SINGLE SIDEBAND RADIO COMMUNICATIONS

TRAINING COURSE

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Equipment Support Division
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Training Course: Single Sideband Radio Communications

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To meet current requirements within the Pacific Missile Range, this training course has been revised by the Technical Data and Documentation Section (33332) in cooperation with Mr. R. F. Chezum of the Maintenance Engineering Section (33333). In the preparation of this revised edition, some of the original material has been deleted, some new data has been added, certain sections have been rewritten for improved clarity, and errors of fact or typography have been corrected.

Requests for further technical information or for additional copies of the training course material should be addressed to the Technical Data and Documentation Section, PMR Code 33332. Inquiries in regard to on-site training classes or training by correspondence should be directed to the Employee Development Division, PMR Code 174.

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FOREWORD

This training course on single sideband radio communications has been prepared to meet specific requirements within the Pacific Missile Range. The original version was developed in 1962 by personnel of the Maintenance Engineering Section, PMR Code 33333, with a portion of the text and illustrative material being contributed by the Collins Radio Company as part of the services rendered the Pacific Missile Range under field service contract No. N123(61756)30702A.

The training course has been slanted intentionally, by selection and organization of subject matter, so as to meet most nearly the needs of PMR mechanics and technicians in the operation and maintenance of the specific models and types of single sideband radio equipment in use throughout the Pacific Missile Range. As new types of equipment are procured, and as operating and maintenance parameters change, pertinent sections of the training course will be revised and updated to keep pace.

D. B. Wright

Head, Equipment Support Division

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SECTION I

INTRODUCTION TO SINGLE SIDEBAND COMMUNICATIONS

1-1. Background.

1-2. As an art progresses and becomes more sophisticated, and as its applications become more diversified, more efficient utilization tends to be made of its principles in order to conserve material, space, and time. This development can readily be traced in the improvement of the steam engine and steam turbine, in the generation and utilization of electric power, and in practically every other line of human endeavor. Hence it is not strange to find similar progress being made in the art of radio and wire communications. Early developments were in the realm of transmission media. The early use of single-wire telegraph lines eventually progressed to the use of two-wire telegraph and telephone lines, the introduction of loading coils and of uniformly loaded cable, and, during the latter period of wire line and cable development, to the utilization of electromagnetic radiation in space, called radio.

1-3. The message space occupied by a communications system can be measured in terms of bandwidth. The number of messages that can be accommodated in a communications medium is equal to the total useful bandwidth of the medium divided by the bandwidth required by each message channel. To gain more message channels consideration was given at first to extending the use of radio equipment to higher frequencies within the electromagnetic spectrum. Advances were made in the short-wave region (through the high-frequency range), then in the VHF and UHF ranges and, finally, in the microwave region.

1-4. Because the upper frequency limit for long-distance radio communications is about 30 Mc, which places a limit on the available channel space for this service, attention has been focused on required channel bandwidth and methods for reducing bandwidth for a given message. Two sidebands and a carrier are generated in amplitude-modulated (AM) radio systems. Since either sideband alone contains all the information necessary for transmission of intelligence, it is a waste of power and spectrum space to transmit both of the sidebands and the carrier. For this reason a method has been devised for transmitting one sideband only. This system is known as "single sideband" (SSB) transmission. SSB systems have been found not only to conserve power and spectrum space but also to pay unexpected dividends in greater freedom from fading in long-distance transmissions involving multi-path conditions.

Because of these advantages, SSB has been adopted by the armed forces and may be expected to supersede other methods of communications in many applications.

1-5. Origin and Historical Development of Single Sideband Communications.

1-6. A widespread belief exists that SSB techniques are of recent origin. Actually, the basic principles have been recognized and used in various commercial applications for well over 30 years. The main reason for the delay in the large-scale development of SSB equipment was the lack of compact and economical high-precision components. The development of effective lattice filters, mechanical filters, and the phasing method of generating the SSB signal now makes possible the economical, wide-scale use of SSB equipment. Constant research on and development of both SSB circuits and components have resulted in extremely compact and highly efficient communications systems which would not otherwise have been possible. It is reasonable to expect that future developments, particularly in solid-state electronics, will bring forth even more compact communications equipment of greater efficiency than the units available today.

1-7. Although SSB transmission has only received publicity in recent years, knowledge, development and use of SSB techniques have progressed over the last 40 years. The acoustical phenomenon of combining two waves to produce sum and difference waves carries over into electric wave modulation. The presence of the upper and lower sidebands, in addition to the carrier frequency, were assumed to exist but were not concretely visualized in the earliest modulated transmissions. Recognition that one sideband contained all the signal elements necessary to reproduce the original signal came in 1915. In that year, at the U. S. Naval Radio Station, Arlington, Virginia, an antenna was tuned so as to pass one sideband well and to attenuate the other.

1-8. From 1915 to 1923 the physical reality of sidebands was vigorously argued, with the opponents contending that the sidebands were mathematical fiction. However, the first transatlantic radiotelephone demonstration in 1923 provided a concrete answer. Single sideband was used in this system because of the limited power capacity of the equipment and the narrow resonance bands of efficient antennas at the low frequency used (57 Kc). By 1927, transatlantic SSB radiotelephony was available as a regular public service.

1-9. The first, low-frequency overseas radio systems were followed by high-frequency systems (in the range of 3 to 30 Mc) which transmitted double sideband and carrier (AM) because SSB development did not permit practical SSB transmission in this frequency range. However, SSB techniques were employed in various telephony applications and multiplexing systems. It has been only recently that equipment development has permitted the advantages of SSB communications to be fully exploited. These developments have been in the fields of frequency stability, filter selectivity, and low-distortion linear power amplifiers and have led to wide military and commercial acceptance of high-frequency SSB communications systems. There are presently available many amateur and commercial SSB radio sets, fixed-station SSB exciters, airborne transceivers, and linear power amplifiers of up to 45 kilowatts. Modern SSB systems are capable of providing reliable radio communications over almost unlimited distances. Some of these equipments, especially the military types, are provided with automatic frequency selection and automatic tuning to further enhance their value as reliable, easily operated systems.

1-10. Carrier-System Telephony.

1-11. The first application of SSB techniques (still most important) was in connection with carrier-system voice transmission over long-distance telephone lines, coaxial cables, and microwave radio links. The principle reason for the development and subsequent use of SSB techniques was economy. In carrier systems, voice frequencies between 300 and 3000 cycles are used to modulate an RF carrier which is usually in the region of 20 to 50 Kc. In the modulation process, additional radio frequencies are created and appear in the composite signal as sidebands. In a very simple carrier telephone system (figures 1-1 and 1-2) the RF carrier and its sidebands are fed into a telephone line, coaxial cable, or microwave system for transmission to a distant point. At the receiving end, the composite RF signal is demodulated and the recovered voice frequencies are fed to local telephone circuits.

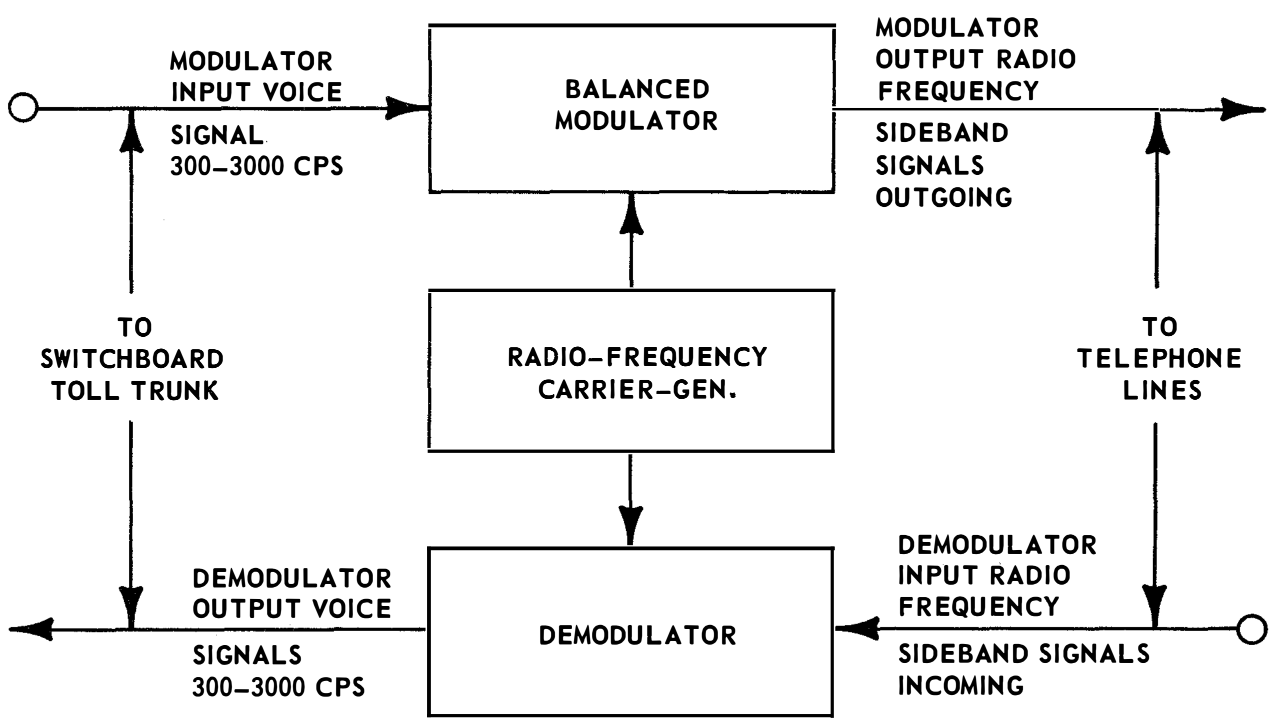


Figure 1-1. Simplified Carrier Telephony System.

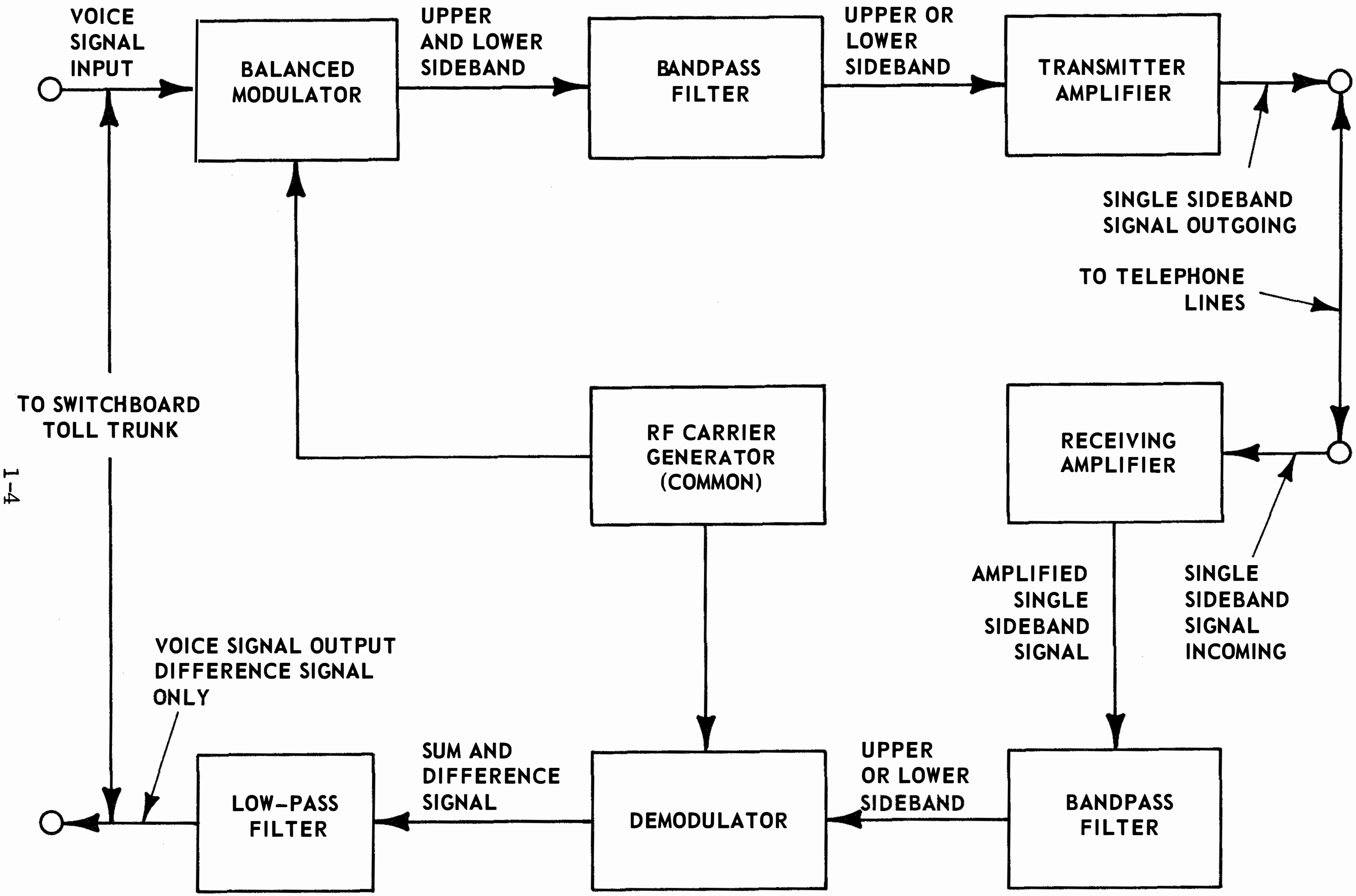


Figure 1-2. Essential Components in a Modern Carrier Telephony System.

1-12. In actual practice, the modern carrier telephony system is much more complex than would appear from the preceding description. The system is similar to ordinary radio broadcasting except that the frequencies are much lower and are transmitted over telephone lines, coaxial cables, or microwave links. Neglecting any consideration of SSB techniques it is apparent that, by careful selection of carrier frequencies spaced at regular intervals throughout the bandpass spectrum of the telephone system, a number of simultaneous long-distance telephone conversations can take place on one circuit without interfering with each other. The number of conversations which can be transmitted over a single circuit is limited only by the channel space required for each conversation, the overall bandpass of the system, and the ability of the receiving equipment to select any one voice channel and reject the others. Present techniques make it possible to provide a frequency band 8 Mc wide, so that two coaxial cables or a microwave complex can handle either 1800 telephone channels or 600 telephone channels and a 4.2 Mc television channel in each direction.

1-13. For many years, the range of telephony was severely limited by the loss of energy along the wires. After the invention of the vacuum tube in 1906, coupled with subsequent research and development, the telephone repeater (amplifier) became available. Through the use of repeater amplifiers, telephone service between New York and San Francisco was inaugurated in 1915. Today, hundreds of thousands of repeaters are in use, making long-distance telephony possible to all parts of the world.

1-14. In the early days of carrier system development, it was discovered that transmission of carrier components overloaded the repeater amplifiers and caused undesirable cross-modulation. Up to that time, the exact nature of amplitude modulation was not thoroughly understood. During the early period of telephone circuit development, however, it was mathematically demonstrated that, along with the carrier, two sidebands existed which contained identical information and were, in fact, mirror images of each other. It was also found that the carrier itself contained no useful information but was used merely as a reference signal for recovery of the sideband intelligence. Needless to say, these conclusions were the subject of much debate at the time and, until a few years ago, it was customary to omit any reference to sideband theory in standard communications manuals. In spite of the controversy, subsequent work by engineers and scientists has proven these theories to be true. As a result, a comprehensive body of literature on theory, techniques, and components of SSB telephony has accumulated over the years. While these papers are of considerable interest from the historical point of view, in most instances they are highly mathematical and deal with extremely complex telephone equipment which cannot be readily duplicated outside a laboratory.

1-15. Except for using much lower frequencies, the early SSB signals transmitted over telephone lines were substantially the same as the signals transmitted by modern military, commercial and amateur SSB transmitters. In modern SSB telephone circuits, these transmission frequencies may be in hundreds of megacycles, although the original carrier and sidebands are sometimes below 100 Kc. It is common practice today to employ SSB techniques to multiplex a large number of voice channels over a band of radio frequencies in the microwave region. The sideband signal is generated at a very low power level, in the low-frequency portion of the RF spectrum. It is then heterodyned, through a number of translator stages, to the transmission frequency. The signal is then amplified to the desired power level in a system of linear amplifiers. It is transmitted by a coaxial cable system and/or radiated through space by a highly directional radio antenna.

1-16. Need for Single Sideband Systems.

1-17. The need for SSB communications systems in the high-frequency spectrum has arisen because present-day radio communications require more effective spectrum-conservation systems. The quantity of commercial and military radio traffic is presently so great in the high-frequency portion of the spectrum (3 to 30 Mc) that it has become necessary to restrict the use of this part of the spectrum to those services which cannot be accommodated by any other means. Landlines, microwave links, and UHF scatter propagation are employed to reduce the load in the high-frequency spectrum. In many instances, these methods provide a better and more reliable service. There are, however, many communications services which require the propagation characteristics obtainable only in the high-frequency range. Among these are ship-to-shore communications, air-to-ground communications, and many naval and other military systems which require independence, mobility, and flexibility. Since high-frequency spectrum space is limited, it is essential that the best possible use be made of the space available. This means that communications systems must use minimum bandwidth, that the width of guard bands between channels to allow for frequency drift and poor selectivity are minimized, and that spurious radiation be kept to a very low level to avoid interference between services. In addition, a more reliable signal is desirable, if not essential. Single sideband communications systems, in their present state of development, provide these assets.

1-18. What Single Sideband Means.

1-19. A single sideband signal is an audio signal converted to a radio frequency, with or without inversion. For instance, an intelligible voice signal contains audio frequencies over the range of 300 to 3000 cps. If this audio signal is converted to a radio frequency by mixing it with a 15 Mc RF signal, the resultant sum frequencies cover the range of 15,000,300 to 15,003,000 cps. Such a signal is an SSB signal without inversion and is referred to as an "upper sideband" because it occupies the spectrum space above the RF conversion frequency. Note that the 15 Mc carrier is not included in the range of the SSB signal. This example does not indicate the presence of a difference frequency. When the voice signal is mixed with the RF, however, a difference frequency is developed which covers the range from 14,999,700 to 14,997,000 cps. This signal is also an SSB signal, but is inverted. This signal is referred to as a "lower sideband" signal because it occupies the spectrum space below the RF conversion frequency. Figure 1-3 illustrates the position of an upper sideband SSB signal in the RF spectrum.

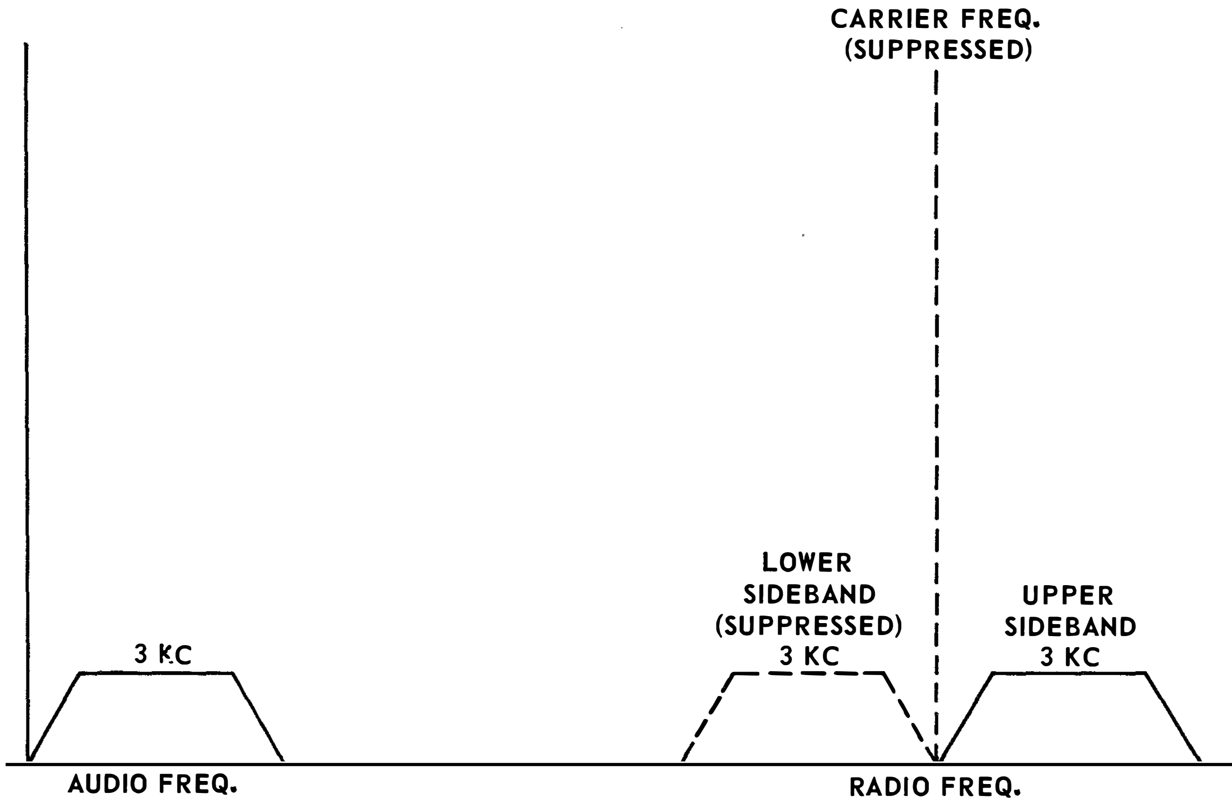


Figure 1-3. Location of an Upper Sideband Signal.

From the preceding description of the SSB signal, it is apparent that only one sideband need be transmitted to convey intelligence. Since two sideband signals are developed in the mixing process, it is necessary to remove one sideband before transmission. In receiving an SSB signal, it is necessary to convert it back to the original audio. This requires identical transmitter and receiver conversion frequencies. In the past, a low-level pilot carrier was transmitted for automatic frequency control (AFC) purposes. With present-day frequency stabilities (1 cps at 10 Mc in ground equipment and 10 cps at 10 Mc in mobile equipment), the need for AFC and pilot carriers is eliminated.

1-20. Several methods of sideband communications are in use or under development. The SSB method (the term used throughout this training course) refers to the method which is more accurately called "single sideband, suppressed carrier." In this method, only one sideband is transmitted and the carrier is suppressed. Demodulation of the SSB signal requires that the received signal be mixed, or heterodyned, with a locally generated signal very close to the proper frequency, with no particular phase relationship being necessary. In the SSB pilot carrier system only one sideband is transmitted, but a low-level carrier of sufficient amplitude for reception is also transmitted. To demodulate this signal, the pilot carrier is separated from the sideband in the receiver, amplified, and used as the conversion frequency to demodulate the sideband signal. In another receiving method, the pilot carrier is used for automatic frequency control of the receiver. In the double sideband (DSB) system, both the upper and lower sidebands are transmitted, with the carrier suppressed. Demodulation of the double sideband signal requires insertion of a locally generated carrier of proper frequency and proper phase. This system depends upon an automatic frequency and phase control, derived from the DSB signal, for control of the locally generated carrier. In the single sideband, controlled carrier system, only one sideband is transmitted but a carrier which varies in level inversely with the signal level is also transmitted. This allows an appreciable average carrier level for automatic frequency control without reducing the sideband power below the full transmitter rating.

SECTION II

COMPARISON OF SSB WITH AM/FM SYSTEMS

2-1. Power Comparison, SSB and AM.

2-2. There is no single method of evaluating the relative performance of AM systems and SSB systems. Perhaps the most straightforward manner in which to make such a comparison is to determine the transmitter power necessary to produce a given signal-to-noise (s/n) ratio at the receiver for the two systems under ideal propagation conditions. Signal-to-noise ratio is considered a fair comparison because it is the s/n ratio which determines the intelligibility of the received signal.

2-3. Figure 2-1 shows such a comparison between an AM system and an SSB system, where 100 percent single-tone modulation is assumed. Figure 2-1 shows the power spectrum for an AM transmitter rated at 1 unit of carrier power. With 100 percent sine wave modulation, such a transmitter will actually be producing 1.5 units of RF power. There is 0.25 unit of power in each of the two sidebands and 1 unit of power in the carrier. This AM transmitter is compared with an SSB transmitter rated 0.5 unit of peak envelope power (PEP). PEP is defined as the power developed at the crest of the modulation envelope. The SSB transmitter rated at 0.5 unit of PEP will produce the same s/n ratio in the output of the receiver as the AM transmitter rated at 1 unit of carrier power.

2-4. The voltage vectors related to the AM and SSB power spectrums are shown in figure 2-1B. The AM voltage vectors show the upper and lower SSB voltages of 0.5 unit rotating in opposite directions around the carrier voltage of 1 unit. For AM modulation, the resultant of the two sideband voltage vectors must always be either directly in phase, or directly out of phase with the carrier, so that the resultant directly adds to or subtracts from the carrier. The resultant shown on the upper and lower sideband voltages are instantaneously in phase to produce a peak envelope voltage (PEV) equal to twice the carrier voltage with 100 percent modulation. The 0.5 unit of voltage shown in each sideband vector produces the 0.25 unit of power shown in figure 2-1A, the 0.25 unit of power being proportional to the square of the 0.5 unit of voltage. The SSB voltage vector is a single vector of the 0.7 unit of voltage at the upper sideband frequency. The 0.7 unit of voltage produces the 0.5 unit of power shown in figure 2-1.

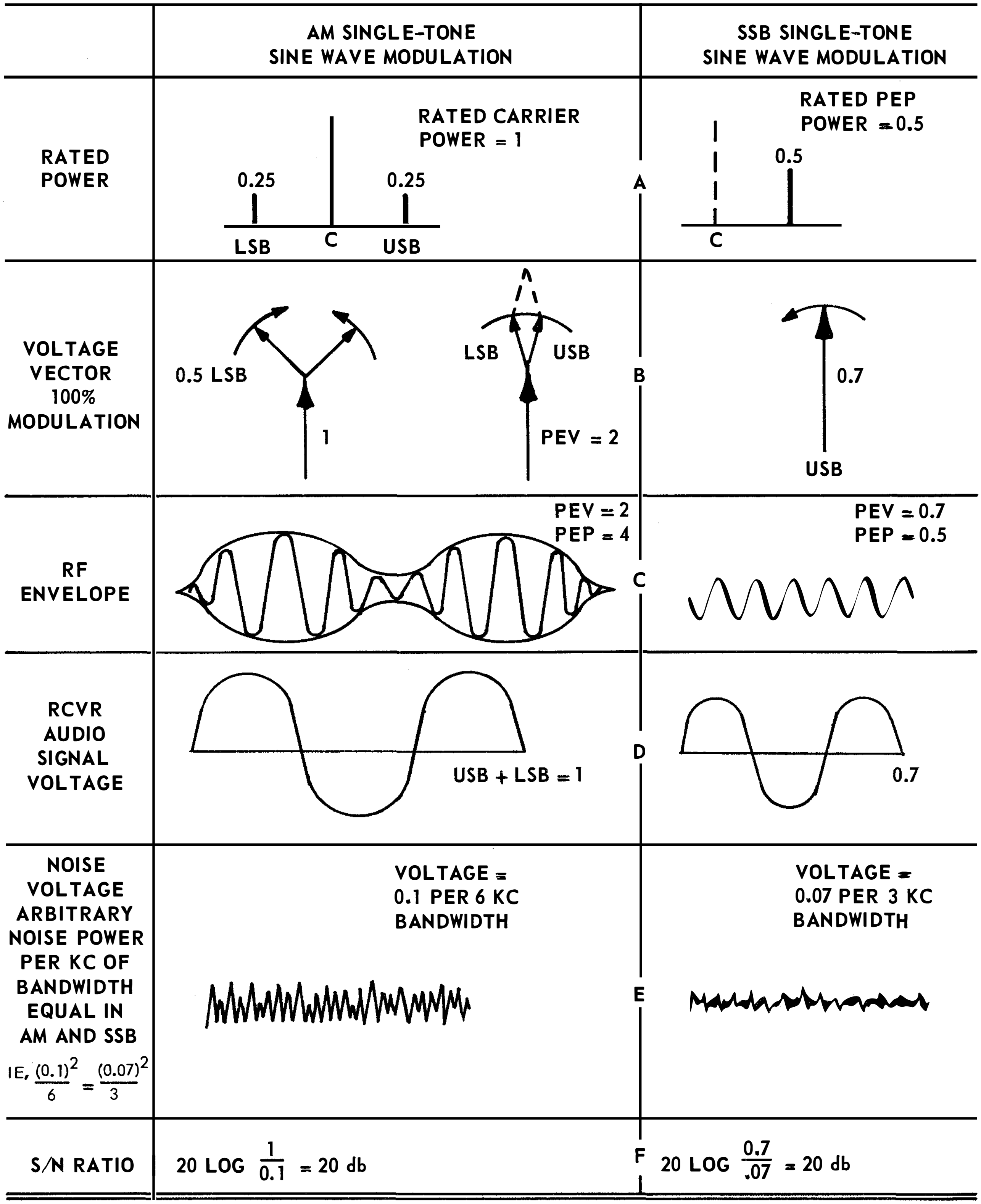


Figure 2-1. Comparison of SSB and AM With Equal S/N Ratio.

2-5. The RF envelopes developed by the voltage vectors are shown in figure 2-1C. The RF envelope of the AM signal is shown to have a PEV of 2 units, the sum of the two sideband voltages plus the carrier voltage. This results in a PEP of 4 units of power. The PEV of the SSB signal is 0.7 unit of voltage with the resultant PEP of 0.5 unit of power.

2-6. When the RF signal is demodulated in the AM receiver, as shown in figure 2-1D, and audio voltage develops which is equivalent to the sum of the upper and lower sideband voltages (in this case, 1 unit of voltage), this voltage represents the output from the conventional diode detector used in AM receivers. Such detection is called coherent detection because the voltages of the two sidebands are added in the detector. When the RF signal is demodulated in the single sideband receiver, an audio voltage of 0.7 unit develops which is equivalent to the transmitter upper sideband signal. A signal is demodulated by heterodyning the RF signal with the proper frequency to move the SSB signal down in the spectrum to its original audio position.

2-7. If a broad-band noise level is chosen as 0.1 unit of voltage per 6-Kc bandwidth (the AM bandwidth), the same noise level is equal to 0.07 unit of voltage per 3-Kc bandwidth, the SSB bandwidth. This is shown in figure 2-1E. These values represent the same noise power level per Kc of bandwidth, that is, 0.1 squared over 6 equals 0.07 squared over 3. (With this chosen noise level, the s/n ratio for the AM system is twenty times the log of the s/n in terms of voltage or 20 db, the s/n ratio for the SSB system is also 20 db, the same as for the AM system). The one-half power unit of rated PEP for the SSB transmitter therefore produces the same signal intelligibility as the 1 power-unit rated carrier power for the AM transmitter. This conclusion can be restated as follows:

> Under ideal propagating conditions, but in the presence of broad-band noise, the SSB and AM systems perform equally if the total sideband power of the two transmitters is equal. This means that an SSB transmitter will perform as well as an AM transmitter of twice the carrier power rating under ideal propagation conditions.

2-8. Antenna Voltage Comparison, SSB and AM.

2-9. Of special importance in airborne and mobile installations where electrically small antennas are required is the peak antenna voltage. (In these installations it is often the corona breakdown point of the antenna which is the limiting factor in equipment power.) Figure 2-1C shows the RF envelopes of an SSB transmitter and an AM transmitter of equal performance under ideal conditions. The PEV produced by these two transmitters is shown to be in the ratio of two for the AM transmitter to 0.7 for the SSB transmitter. This indicates that, for equal performance under ideal conditions, the peak antenna voltage of the SSB system is approximately one-third that of the AM system.

2-10. A comparison between the SSB power and the AM power which can be radiated from an antenna of given dimensions is even more significant. If an antenna is chosen which will radiate 400 watts of PEP, the AM transmitter which may be used with this antenna must be rated at no more than 100 watts.

2-11. Advantage of Single Sideband Under Selective Fading Conditions.

2-12. The power comparison between SSB and AM given in the previous paragraph is based on ideal propagation conditions. With long distance transmission, however, AM is subject to selective fading which causes severe distortion and a weaker received signal. At times, this can make the received signal unintelligible. An AM transmission is subject to deterioration under these poor propagation conditions because all three components of the transmitted signal, the upper sideband, lower sideband, and carrier, must be received exactly as transmitted in order that fidelity and full theoretical power from the signal may be realized.

2-13. The loss of one of the two transmitted sidebands results only in a loss of signal voltage from the demodulator. Even though some distortion results, such a loss is not basically detrimental to the signal, because, as pointed out before, one sideband contains the same intelligence as the other. Since the AM receiver operates on the broad bandwidth necessary to receive both sidebands, the noise level remains constant even though only one sideband is received. This is equivalent to a 6 db deterioration in s/n ratio out of the receiver. Although the loss of one of the two sidebands could be an extreme case, the proportional deterioration in s/n ratio results from the reduction in the level of one or both sidebands.

2-14. The most serious result of selective fading, and one occurring most commonly, is attenuation of the carrier level more than the sidebands. When this occurs, the carrier voltage at the receiver is less than the sum of the two sideband voltages; with the carrier attenuated more than the sidebands, the RF envelope does not retain its original shape and distortion is extremely severe upon demodulation. This distortion results in intermodulation because a carrier voltage at least as strong as the sum of the two sideband voltages is required to properly demodulate the signal. The distortion resulting from a weak carrier can be overcome by use of the exalted-carrier technique, whereby the carrier is amplified separately and then reinserted before demodulation. In using the exalted-carrier, the carrier must be reinserted close to the original phase of the AM carrier.

2-15. Selective fading can also result in a shift between the relative phase position of the carrier and the sidebands. An AM modulation is vectorially represented by two counter-rotating sideband vectors which rotate with respect to the carrier vector. The result of the sideband vector is always directly in phase or directly out of phase with the carrier vector. In an extreme case, the carrier may be shifted 90° from its original position. When this occurs, the result of the sideband vector is ± 90° out of phase with the carrier vector. This results in converting the original AM signal to a phase-modulated signal. The envelope of the phase-modulated signal bears no resemblance to the original AM envelope and the conventional AM detector will not produce an intelligible signal. Any shift in the carrier phase from its original phase relationship, with respect to the sidebands, will produce some phase modulation with a consequential loss of intelligibility in the audio signal. Such a carrier phase shift may be caused by poor propagation conditions; it will also result from using the exalted-carrier technique if the reinserted carrier is not close to its original phase, as previously mentioned.

2-16. An SSB signal is not subject to deterioration due to selective fading which varies either the amplitude or the phase relationship between the carrier and the two sidebands as in AM transmission. Since only one sideband is transmitted in SSB, the received signal level does not depend upon the result of amplitude of the two sideband signals as it does in AM. Since the received signal does not depend upon a carrier level in SSB, no distortion can result from loss of carrier power. And since the received signal does not depend upon the phase relationship between the sideband signal and the carrier, no distortion can result from phase shift. Selective fading within the one sideband of the SSB system only changes the amplitude and the frequency response of the signal. It very rarely produces enough distortion to cause the received signal or voice to be unintelligible.

2-17. Comparison of SSB With AM Under Limited Propagation Conditions.

2-18. One of the main advantages of SSB transmission over AM transmission is obtained under limited propagation conditions over a long range path where communications are limited by the combination of noise, severe selective fading, and narrow-band interference. Figure 2-2 illustrates the results of an intelligibility study performed by rating the intelligibility of information received when operating the two systems under varying conditions of propagation. The two transmitters, when compared, have the same total sideband power; that is, a 100-watt AM transmitter puts one quarter of its rated carrier power in each of two sidebands, while an SSB transmitter puts its full rated output in one sideband. This study shows that, as propagation conditions worsen and interference and fading becomes prevalent, the received SSB signal will provide up to a 9 db advantage over the AM signal. The result of this study indicates that SSB systems will give from zero to 9 db improvement under various conditions of propagation when total sideband power in SSB is equal to AM. It has been found that 3 db (of the possible 9 db advantage) will be realized on the average contact or communications. In other words, in normal use an SSB transmitter rated at 100 watts PEP will give equal performance with an AM transmitter rated at 400-watts carrier power. It should be pointed out that, in this comparison, the receiver bandwidth is just enough to accept the transmitted intelligence in each case and no speech processing is considered for SSB transmission.

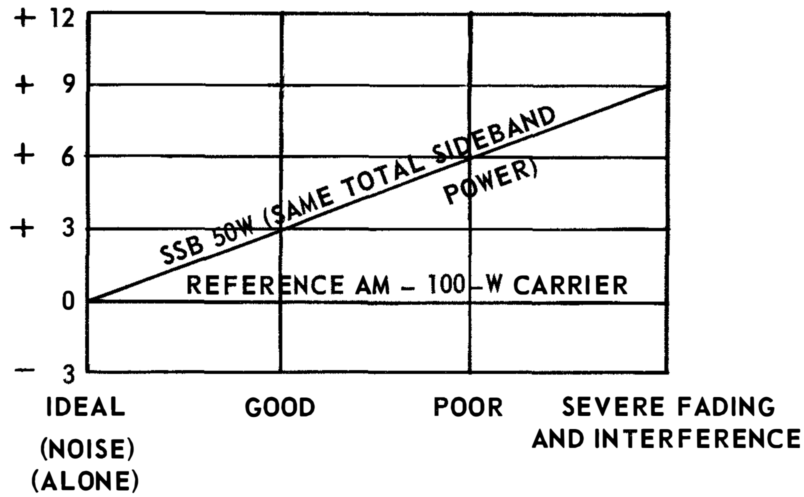


Figure 2-2. Advantage of SSB Over AM, With Worsening Propagation Conditions.

2-19. Comparison of SSB and FM.

2-20. Although much experimental work has been done to evaluate the performance of SSB systems with AM systems, very little work has been done to evaluate the performance of SSB systems with FM systems. Figure 2-3, however, shows the predicted results of one such study based on a mobile FM system as compared to a mobile SSB system of equal physical size. The two systems also use the same output tubes to their full capacity, so that the final RF amplifiers dissipate the same power during normal speech loading. The study is complicated by evaluating the effects of speech processing such as clipping and pre-emphasis with its resulting distortion. Such speech processing is essential in the FM system but has little benefit in the SSB system. Figure 2-3 shows the s/n ratio in decibels on the Y-axis and the attenuation between transmitter and receiver in decibels on the X-axis. This graph indicates that, with 150 to 160 db of attenuation between the transmitter and receiver, a strong signal, the narrow band FM system provides a better s/n ration than the SSB. Under weak signal conditions, from 168 db and more of attenuation between the transmitter and receiver, the s/n ratio of the FM system falls off rapidly and the SSB system provides the best s/n ratio. This fall-out in the FM s/n ratio results when the signal level drops below the level required for operation of the limiter in the FM receiver.

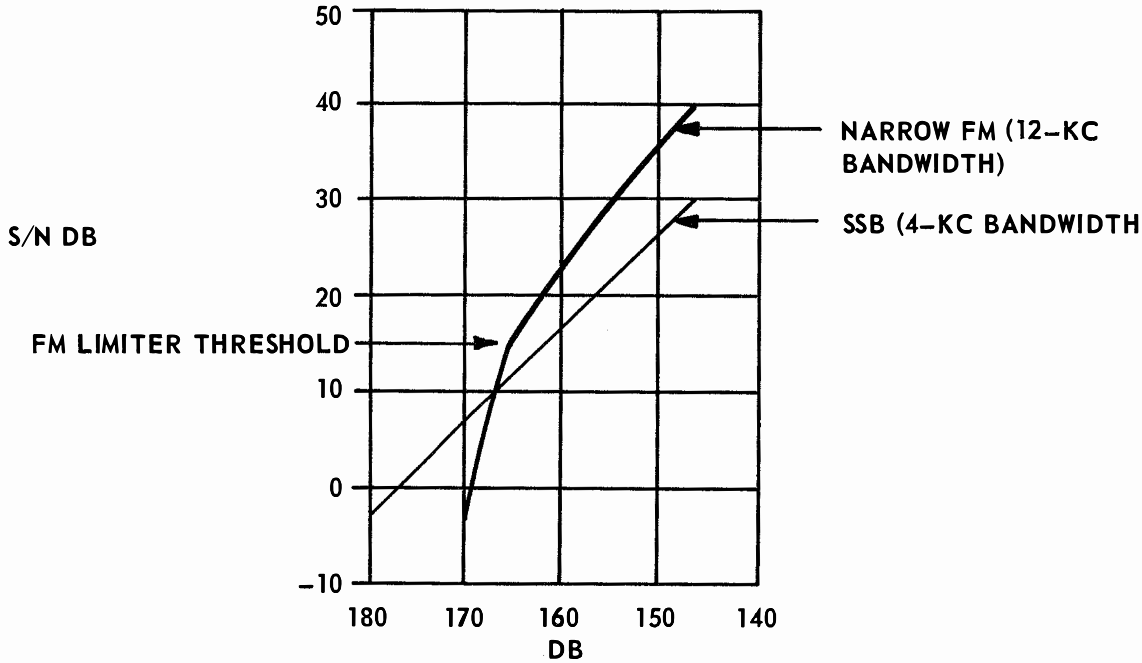


Figure 2-3. SSB Performance Compared With FM.

2-21. The conclusions which can be drawn from figure 2-3 are as follows:

a. For strong signals the FM system will provide a better s/n ratio than an SSB system. However, this is not an important advantage because, when the s/n ratio is high, a still better s/n ratio will not improve intelligibility significantly.

b. For weak signals the SSB system will provide an intelligible signal where the FM system will not.

c. The SSB system provides three times the savings in spectrum space as does the narrow band FM system.

2-22. Derivation of SSB Signals.

2-23. An SSB signal is essentially a modified AM signal from which certain components have been removed. In communications equipment designed for voice transmission, one sideband and the carrier are generally removed. Before we can intelligently discuss SSB theory in practice, however, it will be necessary thoroughly to understand the formation of an ordinary AM signal from which the sideband signal is eventually derived.

2-24. Amplitude Modulation.

2-25. Where the output of a CW transmitter is made to vary in amplitude at a single audio frequency rate, a steady tone will be heard in the receiver tuned to the transmitter frequency. The frequency of the received audio tone will be the same as that of the original audio tone used to modulate the radio frequency output of the transmitter. If instead of a single tone, the modulation consists of a number of tones or audio frequencies such as those of speech or music, then the transmitter output will vary in unison with these frequencies and the sound reproduced at the receiver will correspond to the original voice or music. This process is known as amplitude modulation. The analysis of any modulation process becomes extremely complex when several modulation tones or frequencies are present. The accepted procedure in modulation analysis is to use only one or two tones as the modulating signal. Thus, when a single audio tone is used, the frequency components which appear at the transmitter output are the carrier, the lower side frequency, and an upper side frequency. However, with a complex modulating signal such as speech or music, two bands of frequencies will appear at the output along with the carrier. These two bands are called the lower sideband and the upper sideband. Unlike the lower and upper side frequencies, these lower and upper sidebands consist of many frequencies instead of only one. At this time, however, our discussion will be limited to the side frequencies. The more complex sidebands will be covered later.

2-26. Amplitude modulation is essentially a form of mixing or combining two or more signals in a suitable nonlinear device. When a single audio tone is transmitted by means of amplitude modulation, the tone signal is mixed with a radio frequency carrier to produce a composite signal consisting of the original carrier and two side frequency signals.

2-27. The mixing action can be understood more clearly by the use of a numerical figure in a typical example. Figure 2-4 is a part of the transmitter output frequency spectrum showing the relative characteristics of the carrier and the two side frequency signals. The vertical scale indicates relative voltage (amplitude) in arbitrary units and the horizontal scale indicates frequency in kilocycles. The 5,000-Kc carrier is modulated by pure sine wave tone of 1,000 cycles (1 Kc).

2-28. At 100 percent modulation, two side frequencies will be formed as shown in figure 2-4. One will be located at 4999 Kc, which is equal to the carrier frequency minus the modulation frequency (5000 minus 1 = 4999). The other will appear at 5001 Kc, which is equal to the carrier frequency plus the modulation frequency (5000 + 1 = 5001). The 4999-Kc signal is known as the difference frequency, or lower side frequency and the 5001-Kc signal is the sum frequency, or upper side frequency. Note that the 4999- and the 5001-Kc signal are not referred to as sidebands, since the modulation is a single-frequency audio tone only. At 100 percent modulation the amplitude of each side frequency will be exactly half the carrier amplitude. If a highly selective receiver (one with an overall bandpass of about 500 cycles) is tuned to 4999 Kc the lower side frequency will appear as a steady unmodulated carrier and no tone will be heard. If the signal is removed from the last IF stage of the receiver

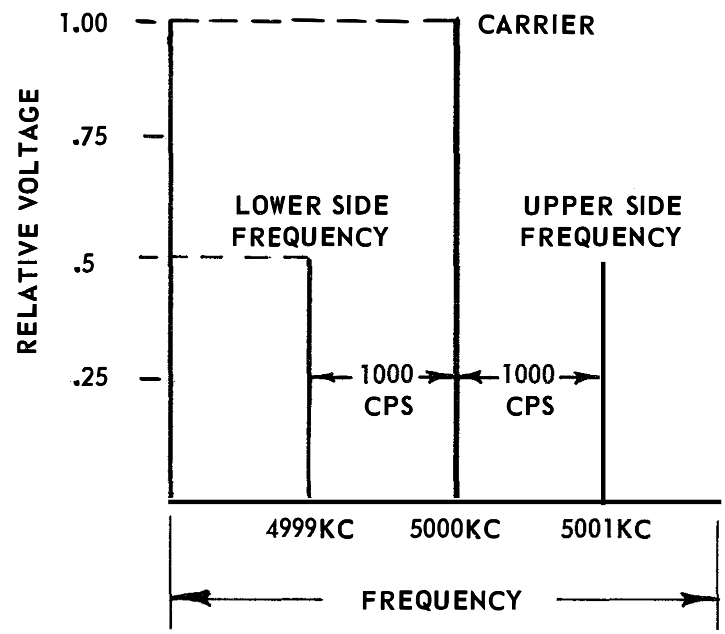


Figure 2-4. Frequency Spectrum Plot of AM Transmitter.

and displayed on an oscilloscope screen, it will have the same wave shape as any other 4999-Kc sine wave RF signal. When the selective receiver is tuned to 5001 Kc, the upper side frequency will also appear as a steady carrier with no audible tone. In order for the 1000-cycle audio tone to be heard on the speaker, or the modulation waveform patterns to be displayed on the oscilloscope, the receiver must have sufficient bandpass to accept at least the carrier and one of the side frequencies.

2-29. The side frequencies appear only when the 1000-cycle tone modulation is applied at the transmitter. When the modulation is removed, the signals at 4999 and 5001 Kc will disappear. If the amplitude of the 1000-cycle modulation signal is varied from zero to maximum, the amplitude of the side frequency will also vary from zero to maximum. It is apparent, then, that the two side frequencies are created by the amplitude modulation process.

2-30. When a receiver with 0.5-Kc bandpass is tuned to a 5000-Kc carrier, no modulation will be heard and no change in the amplitude of the carrier will be observed when the modulation is applied, varied, or removed at the transmitter, since no change occurs in the carrier with modulation. We can reasonably conclude that the intelligence bearing components of this signal are contained in the side frequencies with none in the carrier.

2-31. Power Distribution in the AM Signal.

2-32. As shown in figure 2-4, when a carrier is modulated 100 percent by a pure sine wave tone, the voltage value at either side frequency will be exactly half the carrier voltage. In a circuit where the resistance remains constant, the power will be proportional to the square of the applied voltage. In our earlier example, the power in each side frequency will thus be equal to 0.5 squared, times the carrier power. Assuming the carrier output is 1000 watts, when the power in each side frequency will be 1000 times 0.5 squared, or 250 watts. Since two side frequencies are present, the total side frequency power is 500 watts.

2-33. Figure 2-5 shows the usual modulation envelope presentation of an AM signal as it might be displayed on an oscilloscope screen. This drawing is included in practically all communications texts and is usually labeled "carrier with 100 percent modulation." This type of display has caused considerable confusion since it seems to contradict the previous statements that the carrier amplitude remains unchanged during modulation. What you are actually viewing here is the composite wave which consists of the two side frequencies and the carrier being modulated by 1 cycle of audio frequency. It is evident that during the time required for one half cycle of modulation, the voltage value of the composite signal will vary from a maximum of twice the carrier voltage to zero. Since the power in a resistive circuit is proportional to the square of the applied voltage, a maximum of "peak" power of the 1000-watt transmitter will be equal to 1000 x 2 squared, or 4000 watts. Half of a

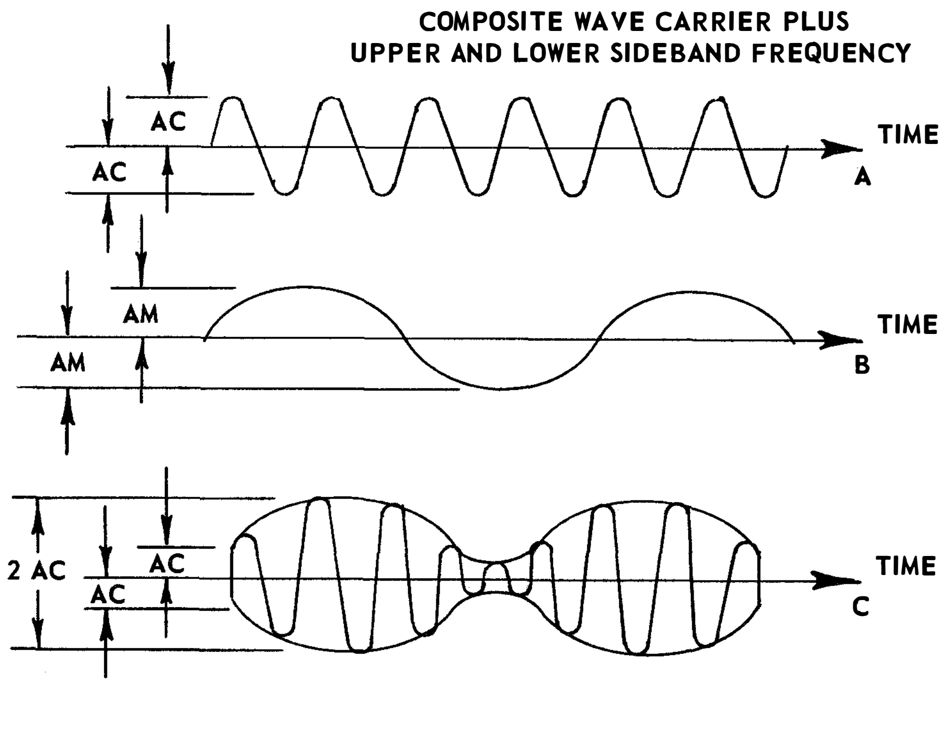


Figure 2-5. 100 Percent Amplitude-Modulated RF Carrier.

modulation-cycle later, the signal voltage and power output will be zero. From this it is apparent that a transmitter rated at 1000 watts must have a peak modulation capacity of 4000 watts if 100 percent modulation, without distortion, is to be realized.

2-34. Formation of the Sidebands.

2-35. Up to this point, the term "sideband" has been avoided as we have been working with single-tone modulation. Strictly speaking, single-tone modulation produces side frequencies, whereas multitone, or complex modulation, is required to produce sidebands. This distinction between side frequencies and sidebands is important because you will later be working with two-tone test signals as well as other modulation waveforms.

2-36. The sidebands exhibit certain characteristics which are not apparent from our study of the simpler side frequencies. It was mentioned previously that the two sidebands are mirror images of each other or, to use a common expression, they are actually back-to-back on either side of the carrier. It is important to keep in mind when discussing the sidebands that we are referring to the amplitude of the output wave plotted against frequency. The very term "sideband" denotes that it occupies a portion of the spectrum. As you know, sidebands or side frequencies exist only during the process of modulation. The unmodulated carrier contains no sidebands and, surprisingly, occupies no channel or spectrum space.

2-37. As an illustration of the formation of sidebands, refer again to the hypothetical 5000-Kc, 1000-watt transmitter where, instead of a single tone, the modulation signal will consist of four sine wave tones at frequencies of 1000, 2000, 3000, and 4000 cycles. For the purpose of discussion, these tones will be considered to be of equal amplitude and any effects of interaction on each other, that is, inter-modulation, will be ignored. Figure 2-6 illustrates the double sideband spectrum form when a 5000-Kc carrier is 100 percent modulated by a four-tone audio signal. The sideband frequency components which appear at 4999, 4998, 4997, and 4996 Kc correspond to the 1000-, 2000-, 3000-, and 4000-cycle tones. The group of frequencies between the carrier and 4996 Kc is known as the lower sideband. Note that the lowest frequency, 1000 cycles, produces a sideband component only 1-Kc lower than the carrier frequency, whereas the highest frequency, 4000 cycles, produces a sideband component 4-Kc lower than the carrier frequency. The complete lower sideband is composed of different frequency signals produced by subtracting the frequency of the audio components from the frequency of the carrier. Thus, the highest audio frequency will produce the lowest sideband frequency and the lowest audio frequency will produce the highest lower sideband frequency, an inversion condition. The upper sideband is exactly like the lower, except that the audio frequencies add to, instead of subtract from, the carrier frequency to produce a series of some frequency sideband components. Referring to figure 2-6 again, this shows the 1000-cycle tone mixed with the 5000-Kc carrier, producing an upper sideband of 5001 Kc. The

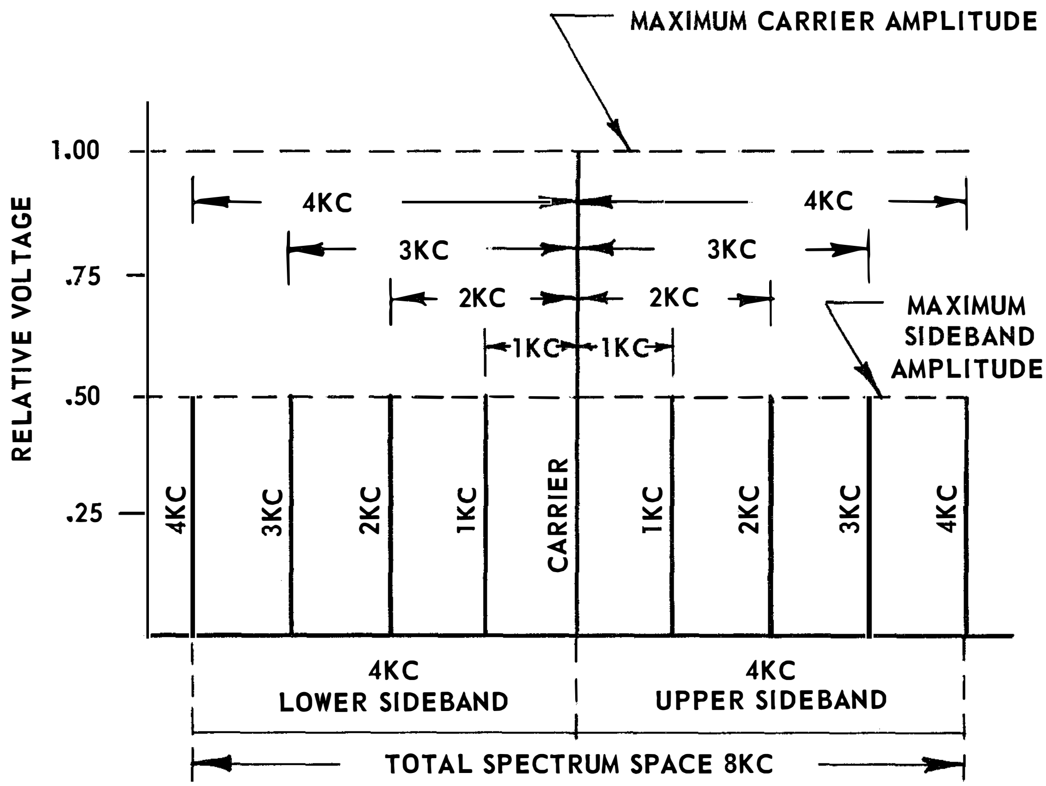


Figure 2-6. Sideband Formation With a Four-Tone
Sine Wave Modulating Signal.

