $\cos 2\pi\mu t$. We can write

$$\frac{1}{2\pi}\frac{d\theta}{dt} = F + \Delta F \cos 2\pi\mu t$$

in which F and ΔF are constants. Integration of the above equation yields

$$\theta = 2\pi Ft + (\Delta F/\mu) \sin 2\pi\mu t + \theta_o$$

Thus the frequency-modulated signal is

$$e(t) = A \sin \theta = A \sin \left[2\pi Ft + (\Delta F/\mu) \sin 2\pi\mu t + \theta_o \right]$$

The quantity $\Delta F \cos 2\pi\mu t$ is called the frequency deviation while ΔF is called the peak frequency deviation. The degree of modulation is usually defined as the ratio of ΔF to the maximum permitted frequency deviation allowed by law or as the ratio of ΔF to the maximum frequency deviation of which the system is capable. The definition of degree of modulation, thus does not depend only upon properties of the signal itself but also involves other things such as equipment or statutes. Block diagram of a frequency modulating system is shown in Fig.9.

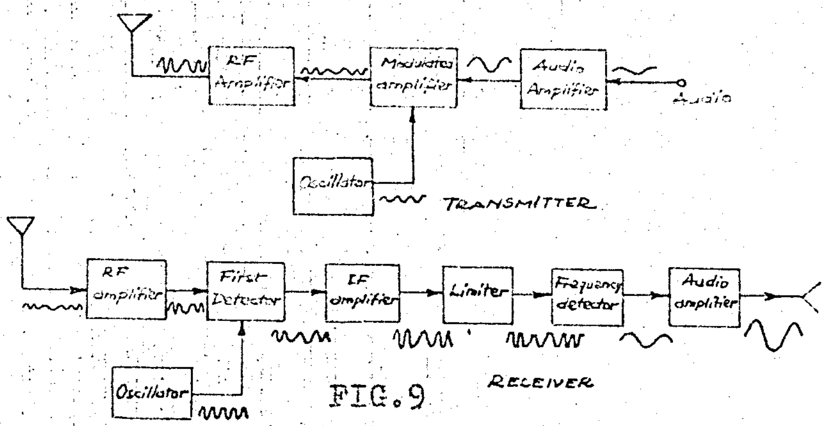


FIG.9

1.7. Pulse Modulation

Communications systems using pulse modulation have found wide application. There are four general kinds of pulse modulation:

1. Pulse-amplitude modulation (PAM)
2. Pulse-width modulation (PWM)
3. Pulse-numbers modulation (PNM)
4. Pulse-phase modulation (PPM)

Pulse-amplitude modulation and pulse-width modulation are illustrated in Fig.10(a) and (b), respectively, where it is seen that the average power involved is a function of the modulating wave, which may be recovered from the pulse-modulated wave by suitable means of demodulations.

(a) (b)

FIG.10

There are several excellent reasons for employing pulse-modulation systems. As a result of radar research, tubes such as magnetrons and klystrons are available which are capable of high peak power. Since the average power involved is equal to the product of the peak power, the pulse duration time, and the pulse repetition frequency, it is evident that by proper choice of operating parameters the average power demands made by the pulse-modulation transmitter on its power supply may be caused to be relatively small, resulting in economy of equipment, size and weight.

1.8. Multiplexing Systems

Multiplexing is the technique of impressing two or more communication channels on a single circuit so that each can be operated simultaneously without interference with the others. There are two methods of multiplexing; time-division multiplexing and frequency-division multiplexing.

Time-division multiplexing is performed using a system with subcarriers, each of which consists of a series of d-c pulses; each pulse subcarrier has the same frequency, but the pulses are spaced in carrier angle so that they do not overlap although interleaved, thus resulting in a series of pulses. Each pulse subcarrier may be amplitude-modulated, phase-modulated, width-modulated, or numbers-modulated and then used to modulate a carrier.

A frequency-division system is one using a separate subcarrier for each channel with spaced subcarrier frequencies. Subcarriers may be amplitude-modulated, frequency-modulated, or single-sideband-modulated, and a group of subcarriers may be used to modulate a higher frequency carrier.

CHAPTER 2

AN INTRODUCTION TO SINGLE-SIDEBAND COMMUNICATIONS

In the previous chapter, various types of modulating systems were discussed. In this chapter, the single-sideband type of modulation system will be undertaken. After having an introduction to the chapter with the "need for single sideband", the nature of the SSB signal will be determined; then, after stating the advantages and disadvantages of the SSB, and comparing the SSB with other types of modulation systems, economy of the SSB system will be discussed, and a short history of evolution of the system will be given.

2.1. Need for Single Sideband

Many of the communication services, such as ship-to-shore communications, air-to-ground communications, and the many military and naval systems which require independence, mobility, and flexibility can satisfy their propagation characteristics only in the high frequency-range. Since the high-frequency range is limited between 3 and 30Mc frequencies, it is essentially required to make best use of the available space. Therefore, the system used must provide a minimum bandwidth. Furthermore, guard bands between channels to allow for frequency drift and poor selectivity be minimized. To avoid interference, spurious radiation be kept to a very low value.

In addition to those stated above, the advantage of the SSB system accrues principally from the higher efficiency of sideband power generation in the SSB system, and takes the form of reduced size and weight of equipment or more effective communication for a given size and weight. Especially, reduction in size and weight is very valuable from the point of airborne communications.

2.2. The Nature of the SSB Signal

In the previous chapter, it was stated that the SSB modulation system appears after the elimination of one of the sideband products in the amplitude modulation. Therefore, it is reasonable to think that SSB type of modulation is a kind of amplitude modulation system in which some additional filters are used to eliminate one of the sidebands. Thus, SSB communication derives its name from the fact that the spectrum of the signal resembles one of the two sidebands that are created in the familiar process of amplitude modulation. In the amplitude modulation, amplitude of the carrier $e_c(t)$ is varied in accordance with a modulating signal, $e_1(t)$. Therefore, the process taking place is expressed by

$$e_2(t) = [1 + ke_1(t)]e_c(t) \tag{1}$$

in which $e_2(t)$ denotes the resulting amplitude-modulated signal. The operation is indicated in Fig.1(a).

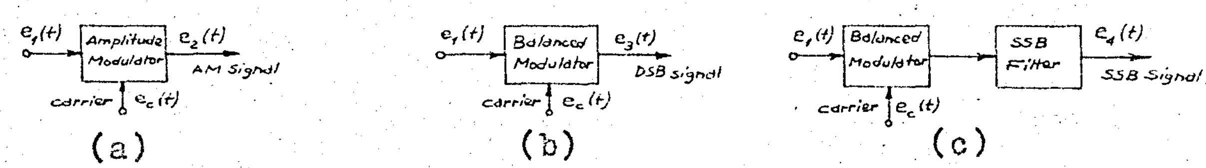


FIG.1

Now, if a carrier wave represented by

$$e_c(t) = \cos(w_c t + \emptyset_c) \tag{2}$$

and a modulating signal which can be described as the sum of a large number of sinusoidal components

$$e_1(t) = \sum_{n=1}^{N} E_n \cos(w_n t + \emptyset_n) \tag{3}$$

are taken and superimposed onto each other, the result is obtained by insertion of (2) and (3) into (1).

$$
\begin{aligned}
e_2(t) = {} & \cos(w_c t + \emptyset_c) \quad \text{carrier component} \\
& + \frac{k}{2}\sum_{n=1}^{N} E_n \cos\left[(w_c - w_n)t + \emptyset_c - \emptyset_n\right] \\
& \qquad \text{lower sideband} \\
& + \frac{k}{2}\sum_{n=1}^{N} E_n \cos\left[(w_c + w_n)t + \emptyset_c + \emptyset_n\right] \qquad (4) \\
& \qquad \text{upper sideband}
\end{aligned}
$$

The manipulation in expression (1) is obvious, since;

$$\cos A \cdot \cos B = \frac{1}{2}\left(\cos\frac{A+B}{2} + \cos\frac{A-B}{2}\right)$$

The expression (4) clearly shows that in the output signal, there is a component due to the carrier frequency while the two bands are symmetrically placed around it. The typical modulating signal as a function of time and of its f spectrum is illustrated in Fig.2(a). The resulting AM signal and its spectrum, showing the two identical sidebands similar to the spectrum of the original modulating signal, are illustrated in Fig.2(b).

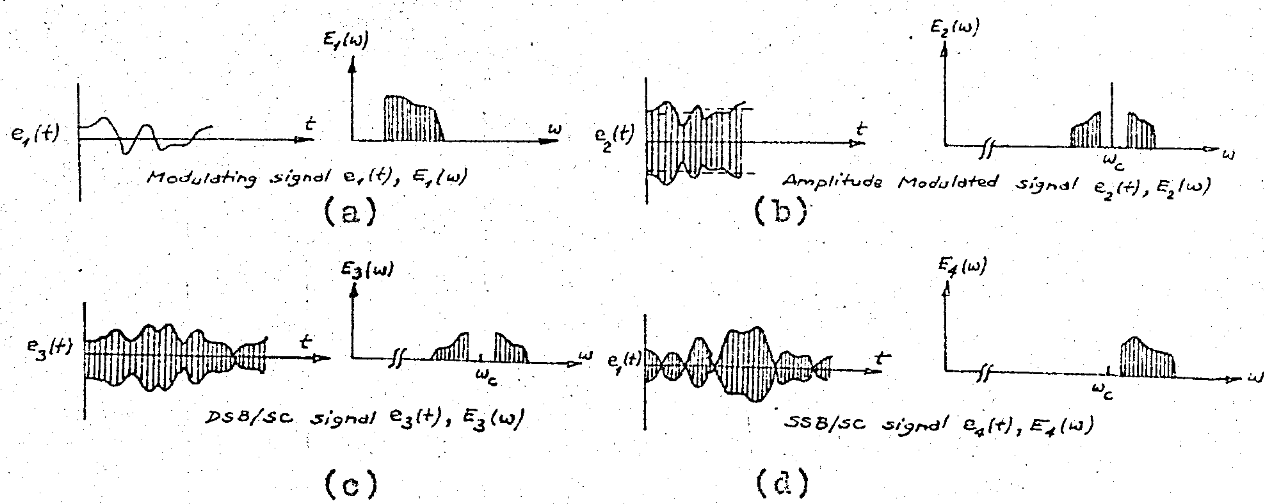


FIG.2

Instead of following the operation given in expression (4), if the carrier wave and the modulating wave are multiplied, the product will be a double sideband signal (DSB) which is identical to the AM signal except that the strong carrier component is not present in the output. This type of signal is illustrated as a function of t and f, Fig.2(c).

$$e_3(t) = e_c(t) \cdot e_1(t) \tag{5}$$

$$\begin{aligned} e_3(t) = {} & \frac{1}{2}\sum_{n=1}^{N} E_n \cos\left[(w_c - w_n)t + \phi_c - \phi_n\right] \quad \text{lower sideband} \\ & + \frac{1}{2}\sum_{n=1}^{N} E_n \cos\left[(w_c + w_n)t + \phi_c + \phi_n\right] \quad \text{upper sideband} \end{aligned} \tag{6}$$

Expression (7) gives a clue to SSB generation. It is very obvious that if one of the sidebands is eliminated by using a SSB filter, a filter which passes only one of the sidebands, a SSB signal is obtained. This operation is shown in Fig.1(c). The expression for such a SSB signal, $e_4(t)$ is given in expression (7), and is illustrated as a function of t and f in Fig.2(d):

$$e_4(t) = \frac{1}{2}\sum_{n=1}^{N} E_n \cos\left[(w_c + w_n)t + \phi_c + \phi_t\right] \tag{7}$$

The expression shows that the selected sideband is the upper sideband. The lower sideband might just as well be selected instead of the upper sideband. But, at this stage it is proper to state that upper sideband is translated in the frequency domain without inversion, whereas lower sideband is translated with inversion. This results from the fact that the two sidebands are symmetrical around the carrier frequency. It is also apparent from expression (7) that the resulting signal occupies only half the total spectrum required for AM communication.

It was stated in the previous chapter that a coherent detection for either SSB/SC or DSB/SC was necessary. In the past, a low-power, pilot carrier was transmitted for synchronization, that is, for automatic frequency control (afc) purposes to provide this end. However, with present day frequency stabilities (1 cps at 10 Mc in ground and 10 cps at 10 Mc in mobile equipment) the need for afc and pilot carriers is eliminated.

Thus, since SSB/SC signal does not contain a high-power carrier signal for demodulation process, this carrier must be supplied by the receiver itself. Now, it is a common practice to use the locally-generated carrier to translate the SSB signal back to its original position in the a-f band using conventional frequency conversion techniques. The block diagram of a SSB demodulator is shown in Fig.3.

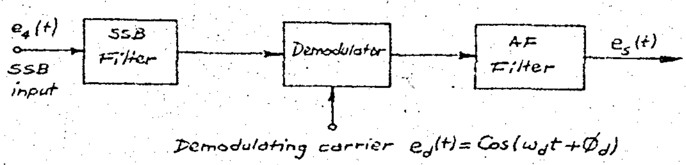


FIG.3

It is quite apparent that, if the demodulating carrier provided by the receiver is not of the correct frequency, the demodulated SSB signal will be shifted up or down by a uniform amount from its original location in the spectrum. If the demodulating carrier $e_d(t)$ is given as

$$e_d(t) = \cos (w_d t + \emptyset_d) \qquad (8)$$

and the demodulated SSB signal, $e_s(t)$ will be given as

$$e_s(t) = \sum_{n=1}^{N} E_n \cos \left[(w_c - w_d + w_n) + \emptyset_c - \emptyset_d + \emptyset_n \right] \qquad (9)$$

It can be seen from the inspection of expression (9) that if $e_d(t)$ is identical to the original carrier frequency, $e_c(t)$, then the demodulation product is simply the original signal frequency. But, if $e_d(t)$ is different from $e_c(t)$, then w_n will be displaced from its original position by an amount equal to $(w_c - w_d)$. Now, from the point of view of phase angles, if only the phase of the demodulating carrier is different from the phase of the original modulating carrier, all components of the output signal will have undergone a constant phase shift equal to the difference $(\emptyset_c - \emptyset_d)$.

It is apparent from the above discussion that to provide a demodulating carrier at the receiver of exactly the right frequency is of supreme importance, and this is the primary reason for the high cost and complexity of SSB equipment in comparison to AM equipment. Therefore the "frequency control techniques for SSB" will be discussed in a later chapter.

2.3. Advantages of the SSB System

As it was discussed earlier in this chapter that the main advantage of the SSB system comes from the reduction or elimination of the high-power carrier. Second advantage is that of the reduced spectrum. A SSB signal is more durable in the presence of selective fading and interference conditions. As an example, deterioration of an AM signal with selective fading is illustrated in Fig.4.

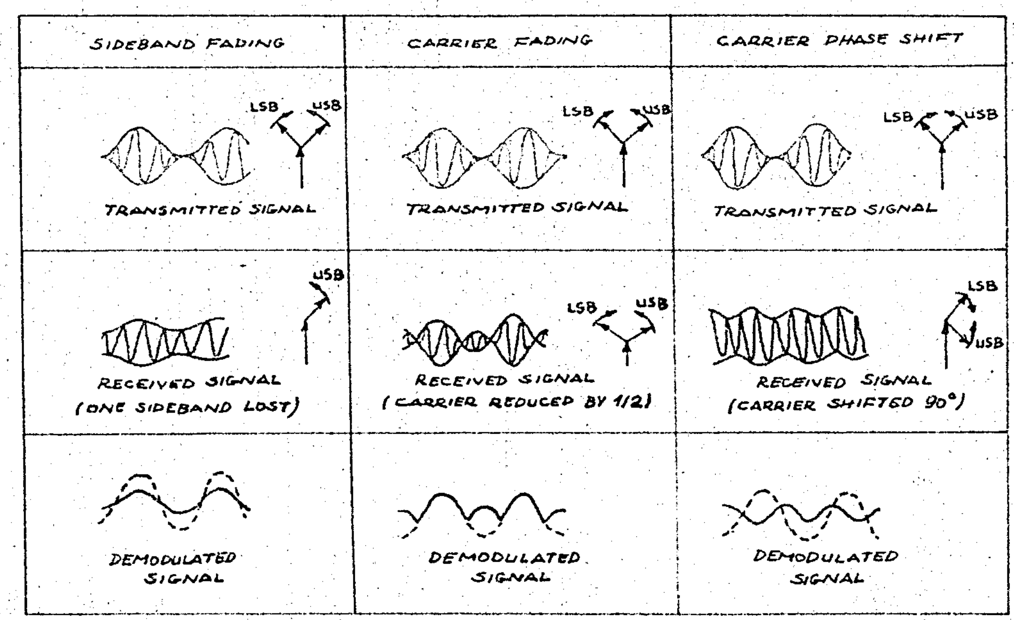


FIG.4

In generation and utilization of sideband power, the SSB overall efficiency is higher than with the AM equipment. A detailed comparison between SSB system and AM system will be given in the section devoted to the comparison of SSB with other type of modulation systems.

2.4. Disadvantages of the SSB System

While discussing the SSB signal generation, it was said that special kind of filters and modulators (balanced modulators) are necessary to pass only one of the sidebands. Therefore, this causes an increase in both complexity and the cost. Furthermore, what is more important is to generate the demodulating carrier which results even in more complexity. In fact, the effect of the latter one is much more pronounced in the complexity problem than that of the first. The SSB system requires frequency control having accuracy and stability on the order of 0.2 to 2 parts per million in the h-f spectrum, whereas present practice in fixed and mobile AM communication in the h-f spectrum ranges from 50 to 200 parts per million. In multichannel operation it is almost nonapplicable to provide large banks of crystals with such high stability, therefore, it becomes unavoidable to employ a very stable crystal oscillator; Which is designed using a highly complex frequency-synthesis technique for deriving any desired operating frequency from the one stable signal. "Stabilized Master Oscillator" design factors will be considered in the chapter of "Frequency Control Techniques for SSB".

2.5. Comparison of SSB with Other Types of Modulation Systems

A. Comparison of SSB with AM

a. Power Comparison of SSB and AM.

There are various ways to compare the operations of AM system and SSB system. But, under ideal propagation conditions it is most convenient to compare the required powers in the transmitters of both AM and SSB systems to produce a given signal-to-noise ratio (s/n) at the receiver for the two systems. The convenience of this way as a comparison basis comes from the fact that it is the signal-to-noise ratio which determines the intelligibility of the received signal. Fig.5 shows the comparison of SSB with AM systems where 100 percent, single-tone modulation is used.(*)

(*) For further information on single-tone, two-tone modulation technique of SSB, Fundamentals of SSB, Collins Radio Company

Fig.5(a) shows the power spectrum for an AM transmitter. The carrier power is assumed to be 1 unit of power. If the degree of sine-wave modulation is 100 percent, the r-f power transmitted will be 1.5 units of power, 1 unit of which is contributed by the carrier, and 0.5 unit of which is contributed by the two sidebands each having 0.25 unit of power. This AM transmitter is compared with a SSB transmitter whose "peak-envelope-power" (PEP) is 0.5 unit of power. Peak-envelope-power is the rms power developed at the crest of the modulation envelope.

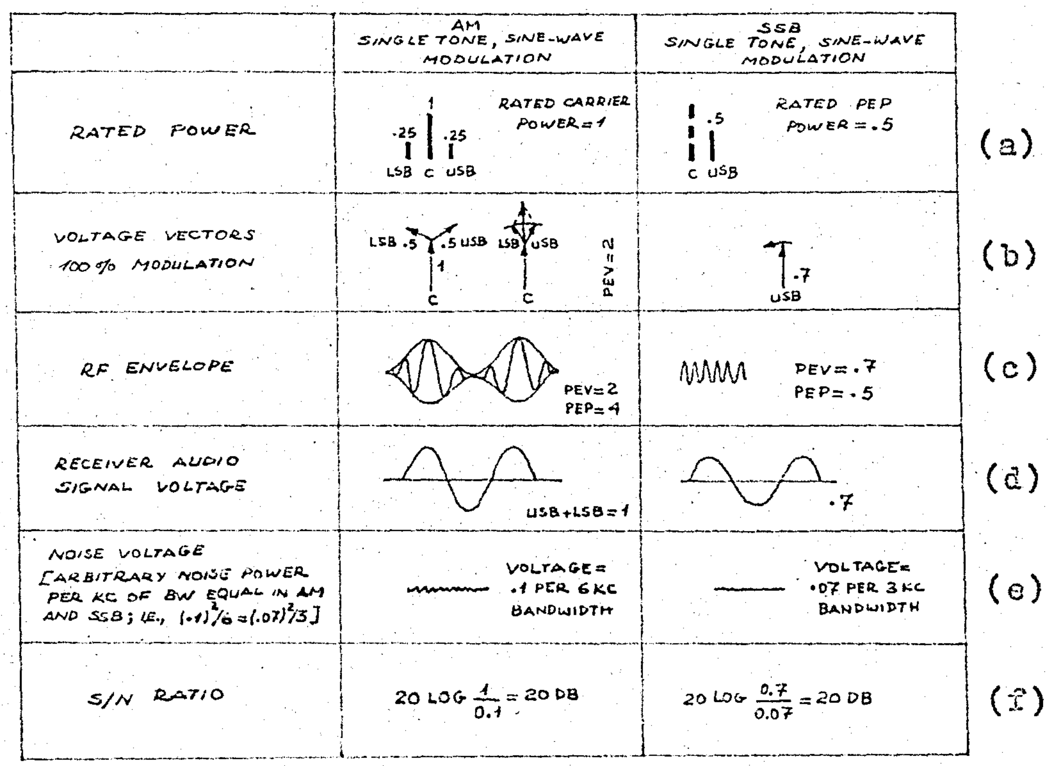


FIG.5

The voltage vectors corresponding to these cases when the identical signal-to-noise ratios are assumed to be the comparison basis are shown in Fig.5(b). An amplitude-modulated wave can be represented as the vector sum of a vector corresponding to carrier, and fixed in direction and magnitude and sense; and two vectors of identical lengths whose magnitudes are one-half of the carrier vector magnitude (since 100 percent sine-wave modulation is assumed) and whose origins are located at the tip of the carrier vector, and rotating in reverse directions with the same angular velocity. It is obviously seen from this representation that the resultant of the two sideband voltage vectors is either in or 180^{o} out of phase with the carrier voltage vector. Peak-envelope-voltage (PEV) is produced when the resultant of the two sideband voltages is in phase with the carrier voltage vector on further condition that the upper and lower sideband voltages are instantaneously in phase with each other. Now, if 100 percent perfect sine-wave modulation is assumed, the voltage vector of the carrier may be taken as 1-volt unit vector in which case each of the two sideband-voltage vectors is 0.5 unit of volt. Therefore, the peak-envelope-voltage in this case will be 2.0 units of volt. If we again assume a unit-ohm resistance basis for comparison purposes, it is seen from the above discussion that 0.5 unit of volt in each sideband produces 0.25 unit of power in each sideband. Following the same reasoning, we can deduce that 0.5 unit of rated peak-envelope-power of a SSB transmitter (which will result in the same signal-to-noise ratio) is generated by $(0.5)^{1/2}$ unit of volt, or 0.7 unit of volt. At this stage, it is emphasized again that 1 unit of rated carrier power in an AM system produces the same signal-to-noise ratio as does a SSB system with 0.5 rated peak-envelope-power.

The r-f envelopes developed by the voltage vectors are shown in Fig.5(c). The r-f envelope of the AM signal has a peak-envelope-voltage of 2 units which produces 4 units of peak-envelope-power. It was stated above that this 2-unit peak-envelope-voltage results in the case when the two sideband voltage vectors are instantaneously in phase. On the other hand the peak-envelope-voltage of the SSB signal is 0.7 unit of voltage with a resultant peak-envelope-power of 0.5 unit of power.

Fig.5(d) shows the demodulation products of AM and SSB r-f signals. In the case of AM demodulation, coherent detection by a conventional diode detector used in AM receivers is assumed. This type of detection takes the name of coherent detection since the voltages of the two sidebands are added in the detector. Audio voltage as a demodulation product of AM receiver is 1 unit of voltage which is equal to the sum of the lower and upper sideband voltages. When the r-f signal is demodulated in the SSB receiver, an audio voltage of .7 unit developes which is equivalent to the transmitter upper sideband (or lower sideband) signal.

If a broadband noise level is chosen as 0.1 unit of voltage per 6 kc bandwidth, the AM bandwidth, the same noise level is equal to 0.07 unit of voltage per 3 kc bandwidth, the SSB bandwidth. This is shown in Fig.5(e). These values represent the same noise power level per kc of bandwidth; that is, $(0.1)^2/6$ equals $(0.07)^2/3$. With this chosen noise level, the signal-to-noise ratio for the AM system is 20 log s/n in terms of voltage or 20 db. The s/n ratio for the SSB system is also 20 db, the same as for the AM system. The 0.5 power unit of rated peak-envelope-power for the SSB transmitter, therefore, produces the same signal intelligibility as the 1 power unit rated carrier power for the AM transmitter.

<u>Conclusion</u>: Under ideal propagating conditions but in the presence of broadband noise (*), an AM and SSB system perform equally (that is, same s/n ratio) if the total sideband power of the two transmitters is equal. This means that a SSB transmitter will perform as well as an AM transmitter of twice the carrier power rating.

b. Antenna Voltage Comparison of SSB and AM

The peak antenna voltage is the governing factor in the airborne and mobile installations where electrically small antennas are required. It was shown in Fig.5(c) and discussed above that the r-f envelope in the AM case has a peak-envelope-voltage of 2 units, whereas in the SSB case it is 0.7 unit of voltage. Therefore the ratio of peak-envelope-voltages is approximately 1/3, that is, the peak antenna voltage of the SSB system is approximately 1/3 that of the AM system. Since the limiting factor is often the corona breakdown point of the antenna in choosing the equipment power, the above ratio must be taken into consideration when selecting the equipment.

(*) For general description of noise bandwidth and noise ratios in Laplace transform treatment; Chapter 10, Harmonics, Sidebands and Transients in Communication Engineering, C. Louis Cuccia, McGraw-Hill Book Company, Inc., 1952.

It is more significant to select the power which can be radiated from an antenna of given dimensions. As an example, an antenna which can radiate 400 watts of r-f power must be connected to a transmitter whose maximum rating is 100 watts of power. This is so, because the peak-envelope-power is 4 times the carrier power in the AM case. On the other hand, a SSB transmitter whose maximum rating is 400 watts can be used with the same antenna.

c. Advantage of SSB with Selective Fading Conditions

Selective fading is due to the phase and power relationships between the carrier and the sidebands in AM type of transmission. Previous two sections are discussed on the basis of ideal propagation conditions. But, serious distortion and weaker signal reception are caused in AM by selective fading over long distances. To realize fidelity and the theoretical power from the signal, the received signal must be the exact replica of the transmitted signal, that is, the upper sideband, lower sideband and carrier must be received exactly as transmitted. Fig.4 shows the deterioration of an AM signal with different types of selective fading. As will be seen from the figure, there are three types of fading, which are sideband fading, carrier fading, and carrier phase shift. Over long distance transmission, one of the sidebands fades out, or even completely disappears. But as long as the other sideband contains the same intelligence this is not so detrimental from the point of reception. However, since the operation of AM is so adjusted to receive both of the sidebands that although the received signal contains the necessary intelligence, it is the level of noise which remains constant. That is, loss of one of the sidebands does not have any effect on the noise level. This is equivalent to a 6 db deterioration in s/n ratio out of the receiver.

Since a carrier voltage must be at least as strong as the sum of the two sideband voltages to demodulate the r-f signal, any attenuation in the carrier level more than the sidebands results in the most serious selective fading.

Carrier phase shift causes also selective fading. In the representation of an amplitude-modulated signal if carrier changes phase by 90 degrees for example from its original position, the resultant of the sideband vectors is $\pm 90^{\circ}$ out of phase with the carrier vector. The envelope of the phase modulated signal which results from this phase change bears no resemblance to the original AM envelope and the conventional AM detector will not produce an intelligible signal. Therefore, any shift in the carrier phase from its original phase relationship with respect to the sidebands will produce some phase modulation with a consequential loss of intelligibility in the audio signal.

Since SSB/SC system does not have any carrier in transmission, and since only one of the sidebands is transmitted, it is very obvious that neither carrier fading nor carrier phase shift influences the transmission characteristics. Selective fading within the one sideband of the SSB system only changes the amplitude and frequency response of the signal. It very rarely produces enough distortion to cause the received signal or voice to be unintelligible.

d. Comparison of SSB with AM under Limiting Propagating Conditions (*)

Communications are limited by the combination of noise, severe selective fading, and narrow-band interference under limiting propagating conditions over a long-range path. In such a case, SSB is more advantageous with respect to AM. This is shown in Fig.6 where the two transmitters compared have the same total sideband power. The figure shows that, the received signal in SSB case will provide up to a 9 db advantage over the AM signal when interference and fading becomes prevalent.

e. Comparison of Airborne High-Frequency Systems

For illustrative purposes an AM set, AN/ARC-38 rated at 100 watts r-f output and a SSB set, AN/ARC-58 rated at 1000 watts r-f output are compared in Fig.7. The frequency range of these two units is from 2 to 30 Mc.

