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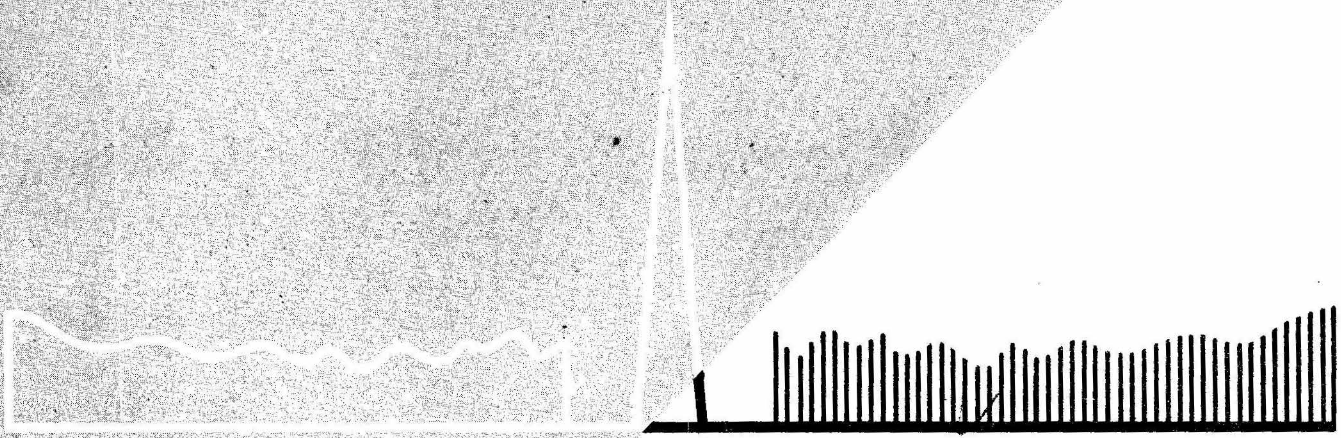
SINGLE SIDEBAND
COMMUNICATIONS



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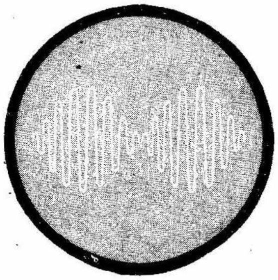

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SINGLE SIDEBAND COMMUNICATIONS

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FOREWORD

The single-sideband method of transmission has been in use for quite some time in fixed ground point-to-point communications. However, since there is much discussion concerning the conservation of the frequency spectrum, and since modern techniques in circuit design and construction now make it possible to extend the use of SSB, this new method of transmission is being considered for standard radio broadcast, mobile vehicular, airborne, and transhorizon-scatter system applications.

The information presented in this manual is not intended to replace the equipment manufacturer's information, but merely to present the general concepts of single-sideband transmitting and receiving techniques. The circuits illustrated in the diagrams can be considered as typical of single-sideband systems; they are not intended to illustrate actual working models of such equipment.

It is important that the reader understand the fundamentals of amplitude modulation before attempting to grasp the techniques of single sideband. For this reason a brief review of amplitude modulation is presented in the INTRODUCTION to this manual.

SAFETY PRECAUTIONS

IT IS THE DUTY of all personnel engaed in the INSTALLATION, OPERATION, and MAINTENANCE of electronic equipment, or engaged in the TRAINING OF OTHER PERSONNEL on electronic equipment, to become familiar with the following SAFETY PRECAUTIONS:

1. DO NOT RELY on safety devices.
2. USE RUBBER GLOVES when applicable.
3. KEEP YOUR FEET CLEAR of objects on the floor.
4. STAND ON A GOOD RUBBER MAT.
5. Whenever possible, KEEP ONE HAND BEHIND YOU, or in your pocket.
6. HAVE ANOTHER PERSON, qualified in FIRST AID FOR ELECTRICAL SHOCK, present at all times.

REMEMBER that men are always injured or killed by HIGH-VOLTAGE EQUIPMENT which is ASSUMED TO BE OFF. TAKE NOTHING FOR GRANTED. Make certain that the POWER IS OFF by securing the POWER-LINE SWITCH in its OFF position. Remove all fuses (including spares) from any circuits where switches might unintentionally be closed.

RESCUE OF SHOCK VICTIMS

1. PROTECT YOURSELF with DRY insulating material.
2. BREAK THE CIRCUIT by opening the power switch or pulling the victim free of the live conductor.

DO NOT TOUCH THE VICTIM WITH YOUR BARE HANDS UNTIL THE CIRCUIT IS BROKEN.

FIRST AID

Do These Three Things First in Any Emergency Requiring First Aid

1. **SEND FOR A DOCTOR OR CARRY THE VICTIM TO A DOCTOR.**
2. **KEEP VICTIM WARM, QUIET, AND FLAT ON HIS BACK.**
3. **IF BREATHING HAS STOPPED, APPLY ARTIFICIAL RESPIRATION. STOP ALL SERIOUS BLEEDING.**

When, from any cause whatever, BREATHING HAS STOPPED, APPLY ARTIFICIAL RESPIRATION IMMEDIATELY and continue WITHOUT STOPPING until normal breathing returns, or a doctor pronounces the victim dead. SPEED IN BEGINNING ARTIFICIAL RESPIRATION IS ESSENTIAL.

ARTIFICIAL RESPIRATION

POSITION OF THE SUBJECT

1. Place the subject in the face down, prone position. Bend his elbows and place the hands one upon the other. Turn his face to one side, placing the cheek upon his hands.

POSITION OF THE OPERATOR

2. Kneel on either the right or left knee at the head of the subject facing him. Place the knee at the side of the subject's head close to the forearm. Place the opposite foot near the elbow. If it is more com-

fortable, kneel on both knees, one on either side of the subject's head. Place your hands upon the flat of the subject's back in such a way that the heels lie just below a line running between the armpits. With the tips of the thumbs just touching, spread the fingers downward and outward.

COMPRESSION PHASE

3. Rock forward until the arms are approximately vertical and allow the weight of the upper part of your body to exert slow, steady, even pressure downward upon the hands. This forces air out of the lungs. Your elbows should be kept straight and the pressure exerted almost directly downward on the back.

POSITION FOR EXPANSION PHASE

4. Release the pressure, avoiding a final thrust, and commence to rock slowly backward. Place your hands upon the subject's arms just above his elbows.

EXPANSION PHASE

5. Draw his arms upward and toward you. Apply just enough lift to feel resistance and tension at the subject's shoulders. Do not bend your elbows, and as you rock backward the subject's arms will be drawn toward you. Then lower the arms to the ground. This completes the full cycle. The arm lift expands the chest by pulling on the chest muscles, arching the back, and relieving the weight on the chest.

The cycle should be repeated 12 times per minute at a steady, uniform rate. The compression and expansion phases should occupy about equal time; the release periods being of minimum duration.

ADDITIONAL RELATED DIRECTIONS

It is all important that artificial respiration, when needed, be started quickly. There should be a slight inclination of the body in such a way that fluid drains better from the respiratory passage. The head of the subject should be extended, not flexed forward, and the chin should not sag lest obstruction of the respiratory passages occur. A check should be made to ascertain that the tongue or foreign objects are not obstructing the passages. These aspects can be cared for when placing the subject into position or shortly thereafter, between cycles. A smooth rhythm in performing artificial respiration is desirable, but split-second timing is not essential. Shock should receive adequate attention, and the subject should remain recumbent after resuscitation until seen by a physician or until recovery seems assured.[*]

OTHER INSTRUCTIONS

1. If a dry blanket is available, slide it under the subject without interrupting the respiration cycles. This may easily be accomplished with the aid of an assistant. Cover the subject loosely by wrapping the ends of the blanket around him.

2. Between cycles, loosen all tight clothing, such as belts or collars.

3. Continue to apply artificial respiration until the subject begins to breathe regularly. Do not give up hope! Sometimes as much as eight hours of continuous artificial respiration is necessary to restore regular breathing. Remember, only a doctor is qualified to pronounce death under circumstances requiring artificial respiration.

4. If the subject should again stop breathing, resume the application of artificial respiration at once.

5. Always wait until consciousness returns before administering liquid stimulants. If liquids are forced down the throat of an unconscious person, he may drown.

6. Recommended stimulants are hot, black coffee; hot, strong tea; aromatic spirits of ammonia, one teaspoonful to a glass of water. Alcoholic drinks are not recommended unless absolutely nothing else is available.

7. Give only one stimulant. It should be sipped slowly.

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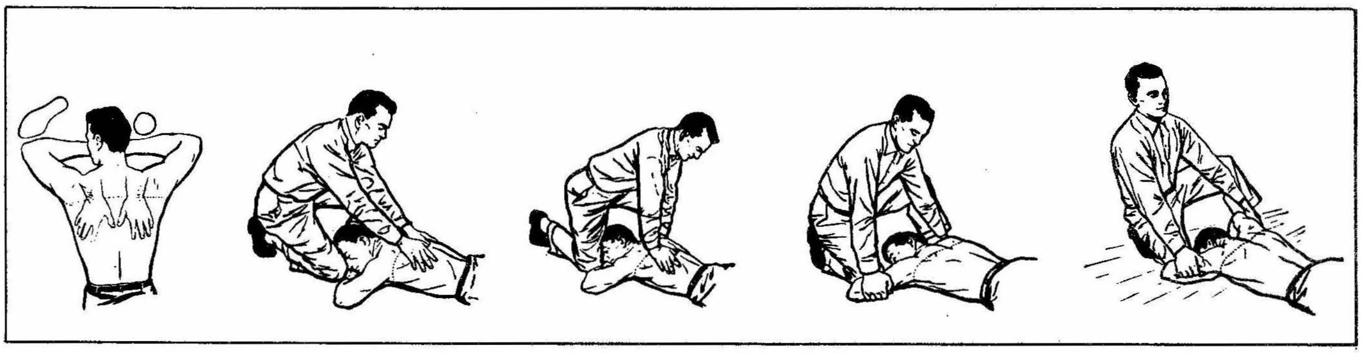

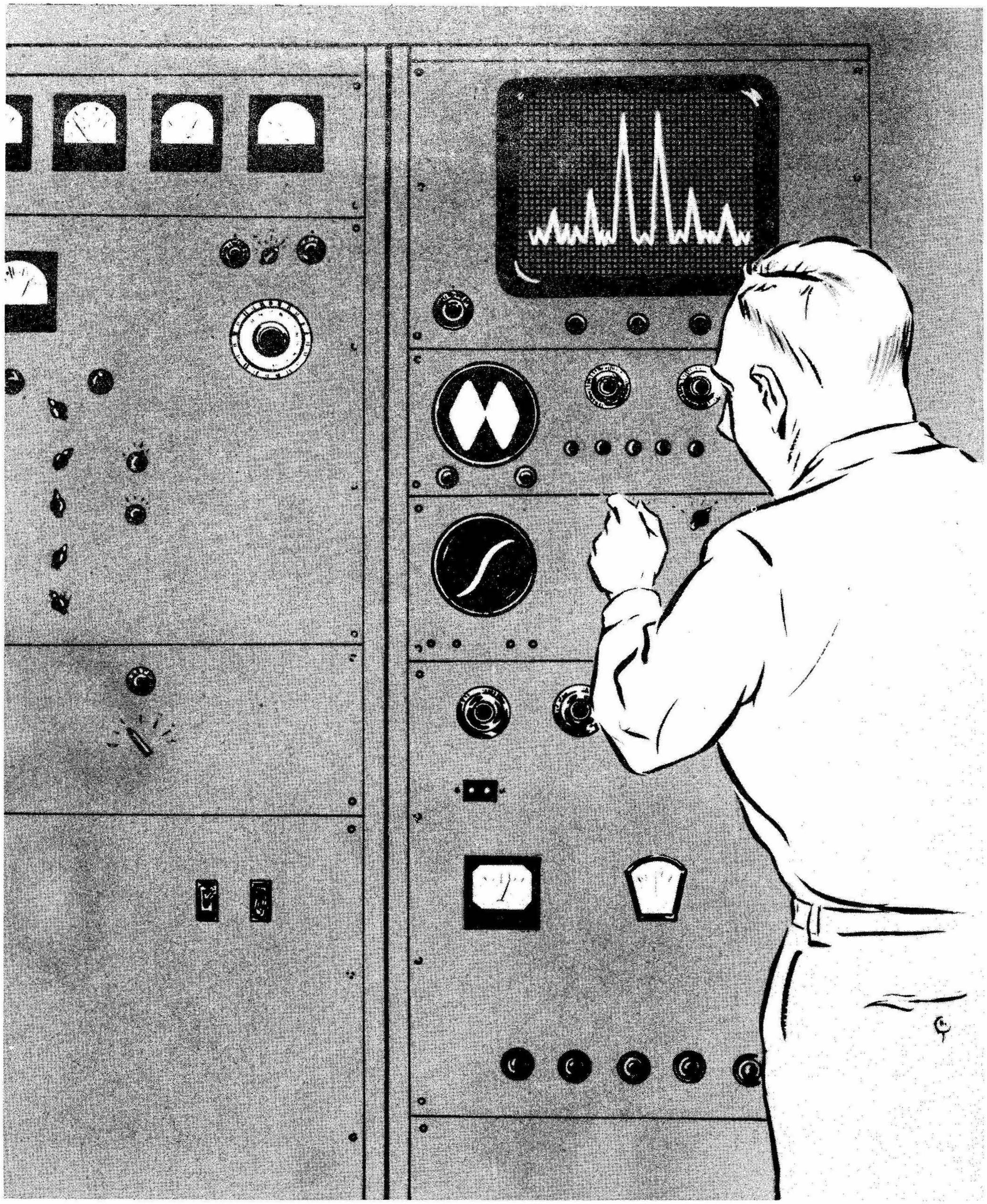

Linearity Testing of SSB Transmitter.

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INTRODUCTION

WHAT IS MODULATION?

Modulation, in the general sense, is the process of translating a message into intelligence-bearing signals so that the message can be transmitted over an intervening medium. Specifically pertaining to the field of radio communications, modulation is the process by which some property of an r-f signal is varied in accordance with the message to be transmitted. The necessity for modulation in radio communications stems from the fact that the message desired to be conveyed is seldom, if ever, in a form suitable for direct propagation over the intervening medium.

One reason for modulation is to prepare the message so that it is in suitable form for transmission. This includes: translating the frequency of the intelligence-bearing signal (message) to a frequency that is more suited for transmission over the desired medium (free space, open-wire lines, coaxial cable, etc) ; increasing the power of the intelligence-bearing signal so that the receiver circuitry can be simplified; and changing the bandwidth (either increasing or decreasing) of the signal so as to obtain optimum balance between the bandwidth and signal-to-noise ratio characteristics of the system.

Another reason for modulation is to translate the original message from one medium to another, as is the case in an ordinary telephone system. In such a system the varying sound intensities produced by speaking are considered to be the intelligence-bearing signals and are translated into electrical signals in the telephone transmitter (mouthpiece). Other examples of systems which translate intelligence-bearing signals from one medium to another are telegraphy and teletype systems, where the mechanical operation of the respective keys is changed to electrical signals; and television and facsimile, where the original message in the form of visible light is translated into electrical impulses.

Modulation also serves to convey messages from one point in space-time to another point in space-time. An example of this is the automatic recording and playback systems in common use. In such systems the message, modulated or impressed on the recording device (phonograph disk, tape, or drum), is "received" at one time and "delivered" at some later time, thus introducing a delay in space-time in the delivery of the message.

HISTORICAL BACKGROUND

Many of the fundamental concepts of modulation have been known for quite some time. As early as 1758 a scheme for a telegraph system was proposed by Marshall (a Scotsman) whereby a separate wire would be used for each letter of the alphabet. This is often considered to be the first practical scheme pertaining to the art of telegraphy. Between 1800 and 1855 the basic ideas of the telegraph were applied to practical use by many experimenters, with the first telegraph message on record being sent by Morse in 1838. During this period the fundamental concepts of time-division and frequency-division multiplexing of telegraph signals were conceived. In approximately 1870 a two-tone telegraph system that used separate frequencies for "mark" and "space" intervals was developed. This was one of the earliest applications of frequency-modulation.

The telephone followed in 1875. Although it was demonstrated experimently that speech messages could be time-division multiplexed, the lack of high-speed mechanical-switching devices at that time prevented the commercial application of such systems. However, approximately 30 years later such switches were available, and time-division multiplexing of voice messages was applied commercially. In 1914 experiments were conducted with frequency-division multiplexing of voice signals; and because of the invention of the vacuum tube and the electronic wave filter such systems were suited for commercial applications. Also in 1914 it was established mathematically that an amplitude-modulated wave consisted of the carrier and two identical sidebands. This led to experiments in single sideband in the following year, with the concepts of single sideband as known today being conceived by Carson in that year. In 1928 means of measuring the bandwidth of a message, and the message-transmitting capacity of a system, were proposed. The advent of pulse modulation followed, with pulse-code modulation being the latest development in this area.

Recent developments in scatter propagation indicate that the basic concepts of modulation and communications theory are still being extended.

AMPLITUDE MODULATION

As stated previously, modulation, pertaining to the radio-communications field, is the process by which some property of an r-f carrier signal is varied in accordance with the message to be transmitted. In amplitude-modulation, as the name suggests, the amplitude of the r-f carrier is the property that is varied. Waveforms of an r-f carrier, audio modulating signal, and amplitude-modulated wave are illustrated in figure 1. Although the modulating signal is very seldom a pure sine wave in actual practice, such a waveform is usually used to illustrate the concepts of amplitude modulation. It will be noted that the outline of the modulated wave (dotted line) resembles the wave shape of the modulating signal. This outline is commonly called the "modulation envelope."

If the modulating signal (defined as the signal

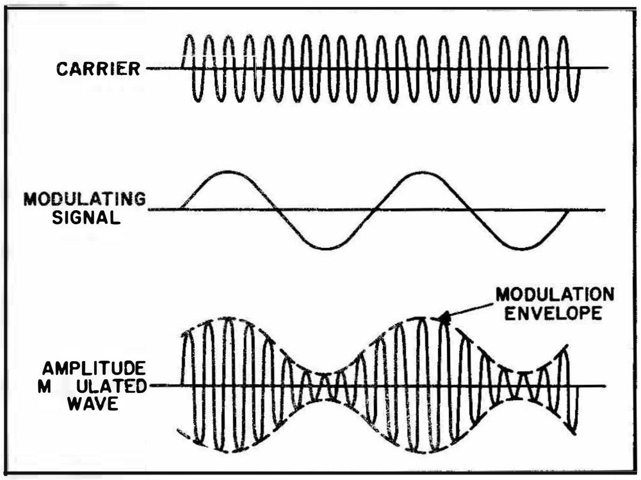

Figure 1. Illustration of Carrier, Modulating Signal, and Amplitude Modulated Wave.

that contains the details of the message) is high in frequency, the amplitude variations of the carrier will be more rapid than if the modulating-signal frequency were low. Similarly, if the volume of the modulating signal is loud, the carrier amplitude variations will be larger than those for low-volume modulating signals. Examples of both these facts are illustrated in figure 2.

Percentage of Modulation

There are certain limitations as to the strength (volume) of the modulating signal. For example, in order to produce an amplitude-modulated wave having maximum modulation and no distortion, the modulating signal should decrease the amplitude of the carrier signal to zero, and raise the carrier amplitude to twice its normal value at the peaks of the modulating signal. The extent to which the modulating signal modulates the carrier is called the "degree of modulation." This degree of modulation is actually the ratio of the peak amplitude of the modulating signal to the peak amplitude of the carrier wave. When the degree of modulation is multiplied by 100, and expressed as a percentage, it is then called the "percentage of modulation." Mathematically the percentage of modulation is expressed:

$$\% \text{ of Modulation} = \frac{\text{Peak Amplitude of Modulating Signal}}{\text{Peak Amplitude of Carrier}} \times 100$$

To obtain 100% modulation, the ratio of the two signals should equal 1. At 100% modulation the amplitude of the modulated wave varies between zero and twice the normal amplitude of the carrier as the modulating signal varies between its most negative and positive excursions. Examples of different percentages of modulation are illustrated in figure 3. It will be noted that when this percentage exceeds 100%, distortion occurs in the modulated wave. This distortion, called overmodulation, will cause a loss of fidelity in the signal by changing the modulation envelope waveshape. Overmodulation also causes a loss in transmitter output power, since maximum undistorted power output of a transmitter is obtained at 100% modulation.

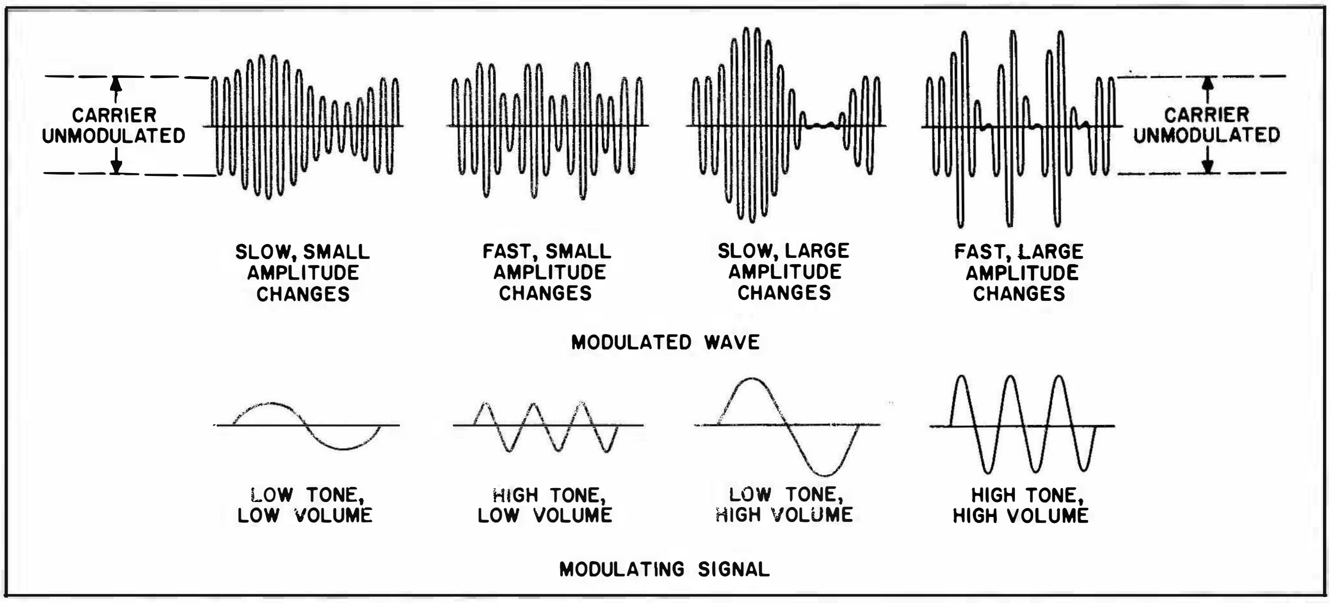

Figure 2. Waveforms of an Amplitude-Modulated Carrier Signal for Varying Types of Modulating Signals.

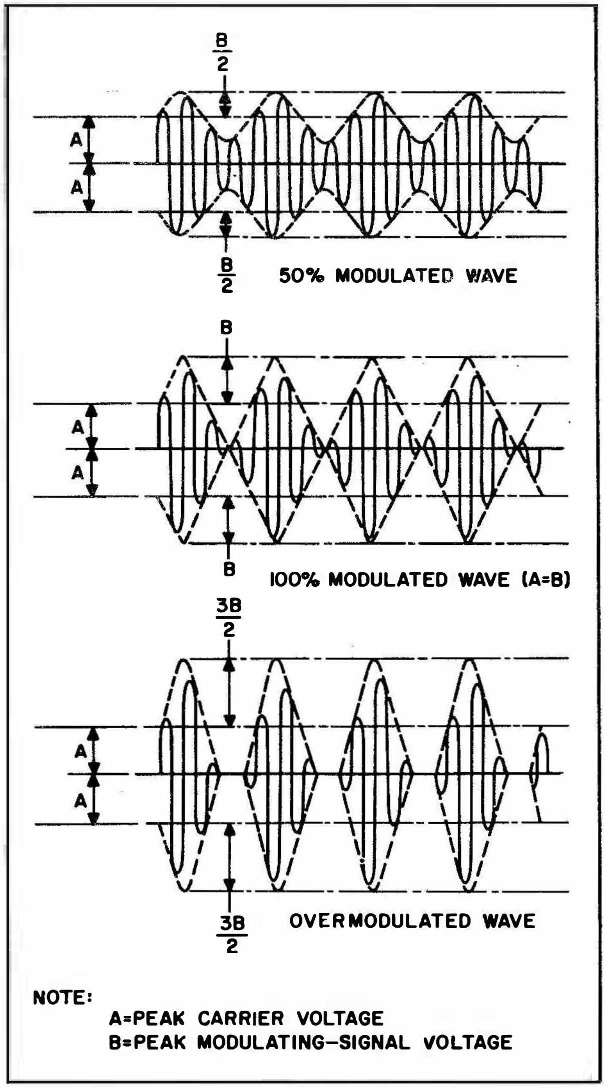

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Figure 3. Examples of Waves Having Different Percentages of Modulation.

Sidebands

In the process of modulating an r-f carrier with a lower-frequency signal, a heterodyning action, which causes the generation of additional frequencies, takes place in the modulator stage. If only a single-frequency signal is used as the modulating signal, two such additional frequencies will be generated (as illustrated in figure 4). One of these, called the upper side-frequency, is the sum of the two beating signals; and the other is the difference between the two beating signals and is called the lower side-frequency. Although it is not readily apparent, the resultant modulated wave produced by this modulation process is actually a combination of the carrier wave and the two side frequencies. If a band of frequencies, instead of a single frequency, is used to modulate the carrier, the additional frequencies generated by the heterodyning action are called upper and lower sideband frequencies. These sidebands will still assume their respective positions in the frequency spectrum above and below the carrier signal, and the resultant modulated wave will be the combination of the carrier and all the frequencies contained in these sidebands.

The carrier wave itself contains none of the intelligence of the modulating signal, all such intelligence being contained in the sidebands. If the modulating signal is removed, the sidebands will no longer be present and only the carrier will remain. Therefore, it can be stated that the sidebands are directly related to the modulating signal. In the process of modulation, all of the intelligence and power of the modulating signal are applied to the generated sidebands, although this fact is not readily apparent upon examination of the modulated wave itself. Another important fact that should be understood is that, although the two sidebands occupy different positions in the frequency spectrum, they both contain the same information (intelligence of the modulating signal) and are actually mirror images of each other.

Methods of Modulation

There are various methods of amplitude-modulating an r-f carrier wave with a lower-frequency signal. Although cathode, screen-grid, and suppressor-grid modulation methods are sometimes used, the most common methods are plate and control-grid modulation. Some of the different modulation methods are illustrated in figure 5.

In the plate-modulation method (figure 5A), the output of an audio-amplifier stage is applied to the plate circuit of an r-f amplifier. The plate-supply voltage of the r-f amplifier will then be varied in accordance with the audio-modulating signal, producing an amplitude-modulated r-f carrier wave in the output of this circuit.

In one form of control-grid modulation (figure 5B), the output of the audio amplifier is applied to the control grids of a Class B push-pull r-f amplifier. The grid bias (and the output signal) of the r-f stage will thus be varied in accordance with the input audio-modulating signal. Adjustments in this circuit are more critical than those required in the plate-modulated circuit.

Cathode modulation is effectively a combination of plate and control-grid modulation. In this circuit (figure 5C) the modulating signal, applied to the cathode of the r-f stage, effectively varies both the control-grid bias and the plate-supply voltage—thus varying the r-f output signal in accordance with the audio-input signal.

In screen-grid modulation the modulating signal is applied to the screen grid of the r-f stage. In such a circuit (figure 5D) the variations in screen-grid voltage, caused by the variations of the input modulating signal, produce large changes in plate

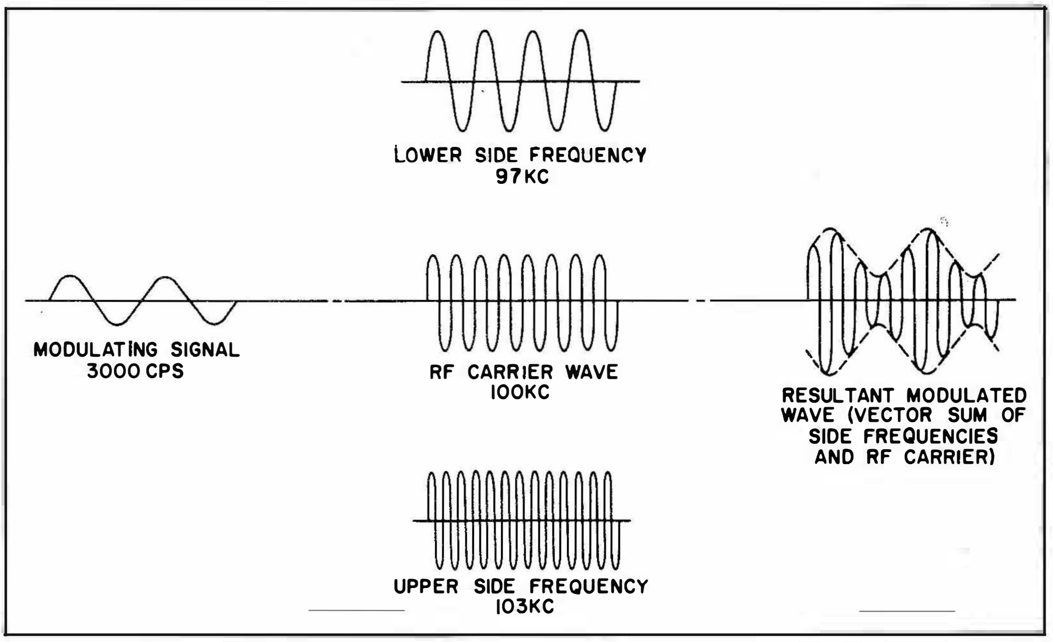

Figure 4. Sideband Frequencies and Resultant Waveform Generated During Amplitude Modulating Process.

current of the r-f stage. This will produce an amplitude-modulated carrier wave at the output of the stage. This form of modulation is usually used instead of plate modulation in pentode type tubes.

Another method of modulating pentode r-f amplifier stages is to use a combination of plate and screen-grid modulation (figure 5E). This circuit employs separate modulation-transformer windings that apply the modulating signal in the same phase to the plate and screen grid of the tube. Using this method, a higher degree of modulation can be obtained than if separate plate or screen-grid modulation methods were used.

Pentode tubes serving as r-f amplifiers can also be modulated by applying the modulating signal to the suppressor grid, which is usually biased by some negative potential (figure 5F). Complete modulation is difficult to obtain with this method; however, 90% modulation with good linearity is possible.

High-Level and Low-Level Modulated AM Transmitters

The block diagram of a simple amplitude-modulated transmitter is illustrated in figure 6. In this circuit arrangement an r-f carrier signal is generated in a stable crystal-controlled oscillator, generally referred to as the "master oscillator." A buffer amplifier amplifies the crystal oscillator output to a level sufficient to drive the r-f power amplifier, which, in turn, provides sufficient amplification to produce the desired transmitter output-power level. The modulating signal, originating as speech frequencies at the input to the microphone, is amplified in the speech amplifier and serves to drive the modulator stage. The modulator stage, in turn, produces an a-f signal sufficient in amplitude to 100% amplitude-modulate the r-f carrier in the power amplifier. The desired output-power level of the transmitter determines the number of stages (and their operating characteristics) preceding the final r-f power amplifier. This type of transmitter is generally referred to as a "high-level" modulated transmitter since the modulation process is effected at a relatively high power in the final power amplifier stage.

A "low-level" modulated AM transmitter (figure 7) is one in which the modulation process is effected in one of the low-power intermediate r-f amplifiers. To obtain the desired transmitter output-power level, the modulated r-f carrier is then amplified in linear amplifiers to preserve the relationships of the modulating components. Because linear amplifiers are used in this system, the efficiency of the low-level modulated transmitter is much lower than that of a high-level system. However, an advantage of the low-level system (over a high-level system) is that lower a-f power is required to drive the r-f amplifier being modulated.

Conventional AM Superheterodyne Receiver

Basically a radio receiver is a device that con-

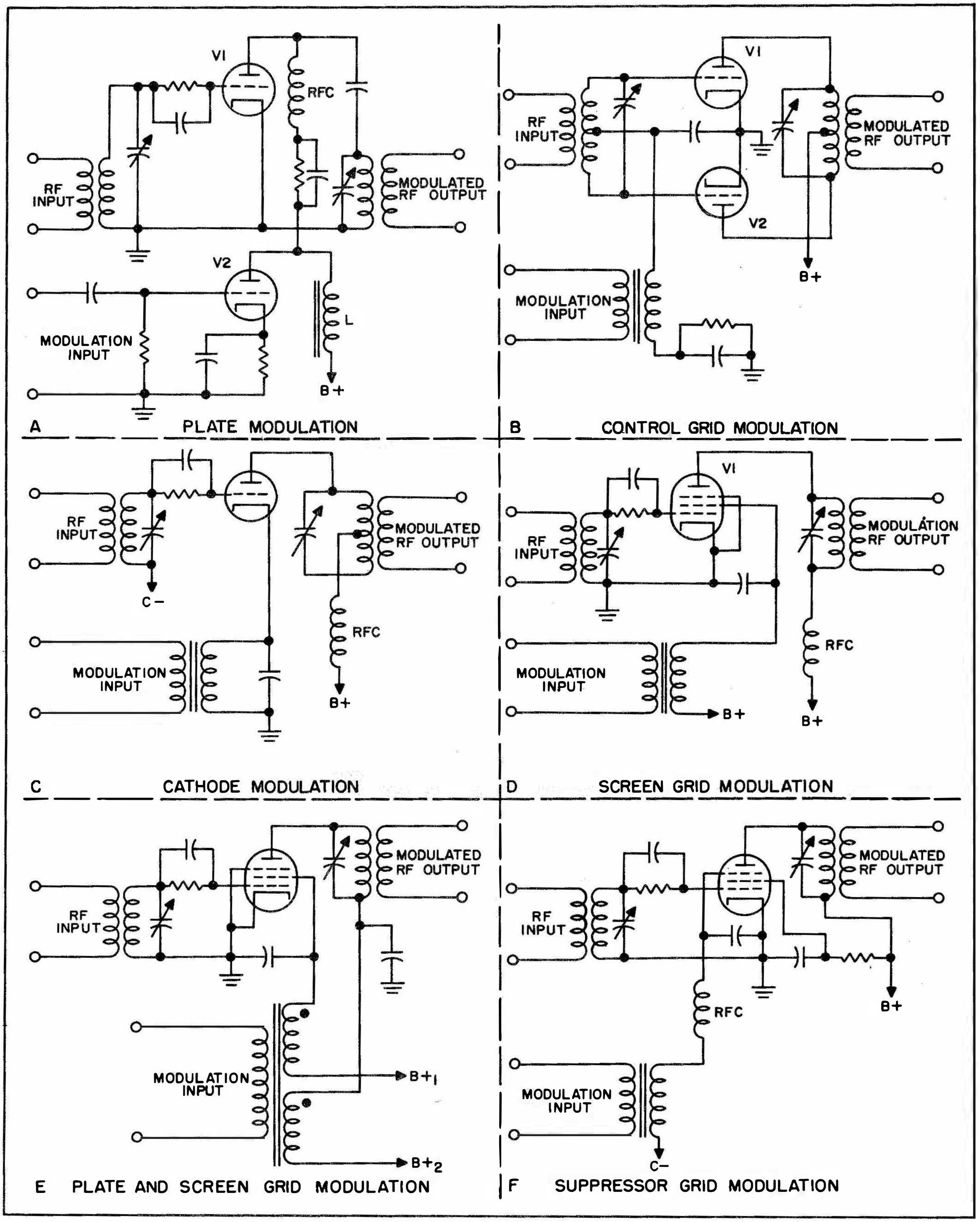

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Figure 5. Different Methods for Amplitude-Modulating an R-F Carrier Wave.

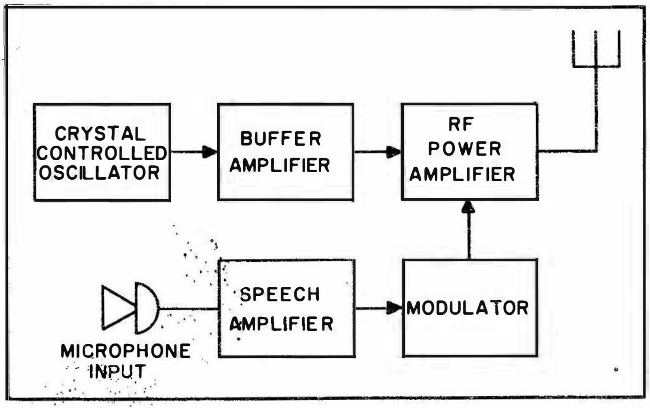


Figure 6. Simple High-Level Modulated AM Transmitter.

verts radio waves into discernible signals. In any radio receiver, however simple or complex it may be, there are certain minimum essential functions that it must be capable of performing. These functions are: (1) reception, the ability to receive radio waves propagated in free space; (2) selection, the ability to select one of a multitude of such propagated radio waves; (3) detection, the ability to detect (or remove) the intelligence from the selected radio wave; and (4) reproduction, the ability to reproduce the detected intelligence so that it can be understood.

A simple AM superheterodyne receiver, one in which one or more frequency changes take place, is illustrated in block diagram form in figure 8. The minimum essential functions mentioned above are also indicated on the diagram. In this receiver, reception of the propagated radio wave is accomplished by the antenna circuit. Selection of one particular frequency is accomplished by the r-f amplifier stage, the amplified output of which is applied to a frequency converter. The frequency converter, local oscillator, i-f amplifier, and AM detector circuits perform the functions of converting the r-f signal to a lower frequency, amplifying the resultant signal, and detecting the intelligence from it. Briefly, these functions are accomplished in the following manner: The local oscillator signal beats with the incoming r-f signal in the frequency converter, the output of which consists of its two input frequencies (r-f and local oscillator) and the sum and difference frequencies between the two input signals. The i-f amplifier selects only the difference frequency from the converter output, amplifies this signal, and applies it to the AM detector. The detector stage then removes the intelligence (modulation) from this intermediate frequency, and applies this detected signal to the a-f amplifier.

The function of reproduction is performed by the a-f amplifier, which amplifies the detected signal, and the speaker, which is driven by the amplified a-f signal to produce audible sound waves. The sound waves produced by the speaker are similar to those that were amplified and used to modulate the r-f carrier signal in the radio transmitter. Thus, through the processes of modulation in the transmitter and demodulation in the receiver, the original message introduced at the transmitter in the form of sound waves is reproduced by the radio receiver.

FREQUENCY MODULATION AND PHASE MODULATION

Two methods of modulation differing from amplitude-modulation, but closely allied to each other, are frequency modulation (FM) and phase modulation (PM), both being forms of angle modulation. These methods differ from AM in that the amplitude of the r-f carrier is maintained constant while the carrier frequency and carrier

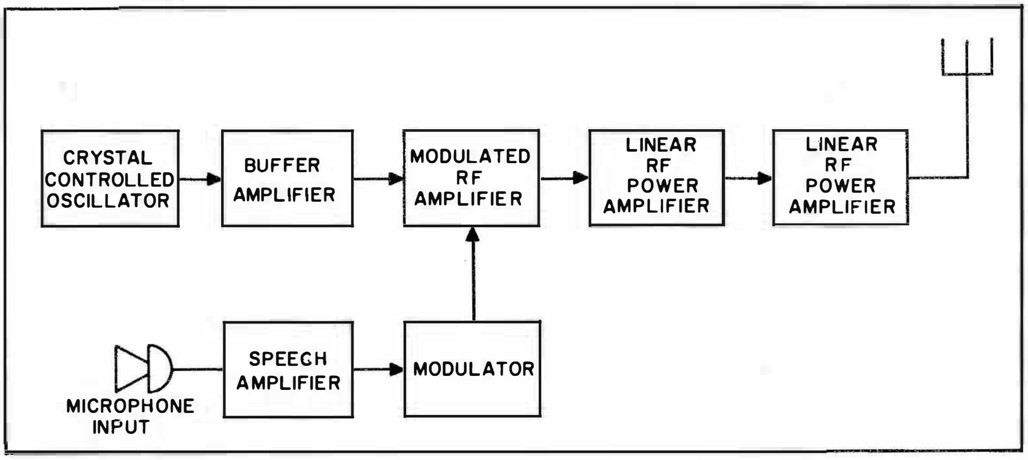


Figure 7. Simple Low-Level Modulated AM Transmitter.

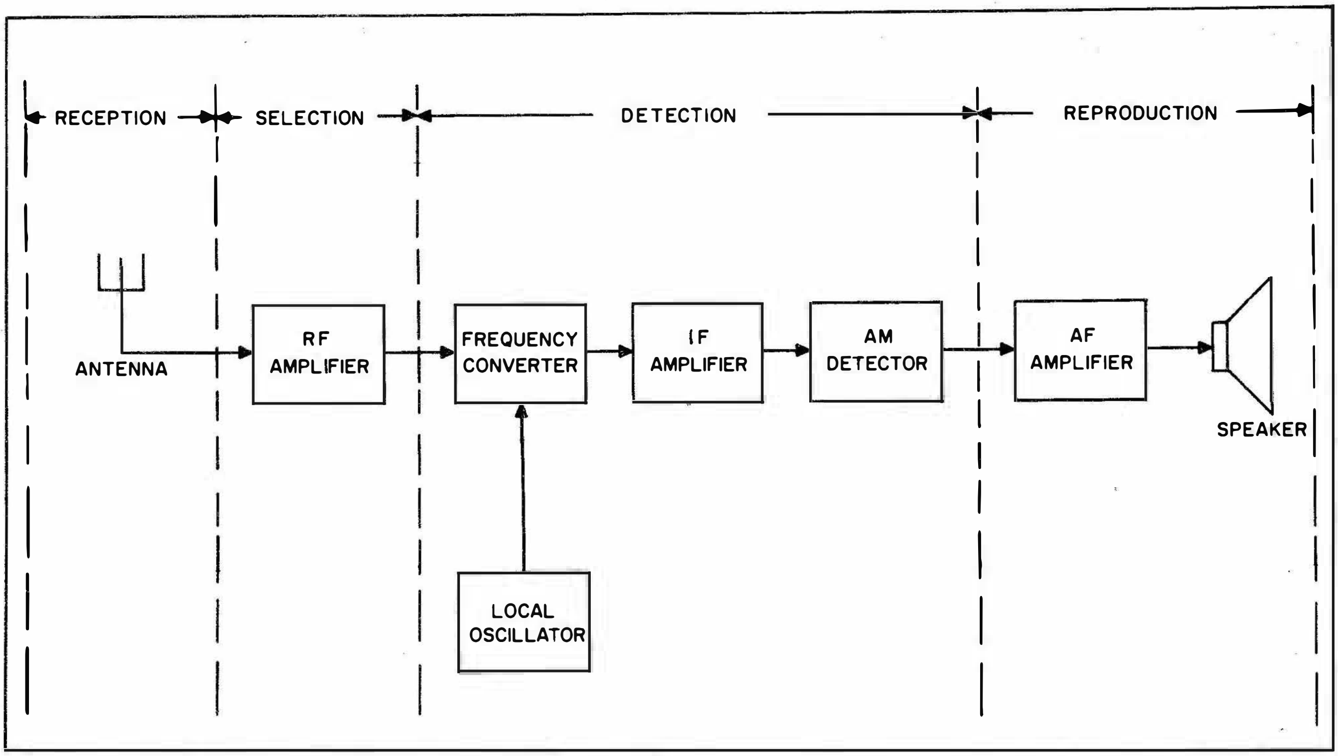


Figure 8. Simple AM Superheterodyne Receiver.

phase are caused to vary in accordance with the modulating signal. In FM the frequency of the carrier is varied in accordance with the amplitude of the modulating signal, with the rate of this frequency change being dependent upon the frequency of the modulating signal. When the phase of the carrier is varied in accordance with the amplitude of the modulating signal, and the rate of this phase change is determined by the frequency of the modulating signal, phase-modulation results. Actually, these two forms of modulation are inseparable since any change in carrier frequency necessarily involves a change in carrier phase, and any change in carrier phase involves a change in carrier frequency.

The assigned carrier frequency of an FM transmitter is called the "resting frequency," or "center frequency." This frequency corresponds to the unmodulated carrier frequency of an AM transmitter. The variation in frequency above or below the resting frequency (caused by the modulating signal) is called the "frequency deviation," with the total frequency variation being called the "carrier swing."

The sideband frequencies present in FM systems are a direct result of the amplitude of the modulating signal. The rate of change of the sideband distribution and the separation between the individual sidebands depend on the frequency of the modulating signal. Because the resting frequency is shifted back and forth by the modulating signal, the number of sidebands thus produced are theoretically infinite. However, there is a practical limit to the number of usable sidebands since the sidebands beyond this practical limit contain so little power they are not usable by the receiver. Because of the fact that the over-all FM signal does not change in amplitude, the sideband power must be obtained from the carrier frequency. In fact, vectorial addition of the amplitudes of all the sidebands and the FM carrier will produce a sum equal to the amplitude of the unmodulated carrier.

It will be recalled that the degree to which the amplitude of the modulating signal modulates the r-f carrier in an AM system is called the percentage of modulation. Since the amplitude of the carrier remains constant in an FM system, the degree of modulation of such a system is expressed in terms of the "deviation ratio," or "modulation index." This ratio is defined as the ratio of the maximum deviation of the resting frequency (frequency deviation) to the maximum modulating frequency causing this deviation. Mathematically this ratio is expressed:

$$\text{Deviation Ratio} = \frac{\text{Max Frequency Deviation}}{\text{Max Modulating Frequency}}$$

The maximum frequency deviation in commercial applications is limited to 75 kc. In military applications this maximum deviation is limited to 40 kc, and is classed as narrow-band FM transmission.

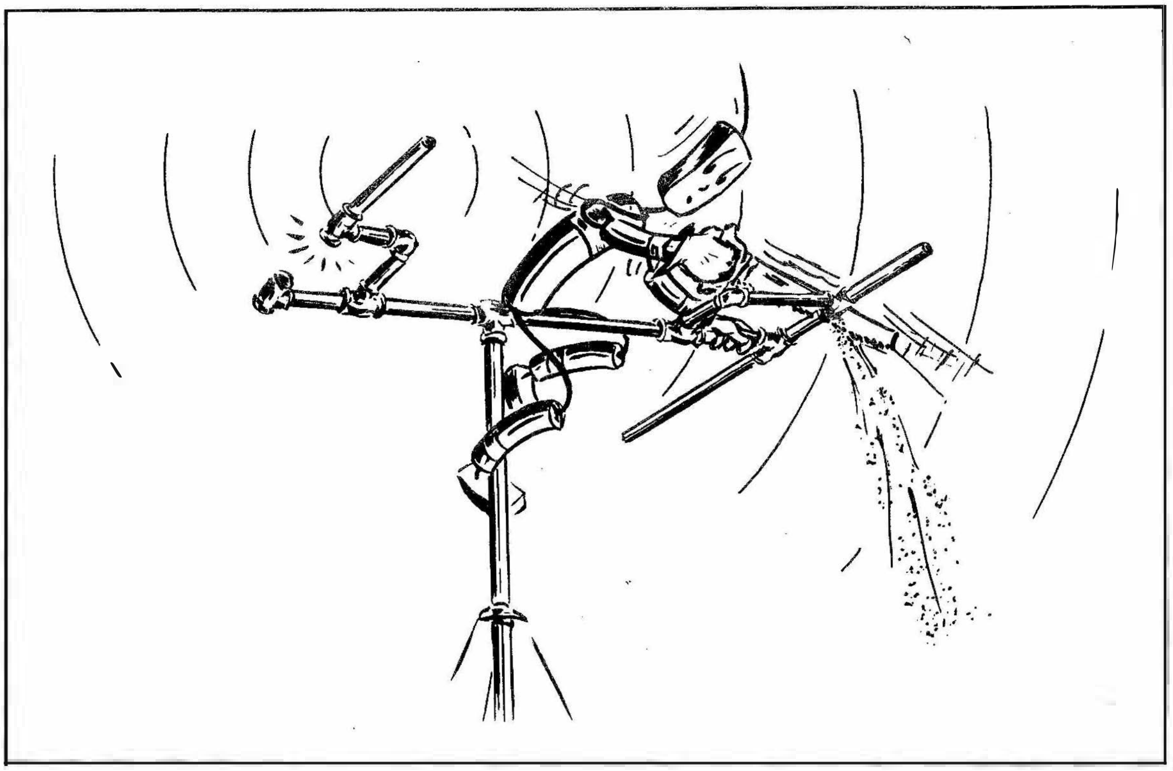
Compared to AM, FM has the disadvantage of using a wider bandwidth for the same fidelity of reproduced signal. The advantages of FM over AM are: less interference from natural sources and nearby transmitters, less noise for a given amplitude of receiver output, and better fidelity for the same transmitter power.

PULSE MODULATION

In one manner of speaking, pulse modulation is a method of modulating an r-f carrier with a pulse train. This form of pulse-modulation is widely used in radar and radar-associated system applications.

In another manner of speaking, pulse modulation is the process in which some characteristic of a pulse carrier is made to vary in accordance with a modulating signal, with the term "pulse carrier" being defined as a carrier consisting of a series of pulses. In such a system the series of pulses, instead of a continuously changing wave, is used to carry the intelligence of the message.

There are various forms of pulse-modulation systems, such as pulse-amplitude modulation (PAM), pulse-code modulation (PCM), pulse-duration modulation (PDM), pulse-position modulation (PPM), etc. It is beyond the scope of this text to elaborate on each of these pulse-modulation methods; such information can be obtained from other publications. At this point let it suffice to say that systems employing time-division and frequency-division multiplexing with these forms of pulse modulation can be used with single-sideband applications. However, because of the fact that flat-topped waveforms having a short rise time (in proportion to the duration of the pulse) produce high peak voltages in single-sideband signals, care must be exercised in the use of pulse-modulation methods with SSB systems.



". . . single sideband transmission——"

CONCEPTS OF SINGLE SIDEBAND

HISTORY AND EARLY USE OF SSB

Single sideband is a method of transmitting intelligence whereby the r-f carrier and one sideband of an amplitude-modulated double-sideband wave are suppressed, and only one of the intelligence-bearing sidebands is transmitted. There are two basic methods of generating such a signal: (1) eliminate the carrier and one sideband by using filters; and (2) balance-out the carrier and one sideband by using balanced modulators with quadrature input signals.

Contrary to common belief, this method of transmission is not new. As early as 1914 it was established analytically that an amplitude-modulated wave consisted of three distinct components: the carrier frequency, and upper and lower side frequencies set apart from the carrier by the frequency of the modulating signal. In 1915 experiments were conducted at the U. S. Naval Radio Station, Arlington, Virginia by H. D. Arnold, who suggested that the equipment be tuned to one side of the carrier frequency to pass only this one sideband while attenuating the other sideband. This is generally recognized as the first actual proof that one sideband contained all the modulation components required to produce the original modulating signal.

Also in 1915, at about the same time that Arnold was conducting his experiments in Arlington, John R. Carson, in analyzing vacuum-tube modulation, "found" the distinct components determined analytically in 1914. Previous to this it was discovered (by B. W. Kendall) that injection of the carrier frequency at the receiver greatly improved detection of the received signal. Knowledge of the facts that the carrier frequency could be injected separately at the receiver, and that one sideband contained all the modulation components required to reproduce the original signal probably prompted Carson's idea to suppress the carrier as well as one sideband. Late in 1915 Carson applied for a patent on his idea of suppression of one sideband, suppression of the carrier, and suppression of the carrier and one sideband. This patent, however, was not granted to Carson until early in 1923 after many interferences. Thus, it is generally recognized that John R. Carson first conceived the idea of single sideband transmission as it is known at the present time.

Single-sideband methods have been in use in wire carrier-telephone systems for approximately 40 years, and in low-frequency transoceanic radiotelephone systems for approximately 30 years. Prior to 1936, high-frequency (3—30 mc) transoceanic telephone systems employed double-sideband with carrier methods because developments of single-sideband equipment in this frequency range had not at that time progressed far enough to allow practical transmission of single-sideband signals. In 1936, however, SSB equipment capable of performing satisfactorily in this high-frequency range was placed in operation in overseas systems. Multi-channel teletypewriter SSB systems were used extensively during World War II by the United States Armed Forces for global communications.

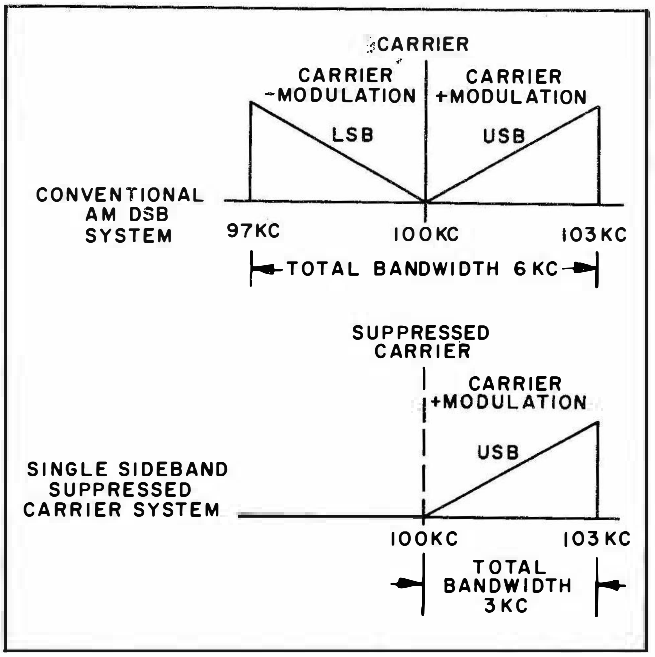
In early voice SSB equipment, provision was made for only one speech channel on one side of the carrier. Later equipment, however, had provisions for transmission of two speech channels at a time—one placed on each side of the carrier. This effectively doubled the number of speech channels over that obtained with the earlier equipment.

At the present time, the SSB method of transmission is generally accepted as being standard for long-range point-to-point communications systems. In fact, it has been suggested that single-sideband methods be utilized for all point-to-point radiotelephone applications at frequencies below 30 mc.

SPECTRUM CONSERVATION

An ordinary amplitude-modulated signal consists of a carrier frequency and side frequencies spaced above and below the carrier by an amount equal to the frequency of the modulating signal. These side frequencies are actually mirror images of each other, only one of them being required at the receiver for demodulation to obtain the original modulating information. Because of this fact, ordinary amplitude modulation is inherently wasteful of the frequency spectrum, since it requires at least twice the bandwidth of the original modulating signal. This is true regardless of the form of modulating signal used; i.e., voice, teletype, video, etc. To transmit a given form of intelligence, single-sideband systems use only one-half the bandwidth required for AM double-sideband systems. One of the important advantages of SSB systems, therefore, is the conservation of the frequency spectrum.

Figure 9 is an example of the spectrum comparison of SSB and ordinary amplitude-modulated DSB systems. With a carrier frequency of 100 kc, and modulating information (voice) from 100—3000 cps, the total bandwidth of the AM double-sideband system is 6000 cycles, or twice the highest modulating frequency. With the same carrier and modulating signals, the bandwidth of the SSB system is only 3000 cycles, or one-half that required for the ordinary amplitude-modulated system. Since the carrier does not contain any of the modulating intelligence, it is suppressed at the transmitter. It is obvious then that the single-sideband method of transmission can provide



TP7-1208

Figure 9. Comparison of AM DSB and SSB Frequency Spectrum Utilization.

twice the number of channels of the ordinary AM system. One method of doubling the number of channels is to transmit intelligence from two unrelated input sources on each side of a suppressed carrier. By using this method (dual channel, illustrated in figure 10), optimum utilization of the frequency spectrum can be realized.

With single-sideband and multiple-tone transmission it would be possible to transmit a minimum of 16 teletype circuits in one 4000-cycle-wide teletype channel. At the present time frequency-shift transmission of only one teletype circuit, or a maximum of four time-division multiplexed teletype circuits, is possible in one normal 4000-cycle-wide teletype channel.

VESTIGIAL SIDEBAND

When the modulating signal includes extremely low frequency components (approaching zero frequency), it becomes more difficult to design SSB systems to properly suppress the undesired sideband. To ease this design problem, a small portion of the undesired sideband is sometimes transmitted with the desired sideband. This method of transmission, called vestigial sideband, is actually a form of single sideband, and is used extensively in the transmission of television signals. The bandwidth of the vestigial sideband signal in standard television broadcasting is approximately one-sixth the bandwidth of the full sideband.

FADING EFFECTS

Multipath or selective fading is caused by the reception of transmitted-signal components from two or more propagation paths of different lengths. When an ordinary amplitude-modulated double-sideband signal is received in this manner, combining of the multipath signals at the receiver can cause the carrier to be partially or totally cancelled, or cause the individual sideband components to be out-of-phase with respect to the carrier or each other. Such phase changes in an amplitude-modulated signal will change the signal from one of pure amplitude modulation to one of

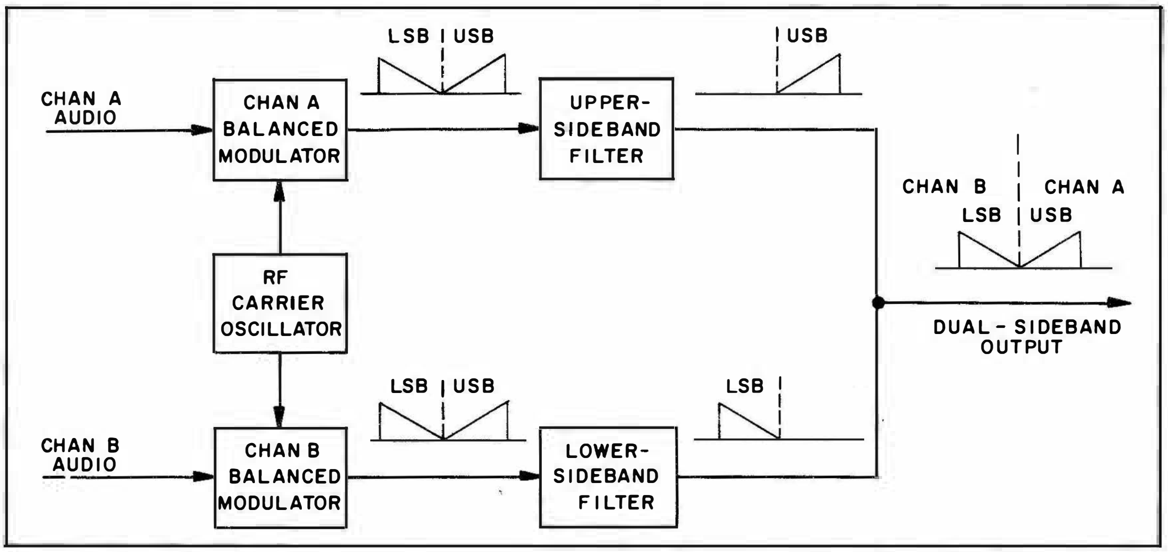


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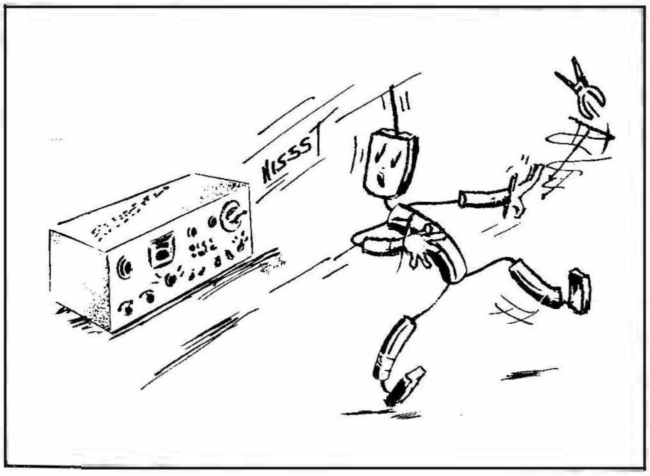
Figure 10. Simplified Example of SSB Dual-Channel System.



"... although receiver hiss increases with bandwidth——"

an amplitude- and phase-modulated combination. If the fading phenomena completely cancels the carrier or individual sideband components, harmonic and intermodulation types of distortion will be generated by the original modulation signals in the process of demodulating this signal in the receiver.

Since only one sideband is transmitted in SSB systems, there is no direct phase or amplitude relationship between the individual sideband components in this type of signal. The effects of fading of one sideband component will not affect any of the other sideband components. Consequently, there is no harmonic or intermodulation distortion generated in the demodulating process in the SSB system. Multipath fading effects on SSB systems produce what is known as "amplitude-vs-frequency distortion." This form of distortion makes voice signals sound odd, but does not materially affect the intelligibility of the message.

POWER COMPARISON OF SSB AND AM DSB

To provide the same coverage as an AM system, an SSB transmitter requires a peak power rating of approximately 9 db less than the AM system. In an AM transmitter modulated 100%, the voltage in each of the sidebands is one-half the carrier voltage. If the resistance of the circuit remains constant, power in each sideband will be one-fourth the carrier power ($P = E^2/R$). With a carrier output of 100 watts, the power in each sideband will be only 25 watts, giving a total sideband power of 50 watts and an average transmitter power rating of 150 watts. Since the peak voltage (carrier plus sidebands) is twice the carrier voltage, the peak power of the transmitter will be four times the carrier power. Therefore, an AM transmitter rated at 150 watts average power must be capable of handling peak power up to 400 watts. Of the 150 watt average power, only one-third (or 50 watts) is used to transmit the intelligence-bearing sidebands, since the other two-thirds (the 100-watt carrier) carries no useful modulating intelligence. To transmit the same intelligence with an SSB system, the transmitter need be designed to handle only 25 watts of peak power, since the high-power carrier and one sideband are suppressed and all the transmitter power is applied to the intelligence-bearing SSB signal. A comparison of the frequency-vs-power spectrum between AM and SSB systems is illustrated in figure 11.

SIGNAL-TO-NOISE RATIO

A theoretical 9 db signal-to-noise ratio advantage is claimed for an SSB system over an equivalent peak power AM system. This 9-db improve-

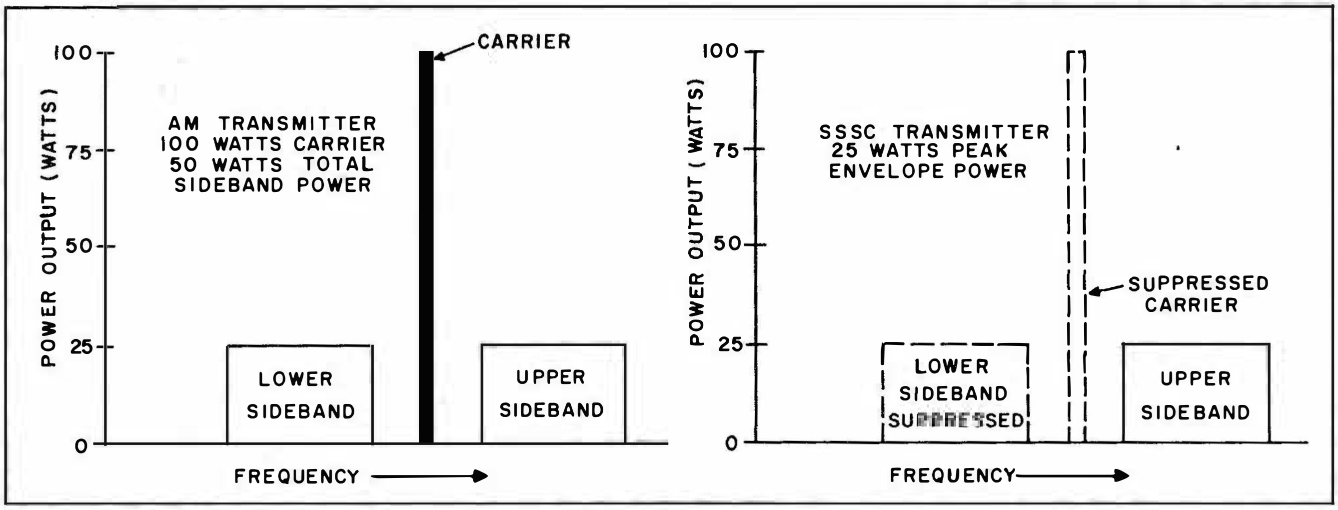


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Figure 11. Comparison of AM DSB and SSB Frequency-vs-Power Spectrum.

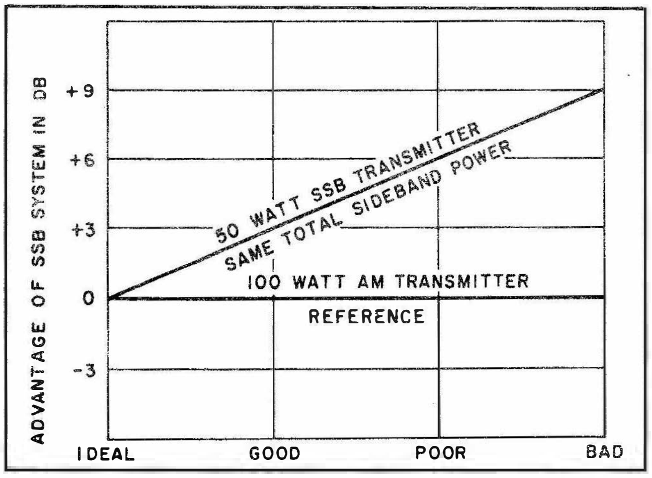

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Figure 12. Relative Comparison of AM DSB and SSB Systems with Changing Propagation Conditions.

ment includes a 3-db gain attributed to the narrower bandwidth of the single-sideband signal, and a 6-db gain obtained from a combination of the SSB transmitter and receiver. However, under actual conditions, the comparison of the two systems depends a great deal on the prevailing propagation conditions. Under ideal conditions of propagation, the two systems will produce effectively the same results, and neither system will have a noticeable advantage over the other. Although noise (receiver hiss) increases with bandwidth, the fact that the two sidebands of the double-sideband AM signal add coherently overcomes the wide bandwidth disadvantage of this system. As the propagation conditions become worse, because of fading or other atmospheric interferences, the signal-to-noise ratio improvement of the SSB system over an equivalent AM system becomes increasingly greater. This relationship is illustrated graphically in figure 12.

Stability Requirements for SSB Systems

The factor which has limited the applications of the single-sideband method of transmission up to the present time has been the lack of precise frequency control of the transmitters and receivers used in this mode of communications. If the demodulated single-sideband signal is expected to assume its proper place in the spectrum with regard to the carrier frequency, the frequency of the carrier re-inserted at the receiver must be within a few cycles of the original carrier frequency which was suppressed in the transmitter. For maximum performance in the high-frequency region, this frequency tolerance must be on the order of 0.2 to 2 parts per million. Frequency errors greater than 30 cycles per second can cause voice transmissions to be unintelligible under certain signal-to-noise ratio conditions. Intelligibility decreases with an increase in frequency error under high signal-to-noise ratio conditions, and can be poor with only a slight frequency error under low signal-to-noise ratio conditions.

Even though the carrier itself contains no intelligence, and all of the modulation products appear in the sidebands, the carrier, or another similar carrier, is necessary for detection of a modulated radio signal at the receiver. In a conventional AM system, the carrier is transmitted and received along with its sidebands, together forming a wave envelope, the carrier of which should normally be in the same frequency and phase relationship to the sidebands as when transmitted.

In single-sideband suppressed-carrier systems (SSSC), the carrier is suppressed approximately 30 to 50 db below the transmitter peak power and (for all practical purposes) is not transmitted with the SSB signal. For demodulation purposes, a local carrier is generated at the receiver. The frequency of this carrier should be identical to that of the carrier used in the modulating process at the transmitter, or have the same frequency relationship to all components of the received sideband. This frequency requirement is essential, since the difference in frequency between the carrier and either sideband is the modulation, or audio, frequency. If the actual sideband frequencies remain the same and the local carrier is increased or decreased in frequency, a change in audio output frequency will result. This change, either up or down in frequency, will be the same number of cycles as the change in oscillator frequency. The same holds true if the local carrier insertion oscillator is maintained in frequency and the sideband is shifted in frequency, caused by either a shift in frequency of the carrier oscillator or of one of the heterodyne oscillators used for frequency conversion in the transmitter or receiver. Thus it can be seen that frequency stability is one of the most critical and important requirements of SSB equipment.

Carrier insertion frequency becomes even more critical in a DSB receiver than in an SSB receiver. Referring to figure 13, it can be seen that a departure in frequency of the carrier away from the sideband frequency will cause a shift, upwards in frequency, in the resultant audio. This is true of either upper or lower sideband reception. A different effect is noted when the carrier is introduced with a frequency shift of the same amount but in the opposite direction in an SSB receiver, that is, so as to fall within the sideband being detected. Frequencies which should result in low-frequency audio components will now be translated into an inverted speech, or audio, and all frequencies beyond the displaced carrier will be translated to a lower frequency, opposite to the

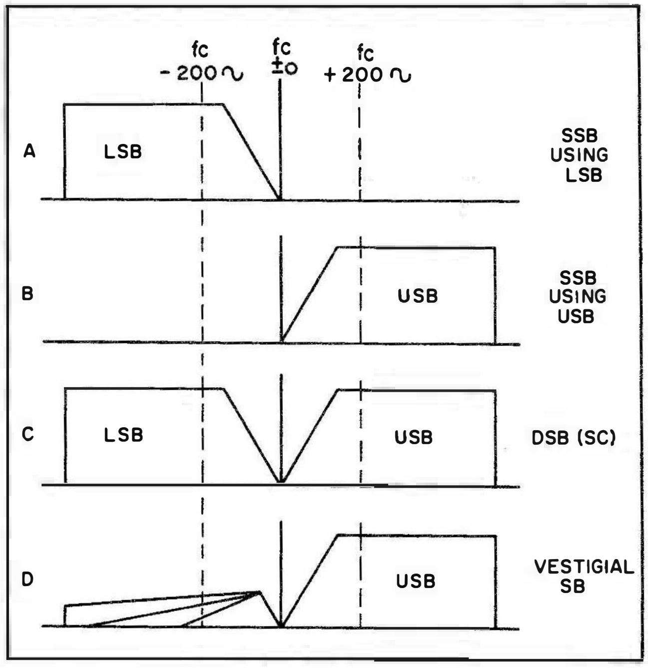

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Figure 13. Comparison of Oscillator Frequency Shifting in Different Systems.

first case. In figure 13A the lower sideband is shown transmitted. If the inserted carrier is 200 cycles higher in frequency (relative to the modulating carrier frequency at the transmitting point), then demodulation will result in all the audio frequencies being 200 cycles higher than normal. In case B, where the upper sideband is being transmitted, the same result will appear when the carrier is 200 cycles lower than normal. In both cases, speech may still be intelligible, although quite unnatural due to the higher frequencies.

When the carrier is shifted in the opposite direction, so as to fall within the sideband, or 200 cycles lower than normal in case A, then that portion of the sideband representing 150 cycles will be demodulated as 50 cycles, and that representing 50 cycles will be demodulated as 150 cycles, or all below 200 cycles will be inverted. 250 cycles will appear as 50 cycles and 350 cycles will appear as 150 cycles. The effect of this combination in speech is to present a "gravelly" speech, if the shift is not too great. At about 200 cycles the shift causes enough inversion to render speech unintelligible in most cases. Both frequency inversion of a portion of one sideband component and decrease of the frequency of the remainder of that sideband component, combined with the increase in frequency of the opposite sideband component, appear when the inserted carrier frequency is improper in double-sideband reception (figure 13C). Double-sideband signals received on an SSB receiver may appear as in figure 13D, depending on the amount of rejection of the unwanted sideband in the receiver. If the rejection is sufficient to reduce the unwanted sideband to a certain level, the garbled speech may appear as background, and the desired information, or speech, may still be understood. This is also true of vestigial sideband, or SSB with a case of poor cancellation of the undesired sideband (at the transmitter).

The effect of inversion can be prevented if the modulation is clipped below 100 cycles, and the carrier-frequency excursion does not exceed this amount in the direction of the sideband in SSB, or in either direction in DSB.

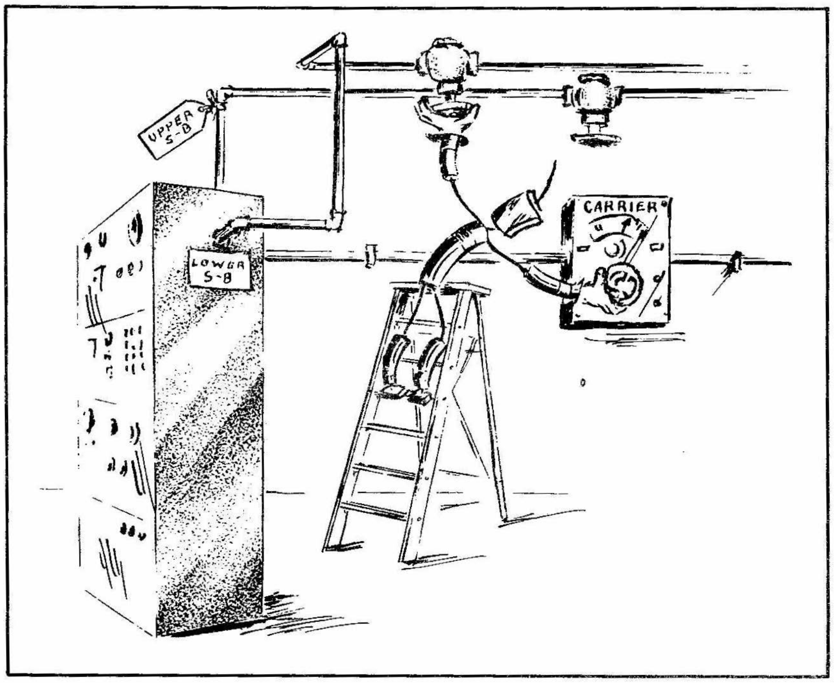
To provide some control or means of reference which enables the frequency relationship between the receiver oscillator and the transmitter oscillator to be maintained at a constant value, a pilot carrier is sometimes transmitted. In this system known as "reduced-carrier SSB" the pilot carrier is transmitted at a level approximately 10 to 20 db below the level of the transmitter peak power. This is accomplished by introducing a portion of the carrier oscillator signal into the output circuits or at some convenient point in the transmitter following the circuits where the carrier was suppressed. The pilot carrier is separated from the sideband signal in the receiver, amplified, and may be used for carrier re-insertion (exalted carrier). It can also be used as a reference to be compared with the local carrier insertion oscillator frequency in a special a-f-c circuit. Automatic-volume-control (a-v-c) voltages may also be derived from the pilot carrier.

Still another form of pilot carrier is known as "controlled carrier." In this system, the carrier rises to approximately full amplitude during the brief pauses in speech, or between syllables of speech, and is reduced to a very low level during actual modulation. The level of this controlled carrier is such that the average power output of the transmitter is maintained effectively constant regardless of the presence or absence of modulation. Slow-acting a-f-c and a-g-c circuits are used in the receiver to hold the receiver circuits on the proper frequency during modulation.

In the full-carrier SSB system, the carrier is transmitted at a level of approximately 4 to 6 db below the peak power of the transmitter. The level of the transmitted carrier will thus be sufficient to allow reception of the single-sideband signal with an AM DSB receiver. This mode of transmission is called "Compatible Single Sideband" (CSSB).

COMPLEXITY AND COST

The requirement of highly stable circuits for exact frequency control of the SSB transmitter and receiver constitutes the major problem in the



". . . reception of single sideband signal is not so simple——"

single-sideband mode of transmission. Besides increasing the complexity of the equipment, this feature also increases the cost of the over-all system. The areas of increased complexity and cost can be attributed to those circuits which directly or indirectly pertain to the demodulation of the single-sideband signal. Specifically these circuits are: the frequency generating and undesired-sideband suppressing circuits in the transmitter, and the receiver demodulating circuits.

At the present time there are relatively few precision standard frequencies, all at relatively low-fixed levels such as 100 kc, 1 mc, etc. For a multi-channel SSB system a great many stable frequencies are required; therefore, in such applications frequency synthesizers or stabilized master oscillators are usually used in conjunction with a low-fixed standard frequency. Some frequency synthesizers use as few as a dozen tubes, while others may use as many as 75 or 80 tubes. Thus one reason for the complexity and cost of an SSB system is readily apparent.

Since the carrier and undesired sideband must be suppressed as much as 30 to 40 db in SSSC systems, the sideband filters employed in filter method SSB applications must meet highly precise and exacting design requirements. In phase-shift SSB applications, the degree of suppression of both the carrier and undesired sideband depends directly upon the design of the phase-shift and the balanced modulator circuits employed. This adds another area of complexity and cost to the SSB transmitter.

The greatest area of complexity and cost, however, is in the demodulating circuits in the SSB receiver. Since the single-sideband suppressed-carrier signal does not include the carrier (which is necessary for the demodulation of the signal), a means of inserting a highly stable carrier at exactly the same frequency as the suppressed transmitter carrier must be provided at the receiver. Because of oscillator drift, it may be quite difficult at times to obtain this exact frequency relationship. When this occurs, it is desirable to transmit a pilot carrier with the desired single-sideband signal. Complex a-f-c systems must then be provided in the receiver to obtain maximum utilization of the reduced-level pilot carrier.

COMPATABILITY OF AM AND SSB

Pure single-sideband suppressed-carrier systems (SSSC) are not compatible with present day AM double-sideband systems. It stands to reason, therefore, that before a change can be made from the existing mode of communication to the pure single-sideband mode, there must be a transition period in which the two systems are compatible. In this transition period technical problems, as well as many other problems, will arise and have to be overcome before any such transition from AM double sideband to pure SSB can be considered complete.

Since pure single-sideband signals cannot be used for general communication purposes during this transition period, the signal used will have to be a compromise between this type of signal and the conventional AM DSB signal; i.e., it will have to be compatible so that it can be used by either system. The only strict requirement of this compromise "bi-mode" compatible signal is that sufficient carrier power be transmitted so that detection of the signal in a conventional AM receiver can be effected properly.

There are various methods of modifying an SSSC transmitter to radiate the necessary compatible signal. One method is to modify only the low-level SSB generating circuits so that they generate a conventional AM DSB signal. This signal will then be heterodyned and amplified to the desired transmitter operating frequency and output power level in the same manner as the pure SSSC signal. This method sacrifices the spectrum reduction and efficiency of the pure SSB system, but retains the SSB system frequency stability.

Another method of converting an SSSC transmitter is to modify the low-level SSB generating circuits so that they generate a signal containing a full carrier and only one sideband, at a slightly reduced level. The remainder of the circuits will not have to be modified; therefore, this change is relatively simple and inexpensive. Although the efficiency of the transmitter is reduced, the frequency stability and spectrum reduction characteristics of the SSB system are retained.

A third method of converting an SSSC transmitter to a compatible system is to retain the low-level SSB carrier generating circuits, and convert the power stages to operate similar to those in a conventional AM double-sideband transmitter. This modification, however, is quite complex and expensive because the power stages and output circuits must be completely redesigned. In effect, this method requires the incorporation of two separate transmitters—one AM, and one SSB—in the same unit.

A single-sideband receiver can easily be converted to receive AM double-sideband signals by including a conventional AM detector in its circuit design. Another point to consider, however, is that ordinary AM detection is usually effected at a higher signal level than SSB signal detection and, therefore, the converted SSB receiver may also require an additional stage of i-f amplification. Receivers using exalted carrier or product detectors are capable of AM reception without modification. A requirement of the i-f amplifiers used is that their bandwidth be sufficient to pass at least one sideband and the carrier of the conventional AM double-sideband signal. The carrier in the DSB signal will provide a reference for the locally generated carrier signal, and the receiver a-f-c circuits will correct for slight frequency errors in the AM transmitter. For substantially large frequency errors, manual tuning must be provided to bring the frequency error within the a-f-c circuit correction range.

The major changes required in existing AM transmitters to make them compatible with SSB systems are primarily concerned with the frequency stability of the AM system. As stated previously, manual tuning is required in a compatible SSB receiver if the AM transmitter frequency errors are substantially large. Transmitters using temperature-controlled crystals, stabilized master oscillators, or frequency synthesizers can produce the required stability characteristics and eliminate the manual tuning obstacle.

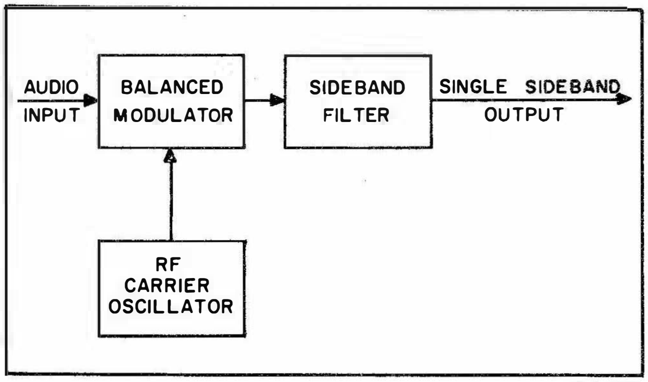
It is also possible to convert an AM double-sideband transmitter into a compatible signal transmitter by converting the low-level circuits of the AM transmitter into SSB generating circuits. In this converted system, only one sideband and full carrier are generated, thus permitting this signal to be received on either AM or SSB equipments. At the output of the SSB generator, the phase and amplitude components of the signal are separated, amplified, and then recombined in conventional AM modulator and power stages. If the proper time relationships are maintained between the separated phase and amplitude components when these signals are recombined, the transmitter output will be a high-power duplicate of the full-carrier SSB signal appearing at the output of the SSB generator.

Present AM receivers having a beat-frequency oscillator can detect SSSC signals. Manual tuning of the receiver with the BFO is required and is quite difficult to accomplish. Since this method of tuning is not practical for commercial applications, SSB adapters to convert existing AM receivers to receive the SSSC signal have been developed. Such an SSB adapter, connected to the output of the i-f stage, will provide the circuits necessary to properly demodulate the single-sideband suppressed-carrier signal. If a compatible signal is received (SSB plus full carrier), the only change desirable in the AM receiver is to reduce the i-f bandwidth to agree with the bandwidth of the SSB signal. The AM receiver special circuits will operate the same as they normally operate when receiving an AM double-sideband signal.

During the transition period from AM to SSB systems, standardization of many of the characteristics of the new system will have to be considered. A few of the more important items are: choice of upper or lower sideband for the "desired sideband," frequency stability and accuracy requirements, bandwidth considerations, receiver a-f-c methods, and testing procedures. These and many other items will have to be considered, evaluated, and adapted before the SSB mode of communications can be accepted as standard.

ADVANTAGES AND DISADVANTAGES OF SSB

The advantages of an SSSC system compared



TP7-1213

Figure 14. Basic Filter System SSB Exciter.

to a double-sideband AM system are summarized as follows: reduced frequency spectrum resulting in more available channels; elimination of the high-power carrier, giving better power efficiency of intelligence-bearing sidebands; more durable signal in the presence of certain interference and fading conditions; and reduced size (volume) and weight.

The disadvantages of the same system are summarized as follows: necessity of extremely stable circuits in the transmitter; requirement of complex a-f-c circuits in the receiver; higher cost and circuit complexity resulting from demodulation difficulties encountered because no carrier frequency is transmitted.

FILTER METHOD OF SSB SIGNAL GENERATION

Of the two basic methods of generating single-sideband signals, the most widely used is the filter method, illustrated in figure 14. Although stability and accuracy of the transmitter are determined mainly by the stability and accuracy of the r-f carrier oscillator, the balanced modulator and sideband filter are the circuits most important to the generation of the single-sideband signal itself.

There are many variations of balanced modulator circuits, some making use of diode rectifiers and others utilizing vacuum tubes. No attempt will be made to cover all such circuit variations in this text. However, the basic functions of all these various balanced modulators are identical: namely, the generation of a double-sideband amplitude-modulated signal, and suppression of the r-f carrier. The amount of carrier suppression is determined by the degree of balance between the two modulator legs (a balanced modulator can be compared to a bridge circuit). Thus the output of the balanced modulator illustrated in figure 15 consists of only the upper and lower sideband frequencies of an amplitude-modulated r-f carrier. The audio input frequency will be rejected by normal tuning of the modulator output circuit.

To obtain a single-sideband signal from the balanced modulator double-sideband output, it is necessary to apply the double-sideband signal to a suitable sideband filter. Either sideband can be accepted or rejected by using an appropriate filter to select one sideband and reject the other. As in the case of balanced modulators, there are various types of filters that can be used in single-sideband applications. Simple L-C filters, which operate in the frequency range of approximately 20 to 100 kc, can be adapted for this type of work. Electromechanical filters, operating over the frequency range of approximately 50 to 500 kc, having their best operation around 250 kc, have recently been designed and developed and can also be adapted for single-sideband use.