2000-, 3000-, and 4000-cycle frequencies produce upper sideband components at 5002, 5003, and 5004 Kc, respectively. In the upper sideband spectrum, the lowest audio frequency will produce the lowest frequency upper sideband component and the highest audio frequency will produce the highest frequency sideband component. Compared with the lower sideband components, precisely opposite frequency conditions prevail in the upper sidebands. This is why the upper and lower sidebands are often called, and correctly so, mirror images of each other.

2-38. Each upper, as well as lower, sideband contains all the intelligence information included in the modulation signal. It can therefore be seen that either or both sidebands will provide faithful reproduction of the transmitted information from a suitable receiver. In figure 2-6 again, the carrier frequency is 5000 Kc; however, the RF carrier may be any frequency. Any RF carrier mixed with a four-tone audio signal will produce a double sideband spectrum where the components are related to each other exactly as shown. For example, if the carrier frequency is 50 Kc instead of 5000 Kc, the lowest sideband components will be located at 49, 48, 47, and 46 Kc.

The upper sideband components then would be 51, 52, 53, and 54 Kc; however, when the sideband signal is generated at a low frequency, such as 50 Kc, it is then heterodyned to the operating frequency and amplified to the desired power level.

2-39. 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 SSB and is used extensively in the transmission of television signals. The bandwidth of the vestigial sideband signal in standard television broadcasting is approximately 1/6 the bandwidth of the full sideband.

SECTION III

NATURE OF SINGLE SIDEBAND SIGNALS

3-1. Introduction.

3-2. As previously defined, an SSB signal is an audio signal converted to a radio frequency with or without inversion. To facilitate illustration of the manner and the results of this conversion, it is necessary to use pure sine wave tones rather than the very complex waveforms of the human voice. For this reason single tones, or combinations of two or three tones, are generally used in the following discussion.

3-3. The SSB Generator.

3-4. The most familiar SSB generator consists of a balanced modulator followed by an extremely selective mechanical filter, as shown in figure 3-1. The balanced modulator produces basically two output frequencies, (1) an upper sideband frequency equal to the injected IF frequency plus the input audio frequency, and (2) a lower sideband frequency equal to the injected IF frequency minus the input audio frequency. Theoretically the injected IF frequency is balanced out in the modulator so that it does not appear in the output.

3-5. It should be noted that the generation of undesirable products occurs in any mixing operation as well as the generation of the desired products. The equipment must be so designated as to minimize the generation of undesirable products and to attenuate those undesirable products which are generated. This is accomplished by designing good linear operating characteristics into the equipment to minimize the generation of undesirable frequencies and by choosing injection frequencies which will facilitate suppression of undesirable frequencies. It should also be noted that the IF carrier injected into the balanced modulator is only theoretically cancelled in the output. Practical design considerations determine the extent to which the carrier can be balanced out. Present balanced modulators, using controlled carrier leak to balance out uncontrolled carrier leak, result in carrier suppression of 30 to 40 db below the PEP of the sidebands. Further suppression of the carrier by the SSB filter results in an additional 20 db of carrier suppression. Total carrier suppression of 50 to 60 db, therefore, can be reasonably expected from the transmitter system.

3-6. Generating the Single-Tone SSB Waveform.

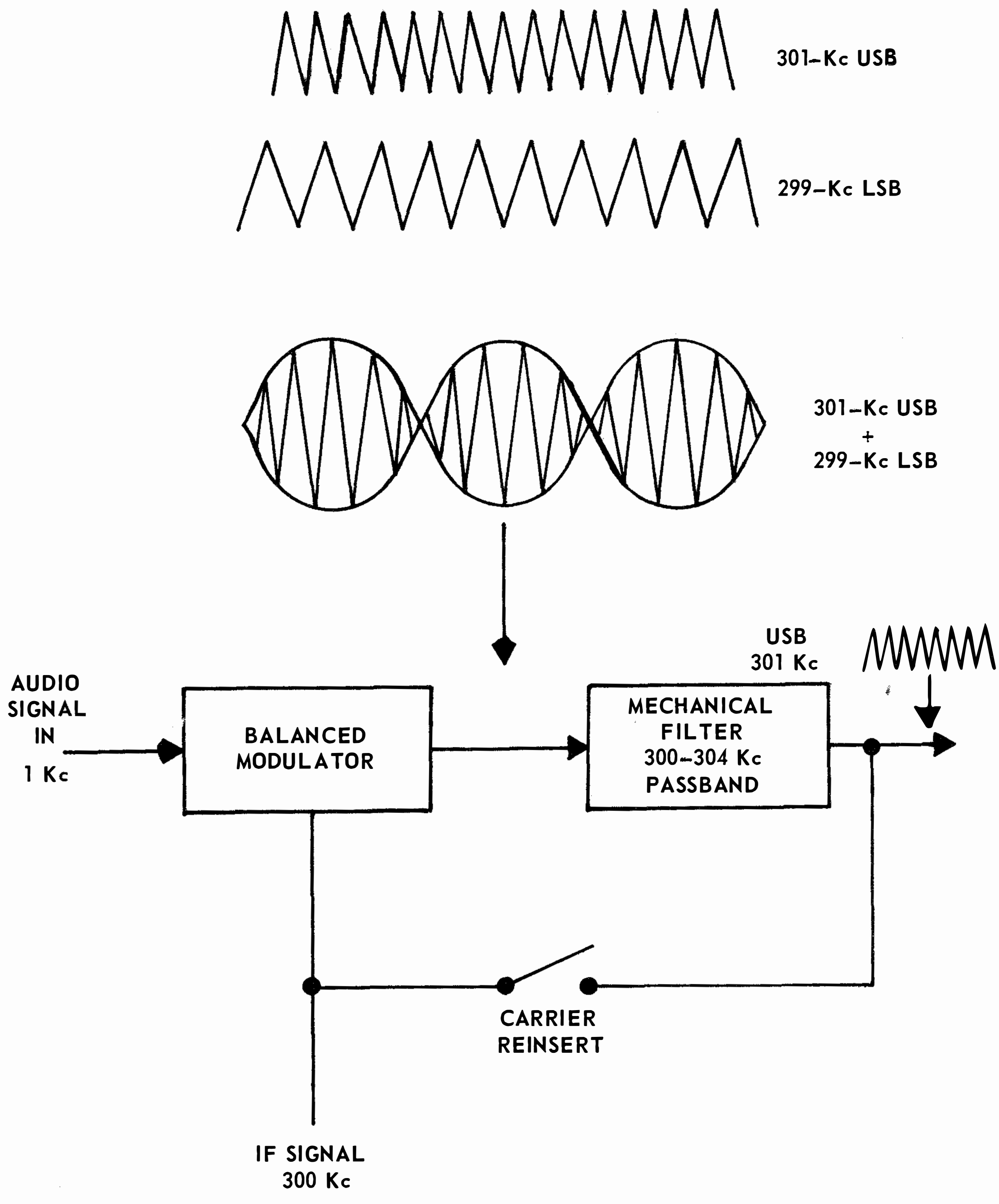


Figure 3-1. Filter-Type SSB Generator.

3-7. The most fundamental SSB waveform is generated from the single-audio tone. This tone is processed through the SSB generator to produce a single IF frequency. As pointed out previously, the SSB signal is actually generated at an IF frequency and is subsequently converted up in frequency to the transmitted RF frequency. It is the generation of the SSB signal at the IF frequency with which we are concerned.

3-8. Figures 3-2 and 3-3 show the waveforms obtained in a filter-type SSB generator. The audio tone injected into the balanced modulator is 1 Kc and the IF frequency injected is 300 Kc. The output from the balanced modulator contains the 299-Kc lower sideband and 301-Kc upper sideband frequencies. These two sideband frequencies, being of equal amplitude, produce the characteristic half sine wave envelope shown in figure 3-2. The repetition rate of this envelope with a 1-Kc tone is 2 Kc, a difference between the two frequencies represented by the envelope. This IF signal, which contains both the upper and lower sideband signals, is called a double sideband (DSB) signal.

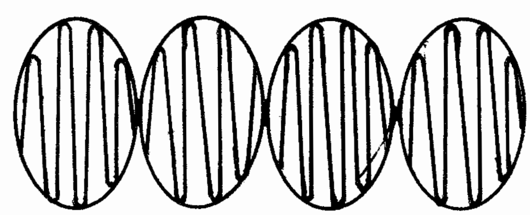


Figure 3-2. Single-Tone Balanced Modulator Output.

Figure 3-3. Single-Tone Balanced Modulator Output After Filtering Out the Lower Sideband.

3-9. By passing the DSB through a highly selective filter with a 300- to 303-Kc passband, the upper sideband signal is passed while the lower sideband signal is attenuated. The 301-Kc signal remains as the upper sideband and appears as shown in figure 3-3. Note that the SSB remaining is a pure sine wave when a single-tone audio signal is used for modulation. This SSB signal is displaced up in the spectrum from its original audio frequency by an amount equal to the carrier frequency, in this case 300 Kc. This SSB signal can be demodulated at the receiver only by converting it back down in the frequency spectrum. This is done by mixing it with an independent 300-Kc IF signal at the receiver.

3-10. Generating the SSB Waveform, Single-Tone With Carrier.

3-11. From the single-tone SSB signal without carrier it is a simple step to generate the single-tone SSB signal with carrier; this is done by reinserting the carrier after the filtering operation, as shown in figure 3-1. When the reinserted carrier is of the same amplitude as the SSB signal, the waveform shown in figure 3-4 results. Note that this waveform is similar to the DSB signal obtained directly out of the balanced modulator, as shown in figure 3-2. However, the frequency components of the two waveforms are not the same. The frequency components of the SSB signal with carrier are 301 Kc and 300 Kc when a 1-Kc signal is used. An SSB signal with full carrier can be demodulated with a conventional diode detector used in AM receivers without serious distortion or loss of intelligibility. If the reinserted carrier is such that the carrier level is less than the level of the single-tone SSB signal, the waveform shown in figure 3-5 results. To successfully demodulate this signal, the carrier must be separated, amplified, exalted, and reinserted in the receiver, or locally supplied. The separate carrier amplification should be sufficient to raise the reinserted carrier to a level greater than the level of the sideband signal. The waveform shown in figure 3-5 represents the waveform used in the SSB-with-pilot-carrier systems. The exalted-carrier technique is used to demodulate such a signal.

3-12. Generating the Two-Tone SSB Waveform.

3-13. The two-tone SSB waveform is generated by combining two audio tones and then injecting this two-tone signal into the balanced modulator. One sideband is then suppressed by the filter, leaving the SSB waveform shown in figure 3-6. This two-tone SSB signal is seen to be similar to the single-tone DSB as well as the SSB signal with full carrier. However, the two-tone SSB signal contains a different two frequencies than either of the other two. In the two-tone SSB signal, as shown in figure 3-6, 1-Kc and 2-Kc audio signals of equal amplitude are injected into the balanced modulator. After filtering, this results in a two-tone SSB signal containing frequencies of 301 Kc and 302 Kc. If a pilot carrier is reinserted with a two-tone test signal, the pilot carrier will be indicated by the appearance of a sine wave ripple on the two-tone waveform. This waveform is shown in figure 3-7. The generation of this two-tone envelope can be shown clearly with vectors representing the two audio frequencies, as shown in figure 3-8.

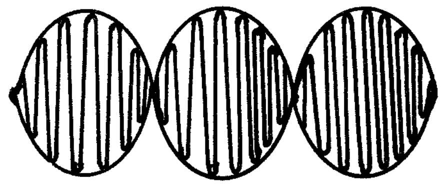


Figure 3-4. Single-Tone SSB Signal With Carrier and Tone Equal in Amplitude.

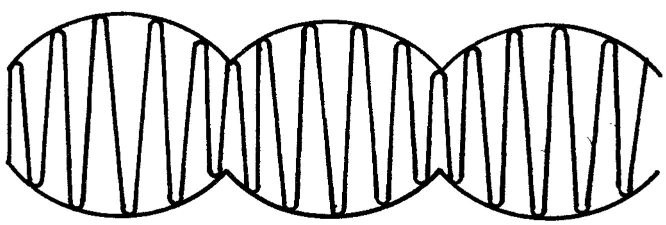


Figure 3-5. Single-Tone SSB Signal With Carrier 10 db Below Tone.

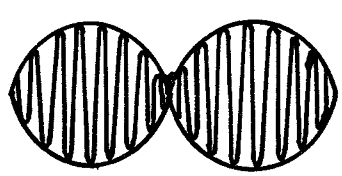


Figure 3-6. Two-Tone SSB Signal With Tones of Equal Amplitude.

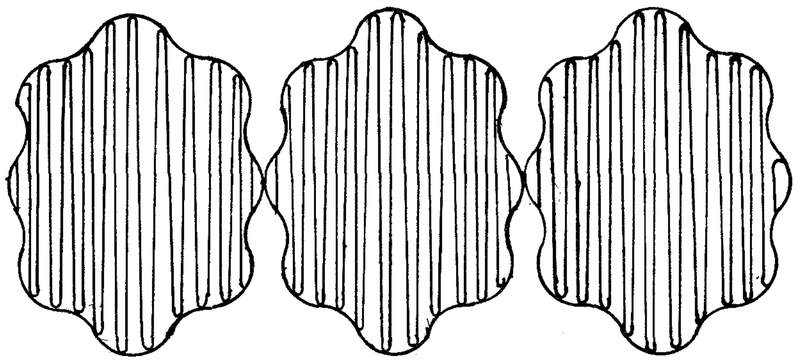


Figure 3-7. Two-Tone SSB Signal With Small Reinserted Pilot Carrier.

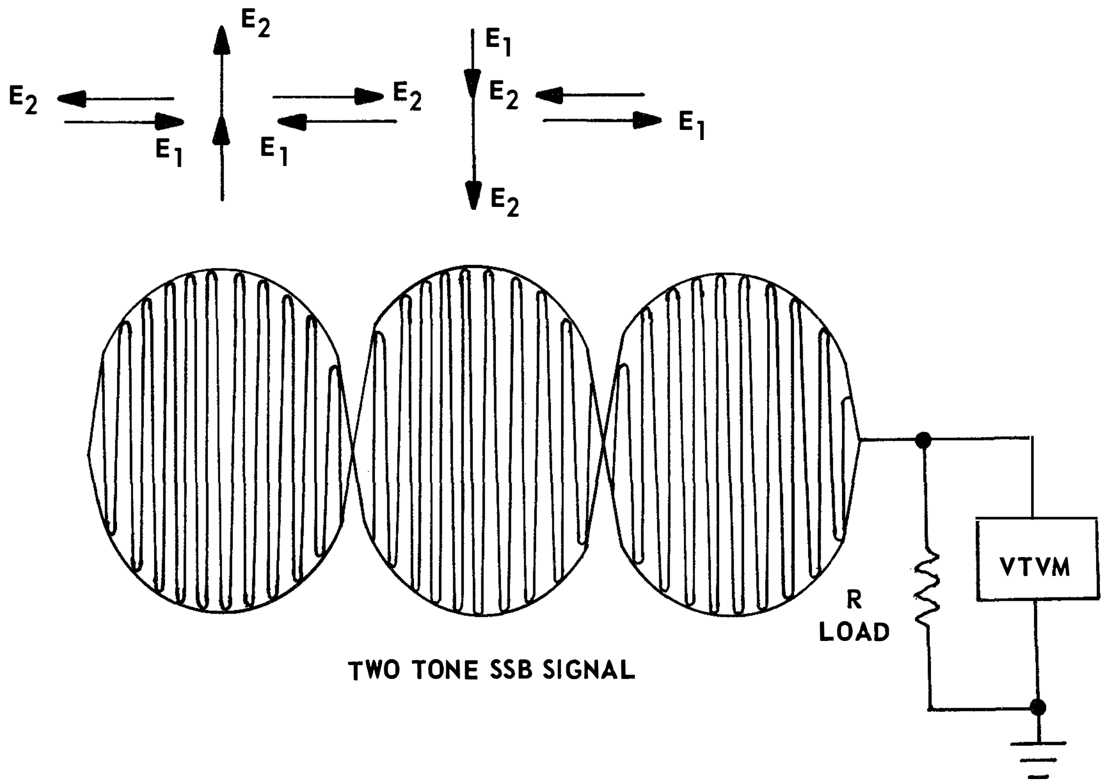


Figure 3-8. SSB Power Measurements With Two-Tone Test Signal.

When the two vectors are exactly opposite phase, the envelope value is zero. When the two vectors are exactly in phase, the envelope value is maximum. This generates the half sine wave shape of the two-tone SSB envelope which has a repetition rate equal to the difference between the two audio tones.

3-14. The two-tone SSB envelope is of special importance because it is from this envelope that power output from an SSB system is usually determined. An SSB transmitter is rated in PEP output with the power measured with a two-equal-tone test signal. With such a test signal, the actual power dissipated in the load is one-half the PEP. This is shown in figure 3-8. When the half sine wave signal is fed into a load, a peak-reading rms-calibrated vacuum tube voltmeter (VTVM) across the load indicates the rms value of the peak envelope voltage (PEV). This load-meter reading is equal to the in-phase sum of E_1 plus E_2, where E_1 and E_2 are the rms voltages of the two tones.

Since in the two-tone test signal E_1 equals E_2, PEP = $\frac{(2E_1)^2}{R}$ or $\frac{(2E_2)^2}{R}$. The average power dissipated in the load must equal the sum of the power represented by each tone, $\frac{(E_1)^2}{R} + \frac{(E_2)^2}{R}$, $\frac{2(E_1)^2}{R}$ or $\frac{2(E_2)^2}{R}$. Therefore, with a two-equal-tone SSB test signal the average power dissipated in the load is equal to one-half the PEP and the power in each tone is equal to one fourth the PEP. PEP can be determined from the relationship PEP = $\frac{V^2 VTVM}{R}$. The average power can be determined from the relationship $P_{average} = \frac{0.5V^2\ VTVM}{R}$. This is true only where the VTVM used is a peak-reading rms-calibrated voltmeter. Similar measurements can be made using an AC ammeter in series with the load instead of the VTVM across the load.

3-15. The preceding analysis can be carried further to show that with a three-equal-tone SSB test signal the power in each tone is 1/9 the PEP and the average power dissipated in the load is 1/3 the PEP. These relationships are true only if there is no distortion of the SSB envelope, but since distortion is usually small its effects are usually neglected.

3-16. Mechanical Filters.

3-17. Both SSB transmitters and SSB receivers require very selective bandpass filters in the region of 100 Kc to 500 Kc. In receivers, a high order of adjacent channel rejection is required if channels are to be closely spaced to conserve spectrum space. In SSB transmitters the signal bandwidth must be limited sharply in order to pass the desired sideband and reject the other sideband. The filter used, therefore, must have very steep skirt characteristics and a flat bandpass characteristic. These filter requirements are met by LC filters, crystal filters, and mechanical filters. Until recently, crystal filters used in commercial SSB equipment were in the 100-Kc range. These filters have excellent selectivity and stability characteristics but their large size makes them subject to shock or vibration deterioration and their cost is quite high. Newer crystal filters are being developed which have extended frequency range and are smaller. These newer crystal filters are more acceptacle for use in SSB equipment. LC filters have been used at IF frequencies in the region of 20 Kc; however, generation of the SSB signal at this frequency requires an additional mixing stage to obtain transmitter output in the high frequency range. For this reason LC filters are not widely used. Recent advancement in the development of mechanical filters has led to their acceptance in SSB equipment. These filters have excellent rejection characteristics, are extremely rugged, and are small enough to be compatible with miniaturization of equipment. Also to the advantage of the mechanical filter is a "Q" on the order of 10,000, which is about one hundred times the "Q" obtainable with electrical elements.

3-18. Although the commercial use of mechanical filters is relatively new, the basic principles upon which they are based is well established. The mechanical filter is a mechanically resonant device which receives electrical energy, converts it into mechanical vibrations, then converts the mechanical energy back into electrical energy at the output. The mechanical filter, basically, consists of four elements:

a. An input transducer which converts the electrical input into mechanical oscillations.

b. Metallic discs which are mechanically resonant.

c. Coupling rods which couple the metal discs.

d. An output transducer which converts the mechanical oscillations back into electrical oscillations.

Figure 3-9 shows the elements of the mechanical filter.

3-19. The transducer which converts electrical energy into mechanical energy, and vice versa, may be either a magnetostrictive device or an electrostrictive device. The magnetostrictive transducer is based on the principle that certain materials elongate or shorten when in the presence of a magnetic field. Therefore, if an electrical signal is sent through a coil which contains a magnetostrictive material as the core, the electrical oscillations will be converted into mechanical oscillations. The mechanical oscillations can then be used to drive the mechanical elements of the filter. The electrostrictive transducer is based on the principle that certain materials, such as piezoelectric crystals, will compress when subjected to an electric potential. In practice, the magnetostrictive transducer is more commonly used. The transducer not only converts electrical energy into mechanical energy and vice versa, but it also

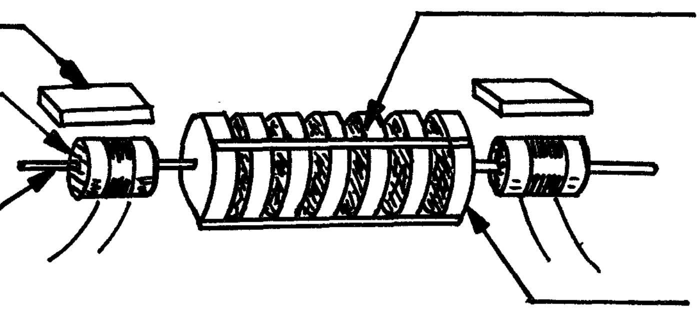


Figure 3-9. Elements of the Mechanical Filter.

provides proper termination for the mechanical network. Both of these functions must be considered in transducer design.

3-20. The center frequency of the mechanical filter is determined by the metal discs which represent series-resonant circuits. In practice, filters between 50 Kc and 600 Kc can be manufactured. This by no means indicates mechanical filter limitations, but is merely the area of design concentration in a relatively new field. Since each disc represents a series-resonant circuit, it follows that increasing the number of discs will increase skirt selectivity of the filter. Skirt selectivity is specified as shape factor, which is the ratio of bandpass 60 db below peak to bandpass 6 db below peak. Practical manufacturing presently limits the number of discs to eight or nine in a mechanical filter. A six-disc filter has a shape factor of approximately 2.2, a seven-disc filter a shape factor of approximately 1.85, and a nine-disc filter a shape factor of approximately 1.5. Future development of mechanical filters promises an even faster rate of cutoff.

3-21. By varying the mechanical coupling between the discs, that is, making the coupling rods larger or smaller, the bandwidth of the filter is varied. Because the bandwidth varies approximately as the total area of the coupling wires, the bandwidth can be increased by using either larger coupling rods or more coupling rods. Mechanical filters with bandwidths as narrow as 0.5 Kc and as wide as 35 Kc are practical in the 100-Kc to 500-Kc range.

3-22. Although an ideal filter would have a flat nose or passband, practical limitations prevent the ideal from being obtained. The term ripple amplitude, or peak-to-valley ratio, is used to specify the nose characteristic of the filter. The peak-to-valley ratio is the ratio of maximum to minimum output level across the useful frequency range of the filter (see figure 3-10). A peak-to-valley ratio of 3 db can be obtained on a production basis by automatic control of materials and assembly. Mechanical filters with a peak-to-valley ratio of 1 db can be produced with accurate adjustment of filter elements.

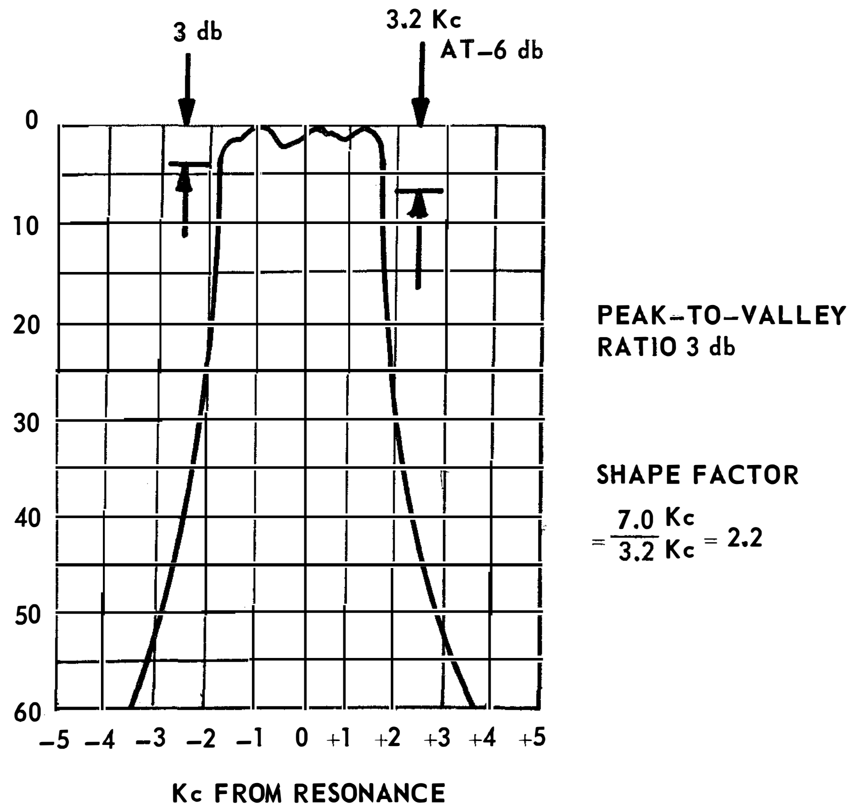


Figure 3-10. Characteristic Curve, Mechanical Filter.

3-23. Spurious responses occur in mechanical filters due to mechanical resonances other than the desired resonance. By proper design, spurious resonances can be kept far enough from the passband to permit other tuned circuits in the system to attenuate the spurious responses.

3-24. Other mechanical filter characteristics of importance include insertion loss, transmission loss, transfer impedance, input impedance, and output impedance. Since the input and output transducers of the mechanical filter are inductive, parallel external capacitors must be used to resonate the input and output impedance at the filter frequency. With such capacitors added, the input and output impedances are largely resistive, ranging between 1000 and 50,000 ohms. The insertion loss is measured with both the source and the load impedance matched to the input and output impedance of the filter. The value of insertion loss ranges from 2 to 16 db, depending upon the type of transducer. The transmission loss is an indication of the filter loss with source and load impedance mismatched. The transmission loss is of importance when using a mechanical filter in pentode IF amplifiers where both source and load impedance are much greater than the filter impedances. The transfer impedance is useful to determine the overall gain of a pentode amplifier stage which utilizes a mechanical filter. The transfer impedance of the filter multiplied by the transconductance of the pentode gives the gain of the amplifier stage.

3-25. The physical size of the mechanical filter makes it especially useful for modular and miniaturized construction. The mechanical filter is about one inch square by three inches long. More recent development has resulted in a smaller tubular filter which is about one-half inch in diameter by one and three-quarter inches long.

3-26. Mechanical filter types other than disc are presently being used. These include the plate type which is a series of flat plates assembled in a ladder arrangement. Another type which has recently been developed is the neck and slug type. This filter consists of a long cylinder which is turned down to form the necks which couple the remaining slugs. All mechanical filters are similar in that they employ mechanical resonance. Mechanical filters differ in that they employ various modes of mechanical oscillation to achieve their purpose. They may also use different types of transducers.

3-27. Compatability of AM and SSB Systems.

3-28. Pure SSB suppressed-carrier systems, abbreviated SSSC, are not compatible with present-day AM DSB. It stands to reason, therefore, that before a change can be made from the existing mode of communication to the pure SSB 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, have arisen and will have to be overcome before the transition from AM DSB to pure SSB can be considered complete.

3-29. Since pure SSB signals cannot be used for general communications 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, that is, it will have to be compatible so that it can be used by either system. The only strict requirement of this compromise is that sufficient carrier power be transmitted so that detection of the signal in a conventional AM receiver may be effected properly.

3-30. There are various methods of modifying an SSB suppressed-carrier 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 SSB suppressed-carrier signal. This method sacrifices the spectrum reduction and efficiency of the pure SSB system but retains the SSB system frequency stability.

3-31. Another method of converting the SSB suppressed-carrier 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 conserving characteristics of the SSB system are, in part, retained.

3-32. A third method of converting an SSB suppressed-carrier 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 DSB transmitter. This modification, however, is quite complex and expensive because the power stages in 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.

3-33. An SSB receiver can be easily converted to receive AM DSB signals by including a conventional AM detector in the 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 IF amplification. Receivers using exalted-carrier or product detectors are capable of AM reception without any modification. A requirement of the IF amplifiers used is that their bandwidth be sufficient to pass at least one sideband and the carrier of the conventional AM DSB signal. The carrier and DSB signal will provide a reference for the locally generated carrier signal and the receiver AFC circuit will correct for any slight frequency error in the AM transmitter. For substantially large frequency error, manual tuning may be provided to bring the frequency error within the AFC circuit correction range.

3-34. 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.

3-35. It is also possible to convert an AM DSB transmitter into a compatible SSB 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 in either AM or SSB equipments. At the output of an SSB generator, the phase and amplitude components of the signals are separated, amplified, and then recombined in conventional AM modulator and power stages. If the proper time relationships are maintained between the separate 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.

3-36. Present AM receivers having a beat frequency oscillator can detect SSB suppressed-carrier 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 SSB suppressed-carrier signals have been developed. Such an SSB adapter connected to the output of the IF stage will provide the circuits necessary to properly demodulate the SSB suppressed-carrier signal. If a compatible signal is received, SSB plus full carrier, the only change required in the AM receiver is to reduce the IF 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 DSB signal.

3-37. 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 AFC 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 a standard throughout the entire industry.

3-38. Advantages and Disadvantages of SSB.

3-39. The advantages of an SSB suppressed-carrier system compared to a DSB AM system are summarized as follows:

a. Reduced frequency spectrum, resulting in more available channels.

b. Elimination of high powered carriers, giving a better power efficiency of intelligence-bearing sidebands.

c. A more reliable signal in the presence of certain interference and fading conditions.

d. Reduced size and weight.

The disadvantages of the same system are summarized as follows:

a. Necessity for extremely stable circuits in the transmitter.

b. Requirement for complex AFC circuits in the receiver.

c. Higher cost and circuit complexity resulting from demodulation difficulties encountered because no carrier is transmitted.

3-40. Exalted-Carrier Receivers.

3-41. 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 over-modulation and distortion in the same manner as in a transmitter. The exalted-carrier receiver is a receiver which insures adequate carrier level for 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 as well. These receivers may not have SSB filters or separate sideband channels. In many receivers, provisions are made for both types of transmission and, with a few additional circuits or modifications, all types of AM transmissions may be received.

3-42. SSB 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 depend on a precise-frequency carrier-insertion oscillator for detecting purposes, rather than the transmitted 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.

3-43. Both types of 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.

3-44. Synchronous Communications.

3-45. 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 3-11. 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 IF system is needed. Detection may be accomplished at a very low level and the resulting signal amplified at the audio frequencies. Filtering at the audio frequencies permits much greater selectivity than is possible at the IF 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 nor 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 must be given to its use in lieu of SSB equipment per se.

3-46. SSB Applied to VHF Scatter Systems.

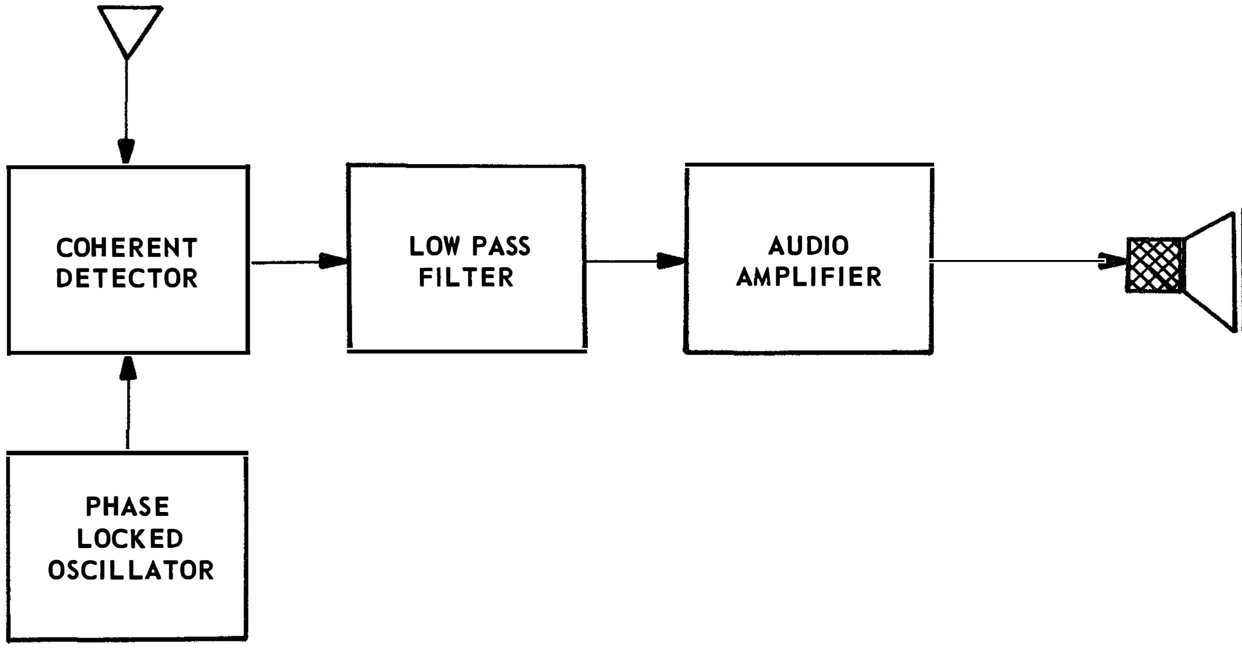


Figure 3-11. Block Diagram, Basic DSB Receiver.

3-47. In VHF ionospheric scatter systems (trans-horizon propagation), a major problem commonly encountered is multipath fading. Because of these fading effects and because VHF scatter systems operate best in the already overcrowded 25 to 60 megacycle spectrum range, broad-band modulation systems are not well adapted for this method of communications. However, SSB 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.

3-48. Although narrow band FM uses a wide bandwidth (a wider bandwidth than SSB), it has the advantage of simplicity and relative insensitivity to rapid fading when the received signal exceeds a certain minimum level. The advantages of SSB compared to FM, in VHF scatter systems, are narrower bandwidth, lower average power requirements, and no minimum threshold level. Reception with SSB can be obtained at much lower signal level than that required when using frequency modulation.

3-49. 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 multiplex teletype or data information. In such applications it is desirable to use the same oscillators for time reference, 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 SSB VHF scatter systems, can be obtained by using RF feedback methods.

SECTION IV

CARRIER SUPPRESSION TECHNIQUES

4-1. General.

4-2. In this section the circuitry for removal of the carrier from the AM signal is covered. The techniques used to remove the undesired sideband will be covered later.

4-3. The device most generally used to suppress the carrier is called the balanced modulator, although it is really better described as a balanced mixer. The balanced modulator may use semiconductor (germanium or silicon) diodes or vacuum tube diodes, triodes, or pentodes. Transistors have been used as balanced modulators in some of the recent circuitry, with good results.

4-4. Push-Pull Balanced Modulator.

4-5. Figure 4-1 shows a simple balanced modulator circuit which uses two screen grid tubes. The two plates are connected to the resonant tank circuit in a push-pull arrangement. The carrier is applied to the parallel control grids and the AF modulating signal is applied to the push-pull operated screens through transformer T1. The RF carrier and audio modulation signals are kept at low peak values (usually a few volts) and B+ may or may not be applied to the screens. With some types of tubes, it may be necessary to apply a low negative bias to the screens as shown in figure 4-1. This aids in selecting the operating points for optimum mixer and carrier suppression action.

4-6. With no audio applied to the screens, the positive excursions of the incoming carrier will cause a simultaneous increase in the plate currents of both tubes; and, during the negative grid swing, both plate currents will decrease simultaneously. These plate current pulses, when applied to the push-pull tank circuit, will generate two RF voltages which are equal in value but opposite in phase displacement. These two out-of-phase voltages will cancel each other and the net carrier output will be zero. The actual amount of carrier suppression obtained depends upon the degree of matching between the two tubes and their associated circuitry. Generally, two selected tubes of the same type will give about 20 db of carrier suppression in the circuit of figure 4-1. In this case, no adjustments other than resonating the balanced tank circuit to the carrier frequency are needed.

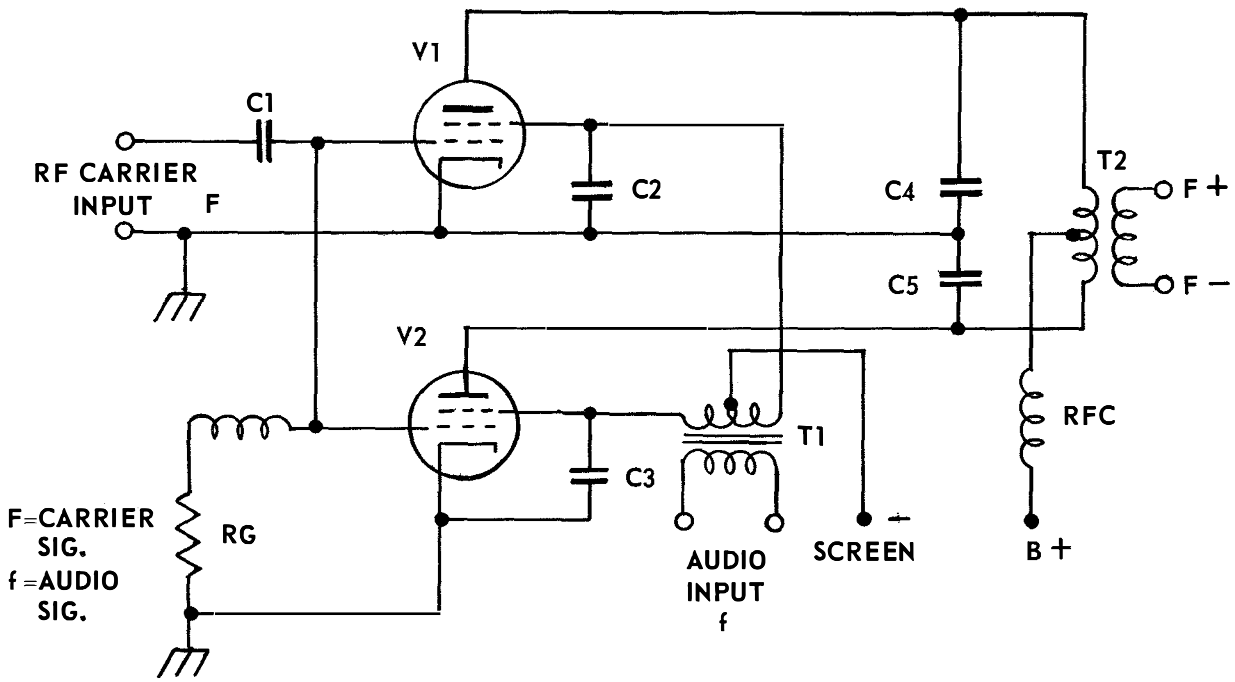


Figure 4-1. Balanced Modulator With Parallel Input and Push-Pull Output.

4-7. Carrier suppression can be further improved by the use of separate bias adjustments to equalize the two plate currents and small trimmer capacitors to balance each grid and plate capacitance to ground. In practical design work, a carrier suppression value of 30 to 35 db is generally considered satisfactory. However, when the balanced modulator is followed by a sideband filter, an additional 20 db of attenuation is usually obtained. The trimmers, tank circuit tuning, and bias values should be adjusted for minimum carrier in the output circuit. These various adjustments will interact to some extent; thus, it may be necessary to trim them several times before the lowest carrier level can be obtained.

4-8. During modulation, the audio voltages on the screens will be 180° out of phase with each other. This will cause the plate current of one tube to increase and that of the other to decrease at an audio rate. Two RF voltages will then appear across the primary of T2 and, not being equal, will result in a net RF voltage in the secondary. Because the modulator unbalance is occurring at an audio rate, the RF and audio frequency signals will mix (combine) to produce sum and difference RF signals (sidebands) in the output circuit. Since the modulator is balanced for the carrier but not for the sidebands, the output signal will consist of a pair of sidebands and a suppressed carrier. Technically, this output is known as a double-sideband suppressed-carrier signal or, more simply, as a DSB signal.

4-9. Parallel-Output Balanced Modulator.

4-10. In figure 4-2, the carrier is applied to the push-pull operated control grids and taken off two parallel-operated plates. Since the carrier signal at the grids is 180° out of phase, the plate current in one tube will be increasing as the plate current of the other is decreasing. These two plate current pulses are applied to a common single-ended tank circuit where each pulse cancels the other. Thus, no carrier output is produced. In this circuit the audio modulation signal is applied to the screens in the same manner as in figure 4-1. Either circuit is equally effective in the production of a suppressed carrier. The choice is determined primarily by constructional, rather than electrical, considerations.

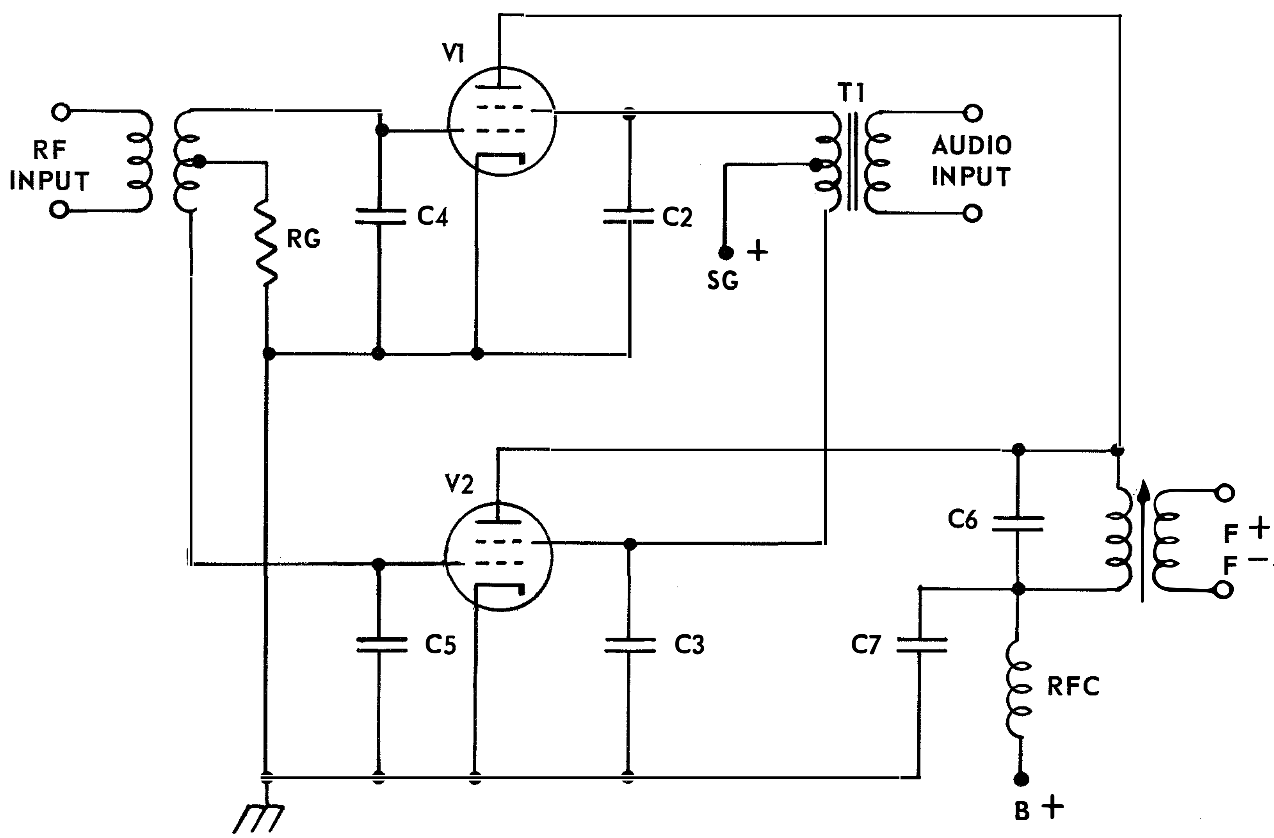


Figure 4-2. Balanced Modulator With Push-Pull Input and Parallel Output.

4-11. Vector Analysis.

4-12. At this time it may be of interest to compare the power vector of a suppressed carrier signal with that of an ordinary AM signal. The conventional vector resultants of an AM signal are shown in figure 4-3A. Here the sideband vectors are considered to revolve around the carrier vector, as indicated by the small curved arrows. The sum resultant of both sidebands and carrier is considered to revolve (at the carrier frequency) around the point represented by the carrier vector origin. The two sidebands are assumed to be of equal amplitude.

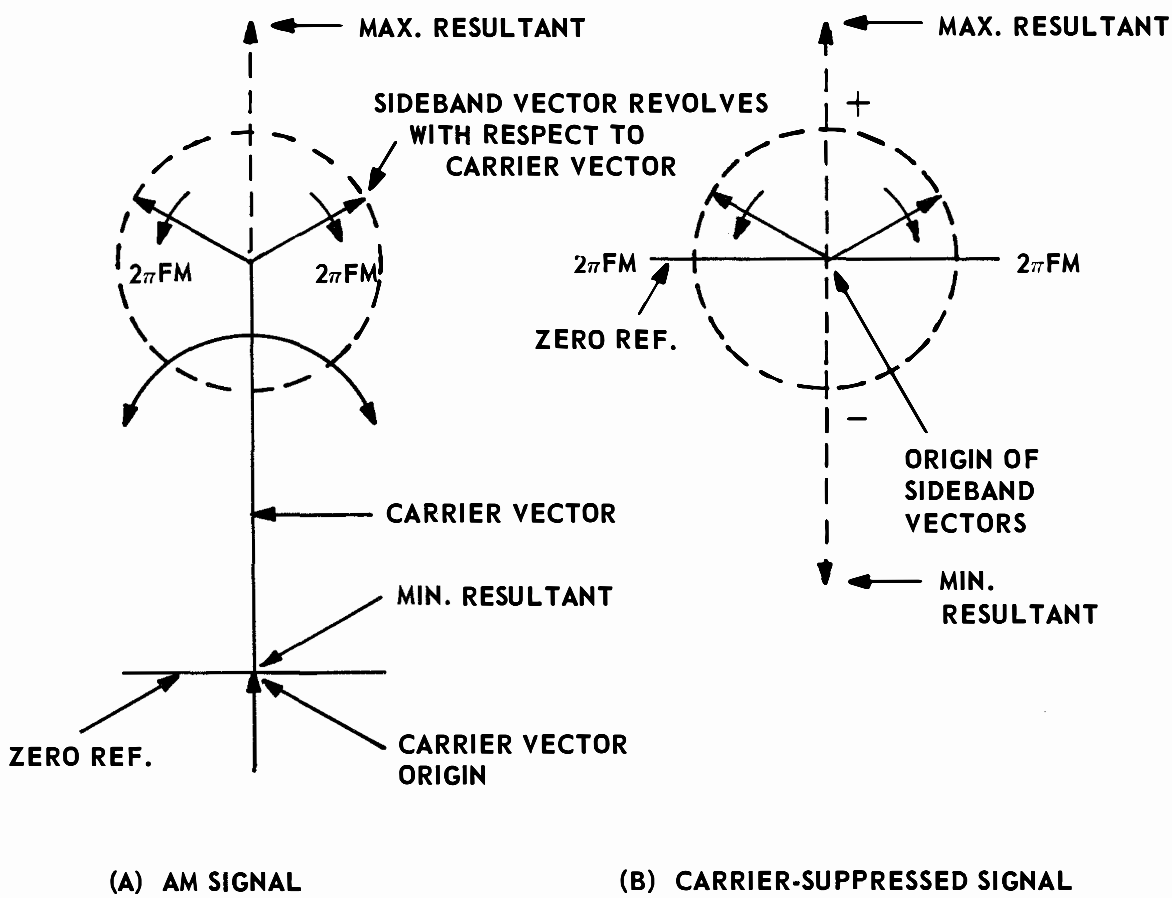


Figure 4-3. Vector Diagrams.

4-13. During modulation peaks, the sidebands and carrier will add vectorially to produce the maximum power resultant as shown. The maximum amplitude of the composite signal will then be twice the amplitude of the carrier alone. During the following half cycle of modulation, the two sideband vectors will have rotated to the point where they are combined. Thus, their resultant will be equal in magnitude but opposite in polarity to that of the carrier. When this point is reached, the sidebands and carrier cancel and the net output at that instant is zero. Note that the resultant will reach zero during the modulation cycle, but does not actually swing negative.

4-14. The vector diagram that represents the suppressed carrier is shown in figure 4-3B. The origin of the sideband vectors is at the zero reference line because no signals are present when there is no modulation. The sideband vectors revolve in the directions indicated by the curved arrows. When the two vectors coincide in the positive direction, they will add and form the maximum resultant (shown by the vertical dotted line). When the two vectors coincide in the negative direction, they will combine to form the minimum resultant, or "negative peak," as indicated by the same dotted line. Either resultant formed by the sidebands is considered to be a constant phase when both sidebands are of equal amplitude. In the suppressed carrier system, the sideband resultant swings both positive and negative during one modulation cycle.

4-15. Push-Pull Balanced Modulator.

4-16. The circuit of a dual-triode balanced modulator is shown in figure 4-4. The audio frequency modulation signal is applied to the grids in push-pull from either a phase inverter stage or the high impedance secondary of an interstage audio transformer. The RF carrier is introduced to the common cathode circuit through a 100-$\mu\mu$f condenser.

4-17. The push-pull output is coupled to a balanced tuned circuit, such as the primary of an IF transformer or the input of a bandpass filter. The two 10,000-ohm resistors and the 5000-ohm pot form an adjustable balanced load for the triode plates. The 5000-ohm pot is used to equalize the two plate currents for minimum carrier in the output circuit. Ordinarily, no balancing adjustments are required when the incoming audio signals are equal in amplitude.

4-18. Diode Modulator.

4-19. The diode modulator was developed during the early years of carrier system telephony but remained comparatively unknown in other communication circles until quite recently. The first balanced modulator used two or more diode rectifiers. The diode or rectifier modulator may be of the ring, series, or shunt type, depending upon the manner in which its diodes are connected. The main advantage of the diode modulator is its greater stability as compared with its vacuum tube counterpart. Moreover, diodes require no power, are very compact, and rarely require maintenance or replacement. Originally, the diodes were copper oxide rectifiers. Today, most are

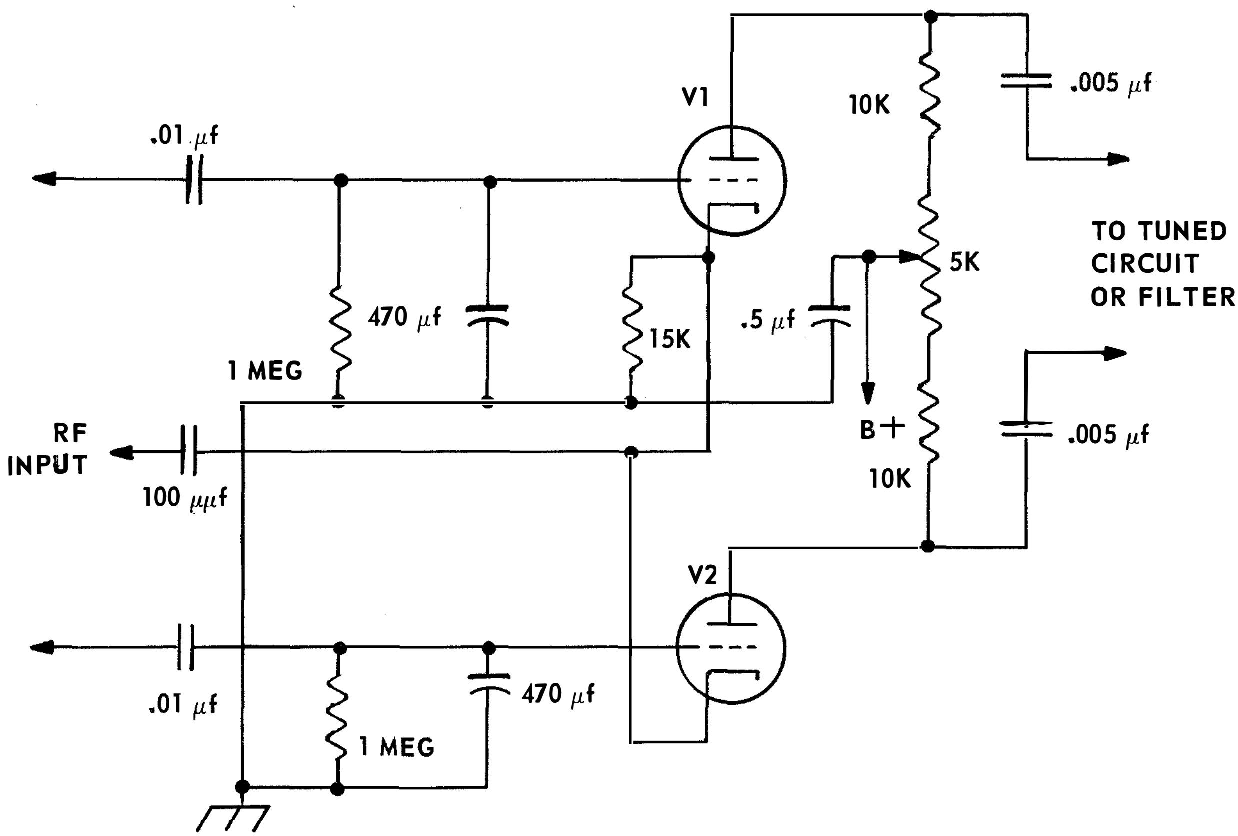


Figure 4-4. Dual Triode Balanced Modulator.

germanium, such as the ordinary type 1N34. When two or more diodes are carefully selected for matched electrical characteristics and mounted in a can or on a terminal board, the unit is known as a "varistor." As an example, the popular 1N34 varistor is simply a pair of closely matched 1N34 germanium diodes.

4-20. Ring Balanced Modulators.

4-21. The ring balanced modulator is the most efficient of the three types, being capable of twice as much output voltage as the shunt or series arrangements. However, the diodes must be very closely matched when used in a conventional ring circuit (figure 4-5). Otherwise, it will be impossible to obtain good carrier suppression in the output circuit.

4-22. The diode elements in any balanced modulator are arranged so that amplification of RF carrier alone does not cause a signal to appear in the output. In figure 4-6, the carrier balancing action can be analyzed by tracing the carrier current paths, as shown. When the carrier voltage is negative at point A, RF current will flow through T2, D1, D2, T1, and back to the generator at point B. The currents through

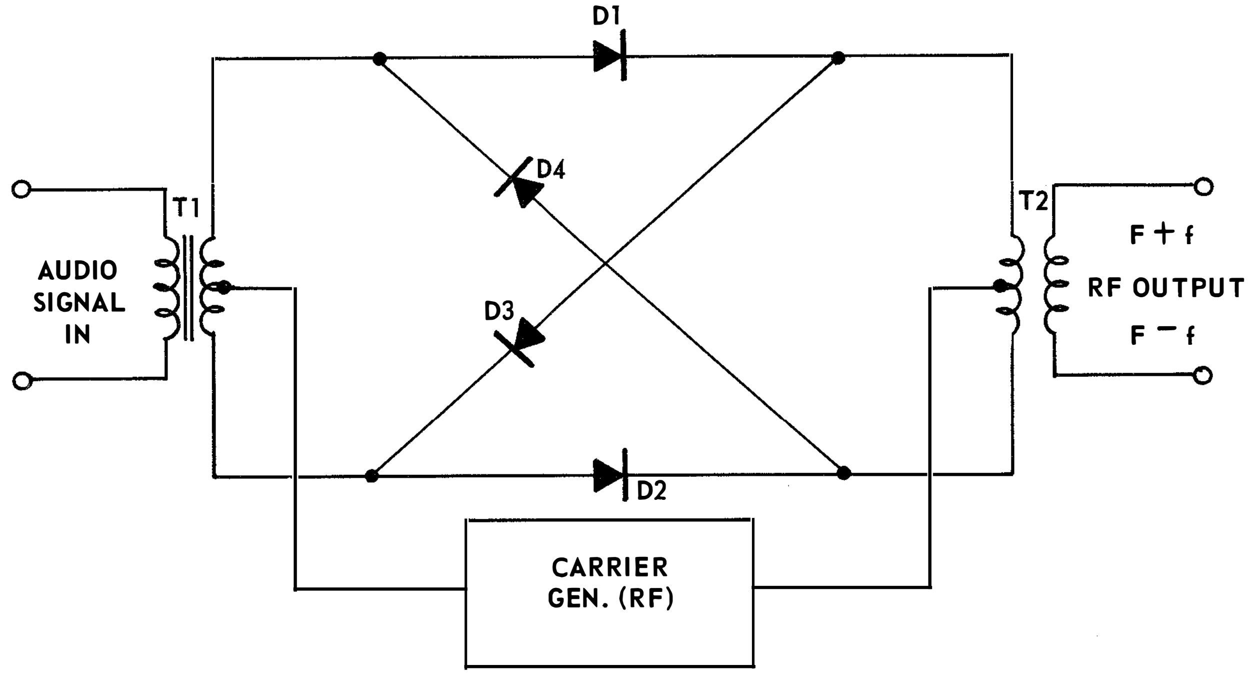


Figure 4-5. Conventional Ring Balanced Modulator.