B. Comparison of SSB with FM

Fig.8 shows a comparison between a mobile FM system and a mobile

(*) For further details; J.F. Honey, "Performance of AM and SSB Communications", Tele-Tech, Sept. 1953.

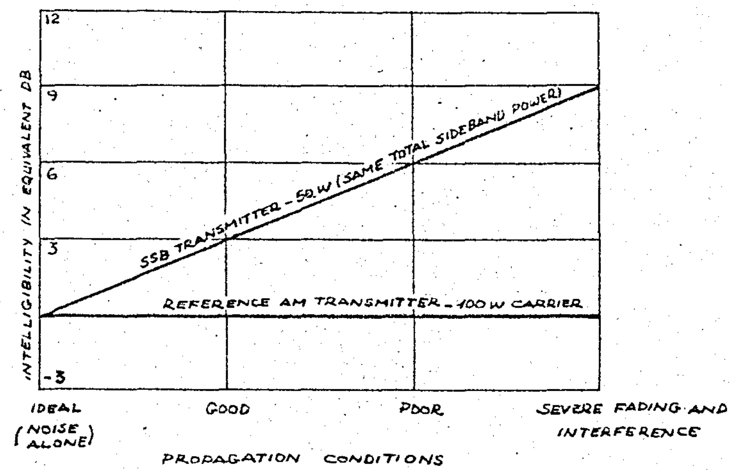


FIG.6

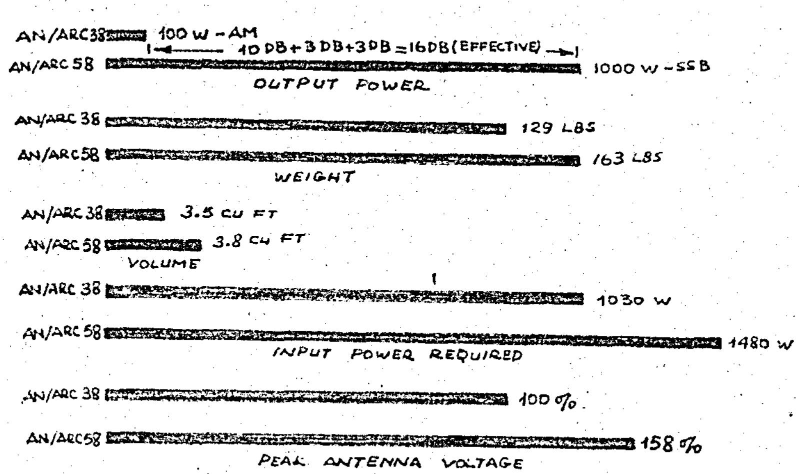


FIG.7

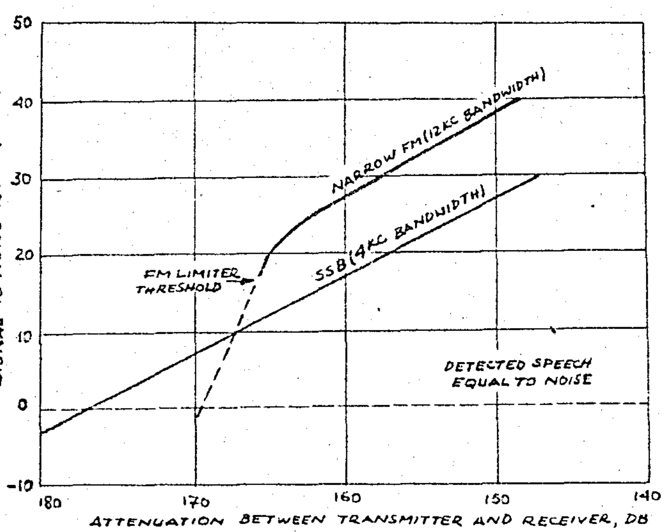


FIG.8

SSB system of equal size. (*) The figure indicates that with between 150 to 160 db of attenuation between the transmitter and receiver, a strong signal, the narrow-band FM provides a better s/n ratio than the SSB system. Under weak signal condition, from 168 and higher db of attenuation between transmitter and receiver, the s/n ratio of the FM falls off rapidly, and the SSB system provides the best s/n ratio.

Furthermore the SSB system, as it will be seen from the figure, provides three times the savings in spectrum space as the narrow-band FM system.

2.6. Economics of SSB

While recent publications indicate a trend toward SSB and forecast a change in the not too distant future to the use of SSB for many services, it will be necessary for SSB to justify its use in terms of the cost factors and economies in operation.

Among the cost factors are those of initial cost of the SSB transmitter and the SSB receiver. Power output is normally considered to be one of the important considerations in analyzing transmitter specifications. Another factor to be considered in the analysis of equipment is the operating economies which can be realized by changing to new, more advanced apparatus.

Without question, the low-power SSB/SC transmitter will sell at a higher price than the comparable AM transmitter. This is brought about by a requirement for a larger number of circuits in the low level stages of the SSB transmitter.

In the SSB transmitter, linear amplifiers must be used throughout, from the very low level stages through the final Linear Power Amplifier (LPA). In general, the r-f portions of a SSB transmitter bring about a moderate increase in the number of components, stages and vacuum tubes over that of an AM transmitter. Furthermore, critically designed sideband separation filters, precise frequency

(*) M. Magnuski and W. Firestone, "Comparison of SSB and FM for VHF Mobile Service", Proc., IRE, Dec.1956.

control devices must be taken into consideration.

The SSB receiver does not differ greatly from a well-designed communications receiver for AM. It differs chiefly in the requirement for reduced bandwidth, which requires a relatively complex IF amplifier or the use of crystal or mechanical filter.

In addition to the advantage of reduced initial cost in the higher transmitter power ratings, there are certain operating economies which can be achieved through the use of SSB transmitter. There is a significant factor of reduced primary power cost. A SSB transmitter utilizing a linear amplifier in the high power range requires much less power from the power lines. This is due to the greatly increased effectiveness of SSB communications which permits a lower power transmitter to do the job of a higher power AM unit.

The cost of the transmitter buildings is increasing continuously so the matter of floor space is achieving increased attention. The SSB transmitter by eliminating the modulator requires much less floor space, and, hence, allows a smaller transmitter building.

Because of the greater effectiveness of the SSB transmitter, transmission line and antenna costs can be reduced because the peak voltages encountered are much less in the SSB system.

2.7. Historical Development of SSB Communication Systems

Although SSB transmission has only received publicity in the last few years, the knowledge of the sideband and the development and use of SSB techniques have progressed over the last 40 years. The acoustical phenomenon of combining two waves to produce sum and difference waves carried over into electric-wave modulation. The presence of the upper and lower sidebands in addition to the carrier frequency were tacitly assumed to exist but were not concretely visualized in the earliest modulated transmissions. Recognition that one sideband contained all the signal elements necessary to reproduce the original signal came in 1915. It was then, that at the Navy Radio Station at Arlington, Va., that an antenna was tuned to pass one sideband well, even though the other was attenuated.

From 1915 until 1923, the physical reality of sidebands was vigorously argued with the opponents contending that sidebands were mathematical fiction. However, the first trans-Atlantic radiotelephone demonstration in 1923 provided a concrete answer. This system employed a SSB signal with a pilot carrier. Single sideband was used in this system because of the limited power capacity of the equipment and the narrow resonance bands of efficient antennas at the low frequency (57 kc) used. By 1927 trans-Atlantic SSB radiotelephony was open for public service.

The first overseas system was followed by short-wave systems, 3 to 30 Mc, which transmitted double sideband and carrier because SSB development did not permit practical SSB transmission in this frequency range. However, SSB techniques were employed in various multiplexing systems. It has not been until recently that equipment developments have permitted the advantages of SSB communication to be fully exploited. These developments have been in the fields of frequency stability, filter selectivity, and low-distortion linear power amplifiers. These developments have led to military and commercial acceptance of SSB communication systems. There are presently available several radio amateur and commercial SSB radio sets, fixed-station SSB exciters up to 45 kw linear power amplifiers (the ones used in İzmir are rated up to 18 kw), and airborne transceivers capable of reliable communications with unlimited range. Some of these equipments, especially the military equipments, are provided with automatic frequency selection and automatic tuning to further enhance their value as reliable, easily operated systems.

CHAPTER 3

SINGLE SIDEBAND GENERATION

After having discussed the SSB characteristics, we are going to undertake the methods of generating the SSB signal. There are two general methods of generating a SSB signal, but also a third method of generation of SSB signal will be examined (*).

3.1. Filter Method of SSB Generation

As it was discussed in Chapter 2, it is possible to eliminate one of the modulation products of amplitude modulation by using filters. The filters used are so designed that they eliminate one of the sidebands, either upper sideband or lower sideband, depending on the type of the filter. But, commonly used filters in SSB generation are bandpass filters, and carrier is placed at one side or at the other of pass region according to which band will be transmitted.

The procedure is simple; an audio signal is fed into a balanced modulator along with the carrier frequency. The output of the modulator is two sidebands while the modulator balances out the carrier. The output of the balanced modulator is fed into a filter in which one of the sidebands is eliminated according to the desire. This stage of balanced modulator and filter is followed by another stage of the same kind. This kind of stage succession is necessary sometimes especially when the carrier frequency is set at a quite high frequency, since then in such a case it is very difficult to design a filter which will completely eliminate one sideband and pass the other without an appreciable attenuation. Therefore, the translation is done in several steps so as to ease the filter requirement. The generation process is shown in Fig.1, in which two stages are

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(*) "A Third Method of Generation and Detection of SSB Signal", Donald K. Weaver, Proc.IRE, pp.1703-1705, Dec. 1956.

shown. The number of translational steps may vary from three to five.

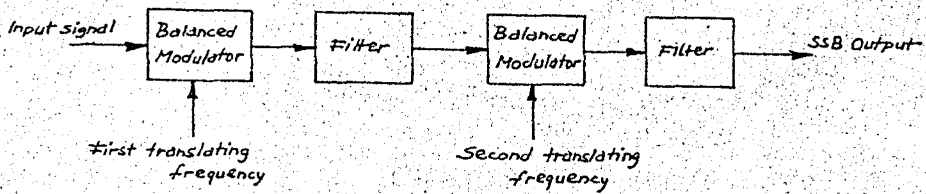


FIG.1

The detection process is simply the reverse of the generation, but in this case the balanced modulators are not needed, instead ordinary converter circuits are used.

In Fig.2 filter passband response is shown for a carrier frequency of 100 kc. As it was mentioned above, the selection of the special band depends on the location of the carrier frequency with respect to the passband characteristic of the filter. Upper sideband selection is achieved by placing the carrier frequency below the filter passband and vice versa. In case the filter is used for reception of the other sideband as well from time to time, the passband characteristic of the filter should be symmetrical, that is, the filter should provide a similar attenuation behavior at both the upper and lower edge of the filter passband. Since the selection of one of the sidebands necessitates the movement of the carrier frequency with respect to the filter passband characteristic, it is necessary to compensate this movement of the carrier frequency in the following translational device so that the output carrier frequency stays fixed.

It must be also emphasized here that the edges of the passband characteristic of the filter should be steep enough to cause the complete elimination of the undesired sideband. Fig.2 shows that a 20-db attenuation is offered for the carrier where a greater degree of attenuation is achieved for the undesired frequency.

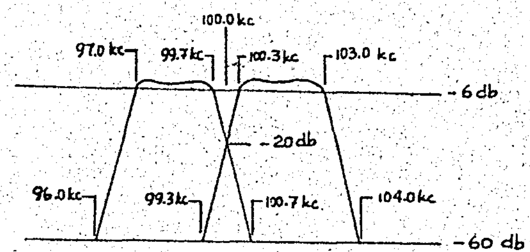


FIG. 2

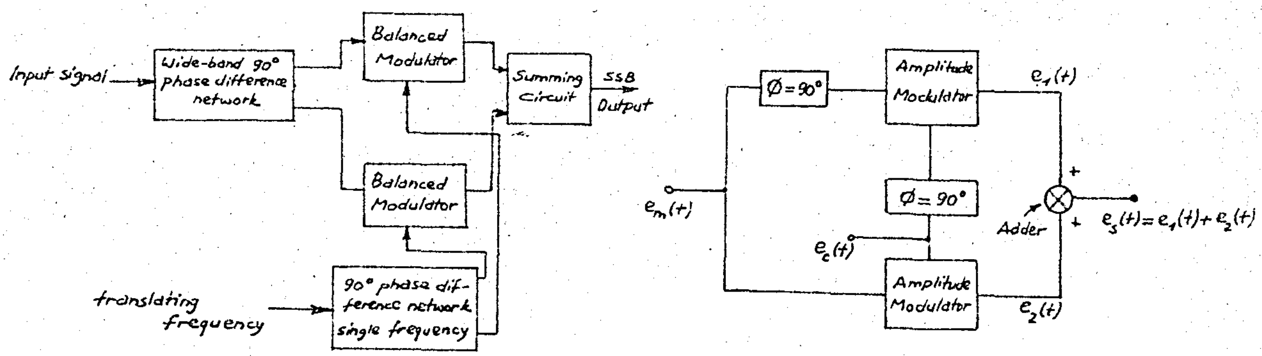
3.2. Phase-Shift Method of SSB Generation

In this method of generation since sharp cutoff filters are not used, it is possible to generate the desired sideband in a single translational step regardless of how high the final signal frequency may be. Actually two ways are used in phasing method of SSB generation. These will be considered below successively.

a. Input is fed into a wideband, 90° phase difference network. This network is so designed that the magnitudes of frequencies do not undergo any distortion, but the output product comprises two components of equal magnitude with 90° phase shift between them. This output signal is applied to two balanced modulators, one component to each, and a carrier frequency containing two products of equal magnitude with 90° phase shift is also applied to these modulators, one component of which to each modulator. When the output signals from these two balanced modulators are added, one set of sidebands will add in phase, generating the desired signal, while the other sideband will cancel itself out. By subtracting instead of adding, it is possible to change sidebands.

The procedure is shown in Fig.3(a) in block diagram. Fig.3(b) is exactly the same procedure, but it shows the operation more clearly. From Fig.3(b) we can write;

$$e_1(t) = E_m \sin w_m t \sin w_c t$$
$$e_2(t) = E_m \cos w_m t \cos w_c t$$



(a) (b)

FIG.3

Since $e_s(t) = e_1(t) + e_2(t)$ we have, after simplification,

$$e_s(t) = E_m \cos (w_c - w_m)t$$

Thus $e_s(t)$ represents the lower sideband.

In this method, suppression degree of the undesired sideband depends upon accurate balancing and requires very careful control of amplitudes and phases. The normal suppression degree practically achieved is 20 db, but 30-db suppression may well be expected. But it is quite difficult to go beyond 40-db suppression.

b. Instead of phase-shifting one component of the audio signal 90^0, for easier control purposes and accuracy in phase shift, one component of the audio signal is shifted by α degrees, whereas the other component is shifted by β degrees. The output of the phase-shift networks are applied to two balanced modulators separately. The carrier frequency is applied to one modulator directly, whereas 90^0 phase-shifted component of the carrier is introduced into the other modulator. The operation is represented in block diagram in Fig.4 below. A mathematical treatment is also given (*).

(*) Single Sideband Principles and Circuits, E.W. Pappenfus, Warren B. Bruene, Edgar O. Schoenike, McGraw-Hill Book Company, 1964, pp.35-37.

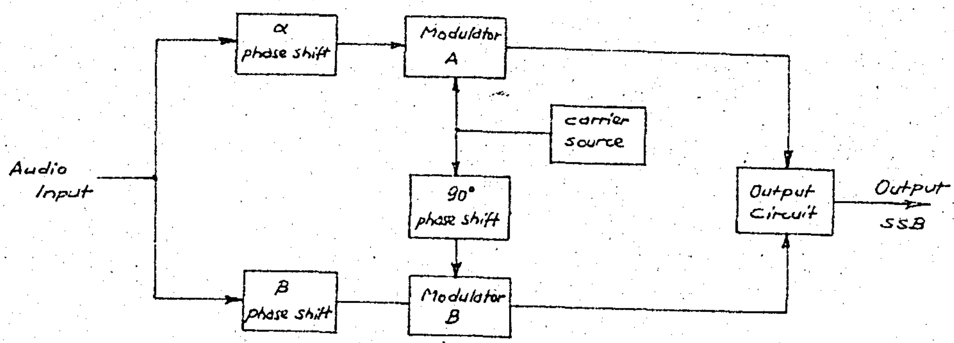


FIG. 4

Input to Modulator A

$$E_c \cos 2\pi f_c t = \text{carrier}$$

$$A \cos (2\pi f_s t + \alpha) = \text{signal through } \alpha \text{ network}$$

Input to Modulator B

$$E_c \cos (2\pi f_c t - 90^\circ + \Delta) = \text{carrier after } -90^\circ + \Delta \text{ phase shift}$$

$$B \cos (2\pi f_s t + \beta) = \text{signal through } \beta \text{ network}$$

Substituting $\beta = \alpha - 90^\circ + \delta$ gives

$$B \cos (2\pi f_s t + \alpha - 90^\circ + \delta) = \text{signal through } \beta \text{ network}$$

where

A = amplitude of signal through α network
α = phase shift of signal through α network
B = amplitude of signal through β network
β = phase shift of signal through β network
Δ = error in carrier phase from 90°
δ = error in signal phase from 90° difference between α and β networks
f_s = signal frequency
f_c = carrier frequency
E_c = carrier amplitude

The output from modulator A, if complete carrier balance is assumed, is

$$E_{AO} = E_c \cos 2\pi f_c t \left[A \cos (2\pi f_s t + \alpha)\right]$$
$$= \frac{E_c A}{2} \left\{\cos\left[2\pi (f_c + f_s)t + \alpha\right] + \cos\left[2\pi (f_c - f_s)t + \alpha\right]\right\}$$

and the output from modulator B, if complete carrier balance is assumed,

$$E_{BO} = E_c \cos (2\pi f_c t - 90^\circ + \Delta) \left[B \cos (2\pi f_s t + \alpha - 90^\circ + \delta)\right]$$
$$= \frac{E_c B}{2} \left\{\cos\left[2\pi (f_c + f_s)t + \alpha - 90^\circ - 90^\circ + \Delta + \delta\right] + \cos\left[2\pi (f_c - f_s)t - \alpha - 90^\circ + 90^\circ + \Delta - \delta\right]\right\}$$
$$= \frac{E_c B}{2} \left\{- \cos\left[2\pi (f_c + f_s)t + \alpha + \Delta + \delta\right] + \cos\left[2\pi (f_c - f_s)t - \alpha + \Delta - \delta\right]\right\}$$

Then the total output from both modulators is

$$E_0 = E_{AO} + E_{BO}$$
$$= \frac{E_c}{2}\left\{A \cos\left[2\pi (f_c + f_s)t + \alpha\right] + A \cos\left[2\pi (f_c - f_s)t - \alpha\right] - B \cos\left[2\pi (f_c + f_s)t + \alpha + \Delta + \delta\right] + B \cos\left[2\pi (f_c - f_s)t - \alpha + \Delta - \delta\right]\right.$$

It is seen from the above equation that if the signal amplitudes to both modulators are equal, and if there are no errors, the total output from both modulators will be

$$E_0 = E_c \, A \cos\left[2\pi (f_c - f_s)t - \alpha\right]$$

and this is the lower sideband component obtained.

c. Modified Phase-Shift Method of SSB Generation

This method of SSB generation uses two sets of balanced modulators instead of one. First audio signal is applied to both of these modulators. Furthermore, an audio frequency f_0 which is placed at the arithmetic mean of the audio band is applied to one of the modulators directly, and to the other one with 90^{o} phase shift. Fig.5 shows this method of SSB generation. In Fig.6(a), the audio input spectrum to the modulators is shown. In Fig.6(b), the spectrum from the modulators is shown. Therefore, we can conclude from this figure that, the modulation results in an inversion of the bottom half of the input spectrum and a shift of the top half of the input spectrum to a position in frequency coincident with the inverted bottom half. There are also sum frequencies as shown which are eliminated by using low-pass filters. Now, after the filters the signals are applied to two modulators, one of which is directly supplied with the carrier frequency, whereas the carrier frequency to the other one is 90^{o} phase-shifted. As a result, the output will have the same audio input spectrum shape with undesired sidebands located inversely below the frequency axis. As it is seen from the figure that, these undesired products are small enough if phase and amplitude balance is good.

The disadvantage of this method is that to avoid loss of signal components to f_o in frequency, d-c coupling should be used. Furthermore, if the balance of the first set of modulators is not achieved accurately, a carrier whistle appears at f_o.

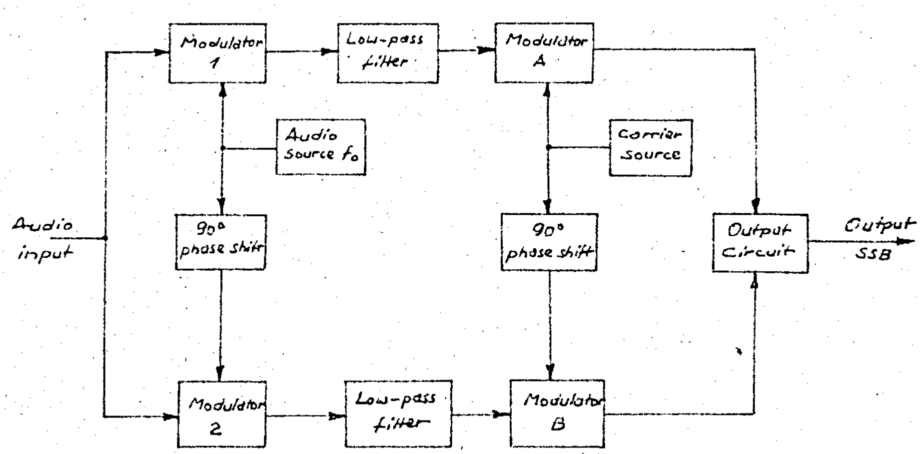
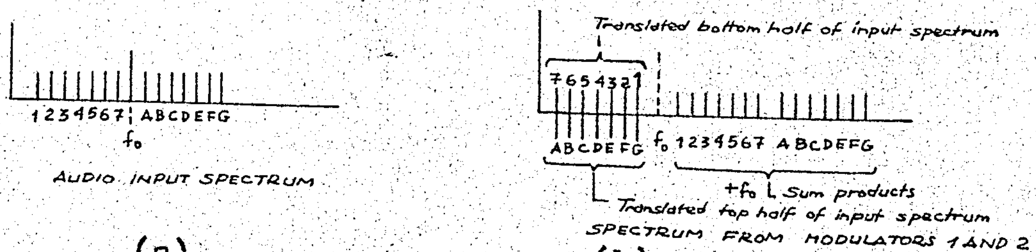


FIG.5



(a) (b)

FIG.6

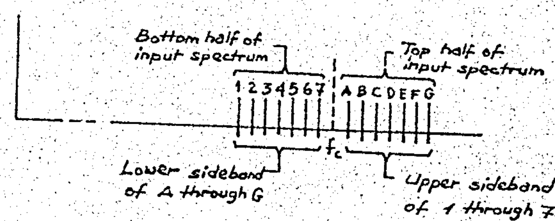


FIG.7

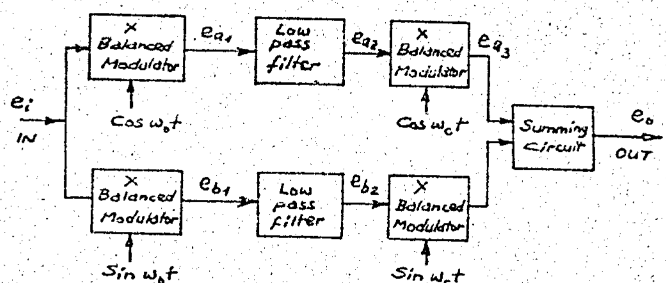


FIG.8

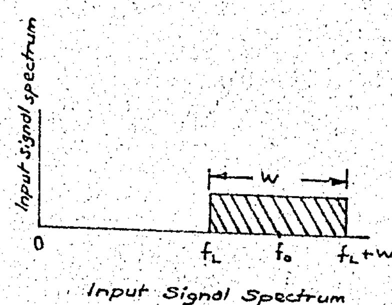


FIG.9

A simplified block diagram corresponding to Fig.5 is shown in Fig.8. If the bandwidth of audio input spectrum is designated by W, and the lowest frequency by f_L, then the center frequency of the bandwidth f_o can be expressed by

$$f_o = f_L + (W/2) \tag{1}$$

Now, if the input signal is a combination of sine terms, it can be expressed as

$$e_i(t) = \sum_{n=1}^{N} E_n \cos (w_n t + \phi_n) \tag{2}$$

As it is seen from the figure, f_o is the modulating frequency of the first set of modulators. Therefore, the outputs of the first two balanced modulators are

$$e_{a1} = 2\, e_i(t) \cos w_o t \qquad (3)$$

$$e_{b1} = 2\, e_i(t) \sin w_o t \qquad (4)$$

where $w_o = 2\pi f_o$. (5)

The coefficient 2 is used for convenience and is assumed to be the property of the balanced modulators. Substituting (2) into (3) and (4) and expanding gives:

$$e_{a1} = \sum_{n=1}^{N} E_n \cos\left[(w_n - w_o)t + \phi_n\right] + E_n \cos\left[(w_n + w_o)t + \phi_n\right] \qquad (6)$$

$$e_{b1} = \sum_{n=1}^{N} -E_n \sin\left[(w_n - w_o)t + \phi_n\right] + E_n \sin\left[(w_n + w_o)t + \phi_n\right] \qquad (7)$$

The frequencies $f_n = w_n/2\pi$ are restricted to the original bandwidth W.

$$f_L \leq f_n \leq f_L + W \qquad (8)$$

Spectrum of signals is shown in Fig.10. The low-pass filter passes the frequencies from zero to $W/2$. From $W/2$ to $2f_o - (W/2)$ there should be no signal energy which provides a convenient transition region for the filter. Above $2f_o - (W/2)$ the filter should have adequate attenuation to eliminate the h-f components from the balanced modulators. Using such a filter the expressions for the filter output voltages are

$$e_{a2} = \sum_{n=1}^{N} E_n \cos\left[(w_n - w_o)t + \phi_n\right] \qquad (9)$$

$$e_{b2} = \sum_{n=1}^{N} E_n \sin\left[(w_n - w_o)t + \phi_n\right] \qquad (10)$$

These two low-frequency functions are then applied to another set of balanced modulators. But, in this case the modulating frequency is the carrier frequency w_c, and this is the band center of the desired SSB signal. The expressions for the outputs of this second set of modulators are

$$e_{a3} = e_{a2} \cos w_c t \quad (11)$$

$$e_{b3} = e_{b2} \sin w_c t \quad (12)$$

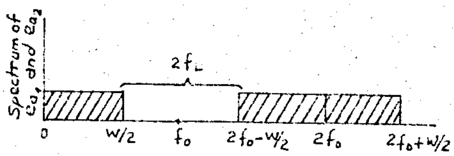


FIG.10

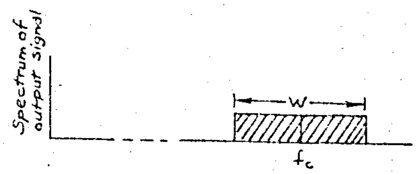


FIG.11

Substituting (9) and (10) into (11) and (12), and expanding gives

$$e_{a3} = \sum_{n=1}^{N} \frac{E_n}{2} \cos\left[(w_c + w_n - w_o)t + \phi_n\right] + \frac{E_n}{2} \cos\left[(w_c - w_n + w_o)t - \phi_n\right] \quad (13)$$

$$e_{b3} = \sum_{n=1}^{N} \frac{E_n}{2} \cos\left[(w_c + w_n - w_o)t + \phi_n\right] - \frac{E_n}{2} \cos\left[(w_c - w_n + w_o)t - \phi_n\right] \quad (14)$$

Finally, adding (13) and (14) gives the desired SSB output

$$e_o = e_{a3} + e_{b3} \quad (15)$$

$$e_o = \sum_{n=1}^{N} E_n \cos\left[(w_o + w_n - w_c)t + \phi_n\right] \quad (16)$$

Frequency normally referred to as the carrier corresponds to $w_c - w_o$ and w_c is the center of the SSB. Fig.11 shows the spectrum of e_o.

CHAPTER 4

SSB EXCITERS

In this chapter, SSB exciters will be discussed shortly, and in the next chapter SSB receivers will be undertaken.