However, crystal-lattice filters (figure 16) employing quartz crystals as the filter elements are most widely used in existing SSB equipment. Although the best operating point of this type of filter network is around 100 kc, crystal-lattice filters can be designed to operate at much higher frequencies.

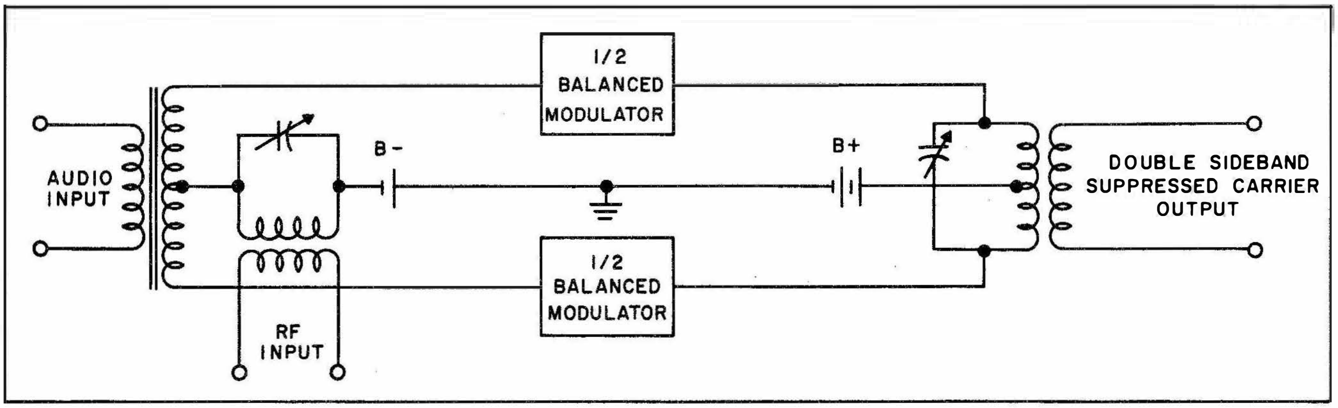


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Figure 15. Simple Balanced Modulator Circuit.

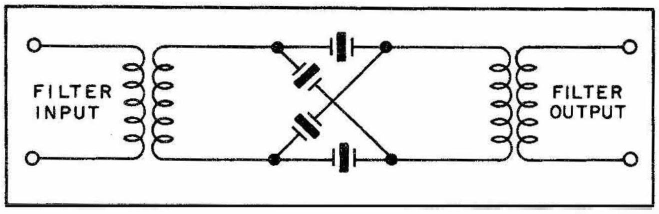

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Figure 16. Simple Crystal-Lattice Filter.

The design requirements of the sideband filter are the primary factors that determine the frequency at which the r-f carrier oscillator operates. Because the percentage of sideband separation is quite small when sidebands are generated at the higher radio frequencies, the requirements of a filter to separate such sidebands are exacting. To ease these exacting requirements, it is common practice to generate the sidebands at a relatively low radio frequency where optimum operation of the sideband filter can be realized.

The most commonly used operating frequency of the r-f carrier oscillator in existing SSB equipment is 100 kc. At this frequency it is relatively easy to construct suitable crystal-lattice sideband filters and highly stable crystal-controlled oscillators. Oscillators having long-time stabilities on the order of 0.2 to 2 parts per million are required in single-sideband applications. If the demodulated signal in the receiver is expected to assume its original position in the spectrum, in regards to the 100-kc carrier, this precise stability must be maintained. Frequency errors over 30 cycles in either the receiver or transmitter oscillators generally will cause voice transmissions to be unintelligible; and variations of only a few cycles will cause distortion in music transmissions. Extremely stable frequency synthesizers, similar to those used in present day AM systems, can also be applied to SSB systems to insure frequency stability of the equipment.

A block diagram of a simplified filter-system SSB transmitter is illustrated in figure 17. Besides being generated at a low frequency, the single-sideband signal is also generated at a low-power level. It is quite evident therefore, that the signal must now be heterodyned to the final transmitter operating frequency and amplified to the desired transmitter output power level.

The circuits following the SSB exciter are in many ways comparable to circuits used in AM systems. However, since frequency multipliers and Class C power amplifiers would impair the frequency relationship between the sideband components, balanced mixer and linear power amplifier circuits are used in the frequency-translating steps in SSB transmitters. Balanced operation of the mixer stages used in these translating steps is not mandatory, unless the frequencies of the two input signals to the mixer circuit are quite close together.

Although only one frequency-translating step is indicated in figure 17, two or more such steps are often employed if the transmitter output frequency is very much higher than the frequency of the r-f carrier oscillator. Regardless of the number of translating steps used, the frequency of the final oscillator is always chosen so as to place the single-sideband output signal in the de-

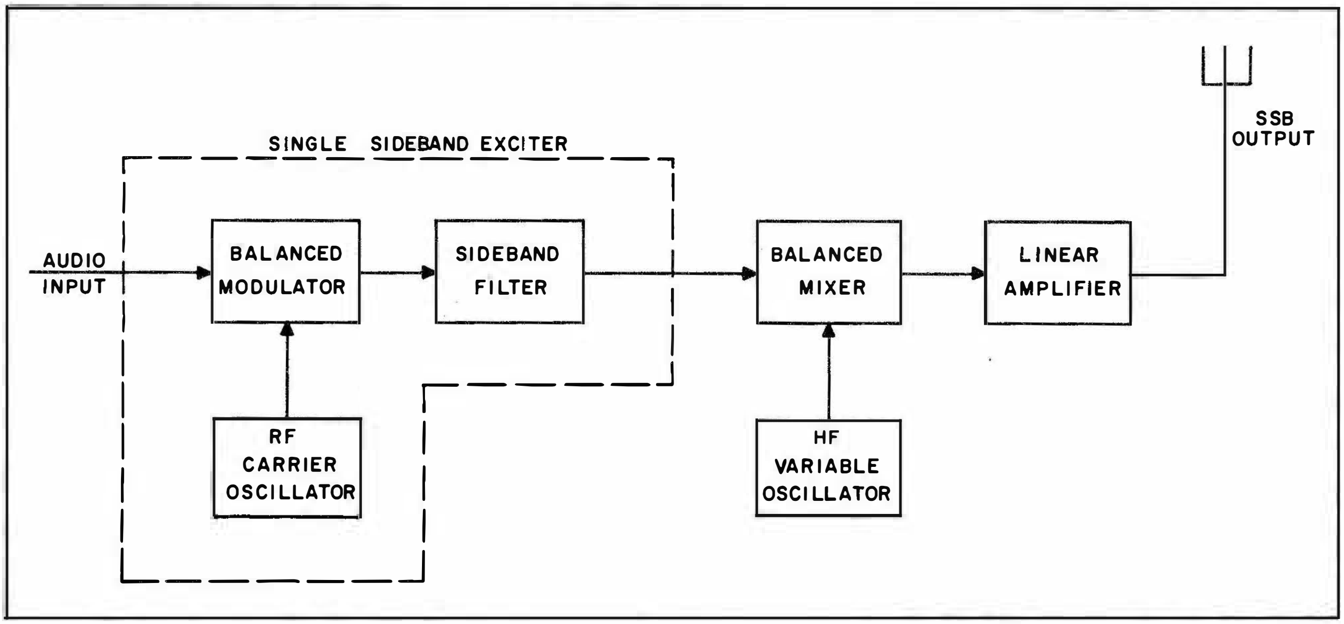

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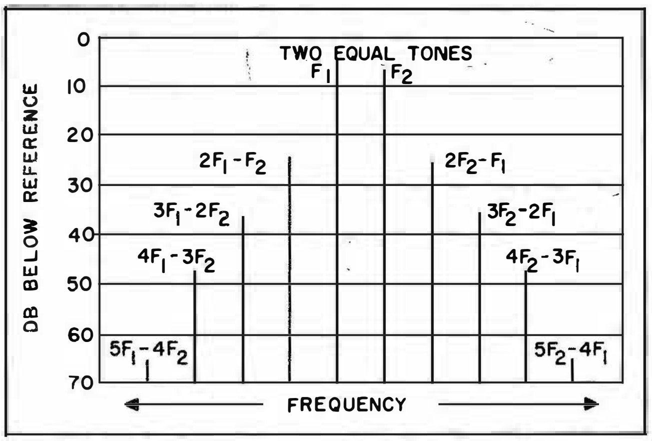
Figure 17. Block Diagram of Filter System SSB Transmitter.



TP7-1217

Figure 18. Spectrum of Intermodulation Components (Splatter) of Two Equal Tones.

sired frequency range for transmission. An important fact to remember about SSB transmitters is that the frequency of the signal radiated from the antenna is not the frequency of the final high-frequency oscillator. Rather, it is the desired sum or difference frequencies (whichever is selected) of this oscillator and the sideband input to the final mixers. The radiated high-frequency carrier as known in AM systems is not present in the radiated signal from an SSB transmitter.

The primary requirements of the output circuits in transmitting systems are usually low distortion (good linearity) and high power gain. Since the modulation process is performed at a low level in SSB transmitters, and since any amplification of a signal containing a modulation envelope must be linear, linear power amplifiers are generally used to satisfy the requirements of the output circuits in SSB system transmitters. Basically a linear amplifier is one in which the output signal is a direct function of, and proportional to, the input signal. However, high gain is not usually obtained with good linearity. Therefore, the output amplifiers in SSB transmitters are generally Class AB or Class B push-pull circuits.

Distortion in SSB system transmitters is mostly associated with the linear amplifiers. The Class B operation, usually at projected cutoff, introduces a slight amount of second-harmonic distortion, but use of the push-pull circuit arrangement reduces this form of distortion to tolerable limits. The most objectionable distortion, therefore, is intermodulation distortion, illustrated in figure 18. This type of distortion is caused by the falling of the odd order products in or near the desired sideband and is more often referred to as "splatter." The amount of this splatter depends on the operational characteristics of the linear amplifiers. Overdriving these amplifiers will cause the splatter to become more objectionable. It is important, therefore, not to overdrive the linear amplifiers.

Linear amplifier circuits designed for r-f applications are comparable to linear amplifiers used in audio systems, the main difference in the two types of circuits being the frequency range over which they are expected to operate.

The advantages of spectrum conservation, elimination of the high-power carrier, and stronger signal in the presence of some types of interference can be realized by any of the methods of generating single-sideband signals. However, individual SSB systems have their own advantages and disadvantages when compared to each other.

The major advantages of the filter systems SSB transmitter are its stability and undesired-sideband suppression. Although the initial design considerations of sideband filters are exacting (a disadvantage), once the filter has been adjusted to its proper operating area it will retain these adjustments for long periods of time. Maintenance requirements on this type of transmitter due to instability are therefore reduced.

Since this sideband filter is designed for a specific band of frequencies, it is easily understood that the suppression of undesired frequencies will be optimum outside this specific frequency range.

The major disadvantages of the filter system are its cost and complex circuitry. Since the modulation process must be performed at a relatively low frequency, heterodyning this frequency to the desired transmitter operating frequency is mandatory. This increases the complexity and cost of the system. The cost is also increased because of the essential design features of the sideband filters. Another disadvantage that must be considered is the difficulty encountered when desiring to select either sideband at will. This can only be accomplished by two methods: switching of the r-f carrier oscillator frequency above or below the existing sideband filter or utilization of two separate filters, one for each sideband, and selection of the appropriate filter for the desired sideband.

PHASE-SHIFT METHOD OF SSB SIGNAL GENERATION

The phase-shift method of generating a single-sideband signal, illustrated in figure 19, removes the undesired sideband by means of a balancing process instead of a filtering process.

In this system an audio input signal is applied to two pairs of balanced modulator circuits through a wide-band audio 90-degree phase-difference network. The r-f carrier signal is applied directly to the first pair of balanced modulators, and to the second pair of modulators through a narrow-band r-f 90-degree phase-shift network. Each of the two input signals is therefore applied to the balanced modulators in quadrature. Since the operation of all balanced modulator circuits is

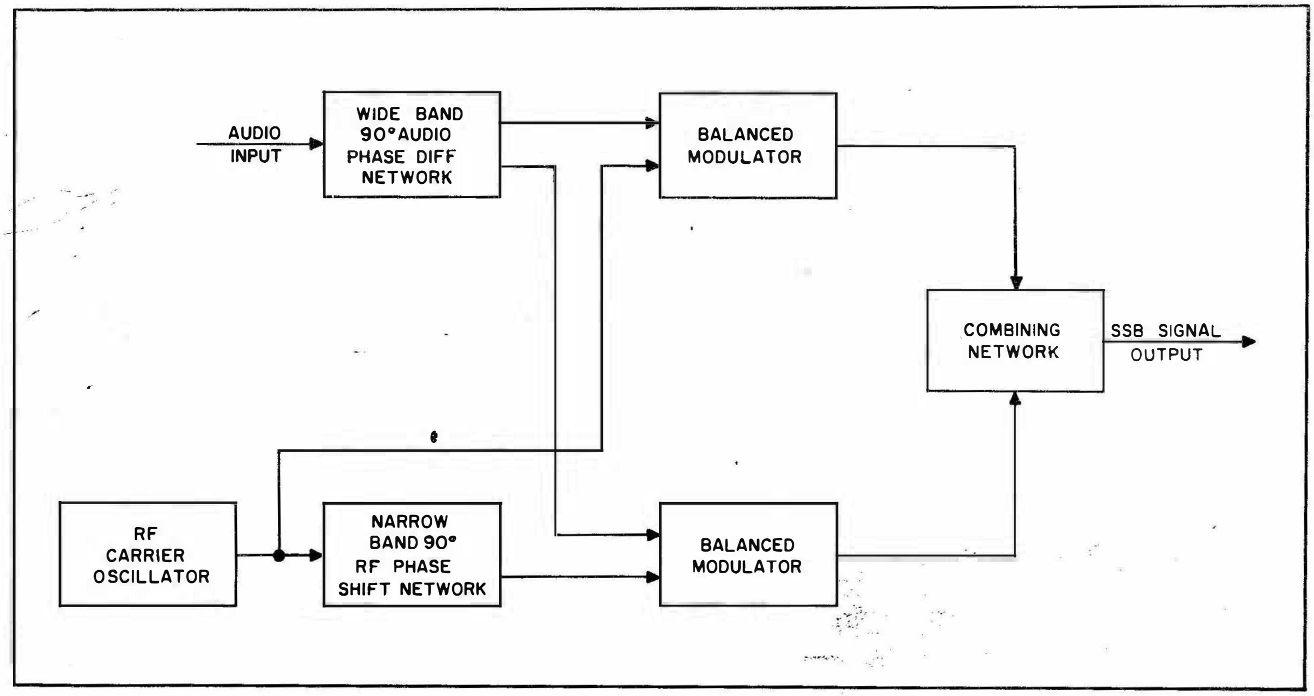


Figure 19. Phase-Shift System SSB Exciter.

basically the same, upper and lower sideband frequencies will be generated and the r-f carrier will be suppressed in these circuits. The double-sideband suppressed-carrier signals thus generated in each balanced modulator will combine in the common plate circuit (combining network). Since the input signals are quadrature-phased, this combining of the sidebands will be in such a manner that one set of sidebands (desired) will add, while the other set of sidebands (undesired) will cancel.

The phase shift system eliminates the necessity of incorporating filters for selection of the proper sideband; therefore, selection of the r-f carrier oscillator operating frequency is not critical. In fact, this oscillator can be operated at the desired transmitter output frequency, an ideal situation for single-channel applications. Unfortunately, however, this advantage is not easily realized for multi-channel equipment. In the latter applications, the narrow-band r-f 90-degree phase-shift network would necessarily have to be made variable in order to cover the same tuning range as that of the variable-frequency oscillator used. Therefore, at least one frequency-translating step is usually employed (following the phase-shift SSB exciter) in multi-channel equipment.

The amount of undesired sideband suppression in the phase-shift system is primarily dependent upon the design and adjustment of the wide-band audio 90-degree phase-difference network. This network must maintain the audio input signal at a constant 90-degree phase difference and amplitude over the entire band of input frequencies to be passed. To satisfy these conditions, audio phase-difference networks have been designed to produce quadrature-phased audio signals with accuracies of less than one degree in phase and one percent in amplitude over the entire audio band.

One major advantage of the phase-shift system is that it is less complex (circuitwise) than the filter system. This is primarily due to the fact that fewer frequency-translating steps (sometimes no steps) are required to raise the r-f carrier oscillator frequency to the desired transmitter output frequency. The cost of this system is less than that of the filter method because of the elimination of the expensive sideband filter and its critical design requirements. Another advantage not to be overlooked is that the selection of either sideband can easily be accomplished by simply reversing the phases of either of the two input signals with a switching device.

Adjustment of the phase-shift system for optimum performance can be quite difficult, however, since the adjustments of the 90-degree phasing networks are highly critical. The system is less stable than one utilizing filters, and maintenance requirements due to instability will therefore be increased if maximum performance is to be consistently achieved.

RECEIVING SINGLE SIDEBAND

General

AM reception requires interception of the desired signal in its carrying medium (normally air) by means of an antenna system, amplification of the desired signal and rejection of as much of the unwanted signals as possible before detection, detection (separation of the modulation from its r-f carrier and the sidebands), and amplification of the resultant audio components to the desired level. If the received signal strength is extremely high, as would be the case when the receiver is in close proximity to the transmitting antenna, most of these functions can be omitted. The detection function, however, must be retained. A diode or crystal rectifier can be satisfactorily used in a simple receiver circuit, which might also include an r-f tuned circuit to improve selectivity. Simple circuits of this type are similar to those used in early radio receivers.

Detection can be accomplished in this manner because of the non-linearity of the rectifier. The sidebands, upper and lower, combine with the carrier in the non-linear device, and the difference frequency (between either sideband and the carrier) appears as the audio component.

Reception of single-sideband signals is not as simple as ordinary AM reception, since detection, which produces the original audio signal, requires the mixing of the sideband (either upper or lower or both) with the carrier, and the carrier is not present (or is reduced so as to render it useless for detection purposes without amplification). Therefore, one of the major differences between SSB receivers and conventional AM receivers is that the SSB receiver incorporates some method of carrier insertion before detection.

The relationship between the inserted carrier frequency and the sideband frequencies must be identical to that between the carrier frequency of the transmitter and the sideband frequencies.

Two methods of accomplishing carrier insertion at the proper frequency are in common use. The first makes use of an extremely stable carrier insertion oscillator, and the second amplifies the pilot, or reduced carrier (if transmitted) to a level sufficiently high for detection purposes. Each of these methods is discussed in more detail in other sections of this manual. A simple SSB receiver is illustrated in block form in figure 20.

Superheterodyne receivers are widely used for AM reception because of the selectivity, high gain, and quality performance inherent with this type of receiver. The same techniques apply to SSB. Multiple conversion considerably improves these features, and is generally used for SSB reception both because filtering may be accomplished more easily at low frequencies and because several steps of translation make possible the reduction of spurious responses by proper choice of heterodyning frequencies.

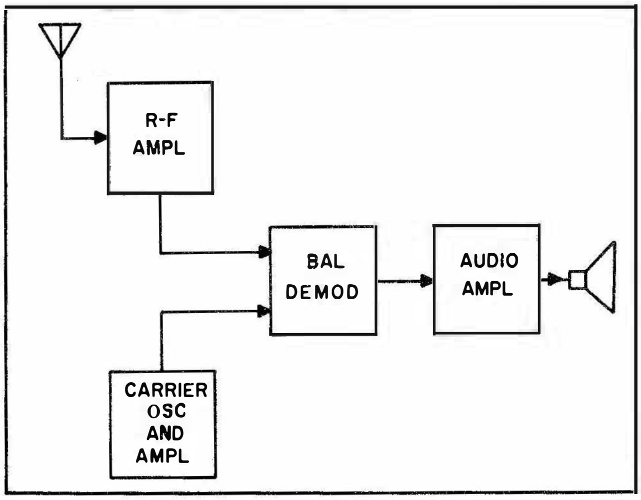


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Figure 20. Simple SSB Receiver.

A typical double-conversion SSB receiver is shown in block form in figure 21. A-F-C and a-v-c circuits are shown; the afc may control either of the heterodyne oscillators, the other of which may be crystal-controlled. These control circuits may be eliminated in an SSSC receiver, or the audio or sideband signal may be detected and used to achieve the desired a-f-c and a-v-c voltages in other ways.

In the following sections, a discussion of the characteristics of these circuits and their concepts will be presented.

R-F and I-F Amplifiers

Single-sideband theory indicates the possibilities of spectrum conservation and narrow bandwidth requirements in signal-handling circuits.

An increase in selectivity, with resultant band narrowing (over normal AM considerations) in r-f amplifiers may render tuning extremely difficult. Results may prove disheartening because of the shape of response curves of simple tuned circuits of the L-C type.

In a double-sideband receiver, or a receiver of the type used to receive double sideband with different information on either sideband (dual channel), the r-f amplifiers and i-f stages preceding the sideband separation filters, must be able to pass both sidebands, as in conventional AM.

Single-sideband receivers, as other AM receivers, attain most of their selectivity in the i-f stages, where tuning is fixed and L-C ratio is held constant.

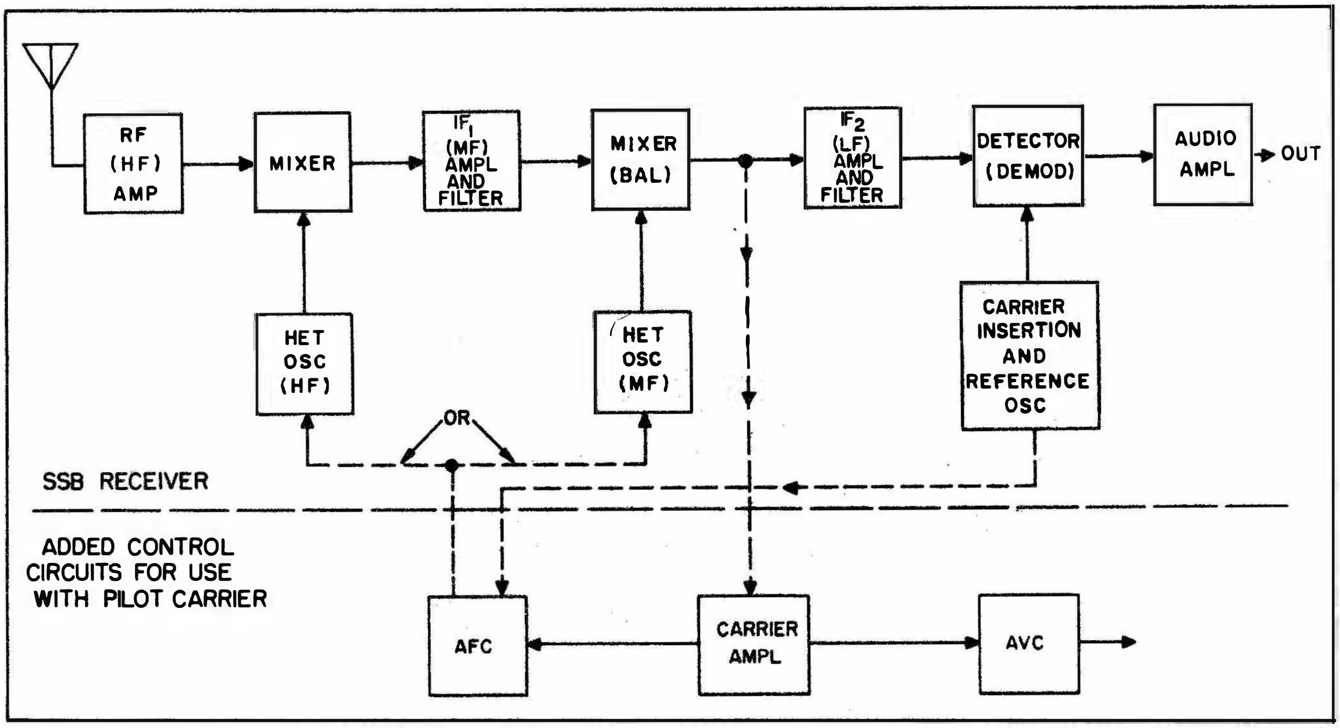

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Figure 21. Double Conversion Superheterodyne SSB Receiver, Showing Added Control Circuits, Using Reduced, or Pilot Carrier.

I-F amplifiers for single sideband make use of filter circuits to limit all but the desired sideband frequency.

Filters and Amplifier Bandpass

The importance of good filter circuits cannot be overlooked in good SSB reception. A narrow bandpass in conventional amplifier circuits alone is not satisfactory because of the selectivity curves attainable. Figure 22A shows typical i-f response curves. To insure that all of the sideband information is passed, a filter circuit must be incorporated. Ideally, the filter curve should be flat-topped with square sides, as in figure 22B. Sharp cutoff at the carrier edge to provide high attenuation of the undesired sideband is the most important factor.

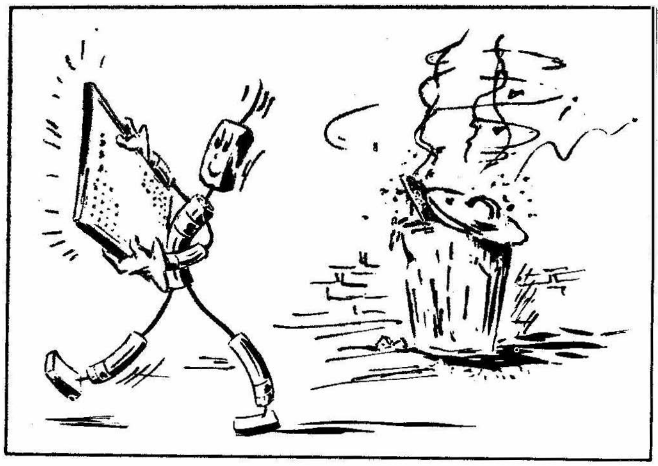
Depending on the number, quality, and type of components used, various approaches to the ideal are obtained. Because of their inherent high Q and stability, crystals have been incorporated in the design of many filter circuits. Special type L-C circuits may be used either alone or in combination with crystals to further increase the effectiveness of the filter. Recently developed mechanical filters have proven practical in the low-frequency (250—350-kc) range.

Typical filter circuits and curves are covered in other sections of this handbook. Filters may be incorporated in receivers to pass either or both sidebands or to pass or reject the carrier. A very narrow bandpass (a few cycles) is used in the carrier filter application.

Unbalanced Detectors

Diode Rectifier

Perhaps the most familiar type of detector is the diode rectifier. A simple unbalanced diode rectifier is shown in figure 23. The rectification process limits all frequencies (wanted and unwanted) present at the detector to half-cycle ex-



". . . the importance of good filters cannot be overlooked——"

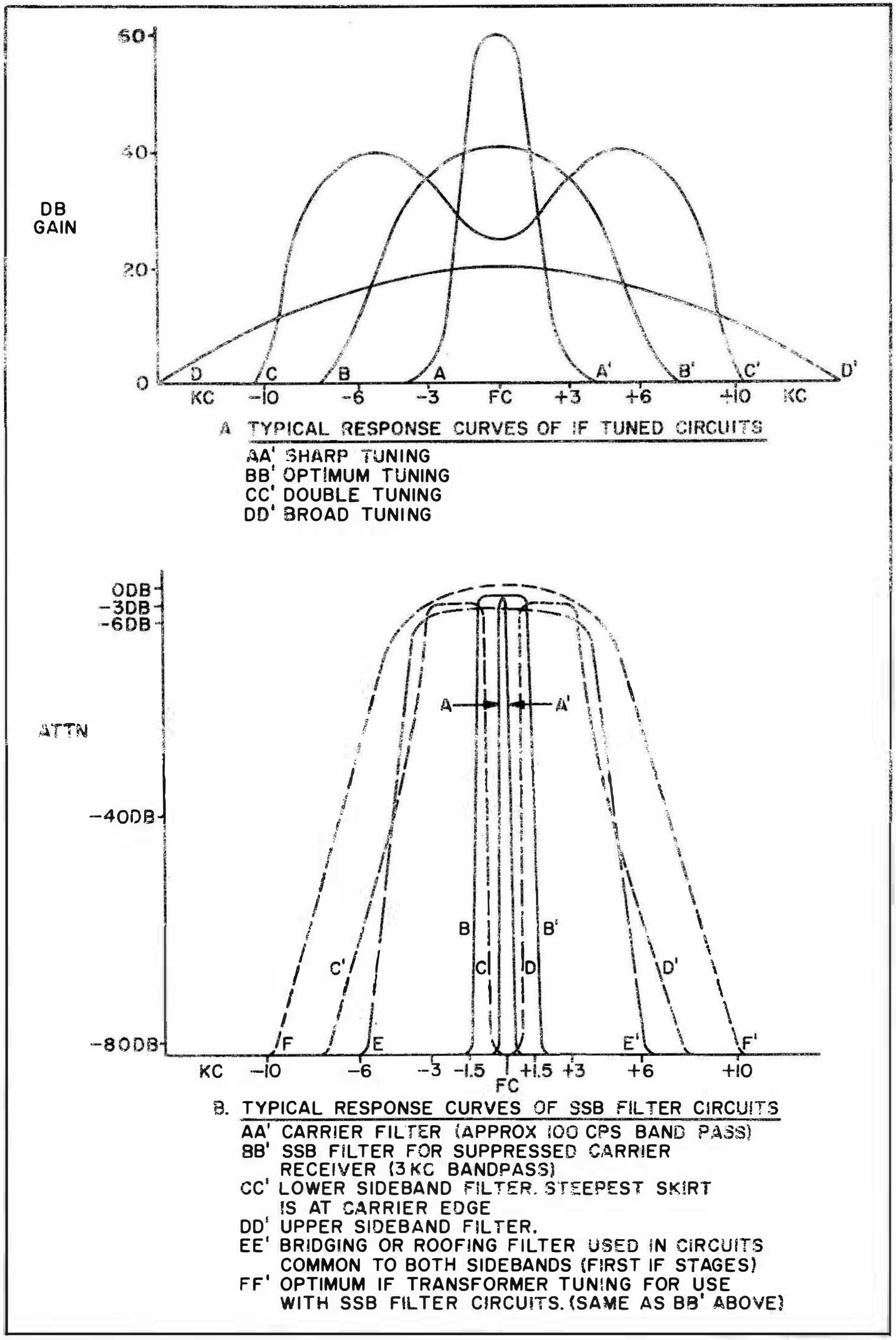


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Figure 22. Typical Response Curves.

cursions. The r-f components are then filtered out with a simple R-C filter. The remaining audio components present at the output are the result of the wave envelope variations. Thus the rectifier type of detector is usually referred to as an "envelope detector." Several other types, such as the plate detector and the grid and slope detector, are also rectifier types of envelope detectors. The major advantages achieved by the use of this form of detection are simplicity, small physical size, relatively few components, and low cost. Disadvantages are the high noise level inherent in rectifiers and the fact that the rectifier does not discriminate between frequencies. Hence, any unwanted signals, carriers, or sidebands present at the detector will be rectified and the resultant beats (intermodulation and cross modulation) will be present in the output.

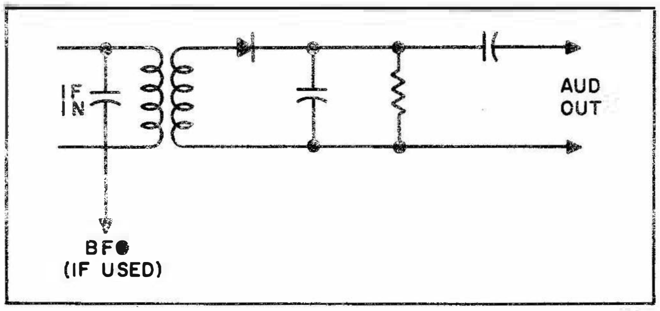

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Figure 23. Unbalanced Diode Detector.

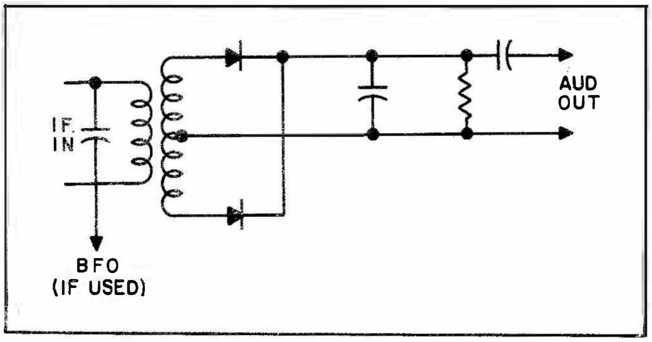

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Figure 24. Unbalanced Diode Detector (Full Wave).

Single sideband can be received using the unbalanced detector, provided that the carrier is added to the sideband before detection. The carrier may be added in any of the stages preceding the detector, or at the detector.

In receivers designed for c-w reception, a bfo is used, and is usually coupled at the input to the detector through a capacitor. This same circuit can be used for single sideband, provided the oscillator signal is kept at a high level with respect to the sideband signal, to prevent overmodulation of the injected carrier. This fact must be remembered in any attempt to use a conventional communications receiver equipped with bfo for c-w reception. Since this type receiver (unless designed for SSB reception) has no provision for control of bfo output level, the incoming carrier level (or r-f gain) must be kept very low. In a receiver designed for SSB, the carrier oscillator level is usually kept at 6 volts and the sideband level at 0.5 volt, corresponding to a ratio of better than ten-to-one.

The unbalanced diode detector may be made to resemble a balanced detector by modification to the full-wave circuit shown in figure 24. The action of this detector is similar to that of the half-wave diode just mentioned, except that alternate half cycles are inverted rather than eliminated. A higher signal-to-noise ratio can be realized from this detector than from the single diode. The action of this detector is similar to that of a full-wave rectifier power supply. Although the full-wave detector resembles a balanced detector in many ways, the two should not be confused.

Mixers

Another unbalanced detector is the mixer, or converter, normally used as a first detector in heterodyne receivers. Although rectification takes place in the plate circuit, the detection is actually a result of the mixing, or heterodyning, of the incoming signal with the local oscillator signal to produce a third, or difference, frequency. This method of detection is based on frequency conversion, therefore, rather than on envelope rectification. However, the circuit normally used operates in a non-linear manner, with some rectification in the plate circuit in order to detect the difference frequency. Both of the input frequencies, as well as the difference frequency, are present in the output. This condition is acceptable provided that the frequencies of the signals are sufficiently far apart to allow a tuned circuit in the output to select only the desired difference frequency, as in r-f to i-f conversion. Since this detector uses a combination of both methods of detection, no advantage would be realized if it were used for audio detection in single-sideband reception.

Balanced Detectors

The balanced detector, or demodulator, is identical to the balanced modulator, except that the frequency translation is reversed (input to output).

One form of a simplified balanced detector circuit using crystal diodes is shown in figure 25. (Tubes may be substituted for the diodes indicated.) As in the balanced modulator, the carrier is applied in such a manner that if no signal is present at the input, the carrier is balanced out, or cancelled, in the output circuit. This cancellation occurs when carrier-signal voltages of the same phase are applied to the diodes in equal

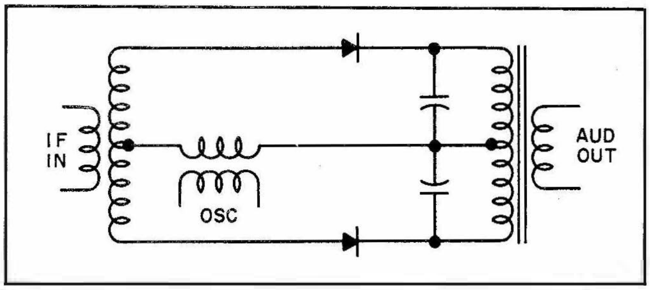

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Figure 25. Balanced Detector (Demodulator).

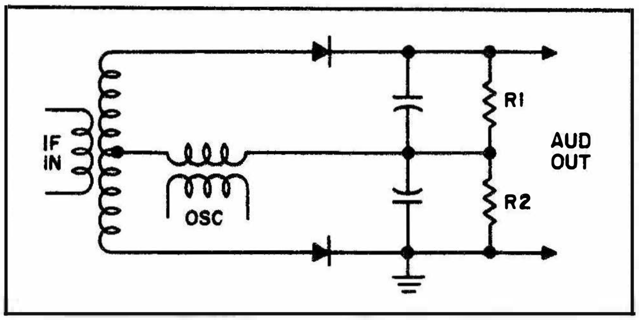

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Figure 26. Balanced Detector, RC Output.

amounts, and the in-phase outputs of the diodes are impressed at opposite ends of the output transformer. When sideband information is present at the input, those portions of the signal that are in phase with the carrier in the secondary of the input transformer add to the carrier voltage and unbalance the detector at the difference frequency. In the absence of either input signal, the output will be zero. The balanced detector may be used for converting r-f to i-f, i-f to audio, or audio to audio. This last type of conversion is encountered in telephone carrier multiplexing applications.

There are as many types of balanced demodulators, or detectors, as there are balanced modulators, each having some advantages over the others. Use of the circuit and availability of circuit components are the major determining factors of the type chosen. A few types are shown in figures 25 through 31. Figure 26 shows a balanced detector with a conventional R-C type of output. Figure 27 shows one connection of a balanced-bridge demodulator, and figure 28 shows the ring or lattice type (also known as a shunt-quad balanced demodulator). Both methods of drawing the ring type (figure 28A and B) are shown for comparison. These types are not normally found in receivers, but are used for audio conversion, as in frequency-division multiplexed telephone circuits, and are desirable for their bilateral characteristics. These are described in detail in the discussion of modulators given in another section of this manual. Figure 29 shows a typical balanced detector using triode tubes with their plates connected in push-pull. This detector, which is one of the most common types found in SSB receivers, is covered in detail under RECEIVER THEORY, in this manual. Figure 30 shows a similar circuit using pentagrid converter tubes, with a variation of input connections. Figure 31 shows a balanced detector using a parallel plate connection. This method eliminates the use of a center-tapped transformer, and R-C coupling may be substituted in the output, eliminating the audio transformer entirely.

The same theory of frequency conversion is applicable to all types. It will be noted that some operate with a push-pull signal input circuit and a push-push (parallel) output, and that others are

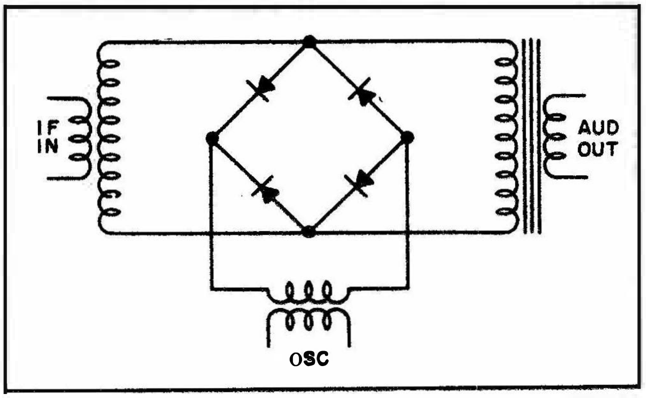

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Figure 27. Balanced Bridge Demodulator.

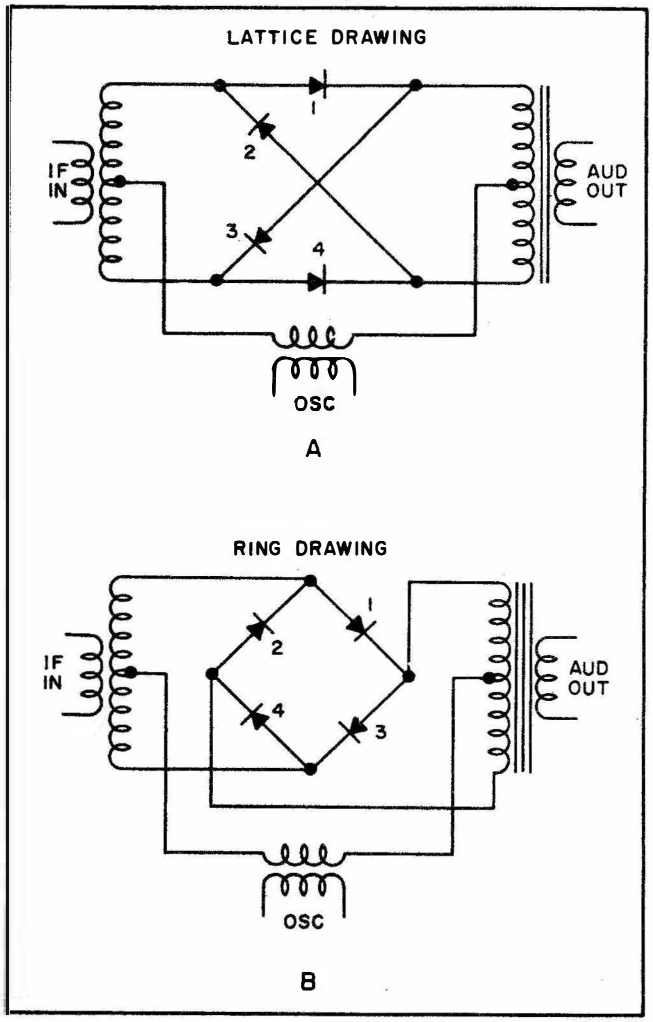

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Figure 28. Ring- or Lattice (Shunt Quad) -Type Balanced Bridge Demodulator.

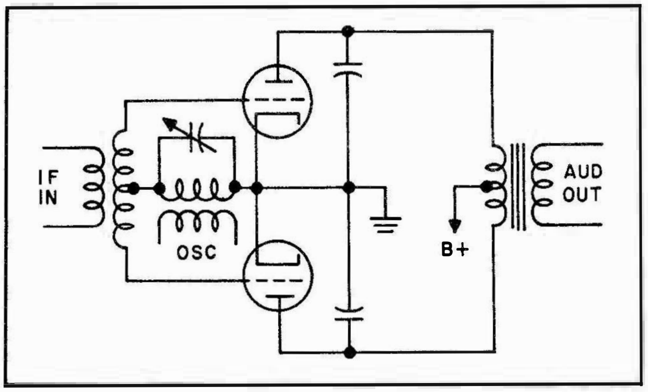

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Figure 29. Triode Balanced Detector with Push-Pull Plates.

reversed. Carrier injection may be push-pull or push-push. A rule to follow is that one of the inputs, either the signal or the carrier, must be push-pull if the output is push-push (parallel), or vice versa, the determining factor being that the input signal which is to be balanced out, or cancelled, is connected in a manner opposite to that of the output. The degree of balance achieved in the over-all circuit determines the amount of suppression, or cancelling, of the undesired components. Thus in the balanced modulator of the SSB transmitter, the degree of balance directly affects the radiated signal, the carrier normally being the suppressed, or cancelled component. In receivers, the degree of balance determines the noise level which results from beats, or heterodynes, of unwanted carriers, and intermodulation products. When the inserted carrier is completely cancelled, the only output will be that which results from the heterodyning of signals with the oscillator or carrier signal.

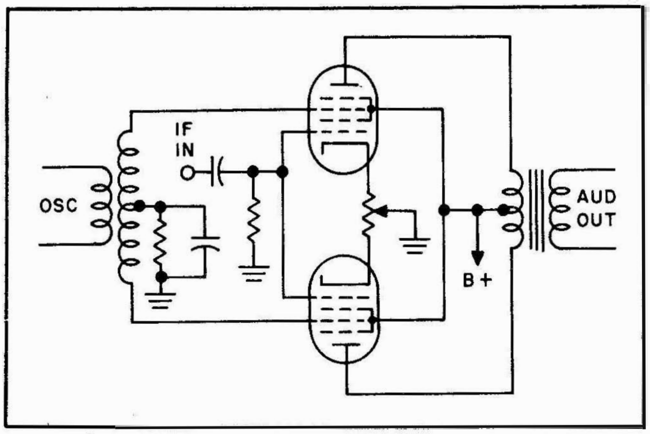

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Figure 30. Pentagrid Balanced Detector.

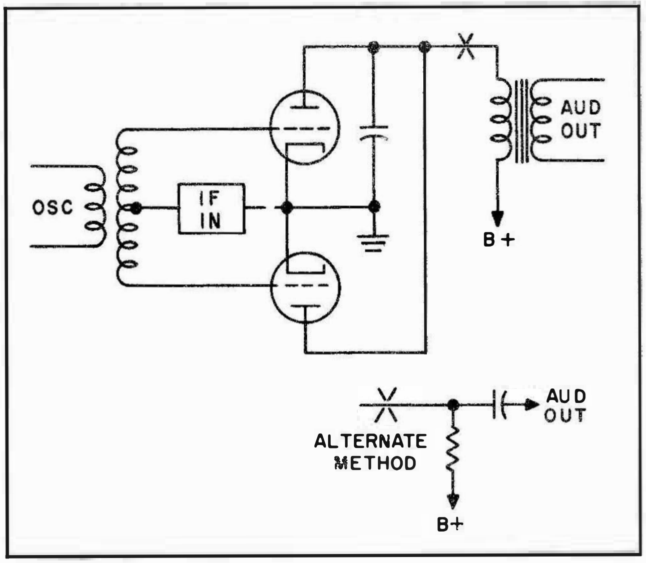

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Figure 31. Triode Balanced Detector with Parallel Plates.

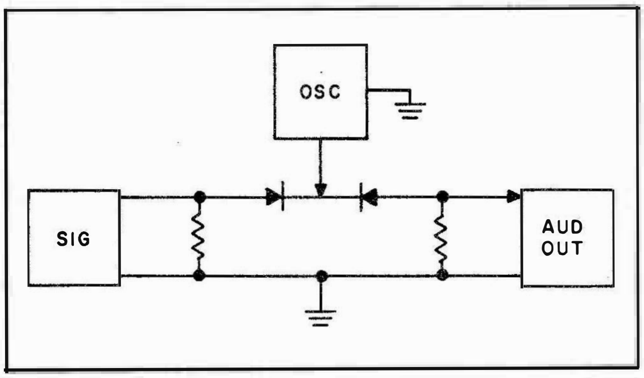

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Figure 32. Coherent Detector.

Several other forms of detectors, have been applied to SSB receivers, especially those possessing characteristics of compatibility with other types of transmission. Two of these are the coherent detector, shown in simplified form in figure 32, and the product detector, shown in simplified form in figure 33.

The coherent, or synchronous, detector utilizes two diodes, which may be either crystals or vacuum tubes. The oscillator, or exalted carrier, signal is applied at the common plate or anode connection. The received signal is applied to one cathode, and the output is taken from the other cathode. For proper action the oscillator and incoming signals must be kept in step and in proper phase (synchronized). When the oscillator signal departs 90 degrees from the carrier phase of the incoming signal, the result is zero output. This detector is often used in phase-locked, or synchronous receivers, and is popular in double-sideband systems.

The product detector is so designed that the output is the mathematical product of the ampli-

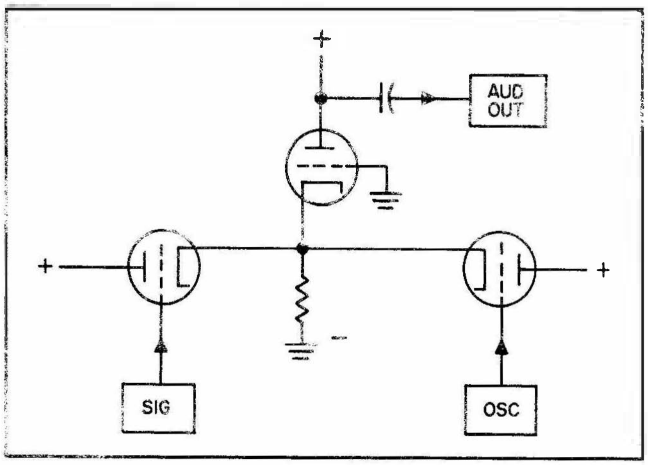

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Figure 33. Product Detector.

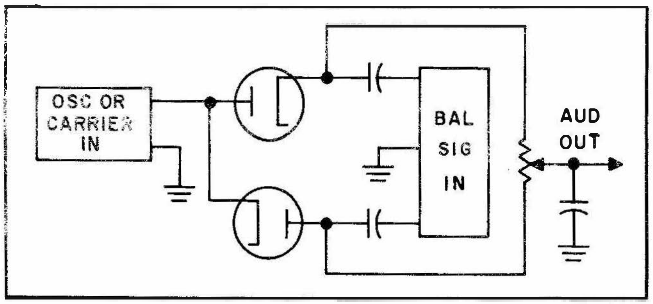

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Figure 34. Peak Conduction Type Detector.

tude of the two inputs. The received signal and the oscillator (carrier) signal are applied to separate grids in such a manner that one signal modulates the other. The output is taken from the common cathodes, which are coupled to a grounded-grid amplifier. The same result may be obtained using a single multi-grid tube. Balance is obtained when both grids are operated in a linear condition. One advantage of the product detector is that the inserted carrier level may be very low without causing distortion.

A peak conduction type of balanced detector is shown in figure 34. This detector is particularly well suited to exalted-carrier reception. A high ratio of carrier-to-signal level is required for proper operation. Carrier and signal inputs may be reversed if proper signal level ratios are maintained. A maximum carrier input level of 200 volts is not uncommon with this type detector.

Oscillators

The most important requirement of oscillators used in single-sideband equipment is frequency stability. This is as true for receiving equipment as it is for the transmitter, where tolerances are usually held very close regardless of the type of transmission. In normal AM reception, with full carrier, drift of the heterodyne oscillator by as much as several kilocycles produces little noticeable effect, except for perhaps a slight decrease in signal amplitude. The frequency of the output (audio) signal will remain the same, because the sidebands still retain the same relationship in frequency to their carrier. In the case of a single-sideband suppressed-carrier signal, no carrier is present, and a local oscillator (carrier insertion oscillator) is used for carrier insertion. This oscillator signal must assume the same frequency relationship to the sideband as was established at the transmitter. Similarly, any oscillator used for frequency conversion, or heterodyning, preceding the carrier insertion oscillator, must not shift in frequency. Such a shift would cause a shift of the intermediate frequency and, consequently, a shift of the sideband with respect to the carrier insertion oscillator. The carrier insertion oscillator signal is often used for detection even when a carrier or pilot carrier, is received. Thus stability is equally important in all SSB receivers.

Frequency stability of a very high degree can be achieved by the use of crystal oscillators for second and third conversion and for carrier insertion. The use of high quality components, good mechanical construction, crystal ovens, and temperature compensation devices will increase the stability of the oscillator.

If the receiver incorporates variable tuning (as is normally the case), then the first conversion (high-frequency) heterodyne oscillator must also be variable, and must track with the incoming signal. Since a crystal oscillator cannot be employed in such a circuit, a highly stable variable frequency oscillator is used. Here again, high quality components and good mechanical construction are essential. A deviation of ten cycles presents noticeable distortion in the resultant audio frequency at the receiver output. At 10 megacycles an error of only one part per million would cause a 10-cycle deviation. It is apparent, therefore, that a good a-f-c circuit is essential.

A-F-C Circuits

Two general types of automatic frequency control may be used in SSB receivers. The first of these is the comparator type, in which the carrier frequency (in the i-f stages) is compared with the frequency of a local reference oscillator (the carrier insertion oscillator). The second type of afc is the phase discriminator type, the operation of which depends on the output voltage of an accurately tuned or crystal-filtered discriminator circuit. An a-c output is provided by the comparator type a-f-c circuit, whereas d-c output is provided by the discriminator circuit.

Two types of heterodyne oscillator control may be used in a-f-c systems. One type is the electronic system, which uses a reactance tube. In this system, the interelectrode capacitance of the reactance tube and, hence, the capacitance in the tank circuit of the heterodyne oscillator is controlled by a voltage applied to the reactance-tube grid. The other type of oscillator control is the electromechanical system. This system uses either an a-c synchronous motor (or servomechanism), or a d-c motor to drive a small variable capacitor (or an inductive device) in the controlled-oscillator tank circuit.

Each of these systems possesses certain advantages, with the reactance tube eliminating some of the disadvantages of the mechanical systems. The comparator type using an a-c motor may be designed to provide frequency control to within one cycle with reference to the carrier insertion oscillator. This control is independent of frequency changes in any of the transmitter or receiver oscillators within the limits of control of the system and the frequency range of the carrier filter circuits. These limits are imposed on any type of afc used in SSB or exalted-carrier receivers. Bulk and weight factors, along with the inability to correct for sudden changes in frequency without some delay, add to the disadvantages of motor systems. This delay may be an advantage in some cases, such as during brief periods when an unwanted carrier close to the operating frequency predominates or overrides the desired carrier, or when noise overrides the received carrier in the system. This delay, however, is a serious disadvantage in high speed aircraft communications, in which the carrier or even the entire transmission may last for only a very brief period and Doppler shift may be introduced. In aircraft and air-to-ground receivers, then, the reactance tube may be preferable, since a faster rate of correction can be achieved with an all-electronic system. The reduction in bulk and weight possible in such a system is of paramount importance in aircraft receivers.

Either a reactance tube or a d-c motor (usually operated in the plate circuit of a control tube) is used with a phase discriminator, since the discriminator output is dc. Phase discriminator circuits are almost exclusively used in exalted-carrier AM receivers, where a separate reference or carrier insertion oscillator is not necessary or not used. In such a system, the received carrier is reinserted at the detector and is always in the proper frequency relationship with its sidebands, regardless of the receiver local oscillator (or heterodyne oscillator) frequency. In exalted-carrier AM reception, therefore, it is only necessary to keep the carrier within the carrier filter frequency limits, which should be rather narrow.

Phase discriminators of this type are usually referenced to a crystal, because the frequency control depends on the tuning of the discriminator and both the discriminator and reactance tube are subject to drift and error. When the system is properly tuned, the output of the discriminator is zero. When the incoming frequency differs from the crystal or discriminator frequency, a d-c voltage is produced at the output. The magnitude of this voltage depends on the amount of error, and the polarity depends on whether the signal is above or below the discriminator center frequency (crystal frequency).

When the carrier inserted for demodulation is the output of the carrier insertion oscillator, it is necessary or desirable to reference the signal (sidebands) to the inserted carrier, because any difference in frequency relationship between oscillators and sidebands will be reflected as a frequency shift (frequency distortion) in the output of the receiver, and a more precise control must be maintained. When a pilot carrier, or carrier of any level, is received with the sidebands, the comparator type a-f-c circuit is generally used.

In one comparator type of system, an output from the carrier insertion or reference oscillator is compared with the filtered and amplified carrier in a balanced modulator. The frequency of the reference oscillator is usually 100 kc, because standard reference oscillators are readily available at this frequency. The same method, however, could apply to any frequency. The output of the balanced modulator is an a-c signal, the frequency of which depends on the frequency difference (error frequency) between the two input signals. The output is applied to an a-c motor or servomechanism, which drives the frequency-correcting capacitor in one of the heterodyne oscillators.

To establish the proper direction of rotation of the motor, as well as to provide a phase shift for a split-phase synchronous motor, two balanced modulators are often used, and a quadrature relationship is established in a 90-degree phase-difference network, through which the oscillator signal is applied to the two balanced modulators. Such a method is similar to the phase-shift method of SSB generation and detection, which is discussed in detail in other sections of this manual.

A comparator-discriminator type of afc may be used with a local reference oscillator in still another combination in which both the received carrier signal (after amplification and limiting) the carrier insertion oscillator output signal are applied to a phase discriminator through appropriate limiters, producing a pulsating d-c voltage at the output. The pulse shape, rate, and polarity of this output depend on the amount and direction of error existing between the compared signals. The output of the phase discriminator is applied through both positive and negative gate circuits to a summing circuit. The average d-c output from the latter circuit may then be applied to either a reactance tube or a d-c motor control circuit.

An advantage obtained by using this type of circuit is that the control does not depend on a separate crystal or on the tuning of the discriminator. Therefore, the tuning of the discriminator is not as critical as in the other circuits mentioned.

Suppressed-carrier systems must depend on highly accurate frequency stability without carrier reference, since no usable carrier is radiated in the transmission of such signals. Crystal oscillators may be used, but the desired accuracy cannot usually be maintained with crystal oscillators alone. Temperature-controlled crystal ovens with carefully chosen crystals in special circuits provide an appreciable reduction in oscillator drift. Slaved oscillators, using crystal frequency synthesis techniques with a common standard reference oscillator of precise frequency, are capable of providing a higher degree of stability. In some suppressed-carrier systems, a constant tone is transmitted, and is used as a reference for a-f-c purposes in the receiver. This method of control is desirable in telegraphy, multiplex, teletype, and data transmission.

A constant tone transmitted on an unused channel in a multiplex system is an aid in the monitoring of speech or broadcast transmissions, because its use makes possible the detection or anticipation of frequency errors. In the case of tone or data transmissions, however, the constant tone is of little or no value because errors are difficult or impossible to detect.

A-V-C and A-G-C Circuits

Various basic a-g-c circuits are used in SSB receivers. In any automatic gain or volume control circuit, some portion of the received signal is rectified and applied as a varying negative d-c voltage to control the gain of one or more amplifiers in the receiver. The gain of the r-f and i-f stages is usually controlled, in order to prevent overloading of the detector stage. A-V-C circuitry is found in most AM receivers, and the operation of such circuits is described in most radio textbooks. However, some variations are often incorporated in SSB receivers. One of the differences is that the filtered carrier is usually rectified and used for a-v-c purposes. The carrier normally has a constant amplitude, regardless of modulation, and thus provides a better reference of signal level. The carrier is conveniently sampled in the SSB receiver, since better detection is possible if the carrier is removed from the sidebands, and in this process the carrier is separately filtered and amplified to be reinserted at the detector. This filtered and amplified carrier is also rectified to provide the a-v-c voltage. In some applications, however, use of the carrier is not practicable or desirable. For example, in suppressed-carrier systems, the carrier is not transmitted and some other form of a-v-c provision must therefore be incorporated. In this case, the audio or sidebands may be rectified by means of circuitry identical to that used for carrier avc. Use of this audio or sideband avc is often more effective than carrier avc in the reception of code, telegraph, or data transmission signals. Sometimes in multiplexed transmissions a constant tone is transmitted over one predetermined channel, and this is filtered from the detected audio, or output signals, and used for both a-v-c and a-f-c purposes. In many receivers, switching is incorporated to provide for any type, or a combination of all types, of a-v-c operation. Such selective switching is desirable when various carrier conditions such as fading or temporary loss of carrier are encountered, or when carrier and sidebands are fading independently of each other.

Long time constants are usually provided in the a-v-c circuits of SSB receivers. This provision is essential when detected audio is used, because otherwise the a-v-c voltage would tend to follow the syllabic variations of speech modulation. Even longer time constants are required in the reception of controlled-carrier SSB, because the carrier level is increased to the peak sideband level during pauses between words and syllables in speech transmission, and is reduced to a very low level during actual modulation. In receivers for this type of transmission, both the avc and the afc must be slow-acting, or contain a memory circuit, in order to hold the proper values during modulation periods. Consequently, time constants of from 8 to 30 seconds are not uncommon in these receivers. Selectable time constants are used in many SSB receivers to provide suitable a-v-c action under varying conditions.

In some receivers, forward-acting avc is incorporated, in addition to conventional avc, to provide compensation for fading conditions. This forward-acting avc is applied to stages subsequent to the point of a-v-c takeoff in the receiver, and therefore is usually found only in carrier-operated a-v-c circuits. Carrier voltage for a-v-c purposes is usually obtained at the output of the second mixer stage in double-conversion receivers using this type of avc.

The pilot carrier is usually transmitted at a lower level than the peak sideband signal. In order to be usable for a-v-c purposes, this carrier must be given additional amplification. If all signals were amplified equally, the pilot carrier would be increased in voltage by only a slight amount, whereas a full carrier would be increased by a large amount, because the signal level is larger

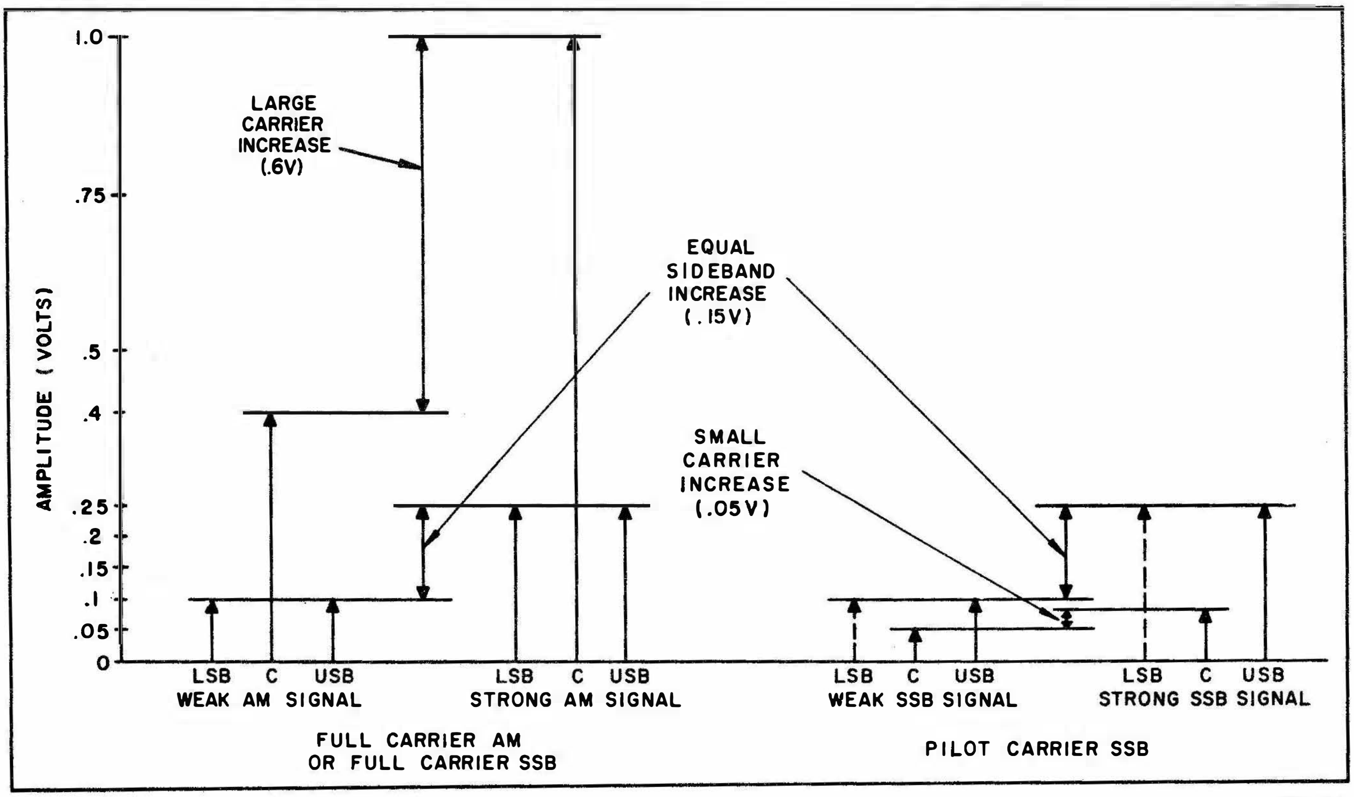

Figure 35. Comparison of Varying Signal Strength in AM and SSB Receivers, Showing Ratio of Change in Carrier Amplitude with Equal Sideband Levels.

at the input. See figure 35. Weak and strong signals may be compared in the same manner. Thus for the same a-v-c voltage as required in both cases, the carrier must be amplified separately, or the a-v-c voltage would have to be amplified after detection to produce the same result. Since the carrier must be amplified before being used for detection, this serves the combined purpose.

When long-time-constant circuits are used in diode detector a-v-c system, proper level of charge may not be realized quickly with sudden increases in signal level. This is especially true when several a-v-c circuits are combined, adding to the number of capacitors to be charged. Triodes, or other grid type tubes, are often operated in conjunction with a negative supply voltage as grid-driven detectors in long-time-constant a-v-c systems. This usually provides rapid charge-slow discharge avc, and is helpful in counteracting sudden increases in signal strength. A combination of these features will be found in the section of this manual which deals with a detailed analysis of the receiver.

EXALTED-CARRIER RECEIVERS

In a conventional AM transmission containing a full carrier, fading conditions caused by multipath reception may result in a carrier too weak for satisfactory detection. Insufficient carrier level at the detector, or demodulator, causes overmodulation and distortion, in the same manner as at the transmitter.

The exalted-carrier receiver is a receiver which insures adequate carrier level by amplifying the carrier more than the sidebands. It does this by separating the carrier from the sidebands before detection, filtering the carrier and amplifying it to the required level, and recombining the amplified signal with the sidebands either at the detector or at a stage preceding the detector. Exalted-carrier receivers may not necessarily be SSB receivers, because exalted-carrier methods are generally used for conventional, full-carrier AM reception in long-distance systems. These receivers may not have sideband filters or separate sideband channels. In many receivers, provision is made for both types of transmission, and with a few additional circuits or modifications, all types of AM transmissions may be received.

Single-sideband receivers, on the other hand, are almost always a type of exalted-carrier receiver, the only exception being those receivers designed for suppressed-carrier reception. The latter receivers depend on a precise frequency carrier insertion oscillator for detecting purposes, rather than on the carrier itself. Reduced-carrier, or pilot-carrier SSB receivers use exalted-carrier circuits and methods because the carrier is always at a low level, regardless of fading conditions.

Both type receivers may be easily converted or modified to provide for either type of reception. Adapters are often used with existing receivers, where space is available. Switching circuits and combining circuits provide additional flexibility.

DOUBLE-SIDEBAND SYSTEMS

Much attention has been given to the use of only one sideband. It has been reasoned that considerable gain can be realized by concentrating all the power of a transmitter in one sideband, since the carrier alone carries no information, and the lower sideband is the mirror image of the upper sideband. Spectrum conservation has been another point in favor of single-sideband transmission.

One of the arguments opposing the use of single sideband is the phase shift inherent in SSB, due to the loss of the opposing phase of the other sideband (see figure 36A). This shift can be minimized to a certain extent by maintaining a sufficiently high level of carrier insertion at the receiver, as illustrated in figure 36B. The phase shift has little noticeable effect on voice communication. Pulse and data transmission may be seriously affected, however, because square waves and pulses may be sufficiently altered in shape by the various resultant phase shifts of the many frequency components of this type of waveform to be detrimental in some applications.

Selective fading has prompted arguments in favor of double-sideband systems over SSB, and the possiblity that jamming or interference may affect one sideband without seriously affecting

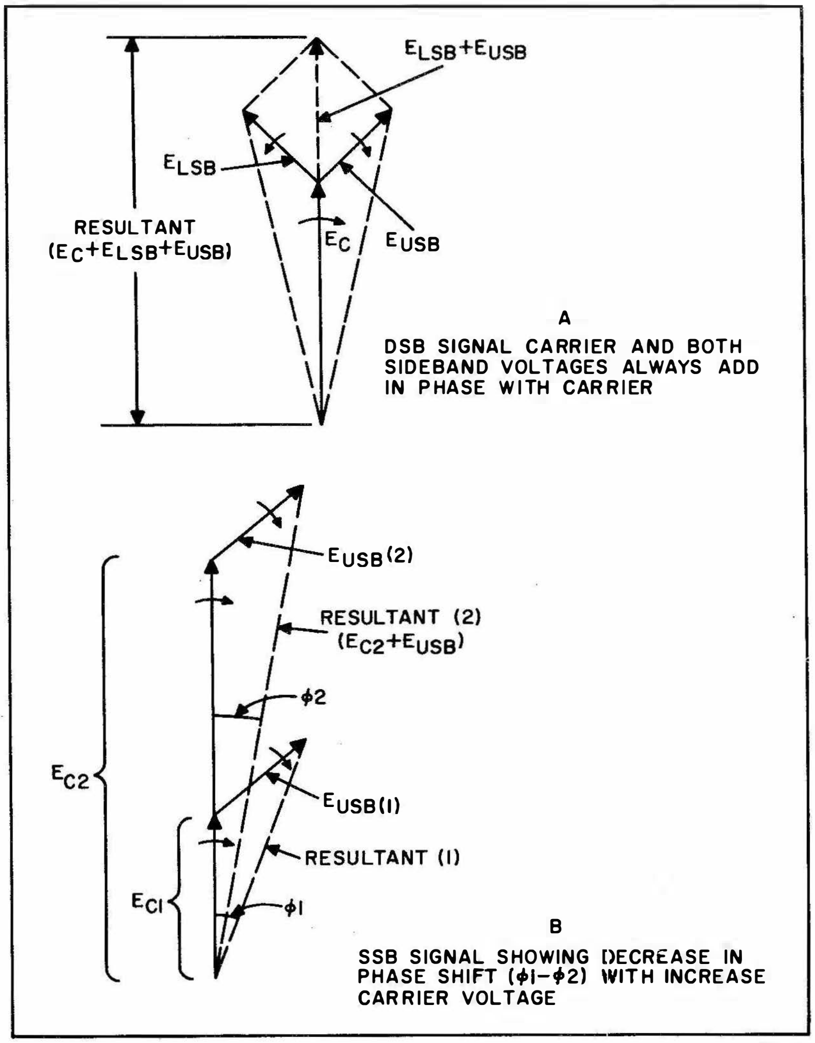

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Figure 36. Phase Shift Due to Loss of Opposite Sideband, Showing Effect of Increased Carrier Insertion Voltage.

the other has also been given considerable thought.

Since the carrier is the most offensive component in causing whistles or beats in the reception of AM transmissions, it has been reasoned that simply removing or eliminating the carrier, while still transmitting both sidebands, might provide the best system. Such a system would eliminate heterodynes, or beating of carriers, or carrier and sidebands, in the receiver, and would result in a power saving, or gain, in the transmitter, while at the same time counteracting the shortcomings of single-sideband systems.

Double-sideband suppressed-carrier transmission is easily accomplished at the transmitter by simply using the balanced modulator or its equivalent circuit without sideband filters (such a transmitter is discussed in the TRANSMITTER THEORY section of this manual). The double-sideband suppressed-carrier receiver, on the other hand, differs quite radically from both the SSB receiver and the full-carrier AM receiver.

In a DSB receiver, since no carrier is received, a locally generated carrier is combined with the sideband signals, as in an SSB receiver. Thus fading of the carrier, or phase differences caused by multi-path effects on the carrier, are eliminated. Frequency stability, therefore, is of great importance in the local oscillator, or carrier insertion oscillator, in the DSB receiver, just as it is in the SSB receiver.

More important, however, is the requirement that the carrier must be inserted in the same phase relationship to the sidebands as in the original modulation at the transmitter. Since the carrier heterodynes with both sidebands, any phase shift between the carrier and one sideband would be accompanied by a like phase shift of opposite polarity between the carrier and the other sideband, and partial or complete cancellation (depending on the phase angles) of the two difference (audio) frequencies would result. Therefore, a phase-locked oscillator must be used for carrier insertion. This feature is not required for SSB reception, because with one sideband absent, no opposing phase shift is developed.

Some phase shift between the two sidebands may be introduced by multi-path reception. However, it is reasoned that since the two sidebands would have to be 180 degrees apart to cause complete cancellation, this condition would be less likely, or at least would present less of a disadvantage, than in full-carrier AM, where the condition would be combined with the phase differences between either sideband and the carrier, all other disadvantages of the carrier transmission being taken into consideration.

Another advantage seen in DSB transmission is the possibility of greater reliability of reception under varying conditions of fading. Under such conditions, one sideband may be phased out by multi-path effect, while the other sideband may be received in some degree of amplitude. (This selective fading of the sidebands, as well as the phase difference between the lower sideband components and the upper sideband components, is caused mainly by the difference in frequency between the two sidebands.)

Therefore, it is probable that one sideband can be received during the time that the other sideband is unsuitable. Also, in the case of interference or jamming one of the sidebands may be usable. These conditions may be corrected manually in some single-sideband receivers, if sideband selective switching is incorporated in both transmitter and receiver.

SYNCHRONOUS COMMUNICATIONS

One method of DSB reception involves the use of a phase-locked oscillator and a coherent detector. A basic receiver of this type is shown in figure 37. Audio output is produced directly by mixing the incoming signal with the output of a coherent local oscillator in the detector, and the audio output is then filtered to eliminate all but the desired signal. Such a receiver possesses many advantages, since no i-f system is needed.

Detection may be accomplished at very low levels and the resulting signal amplified at the audio frequencies. Filtering at the audio frequencies permits much greater selectivity than is possible at i-f frequencies.

Full-carrier AM signals may be received with this type of receiver, although the carrier has no effect on the detection, since it simply is not required or used.

Although the system just described is not an SSB system, some information has been included here, because of the similarity of some of the circuitry to SSB circuitry and because of the fact that some consideration has been given to its use in lieu of SSB equipment.

NEW DEVELOPMENTS IN SSB

Stabilized Master Oscillator (SMO)

In the field of aeronautical communications the effect of Doppler shift becomes more of a problem as the speeds of aircraft are increased. The Doppler shift is caused by the movement of the aircraft toward or away from the station with which it is communicating. The effect of this shift is to change the frequency approximately one cycle per megacycle per Mach number, or one part per million at approximately 670 mph. For example, in a transmission on a frequency of 30 mc from an aircraft flying at this speed, the frequency error due to the Doppler shift will be 30 cycles. If the aircraft is flying directly toward the sta-

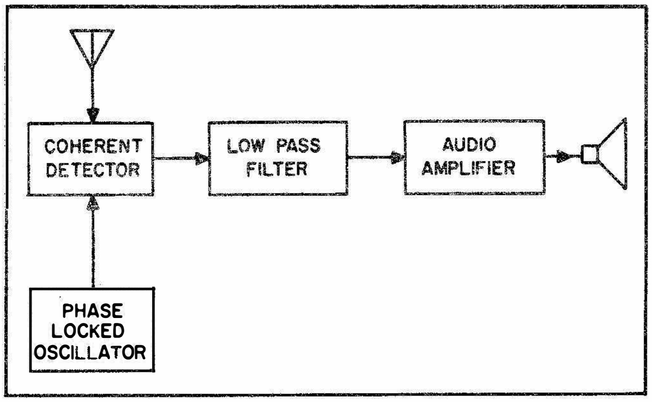

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Figure 37. A Basic DSB Receiver.

tion, the frequency will be increased by this amount; if it is flying directly away from the station, the frequency will be decreased by the same amount. As the speeds of aircraft increase and the operating frequencies increase, the frequency errors produced by the Doppler shift alone may be sufficient to affect the intelligibility of the signal. It is quite apparent, therefore, that the frequency errors caused by the receiver and transmitter units used in such applications must be exceedingly small (on the order of approximately .00005 percent) to keep the over-all system error within tolerable limits.

One method of satisfying the exacting frequency stabilty requirements of such equipment is to use a stabilized master oscillator. Such a circuit uses a frequency synthesizer to provide extremely accurate reference signals to "lock-in" a variable-frequency master oscillator. Besides possessing extreme accuracy, this type of circuit will also provide a multiple number of stable channel frequencies. Since the frequency-synthesis process employed in this SMO is similar to that employed in some existing equipments, only the method of controlling the stability of the reference oscillator will be described.

A block diagram of a typical standard reference frequency oscillator is illustrated in figure 38. Basically, the stabilized master oscillator consists of two sections—the electronic-controlled oven circuits and the oscillator circuits. The oven is required for the resonator circuits because the stability of the over-all oscillator circuit depends upon the stability of the components that form the resonator portion of the oscillator. This electronic-controlled oven (illustrated in figure 39) consists of a temperature-sensitive resistance bridge acting as a heater, and a four-stage audio amplifier. When the bridge network is in an unbalanced state, a feedback signal will be applied to the input of the audio-amplifier. The output of this amplifier will in turn be applied back to the bridge network. The bridge will assume a balanced state when its resistance, due to heating,

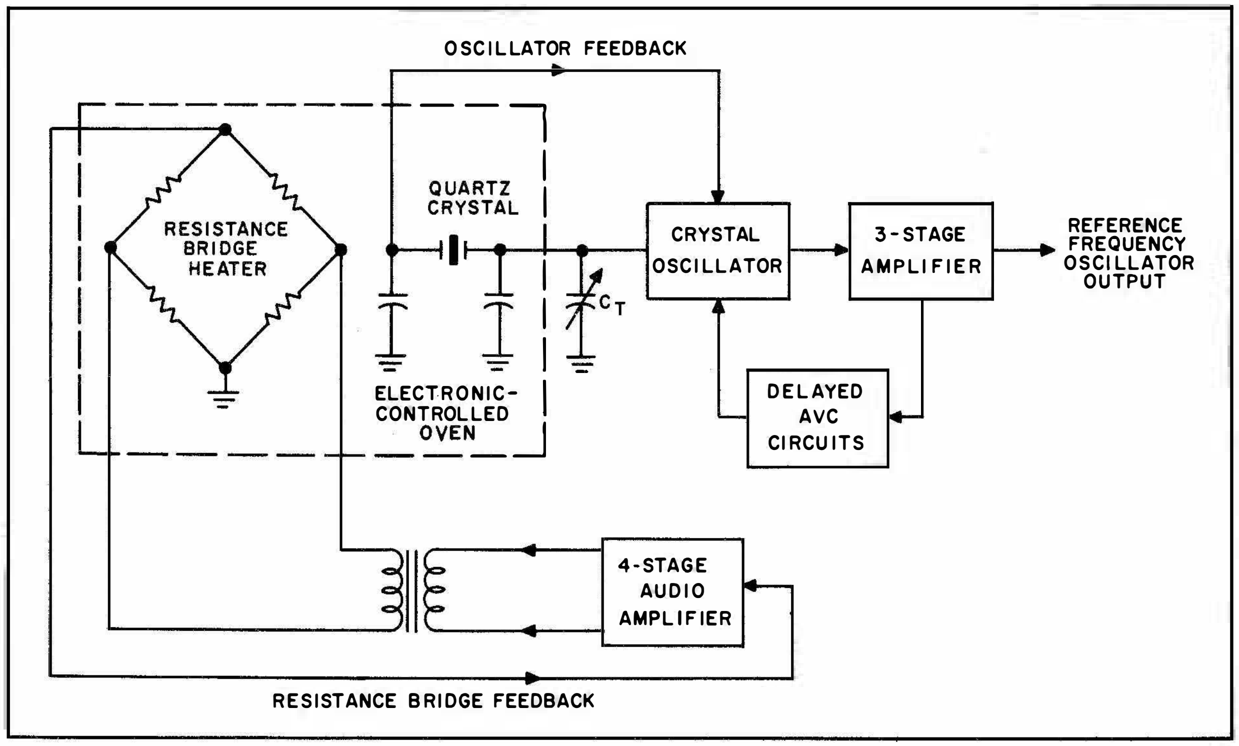

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Figure 38. Block Diagram of Standard Reference Frequency Oscillator.

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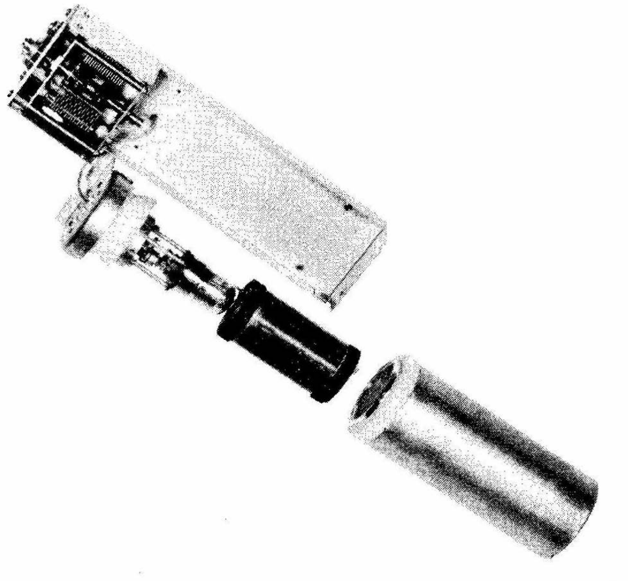
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Figure 39. Electronically Controlled Oven Used in a Ground Station Reference Frequency Oscillator.

equals the gain of the amplifier. At this point stable conduction will be produced in the amplifier, and the oven will be maintained at a steady temperature sufficient to provide long-term frequency stabilty of the quartz-crystal resonator.

The oscillator circuit used in this arrangement is a Pierce type amplitude-controlled crystal oscillator with the quartz crystal sealed inside the oven. Minor adjustments of the oscillator frequency can be accomplished by adjusting the trimmer capacitor (C_t) located outside the oven. The amplitude of the oscillations is controlled by rectifying a portion of the amplified oscillator output signal, and applying the rectified signal back to the control-grid of the oscillator through a delayed a-v-c type of circuit. The reason for this amplitude control of the oscillator is to prevent excessive power dissipation in the crystal resonator, and to reduce harmonic generation in the oscillator circuit.

Photographs of standard reference frequency oscillators for ground-station and airborne applications are shown in figures 40 and 41, respectively. The circuit operation of the airborne oscillator is basically the same as that of the ground-station unit just discussed. However, in order to decrease the size of the airborne oscillator, a slightly higher oscillator frequency is used, and the oven circuits are transistorized. The size of this unit is only 4½ x 4½ x 3 inches, yet it has essentially the same stability as the larger ground-station oscillator unit.

SSB Applied to VHF Scatter Systems

In VHF ionospheric scatter systems (trans-horizon propagation), a major problem commonly encountered is multi-path fading. Because of these fading effects, and because v-h-f scatter

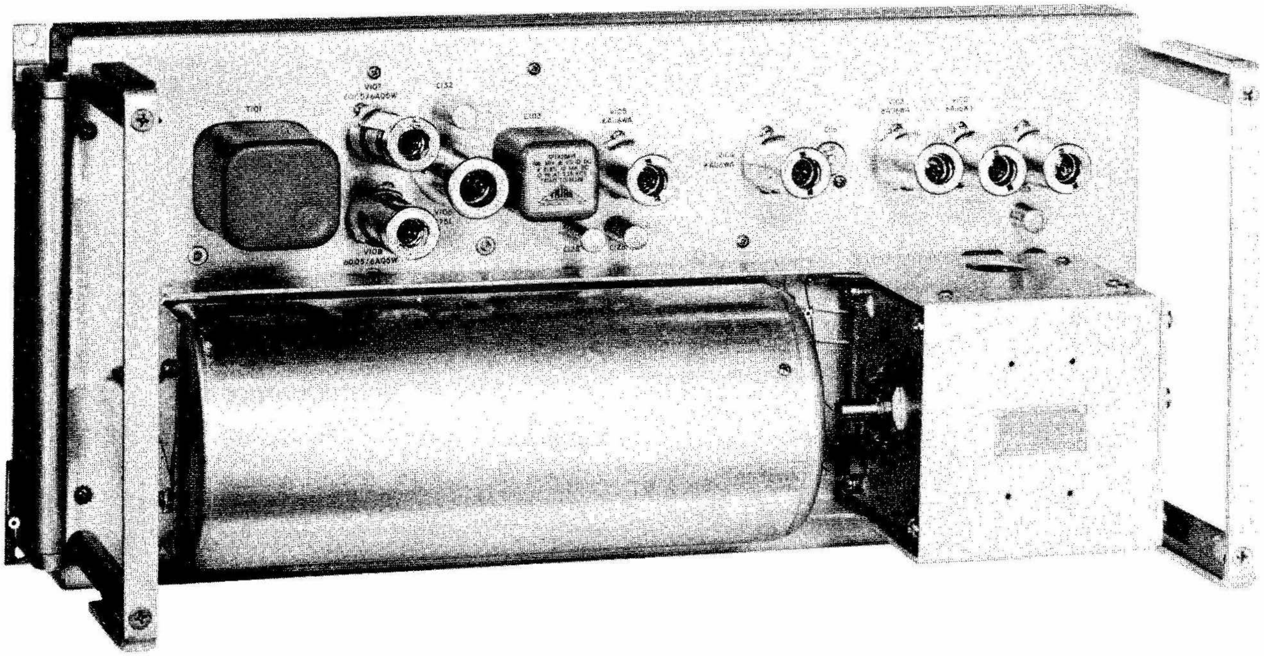
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Figure 40. Ground Station Reference Frequency Oscillator.