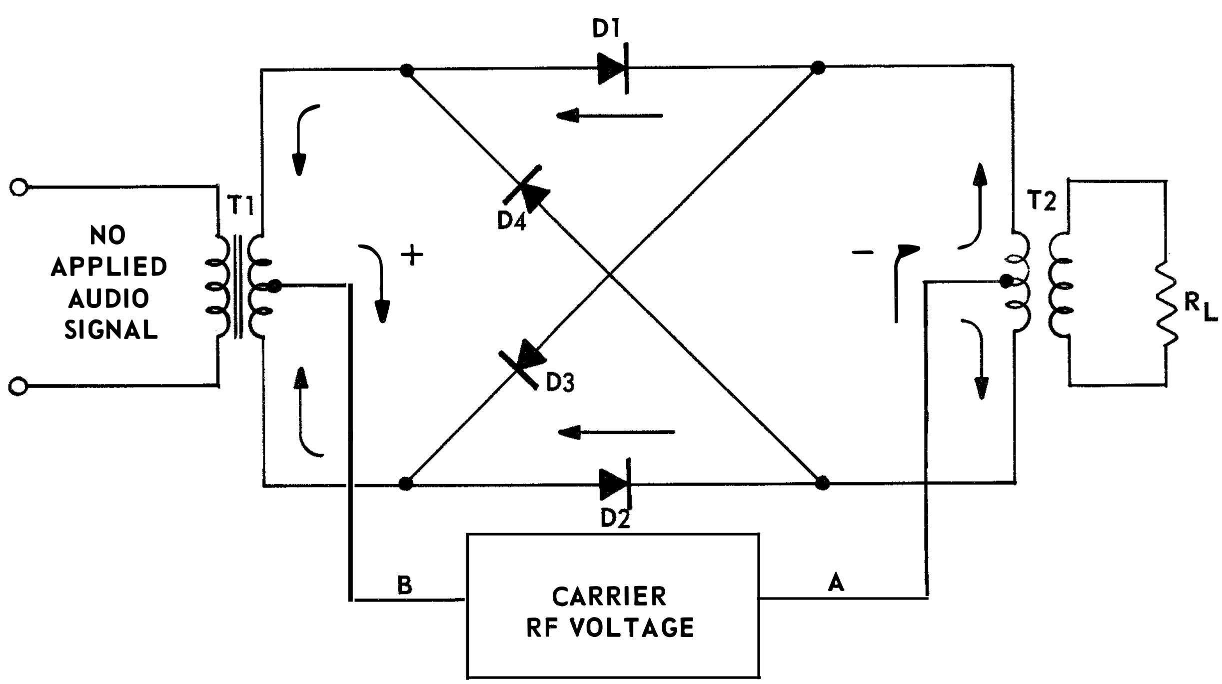


Figure 4-6. Carrier Current in Ring Modulator.

the two halves of the T2 primary are 180° out of phase. As a result, the net RF voltages which are developed cancel each other and no RF is coupled to the secondary. On the next carrier half cycle, the polarity will reverse and diode D3 and D4 will conduct. As shown in figure 4-7, the carrier current will now reverse direction but the currents in T2 primary will still be flowing in opposite directions. Thus, the RF voltages developed across T2 primary will cancel each other and there will still be no carrier voltage in the output. The diode balanced modulator can be more easily understood if the diode elements are considered as switches. A simplified schematic in figure 4-8A shows diodes D1 and D2 as closed switches and D3 and D4 as open switches. In this diagram, the circuit conditions correspond to the action occurring in figure 4-6. Figure 4-8B shows the conditions when the carrier voltage polarity reverses; now D1 and D2 are "open," D3 and D4 are "closed."

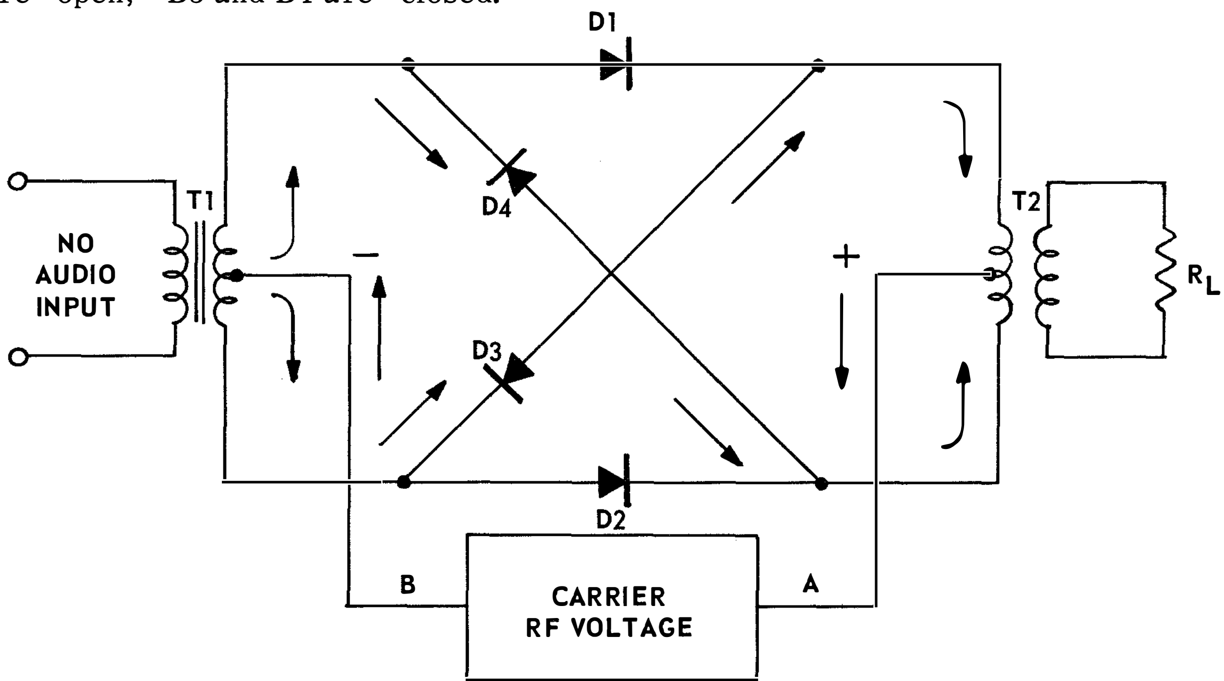


Figure 4-7. Carrier Current Paths When Polarity Reverses.

These conditions correspond to the action taking place in figure 4-7. The switching voltage is the carrier signal and the switching rate is determined by the carrier frequency. In order to obtain good switching action and to reduce distortion in the modulator output, it is general practice to use a carrier signal which greatly exceeds the audio signal level.

4-23. When an audio signal is applied to the primary of T1, the audio voltage across the diodes will unbalance the modulator circuit. The resultant RF voltage across load resistor R3 will consist of a series of pulses whose polarity and repetition rates are determined by the carrier voltage, and whose amplitude is controlled by the audio voltage. When viewed on a spectrum analyzer, the output signal will be seen to contain

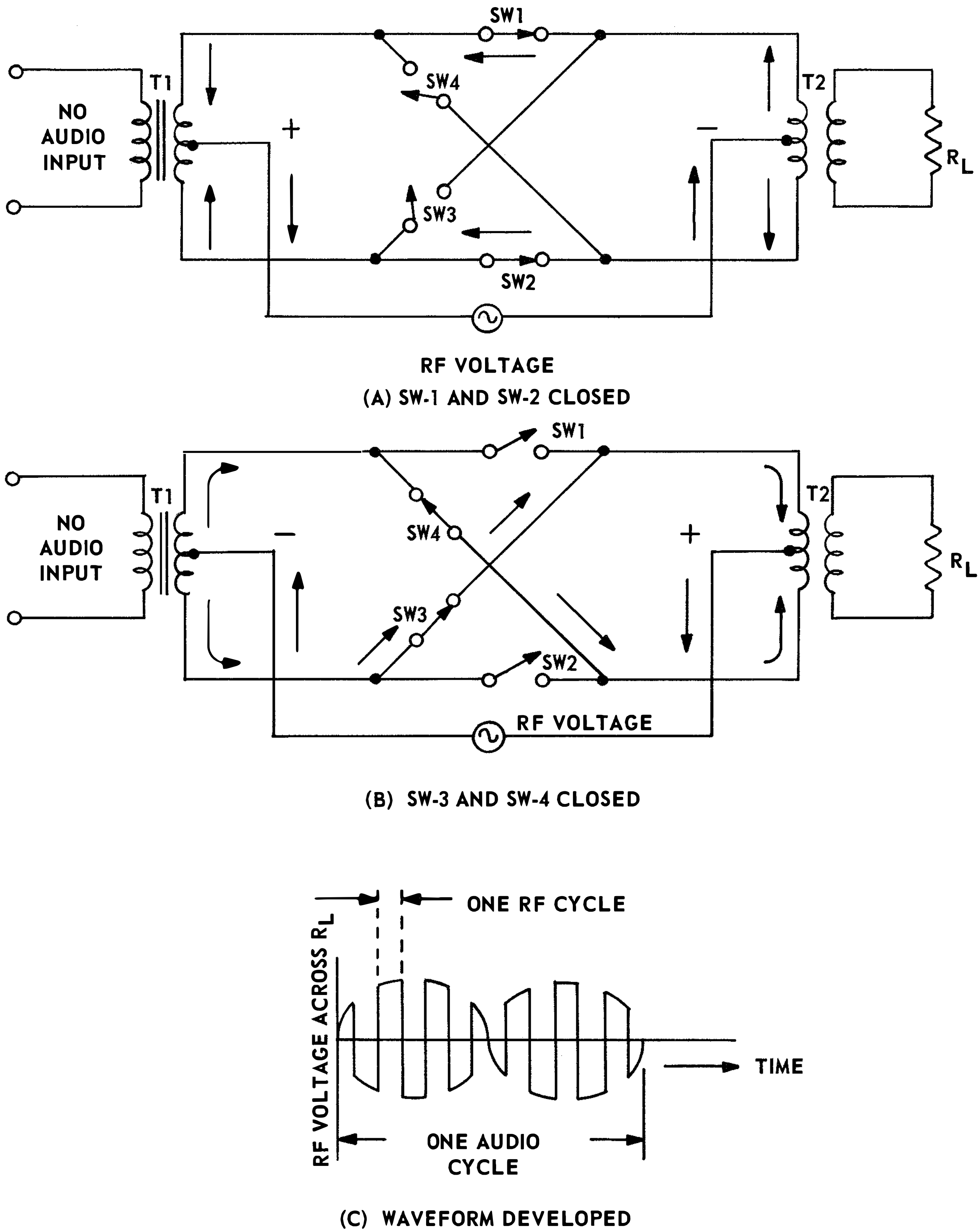


Figure 4-8. Carrier Current Paths in Ring Modulator.

an upper and lower sideband. A similar pair of sidebands (and some other undesired higher frequency products) will be placed about the second harmonic of the carrier frequency. The upper and lower sidebands displaced around the carrier frequency are the sum and difference RF signals described previously. A single tone-modulating signal and the resultant waveform of the balanced modulator output are shown in figure 4-8C.

4-24. Diode balanced modulators are capable of a high degree of carrier balance and will retain this characteristic over long periods of time if the components are of good quality and are accurately matched. In a well designed system, it is not difficult to obtain a carrier suppression of 40 db or better, and the level of third-order intermodulation products can be made 50 db below the desired sideband output signal.

4-25. Four-Diode Shunt Balanced Modulator.

4-26. Figure 4-9 shows a balanced four diode modulator. All four diodes conduct during one half of the RF cycle only; during the other half cycle they do not conduct at all. If you can visualize the diodes as switches controlled by the carrier voltage polarity, the action will be similar to that shown in figure 4-10. While the diodes are at their conducting peak, the input and output circuits are virtually short circuited, as shown in figure 4-10A. Conversely, in figure 4-10B the high impedance conditions are restored while the diodes are "open." If the four diodes are carefully matched, the

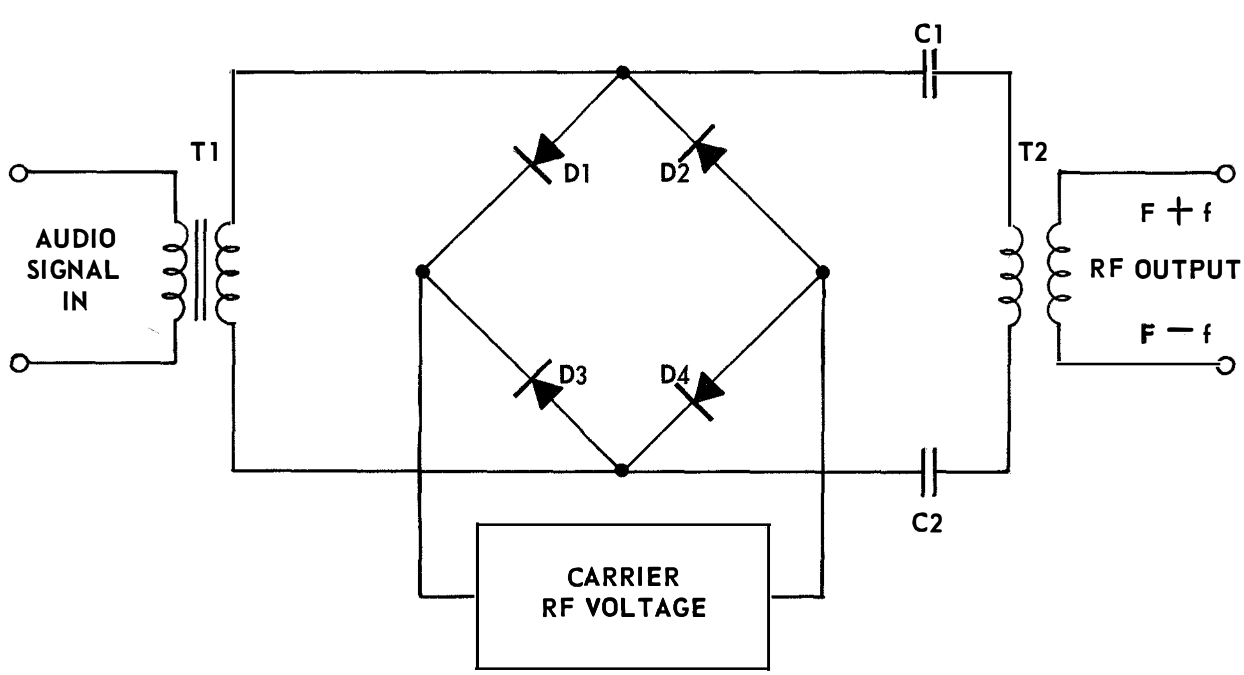


Figure 4-9. Basic Shunt Balanced Mod.

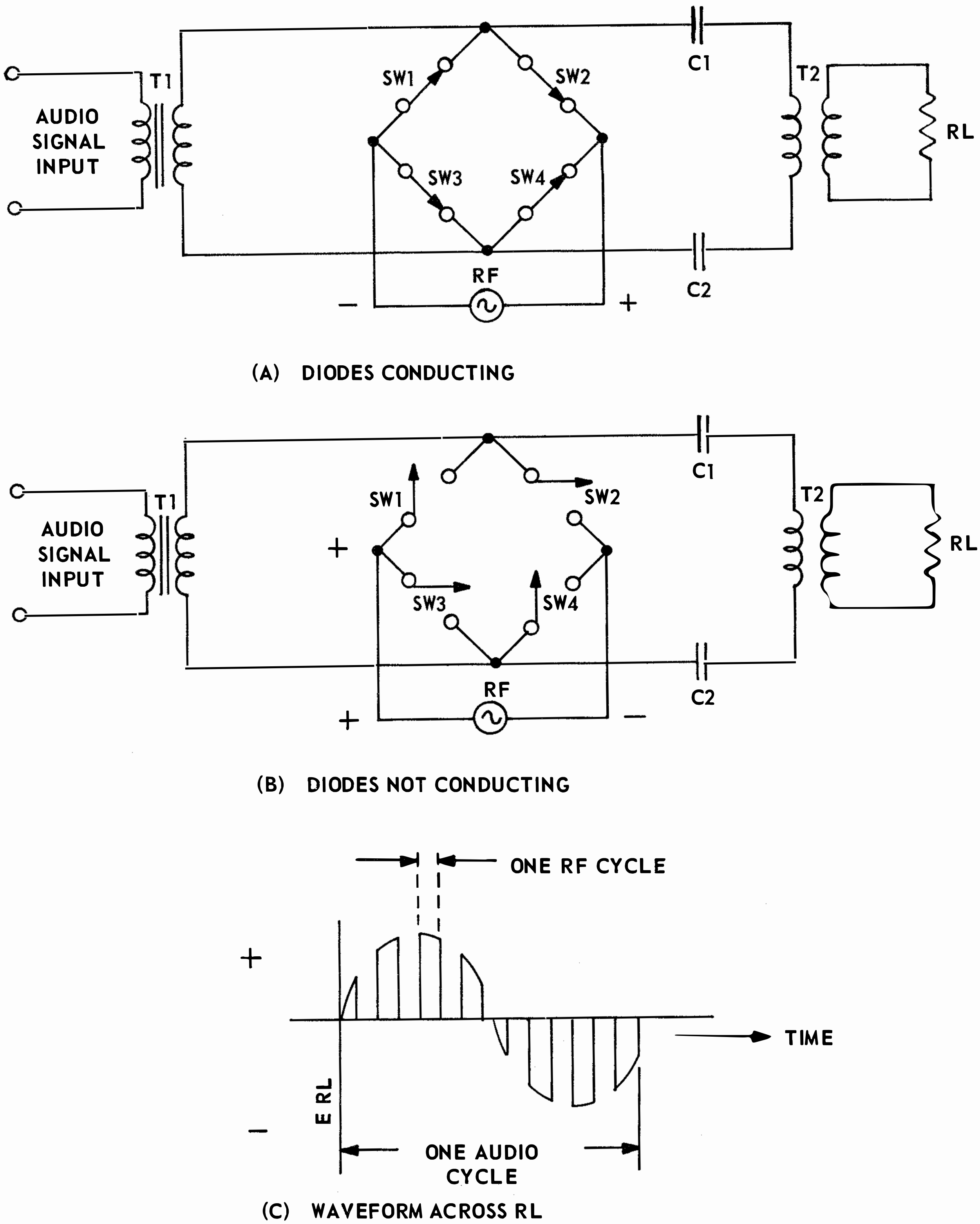


Figure 4-10. Diode Switching Action.

circuit will be balanced for the RF carrier and no modulated signal will appear in the output. Since the diodes conduct only on the negative portions of the RF cycle, the output currents flow for approximately half a cycle of the carrier frequency. However, in a practical circuit these half-wave signals are fed into a resonant tank circuit and appear as normal full cycles at the output. A single-tone audio-modulated signal and the resultant modulated waveform are shown in figure 4-10C. As described previously, the modulated RF signal will contain sum and difference signals known as sidebands.

4-27. Two-Diode Shunt Type Balanced Modulator.

4-28. A shunt diode modulator, using only two diode elements, is depicted in figure 4-11. The 1N35 varistor is used in place of the diode elements and the circuit values are typical for operation in the 450- to 460-Kc region. Notice that D1 and D2 form one-half of the bridge circuit, and the resistors R1, R2, and R3 form the other half. The 250-ohm pot, R2, is used to balance the circuit with the RF carrier applied. Known as the "carrier balance control," it is usually a screwdriver adjustment somewhere on the main chassis. Normally, this adjustment will be made during initial alignment and then left alone. The two 820-ohm resistors, R1 and R3, prevent one side of the carrier source from short circuiting to ground when the potentiometer arm is rotated to either extreme. A 2000-ohm pot is sometimes used across the carrier source instead of the three resistors as shown. The applied carrier signal must be taken from a

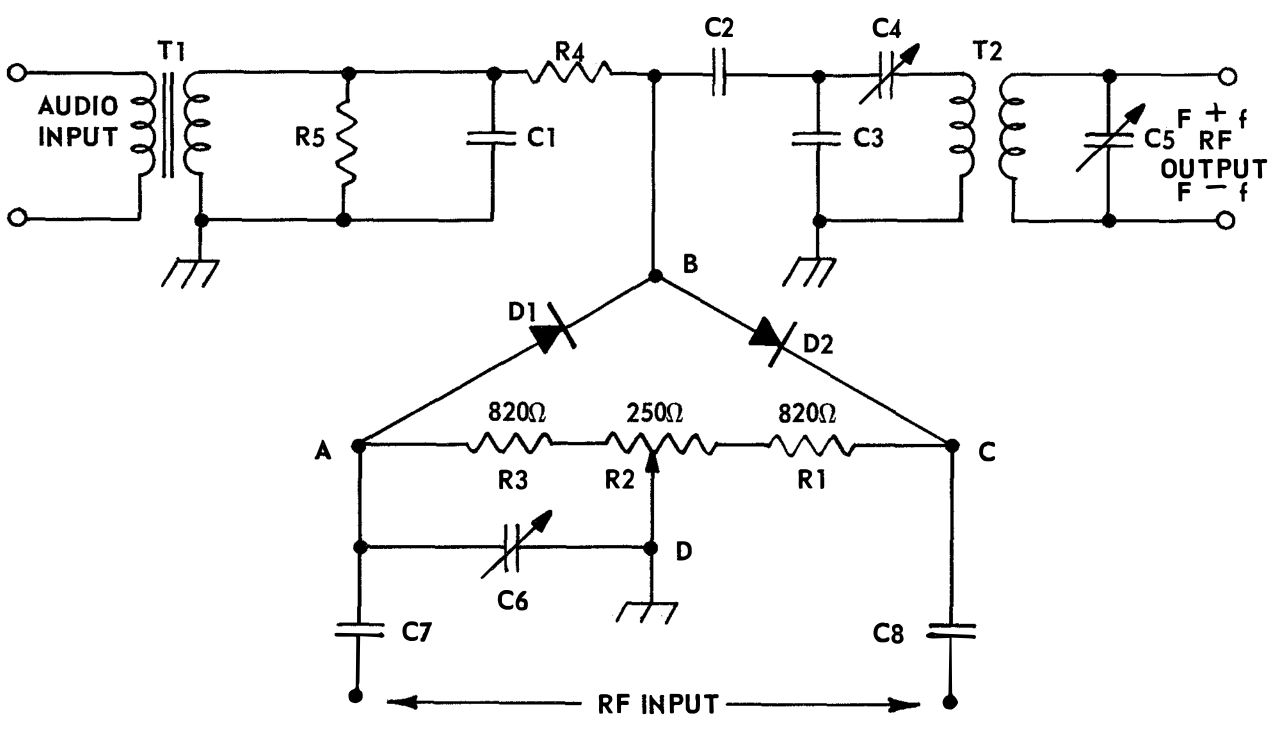


Figure 4-11. Shunt Balanced Modulator.

balanced source. The signal may be supplied to C7 and C8 from an RF transformer secondary winding, or from the output of an RF phase-inverter. This subject will be covered in more detail later.

4-29. Assume, in figure 4-11 that there is a carrier signal of about 10 volts across points A and C, but no applied audio. When point A is more positive than B, diodes D1 and D2 will conduct, but will cut off when point A goes more negative. Current will flow from C, through D2, to B, and from B, through D1 to A and back to the generator. The current will also flow through R1, R2, and R3 as it goes from C to A. The modulator RF output circuit is connected between points B and ground. If the two diodes are perfectly matched and the RF voltage from A to ground and C to ground are equal, the modulator circuit will be well balanced and no RF carrier will appear in the resonant output circuit T2. However, such an ideal condition is rare; in most cases adjustment of the circuit will be required. Carrier balance control R2 is now adjusted to obtain minimum carrier signal in the secondary winding of T2. As the control is rotated from one extreme to the other, a definite null should be obtained at or near the center of the balance adjustment range. If no null appears, or if it appears at one extreme of the balance control, then the two diodes do not match closely enough or unequal RF carrier is being applied at points A and C. Variable capacitor C6 permits adjustment of the carrier so that equal RF voltages will be applied to A and C. It should be turned slowly from its minimum value while balance control R2 is being adjusted. If a satisfactory null is still not obtained, C6 should be removed from point A and connected to point B and the balancing procedure repeated.

4-30. The 1N35 varistor, which consists of a pair of matched 1N34 germanium diodes, is usually sufficient to permit a good balance with the circuit shown. A pair of matched 1N34 diodes, selected at random from dealer's stock, will rarely give optimum balance. If no 1N35 varistor is available, the 1N34's can be roughly matched by selecting those which exhibit similar front and back DC resistances when measured with a simple ohmmeter. When doing this, be sure to measure resistance on a meter scale where the needle is mid-range.

4-31. When the audio signal from T1 is applied to the diode circuit between B and ground, diodes D1 and D2 will alternately pass RF current during the audio frequency cycle. When point B is more positive than ground, due to the audio signal, D1 will be back-biased and will not conduct until the positive-going carrier voltage exceeds the audio voltage. On the other hand, diode D2 is forward-biased by this positive audio voltage and will conduct over slightly more than one-half of the RF cycle. Thus, the modulator is unbalanced by the audio signal, and RF voltage will now appear in the resonant output circuit T2. When the audio signal goes negative at point B, the conditions will reverse; D2 will now be back-biased and D1 will conduct. The modulator is again unbalanced by the audio signal, and RF voltage still appears in the output circuit. Since the modulator unbalance is taking place at an audio rate, the RF carrier and the audio signals will combine, or "mix," to produce sideband signals in the output circuit. If the carrier itself has been satisfactorily suppressed no carrier component should be present in the output during modulation. If there is, this means the frequency stability of the carrier generator is poor under load. As a result, an FM component will be introduced into the output circuit. This condition rarely occurs when the carrier generator is crystal controlled, but it can be a serious problem with generators employing self-controlled oscillators.

4-32. Any shunt-connected diode modulator must be so designed that the proper impedances are presented to the AF and RF signals. In figure 4-11, T1 is a miniature tube-to-line transformer with a primary impedance of 20,000 ohms and a secondary impedance of 600 ohms. The secondary winding is shunted by a 1000-ohm resistor and a 330 $\mu\mu$f capacitor. Another 1000-ohm resistor (R4) is connected in series with the high side of the transformer and the junction point, B, of the two diodes. The two resistors form a voltage divider for the audio signal, and R4 offers a relatively high impedance to the RF signal at point B. In some applications, R4 may be replaced by a suitable RF choke. In the direction of the RF output circuit, the audio signal sees the high impedance of the .001-$\mu\mu$f capacitor, C2. The only load for the audio signal is the diodes themselves. The RF signal at point B, looking toward the output circuit, sees the low impedances of C2 and C3 of the series resonant transformer primary. The two .001-$\mu\mu$f capacitors, C2 and C3, should be of good quality and have silver mica dielectric. The variable capacitor C4 is the IF trimmer normally connected across the primary winding.

4-33. The shunt type of balanced modulator is more suitable for use in circuits where the output impedance can be made low by series tuning, as has been done here. The Collins series of mechanical filters utilizes a tuned input circuit which may be either series or, in special cases, parallel-resonated. This circuit may also be used with certain types of crystal lattice filters which either may employ tuned inputs or be matched by means of a capacitance voltage divider.

4-34. Series Balanced Modulator.

4-35. Some filters must be terminated in a definite impedance value. The Burnell S-15000, for example, must be terminated in a 30,000-ohm input impedance in order to retain the proper response characteristics. This input termination value will not match the output impedance of the shunt modulator; hence some kind of intermediate coupling device would seem necessary to obtain a satisfactory match between the modulator and filter. However, such a coupling device would not only mean higher cost and additional circuit complexity, but might also increase circuit losses. In circuits using terminated or high impedance filters after the balanced modulator, a series modulator like the one in figure 4-12 is more suitable. Another advantage of the series modulator is that no audio input iron-core transformer is needed. The AF cathode follower serves the same purpose, but costs less and requires less space. Two basic requirements of the series balanced modulator are that the impedance across which the audio frequency is developed be low at the carrier frequency, and that the impedance across which the RF is developed be low at the audio frequency. In practice, this requirement

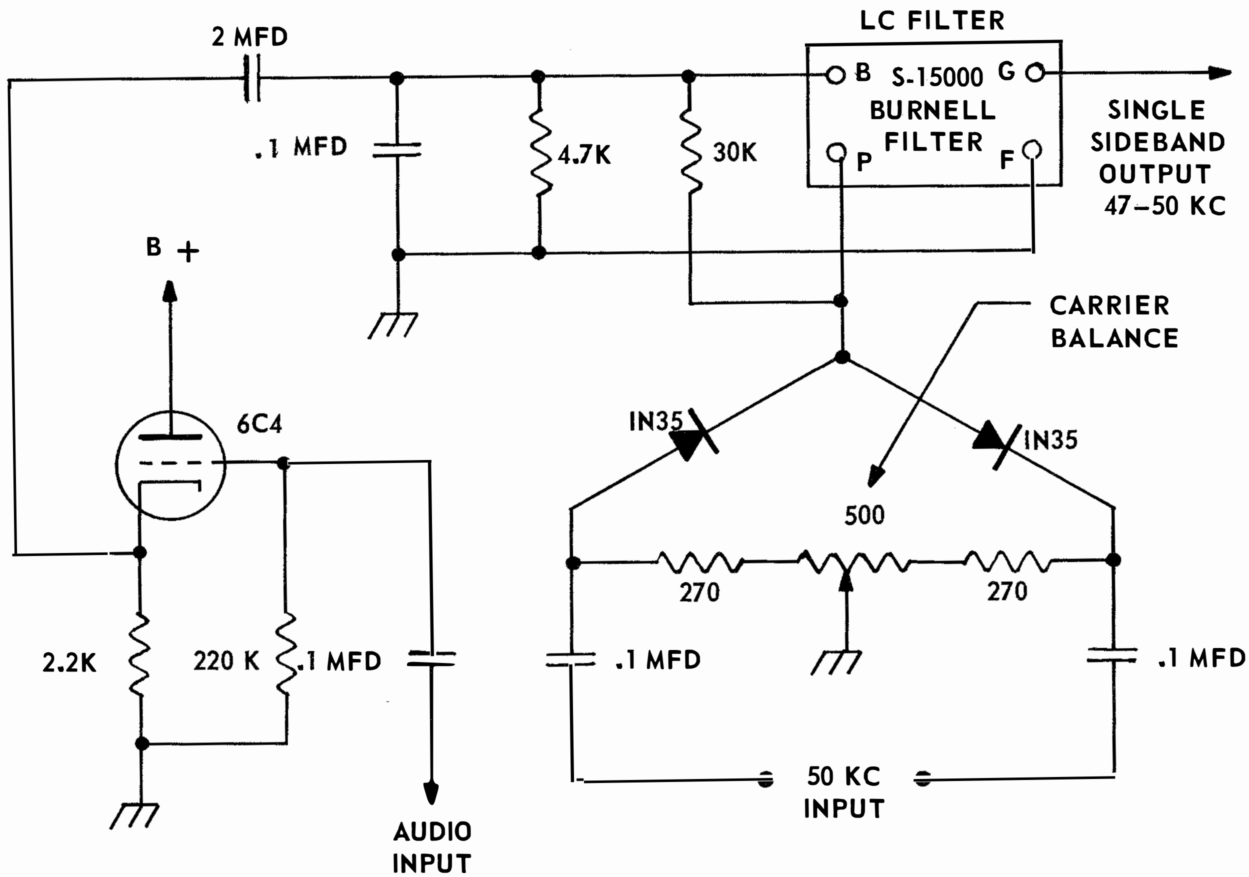


Figure 4-12. Series Balanced Modulator Using L-C Filter, Burnell S-15000.

is satisfied by the circuit arrangement in figure 4-12. The circuit values shown are suitable for use with the 50-Kc Burnell S-15000 filter. Those in figure 4-13 are suitable for the 455-Kc Collins mechanical filters. The general theory of operation of the series balanced modulator is similar to that of the shunt type described previously.

4-36. The balanced modulators described above are considered rather standard arrangements and are used extensively in modern sideband equipment. The actual circuit configurations may vary somewhat with different sets, with transistors instead of tubes, but the basic functions will remain the same. Any of these circuits, when used properly, will produce a satisfactory suppressed-carrier, double-sideband signal. In most cases, the circuit choice will be determined by the type of components used, such as the sideband filter, and not by any electrical differences between circuits.

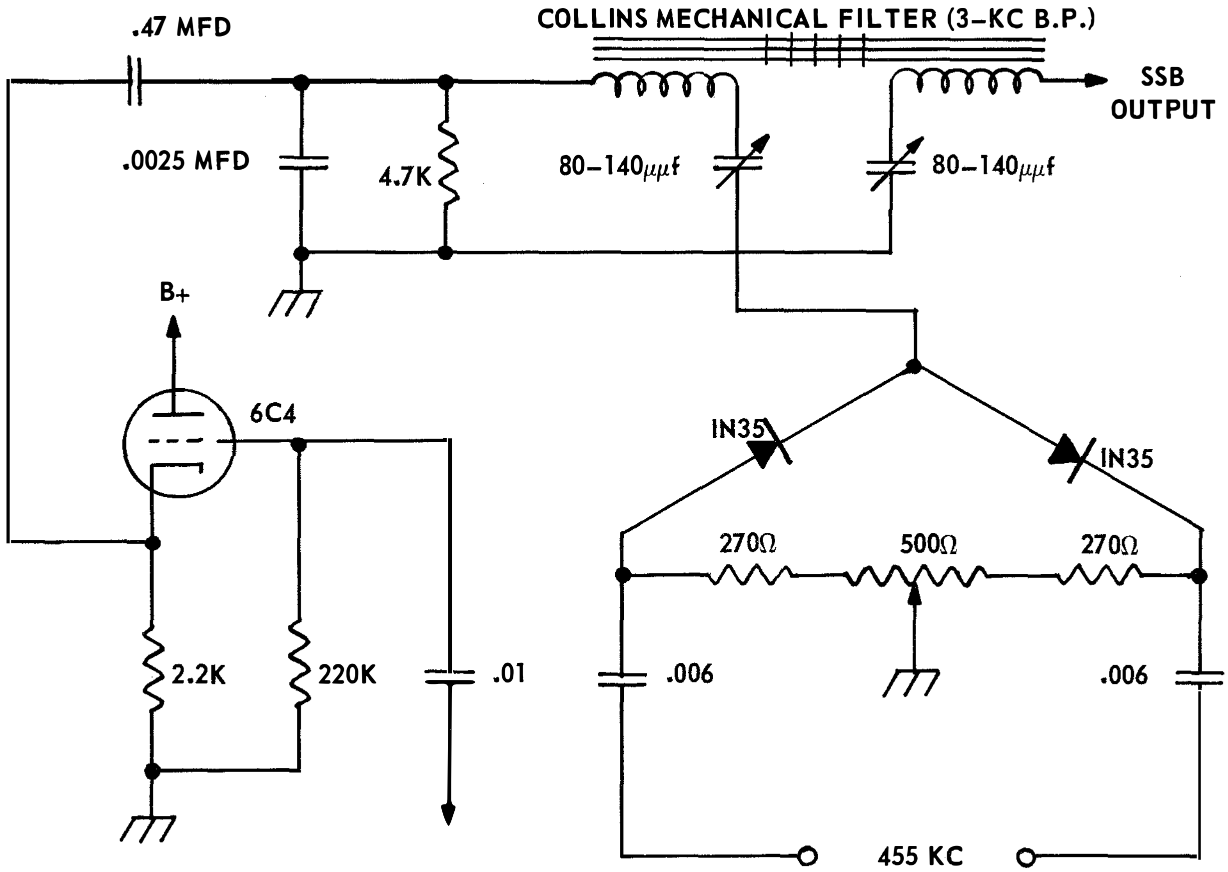


Figure 4-13. Series Balanced Modulator
Using Collins Mechanical Filter.

<u>4-37.</u> <u>Balanced Modulator Using a Beam Deflection Tube.</u>

4-38. Recently, a totally different type of balanced modulator circuit has been developed which uses a beam-deflection tube as the modulator. The sheet-beam switching tube was originally designed for use as a synchronous demodulator in color TV receivers. The first of these tubes produced commercially was the 6AR8, manufactured by GE. The 6AR8 was designed primarily for color TV; the fact that it could also be adapted to sideband generation was merely coincidental. Now, a later version of the basic tube is manufactured by RCA. Known as the 7360, it is designed especially for balanced modulator and converter service and is certain to be widely used in the near future.

4-39. The 7360 is a nine-pin miniature tube featuring a unique electrode structure which consists of two plates, two deflection electrodes, a cathode, and two grids. The total beam current is determined by the potentials applied to the two grids, and the portion of the total beam current collected by each plate is determined by the voltage difference between the two deflection electrodes. The current collected by each plate thus becomes a function of two voltages. Because of this unique design, the 7360 is particularly useful in modulator, demodulator, and frequency-converter service, and in equipment operating at frequencies up to 100 megacycles. The tube also has inherent long-term stability because the electron flow to each plate is derived from a single beam. Other features of the tube are high transconductance (on the order of 5500 micromhos), high deflection sensitivity, low inter-electrode capacity, and good shielding. (This prevents interaction between the deflection electrodes and control grid.) In carefully designed balanced modulator circuits, the 7360 is capable of providing a carrier suppression of 60 db or better. In balanced mixer (frequency conversion) circuits, it is possible to obtain at least 40 db of oscillator signal suppression in the balanced output tank circuit.

4-40. In figure 4-14 the circuit symbols and base connections are shown for the 7360 tube. The balanced modulator circuit recommended by RCA is shown in figure 4-15. In operation, the sheet of electrons emitted by the cathode passes through the two grids and, if no deflection voltages are present, strike the open space between the two plates. Since the electrons do not strike either plate, no plate current flows and there is no output from the tubes. If the RF carrier signal is applied to the control grid, the amplitude of the beam will vary at the RF rate, but there will still be no output since the beam is not striking either plate.

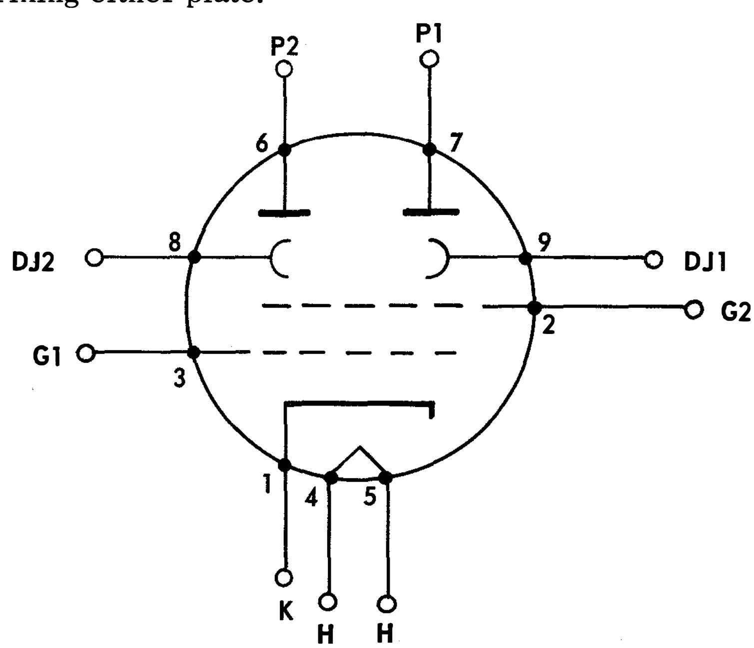


Figure 4-14. Symbols and Connections, RCA 7360.

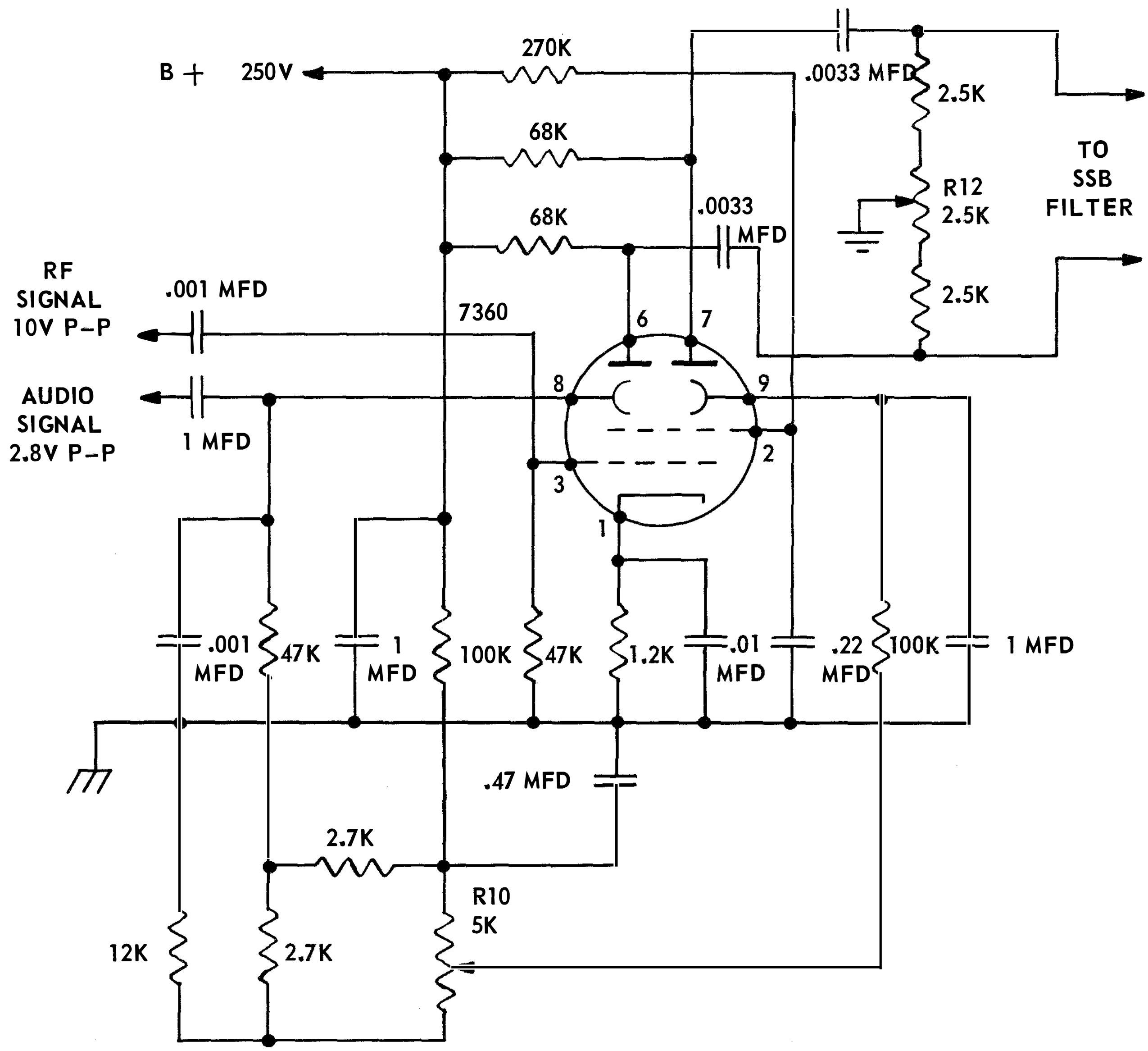


Figure 4-15. 7360 Balanced Modulator Circuit.

4-41. An electron beam may be deflected by either an electrostatic or electromagnetic field. The construction of the 7360 is such that the electrons are made to pass between the two deflection electrodes (but not contact them). Thus, the application of voltage (either AC or DC) to the electrodes will deflect the beam and cause it to strike one of the plates. When the polarity of the deflection voltage is reversed, the beam will be deflected in the opposite direction and strike the other plate. It is obvious that the beam can be swept back and forth at either an audio or radio frequency rate by application of a suitable signal at the deflection electrodes. In balanced modulator service, the audio is applied to the two deflection electrodes. The peak-to-peak audio signal required for full beam deflection is approximately 2.8 volts. The RF carrier signal is applied to the control grid; its peak-to-peak amplitude must be approximately 10 volts. When the 7360 is used as a balanced frequency converter, the conversion oscillator signal is applied to the control grid, and the sideband RF signal is applied to the deflection electrodes. In either case, the RF signal applied to the control grid does not appear in the output.

4-42. Because of physical and electrical variations inherent in the manufacture of vacuum tubes, balancing and centering devices must be used with the 7360 in order to obtain optimum carrier suppression and the desired signal output. In figure 4-15, a positive DC potential of about 25 volts is applied to the second deflection electrode (terminal 8) by means of a voltage-divider network across the B+. The DC voltage applied to the first deflection electrode (terminal 9) is made adjustable from about 20 to 30 volts by R10 (the 5000-ohm carrier-balance pot). Output circuit balance is adjusted by the 2500-ohm pot, R12. In operation, R10 and R12 are alternately adjusted for minimum RF signal in the output circuit. When the 7360 is followed by a sharp-cutoff sideband filter, it is possible to obtain carrier suppression in the order of 70 to 80 db below the desired sideband output voltage.

4-43. Transformerless Balanced Modulator.

4-44. A type of balanced modulator circuit quite different from those previously discussed is one which uses no transformer. This circuit, figure 4-16, 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 RF 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 resistor and effectively apply negative voltage to the grid of V2B. The phase of the voltage developed at 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 RF 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 feedthrough which may occur in V1 and V2B. Upper and lower sideband frequencies will be generated in this circuit while the two input signals will be suppressed. Cathode resistor R6 of V2B provides the proper operating point for this tube with capacitor C4 acting as an RF bypass. If so desired, this RC combination can be eliminated and a separate DC bias supply can be connected to the grid of V2B. 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 percent modulation.

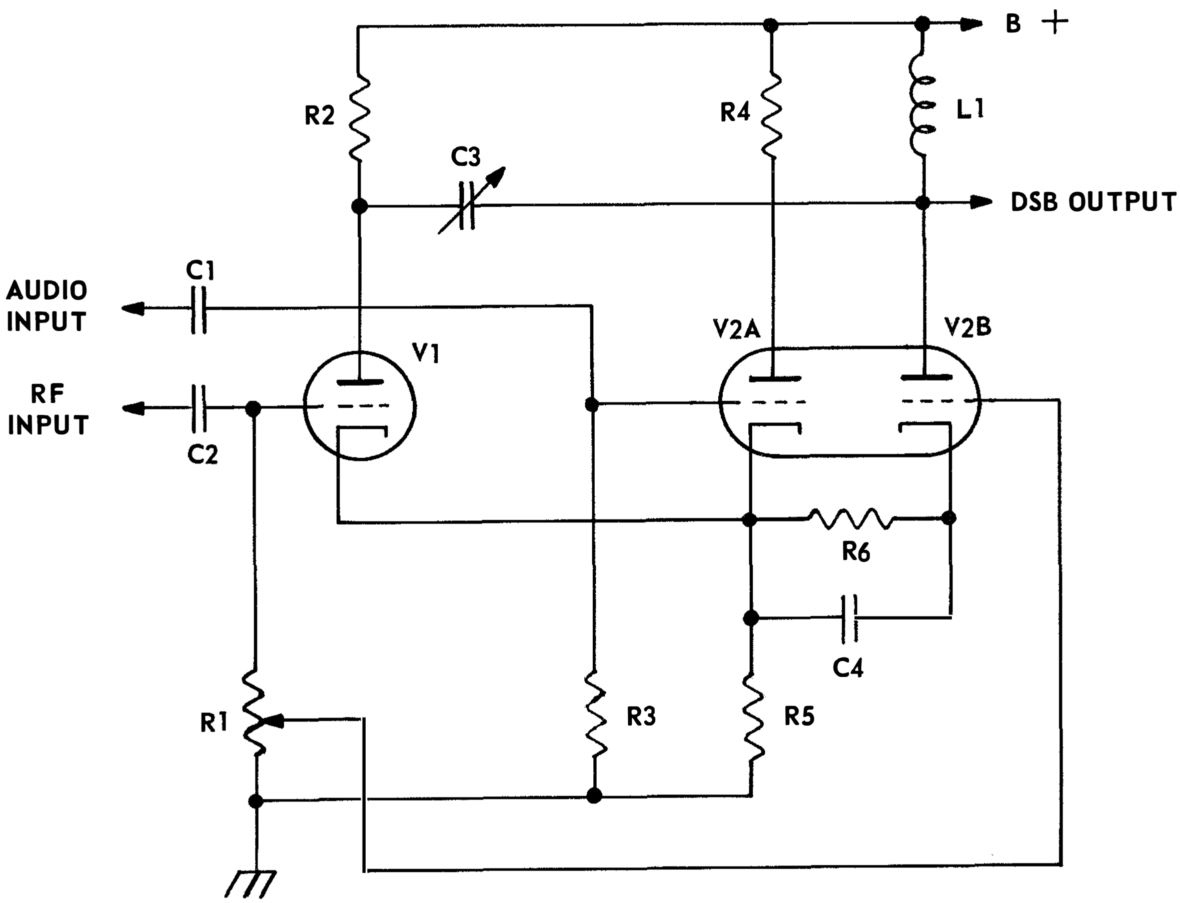


Figure 4-16. Transformerless Balanced Modulator.

SECTION V

SIDEBAND SELECTION

5-1. General.

5-2. The balanced modulators described in the preceding section all produce an output consisting of a pair of sidebands located on either side of the suppressed carrier. Since the objective is to transmit only a single sideband, some means must be used to select one sideband and to suppress the other. In most modern single sideband systems, this is accomplished by either frequency (filtering) or phase (phase shift) discrimination.

5-3. Filter Method of Sideband Selection.

5-4. The filter method of sideband selection originated with the carrier system of wire telephony. The filter itself is simply a device which presents a high impedance to unwanted frequencies and thereby weakens (attenuates) them, but offers a relatively low impedance to the desired frequencies. The band of frequencies passed by the filter is generally called the filter passband. However, no filter is perfect; they all attenuate somewhat the frequencies contained in the passband. This attenuation is called the filter insertion loss. At the upper and lower frequency extremes of the passband, the filter rapidly attenuates those frequencies which lie in both directions away from the filter bandpass. When these frequencies are plotted along the X-axis of an X-Y coordinate system and the attenuation in decibels is plotted along the Y-axis, the ratio of frequency change to attenuation is called the skirt selectivity. A similar plot of the entire filter response yields the characteristic curve. The bandwidth of a filter is the difference in cycles, or kilocycles, obtained when the lowest frequency contained in the passband is subtracted from the highest frequency. The two reference frequencies are usually specified at some attenuation point along the skirt, or slope, of the filter response and will vary with different manufacturers. In the Collins F455Y-31 mechanical filter, for example, the bandwidth at 6-db attenuation is specified as 3.1 Kc; at 60-db attenuation, the bandwidth is 6 Kc. The ratio of the bandwidth at 6- and 60-db attenuation points is called the shape factor. The Collins filter has a shape factor of slightly less than 2.0, which is considered very good.

5-5. The original filters developed by the commercial telephone companies were designed to operate at very low frequencies, 10 to 20 Kc. These early filters invariably were derived from combinations of fixed capacitors and high-Q inductors. The characteristic impedances were generally made low to prevent the distribution capacitance of the inductances from deteriorating the over-all filter performance. Some time later the crystal filter came into wide use. Crystal filters are usually based on full-lattice configurations and might contain as many as 80 to 100 individual units. These filters are carefully designed to present a skirt attenuation of 70 to 80 db in less than 50 cycles.

5-6. Low-Frequency Filters.

5-7. Most of the filters in single sideband equipment are considerably more simple than those used in carrier system telephony. However, it is obvious that designing highly selective RF filters requires extremely complex calculations and very precise measurements beyond the scope of this course. For those interested in this subject, there are a number of texts and scientific papers available. With the exception of the relatively simple crystal-lattice filter described later, our discussion will be confined to the various manufactured filters available as stock items.

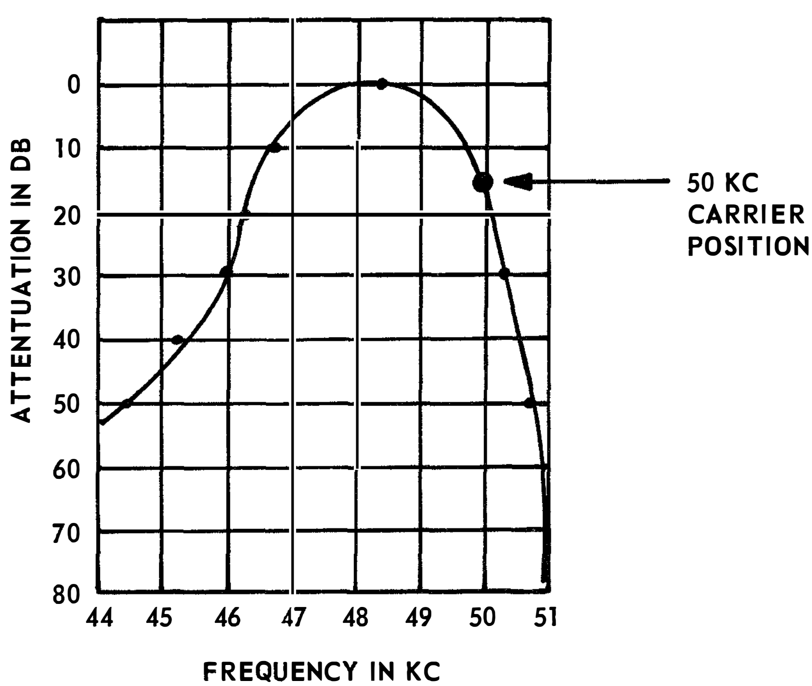


Figure 5-1. Bandpass Characteristics of Burnell S-15000 Filter.

5-8. The most widely known low frequency filter is the Burnell S-15000, designed for use at a carrier frequency of 50 Kc. A careful examination of the response curve in figure 5-1 will reveal that this filter passes only those sideband frequencies contained in the region from the 50-Kc carrier down to 47 Kc. This is the lower sideband. All sideband frequencies higher than 50 Kc are rapidly attenuated to the point where virtually none appear at the filter output. The 50-Kc carrier is placed at approximately the 20-db attenuation point on the slope of the filter response. At 50.5 Kc the upper sideband is down 50 db, and at 51 Kc it is down 80 db. For all practical purposes, the upper sideband frequencies are considered to be completely suppressed, since none appear in the output signal. Notice that the lower sideband frequencies are attenuated more gradually. At 47 Kc, the lower sideband is down about 5 db, and at 46 Kc it is down about 25 db. The 20-db attenuation point on the low frequency slope occurs at about 46.3 Kc. If the filter bandwidth is considered to be the frequency difference between the two 20-db attenuation points, then this filter has a bandpass (or bandwidth) of 3.7 Kc. All attenuation figures are given with respect to the maximum peak signal output from the filter, which in this case is 0 db. The insertion loss is not indicated for the filter, since all measurements are made at its output terminals. The carrier is placed at the 20-db attenuation point on the sharp slope of the filter response. This is done to attenuate the speech frequencies below 300 cycles, and also to give another 20 db of carrier suppression in addition to that of the balanced modulator. If the carrier is placed too high on the filter slope, or if the filter attenuation is too slow in the direction of the unwanted sideband, a portion of the unwanted sideband will appear in the output signal, and we would then have what is called vestigial sideband. Because of the filter-response slope, the two sideband components closest to the carrier will be unequal in amplitude. As a result, angle modulation will take place. In single sideband transmitters this causes a rapid increase in distortion products and can usually be identified by the lack of quality in the transmitted signal. Angle modulation can also be caused by 60- or 120-cycle hum components reaching the audio circuit of the modulator. This condition is especially serious when the carrier-generator frequency stability is poor under load.