4.1. The SSB Exciter Considerations

The SSB exciter translates the incoming audio signal to a band of frequencies in the r-f band. Therefore, the SSB exciter generates an r-f sideband from the audio signal input, translates this sideband to the final output frequency, and provide sufficient amplification to drive the r-f power amplifier. Hence, a SSB exciter, is in fact, a complete transmitter in itself. Functional diagram of a typical SSB exciter is given in Fig.1.

Exciters can be classified in two groups; a. Elementary exciters, and b. Double-conversion exciters. The classification as such depends on the degree of attenuation needed for the undesired spurious mixer products.

A single mixer-exciter which corresponds to the elementary exciters is given in Fig.2. The process in this single mixer-exciter is that an audio signal after being processed (*) is fed into a balanced-modulator-filter SSB generator. The output of this generator is two independent sidebands after the filter, and these two outputs are connected in parallel provided that necessary isolation be achieved. It is necessary that the SSB exciter attenuates the image and injection frequencies sufficiently. If this need is not met by a single mixer-exciter, then double-conversion exciters must be used. It is extremely important to minimize the distortion level due to the translation of the modulating frequencies, and since the mixers with a power output above 0.5 mw are seldom employed, it is also necessary to provide a large amount of amplification at the last stage which creates serious problems from the point of

(*) "Processing an audio signal" will be discussed later in this chapter.

stability and regeneration.

The signals at mixer output is shown in Fig.3. If we assume an X Mc is the desired mixer output, and y is the modulating frequency, there seems to be three frequency products only one of which is the desired frequency output, X Mc. The two other are X + y, and X + 2y. Since, normally the injection of X + y is at much higher than the SSB signal, it is necessary to provide the sufficient attenuation for the undesired frequency products. Therefore the mixing problem is carried out in more than one stage ~~s~~ as compared to the elementary exciters. Such a dual mixer exciter for AN/ARC-58 (*) is shown in Fig.4.

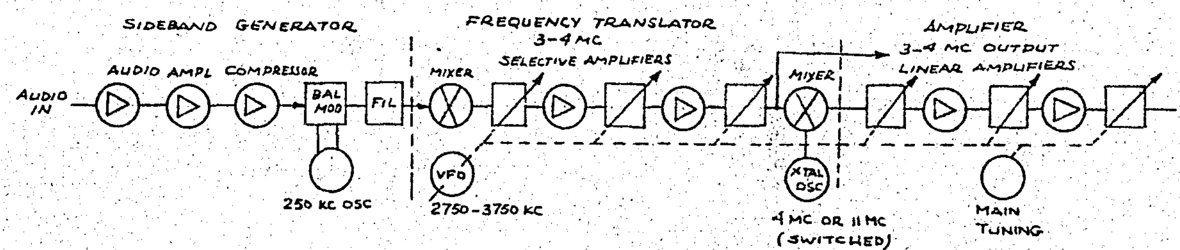


FIG.1

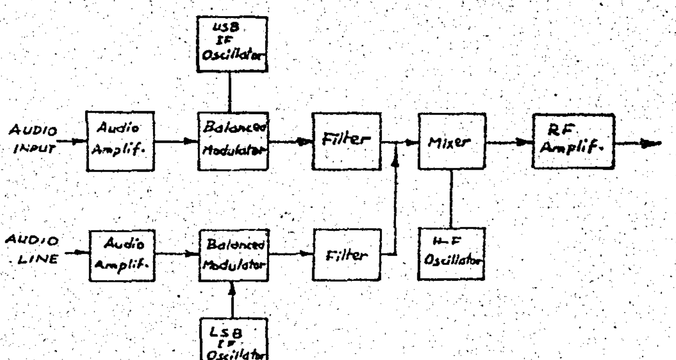


FIG.2

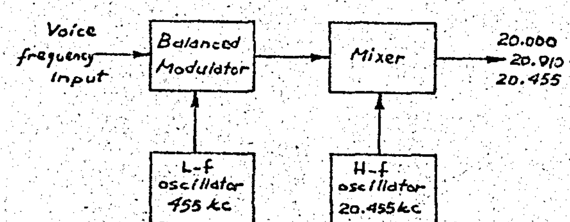


FIG.3

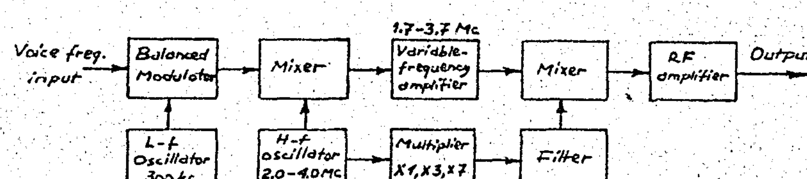


FIG.4

(*) This is a piece of equipment manufactured by Collins Radio Company. AN/ARC-58 was compared against AN/ARC-38 which is an AM set of the same company in the second chapter.

To generate the r-f sideband of frequencies, the SSB exciter uses low-level modulation and obtains the desired output level through the use of linear amplifiers. Low-level modulation is used since the carrier and the undesired sideband must be suppressed, and this can be achieved best by using a fixed low frequency. It is possible by practical designs to achieve a suppression of carrier by 40 db. The band of side frequencies is normally held to 4 kc in single-sideband exciters for communication purposes.

We have in the preceding chapter seen the ways of generating a SSB signal. This is done in the sideband generator part of the SSB exciter in Fig.1. The sideband generator processes the input audio signal, generates the r-f sideband in a modulator, selects the desired sideband while suppressing the undesired sideband, and suppresses the carrier.

4.2. Input Signal Processing

The audio input to the SSB generator must be amplified, amplitude limited, and shaped before being applied to the modulator circuits. Since the applied audio input is generally a voice signal, the processing of the signal is much more complicated as compared with the processing of a single tone, or a group of tones, or a signal from a data gathering device of constant amplitude.

The amount of amplification depends upon the output capability of the source of the audio signal and the input signal requirements of the modulator. Modulators require an audio signal in the range of 0.1 to 1 volt at impedances of 200 ohms for diode modulators, or several hundred-thousand ohms for vacuum tube modulators. The output of a microphone may be from 100 to 1000 times less than the 0.1 to 1 volt range. Therefore, to obtain efficient utilization of the transmitter power amplifier, the applied driving signal should be as close to maximum without exceeding the overload level.

In order to make the human voice more compatible with the electrical characteristics of a communications systems, two methods of reducing the extreme amplitude variations are used. These two methods are the 1) compression, 2) clipping method.

When the input signal is made up of extreme variations, such as a peak level to average level to average level of 4/1, the average transmitted power level will only be 1/4 the maximum output the transmitter is capable of furnishing. Therefore, in compressor circuit, this ratio is somewhat reduced.

A compressor circuit uses a stage of push-pull which is biased with a d-c voltage applied to the control and suppressor grids. Then it is amplified through an amplifier stage which is coupled to a transformer T_3 (Fig.5). According to the polarity of the signal appearing on the center-tapped transformer either $V4_A$ or $V4_B$ conducts. The resultant current flow causes a voltage drop across the corresponding resistor, R_{10}. The negative voltage on the control and suppressor grids of V1 and V2 reduces the gain of the tubes to limit the excursion of the audio signal. Compression of about 10 db is usually considered as an acceptable maximum value.

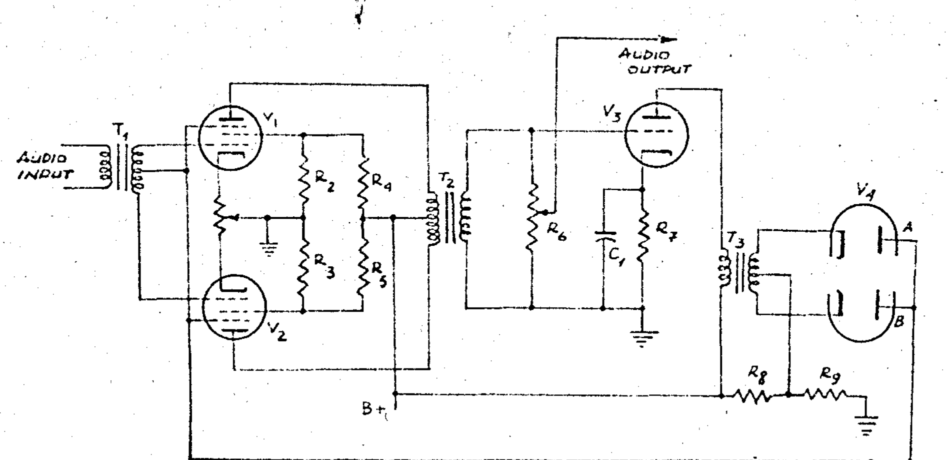
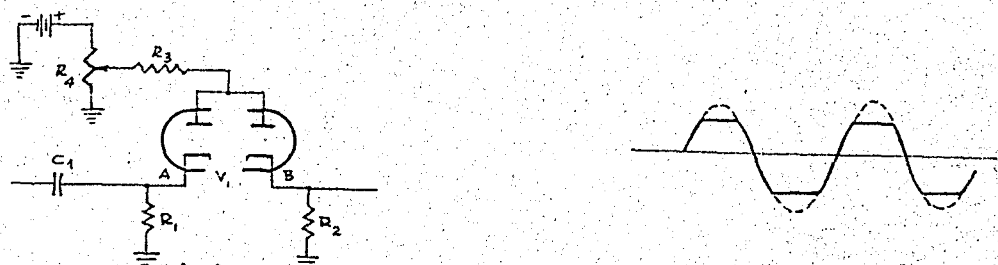


FIG.5

The clipper circuit, on the other hand, reduces the amplitude of a signal, and prevents it from exceeding a preset level. In this type of reduction of the amplitude of the signal the distortion is very high, and results in loss of individuality in speaking and broadening of the spectrum occupied by each speech. Low-pass filters are usually used in conjunction with clippers to limit the spectrum and reduce distortion. In the clipping circuit, a weak consonant that follows a loud vowel in human voice will be given full amplification, although the preceding vowel was severely clipped. This amplifying of weak sounds in relation to soft sounds is referred to as "consonant amplification". A simple clipper circuit is shown in Fig.6.



(a) (b)

FIG.6

4.3. Modulators for SSB

In this section a short discussion of the modulators used in SSB system will be given along with the schematic diagrams. There are three types of modulators used in the SSB system. These are (1) rectifier.modulators, (2) multielectrode vacuum tube modulators, (3) nonlinear reactance modulators. In general, several types of modulators can be classified in two main functional classes; (1) those in which the modulation is dependent on the polarity of the modulating signal, and (2) those where the modulation is dependent on the instantaneous waveform of the modulating signal.

The main advantage of the rectifier modulators comes from the fact of high stability compared to multielectrode vacuum tube modulators. Since no heating elements are present, no power dissipation takes place. The rectifier modulators are of three types, ring, series, or shunt. These are shown in (a), (b), (c) of Fig.7. The rectifier modulators are connected almost always as balanced modulators so that, there will be no output of the r-f switching voltage in the modulator output terminals.

On the other hand, multielectrode vacuum tube modulators are more flexible and used in a wide variety of applications in addition to sideband generation. They are capable of giving conversion gain rather than loss as in the case of rectifier modulators. But the main disadvantage is, as indicated in the rectifier modulators, that they have heating elements. Vacuum tube modulators, employing modulating functions dependent on the instantaneous amplitude of the modulating signals, are basically one of two types: a product modulator, or a square law modulator.

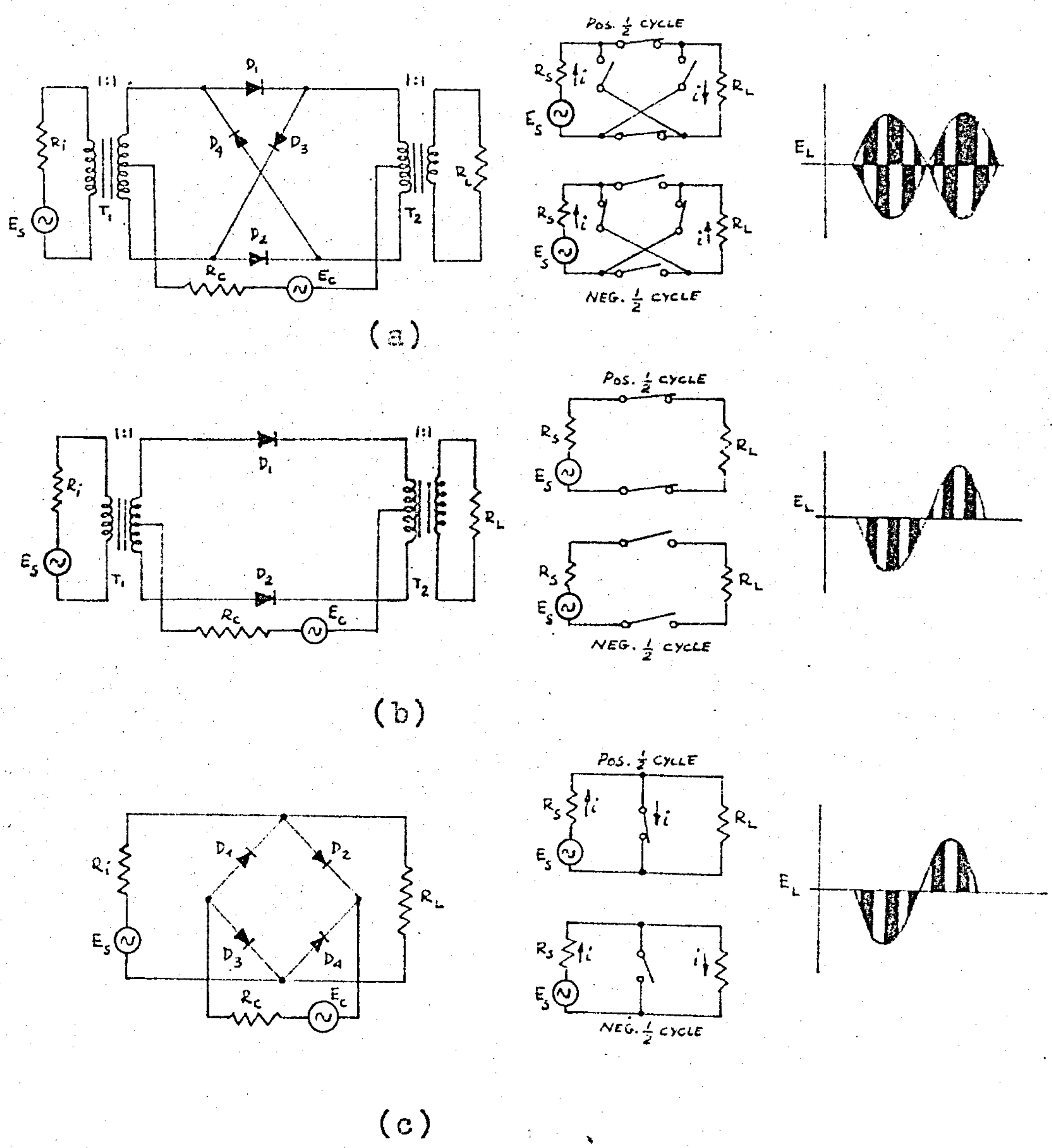


FIG.7

The output signal of a product modulator is proportional to the two input signals, one being the carrier and the other being the modulating signal. An example to this kind of modulator is a double grid vacuum tube, where the carrier voltage is applied to one grid and the modulating signal applied to the other.

On the other hand, the modulation takes place in a square-law modulator directly because of the nonlinearity of the shape of the plate current versus grid voltage curve of a vacuum tube triode.

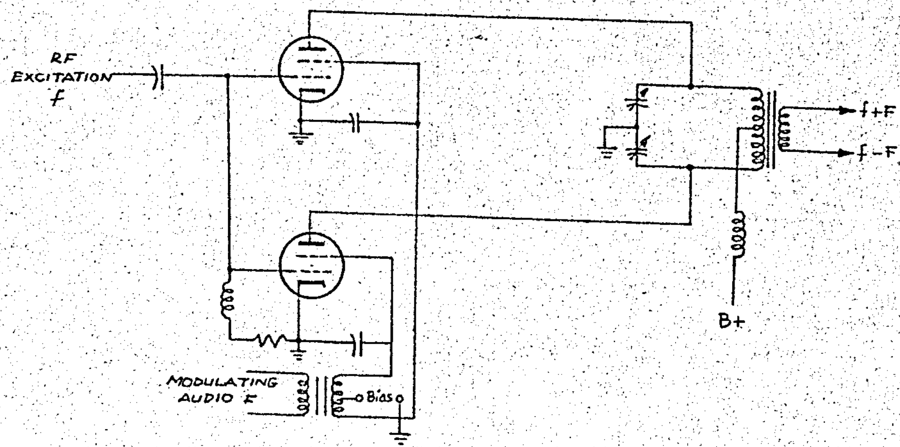
Nonlinear reactance modulators are not used frequently due to the lack of materials usable at high frequencies.

The purpose of the modulators is to obtain an r-f sideband by combining the audio signal obtained from the processing circuits and an r-f carrier wave in an amplitude modulator. Suppression of carrier takes place, however, in the balanced modulators which employ either vacuum tubes or diodes.

In balanced vacuum tube modulators, when no audio signal is applied, the output of the modulator is zero, since, the signal from one tube is canceled out in the output circuit by the signal from the other tube. But when push-pull audio signal is applied, one tube will conduct more than the other, since the modulating voltages are of opposite polarity. Modulator is not balanced for sideband, although it is for carrier; therefore, sideband will appear in the output though the carrier is suppressed. Since the process taking place in the modulator is actually a mixing process, there will be a sum and difference frequencies at the output.

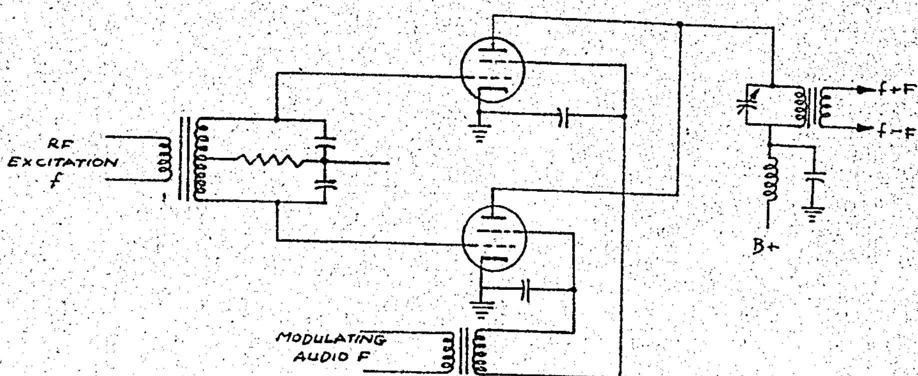
At least minimum 30 db suppression of the carrier must be achieved by using two tubes of the same characteristics. But, since it is necessary to obtain at least 40 db suppression of carrier in single sideband suppressed carrier transmission, further suppression of the carrier is achieved by use of filters following the modulator stage.

Diode balanced modulators employ diodes instead of vacuum tubes. The advantage of this type of modulators, as it was discussed earlier, is the high stability due to non-existence of the heating elements. Two types of balanced modulators are shown in Fig.8 employing vacuum tubes. A diode balanced modulator is shown in Fig.9.



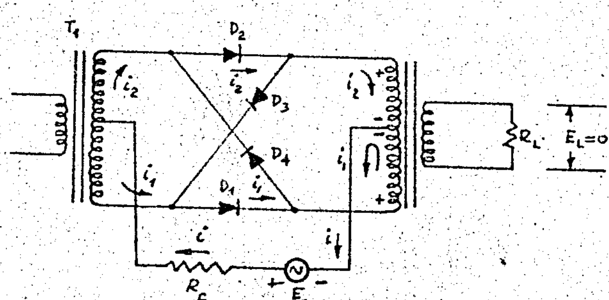
(a)

VACUUM TUBE BALANCED MODULATORS



(b)

FIG.8



RING MODULATOR CARRIER CURRENT PATHS

FIG.9

Table I shows the most significant modulation products (up to and including sixth order) existing in the output of various modulators.

TABLE I. MODULATION PRODUCTS

Single diode	Series two-diode	Ring or four-diode
f_s	f_s	
$2f_s$		
$3f_s$	$3f_s$	
$4f_s$		
$5f_s$	$5f_s$	
$6f_s$		
f_c		
$f_c \pm f_s$	$f_c \pm f_s$	$f_c \pm f_s$
$f_c \pm 2f_s$		
$f_c \pm 3f_s$(*)	$f_c \pm 3f_s$(*)	$f_c \pm 3f_s$(*)
$f_c \pm 4f_s$		
$f_c \pm 5f_s$(**)	$f_c \pm 5f_s$(**)	$f_c \pm 5f_s$(**)
$2f_c$		
$2f_c \pm f_s$	$2f_c \pm f_s$	
$2f_c \pm 2f_s$		
$2f_c \pm 3f_s$	$2f_c \pm 3f_s$	
$2f_c \pm 4f_s$		
$3f_c$		
$3f_c \pm f_s$	$3f_c \pm f_s$	$3f_c \pm f_s$
$3f_c \pm 2f_s$		
$3f_c \pm 3f_s$	$3f_c \pm 3f_s$	$3f_c \pm 3f_s$
$4f_c$		
$4f_c \pm f_s$	$4f_c \pm f_s$	
$4f_c \pm 2f_s$		
$5f_c$		
$5f_c \pm f_s$	$5f_c \pm f_s$	$5f_c \pm f_s$

Single diode	Series two-diode	Ring or four-diode
$6f_c$		

(*) If the signal consists of two frequency components, f_1 and f_2, the modulator output includes $f_c \pm (2f_1 - f_2)$ and $f_c \pm (2f_2 - f_1)$

(**) Likewise the modulator output for an input of two frequencies, f_1 and f_2, includes the following: $f_c \pm (2f_1 - f_2)$, $f_c \pm (2f_2 - f_1)$ $f_c \pm (3f_1 - 2f_2)$, and $f_c \pm (3f_2 - 2f_1)$.

4.4. Translation to the Operating Frequency

Translation to the operating frequency is achieved by use of frequency changers. These frequency changers employ all types of modulators discussed above, and named as "mixers". In this process the SSB signal is used to modulate a high-frequency carrier whose frequency is such that the upper or lower sideband is on the desired operating frequency. This modulation process gives as a result products of frequencies, either the sum or the difference of the carrier and the modulating frequency. The important consideration here is the frequency stability of the carrier and the frequency accuracy since any error in the carrier frequency is passed on to sideband signal in exact proportion. The translation system consists of two basic components: the modulator which is generally called as a mixer, and the carrier commonly called oscillator signal.

4.5. Spurious Mixer Products

A balanced mixer, using a double triode 12AT7 is shown in Fig.10, and another balanced mixer ~~using transistors~~ is shown in Fig.11. As it was discussed earlier, to eliminate the spurious mixer products double-conversion exciters are used. These employ more than one stage of frequency changer. The modulation products are shown in Table I for various types of modulators, and Table II shows the calculated frequency products contained in the plate current of a 12AU7 triode mixer(‡).

(‡) "Dual Triode Mixers Power Series Coefficients", A report by Dr. V.W.Bolie 7/23/53.

TABLE II

$e_{osc} = P \cos pt = 2\ V_{rms}$ $\quad e_{sig} = Q \cos qt = 0.2\ V_{rms}$

$E_b = 250\ V \quad E_k = 10\ V \quad E_{bb} = 415\ V \quad R_L = 10\ K$

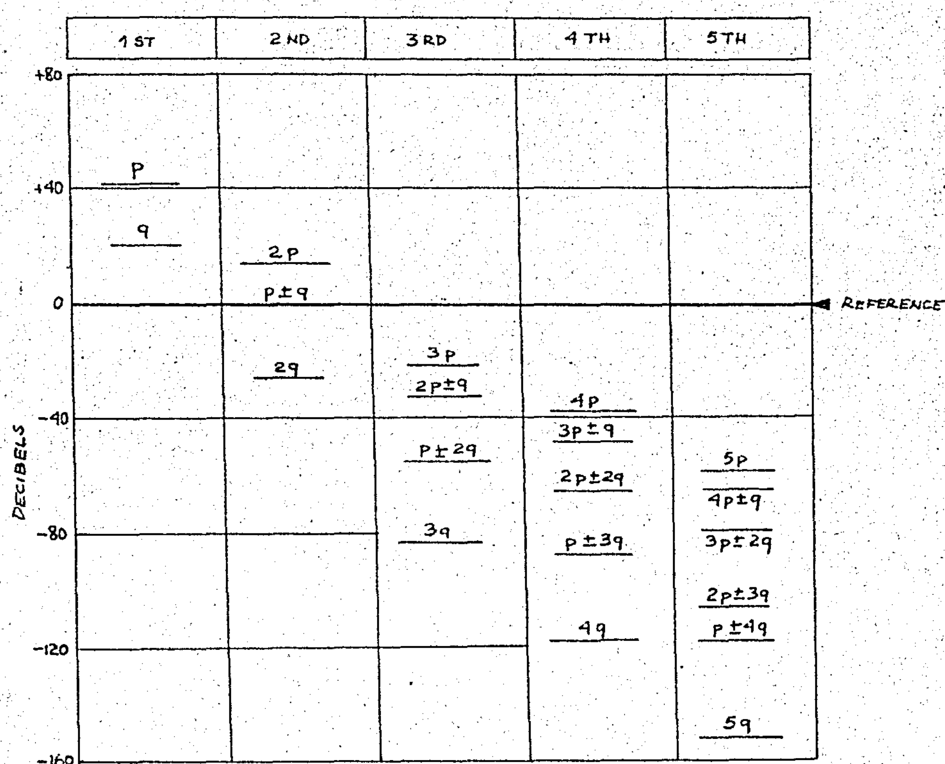
Table derived from power series expansion where

$$e_{in} = P \cos pt + Q \cos qt$$

Zero db reference is magnitude of $(p \pm q)$.

$$a_1=3.47\times10^{-4},\ a_2=1.47\times10^{-5},\ a_3=2.2\times10^{-7},\ a_4=3.7\times10^{-8}$$
$$a_5=5.7\times10^{-9}$$

Product Order



In order to obtain the desired sum or the difference frequency, it would be ideal to use a tube which had a characteristic curve of second degree only. Unfortunately, since all the tubes have characteristics with higher-degree of curvatures, this causes additional undesired components of frequencies. Sometimes these products are far away from the desired frequency that it is relatively easy to eliminate them. But in case that they are close to the desired frequency, or in the passband of the filter used, the problem becomes complicated.

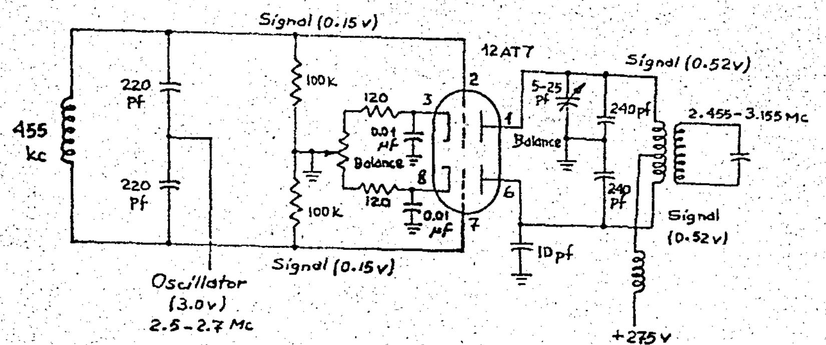


FIG.10

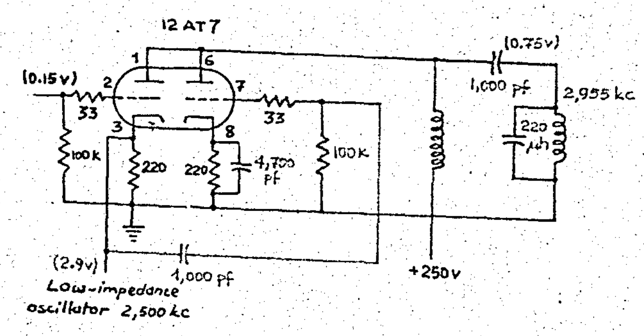


FIG.11

Table II shows that there are many undesired products which are greater in amplitude than the desired one, and also many which are smaller in amplitude than the desired one. Furthermore it shows that as the amplitude of the undesired products decreases, the order of the mixer product involved increases. If the signal and operating frequencies are chosen intelligently, the presence of undesired mixer products in the output of the frequency translation system may be minimized.

4.6. Oscillator Requirements

Since the frequency stability of the output signal depends on accuracy achieved in the carrier frequency and the frequencies of the frequency changers or mixers, the error involved is the sum of the errors present in each stage. As the error increases, a certain point is reached where intelligibility is degraded. This usually occurs at approximately 100 cps. Also the increase in error causes the naturalness of the reproduced speech to suffer first.

For various oscillators, the oscillator stability is given, for the stability is the most important factor in choosing the convenient oscillator. Frequency control techniques will be discussed in Chapter 6.