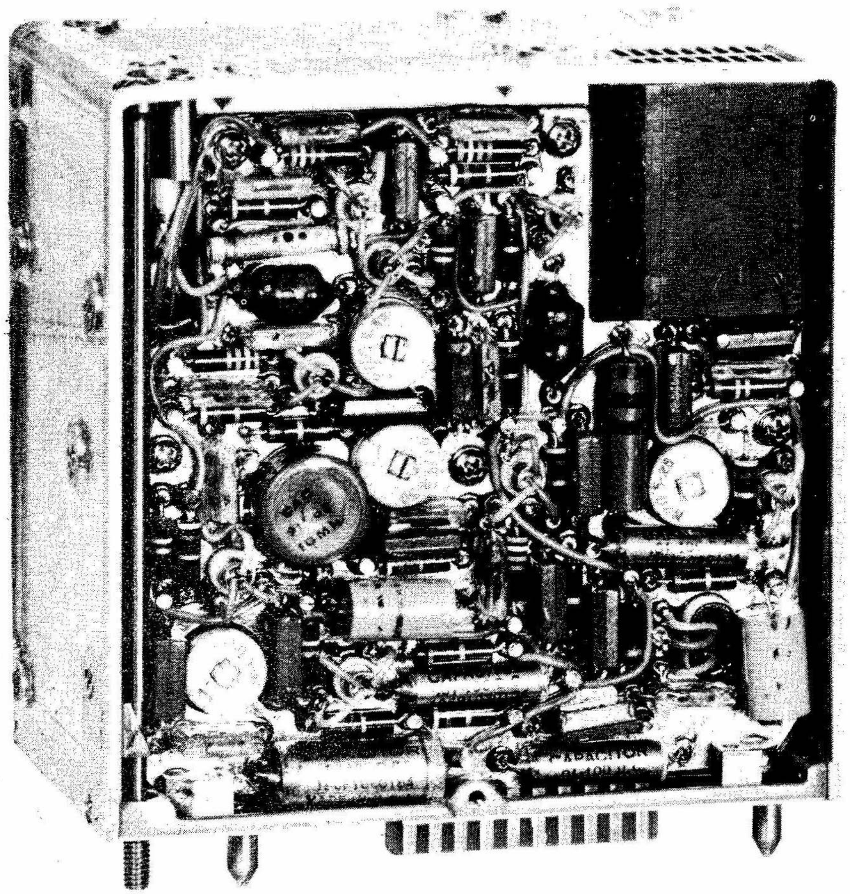

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Figure 41. Airborne Reference Frequency Oscillator.

systems operate best in the already overcrowded 25—60 mc spectrum range, broad-band modulation systems are not well adapted for this method of communications. However, single-sideband and narrow-band frequency modulation can both be used in VHF scatter systems, since both of these modulation methods have certain advantages when used in such applications.

Although narrow-band FM uses a wider bandwidth than SSB, it has the advantages of simplicity and relative insensitivity to rapid fading when the received signal exceeds a certain minimum level. The advantages of SSB (compared to FM) that make it suitable for use in VHF scatter systems are narrower bandwidth, lower average power requirements, and no minimum "threshold" level. Reception with single sideband can be obtained at much lower signal levels than those required when using frequency modulation.

The major problem encountered when using SSB voice communications in the VHF range is frequency stability. This stability problem becomes even more critical in the transmission of synchronous multiplexed-teletype information. In such applications it is desirable to use the same oscillators for time references as well as for frequency generators, instead of transmitting a separate continuous synchronizing signal. Stabilized master oscillators are best suited to meet the strict frequency stability requirements of such SSB transmitters. Good linearity of the power amplifiers, also a requirement in single-sideband VHF scatter systems, can be obtained by using r-f feedback.

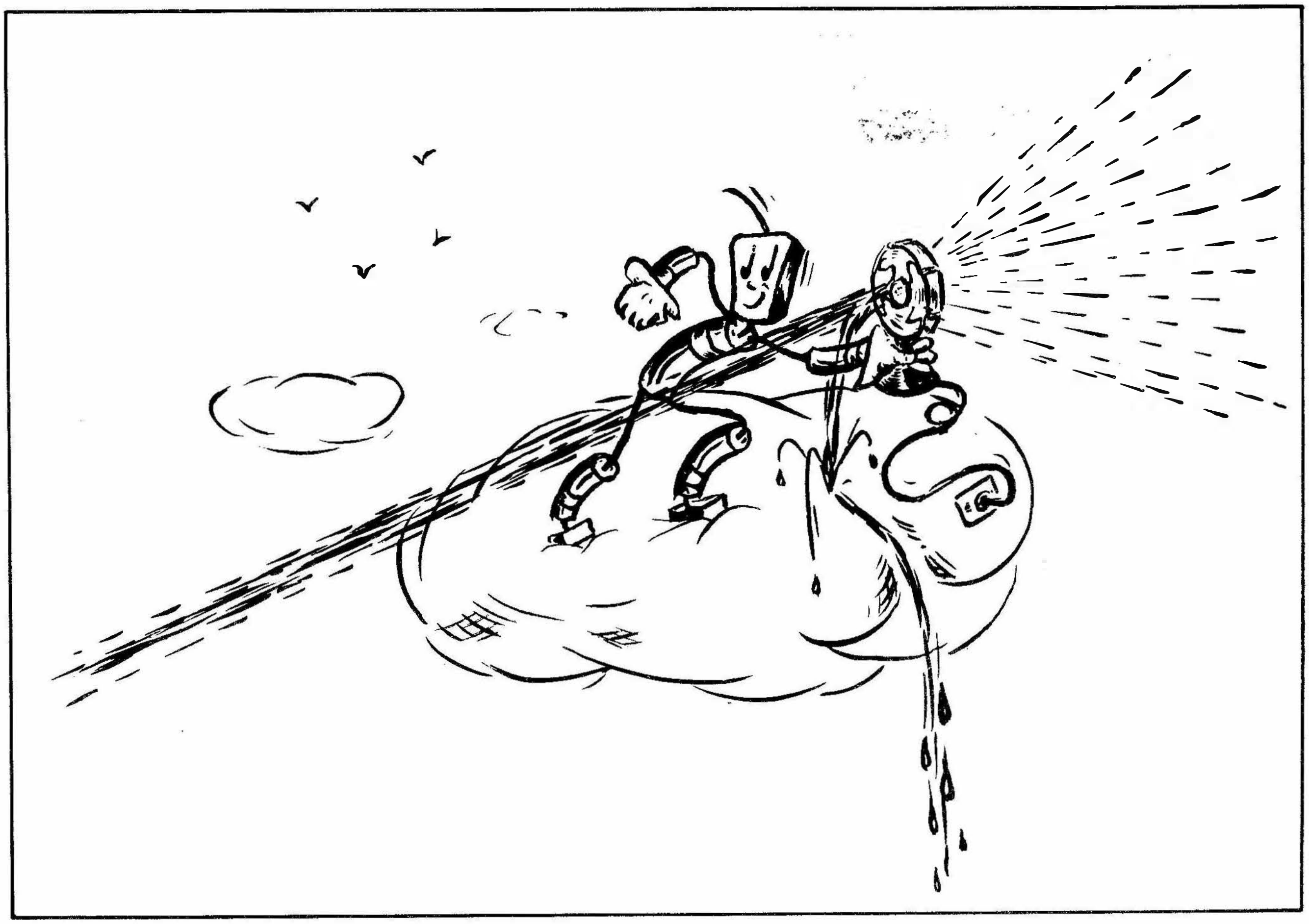
SSB Applied to UHF Scatter Systems

The principles of scatter proprogation applied to VHF communication systems can also be applied to similar UHF systems. Whereas the lower region of the ionosphere is used to scatter the propagated wave in the VHF range from 25—60 mc, the region of the atmosphere below the ionosphere—called the troposphere—is used for the same purpose in some UHF systems operating above 100 mc. Although frequency modulation has been widely used up to the present time, single sideband has many advantages (over FM) that make it readily adaptable for these UHF systems.

One advantage in using SSB is the reduction in the spectrum in terms of the bandwidth of one voice channel. Depending on the tolerable intermodulation, FM systems use bandwidths of approximately 24 to 40 kc for each voice channel. On the other hand, SSB systems require only a 4-kc bandwidth, including the guard band of the channel.

As in the v-h-f ionospheric scatter systems, the effects of multi-path fading are also present in long-range UHF tropospheric scatter systems. Therefore, another advantage of SSB is a reduction in multi-path fading effects, thereby allowing for an increase in bandwidth for multi-channel operation. This will increase the capacity-handling ability of the system, which is usually restricted by the effects of multi-path fading. Also, for the same usable communications, the SSB method uses less average power than that required for the equivalent FM transmitter. It will be recalled that these advantages were also obtained in the VHF scatter systems.

The design of a UHF SSB transmitter is similar to that of the lower-frequency transmitters in present day use. The main difference in UHF equipment is the result of the higher frequencies used. This frequency increase results in the use of smaller parts and different type tubes in these applications.



". . . scattering service——"

TRANSMITTER THEORY

FUNCTIONAL BLOCK ANALYSIS OF A TYPICAL SSB TRANSMITTER

General

As stated previously, there are two basic systems of generating a single-sideband signal—the filter system and the phase-shift system. Although these two systems are basically different in circuit operation, they both produce the same result—the generation and transmission of only one sideband of an amplitude-modulated radio-frequency carrier. Each system by itself is capable of good over-all system performance, and both systems provide an approximate 9 db gain over double-sideband-plus-carrier systems. Compared to each other, however, the filter and phase-shift systems each have certain advantages and disadvantages.

The method of modulating the r-f carrier frequency in a single-sideband transmitter is quite different from that used in an amplitude-modulated transmitter. Either high-level or low-level modulation processes are employed in the usual AM system; in the high-level process the modulating signal is impressed on the final r-f power amplifier in the transmitter, and in the low-level process the modulating signal is impressed on one of the intermediate amplifiers in the transmitter. In the low-level process one or more stages of power amplification follow, to raise the r-f signal to the desired power level for transmission. Thus it can be seen that the modulating process in the AM transmitter is performed at the carrier frequency, and usually at a relatively high-power level. In the filter-system SSB transmitter, the modulation process is accomplished at low-frequency and low-power levels, and the signal is then heterodyned and amplified to the desired frequency and power for transmission. The phase-shift system accomplishes the modulation at a low-power level and, in some cases, at the frequency to be transmitted, with power amplifiers following to amplify the signal to the desired power level.

Filter System of Generating an SSB Signal

Figure 42 shows an example of a typical single-sideband transmitter employing the filter system. The audio amplifier, V1, is a conventional Class A amplifier, operating on the linear portion of its characteristic curve. Normally, audio frequencies below approximately 100 cycles per second and above 3000 cycles per second add little to the intelligibility of speech; therefore, the input audio amplifier may be designed to pass frequencies only

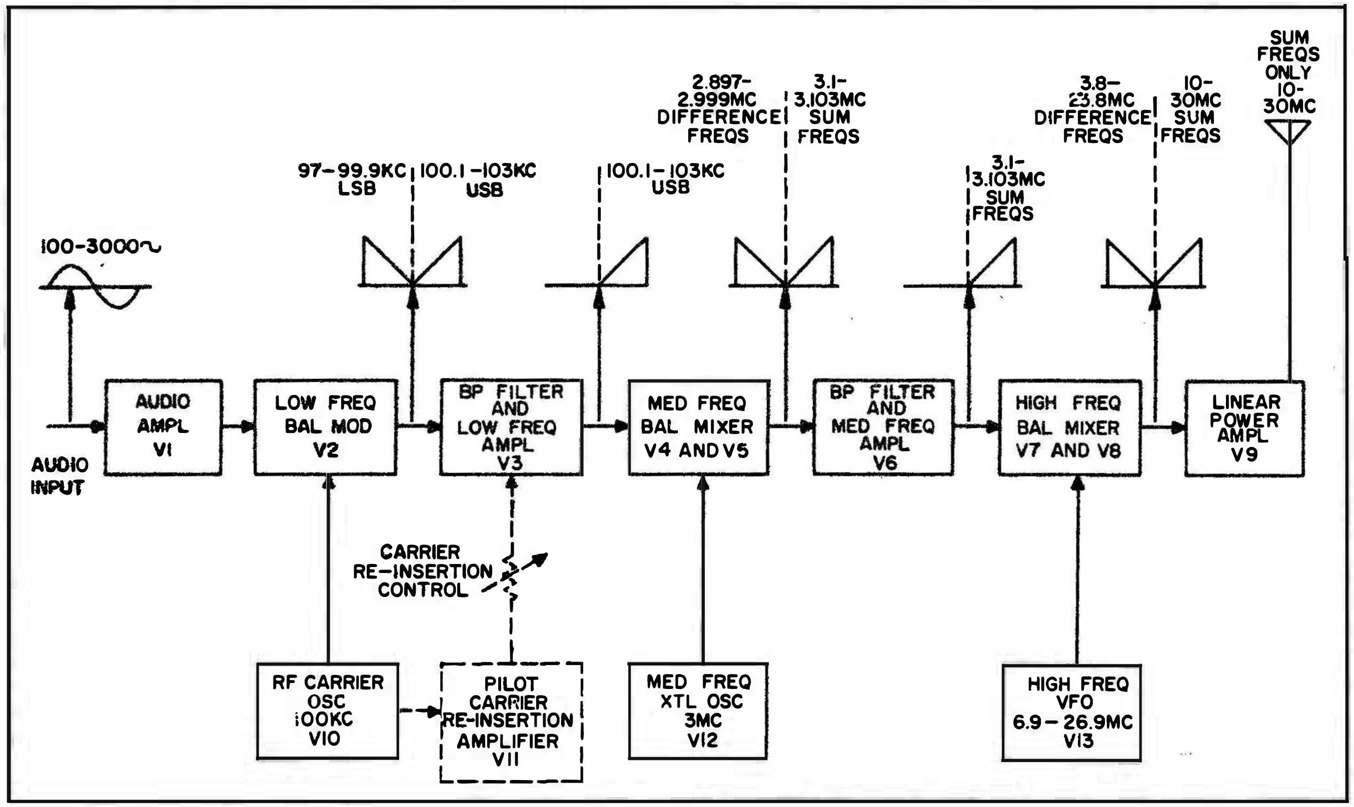


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Figure 42. Block Diagram of a Typical Single Sideband Transmitter (Filter System).

within these limits. The amplified audio output is then applied to the low-frequency balanced modulator, V2, where it amplitude-modulates a low-frequency r-f carrier.

The low-frequency crystal oscillator, V10, uses a highly stable 100-kc crystal, because one of the important requisites of an SSB transmitter is good frequency stability. The frequency of this oscillator is chosen relatively low so that the operating range of the filter (following the balanced modulator, V2) can be placed at its optimum point. The output of the oscillator is applied to the low-frequency balanced modulator.

The primary purpose of any balanced modulator circuit is to produce the sidebands of an amplitude-modulated radio-frequency carrier and to suppress or balance out the carrier. This is exactly what is accomplished in the low-frequency balanced modulator, V2, when the 100—3000-cps audio modulating signal and the 100-kc r-f carrier signal combine in this circuit. The amount of carrier suppression in the balanced modulator depends upon the degree of balance or matching of the tubes and their associated circuit components. Therefore, with an r-f carrier input signal of 100 kc, and audio modulating frequencies of 100—3000 cycles per second, the output of this stage will be a double-sideband signal with the carrier suppressed. The upper sideband will contain frequencies from 100.1 kc to 103 kc, and the lower sideband will contain frequencies from 97 kc to 99.9 kc. These output frequencies are applied to an r-f amplifier, V3, through a bandpass filter.

The bandpass filter in this case is designed to pass only the upper sideband frequencies and eliminate the lower sideband frequencies. The various types of filters which can be employed in SSB systems are discussed in detail in other sections of this manual. However, for the present, let it suffice to say that the purpose of bandpass filters in a single-sideband transmitter employing the filter system is to remove the undesired sideband and to pass the desired sideband. Although the upper sideband has been chosen as the desired sideband in this example, the lower sideband could be just as easily used, with the bandpass filter designed accordingly. For consistency in the discussion, however, the upper sideband is considered to be the desired sideband.

The low-frequency amplifier, V3, is an r-f amplifier operating Class A, and its purpose is to amplify the upper sideband signal. The output is applied to the medium-frequency balanced mixer, V4 and V5, as a heterodyning signal to mix with the 3-mc r-f carrier from the medium-frequency crystal oscillator, V12.

The operation of this oscillator and balanced mixer is similar to the operation of the low-frequency oscillator and balanced modulator circuits. Actually, the medium-frequency balanced mixer performs the first of two heterodyning processes used to raise the transmitter frequency to the value desired for transmission. The output of this mixer will be the sum and difference frequencies of its two input signals, with the 3-mc r-f carrier suppressed. The sum frequencies will extend from 3.1 mc to 3.103 mc, and the difference frequencies will extend from 2.897 mc to 2.999 mc. These signals are applied through a second bandpass filter, where the difference frequencies are eliminated and the sum frequencies are passed to the medium-frequency amplifier, V6. This amplifier is another Class A r-f amplifier, the purpose of which is to amplify the 3.1—3.103-mc sum frequencies and apply them to the high-frequency balanced mixer, V7 and V8, where they are heterodyned with a third r-f carrier frequency.

The third r-f carrier is obtained from a high-frequency oscillator, V13, with a frequency range variable from 6.9 mc to 26.9 mc. The exact frequency from this oscillator depends upon the desired output frequency from the transmitter (10—30 mc).

In the high-frequency balanced mixer, the 6.9—26.9-mc oscillator signal is mixed with the incoming 3.1—3.103-mc signal from the medium-frequency amplifier. The operation at this point accomplishes the second heterodyning process; therefore, V7 and V8 may be referred to as the second balanced mixer. The output of the high-frequency balanced mixer will contain the sum frequencies from 10 to 30 mc (nominal), and the difference frequencies from 3.8 to 23.8 mc (nominal). The absence of a bandpass filter following this stage will be noted. Actually, it is unnecessary to include such a filter at this point, because the output amplifier and antenna circuit will be tuned to the sum frequencies and will not pass the undesired difference frequencies.

It will be noted that up to this point the modulation process (in V2) and the heterodyning processes (in V4 and V5, V7 and V8) have been carried out at a low-power level. The signal at the output of the high-frequency balanced mixer, contains the modulating intelligence at the desired frequency for transmission. In order to increase the power level of the signal, without introducing distortion, Class AB or Class B linear power amplifiers are employed extensively in SSB transmitters. In this transmitter, the linear power amplifier, V10, is operated Class B.

So far the discussion has been concerned with a single-sideband transmitter using a completely suppressed carrier (SSSC). For demodulation at the receiver, a carrier of the same frequency as the original carrier must be introduced by a local low-frequency oscillator. For satisfactory results the oscillators in both the transmitter and receiver must be extremely accurate and stable. Errors as slight as 30 cycles per second in either

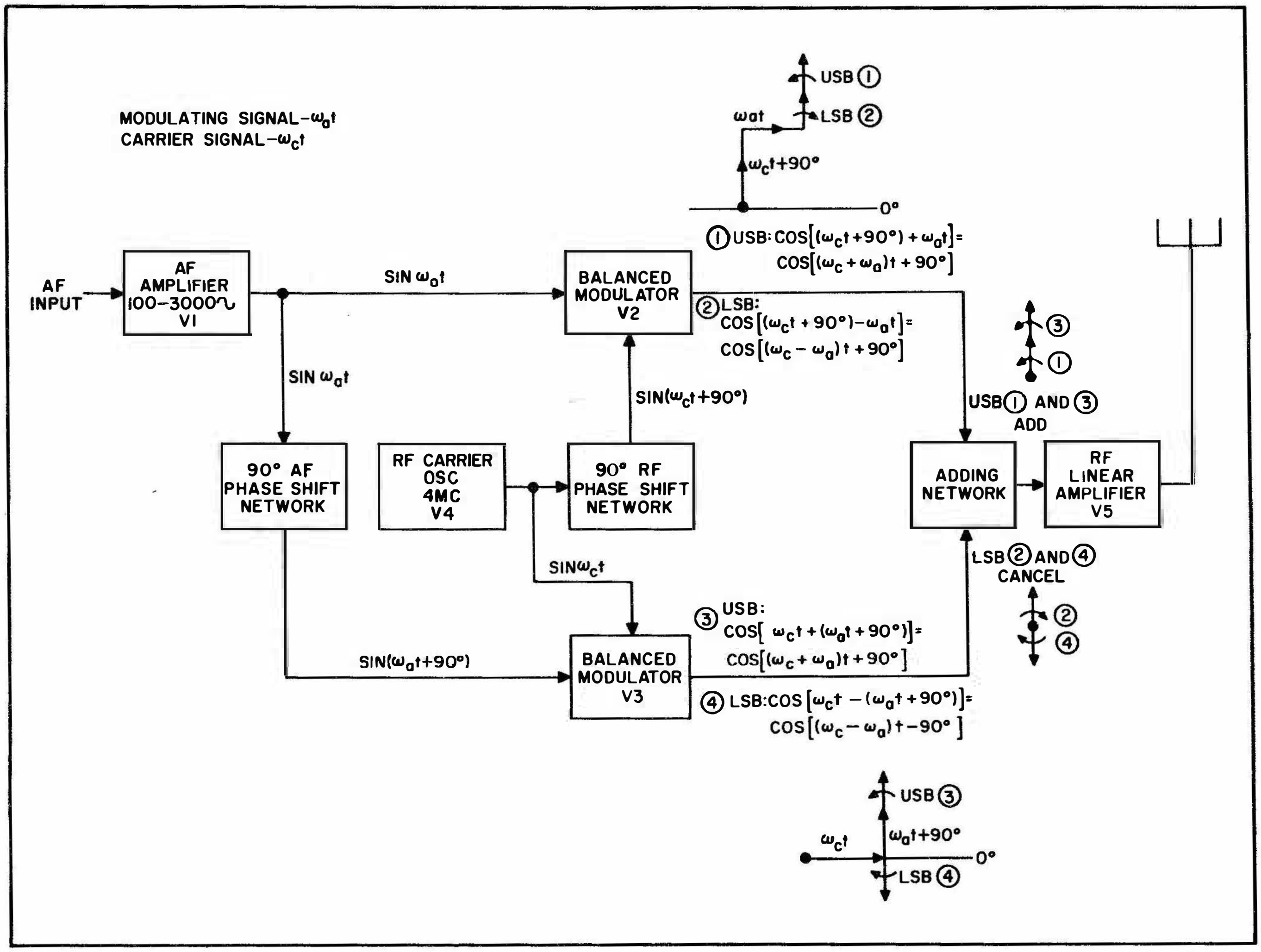


Figure 43. Block Diagram of a Typical Single Sideband Transmitter (Phase Shift System).

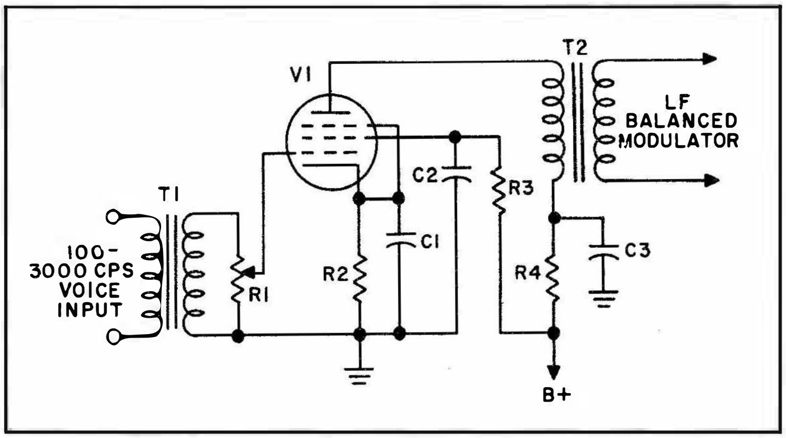
oscillator will cause the intelligibility of speech transmissions to be impaired. To eliminate this possibility of speech distortion, provision can be made in the transmitter for the transmission of a single-sideband reduced-carrier signal. The reduced carrier, often referred to as a "pilot-carrier," is usually transmitted at a level 10 to 20 db below the level of the single-sideband signal.

Referring to the block diagram, figure 42, it can be seen that the pilot carrier is applied from the low-frequency oscillator, V10, through the pilot-carrier re-insertion circuits (dotted block) to the low-frequency amplifier, V3. The pilot-carrier re-insertion control controls the amount of pilot carrier desired for transmission. The SSB receiver receiving this type of reduced-carrier signal must have an a-f-c circuit capable of detecting and utilizing the pilot carrier to maintain accuracy of the frequency-relationship between the transmitter and receiver oscillators.

Phase-Shift System of Generating an SSB Signal

Basically, the phase-shift system removes the undesired sideband by a balancing or cancelling process instead of by a filter. This balancing process is based upon the phase relationships between the sideband signals generated in two separate balanced modulator stages. One set of sidebands from one balanced modulator adds to the same set of sidebands from the second balanced modulator, and the other set of sidebands from one of the other modulators effectively cancels the same set of sidebands from the remaining modulator. Thus the output of a phase-shift system SSB transmitter is essentially the same as the output of a filter system SSB transmitter.

Referring to the block diagram of the phase-shift SSB transmitter, in figure 43, it can be seen that the operation of the input audio amplifier,



TP7-1243

Figure 44. Class A Audio Amplifier (Input Amplifier).

V1, is comparable to that in the filter system. The output of this audio stage is applied directly to the top balanced modulator, V2, and through a wide-band audio 90-degree phase-shifting network to the bottom balanced modulator, V3.

The r-f carrier signal is generated at a low level in a highly stable crystal oscillator, V4, which can be operated at the frequency to be transmitted. The phase-shift system, therefore, sometimes eliminates the use of any heterodyning of the r-f carrier prior to transmission. From the r-f oscillator the carrier is applied directly to the bottom balanced modulator, V3, and through a narrow-band r-f 90-degree phase-shifting network to the top balanced modulator, V2.

The purpose of the balanced modulator in this system is similar to that in the filter system; namely, generation of upper and lower sideband frequencies and cancellation of the carrier frequency. The individual carriers are thus cancelled in their respective balanced modulators, with the output of each balanced modulator then consisting of the upper and lower sideband frequencies only. (The input and output signals of the modulators are noted on the block diagram in figure 43.) These sideband signals are then combined in the adding network in such a manner that the upper sidebands from both modulators aid each other, while the lower sidebands of both modulators oppose each other and are cancelled. The net result is that only the upper sideband frequencies are applied to the linear r-f amplifier, V5, for amplification to the desired power level for transmission. The amplifier is comparable to the one used in the filter system.

For selection and transmission of the lower sideband rather than the upper sideband, the 90-degree phase shift in both the audio modulating signal and the r-f carrier would be applied to the same balanced modulator tube. In such a case, the lower sideband frequencies from both modulators will add and the upper sideband frequencies will cancel.

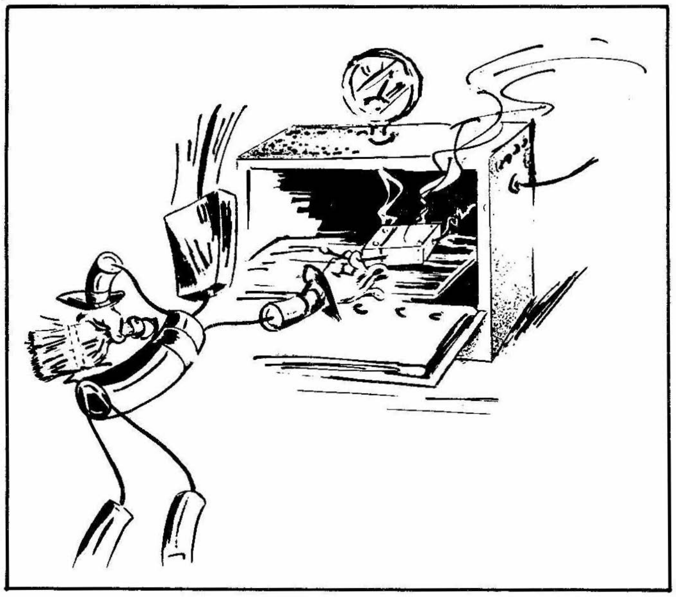
Exact 90-degree phase relationships must be maintained throughout the system, since errors in these phase relationships in either the r-f carrier or audio modulating signals will lead to incomplete cancellation of the undesired sideband. However, with proper phase-shift network design and circuit adjustment, the phase-shift system of SSB can provide a degree of undesired sideband suppression equal to that of the filter system.

FILTER SYSTEM DETAILED CIRCUIT ANALYSIS

In this discussion of the detailed circuit analysis of single-sideband transmitters, particular emphasis will be placed on those circuits peculiar to the single-sideband system of transmission. Since audio amplifiers used in SSB transmitters are comparable to those used in amplitude-modulated transmitters, it is felt that a detailed discussion of their circuit operation is not necessary. However, circuits such as balanced modulators, Class B linear amplifiers, and special type filter and phase-shift networks are peculiar to single-sideband systems and will be covered in detail.

Input Circuits

The purpose of the input-audio amplifier (figure 44) is to amplify the input speech frequencies from a microphone or other source of audio input. Frequencies above 3000 cycles and below 100 cycles are not normally required for the transmission of intelligible speech; for this reason the input amplifier is designed to pass only this band (100—3000 cps) of frequencies. Operation of the amplifier is usually Class A, that is, on the linear portion of its grid-voltage, plate-current characteristic curve.



". . . a temperature controlled crystal is frequently used——"

R-F Carrier Oscillator

The "master oscillator" in the single-sideband transmitter must be extremely accurate and possess a high degree of frequency stability. Because of these requirements, a crystal-controlled oscillator (figure 45) employing a temperature-controlled crystal oven is used quite frequently in SSB transmitters. This is especially true in single-sideband suppressed-carrier systems (SSSC), in which the frequency of the carrier inserted at the receiver must be identical to that of the original carrier which was suppressed in the transmitter. Since quartz crystals are quite fragile, crystal-controlled oscillators, of course, operate with relatively low voltages. Therefore, the necessity for low-level modulation in SSB transmitters is quite evident.

Balanced Modulators

In single-sideband suppressed-carrier transmitters, the carrier can be suppressed either by the use of extremely sharp cutoff filters or by the use of balanced-modulator circuits. The basic principle of a balanced modulator is to introduce the r-f carrier in such a way that it will not appear in the output of the stage. There will be an output signal, however, when both the audio modulation and r-f carrier signals are present simultaneously at the modulator input. This output signal will consist of only the upper and lower sideband frequencies generated in the balanced modulator by the mixing of the two input signals in this nonlinear device. The original audio and r-f inputs will be suppressed because of the operational characteristics of this type of circuit.

There are two basic configurations of balanced-modulator circuits—those using crystal diode rectifiers and those using vacuum tubes. Diode-rectifier balanced modulators will be discussed first, since their operation is relatively simple to understand.

Diode-Rectifier Balanced Modulators

One type of diode-rectifier balanced modulator is illustrated in figure 46. Since the rectifier elements are arranged in the familiar bridge circuit,

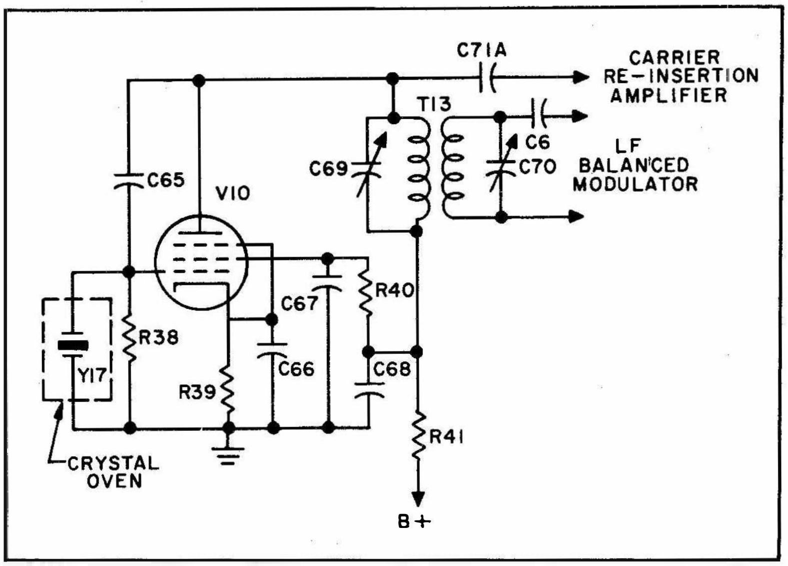


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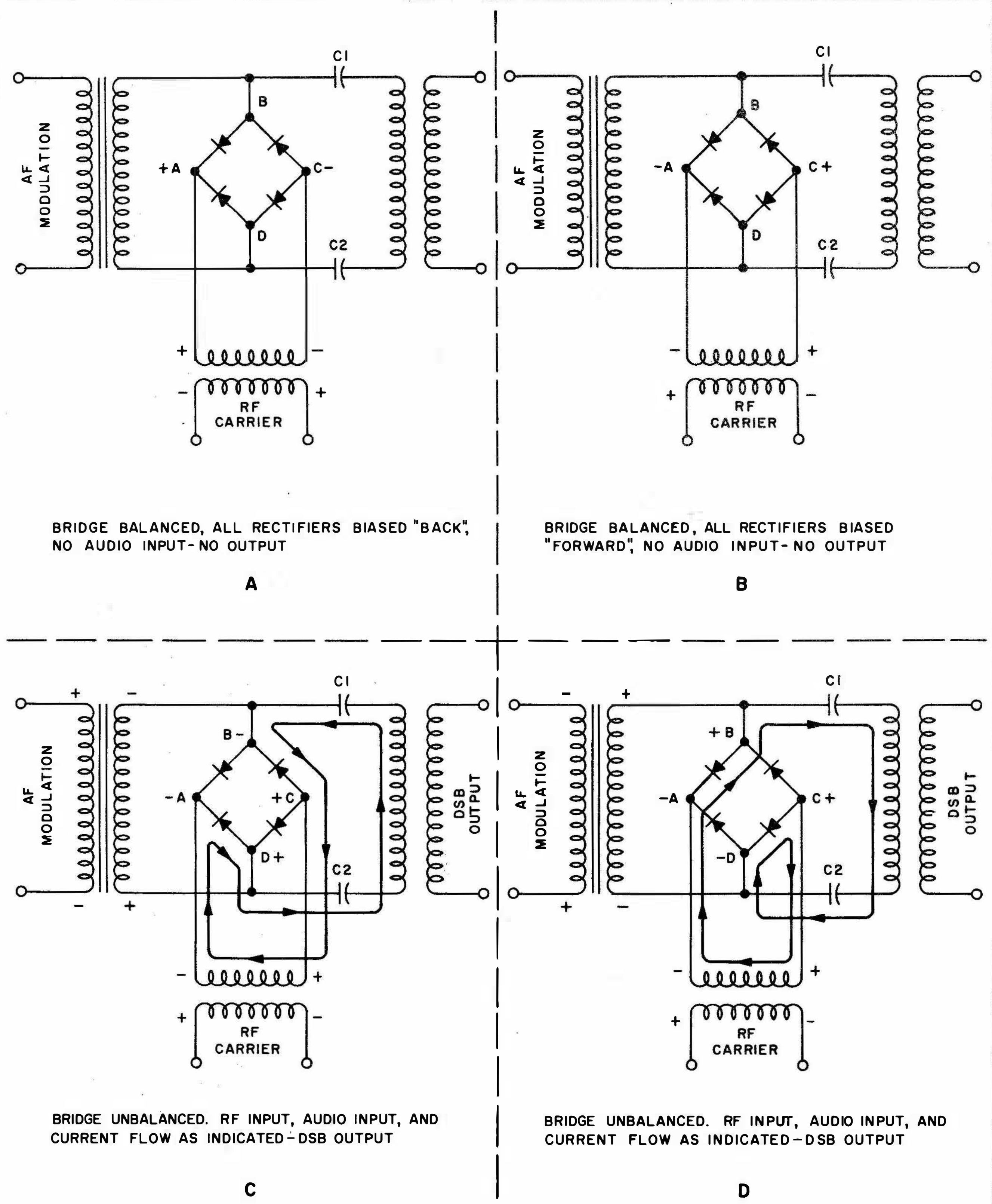
Figure 45. Low-Frequency Crystal Oscillator (100-kc Osc).



BRIDGE BALANCED, ALL RECTIFIERS BIASED "BACK", NO AUDIO INPUT-NO OUTPUT

A

BRIDGE BALANCED, ALL RECTIFIERS BIASED "FORWARD", NO AUDIO INPUT-NO OUTPUT

B

BRIDGE UNBALANCED. RF INPUT, AUDIO INPUT, AND CURRENT FLOW AS INDICATED-DSB OUTPUT

C

BRIDGE UNBALANCED. RF INPUT, AUDIO INPUT, AND CURRENT FLOW AS INDICATED-DSB OUTPUT

D

NOTE: THE FORWARD DIRECTION OF ELECTRON FLOW FOR THE DIODE RECTIFIERS IS ASSUMED TO BE OPPOSITE TO THAT OF THE ARROW HEAD IN THE SYMBOL.

Figure 46. Analysis of Balanced-Bridge Modulator.

TP7-1245A

this type of modulator is called a "balanced-bridge modulator." To analyze the operation of this circuit, assume first that only the r-f carrier is applied at points A and C with the polarity indicated in figure 46A. Since the four rectifiers are now biased "back," there will be no output from the circuit because of the fact that the rectifiers will have maximum impedance and will not permit any current to flow. This action will effectively cause an open circuit to the r-f input. When the polarity of the r-f carrier is reversed on the second alternation, as indicated in figure 46B, the four rectifier elements are biased "forward" and once again there will be no r-f output from the bridge circuit—this time because of the fact that all the rectifiers now offer minimum impedance and effectively create a short-circuit across the output transformer. Thus it can be seen that as long as there is only an r-f signal input to a balanced modulator, there will be no output from this type of circuit.

When audio modulation is applied to the balanced-bridge modulator at the same time that an r-f carrier is being applied, a different situation arises. Assume now that the audio is being applied to points B and D, as indicated in figure 46C, while the r-f carrier is being applied to points A and C. The audio signal will cause the bridge to become unbalanced, and a current will flow through the bridge as indicated in the figure. There will be an r-f output at this time, but it will be only the upper and lower sidebands generated in the rectifiers.

When the polarity of the audio signal reverses, current will flow through the rectifiers as indicated in figure 46D. The reactance of capacitors C1 and C2 will block the low audio frequencies from the output transformer. It now can be understood that there will be an r-f output from the balanced-bridge modulator only when the audio modulation and r-f carrier signals are both applied to the modulator simultaneously. The amplitude relationship between the r-f and audio voltages should be such that the amplitude of the r-f voltage is eight to ten times that of the audio voltage. This relationship is necessary if it is desired to keep over modulation distortion to a minimum. The input and output waveforms of this circuit are shown in figure 47. The shaded area indicates the times during which there will be an output signal from this type of modulator circuit.

A second type of diode-rectifier balanced modulator is the ring or lattice-type modulator. Two variations of this circuit are illustrated in figure 48. In these diagrams, the diode elements are numbered identically, to indicate that the two variations in circuitry will produce the same results.

To understand the operation of the ring-type balanced modulator, assume that the positive alternation of the r-f signal is applied to center tap B in figure 48A; also assume that there is no audio modulation input. At this time, rectifiers 2 and 4 are biased "forward" and show minimum impedance, while rectifiers 1 and 3 are biased "back" and effectively appear as open circuits. For this condition there will be no r-f output from the circuit because the r-f input, which is applied to the center taps of the input and output transformers, will cause equal amounts of r-f current to flow (as indicated in the figure) through each half of these windings, thus cancelling this signal. When the polarity of the r-f input reverses, current will flow as indicated in figure 48B. Now rectifiers 1 and 3 conduct while rectifiers 2 and 4 appear as open circuits. Once again there will be no r-f output from the circuit for the same reason as mentioned above.

Referring to figure 48C, it can be seen that when an audio modulating signal is applied to the input transformer at the same time that the r-f signal is being applied, current will flow as indicated. By noting the polarities only, it appears that rectifier 4 is biased "back" and should not conduct. However, because the r-f signal is much greater in amplitude than the audio signal, this rectifier will actually conduct. In the analysis of diode-rectifier modulators, it is important to consider the relative amplitudes of the two input signals, as well as their polarity.

Because of the diode rectifiers are non-linear devices, mixing of the two input signals in these elements will generate upper and lower sidebands. Since the ring-type modulator circuit is not balanced for these sideband frequencies, they will appear across the output transformer. At no time will there be a complete circuit through the pri-

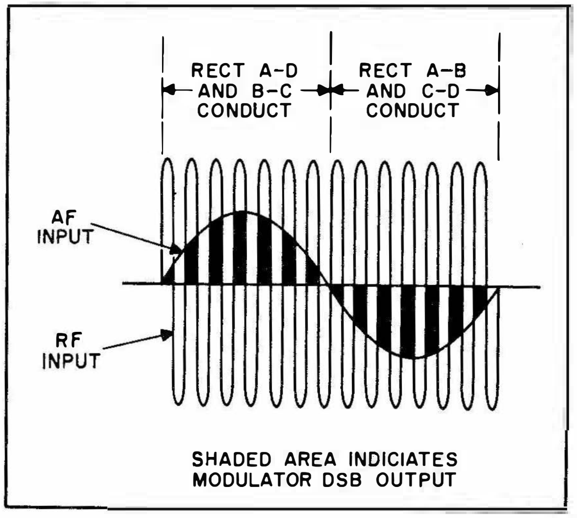


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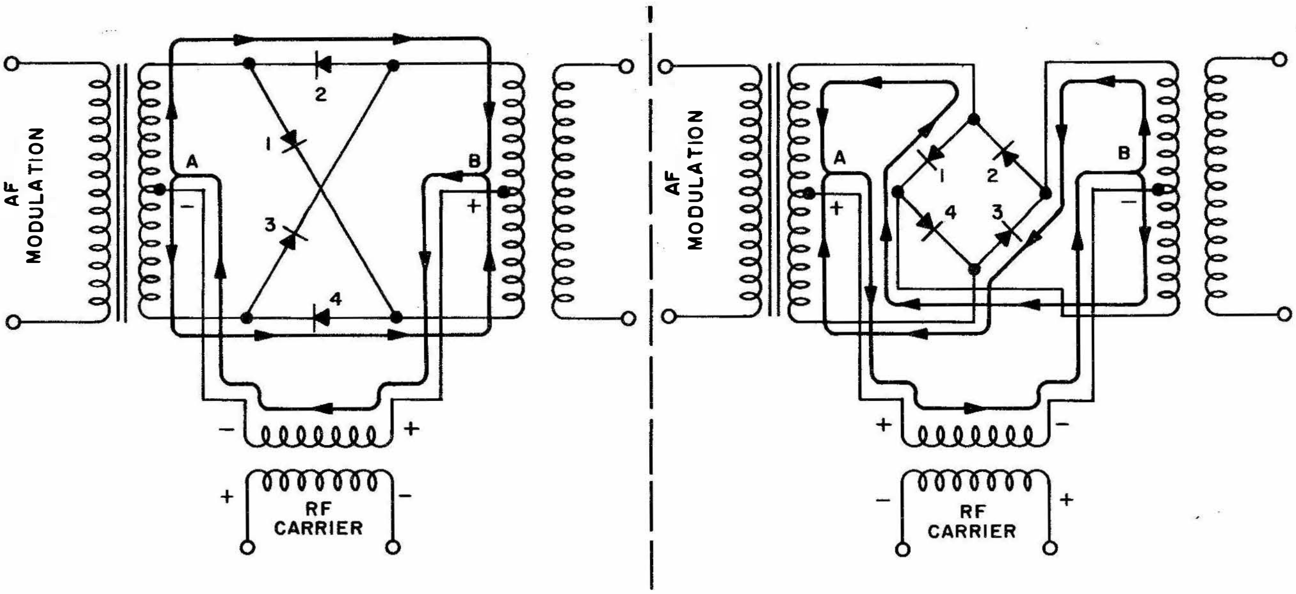
Figure 47. Balanced-Bridge Modulator Waveforms.

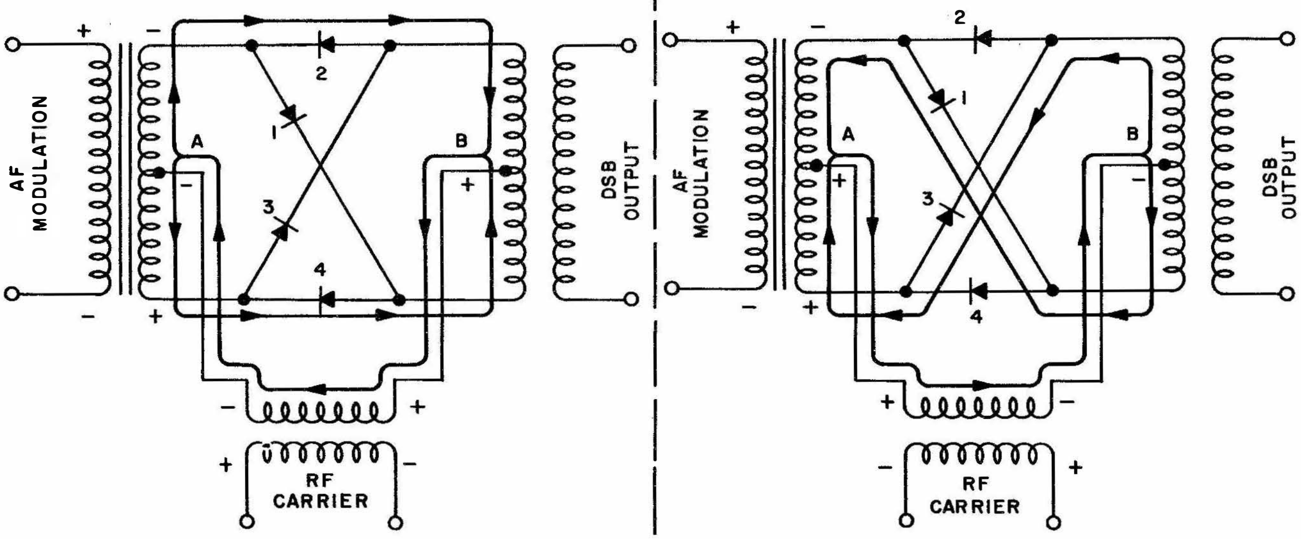


NO AUDIO INPUT. RECTIFIERS 2 AND 4 CONDUCT, BALANCE OUT RF INPUT - NO OUTPUT

A

NO AUDIO INPUT. RECTIFIERS I AND 3 CONDUCT, BALANCE OUT RF INPUT – NO OUTPUT

B



AUDIO INPUT, RF INPUT, AND CURRENT FLOW AS INDICATED - DSB OUTPUT

C

AUDIO INPUT, RF INPUT, AND CURRENT FLOW AS INDICATED- DSB OUTPUT

D

NOTE: THE FORWARD DIRECTION OF ELECTRON FLOW FOR THE DIODE RECTIFIERS IS ASSUMED TO BE OPPOSITE TO THAT OF THE ARROW HEAD IN THE SYMBOL.

Figure 48. Analysis of Ring- or Lattice-Type Balanced Modulator.

TP7-1247A

mary of the output transformer for the low-frequency audio components. Hence, like the r-f carrier input, this signal will never appear in the output of this modulator.

Operation of the ring-type modulator illustrated in figure 48D is identical to that of the one in figure 48C, except that the polarity of the r-f input signal is reversed. This reversal of the r-f signal will cause rectifiers 1 and 3 to conduct, and rectifiers 2 and 4 to be biased "back."

The controlling signal in the ring-type modulator is the r-f carrier, whereas the controlling signal in the balanced-bridge modulator is the audio input signal. By analyzing the waveform illustrated in figure 49, it will be observed that the output of the ring-type modulator is twice that of the balanced-bridge modulator. It will also be noticed that the shaded area indicates only the sideband output of the circuit. Although the shaded area appears to be at the same frequency as the r-f carrier input, it actually represents the combination of the upper and lower sideband frequencies only.

Vacuum-Tube Balanced Modulators

Vacuum-tube balanced modulators perform the same functions as crystal-diode-rectifier balanced modulators, and have similar operational characteristics; that is, they combine the audio and r-f input signals in such a way that only the generated upper and lower sideband frequencies appear in the output. In order for these vacuum-tube circuits to generate these sidebands, they must be operated in a non-linear fashion. Therefore, vacuum-tube balanced modulators are generally operated on the non-linear portion of their characteristic curves. The two input signals will be suppressed in these modulators, just as they are in diode-rectifier circuits. In fact, balanced modulators can be designed to balance out either or both of the input signals; however, there are no balanced modulators that can balance out only one sideband and not the other.

The choice of balanced-modulator circuit to be employed is determined by the equipment designer, and will vary for different applications. Triodes, tetrodes, and pentodes are all found in vacuum-tube modulators. The methods of modulation that may be used include control-grid, screen-grid, suppressor-grid, and plate modulation. Control-grid modulation is the most widely used because this method of modulation usually results in better control of the circuit by the audio signal.

Two types of simple vacuum-tube balanced modulator circuits employing control-grid modulation are illustrated in figure 50. From figure 50A, the first modulator circuit to be considered, it can be seen that the audio and r-f input signals are both introduced in push-pull, and that the modulator plate circuit is connected in parallel. Assume that only the r-f signal is present on the modulator control grids, with the polarity indicated by the waveform in the figure. For this condition the tubes will conduct push-pull; that is, when conduction through the top tube increases, the conduction through the bottom tube decreases. The r-f signal will therefore be cancelled (suppressed) in the parallel-connected plate circuit, and there will be no output from the modulator.

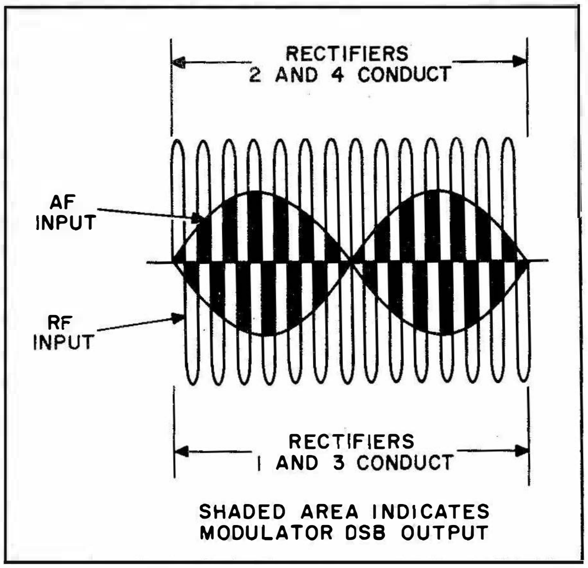


Figure 49. Ring- or Lattice-Type Balanced Modulator Waveforms.

When audio modulation of the polarity indicated in the figure, is applied to the balanced-modulator control grids, upper and lower sideband frequencies will be generated in these non-linear stages. Since the circuit is not balanced for the sidebands, they will appear across the output transformer. The r-f carrier will be suppressed in the manner described above, with the amount of carrier suppression depending upon the degree of balance between the modulator tubes and their associated circuit components. If further carrier suppression is required, R-C balancing or separate bias adjustments may be incorporated in the circuit design. The audio input will also be suppressed in the balanced modulator, because the r-f tuned plate circuit will offer maximum impedance to these low-frequency audio components.

A second simple type of vacuum-tube balanced modulator is illustrated in figure 50B. In this circuit the r-f carrier is introduced in parallel, and the audio modulation in push-pull; the modulator plate circuit is connected in push-pull. A practical

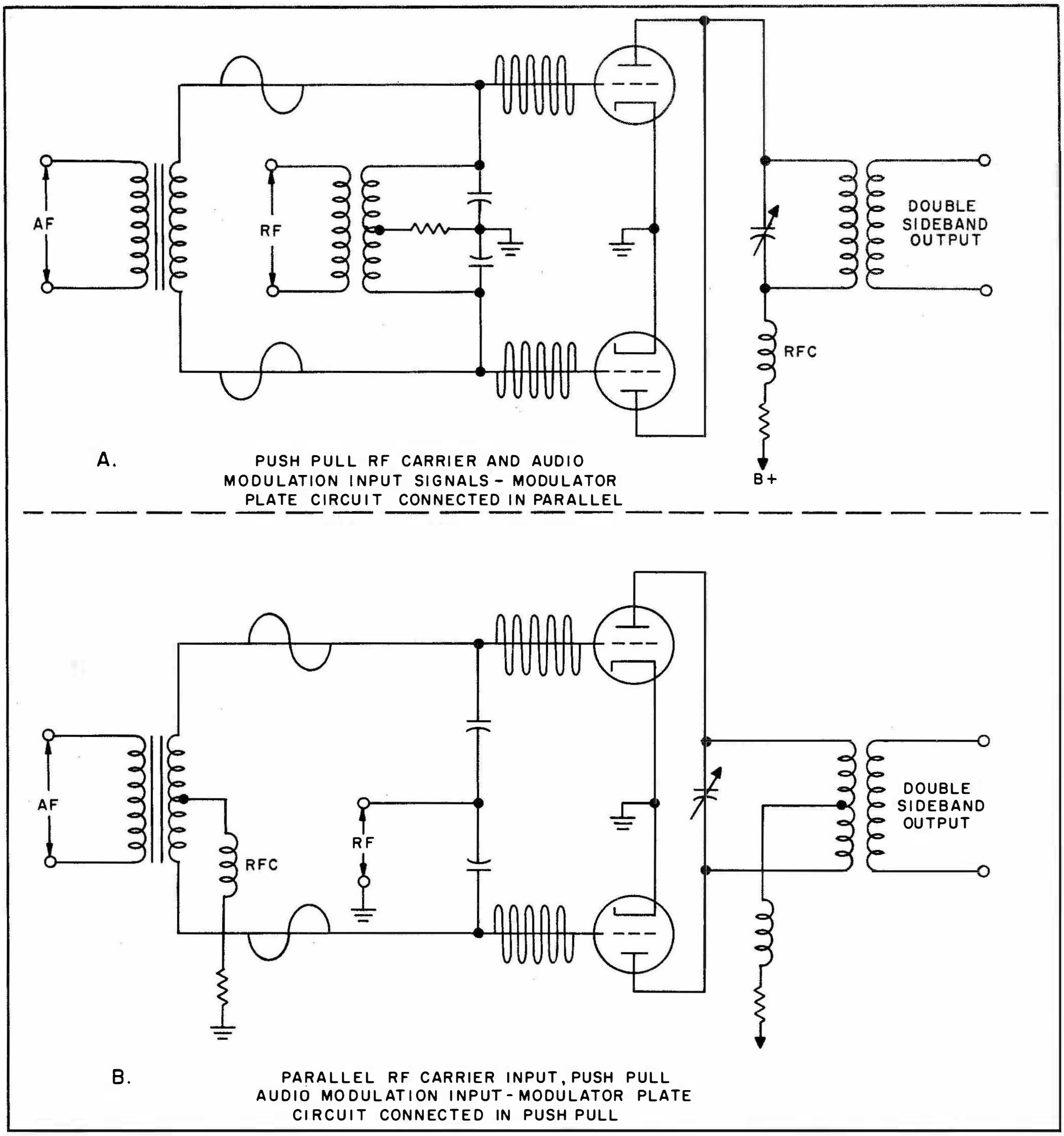
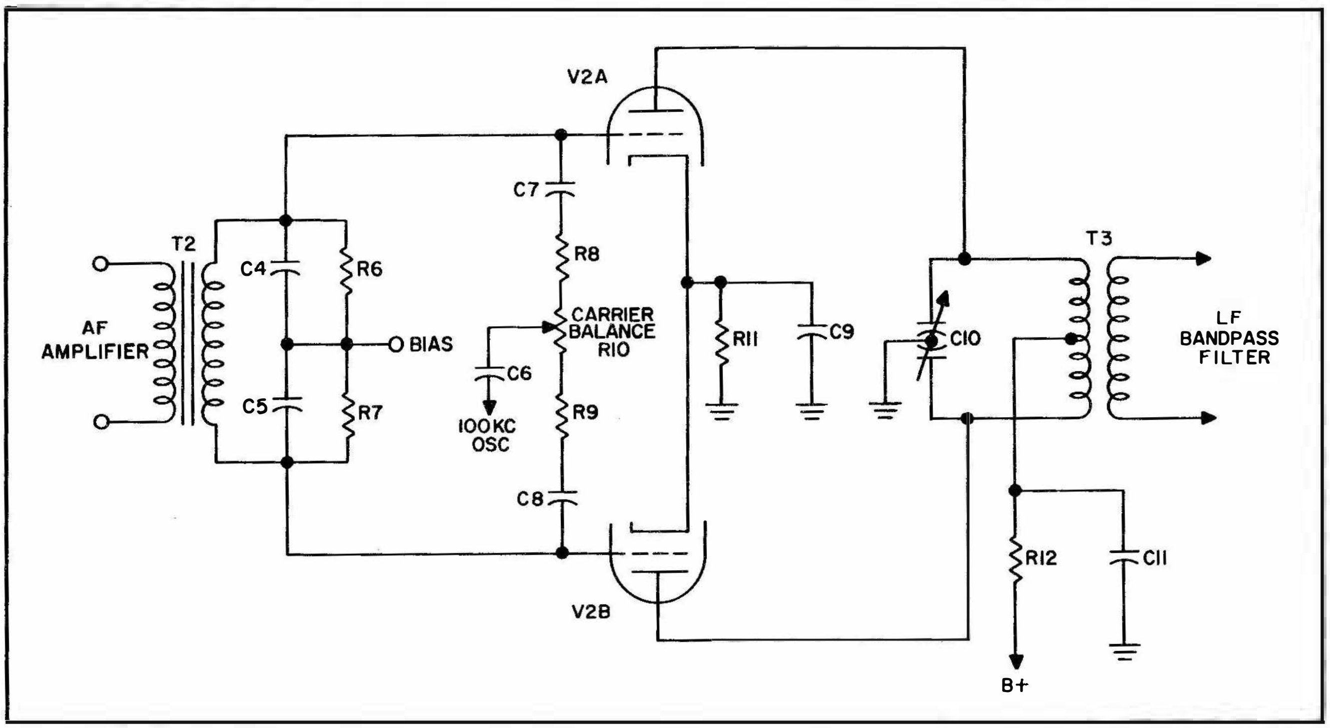


Figure 50. Two Types of Simple Balanced Modulator Circuits.

arrangement of the same circuit is shown in figure 51. In the practical circuit a "carrier-balance" control, R10, and a separate bias supply have both been incorporated in the circuit design. The bias, which is applied to the control grids through equal value resistors, R6 and R7, provides the proper non-linear operating characteristic for the circuit.

Portions of the r-f carrier are applied in equal amplitude and the same phase to both control grids through the carrier balance control, R10. If only the r-f carrier is present, there will be no output from the modulator, because this signal will be balanced out in the push-pull plate circuit. For example, when positive alternations of the r-f signal are applied to both control grids at the



TP7-1250

Figure 51. Low Frequency Balanced Modulator.

same time, both modulator tubes will conduct equal amounts and develop signals 180 degrees out of phase across the top and bottom halves of the primary of the output transformer, T3. The polarity and amplitude of the two signals thus developed will be such that they will oppose, and tend to cancel each other.

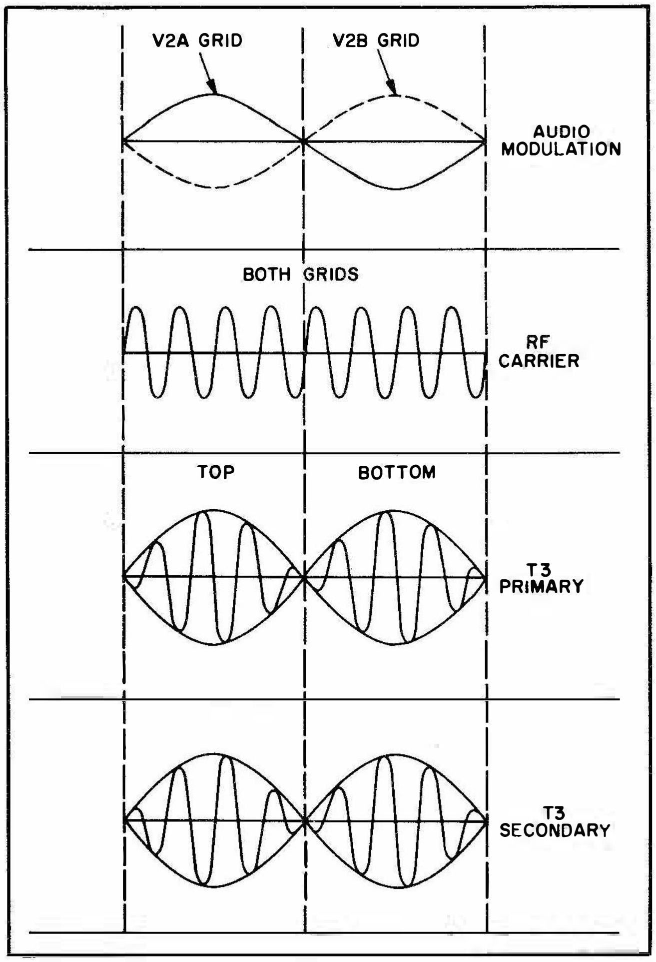
On the other hand, if the audio modulating signal is impressed on the control grids in push-pull at the same time that the r-f carrier is also present on the control grids, a slightly different situation arises. For purposes of illustration, assume that the positive alternation of the audio is applied to the top tube, V2A, and that the negative alternation is applied to the bottom tube, V2B. It can readily be observed that an unbalance of the modulator tubes will exist, with V2A conducting more heavily than V2B. Modulation of the r-f signal by the audio will occur in tube V2A, with upper and lower sideband frequencies being generated accordingly. These frequencies will then be developed across the top half of the primary of transformer T3. On the second half cycle of the audio, tube V2B will conduct more heavily, because the positive alternation of the audio will now appear on this grid. Upper and lower sideband frequencies will be generated (because of the modulation process in tube V2B), and these signals will be developed across the lower half of the modulator plate transformer. Subsequently, the upper and lower sideband frequencies from each modulator tube will add to each other in the secondary of T3, giving a double-sideband amplitude-modulated signal. The carrier will be suppressed in the manner described above and will not appear in the output with the double-sideband signal.

It is understood that the audio modulating signal will also be suppressed in the balanced modulator, since the modulator output circuit is tuned to the sideband frequencies (r-f). No audio modulating components will therefore be developed in this circuit. Critical balance of the audio is not essential in the balanced modulator, as any slight unbalance of this signal will have no effect on the desired double-sideband output. Figure 52 illustrates typical waveforms in the low-frequency balanced modulator circuit.

Transformerless Balanced Modulator

A type of balanced-modulator circuit quite different from those just discussed is one which uses no transformers. This circuit (figure 53) is actually a form of product modulator—a circuit whose output is proportional to the product of the amplitudes of the input signals.

In the operation of this circuit, V1 and V2A act as cathode followers to control the voltage on the cathode of V2B. When the r-f carrier and audio modulation signals are applied to the control grids of V1 and V2A, respectively, current flow through the common cathode resistor (R5) will develop a positive voltage at the top of this



TP7-1251

Figure 52. Balanced Modulator Waveforms.

resistor and effectively apply a negative voltage (bias) to the grid of V2B. The phase of the voltage developed at the load inductor (L1) as a result of the negative grid voltage of V2B will be opposite to that of the voltage developed at the same inductor as a result of the positive grid voltage of V1 and V2A. Cancellation of the r-f carrier will result, with the amount of cancellation depending upon the setting of potentiometer R1. For a completely suppressed carrier, R1 is adjusted to produce minimum carrier signal across L1, and capacitor C3 is adjusted to reduce any capacity feed-through which may occur between V1 and V2B. Upper and lower sideband frequencies will be generated in this circuit, while the two input signals will be suppressed.

The cathode resistor of V2B (R6) provides the proper operating point for this tube, with capacitor C4 acting as an r-f bypass. If so desired, this R-C combination can be eliminated and a separate d-c bias supply can be connected to the grid of V2B (as indicated by the dotted lines). By using this separate bias supply, it is possible to obtain more critical adjustment of the circuit. This modulator circuit can also be used as a low-distortion AM modulator in systems requiring up to 100% modulation.

Bandpass Filters

Crystal-Lattice Filters

The double-sideband output from the low-frequency balanced modulator is applied to a bandpass filter, which allows the desired upper sideband frequencies to pass to the following amplifier

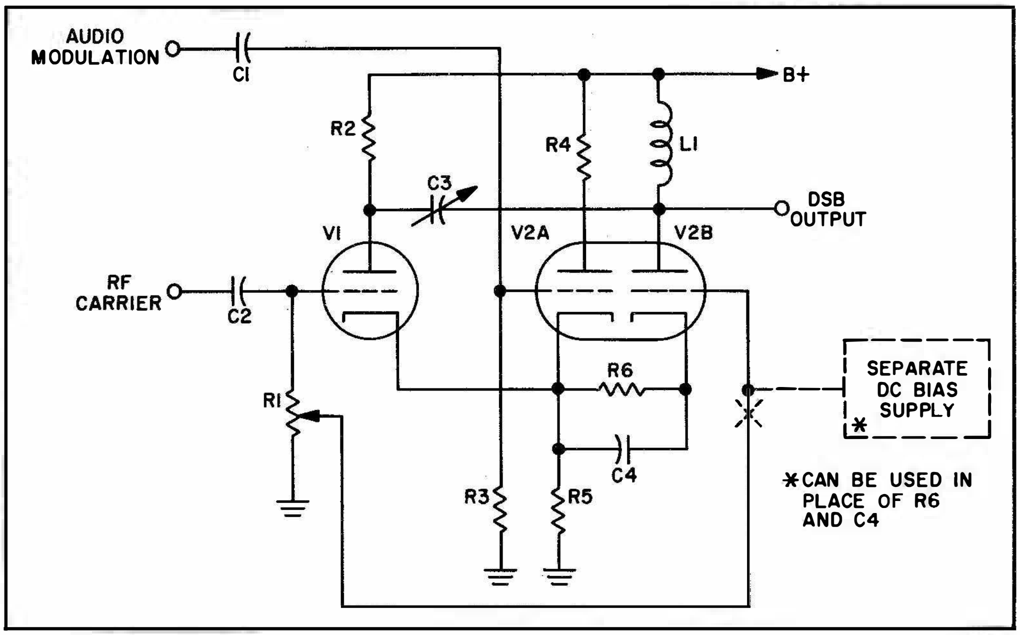


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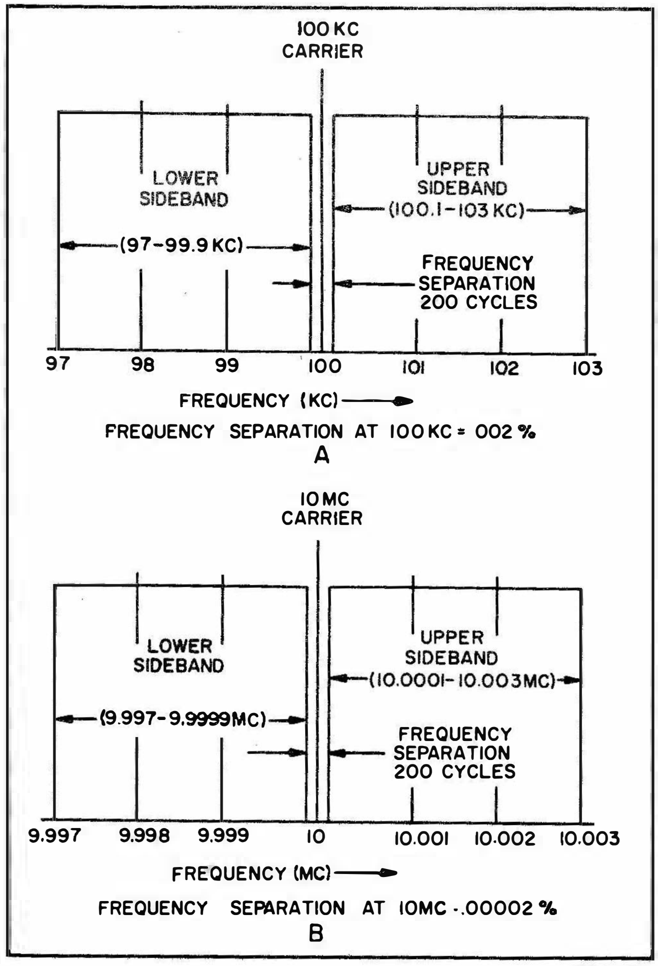
Figure 53. Transformerless Balanced Modulator.



TP7-1253

Figure 54. Frequency Separation of Sidebands at 100 kc and 10 mc, with 100—3000-cps Audio Modulating Signals.

stage, but rejects the lower sideband frequencies. The design of a filter sharp enough to select the desired sideband and reject the undesired sideband would be exacting if the modulation process were performed at a high radio frequency, because the percentage of frequency separation between the two sidebands would be quite small. Therefore, for greater ease of filter design, the modulation process in the filter-system single-sideband transmitter is accomplished at a relatively low radio frequency (100 kc).

Figure 54 illustrates the percentage of frequency separation of sidebands for carrier frequencies of 100 kc and 10 mc. Assuming the same band of audio modulating frequencies (100—3000 cps) for both carriers, the sidebands of the 100-kc carrier will have a frequency separation of .002%, while the sidebands of the 10-mc carrier will have a frequency separation of only .00002%. From these facts it is easy to see that the task of designing a filter to separate the sidebands for the higher r-f carrier would be quite difficult. However, filters of the crystal-lattice type—employing piezoelectric quartz crystals—are relatively easy to design and construct for frequencies in the lower r-f range. A typical attenuation curve for such a filter is shown in figure 55.

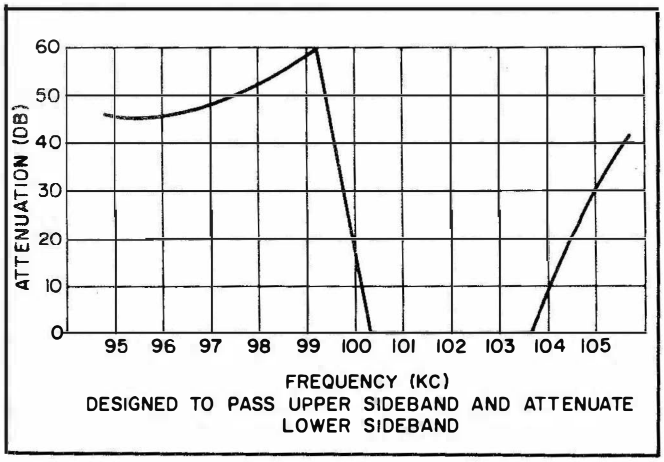


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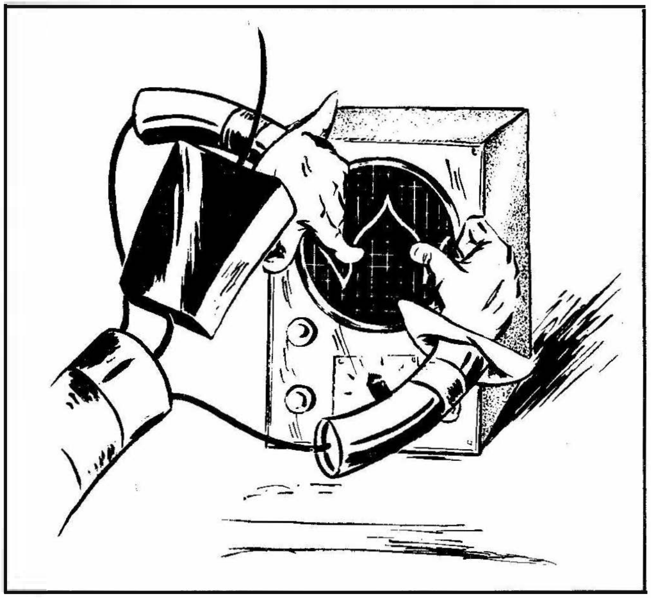
Figure 55. Typical Attenuation Curve for a Crystal Filter in the Range of 100 kc.

Before discussing crystal-lattice filters in detail, it will be of value to briefly review the equivalent electrical circuit and reactance curve of a single quartz crystal. The schematic symbol of a crystal and its equivalent electrical circuit are illustrated in figure 56A. A crystal in its holder is actually a combination of both series and parallel resonant circuits, and as such it has two resonant frequencies, as indicated in figure 56B. The series resonant frequency, f_r, occurs at the point where the reactance curve crosses the zero-reactance line; and the parallel resonant frequency (antiresonant), f_a, occurs at the point where the reactance curve rises to a high inductive reactance and then falls sharply through the zero-reactance line to a high capacitive reactance. In most crystals the two resonant frequency points (f_r and f_a) will occur within a few hundred cycles of each other, and, by spreading these two resonant points, it is possible to use the crystals as elements in a filter network. Such spreading can be accomplished by using an inductor in parallel with the crystal. Spreading the reactance curve in this manner causes the series resonant frequency point (f_r) to remain fixed and moves the parallel resonant point (f_a) to a higher frequency. Also, the spreading creates a new parallel resonant frequency (f_{a2}) at a point lower in frequency than f_r. Figure 56C shows the reactance curve of a crystal whose frequency has been spread by using a parallel inductor.

The crystal-lattice filter, consisting of two pairs of identical crystals (Y_1 and Y_2, Y_3 and Y_4) connected between the input and output transformers

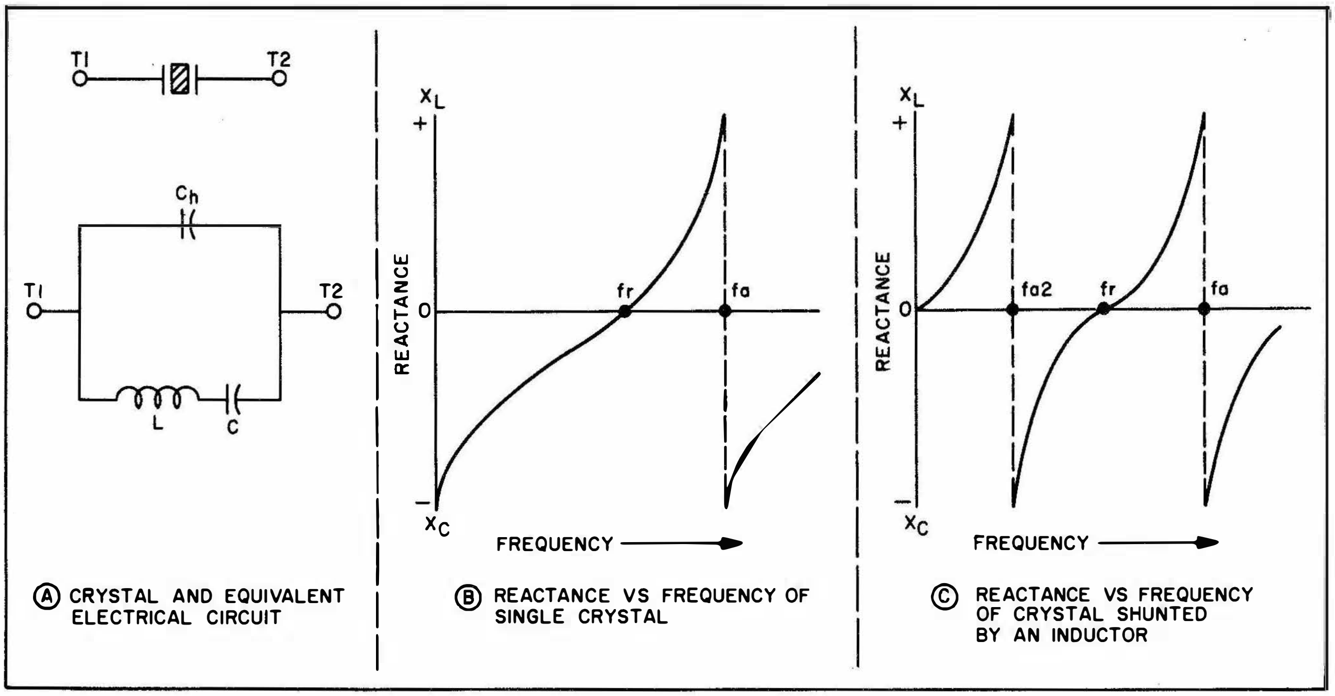


". . . and by spreading these two resonant points——"

of the filter network (figure 57), is the type most commonly used in filter-system single-sideband applications. For optimum filter operation, each pair of crystals should be matched in frequency as close as possible, usually within 10 or 20 cycles of each other. Also, the parallel-connected pair of crystals (Y_3 and Y_4) should be approximately 2 or 3 kc higher in frequency than the series-connected pair (Y_1 and Y_2). Frequency-spreading of the f_r and f_a resonant points of the crystals is accomplished by the filter network input and output transformers, which act as parallel inductors, and any overspreading of the resonant frequency points can be compensated for by the trimmer capacitors in parallel with each transformer. The L-C network is tuned to the center of the pass band, and the bandpass is determined by the frequency separation between the identical pairs of crystals. Although only a one-section crystal-lattice filter network is shown in figure 57, two crystal-lattice filter sections are often used in series in practical equipment circuit design.

The operation of a crystal-lattice filter is based upon essentially the same principle of operation as that of a simple rectifier bridge circuit. When the signals on the bridge legs are equal and have the same polarity (balanced), the signals through the two legs of the bridge will cancel and there will be no output from the circuit. On the other hand, when the signals on the bridge legs have opposite polarity (unbalanced), a signal will be present in the output. The maximum output will occur when the input signals are equal in amplitude but opposite in polarity. A similar action occurs in the crystal-lattice filter.

The reactance curves for the matched pairs of crystals are illustrated in figure 58A, the curve for the parallel-connected pair being shown by the solid line, and the curve for the series-connected



Ⓐ CRYSTAL AND EQUIVALENT ELECTRICAL CIRCUIT

Ⓑ REACTANCE VS FREQUENCY OF SINGLE CRYSTAL

Ⓒ REACTANCE VS FREQUENCY OF CRYSTAL SHUNTED BY AN INDUCTOR

TP7-1255

Figure 56. Quartz Crystal, with Equivalent Electrical Circuit and Reactance Curves.

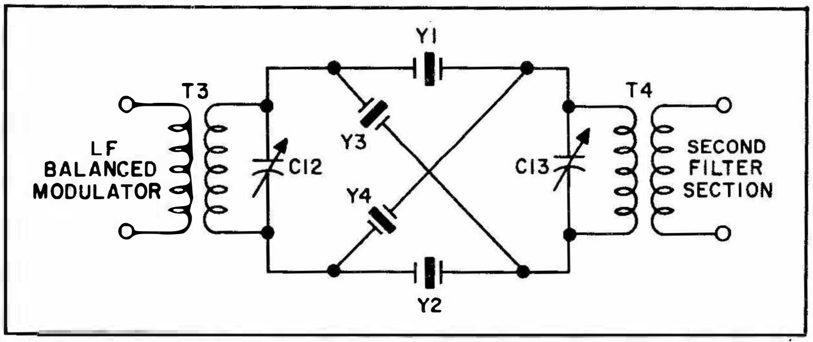


Figure 57. Crystal Lattice Filter (L-F Bandpass Filter).

pair, by the dashed line. Observe that each curve has a series-resonant frequency point and two parallel-resonant frequency points (due to spreading by the transformers). Careful alignment of the network is necessary in order to make the series-resonant point of the series-connected pair of crystals (point R) correspond to the parallel-resonant point of the parallel-connected pair of crystals (point A), and vice versa. The attenuation curve for this lattice network is illustrated in figure 58B. It can be seen that there will be high attenuation at the points where the values of the two reactance curves are equal and have the same polarity (+ or —). For those frequencies where the two reactance curves are opposite in polarity but of equal amplitude, there will be a bandpass.

Beside the band-pass consideration in the design of a crystal lattice filter, another factor usually considered is the "insertion loss" of the filter. This loss is the ratio of the input power to the output power of the filter network, and for wide-band filters it is approximately 10 to 15 db. However, the loss may be reduced to 5 or 6 db by using the proper input and output circuits for a particular filter network.

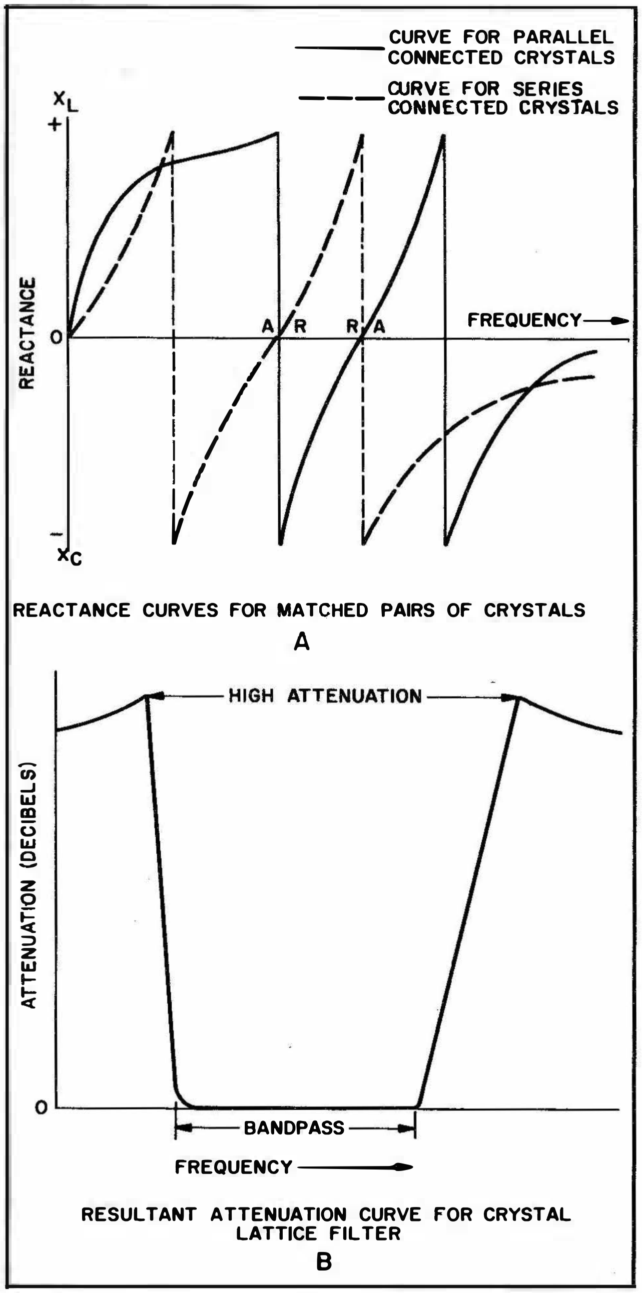
Electromechanical Filters

The crystal-lattice filter is by no means the only type of filter used in single-sideband applications. Electromechanical filters (often referred to as mechanical filters) are also suitable as sideband filters, especially in the frequency range around 250 kc. Below this range, the physical size of the mechanical filter becomes relatively large, and above this range the required physical tolerances become difficult to maintain. However, it is possible to use mechanical filters in almost all applications where other type filters are used.

Various types of mechanical filters have been designed and constructed for use in SSB (and other) applications. The basic operating principle of this type of filter is the use of extremely high-Q (up to 10,000) metallic plates, rods, or disks as resonating elements, together with magnetostrictive transducers to convert the electrical signals to mechanical vibrations and vice versa. The number of resonator elements, the coupling between the resonator elements, and their physical size determine the skirt selectivity, bandwidth, and center frequency, respectively, of the mechanical filter. Spurious frequencies can be held to a minimum by such design features as proper arrangement of the coupling and driving elements and proper selection of resonator shape and size. Because of the extremely high Q of metallic resonator elements, mechanical filters can be designed and constructed to produce sharper cutoff characteristics and narrower bandwidths than is possible with other types of filters.

The internal construction of a disk-type mechanical filter is illustrated in figure 59, and the transducer element for such a filter is illustrated in figure 60. An electrical signal applied to the input transducer inductor excites the transducer element. If minimum distortion is desired, the filter input voltage should be held to a value less than that rated for the filter, because strong input signals can cause non-linear operation of the filter. The purpose of the input transducer is to convert the electrical signals into mechanical vibrations. (The principles of operation of magnetostrictive electromechanical transducers will not be covered here, since such information can be obtained from other publications.) The mechanical energy is then coupled through the resonator elements to the output transducer, where the mechanical energy is converted back to an electrical signal. The operation of the output transducer is similar in all respects to the operation of the input transducer, except for the reversal of the conversion process. The small permanent magnets mounted above the transducers apply a magnetic bias to the transducer elements, in order to effect optimum electromechanical coupling.

Since a filter is a relatively high-impedance device, proper termination of both the input and output of such a device is required. Besides acting as electrical-to-mechanical (and vice versa) con-



REACTANCE CURVES FOR MATCHED PAIRS OF CRYSTALS

A

RESULTANT ATTENUATION CURVE FOR CRYSTAL LATTICE FILTER

B

TP7-1257

Figure 58. Single-Section Crystal Lattice Filter Curves.

verters, the transducers provide the correct values of input and output impedances for filter termination. The proper impedances are obtained by mechanical damping caused by the transducer, and by the resistance of the electrical circuits to which the transducers are connected.