5-9. In addition to extremely accurate and stable carrier generators and filters, most modern single sideband systems also make use of bandpass filters to restrict the audio frequency range. The band of audio frequencies between 300 and 3000 cycles is generally considered adequate for the transmission of intelligible speech. Those below 300 cycles waste transmitter power and contribute virtually nothing to signal intelligibility. The main reason for restricting the upper audio-frequency range to 3000 cycles is to prevent adjacent-channel interference. The design of single sideband audio circuitry will be covered in detail in a later section.

5-10. The Burnell S-15000 sideband filter is of unbalanced input type, where the input must be terminated in a resistance of 30,000 ohms in order to retain the characteristic response. Because of the unbalanced input termination, the filter cannot be used with balanced modulators unless an intermediate balanced-to-unbalanced coupling device is employed. Figure 5-2 shows the method of using the S-15000 filter with a push-pull output type of balanced modulator.

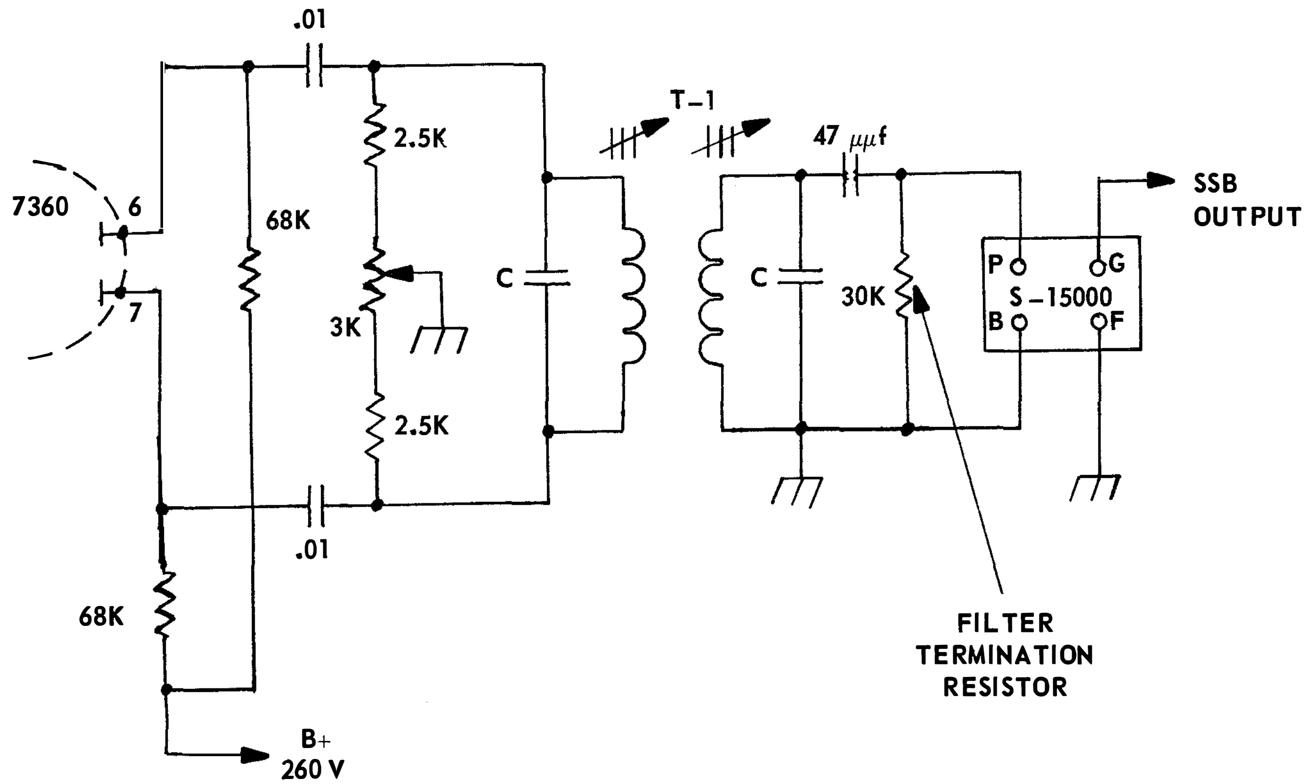


Figure 5-2. Burnell S-15000 Filter Used in 7360 Balanced Modulator Circuit.

5-11. Medium-Frequency Filters.

5-12. The early crystal-lattice filters for high-frequency single sideband use were generally in the 450- to 470-Kc region. The coupling devices were usually made from ordinary IF transformers. Commercially, the most widely known 455-Kc single sideband filter is one made by Collins. Produced in a number of shapes and bandwidths, Collins filters are used extensively in commercial single sideband equipment. The symmetrical bandpass characteristic of the mechanical filter simplifies the over-all single sideband generator design to some extent. To change the signal passed by the filter from an upper to a lower sideband, or vice versa, it is only necessary to change the frequency of the carrier-generator oscillator. As shown in figure 5-3, an upper sideband signal will be passed when the carrier frequency is at the 20-db attenuation point on the low-frequency slope. Likewise, the lower sideband will be passed when the carrier frequency is at the 20-db attenuation point on the high-frequency slope. In single sideband generators designed around the mechanical filter, the carrier frequencies are usually controlled by two crystals in the generator oscillator. The two crystals are usually connected to a sideband selector switch controlled from the front panel. Its two positions are generally marked "Lower sideband" and "Upper sideband," for lower and upper sideband transmission, respectively. Note, however, that the

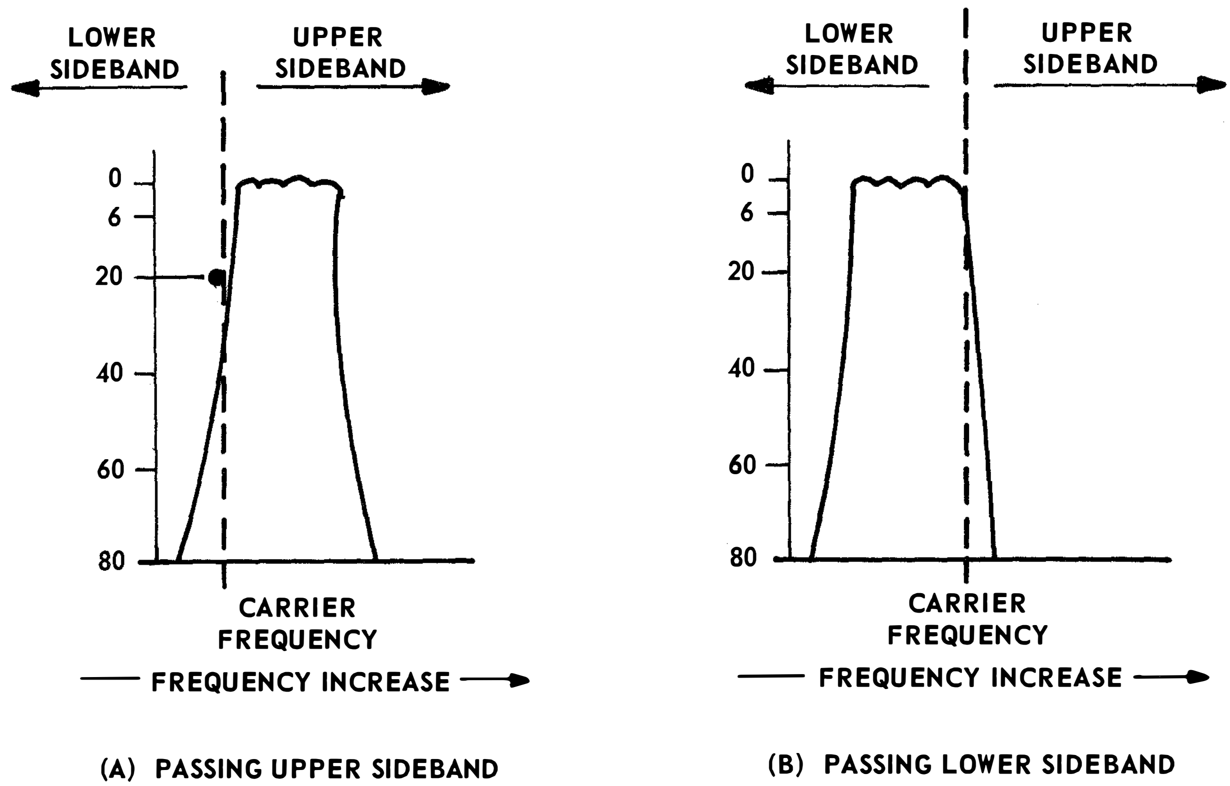


Figure 5-3. Sideband Selection Characteristics of the Mechanical Filter.

panel markings on most commercial equipment refer to the characteristic of the sideband transmission from the output terminals of the transmitter, and not necessarily to the sideband being passed by the filter. This subject will be discussed further when the heterodyning process of the sideband signal is taken up.

5-13. The input and output impedances of the mechanical filter are largely resistive, since the input and output circuits are tuned to resonance at the center of the passband. The actual impedance values will range from 1000 to 50,000 ohms, depending on the type of filter. The impedance and the recommended resonating capacitance are included on the data sheet packed with the filter unit. The tuning capacitors, which may be small ceramic trimmers, are not supplied by the manufacturer. The capacitance required to resonate the transducer coils varies from type to type, and may even vary slightly among filters of the same type. In most mechanical filters, the nominal capacitance is 120 $\mu\mu$f. A suitable value of capacitance can be obtained by connecting a ceramic trimmer adjustable from 8 to 50 $\mu\mu$f and a 100-$\mu\mu$f capacitor (a miniature, silver mica type) in parallel. The use of air-dielectric tuning capacitors, particularly relatively large ones, is not recommended. The large mass of metal in the stator-plate assembly may cause signal radiation or coupling around the filter. Either will result in incomplete suppression of the unwanted sideband, or other deterioration of the filter characteristic response.

5-14. The smallness of the mechanical filter makes it especially attractive for use in compact equipment such as portable or mobile transmitters and receivers. One of the recently developed tubular mechanical filters is only about 1/2 inch in diameter by 1-3/4 inches in length.

5-15. The lattice-type filter using quartz crystals has become very popular during the past few years. There is little doubt that the lattice filter played a large part in the enthusiastic response of users of single sideband equipment and designers of single sideband technology. In the full-lattice filter, crystals are used in a lattice filter configuration, as shown in figure 5-4. The bandwidth of the filter is determined by the frequency separation between crystals X1 and X2. This crystal arrangement is known as a full-lattice filter and is used extensively in single sideband telephone equipment. Probably the greatest difficulty in the construction of a full-lattice crystal filter is obtaining matched pairs of crystals. Each pair designated X1 or X2 must be within 10 to 20 cycles of the same frequency. Although crystals with a slightly greater frequency difference are usable, the filter response characteristic will deteriorate rapidly as the frequency separation increases.

5-16. The frequency separation between crystals X1 and X2 in figure 5-4 determines the bandwidth of the filter. In most lattice filters constructed from crystals and standard IF transformers, the bandwidth has been about 1-1/4 the frequency difference between the X1 and X2 frequencies. If two pairs of crystals 2 Kc apart are selected for the X1 and X2 frequencies, the bandwidth will be approximately 2.3 Kc. This is very close to the 2.1 Kc bandwidth of the filters used in some recent commercial single sideband transmitters. If crystals 1 Kc apart are used, the bandwidth of the filter will be approximately 1.75 Kc; this is generally considered too narrow for readable speech quality. It is possible, by using different separations of crystals, to produce filters with various bandwidths.

5-17. The half-lattice filter in figure 5-5 is similar to the single crystal filter used in older communications receivers, except that the phasing capacitor has been replaced by a second crystal. The shunt capacitances of the two crystals balance each other and produce a bandpass response sinilar to that shown in figure 5-6A. When the tuned circuits are properly adjusted, the response curve should have a flat top. However, its final shape factor will be poor due to a compromise between good skirt selectivity and a flat bandpass response. Most lattice filters will have two peaks, separated by about 2 Kc, with a dip of 3 to 6 db at the center between the two peaks. The shape factor (ratio of the bandwidth at the 6- and 60-db attenuation points) of the filter shown in figure 5-5 is about 4 or 5, which indicates a rather poor skirt selectivity. However, the shape factor can be improved considerably by the use of a 2-$\mu\mu$f trimmer across

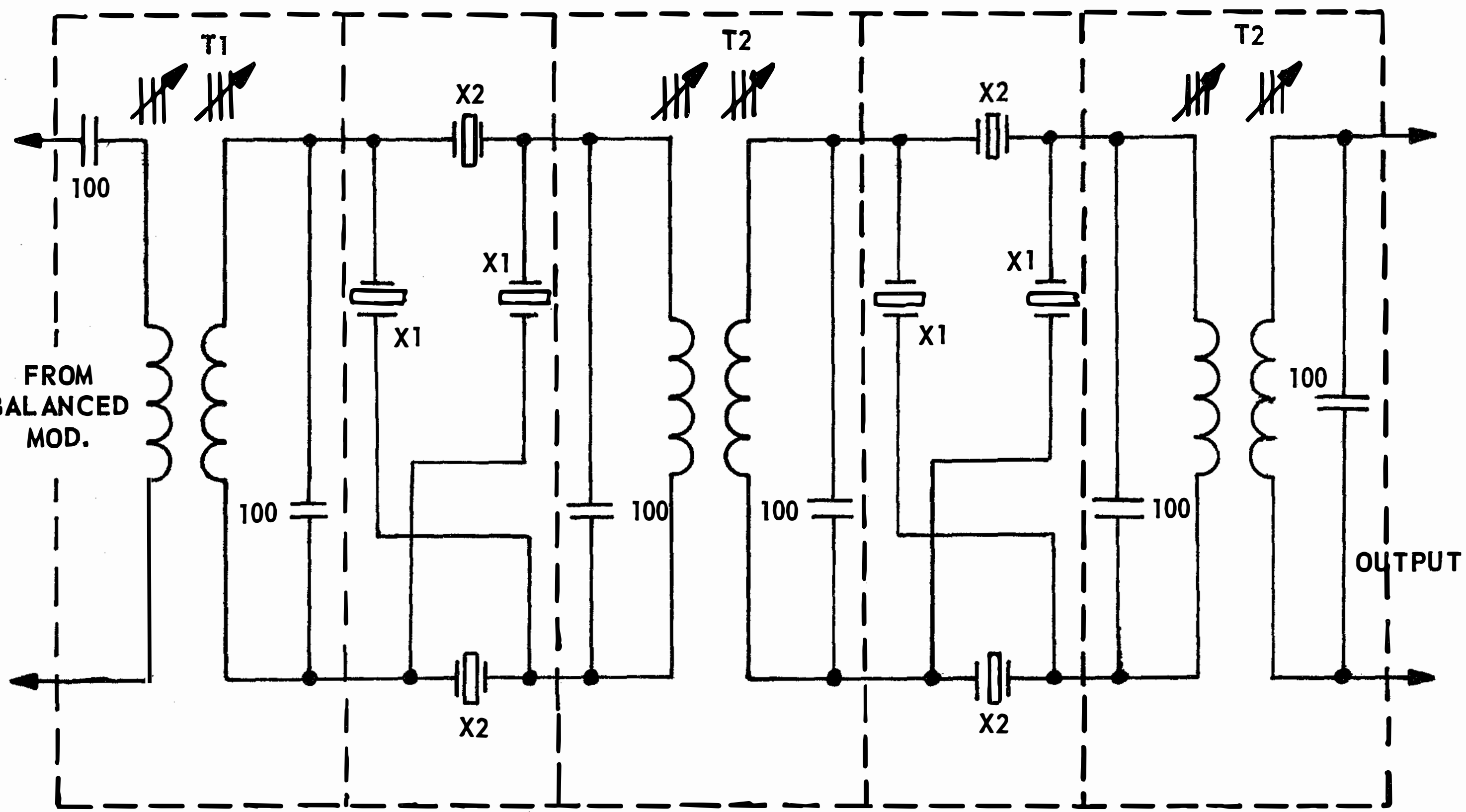


Figure 5-4. Schematic, Full-Lattice Crystal SSB Filter.

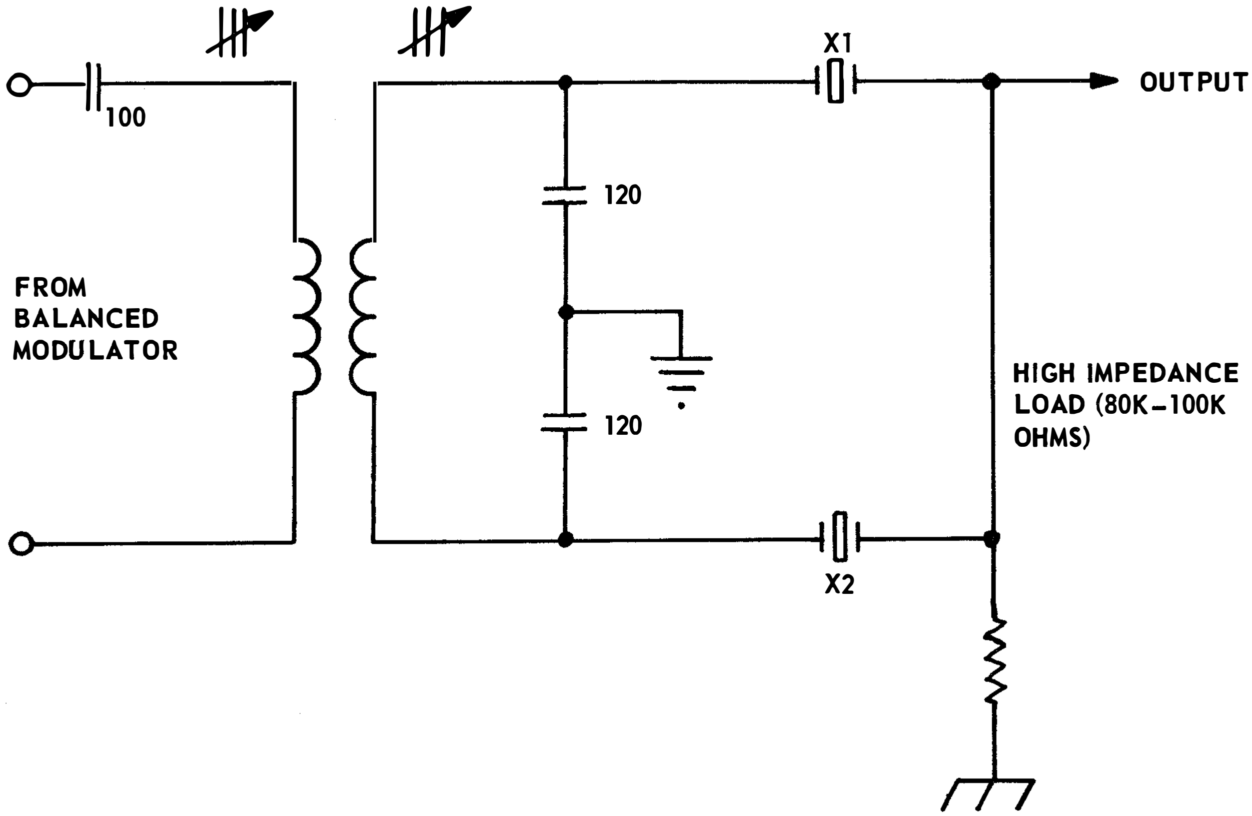


Figure 5-5. Half-Lattice Crystal Filter.

crystal X2, as shown in figure 5-7. The trimmer is adjusted until notches appear at the sides of the response curve, as shown in figure 5-6B. The optimum adjustment is at the point where the side responses just begin to appear. Beyond this point, the skirt selectivity will improve, but the two side responses will rise to unacceptable values. The side responses can be made insignificant by selecting crystals with series-resonant frequencies in the side-response region. When shunted across the tuned circuit as shown in figure 5-8, the crystals will appear as extremely high-Q traps. Generally, four crystals with 1-Kc separation are used, two of which are higher in frequency than X2, and two of which are lower in frequency than X1. The addition of the crystal traps across the tuned circuit will further improve the skirt selectivity of the filter.

5-18. High-Frequency Filters.

5-19. The crystal-lattice filters described in the preceding section have been more or less standard arrangements for a number of years. During the past few years, however, there has been even greater interest in crystal-lattice filters, especially in the 5- to 30-megacycle region. Commercial filters designed for up to 50 megacycles and higher have been available for a short time. The first commercial single sideband

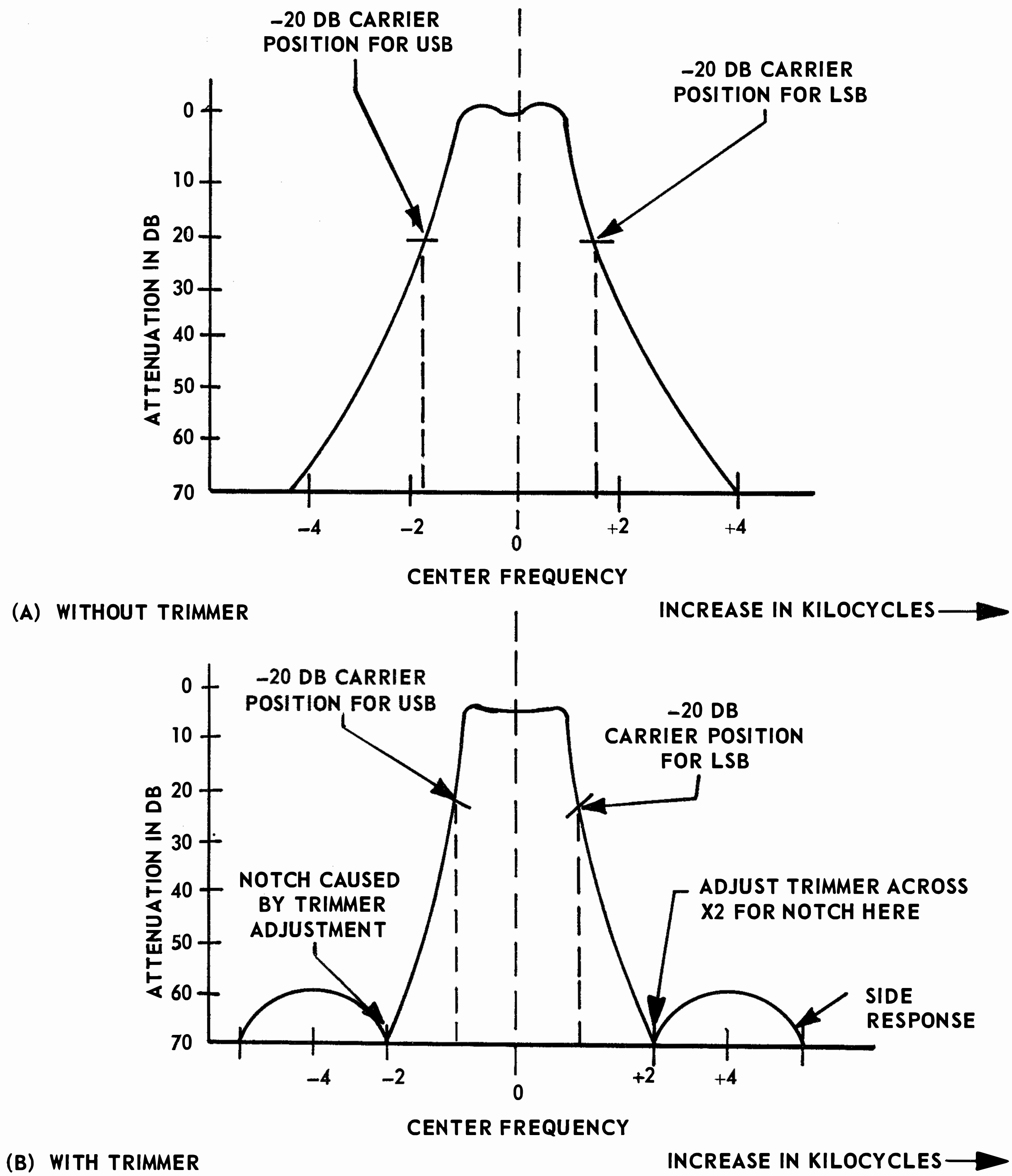


Figure 5-6. Half-Lattice Filter Response Curve.

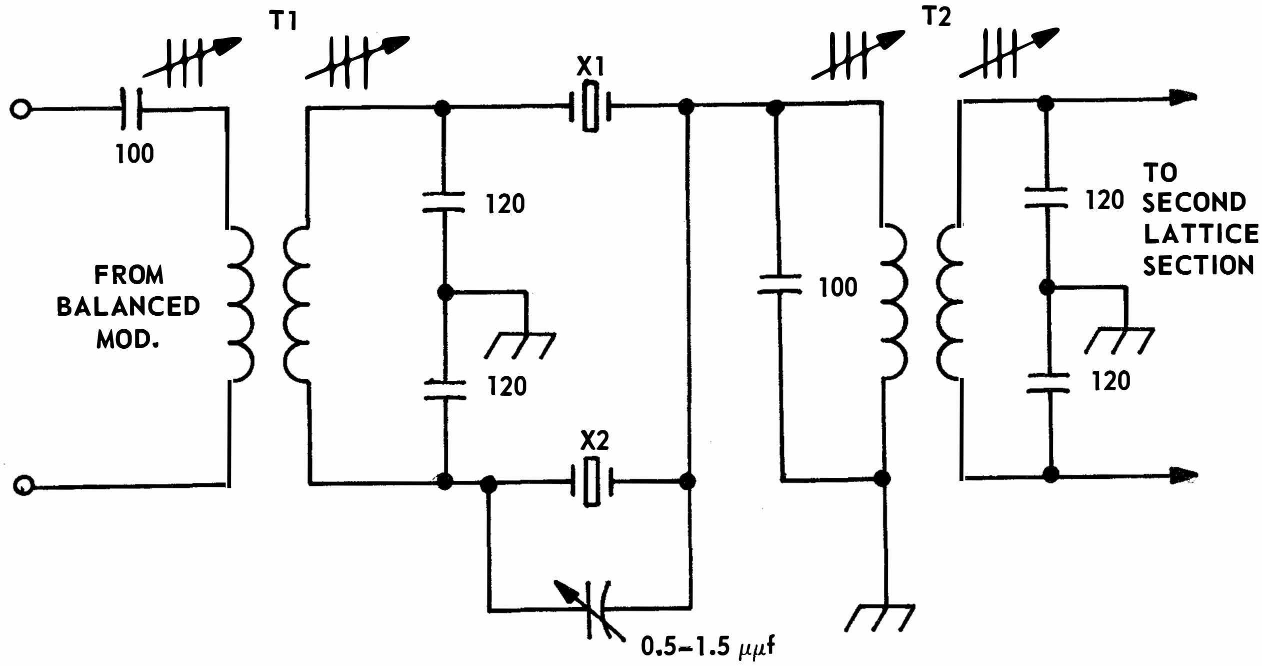


Figure 5-7. Balancing Trimmer Used in Crystal-Lattice Filter.

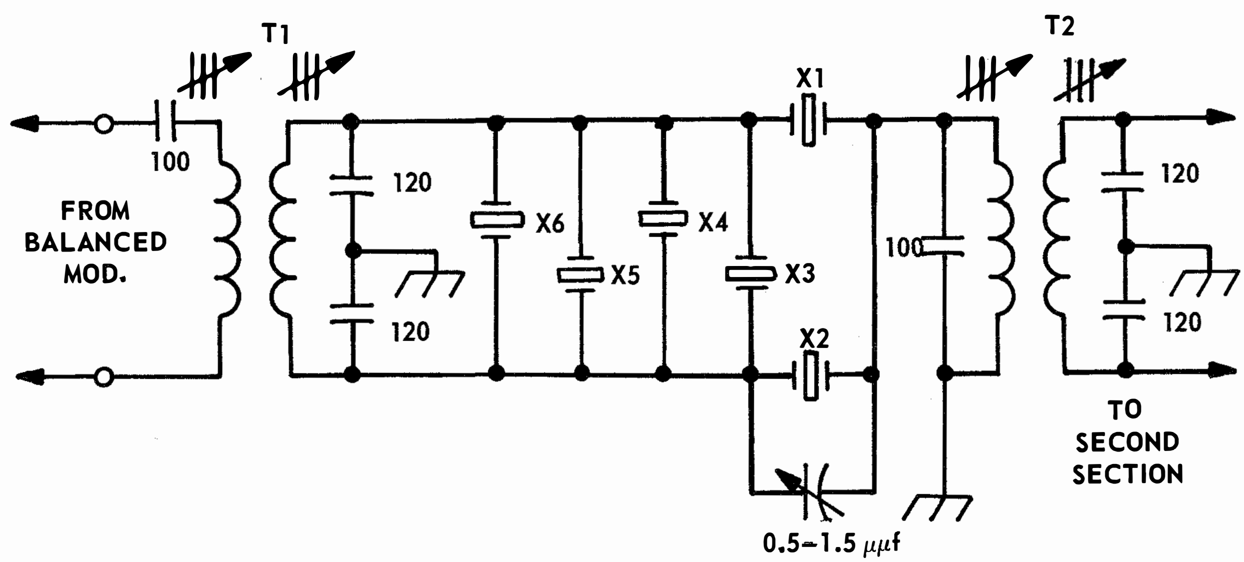


Figure 5-8. Crystals Used as High-Q Traps.

transmitter to use a high-frequency crystal filter was the Hallicrafters company's HT-32. The current version uses a crystal-lattice filter designed for a carrier-generator frequency of 4.95 megacycles. The filter passes the upper sideband of the modulated 4.95-megacycle carrier only. A choice of upper or lower sideband transmission is obtained by mixing the 4.95-megacycle upper sideband with the signal from a 4.05-megacycle crystal oscillator to produce an upper sideband of 9 megacycles, or with a 13.95 megacycle signal to obtain a lower sideband of 9 megacycles. This upper or lower sideband is then heterodyned to the desired operating frequency and then amplified to the specific power level by a linear amplifier. As might be expected, the stray capacitance and leakage of this type of filter proves to be much more critical than those of its low-frequency counterpart. The entire filter is well enclosed in a shielded aluminum box. The input and output terminals are kept well separated, and are shielded from each other by a metal plate. All terminal leads must be kept as short as possible and well shielded, to prevent signal radiation around the filter.

5-20. Summary of the Filter System of Single Sideband.

5-21. The filter system uses a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Filters having such characteristics are normally constructed for relatively low frequencies, below 500 Kc. Recent developments in crystal filter research have produced workable filters at 5 megacycles and beyond. The carrier generator output is combined with the audio output of a speech amplifier in a balanced modulator. The upper and lower sidebands appear in the output but the carrier is suppressed. One of the sidebands is passed by the filter and the other rejected so that a single sideband signal is applied to the following mixers (see figure 5-9). The signal is mixed with the output of a high-frequency RF oscillator to produce the desired output frequency. The problem of undesired mixer products arising in the

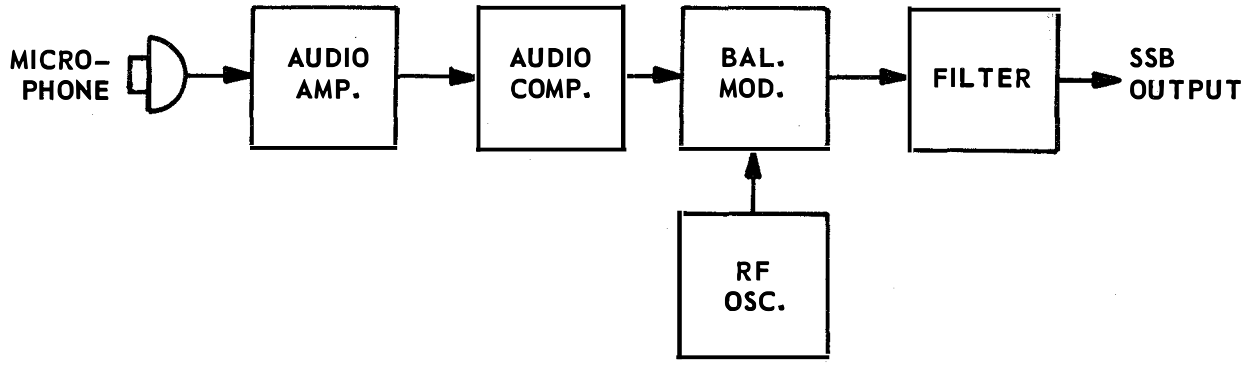


Figure 5-9. Block Diagram, Typical Single Sideband Exciter.

frequency conversion of single sideband signals becomes important. Either balanced modulators or sufficient selectivity must be used to attenuate these frequencies in the output and minimize the possibility of unwanted radiation.

5-22. Phase-Shift Sideband Selection.

5-23. Two balanced modulators are used in the phase-shift method of sideband selection. The exciting signals are made to have phase relationships such that, when the outputs of the two modulators are combined, the desired sideband components are reinforced and the undesired ones canceled out. The phase-shift single sideband generator requires no selective filters. It also has the advantage of being able to generate the single sideband signal at the operating frequency without the frequency conversion required by the filter method. The phase-shift and amplitude properties of the circuit, however, are very critical and must be held to close tolerances. This characteristic, and the requirement for two balanced modulators, tend to make the system somewhat more complex than the filter method.

5-24. Figure 5-10 illustrates the generation of a single sideband signal by the phase-shift method. The carrier is applied directly to balanced modulator A with no phase shift. The carrier signal applied to balanced modulator B is shifted in phase by 90°. In other words, at any time, for any given instant, the carrier signals applied to the two balanced modulators will be 90° out of phase. The 90° carriers are adjusted so that their amplitudes are equal. The two balanced modulators have a common output tuned circuit which is resonated to the carrier frequency. When the usual balancing and tuning procedures are carried out, no RF carrier will appear in the tuned output circuit.

5-25. In order for the system to function as a single sideband generator, it is necessary to divide the audio frequency signal of the balanced modulators into two components 90° out of phase. If the audio signal where a single tone of, say, 1000 cycles, and no other tones were to be transmitted, the 90° audio phase-shifting device would be relatively simple. In this case, the single-tone audio signal would be applied directly to modulator A with no phase change, and the 90° phase-displaced audio signal would be applied to modulator B. In a practical single sideband transmitter, however, it is generally desired to transmit a band of audio frequencies rather than a single tone. At this time, there is no known means for obtaining a 90° phase shift over a wide range of audio frequencies. In actual practice, two phase-shift networks with a differential phase shift of 90° are inserted between the AF source and balanced modulators. This phase-shift difference can be maintained over a frequency range of about seven octaves with a deviation of not more than ±2°. The phase-shift networks are made up of high-precision resistors and capacitors, and require the use of laboratory instruments for adjustments and measurements. Manufactured phase-shift networks, however, are relatively inexpensive. In fact, a complete unit often costs less than the total net price of the individual components.

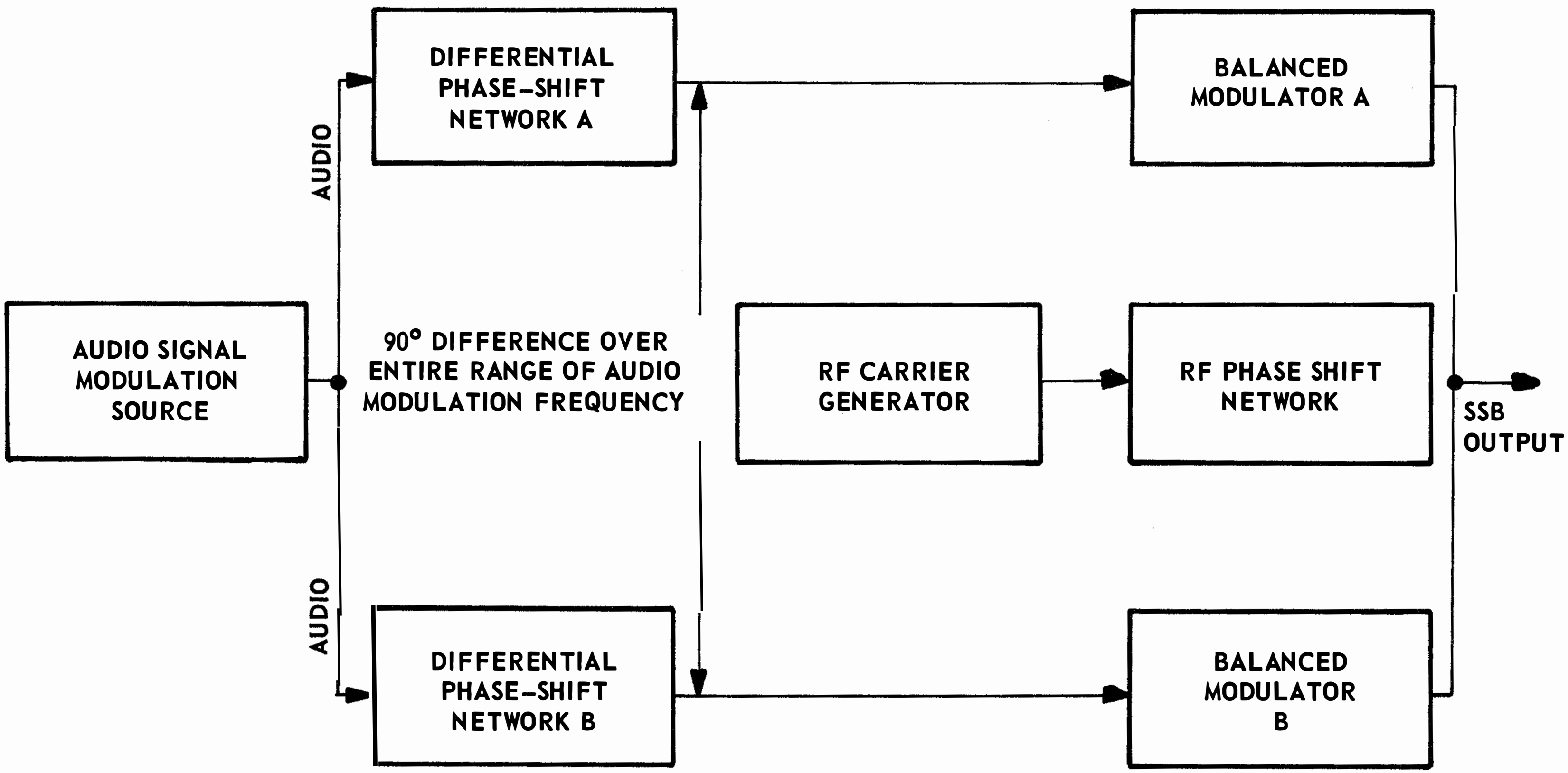


Figure 5-10. Block Diagram, Phase-Shift Single Sideband Generator.

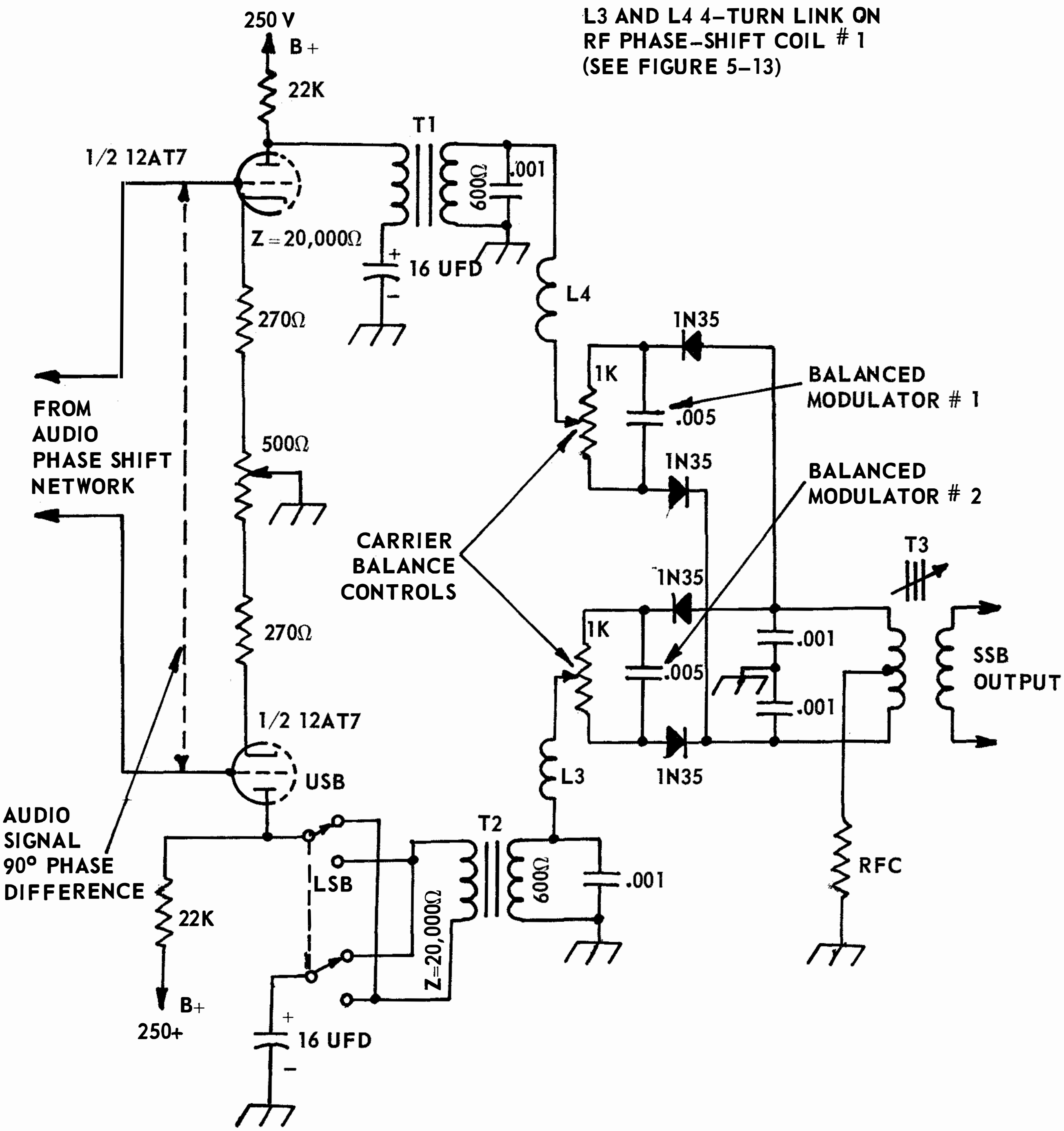


Figure 5-11. Phase-Shift Single Sideband Generator.

5-26. A Practical Generator.

5-27. In figure 5-11 a typical balanced modulator section of a phase-shift single sideband generator is shown. The sideband signal is generated with a 9-megacycle carrier. The 9-megacycle upper or lower sideband signal is then mixed in a frequency conversion circuit, together with an appropriate RF oscillator signal, and heterodyned to the operating frequency.

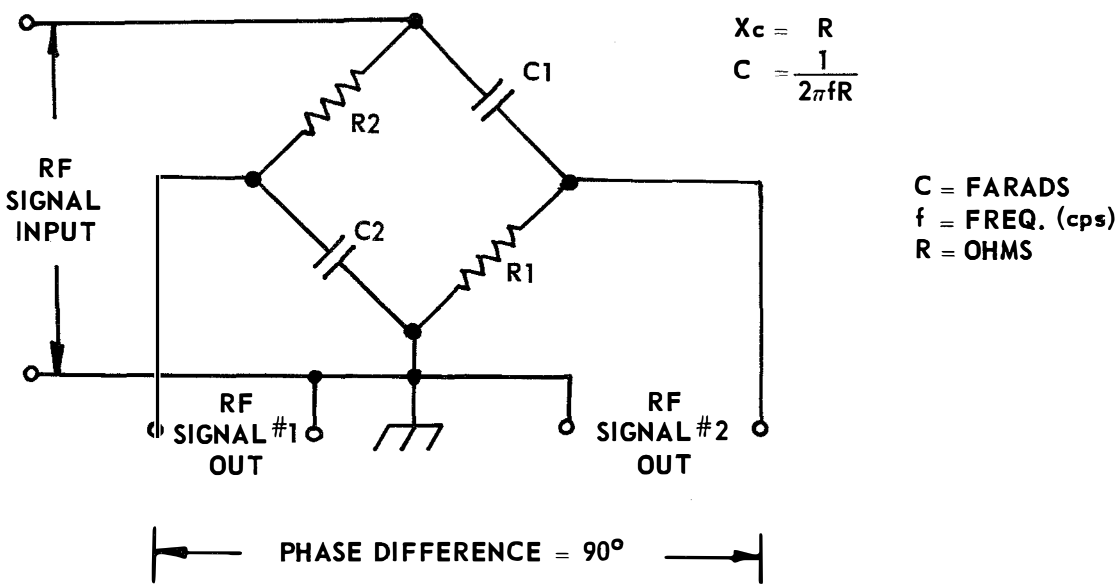


Figure 5-12. Resistance-Capacitance RF Phase-Shift Network.

5-28. The primary function of the RF phase-shift network (see figure 5-12) is to accept a signal from the carrier generator (usually this is a crystal-controlled oscillator) and divide it into two RF signals, which are then applied to the balanced modulators. These two signals must be of equal amplitude and precisely 90° out of phase. The most common RF phase-shift network consists of the double-tuned transformer shown in figure 5-13. Primary winding L1 functions as the usual plate coil with secondary winding L2 loosely coupled to it. These two coils are usually identically wound on slug-tuned forms and are mounted so that the centers of the windings are about 7/8 inch apart. L1 is detuned toward the high-frequency side of resonance until the RF voltage across it is down by 3 db from the value of resonance. L2 is detuned toward the low-frequency side of resonance until its RF voltage drops 3 db from the value at resonance. When these two coils are adjusted as described, the voltages across link windings L3 and L4 will be approximately 90° out of phase.

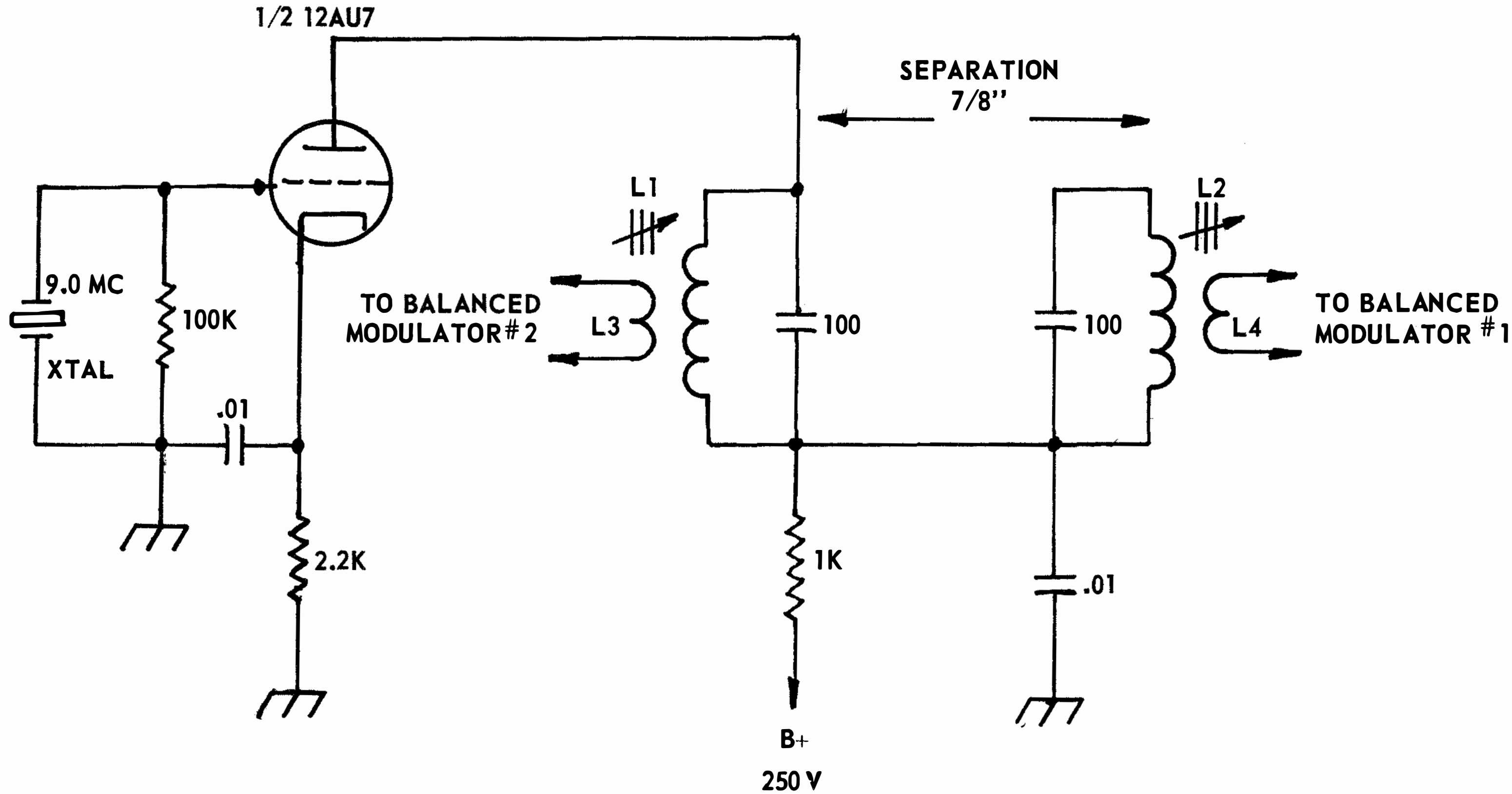


Figure 5-13. Double-Tuned Transformer RF Phase-Shift Network.

5-29. Resistance-Reactance Networks.

5-30. Figure 5-14 shows a simple phase-shift network composed of resistance, capacitance, and inductance. R1 and R2, which are equal in value, determine the impedance of the network. The inductive and capacitive reactances must equal the resistances of R1 and R2, respectively, at the operating frequency. When an AC signal (RF in this case) is applied to a pure inductance, the voltage across the circuit will lead the current by 90°. When the same signal is applied to a pure capacitance, the voltage across the circuit will lag the current by 90°. When a reactance and a resistance are used in combination, the resulting phase-shift of the voltage and current will be somewhere between ±90° and 0°, depending on the relative values of the total reactance and resistance. In this network, where each reactance is equal to each resistance, the voltage across L1 will lead the current by 45°, and the voltage across C1 will lag the current by 45°. The phase difference between the two voltages will then be 90°. Their amplitudes will be equal only at the frequency where the reactance values are equal. When the carrier frequency is increased, the voltage drop will

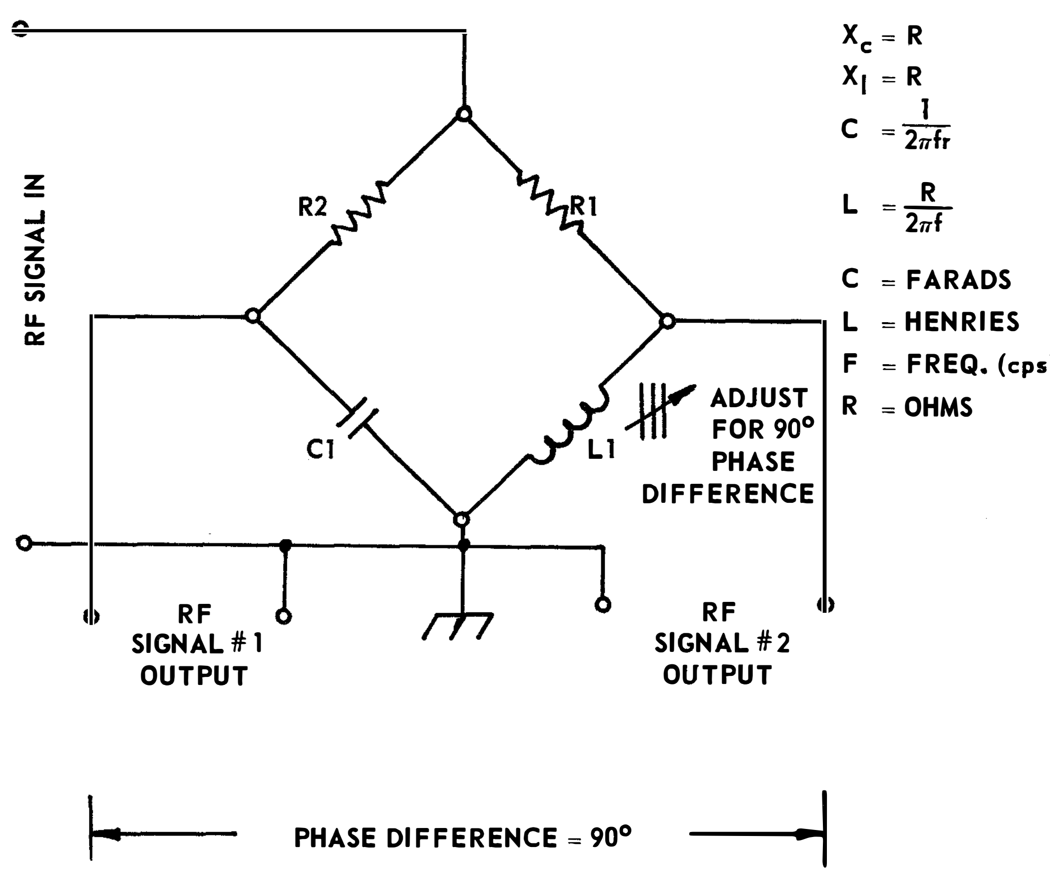


Figure 5-14. Simple Resistance-Capacitance-Inductance RF Phase-Shift Network.

increase across L1 and decrease across C1. Conversely, when the carrier frequency is decreased, the opposite will occur. The phase relationship of this circuit will remain at 90° for a reasonable range of frequencies above and below the design frequency. The output impedance is high and should not be terminated in resistances of less than 1 or 2 megohms. The network is generally used with vacuum tube balanced modulator circuits.

5-31. Comparison of Filter and Phase-Shift Single Sideband Transmitters.

5-32. Since the audio input and linear power amplifiers used in filter and phase-shift transmitters are essentially the same, comparison of the two systems is centered around the method of generating the single sideband signal.

5-33. It will be recalled that in most filter systems the modulation process must be performed at a relatively low RF level 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 an electromechanical filter, 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 comparatively high, more than one heterodyning process is usually required. Selection of either sideband, when desired, is usually quite difficult to accomplish in this system. To have sideband selection, it is necessary to change the filter or to change the operating frequency of the RF 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 variations in loading.

5-34. Adjustment of the filter system transmitter circuits is not critical except for the internal 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 single sideband transmitter is highly stable.

5-35. 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-power 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° shifts from the phase-shifting networks. Exact 90° phase relationship must be maintained throughout the system 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 single sideband transmitter is less stable than the filter-system transmitter and requires more frequent attention if optimum performance is to be obtained.

SECTION VI

SINGLE SIDEBAND EXCITERS

6-1. General.

6-2. The single sideband exciter must translate the incoming audio frequency signal to a band of frequencies in the RF range. A single sideband exciter is, in fact, a complete transmitter in itself. It must generate a radio frequency sideband from an audio input signal, translate this RF sideband to the final output frequency, and provide sufficient amplification to drive the RF power amplifier.

6-3. To generate the RF sideband frequencies, the single sideband exciter uses low-level modulation and obtains the desired output level through the use of linear amplifiers. Low-level modulation is used since the carrier and unwanted sideband must be suppressed. The best suppression is obtained at a fixed low frequency, since the problems involved in building a high-level balanced modulator capable of working over a wide frequency range appear to be insurmountable.

6-4. The most desirable performance characteristics of a single sideband exciter would be the ability to generate the desired sideband, completely suppress the undesired sideband, and suppress the carrier. Practical design permits suppressing the undesired sideband and carrier frequencies by more than 40 db.

6-5. Careful consideration must be given to the amount of frequency spectrum space occupied by the generated signal. The band of side frequencies is normally held to 4 Kc in single sideband exciters for communications purposes.

6-6. The Single Sideband Generator.

6-7. The sideband generator processes the input audio signal, generates the RF sidebands in a modulator, selects the desired sideband while suppressing the unwanted sideband, and suppresses the carrier. The audio input wave must be amplified, amplitude limited, and shaped before being applied to the modulator circuits. Sideband generation is accomplished by using this audio input signal to vary the amplitude of a carrier wave in a modulator. A desired sideband is selected from the modulator output, using frequency discrimination. The carrier wave is suppressed by using balanced modulators or rejection filters.