TABLE III. TYPICAL OSCILLATOR LONG-TERM FREQUENCY ERROR

Oscillator Type	Error cps			
	Error %	3 Mc	10 Mc	30 Mc
Variable Frequency Osc.	0.05	1500	5000	15000
Crystal Oscillator	0.005	150	500	1500
Temp. Controlled Cry.Osc.	0.001	30	100	300
Precision Standard Osc.	0.0001	3	10	30

4.7. Amplification

Amplification in SSB transmitters is usually achieved by high-gain tetrode tubes. Therefore, the output of an exciter may be limited to a fraction of a watt. This power output is applied to linear amplifiers in which it is amplified to such a degree that it is of sufficient magnitude to drive the last stage of linear power amplifier.

Tuned circuits are used in linear amplifiers of SSB exciter, because, on one hand, these tuned circuits constitute a suitable load resistance for the linear amplifier causing a sufficient voltage amplification, and, on the other hand, they act as filters which suppress the undesired frequency products in the frequency translation system.

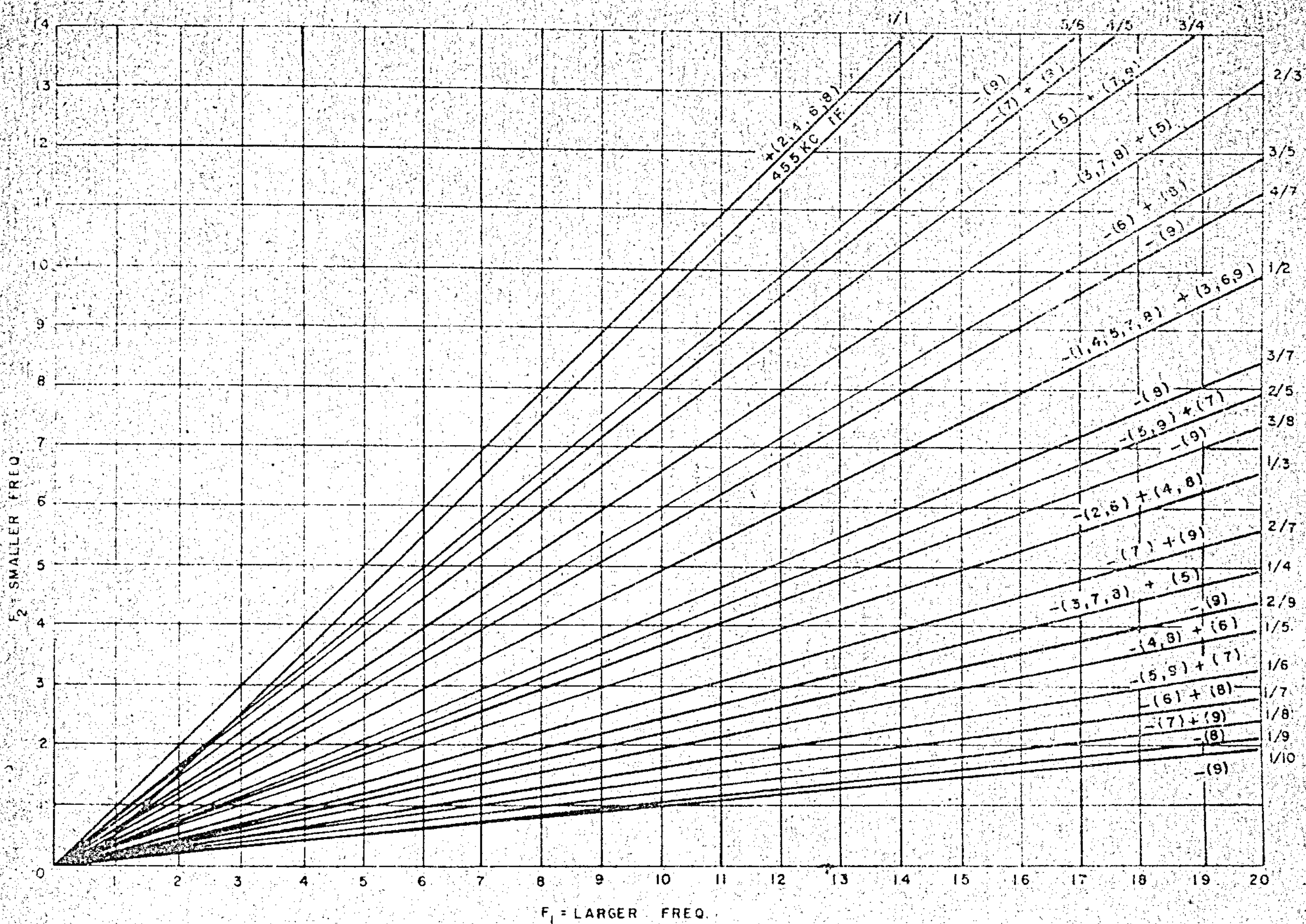
In order to eliminate the third order intermodulation product

Note: A "Spurious Response Chart" is added at the end of this chapter. In this chart spurious mixer products are plotted with respect to the signal and oscillator frequencies.

which is the main source of distortion in tuned linear amplifiers, low level modulation must be used in linear amplifiers of the SSB transmitter. But since the third order intermodulation products are always close to the desired signal or fall in the amplifier pass-band, it is necessary to limit the magnitude of the input signal so that the tube operates on a linear part of its characteristic.

The linear power amplifiers will be discussed in detail in Chapter 8 of this thesis.

SPURIOUS RESPONSE CHART



$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
1/1		•20 •02		•13 •31		•24 •42		•35 •53	
1/2	10		•12 •30	31	32	•33 •51	52	53	•54 •72
1/3		20		•22 •40		42 51		•53 •71	
1/4			30		•32 •50		52	71	
1/5				40		•42 •60		62	
1/6					50		•52 •70		72
1/7						60		•62 •80	
1/8							70		•72 •90
1/9								80	
1/10									90
2/3			21		•23 •41		43	53	

$F_2 \sim F_1$

ORDER	1	2	3	4	5	6	7	8	9
2/5				41			•43 •61		63
2/7							61		•63 •61
2/9									81
3/4					32		•34 •52		54
3/5						42		•44 •62	
3/7								62	
3/8									72
4/5							43		•45 •63
4/7									63
5/6									54

• INDICATES SUM MIXING
OTHER — DIFF MIXING

CHAPTER 5

SSB RECEIVERS

In the previous chapter the SSB exciter was considered. In this chapter we are going to discuss the SSB receiver considerations. As it was already discussed in the first chapter that reception process in SSB system is essentially the same as the demodulation process in the AM system except that a carrier which is identical to the one used in SSB exciter must be reinserted in the demodulation process. It was also said that previously a pilot carrier was transmitted in order to facilitate the reception process, although this is not necessary today with the use of highly precise frequency devices and high selective circuits. Furthermore, it was indicated in the previous chapter that when the frequency error exceeds 100 cps intelligence degradation occurs. Therefore, it is essential that the total frequency error of the system must be less than 100 cps.

Usually double-conversion superheterodyne circuits are employed in the SSB receivers. The purpose of using double-conversion method is, as in the case of double-conversion mixing in the transmitters, ~~is~~ to reduce the spurious demodulation products. High-frequency conversion is achieved by a highly stable crystal oscillator, and the low-frequency conversion employs a tunable oscillator.

The audio signal in the AM system is recovered from the r-f signal by means of an envelope detector. The same method can be also used for detection purpose in the SSB receiver provided that the amplitude of the reinserted carrier be high enough to keep the intermodulation distortion at a low level.

The avc (automatic volume control) circuit used in the SSB receiver is also somewhat different from the one that is used in the AM system. This arises from the fact that the carrier in the AM system is relatively constant and does not vary quickly in the amplitude. Therefore this avc system may have a relatively long time constant. On the other hand, since the signal in the SSB system changes very rapidly, this necessitates the use of quick acting type of agc (automatic gain control) rectifier.

Single-sideband receivers have three main sections: a radio-frequency section, an intermediate frequency section, and an audio-frequency section. This classification is the same as the one used in AM system.

5.1. Radio-Frequency Section

The radio-frequency section of the SSB receiver employs an r-f amplifier and one or more mixer stages. The purpose of the r-f section is to translate the r-f frequency to an intermediate frequency. Increased sensitivity is obtained by the use of an r-f amplifier at the first stage. The second advantage of using the amplifier at the first stage is that it reduces spurious products. Increased sensitivity results from the fact that the noise in the amplifiers is much lower than the noise in the mixers. Also, r-f filtering is used to eliminate the spurious products without much affecting the signal-to-noise ratio.

The main sources of noise in a SSB receiver are the antenna noise the input resistance of the receiver, and the grid circuit of the first amplifier tube. If the gain of the first amplifier tube is not high enough, the grid circuit of the second tube also contributes to the over-all noise.

The antenna noise is due to the thermal noise, which results from the random motion of electrons in the antenna. The noise voltage can be calculated by the following equation:

$$E_n = (4KTBR)^{1/2}$$

where E_n = rms noise voltage

K = Boltzmann's Constant, $1.38X10^{-23}$

T = absolute temperature in ^{o}K

B = bandwidth in cps

R = resistance in ohms

Therefore, it is apparent from this equation that the noise magnitude is proportional to the bandwidth.

The noise figure pertaining to the receiver is expressed as the ratio in decibels between the noise level of the receiver to the noise level of a perfect receiver, in which all the noise is assumed to be generated in the antenna by thermal agitation. A perfect receiver in which the input circuit is designed to match the antenna resistance has a noise figure of 3 db.

The tube noise can also be expressed as being equal to the noise generated in a resistance of the proper value and this is called the equivalent noise resistance of the tube. Equivalent noise resistances of the tubes are given in Table I. A functional diagram of a typical SSB receiver is given in Fig.1, and the noise sources in a receiver are shown in Fig.2.

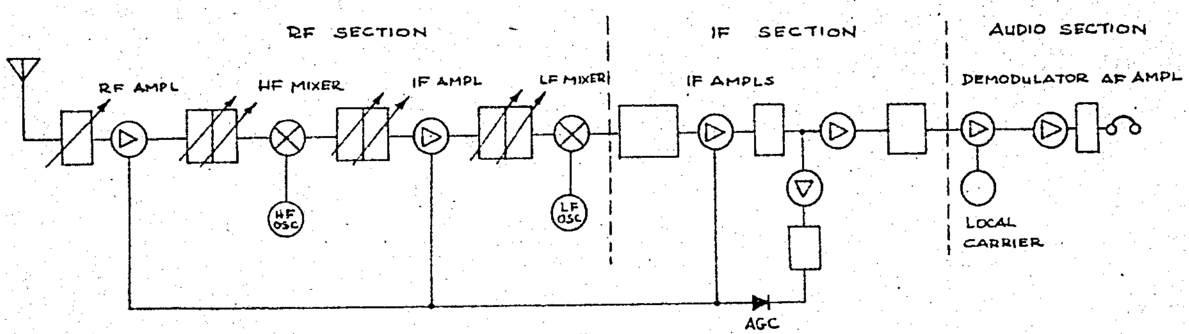


FIG.1

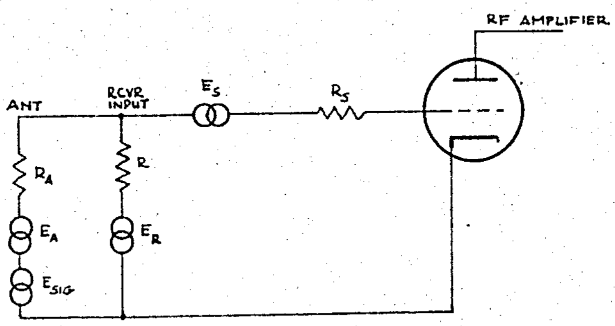


FIG.2

From the table it may be deducted that the triodes have lower noise than pentodes and amplifiers have lower noise than mixers.

TABLE I. EQUIVALENT TUBE NOISE RESISTANCES

Type	Application	gm or gc	Calculated R_{eq}
2C51	Triode Amplifier	5500	455
6AC7		11200	220
6AH6		11000	230
6AN4		10000	250
6BK75		6100	410
6BQ7A		6400	390
6BZ7		6800	370
6J4		11000	230
6J6		5300	470
6T4		7000	360
6U8		8500	295
12AT7		6600	380
12AU7		2200	1140
12AX7		1600	1560
12BH7		3100	810
5687		10000	250
5842		24000	105
6386		4000	625
6AG5	Pentode Amplifier	5000	1650
6AH6		9000	720
6AK5		5100	1880
6AK6		2300	8800
6AU6		5200	2660
6BA6		4400	3520
6BC5		5700	1350
6BD6		2350	13800
6BH6		4600	2360
6BJ6		3800	3860
6BZ6		6100	1460
6CB6		6200	1440
6U8		5200	2280

Type	Application	gm or gc	Calculated R_{eq}
2C51	Triode Mixer	1375	2900
6AN4		2500	1600
6J4		2750	1450
6J6		1575	2540
6T4		1750	2290
12AT7		1650	2430
12AU7		550	7280
12BH7		775	5170
5687		2500	1600
6386		1000	4000
6AG5	Pentode Mixer	1250	6600
6AK5		1280	7520
6BA6		1100	14080
6BC5		1425	5400
6BZ6		1525	5840
6U8		1300	9120
6X8		2100	7780
6BA7	Pentagrid Converter	950	61700
6BE6		475	174000
6SA7		450	240000
6SB7Y		950	61700

The ideal r-f selectivity would be obtained if tunable circuits were placed before the r-f amplifier. But this is impossible for mainly two reasons; first, it is impractical to use bulky tunable circuits to correspond to a large span of h-f frequencies, and the use of filters introduce such a high insertion loss that this degrades the noise figure considerably. Where it is necessary to use filtering before the r-f amplifier, only one tunable circuit is used. High selectivity is obtained by suppressing the spurious frequency products by using a tunable filter between the r-f amplifier and the mixer. The number of elements used in the tunable circuit is determined by the factor Q.

The generalized selectivity curve is given at the end of this chapter.

The r-f signal is translated into i-f signal by means of mixers. The mixers have been discussed in the previous chapter. But the problems encountered in receiver mixers are slightly different from those encountered in the mixers of transmitters. Typical receiver mixer spurious response is shown in Fig.3. A receiver mixer cross-over response is shown in Fig.4.

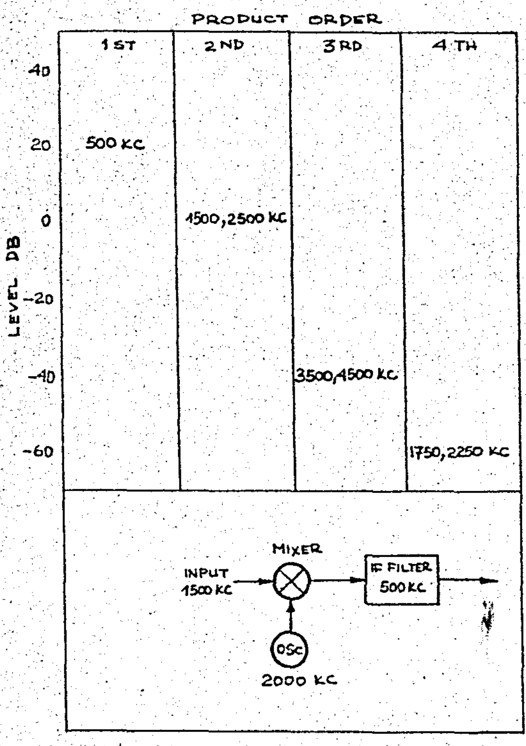


FIG.3

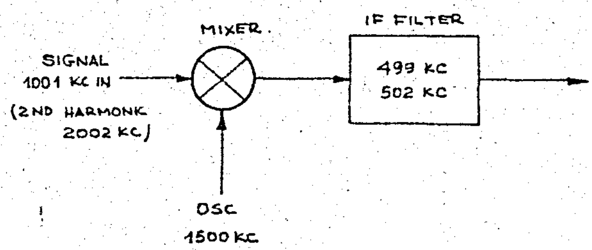


FIG.4

Crossover (tweet, or birdie) type of response is very important in receiver mixers. This is illustrated in Fig.4. Fig.4 shows that a 1001 kc is applied to the mixer along with the translating frequency of 1500 kc. Therefore, at the output of the filter appears the 499 kc desired frequency. But also, due to the second harmonic (2002 kc) of the signal frequency another output of 502 kc appears at the output of the mixer. These two signals separated only by 3 kc are passed by the i-f filter and demodulated to give a tweet along with the desired signal.

The best sensitivity is achieved by as much gain as possible ahead of the mixers. This means that the signal level is so strong to be sufficient to override all the noise from the mixers.

Cross modulation is not so important in SSB receivers as it is in AM receivers. The cross modulation results from a strong signal near the passband of the i-f filter which affects the passed signal when the filter is tuned to a weak signal. Since there is no carrier in the single sideband reception, the modulation is applied to each of the sideband signal components. Furthermore, since the single-sideband signal consists of a number of relatively weak components, this undesired modulation is spread, and the interfering signal is merely recovered as noise.

5.2. Intermediate-Frequency Section

This section is composed of two main sections, frequency selective filter elements and the amplifier stages. The selectivity of the filters used in the SSB receiver must be considerably higher than those used in the SSB transmitter to attenuate the undesired signals more than 60 db. In the case of extremely weak signals, the attenuation degree must be greater than 60 db even. Optimum selectivity occurs when the noise bandwidth is large enough to pass the required intelligence, and the skirt bandwidth is narrow enough to reject an undesired signal in the adjacent communication channel. These correspond to 6 db and 60 db respectively. Therefore this necessitates extremely steep skirts on the selectivity curve. A selectivity comparison is shown in Fig.5. This figure shows that the selectivity curve obtained by a Collins mechanical filter is very close to the ideal selectivity curve. On the other hand, double-tuned circuits employ overcoupled i-f transformers of Q's of 150.

The amplifier stage increases the signal to such a value to drive the demodulator stage. In the amplifier stage cascaded class A linear amplifiers are employed using remote cutoff pentode tubes. Tuned circuits are used as load resistances for these amplifier stages.

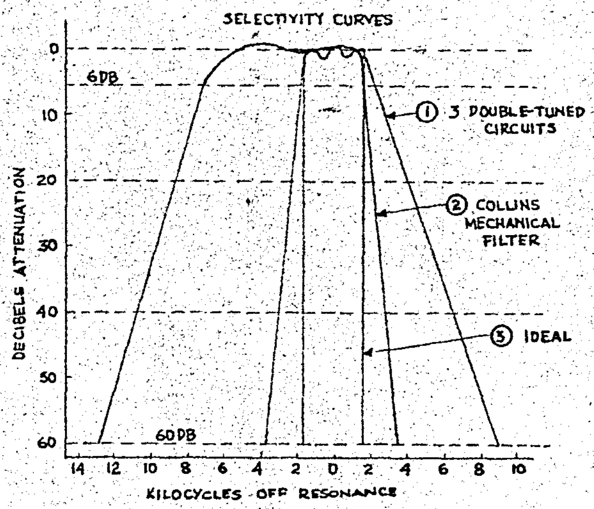


FIG.5

As it was mentioned above that the agc is different in the SSB receivers from that used in the AM receivers. Because conventional AM avc operates on the level of the carrier. Since there is no carrier in the SSB reception except the inserted one or the local carrier, the automatic gain control systems used obtain their information directly from the modulation envelope. On the other hand, since the inserted carrier is of high amplitude, special attention must be devoted to isolate it from the automatic gain control system. This problem is solved by taking the information for the agc directly from the audio signal voltage.

The time constant of the automatic gain control system is very important, because it must limit rapidly strong signal from becoming too loud, and also it must be slow enough to follow the syllabic variation of normal speech. This is satisfied by an RC circuit of 50-msec charge time and 5-sec discharge time. An AGC circuit is shown in Fig.6.

5.3. Audio-Frequency Section

The audio-frequency section employs demodulators to obtain the desired audio-frequency signal from the intermediate frequency along with the amplifiers necessary to amplify the signal to a level suitable for the audio output circuits. The single sideband is first combined with a local carrier of proper frequency relationship with the original audio signal. As it was discussed in the chapter of

Single Sideband Generation, either upper or the lower sideband is obtained according to where the carrier is located with respect to the passband characteristic of the low-pass filter. If the carrier is placed on the high-frequency side of the characteristic the lower sideband is obtained and vice versa.

In order to minimize the intermodulation distortion products present in the audio output signal product modulator circuits are used in the SSB reception. Also they do not require large local carrier voltages. A product modulator used in SSB reception is shown in Fig.7.

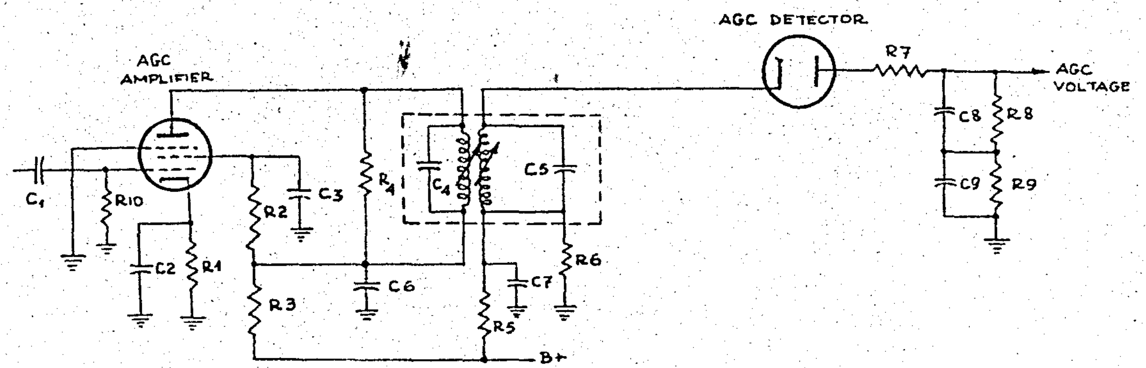


FIG.6

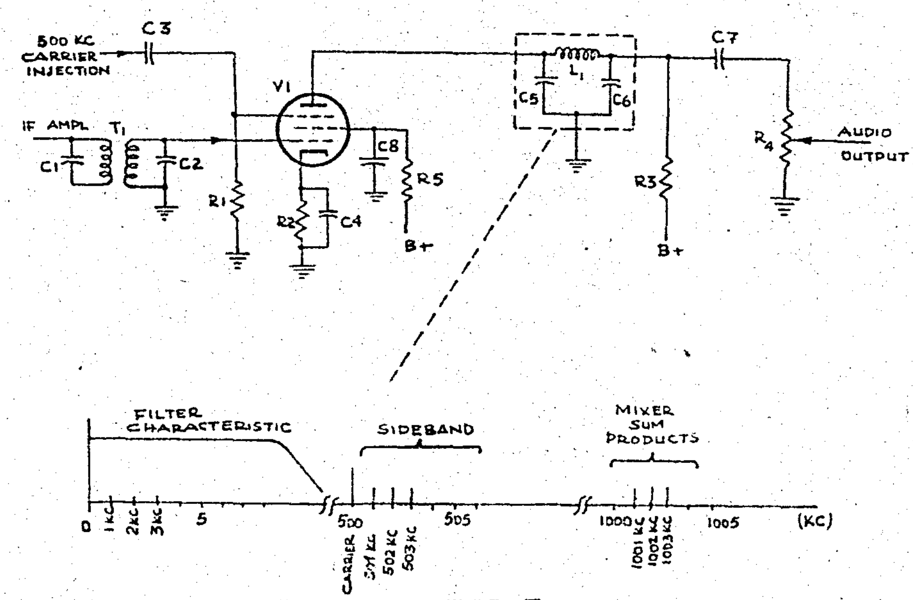
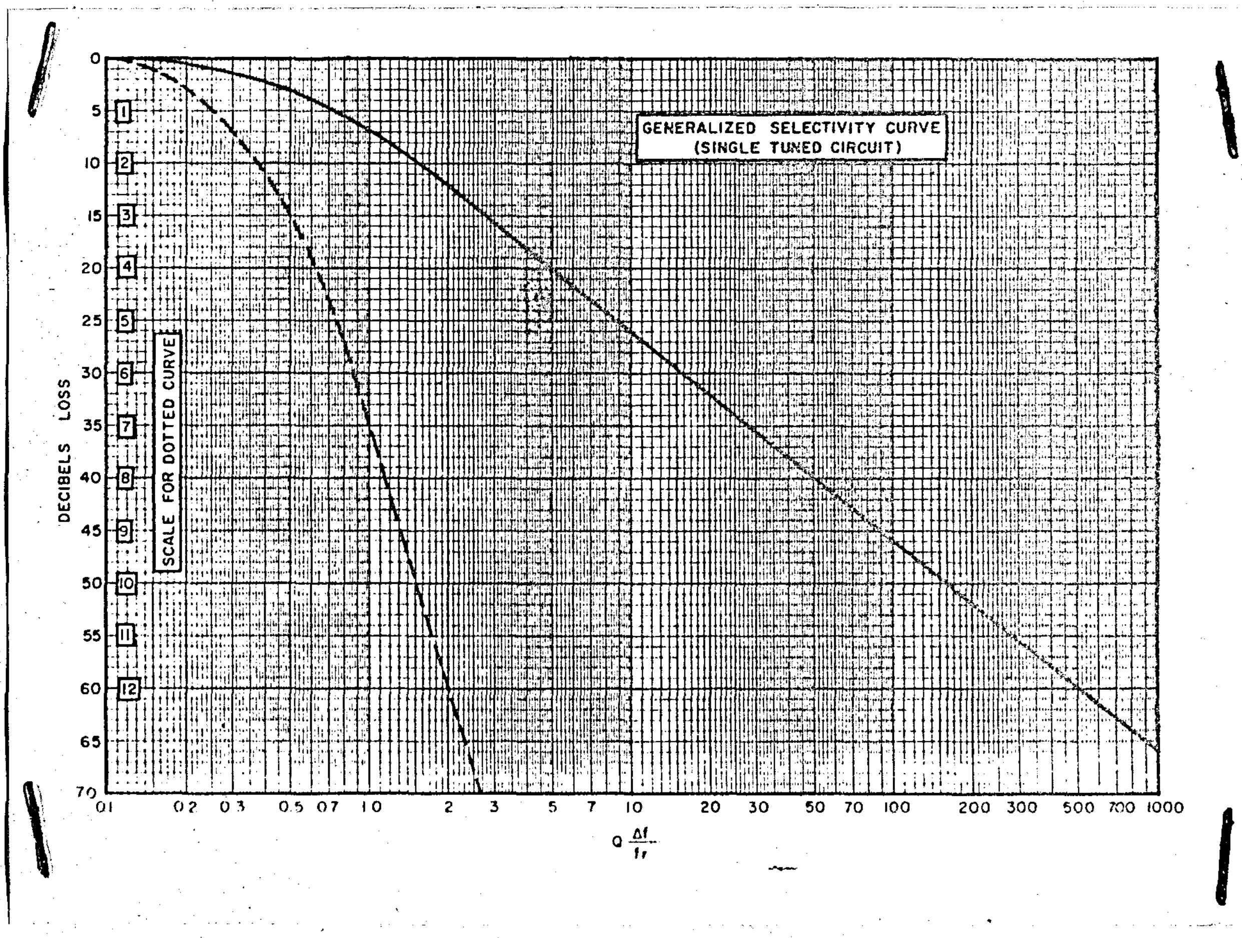


FIG.7


GENERALIZED SELECTIVITY CURVE
(SINGLE TUNED CIRCUIT)
DECIBELS LOSS
SCALE FOR DOTTED CURVE
0
5
10
15
20
25
30
35
40
45
50
55
60
65
70
1
2
3
4
5
6
7
8
9
10
11
12
01
0.2
0.3
0.5
0.7
1.0
2
3
5
7
10
20
30
50
70
100
200
300
500
700
1000
$Q \frac{\Delta f}{f_r}$

SUMMARY

In the first five chapters the general characteristics of the SSB signal are considered. First chapter was devoted to the general comparison of various modulation systems as an introduction. Second chapter is devoted to the consideration of the SSB system in general as compared to the AM and FM system. It was also discussed in this chapter that the SSB system has an over-all advantage over both the AM and FM although the comparison between the SSB and FM has not been fully developed yet. The economics of the SSB system was considered as compared to the AM and a brief history of SSB development was given.

Third chapter undertakes the single sideband generation. In this chapter single sideband generation techniques were discussed. In this chapter also a third method or "modified phase-shift" method of generation of single sideband was given.

The fourth and fifth chapters contain the general descriptions of single sideband exciters and receivers. The basic elements of the transmitters and receivers were given in these two chapters along with the due considerations to important specifications of the systems used especially in Collins equipment.

The following two chapters will be devoted to the frequency stabilization considerations and synthesizer stabilized SSB systems.