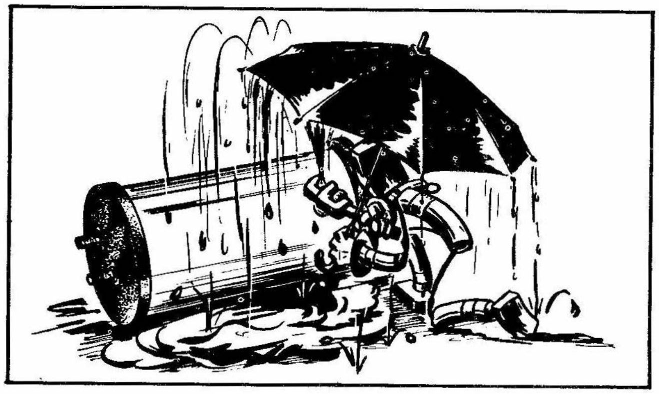
Figure 61 is a photograph of a mechanical filter which has an internal construction similar to that illustrated in figure 59. It can be seen in the photograph that the connecting terminals are located on opposite ends of the filter. This arrangement provides adequate shielding between the external input and output connections, and thus eliminates any stray coupling between these points which might affect the pass band and the skirt selectivity of the filter. It can also be observed that the filter is hermetically sealed; therefore, no adjustments are necessary or possible. The over-all size of the filter is .75 inch in diameter and 2.75 inches in length.

Another type of mechanical filter in use at the present time is the plate-type filter. This filter is sometimes referred to as the "ladder-type mechanical filter" because of its internal construction, which resembles a ladder. This filter uses flat, rectangular metallic plates as the resonator elements, and these plates are connected by wires, which form the couplers, similar to those in the disk filter discussed above. Since the operation of all mechanical filters is basically the same, a detailed discussion of the plate-type filter will not be given.

Two types of mechanical filters using cylindrical metal rods for the resonator and coupler elements are the slug-coupled and the neck-coupled filters. The former consists of a series of small-diameter resonators and large-diameter couplers, while the latter consists of large-diameter resonators and small-diameter couplers. In both types of filters, the resonators are usually one-half wavelength ($\lambda/2$) long, and the couplers are normally one-quarter wavelength ($\lambda/4$) long. The basic operating principles of mechanical filters also apply to these types of filters. A major advantage of the neck-coupled filter over the slug-coupled filter is that its heavier construction will withstand more shock and vibration; therefore, the former is more suitable for the environmental conditions encountered in such applications as airborne or mobile communications.

Low-Frequency Amplifier

From the sideband filter, the single-sideband



". . . proper impedances are obtained by mechanical damping —"

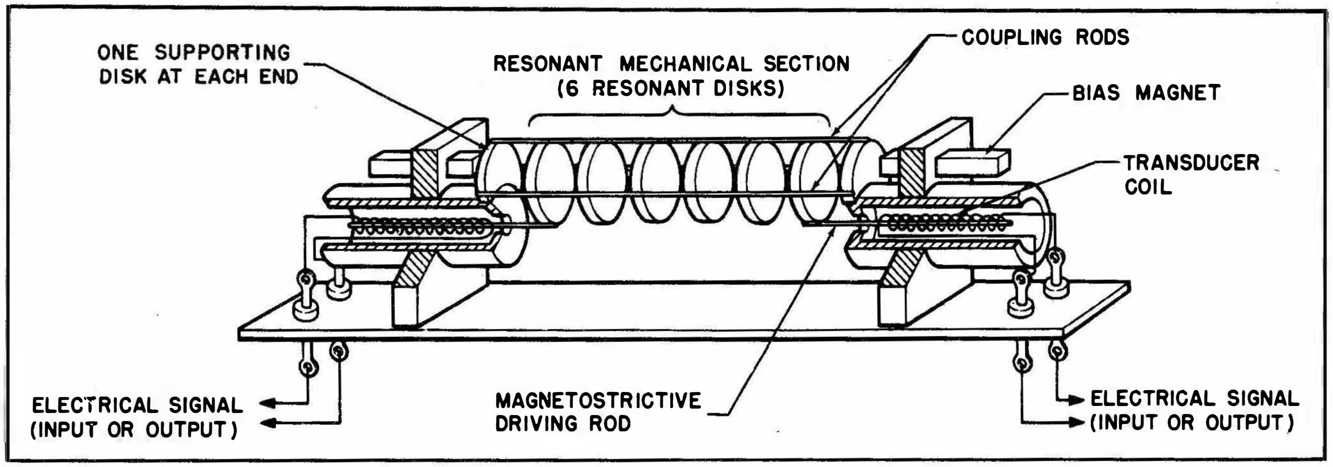


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Figure 59. Electromechanical Filter Using Metallic Disks for the Resonator Elements.

signal is transformer-coupled to a Class A r-f amplifier (V3 in figure 62). Since the purpose and circuit operation of this amplifier are conventional, a detailed discussion of the circuit is unnecessary. Figure 62 is a complete diagram of the low-frequency circuits, which are typical in SSB filter-system transmitters.

Medium-Frequency Circuits

The output of the low-frequency r-f amplifier is applied to the medium-frequency balanced mixer, where it is mixed with the 3-mc r-f carrier from the medium-frequency crystal oscillator. This balanced mixer (V4 and V5 in figure 63) is essentially a balanced-modulator circuit. The tubes in this circuit are biased to operate on the non-linear portion of their respective characteristic curves so that heterodyning (mixing) of the two input signals can take place in the circuit.

The single-sideband signal from the low-fre-

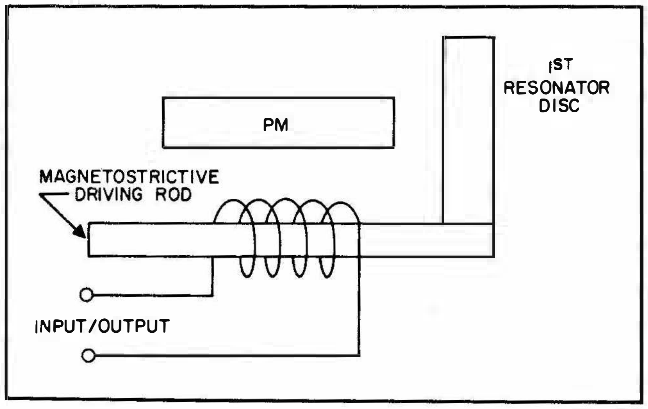

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Figure 60. Transducer Element for Disk-Type Electromechanical Filter.

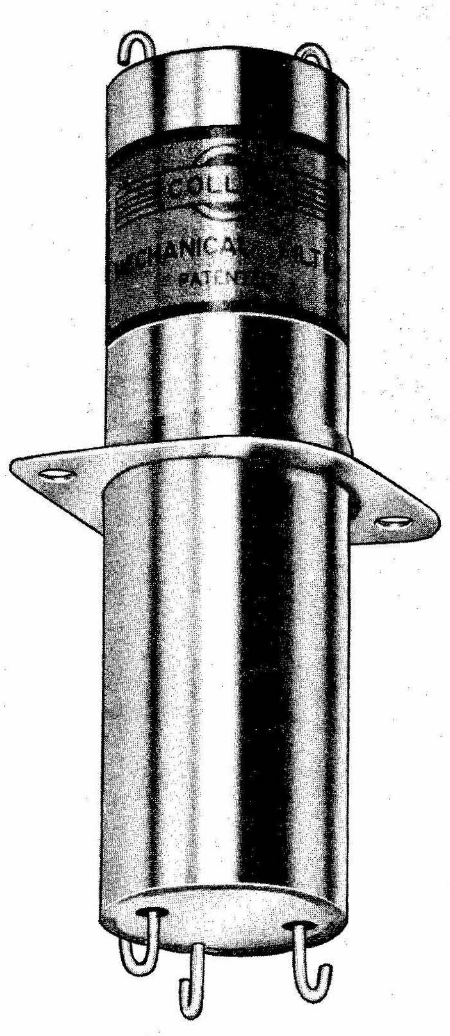

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Figure 61. Photograph of Disk-Type Electromechanical Filter.

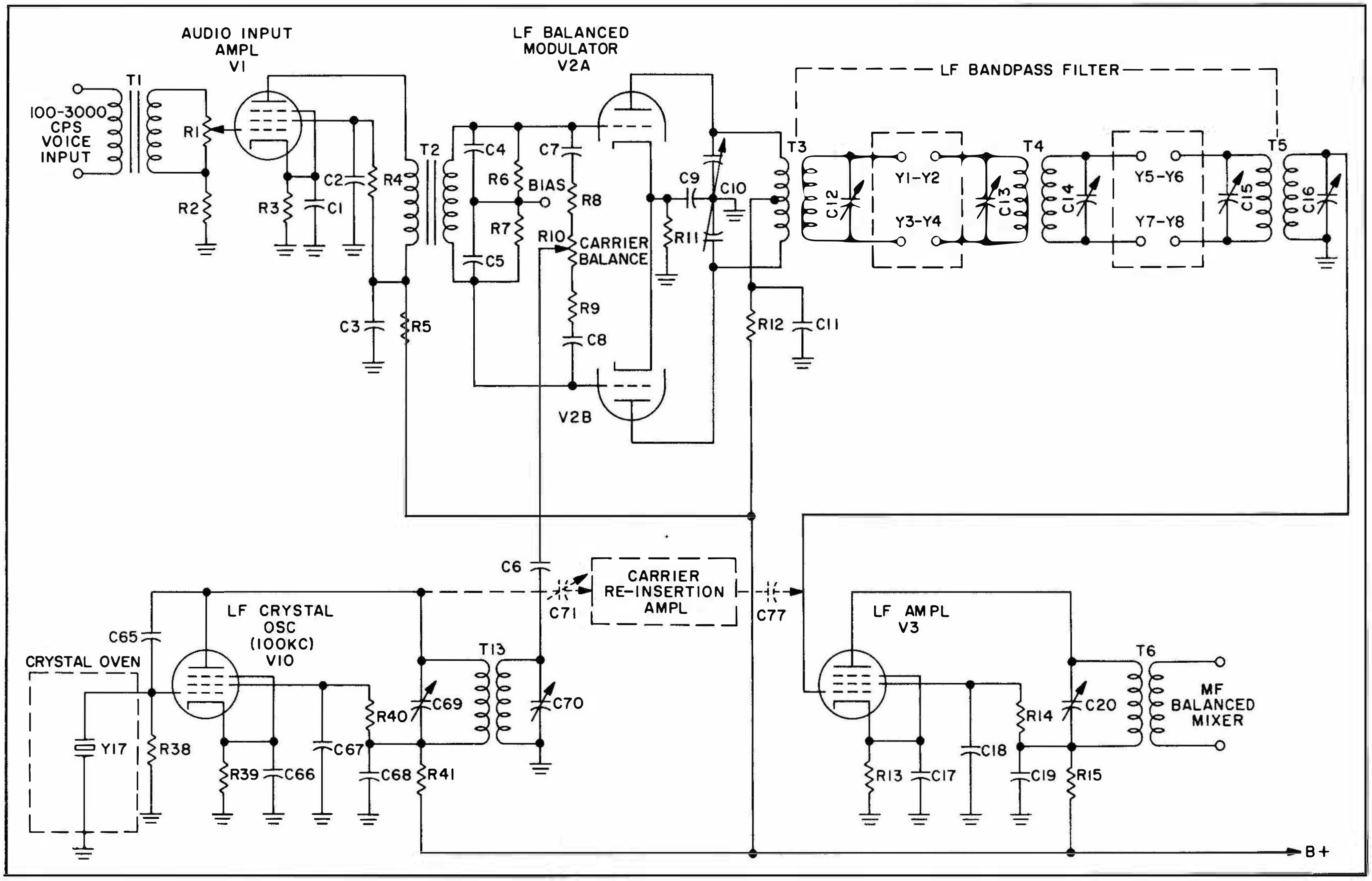

Figure 62. Low-Frequency Circuits.

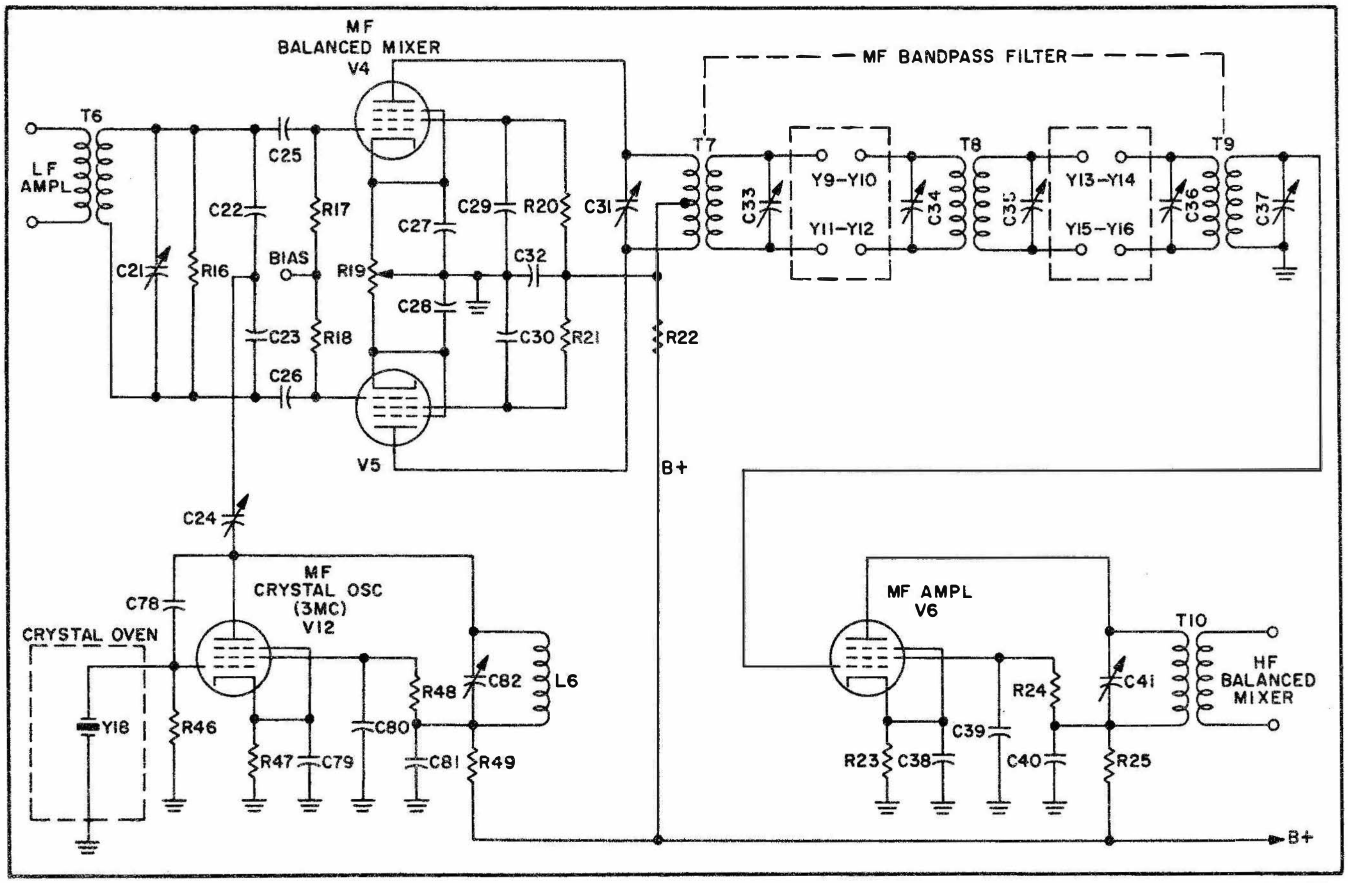

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Figure 63. Medium-Frequency Circuits.

quency amplifier is applied to the control grids of the balanced mixer in push-pull, and the r-f carrier from the medium-frequency crystal oscillator is applied to the same control grids in parallel. The r-f carrier will be balanced out in the push-pull plate circuit of the mixer. As in the balanced modulator discussed previously, the amount of carrier suppression (or balance) largely depends upon the degree of balance between the two tubes and their associated circuit components. Also, critical balance of the low-frequency SSB signal is not necessary, because the mixer plate load is tuned to pass only the sum and difference frequencies generated in the mixer stage.

Other than the higher frequencies involved, the operation of the medium-frequency bandpass filter and r-f amplifier is similar to that of the preceding low-frequency circuits. The output of the amplifier is essentially a single-sideband signal (sum frequencies only), which is applied to the high-frequency balanced mixer. In this circuit, the amplifier output signal is heterodyned with the r-f carrier from the high-frequency variable-frequency oscillator. Figure 63 is a complete diagram of the medium-frequency circuits, which are typical in SSB.

High-Frequency VFO

It will be recalled that the low-frequency and medium-frequency oscillators were tuned to one frequency only, since they were crystal-controlled. The high-frequency oscillator, on the other hand, is made variable over a fairly wide range of frequencies (6.9 to 26.9 mc) so that the transmitter output frequency will cover a considerable portion of the r-f spectrum (10 to 30 mc).

To allow the output frequency to vary over a particular range of frequencies, while still maintaining good stability in the oscillator, an electron-coupled oscillator is used as the high-frequency vfo (V13 in figure 64). The circuit is essentially a series-fed Hartley oscillator, with the feedback being inductively coupled from the cathode by the tap on the control-grid tank circuit inductance. The 6.9-to-26.9-mc range of the oscillator is covered in two bands (LO and HI), with the band-switching being accomplished by selecting the proper capacitors in the control grid and plate tank circuits. As is common in all electron-coupled oscillators, the screen grid acts as the plate of the oscillator and isolates the output plate circuit from the oscillator section of the tube. This isolation prevents load changes in the output plate

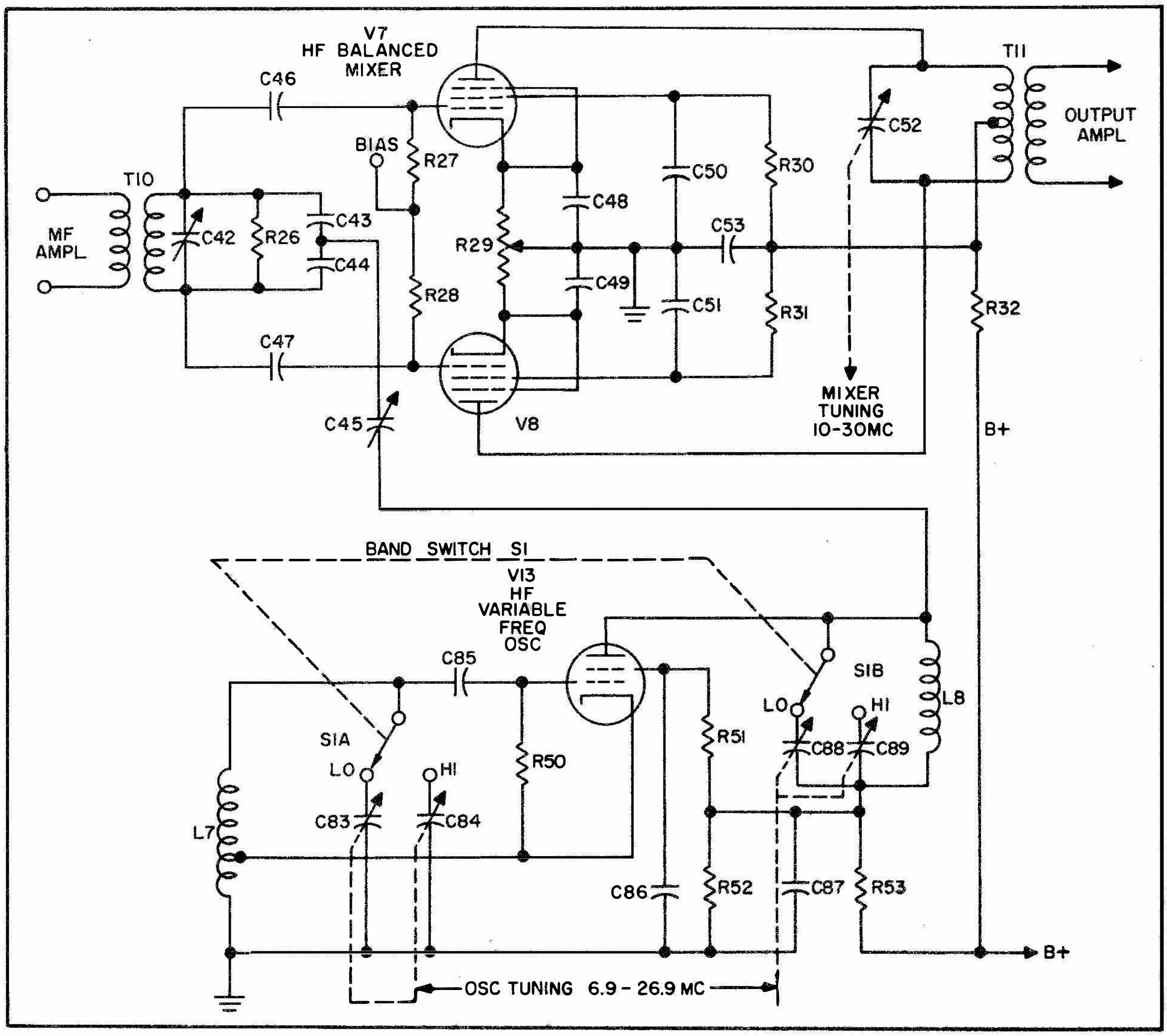


Figure 64. High-Frequency Circuits.

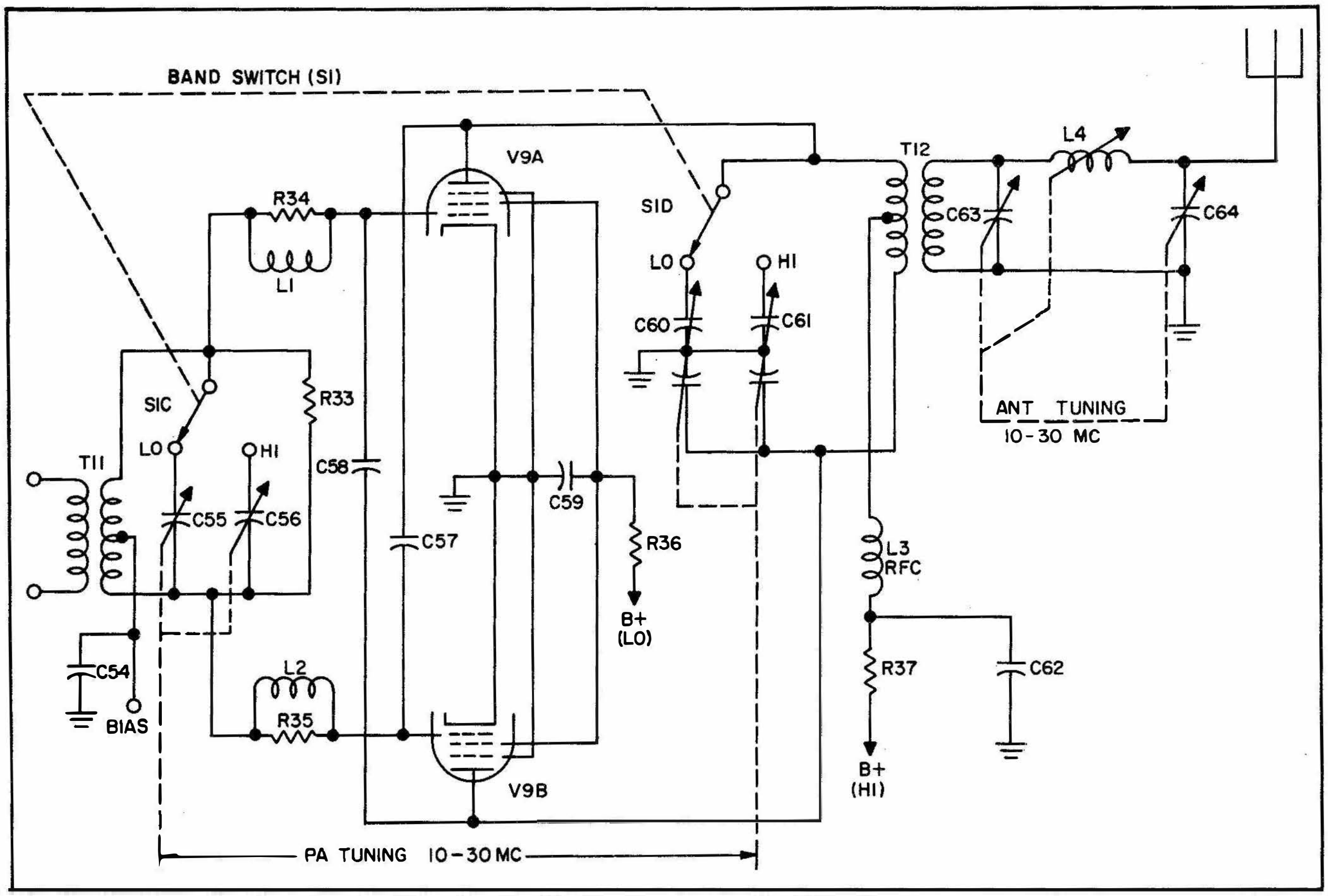
circuit from having any adverse effects on the oscillator. Improved stability of the electron-coupled oscillator is also obtained by using the proper proportions of screen-grid and plate voltages, since changes in these voltages shift the frequency of the oscillator in opposite directions. This is true, of course, only if both voltages are supplied from the same source.

The output of the vfo is applied to the high-frequency balanced mixer (V7 and V8 in figure 64), where it is heterodyned with the signal from the medium-frequency r-f amplifier to obtain the final transmitter output frequency. The balanced mixer is similar in design and operation to the one used in the medium-frequency circuits, and a detailed discussion of its operation is not necessary. Figure 64 is a complete diagram of the high-frequency circuits, which are typical in SSB transmitters.

Output Circuits

Since the modulation process was performed at a low-power level, the single-sideband signal must now be amplified in one or more power amplifiers to bring its power level up to that which is required for transmission. Any such amplification of the signal following the modulation process must be performed in amplifiers which introduce minimum distortion in the amplified signal. In addition to low distortion, the amplifiers must also have high efficiency for economy of operation in high-power circuits. The type of amplifier circuit best suited for SSB applications, therefore, is a linear amplifier having high gain.

A linear amplifier is one whose amplified output signal is proportional to its input signal. An ordinary Class A amplifier would satisfy the linearity requirement, but unfortunately this type of amplifier has low efficiency and its use as a power



TP7-1264

Figure 65. Class B Linear Power Amplifier (Output Amplifier).

amplifier in a high-power transmitter must be ruled out. On the other hand, a Class C power amplifier meeting the high-efficiency requirement is not suitable for use in voice-modulated SSB transmitters, because the Class C operation would cause excessive distortion of the audio components which are impressed on the r-f carrier. Therefore, as a compromise between the low-distortion and high-efficiency requirements, linear power amplifiers operated Class AB (AB_1 or AB_2) or Class B are used as the output amplifiers in SSB transmitter applications.

The Class B linear power amplifiers used in SSB transmitters are usually operated at "projected-cutoff"; that is, the point where the linear portion of the grid-voltage, plate-current curve is projected to zero plate current. This type of operation allows plate current to flow for approximately one half of the input cycle. Altough harmonics of the carrier are developed, with this class of amplifier, the linearity of the input signal is not appreciably distorted. Although some distortion is to be expected in all Class B amplifiers, by properly tuning these amplifiers or by using push-pull circuits, it is possible to reduce such distortion to a minimum. Since each amplifier generates some distortion, it is an advantage to keep to a minimum the number of linear amplifier stages used in a particular transmitter.

A typical Class B push-pull linear power amplifier is illustrated in figure 65. The Class B operating characteristic is obtained by applying a suitable bias to the control grids of the tubes. The resistor-inductor combinations (R34 and L1, R35 and L2 in the top and bottom control grids, respectively) are parasitic suppressors. Neutralization to prevent the occurrence of oscillations in the power amplifier is provided by capacitors C57 and C58. The tuning range of this output amplifier (10 to 30 mc) is covered in two bands (LO and HI), as was the case in the high-frequency vfo. The frequency separation between the sum frequencies (10 to 30 mc) and the difference frequencies (3.8 to 23.8 mc, or 6.2 mc lower) of the high-frequency balanced mixer output will be sufficient to insure effective "filtering" action at this point. With normal tuning, the power amplifier and antenna circuits will pass only the sum frequencies and reject the difference frequencies, since the latter frequencies will not be within the

10-to-30-mc tuning range of these output circuits. For this reason a bandpass filter is not required following the high-frequency balanced mixer.

Pilot Carrier Re-insertion

The SSB transmitter under discussion is of the type which employs the suppressed-carrier system (SSSC), in which the carrier frequency is suppressed as much as practicable in the transmitter. The amount of carrier suppression obtained in this manner is approximately 30 to 50 db with reference to the level of the sideband power. The accuracy with which transmitter and receiver in an SSSC system can be maintained on a given frequency is totally dependent upon the stability and accuracy of the oscillator circuits employed, since afc cannot be used in this type of system. It can be seen, therefore, that any appreciable frequency shift in either the transmitter or receiver oscillators will result in adjacent-channel interference. In order to reduce the exacting requirements of the carrier oscillator stability in the receiver and transmitter and prevent adjacent-channel interference, a small amount of carrier is transmitted at a reduced power level. The level of this reduced carrier, or pilot carrier, is usually 10 to 20 db below that of the sideband power. When the pilot-carrier method of transmission is used, the SSB receiver must be equipped with an a-f-c system to lock-in the receiver local oscillator frequency with the pilot carrier.

From the diagram of the pilot-carrier re-insertion circuits (figure 66) it can be seen that the pilot carrier is applied to the control grid of the low-frequency amplifier from the low-frequency crystal oscillator (100-kc osc) through the carrier re-insertion amplifier, V11. This amplifier is a conventional Class A amplifier and will operate only when the carrier re-insertion switch (S2) in the cathode circuit is closed. Thus the use of the pilot carrier can be selected whenever required. The pilot-carrier amplitude control (R45) controls the level of the carrier with respect to the peak-sideband power.

PHASE-SHIFT SYSTEM DETAILED CIRCUIT ANALYSIS

General

In the phase-shift SSB transmitter the audio input amplifier, r-f oscillator, balanced modulator, and linear power amplifier circuits serve the same basic functions as their equivalent circuits in the filter-system SSB transmitter. The principal difference between the two systems is in the method used to produce the desired sideband and reject the undesired sideband. Instead of using crystal-lattice or electromechanical filters to accomplish this sideband rejection, the phase-shift system operates on the principle of removing the undesired sideband by a balancing or cancelling process based upon the phase relationships between the sideband signals generated in two separate balanced modulators. As an illustration, in figure 43, the audio modulating signal from the input amplifier is applied directly to one balanced modulator, and through a wide-band audio 90-degree phase-shifting network to a second balanced modulator. The r-f carrier is applied directly to the second balanced modulator, and through a narrow-band r-f 90-degree phase-shifting network to the first balanced modulator. Since the purpose

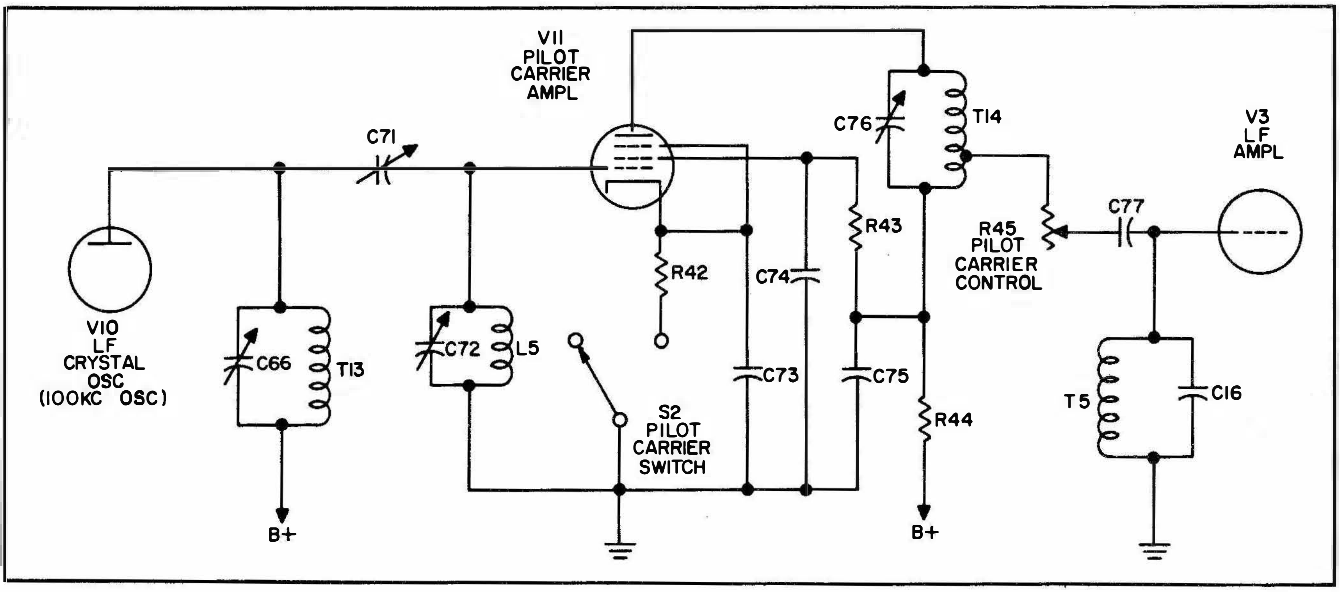
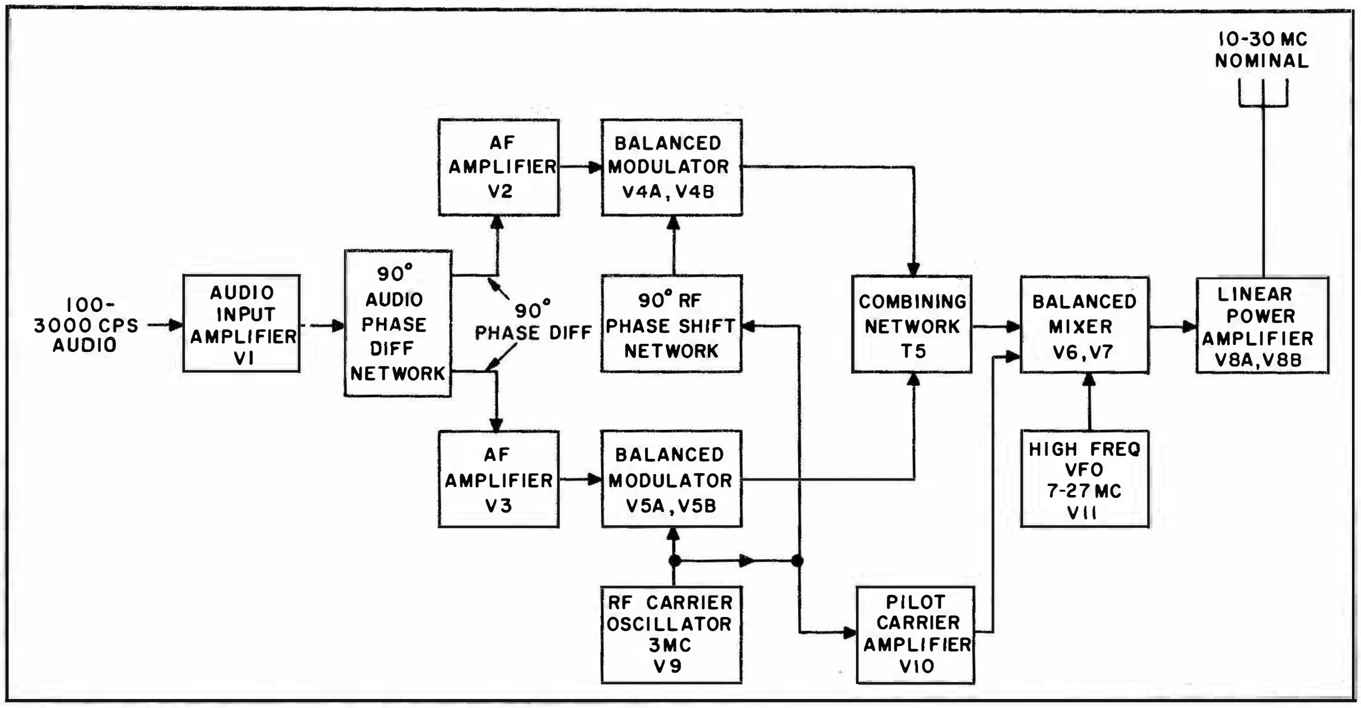


Figure 66. Pilot Carrier Re-insertion Circuits.



TP7-1266

Figure 67. Block Diagram of Practical Phase-Shift System SSB Transmitter.

of a balanced modulator is to suppress the r-f carrier, the output from each modulator contains only the generated upper and lower sideband frequencies. These sideband signals are then mixed in a combining circuit in such a manner that the cancellation mentioned above will be obtained.

Although this simple phase-shift system transmitter (figure 43) will work in theory, it is not a practical arrangement. In actual practice it is difficult to obtain a wide-band audio 90-degree phase-shifting network that will hold the amplitude and phases of all the individual audio frequencies constant over the entire band to be passed, namely, 100 to 3000 cps. Such a network is necessary for the phase-shift SSB transmitter to provide the proper degree of undesired sideband cancellation. A second reason why the system illustrated in figure 43 is not practical is the fact that only one transmitter output frequency is provided. For a practical voice-modulated transmitter, provision must be made for multi-channel operation.

Figure 67 shows the block diagram of a practical phase-shift system SSB transmitter. In this circuit arrangement the output signal of the audio phase-shifting network has a phase difference of 90 degrees, and it is this phase-difference signal that is applied to the balanced modulators through Class A a-f amplifiers acting as isolation circuits. Provisions are made for multi-channel operation by including a high-frequency vfo and a balanced-mixer stage following the balanced modulators. A pilot-carrier insertion circuit is also included so that the carrier may be transmitted at reduced power for pilot-carrier operation.

It will be recalled that the degree to which the undesired sideband is attenuated in a filter-system SSB transmitter depends directly upon the design of the filter network. The degree of undesired sideband attenuation in the phase-shift SSB transmitter, on the other hand, depends primarily upon the function of the wide-band audio and narrow-band r-f 90-degree phase-shift networks, with the audio network being the more critical of the two. Maximum attenuation of the undesired sideband occurs when the phase difference of the audio network is exactly 90 degrees, and less attenuation is obtained when the phase difference deviates from 90 degrees. However, this deviation is not usually greater than one or two percent in a properly designed network, with the sideband attenuation averaging 35 to 40 db over the complete audio band. Since phase-shift networks are so important to this type of SSB transmitter, their design considerations will be discussed in greater detail in the following paragraphs.

Phase-Shifting Networks

Audio Phase-Shift Networks

There are two general classifications of audio phase-shift networks—active and passive—depending on whether or not the network employs vacuum tubes (or transistors). An active network is one in which power, other than the input signal is applied. The response of an active network (figure 68A) is acceptable over a relatively wide audio band (approximately 50 to 5500 cps), in which the amplitude and 90-degree phase difference remain essentially constant. However, since this network employs vacuum tubes, it has the disadvantages of large physical size and power consumption. The input signal to the active network is applied in single phase, and the output is a two-phase balanced signal consisting of four individual voltages having equal amplitude and an over-all 90-degree phase relationship. The actual phase-shifting elements are R-C circuits which are excited by voltages phased 180 degrees and supplied by an associated phase-inverter stage. The individual R-C circuits and phase inverters are tuned to specific frequencies, to give the over-all network the desired response characteristics. For optimum operation, the plate and cathode resistors of each phase-inverter stage should be matched to within one-half percent of each other and should maintain this exacting tolerance for long periods of time. Likewise, the phase-shifting R-C circuits should have the same high order of stability and precise tolerance. The power supply for this active network should be highly regulated to prevent adverse effects on the network due to changes in power-supply loading. Transistors can be substituted if the circuit is altered so that the proper phase shift is maintained.

A passive audio phase-shift network (one which does not use vacuum tubes or one in which the input signal provides the operating power) is illustrated in figure 68B. The response of the passive network, in comparison to the active network, is acceptable over a relatively narrow portion of the audio spectrum (approximately 150 to 2700 cps). This network consists of six individual R-C circuits in a lattice arrangement, and requires a balanced push-pull input signal. Similar to the R-C circuits in the active network, the individual R-C circuits in this network are tuned to specific frequencies, to give the over-all network the desired response characteristics—constant amplitude and constant 90-degree phase difference over the entire audio modulating frequency band. The 90-degree phase difference is obtained through individual 45-degree lead-lag phase shifts, which, when combined, produce the required 90-degree shift of the network.

R-F Phase Shift Networks

Since the r-f oscillator generates an essentially fixed frequency, the narrow-band r-f 90-degree phase-shift network can be relatively simple in design and construction. Two such phase shifters are illustrated in figure 69.

The double-tuned over-coupled transformer r-f phase-shifter (figure 69A) operates on the principle that there will be approximately a 90-degree shift in secondary voltage, with respect to the primary voltage, when both the primary and the secondary are at resonance and the coefficient of coupling slightly exceeds the critical value. This action will produce the required 90-degree shift in the r-f carrier signal so that it can be applied to the separate balanced modulators in quadrature.

The R-L-C phase-shift network (figure 69B) employs separate R-L and R-C circuits to obtain a combined phase difference of 90 degrees at the carrier frequency. When the inductive and capacitive reactances are equal to their associated series resistances (at the carrier frequency), the R-L leg will present a 45-degree lead and the R-C leg will present a 45-degree lag in the r-f signal. The combination of the two individual 45-degree shifts will produce the desired 90-degree over-all network phase shift in the r-f carrier.

Audio Circuits

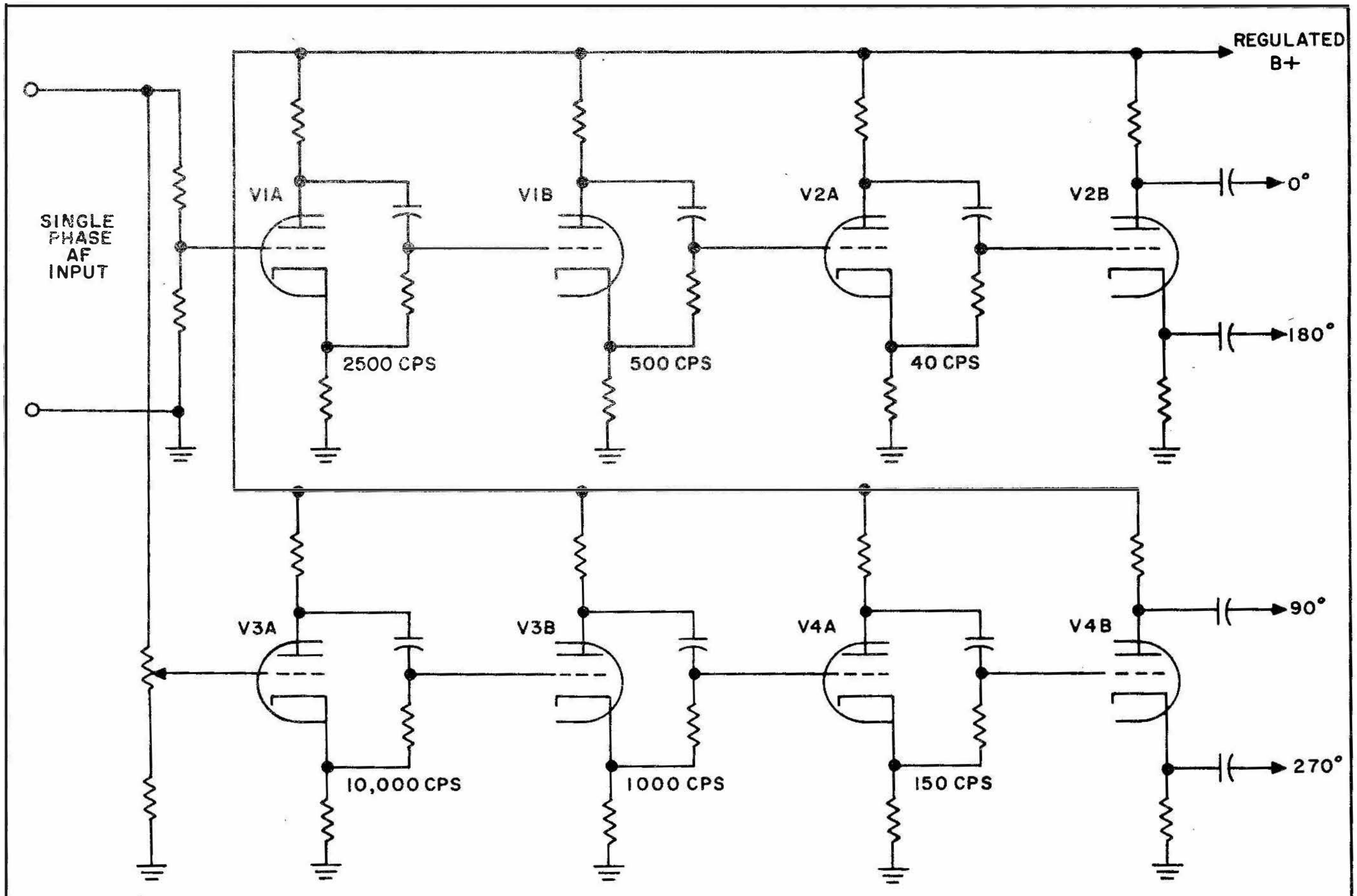
Figure 70 shows a complete diagram of the audio circuits. The input amplifier, V1, is a Class A audio amplifier, and serves to amplify the input speech frequencies. It is similar in operation to the input amplifier in the filter-system SSB transmitter.

Transformer T2 couples the amplified speech input frequencies to the wide-band audio 90-degree phase-difference network in push-pull. As explained previously, the individual R-C combinations in this network (R8 and C4, R9 and C5, R10 and C6, R11 and C7, R12 and C8, and R13 and C9) are tuned to separate frequencies in the audio band. This arrangement introduces separate 45-degree lead-lag phase shifts, which are combined to insure that the over-all network produces the required constant amplitude and constant 90-degree phase difference over the desired audio band to be passed. The 90-degree phase-difference signal is then applied to the control grids of separate Class A audio amplifiers, V2 and V3. Potentiometer R15 in the upper amplifier control grid provides balance of the audio signal at the input to the amplifiers.

The purpose of these audio amplifiers is twofold: they amplify the audio signal, thus compensating for signal losses which occur in the phase-difference network; and they provide isolation of the balanced modulators from the phase-difference network. Output transformers T3 and T4 couple the 90-degree phase-difference audio signal to the separate balanced modulators, where it is used to modulate an r-f carrier signal, which is also applied to the balanced modulators in quadrature.

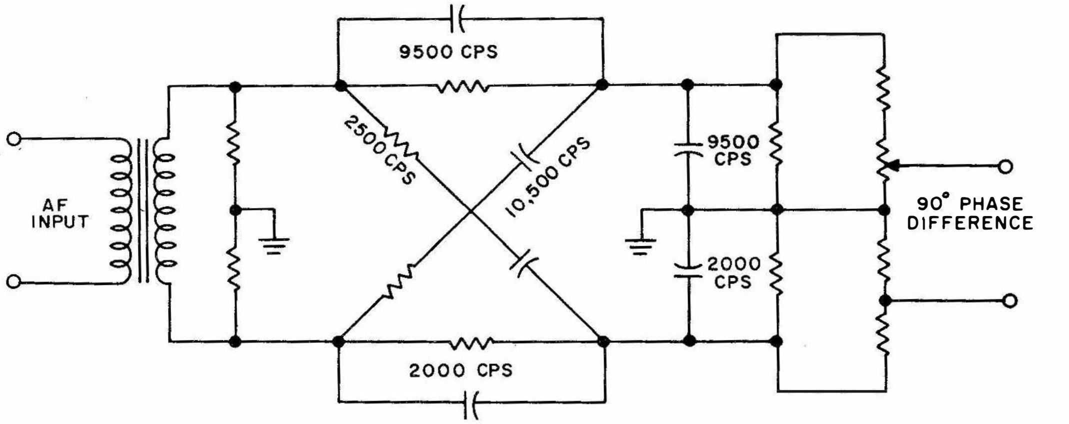
R-F Carrier Oscillator and 90-Degree Phase-Shift Circuits

The crystal-controlled r-f oscillator in the phase-



NOTE: FREQUENCIES LISTED ARE TYPICAL AND NOT NECESSARILY EXACT.

A. ACTIVE 90° AUDIO PHASE SHIFT NETWORK.



NOTE: FREQUENCIES LISTED ARE TYPICAL AND NOT NECESSARILY EXACT

B. PASSIVE 90° AUDIO PHASE DIFFERENCE NETWORK

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Figure 68. Typical Audio Phase-Shift Networks.

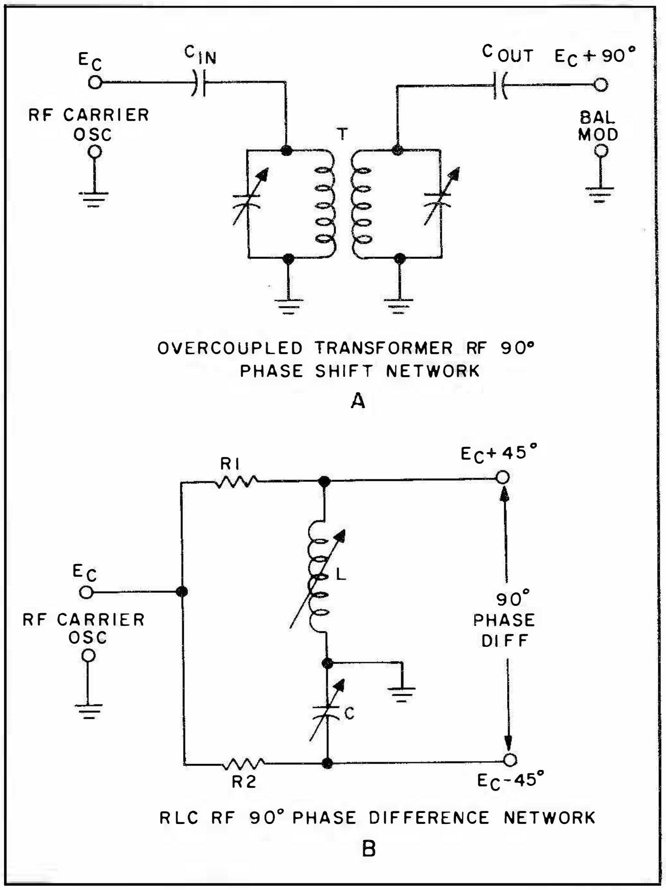

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Figure 69. Typical R-F Phase-Shift Networks.

shift SSB transmitter (V9 in figure 71) serves the same function as its equivalent in the filter-system SSB transmitter. However, since filters that eliminate the undesired sideband are not required here, and consequently the critical design features of such sideband filters are eliminated, this r-f oscillator is operated at a much higher frequency (3 mc) than its counterpart (100-kc r-f carrier oscillator) in the filter system. In fact, the modulation process in the phase-shift transmitter can be performed at the actual transmitted frequency in applications requiring only single-output-frequency operation.

An r-f 90-degree phase-shift network using an over-coupled transformer (T8) is located in the plate circuit of the 3-mc carrier oscillator. The windings of this transformer are slightly over-coupled so that at the carrier frequency the secondary voltage will be shifted 90 degrees with respect to the primary voltage. The 90-degree shifted carrier is coupled through capacitor C47 from the secondary of T8 to the control grids of the top balanced modulator (V4A and B); and the in-phase carrier is coupled through capacitor C46 from the primary of the same transformer to the control grids of the bottom balanced modulator (V5A and B). These r-f carrier voltages are thus applied to the separate balanced-modulator circuits in quadrature.

Balanced Modulators and Combining Circuits

Individually the balanced-modulator circuits perform the functions of generating upper and lower sideband frequencies and suppressing their individual r-f input signals. By connecting the plate circuits of the separate modulators in parallel to a common balanced tuned load, T5, it is possible for these modulators to perform the added function of cancelling the undesired sideband.

Assume first that only the quadrature 3 mc r-f carrier signals are applied to the balanced-modulator system. These signals are fed in parallel to the separate modulators through balance controls R22 and R26, which provide balance of the r-f input signals at the respective modulator control grids. The parallel-applied carrier signals will be balanced out in the push-pull plate circuit of the separate modulators in the same manner as in the filter system previously described. Therefore, when only the r-f carrier is present on the balanced modulators in either the filter or phase-shift SSB transmitters, the results produced by the modulators will be the same—no output. However, since the plate circuits of the separate modulators in the phase-shift system are connected in parallel to the common tuned load, these modulators will cancel the undesired sideband when both the r-f carrier and the audio modulating signals are present in the system. The primary winding of the modulator tuned load, T5, is a slug-tuned split-winding which provides a means of balancing the entire balanced-modulator system.

Referring to figure 72, assume that the r-f carrier and audio modulating signals are both being applied to the modulators. Although the 100—3000-cps audio is applied in quadrature to the balanced-modulator system through input transformers T3 and T4, it is applied in push-pull from the individual secondary windings of these transformers to the control grids of the separate modulators. With respect to a zero-degree horizontal reference signal (f_a), the phase of the audio signal on the primary of T3 has a 45-degree lag ($f_a - 45°$) while the phase of the audio on the primary of T4 has a 45-degree lead ($f_a + 45°$). Since there is a 180-degree difference in polarity between opposite ends of a transformer winding, the signal at the top of the T3 secondary will be $f_a + 135°$ while the signal at the bottom of the T3 secondary will be $f_a - 45°$. Similarly, the signals at the top and bottom of the secondary of T4 will be $f_a - 135°$ and $f_a + 45°$, respectively. These

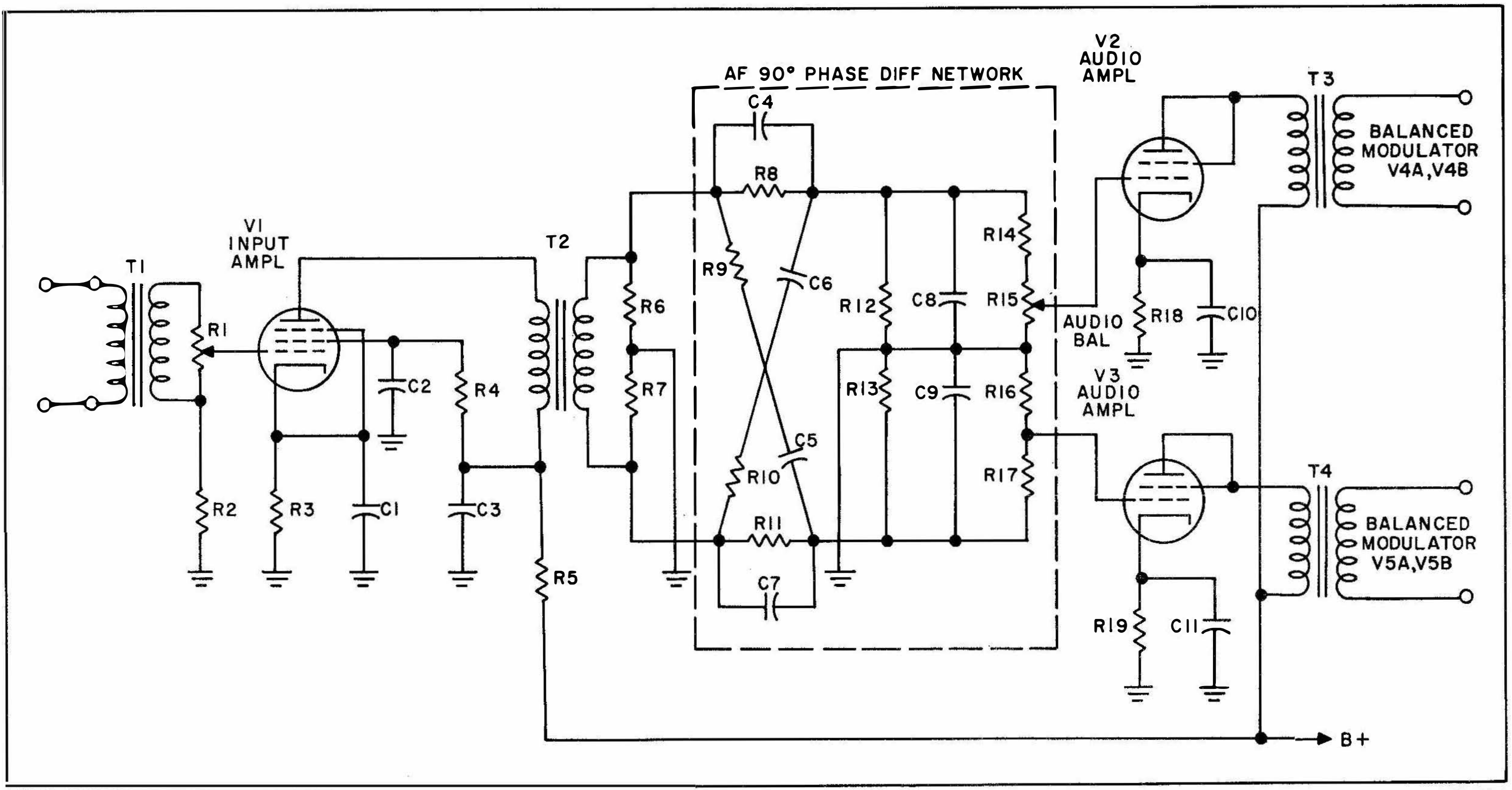

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Figure 70. Audio Input and Phase-Shifting Circuits.

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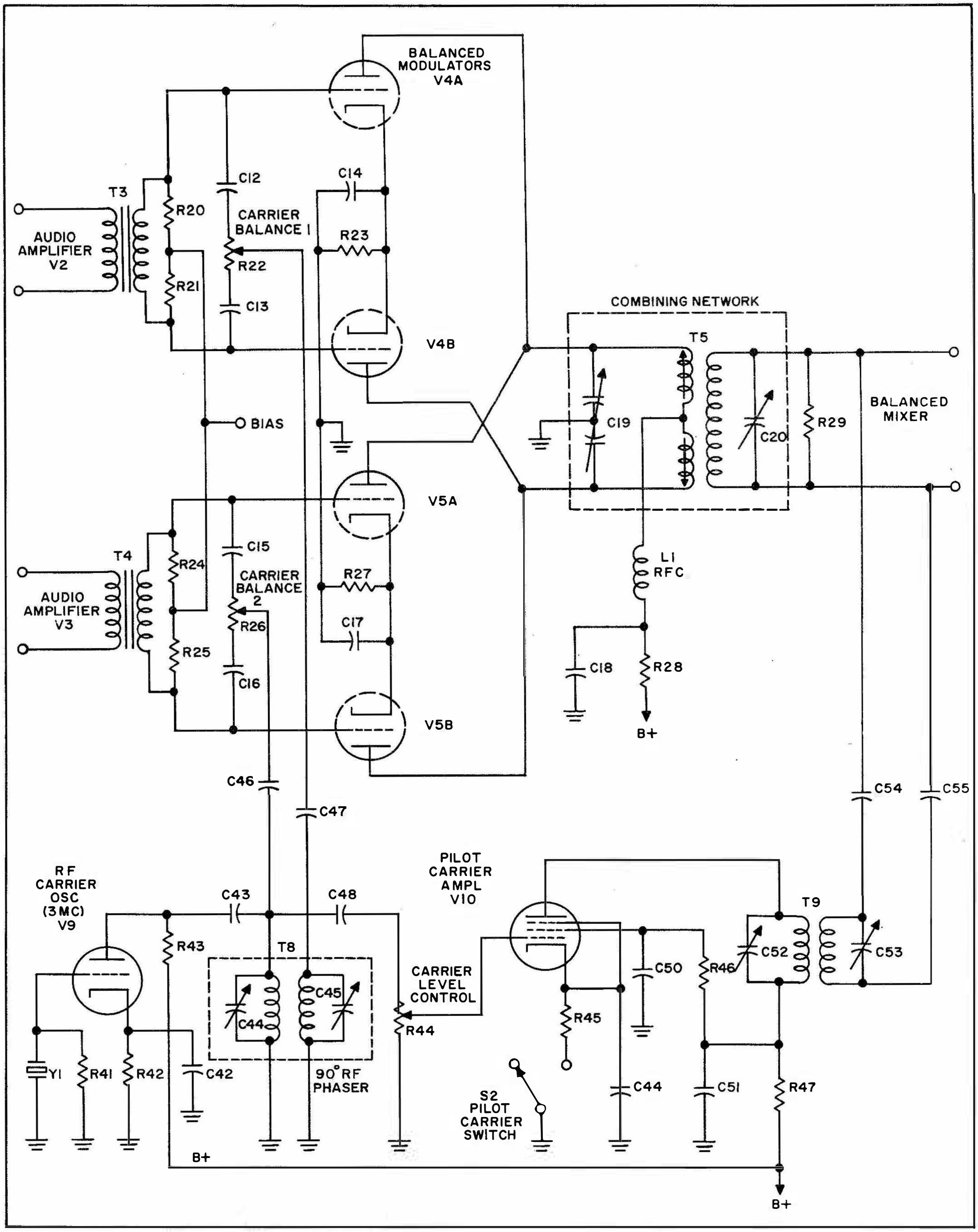

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Figure 71. Balanced Modulators, Combining Network, R-F Oscillator, R-F Phaser, and Pilot Carrier Amplifier.

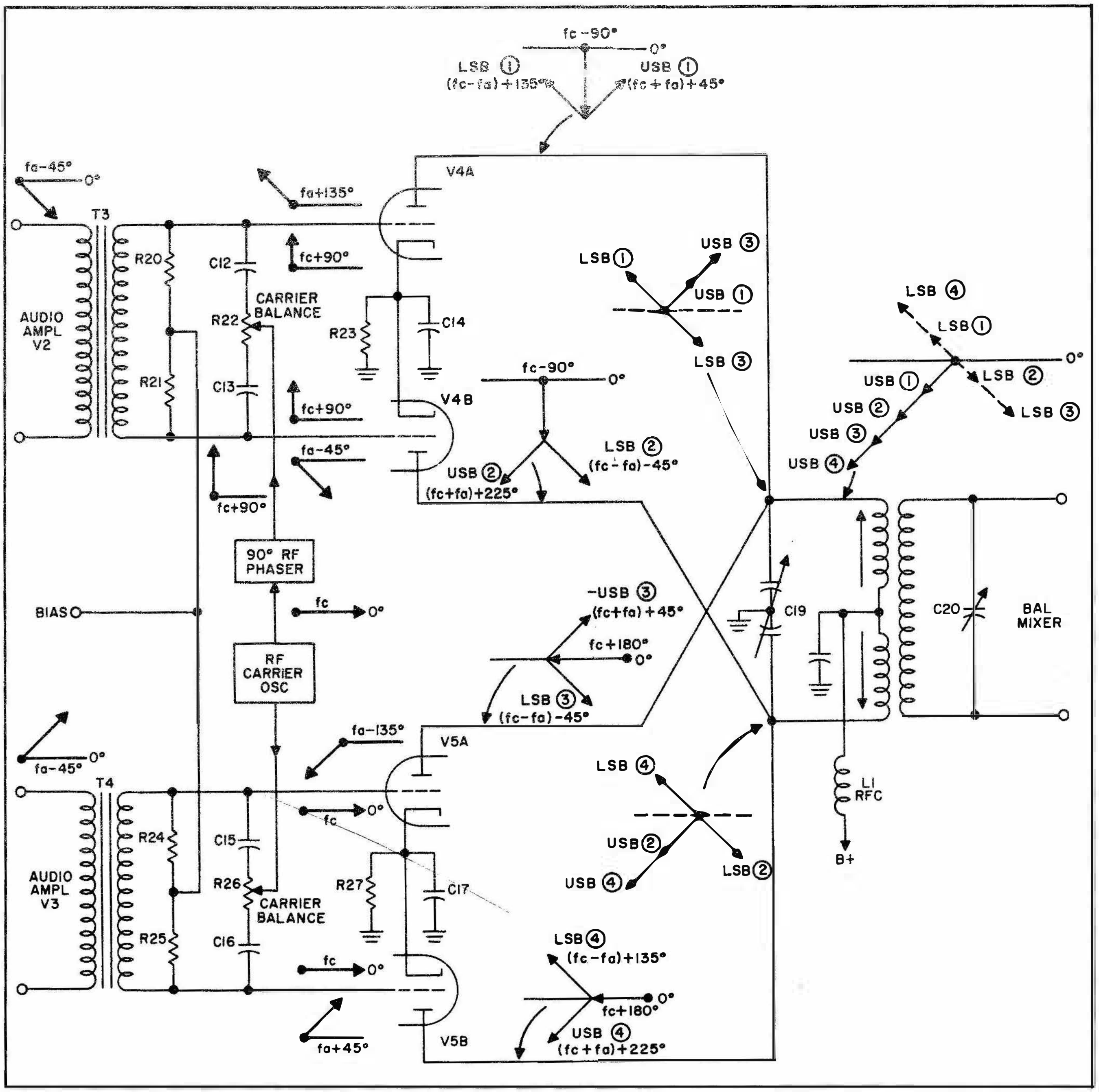

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Figure 72. Balanced Modulators, Showing Vector Analysis of Sideband Generation.

push-pull audio signals are applied to the control grids of the separate modulators (V4A and B, V5A and B), as indicated in figure 72.

As stated previously, the 3-mc r-f carrier is also applied to the balanced-modulator system in quadrature. These r-f signals, f_c and $f_c + 90°$, are applied to the control grids of the separate modulators in parallel, as indicated in figure 72. It can be seen, therefore, that there will be two input signals (3-mc r-f carrier and 100—3000-cps audio modulation) applied to each individual modulator tube. Since the balanced modulators are operated on the non-linear portion of their respective characteristic curves (usually Class B), a mixing of the two input signals will occur in each individual modulator tube. This mixing action will cause the individual modulator tubes to generate upper and lower sideband frequencies, while the push-pull pairs of modulator tubes (V4A and B, V5A and B) will suppress their respective carrier input signals. The sideband frequencies thus generated in each individual modulator tube will combine in the common tuned plate load (split primary of T5) in such a manner that the upper sidebands will aid each other and the lower sidebands will oppose each other; hence, the lower-sideband signal will be cancelled.

The processes of upper and lower sideband generation, upper sideband addition, and lower sideband cancellation are illustrated vectorially in figure 72. A simplified mathematical analysis of the sideband generation in each modulator tube is presented below.

V4A; Upper Sideband, USB 1:
$(f_c + 90^\circ) + (f_a + 135^\circ) = (f_c + f_a) + 45^\circ$
Lower Sideband, LSB 1:
$(f_c + 90^\circ) - (f_a + 135^\circ) = (f_c - f_a) + 135^\circ$
V4B; Upper Sideband, USB 2:
$(f_c + 90^\circ) + (f_a - 45^\circ) = (f_c + f_a) + 225^\circ$
Lower Sideband, LSB 2:
$(f_c + 90^\circ) - (f_a - 45^\circ) = (f_c - f_a) - 45^\circ$
V5A; Upper Sideband, USB 3:
$f_c + (f_a - 135^\circ) = (f_c + f_a) + 45^\circ$
Lower Sideband, LSB 3:
$f_c - (f_a - 135^\circ) = (f_c - f_a) - 45^\circ$
V5B; Upper Sideband, USB 4:
$f_c + (f_a + 45^\circ) = (f_c + f_a) + 225^\circ$
Lower Sideband, LSB 4:
$f_c - (f_a + 45^\circ) = (f_c - f_a) + 135^\circ$

From figure 72 it can be seen that the upper sidebands from V4A and V5A will combine in the top half of the T5 primary, and that the lower sidebands from these same two tubes will oppose each other at this point. In the bottom half of the T5 primary the upper sidebands generated in V4B and V5B will combine, and the lower sidebands generated in these two tubes will oppose each other. Thus the four individual lower sidebands will be cancelled in the primary winding of T5, and the four individual upper sidebands will add to each other and appear as a single upper-sideband signal in the secondary of this transformer.

Since the r-f carrier input to the modulators is 3 mc, and the audio modulating frequencies are in a band from 100 to 3000 cps; the single-sideband output of the balanced-modulator system will contain frequencies from 3.0001 mc to 3.003 mc. This upper-sideband signal is applied in push-pull to the control grids of the balanced mixer, where it is heterodyned with an r-f carrier (from the high-frequency vfo) to provide the desired transmitter output frequency.

The selection of either upper or lower sideband for transmission is easily accomplished in the phase-shift system simply by reversing the phases of either the audio or r-f signals to the balanced modulators. To illustrate this fact, refer to figure 72 and assume that the phases of the audio input signals are reversed while those of the r-f input are unaffected. The $f_a + 45^\circ$ audio signal will then be applied to balanced modulator V4, and the $f_a - 45^\circ$ signal will be applied to balanced modulator V5. Through a vectorial analysis it can be seen that in this case the upper sidebands will cancel in T5 and the lower sidebands will appear in the output of the modulators. Similarly, if it is assumed that the r-f phases are reversed, and the audio phases are unaffected, the same results will be obtained. Thus one advantage of a phase-shift transmitter over a filter-system transmitter is the ease of selecting either sideband when desired.

High-Frequency and Output Circuits

As stated previously, it is possible for the modulation process to be performed at the actual transmitter output frequency in the phase-shift SSB transmitter. Although this fact is true, it is not very practical for a voice communication transmitter. In this type of transmitter it is usually desirable to have multi-channel (variable output frequency) operation, and, if the modulation process were to be performed at the output frequency by using a variable-frequency oscillator for the r-f oscillator, the tuning of the narrow-band r-f 90-degree phase-shift network would also have to be made variable over the entire transmitter output frequency range. Tuning of the r-f 90-degree phaser over such a wide range of frequencies would not produce the best operating characteristics desired from this type of network. It would also be very difficult to maintain exactly 90 degrees of phase shift using such a variable component. Therefore, the most practical and most common arrangement of a phase-shift SSB transmitter is to effect the modulation at a relatively low rf, and then heterodyne the modulator output with a variable high-frequency oscillator to obtain the desired transmitter output frequency range.

Figure 73 shows the high-frequency and output circuits. The balanced mixer (V6 and V7), high-frequency vfo (V11), and linear power amplifier (V8A and B) have essentially the same circuitry, operation, and functions as their equivalents in the filter-system SSB transmitter.

For a nominal 10-to-30-mc output frequency range from the transmitter, the high-frequency vfo in the phase-shift transmitter is operated over a frequency range from 7 to 27 mc, as compared to a frequency range from 6.9 to 26.9 mc for the corresponding oscillator in the filter system. The reason for the .1-mc change in oscillator frequencies is quite evident—only one heterodyning process is employed in the phase-shift system. Instead of the modulation process being performed at 100 kc (.1 mc) as was the case in the filter system, the modulation is effected at 3 mc directly, thus eliminating the first heterodyning process from 100 kc to 3 mc.

The high-frequency vfo is a modified series-fed Hartley circuit operating as an electron-coupled

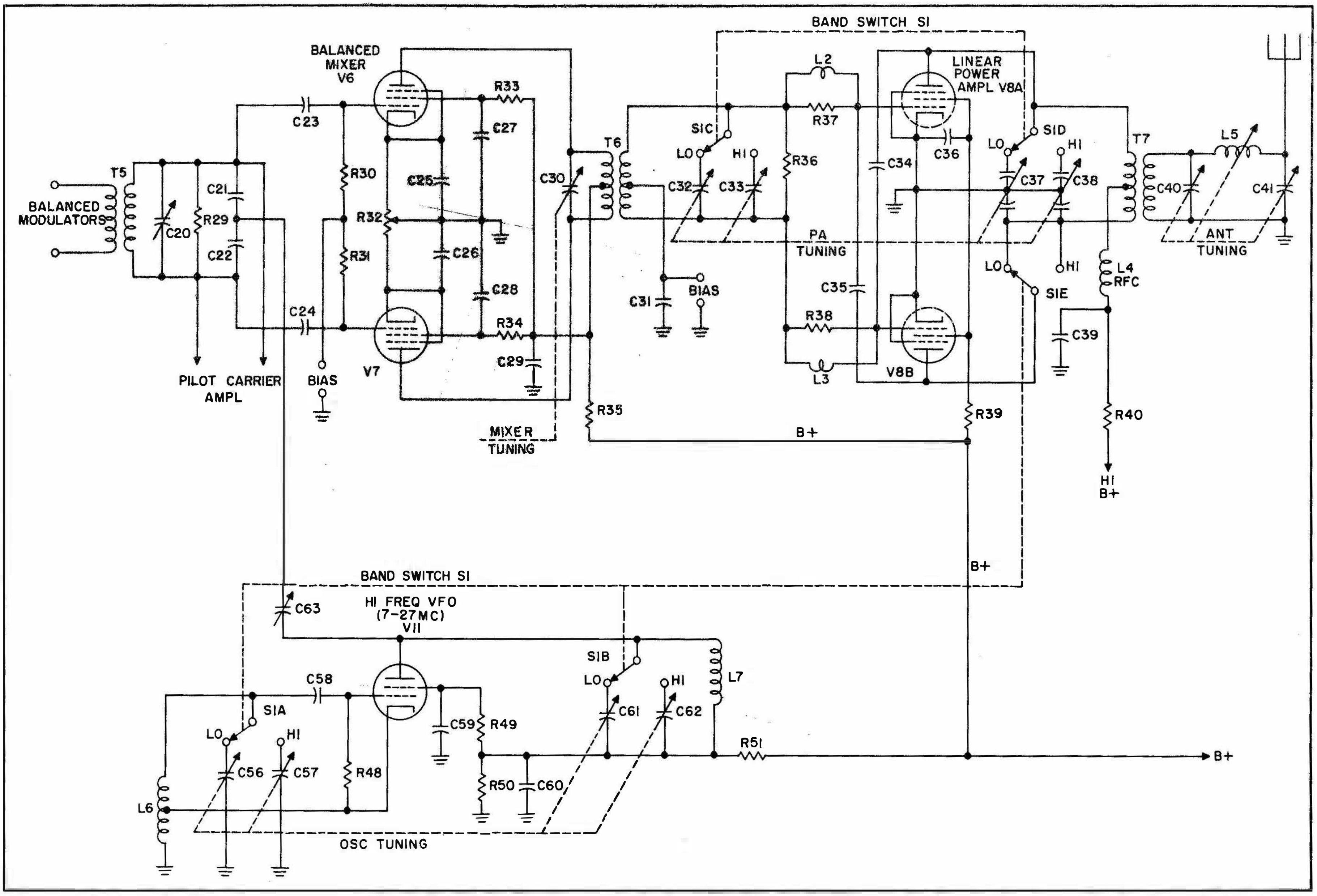

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Figure 73. High-Frequency VFO, Balanced Mixer, and Linear Power Amplifier.

oscillator. The 7-to-27-mc range of this oscillator is covered in two bands (HI and LO), with the desired band being selected by one section of the band selector switch, S1. The 7-to-27-mc output of the high-frequency vfo is coupled through capacitor C63 and applied in parallel to the control grids of the balanced mixer.

The 3.0001-to-3.003-mc upper sideband frequencies from the balanced-modulator circuits are transformer-coupled through T5 and applied to the same control grids of the balanced mixer in push-pull. The mixer circuit will cause a heterodyning action to accur between the two input signals, thus generating sum and difference frequencies. Since the mixer is a balanced circuit, the parallel input 7 to 27 mc r-f carrier signals will be balanced out in the push-pull plate circuit. The plate load (C30 and T6) is tuned to pass only the 10-to-30-mc (nominal) sum frequencies generated in the mixer. This tuning will effectively provide filtering of the 4-to-24-mc (nominal) difference frequencies, which are also generated in the balanced mixer stage.

The 10-to-30-mc sum frequencies are applied in push-pull to the control grids of the linear power amplifier (V8A and B), where the signal is now amplified to the power level desired for transmission. The 10-to-30-mc tuning range of this circuit is also covered in two bands, similar to the high-frequency vfo. Since the desirable characteristics of this amplifier are low distortion and optimum power gain, the linear power amplifier is usually operated Class B, Class AB, or Class AB_2, depending on the tubes used.

Pilot-Carrier Re-insertion

Provision is made for the transmission of a pilot carrier at reduced level through the pilot-carrier amplifier, V10 in figure 71. A portion of the 3-mc r-f oscillator signal is applied directly to the pilot-carrier amplifier control grid through capacitor C48 and "carrier level" control R44. The 3-mc carrier is amplified in V10 (a conventional Class A r-f amplifier), and the output is transformer-coupled by T9, and applied in push-pull through capacitors C54 and C55 to the control grids of the balanced mixer. A pilot-carrier signal will be transmitted only when the pilot-carrier switch, S2, in the cathode of the carrier amplifier is closed.

Comparison of Filter and Phase-Shift SSB Transmitters

Since the audio input and linear power amplifiers used in the filter and phase-shift transmitters are essentially the same, the comparison of the two systems is centered around the method of generating the single-sideband signal.

It will be recalled that in the filter system the modulation process must be performed at a relatively low rf, because of the severe requirements of the filter necessary to suppress the undesired sideband at higher frequencies. The filters used must be highly selective so that they pass only the desired sideband and provide adequate attenuation of all other frequencies. Usually the filters are either high-Q quartz crystals used in a lattice arrangement, or electromechanical filters, both of which are relatively costly. Since the single-sideband signal in this system is generated at a low radio frequency, the signal must be heterodyned to the transmitter operating frequency. If the operating frequency of the transmitter is very much higher than that of the carrier, more than one heterodyning process is usually required. Selection of either sideband when desired is usually quite difficult to obtain in this system. To change the sideband selection, it is necessary to change the filter, or to change the operating frequency of the r-f oscillator so that it is above instead of below the bandpass of the existing filter. Switching may be provided to facilitate either of these changes. However, in the first case, this requires the use of two filters, only one of which will be in the circuit at a time, adding to the cost of the equipment. In the second case, frequency error may be introduced or oscillator stability may be impaired, because of the complex tuned circuitry and switching involved.

On the other hand, adjustment of the filter-system transmitter circuits is not critical except for the initial design and adjustment of the filter itself. However, once the filter is adjusted (at the factory) to its proper operating frequency, it will retain this adjustment for long periods. Therefore, the filter-system SSB transmitter is highly stable.

In the phase-shift transmitter the modulation process can be performed at the transmitter operating frequency, thus eliminating the need for any heterodyning, if only single-frequency output is desired. The modulation can also be performed at any desired power level, but for optimum efficiency it is usually done at a low level and the modulated signal is then amplified to the desired output level. Since suppression of the undesired sideband is accomplished in the balanced modulators in this system, expensive filters having severe design requirements are unnecessary. Thus this system is less costly and less complicated circuit-wise than the filter system. Either sideband can be selected when desired in the phase-shift transmitter by reversing the phase-shift inputs to the balanced modulators by means of a simple switching arrangement.

The degree of undesired sideband suppression in this system is directly affected by unavoidable variations in the 90-degree shifts from the phase-shifting networks. Exact 90-degree phase relationships must be maintained throughout the sys-

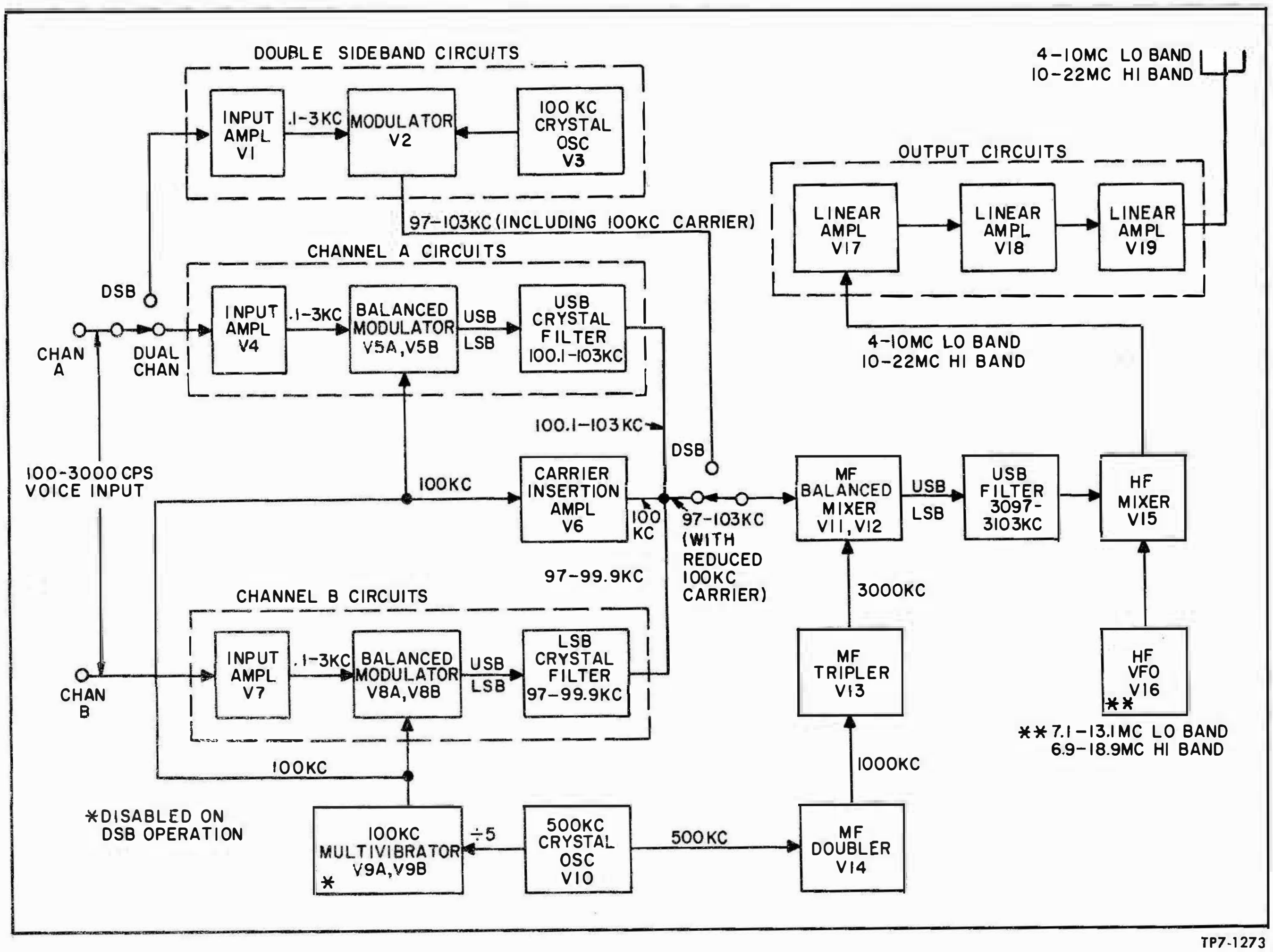

Figure 74. Typical Dual-Channel SSB Transmitter.

tem if complete cancellation of the undesired sideband is to be realized. Adjustment of the phase-shift networks is quite critical, and difficult to maintain over long periods of time. Therefore, the phase-shift SSB transmitter is less stable than the filter-system transmitter and requires more frequent attention if optimum performance is to be obtained.

DUAL-CHANNEL SSB TRANSMITTER

A dual-channel SSB transmitter can be considered as consisting of two separate single-sideband exciters connected in parallel to common heterodyning circuits and output power amplifiers. This transmitter provides for transmission of intelligence from two separate input channels, in two independent frequency bands, on opposite sides of a common reduced r-f carrier. Crystal-controlled oscillators and regulated power supplies are generally used in this type of transmitter to obtain good frequency stability. Provision can also be made for double-sideband operation, with reduced carrier, from only one channel input.

An analysis of the principles of operation of the dual-channel transmitter is given in the following paragraphs. Since the basic circuits of this transmitter are essentially the same as those already considered, or the same as other conventional circuits, the discussion is limited to a block diagram of the transmitter.