6-8. Input Signal Processing.

6-9. Processing of the audio input signal is an important part of single sideband generation. If the input signal is a tone or group of tones, of constant amplitude, such as the signal from a data gathering device, only a limited degree of processing will be required. However, if the input audio signal is a voice signal, rather elaborate input processing circuits must be designed to obtain optimum results.

6-10. The amount of amplification required depends upon the output capabilities of the source of the audio signal and the input signal requirements of the modulator. Modulators require an audio signal in the range of 0.1 to 1 volt and impedances of 200 ohms for diode modulators, or several hundred thousand ohms for vacuum-tube modulators. The output from a microphone may be from 100 to 1000 times less than the 0.1 to 1 volt range. Telephone line levels will also be considerably less than the required level. To obtain efficient utilization of the transmitter power amplifier, the driving signal level should be as high as possible without exceeding the amplifier overload point. To avoid driving the power amplifier into overload, it is necessary to adjust the gain to the point where maximum output is obtained with this optimum input signal.

6-11. When the input signal is made up of extreme variations, such as a peak level to average level of 4 to 1, the average transmitted power level will only be 1/4 the maximum output the transmitter is capable of furnishing. With the average voice, the average level to peak level is considered to be 12 db. An effort must be made to compress the dynamic range of the human voice to make it more compatible with the electrical characteristics of a communications system. The two methods most commonly used to reduce these amplitude variations are compression and clipping.

6-12. The Compressor.

6-13. A compressor is an automatic variable-gain amplifier, the output of which bears some consistent relation to its input; for example, a 1-db rise in output for a 2-db rise in input. This circuit has very low steady-state distortion. Common compressors use some type of feedback loop that samples the output of the amplifier and regulates the gain of the stage. The time constants of this type circuit are necessarily slow to prevent oscillation ("motorboating") and distortion. The attack time, the time necessary to reach steady-state condition after a sudden rise in input level, will be several milliseconds. The release time, the time necessary to reach a steady-state condition after a sudden drop in input level, will be several seconds. Compression of about 10 db is usually considered as an acceptable maximum value.

6-14. The Clipper.

6-15. The clipper circuit prevents the amplitude of a signal from exceeding a preset level. Its time constants are practically instantaneous and it functions on each cycle of a wave. Distortion is very high, which results in loss of individuality in speaking and broadening of the spectrum occupied by the speech. Low pass filters are usually used in conjunction with clippers to limit the spectrum and reduce distortion. The advantage of clipping is simplicity of circuit design and its ability to prevent overmodulation. The ability to prevent overmodulation results from its extremely fast attack on a wave after it exceeds a threshold. A well-designed clipper has no overshoot and an extremely fast release. A weak signal following one cycle after a wave that is heavily clipped will not be limited. This means that a weak consonant that follows a loud vowel in human speech will be given full amplification, although the preceding vowel was severely clipped. This amplifying of weak sounds in relation to loud sounds is referred to as consonant amplification.

6-16. Frequency Response Shaping.

6-17. The energy contained in a voice signal is confined principally to frequencies below 1000 cycles per second. Most of this energy is used to produce the vowel sounds which contribute little to intelligibility. The energy used to produce the consonant sounds is largely high frequency in content and is very important in intelligibility. An improvement in intelligibility will result if the frequency response of the audio input signal circuits is modified to amplify the high frequencies more than the low frequencies.

6-18. Translation to the Operating Frequency.

6-19. The single sideband signal is translated to the operating frequency by the use of one or more frequency changers. These frequency changers perform their functions through the modulation process, which is identical with that used to generate the single sideband signal. The sideband signal is used to modulate a high-frequency carrier whose frequency is such that the upper or lower sideband is on the desired operating frequency. As a result of this modulation process, the sideband signal will be shifted to a new frequency that is either the sum of the carrier and sideband frequencies, or the difference between the carrier and sideband frequencies. It is important to realize that if the lower sideband of the translation modulation process is selected, an inversion of the sideband signal occurs; that is, an upper sideband signal will be converted into a lower sideband signal. Another important consideration is the frequency accuracy and stability of the carrier used in the modulation process since any error in the carrier frequency is passed on to the sideband signal in exact proportion.

6-20. All modulators previously described can be used for frequency changing. Vacuum-tube modulators are used almost exclusively in this service, because their circuits have considerable gain and generate a minimum of spurious products. These spurious products exert the influence on the design of frequency translation systems.

6-21. Spurious Mixer Products.

6-22. In order to show how undesired frequencies may be generated in a mixer stage, consider the case where signal and oscillator voltages are applied to the same grid of a mixer tube. In order that the desired sum or difference frequency be generated, it is necessary that the plate current versus grid voltage characteristics have some non-linearity or curvature. The components of the plate current will be the DC, oscillator, signal second harmonic, oscillator second harmonic, and the sum and difference of the signal and oscillator frequencies. It is necessary to eliminate all the products except the desired sum or difference product by filtering. To obtain the desired sum or difference product, we would desire a tube in which the characteristic curve had only second-order curvature. Unfortunately, all practical tubes have characteristic curves with higher-order curvature. This higher-order curvature contributes additional unwanted frequency components to the output current. Sometimes the frequency of these unwanted components is far removed from the desired output frequency and is easily filtered out, but often these frequencies are very nearly equal in frequency to the desired signal frequency and will fall within the passband of the filter used in the mixer output circuit. The amplitude of these undesired mixer products varies from tube type to tube type and from tube to tube and, with a given tube, will change when the operating point is varied. Consequently, it is not surprising that tube designers have not been particularly successful in designing tubes having the desired second-order curvature to the exclusion of any higher-order curvature. The circuit designer, therefore, must select his mixer tubes by means of a series of experiments in which the amplitudes of these undesired mixer products are measured.

6-23. There are several undesired products which are greater in amplitude than the desired signal and a considerable number that are weaker than the desired signal. Furthermore, as the order of the mixer product involved increases, its amplitude decreases. The presence of undesired mixer products in the output of the frequency translation system may be minimized through intelligent selection of the signal and oscillator frequencies. The problem of frequency selection is relatively simple where the operating frequencies are fixed. Where the operating frequencies must be varied the problem becomes more complex and if continuous operation over wide frequency ranges is required, the problem is exceedingly complicated. In an attempt to simplify the problem, circuit designers have resorted to charts in which the frequencies of the spurious mixer products are plotted with respect to the signal and oscillator frequencies.

6-24. The following example, and figure 6-1, illustrates the spurious product problem. In this example it is desired to produce a single sideband transmitter capable of operating on the amateur 80-, 40-, and 20-meter bands. These bands are 3.5 to 4 megacycles, 7 to 7.3 megacycles, and 14 to 14.35 megacycles. Assume that the single sideband signal has been generated at a 250-Kc carrier frequency. The lower frequency band, 3.5 to 4 megacycles, can be covered by mixing this single sideband signal with the output of a variable frequency oscillator tunable from 3.25 megacycles to 3.75 megacycles. Due to the large difference between the signal and and the oscillator frequencies, there are no difficulties with the undesired mixer

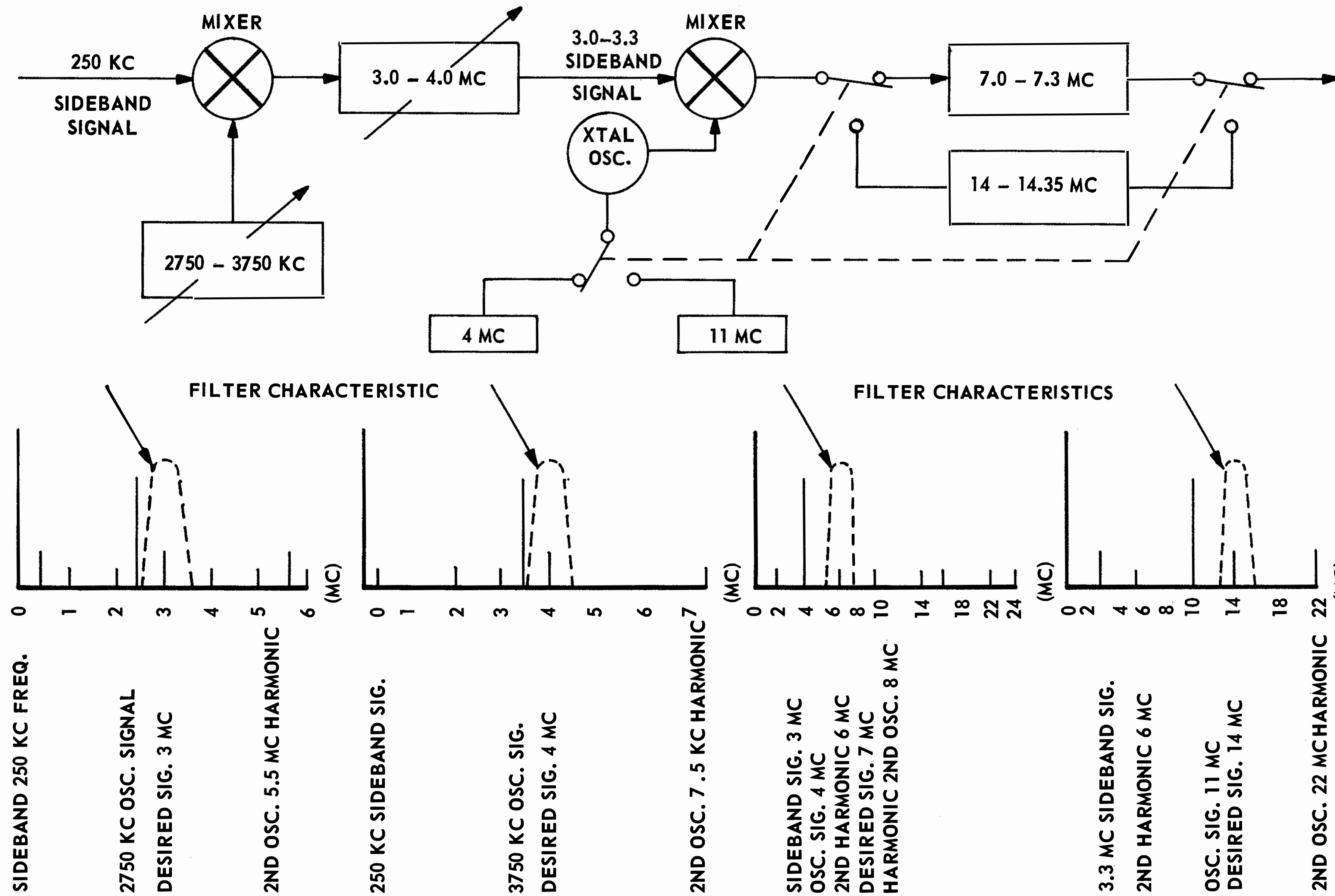


Figure 6-1. Selectivity Considerations in Frequency Translators.

products. However, the oscillator signal is only 250 Kc removed from the output frequency and must be filtered by means of a bandpass filter or balanced out through the use of a balanced modulator. As it is quite difficult to build a modulator which can retain balance over a wide frequency range, it is necessary to resort to a combination of both methods to obtain suppression of this spurious frequency by at least 60 db. As the operating frequency increases, it becomes difficult to suppress this product and some other method must be found. This is because the selectivity required in the tuned circuit is so high as to be impractical. A solution to the problem is to use a second conversion following the first. In this mixer, the 3.5 to 4 megacycle output of the first conversion is mixed with the output of a crystal oscillator at 3.3 megacycles. This frequency is chosen rather than 3.5 megacycles because, with a 7-megacycle output frequency, there is a crossover of the second harmonic of the 3.5-megacycle signal with the desired output. A closer look at the frequencies involved, however, reveals that even with this crystal frequency, the second harmonic of the crystal at 6.6 megacycles is only 400 Kc removed from the low frequency desired output, and the 7.4 megacycle second harmonic of the input signal is only 400 Kc on the other side of the desired output frequency. Selectivity of a higher order would be required to reduce these spurious signals to a satisfactory level. Some relief can be obtained by extending the range of the variable frequency oscillator used in the first mixer so that the output frequency from the first conversion runs from 3 to 4 megacycles. Now, a crystal oscillator frequency at the second mixer or 4 megacycles can be used. With this frequency, the range of the first converter system of 3 to 3.3 megacycles can be used to cover the 7 to 7.3 megacycle band. With this frequency scheme, the second harmonic of the crystal oscillator is at least 700 Kc removed from the desired operating range and the second harmonic of the signal frequency ranges from 100 to 700 Kc below the desired output frequency range. These can be filtered adequately with relatively simple filters. A seventh-order crossover occurs at the low end of the band when the third harmonic of 3 megacycles mixes with the fourth harmonic of 4 megacycles to yield a frequency equal to the output frequency. However, the level of the seventh-order spurious signal can be disregarded if sufficient attention is paid to the selection of the mixer tube and its operating point. In considering frequencies to be used to cover the 14 to 14.35 megacycle band, it is possible to use the sum product if a crystal frequency of 11 megacycle is used. The ninth-order crossover occurring near the low end of the range is of no consequence, since it is of negligible amplitude.

6-25. Oscillator Requirements.

6-26. It must be realized that the frequency stability of the output signal is dependent on the frequency stability of the carrier and oscillator outputs used in the frequency changers. The total frequency error is the sum of the errors in all three of these oscillators. Oscillator frequency error has two aspects. First, there is the accuracy of the calibration of the frequency involved. The second aspect of the frequency error is that of stability or of drift. If the equipment is to be operated and continuously monitored by skilled operators, it is possible to get by with rather large errors in both calibration and drift. In some cases, it is possible to use equipment in which the calibration error is relatively large, but the frequency drift is quite small. In this case, it may be possible to carry out effective communications with such equipment if an operator is available to make the initial tuning adjustment. In some cases, where it is desired to operate the equipment on many channels by remote control and with relatively unskilled operators, it is necessary to provide equipment with a high degree of performance, both with respect to calibration and drift. Authorities tend to disagree as to the frequency error which can be tolerated in a single sideband communications system used for voice communications. As the error increases, the naturalness of the reproduced speech suffers first. If the error is such as to place the reinserted carrier on the high side of the original frequency, the voice becomes lower in pitch. If the error is such as to place the reinserted carrier on the low frequency side, the voice becomes higher in pitch. As the error is increased, a point is reached where intelligibility is degraded. The frequency error at which this occurs is approximately 100 cycles per second. When the single sideband transmitter is used to transmit narrow-band telegraph or teletype signals, it is sometimes necessary to maintain accuracy considerably higher than that required for voice transmissions. For example, in narrow-band radio-teletype signals using a total shift of 170 cycles, a 10-cycle drift will cause the signal to deteriorate so that hits appear on the teletype circuit; thus, much greater accuracy must be maintained.

6-27. The stability which can be obtained from oscillators is an important factor in the design of a frequency translation system. Typical oscillator frequency errors are shown in tabular form in tables 6-1 and 6-2. In table 6-1 the frequency error is the long-term frequency error. Calibration, drift, and aging all contribute to the long-term frequency error. The errors listed in table 6-1 are typical for a term of several months. The short-term frequency errors are shown in table 6-2. The short-term frequency error is principally that of frequency drift, although, in some cases, aging is rapid enough to have some effect. From the data shown on these tables it can be seen that single sideband equipments using variable-frequency oscillators would require manual operation and frequent attention. For an equipment to meet the stability and accuracy requirements for quick frequency selection by remote control, by

TABLE 6-1. TYPICAL OSCILLATOR LONG-TERM FREQUENCY ERROR

Oscillator Type	Error %	Error cps		
		3 Mc	10 Mc	30 Mc
Variable Frequency Osc.	0.05	1,500	5,000	15,000
Crystal Osc.	0.005	150	500	1,500
Temperature Controlled Crystal Osc.	0.001	30	100	300
Precision Standard Oscillator	0.0001	3	10	30

TABLE 6-2. TYPICAL OSCILLATOR SHORT-TERM FREQUENCY ERROR

Oscillator Type	Error, ppm	Error cps		
		3 Mc	10 Mc	30 Mc
Variable Frequency Osc.	20	60	200	600
Crystal Osc. and Temperature Controlled Crystal Osc.	1	3	10	30
Precision Standard Oscillator	0.01	0.03	0.1	0.3

unskilled operators, it is necessary to use the form of oscillator known as a stabilized master oscillator. In it, a variable-frequency oscillator is stabilized by comparing its frequency with that of a frequency derived from a reference standard oscillator. Oscillators of this type are described later in another section.

6-28. Amplification.

6-29. In order that the single sideband exciter have useful output, it is necessary to provide amplification of the single sideband signal. Since the output power will be used to drive the power amplifiers of the system, the power output of the exciter is determined by system considerations. The driving power required is usually quite small since it is customary to use high-gain tetrode tubes in most linear amplifiers. As a result, the power output of the single sideband exciter may be limited to a fraction of a watt or it may be at the 100- to 150-watt level. This power level can be readily obtained through the use of receiving type tubes.

6-30. Pentode receiving tubes, designed for use as RF or IF amplifiers in receivers, may be used for low-level voltage-amplifier stages. Low grid-to-plate capacitance is a necessary requirement for linear RF amplifiers since positive feedback through this path increases distortion. High mutual conductance is a useful characteristic because the required gain is then obtained with a minimum of stages. Receiving-type power pentodes may be used to obtain moderate power output, although most types suffer from having relatively large grid-to-plate capacitance. Tubes designed for video power amplifiers are best suited for use in linear RF power amplifiers.

6-31. Tuned Circuits.

6-32. Tuned circuits used in single sideband linear amplifiers perform a dual function. A tuned circuit provides a suitable load impedance for the amplifier stage so that the amplifier may provide sufficient voltage amplification. Secondly, this tuned circuit acts as part of a selective filter which acts to suppress unwanted mixer products generated in the frequency translation system. To obtain sufficient selectivity, it is often necessary to use double-tuned and even triple-tuned circuits.

6-33. Linear Amplification.

6-34. It is necessary to use linear amplifiers in a single sideband transmitter in which low-level modulation is used. The single sideband system is an amplitude modulated system, and once the modulation is performed the amplitude relationships of the sideband components must be faithfully maintained. The principal distortion component encountered in tuned linear amplifiers is the third order intermodulation product. This product is so called because its generation depends on the third order curvature of the input-output amplifier characteristics. Unlike audio linear amplifiers, which must handle a wide frequency range, tuned radio frequency linear amplifiers seldom have difficulty with products generated due to second order curvature, such as sum and difference frequencies and their harmonics. These frequencies usually fall far outside the tuned passband and are suppressed accordingly. On the other hand, the third order intermodulation products are always close to the desired frequency band and many of the products actually fall within the desired passband. These intermodulation products are generated whenever there are two tones or frequencies within the amplifier passband whose frequencies are sufficiently close together that the second harmonic of one will mix with the other to yield a third frequency within the tuned amplifier passband. The amplitude of these spurious products can be controlled by limiting the input signal amplitude so that operation of the tube is always over a linear part of its input-output characteristic. It is readily possible to obtain sizable voltage and power amplification, using receiving type tubes, and still limit the amplitude of the third order intermodulation product to a level of more than 50 db below the desired signal.

6-35. An important consideration in transmitters using linear power amplifiers is that the amplifier be driven with sufficient signal and yet not be over-driven to cause excessive intermodulation distortion. There are many factors which tend to cause the output of a single sideband exciter to vary. The gain of the amplifier stages changes from one frequency channel to another, as the load impedance in these stages varies with frequency. Tube gain characteristics change from tube to tube, and from time to time as the tubes age. Changes in temperature and other environmental factors can also cause changes in amplifier gain. An effective way to cope with these variations is to sample the driving voltage of the power output amplifier with a rectifier and to use the resulting DC to control the gain of one or more amplifier stages in the exciter. This control voltage may be used to control the gain of amplifier stages using remote-cutoff tubes similar to those used in receiver RF amplifiers on which automatic gain control is used.

6-36. Summary.

6-37. The single sideband exciter consists of three major sections: a single sideband generator, a frequency translator, and a power amplifier. In the sideband generator, the audio input signal is processed by the use of amplification, amplitude limiting and frequency energy distribution. The processed signal is then converted into an RF sideband in a balanced modulator. The desired signal or sideband is selected and the unwanted sideband suppressed, using the technique of frequency discrimination or phase discrimination. The desired sideband is then translated to the desired operating frequency by means of a frequency translation system. The desired output level is obtained through the use of a linear amplifier.

6-38. Third Method of Generating Single Sideband Signals.

6-39. Although the filter and phase-shift methods of generating single sideband signals are most commonly used in existing single sideband 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 eliminated. Thus, this third method of generating a single sideband signal differs from the filter and phase-shift method 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.

6-40. A block diagram of the third method of single sideband signal generation is illustrated in figure 6-2. 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, RC phase-shifting networks, is a quadrature, 1900-cycle frequency (the first carrier) from an audio oscillator. The frequency of this oscillator is selected to fall in the center of the audio spectrum, as illustrated in figure 6-3A. 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 6-3B. It can be seen that the lower sideband generated in these modulators (first pair) contains frequencies from 0 to 1600 cps, and that the upper sideband contains frequencies from 2200 to 5400 cps. Since the lowest audio modulating frequency is 300 cycles, no frequencies between 1600 and 2200 cycles will be generated in this modulator system.

6-41. The output of the first pair of modulators is applied to separate low-pass filters which are designed to pass the 0- 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 cycles, 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 RF oscillator. This RF frequency (second carrier) is applied in quadrature through separate 45-degree, lead-lag, RC phase-shifting circuits similar to the type

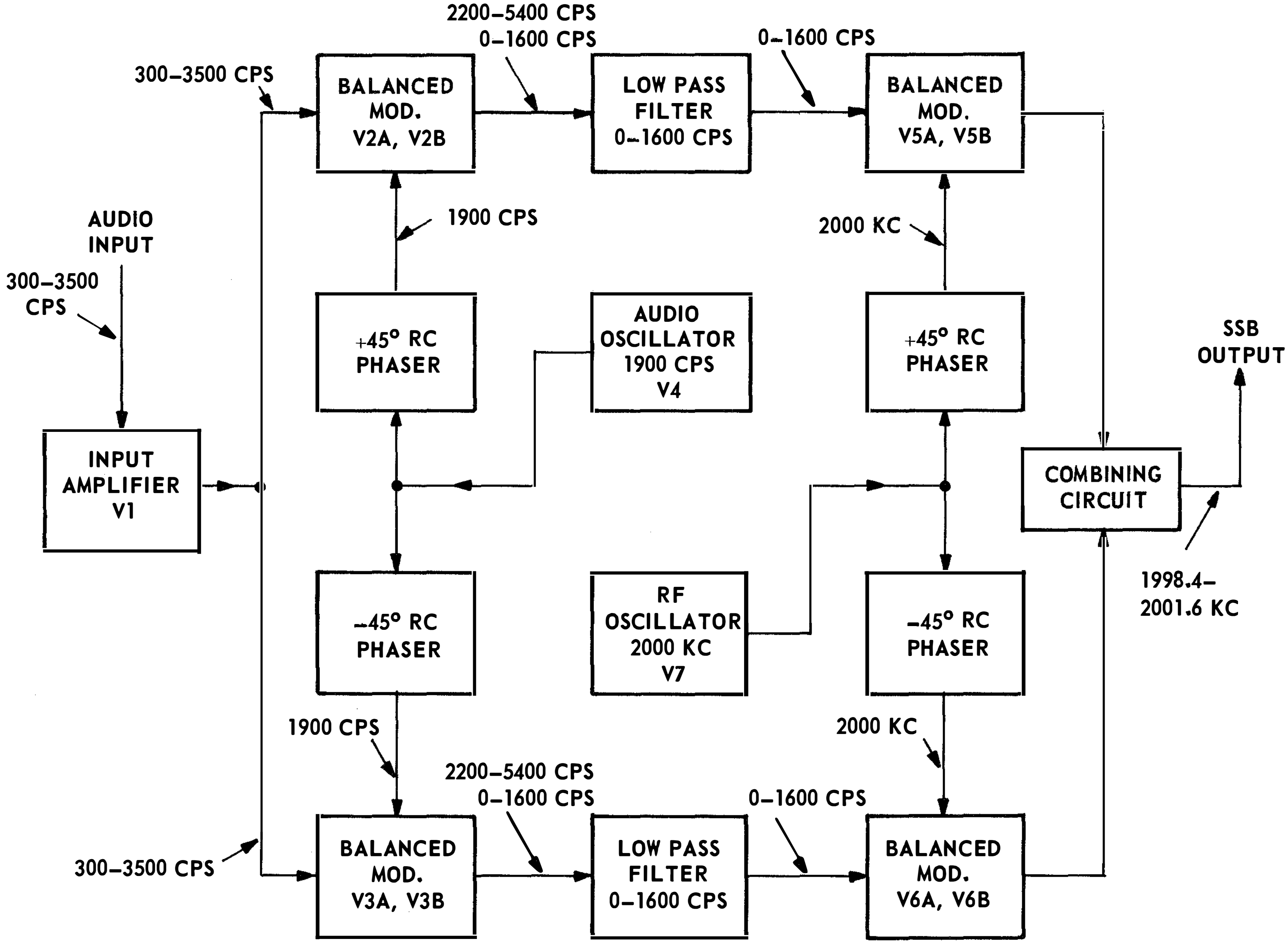


Figure 6-2. Block Diagram, Third Method of SSB Generation.

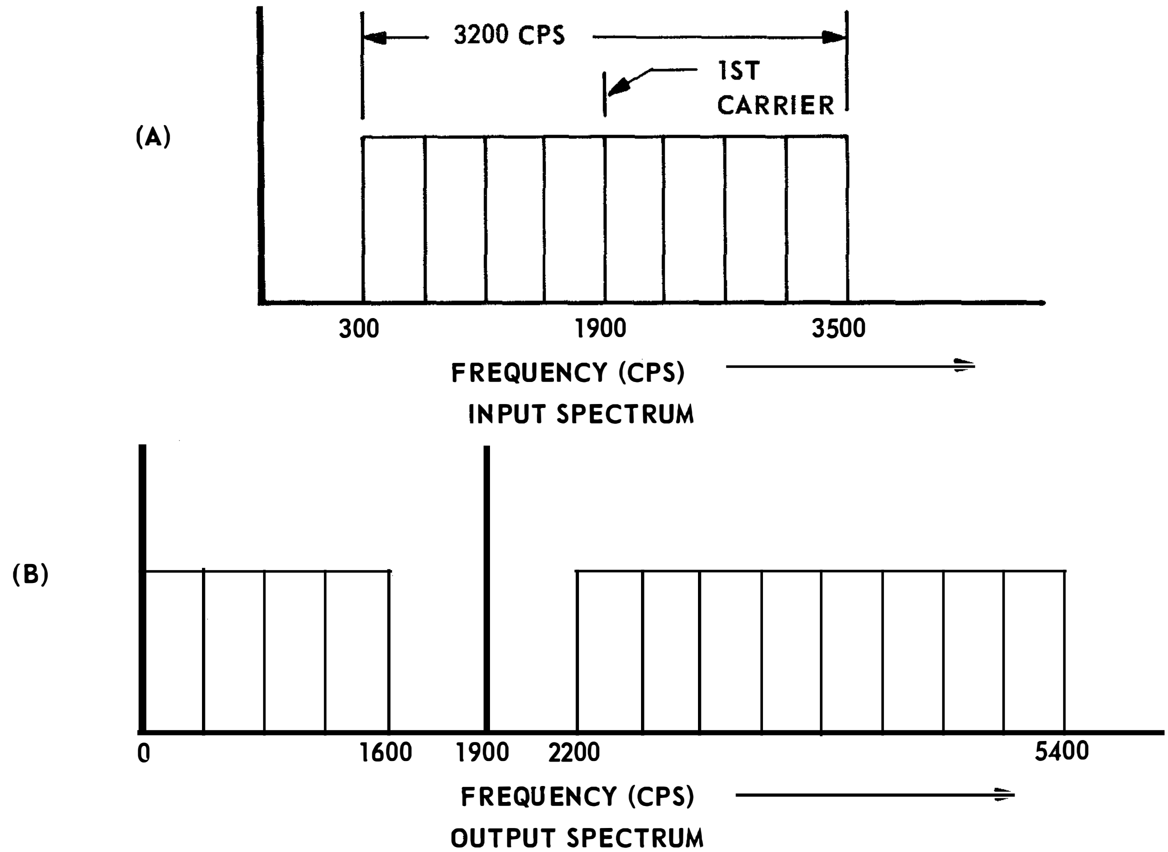


Figure 6-3. First Balanced Modulator Signal Spectra.

used in the low-frequency audio system. The RF 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.

6-42. The second pair of balanced modulators will generate upper and lower sideband frequencies and suppress their individual RF 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.

6-43. The circuitry of the third method of generating a single sideband signal is bilateral in operation, that is, a circuit can be used for demodulating as well as generating a single sideband signal. Since the circuit does not use any expensive, sharp-cutoff filters, or require any critical balance of balanced modulators, it has the advantages of being less costly than the filter system and less critical in adjustment than the phase-shift system.

SECTION VII

SINGLE SIDEBAND RECEIVERS

7-1. Single Sideband Receiver Considerations.

7-2. The operation of a single sideband receiver is, in a limited sense, the reverse of the process carried out in a single sideband exciter. The received single sideband radio-frequency signal is amplified, translated to a low IF frequency, and converted into a useful audio-frequency signal. The reception of a high-frequency single sideband signal is essentially the same as receiving a high-frequency AM signal. Receivers are invariably of the superheterodyne type, providing high sensitivity and selectivity. The absence of a carrier in the received single sideband signal accounts for the principal difference between single sideband and AM receivers. In order to recover the intelligence from the single sideband signal, it is necessary first to restore the carrier. This local carrier must have the same relationship with the sideband components as the initial carrier used in the exciter modulator. To achieve this, it is a stringent requirement of single sideband receivers that the oscillator which produces the reinserted carrier have extremely good frequency accuracy and stability. The total frequency error of this system must be less than 100 cycles per second, or the intelligibility of the received signal will be degraded. It should be noted that this 100-cycle error is for voice transmission only; for data transmission, a much closer tolerance is required.

7-3. The single sideband receiver must be able to select a desired signal from among the many signals which occupy the high-frequency band. Good selectivity becomes an essential requirement when signals of considerable variance in amplitude are spaced close together in the frequency spectrum. The sensitivity of the receiver must be sufficient to recover signals which are of very low amplitude. These requirements determine the design of the front end for the RF section of the receiver.

7-4. Double-conversion superheterodyne circuits are used in present-day receivers. The principal advantages of such circuits are the extra image rejection obtained and the decided improvement in frequency stability. This can be achieved by using a crystal oscillator in the high-frequency conversion and injecting the tunable oscillator signal at a lower frequency conversion, where its error has less effect.

7-5. In a conventional receiver, the audio intelligence is recovered from the radio frequency signal by means of an envelope detector. This detector may be a simple diode rectifier. This same diode detector, provided with a local carrier, can also be used to recover the audio signal from a single sideband suppressed-carrier signal; however, the amplitude of the local carrier must be quite high in order that the intermodulation distortion may be kept low. Better performance, particularly with respect to distortion, may be obtained by using product demodulators to recover the audio signal.

7-6. The characteristics of the automatic gain control system of a single sideband receiver must be somewhat different from those of a receiver designed for conventional AM signals. The conventional AVC system provides the automatic gain control by rectifying the carrier signal since this carrier is relatively constant and does not vary in amplitude rapidly. This AVC system can have a relatively long time constant. In a receiver for single sideband suppressed-carrier signals, the AGC rectifier must be of a quick acting type, because the signal amplitude is changing very rapidly and frequently drops to zero.

7-7. Single sideband receivers have three main sections: a radio frequency section; an intermediate frequency section; and an audio frequency section. The sections of a typical single sideband receiver are shown in figure 7-1. The principal requirement of the RF section is to select the desired signal and translate this signal to a lower RF frequency (intermediate frequency) with minimum distortion and minimum generation of spurious signals. The intermediate frequency section provides selectivity and amplification; the audio section recovers the audio-frequency intelligence and provides the necessary audio-frequency amplification.

7-8. The RF Section.

7-9. The RF section consists of an RF amplifier and one or more mixers which translate the signal to the intermediate frequency. The use of an RF amplifier, as the first stage of a receiver, provides increased sensitivity and reduction in spurious response. Increased sensitivity results from the lower inherent noise of amplifiers, as compared with mixers. Spurious signals are reduced because increased RF filtering can be used without degrading the signal-to-noise ratio. The amplification provided by the RF amplifier offsets the losses inherent in the passive filter circuits.

7-10. The sensitivity of a receiver is usually defined as the minimum signal with which a 10-db signal plus noise-to-noise ratio may be obtained. This definition has a practical basis because it recognizes the fact that, ultimately, the noise level is the limiting factor in readability and that a signal 10-db stronger than the noise level is acceptable for voice communications. Maximum receiver sensitivity is not determined by the gain of the receiver, but by the magnitude of the receiver noise. The three sources of noise which contribute to the noise level of a receiver are: the antenna to which the receiver is connected; the input resistance of the receiver; and

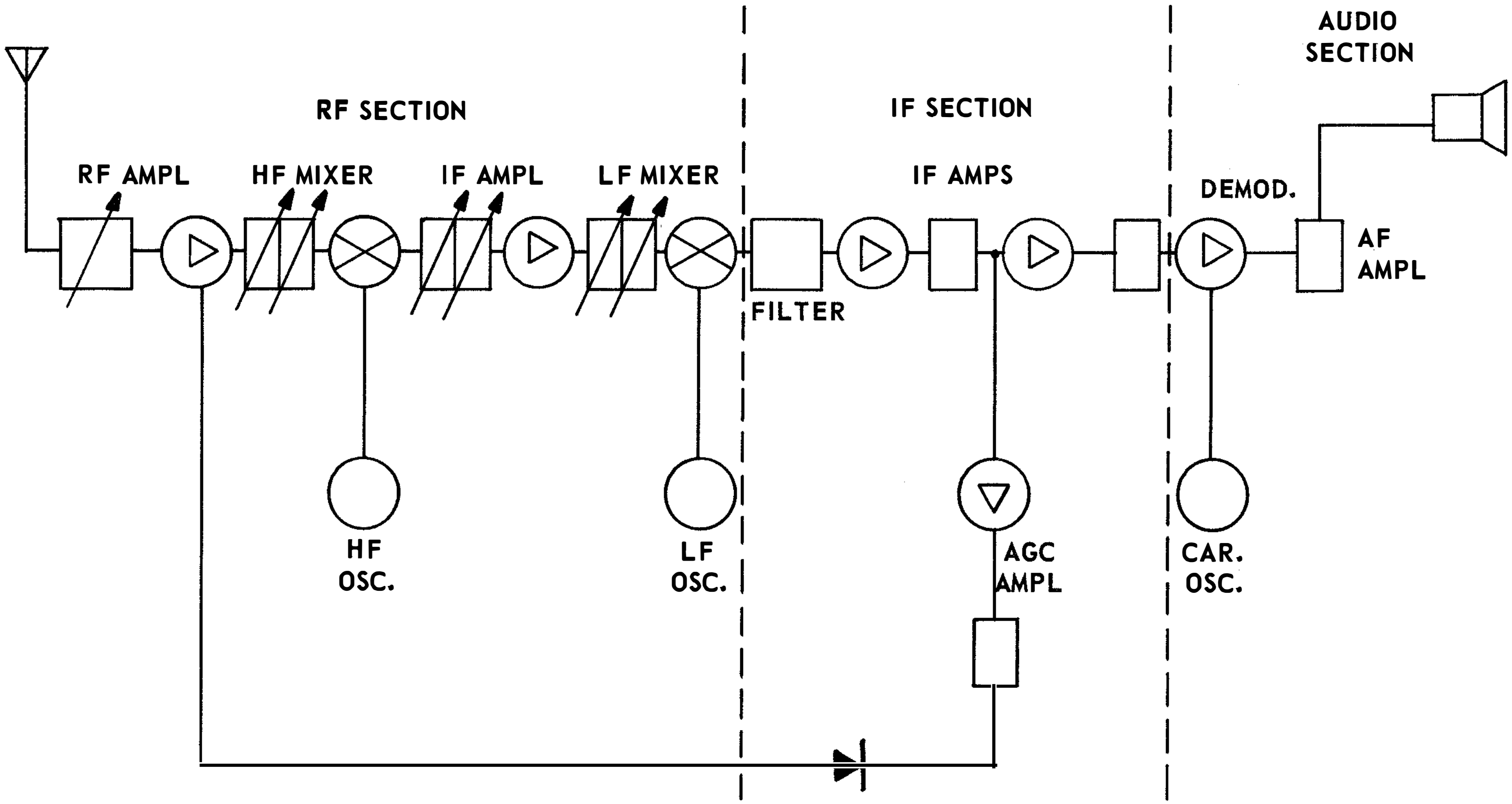


Figure 7-1. Functional Diagram, Typical Single Sideband Receiver.

the grid circuit of the first tube used in the receiver. If the gain of the first amplifier is low, it is possible that the noise of the second tube in the receiver can also have some effect on the receiver overall noise level.

7-11. Noise Sources.

7-12. A noise voltage will be present across the terminals of any conductor due to the random motion of electrons. This random electron motion is known as thermal agitation noise and is proportional to the resistance of the conductor and its absolute temperature. All noise currents and voltages are random fluctuations and occupy an infinite frequency band. The actual magnitude of noise voltage which affects the device is proportional to the bandwidth of the device. Noise voltage is calculated with the following formula:

$$E_n = 4KTBR$$

in which

E_n = rms noise voltage

K = Boltzmann's Constant, $1.38(10^{-23})$

T = absolute temperature

B = bandwidth in cycles per second

R = resistance in ohms

7-13. If the antenna to which the receiver is connected could be placed in a large shielded enclosure, there would exist at the terminals of this antenna a noise voltage equal to the thermal agitation noise of a resistor which is equal to the radiation resistance of the antenna. Even if a receiver could be built with no internal sources of noise, noise would still be introduced into the receiver from the antenna, and weak signals would have to compete with this noise.

7-14. Additional noise signals originating in atmospheric disturbances, the sun and other stellar sources, and in electrical machinery, increase the noise threshold below which even a perfect receiver could not detect signals. In the high-frequency band, from 2 to 30 megacycles, this threshold is usually much higher than that set by receiver internal noise sources. However, the external noise threshold is subject to many variations and it is possible that, under certain favorable combinations of conditions, the receiver noise could be a factor at frequencies in the upper half of the band. For this reason, the noise generated in the receiver circuit is an important consideration.

7-15. The internal noise generated in a receiver is conveniently described by a number called the "noise figure." The noise figure is expressed as the ratio in db between the noise level of the receiver and the noise level of a theoretically perfect receiver in which all the noise is assumed to be generated in the antenna by thermal agitation. A perfect receiver, in which the input circuit is designed to match the antenna resistance, has a noise figure of 3 db.

7-16. The sources of noise in a receiver are the input circuit resistance, the first tube grid circuit, and the second tube grid circuit (if the first tube gain is low). These noise sources are shown in schematic form in figure 7-2. For convenience, the tube noise is usually expressed as being equivalent to the noise generated in a resistance of the proper value, referred to as the equivalent noise resistance of the tube. A tube having a low value of equivalent noise resistance is a low noise tube. It will be found that triodes have lower noise than pentodes, and amplifiers have lower noise than mixers. Tube noise, although important, is not a decisive factor in tube selection. Pentode tubes offer advantages over triodes as amplifiers, since they have very low grid-to-plate capacity and give large gain without neutralization. Pentagrid mixers require less oscillator power, have excellent isolation between signal and oscillator circuits, and give high conversion gain.

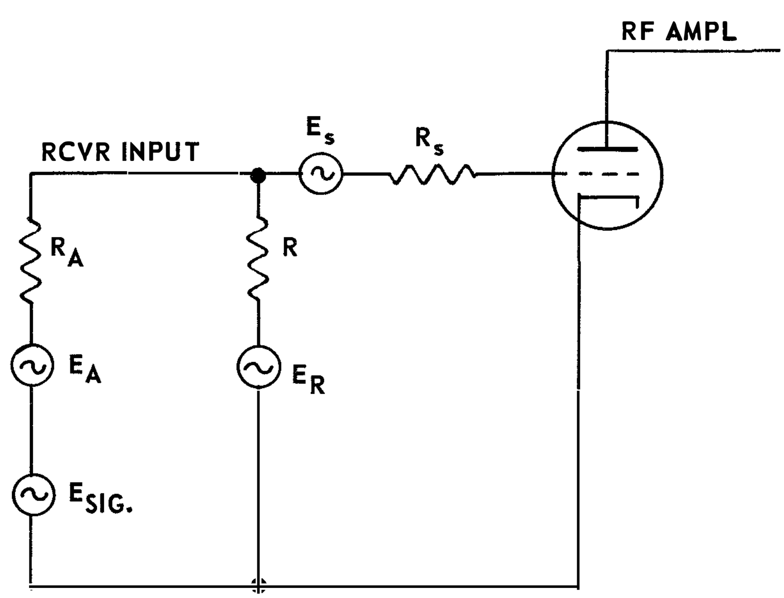


Figure 7-2. Noise Sources in the RF Section of a Single Sideband Receiver.

7-17. RF Selectivity.

7-18. For minimum spurious response, it would be best to provide all the selectivity ahead of the amplifiers in the receiver. This is impractical for several reasons. First, the high-frequency band spans a range of frequencies in which filters having the required selectivity would be large and difficult to tune. Furthermore, they would have such high insertion loss that the noise figure would be seriously degraded. In some applications, where degradation of the noise figure can be tolerated, and preselection is a necessity, RF filters are used. Usually a single tuned circuit is used between the antenna and the RF amplifier grid. The selectivity required adequately to suppress the various spurious signals is provided by a tuned filter between the RF amplifier and the mixer. The tuned filter may consist of several parallel-tuned LC circuits, interconnected by mutual inductance or capacitance. The number of tuned elements required depends on the Q-factor, frequency, and attenuation required.

7-19. Mixers.

7-20. The RF signal is translated in frequency from the operating frequency to the intermediate frequency by means of modulation in circuits commonly called "mixers." The problems encountered in using mixers in receivers are slightly different from those encountered in exciters. Referring to figure 7-3, it can be seen that a desired signal of 1500 Kc can be translated to an intermediate frequency of 500 Kc with a local oscillator having a frequency of 2000 Kc. Going further into the example, it can be seen that there are several signals that can enter the IF amplifier through the mixer. Some of these signals are listed in figure 7-3. The response at 2500 Kc is called the "image response," and is usually the most troublesome. The higher-order responses are attenuated in the mixer tube. Careful selection of the tube and its operating point is necessary to obtain maximum suppression of these responses.

7-21. IF Section.

7-22. Careful selection of frequencies used in the IF amplifier is necessary to avoid spurious responses. These responses occur whenever the spurious-response frequency coincides with the desired frequency. This type of response is referred to as a crossover, tweet, or birdie, and is illustrated in figure 7-4. A signal of 1001 Kc, when mixed with an oscillator signal of 1500 Kc, yields a desired signal of 499 Kc. Due to the nonlinearity of the mixer, another product is generated which has a frequency equal to the difference between the second harmonic of the signal and the oscillator frequency. Both of these signals are passed by the IF filter because they are only 3 Kc apart. These signals will be demodulated by the audio section to yield an audio output tweet of 3 Kc in addition to the desired output. Spurious responses are minimized if the intermediate frequency is kept as low as possible consistent with good image rejection.

7-23. As the range over which the receiver operates is increased, it becomes increasingly difficult to find frequency schemes which are reasonably free of spurious responses.

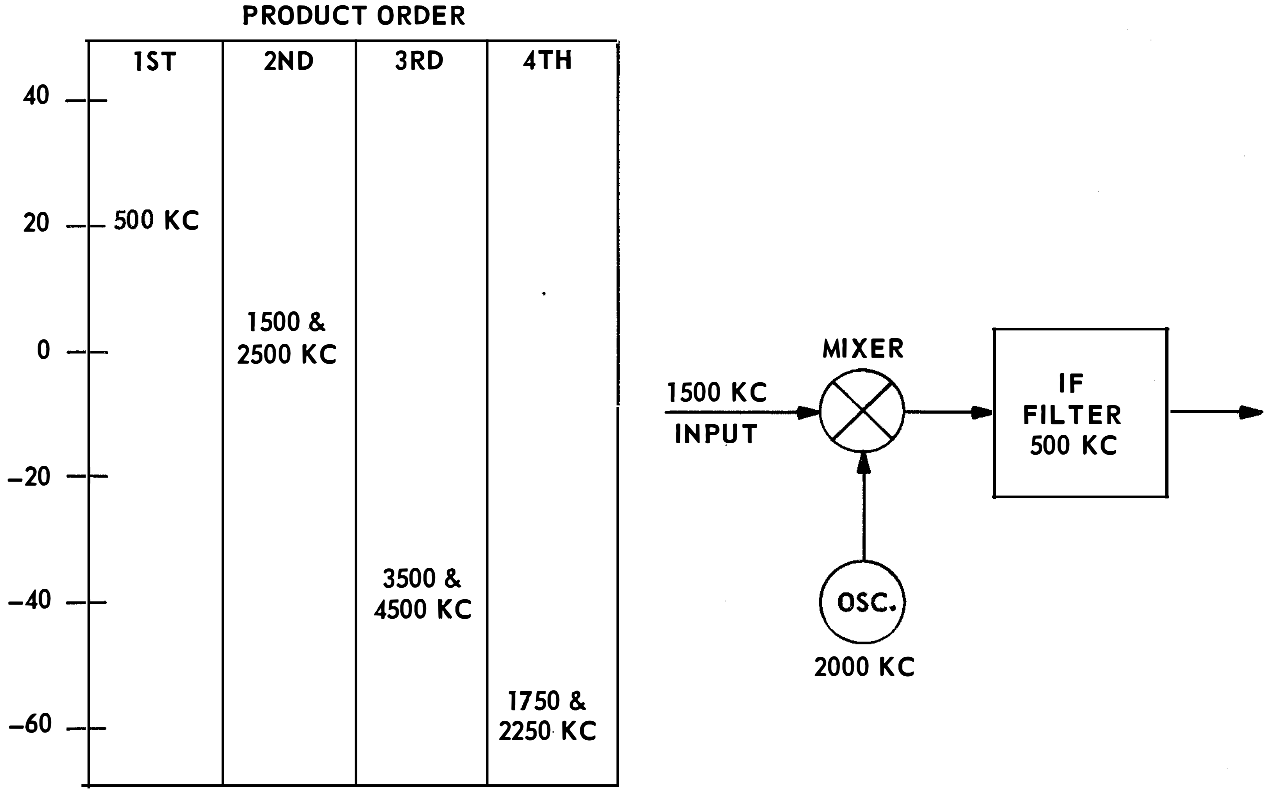


Figure 7-3. Typical Receiver Mixer Spurious Response.

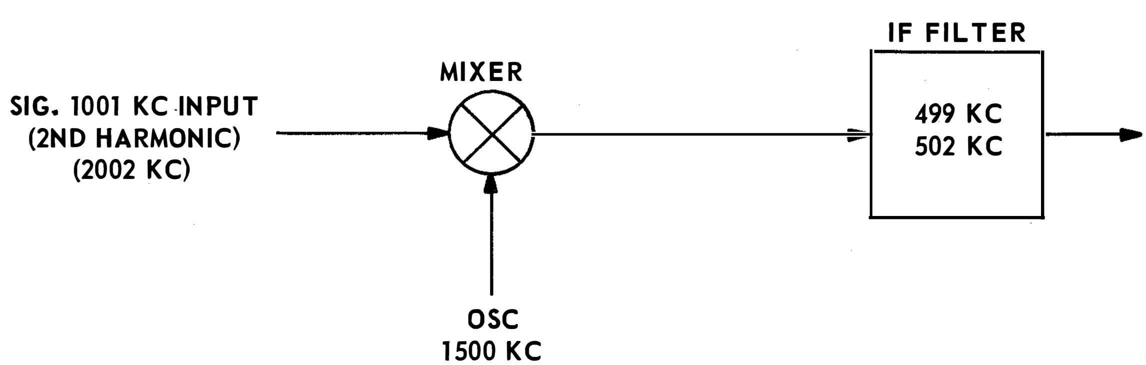


Figure 7-4. Receiver Mixer Crossover Response.

In order to keep these responses attenuated when covering the high-frequency band (2 to 30 megacycles), it is necessary to resort to double conversion, or the use of two intermediate frequency sections. Single conversion is then used on the low frequencies, and the second conversion is brought into use at high frequencies.

7-24. The use of double conversion makes possible an improvement of frequency stability through the use of a crystal-controlled, high-frequency oscillator. Tuning is accomplished by providing a variable first intermediate frequency ganged to the tunable low-frequency oscillator. As shown in table 6-1, the frequency stability of crystal oscillators is many times better than that obtained from tunable oscillators. Furthermore, the tuning rate remains the same on the high-frequency bands as it is on the low-frequency bands. On the low-frequency bands, the RF amplifier feeds directly into the tunable IF circuit, retaining the favorable ratio of signal-to-IF frequencies.

7-25. For the best sensitivity, it is desirable to have as much gain as possible ahead of the mixers. This would ensure that the signal level would be high enough to override completely the noise from the mixer. From the standpoint of strong signals, it is desirable to have low amplification until the selectivity of the receiver is effective. This would keep the level of strong adjacent-channel signals from becoming high enough to overload the initial stages of the receiver. These requirements for no amplification ahead of the selective filter for strong signal reception, and high gain in the RF amplifier for weak signal reception, conflict, so that a compromise is necessary.

7-26. When a receiver is tuned to a weak signal, and a strong signal is present outside the passband of the IF-selective filter, a type of interference known as cross modulation can exist. The selectivity of the RF-section circuits is not as good as the IF-selective filter, and there is a region near the operating frequency in which strong signals are accepted by the RF section. Due to the sharp selectivity of the IF circuits, these signals are not passed by the IF amplifier and, therefore, do not produce automatic gain control voltage. As a result, these large, interfering adjacent-channel signals are amplified by the RF amplifier along with the weak desired signals. When these interfering signal voltages are large enough to drive the amplifier and mixer tubes into nonlinear operation, they cause modulation of the desired signal. To minimize the generation of cross modulation interference, it is necessary to use care in selecting the tubes used in the RF section. The application of automatic gain control bias is helpful. As the desired signal level increases, the gain of the RF amplifier can be decreased, reducing the amplification of the interfering signals as well as that of the desired signal. It is necessary that the tube used in the RF amplifier retain its linearity with the application of variable bias. It is interesting to note that cross modulation is not as troublesome in single sideband reception. As an example, if both the undesired adjacent-channel signal and the desired signal are conventional AM signals with full carrier, the modulation of the undesired signal is readily transferred to the desired signal through the process of cross modulation. Effectively, the modulation on the undesired signal modulates the carrier of the desired signal. This undesired modulation is passed through the receiver as readily as the desired modulation and considerable interference results. In the case of single sideband suppressed-carrier reception there is no carrier present to be modulated and, therefore, the modulation is applied to each of the sideband signal components. As the single sideband signals consist of a number of relatively weak components, this undesired modulation is spread. Furthermore, when the single sideband is demodulated, the interfering signal is merely recovered as noise and is not as troublesome.

7-27. The intermediate frequency section contains the frequency selective filter elements and the principal amplifier stages.

7-28. Selectivity.

7-29. Consideration must be given to the bandwidth of the receiver as well as to that of the transmitter if the advantages offered by single sideband communications are to be realized. Optimum receiver selectivity occurs when the nose bandwidth (6-db point) is wide enough to pass the required intelligence, and the skirt bandwidth (60-db point) is narrow enough to reject an unwanted signal in the adjacent communications channel. Extremely steep skirts on the selectivity curve are required to obtain this optimum passband. Ideally, the ratio of the 60-db to 6-db bandwidth should be one-to-one.

7-30. The selectivity characteristic has generally been determined by comparing the shape factor, which is the ratio of the 60-db to the 6-db bandwidth. This basis of evaluation has developed from the problem of avoiding adjacent channel interference. While it is customary to define receiver performance in terms of shape factor, it is not always adequate. It can be shown that better shape factors are easier to obtain in wideband systems than in narrowband systems. The shape factor is a good comparison if the selectivity curves being compared have the same nose bandwidth. A better method of specifying the performance of a selective system is to define the selectivity in terms of nose bandwidth and the attenuation in db per kilocycle on the slopes of the selectivity curve.

7-31. It is desirable to place a selective filter in the circuit ahead of the amplifier stages, so that strong adjacent channel signals are attenuated before they can drive amplifier tubes into the overload region. These filters are very similar to the filters used in sideband generators to select the desired sideband while rejecting the undesired sideband. Electromechanical elements, piezoelectric elements, and inductance-capacitance elements can be used in these filters. In one respect, the requirements for these filters are different from those of sideband selecting filters used in the exciter. In order for the receiver to have good rejection of strong adjacent-channel signals, it is necessary for the filter used in the receiver to have the ability to reject signals outside the passband to a much higher degree. Attenuation of 60 db, or more, is necessary for this purpose and greater rejection is required under some conditions when receiving extremely weak signals. Since single sideband transmission occupies one-half the bandwidth of a conventional AM signal, the IF filter need be only one-half the bandwidth needed for AM.

7-32. Amplification.

7-33. The amplifier portion of the intermediate frequency section consists of the necessary amplifiers to build up the signal to a level suitable for the demodulator. This amplifier consists of cascade Class A linear amplifier stages using remote-cutoff pentode tubes. Tuned circuits may be used to provide the load resistance for these stages. The selectivity of these tuned circuits is helpful in improving the overall receiver selectivity, especially at frequencies which are down on the skirt of the selectivity curve. Some types of filters have spurious responses, outside the passband, which can be suppressed in this manner.

7-34. Automatic Gain Control (AGC).

7-35. A factor to be carefully considered in single sideband receiver design is the use of AGC. The basic function of AGC is to keep the signal output of the amplifier constant and thus hold audio output constant for changing signal levels. AGC is also applied to amplifiers in the RF section. However, it is important to delay the application of AGC voltage to the RF amplifier until a suitable signal-to-noise ratio has been reached. Conventional AM systems are generally not usable, since they operate on the level of the carrier and this carrier is suppressed in single sideband. AGC systems must be used which obtain their information directly from the modulation envelope. This can be done with conventional diode rectifiers and additional amplification. This may be a DC amplifier or an AC amplifier using the IF frequency. Special care must be taken to isolate the AGC system from the reinserted carrier, since it is a large signal of the same carrier frequency as the IF signals. This problem can be avoided by developing the AGC voltage from the audio signal. In either case, the time constant of the system is very important. The control must be rapid enough to prevent strong signals from coming through too loud initially, yet slow enough to prevent the following of syllabic variations of normal speech. One solution to this time constant problem is to use a fast-charge, slow-discharge type of circuit. Circuits having a charge time of 50 milliseconds and a discharge time of 5 seconds have proven quite successful. Consideration should also be given to dual time constant circuits having a ratio of 100 to 1. Such a circuit allows the rapid signal changes to develop a control voltage across one RC network, and the slow signal variations to develop a control voltage across another RC circuit. These two voltages can then be applied in series, or to different stages, to give the desired control characteristics. Such a dual time constant circuit is similar to a rapid AGC system used in conjunction with a manual gain control.

7-36. The Audio Section.

7-37. The information carried by the single sideband signal is recovered and amplified to a level suitable for the audio output circuits. The circuits used to recover this audio intelligence perform the same function as the modulator in the exciter; therefore, the same circuits can be used. The single sideband is first combined with a local carrier. The local carrier must have a proper frequency relationship with the sideband components for faithful reproduction of the original audio signal. In the demodulator the single sideband signal is used to modulate the local carrier. The demodulator output consists of an audio signal and several RF outputs. These signals are easily filtered by passing the output of the modulator through an audio low-pass filter. It is necessary to maintain the proper frequency relationship between the sideband signal and the local carrier. If the received signal is an upper sideband, the carrier frequency is below the sideband, and if the received signal is a lower sideband, the carrier frequency is above the sideband signal. If a receiver must be capable of receiving either upper or lower sideband, it is necessary to provide a means of changing the relative position of the carrier with respect to the sideband. One way of accomplishing this is to use two sideband filters in the IF section and a single carrier at the demodulator. The desired sideband is then selected by switching in the proper filter or using both for dual single sideband. A single filter can be used for upper or lower sideband reception by providing a means of shifting the local carrier from one side of the IF filter passband to the other and retuning the oscillators in the RF section.