The last chapter will be on the subject of linear power amplifiers employed in the last stage of SSB transmitters. In this chapter linear power amplifiers will be discussed, and means of reducing the distortion in such amplifiers will be determined. As a specific example 208U-10 employed in the Transmitter Site of NATO, İzmir, is chosen. Furthermore, 30S-1 Linear Amplifier of Amateur Sideband Equipment of Collins is described shortly with its circuit diagram given.

CHAPTER 6

CONTROL OF FREQUENCY IN THE SSB SYSTEM

It was indicated throughout the five chapters that the frequency control in the SSB system is very important. This results from the fact that the single sideband suppressed carrier communication system does not use a carrier. Therefore, stability of the reinserted frequency at the receiver is the main point to be obtained. A frequency error in carrier reinsertion of 20 cps or less will give good voice reproduction. Frequency errors of around 50 cps will cause an appreciable distortion, and the reproduction will be impaired at 100-cps frequency error.

Before the World War II, the increased frequency accuracy was provided by crystal oscillators and a multiplicity of channels was provided by a like number of crystals. This is very impractical, and it was required that at a flick of a switch one of hundreds of channels is chosen immediately. Furthermore, the need for having narrower guard bands and utilizing the vhf bands in more extensive service necessitated the development of multiple crystal synthesizer soon. The operation principle was simple and depended upon the procedure of mixing the output frequencies of several crystal oscillators together to produce the desired output frequency. Each oscillator contained ten or more crystals to provide a multiplicity of channels.

6.1. Achievement of Frequency Stability in the SSB System

The first single sideband h-f systems did not use the techniques of frequency stability that are used today. In order to achieve a coherent detection either a pilot tone or carrier was transmitted along with the sideband components, and the receiver frequency was synchronized with the transmitting frequency. The use of crystal-controlled oscillators was sufficient to give the required amount of stability.

Automatic frequency control technique was not used in the first single-sideband radiotelephony because the operating frequency was as low as 60 kc so that the frequency stability requirement was met by the available oscillators. As the stability and the accuracy of the frequency were increased, the oscillators became bulky and fragile and limited in frequency channels. Therefore, the need for stability and accuracy was satisfied later by the development of the crystal frequency synthesizer.

Basically, the single crystal frequency synthesizer is a circuit in which harmonics and subharmonics of a single-standard oscillator are combined to provide a multiplicity of output signals which are all harmonically related to a subharmonic of the standard oscillator. The block diagram of such a synthesizer is given in Fig.1. The great advantage of this circuit comes from the accuracy and stability of the output signal which is essentially equal to that of the standard oscillator.

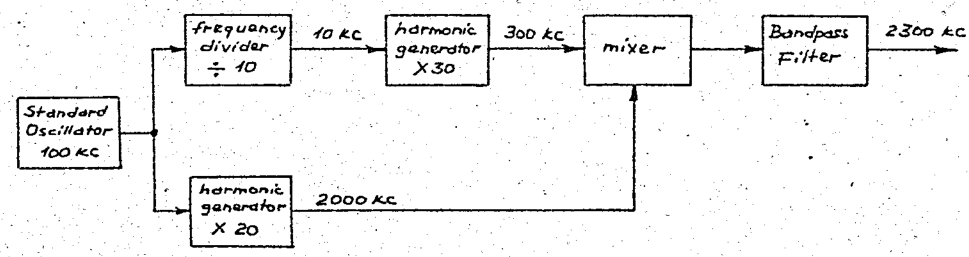


FIG.I

As the frequency increases the spurious frequency problems increase proportionally and the channel spacing decreases. Therefore, even in the simplest circuits, extensive filtering and careful selection of operating frequencies are required. This is the main difficulty encountered in the design of the frequency synthesizer.

If the synthesizer is to be used in the h-f range, the circuit is considerably simplified. Particularly in such cases, universal superheterodyne circuits are used. The block diagram is fiven in Fig.2.

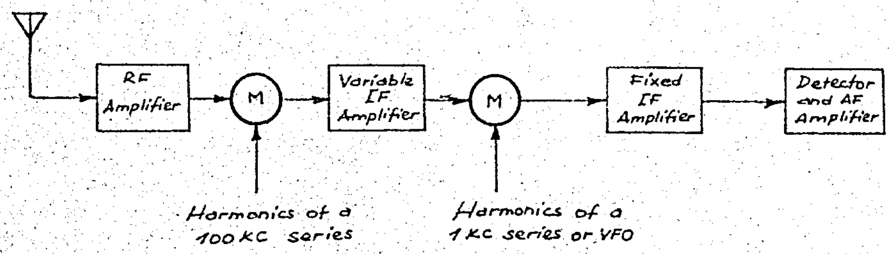


FIG.2

6.2. Variable-Frequency Oscillators (VFO)

The desired frequency channel is obtained by using a variable-frequency oscillator (VFO). The variable element in a VFO may be either a capacitor or an inductor. The main problem in the VFO is to obtain a linear frequency variation by shaft rotation. This requirement is met in variable-capacitor VFO by adding a large-value fixed capacitor in parallel with the variable capacitor. In a variable-inductance type of VFO a magnetic core is inserted and moved in or out by means of a lead screw. The basic principle is that the permeability of the flux path depends upon the core position. The two types of circuits are shown in Fig.3 and 4.

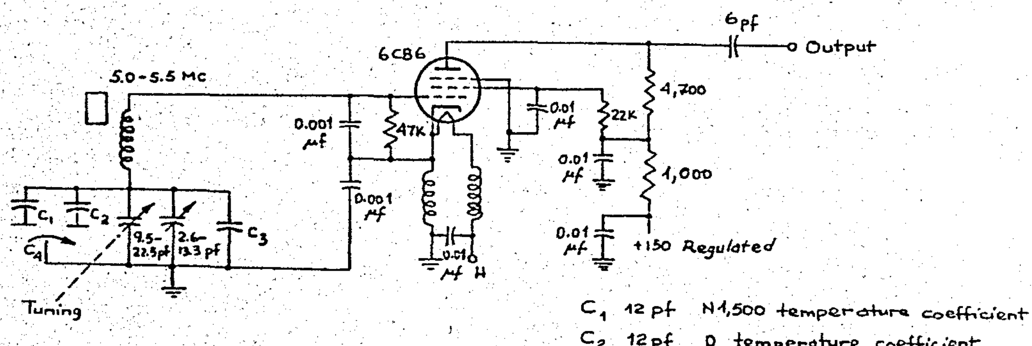


FIG.3

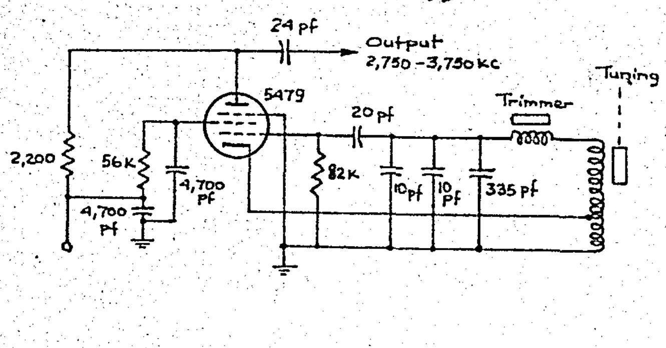


FIG.4

The capacitor C_3 provides a temperature compensation, and C_4 which is called a butterfly capacitor keeps the total capacitance across the tuning capacitor constant.

Supply-voltage variation, temperature change, michrophonics cause a shift in the selected frequency. By selecting a proper coil tap position and selecting a proper ratio screen to plate voltage the frequency variation is decreased.

6.3. The Stabilized Master Oscillator (SMO)

In present-day demands the figure of $\pm\frac{1}{2}$ part in one million presents the required absolute accuracy over long periods of time rather than short-term stability. Furthermore, providing either continuous coverage or channelized coverage in steps no greater than 4 kc is required in many SSB systems.

The variable frequency master oscillator is the main part of the stabilized master oscillator. This oscillator is capable of being locked to a reference signal derived from a standard oscillator of extremely high accuracy and stability. The master oscillator is stabilized by means of a feedback servo system deriving its error signal from the comparison of the phase of the master oscillator and the phase of the signal derived from the standard reference oscillator. The block diagram of a stabilized master oscillator is shown in Fig.5.

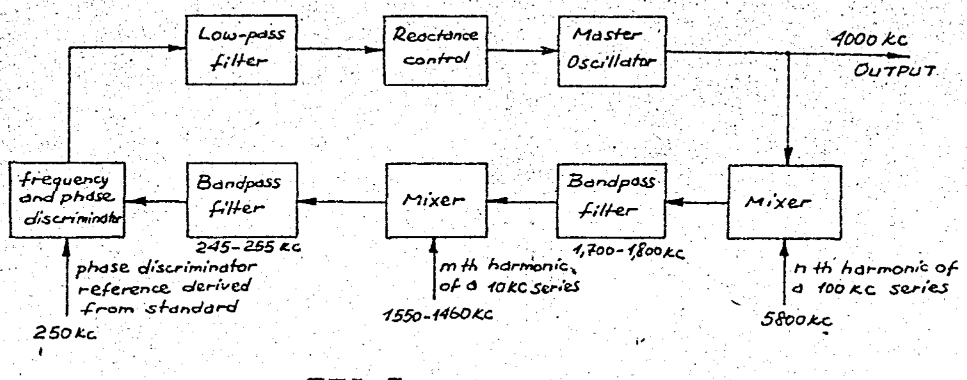


FIG.5.

The basic elements of a stabilized master oscillator are the master oscillator, reactance control and discriminator. These elements are shown in Fig.6. Inductance L_1 and capacitance C_1 determine the frequency of the master oscillator. The frequency of oscillation may be manually changed by varying the capacitance of C_1, or electronically changed by varying the permeability of the core on which L_1 is wound.

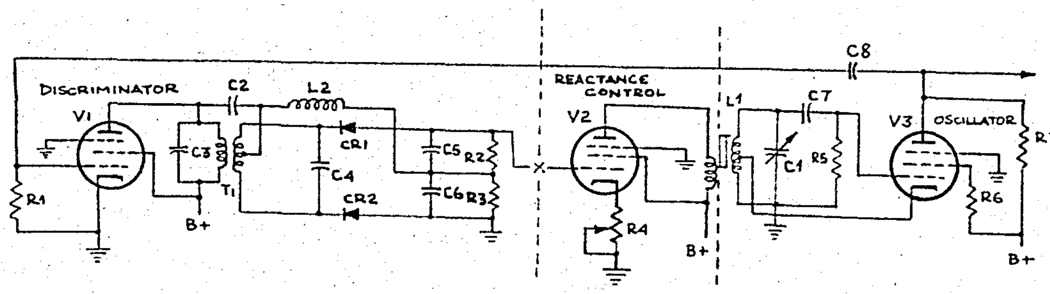


FIG.6.

Frequency discriminator provides a d-c output signal whose amplitude and polarity are determined by the relationship between the input signal frequency and the frequency to which the discriminator is tuned. The frequency discriminator consists of a double-tuned transformer and two diode rectifiers. The action of the discriminator depends on the fact that the phase of the voltage developed across the secondary of the discriminator transformer will vary as the frequency of the applied signal is varied above and below the transformer resonant frequency. The frequency discriminator circuit and the resultant output error voltage are shown in Fig.7.

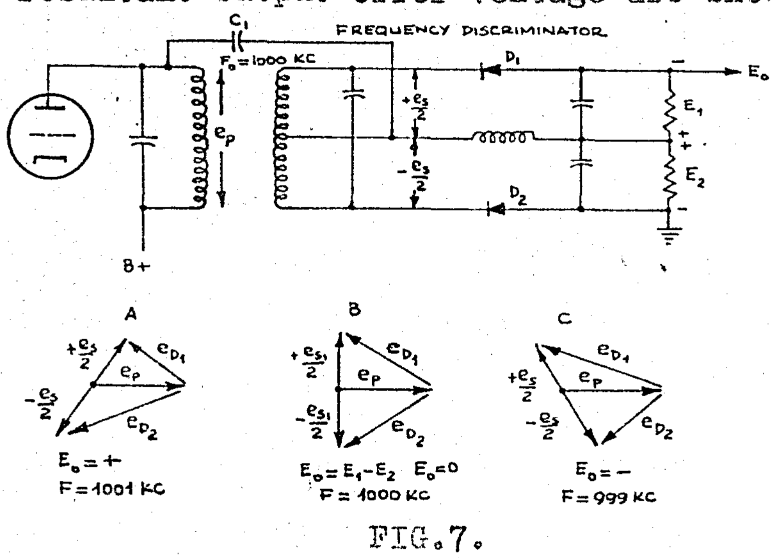


FIG.7.

The reactance control provides the means by which the direct current output of the discriminator is made to alter the inductance or capacitance of the tuning elements of the master oscillator. Reactance tube circuits, saturable reactors, voltage-sensitive capacitors, and motor-driven variable capacitors are used for this purpose.

The manner in which the stabilized master oscillator circuit operates may be described in two conditions, open-loop and closed-loop. If the control is opened at the grid of the reactance control tube and the tuning of the oscillator varied with the discriminator tuning fixed at f_o, the output voltage of the discriminator will follow the open-loop curve shown in Fig.8. If the master oscillator frequency differs from the discriminator frequency when the loop is closed, the master oscillator frequency will be pulled toward the discriminator frequency provided the proper polarity of discriminator and control device has been observed. It is important to realize that perfect correction cannot be achieved unless there is an infinite amount of amplification of the error signal from the discriminator. This can be seen by examining the discriminator output when the loop is closed. If perfect correction had somehow been achieved, the discriminator output would be zero. Obviously such cannot be the case as there must be some signal applied to the reactance control to correct the oscillator frequency error. The closed-loop frequency error will depend on the master oscillator error and on the gain of the control loop. The performance of the closed-loop is shown by the dashed curve in Fig.8.

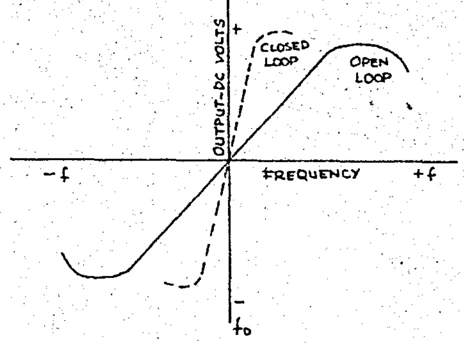


FIG.8.

In stabilized master oscillators stable operation can be obtained with loop bandwidth exceeding 400 cps, and enough gain to suppress microphonic disturbances arising from vibration and shock.

Careful attention must be paid to the frequencies used in the synthesizer mixer although the SMO synthesizer avoids many spurious frequencies. Otherwise, spurious frequencies falling within a few kc of the desired frequency will phase-modulate the master oscillator. The spurious frequencies outside this range will be easily suppressed and will not appear in the master oscillator output.

6.4. The Standard Reference Oscillator

Since frequency accuracy and stability of the stabilized master oscillator depend completely upon the standard reference oscillator, it is very important to employ a reference oscillator of the greatest precision obtainable within the limitations of the over-all equipment specification. The important factors in order to obtain a high frequency stability are various. First of all, time and changes in the environmental conditions must not effect the elements of the resonator circuit. A high degree of cleanliness is necessary in order to minimize the aging rate of the resonator. Therefore, the resonator should be sealed in an evacuated glass envelope. Furthermore, the temperature of the quartz crystal must be extremely closely controlled since the resonant frequency of the quartz resonator is dependent on the r-f power dissipated in it, especially at higher powers.

The coupling between the resonator and the active or amplifying portion of the circuit should be as small as possible while still maintaining sufficient coupling for sustained oscillations. The Q of the resonance network must be at least one million so as to allow very light coupling to the active network.

The ratio of the gain to the phase instabilities through the amplifying network must be maximized. Care in shielding is required between the oscillator stage and the following buffers so as to eliminate undesired feedback from these stages since this feedback generally has poor phase stability.

The block diagram of a typical reference oscillator is shown in Fig.9. The temperature of this resonator and allied critical components are held constant to better than 0.01°C during normal fixed station operation by means of an electronically-controlled oven. The 1-mc quartz resonator is a fundamental AT cut crystal sealed in an evacuated glass envelope. The temperature coefficient of this crystal resonator is only several parts in 10^7 per degree C. By varying the precision trimmer capacitor C-104 small adjustments of the oscillator frequency are accomplished. The amplitude of oscillation is controlled so that the resonator power dissipation is less than 0.1 μwatt, and so that no appreciable harmonics are present in this circuit.

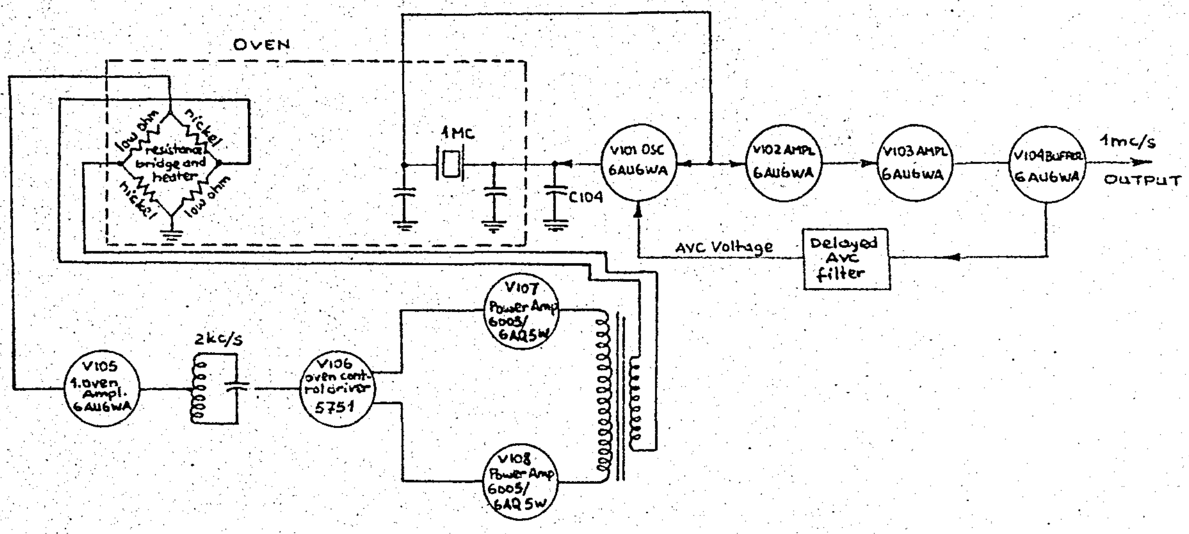
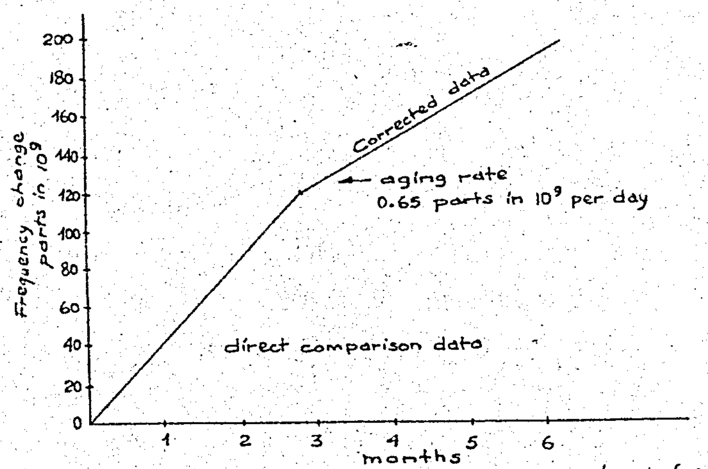


FIG.9.

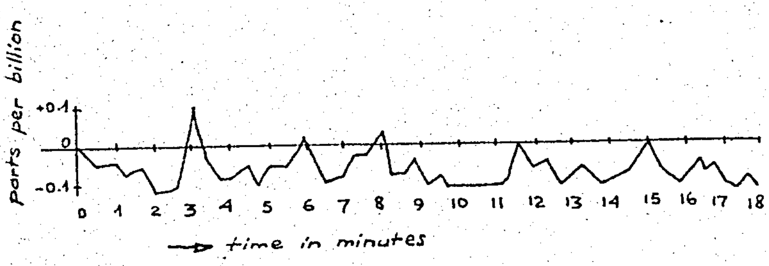
The temperature control oven is the basic part of the reference frequency oscillator. This oven maintains the resonator and other critical components at a constant temperature within very close limits. The oven is controlled at approximately 65°C, and is capable of delivering 8 w of heater power.

Typical short-term and long-term stability curves are shown in Fig.10 and Fig.11.



Long-term frequency stability

FIG.10



Short-term frequency stability

FIG.11

CHAPTER 7

SUMMARY ON FREQUENCY STABILITY

This chapter is a short summary of single-sideband system frequency stabilization, and sigle-sideband system operation in general.

A vector representation of sidebands of an amplitude modulated signal with respect to carrier is given in Fig.1. It was already discussed that the two sidebands in an amplitude modulated signal contain~~s~~ the identical information, and the carrier is only a means of translation of frequency from one band to the other. Nevertheless, the phase relations between the carrier and the sidebands are important from the intelligibility of the signal. In Fig.1, the amplitudes of the signal are represented by vector length, while vector direction indicates instantaneous phase with respect to the carrier. It is seen from the figure that the two sidebands are identically symmetrical to each other in both amplitude and phase about the carrier. If there is any deviation from this symmetry, the signal is no more amplitude modulated.

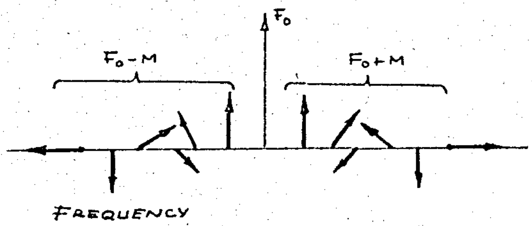


FIG.1.

7.1. Advantages of SSB Systems

As it was discussed in Chapter 2, the ionospheric return effects the amplitude modulated communication resulting in a kind of distortion known as selective fading. This selective fading was examined in detail with the results on the amplitude-modulated signal. The main discrepancy here arises from the fact that in amplitude modulation the carrier phase with respect to the sidebands is an important point to draw full attention. But in a single-sideband signal since the carrier is suppressed, there is no problem whatsoever concerning this point. Of course, single-sideband signals are subject to selective fading as are other types of modulated signals, but this fading does not generate intelligence destroying harmonic distortion and intermodulation in the SSB modulation. Selective fading produces what is known as amplitude versus frequency distortion in SSB systems, but this type of distortion only makes a voice signal sound peculiar and has little effect on the intelligibility.

Second advantage of the SSB systems over AM as was discussed is the channel size reduction. The SSB system utilizes only half of the bandwidth assigned to AM system. Furthermore, narrow-band phase and frequency modulation voice circuits usually require more than 8000 cycles because significant second- and third-order sidebands are produced as the modulation index approaches one; therefore SSB is considerably more economical from the point of bandwidth than either of these.

A third advantage of the SSB systems comes from the power requirement. Low power requirement reduces the transmitter size and cost considerably. For example, an AM transmitter, its transmission-line system, and antenna system would need to be capable of handling four times as high r-f voltages and currents as the SSB equipment capable of doing the same job. From the standpoint of a-c input required to operate the two transmitters, the SSB transmitter should require only approximately one-sixth as much power as the AM transmitter.

7.2. SSB Transmitter and Receiver

In AM transmitter frequency translation is performed at the frequency of the transmitted signal, and this may be either at high level as in the case of the plate modulated signal, or at a low level followed by several stages of linear amplification. Because of relatively complex filtering problems, the modulation problem is carried out at a relatively low fixed frequency in the SSB transmitter. This facilitates the design of satisfactory sideband filters. A simplified block diagram of a typical SSB transmitter is given again in Fig.2. The characteristic of the designed filter should be so as to provide moderately flat pass-band and a very sharp cut-off at each side of the passband, with especially high attenuation at the carrier frequency. As it was discussed earlier, the frequency translation process is carried out in two or more stages to eliminate the spurious mixer products. If the pilot carrier is transmitted, the level of this carrier is usually 10 to 20 db below the level of a normal AM carrier. All stages up to the linear power amplifier operate at about only 0.1 watt and linear power amplifier requires a peak input of only approximately 0.1 watt.

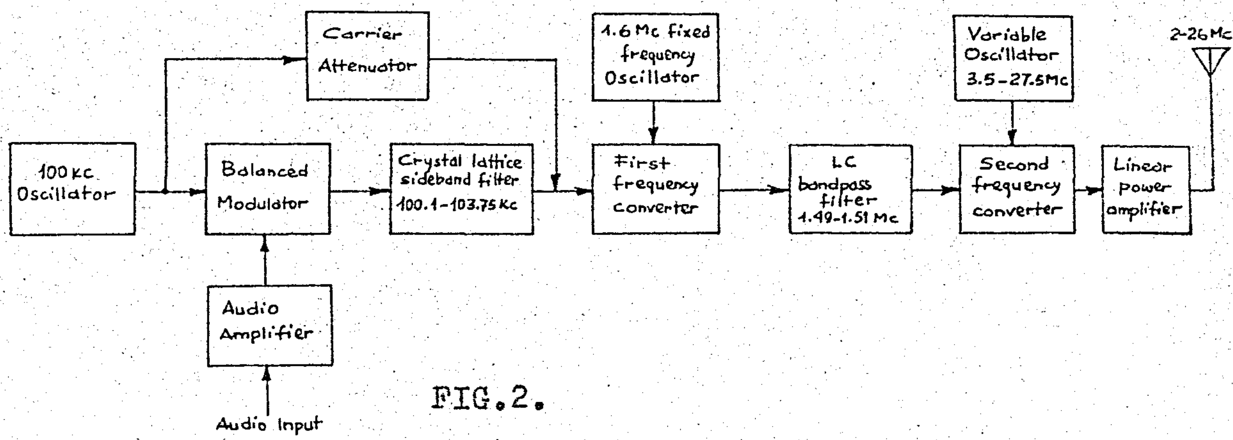


FIG.2.

A simplified block diagram of a SSB receiver is given in Fig.3. Usually the receivers employ more than one stage of frequency conversion so as to keep the level of the spurious mixer products sufficiently low. As it will be seen from Fig.3 that special parts employed in a SSB receiver are sideband and carrier filters,

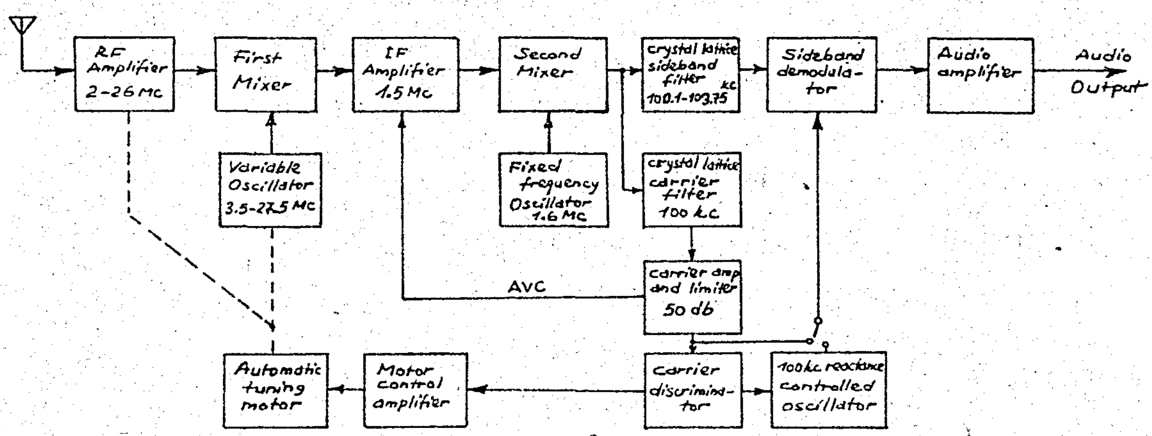


FIG.3.

carrier amplifier and limiter, carrier discriminator, automatic tuning mechanisms, and a special audio demodulator. All of these features are necessary to separate sidebands from the carrier, maintain the receiver in exact tune with the transmitted signal, and produce a signal that has a very low distortion and intermodulation level.

7.3. Precision Frequency-Controlled SSB Systems

It is very important to obtain a frequency stability and accuracy in both SSB transmitters and receivers. The inserted carrier in the receiver must be within a few cycles of the original modulation frequency carrier, since any instability of this carrier is reflected directly in the output. But at present time only the frequency stability and accuracy are not sufficient to establish and maintain a receiver in sufficiently accurate tune with transmitter to give satisfactory communications. Therefore, in order to avoid this difficulty, it is necessary to include some sort of automatic transmitter tracking mechanism that is capable of maintaining the receiver in proper tune once the receiver is accurately manually tuned to the desired signal. In order to obtain this automatic tracking, some continuous signal which is called the pilot carrier is transmitted along with the normal modulation products. But this pilot carrier is at level 10 to 30 db below normal AM carrier level. Fig.4 shows a block diagram of a precision frequency-controlled SSB transmitter. The stability and accuracy of the output frequency are determined entirely by the characteristics of the 100-kc precision oscillator, whereas in Fig.2 the accuracy and stability depend upon the accuracies and stabilities of the three oscillators employed.