Low-Frequency Circuits

Figure 74 shows a block diagram of a typical dual-channel SSB transmitter. The low-frequency circuits in this system include the following blocks: 500-kc crystal oscillator (V10), 100-kc multivibrator (V9), carrier re-insertion amplifier (V6); Channel A and Channel B input amplifiers (V4 and V7), balanced modulators (V5 and V8), and crystal-lattice filters. The Channel A and B circuits are identical in design and operation except that the Channel A balanced modulator plate transformer and crystal-lattice filter are designed to pass only the upper sideband frequencies generated in Channel A, and the same circuits in Channel B are designed to pass only the lower

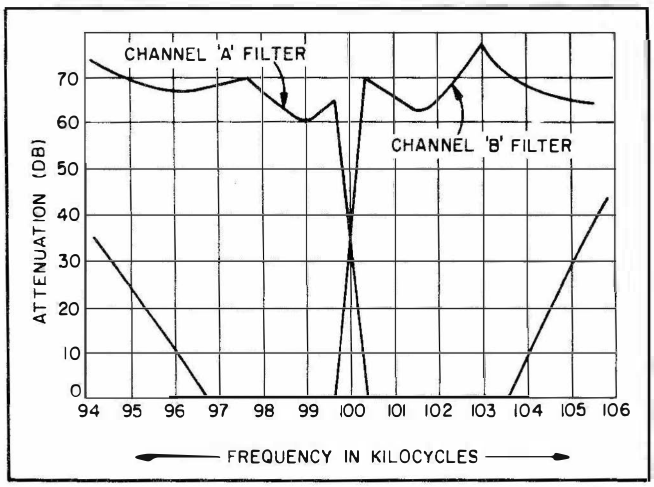

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Figure 75. Typical Bandpass Curves for Channel A and B Crystal Filters.

sideband frequencies generated in this channel.

The 500-kc crystal oscillator (V10) employs a 500-kc low-temperature quartz crystal and regulated voltages for good frequency stability. Its output is applied to the medium-frequency doubler (V14), and through a 5:1 count-down circuit to the 100-kc multivibrator (V9). The 100-kc output of this multivibrator, which is actually controlled by the 500-kc crystal oscillator, is applied in parallel to the Channel A and Channel B balanced modulators and the carrier re-insertion amplifier. In the balanced modulators this 100-kc r-f carrier signal is modulated by push-pull audio signals from the respective Channel A and B input amplifiers. These input-amplifier and balanced-modulator circuits are conventional, and their operation is similar to that of the corresponding circuits used in the filter-system SSB transmitter. The output from the balanced modulators in each channel consists of upper and lower sideband frequencies spaced symmetrically about a suppressed 100-kc carrier.

Note that the crystal oscillator used in this circuit operates at 500 kc, rather than 100 kc as used in the previously discussed transmitter. By operating at a higher frequency, the oscillator serves as a master oscillator to control both the low-frequency carrier oscillator signal (the 100-kc multivibrator) and the medium-frequency (3-mc) oscillator signal (applied to the mixer stage). Either system could be used with either transmitter; however, a dual-channel transmitter requires even greater frequency stability than that previously discussed, to prevent spurious products of one channel from affecting the other channel.

The Channel A crystal-lattice filter is designed to pass a band of frequencies from 100.1 to 103 kc, the upper sideband frequencies generated in the Channel A balanced modulator. Similarly, the Channel B filter is designed to pass a band of frequencies from 97 to 99.9 kc, the lower sideband frequencies generated in the Channel B balanced modulator. Typical bandpass curves for the crystal filters are illustrated in figure 75. For transmission of a high quality single-sideband signal, the filter used must have a sharp cutoff characteristic. When a second channel is used, as is the case in the dual-channel system, a high degree of filter accuracy is very important to the over-all performance of the system. The output of the two individual channel filters is combined in a simple resistance combining network, from which they appear as one 6-kc band of frequencies (97 to 103 kc). However, it will be well to keep in mind that this 6-kc band actually consists of frequencies from two unrelated sidebands, one on each side of a suppressed 100-kc carrier. This combined signal is then applied to the medium-frequency balanced mixer, where it is heterodyned with an r-f carrier signal from the medium-frequency carrier-generating circuits.

A carrier re-insertion amplifier circuit (V6) is incorporated in the transmitter to permit transmission of the r-f carrier at a reduced level. The gain of this amplifier, which is a conventional Class A r-f amplifier, is made variable so that the level of the reduced carrier can be controlled. The carrier re-insertion amplifier receives its input signal from the 100-kc multivibrator, and applies its amplified output to the m-f balanced mixer. This reduced-level 100-kc carrier frequency will appear in the center of the 97-to-103-kc band of frequencies applied to the m-f balanced mixer from the Channel A and B crystal-lattice filters.

Medium-Frequency Circuits

One of the outputs of the 500-kc crystal oscillator is applied to the m-f doubler circuit (V14). This stage, which is a conventional frequency-doubling circuit, applies its 1000-kc output to the m-f tripler (V13). The m-f tripler is a conventional frequency-tripling circuit, and applies its 3000-kc output to the m-f balanced mixer (V11 and V12).

In the m-f balanced mixer, the 3000-kc output of the tripler is heterodyned with the 97-to-103-kc signal from the Channel A and B circuits, the latter signal being placed symmetrically about a 100-kc reduced-carrier frequency. Since this mixer is a balanced circuit, it will suppress the 3000-kc input and generate sum and difference frequencies. The difference frequencies cover a 6-kc band from 2897 kc to 2903 kc, and the sum frequencies cover a 6-kc band from 3097 kc to 3103 kc. Since the plate circuit of this mixer and the m-f bandpass filter are both tuned to pass only the sum frequencies, the 2897-to-2903-kc difference frequencies from the mixer will be rejected at this point.

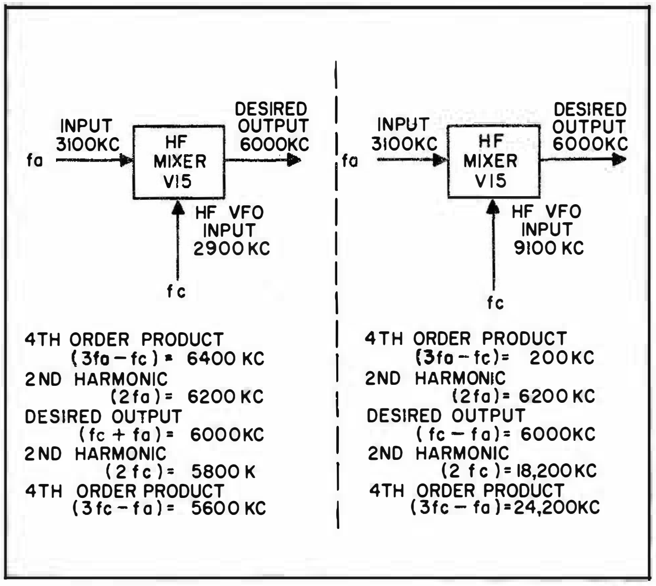

Figure 76. Selection of Heterodyning Frequency for H-F Mixer.

The 6-kc band of sum frequencies contains the lower sideband frequencies of Channel B and the upper sideband frequencies of Channel A. Centered in this 3097-to-3103-kc band is a 3100-kc reduced-carrier signal, which is produced when the 100-kc l-f carrier and the 3000-kc m-f carrier are heterodyned in the m-f balanced mixer. Therefore, the output of the m-f circuits will contain frequencies from two unrelated input channels placed 3 kc on either side of a 3100-kc reduced carrier. This signal is then applied to the high-frequency mixer, where it is heterodyned to the final desired transmitter output frequency by the h-f variable-frequency oscillator. A transition from push-pull operation to single-ended operation occurs at the input to the h-f mixer.

High-Frequency Circuits

In any single-sideband transmitter designed to operate over a range of output frequencies, it is necessary to use a heterodyning process to obtain the final transmitter output frequency after the SSB signal has been generated. The proper frequencies for this heterodyning must be selected with care, because selection of the wrong frequencies can cause many undesirable spurious signals. Considerable filtering will remove most spurious signals; but, if they are at or near the operating frequency, no degree of filtering or selectivity will satisfactorily eliminate them. To avoid this problem in h-f mixers, it is customary to use the sum frequencies from the mixer for transmitter output frequencies above 10 mc, and the difference frequencies for output frequencies below 10 mc. Figure 76 illustrates examples of proper and improper heterodyning frequency selection. The transmitter is arranged so that the Channel A signal is the upper sideband of the dual-channel output when the operating frequency is above 10 mc, and the lower sideband when the operating frequency is below 10 mc. The Channel B signal will then be the opposite sideband in both cases. This same sideband inversion at 10 mc must also occur in the equipment receiving this transmitted signal if the demodulation process is to be performed properly. Otherwise Channel A at the transmitter location would communicate with Channel B at the receiver location and vice versa.

The over-all frequency range of the transmitter is 4 to 22 mc (nominal) and is covered in two bands—a 4-to-10-mc LO band and a 10-to-22-mc HI band. Since the frequency output of the h-f vfo depends upon the desired transmitter output frequency, this circuit (V16) is made variable from 6.9 to 18.9 mc. With the exception of its frequency range, this oscillator compares in all respects to the h-f oscillators described previously—it is a modified series-fed Hartley circuit operating as an electron-coupled oscillator, with regulated voltages for good frequency stability. The output of the h-f vfo is applied to the h-f mixer (V15), where it is heterodyned with the 3100-kc carrier (and the two unrelated sidebands) from the m-f circuits.

When the transmitter is operating on the LO band (transmitter output frequency below 10 mc), the h-f mixer output will be the difference frequency between the two incoming signals. At this time the h-f vfo will cover the frequency range from 7.1 to 13.1 mc only, with the output plate circuit of the h-f mixer, and the following output amplifiers, being tuned to pass only the 4-to-10-mc difference frequencies from the mixer. On the other hand, when the transmitter is operating on the HI band (transmitter output above 10 mc), the h-f mixer output will be the sum frequencies of the two incoming signals, and at this time the h-f vfo will be operated over its entire 6.9-to-18.9-mc frequency range. On this band the mixer plate circuit and the following output amplifiers will be tuned to pass only the 10-to-22-mc sum frequencies from the mixer. To shift the transmitter output frequency, it is necessary, therefore, to change the final heterodyning carrier frequency applied to the h-f mixer, and retune the transmitter accordingly.

Output Circuits

Since the output circuits of an SSB transmitter must provide maximum power gain with minimum distortion, the three output amplifiers in this transmitter are operated as Class B linear amplifiers. All precautions pertaining to neutralization and suppression of parasitic oscillations in

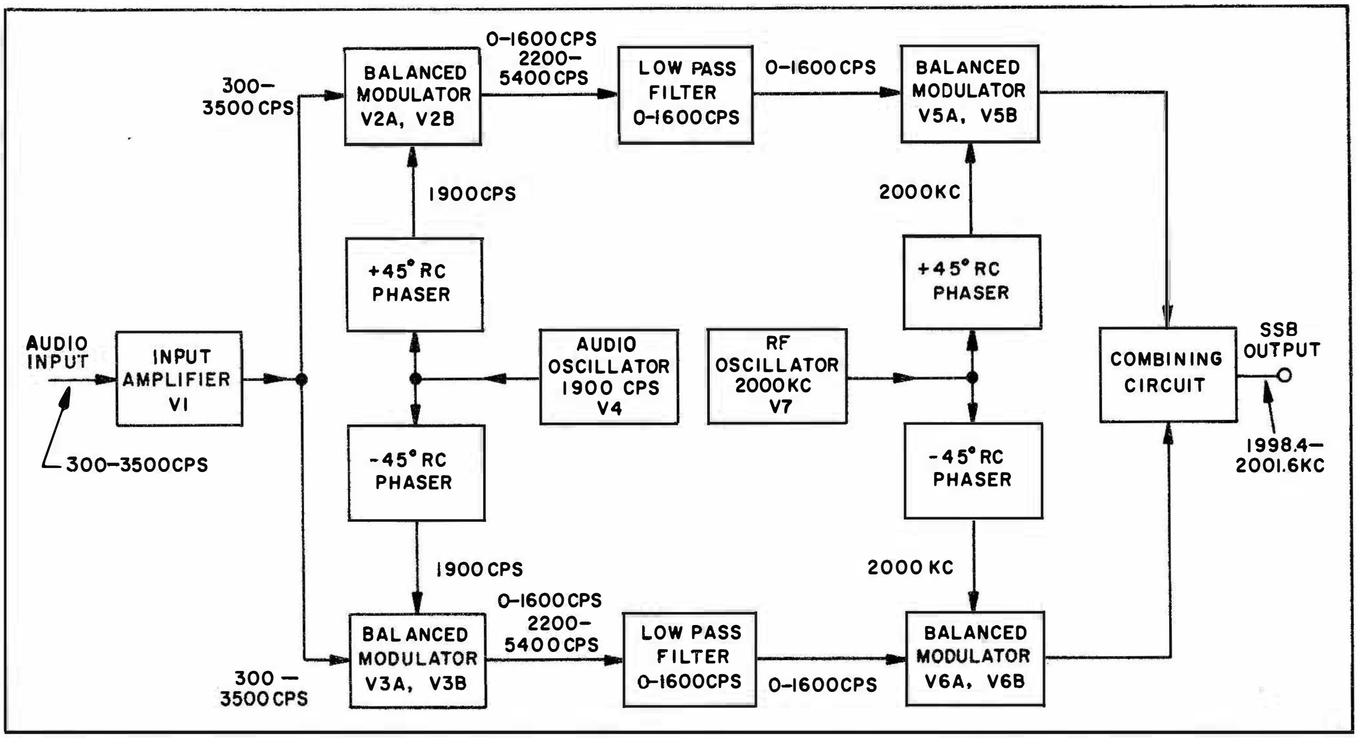

Figure 77. Block Diagram of Third Method of SSB Generation.

power amplifiers also apply to these circuits.

The first amplifier (V17) has a single-ended input (from the h-f mixer) and a push-pull output. A transition is made, therefore, from a single-sided circuit back to a push-pull arrangement at this point. The second output amplifier (V18) is effectively a push-pull driver stage for the final output amplifier (V19), which is also operated push-pull. Tuning of these three stages is made variable in each of the two frequency bands covered by the equipment.

Double-Sideband Circuits

The double-sideband circuits, consisting of a 100-kc crystal oscillator, an audio input amplifier, and a low-level modulated amplifier, are placed in operation by means of a simple switching arrangement. At the same time, the 100-kc multivibrator (V9) is disabled to prevent that stage from operating when the double-sideband circuits are being used. Operation of the transmitter for double sideband transmission is usually used when the equipment receiving the transmitted signal (conventional AM receiver) is not designed for single-sideband reception.

When the transmitter is switched to double-sideband operation, the input signal normally applied to the Channel A input amplifier (V4) is transferred to the double-sideband input amplifier (V1). The 100-to-3000-cps audio input frequencies are amplified in this stage and applied to a low-level modulator (V2). In the modulator the audio signal modulates a 100-kc r-f carrier, which is applied to this same stage from the double-sideband 100-kc crystal oscillator (V3). The amplitude of the output signal from the double-sideband circuits is controlled by an output level control placed in the output circuit of the modulator. On DSB operation, therefore, the input to the m-f balanced mixer will be a controlled-level, double-sideband, amplitude-modulated signal containing lower sideband frequencies from 97 to 99.9 kc and upper sideband frequencies from 100.1 to 103 kc, spaced symmetrically about a 100-kc r-f carrier. These sideband frequencies are related to each other, whereas on dual-channel SSB operation the sidebands were from two unrelated input sources.

The operation of the transmitter circuits from the m-f balanced mixer to the final push-pull output power amplifier is the same on double-sideband operation as on dual-channel single-sideband operation, and no further explanation of these circuits is necessary.

Although the filter-system method of generating a single-sideband signal has been used in this example of a dual-channel SSB transmitter, phase-shift exciters could just as easily have been employed. The only change in the transmitter circuitry would be the use of individual phase-shift circuits instead of the filter-system circuits shown as the Channel A and Channel B exciters.

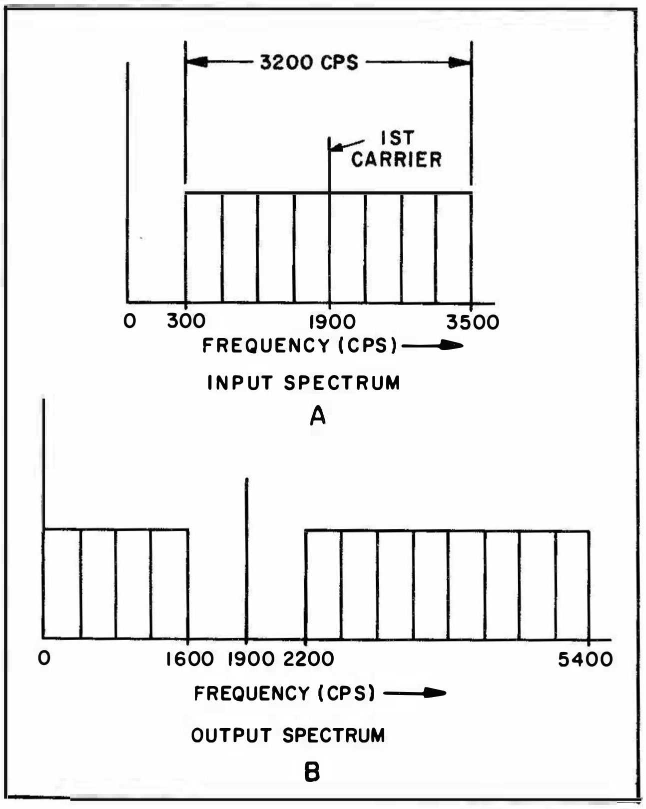

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Figure 78. First Balanced Modulator Signal Spectra.

THIRD METHOD OF GENERATING SSB SIGNALS

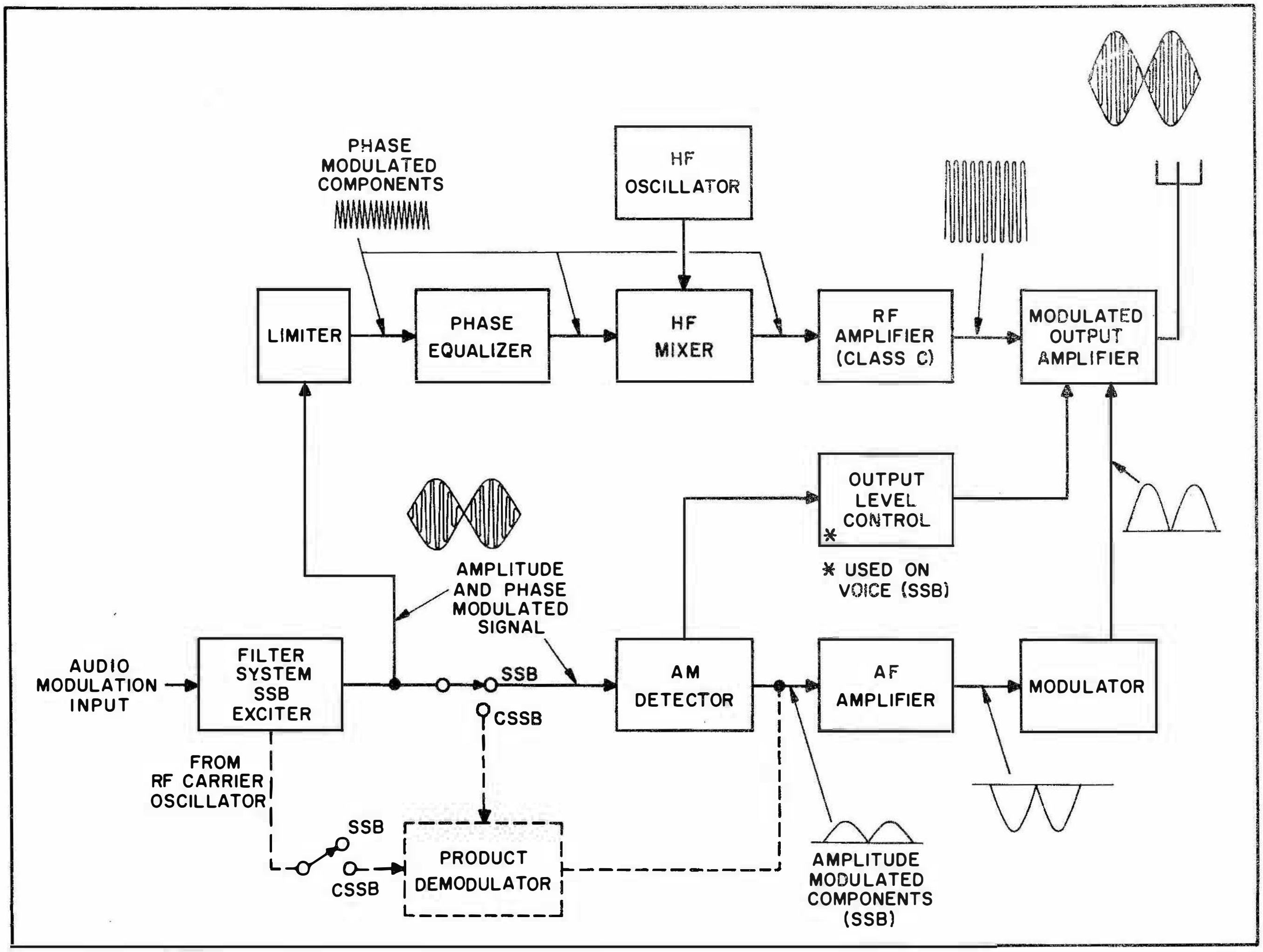
Although the filter and phase-shift methods of generating single-sideband signals are most commonly used in existing SSB applications, a third method of generating such a signal has been developed. Instead of using wide-band 90-degree phase-difference networks, this system uses balanced modulators with quadrature carriers to generate 90-degree phased audio signals. Also, by positioning the first carrier frequency in the center of the audio spectrum, both sidebands are made to fall in the same band. Hence, in this system, there is no undesired sideband, and the need for sharp-cutoff filters is therefore eliminated. Thus this third method of generating an SSB signal differs from the filter and phase-shift methods in that it does not use any sharp-cutoff filters, or wide-band 90-degree phase-difference networks. However, if it is desired to use filters to provide added suppression of the residual sideband, the first carrier frequency can be positioned at the high end of the audio spectrum.

A block diagram of the third method of SSB signal generation is illustrated in figure 77. The audio modulating signal from the input amplifier, confined to a 300-to-3500-cps band of frequencies, is applied to the first pair of balanced modulators (V2A and B, V3A and B). Also applied to these modulators, through simple 45-degree lead-lag R-C phase-shifting networks, is a quadrature low-frequency 1900-cps carrier frequency (the first carrier) from an audio oscillator. The frequency of this oscillator is selected to fall in the center of the audio input spectrum, as illustrated in figure 78A. The first pair of balanced modulators is designed to suppress the audio input signal, thus preventing the input frequencies from appearing in the output of this circuit. The output spectrum of the first pair of modulators is illustrated in figure 78B. From the figure it can be seen that the lower sideband generated in these modulators (first pair) contains frequencies from zero to 1600 cps, and that the upper sideband contains frequencies from 2200 to 5400 cps. Since the lowest audio modulating frequency is 300 cps, no frequencies between 1600 and 2200 cps will be generated in this modulator system.

The output of the first pair of modulators is applied to separate low-pass filters, which are designed to pass the zero-to-1600-cps band of frequencies and to provide adequate attenuation of all the high-frequency components above 2200 cps. The outputs of these filters, containing only the low-frequency components below 1600 cps, are then applied to a second pair of balanced modulators (V5A and B, V6A and B). Also applied to this second pair of modulators is the final heterodyning frequency from the high-frequency r-f oscillator. This r-f frequency (second carrier) is applied in quadrature through separate 45-degree lead-lag R-C phase-shifting circuits similar to the type used in the low-frequency audio system. The r-f oscillator frequency is positioned in the center of the desired single-sideband signal, and is usually a much higher frequency than any of the frequencies in the original signal.

The second pair of balanced modulators will generate upper and lower sideband frequencies, and suppress their individual r-f carrier inputs. Through the action of a combining circuit, located in the plate circuit of the second pair of balanced modulators, the output signal from these modulators will contain only the generated upper sideband frequencies. Critical balance of the balanced modulators is not an absolute must, because in this method of generating a single-sideband signal the "undesired" sideband occupies the same band as the desired sideband, except that it is inverted.

The circuitry of the third method of generating a single-sideband signal is bilateral in operation; that is, the circuit can be used for demodulating, as well as generating, a single-sideband signal. Since the circuit does not use any expensive sharp-cutoff filter or require any critical balance of the balanced modulators, it has the advantages of being less costly than the filter system and less critical in adjustment than the phase-shift system.



TP7-1278

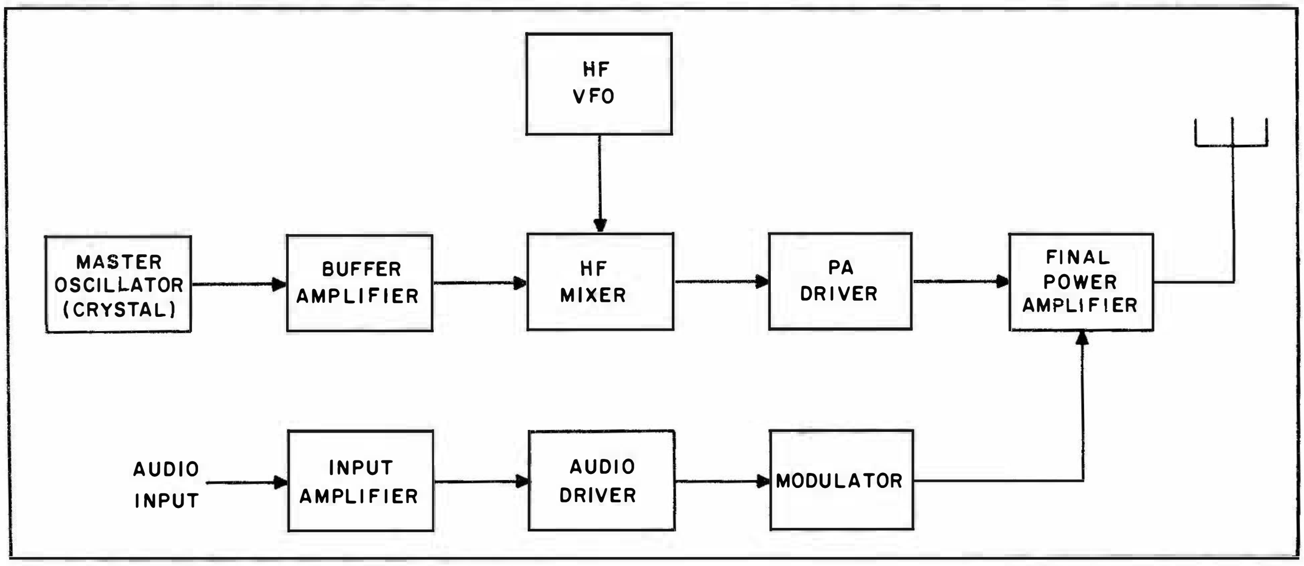
Figure 79. Block Diagram of Envelope Elimination and Restoration SSB Transmitter.

ENVELOPE ELIMINATION AND RESTORATION SSB TRANSMITTER

The single-sideband transmitters discussed up to this point are concerned primarily with the method of generating the single-sideband signal. The envelope elimination and restoration SSB transmitter, however, is concerned primarily with the amplification of the single-sideband signal after it has been generated in a suitable SSB exciter. Either filter or phase-shift exciters may be used in this transmitter, but for purposes of discussion it will be assumed that the exciter is of the filter type.

It will be recalled that in the filter-system SSB transmitter the modulation process was performed at a low-power level, with the single-sideband signal then being amplified to the desired transmitter output power level in one or more linear power amplifiers. Since all linear amplifiers introduce some distortion, the higher the power desired from the transmitter, the higher this spurious output will be. The envelope elimination and restoration system does not use linear amplifiers in its output circuits; therefore, the spurious output in this transmitter is reduced, and is independent of the transmitter output power level. It is felt that a detailed analysis of the circuits used in this system is unnecessary since they are conventional; therefore, only a block-diagram analysis of this transmitter will be presented.

Figure 79 shows a block diagram of an envelope elimination and restoration SSB transmitter. The SSB exciter produces a low-frequency and low-power suppressed-carrier SSB signal consisting of both amplitude- and phase-modulated components. One portion of the exciter output is applied to a limiter stage, while another portion of the same signal is applied to an AM detector. In the limiter the amplitude-modulated components are removed from the SSB signal, leaving only a constant-amplitude phase-modulated signal as the output of this stage. This phase-modulated signal is then applied to a mixer stage through a phase-equalizer circuit. The purpose of the phase equalizer is to maintain the proper time relationships



TP7-1279

Figure 80. Block Diagram of a Synchronous Communications Transmitter.

between the amplitude and phase-modulated components so that they can be properly recombined.

The h-f oscillator and h-f mixer circuits translate the low-frequency phase-modulated signal to the frequency desired for transmission. For single-channel operation the oscillator may be of the stable crystal type, and for multi-channel operation it may be of the v-f-o type. Care should be taken in the design of the mixer stage to prevent intermodulation distortion.

From the h-f mixer the signal is applied to a Class C r-f amplifier. Since the phase-modulated components will not be affected by any amplitude non-linearity in the r-f amplifier, the use of a highly efficient Class C amplifier circuit is permissible in this system. The amplified phase-modulated signal is then applied to the final modulated output amplifier.

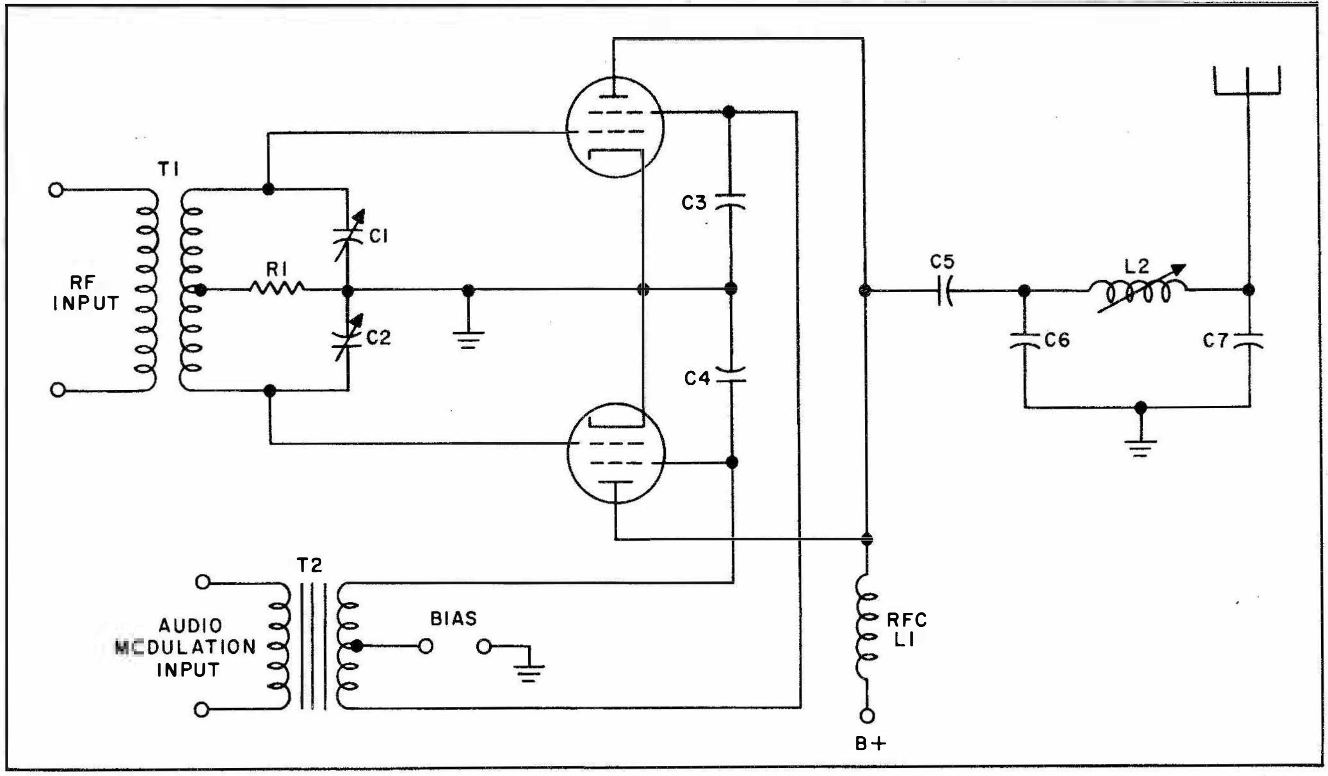
The AM detector, which receives a portion of the signal from the SSB exciter, is a conventional diode detector. This circuit removes the phase-modulated components of the single-sideband signal, leaving only the amplitude-modulated components as the output of this detector. These components, which are identical to the AM components at the output of the SSB exciter, are applied to an a-f amplifier. The amplified signal from this stage is then applied to the final modulated output amplifier through a conventional modulator circuit.

In the final output amplifier, the amplitude-modulated components modulate the phase-modulated signal from the Class C r-f amplifier. If the exact time relationship between the amplitude- and phase-modulated signals is maintained throughout the system (by the action of the phase equalizer), the output of this final amplifier will be a highly amplified version of the single-sideband signal appearing at the output of the SSB exciter. Thus it can be seen that the envelope elimination and restoration SSB transmitter can produce a higher output signal, with better efficiency, than an SSB transmitter using linear amplifiers.

To maintain the average power output of the transmitter constant during the transmission of variable average-amplitude signals (voice), an output level control circuit is included in the transmitter design. This circuit consists of one or more d-c amplifiers which control one of the d-c voltages applied to the final modulated output amplifier, and thus control the level of the output signal from this stage.

This system can easily be converted into a compatible single-sideband transmitter (CSSB) by a few minor modifications to the circuitry. The first of these is to alter the SSB exciter to generate a full-carrier single-sideband signal instead of a suppressed-carrier type signal.

In the modified transmitter the output of the exciter is still applied to the detecting, heterodyning, and amplifying circuits for the phase component as described previously. However, instead of being applied to the AM detector, the output of the exciter is now applied to a product demodulator (as indicated by the dotted circuits in figure 79). A portion of the carrier signal from the r-f carrier oscillator in the SSB exciter is also applied to the product demodulator. The full-carrier SSB signal and the r-f carrier signal are combined in the circuit in such a manner that the resultant output signal consists of only the original audio modulating components, as they appeared at the input to the SSB exciter. This signal



TP7-1280

Figure 81. Final Power Amplifier for a Synchronous Communications Transmitter.

is then amplified and used to modulate the phase components in the modulated output amplifier, as discussed previously.

To produce a relatively distortion-free signal with this compatible single-sideband transmitter, it is necessary for the modulation process to be linear. Conventional AM receivers will then be capable of receiving and properly demodulating the signal. Also, since the carrier is transmitted in this system, there is no need for the output level control circuits and they can be eliminated from the CSSB transmitter circuitry.

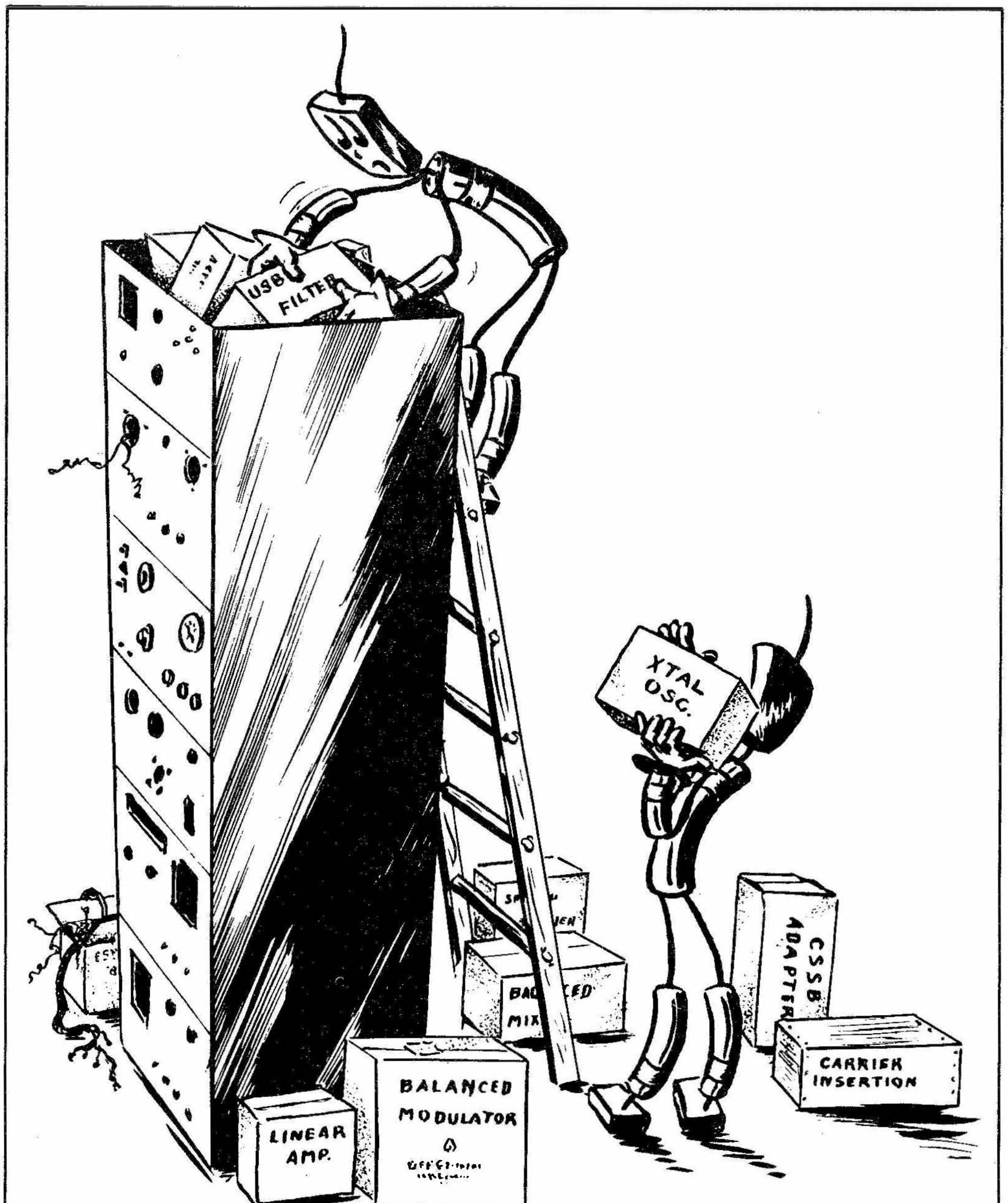
DOUBLE-SIDEBAND SUPPRESSED-CARRIER TRANSMITTERS

Although this manual is primarily concerned with discussing single-sideband concepts and techniques, brief mention will be made of an amplitude-modulated double-sideband suppressed-carrier transmitter, often referred to as a "synchronous communications" transmitter. The basic concept of this method of transmission is that if the maximum capabilities of present AM DSB communications systems were realized, the present mode of communications would produce results equal to, and sometimes better than, SSB systems. This concept, of course, is based on the assumption that both systems operate under identical conditions (use the same type of modulating signal).

For present AM double-sideband systems to realize their maximum capabilities, two important changes must be made in existing transmitting and receiving equipment. The first change is the inclusion of carrier suppression in the transmitter. With the carrier suppressed, all the transmitter power will be used to transmit the intelligence-carrying sidebands, and none will be expended in a carrier which does not convey any of the signal intelligence. The second change is the incorporation of a phase-locked oscillator and synchronous detector (or similar circuits) in the receiver to insure proper demodulation of the transmitted AM double-sideband suppressed-carrier signal. The synchronous communications receiver will be discussed in the receiver portion of this publication so no detailed analysis of its operation will be presented here.

A major advantage claimed for a synchronous communications transmitter over an SSB transmitter is its simplicity. Figure 80, a block diagram of this system, illustrates the similarity between a synchronous AM transmitter and a conventional high-level modulated AM DSB transmitter. The r-f circuits from the master oscillator to the power-amplifier driver are conventional, and so are the audio circuits. To convert an existing AM DSB transmitter into a synchronous AM transmitter, it is necessary to incorporate a circuit that will provide adequate suppression of the carrier frequency. The final power amplifier in this transmitter is such a circuit, and is illustrated schematically in figure 81.

The final power amplifier is a Class C stage using push-pull tetrodes, with the r-f applied to the control grids and the audio modulation applied to the screen grids, both signals being applied in push-pull. The purpose and operation of this stage are similar to those of a balanced modulator. When no audio modulation is present, the push-pull r-f input will be cancelled in the parallel-connected plate circuit, and there will be no output from the transmitter. When an audio modulating signal is applied to the screen grids, on the other hand, upper and lower sidebands will be produced, and both the audio and r-f push-pull input signals will be cancelled in the parallel-connected plate circuit of the amplifier. The output of this synchronous communications transmitter, therefore, will be an amplitude-modulated double-sideband signal with the carrier suppressed. This output will be present only when there is an audio modulating signal present at the screen grids of the final power amplifier; hence, there will be brief periods when the transmitted signal is absent. This absence of the signal, however, produces no noticeable distortion in the receiver output because of the rapid operation of the phase-locked oscillator.



FUNCTIONAL BLOCK ANALYSIS OF TYPICAL SSB RECEIVER

The single-sideband receiver circuit in many respects resembles an AM receiver circuit. The use of high-stability oscillators, a balanced detector with sharply tuned filters, and multiple (double or triple) conversion is indicative of the major differences between AM receiver circuits and circuits common to single-sideband receiving systems.

Many variations in circuitry may be used, each of these variations having specific advantages and disadvantages. Indeed, an AM receiver may be used for single-sideband reception if a bfo or an external oscillator of good stability is utilized for carrier insertion (figure 82). However, these methods are obviously not suitable for military or commercial applications. Various types of detectors have been developed, and filter circuits have been improved or simplified in many cases. The exact circuitry used in a specific receiver depends on the type of service, or end use of the intelligence to be received, and the required fidelity of reproduction. Some of these circuits will be discussed in greater detail in other sections of this manual.

Consider a typical triple-conversion superheterodyne receiver suitable for single-sideband reception of single-channel voice communications in the frequency range of 10 to 30 megacycles. Such a receiver is shown in block form in figure 83. Assume that the upper sideband is being received, with or without a pilot carrier. If transmitted, the lower sideband could be received or selected by switching the band filters or changing the oscillator frequencies.

The high-frequency (r-f) amplifier, V1, is tuned to the desired incoming signal frequency received from the antenna circuit, and performs the functions of preselection and amplification.

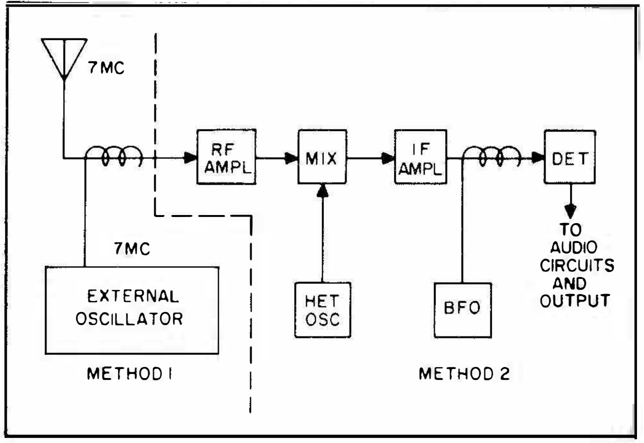
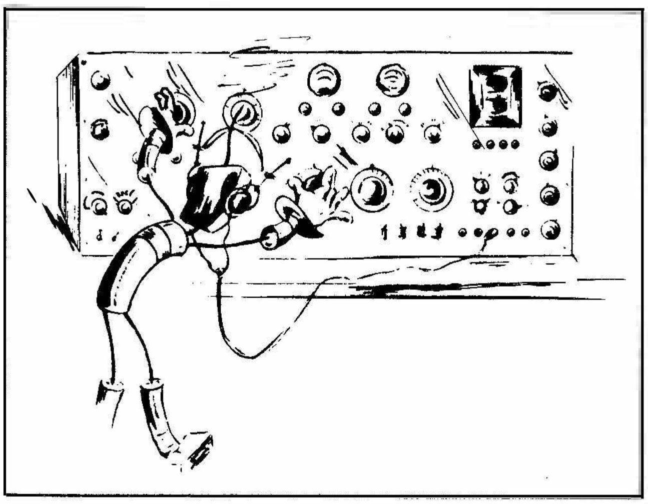


Figure 82. Methods of Using Existing AM Equipment for Receiving SSB.

The amplified sideband signal is then applied to the high-frequency mixer, V2. This circuit mixes the signal from the high-frequency amplifier with the output of the high-frequency oscillator, V3. In the mixing process, heterodyning action between the two signals produces a third, or difference, frequency. Since this frequency results from the first conversion process of a triple-conversion receiver, it is called the medium frequency (high i-f or first i-f). This process is the same as that used for normal AM reception by the frequency-conversion (heterodyning) method.

The high-frequency oscillator, V3, is variable-tuned in the range of 6.9 to 26.9 mc. Thus this oscillator is operated below the signal frequency, with a medium-frequency difference of 3.1 mc, or 3100 kc. This frequency, incidentally, is the same as the medium frequency used in the typical transmitter circuit discussed in the transmitter section of this manual. This need not be the case, however. The frequency may be chosen within limits to conform to easily obtainable filters.

The oscillator tuning is extremely critical, and frequency stability is essential, since the full wave envelope is not present. Any deviation of oscillator frequency with respect to sideband frequency will result in frequency distortion in the output of the receiver.



". . . tuning becomes quite critical——"

The medium frequency (high i-f) resulting from heterodyning of the incoming signal and the local-oscillator output is still only the upper sideband, converted to a lower frequency. The carrier may be present, if it is being transmitted; however, it has not yet been added.

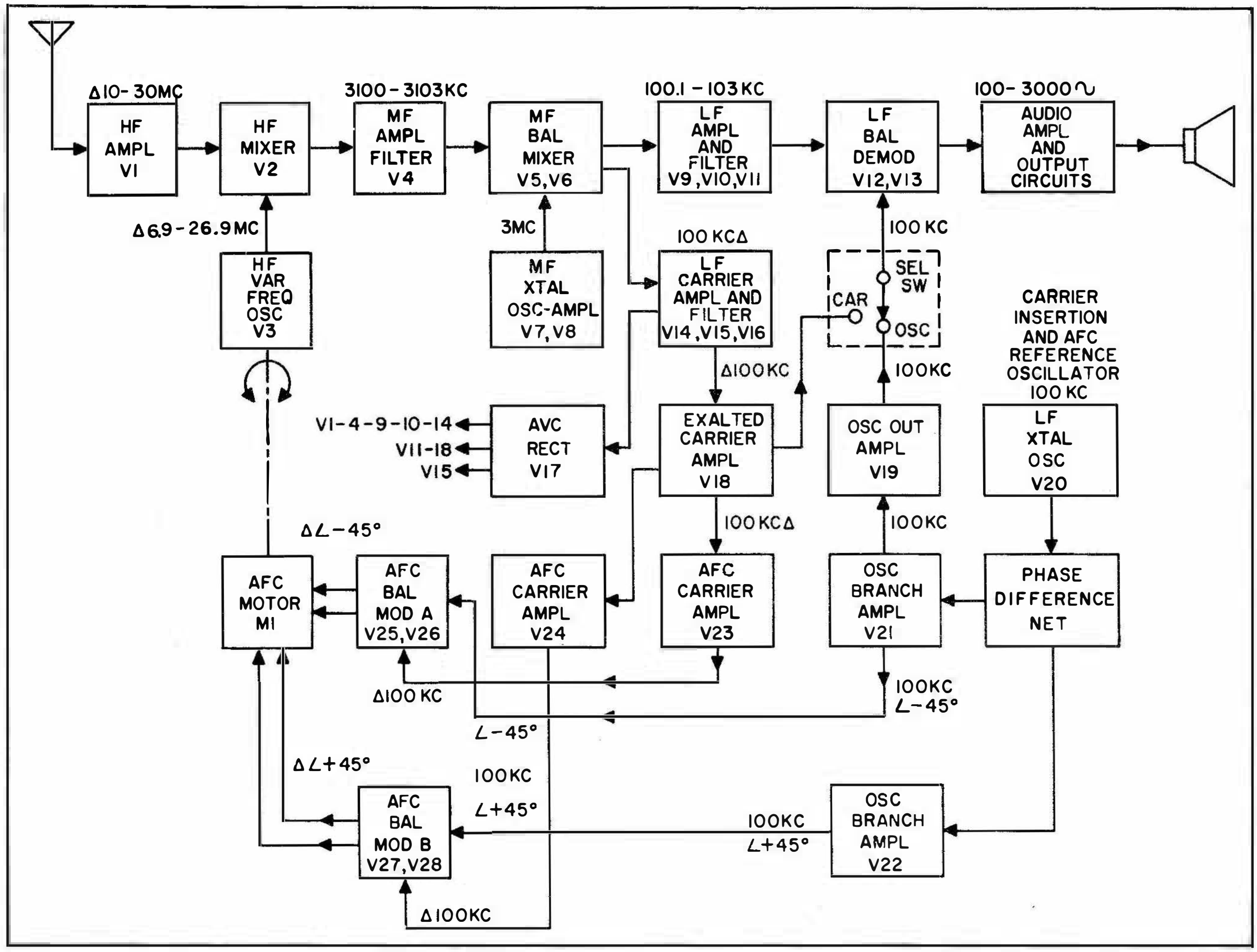


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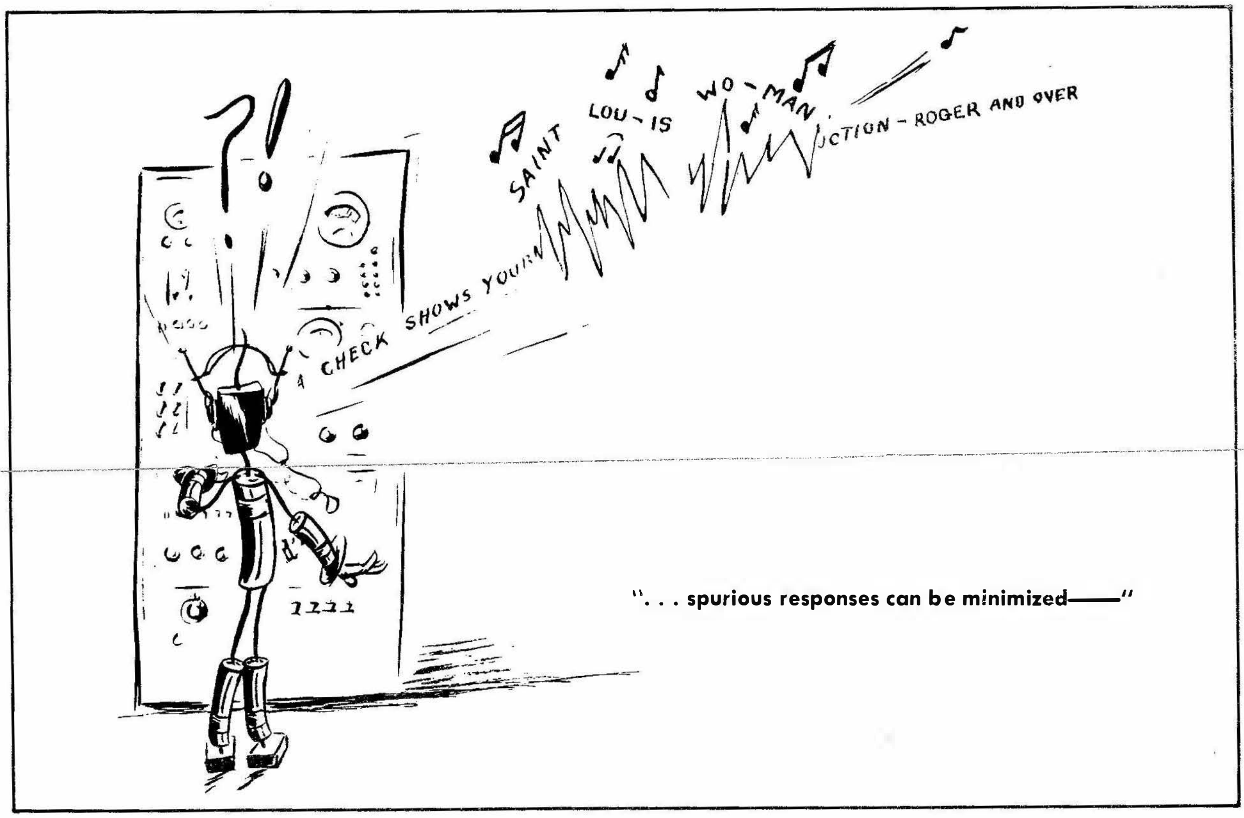
Figure 83. Block Diagram of a Typical SSB Receiver.

The output of the high-frequency mixer, V2, is filtered to eliminate all but the desired upper sideband information converted to a medium frequency which may or may not include the carrier. The filtered signal is then amplified in the medium-frequency stage, V4. In this particular receiver, the medium frequency chosen was 3100 to 3103 kc. The filter used has sharp attenuation on either side of the upper sideband; however, it must also pass the carrier frequency of 3100 kc if used. In any case, the bandwidth is determined exclusively by the filters employed. Typical filter circuits for single-sideband reception are discussed elsewhere in this manual. Since only one sideband is being received, the filter and amplifier require only one half of the bandwidth that would be required for normal AM reception. This reduction in bandwidth usually results in an improved signal-to-noise ratio. However, the i-f transformer tuning is usually broadened slightly to insure reasonably flat response to all frequencies in the pass band of the filter.

As an example of these principles, assume that a carrier is transmitted at 10 mc. This carrier, heterodyning with the local-oscillator signal at 6.9 mc, would then produce an i-f signal of 3100 kc. Now assume that the same carrier is voice-modulated with frequencies limited to a range of from 100 to 3000 cycles per second. If this were an AM transmission with both sidebands, a bandwidth of 6 kc would be required in the r-f and i-f sections, and the i-f stage would be center-tuned to 3100 kc. If the modulation of the received signal included music, or frequencies up to 10,000 cycles, then for full reproduction, the receiver bandwidth would have to be at least 20 kc wide for AM, or 10 kc for SSB. Since this receiver is designed for single sideband, voice frequencies limited to from 100 to 3000 cycles, the bandwidth need only be one-half that required for double sideband, or 3 kc instead of 6 kc. Also, since the filters are fixed tuned (crystals), it is desirable to depend on the filters alone to limit the bandpass, and to tune the



". . . spurious responses can be minimized——"

i-f transformers (and the beating oscillators) to correspond to the filter frequencies.

The output of the medium-frequency amplifier is applied to the medium-frequency balanced mixer, V5 and V6. Here the heterodyning of the m-f signal of 3100 to 3103 kc with the 3-mc oscillator output from V7 and V8 produces the low frequency of 100 to 103 kc. The action of the balanced mixer is the same as that of a balanced modulator, and the basic circuits are the same, the difference being only in its use. The balanced mixer is used because it possesses certain advantages favorable to single-sideband circuits. It provides cancellation of the 3-mc oscillator frequency, and thus prevents possible overloading of the following stage. Also, it permits only the results of the heterodyning of the signal with the 3-mc oscillator frequency to appear in the output. This is important because of the close proximity of 3 to 3.1 mc. In addition, spurious responses and third order action can be minimized by using the balanced mixer.

To avoid confusion in the use of the balanced modulator, or mixer, remember that sidebands are sum and difference frequencies. The output of the medium-frequency balanced mixer is tuned to the 100-to-103-kc difference frequency, or the lower sideband of the 3-mc oscillator and the 3.1-to-3.103-mc signal. This should not be confused with the fact, however, that the frequencies between 100.1 and 103 kc are still the upper sideband of the original received pilot carrier, the frequency of which is now 100 kc. All have simply been converted to a lower frequency, in much the same manner as they were in the high-frequency mixer, V2.

The 3-mc, or medium-frequency oscillator, V7, must also be extremely stable, to prevent any shift in the sideband frequencies. Since the preceding m-f amplifier is fixed tuned, a crystal oscillator is conveniently used in this circuit. The use of a crystal oscillator which may include a heated, or temperature-controlled crystal oven, insures an extremely stable frequency. In other receivers this second beating oscillator may be a vfo, particularly in cases where the high-frequency oscillator is crystal-controlled, or where variable-tuned i-f stages are employed, as in some aircraft communication transceivers.

The value of 3 mc for the oscillator frequency was chosen in this discussion of an SSB receiver so as to place the oscillator frequency below the signal frequency. This need not be the case in practical circuitry, since standard i-f considerations and design practices apply, as in other AM circuitry. However, sideband transposition and inversion must be taken into account, if the oscillator is operated above the signal frequency, with SSB.

An amplifier, V8, is used to insure that the oscillator voltage at the mixer is higher than the signal voltage, in order to prevent distortion, as well as overloading of the oscillator.

The output of the medium-frequency mixer, V5 and V6, is applied to the low-frequency amplifier, V9, V10, V11, with its bandpass (sideband) filter circuit, and the carrier amplifier, V14, V15, and V16, with its sharply tuned (carrier) filter circuit. Here the pilot carrier is separated from the sideband signal, if carrier circuits are included in the receiver.

Conventional i-f circuitry is used in both amplifiers, except for the filters. The filter in the sideband channel (V9, V10, V11) is tuned to provide a bandpass of from 100.1 to 103 kc, corresponding to the upper sideband frequencies, and has a very sharp cutoff at the lower side of its frequency range, so as to reject the carrier frequency. The carrier channel (V14, V15, V16) and its filter are sharply tuned to 100 kc.

The output of the low-frequency amplifier (sideband channel), V11, is applied to the low-frequency balanced demodulator, V12 and V13, together with the amplified output of the 100-kc oscillator, V20. Note that this oscillator frequency of 100 kc corresponds to the frequency of the carrier, missing at the output of the sideband channel i-f amplifier, V11. Since the sideband-channel signal contains only sideband frequencies, it cannot be detected properly with the conventional diode, or envelope detector, because the complete wave envelope is not present. To permit proper detection, the 100-kc oscillator re-inserts the necessary carrier.

If a pilot carrier is received, it may be used, instead of the output of the 100-kc oscillator, for carrier re-insertion. However, since the pilot carrier is greatly reduced in amplitude, it must first be reconditioned and amplified, or its use will produce the same effect as overmodulation at the transmitter and thus cause serious distortion. It is amplified to the correct level by the exalted-carrier amplifier, V18, which provides an output voltage of constant amplitude (usually ten times the sideband voltage). This output is then applied to the carrier input of the balanced demodulator.

Under conditions of selective fading, when the carrier is subject to serious fading while the sidebands are not affected, it is preferable to use the 100-kc oscillator for carrier re-insertion. Use of this oscillator is of course a necessity for the reception of suppressed-carrier signals. In the receiver shown, a switch is used to select the reconditioned carrier output of V18, or the amplified low frequency oscillator output from V19.

Since the 100-kc oscillator, V20, is used for carrier re-insertion, it is also called the carrier re-insertion oscillator. A highly stable crystal oscillator is customarily used in this circuit for the same reasons given for the oscillators (V3 and V7) discussed earlier. The output of this oscillator must be high with respect to the sideband signal level at the demodulator, so that proper envelope shape will result and serious distortion will be prevented. For this reason, a stage of amplification, V19, is employed.

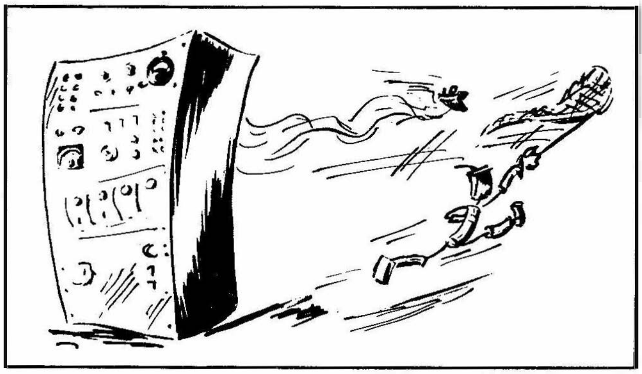
After carrier re-insertion, the signal may be detected with a conventional diode or other type of AM detector. However, the use of frequency conversion results in a higher signal-to-noise ratio and less cross-modulation and intermodulation from unwanted carrier and/or sideband beats than detection by rectification. This method also corrects, to some extent, the effect of the inherent phase distortion common to SSB. A convenient method of detection by frequency conversion is the use of a balanced detector.

The low-frequency demodulator, V12 and V13, is a balanced detector. The theory and design considerations of balanced detectors (demodulators) are the same as those of the balanced modulators discussed elsewhere in this manual, except that the frequencies are reversed.

The output of the low-frequency demodulator, V12 and V13, is applied to a conventional audio-amplifier circuit. Since this circuit and the power-supply circuits are conventional, an analysis of these circuits will not be given. In the audio-amplifier circuit audio filters may be used, if necessary, to further separate audio components. The power supply should include a well regulated plate voltage for the oscillator circuits, to provide the required frequency stability.

A specially designed a-f-c circuit is incorporated in this receiver to control the high-frequency oscillator. The frequency stability of this oscillator is an important factor, because any frequency shift of the oscillator with respect to the frequency of the transmitter carrier oscillator, unless compensated for elsewhere, will cause a shift in the audio output of the receiver, which will be apparent as frequency distortion.

If a pilot carrier or a controlled carrier is received, the amplified carrier (since bandpass considerations in the m-f stages included it), is taken from a convenient point, usually the output of the medium frequency mixer, sharply filtered and amplified in the carrier filter and amplifier circuit,



". . . due to oscillator drift——"

V14, 15, 16 and V18, and applied to the comparator type a-f-c circuit, along with the output of the carrier oscillator, V20. A shift in carrier frequency (if not too great) or a shift in the local low-frequency oscillator, V20, or any difference between the two, would then result in a corresponding change in the frequency and phases of the control voltages in the a-f-c circuit, and a subsequent change in frequency of the high-frequency oscillator, V3.

A change of 100 cycles, at 10 mc is a change of only ten parts per million, or a change of .001 percent. While this change would probably be unnoticed in a conventional AM receiver, which uses full carrier and envelope reception, the SSB receiver output would suffer a 100-cycle shift in audio frequency, since the sideband would be displaced from the re-inserted carrier by that amount. This displacement is the main cause of frequency distortion in the SSB receiver. Expressed another way, a shift in frequency of any heterodyne oscillator in either the transmitter or the receiver, unless accompanied by a like shift, in the proper direction, of a subsequent oscillator before detection, will produce an audio-frequency change of the same number of cycles as the original shift.

Since .001 percent of 100 kc is only 1 cycle, and the same percentage of 10 megacycles is 100 cycles, it can be seen that the higher-frequency oscillators are more critical in this type of reception.

Although other types of a-f-c circuits could be used, a more positive control circuit, able to correct the oscillator frequency to within 1 cycle, makes use of a motor to mechanically drive a variable capacitor in the h-f oscillator. This type of electromechanical circuit has been in common use in trans-oceanic telephone service, and its reliability is established and recognized in this field.

A phonic, split-phase, synchronous motor is used to drive a small variable capacitor in the high-frequency oscillator. An output from the exalted-carrier amplifier, V18, is applied, after further amplification in amplifiers V23 and V24, to the two a-f-c balanced modulators, (V25, V26, V27, and V28).

The 100-kc output of the low-frequency oscillator, V20, is applied to an R-L-C phase-shift network which divides the signal into two components 90 degrees apart in phase; one component lags the oscillator output by 45 degrees, and the other leads by 45 degrees. These two outputs are then amplified in V21 and V22, and applied to the two balanced modulators. Thus there will always be a difference in the outputs of the two balanced modulators of 90 degres, to provide the necessary phase shift for the split-phase motor, M1.

When a difference in frequency occurs between the received carrier and the output of the low-frequency oscillator, the balanced modulators become unbalanced and produce two output voltages. The frequency of each voltage is the same as the difference frequency of the oscillator and carrier, and the direction of phase rotation of one with respect to the other is determined by whether the carrier is higher or lower than the oscillator frequency. These voltages are applied to motor M1, which in turn mechanically drives the capacitor in the proper direction to correct the frequency of the h-f oscillator. When the oscillator frequency becomes the same as the carrier frequency, the balanced modulators regain their state of balance, their output voltages decrease to zero, and the motor stops.

Another output is taken from the carrier amplifier, V16, and applied to the a-v-c rectifier, V17. This circuit differs from conventional a-v-c circuits in that it is carrier-operated, and long time constants (on the order of from 1 to 8 seconds) are used. A forward-acting avc is included, along with a separate short-time-constant circuit, for compensatory purposes.

The a-v-c circuit uses several R-C networks, to apply the a-v-c voltage to various sections of the receiver. V17, the a-v-c rectifier, is used to charge the capacitors of these networks in accordance with the peak level of the amplified carrier voltage. During brief absence of the carrier from the received signal (as under conditions of fading) or during reception of controlled-carrier transmissions, the a-v-c circuit remains effective because of the long time constants used. The circuit is of the fast-charge, slow-discharge type.

In the complete absence of a received carrier, as in suppressed-carrier operation, both the a-v-c and a-f-c circuits remain inoperative, except when operated erratically by bursts of noise.

Although these a-v-c, a-f-c, and carrier circuits are not used in receivers designed for only SSSC operation, they have been discussed to associate the variations in receiver circuitry. Other types of these circuits are discussed separately in another section of this manual.

In a suppressed-carrier receiver, or when no pilot carrier is received, the a-f-c circuits described are inoperative or may be omitted entirely from the receiver design. A crystal oscillator is usually used in place of the variable oscillator, V3, in such receivers. The use of a crystal oscillator presents no problem when single-frequency operation is desired, and crystals may be switched for multi-frequency operation. However, switching circuits introduce factors disturbing to crystal circuits at high frequencies, especially if ovens are used for temperature control. Synthesized crystal frequencies, using multiple crystal circuits such as found in certain aircraft transceivers, can be used. In any case, both the transmitter and the receiver oscillators must be extremely stable, because no

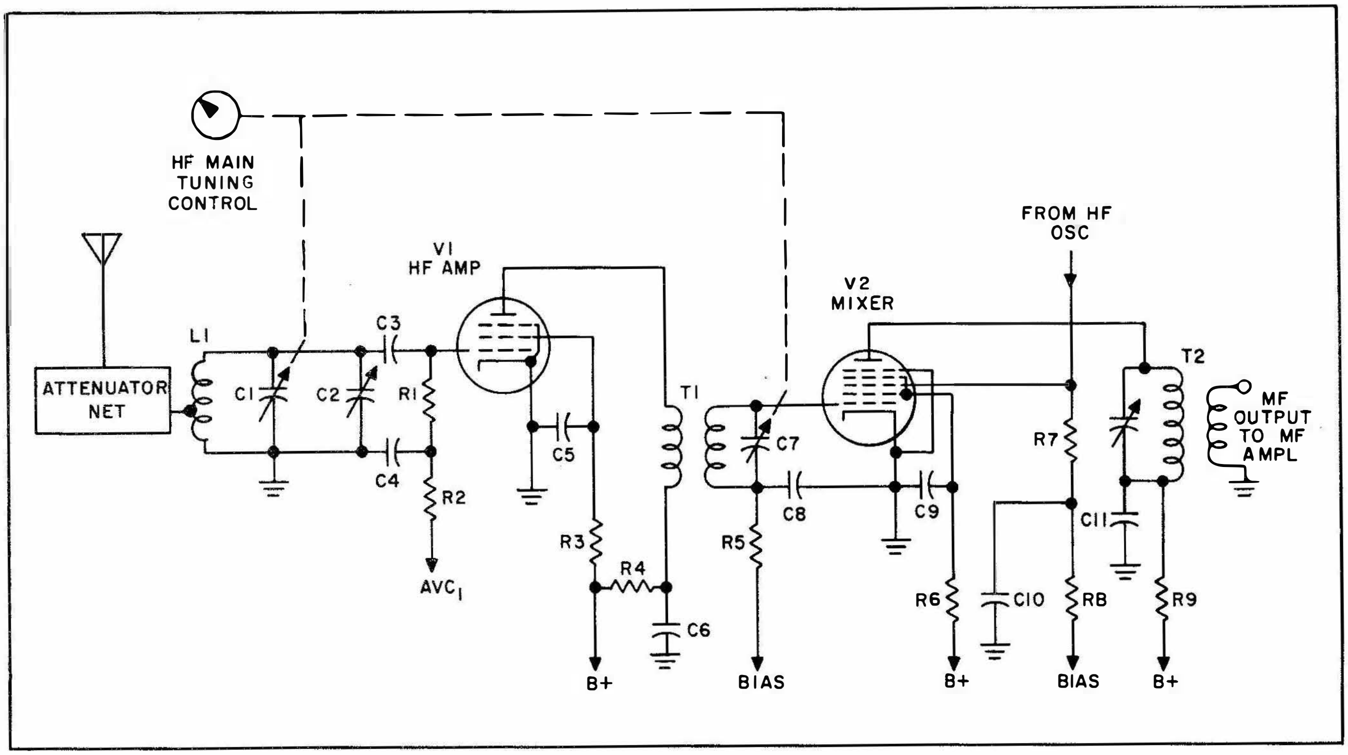


Figure 84. High-Frequency Amplifier and Mixer.

means of locking them together can be employed.

DETAILED CIRCUIT ANALYSIS OF TYPICAL SSB RECEIVER

The typical SSB receiver to be discussed in the following paragraphs incorporates some circuitry which may not necessarily be considered typical of the circuitry found in compact, modernized communications equipment. It is typical of good quality, fixed-station, transoceanic equipment, however, and, since single sideband has been in use in this type of service for a long period of time, circuitry for this type of equipment has been standardized to some extent.

No attempt is made to present a complete schematic of this receiver, since this would be found in any maintenance manual for a particular equipment; therefore, power supplies, metering circuits, and multiple decoupling circuits have been eliminated for simplicity. Since a negative power supply is required for the a-f-c circuit, fixed bias is also used in many of the stages that might otherwise use cathode-resistor bias.

Transformer coupling is shown in much of the circuitry, a method of coupling often used in rack-mounted equipment with various sections on separate chassis, to provide low-impedance coupling so that suitable coaxial cable may be used between the various units. This was found a useful means of isolating the circuits for purposes of discussion. It is not intended to indicate the ideal coupling method, however.

High-Frequency (R-F) Amplifier

The high-frequency amplifier, V1, is shown with the mixer circuit in figure 84.

The incoming signals from the antenna circuit are applied to the antenna coil, L1. The impedance of the antenna circuit input is matched by use of a tap on the antenna coil. In many SSB receivers, an attenuator is used in the antenna circuit, because these receivers are more susceptible to distortion due to overloading than are conventional AM receivers. This attenuator is represented in the diagram in block form. C1 and C7 are the main tuning capacitors, and C2 is a trimmer capacitor across C1 and L1. Coupling capacitor C3 is used to isolate the a-v-c voltage from the d-c ground path through L1. The a-v-c voltage is applied to the control grid through grid resistor R1 and decoupling resistor R2. C4 is the decoupling capacitor for the grid circuit, isolating the avc. R3 is the screen voltage dropping resistor, and also serves as a decoupling resistor. C5 is the screen-decoupling, or bypass, capacitor.

The antenna coil may be tuned to resonate at

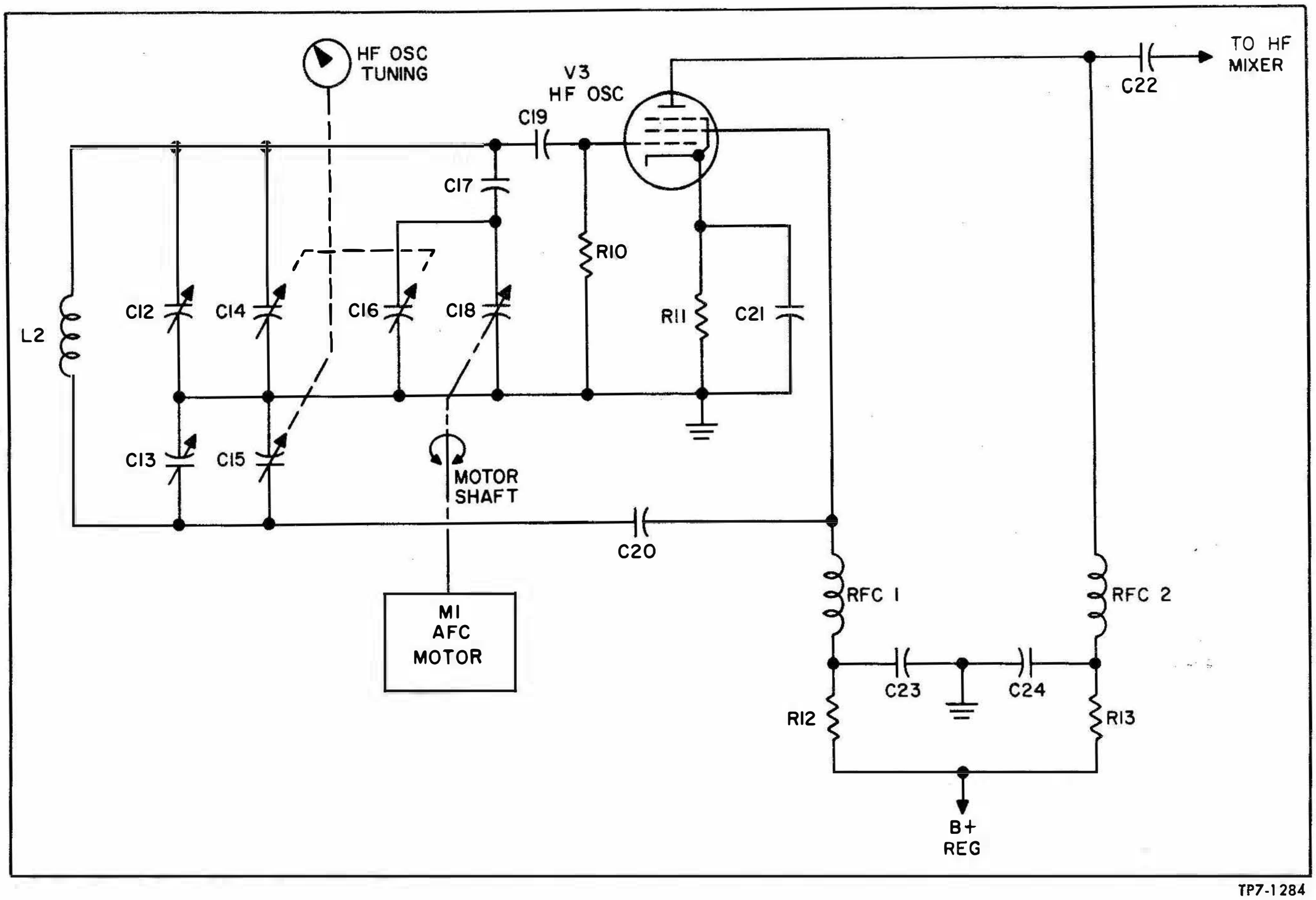


Figure 85. High-Frequency Oscillator.

any frequency between 10 and 30 megacycles, by tuning capacitor C1. The signal selected is amplified in tube V1, and the amplified output appears across the plate load, the primary of T1. C6 and R4 decouple the signal from the B+ supply.

Operation of the high-frequency amplifier in the SSB receiver is identical to that of the r-f amplifier in a conventional AM receiver.

High-Frequency Mixer

The high-frequency mixer, V2, shown in figure 84, mixes the incoming sideband signal with the output of the h-f oscillator, V3, to produce a medium-frequency (or difference-frequency) output of 3.1 to 3.103 mc.

The output of the h-f amplifier is transformer-coupled to the control grid of the mixer. Signal voltages appearing in the primary coil of T1 are impressed on the secondary, which is tuned with variable capacitor C7. This capacitor is ganged with C1, and together they form the main high-frequency tuning control. Thus T1 is tuned to the frequency of L1 and C1, and aids in the preselection of the desired signal, improves image rejection, and helps prevent radiation of the h-f oscillator signal.

The signal from the h-f oscillator is R-C coupled to the injector grid of the mixer tube, V2. Fixed bias is applied to this grid through grid resistor R7, and is also applied to the control grid through the secondary of T1, to establish the proper operating conditions for the tube. R5, R6, R8, and R9 are decoupling resistors, and C8, C9, C10, and C11 are decoupling or bypass capacitors.

The received signal and the oscillator signal and their sum and difference frequencies appear at the plate circuit of the tube. The primary of T2, which forms the plate load, is fixed-tuned to the difference frequency. Therefore only this difference frequency, or medium frequency, appears across the secondary and the output.

High-Frequency Oscillator

The high-frequency oscillator produces the 6.9-to-26.9-mc output applied to the injector grid of the h-f mixer. The oscillator, shown in figure 85, is a modified Colpitts oscillator suitable for use with the receiver circuits discussed.

The oscillator coil, L2, is tuned to resonance at a frequency 3.1 mc below the received signal by the constant-ratio tuning capacitors, C14 and C15. These capacitors are ganged, and may be operated by a separate control knob or may be ganged with h-f tuning capacitors C1 and C7 in the h-f amplifier. C12 and C13 are trimmer capacitors for C14 and C15. C19 and R10 are the coupling capacitor and grid resistor, respectively, which together develop grid-leak bias for proper operation of oscillator tube V3. Additional protective bias is provided by cathode resistor R11. C21 is the cathode-bypass capacitor.

Oscillation is sustained by feedback from the screen grid circuit of the tube through feedback capacitor C20. C22 in the plate circuit capacitively couples the output to the mixer. R-F variations or the oscillator signal is prevented from affecting the power supply and other associated circuits by r-f chokes RFC1 and RFC2, resistors R12 and R13, and bypass capacitors C23 and C24.

C18 is the a-f-c tuning capacitor, and is mechanically coupled (usually through reduction gearing) to the a-f-c motor, M1. A difference in frequency between the received pilot carrier and the h-f oscillator of other than exactly 3.1 mc will result in a change in the carrier frequency at the medium-frequency level (3.1 mc) and, consequently, a difference in frequency between the 100-kc carrier and the 100-kc reference frequency of the l-f oscillator. Since these frequencies operate the a-f-c circuit of which motor M1 is a part, the motor rotates capacitor C18 in the proper direction to shift the frequency of the h-f oscillator to the point where the output of the h-f mixer is exactly 3.1 mc. When the frequency is reached that results in the 100-kc carrier frequency being exactly the frequency of the 100-kc l-f reference oscillator, the output voltage to a-f-c motor M1, drops to zero, and the motor stops.

The tuning range of C18 need not be greater than about 5 kc, because too great a change in the difference frequency at the m-f or l-f stages, especially in the carrier amplifier, will cause the carrier to be outside the narrow pass band of the filters. Therefore, C18 has a relatively small value. To maintain the proper ratio of C18 to the tuned circuit consisting of L2, C14, and C15, a variable capacitance-divider circuit, C16 and C17 is added. C16 is ganged to oscillator tuning capacitors C14 and C15. C16 is in parallel with C18, and therefore maintains the proper L-C ratio as the tuning is varied. C17 is in series with this combination, and its value affects the range of C18.

Medium-Frequency Amplifier

The medium-frequency output of the mixer, V2, figure 86, is taken from the secondary winding of T2, the plate load and output transformer for the h-f mixer stage, and is applied directly to the lattice-type crystal filter, F1. Proper impedance matching is provided by the secondary of T2, and the input transformer incorporated in the filter. This filter is normally composed of two or three sections. It must pass the upper sideband frequencies of 3100 to 3103 kc and must have sufficient response at frequencies slightly below 3100 kc to allow proper a-f-c action. If a change in the carrier frequency or the h-f oscillator frequency causes the m-f carrier to decrease 100 cycles to 3099.9 kc, the a-f-c circuit should bring the oscillator back to the proper frequency. However, if the filter cut off sharply at 3100 kc, then the carrier may be

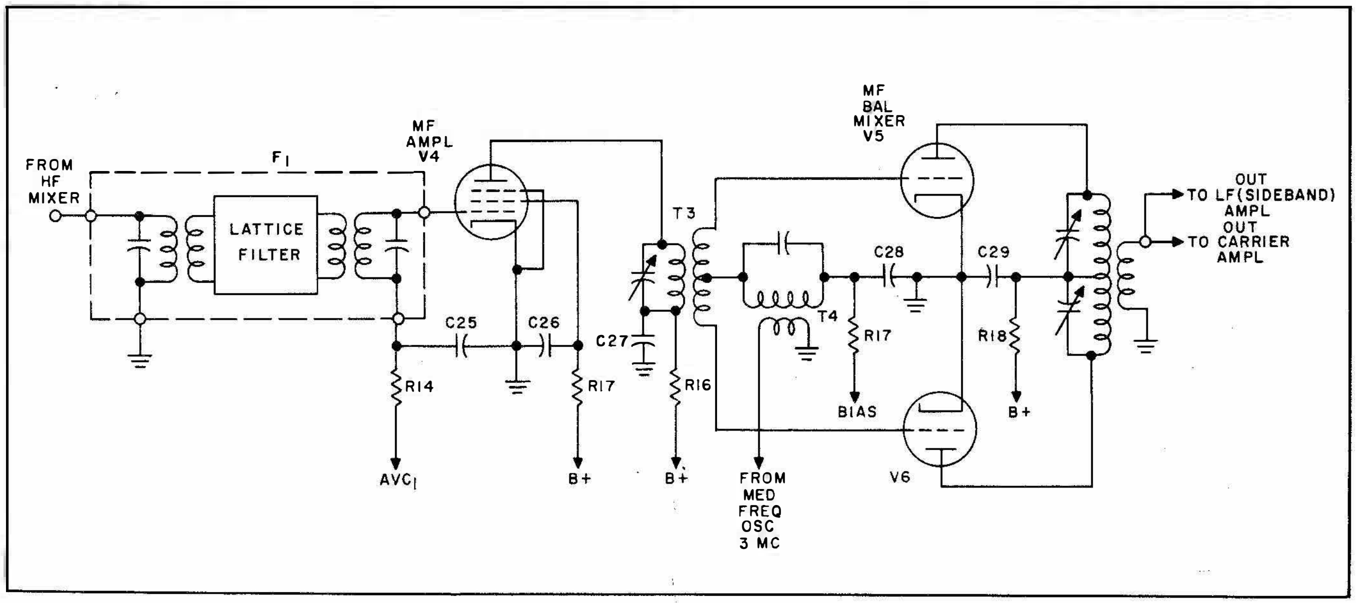


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Figure 86. Medium-Frequency Filter, Amplifier, and Balanced Mixer.

severely attenuated at a slightly lower frequency and cause the a-f-c circuit to drop out of control. Thus the desired limits of a-f-c operation are also considered in the bandpass of the filter, and vice versa. Typical filters and their curves are discussed in detail in another section of this handbook; therefore, only those frequency requirements that apply to the receiver being discussed are mentioned here. The output impedance of the filter must be matched to its load. The impedence of the filter used is sufficiently high to permit it to be connected directly to the grid of the medium-frequency amplifier, V4. A-V-C voltage is applied to the grid of this tube through the secondary of the filter transformer. Decoupling of the signal from the a-v-c circuit is accomplished by decoupling resistor R14 and bypass capacitor C25. V4 is the m-f amplifier tube; a pentode type is normally used to provide sufficient gain, with typical considerations common to i-f amplifiers operating in this frequency range. R15 and C26 are the screen grid decoupling resistor and capacitor. The primary of T3 forms the plate load for tube V4, and plate voltage is applied through this transformer and plate decoupling resistor R16. C27 is the plate decoupling, or bypass, capacitor. The secondary of output transformer T3 forms the balanced input to the m-f balanced mixer.

Tuning of the m-f amplifier may be adjusted slightly with the small trimmer capacitors in the filter circuit, and the primary of output transformer T3 may be peaked with a small variable capacitor in parallel; however, the over-all frequency response of the stage is determined almost exclusively by the frequency characteristic of the filter.

Medium-Frequency Balanced Mixer

The secondary of interstage transformer T3, figure 86, is connected to the grids of balanced mixer tubes V5 and V6 in a push-pull arrangement. The input signal of 3100 to 3103 kc appears in the primary of T3 as the output of the m-f amplifier. The 3-mc oscillator signal from medium-frequency oscillator V7 and V8 is impressed across the primary winding of the oscillator coupling transformer, T4. The secondary of T4 is connected to the center tap of the secondary of T3, capacitor C28 completing the signal path to ground. A fixed bias is applied to the grids through decoupling resistor R17 to the ground terminal of T4. The cathodes of V5 and V6 may then be grounded.

The 3-mc medium-frequency, beating or heterodyne oscillator signal is thus parallel-connected to the grids of V5 and V6, and the input m-f signal of 3100 to 3103 kc is applied in series with the oscillator signal, which it alternately aids and opposes at the opposite, push-pull grids of the balanced mixer.

The primary of the low-frequency output transformer, T5, is connected to the plates of V5 and V6 in push-pull. Plate voltage is applied through plate decoupling resistor R18 to the center tap of the primary of T5. Signal voltage is bypassed to ground by decoupling capacitor C29.

In a carefully balanced circuit, the 3-mc oscillator signal is cancelled out in output transformer T5; since this signal is applied in parallel at the grids of V5 and V6, it appears at the opposite ends of the primary of T5 in like polarity, or 180 degrees out of phase in each half of the winding.