7-38. It seems likely that, for commercial communications, a standard may be set using upper sideband only. For military and amateur receivers, provision should be made for the selection of either upper or lower sideband. In a receiver, the problem is to reinsert the carrier on the proper side of the signal. In stabilized, step-tuned systems, it is usually easier to move the passband to the proper side of the inserted carrier by switching IF selectivity. This is especially easy if the IF selectivity is of the concentrated (filter) type. However, if the local carrier is tunable, as with a single BFO, it would be easier and more economical to move the carrier to the opposite side of the passband. Another method of selecting sidebands in a receiver is referred to as "passband tuning." In this system, the BFO is mechanically coupled to the main tuning oscillator so that they track together, moving cycle for cycle. A CW beat note will not change as the BFO is tuned but will move across the passband of the receiver. This allows the BFO to be set on the carrier frequency of a station and either sideband to be received by simply tuning the passband control from one side of the passband to the other. This is also useful for orienting the single sideband signal to a position in the passband with minimum interference.

7-39. If a receiver is to be used for single sideband reception only, and is to be used with a synthesized frequency control system, it is desirable to use a fixed-frequency oscillator for the local carrier, placed in the correct relationship to the IF passband. The required oscillator output voltage depends upon the demodulator circuits used and the input signal level to the demodulator. To reduce hum and noise products, the detector levels are often in the range of 0.05 to 0.5 volt. To reduce distortion, carrier levels at least 20 db higher than this are preferable (a minimum of 0.5 to 5 volts). Care should be taken to keep the shields and filter assemblies of the oscillator circuits intact when repairing these systems; this is to prevent coupling to any circuit of the receiver other than to that intended.

7-40. The Product Demodulator.

7-41. Product demodulator circuits are preferred in SSB reception because they minimize intermodulation distortion products present in the audio output signal and do not require large local carrier voltages. See figure 7-5. The sideband signal from the IF amplifier is applied to the control grid of the dual-control pentode tube V1 through transformer T1. The carrier is applied to the other control grid. The desired audio output signal is recovered across resistance R4 in the demodulator plate circuit. Since this circuit is controlled by both grids acting simultaneously, the plate current will contain frequencies equal to the sum and difference between the sideband and carrier frequencies. There will also be components of plate currents having a frequency equal to the carrier frequency and the sideband frequency. These components are suppressed by a low-pass filter (L1, C5, and C6) and the desired audio signal is passed to the audio amplifier. The frequency spectrum presentation in figure 7-5 shows the principal components present in the demodulator plate current. It is assumed here that the sideband signal consists of three components having frequencies of 501, 502, and 503 Kc. The carrier frequency is 500 Kc. The plate current components consist of three AF components of 1, 2, and 3 Kc, and three RF components of 1001, 1002, and 1003 Kc, as well as a carrier and the original sideband frequencies. By constructing a low-pass filter in the plate circuit consisting of L1, C5, and C6, it is possible to filter out all frequencies except the difference frequencies. By this method, the audio frequency can be recovered from the RF sideband signal.

7-42. Noise Reduction.

7-43. Since there is no general method of reducing the effects of random noise in radio communications systems other than by using higher transmitter power, more sensitive receivers, and complex modulation schemes, much interference is caused by impulse noise, such as that produced by automobile ignition systems and switch contacts. This type of noise must have a high amplitude to contain enough average energy to cause interference. It is, therefore, possible to discriminate between noise and signal on the basis of amplitude. The greater the noise amplitude and the shorter the noise pulses, the easier it is to reduce or eliminate the noise without affecting the desired signal.

7-44. Successful SSB reception requires a fast AGC attack and a slow decay. From the standpoint of impulse noise interference, this may be unfortunate, because the AGC is susceptible to loading up in the presence of noise, thus reducing receiver sensitivity. An AGC detector noise limiting circuit provides some relief. However, the narrower bandwidth and the steeper selectivity skirts usually employed in single sideband reduce the amplitude of the noise pulses and stretch them out in time so that it is difficult for the AGC noise limiter to discriminate on the basis of amplitude. Moreover, there is no carrier to set the clipping level automatically. In addition to causing AGC line loading, impulse noise also reduces the receiver signal-to-noise ratio. Attempts to use conventional noise limiters at the input or output of the demodulator meet with the same objection as the AGC limiter, because impulse noise is reduced in amplitude and

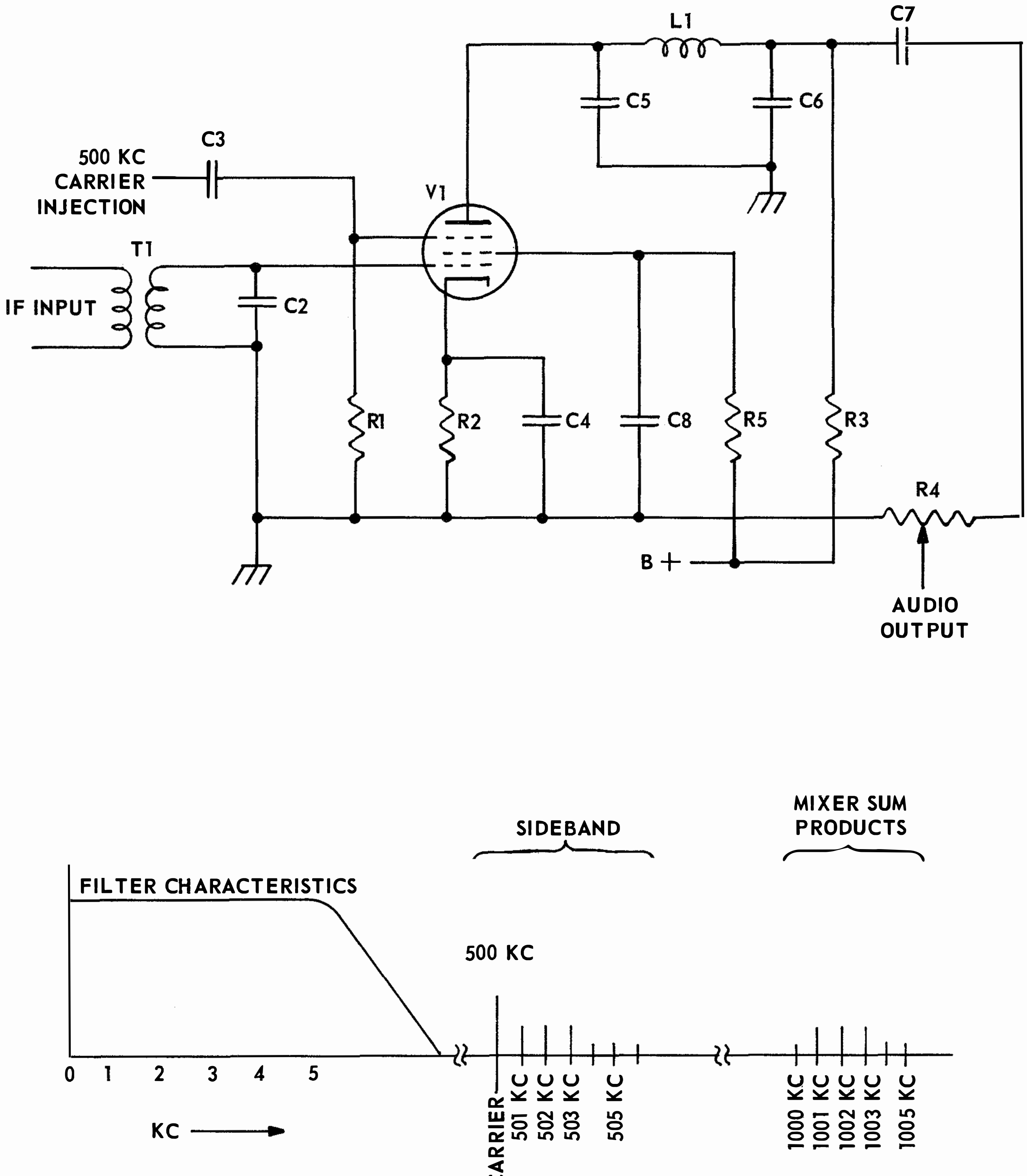


Figure 7-5. Typical Product Demodulator.

stretched out in time. In both cases, it is desirable to reduce or eliminate the noise before the bulk of the receiver selectivity. This, in a filter-selectivity receiver, is just before the IF filter.

7-45. There are two principle methods of impulse noise reduction. These are silencing (blanking), and limiting (clipping). A silencer (blanker) interrupts the signal for the duration of the noise pulse, removing both signal and noise for a short period of time. In the presence of an AM carrier, this constitutes downward modulation and produces a spectrum of sidebands, some of which pass through the IF, are detected, and cause a residual audio noise output. In suppressed-carrier SSB, there is no carrier to modulate, so the only effective blanking is to remove small portions of the signal. If the blanking periods are short, these small holes in the signal are not noticeable.

7-46. The second method of reducing noise in the IF amplifier is by limiting (clipping). Ideally, the clipper would have no effect on the desired signal but would simply clip off noise peaks which were higher in amplitude than the signal. In designing a clipper, a choice must be made as to the location of the clipper in the circuit. It is desirable to have sufficient selectivity before the clipper so that it does not clip adjacent-channel signals and produce intermodulation products in the desired signal channel. Clipping should be done before the bulk of the selectivity for much the same reason as with blanking. This is particularly important in receivers employing an IF filter with a flat-topped and steep-sided selectivity curve.

7-47. Phase Shift Method of Detection.

7-48. A phase shift method of detection may be used in single and double sideband receivers to eliminate the need for sideband filters. This method is similar to the phase shift method of single sideband 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 7-6. Since no filters are necessary in the IF stages, these stages may employ conventional AM circuitry, and the chief requirement in the circuits preceding the detector stage is oscillator stability. Thus, the output of a conventional AM receiver IF section may be applied to an adapter using the circuitry shown.

7-49. 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 wideband 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. Frequencies slightly above and below the desired limits of the filter will be passed but may be somewhat distorted because of a shift in phase greater or less than 90 degrees.

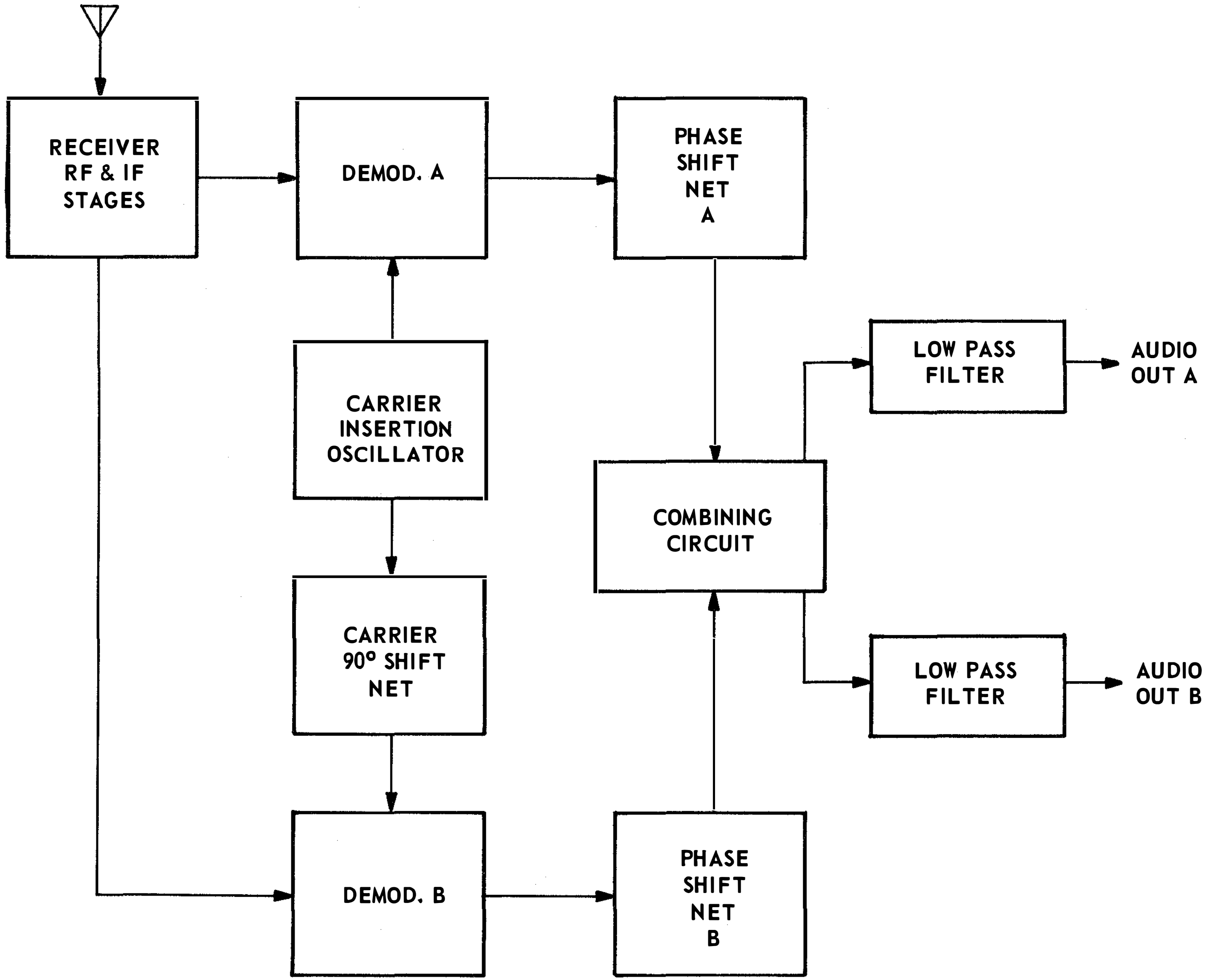


Figure 7-6. Phase-Shift Method of Detection.

7-50. The 90-degree phase-shift network in the carrier-oscillator signal input circuit to the demodulators and the two diode 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. 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 circuit in the tube is added to the total phase shift appearing in 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, appears in the output of the B channel in the same amplitude as in the output of A channel but exactly 180 degrees out-of-phase with it (or both of the same polarity at opposite ends of the resistive combining networks, since the circuit arrangement is push-pull). This method may be used for dual-channel single sideband demodulation by variations 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.

7-51. 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 previously described 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 obtained in the amplitude of the signal from both A and B audio networks so that only vectorial addition 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 shifts are maintained in the carrier-to-detector circuit.

7-52. Single Sideband Dual Channel Receivers.

7-53. For many years single sideband receivers have been used for long range point-to-point fixed ground communications. 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 combinations are expensive and require the allocation of additional frequency channels because each transmitter would have to operate at a different frequency.

7-54. In single sideband 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 carriers, overseas communications, and relay links. With the dual-channel system, twice as many channels are available for a given number of transmitters and companion receivers. Dual sideband signals are received, amplified, and heterodyned in the same manner as in the single sideband 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 at this point. Actually, both sidebands and their common carrier are applied simultaneously to all three amplifiers. The filter included in each of these 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.

7-55. 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 communications may be impaired. This precaution is as important in the receiver as it is in the transmitter.

7-56. Double sideband single-channel signals may also be received with this receiver. For such reception, 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. If the lower sideband is received, it is taken from channel B. In either case, the opposite channel is merely inoperative. The dual channel, single-sideband receiver then may be considered as two single sideband receivers combined into one.

7-57. Diversity Receiving Systems.

7-58. In long range communications, fading becomes a serious problem because of the effect 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. 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.

7-59. The distance between points where a sky wave strikes the earth, or between the transmitting 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 the signal from the direct ground wave, if the receiving stations are within the relatively short range of the ground wave.

7-60. Propagation conditions affect different frequencies differently and play a prominent role in medium 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 difference in fading of the received sidebands, and the use of exalted-carrier reception, with single or double sideband 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 communications and may be used to great advantage in single sideband reception.

7-61. 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 7-7. When the signal arriving at one antenna is in a fading condition, the same signal received at another antenna some distance away may be of maximum amplitude. When more antennas are added, this condition becomes 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, AVC and control circuits, may be used.

7-62. Advantages of several methods are combined to produce even greater flexibility when single sideband receivers are used in space diversity systems. Since dual-channel single sideband receivers provide for simultaneous reception of both sidebands separately, this feature is ideally suited for diversity systems. Also, since single sideband receivers may use exalted carrier in lieu of added carrier, double sideband 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.

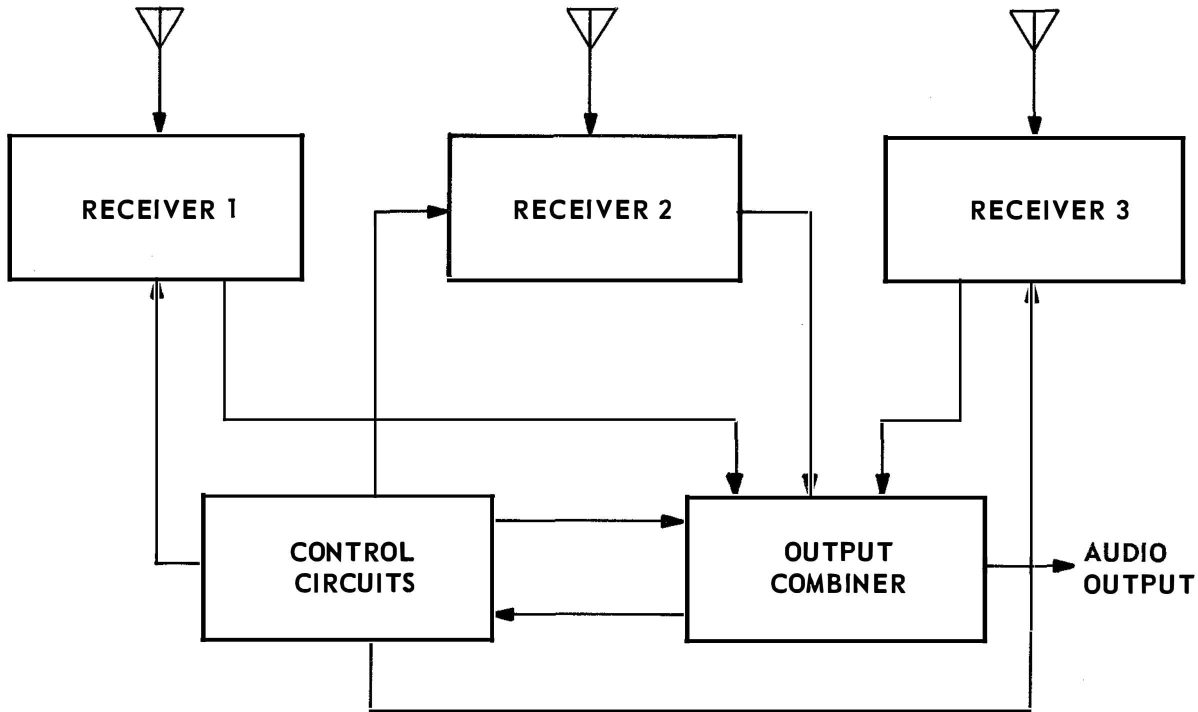


Figure 7-7. Diversity Receiving System.

7-63. Factors of reliability, flexibility, cost, or type of transmission will govern the selection of the various combining networks. Three basic types are in use. One is basically a common load resistor, across which all the received outputs are applied. This is called the additive type. Another uses some form of gathering 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 and may take a variety of forms. Combiners for data transmission may be included in this third category.

7-64. 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 B lower sideband channels are also combined in a separate circuit so that dual-channel single sideband (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 AFC 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 AFC circuit. On some equipments a switch is provided which selects the reconditioned ("exalted") carrier to demodulate the sidebands instead of the carrier oscillator. AFC may be applied to the medium-frequency 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.

7-65. The AVC may operate from the combined carrier, or it may be switched, permitting operation from the combined outputs of all sideband channels or from 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 in the event of excessive noise, interference, or jamming of the carrier.

7-66. This type of receiving equipment is usually rack mounted, and much flexibility can be incorporated in the design of such equipment. Unfortunately, the requirement for wide spacing of antennas limits the use of diversity systems to large installations, especially at the lower frequencies. The same fading conditions that render a space diversity system advantageous at the 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 exist. Spacing of antennas more than one wavelength apart does not present as much of a problem in the VHF and UHF range of frequencies. Another form of diversity quite often used on shipboard installations is frequency diversity. This is used in lieu of space diversity and, of course, is next best.

SECTION VIII

STABILIZED MASTER OSCILLATORS

8-1. Technical Requirements.

8-2. The frequency accuracy requirements for single sideband communications are very precise when compared with most other communications systems. A frequency error in carrier reinsertion of 20 cps or less will give good voice reproduction. Errors of only 50 cps result in noticeable distortion, and intelligibility is impaired when the frequency error is 150 cps or greater.

8-3. Significant frequency errors are introduced by the propagation medium and by Doppler shifts due to relative motion between transmitter and receiver in aircraft communications. In HF skywave transmission, the Doppler shifts caused by the motion of the ionosphere introduce frequency shifts of several cycles per second. Doppler shift due to relative motion amounts to one part in 10^6 for every 670 miles-per-hour difference in velocity between the transmitting and receiving station. At a carrier frequency of 20 Mc, communicating from a jet aircraft to ground, the frequency shift will be approximately 20 cps. Inasmuch as this represents approximately one-half of the nominal maximum permissible frequency error, the errors introduced by the transmitting and receiving equipment must be comparatively small. This dictates a design goal in the vicinity of $\pm 1/2$ part in 10^6 in both ground and aircraft installations.

8-4. Present day trends demand that communications be established on prearranged frequencies without searching a portion of the spectrum in order to obtain netting; therefore, the figure of ± 0.1 part in 10^6 represents the required absolute accuracy rather than short-term stability. Most military and some commercial applications demand that operation be obtained on any one of the seven thousand SSB voice channels in the HF band. A channel frequency generator having an absolute accuracy of ± 0.1 part in 10^6 ($\pm 0.00001\%$) and providing either continuous coverage or channelized coverage in steps no greater than 4 Kc is required in many SSB systems.

8-5. AFC vs Absolute Frequency Control.

8-6. To meet the stringent frequency control requirements, early HF single sideband systems utilized various methods of automatic control of the reinserted carrier at the receiver. Either a pilot tone or carrier was transmitted along with the sideband components and the receiver frequency was synchronized with the transmitter frequency. No stabilization of the transmitter frequency was used other than that obtained by using crystal-controlled oscillators.

8-7. The first single sideband radiotelephony system did not use automatic frequency control but was able to accomplish its purpose because the carrier frequency employed, about 60 Kc, was in the low-frequency portion of the spectrum for which oscillators of sufficient stability were available. Although oscillators have long been available with sufficient frequency stability and accuracy for use in high-frequency single sideband equipment, these oscillators were bulky, fragile, and limited in frequency channels. They were used principally as laboratory frequency standards. Improvements in the crystal art, development of circuit techniques, and new components have made available the means to obtain HF receivers and transmitters capable of multichannel operation with sufficient frequency accuracy and stability for independent operation of the receiver.

8-8. The advantages obtained through the use of independent, absolute frequency control are considerable. The bandwidth required for a communications channel is minimized since there need be no allotment for the synchronizing signal and the frequency tolerance. The relationship between transmitter and receiver carriers is absolute and indestructible and is immune to any type or degree of interference, resulting in maximum fidelity of the received signal. Even in the extreme cases where Doppler effects introduce sufficient frequency shift to upset the system, making some form of automatic frequency correction necessary, the use of absolute frequency control assures that the bandwidth and, therefore, the interference susceptibility of the AFC circuit will be minimized.

8-9. Development of Frequency Control.

8-10. It is of some interest to trace the development of frequency control circuits and the technical and economic forces that caused their evolution. In the early days of radio, the tunable LC oscillator provided a simple and serviceable answer to the problem of generating channel frequencies. The lower end of the frequency spectrum and amplitude modulation were in use, and the spectrum was not unduly crowded. Later, crowding of the spectrum was alleviated by closer channel spacing and expansion into the higher frequency regions. The increased frequency accuracy required was provided by crystal oscillators, and a multiplicity of channels was provided by a like number of crystals. In World War II, the logistics of delivering the right crystal to the right place at the right time became untenable.

8-11. It became apparent to those involved in multichannel equipment that the simple master-oscillator power-amplifier (MOPA) circuit would no longer provide the desired flexibility. A choice of one of hundreds of channels was required at the flick of a switch, guard bands were narrowed, VHF bands were pressed into more extensive service and, under these circumstances, the multiple-crystal synthesizer was evolved. The principle was simple: the output frequencies of several crystal oscillators were mixed together to produce the desired output frequencies. Each oscillator was provided with a means of selecting one of ten or more crystals so that a large number of channel frequencies might be synthesized. This principle is illustrated in figure 8-1. It would be technically and economically unfeasible to maintain all the crystals in a multiple-crystal synthesizer to the required accuracy. It would be more practical to place all the stability requirements in one or, at the most, several highly stable oscillators. From this challenge has emerged several operationally satisfactory types of single-crystal synthesizers.

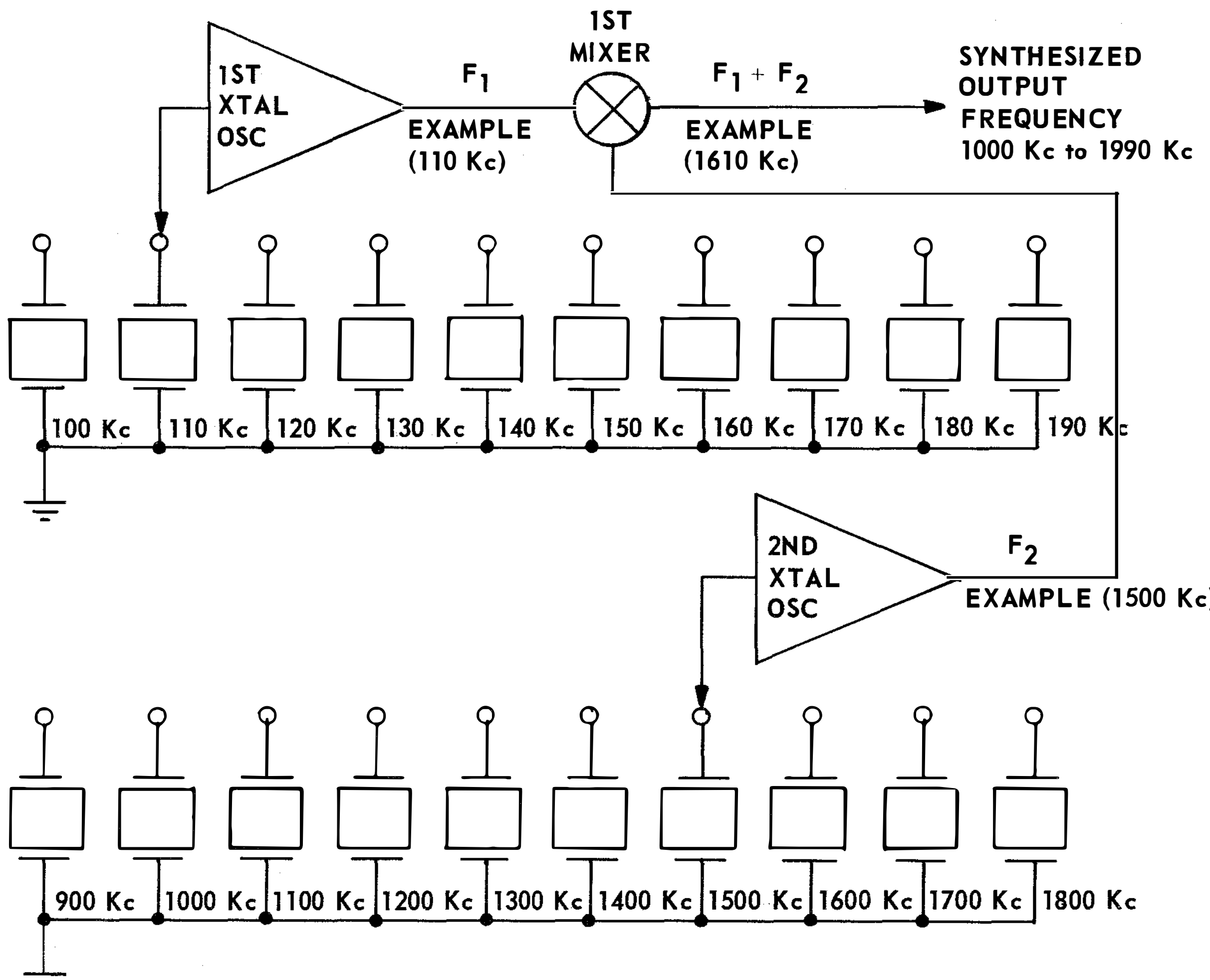


Figure 8-1. Multiple-Crystal Frequency Synthesizer.

8-12. Frequency Synthesizers.

8-13. The frequency synthesizer is basically a circuit in which harmonics and sub-harmonics of a single standard oscillator are combined to provide a multiplicity of output signals which are all harmonically related to a subharmonic of the standard oscillator. A simple block diagram of such a synthesizer is shown in figure 8-2. A great advantage of this circuit is that the accuracy and stability of the output signal is

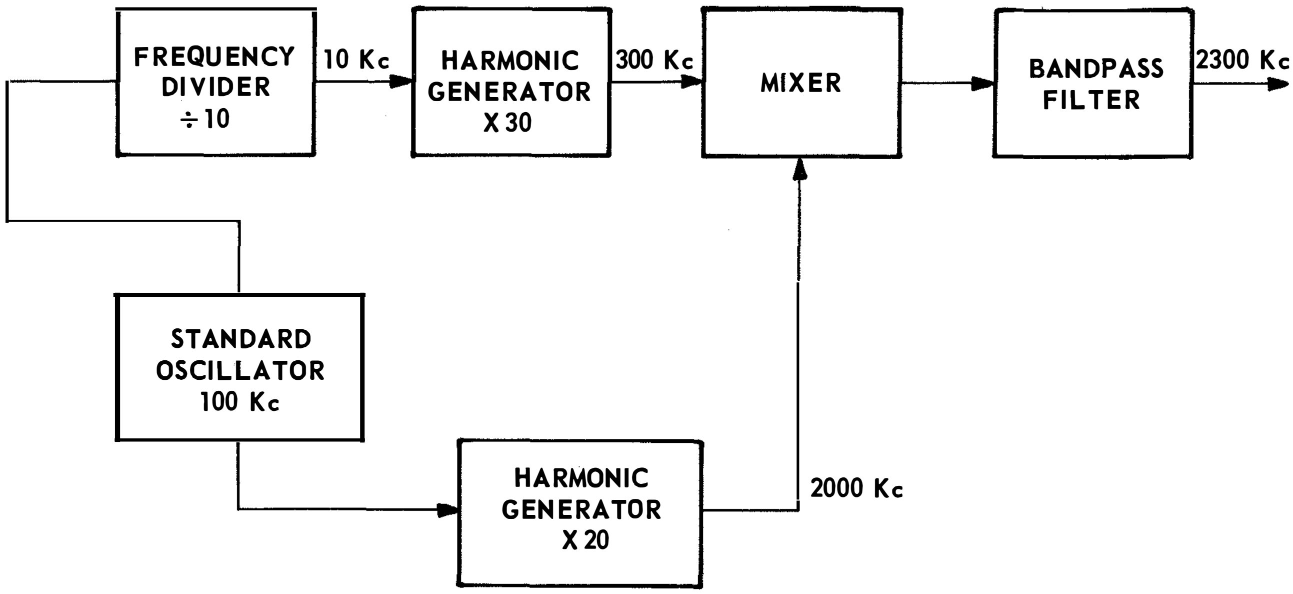


Figure 8-2. Single-Crystal Frequency Synthesizer.

essentially equal to that of the standard oscillator. The problems involved in building a single-frequency oscillator of extreme precision are much more simple than those associated with multifrequency oscillators. Furthermore, as techniques improve, the stability of the synthesizer is readily improved, because it is necessary only to replace the standard oscillator to obtain improved precision. The primary difficulty encountered in the design of the frequency synthesizer is the presence of spurious signals generated in the combining mixers. Extensive filtering and extremely careful selection of operating frequencies are required for even the most simple circuits. Spurious frequency problems increase rapidly as the output frequency range increases and the channel spacing decreases.

8-14. Harmonic Generators.

8-15. The generation of harmonics of signals from low-frequency sources is a rather difficult problem when carried to higher order harmonics. To obtain stable signals which are the exact multiples of a low frequency, several schemes can be used. An ordinary class C amplifier can be used for harmonics up to the ninth. Diode clippers, yielding square or rectangular waveforms, provide much higher harmonics but have limited amplitude capability. A blocking oscillator synchronized to the reference frequency generates short, sharp pulses which contain considerable harmonic energy. A particularly effective harmonic generator can be devised using a keyed oscillator (see figure 8-3). In this circuit, the low-frequency reference signal is shaped by a clipper to provide an on-off keying signal to control a free-running oscillator tuned to

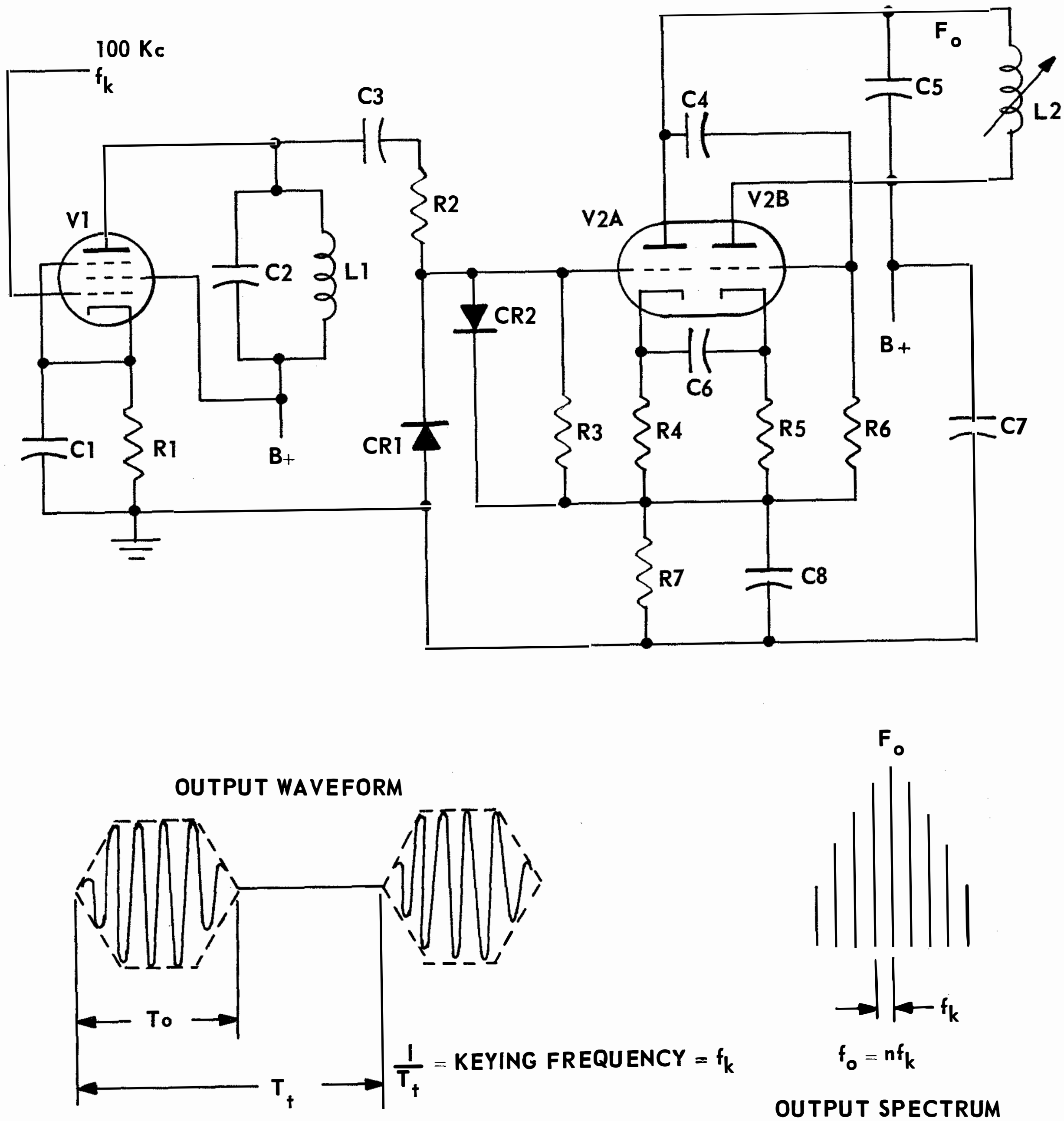


Figure 8-3. Harmonic Generator.

the approximate frequency of the desired harmonic of the keying signal repetition frequency. The resulting oscillator output is a train of RF pulses. If the keying signal is sharply defined and the oscillator starts oscillation uniformly, each pulse will begin on the same RF phase. The output waveform will then be as shown in figure 8-3. The spectrum of this wave consists of a number of components having various amplitudes grouped around the oscillator free-running frequency. The frequency of each component is an exact integral multiple of keying-signal repetition frequency.

8-16. Harmonic Frequency Selectors.

8-17. The problem now is to select the desired harmonic while rejecting the adjacent, undesired harmonic. Such a selection requires very sharp filters as the frequency range increases and the spacing between harmonics decreases. By means of an additional mixer, it is possible to relieve this situation. Such an arrangement is shown in figure 8-4. In this case, it is desired to select higher-order harmonics from a 1-Kc source. The 1-Kc reference signal is applied to a harmonic generator, the output of which is tuned to approximately 2.4 Mc. A considerable number of 1-Kc harmonics will be contained within the harmonic generator output. This harmonic generator output is fed to mixer Number 1, along with a local oscillator signal at 1945 Kc. The desired output of 455 Kc is selected by means of a mechanical filter having a bandwidth of less than 1 Kc. This signal is fed to mixer Number 2, along with the output of the same oscillator used to drive mixer Number 1. The desired product of 2400 Kc is selected in the output filter. To select the adjacent 1-Kc harmonic of the reference signal, the local oscillator is moved to a frequency 1 Kc higher. The desired output is then 2401 Kc. The stability of the output signal is dependent entirely upon the stability of the 1-Kc reference signal. The local oscillator frequency accuracy need only be such as to keep the desired 455-Kc signal within the passband of the mechanical filter.

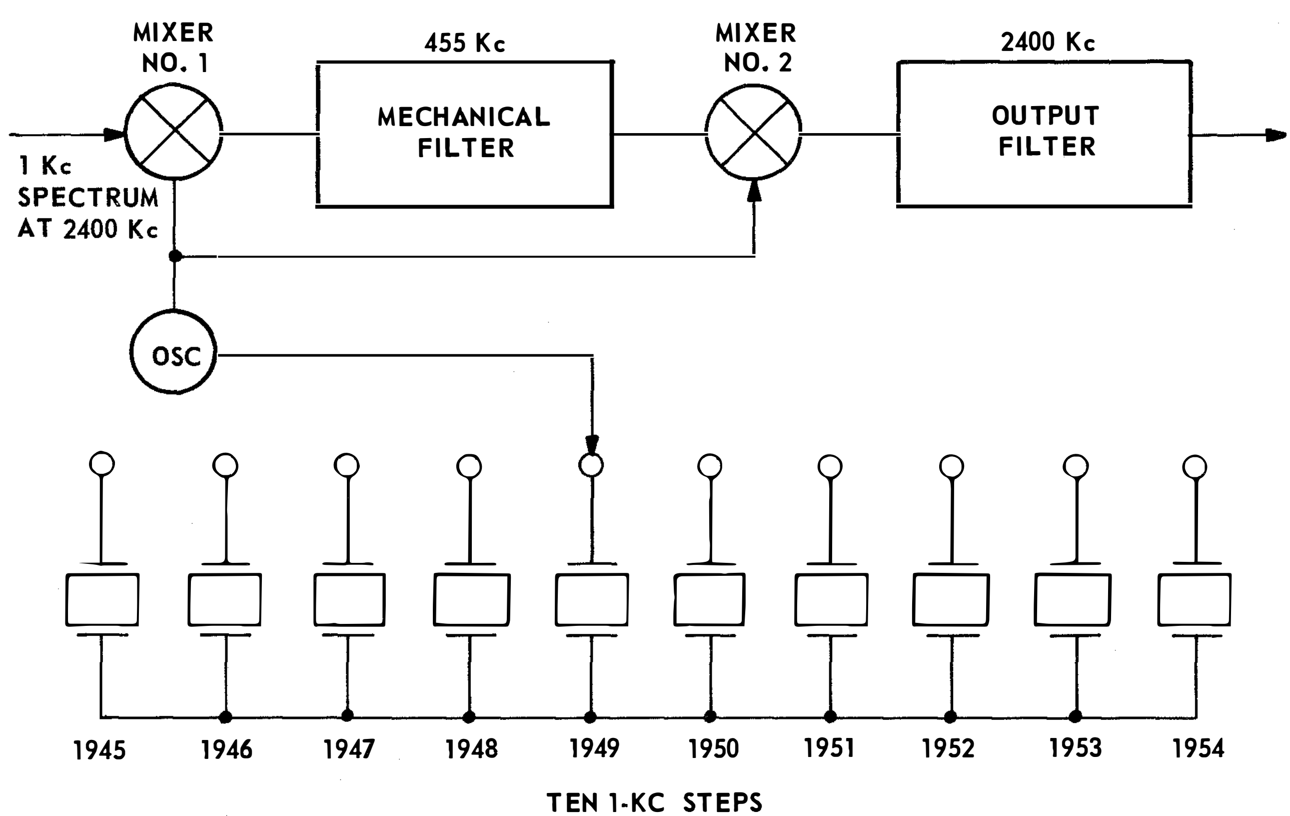


Figure 8-4. Harmonic Frequency Selector.

8-18. The Stabilized Master Oscillator.

8-19. It is possible to retain the advantage of the frequency synthesizer and avoid many of the spurious frequency problems by using the synthesizer to provide a reference signal to control the frequency of a variable-frequency master oscillator. Such a circuit has come to be known as a stabilized master oscillator (SMO). The basic elements of an SMO circuit are the master oscillator, reactance control, and discriminator (see figure 8-5). The frequency of the master oscillator is determined by the inductance and capacitance of elements L1 and C1. The frequency of oscillation may be changed manually by varying the capacitance of C1, or electrically by varying the permeability of the core on which L1 is wound.

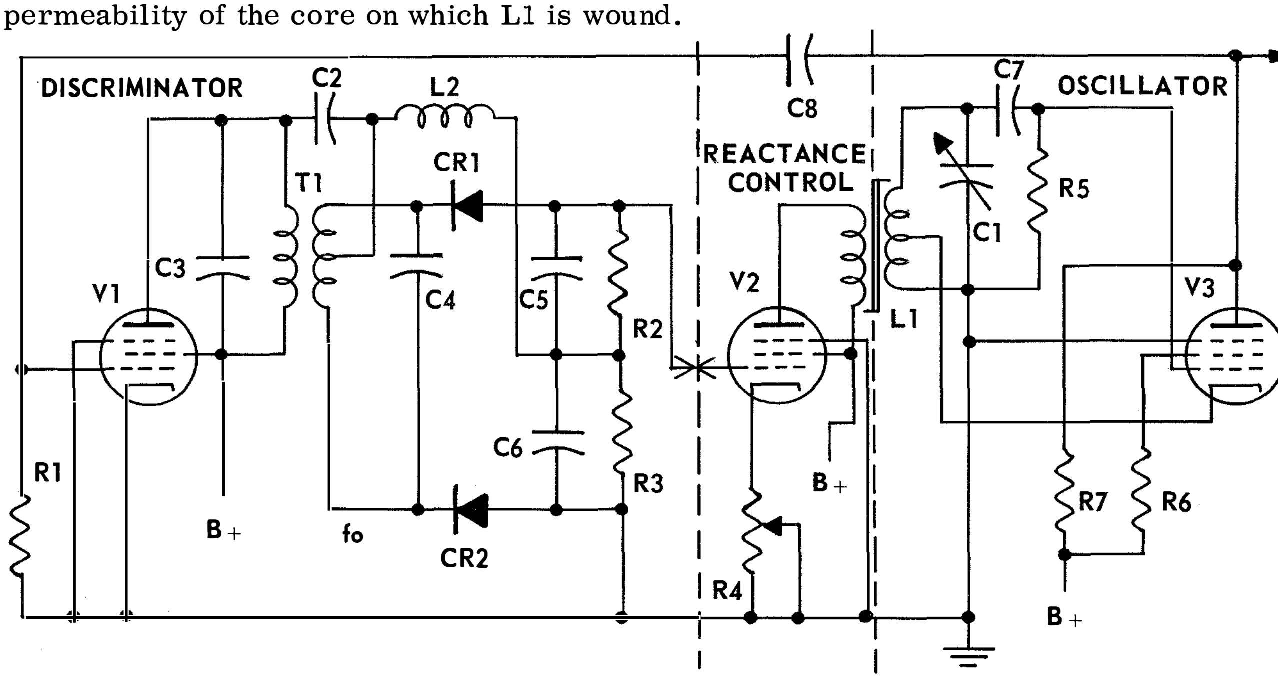


Figure 8-5. Simple Stabilized Master Oscillator With Frequency Discriminator.

8-20. Frequency Discriminator.

8-21. The operation of the frequency discriminator is such as to provide a DC output signal whose amplitude and polarity are determined by the relationship between the input signal frequency and the frequency to which the discriminator is tuned. The frequency discriminator consists of a double-tuned transformer and two diode rectifiers (see figure 8-6). The transformer is used to supply signals to the two rectifiers, the outputs of which are series connected. Coupling capacitor C1 places the centertap of the secondary at the same RF potential as the plate end of the primary winding. As a result of this connection, the voltage applied to each diode is the sum of one-half the secondary voltage and the voltage appearing across the primary. The voltage applied to each diode is shown in the vector diagrams following the circuit diagram. The action of the discriminator depends on the fact that the phase of the voltage developed across the secondary of the discriminator transformer will vary as the frequency of the

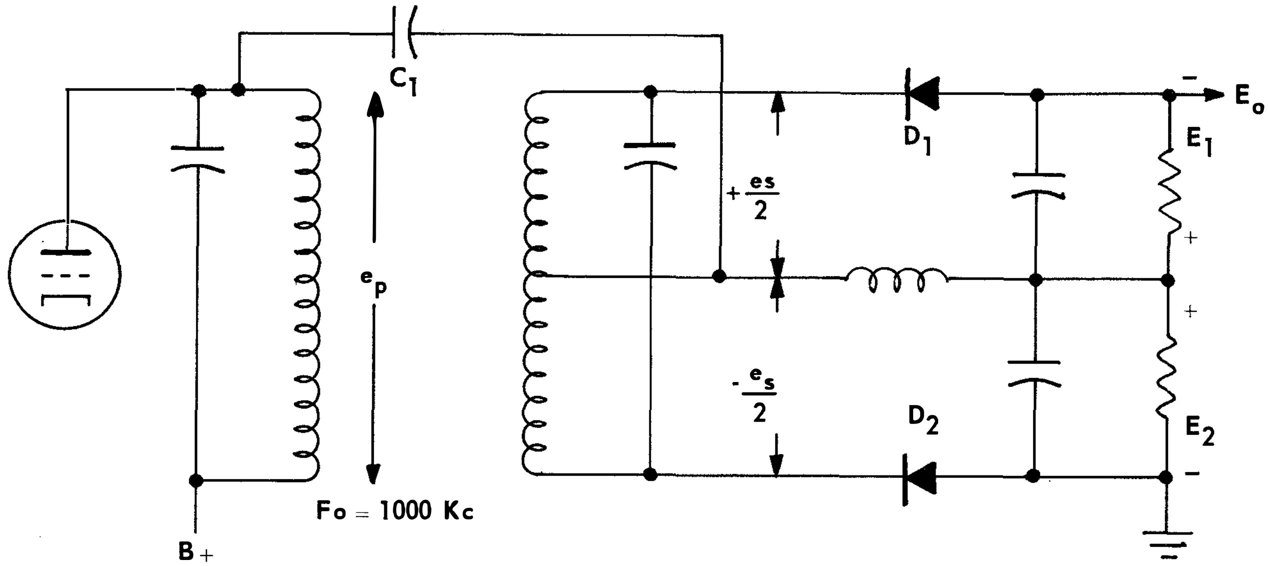


Figure 8-6. Frequency Discriminator Circuit Diagram.

applied signal is varied above and below the transformer resonant frequency. Referring to the vector diagrams (figure 8-7) it can be seen that if the applied signal frequency is equal to the discriminator frequency, equal voltages are applied to each discriminator diode and the DC output of the discriminator is zero (figure 8-7B). If the applied frequency is higher than the discriminator frequency, the voltage applied to diode No. 1 exceeds that applied to diode No. 2, and the resulting DC output is positive (figure 8-7A). If the applied signal is lower than the discriminator frequency, the voltage applied to diode No. 1 exceeds that applied to diode No. 2, and the resulting DC output is negative (figure 8-7C). This DC output is applied to a reactance control device.

8-22. Reactance Control.

8-23. The reactance control provides the means by which the DC output of the discriminator is made to alter the inductance or capacitance of the tuning elements of the master oscillator. Devices that have been used for this are reactance-tube circuits, saturable reactors (current-sensitive inductors), voltage-sensitive capacitors, and motor-driven variable capacitors. The saturable reactor is used in the example given, as its operation is easily understood. The saturable reactor consists of an inductor wound on a core material having magnetic permeability which is a nonlinear function of the magnetizing force. Such a reactor will have inductance which can be changed through an auxiliary control winding or by varying the current in its winding (see figure 8-8). The change in permeability will be the same for either polarity of magnetizing force. For this reason, it is necessary to resort to fixed-magnetic bias to obtain an

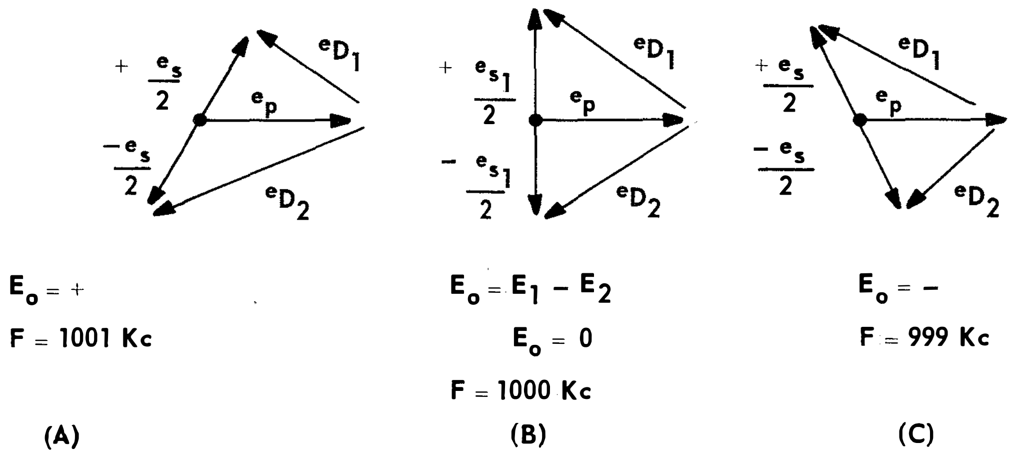


Figure 8-7. Frequency Discriminator, Vector Diagrams.

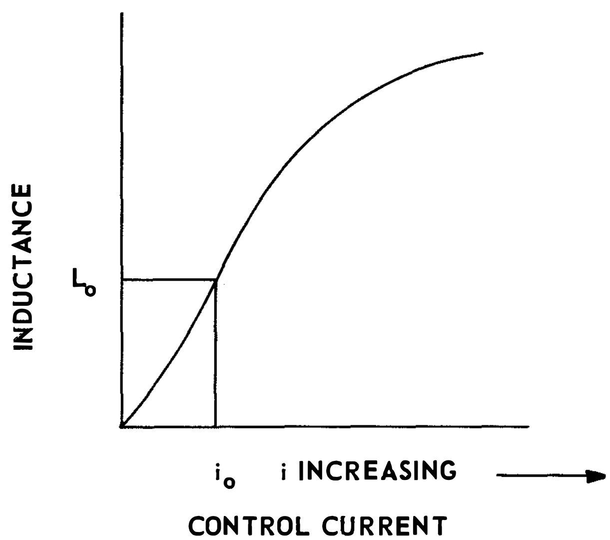


Figure 8-8. Variable Inductor Response Curve.

inductance change that will reverse polarity when the external magnetizing force polarity reverses. The magnetic bias may be obtained from a permanent magnet or from a bias current in the control winding as is the case in the example shown.

8-24. Basic SMO Operation.

8-25. The manner in which the SMO circuit operates may be described in two conditions, open-loop and closed-loop. If the control is opened at the grid of the reactance control tube and the tuning of the oscillator varied with the discriminator tuning fixed at F_O, the output voltage of the discriminator will follow the open-loop curve shown in figure 8-9. In this curve, the discriminator voltage is plotted on the vertical scale

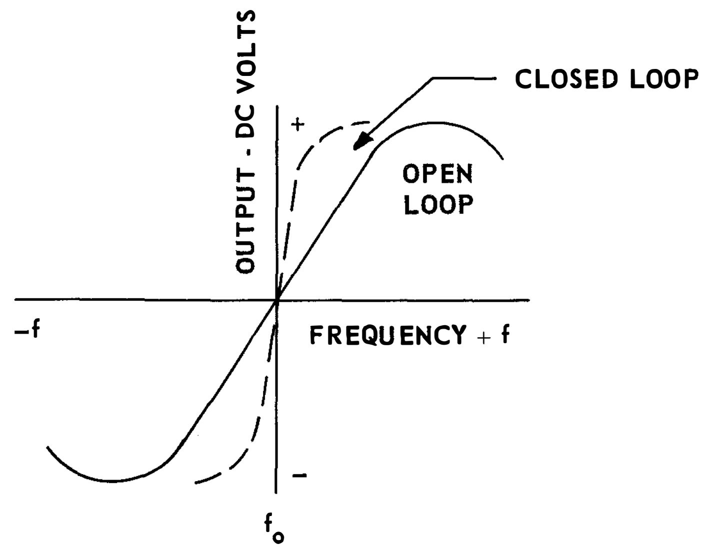


Figure 8-9. Discriminator Frequency Response Curve.

versus oscillator frequency on the horizontal scale. If the master oscillator frequency differs from the discriminator frequency when the loop is closed, the master oscillator frequency will be pulled toward the discriminator frequency, provided the proper polarity of discriminator and control device has been observed. It is important to realize that perfect correction cannot be achieved unless there is an infinite amount of amplification of the error signal from the discriminator. This can be seen by examining the discriminator output when the loop is closed. If perfect correction had somehow been achieved, the discriminator output would be zero. Obviously, such cannot be the case, as there must be some signal applied to the reactance control to correct the oscillator frequency error. The closed-loop frequency error will depend on the master oscillator error and on the gain of the control loop. The performance of the closed-loop is shown by the dashed curve in figure 8-9.