In the SSB receiver, a carrier is inserted or the pilot carrier is first separated from the remainder of the signal by means of a very sharp filter, then amplified and limited. A block diagram of a precision frequency-controlled SSB receiver is given in Fig.5. Since a SSB system with precision frequency control of both transmitters and receivers has no need for automatic tracking of receivers, the elements of a conventional SSB receiver that are employed for tracking have been left out in the receiver of Fig.5.

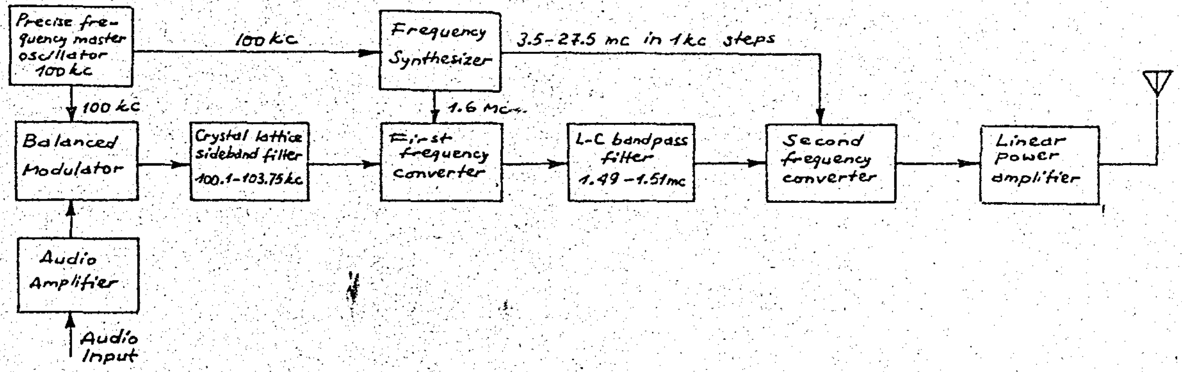


FIG.4.

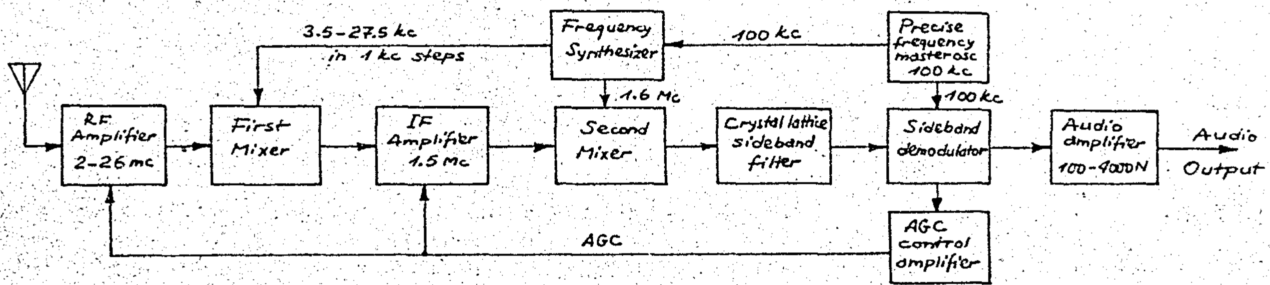


FIG.5.

The precision frequency control for SSB transmitters and receivers is derived through the use of precision master oscillators, followed by suitable frequency synthesizers. Today, precise frequency standards are designed to operate at only a few frequencies such as 100 kc, 1 mc, 2.5 mc, 5 mc. Therefore to convert these frequencies to other accurate frequencies some sort of frequency converter named synthesizer is used. In the design of synthesizers for h-f SSB use it has been suggested that it is only necessary for these devices to be able to generate accurate output frequencies at the 1000-cycle points. This type of design results in considerable simplification over one that would be capable of delivering continuous accurate frequency control over the complete range.

CHAPTER 8

LINEAR POWER AMPLIFIERS

This chapter is about the last stage of SSB transmitters. Since it is very important to sustain the frequency stability and accuracy, this stage should be given some consideration.

For r-f linear amplifier operation, the following features are desirable in the power amplifier tubes:

a. High gain
b. Low plate-to-grid capacitance
c. Good efficiency
d. Linear characteristics which are maintained without degradation at all frequencies in the desired operating range.

The needs for power amplifier tubes in the v-h-f and u-h-f ranges have spurred development of tubes suitable for operation at those frequencies. This has resulted in tubes with better performance in the h-f (3 to 30 mc) range. A typical comparison can be made between type 813 tube and the type 4X250B tube which are in the same power class. The small compact design of the 4X250B tube results in short lead lengths, better screening, closer element spacing and much higher performance which can be maintained easily over the h-f range. The ceramic construction, rather than glass, of an increasing number of new tubes promises to result in a more rugged and longer lifed tube. Ceramic sealed tubes which are now available include the RCA-6118 which is smaller than the 4X150A, the Eimac 4CX300A which has characteristics similar to the 4X250B, and all ceramic version of the 4X250B, the Eimac 4CX5000A which is capable of 10 kw of r-f output, and an RCA super-power, shielded-grid tube that will deliver 500 kw of r-f output. Tube manufacturers have additional types of power amplifier tubes under development which promise better performing tubes for the near future.

The Collins Radio Company has chosen to use high gain tubes of those types considered to be the best compromise of desired characteristics. At low signal levels, such as exist in exciters, conventional receiver-type r-f amplifier tubes are used. For delivering 0.1 watt output from exciters, the type 6CL6, which is a miniature 9-pin tube, is generally used. The 6CL6 is also frequently used to excite type 4X250B power amplifier tubes. The 4X250B tube is used in small, compact equipment for power levels of 1 kw by parallelling three, and for power levels of 500 watts by parallelling two. The type 4CX5000A is used for power levels of from 5 kw to 10 kw. This tube is used to obtain power levels up to 45 kw by parallelling four of them.

8.1. Basic Linear Power Amplifier Circuits.

For linear operation, r-f power amplifiers may be operated class A or class AB. The amplifiers used are quite conventional, being either grid driven or cathode driven (grounded grid) type amplifiers. However, the design considerations are extremely stringent to produce maximum linearity for a given tube in a given circuit. The tube operating point must be discreetly chosen and precisely maintained, neutralization must be as effective as possible, r-f feedback circuits are often used, and input and output impedances must be held as constant as possible. Generally class A pentode amplifiers are employed in low-level power stages to preserve linearity in these stages while producing enough power to drive the higher level stages. Class AB_1 or AB_2, triode or tetrode power amplifiers are employed in the high-level power stages to obtain the desired power output.

a. Grid Driven Triode Power Amplifier

Fig.1 is a simplified schematic of a typical grid driven triode power amplifier. This amplifier, operating class AB_1, produces up to 2.5 kw using the 3X3000A-1 triode. The triode tube, having a large plate-to-grid interelectrode capacitance, always requires neutralization to prevent oscillation when used in the grid-driven circuit. The only types of triodes capable of class AB_1 operation are the low amplification factor types, such as 3X3000A-1. Due to the low amplification factor, very high r-f grid excitation voltage is required, on the order of 1000 volts for 3X3000A-1. A similar

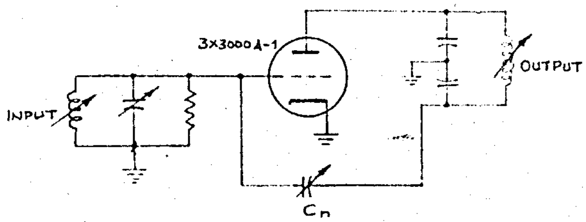


FIG.1

tube suitable for class AB_2 operation is the 3X2500A-3 which has an amplification factor of 20. This medium-μ triode requires less grid swing, but it requires grid driving power for class AB_2 operation. Neutralization, of course, is still required.

b. Cathode Driven Triode Power Amplifier

Fig.2 is a simplified schematic of a typical cathode driven (grounded grid) triode power amplifier. This amplifier, operating

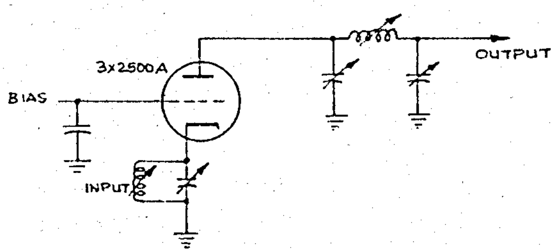


FIG.2

class AB_2 produces 4 to 5 kw using the type 3X2500A triode. In the cathode driven amplifier, the control grid is at r-f ground and the signal is fed to the cathode. The main advantage of operating the triode in this manner is that the control grid becomes an effective screen between the plate and the cathode making neutralization seldom necessary. The small values of plate-to-cathode capacity have very little effect on the input signal because the imput circuit impedance is usually quite low. Since neutralization is not required, triodes with an amplification factor of 20, such as the 3X2500A, can be used.

c. Grid Driven Tetrode Power Amplifier

Fig.3 is a simplified schematic of a grid driven tetrode power amplifier. This amplifier operating class AB_1 produces 250 watts per tube using the type 4X250B tetrode. In general the same design considerations exist for tetrode amplifiers as for triode amplifiers. That is, grid circuit swamping is required to hold the input impedance constant if the tetrode is driven into the grid current region, and neutralization is generally required if the tube is to operate over the entire h-f range. However, since the plate-to-grid capacitance is small in the tetrode, neutralization is much simpler. The

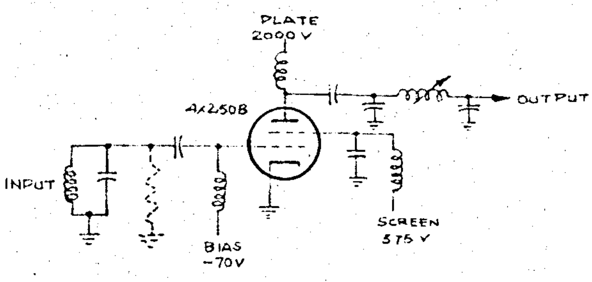


FIG.3

tetrode amplifier, being a high gain tube, requires relatively little driving power and a relatively small grid swing for operation. This permits the parallelling of tubes with a common input network and a common output network which reduces the number of stages and simplifies tuning. In the tetrode power amplifier, the screen voltage has very pronounced effects on the dynamic characteristics of the tube. By lowering the screen voltage, the static current required for optimum linearity is lowered. This permits greater plate r-f voltage swing which improves efficiency. The use of lower screen voltage has the adverse effect of increasing the grid drive for class AB_2 operation and lowering power output for class AB_1 operation. The tetrode tube can be used without neutralization in the high-frequency range.

8.2. Power Amplifier Output Networks

a. Tank Circuit Considerations

The plate tank circuit of an r-f power amplifier must perform four basic functions:

1. It must maintain a sine wave r-f voltage on the plate of the tube.
2. It must provide a low impedance path from plate to cathode for harmonic components of the plate current pulses.
3. It must provide part or all of the necessary attenuation of harmonics and other spurious frequencies.
4. It must provide part or all of the impedance matching from the tube plate to the antenna.

In addition, for many uses the output circuit should be single ended so that it will feed into a 52-ohm coaxial transmission line. A 52-ohm coaxial transmission line is desirable because it prevents stray r-f radiation near the transmitter; it is convenient for coaxial r-f switching; it is a convenient impedance for additional r-f filtering, and because it is ideal for directional wattmeter installation. For simplicity of operation, the output circuit should require a minimum of tuning controls. A direct-coupled network, such as the π-L network, is the most suitable network to meet these requirements.

The Q of the plate circuit, of which the tank is a part, must be sufficient to keep the r-f plate voltage close to a sine wave shape. If the plate circuit Q is insufficient, the r-f waveform may be distorted which will result in low plate efficiency. This loss of efficiency is seldom noticed unless the plate circuit Q is less than 5. A plate circuit Q of at least 10 is known to be sufficient for linear operation and is a recommended minimum.

In a linear power amplifier, the second harmonic component can be as great as 6 db below the fundamental at full peak envelope power. The higher order harmonic components drop off rapidly but their magnitude varies greatly, depending upon the pulse shape. These harmonics must be attenuated in the otput network so that they are 50 db, 80 db, or even further, below the fundamental component. The π-L network will attenuate the second harmonic to about 50 db lelow the fundamental, which is from 10 db to 15 db more attenuation than can be obtained from the simple π network. Where more attenuation is required, external filters of either the low-pass or band rejection type are added. Increasing plate circuit Q increases harmonic attenuation, but since doubling the Q results in only 6 db more second harmonic attenuation, Q's above 20 are seldom used below 30 mc.

The π-L output network is ideally suited to matching a tube load to a 52-ohm coaxial transmission line. Loads with a standing wave ratio as high as 4 to 1 can be matched easily. This can be done with any value of tube load impedance, whereas the simple π network has difficulty matching to low load impedance when the tube plate load resistance is high.

The π-L network has only four variable elements, and they can be ganged to have only a tuning control and a loading control, as shown in Fig.4. Since in the π-L network, C_2 and L_2 affect loading in the same direction, the extra capacity and inductance range of the elements required to extend the loading range of the circuit is relatively small. For example, the loading control varies about ±25 per cent to match a 52 ohm load with a 4:1 swr. The tuning control varies about ±10 per cent.

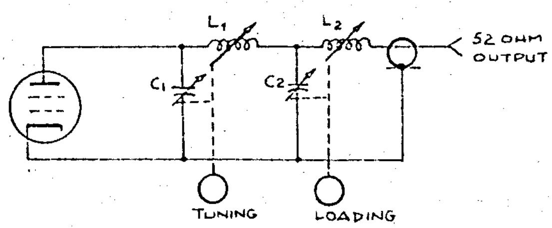


FIG.4

b. Circuit Losses

Nearly all of the tank circuit loss occurs in the coils. These losses are closely related to the ratio of plate circuit coil Q to plate circuit Q, but other design considerations enter in. These circuit losses are shown in Fig.5 for a π-L network, which has lower losses than other networks for 50 db of second harmonic attenuation. Resistances r_1 and r_2 represent the equivalent series resistance of the coils determined from coil Q and reactance. Resistance r_q is the equivalent load resistance in series with L_1 and is determined from the relationship

$$r_q = \frac{R_L}{Q^2 + 1} = \frac{R_L}{(R_L/X_C)^2 + 1}$$

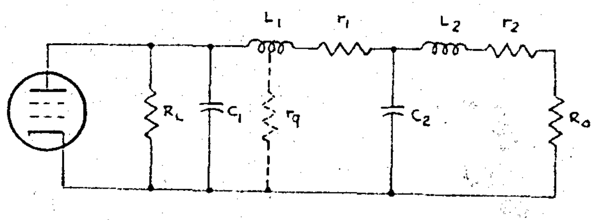


FIG.5

Resistance R_a is the series resistive component of the load. The π-L network loss is given by the equation:

$$\text{Per cent loss} = \left(\frac{r_1}{r_q + r_1} + \frac{r_2}{R_a + r_q} \right) \times 100$$

c. Tank Coil and Capacitor Requirements

The frequency range and method of tuning are major factors in determining tank circuit components. Continuously variable coils and capacitors which will cover the entire frequency range without any band switching are the most desirable. However, this is not practical in Autotune transmitters because of the limited torque available to drive the tuning elements and the often short repositioning time specified. With these limitations, bandswitching is almost essential. Where instantaneous frequency change is specified, it is common to switch from one pretuned r-f unit to another and manually tuned circuits are suitable for this purpose.

The use of continuously variable elements has the following advantages:

1. The circuit Q can be kept more uniform across the frequency range.
2. The circuit losses can be kept to a minimum.
3. The range of variable coils and capacitors can be less.
4. A maximum amount of harmonic attenuation is more easily maintained across the frequency range.

Variable vacuum capacitors are widely used in transmitters with power levels of 1 kw and higher. Their added expense is often justified by the added capacity range, small size, and low series inductance, especially where voltages above 2500 volts are employed. Variable tank coils are usually constructed with a rotary coil and either a sliding or rolling contact that traverses the length of the coil as it is rotated. The unused turns are shorted out to keep high voltages from developing in them. The series self-resonant frequency of the shorted-out section must not be near the operating frequency or high circulating currents will develop and cause, appreciable power dissipation.

8.3. Neutralization

a. Effects of Plate-to-Grid Capacitance

The purpose of neutralization is to balance out the effect of plate-to-grid capacitive coupling in a tuned r-f amplifier.

In a conventional tuned r-f amplifier using a tetrode tube, the effective input capacity is given by the following equation:

$$\text{Input capacitance} = C_{in} + C_{gp}\,(1 + A \cos\theta)$$

where C_{in} = tube input capacitance,
C_{gp} = plate-to-grid capacitance,
A = voltage amplification from grid to plate,
θ = phase angle of the plate load.

In an unneutralized, 4-1000A tetrode amplifier with a gain of 33 the input capacity of the tube with the plate circuit in resonance is increased 8.1 $\mu\mu F$ due to the unneutralized plate-to-grid capacity. This small increase in capacitance is not particularly important in amplifiers where the gain remains constant, but if the gain does vary, serious detuning and r-f phase shift can result. The gain of a tetrode or pentode r-f amplifier operating below plate saturation does vary with loading so that if it drives a following stage into grid current, the loading increases and the gain falls off.

The input resistance of the grid is also affected by the plate-to-grid capacitance. The input resistance is given by the following equation:

$$\text{Input resistance} = \frac{1}{2\pi f \; C_{gp}(A \; \text{Sin} \; \theta)}$$

b. Neutralizing Circuits

Most of the neutralizing circuits developed for use with triodes may be used equally successfully with tetrodes. However, those circuits which require balanced tank circuits for neutralizing purposes only, are undesirable because the trend in r-f power amplifier design is toward single-ended stages.

A conventional grid neutralized amplifier is shown in Fig.6. Capacitor C_3 balances the grid-to-filament capacity to keep the grid circuit in balance. When $C_1 = C_2$ and $C_n = C_{gp}$, it is readily seen that a signal introduced into the grid circuit will not appear across the plate circuit because the coupling through C_n is equal and opposite to the coupling through C_{gp}. The relationship for no coupling from the grid circuit to the plate circuit is given by the relationship

$$C_1/C_2 = C_{gp}/C_n$$

This indicates that the grid tank circuit need not be balanced to ground. If C_2 is made larger, then C_n must be made correspondingly larger. In a tetrode amplifier, C_{gp} is very small (approximately 0.1 $\mu\mu f$) so that practical values, 5 $\mu\mu f$, can be used for C_n when C_2 is very much larger than C_1.

By placing most of the grid tuning capacitance across the grid tank coil, using the bypass capacitor C from the bottom end of the grid tank circuit to ground for C_2, and using the grid-to-filament capacity for C_1, the modified grid neutralized circuit shown in Fig.7 results. The relationship for neutralization of this circuit

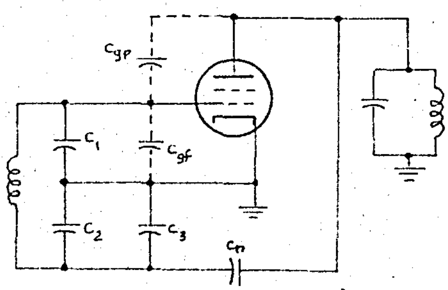


FIG.6

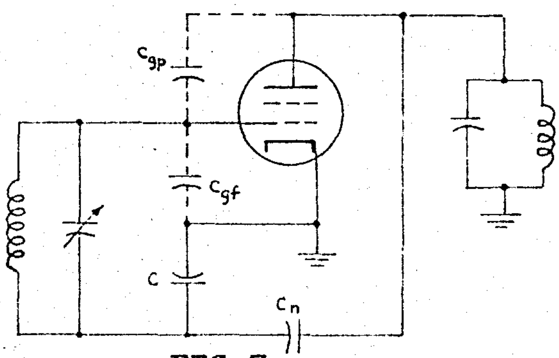


FIG.7

is given by the relationship

$$C_n/C = C_{gp}/C_{gf}$$

This relationship assumes perfect screen and filament bypassing and negligible effect from stray inductance and capacity. This modified grid neutralizing circuit is very effective for neutralizing tetrode power amplifiers and is accomplished with single-ended tuning elements.

8.4. r-f Feedback Circuits

An r-f feedback is a very effective means of reducing distortion in a linear power amplifier. 12 db of r-f feedback produces nearly 12 db of distortion reduction, and this distortion reduction is realized at all signal levels. However, voltage gain per stage is reduced by the amount of feedback employed, so that with 12 db of feedback the gain is reduced to one-quarter.

a. Feedback Around One Stage

Fig.8 shows a negative feedback circuit around one-stage r-f amplifier. The voltage developed across C_4 is introduced in series with the voltage developed across the grid tank circuit and is in phase opposition to it. The feedback obtainable with this circuit can be varied between zero and 100 per cent by properly choosing the values of C_3 and C_4. It is necessary to neutralize this feedback amplifier, the neutralization requirements being

$$C_{gp}/C_{gf} = C_3/C_4$$

To satisfy the neutralization requirement, it is usually necessary to add capacity from the plate to the grid.

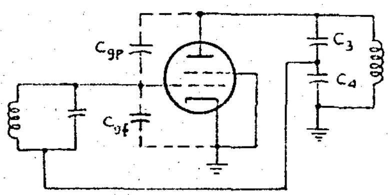


FIG.8

b. Feedback Around Two Stages

Feedback around two r-f stages has the advantage that more of the tube gain can be realized while nearly as much distortion reduction can be obtained. For instance, 12 db feedback around two stages provides about the same distortion reduction as 12 db around each of two stages separately. Fig.9 shows a negative feedback circuit around a two stage amplifier with each stage neutralized. The small feedback voltage required is obtained from the voltage divider C_6 and C_7. This feedback voltage is applied to the cathode of the first stage. The feedback divider can be left fixed for a wide frequency range since C_6 is only a few micromicrofarads. For example,

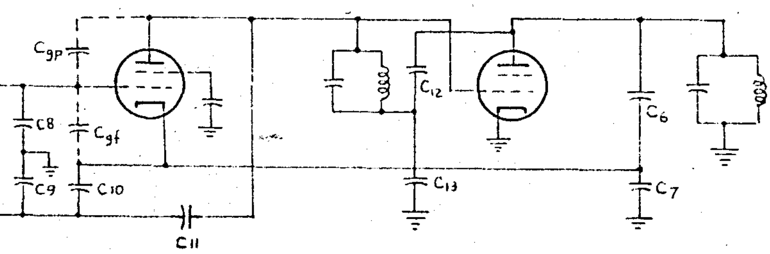


FIG.9

if the combined tube gain is 160 and 12 db of feedback is desired, the ratio of C_7 to C_6 may be 400 $\mu\mu f$ to 2.5 $\mu\mu f$. Either inductive input coupling or direct capacitive coupling may be used with this circuit, and any form of output coupling can be used.

It is necessary to neutralize the cathode-to-grid capacity of the first tube in the two stage feedback circuit to prevent undesirable feedback coupling to the input grid circuit. The relationship for the circuit which accomplishes this cathode-to-grid neutralization is

$$C_8/C_9 = C_{gf}/C_{10}$$

In a two stage feedback amplifier, the voltage fed back to the cathode of the first stage must be in phase with the grid input signal, the resultant grid-to-cathode voltage increases as shown in Fig.10. When the output circuit is properly tuned, the resulting grid-to-cathode voltage on the first tube is minimum which will make the voltage across the interstage tank circuit minimum also.

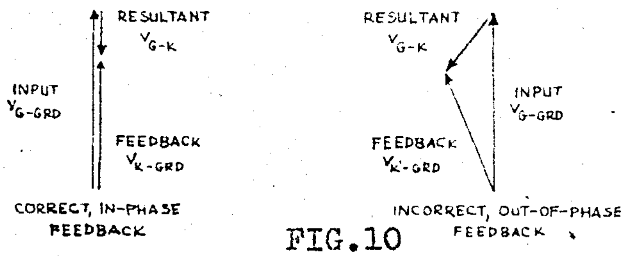


FIG.10

8.5. Automatic Load Control

Automatic load control is a means of keeping the signal level adjusted so that the power amplifier works near its maximum power capability without being overdriven on signal peaks. In AM systems it is common to use speech compressors and speech clipping to perform this function. However in a SSB system these methods are not equally useful because the peaks of the SSB signal do not necessarily correspond with the peaks of the audio signal. Therefore, the most effective means of control is obtained by a circuit which receives its input from the envelope peaks in the power amplifier and its output to control the gain of the exciting signal. Such a circuit is an automatic load control (alc) circuit.

Fig.11 is a simplified schematic of an alc circuit. This circuit uses two variable gain stages of remote cutoff tubes, such as 6BA6, operating very similarly to the i-f stages of a receiver with automatic volume control. The grid bias voltage of the variable gain amplifiers is obtained from the alc rectifier connected to the power amplifier plate circuit. The capacity voltage divider steps down the r-f voltage from the power amplifier plate to about 50 volts for the rectifier. A large delay bias is used on the rectifier so that no reduction of gain takes place until the signal level is nearly up to full power capability of the power amplifier. The output of the alc rectifier passes through RC networks to obtain the desired attack and release times. Usually a fast attack time, about two milliseconds, is used for voice signals so that the gain is reduced rapidly to remove the overload from the power amplifier. After a signal peak passes, a release time of about one-tenth second returns the gain to normal. A meter calibrated in decibels of compression is used to adjust the gain for the desired amount of load control.

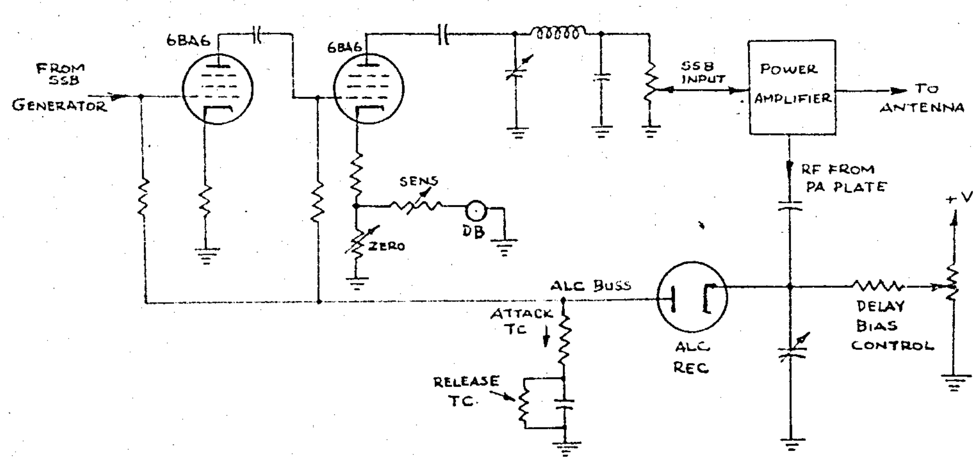


FIG.11

In a single channel speech transmission, the alc circuit performs the function of a speech compressor. To do this a range of 12 db is usually provided with control maintained on input peaks as high as 20 db above the threshold of compression. Since the signal level should be fairly constant through the preceding SSB generator, it is unlikely that more than 12 db for the SSB generator, a speech compressor in the input audio amplifier is usually used to limit the range of the signal fed into the SSB generator.

Fig.12 shows the effectiveness of the alc circuit in limiting the output signal to the capabilities of the linear power amplifier. An adjustment of the delay bias will put the threshold of compression at the desired level.

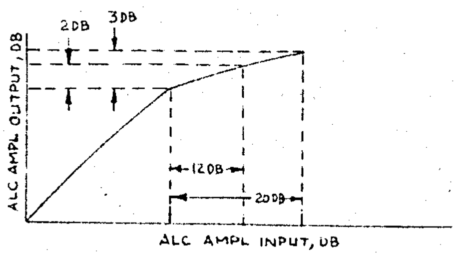


FIG.12

8.6. Linear Power Amplifier Tuning

When a power amplifier is operated class C, a pronounced plate current dip and grid current peak are fairly accurate indications of proper tuning. In a linear power amplifier, the use of these indications are limited. For instance, in a class A amplifier there will be no plate current dip; therefore, the class A amplifier output circuit must be tuned for an indication of maximum input to the next stage. In class AB amplifiers, the plate current dip is not always readily detected. This does not mean that conventional tuning procedures will not properly tune a linear amplifier, but tuning a linear amplifier with conventional procedures is much more exacting. One procedure commonly used is to increase the drive to a stage in order to obtain a good plate current dip indication.