When an incoming upper sideband signal with pilot carrier, or a signal in the filter range of F1 (3100 to 3103 kc), appears at input transformer T3, it is impressed on the grids of V5 and V6 in push-pull, upsetting the balance of the modulator tubes alternately, and at the rate of the signal frequencies. Since the primary of T5 is connected in push-pull, two signal components are produced 180 degrees out of phase at the opposite (or plate) ends of the primary. In the process, because of the non-linear action of tubes V5 and V6, sum and difference frequencies (or upper and lower sidebands) of the two input signals are also produced, and appear 180 degrees out of phase at opposite ends of the primary of T5, so that signals appearing in both halves of the winding will add in amplitude.

The sum, or upper sideband of the two input signals (3 mc and 3.1 to 3.103 mc) is 6.1 to 6.103 mc. Output transformer T5, however, is tuned to the lower sideband, or difference frequency of 100 to 103 kc; therefore, only the difference (since the signal frequency is also out of the frequency range) will appear in the secondary of T5. It must be remembered, however, that although this low-frequency of 100 to 103 kc is the difference frequency, or actually the lower sideband of the heterodyned input signals, it is still the upper sideband of the original carrier, which is now 100 kc at this low-frequency point. This carrier will be present at the output of T5 (assuming pilot-carrier reception), and so will the upper sideband of 100.1 to 103 kc. Here the pilot carrier is applied to the carrier branch amplifier, V14, V15, and V16, and the upper sideband is applied to the low-frequency amplifier V9, V10, and V11. Both amplifiers are actually connected in parallel, but the filters in each amplifier will pass only the applicable frequencies.

Medium-Frequency Oscillator

A modified Pierce-type oscillator mounted in a temperature-controlled oven is used in the medium-frequency, 3-mc oscillator circuit, shown in figure 87.

A 3-mc crystal, Y1, is connected in the grid-to-plate circuit of the m-f oscillator tube, V7. Grid

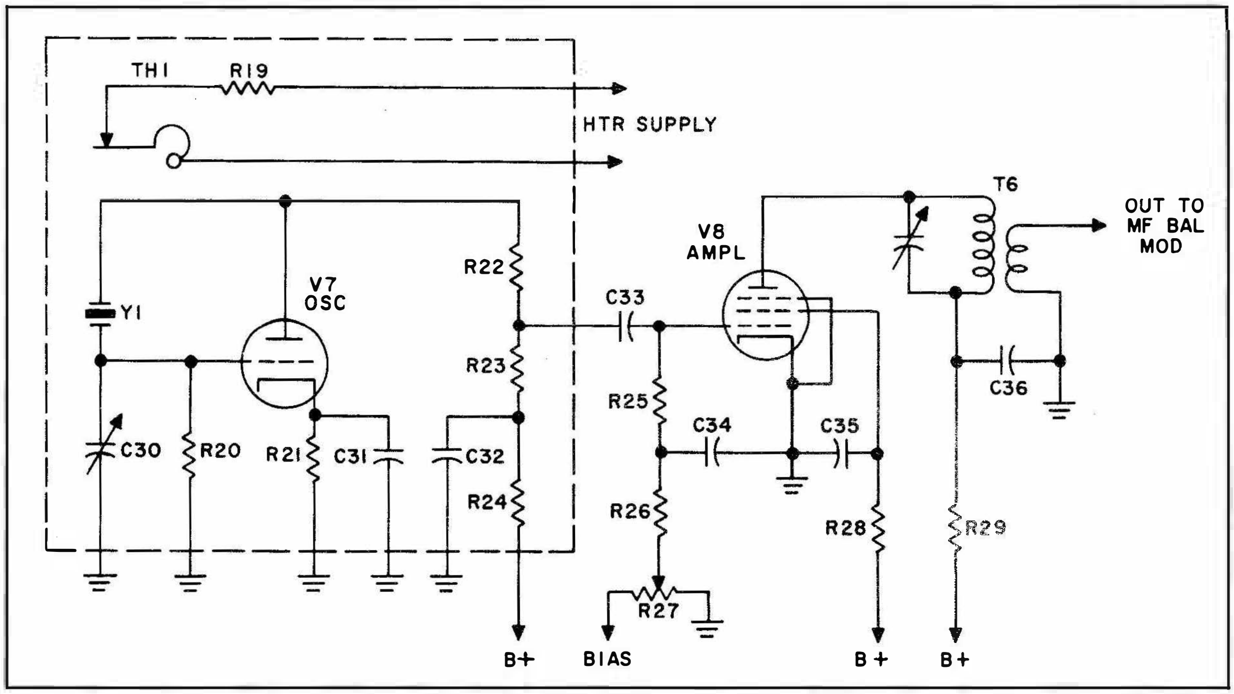

Figure 87. Medium-Frequency Oscillator (3 mc).

resistor R20 with capacitor C30 in parallel provides operating bias. C30 is made variable, to provide a small range of frequency adjustment. Cathode resistor R21 provides additional protective bias, and C31 serves as a bypass to ground for the r-f components.

In the plate circuit of V7, the 3-mc signal is developed across a voltage divider consisting of plate load resistors R22 and R23. Plate voltage for V7 is applied through these resistors, and through decoupling resistor R24, which is bypassed to ground by C32. The output of the oscillator is taken from the junction of R22 and R23, and applied to the grid of buffer-amplifier V8, through coupling capacitor C33. The use of the voltage divider helps to prevent overloading of the oscillator and increases stability. Heater R19 and thermostat TH1 are included in the oscillator compartment, or oven, for increased frequency stability.

The oscillator signal is amplified in buffer-amplifier tube V8. This stage prevents overloading of oscillator V7 and provides a sufficiently high amplitude signal (approximately six volts) at the input to the medium-frequency mixer, or transformer T4. T6, in the plate circuit of amplifier V8, is used for impedance-matching, the secondary of T6 and the primary of T4 having a low impedance, so that suitable coaxial cable may be used. For high-impedance coupling, the primary of T4 could be connected in the plate circuit of V8, eliminating T6. Plate voltage is applied through the primary of T6. R29 and C36 decouple the plate circuit, and R28 and C35 decouple the screen grid circuit. R28 may also serve as the screen voltage dropping resistor. R25 is the grid resistor for the amplifier. Bias may be applied to the stage through R25 and decoupling resistor R26; a bias adjustment, R27, is used to control the gain of the stage, so that the level of the oscillator voltage applied to the balanced modulator may be adjusted. C34 is the grid decoupling capacitor.

A trimmer capacitor across the primary of output transformer T6 serves to peak the output circuit, without affecting the frequency of the oscillator signal.

Low-Frequency Amplifier (Sideband Channel)

The low-frequency output of the medium-frequency mixer, taken from the secondary of T5, is applied to the low-frequency amplifier, V9, V10, and V11, shown in figure 88.

An attenuator is provided at the input of the l-f amplifier, so that proper signal level can be established to prevent overloading of the amplifier. This attenuator may consist of a simple potentiometer circuit, or, more commonly, a step attenuator of the resistor lattice type, calibrated in 6-db steps. A matching transformer, T7, may be used, or, if high-impedance coupling is desired, the output of the attenuator circuit may be applied directly to the grid of the first l-f amplifier, V9.

The output of amplifier V9 is applied to l-f bandpass filter F2. This filter may be of the crystal-lattice type, since crystals usually are employed

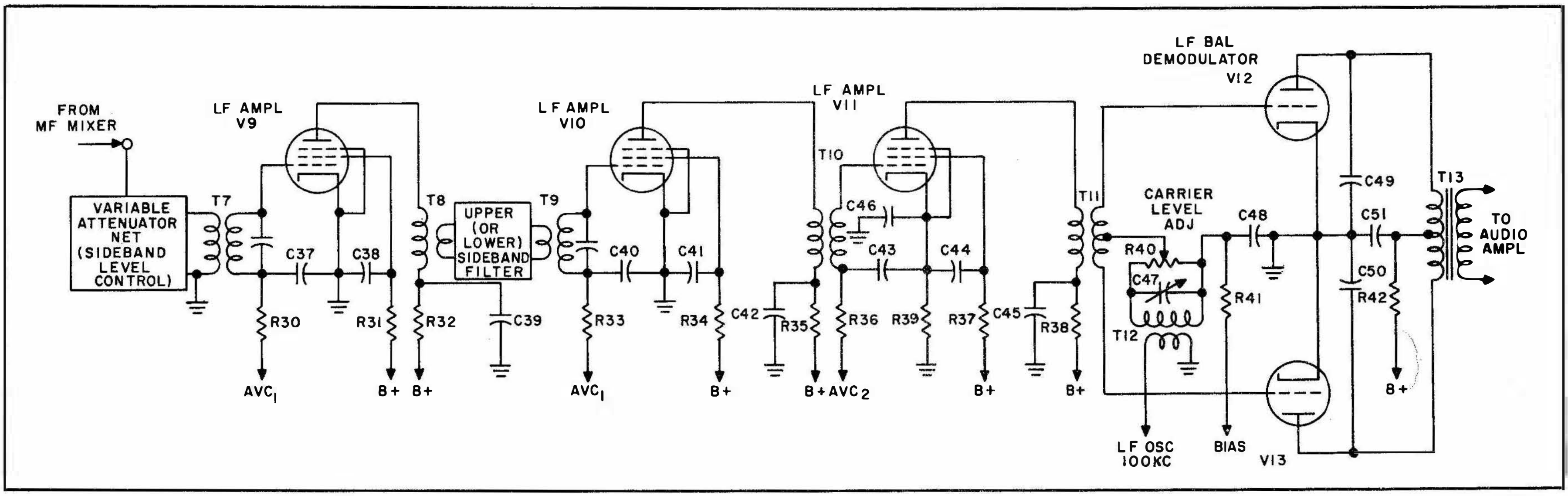

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Figure 88. Low-Frequency Amplifier, Filter, and Balanced Demodulator (100 kc).

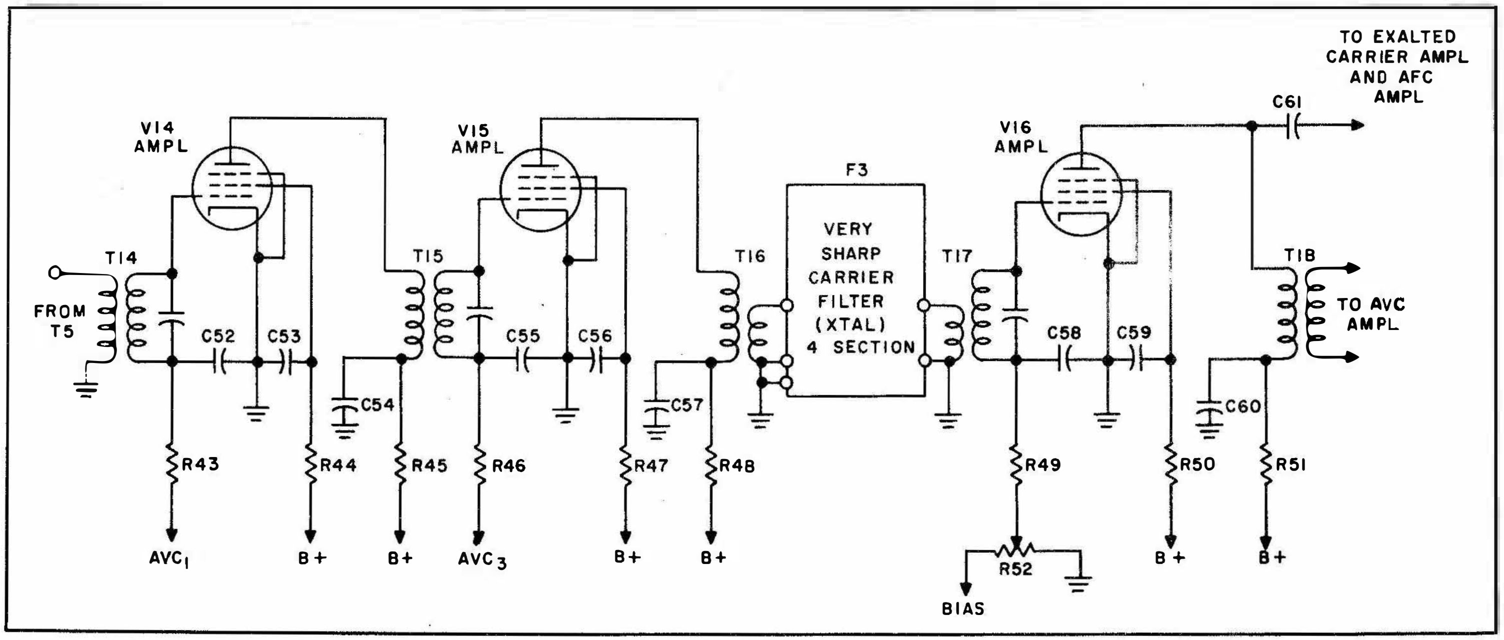

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Figure 89. Carrier Amplifier and Filter (100 kc).

in the 100-kc region. (Mechanical filters are also employed in low frequency circuits, but are presently designed for the 250-kc region.) Filter F2 is designed to pass the frequencies of the upper sideband, or 100.1 kc to 103 kc with a sharp cutoff below 100.1 kc. This sharp cutoff is necessary to insure that the 100-kc carrier is rejected, or suppressed to such a low level that it will not interfere with the proper operation of the low-frequency balanced demodulator.

The frequency response of the low-frequency amplifier depends almost exclusively on the frequency characteristics of filter F2.

Transformer coupling is employed at the input and output of filter F2; the transformers may be included in the filter, or separate transformers may be used if low-impedance filter circuits are used, as shown in this circuit. The primary of T8 forms the plate load of V9, and the secondary matches the input impedance of the filter, while the primary of T9 matches the output of the filter. The secondary of T9 is in the grid circuit of the second l-f amplifier, V10. Here the filtered signal is again amplified, and the output is transformer-coupled to the third l-f amplifier, V11. T10 serves as the interstage transformer, with the plate voltage for V10 being applied to the primary and a reduced a-v-c voltage (avc-2) being applied through the secondary winding to the grid of V11. Plate voltage is applied to V11 through the primary of T11, the input transformer to the low-frequency balanced demodulator, the primary of T11 serving as the plate load for amplifier V11.

The long-time-constant a-v-c voltage (avc-1) is applied to the first two amplifiers, V9 and V10, and a separate reduced a-v-c voltage is applied to V11, to which additional bias is provided by cathode resistor R39. Some effects of fading are reduced by the use of this separate a-v-c voltage.

Resistors R30 through R38 are decoupling resistors, and capacitors C37 through C45 are decoupling (bypass) capacitors for the grid, screen, and plate circuits. Cathode resistor R39, in the cathode circuit of V11, is bypassed with capacitor C46.

Some trimming of tuned circuits may be required; however, the use of a filter circuit permits rather broadly tuned transformers, trimmed only to insure adequate gain over the relatively narrow pass band.

Low-Frequency Balanced Demodulator

The low-frequency balanced demodulator, or detector, shown in figure 88, functions in the same manner as the balanced mixer shown in figure 86. Both operate the same as a balanced modulator, except that input and output frequencies are reversed. All are commonly called "balanced modulators;" however, since the purpose of such a circuit in the receiver is to extract the audio, or modulation, components from the sideband, the circuit is more appropriately called a "demodulator." In conventional AM receivers it is commonly referred to as a "detector." The principal difference between the balanced demodulator and an AM detector is that the former makes use of the frequency-conversion process, rather than rectification, as pointed out previously.

The center-tapped secondary of input transformer T11 is connected to the grids of tubes V12 and V13 in a push-pull arrangement. Audio transformer T13, in the output of the demodulator, is also connected in push-pull in the plate circuit of V12 and V13.

Either the amplified 100-kc output of the l-f carrier insertion oscillator (V20) or the reconditioned 100-kc carrier, whichever has been selected, is introduced at the primary of the carrier oscillator coupling transformer, T12. One end of the secondary of T12 is connected to the center tap of input transformer T11, and the other end of the secondary is bypassed to ground through capacitor C48. Carrier insertion voltage may be adjusted by means of potentiometer R40, across the secondary of T12. Trimmer capacitor C47 may be used to peak the transformer output. Bias is applied through grid decoupling resistor R41 and input transformers T11 and T12.

The cathodes of both tubes are grounded, and plate voltage is supplied through decoupling resistor R42, and the center tap of the primary of T13. Capacitor C51 bypasses or decouples the signal from the power-supply circuit.

The balanced bridge circuit consisting of the center-tapped input and output transformers and the equal tube circuits, present equal, but opposite, paths to the 100-kc carrier, present at transformer T12. Upper sideband signals within the frequency range of 100.1 to 103 kc are impressed across input transformer T11, the primary of which is the plate load of the last l-f amplifier, V11. These signals appear at the grids of tubes V12 and V13 in opposite polarity, and are therefore present in the plate circuit of the tubes in opposite polarity, so that the phase of the signal in one half of the output transformer primary will aid the phase in the other half. This action is similar to that of a push-pull amplifier.

In the process, the 100-kc carrier frequency is heterodyned with the incoming upper sideband signal of 100.1 to 103 kc to produce the upper and lower sidebands, or the sum and difference frequencies of the two signals. Since T13 is an audio transformer, the signal frequencies and the sum frequencies are simply bypassed to ground with capacitors C49 and C50 across the primary of the transformer. The difference frequencies are 100 to 3000 cycles, or the desired audio modulation. This output is taken from the secondary of T13 and handled in conventional audio circuits.

Carrier re-insertion voltage is usually on the order of 6 volts, and incoming sideband signals are usually maintained at a maximum of ½ volt, or a ratio of at least ten to one, so that overmodulation will be prevented. Overmodulation in a balanced modulator, or demodulator, produces distortion, or has the same effect as overmodulation of a conventional AM transmitter.

Carrier Amplifier

The low-frequency output of the medium-frequency balanced modulator V12 and V13, taken from the secondary of output transformer T5, may be applied also to the 100-kc carrier amplifier, in parallel with the 100-to-103-kc sideband channel l-f amplifier. The carrier amplifier is shown in figure 89.

Interstage transformer T14 couples the 100-kc signal to the grid of the first carrier amplifier, V14, where it is amplified and coupled through another interstage transformer, T15, the primary of which is in the plate circuit of V14, and the secondary of which is in the grid circuit of the second carrier amplifier, V15. Here the signal is again amplified and appears in coupling transformer T16, the primary of which is the plate load for V15, and the secondary of which is applied to carrier filter F3. The fact that some of the sideband information is also amplified along with the 100-kc carrier is insignificant, because of the extremely narrow bandpass (usually under 100 cycles at the —6-db points) of the carrier filter, with its sharp rejection characteristics for frequencies either side of this bandpass. As in most SSB filter applications, the bandpass of the amplifier is determined almost exclusively by the filter employed. A four-section crystal filter is used in this case so that the proper frequency response curve with sharp cutoff can be obtained.

Proper impedance match should be maintained for the input and the output of the filter. This is accomplished through proper choice of the coupling transformers, T16 and T17. T17 couples the output of the filter, which for all practical purposes now contains only the carrier, to the grid of the third carrier amplifier, V16. Here the carrier is again amplified, and the output appears across the plate load of V16, or the primary of coupling transformer T18. The carrier is applied to the a-v-c circuit and V17 from the secondary winding of T18, and another output is taken directly from the plate of V16, through coupling capacitor C61, and applied to the a-f-c circuit and exalted-carrier amplifier V18.

Decoupling of the various stages is accomplished by resistors R43 through R51 and capacitors C52 through C60. Regular a-v-c voltage (avc-1) with the long-time-constant feature (on the order of 8 seconds) is applied to the grid of the first carrier amplifier tube, V14, and a selectable-time-constant a-v-c voltage (avc-3) is applied to the grid of the second carrier amplifier tube, V15. This time constant, which is selectable between .1 second and 2 seconds by means of a front panel selector switch, may be varied to suit particular fading conditions. This provides a means of compensation because of the fact that the output of the carrier amplifier is applied to the circuit where the a-v-c voltages are developed.

An adjustable bias is applied to the third carrier amplifier tube, V16; this bias is controlled by means of potentiometer R52.

The A-V-C Circuit

The a-v-c circuit in this receiver uses a slow-acting long-time-constant circuit, a compensating short time constant circuit, and a reduced level, forward-acting avc, for additional compensation (figure 90).

The amplified carrier input for the a-v-c circuit is conveniently taken from the secondary of coupling transformer T18 in the output circuit of the carrier amplifier, or plate circuit of V16. This carrier is impressed on the grid of a-v-c rectifier tube V17.

The plate circuit of V17 is connected in series with resistance network R56 and R57 to ground, and the cathode is connected through potentiometer R54 to a negative supply. Thus voltages appearing in the plate circuit will be of a negative value with respect to ground, the value of this voltage depending on the signal level at the grid of V17 and the setting of bias control R54. One end of bias control R54 is connected to the negative supply, and the other end is connected to ground through R53, which together with R54 forms a voltage divider between the negative supply and ground. The negative voltages appearing in the plate circuit of V17 are applied to the grids of the various tubes in the receiver for automatic volume control.

A tuning meter, M, is connected in the plate circuit of V17. The maximum carrier signal appearing at the grid of this tube causes increased conduction, and a rise in plate current. Thus a plate current meter is conveniently used as the tuning meter. The rise in plate current causes increased voltage drop across R56 and R57, the plate load resistors, and thus develops the a-v-c voltage. C63 is a bypass capacitor for bias resistor R54. Screen voltage is applied from ground through screen resistor R55, and C62 bypasses the screen grid to the cathode (since the cathode is the negative reference point in the signal circuit).

The main slow-acting a-v-c voltage (avc-1) is

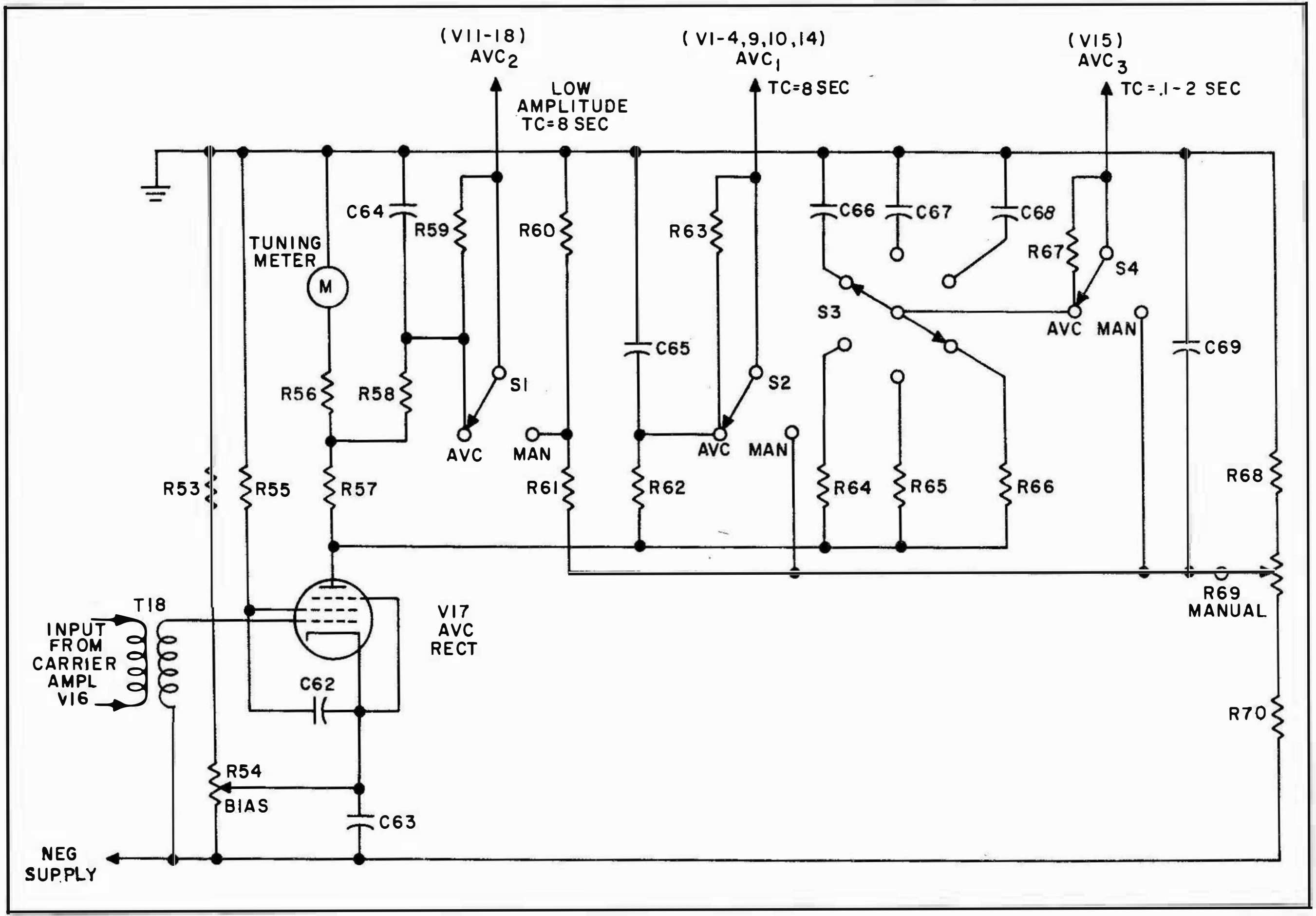

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Figure 90. A-V-C Circuit.

taken directly from the plate of V17, and applied through R62 and switch S2 to the grid circuits of the h-f amplifier, V1, the m-f amplifier, V4, the first and second l-f amplifiers in the sideband channel, V9 and V10, and the first carrier amplifier, V14. A large-value capacitor (on the order of several microfarads), C65, is connected from the grid end of R62 to ground. Thus C65 and R62 are the principal components which govern the time constant of the a-v-c circuit for these five stages, the discharge path for C65 being completed through R62, plate load resistors R56 and R57, and the tuning meter to ground. A time constant of about 8 seconds usually proves satisfactory in this type of receiver.

Another a-v-c branch (avc-3) is also taken from the plate of V17 and applied to the grid of the second carrier amplifier tube, V15. This circuit functions in the same manner as the normal a-v-c circuit, except that selectable time constants of shorter duration (from .1 second to 2 seconds) are employed. The selector switch, S3, selects either capacitor C66 and resistor R66, or C67 and R65, or C68 and R64. More combinations are frequently used, although only three are shown here to indicate the method. Various values of resistors and capacitors may be used, separately or in parallel, by use of the selector switch. Some conditions of flutter caused by fading can be overcome by proper selection of the time constant in this circuit, since it operates in the carrier circuit ahead of the a-f-c rectifier, and has a bootstrap effect on the a-v-c voltage itself.

A low-amplitude a-v-c voltage (avc-2) is taken from the junction of the two series plate load resistors, R56 and R57. C64 and R58 are the time-constant-governing components, and the voltage is applied through resistor R58 and switch S1 to the grids of the third low frequency amplifier, V11, in the sideband channel and to the exalted-carrier amplifier, V18. Since V11 and V18 operate in circuits following the a-v-c take-off point, this type of a-v-c circuit is usually called a "forward-acting a-v-c circuit." With this circuit the control of tubes V11 and V18 by the a-v-c voltage has no effect on the signal applied to a-v-c rectifier V17. Better control can be obtained in this manner, and a more uniform output may be realized by using this or a similar combination of separate a-v-c circuits. The time constant of this circuit is also long (about 8 seconds) for the same reasons given in the discussion of the avc-1 circuit.

A manual volume control, R69, is connected in a series divider network with resistors R68 and R70, between the negative supply and ground. The AVC-MANUAL switches, S1, S2, and S4 (which may be ganged), provide a means of selecting either the a-v-c voltage or the manually adjustable voltage determined by the setting of R69. Primarily the latter voltage is used during tuning or in the absence of a carrier.

Resistors R60 and R61 form a voltage divider for the low-amplitude circuit for the grids of V11 and V18. The high-value resistors, R59, R63, and R67, provide a d-c grid return path during switching, to help minimize switching noise in the output of the receiver, and capacitor C69 also prevents switching transients from affecting the negative-supply circuits.

The Low-Frequency Oscillator

The importance of frequency stability in the single-sideband receiver may be reflected in the chioce of a good low-frequency oscillator circuit. Since the low-frequency oscillator serves the twofold purpose of providing a reference frequency for the a-f-c circuit, or control of the h-f oscillator frequency, and providing the carrier for reinsertion with the signal at the demodulator, it may well be regarded as the heart of the SSB receiver. Hence, it may be termed the "reference oscillator," or the "carrier re-insertion oscillator," or simply the "low-frequency, or 100-kc oscillator."

Fortunately, the use of a relatively low frequency makes requirements for lead length and parts layout or placement less critical than would be found in the use of a higher frequency, although, on the other hand, the tolerance and stability required tend to offset this factor. The frequency of 100 kc is well chosen, as this is the standard frequency used in many types of test equipment and frequency-controlling devices. Crystals are easily obtained at this frequency, and accuracy may be combined with ruggedness in all associated components.

Figure 91 shows a Meacham oscillator in a form applicable to the receiver being considered. This oscillator is of the bridge-stabilized, linear type, which is frequently used in long-range, fixed, ground or transoceanic telegraph equip-

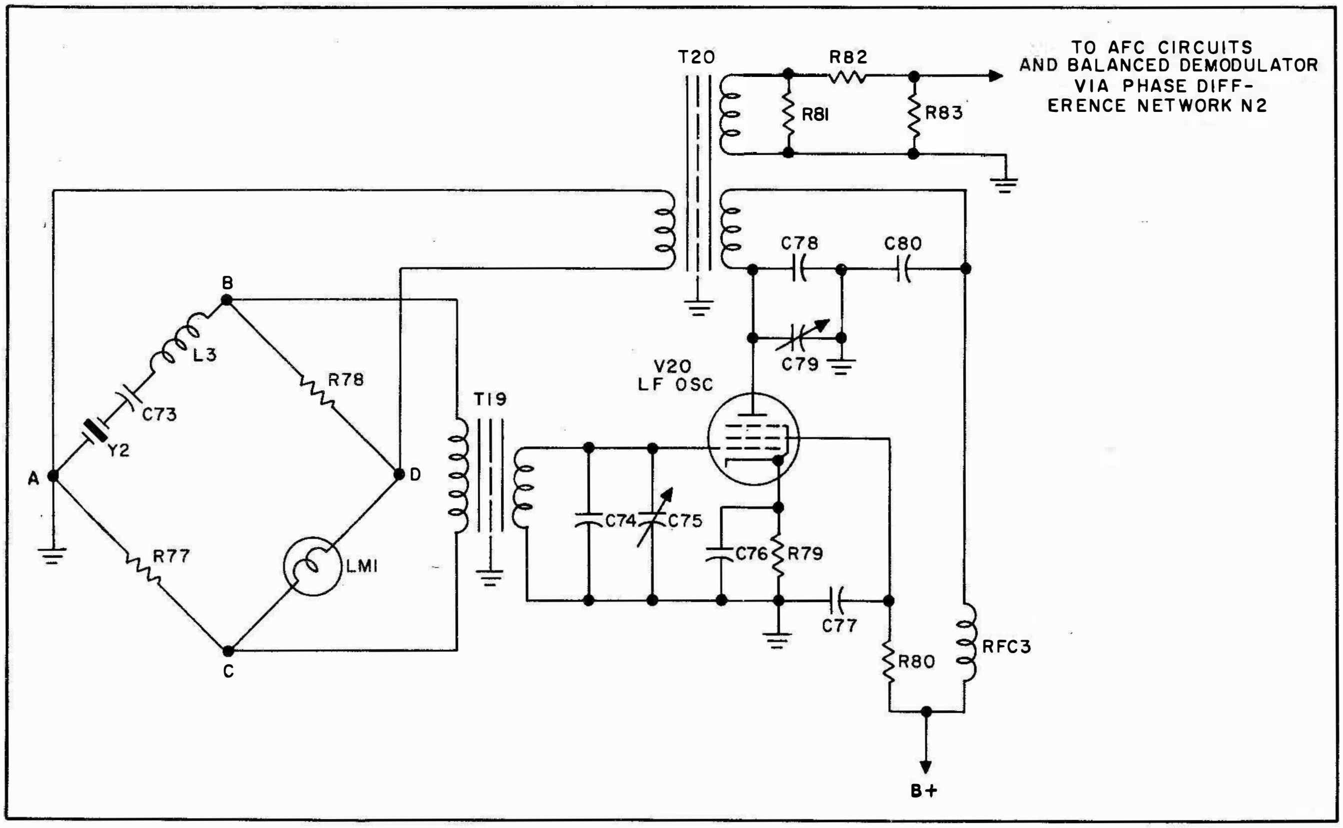

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Figure 91. Low-Frequency Oscillator (100 kc) (Meacham Circuit).

ment. It is claimed to possess very great frequency stability.

A GT cut crystal, Y2, is used in a bridge circuit consisting of: one arm, A-B, containing the crystal, a capacitor, C73, and an inductor, L3, in series connection; an opposite arm, C-D, consisting of a small incandescent lamp, LM1; and the other two bridge arms, A-C and B-D, consisting of equal, low-value resistors. The output of the bridge is taken from junctions B and C, and applied to the low-impedance primary of grid transformer T19. The secondary of T19 is matched to the grid of oscillator tube V20.

Oscillator tube V20 operates as a linear amplifier, rather than in the limiting, or non-linear condition usually found in conventional oscillators. Bias is obtained by means of cathode resistor R79, and the tube should operate Class A. Of interest is the absence of a grid resistor, which is usually employed in other types of oscillators to develop bias from the grid current.

The primary of plate transformer T20 is in the plate circuit of tube V20, and plate voltage is applied through this winding. R-F choke RFC3 prevents the oscillator signal from appearing in the power-supply circuits, and capacitor C80 provides a bypass to ground. Screen voltage is supplied through screen voltage-dropping resistor R80, and is bypassed to ground by decoupling capacitor C77. C76 serves as a cathode resistor bypass.

The feedback loop, to sustain oscillations, is provided by a low-impedance winding on plate transformer T20. The loop is connected to the oscillator bridge network at points A and D, of which point A is grounded.

Lamp LM1 provides a thermistor action as a result of its positive temperature coefficient; that is, the resistance of its filament increases with an increase in temperature. The combination of circuit elements and the lamp type are so chosen that the lamp normally does not reach full brilliance, but usually glows a dull red, corresponding to a temperature of about 900° K. Its resistance at this temperature is about four times that of its resistance when cold (room temperature).

Initially the bridge is in a state of unbalance. Three of the arms contain pure resistance, while the series-resonant arm, L3, C73, and Y2 (A-B) offers reactance. The bridge approaches balance only at the resonant frequency (100 kc), where the reactance of arm A-B vanishes, and the bridge becomes purely resistive.

During oscillation, the amplitude adjusts itself until the loss of the bridge is equal to the input, depending on the gain of the amplifier. Thus the bridge, with its thermistor (lamp), serves as both resonator and limiter. A further increase in amplitude results in an increase in current through lamp LM1, directly and also as a result of the output of the amplifier applied back through the feedback loop. The balancing action of the bridge circuit tends to increase the effective Q of the series-resonant circuit by magnifying the phase shift produced as a result of any frequency deviation.

The Meacham oscillator is similar to the more familiar Wien-bridge oscillator, in which the bridge is directly connected to one of the two tubes. However, the stability is greatly increased in the Meacham circuit, since none of the bridge elements are directly in the tube circuits.

The frequency of oscillation of the Meacham circuit is also self-adjusted to the point where the phase shift of the bridge is equal and opposite to that of the amplifier. Powdered-iron-core toroidal transformers are usually used at this frequency (100 kc) for T19 and T20. This type of core provides tight coupling so that minimum phase shift may be attained. The transformers are tuned to resonance at the operating frequency by means of capacitors C74 and C78 and their trimmers, C75 and C79.

Output is taken from the oscillator from a third winding on T20. This winding may be designed to properly match the load, or an impedance-matching network consisting of resistors R81, R82, and R83 may be employed. The output is applied through a phase-difference network to the a-f-c circuits and the low-frequency balanced demodulator.

Carrier Insertion and the A-F-C Circuit Oscillator and Exalted Carrier Amplifiers

The 100-kc output of the low-frequency oscillator, V20, might be applied directly to the low-frequency balanced demodulator if it were of sufficient amplitude. However, to prevent overmodulation of the carrier (oscillator signal), it should be reasonably greater in amplitude than the sideband signal at the demodulator, and is usually inserted at a level at least ten times greater than that of the sideband voltage. Since the sideband voltage is about ½ volt at the demodulator input, the inserted carrier is usually 5 or 6 volts. This level might be obtained from the output transformer of the oscillator, but not without serious loading. Therefore, a stage or more of amplification is usually necessary.

If the output of oscillator V20 were amplified only by the oscillator output amplifier, V19, sufficient amplitude would be attained. However, since the a-f-c circuit requires two components of the oscillator signal, shifted 90 degrees apart in phase from each other, advantage is taken of additional amplification in one of these branches, before application of the 100-kc signal to V19.

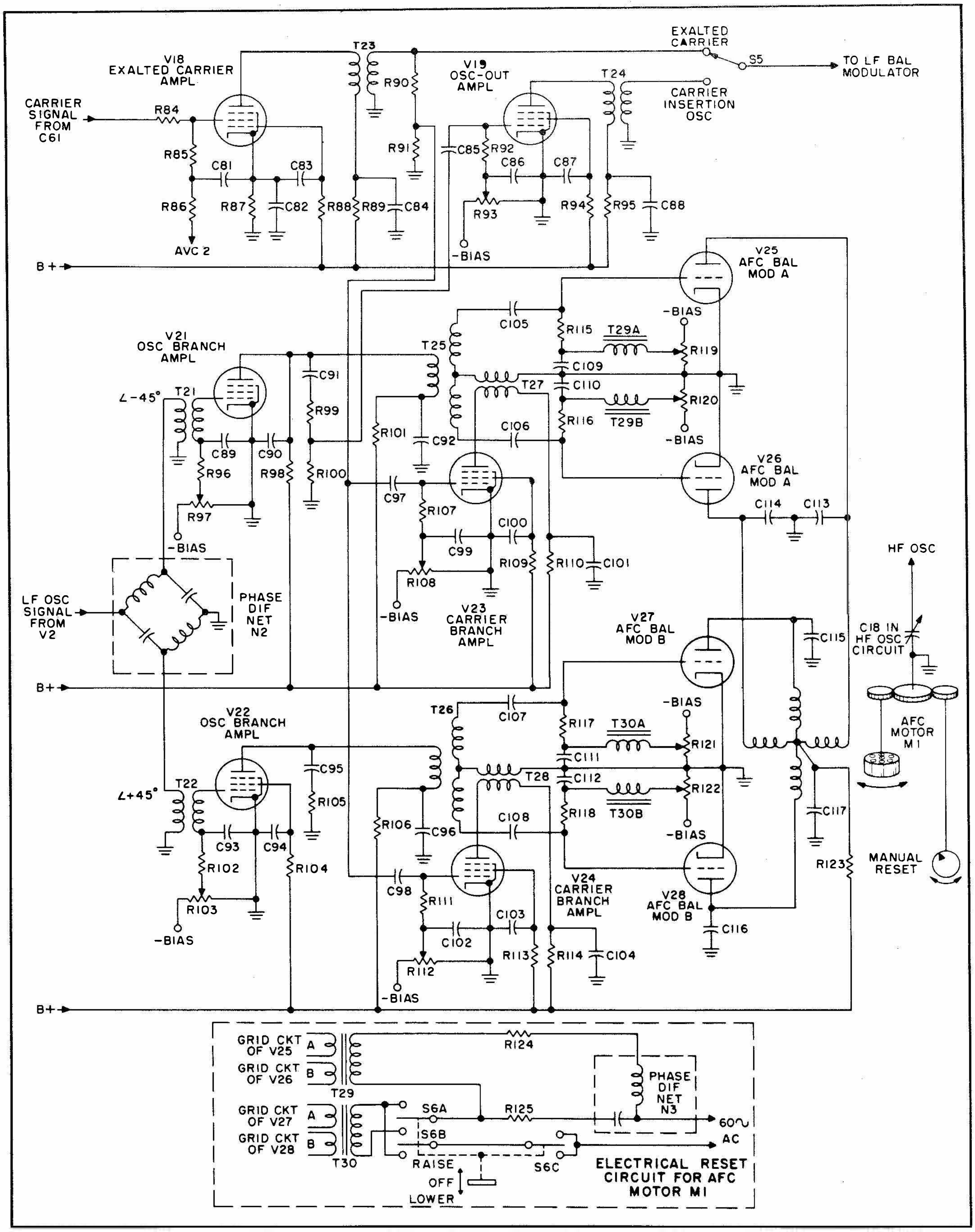

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Figure 92. A-F-C Circuit (Comparator Type).

The shift in the phase of the signal is of no consequence, since the phase of the inserted carrier in SSB detection is unimportant.

Refer to figure 92 for an over-all schematic of the a-f-c circuit, and to figure 93 for the same circuit simplified, showing the signal paths.

The output of l-f oscillator V20, taken from plate transformer T20, is applied through an impedance-matching network (R81, R82, and R83), to a phase-difference network (or a split phase-shift network), N2, consisting of capacitors and inductors (shown simplified). Any of several different types of networks may actually be used. This network shifts the phase of the 100-kc signal to 45 degrees leading in one branch and 45 degrees lagging in the other branch, for a total phase shift of 90 degrees at the output of the network. The leading (inductive) phase is applied to the grid of amplifier V21 through impedance-matching transformer T21, and the lagging (capacitive) phase is applied to amplifier V22 through transformer T22.

The output of amplifiers V21 and V22 is of sufficient amplitude for use in the balanced modulators; therefore, the output of V21 is applied to the primary of T25, the input transformer of the a-f-c balanced modulator (A), and the output of V22 is applied to T26, the input transformer of the other a-f-c balanced modulator (B). The primaries of these transformers serve as the plate loads for V21 and V22, and plate voltage is applied through decoupling resistors R101 and R106 and the primaries of T25 and T26. C92 and C96 are plate decoupling or bypass capacitors. Screen voltage is applied to tubes V21 and V22 through resistors R98 and R104, which are bypassed by capacitors C90 and C94. The cathodes of the tubes are grounded, and bias is applied from the negative supply through bias adjustment potentiometers R97 and R103 and grid decoupling resistors R96 and R102 to the grid winding of transformers T21 and T22. C89 and C93 serve as decoupling capacitors for the grid circuits.

An output is taken from the plate of amplifier V21 and applied through blocking capacitor C91 to a voltage divider consisting of R99 and R100 connected to ground. A reduced oscillator signal is taken from the junction of resistors R99 and R100, and applied through coupling capacitor C85 to the grid of the oscillator output amplifier tube, V19. The use of the resistance divider network, R99 and R100, provides some isolation of V19 from the signals in the a-f-c balanced modulator which may be reflected into the primary of T25, also in the plate circuit of V21.

Since R99 and R100 represent sufficient load on amplifier V21 to upset the equality of the two carrier branches, and thus upset the desired balance of the two a-f-c modulators, a similar load is placed on amplifier V22, by capacitor C95 and resistor R105 in the plate circuit of this tube.

The phase of the re-inserted carrier, or oscillator signal, has no effect on the demodulation in SSB receivers (not the case in some types of DSB reception); therefore, the fact that a phase shift has been introduced in this branch of the oscillator signal path is insignificant. In fact, V19 may just as well have been connected in the other branch, or plate circuit of V22.

The oscillator signal applied to the grid of the oscillator output amplifier, V19, is amplified and appears in the plate circuit, of which the primary of coupling transformer T24 is a part. The output, taken from the secondary of T24, may be applied to the low-frequency balanced demodulator at the primary of its carrier input transformer, T12, when selector switch S5 is in the oscillator position. The level of the oscillator signal applied to the demodulator may be equalized, or the gain of V19 may be set by means of bias adjustment R93 in the grid circuit of V19, which is across the negative supply. R92 is the grid resistor, C85 is the coupling capacitor, and C86 is the decoupling capacitor. Screen voltage for the tube is furnished through screen resistor R94, which is decoupled with bypass capacitor C87. Plate voltage is connected through the primary of T24 and decoupling resistor R95. C98 is the plate decoupling capacitor.

If conditions are favorable, the reconditioned carrier may be used for demodulation, rather than the locally generated oscillator signal. The output of the carrier amplifier, at the plate of V16, may be coupled by means of coupling capacitor C61 to the grid of exalted-carrier amplifier V18, where it is amplified and fed through coupling transformer T23 to the carrier input transformer, T12, at the l-f demodulator. The selection of either the exalted carrier or the oscillator signal is made possible by selector switch S5. The output of the carrier amplifier at V16 is of insufficient amplitude to be used directly for demodulation, since an equal number of stages have been used for carrier amplification to this point as were used for side-band signal amplification to the demodulator, although additional gain can be realized in the narrow-band carrier amplifier. The additional stage, V18, also provides a point, following the a-v-c takeoff (T18 in parallel with C61), where the forward-acting a-v-c voltage (avc-2) may be applied, for compensation of varying signal strength due to fading, etc, of the received carrier. AVC-2 is applied to the grid of V18 through decoupling resistor R86 and grid resistor R85. Capacitor C81 decouples the a-v-c circuit at this stage. Cathode bias is provided by cathode resistor R87 and bypass capacitor C82. Screen and plate voltages are applied through resistors R88 and R89. C83 and C84 are decoupling capacitors for the screen and plate circuits,

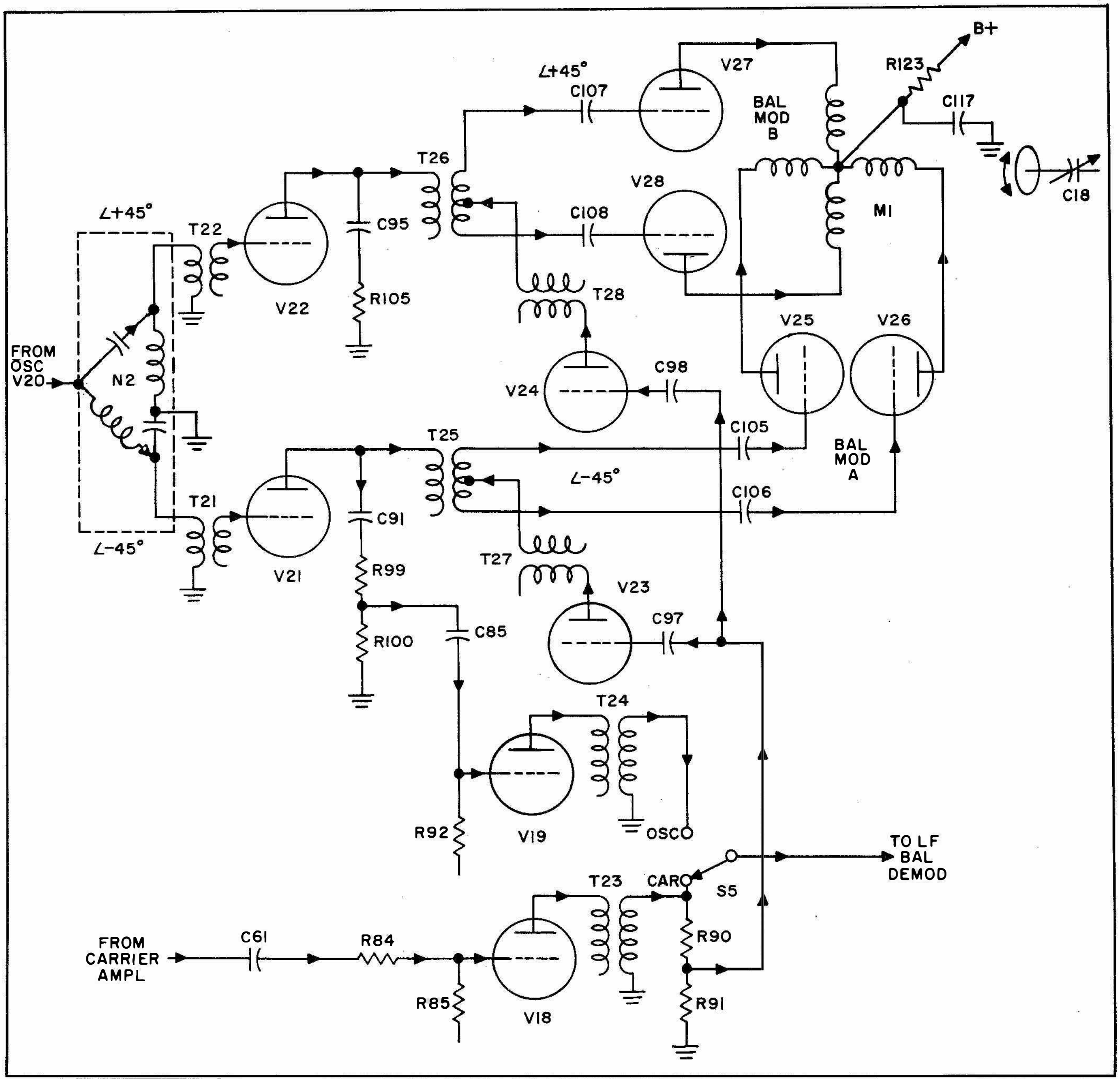

Figure 93. Simplified Diagram of A-F-C Circuit.

respectively. R84, in the grid circuit of V18, and grid resistor R85 form a voltage divider for signals coupled from the plate circuit of V11, in the carrier amplifier. This divider provides some isolation between the two outputs of V11.

A voltage divider, consisting of R90 and R91, is connected across the secondary of the exalted-carrier coupling transformer, T23. A second output is taken from the junction of these two resistors and applied, through coupling capacitors C97 and C98, to the grids of the carrier branch amplifiers, V23 and V24. This voltage divider (R90 and R91) provides some isolation between the l-f balanced demodulator and the carrier branch amplifiers. V23 and V24, the inputs of which are in parallel, separately amplify the carrier signal and apply it to the two a-f-c balanced modulators through carrier input transformers T27 and T28, the primary windings of which are in the plate circuits of V23 and V24, respectively. Fixed bias is applied to the two carrier branch amplifiers through potentiometers R108 and R112 and grid resistors R107 and R111. Capacitors C99 and C102 provide grid decoupling. Screen

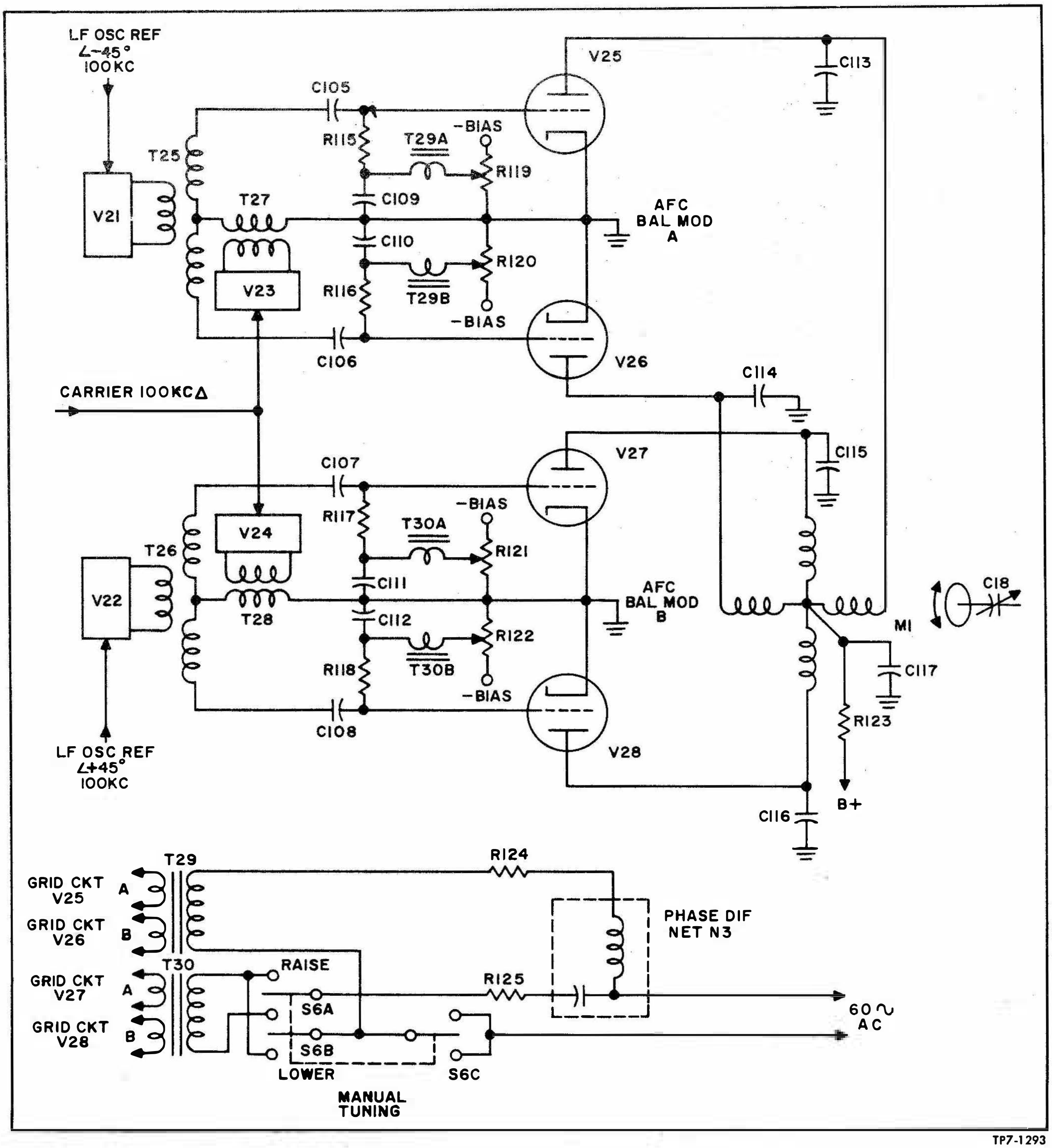

Figure 94. A-F-C Balanced Modulators.

voltage is supplied to the two tubes through screen resistors R109 and R113. C100 and C103 are screen bypass capacitors. Plate voltage is applied through decoupling resistors R110 and R114 and the primaries of carrier input transformers T27 and T28 in the plate circuit of the tubes. C101 and C104 are plate decoupling capacitors. The cathodes of the tubes are grounded.

A-F-C Balanced Modulators

In the comparator type of a-f-c circuit shown in figure 92, a motor is used to drive a small variable capacitor in the high-frequency oscillator

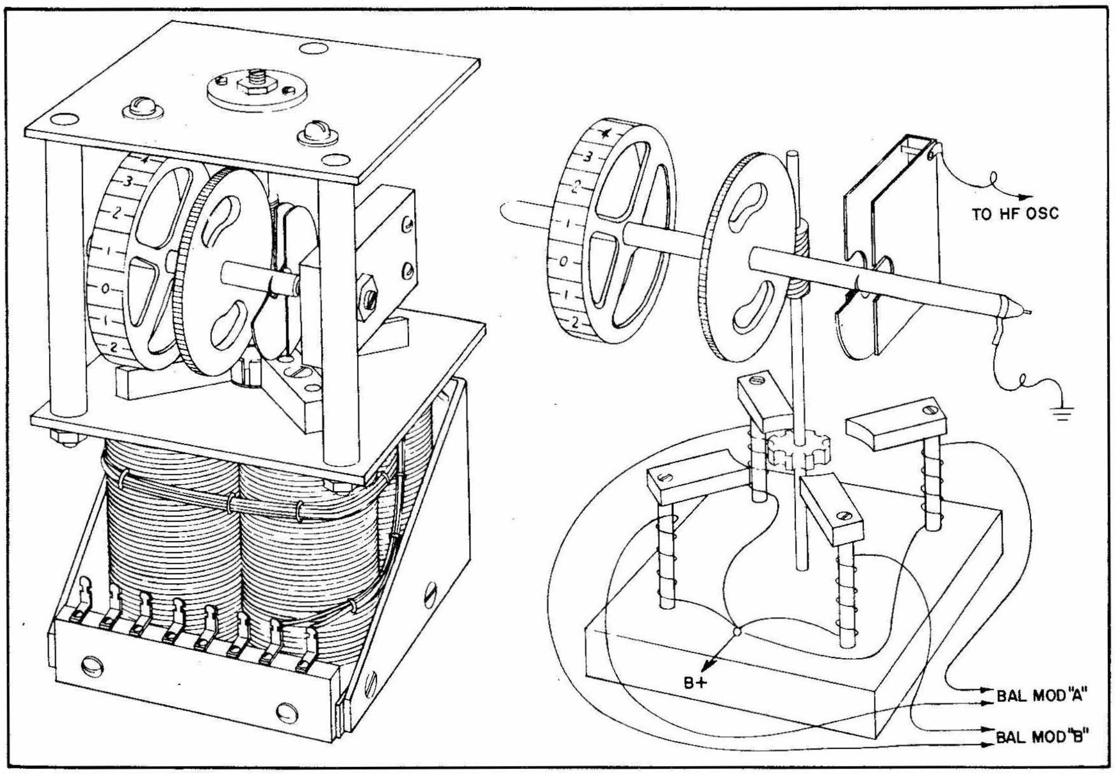

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Figure 95. Typical A-F-C Motor.

circuit. The amplified signal from the received carrier is compared with the quadrature signals from the reference oscillator, and the difference is established. This difference is represented in the outputs of two balanced modulators, which supply these outputs to the split-phase type of motor, M1. Power to drive the motor is obtained from the plate current of the modulator tubes, which may be of the triode or tetrode power-handling type. Plate voltage (B+) is applied at the center tap of the four motor windings. Two of the windings appear in the push-pull plate circuit of balanced modulator A, V25 and V26, and the other two windings, which are electrically 90 degrees apart from the A windings, are connected in the push-pull plate circuit of balanced modulator B, V27 and V28. Figure 94 shows the balanced modulator circuit, without the amplifiers.

The separately amplified 100-kc oscillator signals, which are in phase quadrature (90 degrees apart in phase from each other), are applied to the two push-pull input transformers, T25 and T26. The amplified carrier signal is applied, without any phase shift, to the two carrier input transformers, T27 and T28. Since T27 and T28 are connected to the center taps of input transformers T25 and T26, or in series with the oscillator signals but in parallel at the modulator grids, the carrier signal will be cancelled in the output circuit, because the output circuit is push-pull for both modulators. The 100-kc oscillator signals are not cancelled, because they are applied through T25 and T26 in push-pull at the grid circuits, and the output is push-pull so that the modulators look like a push-pull amplifier to the oscillator quadrature signals. However, the motor will not respond to the 100-kc signal because its frequency is too high; hence, this signal is conveniently bypassed to ground with capacitors C113, C114, C115, and C116 in the plate circuits of V25, V26, V27, and V28, respectively.

In the process, through heterodyne action of the balanced modulators, sum and difference frequencies are produced in both modulators. Since the sum frequencies are also too high for motor operation (about 200 kc), these are also bypassed to ground, leaving only the difference between the reference oscillator frequency, fixed at 100 kc, and the received carrier frequency, which may be 100 kc or slightly higher or lower. If a difference of 100 cycles exists, two 100-cycle voltages, 90 degrees apart in phase, are applied to the motor, M1. The motor operates, then, as a 100-cycle split-phase motor. A small variable capacitor, C18, which is electrically connected in the high-frequency oscillator of the receiver, is mechanically coupled to the motor. A reduction gear train is used so that finer control can be realized. (A detailed drawing of a typical a-f-c motor is shown in figure 95.) If the connections to the modulators and to the motor have been made in proper polarity, the motor will run in a direction so as to increase or decrease the frequency of the l-f oscillator, whichever is necessary. (The motor operates as a synchronous motor at slow speeds, up to about 20 cycles, and as an asynchronous motor at the faster speeds.) This action, in turn, varies the resulting medium-frequency carrier, and hence, the low-frequency carrier. As the oscillator frequency is corrected, the voltage to the motor will decrease in frequency until zero frequency is reached, when the motor stops. Thus the oscillator frequency is tracked to the carrier frequency, or vice versa.

The fixed phase shift of the oscillator signal might seem to indicate only one direction of rotation. This would be true if the oscillator signal were the driving signal. It must be remembered, however, that it is the difference frequency which is applied to the motor, and that the phases of the two outputs of the modulators reverse as the frequency goes through zero, i.e., as the carrier goes through the frequency of the oscillator. Although the reference oscillator used is highly stable, a condition could exist where it may be operating at a higher or lower frequency than 100 kc. It makes no difference whether this is the case, or whether the h-f (or m-f) oscillator or the received carrier has shifted in frequency, as long as the proper relationship is maintained between the re-inserted carrier and the received sideband signal at the demodulator.

If the reference oscillator were low in frequency, for instance, 10 cycles, the afc would correct the h-f oscillator frequency so that the 100-kc carrier and the sideband would also be 10 cycles low in frequency. Limits of a-f-c control may not exceed the frequency (by multiplication) response of the carrier filter in the carrier amplifier. This response is usually on the order of 100 cycles per second, which is equivalent to a range of 10 kc carrier or h-f oscillator shift at 10 mc.

Bias is applied through grid resistors, R115, R116, R117, and R118, to the grids of balanced modulator tubes V25, V26, V27, and V28. Coupling capacitors C105, C106, C107, and C108 block the d-c bias from the ground path through the transformers. Balance of gain of the individual tubes is controlled by separate potentiometers, R119, R120, R121, and R122, across the bias supply. C109, C110, C111, and C112 are bypass or decoupling capacitors for the bias potentiometers.

Plate voltage for the tubes is applied through the motor windings of M1 from the center tap, which is decoupled by resistor R123 and capacitor C117. In the state of equilibrium, plate current provides an equal flux in the motor windings, which acts as an electrical brake, to prevent mechanical drift of the motor.

Manual reset may be provided in the gear train of the motor-capacitor connection, or an electrical reset may be provided, or both. One form of electrical reset is shown, consisting of transformers T29 and T30, the secondaries of which are in the grid circuits of the modulators. A phase-difference network, N3, provides a lead-lag phase difference to the primaries of the two transformers. A switch, S6, determines the polarity of the voltage on the primary of T30, and the switch is spring loaded in the off position. Line voltage may be dropped with resistors R124 and R125. An alarm system is usually connected with contacts

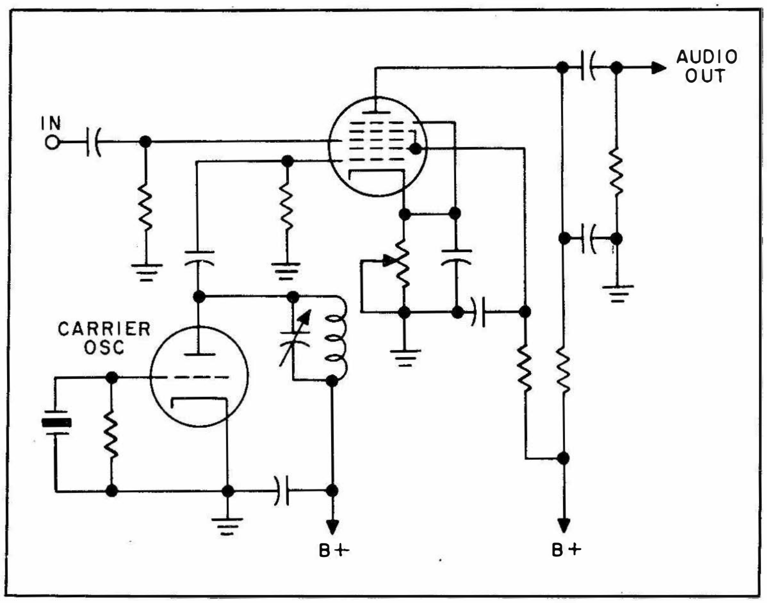

Figure 96. Single-Tube Product Detector (Pentagrid Type).

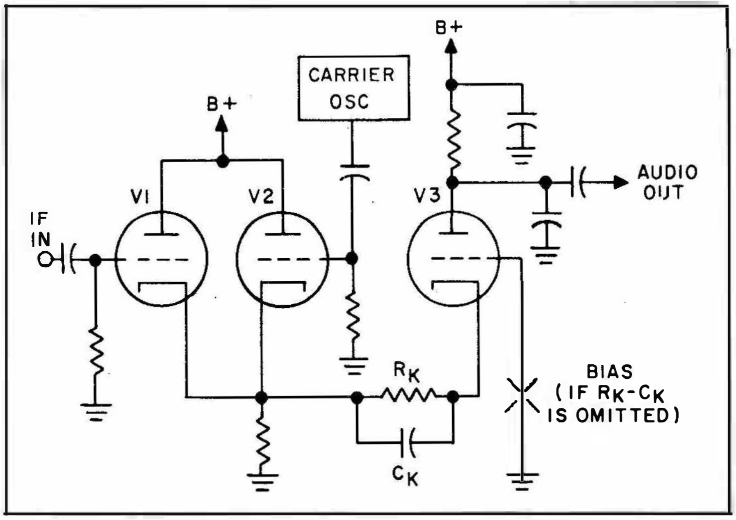

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Figure 97. Triple-Triode Product Detector.

on the capacitor shaft, to warn the operator when the capacitor is nearing either end of its rotation. Operation of this alarm indicates a need for retuning the h-f oscillator and resetting the a-f-c motor, M1.

DIFFERENCES IN RECEIVER CIRCUITRY

Product Detectors

A detector which possesses advantages favorable to single-sideband reception is the multi-grid product detector. The term "product detector" is used because the output (audio) is a mathematical product of the two inputs. Thus, in this type of detector, when either of the two inputs (carrier oscillator or signal) is zero, the output is also zero. There is no detection of the signals applied to the signal grid when the carrier oscillator is off, and vice versa. This detector works equally well for AM, single or double sideband, PM and c-w reception, and is less susceptible to selective fading than other types of detectors. The main advantage is that the output consists mainly of beats with the carrier-oscillator signal, and cross modulating beats between the incoming signals or sidebands are minimized.

One form of this detector is the single-tube, or pentagrid-converter product detector, shown in figure 96. The sideband signal is applied to the injector grid, and the oscillator output is coupled to grid number one. Both grids are biased to the linear portions of their characteristic curves, so that there is no detection when a signal is applied to one grid alone. This detector is of the linear type in which the signal at one grid controls the gain of the signal at the other grid.

The output of a product detector is proportional to the product of the amplitudes of the two inputs, or the output is proportional to the square of the input voltages if both inputs are varied by the same amount. The latter relationship may exist when the received carrier or pilot carrier is filtered and used for re-insertion. Limiting of the carrier, in combination with a good a-v-c system, or use of a local carrier insertion oscillator overcomes the square-law characteristic. In other words, if one of the inputs to a square-law detector is maintained at a constant level, the output has a linear relationship to the other input.

Since the tube is operated in a linear condition, proper bias is established easily by means of a cathode resistor. One disadvantage of the multigrid tube is that proper bias for one grid may not necessarily be proper for the other grid. Aging or replacement of the tube may require readjustment of element voltages. This effect, however, can be minimized by keeping signal and oscillator voltages low.

Triple-Triode Product Detector

The detector shown in figure 97 is a triple-triode product detector. It uses three triodes rather than a single tube. A convenient arrangement is to use two dual triodes, with three of the triodes in the detector and the fourth either as the carrier oscillator or the audio amplifier. The use of separate triodes overcomes the differences of the E_g-I_p curves of the separate grids of a pentagrid tube, since each should be operated on the linear portion of their curves. The one shown in figure 97 uses a common cathode resistor connecting the output triode with the two cathode-follower input triodes. In effect, the arrangement consists of two cathode followers, one of which receives the sideband input, and the other the carrier oscillator input, and a cathode-driven output tube, which provides the detected audio output.

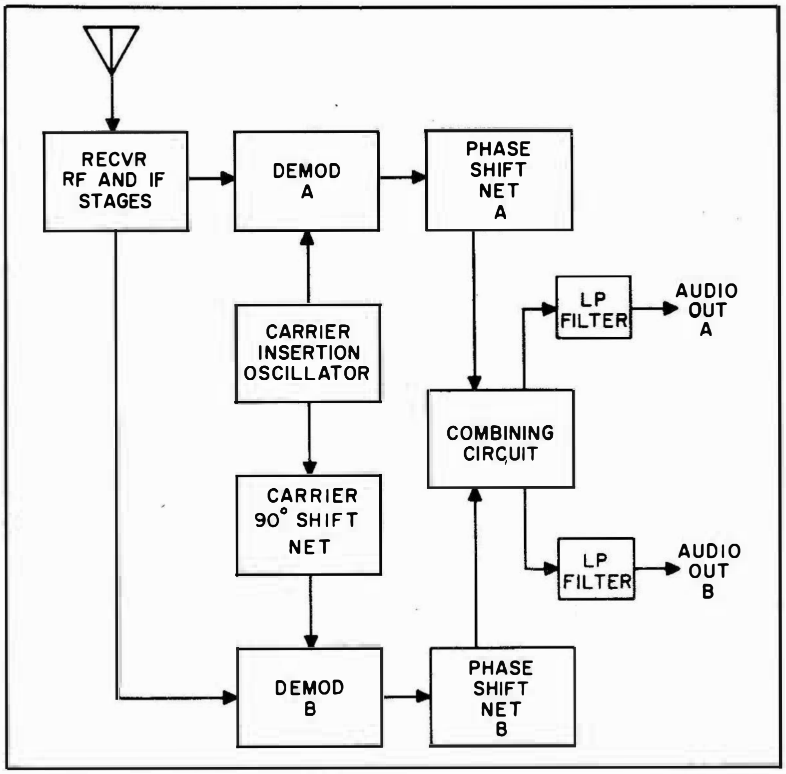

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Figure 98. Phase-Shift Method of Detection.

The self-biasing resistor in the output tube cathode circuit may be omitted and all cathodes tied together if a bias is applied to the output triode grid. The bias is adjusted for minimum output with the carrier oscillator off. One advantage of the single-tube or three-tube product detector is that no transformers are required. Hence, their use can result in a saving in both space and cost. This feature makes the product detector a desirable circuit for single-sideband adapters, used in conjunction with existing AM communications receivers. Another advantage is that only a low carrier insertion voltage is required.

Phase-Shift Method of Detection

A phase-shift method of detection may be used in SSB and DSB receivers to eliminate the need for sideband filters. This method is similar to the phase-shift method of SSB modulation at the transmitter and is well suited for the reception of signals transmitted by this method. The principles involved are the same in both cases. Such a demodulator is shown in block form in figure 98. Since no filters are necessary in the i-f stages, these stages may employ conventional AM circuitry, and the chief requirement in the circuits preceding the detector stages is oscillator stability. Thus the output of a conventional AM receiver i-f section may be applied to an adapter using the circuitry shown.

In the adapter, two balanced demodulators are used in parallel, and the inserted carrier is shifted 90 degrees before being applied to one of the demodulators. The outputs of the two balanced demodulators will therefore have a quadrature relationship to each other. Each output is then applied to its own wide-band-audio phase-shift network. The degree of phase shift in each network (input to output) will vary up to several hundred degrees; however, a 90-degree shift is maintained between the outputs of the two networks over the entire desired audio-frequency bandpass of the receiver (usually 100 to 3000 cycles). Frequencies slightly above and below the design limits of the filter will be passed, but may be somewhat distorted because of a shift in phase greater or less than 90 degrees.

The 90-degree phase-shift network in the carrier oscillator signal input circuit to the demodulators and the two audio phase-shift networks in the output circuits of the balanced demodulators are similar to circuits used for the same purpose in the phase-shift type of modulator. For the design of such networks, or a detailed description, refer to the TRANSMITTER THEORY section of this manual (Phase-Shifting Networks).

The outputs of the two audio phase-shift networks contain in-phase and out-of-phase components of the audio derived from either or both sidebands. These are applied to a resistive combining network. Sideband selection from the combining network is determined by the connection of the combining network to the outputs of the audio phase-shift networks. A simple arrangement is possible when these outputs are cathode followers. When both ends of the combining network are connected to both cathode circuits, the components of one of the sidebands add vectorially in the network, and the components of the other sideband cancel. Which sideband cancels depends on the polarity of the carrier phase shift, which demodulator the shift is applied to, and the polarity of the audio phase shifts in relation to their inputs and in relation to each other. However, selection of the other sideband is easily accomplished by reversing any one of these phase shifts, or by selecting the output of the plate circuits of one of the audio networks, rather than its cathode. In this process a 180-degree phase shift between cathode and plate circuits in the tube is added to the total phase shift appearing at the output. The latter method of sideband selection is preferable, and both are accomplished by switching.

Complete cancellation occurs when one sideband or a component of a sideband (or some unwanted signal) appears in the output of the B channel in the same amplitude as in the output of the A channel, but exactly 180 degrees out of phase with it (or both of the same polarity at opposite ends of the resistive combining network, since the circuit arrangement is push-pull).

This method may be used for dual channel SSB demodulation by variation of the combining network so that outputs may be taken from both the cathode and plate circuits of one of the audio networks and applied to a divided combining (sum and difference) network.

Double-sideband reception may be readily accomplished with this system when the carrier oscillator is operated in a phase-locked (synchronous) condition. In this method of operation, however, both sidebands add and combine, but noise or unwanted signals appear either out-of-phase or only in one sideband, and cancel. Switching in the manner described above, selects the sideband in which noise will be cancelled. When potentiometers are used as a portion of the resistive combining network, and the output is taken from the wiper of the potentiometer, balance can be attained in the amplitude of the signals from both the A and B audio networks so that vectorial addition solely determines the amount of cancellation and maximum output.

Low-pass filters may be used in the audio output circuits to enhance selectivity, but may not be necessary if balance and accurate 90-degree phase shift are maintained in the carrier-to-detector circuit.

Discriminator Type A-F-C Circuits

An a-f-c circuit suitable for exalted-carrier AM and SSB receivers is shown in figure 99. In this circuit, the amplified carrier signal is applied across the center-tapped primary of the discriminator transformer. A carrier frequency crystal, Y1, is used as a crystal filter, connected between the plate end of the transformer primary and the center tap of the transformer secondary. A neutralizing capacitor, CN, is connected in parallel with the crystal but to the opposite end of the transformer primary. With this arrangement, the capacitance of the crystal and holder may be phased out—a method similar to that used in crystal phasing circuits for c-w reception. An output is taken at the secondary crystal connection to provide the filtered carrier for demodulation purposes. This output is also applied across the tuned circuit, L1 and C1, which is tuned to the crystal (carrier) frequency. When the incoming carrier is at the frequency of crystal Y1, no phase shift appears across the crystal because it represents a series resonant circuit. The phase of the carrier voltage at the output of the crystal filter and center tap of the secondary is represented as vector EZ in figure 99A. This voltage is impressed in the same phase across L1 and C1, the crystal input circuit to the discriminator. The voltages induced at opposite ends of the transformer secondary are in quadrature phase relationship to that of L1 and C1, and are 180 degrees out of phase with each other. These are represented as E_s1 and E_s2. At resonance, the resultant voltages, E_r1 and E_r2 are impressed on diodes V1 and V2, and the voltages across diode load resistors R2 and R3 are equal and of opposite polarity; hence, a balanced condition exists in the circuit with zero d-c output. Carrier frequencies are bypassed at the output through C2 and C3.

A shift in frequency will cause the crystal to appear either capacitive or inductive, depending on the direction of carrier-frequency departure, and the resulting voltages E_r1 and E_r2 appear as in vector diagram B (figure 99) because of the shifting of the phase of E_z by the reactance of the crystal. The vector difference between E_r1 and E_r2 appears at the output as a difference in the otherwise equal but opposite d-c voltages across load resistors R2 and R3. A typical voltage-vs-frequency curve for the output of this type discriminator is shown in figure 99C.

The sensitivity of such a circuit depends on the selectivity characteristics of the crystal. The degree of selectivity and the resultant sharpness of control of the discriminator may be controlled by variation of a resistor, R1, either in series or in parallel with L1 and C1. Decreasing the value of R1 increases the selectivity or sensitivity of the discriminator.

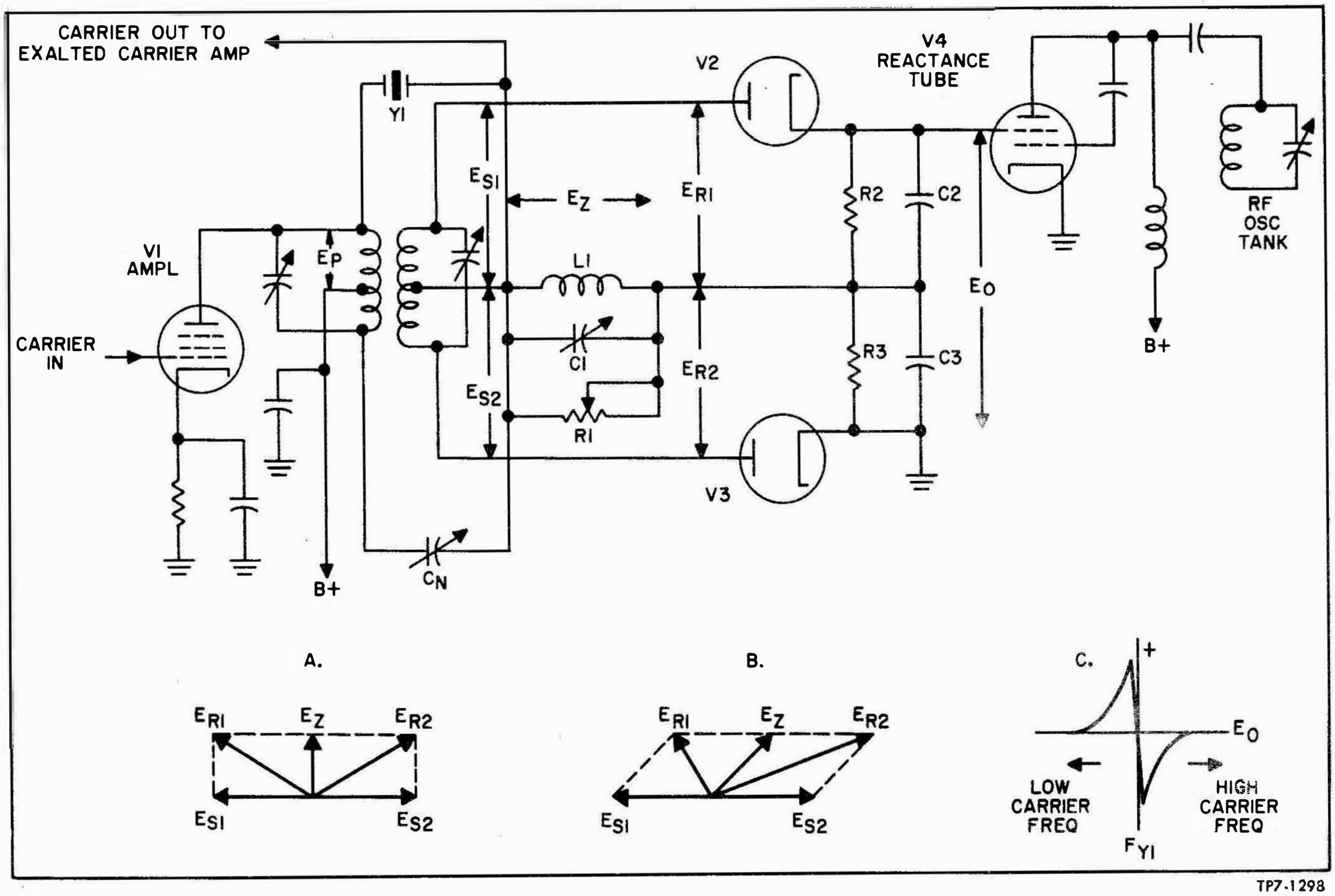

Figure 99. Carrier Filter and A-F-C Phase Discriminator.

The output is shown connected to a representative reactance tube circuit, which controls the tank circuit of the heterodyne oscillator. Variations in the grid (or other element) voltage of the reactance tube vary the effective interelectrode capacitance of the tube, and thus cause the tube to function as a variable capacitor in the oscillator tuned circuit. A d-c motor circuit, together with proper control tubes or a servomechanism, could be used with the discriminator in place of the reactance tube.

A variation of the discriminator circuit is shown in figure 100. Two crystal filters are used in this circuit in a different arrangement. The amplified signal is applied to the first carrier crystal filter, and an output is taken from the filter and may be either amplified or, preferably, limited in the circuit of V3. The output of this stage is then applied to the center tap connection, T4-L1, of the discriminator. An output is also taken at the first carrier filter for application to the a-v-c or a-g-c detector. Better a-v-c action will result if the signal is taken from a point prior to amplitude limiting. V2 and V3 may be either linear amplifiers or limiters, but the latter is preferable, especially in the case of the V2 circuit, the output of which, after filtering in the second carrier crystal filter, is also used for demodulation. Output of the second carrier crystal filter is amplified in the circuit of V4 and applied to the primary of the discriminator transformer, T4.

Different phase responses appear at the two inputs of the discriminator under off-resonance conditions, because of the characteristics of both crystal filters. The difference in phase is converted by discriminator action to a d-c voltage across the load resistors in the cathode circuits of the discriminator diodes, V5 and V6. At resonance the phases of the output signals are equal, and the d-c output of the discriminator is zero. Sensitivity of the controlling circuit may be varied by the use of a potentiometer, R5, across the load resistors. The time constant of the R-C integrator network, R6 and C8, may be made long, to make the afc insensitive to rapid changes in frequency, but responsive to long-term drift.