8-26. In the example shown, the oscillator frequency is effectively compared with the discriminator frequency. There are two fundamental defects in this stabilizing system: there is a residual frequency error, and the stability obtainable from the discriminator is limited. The overall accuracy of a system using this principle in the HF band is approximately 50 parts per million, insufficient for SSB service. Both of these shortcomings can be eliminated by utilizing phase deviation rather than frequency deviation error signals in the control loop. To do this, the frequency discriminator is replaced by a phase discriminator with a reference signal derived from a standard oscillator (see figure 8-10). The SMO is then locked in frequency synchronization with the reference oscillator, and the error signal is the phase angle between the two oscillator voltages. As the frequency of the controlled oscillator drifts away from the reference

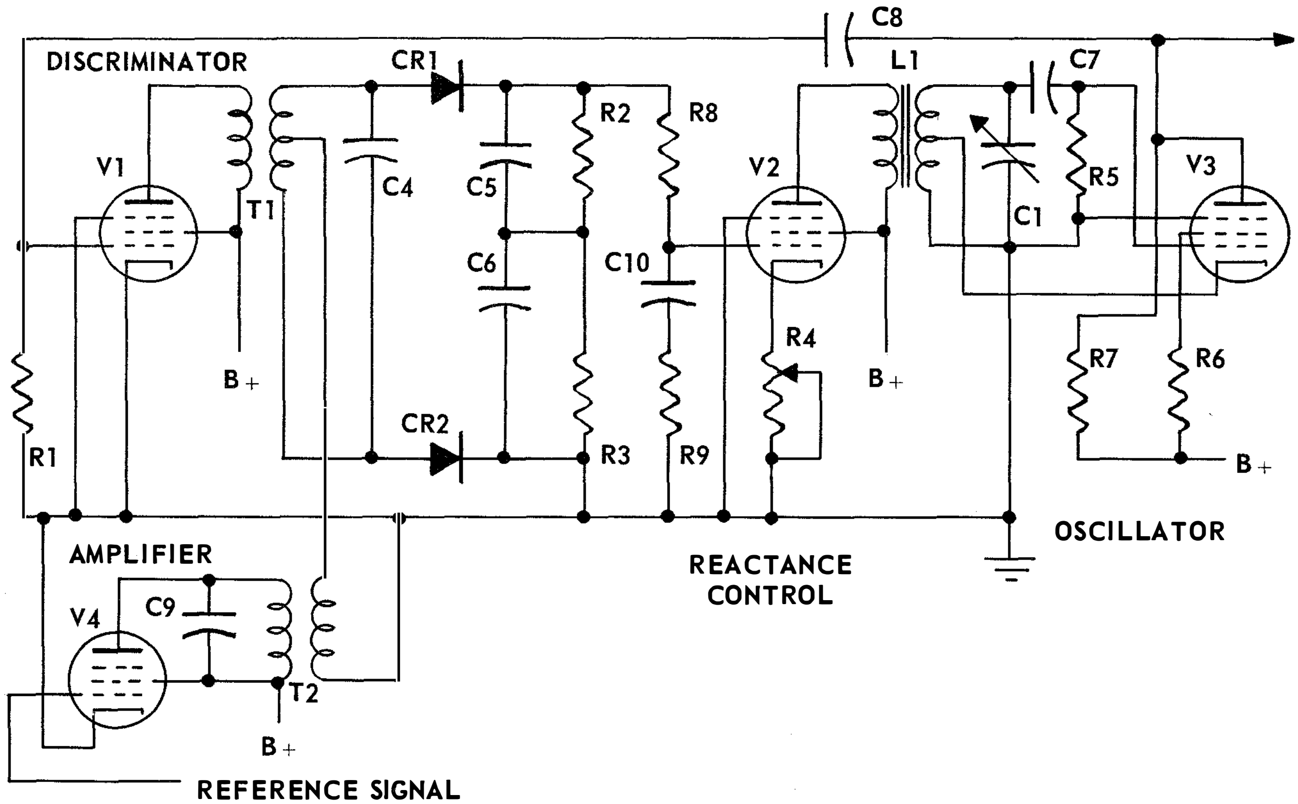


Figure 8-10. Simple Stabilized Master Oscillator With Phase Discriminator.

frequency, the phase angle between the two voltages increases to provide the correcting voltage necessary to operate the reactance control. The stability of the system now is completely dependent on the stability of the reference oscillator. The stabilization loop is a feedback system and, as a result, careful attention must be paid to gain and phase shift if stable operation is to be obtained. If the gain is unity at the frequency at which phase shift around the loop is 180°, oscillation will result. By reducing the gain at the critical frequency, the low-pass filter network in the grid of the reactance control tube provides the necessary control of gain to avoid oscillation.

8-27. Multiple Frequency SMO Operation.

8-28. Although the SMO described is capable of operating on one frequency only, the circuit can be extended to operate on additional frequencies. To accomplish this, a double-mixer frequency translation system is designed to feed the discriminator (see figure 8-11). By this means, the reference frequency is translated upward in frequency to the range 16 to 32 Mc. As long as oscillator No. 2 is fixed in frequency, the reference signal can be translated to frequencies separated by 100-Kc increments between 16.0 and 32.0 Mc: 16.0, 16.1, 16.2, etc. If the frequency of oscillator No. 1 is fixed and oscillator No. 2 is varied, the reference signal will be translated in steps

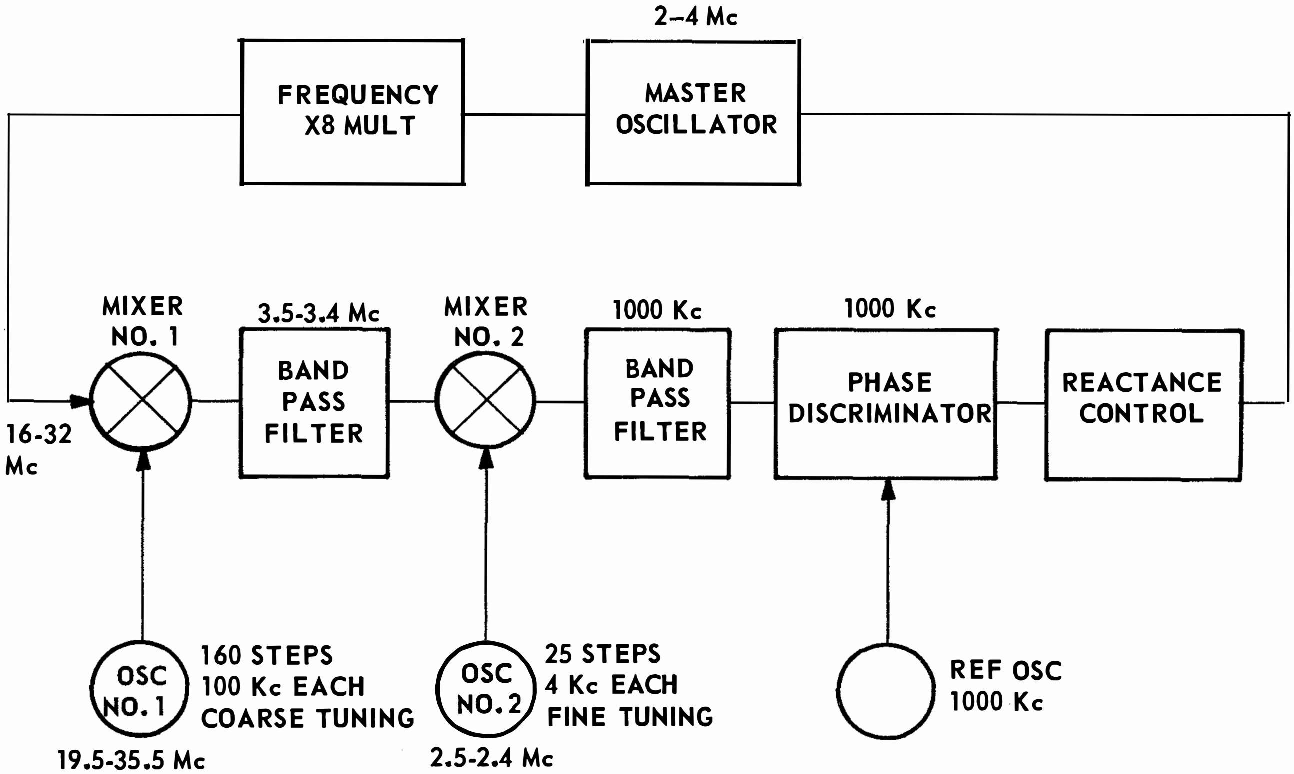


Figure 8-11. Block Diagram of Basic Stabilized Master Oscillator.

of 4 Kc over an interval of 96 Kc (16,000, 16,004, 16,008, etc.). In this way, the reference frequency can be translated to any one of 4000 frequencies between 16 and 32 Mc, spaced 4 Kc apart.

8-29. The master oscillator operates from 2 to 4 Mc, this range being best suited to covering the HF band, using both fundamental and harmonics. For synchronization with the reference, the master oscillator output frequency is multiplied by eight so that the master oscillator signal frequency range corresponds to the reference frequency range. Under these conditions, the master oscillator fundamental output frequency will be stabilized on 1/2-Kc intervals over the range 2 to 4 Mc. More channels can be synthesized by adding another mixer stage or by increasing the number of steps used at each mixer. The accuracy of the stabilization obtained by this system depends on the accuracy of the frequencies used at the translating mixers. To obtain the greatest accuracy, all of these frequencies are derived from a single source: a standard reference oscillator of extremely high accuracy and great stability. The use of a phase error signal in the control loop ensures that the residual error of the stabilized oscillator will be measured in terms of degrees of phase angle between controlled and reference oscillators rather than in terms of cycles of frequency difference where only a frequency discriminator is used.

8-30. SMO Tuning.

8-31. A better understanding of the SMO requirements calls for an explanation of the frequency content in the four output frequency bands of the SSB system. The four bands are as follows:

Band A. ---1.7 to 3.699 Mc, in 1-Kc steps, for a total of 2000 channels.
Band B. ---3.7 to 7.699 Mc, in 1-Kc steps, for a total of 4000 channels.
Band C. ---7.7 to 15.699 Mc, in 1-Kc steps, for a total of 8000 channels.
Band D. ---15.7 to 31.7 Mc, in 1-Kc steps, for a total of 16,000 channels.

Here you can see that the master oscillator is required to supply a different number of frequency increments for each band. The master oscillator may be positioned mechanically to 4000 increments of 0.5 Kc (refer to figure 8-12). This condition permits injections for output bands A and B to be supplied directly by use of mechanical tuning only. The stabilizing loop then serves only to correct any master oscillator error that may occur due to frequency drift. Output bands C and D require a larger number of increments than the master oscillator can position mechanically. The stabilizing loop serves to supply the required additional increments by deriving an error signal. The error signal is converted to information that will cause electronic positioning of a master oscillator to the additional increments and will maintain correction for the master oscillator drift error. Positioning of the master oscillator is determined automatically when a frequency selection is made at the control unit.

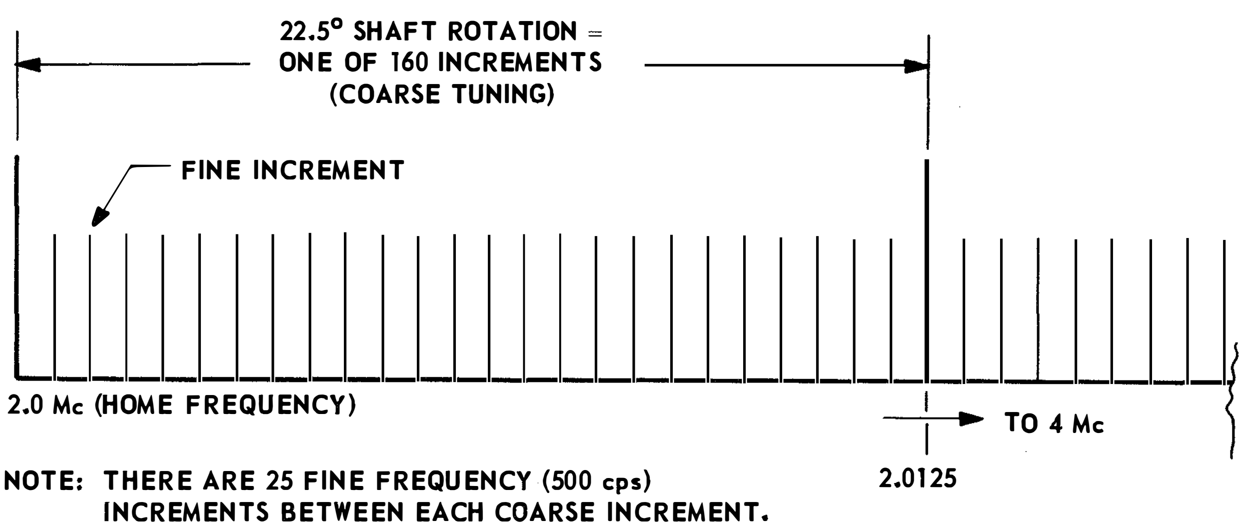


Figure 8-12. Graphic Presentation of SMO Tuning.

8-32. Figures 8-13 and 8-14 are block diagrams of a typical SMO, containing example frequencies to explain how the master oscillator may be positioned to increments smaller than mechanical positioning provides. In addition, figure 8-13 illustrates the drift cancelling action of the stabilizing loop. The operating frequency of the master

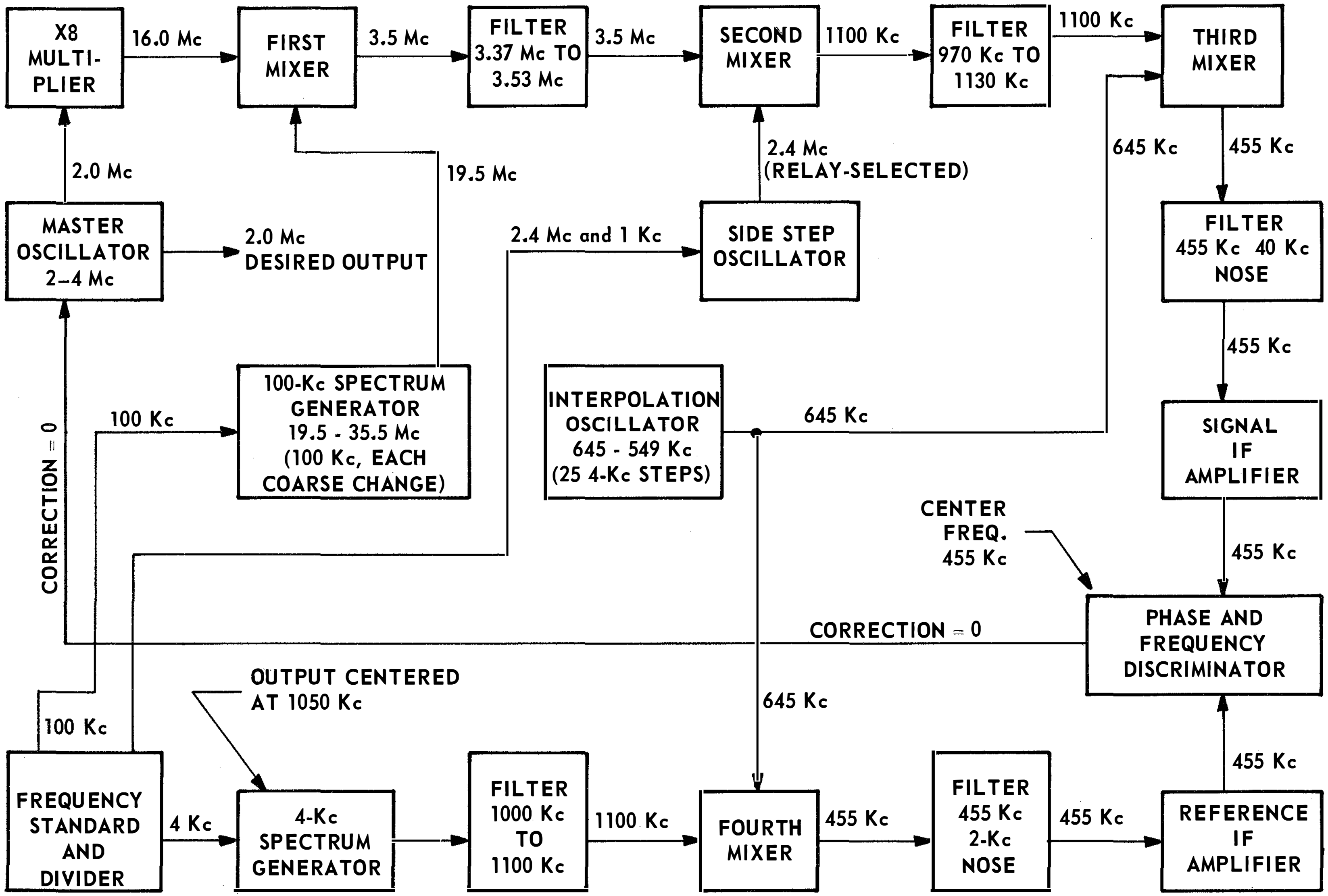


Figure 8-13. Case 1, Frequency Generation, Stabilized Master Oscillator.

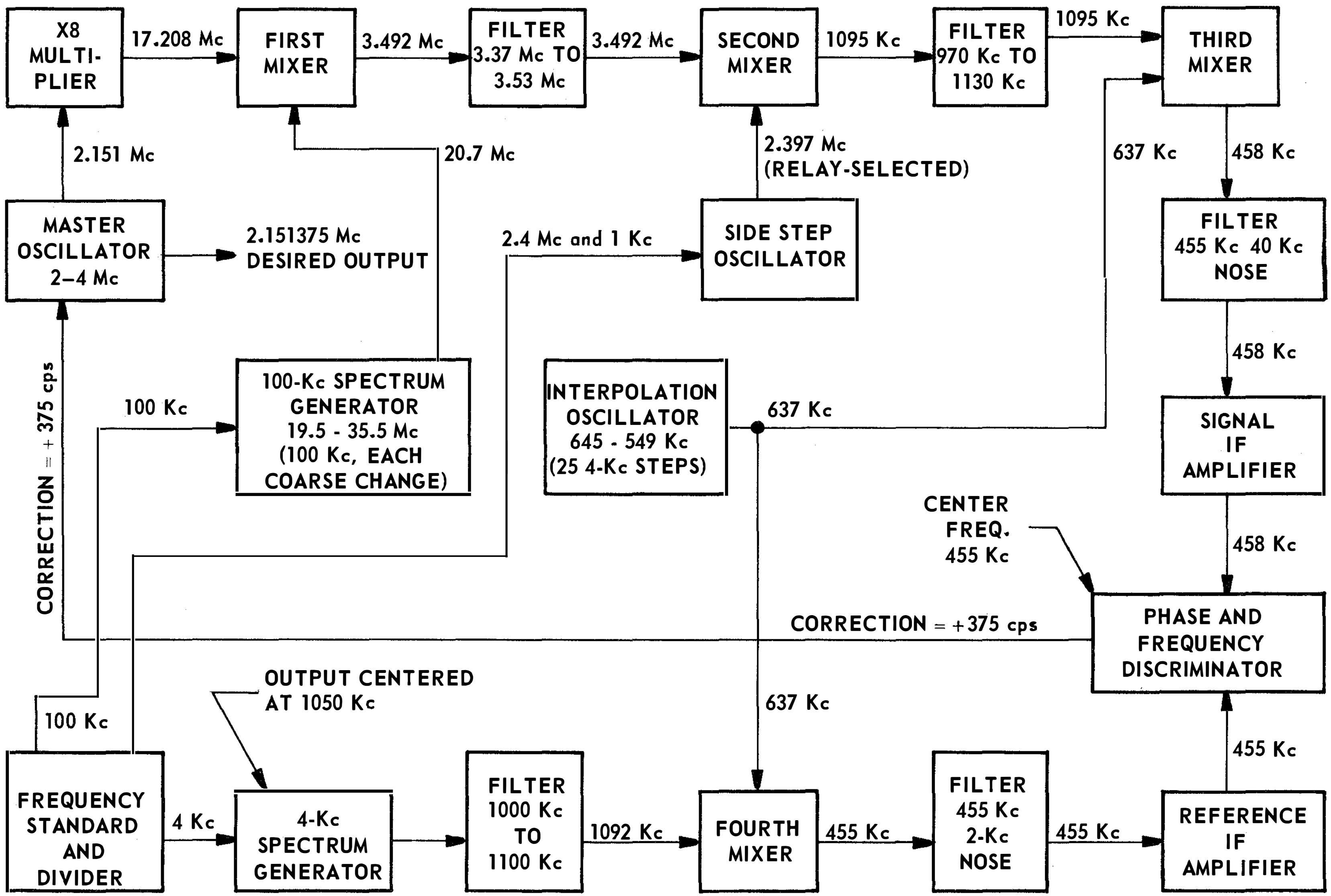


Figure 8-14. Case 2, Frequency Generation, Stabilized Master Oscillator.

oscillator is determined by a tuned circuit located in the control grid circuit of the oscillator. The main tuning inductance of the tuned circuit is a permeability-tuned coil, which has its variable core geared to a servomotor. This system is capable of mechanically positioning the oscillator to any one of 4000 frequency increments spaced at 0.5-Kc intervals.

8-33. Stabilization of the SMO.

8-34. Stabilization of the master oscillator takes place at a low frequency; therefore, a frequency conversion system is set up to translate the master oscillator output to the frequency range of the error-detecting frequency and phase discriminator. This is accomplished by a multiplier stage which operates on the eighth harmonic of the master oscillator frequency. The first mixer of this system is provided with an injection signal from a circuit that is operating on multiples of 100 Kc. The second and third mixer stages are furnished with injection signals from circuits operating in increments of 1 Kc and 4 Kc, respectively. These injections are necessary to cause the stabilizing loop tuning elements to follow the mechanical positioning so that completion of mechanical positioning will leave the circuits of the stabilizing loop at a corresponding frequency which permits electronic insertion of any additional increments required. The control unit must cause setup of the proper injection at each mixer and must set the master oscillator mechanical tuning control to the proper frequency for the conversion system.

8-35. The 2.0- to 4.0-Mc output signal originates in the 2.0- to 4.0-Mc master oscillator. The dial frequency selected at the control unit determines initial positioning of the master oscillator and consequent selection of the injection frequencies necessary to cause the frequency correction and positioning circuitry to operate. The master oscillator output is multiplied by eight, and the multiplied frequency is applied to the first mixer where it is mixed with a 100-Kc spectrum signal in the range of 19.5 to 35.5 Mc. The difference output signal of the second mixer is passed through a filter, which has a passband of 970 to 1130 Kc, and is applied to a third mixer stage. In the third mixer, the signal is mixed with an injection signal in the frequency range of 549 to 645 Kc supplied by an interpolation oscillator. The difference frequency is passed through a 455-Kc filter having a 40-Kc bandpass and applied to a phase and frequency discriminator. If this signal is other than 455 Kc, the phase and frequency discriminator detects a correction or positioning error and converts it to the information that is fed to the master oscillator. The information derived in the phase and frequency discriminator is used to control the saturation condition of a saturable reactor, located in the tuning circuit of the oscillator, with consequent control of the output frequency. To provide the phase and frequency discriminator with a reference, the 4-Kc signal supplied by the frequency divider is fed to a 4-Kc spectrum generator. The output of the spectrum generator is passed through a 1000- to 1100-Kc bandpass filter and applied to the fourth mixer stage. The fourth mixer is supplied with the same interpolation oscillator frequency that was used at the third mixer. The resultant output of the fourth mixer is passed through a 455-Kc filter having a 2-Kc passband and applied to the phase and frequency discriminator for use as the reference.

8-36. Frequency Generation.

8-37. To illustrate the frequency generation scheme, two typical cases of frequencies have been selected. Block diagrams and descriptive text are provided for each case.

8-38. Case 1. (There is no frequency correction to the master oscillator.) See figure 8-13. The 2.0-Mc master oscillator frequency is chosen to simplify our explanation of the home frequencies of the stabilizing loop circuits. No drift error is indicated from the master oscillator, and the mechanical positioning circuits have placed the master oscillator at 2.0 Mc. The 2.0-Mc output is fed to a times-eight multiplier circuit to translate any master oscillator error to the range of a phase-and-frequency discriminator circuit. Therefore, the multiplier output is 16.0 Mc, fed to a first mixer stage. The injection signal for the first mixer is obtained from a 100-Kc spectrum generator geared to the 160-step coarse system. The spectrum generator operates on any of 160 steps, at 100-Kc intervals, in the range of 19.5 to 35.5 Mc, with 19.5 Mc as its home frequency. This circuit serves to translate coarse mechanical positioning into stabilizing loop frequency. At a master oscillator frequency of 2.0 Mc, the spectrum generator is at its home frequency of 19.5 Mc, which is injected into the first mixer.

8-39. Beating of the 19.5 Mc spectrum generator frequency and the multiplied, 16-Mc, master oscillator frequency produces a difference frequency of 3.5 Mc, which is passed through a 3.37 to 3.53-Mc filter and applied to a second mixer stage. The injection signal for the second mixer is provided by the sidestep oscillator. Home frequency for the amplifier-oscillator is 2.4 Mc. The sidestep oscillator output is used for injection of a frequency that causes supplementing of the basic 0.5-Kc mechanical positioning of the master oscillator to achieve electronically the 0.125- and 0.250-Kc increments necessary for output bands C and D. In this example, no additional increments are required, since the master oscillator has been set mechanically to the 2.0-Mc output frequency. Therefore, the sidestep oscillator is at its home frequency of 2.4 Mc, which is injected in the second mixer.

8-40. Heterodyning between the 2.4-Mc signal and the 3.5-Mc signal output of the first mixer filter produces a difference frequency of 1100 Kc, which is passed through a 970- to 1130-Kc bandpass filter and applied to a third mixer. The injection frequency for the third mixer is supplied by an interpolation oscillator which is geared directly to the fine mechanical positioning elements of the master oscillator. The interpolation oscillator operates over a 25-step range, in 4-Kc increments, between the frequencies of 645 and 549 Kc, thus translating the fine positioning frequency to stabilizing loop frequency. Home frequency for the interpolation oscillator is 645 Kc, which is used for this example. The 645-Kc signal is injected into both the third and fourth mixers.

8-41. Beating between the 645-Kc interpolation oscillator frequency and the 1100-Kc frequency from the second mixer filter produces an IF signal of 455 Kc, which is applied to a phase-and-frequency discriminator circuit. Center frequency of the discriminator is 455 Kc; therefore, no error is detected, and no correction is applied to the master oscillator. A reference is required by the discriminator and is derived from the reference signal channel. The 4-Kc input to the spectrum generator from the frequency standard is applied to a 4-Kc spectrum generator. The spectrum is centered at 1050 Kc. The purpose of this spectrum is to provide an injection for the fourth mixer that permits use of the interpolation oscillator frequency in deriving a 455-Kc reference signal that is passed through a 455-Kc mechanical filter, amplified by a reference IF amplifier, and applied to the discriminator. The interpolation oscillator can be in error ±0.5 Kc at its worst; but the method of injecting its output into both the signal and reference IF channels maintains any drift by the same amount in each channel. Thus, no phase difference is present in the discriminator due to interpolation oscillator drift.

8-42. Case 2. (A frequency correction is required for the master oscillator.) See figure 8-14. The dial frequency, or frequency selected through the use of the control unit, is 16.911 Mc. The master oscillator frequency necessary to produce this output is determined as follows:

$$\text{MASTER-OSCILLATOR FREQUENCY} = \frac{F_O + 300\text{ Kc}}{\text{BAND MULTIPLIER}}$$

Where

F_O = Frequency Selected at Control Unit

300 Kc = IF frequency subtracted in the low-frequency mixer (refer to figure 3-1 of the section on SSB Exciters).

BAND MULTIPLIER = RF Tuner low-frequency mixer injection (one) plus band multiplication of the master oscillator frequency for the high-frequency mixer.

Band A = 1 + 0 = 1

Band B = 1 + 1 = 2

Band C = 1 + 3 = 4

Band D = 1 + 7 = 8

In this case $M_O = \dfrac{16.911\text{ Mc} + 300\text{ Kc}}{8} = 2.151375\text{ Mc.}$

8-43. The master oscillator servo-loop follow-up system provides mechanical tuning of the master oscillator. Since mechanical tuning of the master oscillator is limited to 0.5-Kc steps of the master oscillator frequency, the additional master oscillator frequency increments must be achieved by another method. This is accomplished by operation of the stabilizing loop which derives a fine tuning error which operates the electronic positioning circuit. In this example, the master oscillator frequency required is 2.151375 Mc. The master oscillator can be tuned mechanically to 2.151 Mc only by the coarse and fine servomechanism, since this is a multiple of the 0.5-Kc capability. This 2.151-Mc frequency is multiplied by eight in a multiplier stage to place the signal within the operating range of the phase-and-frequency discriminator. The resultant 17.208-Mc output of the multiplier stage is applied to the first mixer, where it is beat with a signal from the 100-Kc spectrum generator in the frequency range of 19.5 to 35.5 Mc. The spectrum generator frequency can be determined with the following formula:

$F_{sg} = 8(F_{mo} - 2) + 19.5$ (drop all digits beyond 0.1 Mc digit. The remainder is used to determine the interpolation oscillator frequency.)

Where

F_{sg} = spectrum generator frequency and 8 is a constant.
$F_{mo} - 2$ = frequency output of the master oscillator minus 2 Mc (home frequency of MO).

The interpolation oscillator frequency can be determined as follows. If there are digits beyond the 0.1-Mc digit, the next two digits equal the frequency in Kc which must be subtracted from the interpolation oscillator home frequency (645 Kc). This represents the fine mechanical tuning. Any digits beyond this are compensated for by the electrical stabilizing loop. In this example, the 100-Kc spectrum generator frequency is 20.7 Mc. This signal is applied to the first mixer and beat with the multiplier master oscillator frequency of 17.208 Mc to produce a usable output of 3.492 Mc. This signal is passed through the 3.37- to 3.53-Mc bandpass filter and applied to the second mixer. The injection signal for the second mixer is a signal from the sidestep oscillator. This signal is one of the four relay-selected frequencies previously discussed, consisting of 2.4, 2.399, 2.398, or 2.397 Mc. These frequencies are necessary to cause electronic positioning of the master oscillator to the 250-cps increments necessary in output band C and 125-cps increments necessary in output band D. The output frequency of the amplifier mixer is selected automatically when a frequency is selected at the control unit. The selected output frequency of the sidestep oscillator can be determined by examining the final required output frequency of the master oscillator; 500 should be subtracted from the "cycles" portion of the master oscillator frequency, if possible. This subtraction accounts for the 0.5-Kc mechanical positioning capability of the master oscillator. The following information is then derived from the remaining "cycle" portion of the master oscillator frequency:

REMAINING "CYCLE" PORTION	SIDESTEP-OSCILLATOR OUTPUT FREQUENCY
000	2.400 Mc
125	2.399 Mc
250	2.398 Mc
375	2.397 Mc

The sidestep oscillator output, for the example being discussed, is 2.397 Mc, since the "cycles" portion of the master oscillator frequency is 375. The 2.397-Mc signal is applied to the second mixer and beat with the 3.492-Mc signal derived from the first mixer stage and filter. The difference frequency of 1095 Kc is passed through a 970- to 1130-Kc bandpass filter and applied to the third mixer, where it is mixed with a signal from the interpolation oscillator. The interpolation oscillator operates in the frequency range of 645 to 549 Kc and is positioned between those frequencies in 4-Kc steps. The frequency of the interpolation oscillator is determined by the servo loop circuitry which controls the mechanical tuning of the master oscillator. The 25-step, fine tuning of the master oscillator is provided by the fine follow-up potentiometer located in the interpolation oscillator. The 160-step, coarse tuning occurs at increments of 12.5 Kc of master oscillator frequency, while the 25-step fine tuning occurs at increments of 0.5 Kc of master oscillator frequency. The 25-step system completes its range for each coarse increment of the 160-step system, thus providing a total of 4000 mechanically positioned increments at intervals of 0.5 Kc at master oscillator frequency. The interpolation oscillator is positioned directly with the 25-step system, in 4-Kc steps, from 645 to 549 Kc, with 645 Kc as home position. To determine the interpolation oscillator frequency, examination of the final tuned master oscillator frequency is necessary which, in this example, is 2.151375 Mc.

8-44. Referring to figure 8-12, it can be seen that each coarse increment of master oscillator tuning contains the twenty-five 0.5-Kc steps inserted by the fine positioning follow-up potentiometer in the interpolation oscillator. The 151 portion of the master oscillator frequency represents 12 coarse increments of 12.5 Kc plus 1 Kc away from 2.0 Mc. The 1-Kc increment represents two 0.5-Kc fine steps of master oscillator frequency past the last completed coarse increment. Since the fine follow-up potentiometer in the interpolation oscillator is responsible for the 0.5-Kc mechanical tuning of the master oscillator and since the fine follow-up potentiometer and interpolation oscillator are geared together directly, the 1-Kc increment represents two steps of interpolation oscillator frequency. The interpolation oscillator is positioned at 645 Kc at the start of each master oscillator coarse increment and is positioned in decreasing frequency steps for any change; therefore, since the interpolation oscillator operates in the frequency range of 645 to 549 Kc, in 4-Kc steps, its frequency appears 8.0 Kc below 645 Kc, at 637 Kc. This 637-Kc signal is beat with the 1095-Kc signal produced at the second mixer and filter. The difference frequency of 458 Kc is passed through the 455-Kc filter and the signal IF amplifier and applied to the phase-and-frequency discriminator.

8-45. In the fourth mixer, the interpolation oscillator frequency is combined with a signal from the 4-Kc spectrum generator. The spectrum generator signal consists of a 4-Kc spectrum centered at 1050 Kc. This permits the interpolation oscillator signal to beat with a spectrum point which will produce a difference output of 455 Kc. This signal is passed through a 455-Kc mechanical filter and applied to the phase-and-frequency discriminator for use as the reference. The phase-and-frequency discriminator derives a DC current in proportion to the error and applies this current to a saturable reactor located in the master oscillator tuning circuit. This causes the inductance of the AC winding of the saturable reactor to change in proportion to the error, changing the master oscillator operating frequency to correct the initial error present at the master oscillator. In this example, the master oscillator frequency is corrected by an additional 375 cycles to obtain the final, tuned master oscillator frequency of 2.151375 Mc. Operation of the stabilizing loop, as just described, occurs in identical fashion for any errors at the master oscillator because of oscillator frequency drift. This causes the master oscillator output to be as precise as the frequency standard.

SECTION IX

FREQUENCY STANDARDS

9-1. Introduction.

9-2. Because frequency is defined in terms of cycles per second or events per unit time, frequency control and timekeeping are inseparable. Any measurement of frequency can be only as accurate as the time unit used. Thus, in order to determine the accuracy of a frequency standard, its period of oscillation must be compared to a time standard of known accuracy. The best secondary time standard consists of a frequency standard which drives a cycle counter or clock. The accuracy of this time standard can then be determined by comparing it with the primary time standard, the mean solar day (the time required for the earth to complete one revolution about its axis).

9-3. Time measurement has always been based on astronomical phenomena. Days and years are determined by the relative motion of the earth with respect to the sun. However, to coordinate events and to make precise measurements of physical phenomena, a device which can divide the day into accurate, shorter intervals is required. The search for accurate timing devices started in prehistoric times. It followed two separate lines: first, those devices which derive time directly from astronomical observations; second, independent mechanisms and devices for measuring time intervals. The first type started with the casual observation of the position of the sun, progressed to the sundial, and culminated in the modern Zenith tube. The second type started with devices based on restricted flow. The first of these were the noncycling types, such as sand clocks, water clocks, time candles, and time lamps (typically, sand clocks or hourglasses had inaccuracies of 4,000 seconds per day). These were followed by automatic recycling types using escapement mechanisms controlled by friction and inertia, such as the Verge and Foliot balance. Clocks of this type varied 1,000 seconds per day. With the discovery of resonance phenomena, that is, an oscillating system in which energy is alternately stored in the form of kinetic and potential energy, much more accurate time measurements were made possible. The first device using resonance phenomena to measure time was the pendulum clock which ultimately attained an accuracy of 0.002 second per day. The pendulum was followed by the hairspring and balance which attained an accuracy of 0.2 second per day, and the electrically activated tuning fork which attained an accuracy of 0.008 second per day. The quartz crystal, which followed the tuning fork as the resonator in time and frequency standards, has an accuracy of 1 part in 10^9 per day or 0.0001 second per day and is the most widely used control element in modern time and frequency standards. Due to variations in the rotation of the earth, the short-term accuracy of the quartz crystal is better than that of the mean solar day. Seasonal variations of several milliseconds and yearly variations as great as 1.6 seconds in the mean solar day have been observed. On a long-term basis, however, the length of the mean solar day is increasing at the rate of only 0.00164 second per century. In order to achieve long-term accuracy, the standard must remain in constant operation; and since mechanical devices such as the quartz crystal clock will run for only a few years, they will probably not replace astronomical phenomena as a primary time standard. The most recent development is the use of devices based on atomic or molecular resonance. These have attained short-term accuracy equal to that of the quartz crystal, but their long-term accuracy is expected to be considerably better.

9-4. Technical and economic forces have led to the development of more and more accurate frequency control circuits. In the early days of radio, the tunable LC oscillator provided a simple and serviceable answer to the problem of generating channel frequencies. The lower frequency end of the spectrum and amplitude modulation were used, and the spectrum was not unduly crowded. Later crowding of the spectrum led to closer channel spacing and expansion into the higher frequency regions. This in turn required more accurate frequency control. Crystal oscillators provided the required accuracy, but many crystals were required to provide the required number of channels. During World War II, it became almost impossible to deliver the right crystal to the right place at the right time. After the war, users of communication equipment demanded a choice of hundreds of channels at the flick of a switch. In order to meet the demand for spectrum space, guard bands were narrowed, and the VHF bands were put to more extensive use. All of these forces led to the development of the multiple-crystal frequency synthesizer (figure 9-1) in which the output frequencies of several crystal oscillators were mixed together to produce the desired output frequencies, providing many more channels than the number of crystals used.

9-5. Present crowding of the spectrum and increasing demand for communication channels now indicate that some method of further decreasing the spectrum space required for each channel must be found. Single sideband is a solution to this problem. The use of single sideband, however, requires that channel frequencies be maintained within ±1/2 part per million. Maintaining all of the crystals in a multiple-crystal synthesizer to the required accuracy is impractical; therefore, all the stability requirements must be concentrated in one or more highly stable oscillators. The solution to this problem is the single-crystal frequency synthesizer. Figure 9-2 is a block diagram of a typical single-crystal frequency synthesizer. Basically it is a circuit in which harmonics and subharmonics of a single standard oscillator are combined to provide a number of output signals which are all harmonically related to a subharmonic of the standard oscillator. In this system the accuracy and stability of the output signals are equal to that of the standard oscillator and as techniques improve, the stability of the synthesizer can be improved by replacing only the standard oscillator.

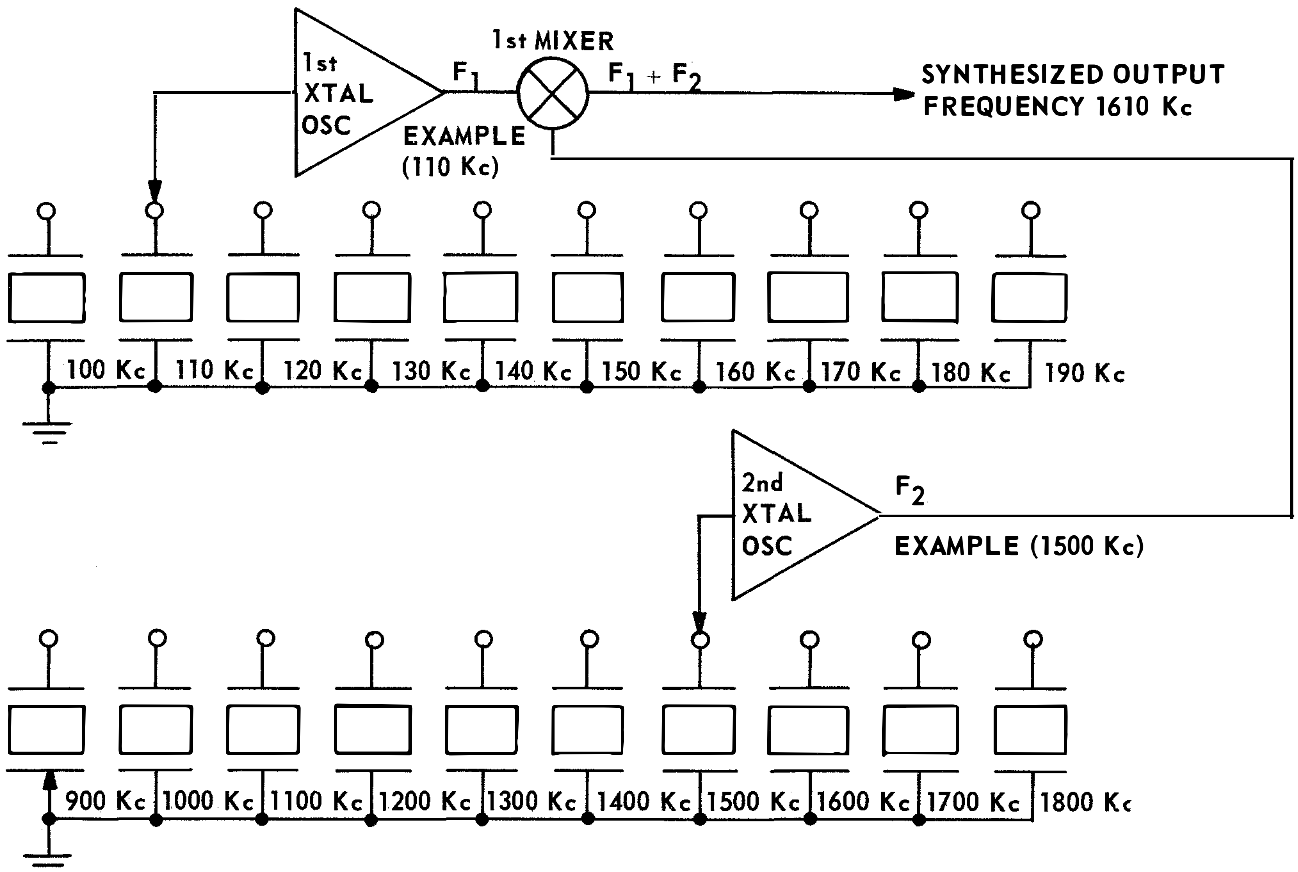


Figure 9-1. Multiple-Crystal Frequency Synthesizer.

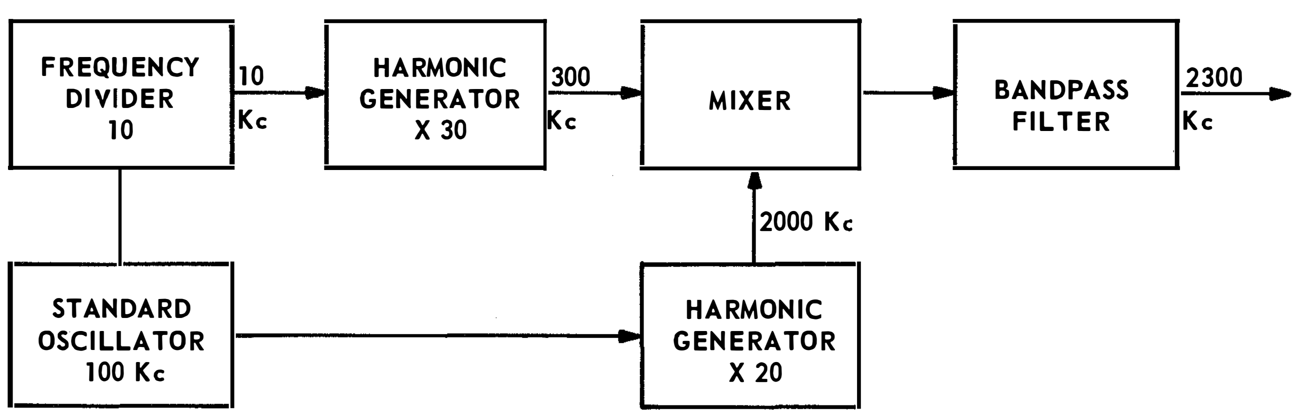


Figure 9-2. Single-Crystal Frequency Synthesizer.

9-6. The best laboratory standards now available have aging rates of approximately 1 part in 10^9 per month and short-term variations of several parts in 10^{11}. Operational standards have an instability several orders of magnitude greater than this. Typical examples are shown in the curves of figures 9-3 and 9-4. In figure 9-3, the dots represent errors derived from direct time comparison with WWV, and the crosses represent errors derived from time comparison with WWV after correction according to the WWV time correction bulletin.

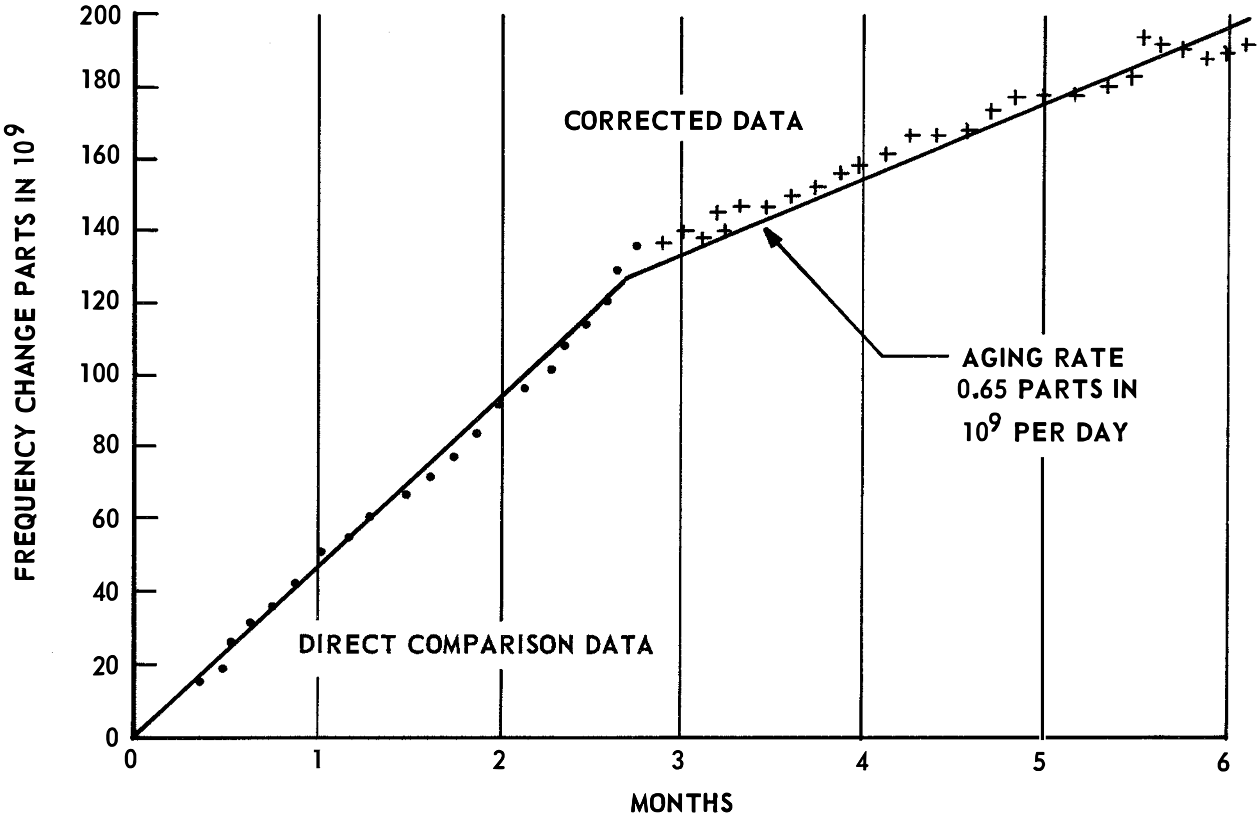


Figure 9-3. Typical Long-Term Stability.

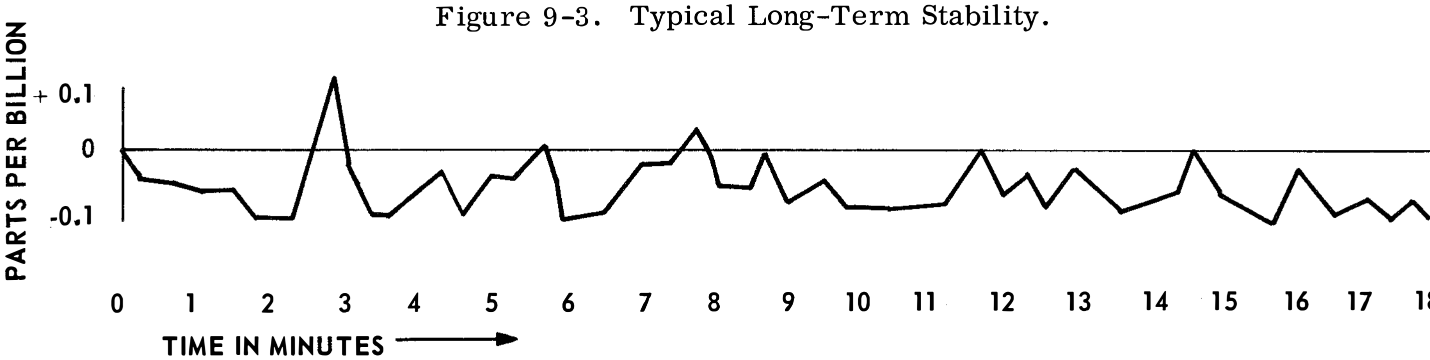


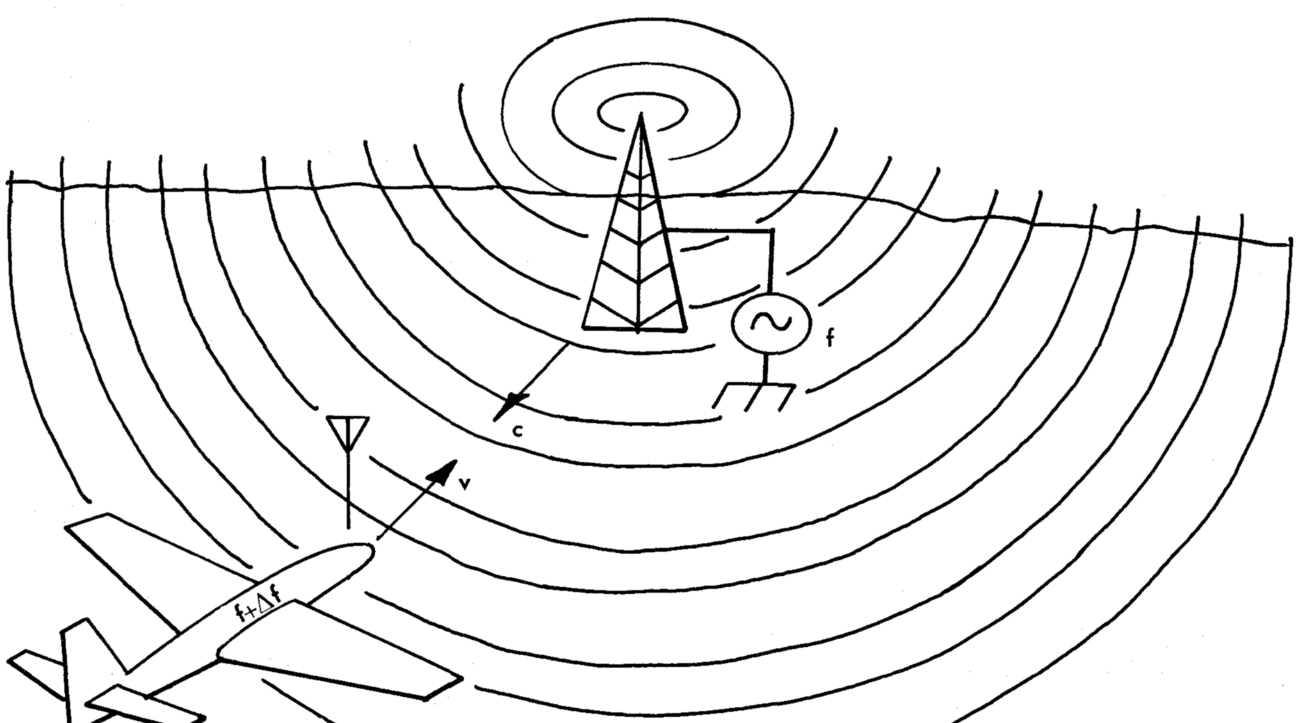
Figure 9-4. Typical Short-Term Stability.

9-7. Deterioration of Generated Frequency.

9-8. Although frequency standards in use today have accuracies of 1 part in 10^8 or better, serious errors can be introduced in the transmission and reception of the signal. These errors are caused by Doppler shift, shifts due to propagation characteristics, and shifts due to equipment circuitry.

9-9. Effect of Doppler Shift.

9-10. Relative motion between receiving and transmitting stations causes premature or delayed reception of individual cycles of the transmitted signal. Since the speed of propagation of radio signals is equal to the speed of light or 186,000 miles per second, the transmitted signal will be received 1 millisecond earlier or later for every 186 miles of change in transmission path length. A change in transmission path length at a rate of 670 miles per hour results in a frequency shift due to Doppler effect of 1 part in 10^6. Figure 9-5 shows an aircraft approaching a radio transmitter. In the formula shown in the figure, v = velocity of the aircraft and C = speed of light. If the aircraft



$$\frac{\Delta f}{f} = \frac{v}{c}$$

Figure 9-5. Doppler Frequency Shift in Aircraft Communications.

is approaching at a velocity of 670 miles per hour or 0.186 miles per second, then the ratio v/C = 0.186/186,000 or 1/1,000,000. Thus the ratio of frequency change to transmitted frequency ($\Delta f/f$) is $1/10^6$. If the transmitter is operating on a frequency of 10 Mc, then the frequency as received at the aircraft will be 10 Mc + 10 cps. If the transmitter were also in an aircraft flying toward the first aircraft at a velocity of 670 miles per hour, the frequency error would be doubled because the relative velocity would be the sum of the velocities of the two aircraft or 1,340 miles per hour. In ship-to-ship communication or in communication between ground vehicles, Doppler shifts of 1 part in 10^7 or greater are possible. Doppler shifts due to antenna sway caused by the pitch and roll of a ship are of the order of ±3 parts in 10^9. Extreme examples of Doppler shift are the case of back-pack radios in which the Doppler shift while the operator is walking is 5 parts in 10^9, and the case of the IGY satellite, where signals transmitted by the satellite suffered frequency shifts of up to 30 parts per million. Signal transit time in the case of a jet aircraft traveling 670 miles per hour and communicating with a fixed station changes at the rate of 3.5 milliseconds per hour, and in the case of battleships communicating with each other the signal transit time can change 0.2 milliseconds per hour.

9-11. Effect of Propagation Characteristics.

9-12. Low-frequency waves tend to follow the curvature of the earth, and the length of the transmission path is not seriously affected by atmospheric or ground conditions. Errors introduced by the propagation medium at low frequencies are only about ±3 parts in 10^9 in frequency and ±40 microseconds in transit time. In the high-frequency bands, however, reflections from the ionosphere are used for long-range communications. Frequency variations of ±2 parts in 10^7 and transit time variations of ±1 or 2 milliseconds can be introduced by changes in path length due to movement of the reflection point in the ionized layer and variations of the skip distance. Errors introduced in VHF and UHF scatter propagation are not well known, but available data indicate that they may be several parts in 10^8 in frequency and several hundred microseconds in transit time.

9-13. Effects of Equipment Circuitry.

9-14. Transmitter or receiver circuit elements, when subjected to mechanical vibration or temperature changes, can cause temporary frequency shifts by temporarily shifting the phase of the signal. Phase advancement of 360 degrees in one second adds 1 cycle per second to the frequency of a signal. Thus, a phase shift change of 1 degree per second imposed on a 100-kilocycle signal would cause a temporary frequency shift of 3 parts in 10^8. Mechanical vibration of tuning elements causes phase shifts which, even under laboratory conditions, may cause frequency shifts as great as 1 part in 10^8. Under operating conditions, severe mechanical vibration and temperature changes may be encountered which, if not compensated for, would cause excessive frequency errors.

9-15. Measurement Techniques.

9-16. Time Comparison.

9-17. Accurate comparisons of time and frequency using radio communications are difficult because of variations in propagating mediums. Present methods are based on time measurements taken over a long period so that these variations average out. By taking time measurements from WWV over a period of 20 days, accuracies of 1 part in 10^9 can be attained in the 2-Mc to 30-Mc bands. At 16 Kc, the same accuracy can be attained in approximately one day because the variations in the propagating medium have less effect on the low-frequency signal. Figures 9-6 and 9-7 are block diagrams of time comparison systems suitable for fixed station use.

9-18. On shipboard, errors introduced by changes in signal transit time due to relative motion between stations must be taken into account to achieve the same accuracies as are attained in fixed station use.