In low Q tank circuits, the point of plate current dip is not a true indication of exact resonance because the plate current dip occurs at maximum impedance rather than when the tank circuit is pure resistive. This is especially true for π networks and π-L networks. For instance, in a network with a Q of 10, the phase angle at maximum impedance is about 17^{o} from unity. Tuning this far from resonance in a linear amplifier with r-f feedback can be much more serious than in a class C amplifier because the phase angle of the feedback voltage is critical.

a. Phase Comparison Tuning

Use of a phase comparator circuit to compare the phase of the input signal to the phase of the output signal affords the most sensitive means of tuning a linear power amplifier stage. This circuit employs a phase discriminator, such as shown in Fig.13, for phase comparison.

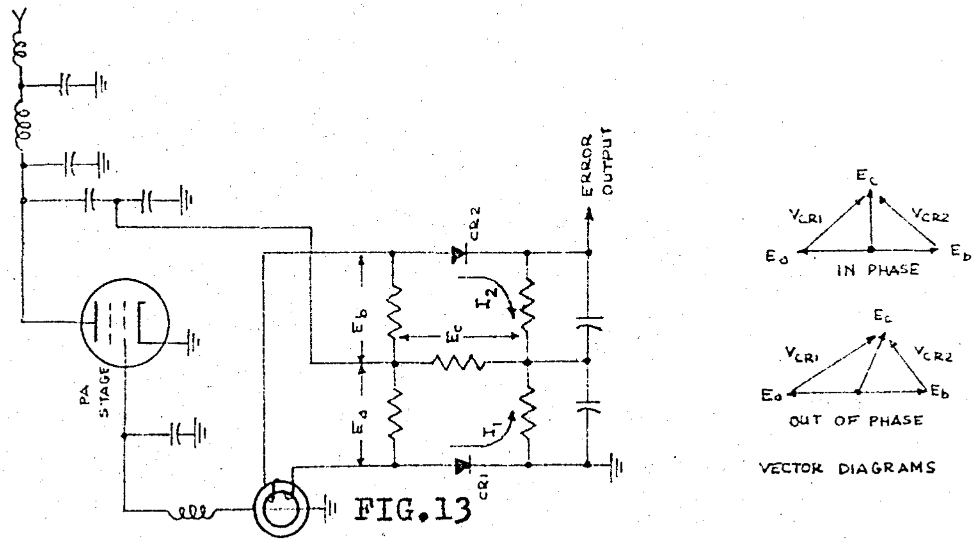


FIG.13

The phase discriminator can also be used to obtain an error signal for servo tuning the stage. However, for servo tuning, coarse positioning information is necessary because the phase discriminator responds to harmonic tuning points and because there is insufficient output from the phase discriminator over much of the frequency range. This coarse positioning information can be provided with a coarse follow-up potentiometer which receives information from the exciter frequency control circuits. Such a system requires that the master potentiometer track the tuning curves of the amplifier tank circuits and that sequencing controls be used to initiate and halt coarse positioning at the proper times. Pretuning information can also be derived from the exciter r-f output signal by using a coarse discriminator circuit. This circuit is a series RC network fed with r-f voltage from the exciter.

b. Loading Comparator Circuit

Since the voltage gain of a tube is dependent upon the load resistance, a loading comparator circuit, as shown in Fig.14, can be

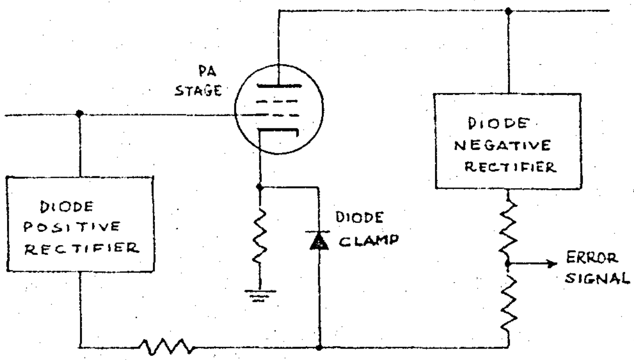


FIG.14

used to determine proper loading. The loading comparator is designed so that a predetermined ratio between positively rectified grid voltage and negatively rectified plate voltage produces zero error signal output. The power amplifier is then manually or automatically loaded until the error signal output goes to zero. The clamping diode is required so that the circuit will maintain control under light load when the amplifier is driven into plate saturation. In plate saturated operation, the rectified grid voltage will continue to rise with reduced loading while the rectified plate voltage remains relatively constant. This will cause the circuit to lose its sense of direction and result in reducing the load even further. To maintain the sense of direction under this condition, the clamping diode prevents the rectified grid voltage from exceeding a voltage which is proportional to plate current. Therefore in plate saturated operation, which is similar to class C operation, loading is determined by the ratio of plate current to r-f plate voltage. Proper compromise of the magnitude of the plate, grid, and clamping signal voltages results in a loading comparator that produces proper loading information regardless of the operating conditions, provided the plate circuit is held at resonance.

c. Antenna Tuning and Loading

The output network of a variable frequency transmitter must be capable of tuning and loading into a transmission line which presents different impedances at different frequencies. This requires output networks which will match a wide range of load impedances with the power amplifier output. In fixed-station equipment, the power amplifier usually works into a transmission line and antenna designed so that the load impedance presented to the amplifier varies over only a limited range. In this case the output network is designed to match the load impedance directly. In mobile and airborne equipment, the power amplifier usually works into a coaxial transmission line terminated with a wide variety of antennas that present unwieldy terminating impedances. In this case an antenna coupler is used which can be located in one of two positions: (1) It can be located near or in the transmitter to provide proper coupling between the transmitter output network and a transmission line which is terminated with a mismatched antenna; (2) It can be located near the antenna to terminate the transmission line properly and provide coupling for maximum power transfer to the antenna. The first method is commonly used in mobile transmitters, and the second method is used in airborne transmitters.

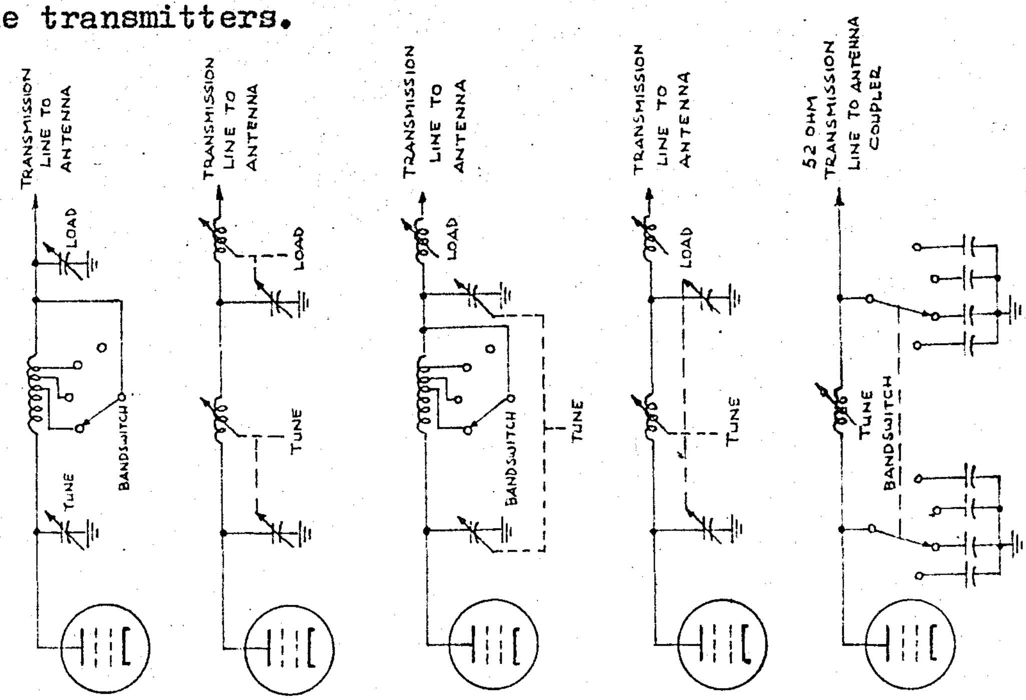


FIG.15

Two power amplifier control functions are required to match properly the load impedances presented to the power amplifier with the power amplifier network. One is a phasing control, or tuning control, which will balance out desirable reactance and make the load resistive or as nearly resistive as is possible. The other is a load control which will provide the proper terminating impedance. Fig.15 shows several ways that the output network components can be ganged to provide tuning and loading with two controls. The tuning control is adjusted to produce a plate current dip, which indicates maximum impedance. For more precise tuning and automatic tuning, the phase discriminator circuit is used. The loading control is adjusted to produce a pre-established value of grid voltage and plate current or, in some cases, a pre-established value of screen current and plate current. For more precise loading and automatic loading, the loading comparator circuit is used. The loading and tuning circuits must be so designed that the controls will not lose sense of direction under any circumstances. This is absolutely essential for automatic loading and tuning and is highly desirable for manual loading and tuning.

8.7. Power Supplies for Power Amplifiers

Fixed transmitters up to 1 kw usually use a single- phase a-c power source, and larger fixed transmitters usually use three-phase a-c power source. Mobile equipment may operate from a 6-volt to 28-volt d-c power source using dynamotors or vibrator power supplies to obtain the required high voltages. Airborne equipment usually uses the 400-cycle a-c power source of the aircraft.

In addition to supplying the required d-c voltage and output current, the power supply must have adequate d-c regulation, good dynamic regulation, and low ripple or noise output. Most high-voltage power amplifiers have a varying load characteristic so that good d-c regulation is essential. To reduce ripple and noise, high-voltage filters are used between the rectifier circuit and the power supply load. The filter chokes place a high impedance between the rectifier and the load, making large capacitors necessary in the output side of the filter. These output capacitors supply the rapid variations in load current which are impeded by the filter choke. This is particularly necessary in high-voltage power supplies for linear power amplifier stages.

Vacuum rectifiers can be used for small, low-voltage power supplies which have relatively constant load. Gas-type rectifiers are required where better regulation is necessary. The mercury-vapor rectifier is the most common gas-type rectifier used because it has long life when properly operated. Operating a mercury-vapor rectifier above or below its rated temperature, changes the vapor pressure in the tube and reduces its peak-inverse-voltage capability, making the rectifier more susceptible to arc-back. Equipment which is subject to wide ambient-temperature variations, such as military equipment, uses inert gas rectifiers such as the 3B28 and 4B32. These tubes can be operated in ambient temperatures from $-75^{\circ}C$ to $+90^{\circ}C$, which is frequently a necessary feature. The tube life of the inert gas rectifier, however, is only about one-third of the tube life of an equivalent mercury-vapor rectifier. Metallic rectifiers, such as selenium and copper oxide, are frequently used in power supplies delivering less than 100 volts for relay operation, etc.

Rectifier tube life is increased by operating the filaments 90° out of phase with the plate voltage. This minimizes the difference in voltage from each end of the filament to the plate and allows a more uniform emission over the entire filament. A 60° phase difference between the filament and the plate voltage is often used when it is more easily obtained because almost the full advantage of quadrature operation is realized. Tube ratings of some of the larger rectifier tubes are increased for quadrature operation.

Transient voltages and currents which far exceed the steady-state values occur in power supplies when the supply is energized. If these transient peaks exceed the peak-inverse-voltage rating of the tube, an arc-back may result. For this reason, rectifier tubes are often operated so that the normal peak-inverse-voltage does not exceed one-half of the rated peak-inverse voltage. If this is not possible, a step-start circuit is used which starts the transformer with resistors in series with the primary. After a short time delay, these resistors are shorted out. Some high-voltage rectifiers are started with a resistor in series with the filter capacitor, with the resistor being shorted out after a short time delay. This prevents a transient due to the charging current required to bring the voltage up on the filter capacitor. The added resistance in the circuit prevents excessive current in the rectifier.

8.8. Control Circuits

Power amplifier control circuits must perform three functions:

1. They must supply circuit control,
2. They must provide equipment protection,
3. They must provide personnel protection.

In small transmitters, the control circuits may consist of nothing more than an on-off switch to supply heater power and a push-to-talk button to apply plate voltage and put the transmitter on the air. In larger equipment, push buttons are usually used to initiate a certain sequence of relay operations which complete a function in the proper manner. Many transmitters, particularly those suitable for remote control, are capable of complete energization from a single push-button control.

The filament on-off switch, or push button, initiates a sequence of functions that applies power to the filaments, starts the cooling system, and energizes time delay circuits that make the power amplifier ready for the application of plate power. When operated to the off position, the power amplifier is shut down.

Filaments of high-power amplifier tubes are energized separately, and, in the case of mercury-vapor tubes, a time delay allows warmup time. The blower is started at the beginning of the starting sequence because the life and reliability of many components is greatly dependent upon operating temperature control. Air interlocks prevent the application of power to high-power tubes before cooling air is present and a blower-off delay maintains cooling air after shutdown. In various power amplifier stages, it is essential that bias voltage be applied before plate or screen voltage is applied. This requires sequencing the application of bias voltage and the plate voltage as well as interlocks between the two so that the loss of bias voltage will result in removing the plate voltage.

Power amplifier control circuits are sequenced and interlocked so that everything else must be on and functioning before the high-voltage plate transformer is energized. Certain power tetrodes require that screen voltage be applied simultaneously with plate voltage to prevent excessive screen dissipation. To prevent high-current and high voltage transients, plate voltage is often applied through step-start circuits which place resistors for a short time in the power supply circuit.

Medium-power and high-power tubes are nearly always protected from excessive plate current by overload relays. These relays remove the high-voltage primary power if the plate current exceeds a preset value. Many overloads that occur during normal operation will clear themselves when the high voltage is removed. For this reason, large power amplifiers are usually provided with an overload recycle circuit. This circuit brings the power amplifier back on after an overload. If the overload reoccurs, the power amplifier will again shut down. The number of recycles before shutdown can generally be preset with a recycle counting switch.

8.9. Tube Operating Conditions for R-F Linear Power Amplifiers

SSB amplifiers provide linear amplification and operating conditions similar to those of audio amplifiers. There is one fundamental difference, however, between audio and r-f linear amplifiers. This is that the input and output voltages of a tuned r-f amplifier are always sine waves because the tuned circuits, if they have adequate Q, make them so. Therefore, the distortion in an r-f amplifier results in distortion of the SSB modulation envelope and not in the shape of the r-f sine wave. This can be restated that distortion in an r-f linear amplifier causes a change in gain of the amplifier when the signal level is varied. The greatest difference between an audio amplifier and an r-f linear amplifier is in the grid driving power requirements when driving into grid current. In the audio amplifier, the driver must supply all of the instantaneous power required by the grid at the peak of the grid swing. To deliver this peak power the audio driver must be capable of delivering average sine wave power equal to one-half of the peak power. In an r-f linear amplifier, the tank circuit averages the power of the r-f cycle due to its "flywheel" effect so that the driver need only be capable of delivering the actual average power required, and not the peak. With these reservations in mind, examination of the audio or modulator data of a tube will give a good idea of its r-f linear power amplifier operating conditions.

a. Class A R-F Linear Amplifiers

In low-level amplifiers, where the output signal voltage is less than 10 volts, small receiving type tubes, such as the 6AU6, are very satisfactory for class A service. For voltage levels above 10 volts, the 4X150A is the best choice for class A operation because it has short leads, low plate-to-grid capacitance, and high transconductance. Class A amplifier tubes should be operated in as linear a portion of the plate characteristic curves as is practical. This can be done by inspecting the plate characteristic curves of the tube. Usually the static plate current which results in near maximum plate dissipation is the best. The maximum output voltage should be kept to about one-tenth the d-c plate voltage or less to obtain signal-to-distortion ratios of 50 db or better. The d-c plate voltage regulation for class A operation is seldom of importance and cathode bias and screen dropping resistors are commonly used. Even with tubes such as the 6AU6 and the 4X150A which have short leads and low grid-to-plate capacitance, it is desirable to load the input and output circuits to 5000 ohms when operating up to 30 mc.

b. Class AB R-F Linear Amplifiers

In the power range from 2 watts to 500 watts, class AB_1 is normally used. This class of operation is very desirable because distortion due to grid current loading is avoided and because high power gain can be obtained. At present, tubes are not available which will give low distortion with good plate efficiency operating class AB_1 at power levels above 500 watts. Therefore, for higher power levels class AB_2 operation is used.

For class AB operating conditions with a given screen voltage and given plate load, there is one value of static plate current which will give minimum distortion. The optimum value of static plate current for minimum distortion is determined by the sharpness of cutoff of the plate current characteristic. Grid bias is then set to produce the optimum static plate current. This optimum point is determined from the load line on a set of constant current plate current curves. Values obtained from this curve are then plotted to obtain the plate current vs grid voltage curve shown in Fig.16. This curve is the dynamic characteristic of the tube. By projecting the most linear portion

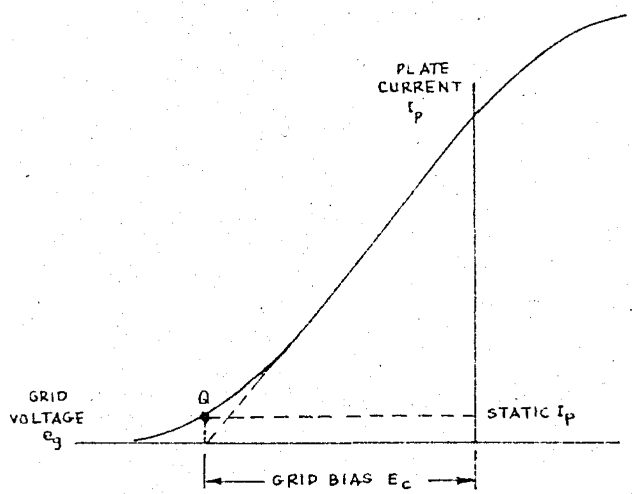


FIG.16

of the curve to intersect with the zero plate current line, the grid bias is determined. This point of intersection is often referred to as the projected cutoff. The static plate current which will flow with this grid bias is the proper static plate current for minimum distortion. This procedure is used in audio amplifier design and is nearly correct for r-f linear amplifier design. Perhaps a more accurate procedure for determining the proper bias for r-f amplifiers is to choose the point Q so that the slope of the curve at Q is one-half the slope of the major linear portion. This will allow the amplifier to operate class A with small signals and deliver power over both halves of the cycle. With a large signal, the tube delivers power over essentially one-half the cycle. Then the change in plate current relative to plate voltage swing over half the cycle will be half as much for small signals as it is for large signals and linear operation is obtained at all signal levels.

The screen voltage of a tetrode tube has a very pronounced effect on the optimum static plate current, because the plate current of a tube varies approximately as the three-halves power of the screen voltage. For example, raising the screen voltage from 300 to 500 volts will double the plate current. The shape of the dynamic characteristic will stay nearly the same, however, so that the optimum static plate current for minimum distortion is also doubled. A practical limit is reached because high static plate current causes excessive static plate dissipation.

In practice, it is found that the static plate current determined by the above method is so high that plate dissipation is near or beyond the maximum rating of the tube when using desired d-c plate voltage. For this reason, it is often necessary to operate the tube below the optimum static plate current, which can be done without causing appreciable distortion. In tetrodes, the optimum static plate current is a function of screen voltage, and the high screen voltages required for class AB_1 operation usually require an excessive amount of plate current for minimum distortion. A choice must then be made between operating the tube at lower than optimum static plate current or using a lower screen voltage and driving the tube into the grid current region, a second principal cause of distortion.

c. Estimating Tube Operating Conditions

The operating conditions of a tube operating class AB in an r-f linear power amplifier can be estimated from the load line on a set of constant plate current curves for the tube, as shown in Fig.17. From the end point of the load line, the instantaneous peak plate current, i_p, and the peak plate voltage swing, e_p, can be established. From these two values, the principal plate characteristics can be estimated by using the following relationships for a single-frequency test signal:

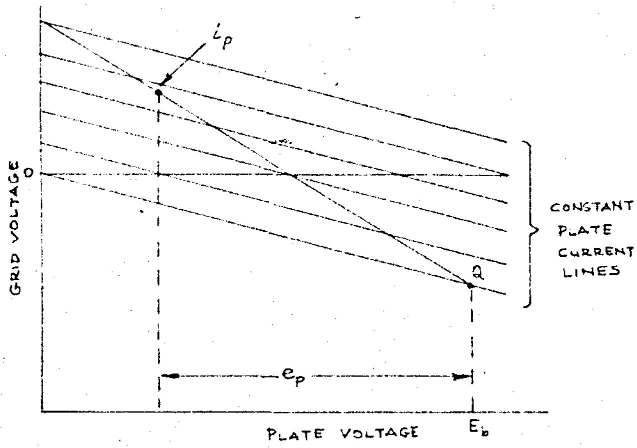


FIG.17

d-c plate current, $I_B = i_p/\pi$

plate input watts, $P_{in} = i_p E_B/\pi$

average output watts and PEP, $P_o = i_p e_p/4$

plate efficiency, $Eff = \pi e_p/4E_B$

For a standard two-frequency test signal the relationships are:

d-c plate current, $I_B = 2i_p/\pi^2$

plate input watts, $P_{in} = 2i_p E_B/\pi^2$

average output watts, $P_o = i_p e_p/8$

PEP watts, $P_o = i_p e_p/4$

plate efficiency, $Eff = (\pi/4)^2 e_p/E_B$

An actual tube with moderate static plate dissipation will have operating characteristics similar to those shown in Fig.18 for the single tone and two-tone signals. Plate dissipation and efficiency at maximum signal level are effected little by even rather high values of static plate dissipation. In practice, the peak plate swing is limited to something less than the d-c plate voltage in order to avoid excessive grid drive, excessive screen current, or operation in the nonlinear plate current region. Most tubes operate with an efficiency in the region of 55 to 70 per cent at peak signal level.

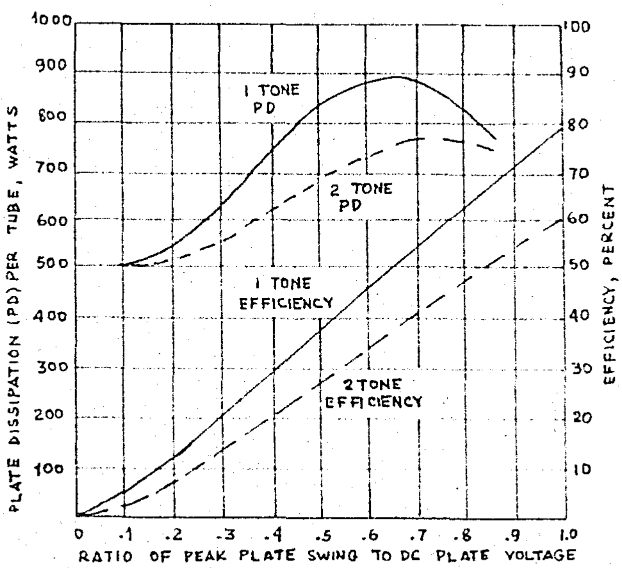


FIG.18

8.10. Distortion

a. Causes of R-F Linear Power Amplifier Distortion

The principal causes of distortion are nonlinearities of the amplifier tube plate current characteristic and grid current loading. In order to confine distortion generation to the last stage or two in a linear power amplifier, all previous stages are operated class A.

The generation of distortion products due to the nonlinear characteristics of the amplifier tubes can be derived from the transfer characteristic of the tubes, also called the dynamic characteristic, as shown in Fig.19. The shape of this curve and the choice of the zero signal operating point Q, determine the distortion which will be produced by the tube. A power series expressing this curve, written around the zero signal operating point, contains the coefficients of each order of curvature, as shown in the following expression:

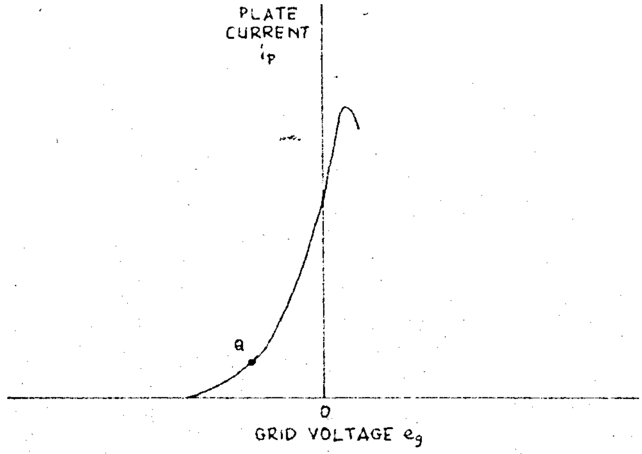


FIG.19

$$i_p = k_0 + k_1 e_g + k_2 e_g^2 + k_3 e_g^3 + k_4 e_g^4 + k_5 e_g^5 + \dots + k_n e_g^n$$

In this expression, i_p represents instantaneous plate current, k_1, k_2, etc., the coefficients of their respective terms, and e_g the input grid voltage signal. The values for the coefficients are different for every power series written around different zero signal operating points. By making the input signal, e_g, consist of two equal amplitude frequencies with a small frequency separation, the distortion products of concern in linear amplifiers can be obtained. Fig.20 shows the spectrum distribution of the stronger plate current components. It is seen that tuned circuits can filter out all products except those which are near the fundamental input frequencies. This removes all of the even-order intermodulation products and the harmonic products. The odd-order intermodulation products fall close to the original frequencies and cannot be removed by selective circuits. Fig.21 shows these odd-

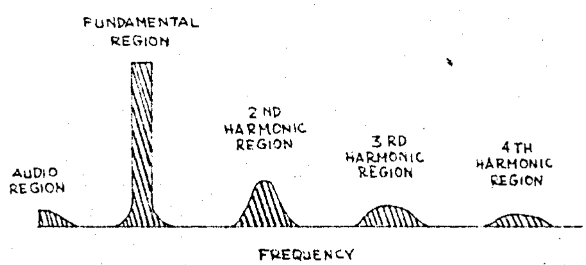


FIG.20

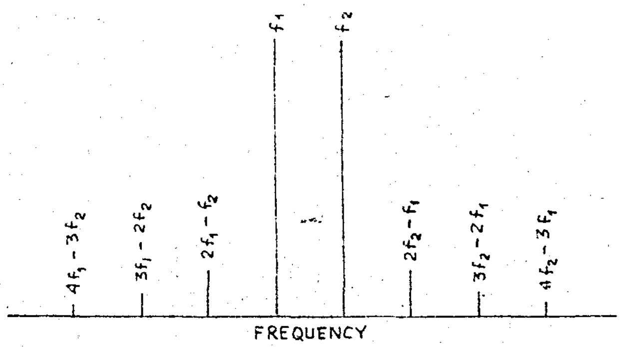


FIG.21

order product which fall within the pass-band of selective circuits. The inside pair of intermodulation distortion products are third-order, the next fifth-order, seventh-order, etc. The first and most important means of reducing distortion, then, is to choose a tube with a good plate characteristic and choose the operating condition for low odd-order curvature.

The nonlinearity caused by grid current loading is a function of the regulation of the grid driving source. In general, this regulation with varying load is poor in linear amplifiers. It is common practice to use swamping resistors in parallel with a varying grid load so that the resistance absorbs about ten times the power that the grid of the tube requires. This provides a low, constant driving source impedance and improves linearity at the expense of increased driving power.

The instantaneous plate current of all tubes drops off and causes distortion when the instantaneous plate voltage is low. The main reason for this drop is that current taken by the grid and screen is robbed from the plate. In all but a few transmitting tetrodes, the plate can swing well below the screen voltage before plate saturation occurs. However, when the plate swings into this region, the instantaneous plate current drops considerably. If distortion requirements are not too high, the increased plate efficiency realized by using plate swings can be realized. However, to minimize distortion, the allowable plate swing may have to be reduced.

b. Distortion Reduction

There is a need for reduced levels of intermodulation distortion from r-f linear power amplifiers used in SSB systems. This need exists not because the distortion noticeably reduces the intelligibility of the SSB signal, but because distortion products outside of the channel width necessary for transmission of intelligence interferes with adjacent channel transmission. The distortion of some of the early SSB power amplifiers was so great that voice channels were placed a full channel width apart to avoid adjacent channel interference. Recent power amplifier developments permit adjacent channel operation, using power amplifiers with signal-to-distortion ratios of from 35 db to 40 db. However, power amplifiers with signal-to-distortion ratios of from 45 db to 50 db would further increase the utility of single sideband.

There are two basic means of reducing distortion to levels better than is obtainable from available tubes. These are r-f feedback and envelope distortion canceling. An r-f feedback is very effective and quite easy to obtain. Ten decibels of r-f feedback will produce nearly 10 db of distortion reduction which is realized at all signal levels. Envelope distortion canceling has an inherent weakness because it depends upon envelope detection for its feedback signal. This means that distortion canceling must be instantaneous to be perfect. Since some delay is inherent in the envelope detector and feedback loop, the effectiveness of this circuit depends upon how short the time delay can be made. Development work indicates that a combination of r-f feedback and envelope distortion canceling will provide more distortion canceling than either method separately. Using 10 db of r-f feedback around all three stages of a 20-kw PEP power amplifier, and a signal synthesized from the input envelope to grid modulate the first stage, a better than 50 db signal-to-distortion ratio has been obtained for all distortion products at any signal level up to 20-kw PEP.