A d-c motor, with control tubes V7 and V8, is shown as the frequency correction device. This motor circuit may be used with any form of a-f-c discriminator, and is shown as an alternate method. With no signal input, the grids of both tubes are at ground potential, and equal currents

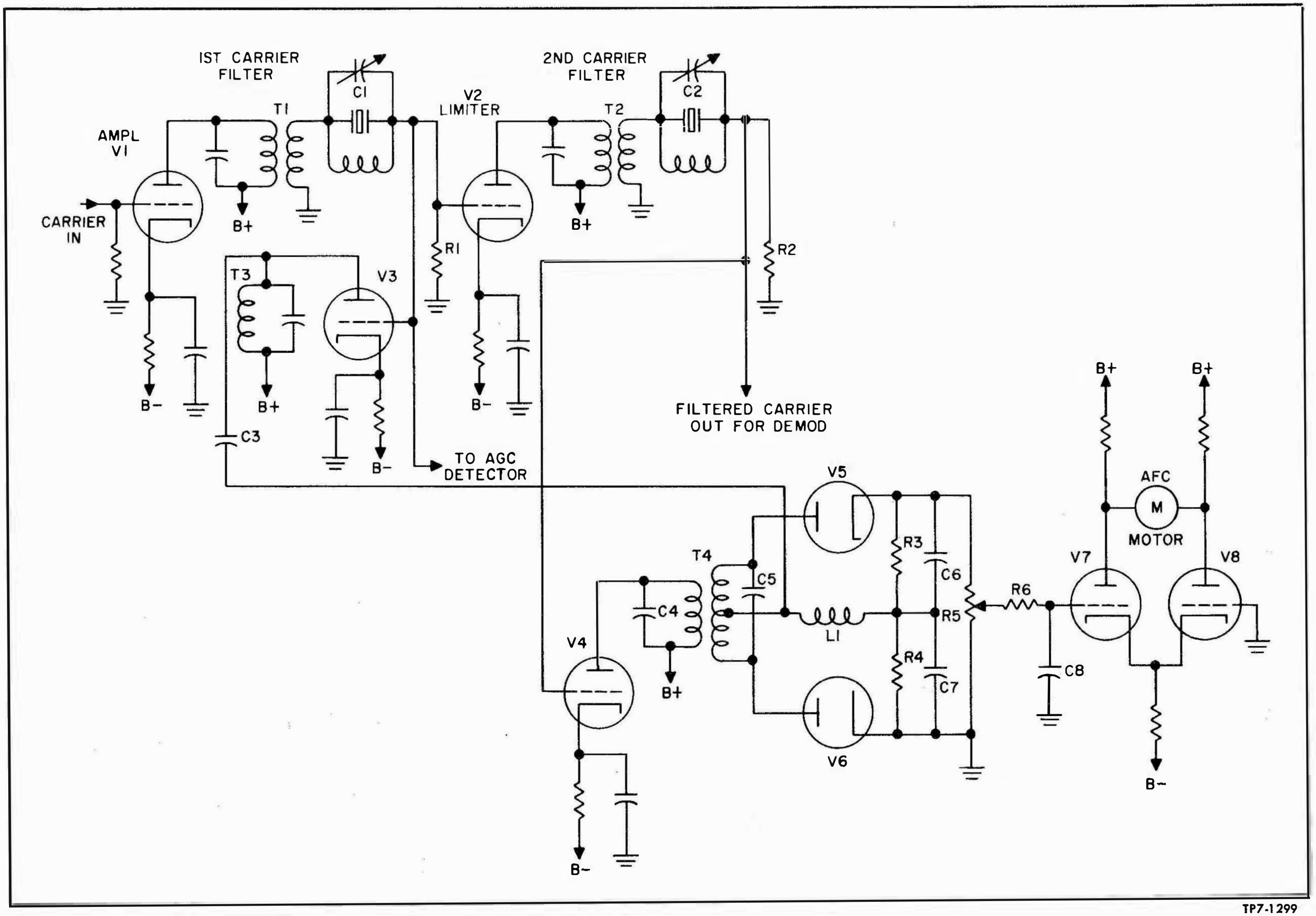

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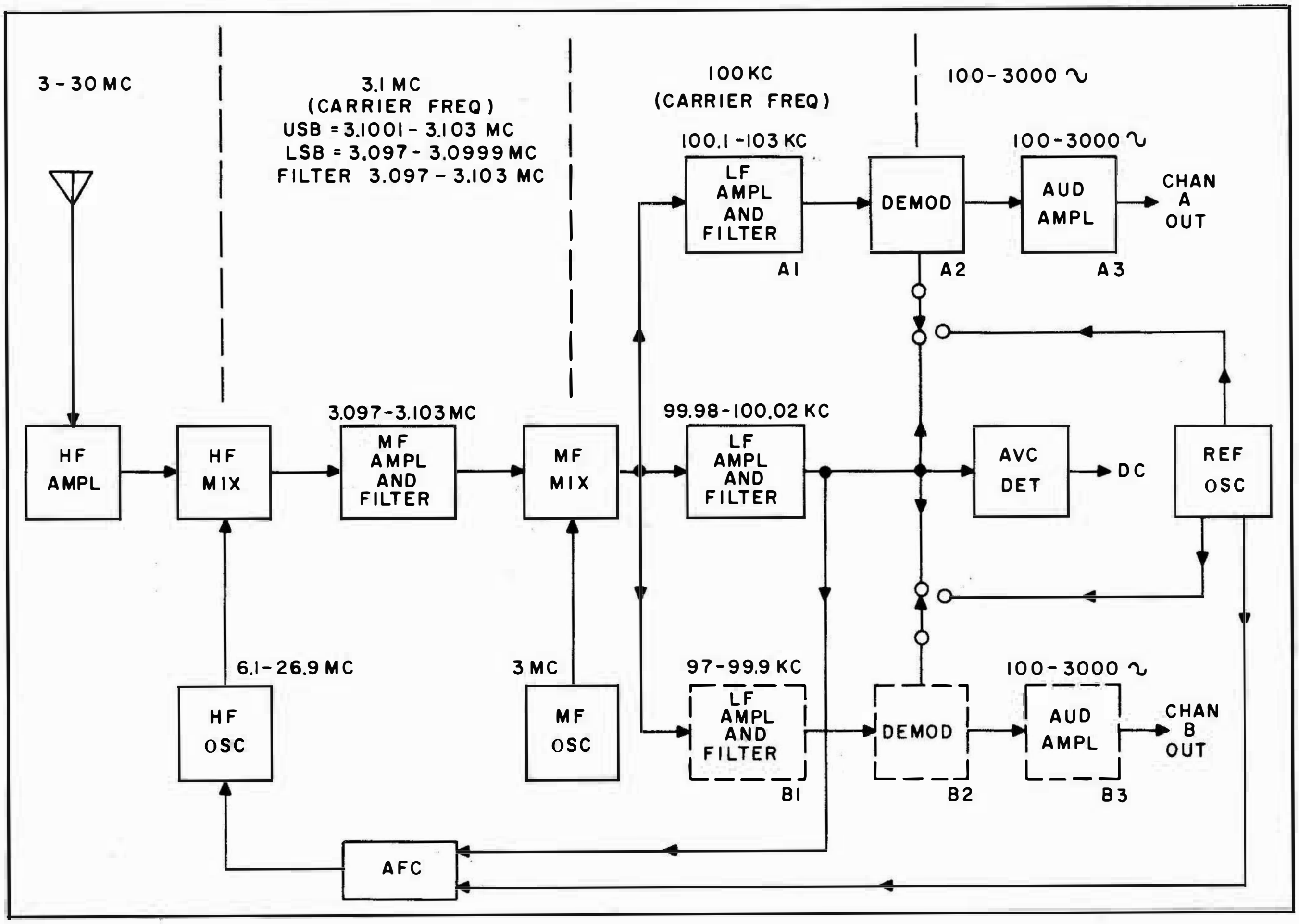
Figure 100. Two-Stage Carrier Filter and A-F-C Circuit.



TP7-1300

Figure 101. Simplified Block Diagram of a Typical Dual-Channel SSB Receiver.

flow in both tubes. These currents produce equal voltage drops across both plate load resistors and equal plate voltages at both tubes. The small d-c motor is connected between the two plates of the tubes. When a positive d-c voltage is applied to one of the grids, more current flows in the plate circuit of that tube, with a resulting decrease in plate voltage, or increased voltage drop across the plate load resistor of that tube. This action also results in an increased voltage drop across the common cathode resistor. The second tube is cathode-coupled to the first tube; therefore, its cathode goes more positive with respect to the grid, and, with the decreased current flow in this tube, the plate voltage increases. The motor is operated by the d-c voltage difference between the plates of the two tubes, the direction of rotation depending on the polarity. The action is reversed when a negative voltage is applied to the control grid connected to the discriminator. The motor may be mechanically connected to drive a small variable capacitor (or inductive device) in the controlled-oscillator tuned circuit.

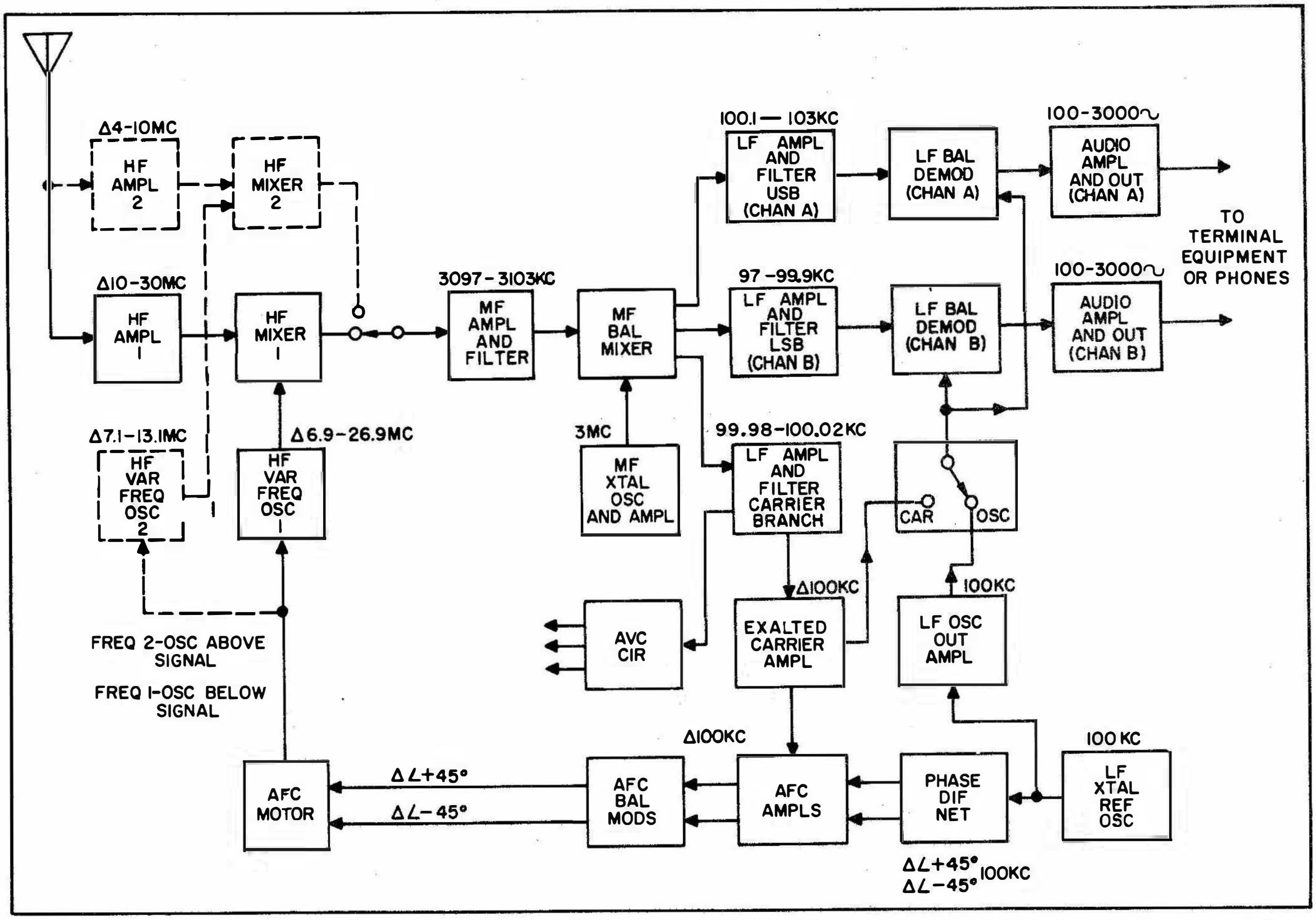
A variable bias applied to the grid of the second tube may be adjusted for balance, to compensate for variations in the tubes and for tube aging.

Single-Sideband Dual-Channel Receivers

Single-sideband receivers have long been used for long-distance, point-to-point, fixed ground communication. A transmitter and receiver may be used at each point to provide single-channel service in both directions. Two transmitters and two receivers could be used to provide two separate channels. Additional channels could be provided in a similar manner. However, such a combination is expensive, and requires the allocation of additional frequency channels, because each transmitter would have to be operated at a different frequency.

In SSB equipment, one sideband is removed. With the addition of a few circuits and components, therefore, a second channel can be added, using the space in the frequency spectrum left by the removed sideband of the first channel.

Such equipment is in common use in such applications as telephone and telegraph carrier, over-



TP7-1301

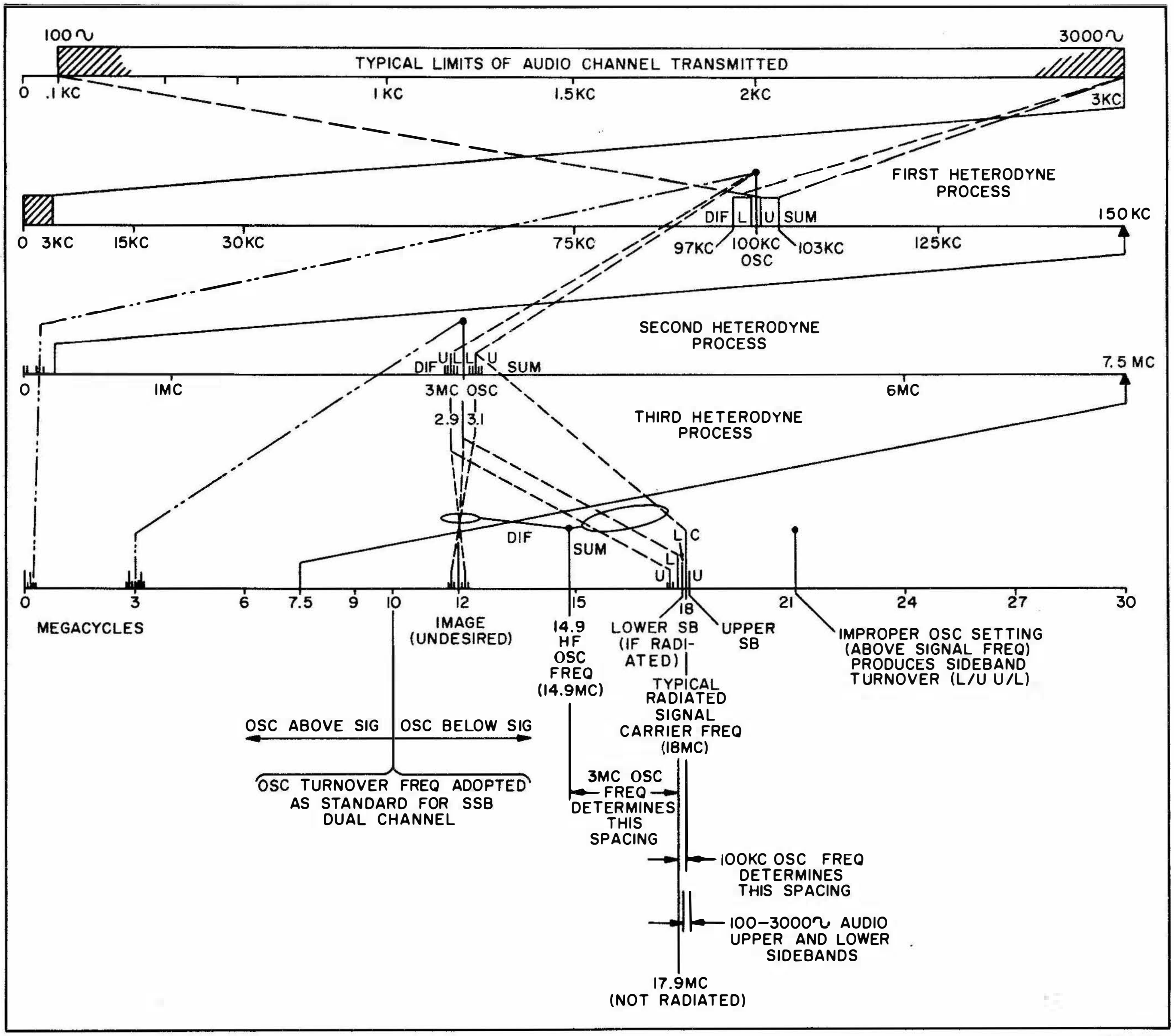
Figure 102. Block Diagram of a Dual-Channel SSB Receiver, Incorporating 10-mc Oscillator Turnover, with Separate Tuners for Both Ranges.

seas communication, and relay links. With the dual-channel system twice as many channels are available with a given number of transmitters and companion receivers. A simplified block diagram of such a receiver is shown in figure 101. The actual circuitry is not shown, since the receiver may be identical to that shown in block form, figure 83, except that the bandpass filter in the medium-frequency amplifier has twice the bandwidth, so as to include the lower sideband, and three circuits have been added (B1, B2, and B3). These circuits constitute the lower sideband channel, or channel B, and are identical to A1, A2, and A3, or the upper sideband channel, channel A, except that the frequency range of the sideband filter in the l-f amplifier, B1, is designed to pass the lower sideband frequencies (97 to 99.9 kc).

Dual-sideband signals are received, amplified, and heterodyned in the same manner as in the SSB receiver described previously, except that up to the output of the medium-frequency mixer, the signal is actually a double-sideband signal (the opposite sidebands containing information from two unrelated input sources) and the bandwidth must be taken into account. Since the filter almost exclusively determines the bandwidth, the increased bandpass of this component usually constitutes the only difference up to this point. The sidebands and the carrier are separated following the output of the balanced medium-frequency mixer; that is, three separate channel amplifiers are connected in parallel to this point. Actually, both sidebands and their common carrier are applied simultaneously to all three amplifiers, and the filter included in each of the three amplifier channels determines which frequencies will be passed and which will be rejected. The signal in each channel is then amplified, detected (or demodulated), and further amplified.

Some care is necessary in the design and maintenance of this type of receiver, to prevent the partial passing or coupling of one sideband into the channel for the other sideband; otherwise, privacy of communication (especially in telephone circuits and broadcast relay) may be impaired. This precaution is as important in the receiver as in the transmitter.

Double-sideband single-channel signals may also be received with this receiver. For such re-



TP7-1302

Figure 103. Analysis of Sideband Conversion with Expanded Low-Frequency Scales.

ception the output may be taken from either sideband channel, or the combined outputs may be used if a combining network is incorporated. Single-sideband, single-channel signals may be received with no changes in this receiver. If the upper sideband is received, the output is taken from channel A, and if the lower sideband is received, it is taken from channel B. In either case, the opposite channel is merely inoperative.

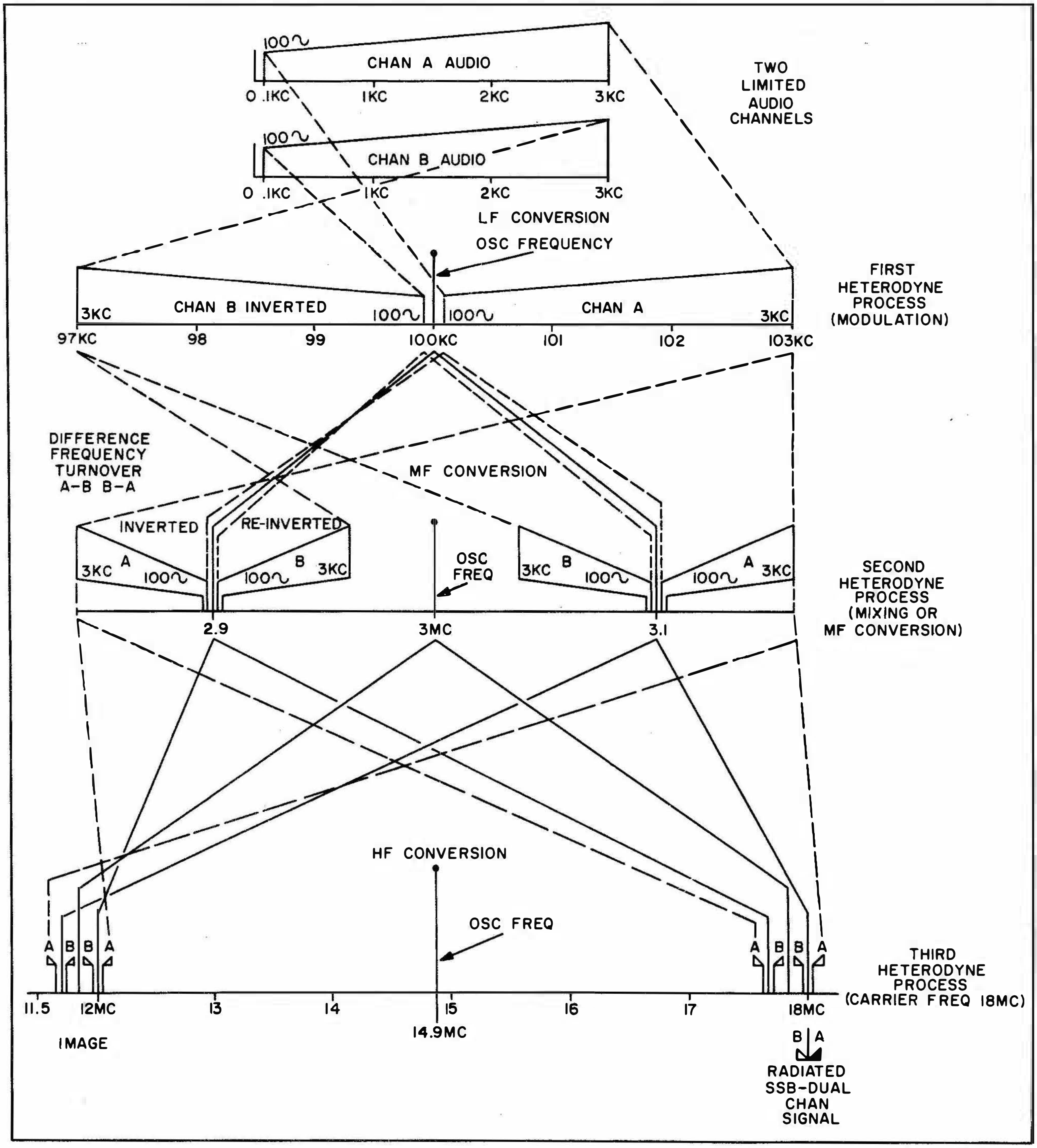
The dual-channel SSB receiver, then, may be considered as two SSB receivers combined in one.

A more complete block diagram of the same type of receiver is shown in figure 102. The functional analysis of this type of receiver is similar to that of the receiver shown in figure 83, except for the addition of the lower sideband channel.

Operation Above and Below 10 Mc

In both transmitting and receiving, frequency translations, or heterodyning of signals involving the use of oscillator frequencies close to the modulating signal frequency, produce spurious responses which are undesirable (the relationships of these frequencies, and their responses have already been discussed).

In the design of SSB equipment using the h-f range of 3 to 30 mc, a received frequency of 10 mc has been adapted as the turnover point of heterodyne oscillators; that is, for transmissions below 10 mc, the oscillator is operated above the signal frequency, and for transmissions above 10 mc, the oscillator is operated below the signal



TP7-1303

Figure 104. Analysis of Sideband Conversion, Showing Upper and Lower Sideband Relationship and Turnover.

frequency. When the oscillator is operated above the signal frequency, a transposition of the sidebands results (upper becomes lower and vice versa). An analysis may be made of this mathematical phenomenon by referring to figures 103 and 104. However, as long as the transposition occurs in both the transmitter and the receiver, the sidebands are repositioned in their respective frequency channels, or the original order. This is true of both SSB and dual-channel SSB (also true of DSB; however, in DSB signals, both sidebands contain the same signal so that transposition makes no difference). The 10-mc turnover point has been chosen to insure that the receiver oscil-

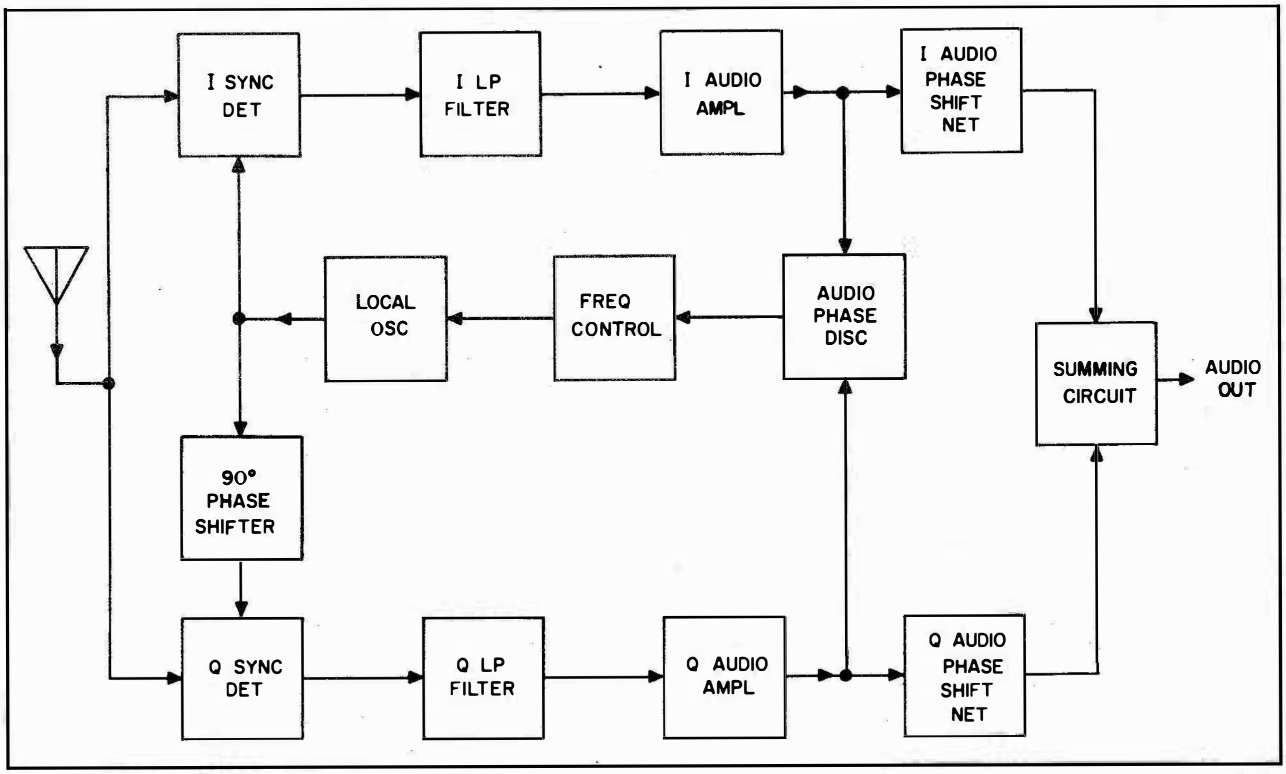


Figure 105. Synchronous Receiver (Two-Phase).

lator operating point corresponds to that of the transmitter, and thus to insure compatibility of equipment. If this were not the case, party A at the transmitter end would be in communication with party B at the receiver. Figure 103 shows the oscillator and signal conditions for covering the range of from 3 to 30 mc. Actually, switching the oscillator frequency by replacing an inductance or capacitor will accomplish the desired result, although this would inject the possibility of tuning errors. Many receivers of this type use two separate oscillators, and often two separate h-f amplifiers and mixers, so that instantaneous changeover is accomplished in tuning with a degree of continuity, by merely switching the outputs of the mixers. Plate voltage may be removed in the unused circuits to prevent interaction, and filaments may be left on to prevent warm-up drift. Such circuitry could have been added to the receiver shown in figure 83, to extend the frequency coverage, but was omitted for simplicity of presentation.

Double-Sideband Reception (Synchronous AM)

Supressed-carrier AM, or double-sideband, reception and detection can be accomplished with SSB receivers if the opposite sideband is rejected sufficiently in the receiver filter circuits. However, full advantage of DSB cannot be realized in this manner. For better results, a local oscillator and a synchronous (or coherent) detector are used in one type of DSB receiver, which utilizes the phase-shift method of detection. Such a receiver is shown in the block diagram of figure 105. Use of the synchronous detector permits direct translation or detection at the receiver frequency and thus provides circuit simplicity and fewer spurious responses because no i-f stages are used. Also, since filtering and amplifying are done at the audio frequencies, easier and closer filtering is possible. This type receiver functions equally well on DSB, SSB, AM, narrow-band FM, CW, and phase-modulated signals.

A similar circuit is shown in simplified form in figure 106. In this receiver a means of post-detection sideband selection has been incorporated. Either the upper or lower sideband resultants are selected at position A of the audio selector switch, and the other sideband resultants at position B, or vice versa depending on whether a leading or lagging phase shift is applied to the B channel oscillator injection signal. The audio phase-shift networks are designed so that there may be any amount of shift from zero to several hundred degrees, depending on the audio frequencies passed (the higher the audio frequency the greater the amount of shift), but 90 degrees of audio shift is maintained between the outputs of

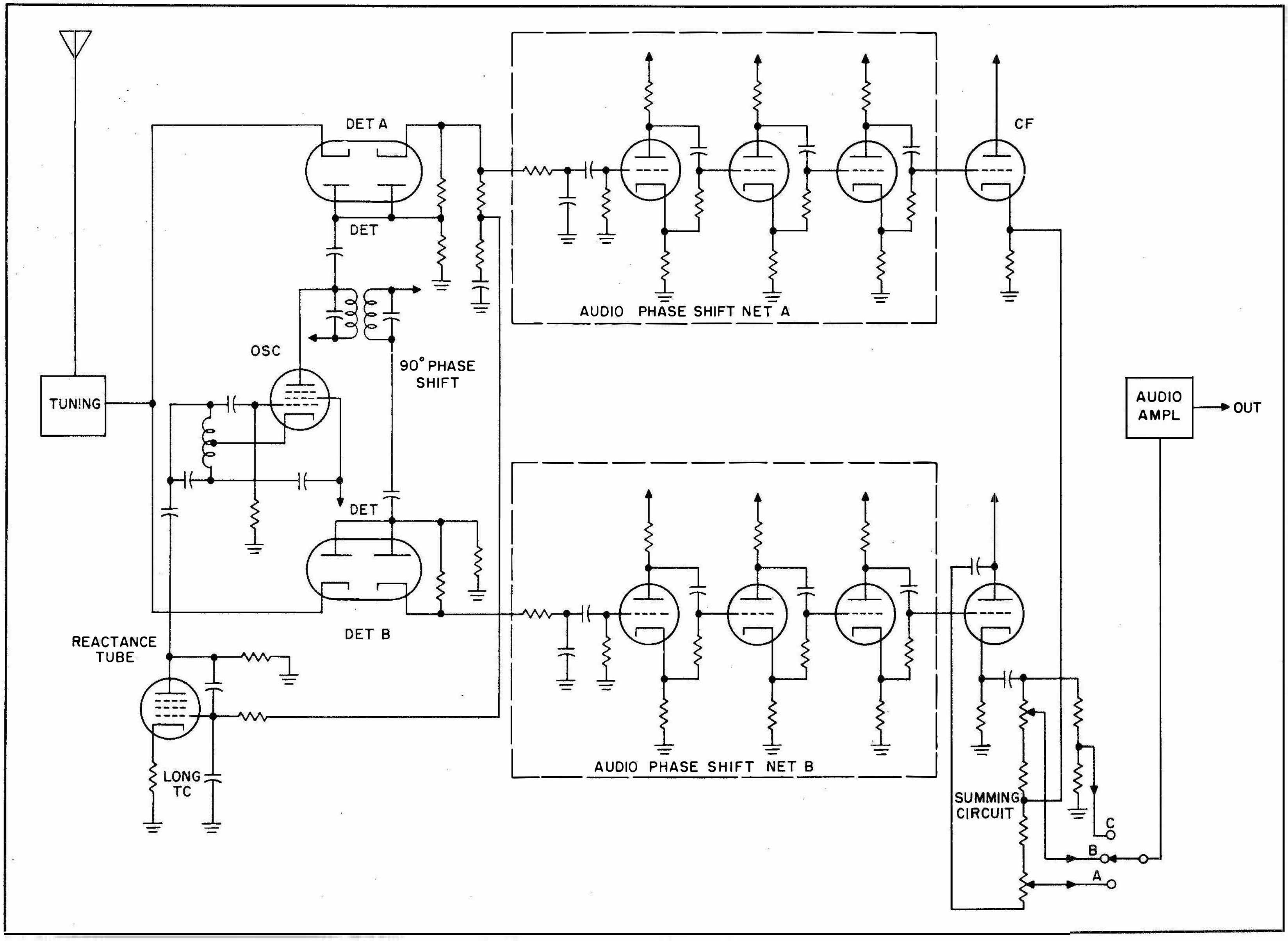


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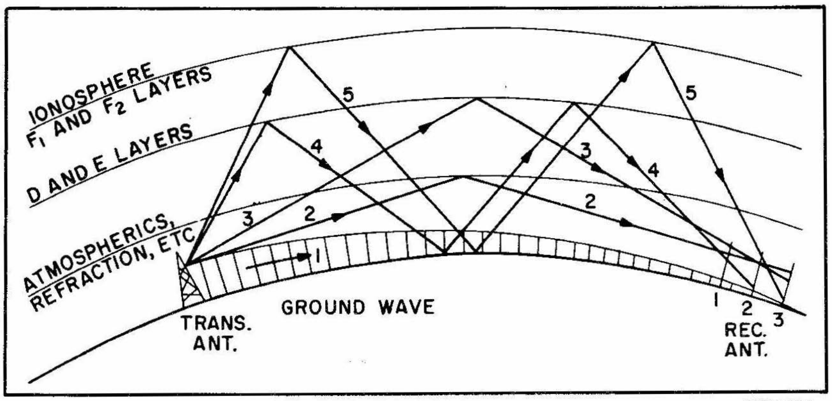
Figure 106. Simplified Diagram of a Two-Phase Synchronous Receiver.



TP7-1306

Figure 107. Effects of Multipath Transmission.

these networks. In the combining circuit an output is taken from the B channel at both the plate and cathode of the last stage. These signals, which are 180 degrees apart, are applied to opposite ends of the resistive combining network. An output from the A network is taken from the cathode of the last stage and applied to the center tap of this resistor network. Output may be taken directly from one of the cathodes (normally at the output of the A network) when the receiver is operating in a phase-locked condition. Position C of the audio selector switch provides for this type of reception.

Long time constants are used in the reactance tube charging circuit, so that the oscillator will remain in approximate synchronization during pilot-carrier loss, or fading.

Although this receiver is similar to the receiver shown in figure 105, it is not the same. The oscillator is referenced to the incoming pilot carrier, rather than to the phase discrimination of the two phase-different channels. Whenever a difference occurs in the phases of the pilot carrier and the oscillator signal, a d-c voltage is developed in the output of detector A. This voltage is applied as a correction voltage to the reactance tube, which corrects the oscillator frequency and phase.

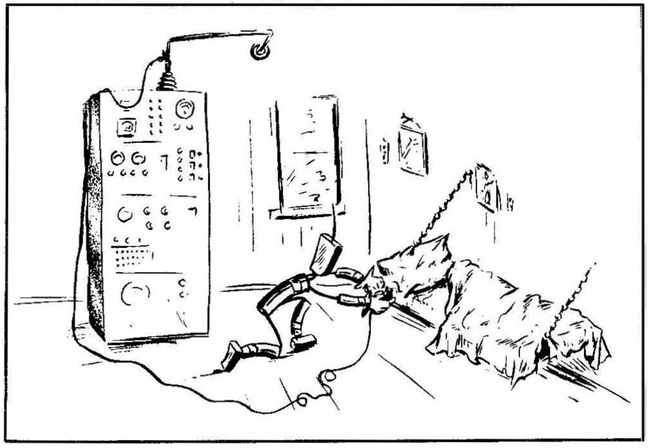
This receiver may be used to receive either single- or double-sideband transmissions, but for the circuit shown such signals must contain some form of carrier. If the circuit were modified to include an a-f-c circuit operated by the audio or sidebands, suppressed-carrier reception would be possible.

Diversity Receiving Systems

In long-range communication, fading becomes a serious problem, because of the effects of multipath transmission. So-called ground waves are predominant to a maximum distance of a few hundred miles. Sky waves are bounced off various atmospheric and ionospheric layers, especially E and F layers, as illustrated in figure 107. It will be noted that many possible propagation paths can be constructed, depending on the distances involved and the height of various layers, clouds, etc, above the earth's surface. It can be seen that radio waves travelling to a given point over one path and those travelling to the same point over another path may arrive at different times. Variations in this time difference result in a constantly changing phase difference of the signals arriving at a specific point (the receiving antenna), and thus cause alternate in-phase and out-of-phase signal conditions, or fading, at the receiver.

The distance between points where a sky wave strikes the earth, or between the transmitter antenna and one of these points, is called the skip distance. Receiving stations located between these points may be unaffected by the sky wave, and receive all, or nearly all of its signal from the direct ground wave, if the receiving station is within the relatively short range of the ground wave.

Propagation conditions affect different frequencies differently, and play a prominent role in the medium-frequency and medium-high-frequency ranges. Radio waves are reflected, or refracted, at different angles at different frequencies. Conditions may change continuously because of weather, sunspot activity, etc. Several methods are used to compensate for fading caused by these conditions. Among these methods are the use of single-sideband reception and synchronous double-sideband reception to compensate for differences in phasing of the received sidebands, and the use of exalted-carrier reception, with SSB or DSB to compensate for carrier fading. Another important method to be considered is the use of diversity reception, which is the standard method used in AM systems for long-range, point-to-point communication.



". . . space diversity receiving systems——"

Space diversity receiving systems use two or more (usually three) antennas, spaced some distance (at least over one wavelength and usually several wavelengths) apart, with a separate receiver for each antenna (see figure 108). When the signal arriving at one antenna is in a fading condition, the same signal received on another antenna some distance away may be of maximum amplitude. When more antennas are added, these conditions become more probable. The outputs of the receivers are fed into a combining network, and some combination (additive or strongest) of the signals is used in the output. Either completely separate receivers or a combination of separate receiver channels, with common oscillators, a-v-c, and control circuits may be used.

Advantages of several methods are combined to produce even greater flexibility when SSB receivers are used in space diversity systems. Since dual-channel SSB receivers provide for simultaneous reception of both sidebands separately, this feature is ideally suited for diversity systems. Also, since SSB receivers may use exalted carrier in lieu of added carrier, DSB reception and exalted-carrier AM may also be received with the same equipment, if suitable switching arrangements are incorporated. With the addition of a carrier phasing network in the receiver carrier circuit, phase-modulated signals may also be received and detected.

Various combining networks may be used for different types of transmission, improved flexibility, and cost factors. Three basic types are in use: One is basically a common load resistor across which all the receiver outputs are applied. This is the additive type. Another uses some form of gating circuit, automatically selecting the output of the receiver or receivers with the strongest signal. The third type is used for telegraph and frequency-shift keying (fsk) and may take a variety of forms. Combiners for data transmission may be included in this third category.

A typical combination receiver for dual-diversity reception of SSB, DSB, and exalted-carrier AM is shown in block form in figure 109. No discussion of individual circuits will be presented, since they may be assumed to be conventional, and have been covered in detail in the receiver analysis on preceding pages of this manual. The block diagram in figure 109 shows the relationship of the various circuits. The circuitry includes practically two dual-channel SSB receivers, with common oscillator and control circuits. The common high-frequency oscillator output is applied to both high-frequency (r-f) mixers, and the common medium-frequency oscillator output serves both m-f mixers.

Individual upper and lower sideband components are applied to separate filters, amplifiers, and demodulators in each receiver channel, and the demodulated outputs are applied to separate audio amplifiers. The outputs of both A (upper sideband) channels of the two receivers are added or otherwise combined in a separate unit, or combiner. The outputs of both B (lower sideband) channels are also combined in a separate circuit, so that dual-channel SSB (different information on opposite sidebands) operation may be received, or choice of sidebands may be provided. A common low-frequency reference oscillator (usually 100 kc) is included, and its output is applied to the common a-f-c circuit, which is of the comparator type. Output from this oscillator may also be selected, through switching, for demodulation purposes. The demodulators may be of the balanced type, or product detectors may be used.

Carriers from both receiver sections are applied to a combining network, and the output of the network is filtered and amplified, and applied to the a-f-c circuit. Reconditioned, or exalted, carrier may be selected, by switching, for demodulation of the sidebands, in lieu of the carrier oscillator. AFC may be applied to the medium-frequency

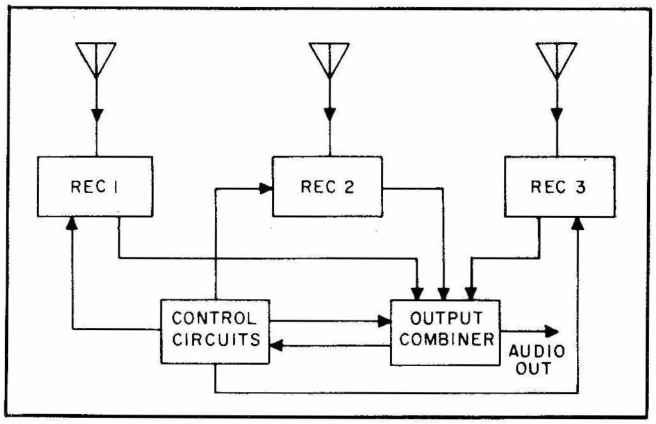


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Figure 108. Diversity Receiving System (Simplified).

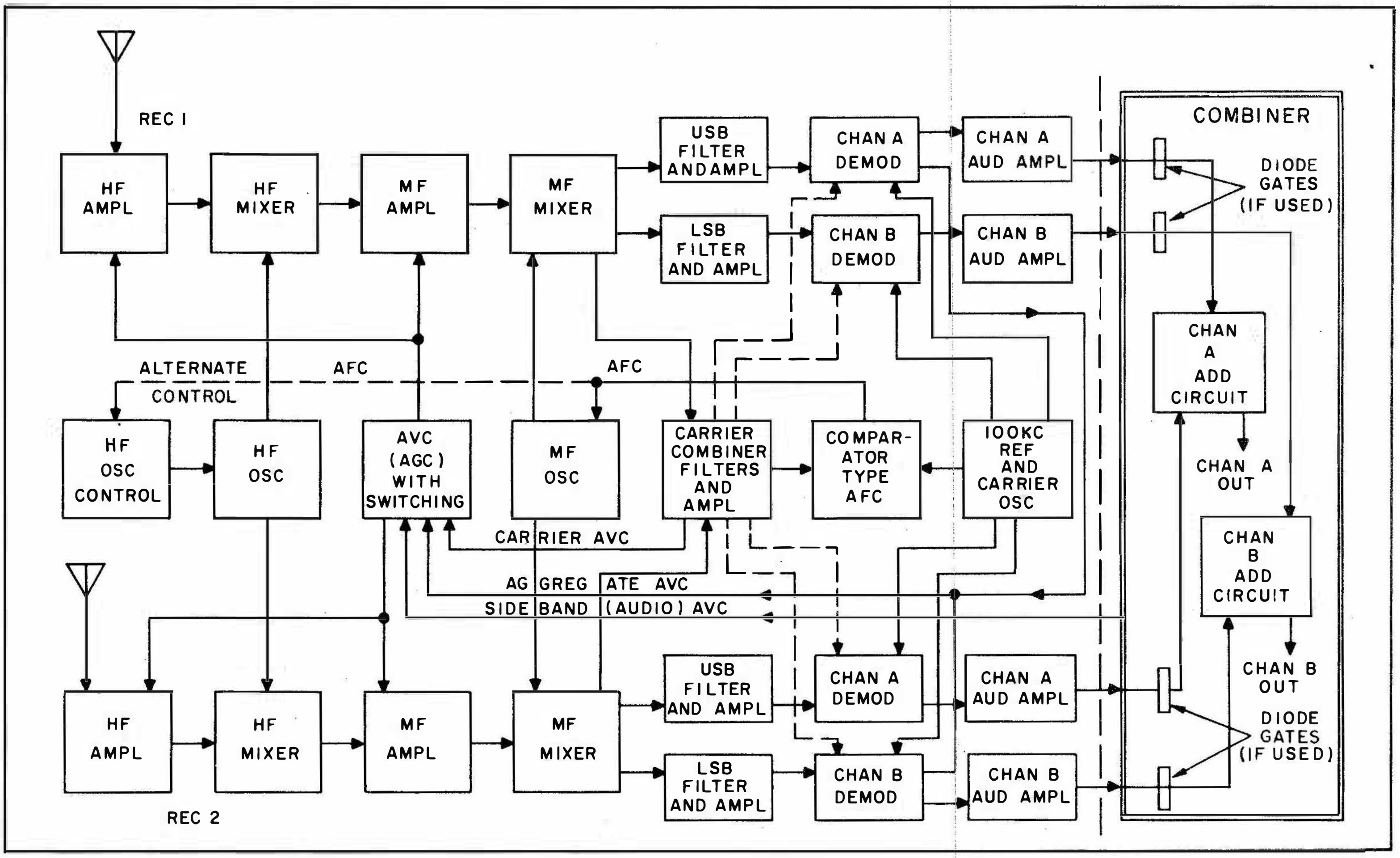

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Figure 109. Exalted Carrier and SSB Dual-Diversity Receiver, Showing Common Oscillators, Carrier, A-V-C Circuits, and A-F-C Circuits.

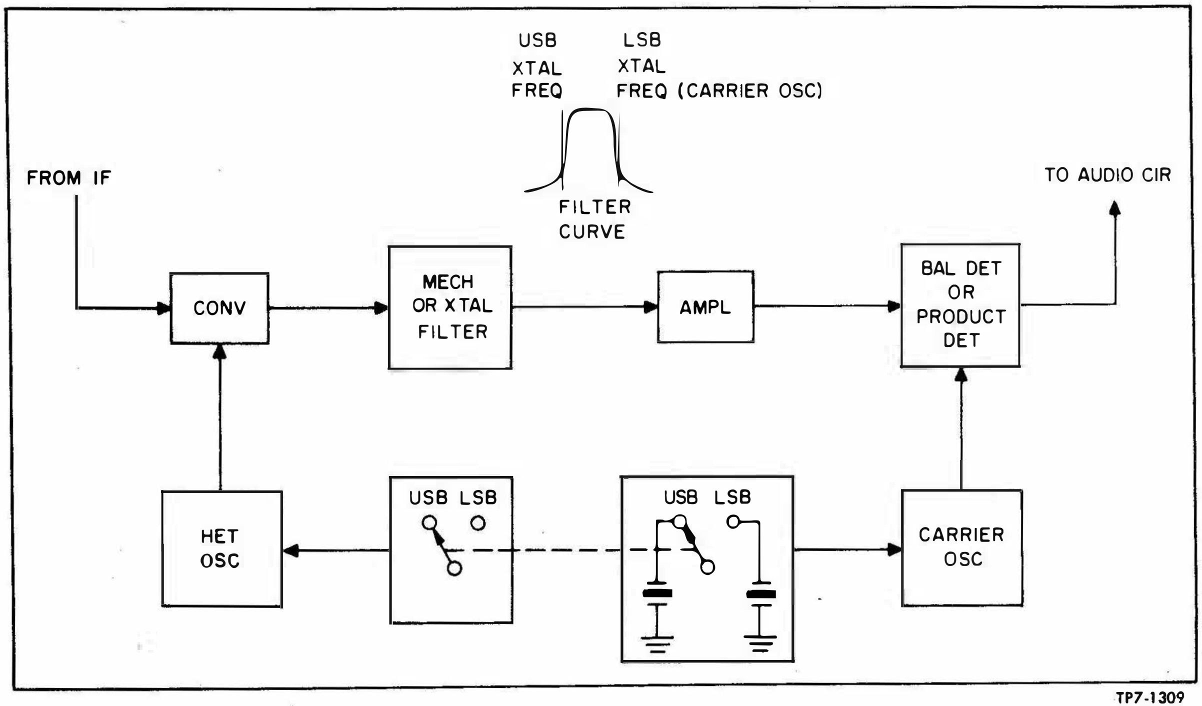


Figure 110. **Simple SSB Adapter.**

oscillator in some systems or to the high-frequency oscillator in other systems. In either case, the oscillator without afc may be crystal-controlled or slaved.

The a-v-c may operate from the combined carrier, or switching may be more frequently found, permitting operation from the combined outputs of all sideband channels (aggregate avc) or the rectified audio from either sideband output of the combiner. Aggregate avc is more often used for telegraph reception, and either aggregate or rectified-audio avc is useful in the absence or loss of carrier or under noisy carrier conditions. Often a noise-operated squelch system is included in the carrier conditioner to disable the carrier-operated circuits (avc and afc) in the event of excessive noise, interference, or jamming of the carrier.

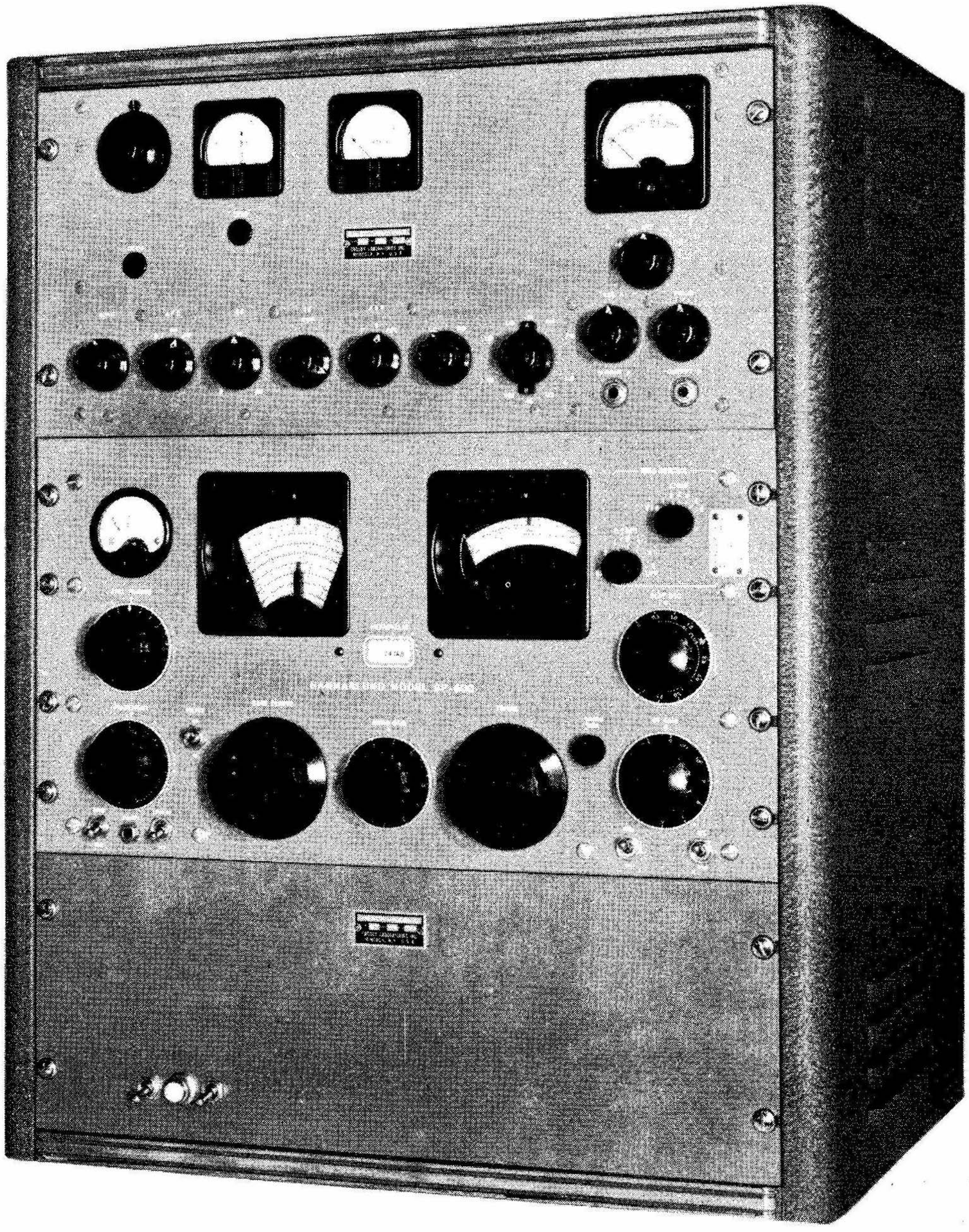
This type of receiving equipment is usually rack-mounted, and much flexibility can be incorporated in the design of such equipment. Unfortunately the wide spacing of the antennas limits the use of diversity systems, especially at the lower frequencies, to large installations. The same fading conditions that render space-diversity systems advantageous at these low and medium-high frequencies are not as predominant at the very-high and ultra-high frequencies, where a different set of conditions and problems exists. Spacing of antennas more than one wavelength apart does not present as much of a problem in the VHF and UHF range of frequencies.

Adapting AM Receivers to Single Sideband

Many high quality AM communication receivers are presently in use. Most of these receivers will not be replaced immediately in a general transition to SSB or DSB suppressed-carrier transmissions. Fortunately, not too great a change is required to render most of these receivers usable for SSB reception. As mentioned previously, one method of receiving SSB signals with an AM receiver makes use of the bfo of the receiver for carrier insertion. This method is possible, but its use requires too critical adjustment of the bfo for rapid communications, especially when non-technical personnel are involved.

An easy, better method of utilizing existing equipment is to use an adapter, or outrigger, similar to the "signal slicer" or Q multiplier so familiar to amateurs. One such adapter is shown in block form in figure 110. Many adapters of this type are being supplied as companion equipments with current models of AM receivers in a rack mounting for SSB use. See figure 111.

An output is taken from the last i-f stage of the receiver used and applied to a converter in the adapter unit. A heterodyne oscillator signal, from a local oscillator in the adapter, is also applied to the converter stage. The input to the converter is tuned to the i-f frequency of the receiver (usually on the order of 455 kc), and the signal is heterodyned to a lower i-f frequency, usually 85 kc. The



TP7-1310

Figure 111. A Commercial AM Receiver and Single-Sideband Adapter in a Rack-Mounted Combination.

use of this low frequency provides the advantages of one more conversion process and results in better selectivity; it permits tuning of the oscillator and input to accommodate a range of receiver intermediate frequencies (within limits); and it provides a local oscillator that may be switched in frequency, so that either the upper or lower sideband will be passed, or fall into the 85-kc filter range, without making any change in the receiver local oscillator. Crystal filters are popular in the 85-kc range, although mechanical and L-C filters may also be used. The filter passes only one sideband, and although most services have adopted use of the upper sideband, many of these adapters incorporate a simple switching arrangement to permit the operator to switch from the upper to the lower sideband. A double-pole, double-throw switch changes the frequency of the heterodyne oscillator, and simultaneously switches crystals in the carrier-oscillator circuit. Two crystals are incorporated in the carrier-oscillator circuit. One of these is of a frequency just above that of the filter range, and the other, just below.

After filtering, the sideband signal is amplified in a low-intermediate-frequency amplifier stage (of the frequency of the filter), and the output is applied to a balanced demodulator or, in some adapters, to a product detector. The carrier-oscillator signal is used for demodulation, in the same manner as in an SSB receiver. The audio output may then be applied to the audio circuits of the communication receiver.

The choice of intermediate frequency used in the adapter is limited principally by the availability of components, mainly the filter. 85-kc i-f transformers are commonly used in low-frequency navigation and range receivers. 250-kc mechanical filters are presently quite popular. 100-kc crystals have long been used as standards and are readily available. 25 kc and 17 kc are also used for filter and oscillator frequencies, but special consideration must be given in the use of such low frequencies. Excellent selectivity can be obtained, but sharp-cutoff, low-pass filters are required because these frequencies approach the audio range. An economy can be realized in the use of the lower frequency filters, since simple L-C circuits, employing toroids, provide ideal filtering in this range.

Adapting Diversity Receiving Systems to SSB

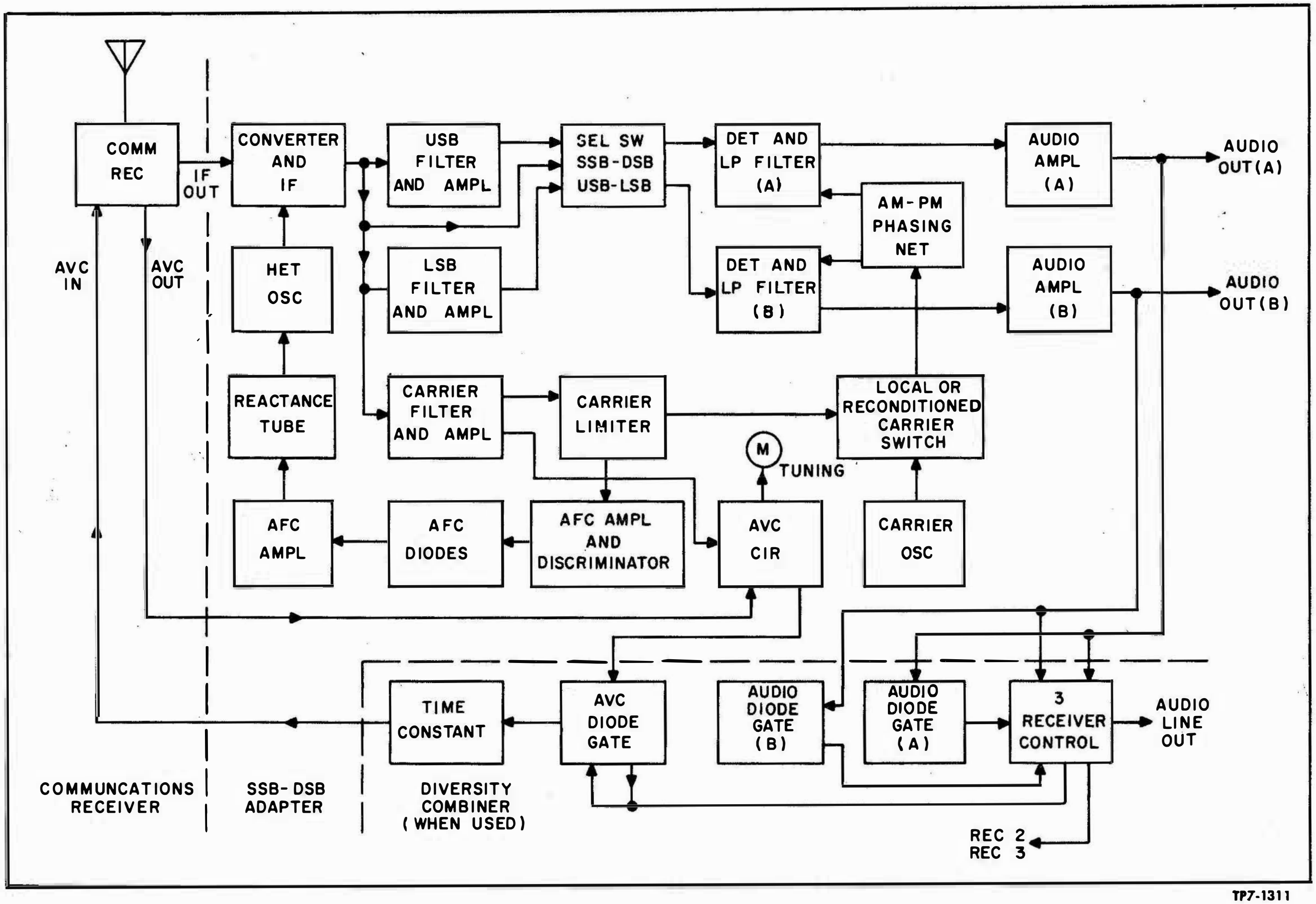
A more complex type of SSB adapter, suitable for use in dual- and triple-space-diversity receiving systems, is shown in block form in figure 112. Both upper and lower sideband filters and amplifiers are included, and two channel detectors and audio circuits are also included, with proper switching to permit both upper and lower sideband reception, separately or combined. In addition, carrier amplifier and filter circuits are included, with a-v-c and a-f-c circuits for more precise and automatic stability features. A crystal-discriminator and reactance-tube a-f-c circuit is used to control the heterodyne oscillator. Exalted carrier for detection purposes is provided from either the amplified and reconditioned carrier or the local carrier insertion oscillator. A phasing network is used to correct the carrier phase for suppressed-carrier DSB and AM or PM reception.

A-V-C action in the adapter is combined with that in the receiver, and an output is taken from the receiver avc. The combined a-v-c voltage from the receiver and the adapter may be fed through a diode gate in a diversity control system (when diode gate circuits are used), and a combined a-v-c voltage may be applied through a selected time constant to the a-v-c system in the receiver.

Audio output may be taken directly from the audio amplifier in either the A or B channel, or both, separately or combined, or, in a dual- or triple-diversity receiver system, it may be applied through diode gate circuits to a combining network. Diode gates are used to provide diversity selection in the type of diversity system shown in figure 113. The gates may be controlled by either the rectified filtered carrier or the rectified total signal. Use of the rectified total signal is an aid in tuning. This method provides rejection of a signal from a receiver or channel when the signal has faded to such a low level that only noise is present in the output circuits. A separate set of gates is incorporated in the combiner for each of three receivers. Selector switches may provide additional combinations. The audio output from each channel of each receiver is fed through its individual diode gate to a common load resistor. D-C voltage from the a-v-c circuit of each receiver is fed through isolating resistors to the respective audio diode gate of each channel, and directly to an a-v-c diode gate. The diode gates act as controlled resistances, the resistance being low when the a-v-c voltage is high and vice versa. As a result, the audio voltage applied to the common load resistor through the diodes is mainly the output from the receiver with the strongest signal, or the highest a-v-c voltage. D-C voltage from the strongest signal can then be fed back to apply a common a-v-c voltage to all three receivers. In other systems audio may be applied directly to the common load resistor from all receivers without the use of diode gates. This method provides direct addition of the audio outputs of the receivers, but has the disadvantage of also adding the noise level of all receivers to the common output along with the signal. Some systems provide for both types of output. Diversity receiving systems may be readily converted to SSB and exalted-carrier systems by the addition of two (in dual diversity) or three (in triple diversity) SSB adapters and a dual-channel combiner. Such a completed system is shown in figure 114.

Adapting AM Receivers to DSB

Double-sideband suppressed-carrier signals cannot be successfully received on conventional AM receivers unless a locally generated carrier is reinserted prior to detection, in a process similar to that used for SSSC reception. Even then a slight shift in frequency results in frequency distortion, just as in SSB, with the partially inverted opposite sideband (when the carrier falls into the frequency range of that sideband) causing further distortion. In addition, a shift in phase of the local carrier with respect to its proper phase relationship to the sidebands causes cancellation of the signal, because of the vectorial addition of the difference between each sideband and the carrier. A carrier phase shift of 90 degrees causes complete cancellation of the resultant audio. For this reason, a phase-locked oscillator is usually employed. One method of phase-locking the oscil-



TP7-1311

Figure 112. SSB Adapter for Diversity Systems, Showing Use of Diode-Gate Diversity Combiner.

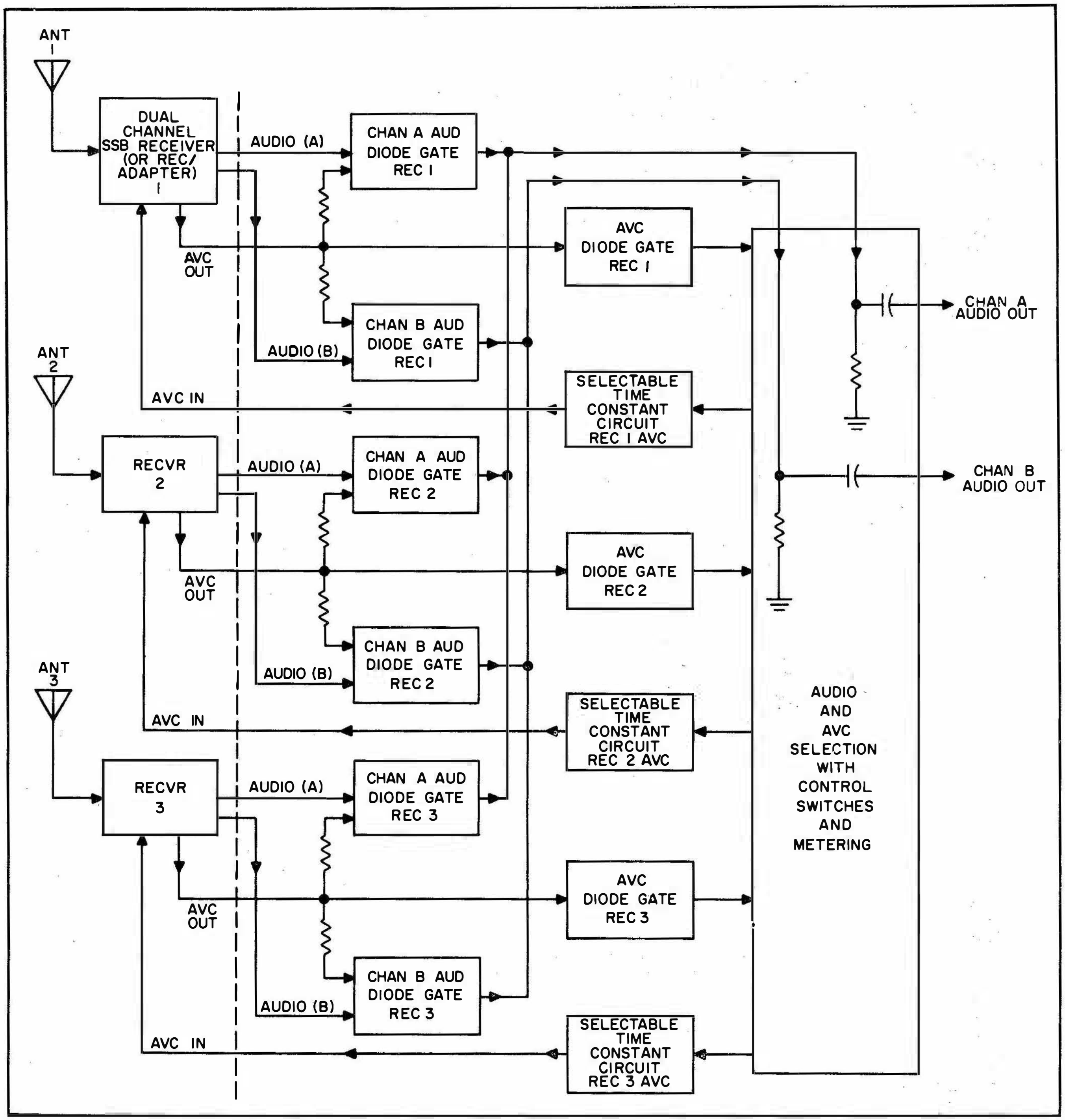


Figure 113. Triple Diversity SSB Receiving System, Showing Combining Network.

lator is to use a two-channel or two-phase (synchronous) detector and audio circuit employing the phase-shift method of detection and a phase discriminator and reactance tube for oscillator control.

Synchronous Detection

A typical DSB adapter employing synchronous detection is shown in figure 115. The block diagram of this circuit is similar to one shown in another section of this manual; however, a variation in circuitry is shown here. Actually, there are many variations in circuitry possible. This adapter uses a product detector in both the I and Q demodulators. The i-f signal from a conventional receiver is applied to one grid of each de-

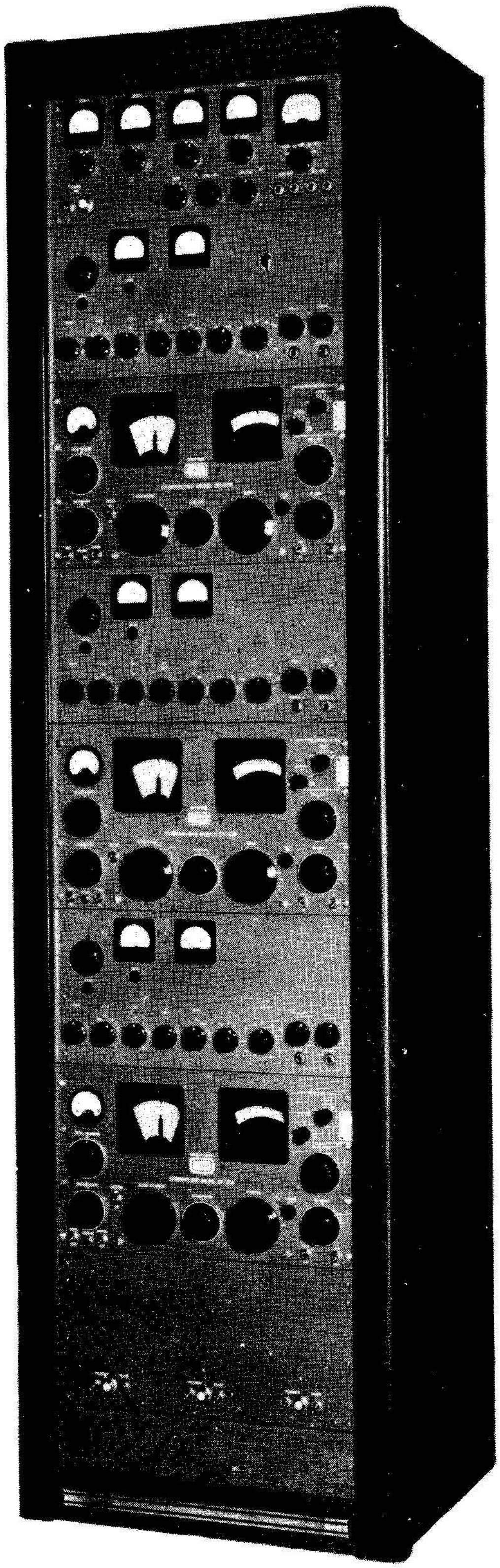


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Figure 114. A Triple-Diversity Receiving System, Consisting of Conventional AM Communications Receivers with SSB Adapters, Diversity Combiner, and Power Supply, in a Rack-Mounted Package.

tector, and the oscillator signal is applied to the other grid through an R-C phase difference network, which establishes the quadrature relationship at the detectors.

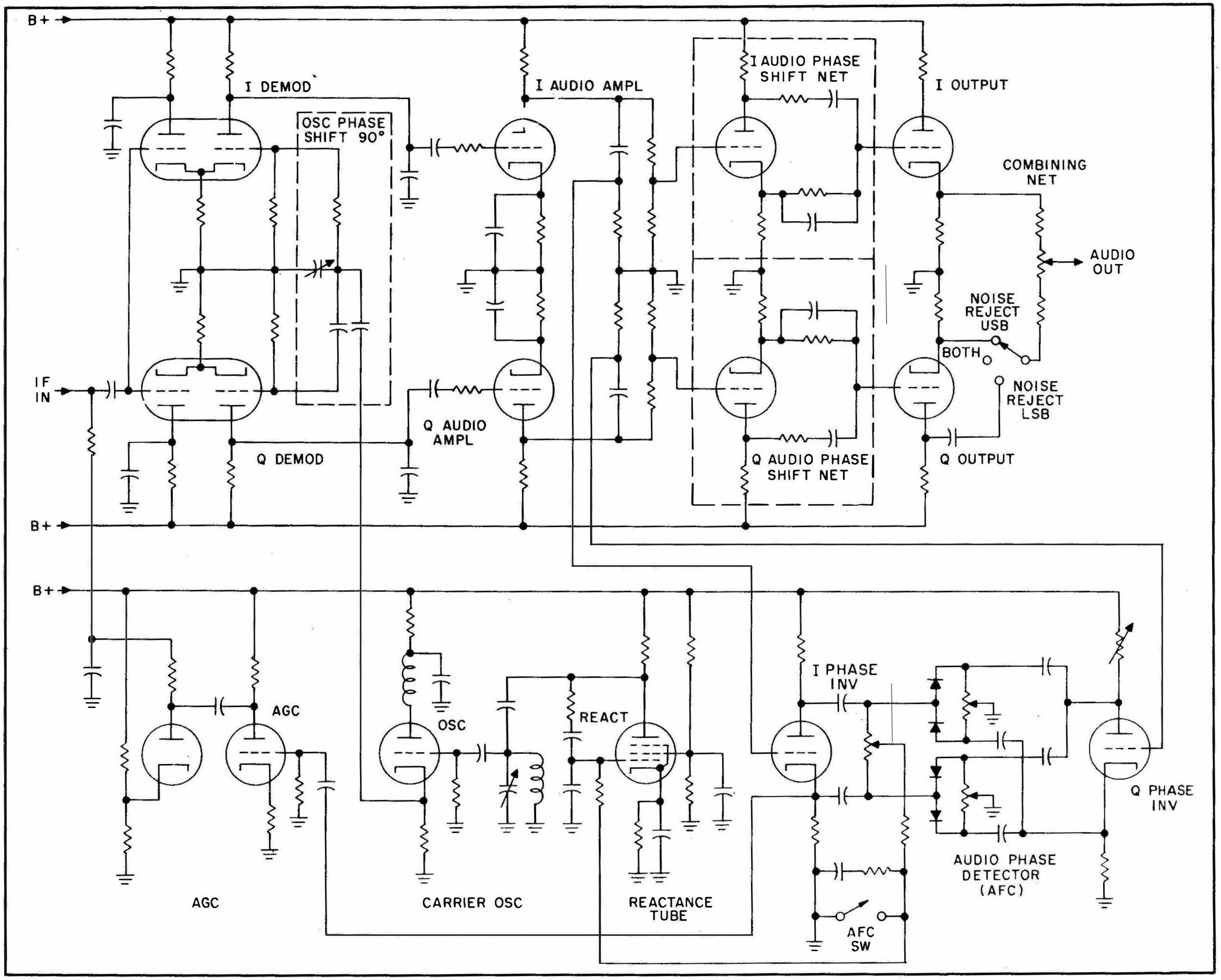
Audio output is amplified in both channels and applied to audio phase shift networks designed to maintain a phase difference of 90 degrees between their outputs over a wide band of frequencies. These circuits were discussed in detail previously, except for a minor variation in circuitry shown. R-C values are extremely critical in these networks, even though the circuit looks quite simple. Values are not the same for both networks, although the circuitry is identical.

The output circuit of the networks is cathode-coupled from one channel (I) and cathode- or plate-coupled from the other (Q) channel, depending on the setting of a selector switch. Use of this switch provides a means of cancelling noise or unwanted signals appearing in either sideband. In double-sideband reception, the signals in both sidebands will have the proper relationship to add, regardless of the setting of the selector switch. Both sidebands appear in the I channel when the oscillator is in proper phase, and only noise and out-of-phase components appear in the Q channel. A combining network adds the signals from both channels vectorially, and the desired audio output is taken from the point of balance in the circuit through a potentiometer.

An audio phase detector is coupled through two phase inverters from the outputs of both audio channels. When the oscillator is in correct phase, no signal appears at the Q phase inverter, and zero voltage is present at the output of the discriminator, because of the potentiometer circuit connected between the grid and plate of the I phase inverter, 180 degrees apart.

Whenever the oscillator output shifts in phase, some of the signal appears in the Q channel. This signal is coupled to the Q phase inverter, which upsets the balance of the phase relationships present in the phase detector due to the I channel signals. As a result, a d-c voltage (a-f-c) which is proportional to the amount of phase shift, and the polarity of which depends on the direction of phase shift, is applied to a reactance tube. This voltage causes the reactance tube to change the capacitance of the oscillator tank circuit in the proper direction to correct the phase of the oscillator. An audio-derived a-g-c voltage is applied back to the receiver i-f stages. Audio to operate the a-g-c circuit is taken from the cathode of the I phase inverter. Amplified agc is used to provide adequate level.

This type of adapter is compatible with AM, PM, DSB, SSB, narrow-band FM, and c-w receivers and for some types of reception it may be operated in a non-locked condition.



TP7-1314

Figure 115. Two-Phase Synchronous DSB Adapter.

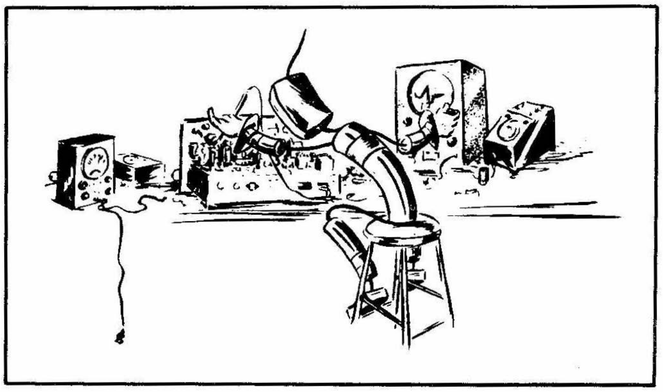
TRANSMITTER MAINTENANCE TECHNIQUES

GENERAL

For proper operation of a communication system, the associated receivers and transmitters must (1) be accurately adjusted to certain assigned frequencies; (2) have the ability to maintain these assigned frequencies after the initial adjustment; and (3) introduce a minimum of distortion to the over-all system. Periodic checks and adjustments of the equipment by qualified personnel will insure that the over-all system performance requirements are maintained. Such periodic checks and adjustments are necessary because all electronic systems are subject to errors introduced by environmental and operational characteristics. These errors can be minimized to a certain extent by proper equipment design, choice of proper components for a particular application, and proper maintenance of the equipment. The areas of equipment design and choice of components are usually determined by the equipment designer (or manufacturer) and cannot be controlled by the maintenance technician. However, by employing the correct maintenance techniques (alignment, system measurements, parts replacement, etc), the maintenance technician can insure that the performance of the equipment will be maximum consistent with equipment design.

The majority of maintenance personnel are familiar with the importance of accurate frequency adjustments and output monitoring of existing AM communication transmitters. Transmitter output monitoring and accurate transmitter frequency adjustment are even more important in SSB than in AM applications. Not only must the SSB transmitter be adjusted to operate within tolerances specified by the FCC; but it must also be adjusted to operate within its own strict tolerances, because intelligibility of the reproduced single-sideband message depends upon such adjustments.

The method employed in testing an SSB transmitter will vary slightly for different equipment applications. For example, some permanent ground installations have special test units incorporated in the over-all system configuration. These test units, often referred to as "distortion measuring equipment," are usually designed for the specific applications where used, and provide a means for making such measurements as distortion, gain, frequency response, noise, and modulation level. It is highly recommended that such distortion measuring equipment be used whenever applicable. If no such equipment is provided, suitable individual test instruments (signal generators, vacuum-tube voltmeters, oscilloscopes, etc) can be used for making these tests.



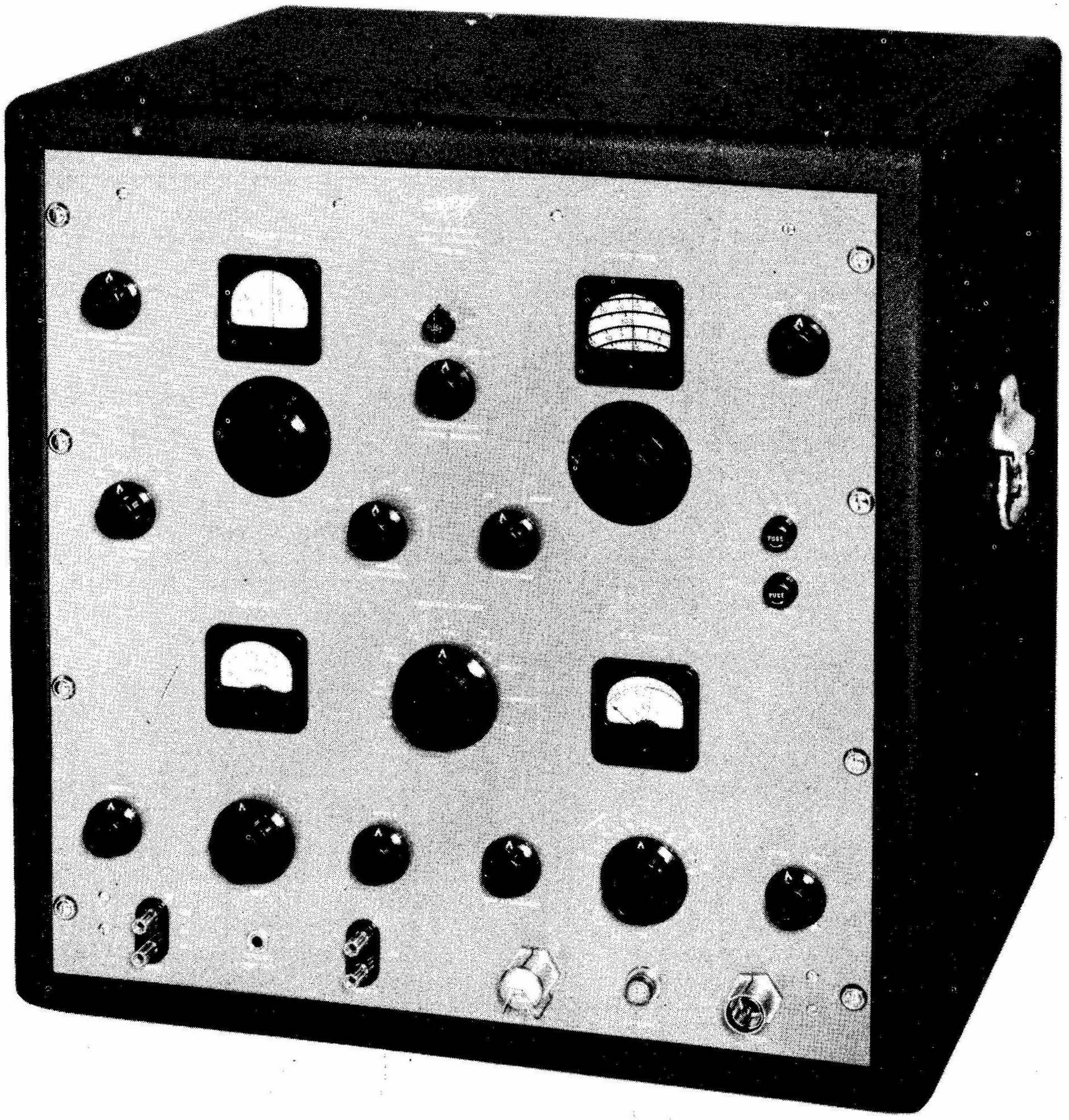
maintenance——

No attempt will be made in this text to present a detailed step-by-step alignment procedure for a particular SSB transmitter because such information can be obtained from the equipment manufacturer's publications. Instead, the information presented will cover some of the maintenance techniques peculiar to SSB transmitters in general.

R-F CARRIER OSCILLATOR

The relationship of the sidebands to the carrier is determined by the modulation process used in the transmitter. For proper demodulation of the signal at the receiver, this relationship of the sidebands to the carrier must be maintained. Although the carrier in a normal AM double-sideband signal contains none of the information of the message, the carrier is still required at the receiver for demodulation of the signal. In SSSC systems (systems in which no carrier is transmitted with the single-sideband signal), a locally generated carrier is inserted into the single-sideband signal at the receiver. This locally generated carrier must be at exactly the same frequency as the carrier that was suppressed in the transmitter if the demodulation process is to be performed properly. Consequently, for proper operation of an SSSC system, the 100-kc oscillators in both the receiver and the transmitter must operate at the same frequency at all times. Therefore, the importance of stability and accuracy of frequency adjustment of the 100-kc r-f oscillator (or its equivalent circuit) in the transmitter cannot be stressed too strongly.

The stability of the 100-kc oscillator is determined by the quality of the resonator used, the method employed to maintain this quality, and the oscillator circuit design. The use of high-Q quartz-crystal resonators enclosed in temperature-controlled ovens and operating in circuits using



TP7-1315

Figure 116. SSB Signal Generator.

regulated power supplies is the most common method used to obtain good frequency stability of such oscillators. The maintenance technician has no direct control over these factors, since they are determined by the equipment designer. The primary concern of the maintenance technician is to insure that the strict requirements established by the circuit designer are maintained in actual equipment operation.

The frequency of the 100-kc oscillator can be checked by comparison with the frequency of an extremely accurate signal generator or by the use of a direct-reading frequency meter. To properly check and adjust this circuit, it is necessary for the accuracy of the test equipment to equal or, preferably, exceed the accuracy of the circuit being tested. Some of the signal generators and frequency meters used in AM applications do not meet the strict requirements of single-sideband systems. Therefore, caution must be exercised when using such instruments to check the frequency of the 100-kc oscillator. Test instruments specifically designed for single-sideband applications should be used whenever such instruments are available. Sufficient warm-up time should be allowed for all test instruments, to insure stable operation of such equipment. A typical signal generator designed for SSB applications is shown in figure 116.

If a check, using the proper test equipment and procedures, reveals that the frequency of the 100-kc oscillator is in error, the first step in maintenance should be to check and adjust the regulated voltages applied to the oscillator circuit. Accurate vacuum-tube voltmeters should be used to make these voltage checks and adjustments. Incorrect voltages applied to this circuit are a major source of oscillator frequency errors. Whenever a tube change is made in the 100-kc oscillator, the applied voltages should be checked and adjusted to the recommended values.

After the applied voltages have been adjusted, the frequency of the oscillator can be properly adjusted. Usually, a small trimmer capacitor is incorporated into the oscillator circuit design to effect minor adjustments of the oscillator frequency. The final adjustment to the oscillator should be the setting of the output signal voltage to the level prescribed by the equipment manufacturer. If a stabilized master oscillator or frequency-synthesized circuit is used, adjustments should be made as prescribed by the manufacturer. The precautions stressed by the manufacturer, and those noted above, should be strictly observed for proper maintenance of the r-f carrier oscillator.

CARRIER BALANCE IN BALANCED MODULATORS

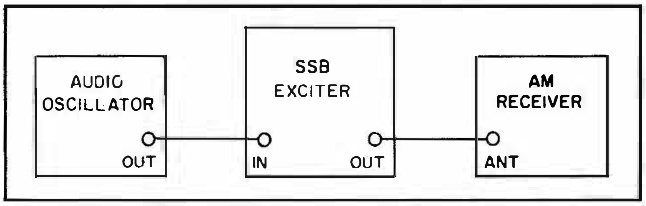
In the TRANSMITTER THEORY section, it was stated that the primary purpose of any balanced-modulator circuit is to produce the sidebands of an amplitude-modulated r-f carrier and to suppress or reject the carrier. The amount of carrier suppression depends upon the degree of balance between the two legs of the balanced circuit. In such circuits using vacuum-tubes, two tubes of the same type will generally balance close enough to suppress the carrier approximately 10 to 15 db without any external adjustments. Since a carrier suppression of at least 35 to 40 db is desirable in SSSC systems, separate bias supplies and R-C balance adjustments are usually provided in the modulator to insure correct balance of these circuits and thus provide adequate suppression of the carrier. If a pilot carrier is transmitted with the single-sideband signal, the level of the reduced carrier is normally 10 to 20 db below the level of the sideband signal. In such systems it is desirable that all of the reduced-carrier signal be applied through the pilot-carrier re-insertion circuits, and that the amount of carrier due to unbalance of the balanced modulator be kept to a minimum. Attainment of this condition insures more accurate control of the pilot-carrier level by the carrier-level control.

A certain degree of balance is "built-in" a balanced-modulator circuit by the equipment designer in the choice of tubes, components, and circuit arrangement. To compensate for changes in values of components due to environmental and operational conditions, separate balance controls are usually incorporated into the circuit design. The maintenance technician is responsible for the proper adjustment of these controls, to insure that the proper degree of balance is maintained in the modulator circuit under operational conditions.

The first step in adjusting a balanced modulator is to check the circuit operating voltages, and to make the necessary corrections (parts replacement, bias supply adjustments, etc) for any errors thus found. When a change of parts is required, caution should be observed to insure that the replacement parts comply with the value and tolerance specified for the circuit. If it is found necessary to change one tube in a circuit using separate balanced modulator tubes, care should be exercised in determining that the replacement tube matches the tube remaining in the circuit. This can be done by checking the mutual conductance of both tubes (the remaining tube and the replacement tube) on a reliable tube checker. After it is ascertained that the tubes are properly matched (by double-checking their operation in the circuit) and that all operating voltages are in compliance with the manufacturer's specifications, the modulator balance-controls can be properly adjusted.

The carrier balance can be adjusted by applying a test tone of constant amplitude and frequency (usually around 1000 cps) to the input of the SSB exciter, and coupling a suitable AM receiver to the output of the exciter (figure 117) or some other appropriate point in the low-level circuits following the balanced modulator. The source of the test-tone signal can be an extremely stable and distortion-free audio oscillator with means of controlling the frequency and amplitude of the output signal. The only requirements of the AM receiver are that it have suitable response characteristics and the proper frequency range to make the test.

After all connections are made, sufficient time should be allowed for warm-up of the equipment. The carrier-level control should be turned to its minimum position; i.e., it should be adjusted so that no carrier is applied to the single-sideband signal through the carrier re-insertion circuits. With a test-tone of sufficient amplitude and suitable frequency applied to the input of the exciter, the AM receiver should be able to receive the single-sideband signal if an unbalance in the mod-



TP7-1316

Figure 117. Equipment Setup for Carrier Balance Adjustment.

ulators exists. The carrier-balance control is then adjusted for minimum tone output from the receiver. During this adjustment the level and frequency of the tone from the audio oscillator should be maintained constant. The volume control on the AM receiver should be adjusted only if the level of the tone from the receiver becomes intolerable or becomes too weak to hear. The exact settings of the audio-oscillator output and receiver volume controls are left to the discretion of the maintenance technician, since each person's hearing is slightly different. The only strict requirement in the setting of these two controls is that they allow sufficient amplitude of tone signal to indicate a definite null point when the carrier-balance control is being adjusted for minimum signal.

The reason for using an AM receiver instead of an SSB receiver for making this adjustment is obvious: Since a carrier is required in an AM receiver for demodulation of the received signal, adjustment of the carrier balance of the modulators for minimum carrier will remove the carrier and thus cause a null in the received signal to be noted. If an SSB receiver were used to make this adjustment, demodulation of the received signal would be performed by the insertion of a locally generated carrier in the SSB receiver. In this case the carrier-balance adjustment would have no effect on the received test tone; therefore, such carrier-balance adjustments would be meaningless. The importance of setting the carrier-level control to its minimum position can also be understood at this time.

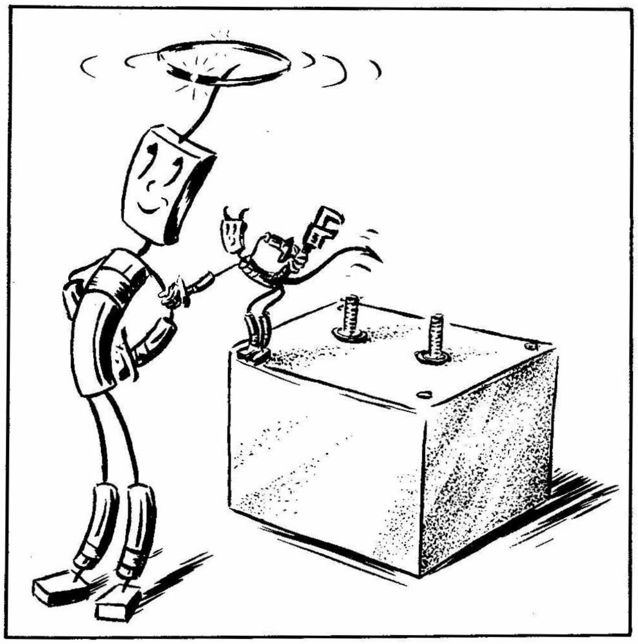
In diode-rectifier balanced modulators (balanced-bridge, ring, or lattice-type modulators), an important factor that must be considered is the proper ratio of the r-f carrier voltage to the modulating-signal voltage. If the distortion products generated in such modulator circuits are expected to be kept to a minimum, this ratio should be kept high. For circuits employing germanium crystals or copper-oxide rectifiers as the diode-modulator elements, the level of the r-f voltage should be on the order of 3 to 6 volts, which is at least eight to ten times the level of the peak modulating-signal voltage. In fact, approximately the same voltage ratio should be observed in all balanced-modulator circuits regardless of the type of device used as the modulator element.

Although strict adherence to the equipment manufacturer's adjustment procedures should be observed and followed at all times, the procedure presented here can be used when such information is not readily available.

SIDEBAND FILTERS

The purpose of the sideband filters in an SSB transmitter is to pass only the desired band of frequencies with minimum distortion and loss. The filter characteristics are primarily determined by the filter design engineer and are usually permanent once the initial design is completed. Not only is it desirable to pass a certain band of frequencies with minimum loss and distortion, but it is equally important that all frequencies outside this desired band be attenuated sufficiently so that they do not appear in the output of the transmitter. This fact is most important in dual-channel SSB systems where information from two unrelated sources is transmitted on opposite sides of a suppressed, or reduced, carrier.

As a general rule, the maintenance technician should not attempt to adjust sideband filters. The necessary adjustments are made and sealed-in during the construction of the filter unit. This fact is especially true of mechanical filters, because some types of these filters are hermetically sealed and no external adjustments are possible. Attempts to adjust mechanical filters should not be undertaken under any conditions.



". . .attempts to adjust mechanical filters should not be undertaken under any conditions——"

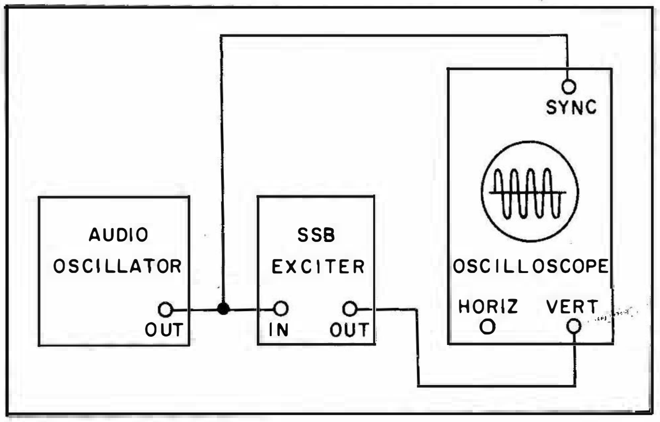


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Figure 118. Equipment Setup for Measurement of Sideband Suppression.

The trimmer adjustments associated with some types of crystal-lattice filters should not be altered by maintenance personnel in the field. If all other attempts at alignment of a transmitter fail, and it is assumed that the trimmer adjustments have been tampered with, only then should an attempt be made to adjust these trimmers. Even under these circumstances only highly skilled technicians having a thorough understanding of crystal filters, and working at activities that have the proper test equipment, should attempt adjustment of these filters.

The importance of the sideband filters in regard to the over-all frequency response and proper performance of a single-sideband transmitter cannot be stressed too strongly. Therefore, strict adherence to the manufacturer's instructions concerning the sideband filters used in his equipment is essential.

SIDEBAND SUPPRESSION

By using an oscilloscope and an audio oscillator, it is possible to closely approximate the amount of undesired sideband suppression. In this procedure (figure 118) a single-tone signal (usually 1000 cps) from the audio oscillator is applied to the SSB exciter under test, with the output waveform of the exciter being observed on the oscilloscope. If the exciter is operating properly, a single-tone input will produce a single-frequency output from the exciter, and the output waveform will be similar in appearance to a normal unmodulated carrier (figure 119A). If the SSB exciter is not operating properly, multiple frequencies will be present in the output, and the waveform observed on the oscilloscope will have the appearance of an amplitude-modulated carrier (figure 119B). The shape of the multiple-frequency envelope will depend upon the number and relative amplitudes of the separate frequencies that are present.

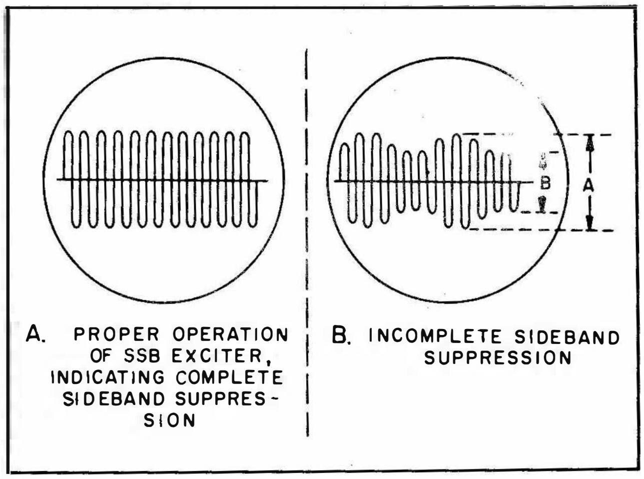
Assuming that the observed signal has the appearance of figure 119B, the suppression ratio between the desired and undesired sidebands can be expressed as:

$$\text{Suppression Ratio (db)} = 20 \log \frac{A + B}{A - B}$$

If more than one audio signal is applied (because of the presence of distortion in the audio amplifier or input signal), the envelope waveshape will be highly complex. However, a relative approximation of the sideband suppression will be possible if the input signal is maintained at a low level to avoid excessive distortion. The r-f carrier must also be sufficiently suppressed to prevent it from appearing in the output and further complicating the output waveform. The amount of undesired sideband suppression should be approximately 35 to 40 db for proper operation of the system.

GAIN OF L-F CIRCUITS

Generally speaking, testing the gain of the l-f section of an SSB transmitter provides a means for determining the gain of the circuits between the input to the transmitter and the m-f circuits. The exact procedure used in making this test will vary for different equipment designs and applications. However, the over-all procedure used is basically the same for all types of equipment.



A. PROPER OPERATION OF SSB EXCITER, INDICATING COMPLETE SIDEBAND SUPPRESSION

B. INCOMPLETE SIDEBAND SUPPRESSION

TP7-1318

Figure 119. Waveforms Observed During Measurement of Sideband Suppression.

In order to test these circuits properly, it is necessary to allow a few minutes for the equipment to warm-up after power is applied. To prevent any signal from being transmitted while making this test, either the m-f or h-f heterodyne oscillators should be disabled. This can be done by removing the crystal corresponding to the carrier frequency being used; or if the oscillator does not use crystals, the circuit can easily be disabled by removing tube from its socket.

After one of the oscillators has been disabled, and the equipment has warmed-up sufficiently, the operating voltages of the l-f circuits should be checked with an accurate and reliable voltmeter—preferably a vtvm. The next step is to apply a test tone (usually 1000 cps) to the transmitter input. The source of this test tone may be the distortion measuring equipment (if available) or an external audio oscillator possessing good accuracy and stability. The gain of the l-f circuits and, in particular, the level of the r-f carrier applied to the l-f balanced modulator should then be adjusted to obtain the proper readings on the test instruments, as recommended by the manufacturer. For example, if the level of the audio input signal is not kept within the proper ratio to the 100-kc r-f carrier oscillator signal voltage, the balanced modulator may produce a sideband signal with the carrier signal not suppressed by the proper amount, and may also produce excessive distortion and high noise levels. In the amplifiers preceding the m-f and h-f mixers, the gain must be controlled so that the signal-to-oscillator injection voltage ratio is of the proper value to insure linear mixer operation, low noise levels, and minimum spurious products.

Some procedures use voltmeters to measure the signal voltage, while others use ammeters to measure grid and plate current in the l-f circuits. Regardless of the method used, the test should always be made according to the recommended procedure and specifications. Upon completion of the test, care should be taken to replace the crystal or tube removed from the equipment at the beginning of the test.

DISTORTION IN SSB TRANSMITTERS

An important requirement of SSB systems is low intermodulation distortion. In such systems the linear power amplifiers are the main source of this form of distortion. Usually the distortion caused by the even-order products (2nd, 4th, etc) are sufficiently removed from the desired signal that normal tuning will eliminate them. Most of the intermodulation distortion in the linear amplifiers is caused by the odd-order products (3rd, 5th, etc) that fall in or near the desired frequencies. Of equal importance are the distortion products that fall outside the assigned SSB channel, because these products may cause interference in the reception of weak signals in equipment operating on the adjacent channel.

A form of distortion peculiar to phase-shift SSB transmitters is called "post-phasing distortion." This form of distortion is caused by harmonics generated in the audio amplifiers following the audio phase-difference networks and by improper adjustment of the balanced modulators. When the signals (including the harmonics) from the two audio amplifiers are applied to the balanced modulators, the distortion products will not be in the correct phase to be cancelled in the output circuit of the modulators. Specifically, the third-order products will cause a single-sideband signal to be present on the undesired side of the suppressed carrier; the fifth-order products will cause distortion to be present in the desired sideband; and all even-order products will be transmitted as double-sideband signals with no carrier. If the carrier is not completely balanced and appears in the output, phase modulation, as well as amplitude modulation, will result.

Proper choice of tube characteristics and operating conditions by the circuit designer can keep the distortion products to a minimum. Although the maintenance technician has no control over these factors, he can insure that the distortion is kept low by observing proper maintenance procedures. These procedures include proper adjustment of element voltages, balance controls, and neutralizing controls, and proper replacing of shields, shielded leads, and feed-through capacitors.

TWO-TONE TEST SIGNAL

The most widely used method of testing SSB transmitters is the two-tone test. This test involves the application of two separate tone signals to the input of a system or circuit and observing of the results on an oscilloscope, spectrum analyzer, or some other indicating device. The two tones should be equal in amplitude and have a difference in frequency of about 1000 cps, in order to achieve the results desired from the test. Typical examples of two-tone test waveforms are illustrated in figure 120.

The sources of the test tones can be either r-f signal generators or audio oscillators, whichever are applicable to the test being performed. In filter-system transmitters, the two-tone test signals can sometimes be obtained by applying a 1000-cps audio tone to the transmitter input and slightly unbalancing the l-f balanced modulator to allow a portion of the carrier to "feed-through." A similar method of obtaining the two tones can be used in phase-shift-system SSB transmitters. In these applications it is only necessary to disable one of the balanced modulators, instead of unbalancing the circuit. When these methods are used, care should be exercised to insure that the amplitudes of the two tones are maintained constant. Upon completion of the two-

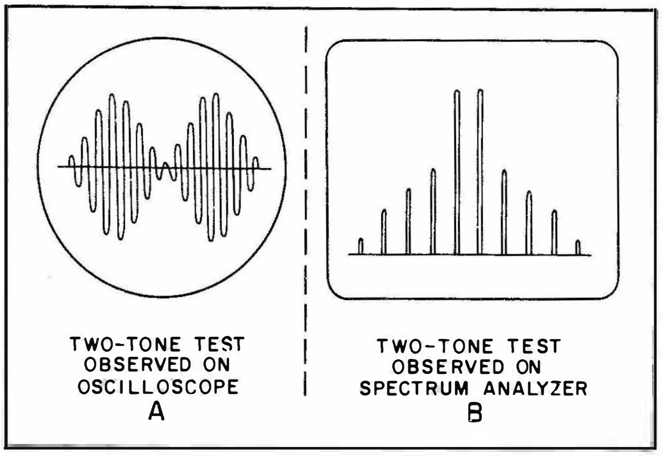


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Figure 120. Examples of Ideal Two-Tone Test Waveforms.

tone test (using these methods), the carrier balance should be re-set and checked to insure that it is at the proper point for operation of the equipment.

SIGNAL-TO-DISTORTION RATIO

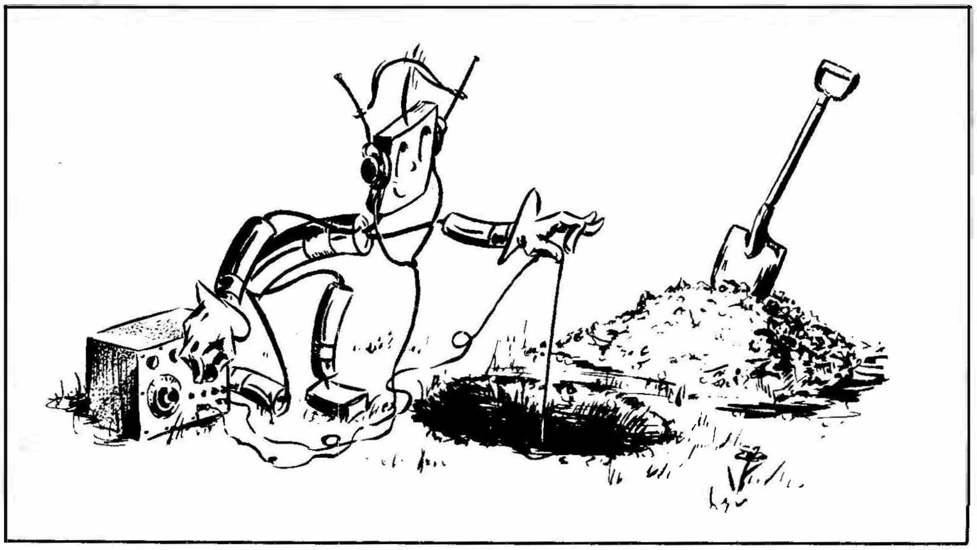
The signal-to-distortion ratio, given in db, is the ratio of the amplitude of one test tone to the amplitude of the third-order product, and is usually determined by using the two-tone test. Although present designs of linear amplifiers can produce signal-to-distortion ratios of approximately 35 to 40 db, lower over-all system distortion values can be obtained by using same form of distortion cancellation, such as r-f feedback. Since the principal causes of distortion in linear amplifiers are non-linearity and grid-current loading, care must be exercised to insure that the linear amplifiers are not overdriven in SSB systems.

The signal-to-distortion measurement can be made by applying the two tones (at the same level) to the input of the transmitter (figure 121) and measuring both the direct output level and the level of the intermodulation products obtained through a bandpass filter and voice-frequency amplifier connected to the monitor output. Since the method of connecting the test equipment will vary for different system applications, the manufacturer's publications should be consulted for each individual system.

SIGNAL-TO-NOISE RATIO

The signal-to-noise ratio, usually given in db, is defined as the value of the signal to the value of the noise in a particular application, and is generally a function of the bandwidth of the over-all transmitting system. In cases where impulse noise is to be measured, the ratio is usually expressed in terms of peak values, i.e., peak-signal to peak-noise ratio. In random noise measurements, rms values are generally used (root-mean-square signal to root-mean-square noise ratio).

To determine the signal-to-noise ratio of a system, it is necessary to measure the output of the system, using an output indicator, both with and without a test tone. Low-level test signals are generally used in making this test so that proper demodulation of the signal can be performed by the output meter. The signal-to-noise ratio is usually expressed in db with respect to the peak envelope power of the transmitter. However, in all cases the equipment handbook should be con-



. . . "low level signals are generally used in making this test——"

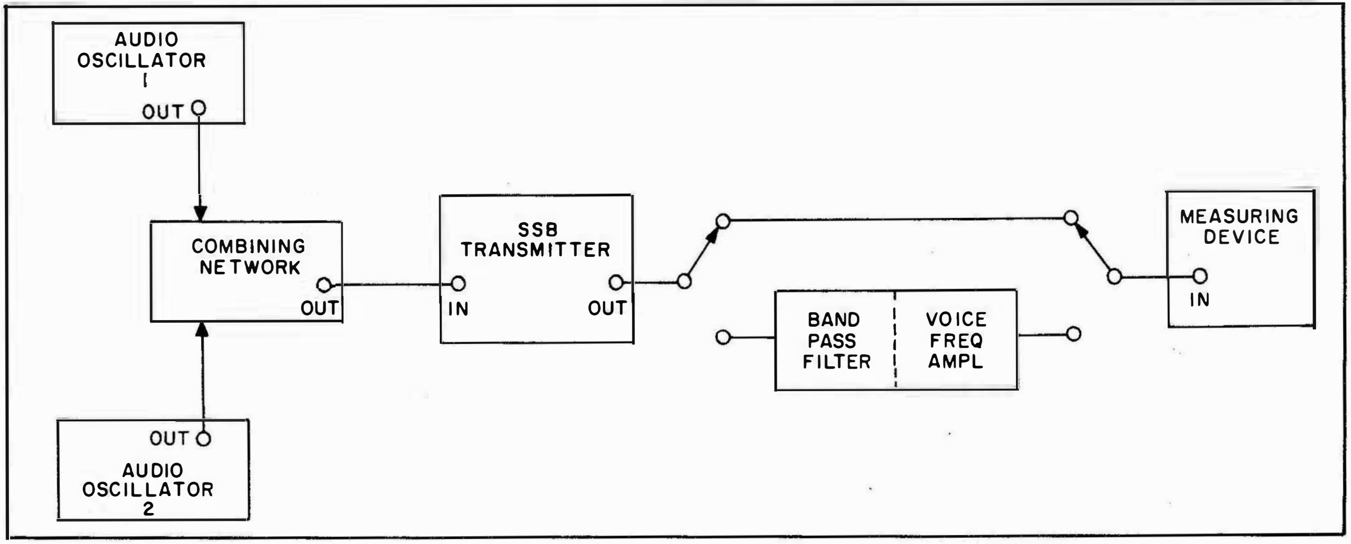


Figure 121. Equipment Setup for Measurement of Signal-to-Distortion Ratio, Using the Two-Tone Test.

sulted for the exact procedure to use in checking a particular system.

AUDIO FREQUENCY RESPONSE

Frequency response is defined as the range of frequencies passed with a specified allowable loss by a system or circuit. In SSB transmitters the frequency response is primarily determined by the insertion-loss and frequency characteristics of the sideband filters associated with the l-f balanced modulator. Proper impedance-matching of the sideband filter input and output to the associated circuit is also an important factor in determining the over-all system frequency response.

The methods of measuring the a-f response of SSB systems are similar to those used in present AM applications. One method is to apply constant-level input signals at different frequencies in the range to be covered and measure the output signal levels for each frequency. Another method is to measure the levels of the input signals at different frequencies in the required range that give a constant output signal level. The usual requirement is that response should be fairly constant over most of the desired band of frequencies to be passed. The exact procedure for making this test will vary for different equipments. Therefore, it is recommended that the equipment test procedures outlined in applicable publications be observed at all times.

TRANSMITTER POWER OUTPUT

There is a direct relationship between the maximum power output of a transmitter and the amount of distortion that is permissible in the system. For this reason, the power output rating of a transmitter is usually given in terms of the maximum power that can be delivered with respect to a specified amount of distortion that can be tolerated. If no value of distortion is specified, it is understood that the distortion will be kept within limits considered to be acceptable for the system. The direct relationship, as well as the importance of this relationship, between the transmitter power output and permissible distortion should not be overlooked when considering the power output of a transmitter.

The power output of an SSB transmitter is usually given in terms of peak-envelope power (PEP), with this term being defined as the r-m-s power developed during the peak r-f cycle. Peak-envelope power is equal to the sum of the amplitudes of the sideband components and the pilot carrier. The measurement of PEP is usually made using the two-tone test procedure with the carrier turned "on". During the two-tone test, the PEP occurs when the peaks of the two tones are in coincidence.

One method of measuring the peak-envelope power of an SSB transmitter is shown in figure 122. In this method, the peak voltage (E_p) of the two-tone test signal developed across a resistive load is observed and measured on an oscilloscope. Since power is equal to the voltage squared divided by the resistance, the PEP can be calculated in the following manner:

$$\text{PEP} = \frac{(.707\ E_p)^2}{R_L}$$

If the voltage developed across the load resistor (E_L) is measured on a vtvm that is calibrated in r-m-s volts, the PEP can be calculated as follows:

$$PEP = \frac{(E_L)^2}{R_L}$$

It is well to keep in mind that the peak-envelope power rating is given in r-m-s values; hence, all such calculations must be in terms of r-m-s volts.

Another rating sometimes given to SSB transmitters is called "peak-sideband power" (PSP). Peak-sideband power is similar to peak-envelope power except that the measurement of this power term is made with the carrier turned "off"; i.e., all the transmitter power is applied to the sidebands and none is applied to the carrier.

"Talking power" is a term often used in reference to SSB transmitters. Talking power is defined as the portion of the transmitter output power that carries the intelligence of the message. Since only the desired sideband is radiated in SSSC systems, the talking power and peak-sideband power of such systems are closely related. As the carrier suppression increases, the amount of available sideband power (and the talking power) will increase. If the carrier is suppressed completely, the total output energy of the transmitter will be applied to the sideband signal. The importance of complete carrier suppression in SSSC systems can now be appreciated.

MONITORING AND TESTING OF SSB TRANSMITTERS

Transmitter monitoring simply means the checking of a transmitter to determine the quality of its transmission. Some transmitters have separate monitor panels or units built into the system, to provide a continuous check of the transmitter while it is in operation. The monitor panel, or unit, samples voltages or currents at different points in the transmitter to ascertain that particular circuits are functioning properly. Monitoring of a portion of the transmitter output signal (or specific circuits) by using an oscilloscope can give a visual indication of the transmitter performance.

The methods used to monitor AM transmitters are well known to most maintenance personnel. These same methods, to a great extent, can be applied to SSB systems. The main differences are in the type of waveforms observed on the oscilloscope and in the interpretation of these waveforms. In all cases, however, the procedures recommended by the equipment manufacturer should be closely followed where available.

Since the linear amplifiers in SSB transmitters produce most of the distortion in such systems, a constant check should be maintained on the linearity of these circuits. An oscilloscope properly connected to the input and output of the linear amplifier is a simple means of performing this check. One method of connecting the oscilloscope to the circuit is illustrated in figure 123. For proper indications using this method, care should be taken to insure that the connections are made properly and that sufficient deflection voltage is applied to the deflection plates. To prevent dam-

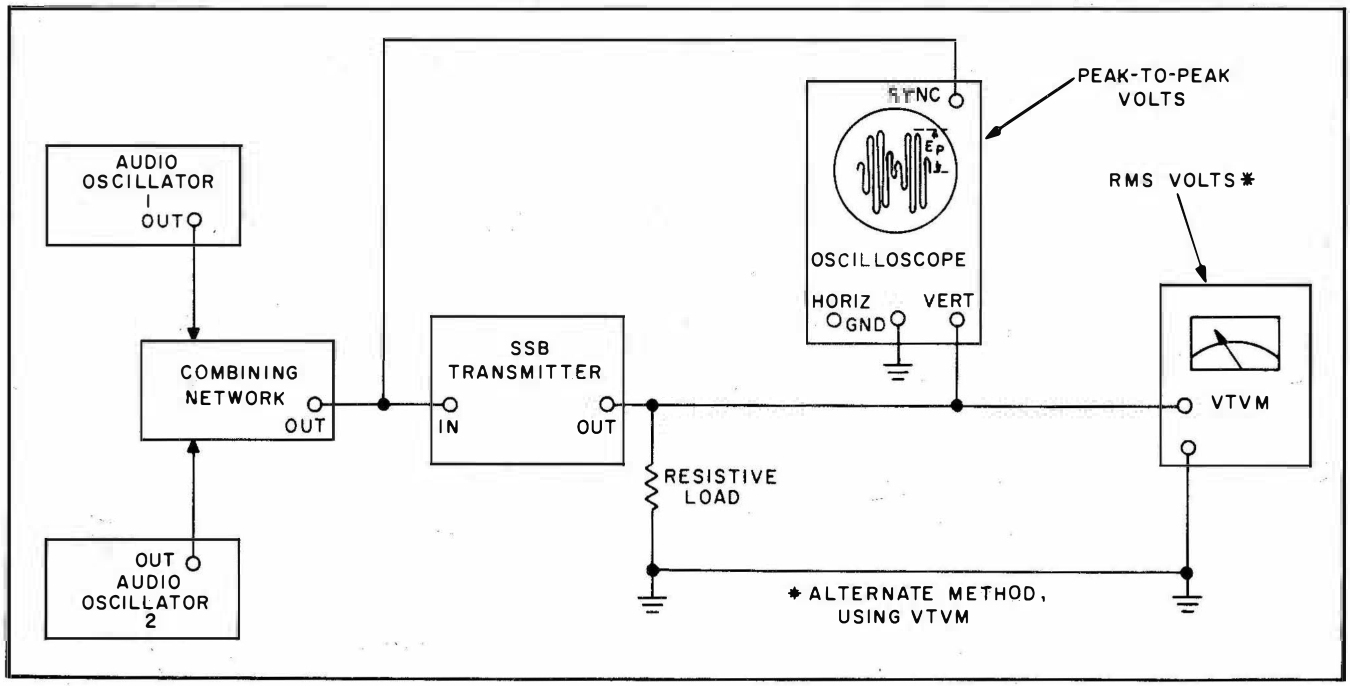


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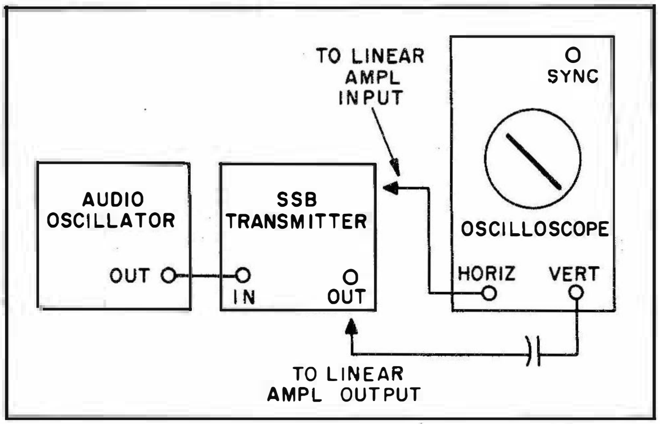
Figure 122. Equipment Setup for Measuring Peak-Envelope Power, Using the Two-Tone Test.



TP7-1322

Figure 123. Equipment Setup for Monitoring a Linear Amplifier with an Oscilloscope.

age to the test equipment, it is important to observe all precautions pertaining to the use and operation of oscilloscopes in circuits where high voltages are present.

The waveforms illustrated in figure 124 are typical examples of the indications that may be expected when a linear amplifier is monitored using an oscilloscope. The waveforms indicated are those that can be expected when using a single tone (including the inserted carrier).

Figure 124A shows the waveform presented when perfect linearity exists between the input and output of the linear amplifier.

Figure 124B indicates a phase shift between the input and output signals. Although a phase difference of 180 degrees between the two signals is understood, any deviation from this value would give a pattern similar to this. This pattern may sometimes appear during initial adjustment of the oscilloscope and may not necessarily be the fault of the transmitter. In this case, different R-C combinations across one set of the oscilloscope plates should be tried until the desired waveform is obtained.

Figure 124C indicates peak limiting caused by too much grid drive or not enough amplifier loading.

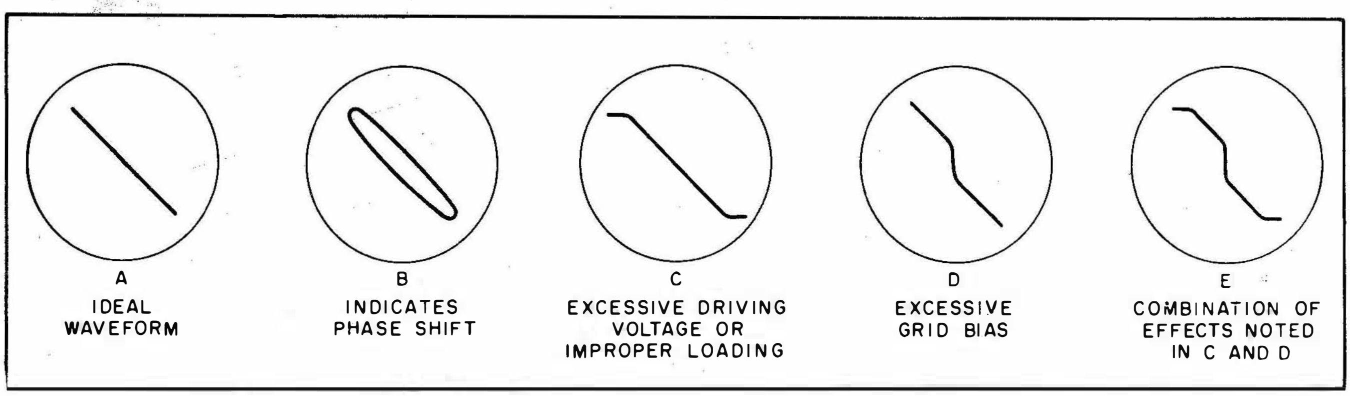
Figure 124D indicates too much grid bias or operation of the amplifier on an improper portion of its operating curve.

Figure 124E indicates a combination of the effects noted in figures 124C and D.

When the linear power amplifiers are monitored or tested using the two-tone test method perfect linearity is indicated by the waveform in figure 125A. Some typical distorted waveforms and their causes are illustrated in the remaining portions of figure 125. The ideal waveform is illustrated by the dotted outline in these examples.

A form of test pattern peculiar only to phase-shift SSB transmitters is the double trapezoidal pattern. This pattern is obtained by disabling the input to one of the balanced modulators in the phase-shift transmitter and making connections as indicated in figure 126. These connections are similar to those made in AM systems to obtain the single trapezoidal pattern used to measure modulation percentage in those systems. The oscilloscope vertical input can be connected to any point in the m-f or h-f circuits. The individual triangles should have the same characteristics as the regular single trapezoidal pattern; that is, straight sloping sides to indicate proper circuit operation. Double-trapezoid patterns indicating some typical troubles in phase-shift transmitters are shown in figure 127.

The importance of utilizing the manufacturer's publications, data, and procedures in all maintenance, testing, and monitoring of electronic equipment cannot be stressed too strongly. Even with such information on hand, the degree of maintenance of such equipment depends directly upon the skill of the maintenance technician in the proper use of this information and the proper use of the test equipment provided.



TP7-1323

Figure 124. Typical Waveforms Observed when Monitoring a Linear Amplifier with an Oscilloscope.

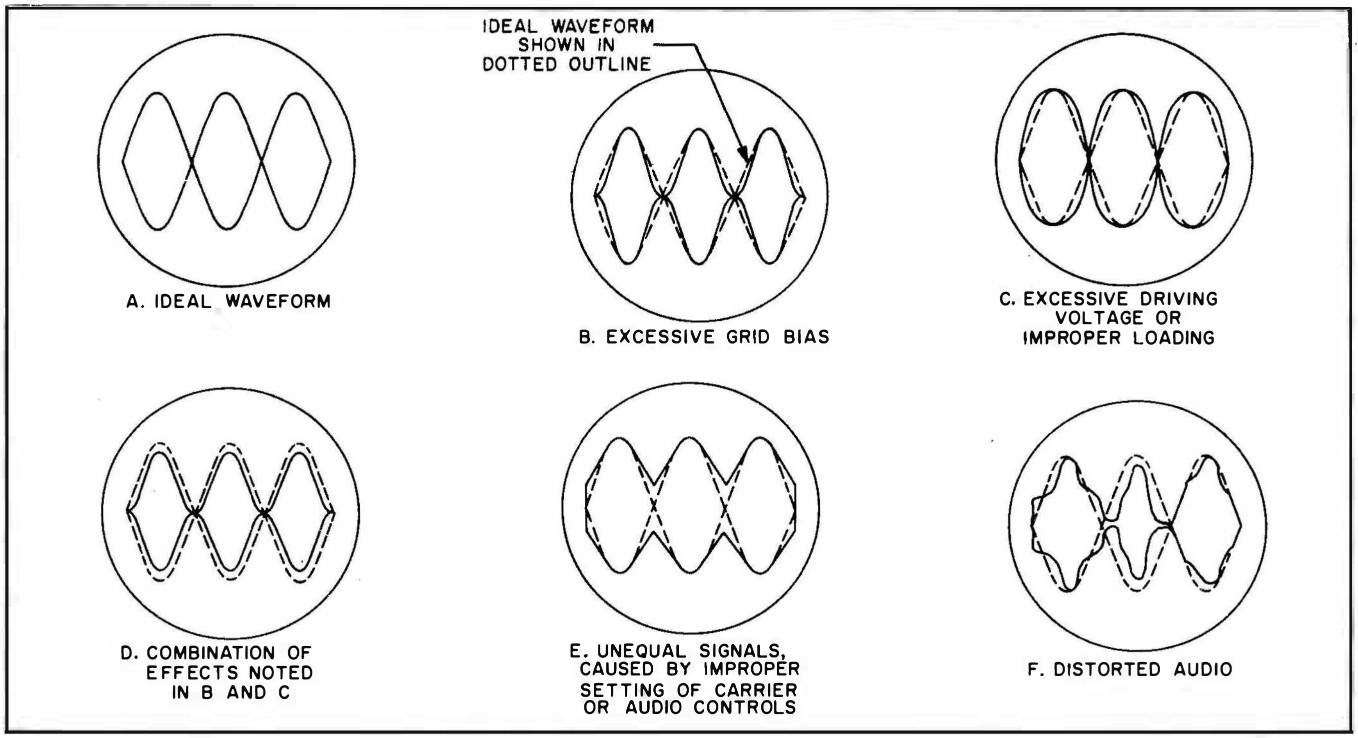


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Figure 125. Waveforms Observed on an Oscilloscope when Monitoring a Linear Amplifier, Using the Two-Tone Test.

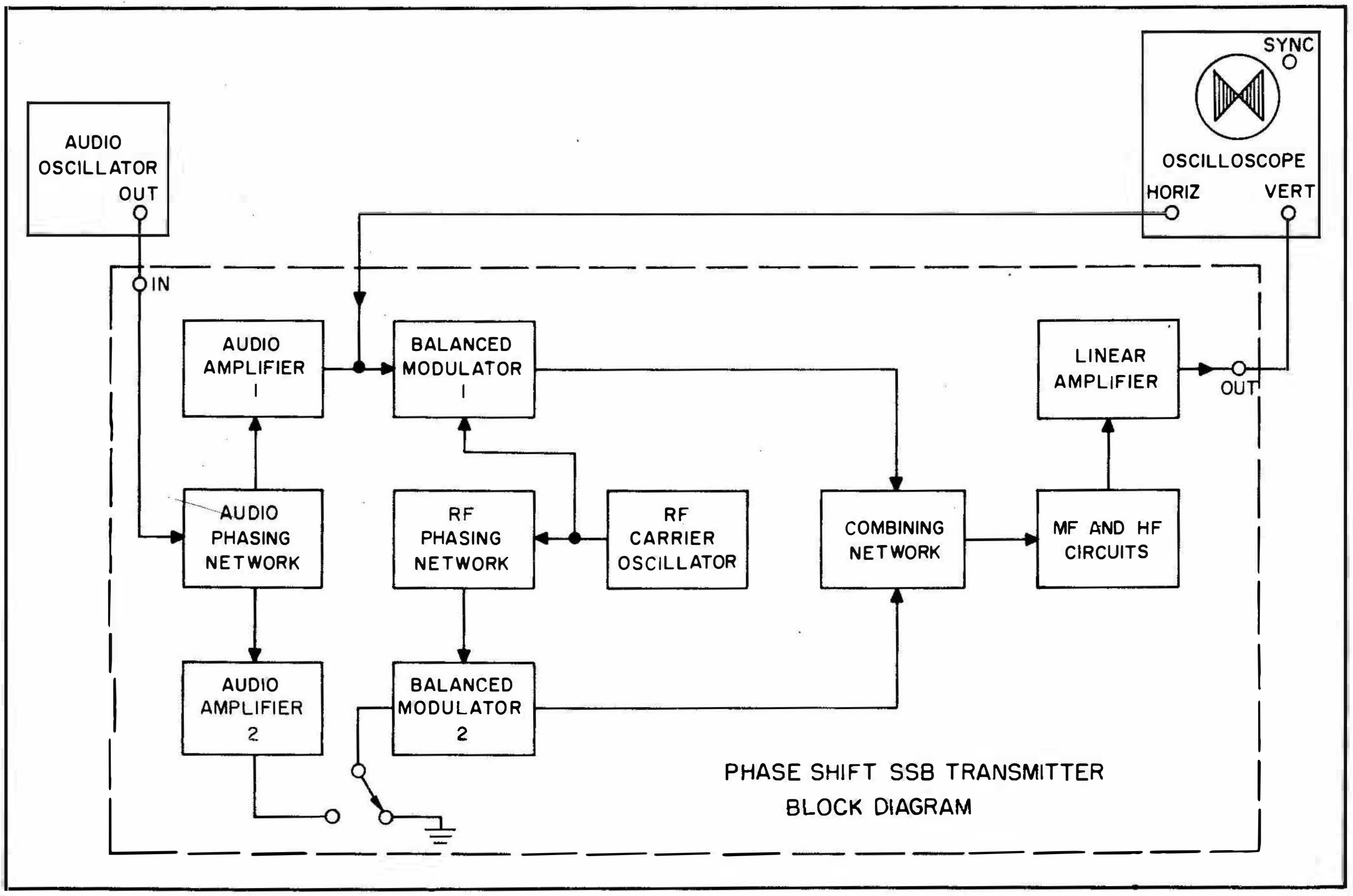

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Figure 126. Method of Obtaining Double-Trapezoid Test Pattern.

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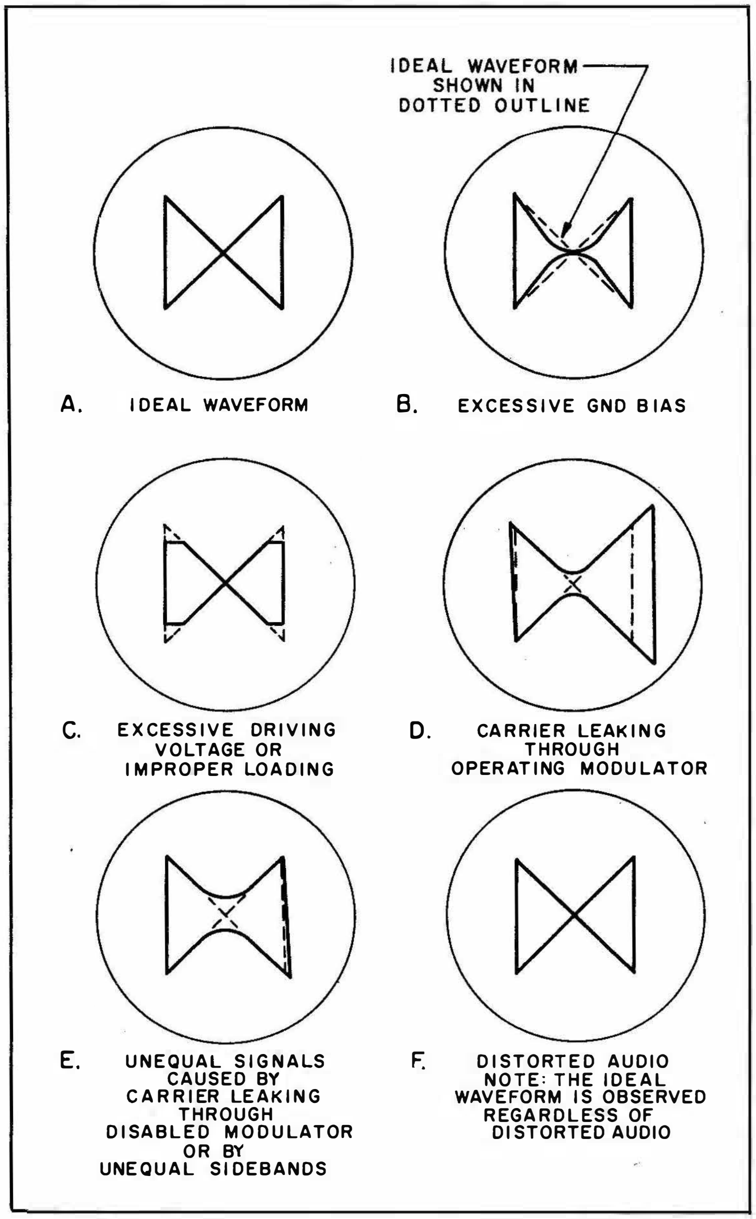


A. IDEAL WAVEFORM

B. EXCESSIVE GND BIAS

C. EXCESSIVE DRIVING VOLTAGE OR IMPROPER LOADING

D. CARRIER LEAKING THROUGH OPERATING MODULATOR

E. UNEQUAL SIGNALS CAUSED BY CARRIER LEAKING THROUGH DISABLED MODULATOR OR BY UNEQUAL SIDEBANDS

F. DISTORTED AUDIO NOTE: THE IDEAL WAVEFORM IS OBSERVED REGARDLESS OF DISTORTED AUDIO

TP7-1326

Figure 127. Double-Trapezoid Waveforms Observed During Phase-Shift SSB Transmitter Testing.

RECEIVER MAINTENANCE TECHNIQUES

GENERAL

Quality and precision may be built into a receiver, through careful design practices and controlled manufacturing. The quality, to a large extent, determines the ability to retain this precision for an expected period of time. Beyond this point, only proper maintenance can insure the original performance of the equipment.

Proper maintenance of single-sideband and exalted-carrier receivers should include the use of test equipment designed for the purpose, when possible, or test equipment with an accuracy equal to, or better than, the accuracy to be maintained in the receiver. The quality of the results of maintenance can never exceed the degree of skill and accuracy with which the maintenance is performed, regardless of the quality of the receiver or the test equipment used. It is important, therefore, that this maintenance be performed by competent technicians with an understanding of the equipment involved.

RECEIVER ALIGNMENT

Sensitivity and selectivity may be affected in alignment of all types of receivers. As receivers become more complex, alignment becomes more of a problem. In AM receivers using conventional full-carrier signals, improper alignment may result in loss of weak signals, through loss of sensitivity, and the ability to select the desired signal may be impaired. If the oscillator is shifted off frequency, dial error will be introduced; and tracking error produces a varying intermediate frequency, which results in loss of signal over portions of the frequency range of the receiver. In FM receivers the discriminator tuning becomes somewhat critical, and in PM receivers, phasing of the carrier must be correct, adding to the alignment problem. When avc and afc are added to a receiver, proper alignment procedures must be followed, or serious errors may be introduced. When multiple conversion is incorporated in the receiver, with two or more heterodyne oscillators, additional variables are introduced, further complicating alignment.

In equipment employing crystals, as either reference generators, oscillators or filters, the alignment must center around the crystals, since the frequencies of crystals are not variable.

All of these factors must be considered in the alignment of single-sideband receivers. In addition, the oscillator frequencies must be precisely adjusted, because demodulation is directly affected by the oscillator frequency and any associated error. Filters must also be considered, since these affect the bandpass and rejection of undesired frequencies. Most of the advantages of single-sideband reception depend on the use of these filters. Some of the more important considerations will be discussed in the following paragraphs. Because of the wide variation in circuitry and layout, actual procedures and alignment specifications must be obtained from manufacturer's data for the particular equipment involved.

Test Equipment Requirements

Two very important factors must be borne in mind in the choice of test equipment for maintenance or adjustment of SSB receivers: (1) the close tolerance to which the oscillator frequencies must be held, and (2) the sharp edges of bandpass which must not be defeated.

Signal generators designed for single-sideband applications are presently available. Such a generator should include a single-sideband output signal, with provision for either upper or lower sideband, or both. It should also include a carrier output, with variable (or selectable) frequency and level, both of which should be accurately calibrated. Alternate methods of alignment, using available signal generators can be used; however, the limitations that may be imposed by the use of these methods should be investigated before a high degree of system analysis or final receiver evaluation is attempted.

Signal generators should never be used as reference for oscillator adjustments in either SSB transmitters or receivers, because of the allowable tolerances of such signal generators. Frequency meters or standard reference oscillators with a known accuracy must be used for such adjustments. The reference used must have an accuracy at least equal to, or greater than, the oscillator to be adjusted. Preferably the accuracy should be greater than ten-to-one; that is, the percentage of tolerance of the reference oscillator should be less than the allowable percentage of tolerance of the oscillator to be adjusted, by a ratio of at least ten-to-one.

Low-Frequency Reference Oscillator

No adjustment should be made to the low-frequency reference oscillator during routine alignment checks, unless satisfactory operation cannot be obtained by other adjustments in the receiver, and incorrect l-f oscillator frequency is indicated. However, tube replacement in the oscillator circuit should be followed by a frequency check; likewise, a frequency check should follow any circuit changes or repairs to the oscillator. In any event, the highest degree of skill should be exercised in performing any repairs or adjustments to the reference oscillator, since both the tuning and demodulation may be affected.

The normally accepted standard frequency for low-frequency reference oscillators or carrier-insertion oscillators in SSB transmitters and receivers is 100 kc, although several other frequencies (25 kc, 75 to 85 kc, and 250 or 300 kc) are becoming popular. Reference standards of 100 kc are readily available in both commercial and military activities, and these oscillators may be used in the adjustment of receiver reference oscillators, provided the frequency accuracy is greater than that of the receiver oscillator to be adjusted. Such oscillators should be checked against the radio standard-frequency signals of WWV prior to use. It must be remembered, however, that the reception of standard WWV signals on the SSB receiver alone does not constitute a check of the receiver oscillators, because manual tuning or a-f-c action will correct the receiver tuning to compensate for low-frequency-oscillator error, the error then being reflected to appear as a high-frequency oscillator error or, possibly, as an a-f-c error. Therefore, either an external receiver or a high-gain, wide-band oscilloscope should be used to compare the signals of the oscillator to be corrected with the frequency standard, by a visual indication method.

A small variable trimmer capacitor is usually included in the oscillator circuit, either in series or in parallel with the crystal. The oscillator frequency may be adjusted, within a range of a few cycles, by means of this capacitor. In circuits where no variable capacitor is included, actual replacement of a small fixed capacitor may be required.

In receivers designed for pilot-carrier reception, employing crystal carrier filters and comparator-type a-f-c circuits, a different approach should be used. In these receivers, the reference or carrier-insertion oscillator frequency must be the same as the center frequency of the carrier filter, even though the carrier-filter frequency may not be exactly 100 kc. In this type of receiver, the carrier-oscillator frequency is adjusted to correspond to the mid-band noise passed by the carrier filter in the absence of a signal. This adjustment may be readily determined by the rate of drift of the a-f-c motor, or rate of voltage change at the grid of the reactance tube, depending on which type of control is used. The rate of drift or the rate of voltage change increases as the error increases.

Sufficient warm-up time must be allowed for both receiver oscillators and the test equipment before any oscillator adjustments are made, and drift should be checked over a period of at least several hours before a final adjustment is made.

Heterodyne Oscillators

The actual frequency of the heterodyne oscillators is not as important as the frequency stability of these oscillators. This requirement may be eased considerably in receivers using a-f-c circuits and pilot-carrier reception. Adjustment of the high- and medium-frequency oscillators may be satisfactorily accomplished following normal AM alignment techniques. It must be remembered, however, that the use of crystal or mechanical filters in SSB receivers demands that the final setting of the heterodyne oscillators result in an intermediate frequency centered within the filter frequencies, regardless of the actual filter frequency.

Filters and I-F Stages

No attempt should be made to adjust the filters in SSB equipment except by highly skilled personnel with a thorough understanding of crystals and filters. Such attempts should be made only when proper facilities and specialized test equipment are available.

Credit must be given to licensed radio amateurs who are constantly seeking to improve radio circuits and equipment under extreme operating conditions with the severe handicaps of limited facilities and equipment and usually a very limited economy. Time consumption and disheartening results seldom exceed the patience and fortitude of these skilled technicians. Many filters and filter designs may be credited to economy-minded amateurs.

These filters, however, are not typical of the filters found in military or commercial SSB equipment. Such filters are usually hermetically sealed or potted and should not be tampered with. After all other attempts at proper alignment or circuit tests and adjustments have failed, replacement of a filter may be deemed necessary. Some lattice or half-lattice type crystal filters are provided with several adjustable trimmers, accessible as screwdriver adjustments. Some of these are labeled as factory adjustments and should never be disturbed. In any case, the manufacturer's data should always be consulted before any adjustments are attempted. Plug-in filters are readily replaced. Mechanical filters should never be tampered with or repairs attempted under any condition.

Schematic diagrams of crystal filters usually indicate variable capacitors and often variable inductances. Such diagrams may be misleading to those unfamiliar with filter circuits. Capacitors in parallel with filter crystals are usually of very small value, on the order of one to several micro-microfarads. These often consist of a pair of leads given a slight wrap or twist, or may be a piece of wire bent near the crystal holder or electrode. Such capacitors are factory-adjusted and are usually not accessible without dismantling the filter.

If attempts are made to repair or adjust a filter, or if a filter has been tampered with, it must be remembered that a filter not only must pass the entire range of frequencies within its bandpass with uniform response, but also must have sharp cutoff and offer great attenuation to frequencies outside its band. High attenuation is especially important in the region of the carrier frequency, and in the case of the sideband filter, the undesired sideband must be rejected in its entirety. Perhaps of equal importance in SSB receivers is the shielding, especially in proximity to the filters. Good filtering will be defeated if feed-through of the undesired frequencies, or stray coupling is permitted. When filtering faults are suspected, the shielding should not be overlooked. Removable covers and interstage shields should be inspected for placement and for adequate bonding or grounding. Shielded leads should not be replaced with unshielded leads, and decoupling resistors, capacitors, and r-f chokes should be inspected when feed-through is evident. The elimination of fed-through is of utmost importance in dual-channel systems to prevent cross-talk or cross-modulation.

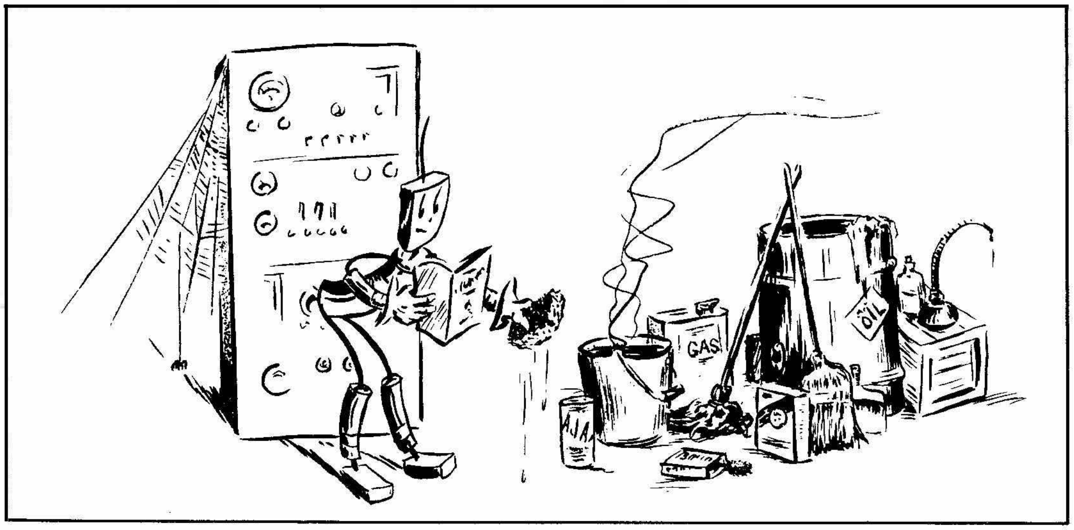
I-F alignment in SSB receivers may follow a pattern similar to that in the i-f alignment of standard AM receivers, with the exception that the bandpass must be determined by the characteristics of the filters involved. I-F transformers are usually tuned so that all of the frequencies within the range of the filters are passed; that is, the selectivity curves of the transformers are broader than those of the filters. Correct tuning of the transformers, therefore, is that which insures maximum signal at the output of the receiver, or, in the case of the carrier amplifier, maximum a-v-c voltage.

A-F-C Systems

A-F-C systems should be checked for proper control and range. Balance must be maintained in discriminators and balanced modulators, to provide the proper range of control and freedom from drift. Systems using motors or servomechanisms must be checked for mechanical freedom and for sticking or binding. Cleaning and lubrication should be performed according to the manufacturer's specifications. Range and control, or over-all performance, of the a-f-c system can be checked by manually detuning the receiver suddenly and checking the a-f-c response. Some mechanical systems employ a damping device, such as an oil-filled dashpot or a flywheel, to prevent oscillations of the mechanical drive, and provide a hold feature to prevent sudden changes in frequency (from noise or jamming) or deep carrier fading from affecting the afc. These components should be inspected if included in the system.

The frequency limits of the carrier filter almost exclusively determine the range of the a-f-c system in pilot-carrier systems. Carrier filters must of necessity have a relatively narrow bandpass, usually less than 100 cycles. When the carrier in the i-f amplifier falls outside the frequency range of the carrier filter, the carrier will be attenuated far below the level required for a-f-c operation, and the system will lose reference of the carrier. It should, however, maintain control over the range of the carrier filter.

In the event of improper a-f-c operation in comparator-type circuits, the frequency of the low-frequency reference oscillator should be checked. This adjustment should be such that the a-f-c system will align on the mid-band noise passed by the carrier filter in the absence of a signal. If the a-f-c control drifts during this check, the fre-



". . . clean and lubricate according to manufacturer's specifications——"

quency of the reference oscillator may be suspected as being incorrect, after it is ascertained that the a-f-c balanced modulator or discriminator is properly balanced and adjusted.

Balanced Detectors (Demodulators)

Proper operation of balanced detectors, used extensively in single-sideband receivers, depends primarily on the degree of balance achieved in the detector. Balanced detectors, or demodulators, are essentially the same as balanced modulators, except that input and output are reversed. In balanced detectors, the balancing is not as important as in balanced modulators in transmitters, where the carrier suppression depends on the degree of balance in the modulator. Balanced detectors are used rather to insure that only the results of heterodyning of the input signal with the inserted carrier will appear at the output. Carrier suppression is not as important as it is in a transmitter, because any r-f carrier appearing in the output will be bypassed in the audio circuits. Therefore, the only noticeable effect resulting from an unbalanced condition is the increased noise nevel and spurious beats, or heterodynes, caused by the beating together of undesired components by rectification. Balance can be checked easily in balanced detectors by simply disabling the signal input and adjusting the carrier balance for minimum noise in the output. This noise is the result of the heterodyning of the carrier with the noise components adjacent to the carrier frequency in the carrier circuits, or channel. Disabling the carrier input to the detector and applying a signal to the receiver (or detector input) should result in minimum output with the same adjustment settings. Matched tubes or diodes should be used, and the balance should be checked whenever tube or diode replacement is made. Tubes should be checked for equal transconductance, and diodes (crystals or copper-oxide rectifiers) should be checked for equal forward resistance. Often, potentiometers are included in the cathode, grid, or plate circuits of the demodulator so that element voltages may be adjusted for balance or to compensate for differences in tubes.

Balanced modulators in comparator-type a-f-c systems should be balanced to produce minimum output voltage in the absence of either input signal. Product detectors, or multi-grid detectors, are balanced in much the same manner as balanced detectors. Each signal grid should be operated on the linear portion of the characteristic E_g-I_p curve for the tube. Balance may be checked by removing or disabling the signals or carrier from the opposite grid and adjusting the element voltages for minimum signal, and then repeating the procedure for the opposite grid.

Carrier Circuits

In exalted-carrier and SSB systems using carrier circuits, the carrier must be separately filtered and amplified to produce sufficient level for demodulation of the sidebands at the demodulator.

The carrier is subject to fading and noise conditions which may vary its amplitude. Variations in amplitude would affect the effective percentage of modulation at the demodulator. Increases in carrier level beyond the point required for proper demodulation merely adds to the noise level, because of the thermal and other noise, and thus results in a lower signal-to-noise ratio in the output of the demodulator.

Most carrier circuits employ limiting stages combined with several modes of a-v-c operation, to maintain a constant carrier level. The bandpass is usually extremely narrow, and is the result of a compromise between the ideal, or narrowest possible bandpass, and that which will allow for carrier output when the receiver is detuned, so that the carrier-operated a-f-c circuits will receive the carrier reference signal. If the carrier bandpass filter circuit is too wide in frequency range, some of the sideband signal may be passed, and this may take control of the a-f-c circuits, causing erratic operation. No attempt should be made to change the bandpass, because it must be determined exclusively in the filter. However, the carrier level at the demodulator, which is usually about ten times the sideband voltage, should be checked and adjusted according to the manufacturer's data. Proper limiting action should also be assured, and a-v-c and a-f-c voltages should be checked for amplitude and range.

Gain and Distortion

R-F Stages

The gain of the r-f stages must be kept low in an SSB receiver, to prevent overloading of the receiver. The primary functions of r-f stages are: (1) to provide preselection of the desired signal; (2) to reject the undesired signals; and (3) to prevent radiation of the receiver local oscillator signal and its harmonics. Several detrimental effects account for various forms of distortion which may be accredited to r-f stage malfunctions. These are listed as follows:

Receiver intermodulation. This type of distortion is the result of two or more undesired signals, or components of the desired sidebands, or both, mixing in the r-f or mixer stages and producing undesired products at the receiver response frequency. It is caused by non-linearity in the r-f circuits or tubes or by overloading.

Receiver cross modulation. This type of distortion is the result of modulation of the desired carrier or sideband components by the sidebands of a nearby undesired carrier. It is caused by vacuum-tube third-order action, poor r-f selectivity, or antenna isolation.

Receiver desensitization or blocking. This condition is caused by overloading of the r-f and mixer stages by a nearby off-channel signal of high level, and is a result of inadequate preselection. It is a form of adjacent-channel interference.

Receiver spurious response. A spurious response results when frequencies other than the desired signal are converted to the i-f frequency. Such a response is due to insufficient r-f preselection. Receiver spurious responses include the following:

1. Intermediate frequency response. This response results when signals at the intermediate frequency of the receiver feed through the r-f stages or enter the i-f stages through improper shielding, etc.

2. Image response. This response consists of an undesired i-f signal which is produced by heterodyning of the oscillator frequency with a frequency located on the opposite side of the oscillator frequency from the true signal, and having the same frequency separation from the oscillator frequency as the true signal. When the oscillator is operated above the signal frequency, the image is the undesired signal below the oscillator frequency. One of the most desirable characteristics of a receiver's r-f system is good image rejection.

3. Harmonic functions. These functions include: responses at submultiples of the operating frequency; responses produced when harmonics of the receiver local oscillator heterodyne with undesired signals, or when harmonics of lower-frequency signals heterodyne with the oscillator signal; and those produced when various other harmonic combinations heterodyne to produce the intermediate frequency or the input frequency.

Most of these effects can be minimized by preventing overload at the r-f stages. All of these effects must be eliminated in the r-f stages, because they cannot be eliminated after introduction in the i-f stages. An attenuator is used in many SSB receivers in the input or antenna circuit. Proper setting of the attenuator is an important aid in preventing overload in most cases. Proper a-v-c time constants and voltages also help prevent overload.

Moisture content, dirt, corrosion, and poorly soldered joints all affect the selectivity (Q) of r-f and i-f circuits. These conditions should be checked during maintenance. Proper tube element voltages are necessary to prevent non-linearity in these circuits.

Elimination of these forms of distortion is important in all forms of AM receivers, and is even more important in SSB receivers, where improved reception is the intended objective. Overload is especially serious in SSB receivers, because the carrier-insertion signal level is held constant, and excessive sideband signal level at the detector will result in overmodulation distortion.

I-F Stages

As in AM superheterodyne receivers, most of the gain and selectivity of SSB receivers are provided by the i-f stages. Also, as in AM receivers, overload and non-linearity must be prevented in the i-f stages. Overload in either the r-f or i-f stages of an SSB receiver produces large amounts of intermodulation distortion, which is considered to be the most serious form of distortion affecting SSB. Intermodulation in i-f stages is produced by the mixing of the individual sideband components in the desired signal, or a combination of the components of the desired signal, with other undesired responses which may be within the i-f frequency range due to the aforementioned r-f malfunctions.

Harmonic distortion, also produced by overload or non-linearity, is produced in greater amounts in full-carrier systems. It is more pronounced than intermodulation distortion in full-carrier (AM or CSSB) receivers because of the presence of the carrier, which beats with the sidebands to produce second- and third-order products and harmonics. As the carrier level is reduced, harmonic distortion decreases, and the intermodulation distortion becomes the more serious effect. It has long been recognized that the human ear is more sensitive to intermodulation distortion than to harmonic distortion. This fact accounts for the more rapid loss of intelligibility in reduced- or suppressed-carrier SSB receivers under overload conditions.

Other malfunctions, such as motorboating or oscillation, feedback and feedthrough, and improper coupling or decoupling, are common to i-f stages in all types of receivers, and are corrected or prevented in SSB receivers by conventional methods.

Audio and Output Circuits

Distortion in audio circuits should not be overlooked, especially in telephone and data systems where continuous monitoring is not always possible. Sine-wave and square-wave tests using oscilloscope presentations and distortion measuring equipment, when available, are most desirable, and pulse test equipment should be used on data and pulse or multiplex equipment. The outputs of receiving equipment handling such services are usually fed into auxiliary equipment, and a thorough knowledge of this equipment is essential in the maintenance of such systems.

Dual-Channel Systems

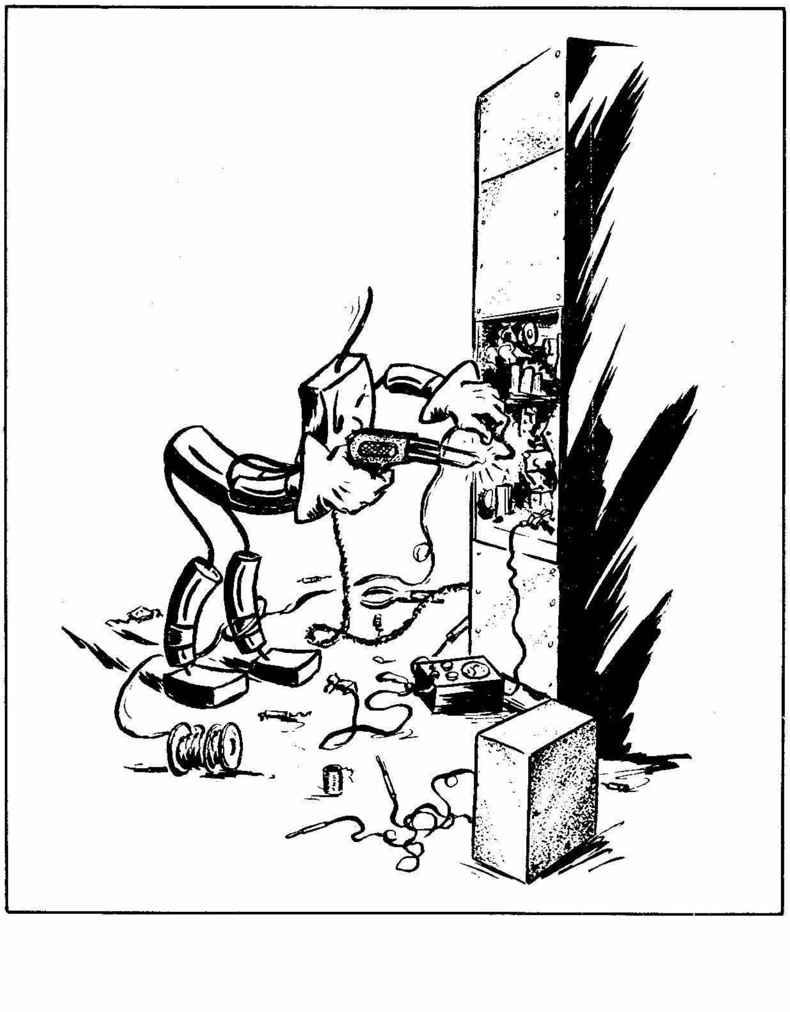
All of the information above also applies to dual-channel systems or DSB receivers. It is also important in such systems that both sideband channels have approximately the same gain. Tube transconductance should be checked in the m-f and l-f stages, and both channels of dual-channel systems should be balanced during maintenance.

Crosstalk is perhaps the most serious effect in dual-channel or multiplexed systems. This can be produced by direct feed-through or coupling between channels due to improper shielding or isolation of circuits. Shielding should always be replaced if removed, and bonding or grounding should be checked if feed-through is evident or suspected. Crosstalk can also be produced by cross modulation. When the offending channel is the lower sideband of a carrier, and the upper sideband of the same transmission is the affected channel, or vice versa, the crosstalk will be intelligible, and will appear as a conversation in the background of the desired signals. Privacy of the channels is thus impaired. The same holds true of two transmissions on or very near the same frequency. In this case, four combinations (upper and lower sidebands, or A and B Channels, of both transmissions) of crosstalk may be produced as a result of co-channel interference.

When the two transmissions are separated in frequency, the frequency of the background signal on the affected channel increases (or decreases) to a point where it becomes unnatural, and, as the frequency separation increases, it becomes unintelligible. The sound heard is commonly called "monkey chatter," since it resembles the chattering of monkeys.

A form of adjacent-channel interference occurs when the carriers of two transmissions are so close together in frequency that the lower side-frequency spectrum that is occupied by the upper band of carrier A occupies the same space in the sideband of carrier B or vice versa; i.e., when carrier A is approximately 3 kc higher (or lower) than B. Under these conditions cross modulation between the two exists, and the background consists of inverted speech (also a form of "monkey chatter"). Although inverted speech is unintelligible in most cases, it might be interpreted and, therefore, should be eliminated when secrecy is involved.

Inverted speech occurs whenever a sideband which is inverted (high-frequency and low-frequency audio components interchanged in positions relative to the carrier) with respect to its own (or a different) carrier is demodulated in that form.



GLOSSARY OF TERMS

active network A network which receives power from both the input signal and other external sources.

balanced detector (demodulator) A detector so connected that the circuit is in a state of balance with respect to one or both of its two inputs, the degree of balance determining the amount of suppression of one or both input signals, so that only the sum and difference frequencies of the two input signals appear in the output. The circuit and its operation are identical to that of a balanced modulator except that the input and output are reversed.

balanced mixer A mixer connected as a balanced modulator; the circuit and its operation are similar to that of the balanced modulator or detector, the difference being in its use.

balanced modulator A modulator so connected that the circuit is in a state of balance with respect to one or both of its two inputs, the degree of balance determining the amount of suppression of one or both input signals, so that only the sum and difference frequencies of the two input signals appear in the output.

bandpass A band of frequencies passed with little or no attenuation.

bandwidth The number of cycles per second expressing the difference between the limiting frequencies of a frequency band.

beating oscillator (heterodyne oscillator) An oscillator used for frequency translation or conversion in transmitters and receivers.

bi-lateral circuit A circuit which will properly function in either direction of signal path.

bi-mode Having two modes of operation.

bridging filter (roofing filter) A filter used in SSB-DSB systems, the bandpass of which encompasses all of the desired frequencies, or the frequency ranges of two or more channel filters. It is used to enhance the rejection characteristics at and beyond the upper and lower frequency limits of the independent channel filters.

carrier filter A filter designed to pass only the carrier frequency, with steep characteristic response curves and adequate attenuation on either side of the carrier frequency.

carrier insertion The process of inserting, or injecting, either the reconditioned pilot carrier or a locally generated carrier into the circuitry for combining with the sideband, or sidebands. This process is used for the purpose of demodulation in a single- (or double-) sideband receiver, or exalted-carrier receiver.

carrier-insertion oscillator The oscillator which generates the local carrier to be combined with the sideband in suppressed-carrier receivers. It may also be called the low-frequency, or reference, oscillator when applicable.

carrier re-insertion The process used in SSB or DSB transmitters for re-inserting a controlled amount of carrier in the output signal (sidebands) subsequent to the modulation process, where the carrier was suppressed. This process insures the proper level of carrier in the transmitter output for reduced pilot carrier or controlled carrier transmission.

channel A limited band of frequencies, or a separate circuit, for conveying a signal so as to be isolated from other signals in the same system or frequency range.

coherent detector A form of synchronous, or phase-locked, detector used in double-sideband (DSB) suppressed-carrier receivers.

combiner A unit used for combining the outputs of diversity receivers in diversity systems, or for combining phone, telegraph, teletype, facsimile, or data signals in any of such systems.

combining network A network for vectorially combining the outputs from two or more separate circuits.

comparator A type of a-f-c circuit which compares the incoming pilot carrier with the signal from the local reference oscillator in the receiver, and supplies an error signal which may be used as an oscillator frequency correction voltage.

compatibility The ability of a transmitted signal to be received by more than one specific type of receiver, or the ability of a transmitter to transmit such a signal, or of a receiver to receive more than one specific type of transmission. This is generally referred to in the use of SSB or DSB equipment operating in conjunction with existing conventional AM equipment.

compatible single sideband (CSSB) A single-sideband transmission in which the full carrier is transmitted, so that the signal may be received on conventional AM receivers. The carrier is maintained at a level of from 4 to 6 db below the peak power output of the transmitter.

conditioned carrier (reconditioned carrier, exalted carrier, or enhanced carrier) The received carrier after being separated from the sidebands, filtered, amplified, and otherwise prepared for insertion or combining with the sidebands for demodulation purposes.

controlled-carrier SSB A single-sideband transmission in which the carrier rises to approximately full amplitude during brief pauses in speech, between syllables, or when no modulation is present, and is reduced to a very low level during actual modulation. The level of the controlled carrier is such that the average power output of the transmitter is maintained effectively constant with or without modulation.

cross modulation A type of distortion produced by modulation of the desired carrier (or a component of its sidebands) by the sidebands of an adjacent or other undesired carrier or signal.

crystal-lattice filter A wave, or bandpass, filter employing piezoelectric (quartz) crystals in a lattice network arrangement.

doppler shift The change in frequency of a signal due to the motion of the transmitter toward or away from the receiver, or the change in frequency due to the change in the length of the transmission path, as in aircraft and vehicular communications.

double conversion The application of two steps of frequency conversion (translation) in a superheterodyne receiver, producing two consecutive intermediate frequencies prior to the actual demodulation, or third conversion, step. Also applicable to a transmitter using two steps of frequency translation.

double sideband (DSB) The transmission of both sidebands. Normally accepted as being a suppressed-carrier transmission with the same information on both sidebands.

dual-channel SSB Actually a double-sideband transmission with different information on either sideband. A two-channel system, each channel being a single-sideband transmission; one channel occupies the upper sideband, and the other the lower sideband. Two unrelated single-sideband transmissions on opposite sidebands of a common carrier.

electromechanical filter A filter in which electrical energy to be filtered is transduced (changed) into mechanical energy, and the filtering process is performed by a mechanical device, such as a specially designed metal rod, disk, or plate, all or portions of which vibrate at the specific frequency, or band of frequencies, to be coupled or transduced into another electrical, or output, circuit.

electromechanical transducer A device so constructed that energy from an electrical system is coupled to a mechanical system, or vice versa. A component of a mechanical filter used in SSB systems. It may take one of four forms: electromagnetic or electrostatic, both of which consist of lumped constants, or magnetostrictive or piezoelectric, both of which consist of distributed constants.

envelope detection Demodulation by rectification of the complete wave envelope (carrier and sidebands).

even-order products Distortion components resulting from heterodyne action between odd or even harmonics of one input signal and the fundamental or harmonic of the second input signal, or separate components of the same signal, to the second-order, fourth-order, etc.

exalted carrier The received carrier, or locally generated carrier, amplified separately to a high level and maintained at that level independent of fading of the received signal or sidebands. It is used to decrease the effects of selective fading and detector distortion in detectors of AM, PM, and SSB receivers. (Conditioned carrier, reconditioned carrier, enhanced carrier, syn.)

fading Decreasing (and increasing), or variations in amplitude of an r-f signal at a receiver antenna due to in- and out-of-phase conditions (vectorial addition) of the signal components received from two (or more) paths of transmission in the atmosphere, one or more paths of which differ in length, causing a time or phase difference at the received point. Usually a constantly changing phenomenon.

forward-acting avc An automatic volume control, which acts on circuits (amplifiers) subsequent to the point of input to the a-v-c circuit in the receiver. Usually used in combination with conventional avc as additional compensation.

frequency conversion (translation) Heterodyning (mixing) of a signal of one frequency (or frequencies) with another signal of a different frequency to produce sum and difference frequencies, one of which is the desired frequency.

frequency discrimination Detecting changes, or variations, in frequency.

frequency distortion A form of distortion; impairment of the fidelity of a signal as a result of the unequal transfer of frequencies, or unequal amplification of frequencies, within the passband of an amplifier.

frequency division (multiplex) Process of transmitting two or more information-bearing signals over a common path by using a different frequency band for the transmission of each signal.

frequency synthesis Producing a signal frequency by heterodyning and otherwise combining frequencies not necessarily harmonically related to each other or the frequency produced.

frequency translation (conversion) Moving a signal or channel to another portion of the frequency spectrum by heterodyning. The sum or difference frequency of the original signal and the heterodyning signal will be the desired channel at another point in the spectrum.

full-carrier SSB (See compatible SSB.)

harmonic distortion Impairment of fidelity caused by the generation of new frequencies that are harmonics of the frequencies contained in the applied signal.

high-level modulation Modulation produced at a point in a system where the power level approximates that at the output of the system.

intelligence The message to be transmitted, such as speech, music, data, etc, or its code equivalent.

intermodulation (distortion) Impairment of fidelity resulting from the production of frequencies that are the sum of, or the difference between, frequencies contained in the applied waveform or channel.

linear amplifier An amplifier that develops an output directly proportional in amplitude to that of the input signal; for example, a Class A or B amplifier. Usually the term "linear amplifier" is used in connection with tuned amplifiers.

linear detection Detection producing an audio output directly proportional in amplitude to the variations of the r-f input.

lower sideband (LSB) The spectrum of the modulating signal displaced by an amount equal to the carrier frequency, the frequency of which is the difference between the carrier and the modulating signal; hence, the lower sideband is located on the lower side of the carrier and inverted, and its bandwidth is the same as the bandwidth of the modulating signal.

low-level modulation Modulation produced at a point in a system where the power level is low compared with the power level at the output of the system.

magnetostrictive transducer An electromechanical transducer in which the magnetic field produced by the electrical (input) signal causes the transducer to expand and contract by magnetic coupling, or vice versa (in the case of the electrical output circuit). A distributed-constant system ideally suited for mechanical filters in SSB applications.

multi-channel More than one branch or path over which signals may be transmitted.

multiple conversion More than one frequency-conversion step in a transmitter or superheterodyne receiver.

multiplexing (multiplex transmission) The simultaneous transmission of two or more channels of intelligence over a single circuit or path, or the preparation of the intelligence for such transmission.

odd-order products Distortion components resulting from heterodyne action between odd or even harmonics of one input signal and the fundamental or harmonic of the second input signal, or separate components of the same signals to an odd order, such as third order, fifth order, etc.

passive network A network which receives all of its operating power from the input signal.

peak-envelope power (PEP) The r-m-s power developed during the peak r-f cycle occurring in the transmitter. Equal to the sum of the amplitudes of the sideband components and the (pilot) carrier.

peak-sideband power (PSP) The r-m-s power developed during the peak r-f cycle occurring in the transmitter without (or minus) the carrier. Equal to the sum of the amplitudes of the sideband components only.

phase-difference network A phase-shift network which establishes a given phase difference between two points, regardless of the actual phase shift of the network itself.

phase discrimination Detecting changes, or variations in phase.

phase discrimination multiplex (Day's system). Process of transmitting two or more information-bearing signals by means of channels superimposed within the same frequency band, but shifted in phase with respect to each other by a predetermined phase angle (90 degrees).

phase distortion Impairment of fidelity due to nonlinear phase characteristics, which cause various frequencies of an applied waveform to be delayed disproportionately.

phase-locked receiver A receiver in which the local oscillator is synchronized and maintained in phase with the received carrier (or its equivalent in the case of suppressed-carrier reception).

phase modulation Variation of the phase of an r-f signal in accordance with the intelligence to be transmitted. A form of angle modulation.

phase-shift network A network in which an applied signal is shifted in phase by predetermined angle.

pilot carrier A reduced carrier, the amplitude of which is maintained at a level of from 10 to 20 db below the peak-sideband power. A carrier often used in SSB transmissions as a reference for a-v-c and a-f-c systems in receivers.

post-phasing distortion A form of distortion due to harmonics generated in the audio amplifiers following the audio phase-difference networks, or due to improper adjustment of the balanced modulators in phase-shift SSB transmitters.

product detector (demodulator) A detector, the amplitude of the output (audio) of which is proportional to the mathematical product of the amplitudes of both of its two inputs (carrier and sidebands).

product modulator A modulator, the amplitude of the output of which is proportional to the mathematical product of the amplitudes of both of its two inputs (carrier and modulating signals). A form of balanced modulator in that the carrier is normally suppressed.

pulse modulation Either the modulation of an r-f signal by a sequence (train) of pulses, or the modulation of pulses by variation of one or more parameters of the series of regular recurrent pulses, such as pulse amplitude (PAM), pulse time (PTM), pulse duration (PDM), pulse position (PPM), or pulse code (PCM).

quadrature input The use of two separate input signals, one of which is shifted in phase 90 degrees with respect to the other.

quadrature modulation The phase-shift method of modulation using two of each of the input signals in quadrature phase relationship with each other. Normally two balanced modulators are used in SSB applications.

reconditioned carrier (conditioned carrier, exalted carrier, enhanced carrier) The received carrier after separation from its sidebands, filtering, and amplification, used for insertion with the sidebands for demodulation.

reduced-carrier SSB **(pilot-carrier** SSB**)** A single-sideband signal with a reduced or pilot carrier, in which the carrier is reduced 10 to 20 db below the peak sideband amplitude.

reference oscillator An oscillator used for automatic-frequency-control reference in SSB transmitters and receivers. Usually a low-frequency oscillator, and ordinarily serves the dual function of frequency control and carrier insertion.

roofing filter (bridging filter) See bridging filter.

sidebands The frequency bands on either or both sides of the carrier frequency, or the components of the signal within such bands, which are the sum (upper sideband) and difference (lower sideband) frequencies produced by modulation of the carrier with the modulating signal.

sideband channel A channel, or signal path, within a single- or double-sideband transmitter or receiver, restricted to the frequencies of either (or both) sidebands without the carrier.

sideband filter A filter designed to pass the sideband frequencies with little or no attenuation and to reject all other frequencies, especially those adjacent to the sideband frequencies, and the carrier and opposite sideband.

side frequency One of the frequencies of a sideband. The sum or difference frequency resulting from the modulation of a carrier with a single tone.

signal-to-distortion ratio The ratio of the amplitude of the desired signal to that of the distortion products, usually expressed in db.

single-sideband exciter A unit containing all the frequency-generating components and modulation components of a single-sideband transmitter. The output of the exciter is the desired signal, suitable for direct radiation, except that the power level is relatively low. The output is normally applied to power amplifiers, suitable for amplifying the signal to the desired amplitude of transmission, or desired power output.

single-sideband modulation (SSB) A form of amplitude modulation in which one sideband and the carrier are suppressed.

single-sideband suppressed carrier (SSSC) A single-sideband signal in which the carrier is (ideally) completely suppressed. Normally the carrier is suppressed to a level at least 40 db below the peak sideband power level.

slaved oscillator An oscillator controlled by, or synchronized with, another oscillator or oscillatory circuit.

splatter A term used to define the distortion products of a transmitter that fall outside of the frequency limits of the desired transmission.

stabilized master oscillator (SMO) A variable-frequency oscillator, in a special circuit employing crystal frequency synthesis, providing a multiple number of stable channel frequencies.

stability The degree to which a specified frequency may be maintained. Usually expressed as a tolerance in percent.

square-law detection Detection which produces an output proportional to the square of the input, or inputs.

synchronous reception Reception of signals by the use of a phase-locked oscillator in a special receiver circuit. Ideally suited for DSB reception, or synchronized AM (SAM).

time-division multiplex The simultaneous transmission of two or more channels of intelligence over a single circuit or path by using different time intervals for the transmission of each channel of intelligence.

two-tone test A test involving the application of two separate tone signals, equal in amplitude and having a difference in frequency of about 1000 cps, to the input of a system or circuit and observing the results on an oscilloscope, spectrum analyzer, or other indicating device.

unbalanced detector A detector in which no component of the input signals is cancelled, or the circuit of which is unbalanced.

upper sideband The spectrum of the modulating signal displaced by an amount equal to the carrier frequency without inversion, the frequency of which is the sum of the carrier and modulating signal; hence, the upper sideband is located on the upper side of the carrier frequency and its bandwidth is the same as the bandwidth of the modulating signal.

vestigial sideband A form of single-sideband transmission wherein one sideband and that portion of the opposite sideband nearest the carrier frequency are transmitted. That portion of the sideband so transmitted is termed the "vestigial sideband." Filtering requirements are eased in this type of sideband transmission, and very low frequencies (down to zero) may be transmitted without loss. Typical applications are television and pulse systems.

wide-band phase-shift network A network designed to provide a uniform phase shift over a wide band of frequencies.

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