CHAPTER 9

9.1. 208U-10 Linear Power Amplifier

The Purpose of Equipment. The 208U-10 Linear Power Amplifier is an automatically tuned r-f amplifier intended for communications in fixed or transportable radio stations. The function of a linear power amplifier is to raise the power level of an r-f input signal without changing any other characteristic of the input signal. Not more than 0.2 watt r-f input power from an external exciter is required to drive the 208U-10 Linear Power Amplifier to a rated power output of 10 kilowatts peak envelope power (PEP) or continuous average power. The linear power amplifier will amplify any type of r-f input signals within the 2.0-to 30.0-megacycle range which do not exceed the bandwidth and power-handling capacities of the equipment. The 208U-10 Linear Power Amplifier can be locally or remotely controlled for continuous duty, unattended operation.

TABLE I. EQUIPMENT REQUIRED

Quantity	Item	Required Characteristics
1	r-f exciter	Capable of delivering 0.2 watt r-f output into 50 ohms from 2.0 to 30.0mc.
1	h-f antenna	Capable of radiating 10 kw r-f power at the desired frequencies within the 2.0- to 30.0-mc frequency range.
3	Power conductors	Rated at 250 volts, 100 amp for 220-volt operation.
3	Coaxial cables	Type RG-58C/U, for r-f input, alc and tgc outputs. Length to be determined by particular installation.
1	Coaxial cable	50-ohm, 1-5/8-inch, coaxial fittings and transmission line to system antenna.

TABLE 2. TEST EQUIPMENT REQUIRED

Item	Type
Signal Generator	Hewlett-Packard 606A or equivalent
Vacuum-Tube Voltmeter	Hewlett-Packard 410B or equivalent
Volt-Ohm-Milliammeter	Triplett Model 630 or equivalent
Capacitive Voltage Divider	Hewlett-Packard 453A for HP-410B
Oscilloscope	Tektronix 533 with type CA plug-in unit
R-F Dummy Load	Bird Model 502 Electro-Impulse or equivalent
Probe Coaxial T Connector	Hewlett-Packard 455A
Probe coaxial N Connector	Hewlett-Packard 458A
Electronic Voltmeter	Ballantine 300G or equivalent a-c voltmeter with 1-percent accuracy.

Electrical Description. The 208U-10 Linear Power Amplifier is an automatically tuned linear power amplifier which is driven by an external exciter. Not more than 0.2 watt r-f input signal is required to drive the linear power amplifier to rated 10.0 kilowatts r-f output power. The linear power amplifier is capable of accepting 2 watts average power input without damage. The equipment will deliver 10 kilowatts r-f output power into a 50-ohm load. The r-f amplifier stages consist of two-level, wideband r-f amplifiers; a plate-tuned driver amplifier; and the power amplifier which operates into a tuned pi-L output network. The linear power amplifier r-f circuits are tuned to r-f input signals within the 2.0- to 30.0-mc range during an automatic tune cycle. The automatic tune cycle program provides for coarse positioning the r-f tuned circuits, conditional fine tuning of optional external output equipment, and final fine tuning or the r-f circuits. The gain of the linear power amplifier is controlled during the tune cycle by internal gain control circuits. The operating point is carefully controlled to provide high efficiency and low intermodulation distortion. A broadband capacitance bridge circuit maintains accurate PA grid neutralization over the entire frequency range. The linear power amplifier develops automatic load control (alc) and transmitter gain control (tgc) voltages which can be applied to the external system exciter to control the system gain.

The linear power amplifier is capable of operation on 195- to 255 volts a-c, 3-phase, 47- to 63-cps primary power. All operating voltages are provided by power supplies within the power supply unit. The power supplies provide +6250 volts d-c for the power amplifier tube plate, -1250 volts d-c for the power amplifier tube cathode and for the broadband and driver amplifier tube cathodes, +150 volts d-c B+ for the igc amplifier and for signal control bias, 225 volts a-c (adjusted) tube filament voltage, 115 volts a-c servo-motor reference voltage, +28 volts d-c for servo-amplifier B+ and relay control, and 600 volts d-c keying bias.

All power supplies are protected by fuses and a main circuit breaker. Unit and compartment door interlocks are provided for safety and several overload protective circuits are provided to protect the equipment against malfunction. The equipment can be locally controlled from the cabinet meter and control panel unit 4, which is provided as part of several application groups. System connections can be made at the top of the cabinet for additional remote-controlled operation. Controls are provided on the front panels of the units for local test operation.

TABLE 3. ELECTRICAL CHARACTERISTICS

Characteristic	Description
r-f output power	10 kilowatts peak envelope power or 10 kilowatts continuous average power.
r-f input power	0.2 watt PEP or average, 2.0 watts maximum (during tune cycle only)
Frequency range	2 to 30 megacycles
Tuning	Automatically tuned to resonance by servomechanical means during an automatic tune cycle. Tune cycle manually positioned for local test purposes. Tuning time is not more than 25 seconds with 60-cps primary power input.

TABLE 3. (CONTINUED)

Characteristic	Description
Type of emission	SSB, AM, CW, or any other type within the bandwidth and power capabilities.
r-f bandwidth	12 kc with not more than 0.2 db variation from 4 to 30 mc, and not more than 0.1 db from 2 to 4 mc.
Output load impedance	50 ohms: maximum SWR of 2 to 1 from 2.0 to 2.5 mc. maximum SWR of 3 to 1 from 2.5 to 30.0 mc.
Input impedance	50 ohms: maximum SWR of 1.5 to 1 from 2.0 to 30.0 mc.
Intermodulation distortion	Third and higher order products not less than 35 db below one tone of a two-tone signal-producing rated power.
Harmonic output	Second harmonic not less than 55 db below fundamental. Higher order harmonics not less than 60 db below fundamental.
Input power requirements	195 to 225 volts line-to-line, 3-phase, 47 to 63 cps. Taps for line voltage compensation are provided.
Power consumption	
Condition (nominal at 8.0 mc)	
Filaments on	2.5 kw (nominal), 0.95 power factor
Plate on	3.3 kw (nominal), 0.96 power factor
Keyed (static plate currents)	10.0 kw (nominal), 0.98 power factor
Two-tone test at 10 kw PEP	20.0 kw (nominal), 0.98 power factor
Single tone CW at 10 kw	22.0 kw (nominal), 0.98 power factor

TABLE 4. TUBES

Designation	Type	Function
1A8V1	6S4A	PA keyer
2V1	7788	Igc amplifier
2V2	4CX350A	Broadband amplifier
2V3	4CX350A	Driver amplifier
2V4	4CX350A	Driver amplifier

TABLE 4. (CONTINUED)

Designation	Type	Function
2V5	4CX10,000D	Power amplifier
2V6	12AL5	Alc/sidetone detector

Fig.1 shows the overall appearance of 208U-10 Linear Power Amplifier.

Fig.2 shows the same Linear Power Amplifier with the doors open.

Fig.3 shows the same Linear Power Amplifier with the compartment doors open.

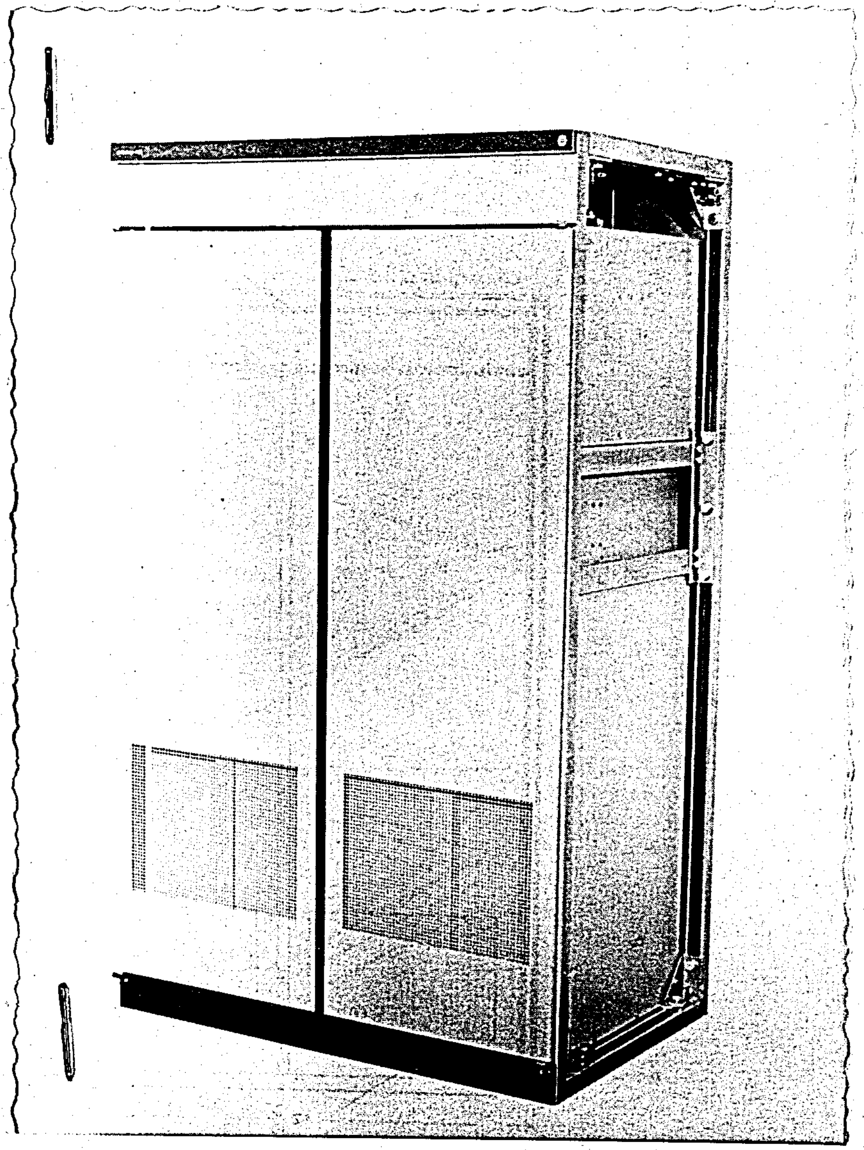


FIG.1. Overall appearance of 208U-10
Linear Power Amplifier

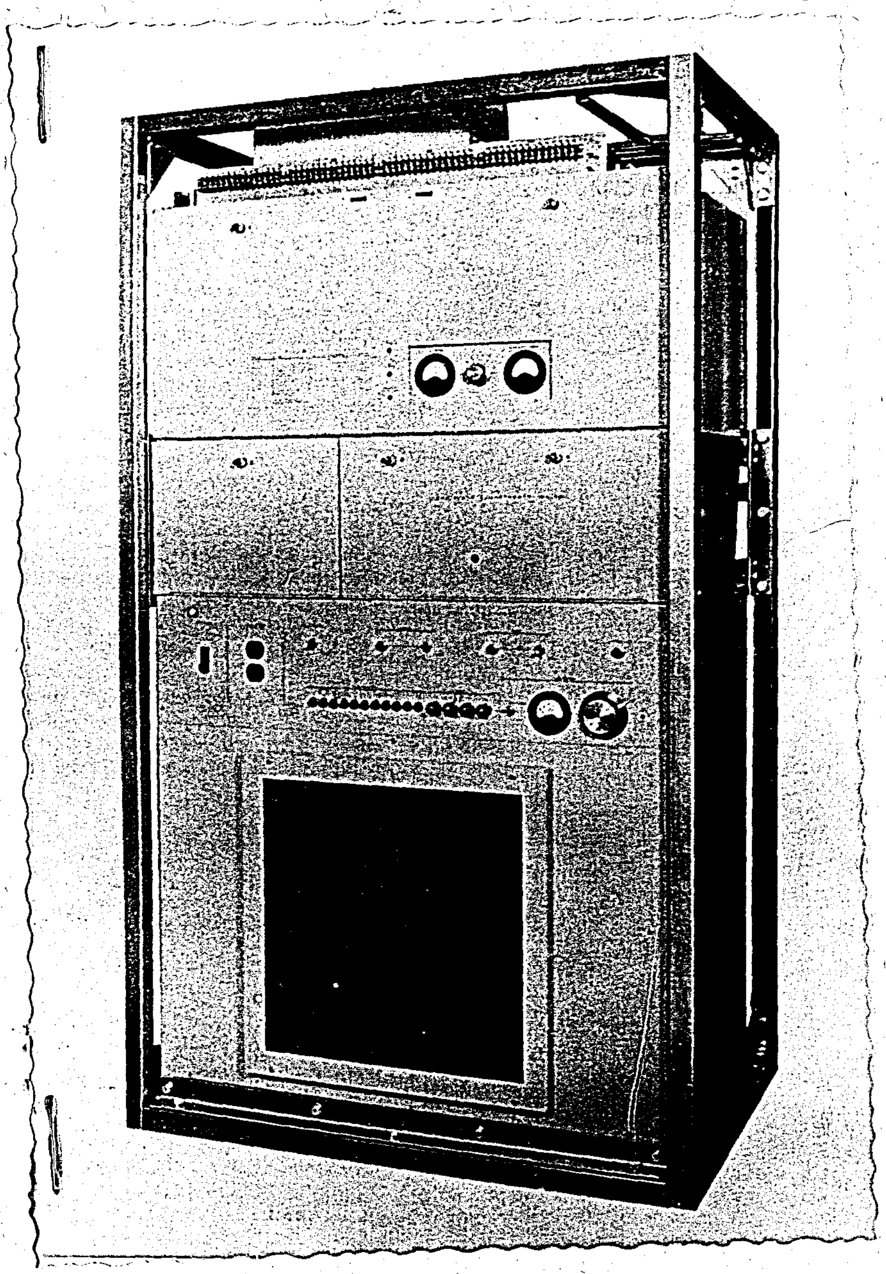


FIG.2. 208U-10 Linear Power Amplifier
with the doors open

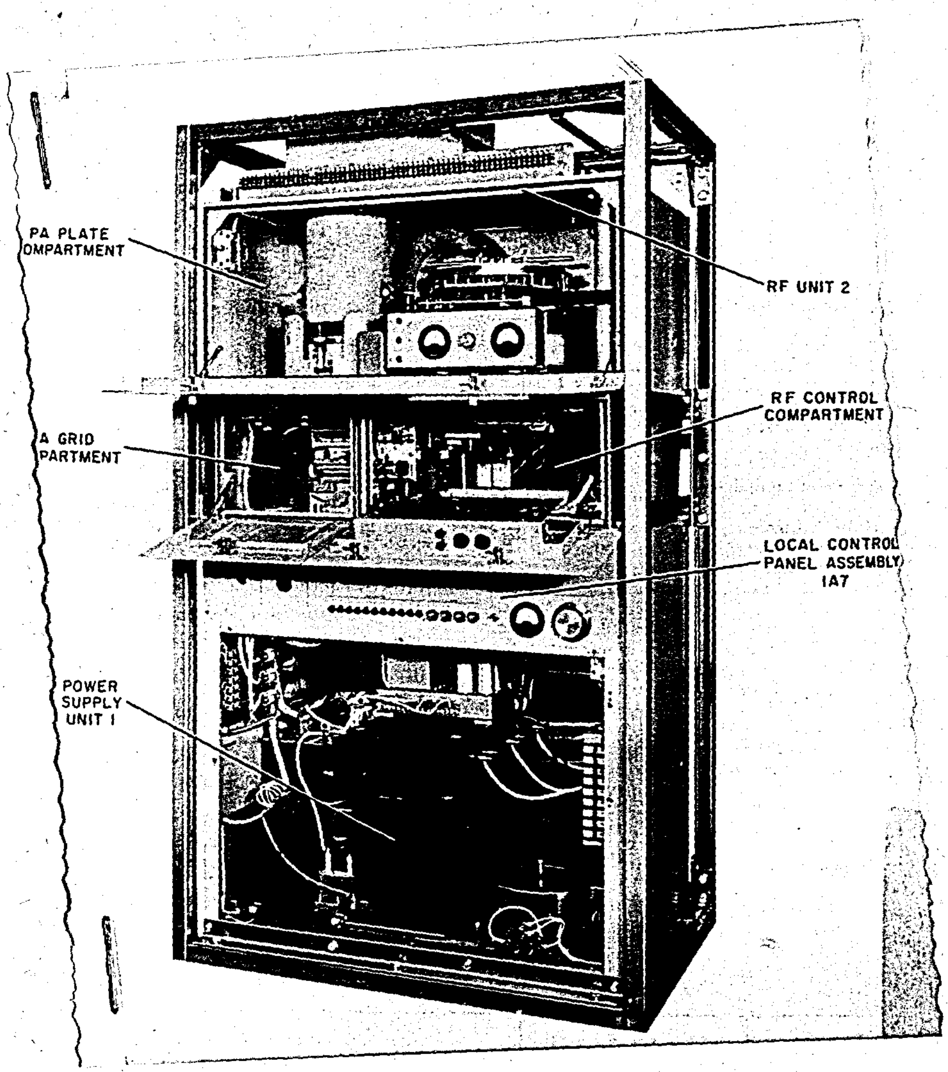


FIG.3. 208U-10 Linear Power Amplifier
with the compartment doors open.

9.2. 30S-1 Linear Amplifier

The 30S-1 Linear Amplifier (Fig.5) consists of a one-stage linear amplifier and the necessary power supplies. It is capable of maximum legal input power in the amateur bands between 3.5 and 29.7 mc. It operates either CW or SSB service with any exciter (such as KWM-1, KWM-2, or 32S-1) capable of 80 watts PEP output. In addition, the amplifier may be operated outside the amateur bands at any frequency between 3.4 and 30 mc by retuning the input circuits. Fig.4 is the block diagram of 30S-1 and Fig.6 is the schematic diagram.

r-f Circuit Description. The 30S-1 uses a 4CX1000A tetrode as the r-f amplifier. The cathode of the amplifier is driven, requiring some 80 watts PEP for proper operation. The screen grid is grounded directly to the chassis for better grid-plate isolation. The biased control grid is bypassed so that it is at r-f ground. A 12AL5 automatic load control rectifier monitors the grid circuit and applies alc to the exciter the moment the r-f amplifier draws any grid current, thus maintaining class AB_1 operation. The minute amount of grid current drawn before alc acts does not degrade linearity. Rather these few microvolts actually improve the linearity curve; only after appreciably grid current is drawn is linearity affected. A plate overload relay, grid overload relay, and thermal overload relay are included for protection of the 4CX1000A. r-f output is coupled through a pi network into a 50-ohm load (with a maximum permissible SWR 2:1).

Power Supply. Since the power amplifier screen grid is at d-c ground potential, it is necessary to provide the cathode with negative 200 volts in order to supply screen voltage which is 200 volts higher than the cathode potential. Effective plate-cathode voltage is the sum of the screen-plate supply (2800 volts) and the cathode-screen supply (200 volts). Control grid bias is referenced to the cathode. All plate and screen current passes through the 200-volt supply, and only plate current through the 2800-volt supply. All relays are operated from d-c sources except the time-delay relay

K202 and thermal overload relay K102. Switching from SSB to CW operation automatically lowers the plate voltage and changes grid bias. The amplifier operates with approximately 3000 volts plate-to-cathode in SSB service and approximately 2400 volts plate-to-cathode in CW service. The power supplies may be connected to either 115-volt lines or 230-volt, 3-wire service lines. The 230-volt, 3-wire connection is preferred.

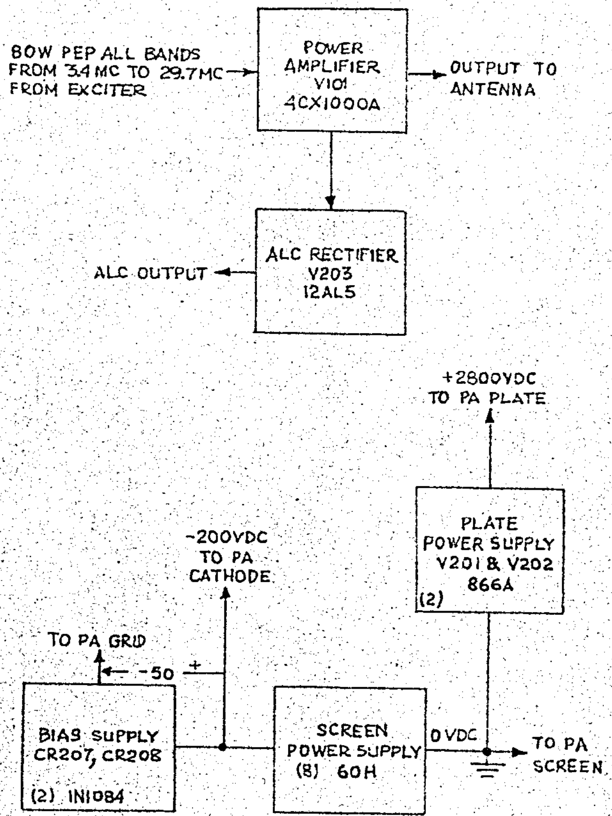


FIG.4.

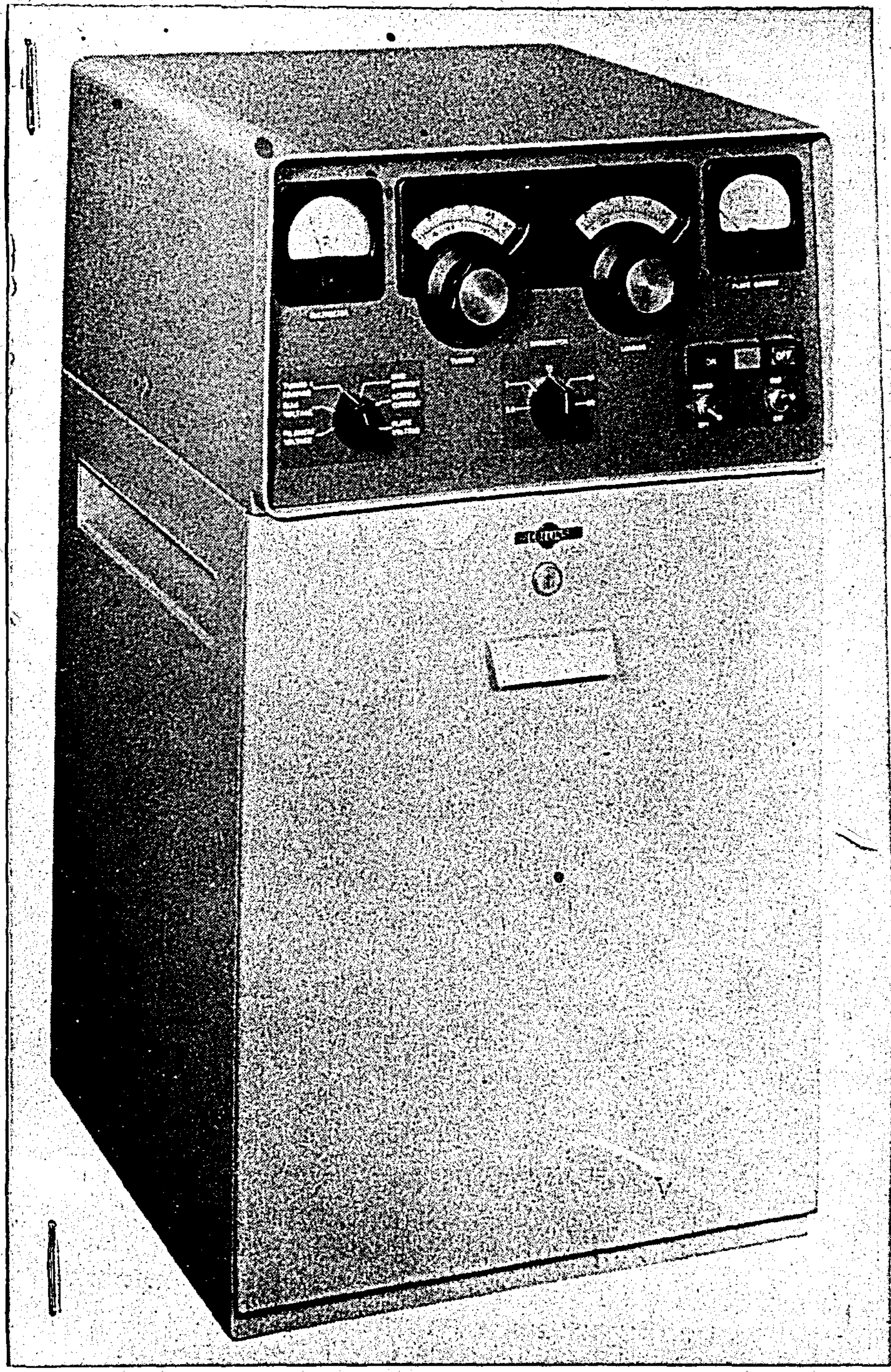
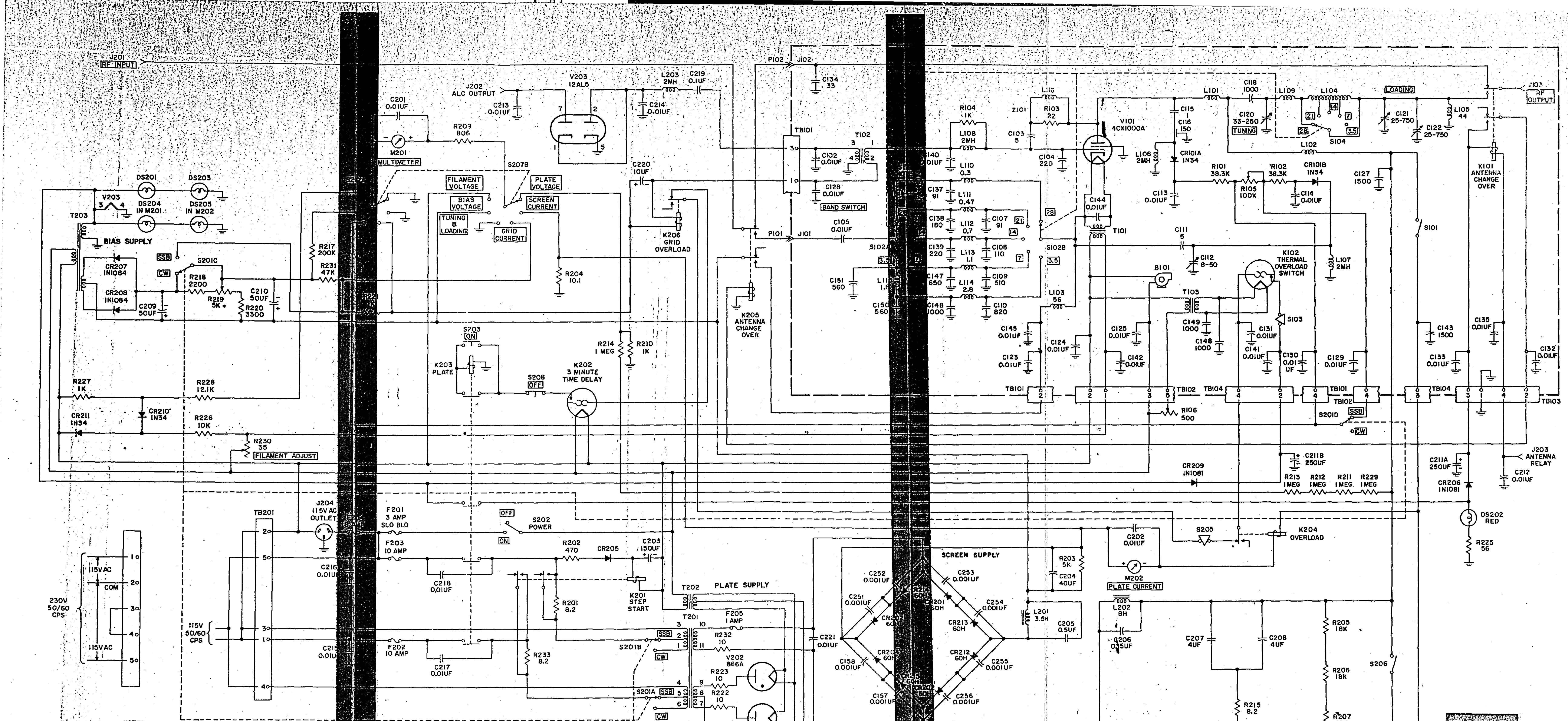


FIG.5. 30S-1 Linear Amplifier



NOTES:
1. UNLESS OTHERWISE INDICATED, ALL RESISTANCE VALUES ARE IN OHMS, ALL CAPACITANCE VALUES ARE IN MICROMICROFARADS, ALL INDUCTANCE VALUES ARE IN MICROHENRYS.
2. LENGTH OF THIS COAXIAL LEAD IS CRITICAL. IT IS PART OF IMPEDANCE MATCH.

FIG. 6.1

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