9-19. Figure 9-6 shows a system using aural indication of the synchronization of a clock, controlled by a local oscillator, with the time signals transmitted by WWV. The oscillator operates at 100 Kc. This frequency is divided by 100, and the resulting 1000-cps signal operates the synchronous clock. The clock operates a switch which closes once each second. A receiver tuned to the WWV signal is used to detect the time signals, in the form of clock ticks. These clock ticks consist of 5 cycles of a 1000-cps tone, transmitted at the rate of one tick per second. The ticks are coupled to a loudspeaker through the clock-operated switch which can be adjusted to close each time a tick is received. Once adjusted, the switch will continue to close in synchronism with the reception of the clock ticks as long as the frequency of the oscillator remains exactly 100 Kc. If the oscillator frequency changes, the speed of the clock will also change, and the switch closures will slowly drift out of synchronization. A calibrated dial is used to adjust the synchronization of the switch daily to permit only the last cycle of the clock tick to pass. The stability of the oscillator can be determined to an accuracy of 12 parts in 10^8 by calculations based on the amount of adjustment required in one day. The accuracy of measurement can be increased to 1.2 parts in 10^9 by basing the calculations on the amount of adjustment required in a period of 100 days.

9-20. Figure 9-7 is a block diagram of a chronoscope. The 100-Kc signal from the oscillator is divided to 10 cps and applied to the vertical and horizontal plates of the cathode ray tube through phase-shifting networks to produce a circular trace on the scope screen. A 1-Kc signal derived from the same oscillator is applied to the intensity control grid to break this solid circle into 100 dots. A receiver tuned to WWV supplies the clock ticks to the same control grid. Five additional dots are produced somewhere on the 100-dot circle, depending upon the relative phase of the 1-Kc signal from the oscillator and the 0.5 cycle of 1 Kc which comprise the clock tick. If the phase relationship remains constant, the 5-dot pattern on the screen will remain fixed; but if the phase changes, the pattern will move around the 100-dot circle at a rate

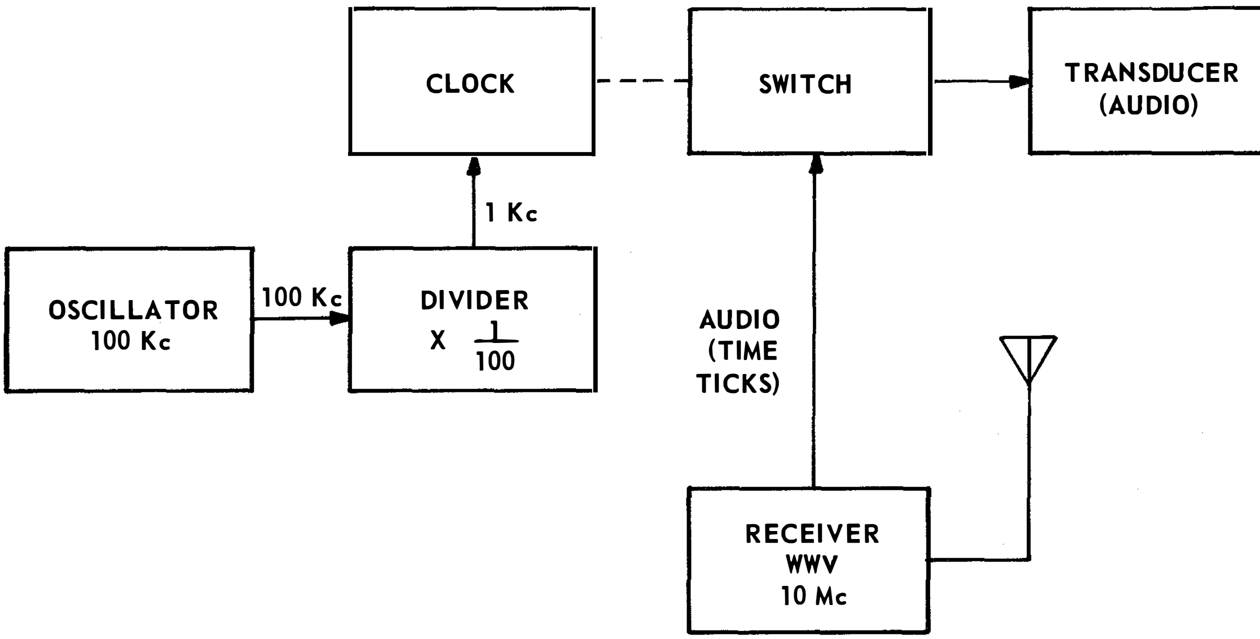


Figure 9-6. Time Comparison System, Aural Indication.

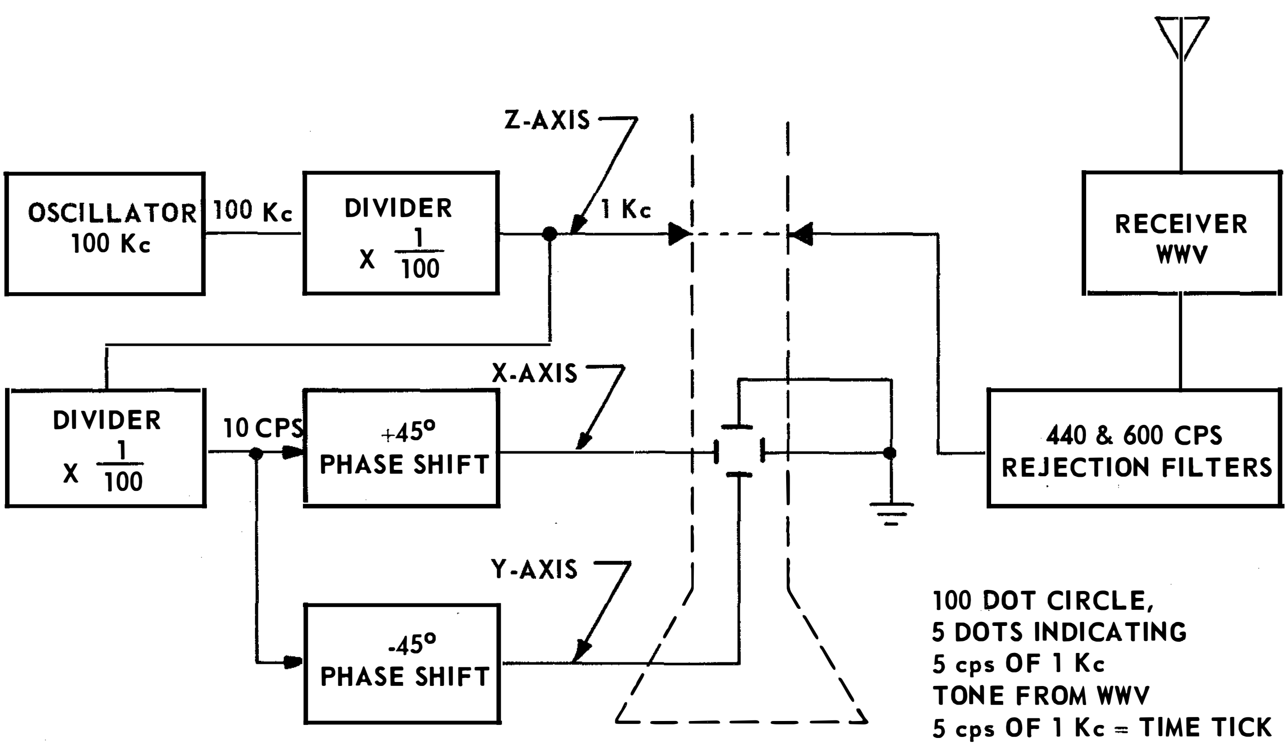


Figure 9-7. Time Comparison System, Visual Indication, Chronoscope.

determined by the rate of phase change. The rate of movement, in turn, indicates the magnitude of frequency error. With this system, the frequency of the oscillator can be determined to an accuracy of 1.2 parts in 10^8 in one day or 1.2 parts in 10^9 in 10 days.

9-21. Frequency Intercomparison.

9-22. Short-term stabilities of oscillators can be determined by intercomparison of the frequencies of two or more oscillators. When only two oscillators are compared, only the relative stabilities of the oscillators with respect to each other can be determined. Statistical data which will indicate the short-term stability of an individual oscillator can be obtained by intercomparing the frequencies of three or more oscillators, two at a time.

9-23. Figure 9-8 illustrates a system using two oscillators operating at frequencies differing by, nominally, 1 cps. One oscillator operates at 1 Mc and the other operates at 1 Mc plus 1 cps. Their outputs are mixed and the difference frequency, 1 cps, is used to control a gate circuit. A 100-Kc standard frequency is applied to the gate and the number of cycles of this standard frequency which are counted at the output indicates

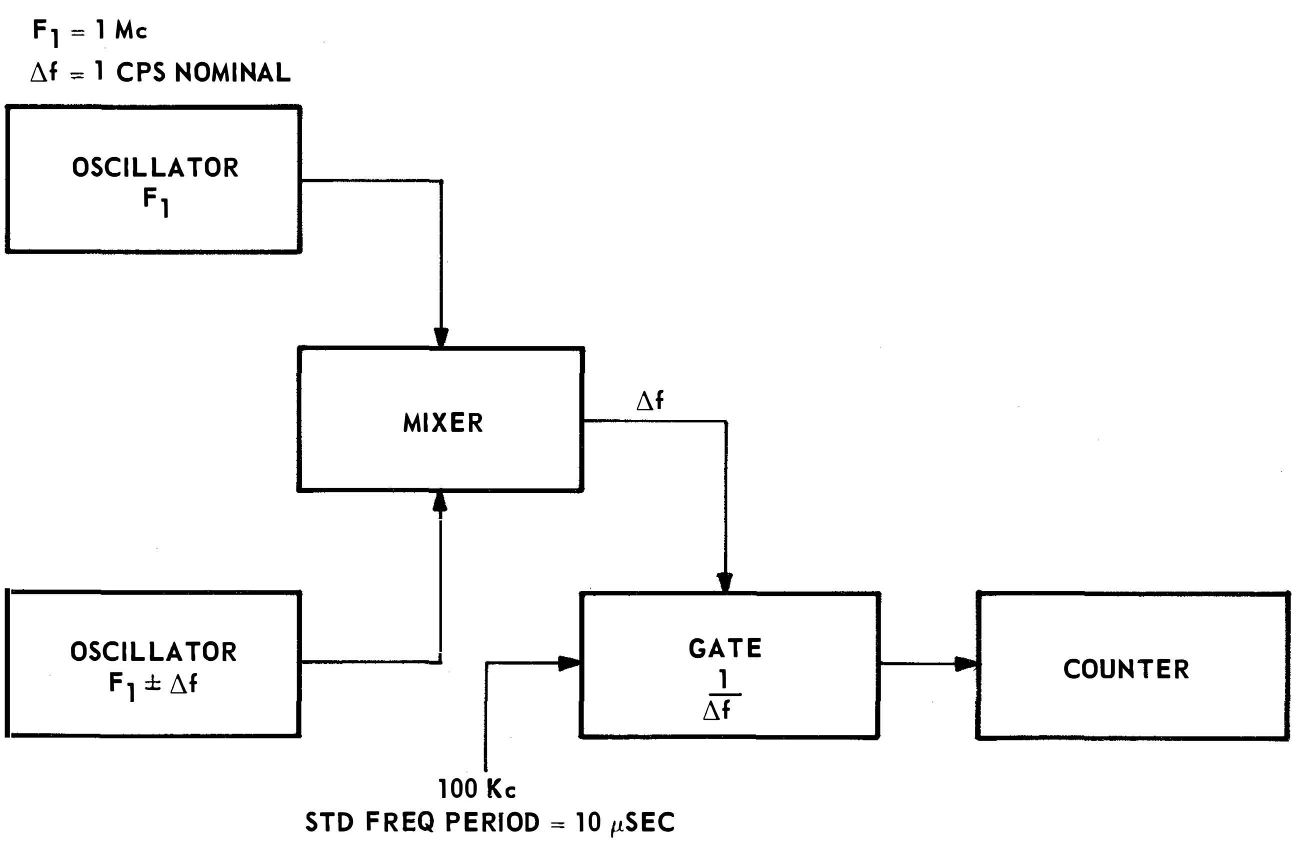


Figure 9-8. Frequency Intercomparison System Using a Frequency Counter.

the length of time the gate is open. This, in turn, is the period of the difference frequency controlling the gate. Thus, if the difference frequency is exactly 1 cycle per second, the gate will be open exactly 1 second and the counter will count exactly 100,000 cycles. If the difference frequency is more than 1 cycle per second, the counter will count less than 10^5 cycles; if the difference frequency is less than 1 cycle per second, the counter will count more than 10^5 cycles. This system will indicate the relative stability of one oscillator with respect to the other to 1 part in 10^{11}.

9-24. Figure 9-9 illustrates a system using two oscillators adjusted to operate at frequencies differing by 0.6 cps. The frequency of each oscillator is multiplied by 100 before mixing so that the resultant beat note is 60 cps. This beat note is recorded on a power line frequency recorder to give a continuous indication of relative stability between the two oscillators with an accuracy of 5 parts in 10^{10}.

9-25. Phase Intercomparison.

9-26. Figure 9-10 illustrates a system wherein the relative phase of two oscillators operating at the same frequency is measured. If the relative phase as indicated by the phase comparison meter changes 360 degrees in one second, the difference in frequency of one oscillator with respect to the other is 1 cps; or, if the oscillators are operating on a nominal frequency of 1 Mc, the difference is 1 part in 10^6. If the phase changes only 3.6 degrees in 10 seconds, the difference is only 1 part in 10^9. Thus, small differences in frequency can be measured and recorded. The resulting record will be an indication of the relative short-term stability of the oscillators.

9-27. Quartz Resonator Theory.

9-28. Construction and Operation.

9-29. Quartz resonators are electromechanical devices, having extremely high Q's and stable resonant frequencies, used as resonant circuits in electronic oscillators and filters. Quartz is a piezoelectric material; that is, mechanical deformation of the quartz causes an electric charge to appear on certain faces and, conversely, application of voltage across the quartz causes a mechanical deformation. The quartz resonator unit generally consists of a crystalline quartz bar or plate provided with electrodes and suitably mounted in a sealed holder. The mounting structure supports the bar or plate at nodal points in its vibrational pattern so that damping of the mechanical vibrations with resultant degradation of Q is minimized. The bar or plate is cut from the mother crystal at a carefully controlled angle with respect to the crystallographic axes and finished to close dimensional tolerances. The quartz bar or plate has a mechanical resonant frequency determined by its dimensions. This resonant frequency changes with temperature; but, by properly orienting the angle at which the blank is cut from the mother crystal, this temperature coefficient can be minimized. Commonly used orientations have been given designations, such as AT, CT, and DT cuts. Figure 9-11 illustrates the relationship between these cuts and the crystallographic axes. Proper

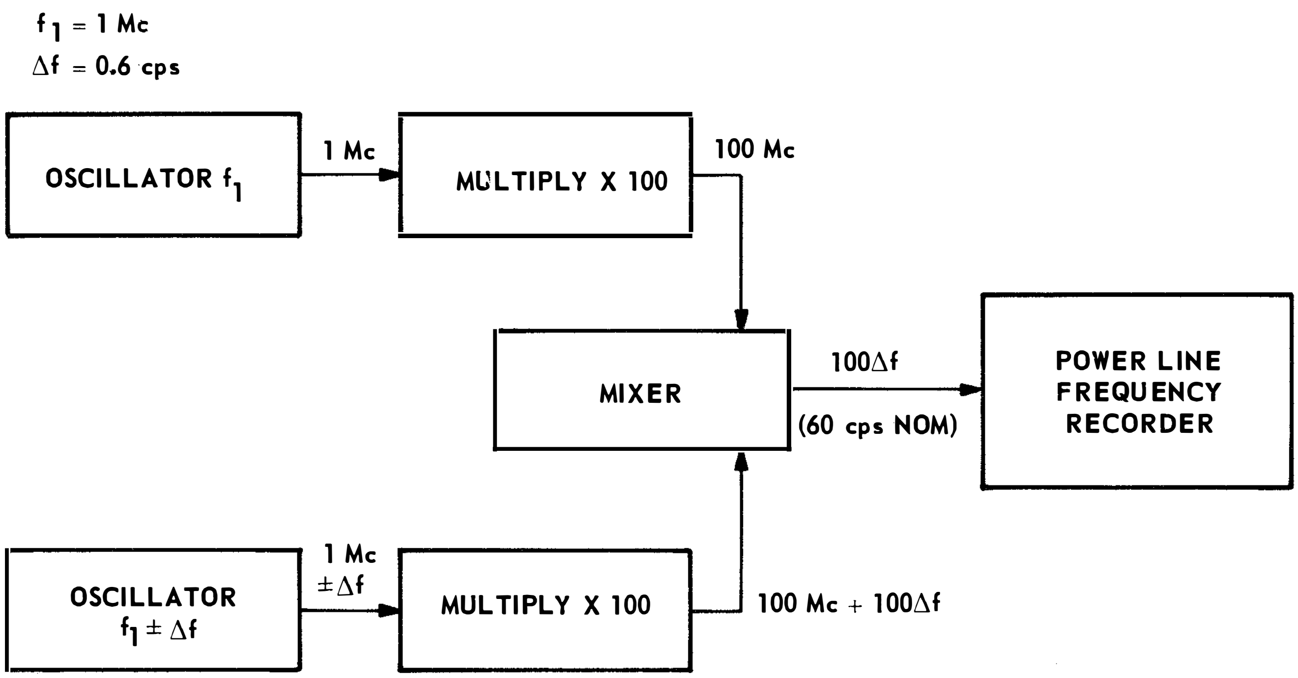


Figure 9-9. Frequency Intercomparison System Using Power Line Frequency Recorder.

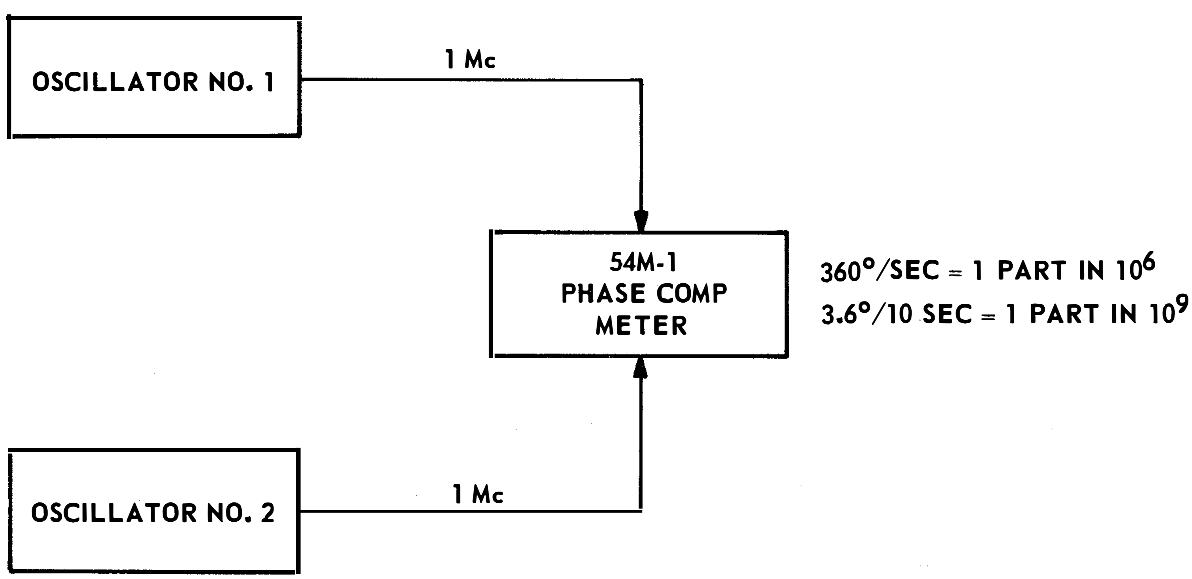


Figure 9-10. Phase Intercomparison System.

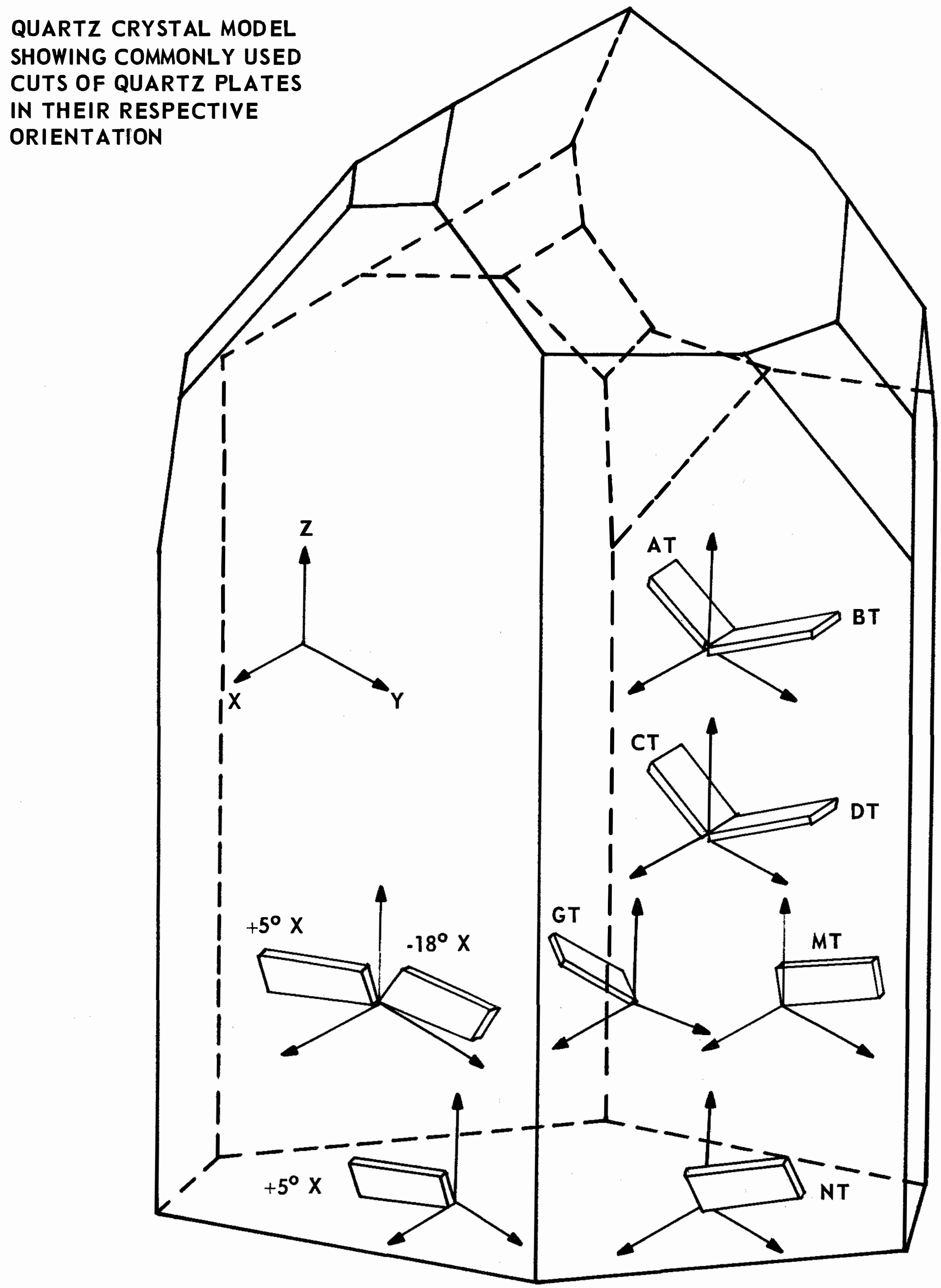


Figure 9-11. Quartz Crystal Showing Types of Cuts.

orientation of electrodes on the quartz plate provides electric coupling to its mechanical resonance. The electrodes usually consist of a metal plating which is deposited directly on the surface of the quartz plate. Connections from these electrodes to external circuits are usually made through the mounting structure. After the blank has been cut from the mother crystal, it is reduced in thickness by successive stages of lapping until it is within etching range of the specified frequency. After thorough cleaning, the blank is etched to final frequency for pressure mounting, or to the preplating frequency if the electrodes are to be metal plated and the unit wire mounted. After the etching process, the blank is again thoroughly cleaned. After cleaning, the blank is base plated, recleaned, and wire mounted in a clean, moisture-free, hermetically sealed holder designed to support the crystal unit against the effects of vibration. After mounting, the metal-plated quartz plate is adjusted to the precise final frequency by additional plating. Typical metals used for plating are gold, silver, aluminum, and nickel. Silver is the metal most often used. Quartz crystal units vibrate in different modes depending upon the principal resonant frequency. The three most common modes being flexure, extensional, and shear. In high-frequency precision-type units, the shear mode is used. Figure 9-12 illustrates the various modes of vibration.

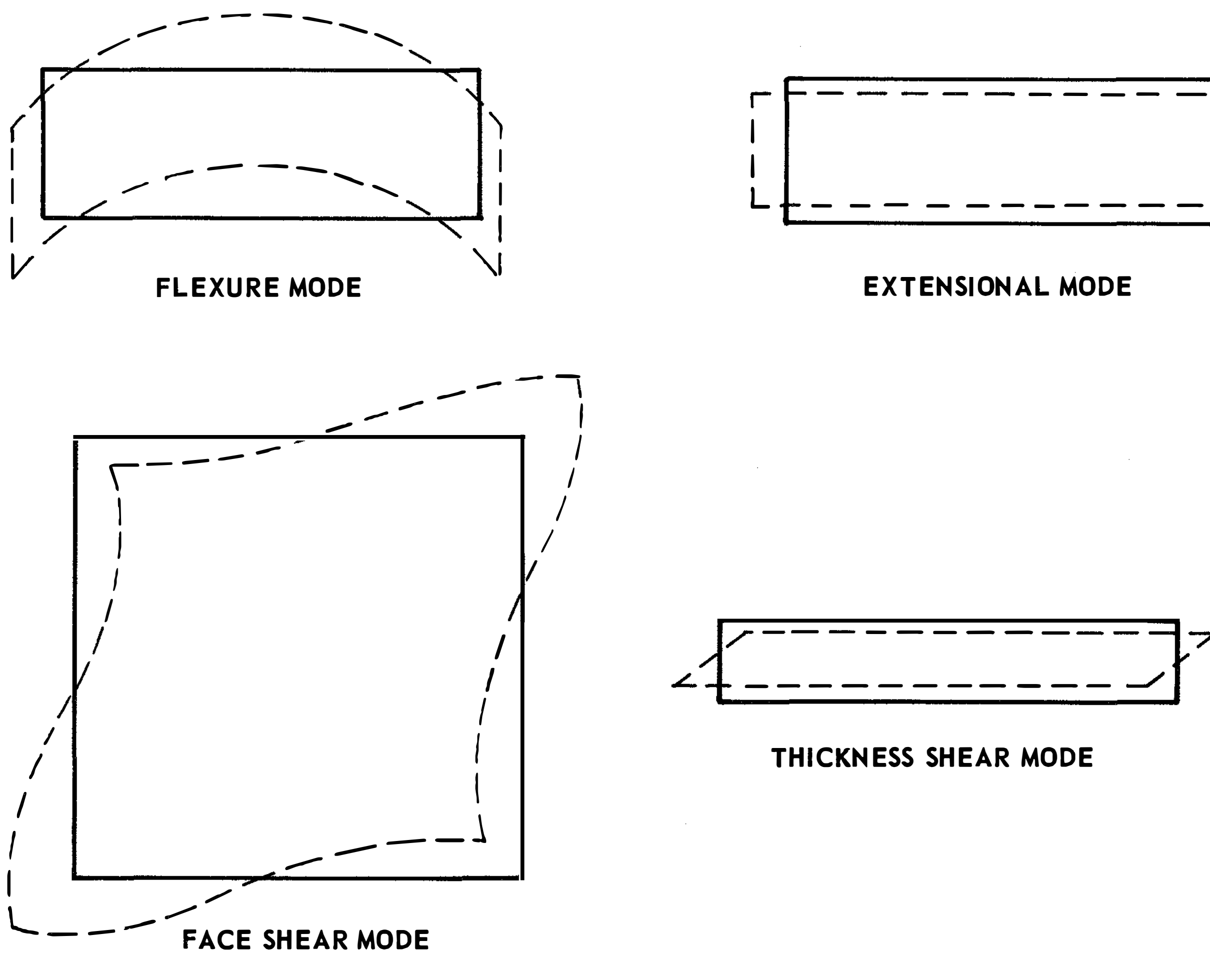


Figure 9-12. Modes of Vibration.

9-30. Characteristics.

9-31. Two-terminal plated quartz resonators may be represented by the electrical equivalent circuit shown in figure 9-13. The series arm consisting of R_1, L_1, and C_1 represents the motional impedance of the quartz plate, while C_o represents the electrode capacitance, C_2, plus the holder capacitance, C_h. At a single frequency, this can be simplified to an effective reactance, X_e, in series with an effective resistance, R_e. These impedances are a function of frequency as shown in the impedance-versus-frequency curve illustrated in four views in figure 9-14. The frequency F_s is the resonant frequency of the series arm and F_r is the resonant frequency of the quartz resonator unit. The antiresonant frequency of the resonator, F_a, is only a fraction of one percent higher than F_r. At frequencies removed from F_a by about one percent, the resonator appears to be a capacitor having a value C_o. Resonators have a number of responses of lesser degree which are usually called unwanted responses. However, certain responses that are approximately harmonically related to the main response are called overtones and are used to control the frequency of VHF oscillators. The equivalent circuit values of a resonator can be controlled to about ±10 percent except for the series arm resistance, R_1. The resonant frequency can be controlled to close tolerances by close dimension control in the construction of the quartz plate. The resonator

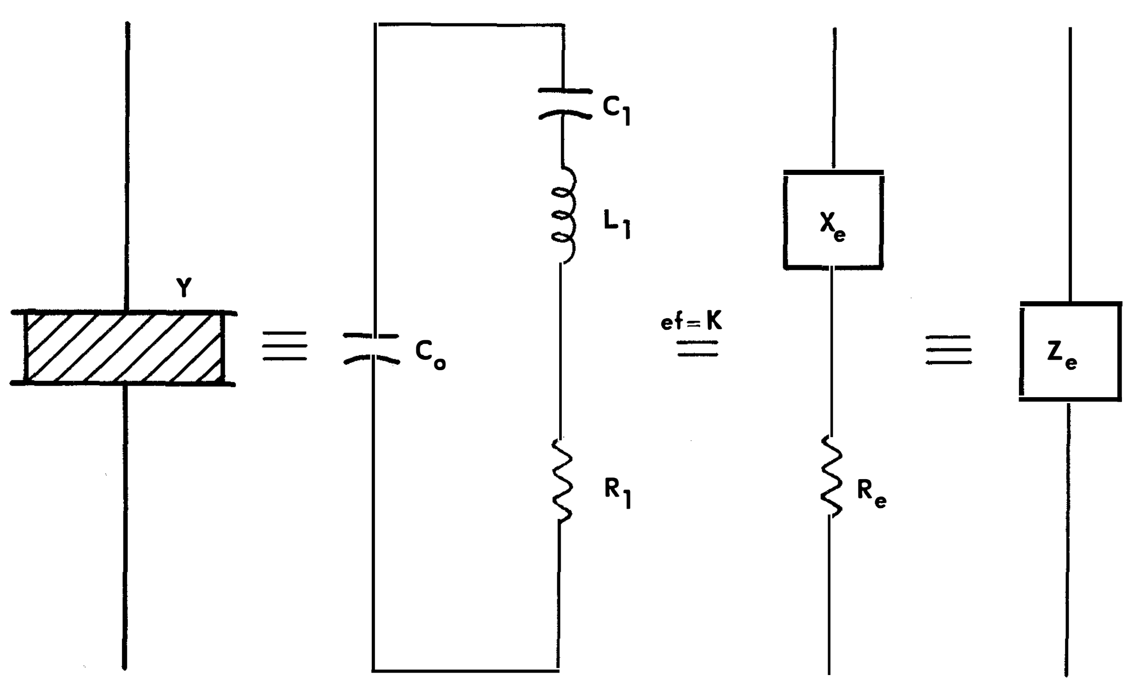


Figure 9-13. Quartz Resonator Equivalent Circuit.

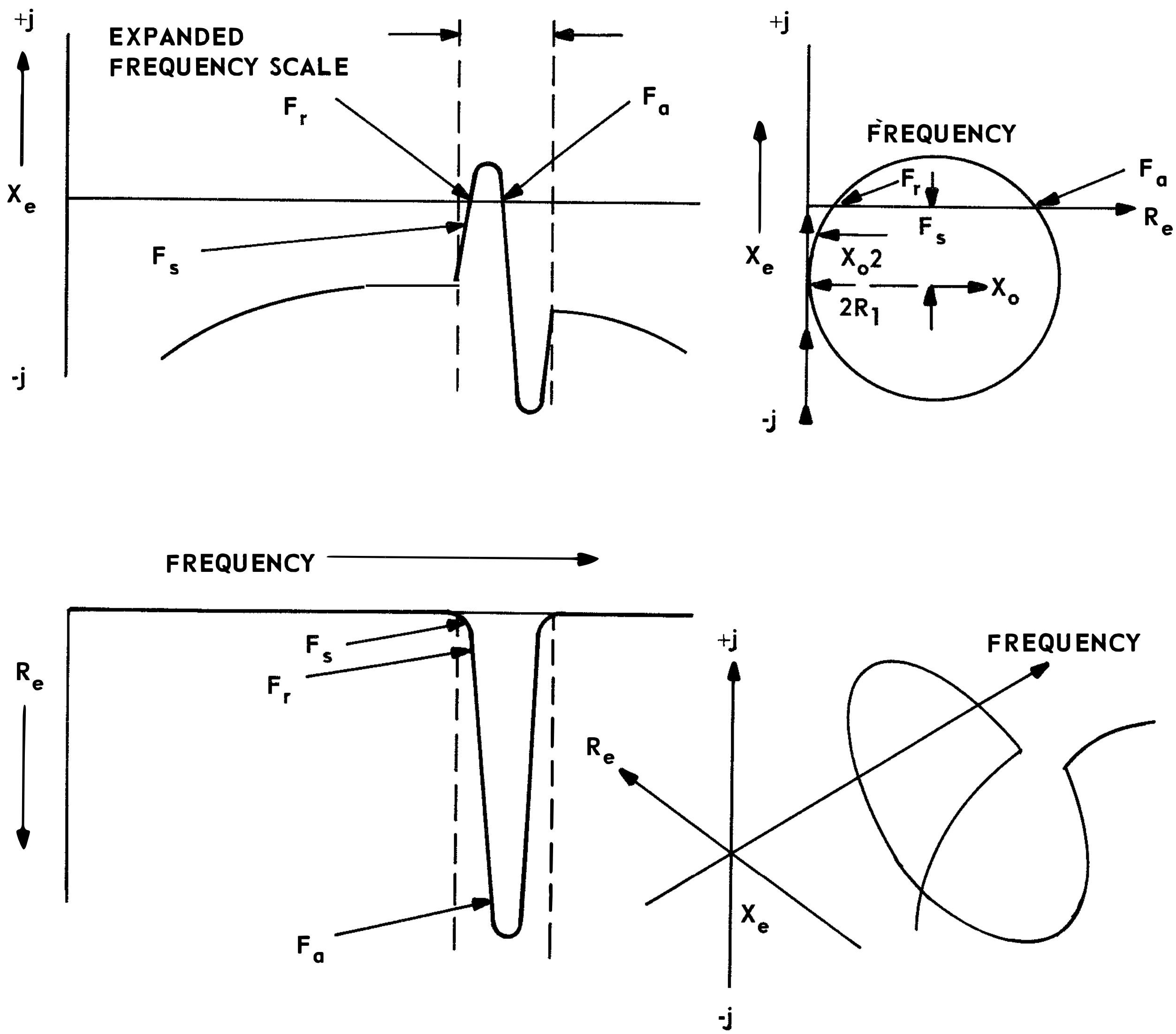


Figure 9-14. Impedance-Versus-Frequency Curve.

performance in a particular application can be calculated if the values of the equivalent circuit are given. If a capacitance, C_X, is added in series with the resonator and this combination operated at its series resonant frequency, F_X, the following formulae hold.

$$X_e = \frac{1}{2\pi F_X C_X}$$

$$R_e = \frac{X_o}{2R_1} - \sqrt{\frac{X_o^4}{4R_1^2} - (X_o + X_x)^2} \quad \text{if } X_o^2 \gg R_1^2$$

$$R_e \approx \left(\frac{C_o + C_x}{C_x}\right)^2 R_1$$

$$R_1 \approx \left(\frac{C_x}{C_o + C_x}\right)^2 R_e$$

$$F_x \approx F_s \left[1 + \frac{C_1}{2(C_o + C_x)}\right]$$

$$\frac{dF_x}{F_x} \approx - \frac{C_1}{2(C_o + C_x)^2} dC_x$$

9-32. The frequency range over which a quartz resonator operates best is determined by the type of cut. Each type of cut has its own optimum frequency range as determined by the physical dimensions of the resonator plate. The following table lists the different cuts and their normal frequency ranges:

Table 9-1. Frequency Range of Quartz Resonators

Fundamental AT	500 Kc to 20 Mc
3rd Overtone AT	10 Mc to 60 Mc
5th Overtone AT	30 Mc to 80 Mc
7th Overtone AT	60 Mc to 120 Mc
CT	300 Kc to 800 Kc
DT	200 Kc to 500 Kc
NT	16 Kc to 100 Kc
+5°X	90 Kc to 300 Kc
Bounded +5°X	1.2 Kc to 10 Kc

9-33. Temperature characteristics of quartz resonators are determined mainly by the orientation of the cut with respect to the crystallographic axes. The frequency versus temperature characteristics are shown in figure 9-15. The peaks of the parabolic shaped curves can be moved so as to appear at any desired temperature by changing

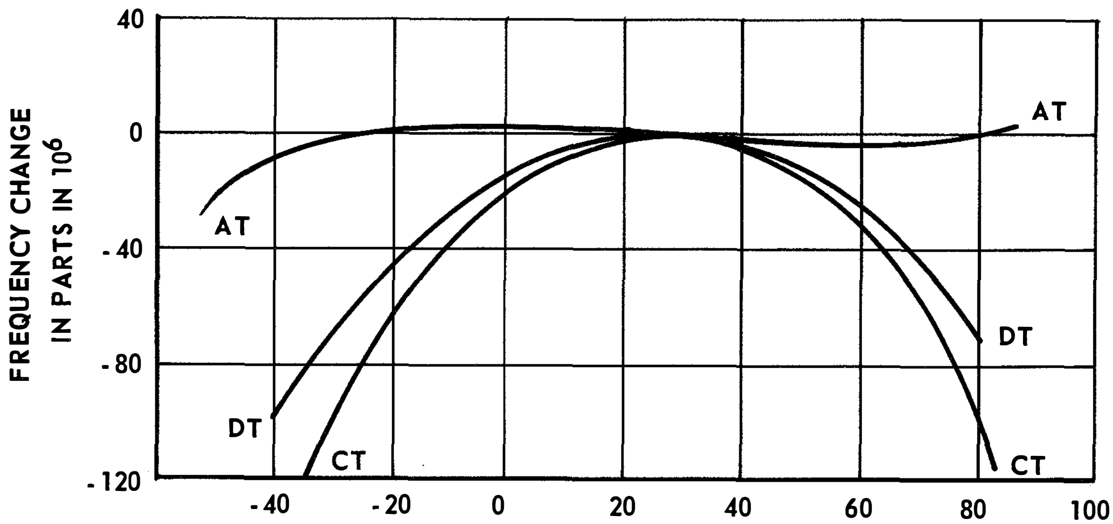


Figure 9-15. Frequency-Versus-Temperature Characteristics.

the orientation of the cut slightly, and the S-shaped curve of the AT-cut resonator can be tipped up or down by the same technique. It is seldom possible to adjust a quartz resonator to an exact resonant frequency at a specified temperature. Normal finishing tolerance for commercial units is about ±20 parts in 10^6. However, in precision resonators, finishing tolerances as low as 1 part in 10^6 have been achieved.

9-34. The resonant frequency of a resonator and the resistance of the series arm are, to some extent, a function of the amplitude of vibration or the power dissipated in the resonator. Below a current of about 100 microamperes, the frequency and resistance are essentially constant. As the current exceeds this critical value, the series arm resistance, R_1, increases and the resonator frequency changes as the square of the current. In AT-cut elements, the frequency increases about 0.1 part in 10^6 per milliwatt per Mc. At still higher values of current, the frequency drifts considerably because of self-heating and finally the resonator fractures because of the large amplitude of vibration. Also, coupling of harmonically related modes of vibration can occur because the vibrations are not linear at the higher amplitudes. This coupling degrades the Q of the wanted response, and since these other modes usually have poor temperature coefficients, the Q depends upon both ambient temperature and resonator current. This Q degradation is known as an activity dip. In AT cuts, the unwanted responses within several percent of the desired frequency are usually higher in frequency than the desired response. These can become prominent enough to control the frequency in oscillator applications. The resonator frequency also changes some with time due to surface contamination of the quartz and to sublimation of the plated electrodes. The actual amount of change depends on the cut, design, cleanness, and construction of the resonator unit. The rate of aging generally increases rapidly with temperature and is sometimes 100 times greater at 40° C than at 0° C. Therefore, aging is more rapid in oven-controlled units. At present, the aging in commercial high-frequency AT-cut crystal units is about 40 parts in 10^6 per year at 85°C. However, in precision, oven-controlled resonators aging rates as low as 1 part in 10^9 per month have been achieved. Normal aging rates for precision units are 1 part in 10^8 per day. Recent studies on the aging rate of quartz resonators indicate that their stability is improved by very low temperature operation. Figure 9-16 shows that stability on the order of 1 part in 10^{10} per day can be achieved by operating commercial grade crystals at 4° K.

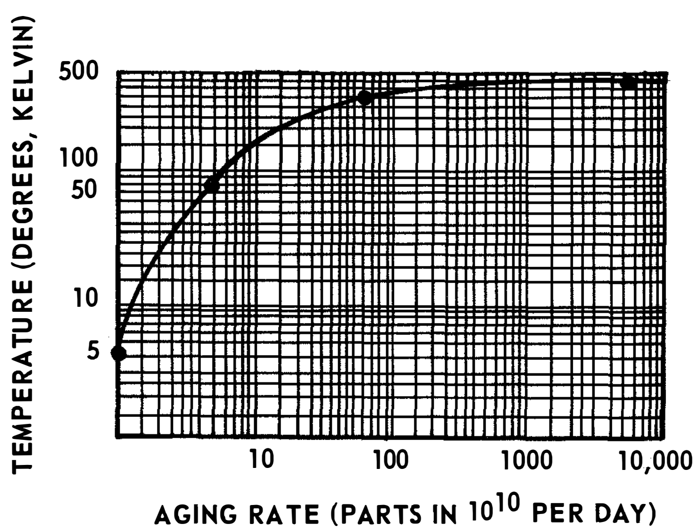


Figure 9-16. Quartz Resonators, Aging Rate Versus Temperature.

9-35. Construction of Precision Resonators.

9-36. Figure 9-17 shows the construction and mode of vibration of a precision crystal resonator, 5th-overtone AT cut. The blank is made circular with one spherical surface and one flat surface. In a crystal of this shape, all of the mechanical vibration takes place near the center of the plate and the edges remain dormant. Thus, supports can be attached to the edges of the plate without degrading Q through damping of the vibrations.

9-37. The quartz plate is usually given a high polish which may be followed by a brief etching operation before the electrodes are plated on. The plating operation is performed in a vacuum in order to minimize contamination and, after plating, the unit is sealed in an evacuated glass or metal envelope. The mode of vibration used is the 5th-overtone in thickness shear. Use of this mode of vibration greatly decreases the volume-to-effective surface ratio and at the same time reduces the effective surface

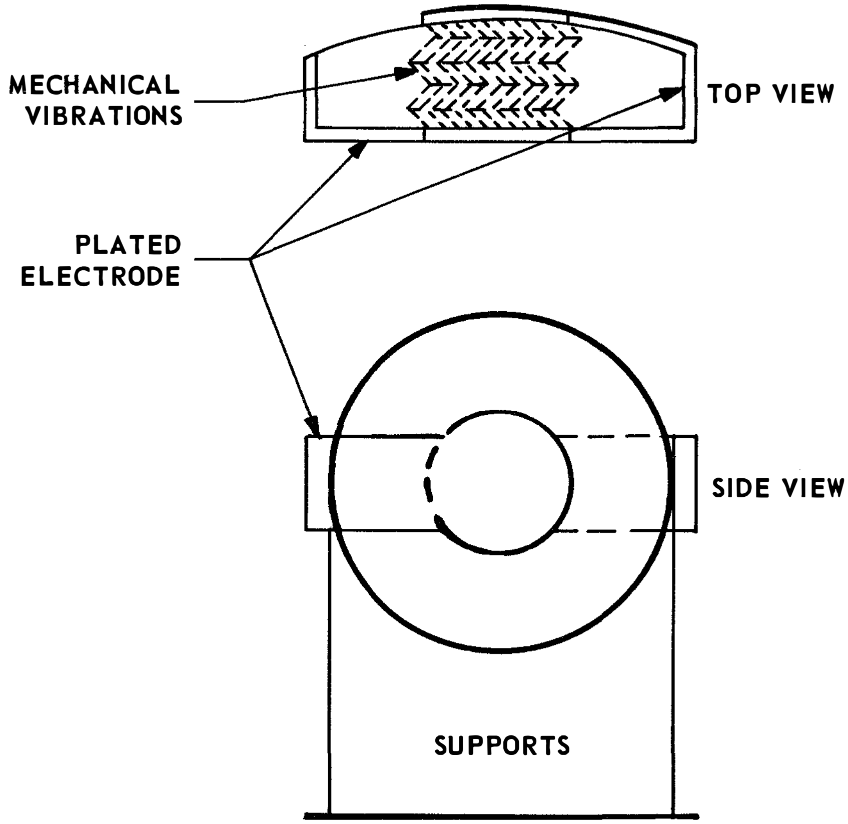


Figure 9-17. Precision Quartz Resonator, Construction and Mode of Vibration of 5th-Overtone AT Cut.

area exposed to contamination since ten effective surfaces, consisting of the five interfaces resulting from 5th-mode operation, are inside the crystal. Typical applications for these units are in 2.5-Mc and 5-Mc frequency standards. The resonant frequency of 5th-overtone AT cut crystals is not affected by shock and vibration below the level that permanently damages the mounting structure.

9-38. Oscillator Theory.

9-39. General Theory.

9-40. Oscillator operation can be analyzed on a feedback basis wherein the oscillator consists of an amplifier with a frequency selective device which couples energy at the desired frequency from the output back to the input. When the circuit is adjusted so that the amplifier supplies sufficient energy at the desired frequency to overcome the losses in the feedback path, the circuit oscillates and generates a signal at a frequency controlled by the resonant frequency of the feedback path. If energy is to be coupled out of the oscillator and used to drive other devices, the amplifier must supply this energy in addition to that required to overcome the losses in the feedback path. In crystal-controlled oscillators, a quartz resonator network provides the coupling from the output of the amplifier to its input. Because of its high Q, the resonator operates as a highly selective feedback network with extremely high attenuation of frequencies on either side of its resonant frequency. Thus, the frequency of oscillation cannot deviate appreciably from the resonator frequency. Since the output of the feedback network is the input of the amplifier, the total phase shift around the loop must be zero. For this reason, the resonator must compensate for phase shifts in the rest of the oscillator circuit and these phase shifts will affect the frequency stability of the circuit.

9-41. Another method of analysis is that based on the negative resistance theory. Figure 9-18a is an equivalent circuit of an oscillator operating at series resonance. The input impedance of the oscillator is a negative resistance, R_{in}, and the resonator has an effective resistance, R_e. The power loss in the resonator is I^2R_e, and the power supplied by the oscillator is I^2R_{in}. If the power gain is greater than the power loss, oscillations will build up; if the power gain is less than the power loss, oscillations will die out. The negative oscillator input resistance is a function of the current I so that R_{in} will decrease as oscillations build up, until R_{in} equals R_e and a stable amplitude is reached. A more general case is illustrated in figure 9-18b in which the oscillator input impedance has a capacitive component. It is standard practice to make this input capacitance, C_{in}, 32 $\mu\mu$f at frequencies above 500 Kc and 20 $\mu\mu$f at frequencies below

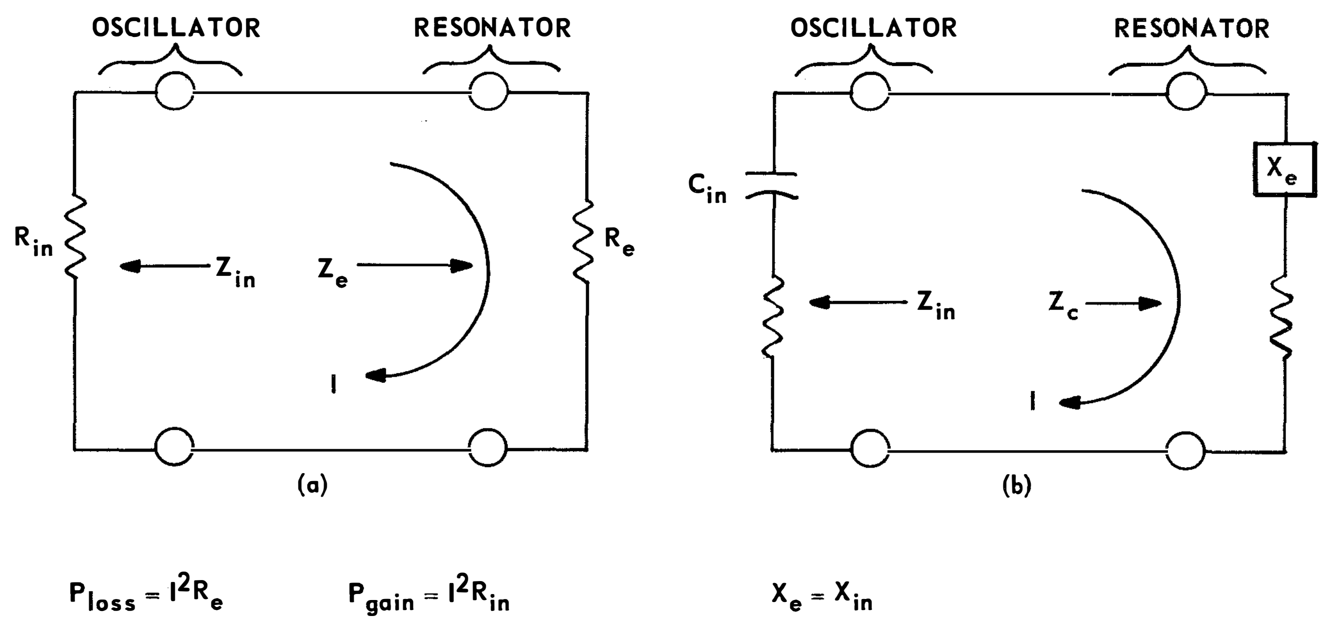


Figure 9-18. Equivalent Circuit of a Crystal-Controlled Oscillator.

500 Kc. The entire network oscillates at a frequency such that $X_e = -X_{in}$, and the equations for power loss and gain given for figure 9-18a still apply. All types of oscillators can be analyzed in this manner except that, in a few special cases, the reactive component of oscillator input impedance may be inductive. Mathematical analysis consists of replacing the resonator with an imaginary test voltage generator and solving for $Z_{in} = \frac{E_{in}}{I_{in}} = R_{in} + JX_{in}$. Figure 9-19 illustrates a method of calculating the power dissipation in the resonator for vacuum-tube saturation limiting. In the first equivalent circuit (a), the oscillator is represented by a generator, E_g, with a series generator resistance, R_g. Since the resonator reactance, X_e, equals the negative reactance, $-X_{in}$, of the input capacitance, C_{in}, these two reactances cancel, and the equivalent circuit is reduced to the second circuit (b) shown. After the grid voltage on the oscillator tube reaches the value that saturates the tube, the generator voltage E_g remains relatively constant and independent of grid drive. Then resonator current is given by $I = \frac{E_g}{R_g + R_e}$, and the power dissipated in the resonator is given by

$$P = I^2 R_e = \frac{E_g^2 R_e}{(R_g + R_e)^2} .$$

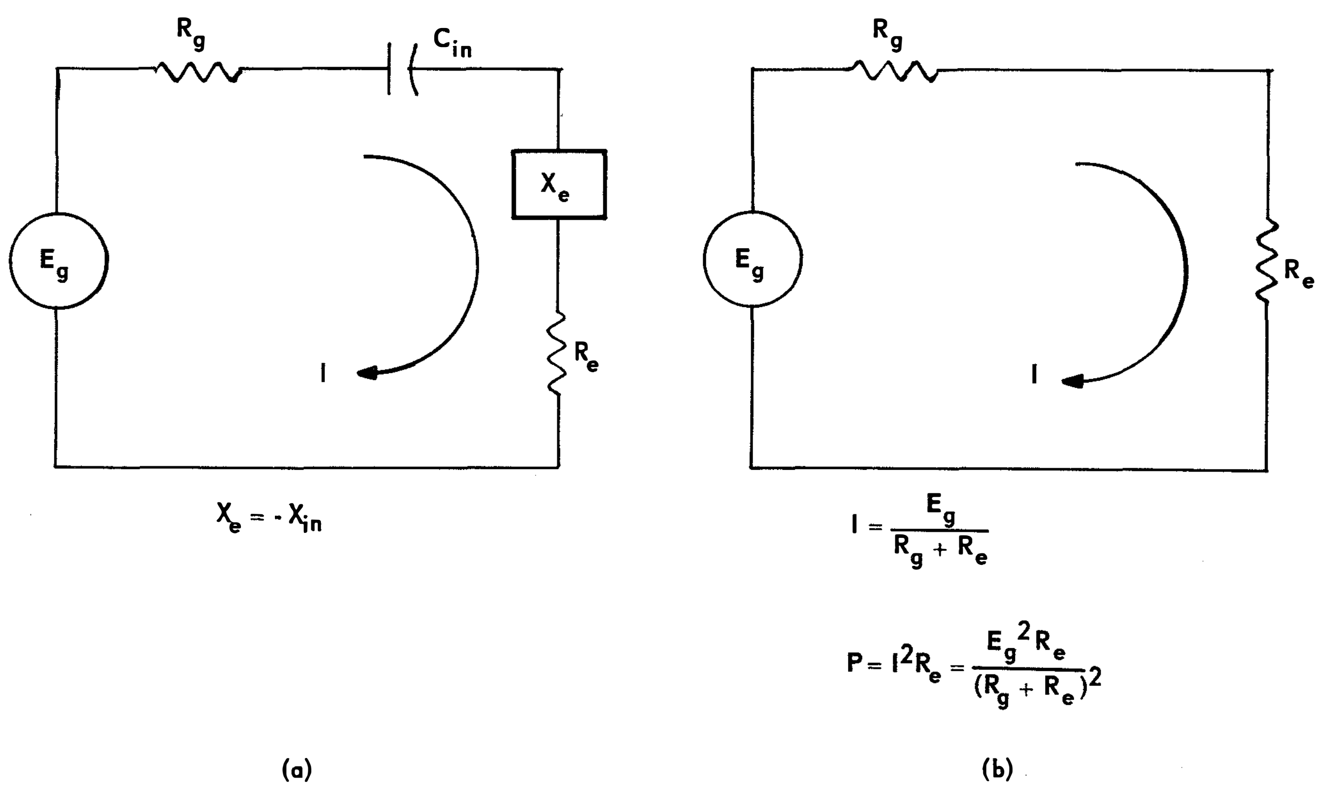


Figure 9-19. Calculation of Resonator Power Dissipation.

9-42. Typical Oscillator Circuits.

9-43. Quartz crystal resonators have two resonant frequencies. At one frequency they exhibit antiresonant characteristics, and at a slightly lower frequency they exhibit series resonant characteristics. At frequencies between these two the resonator reactance is inductive, and at frequencies outside this range the reactance is capacitive. The design of the oscillator circuit determines in which part of the reactance characteristic it will be used. In the oscillator represented by figure 9-20, the series resonant response is used. The amplifier is designed so that the total phase shift from amplifier input to output is zero. Since the total phase shift around the complete loop, including the feedback network, must be zero, the feedback network must also have zero phase shift. If the feedback network is to have zero phase shift, it must be resistive at the frequency of oscillation. The quartz resonator which forms the feedback network is resistive at two frequencies, its antiresonant frequency and its series resonant frequency. Since the resonator is in series with the feedback path, the frequency at which it offers the least resistance to the signal is its series resonant frequency, and this will be the frequency of oscillation.

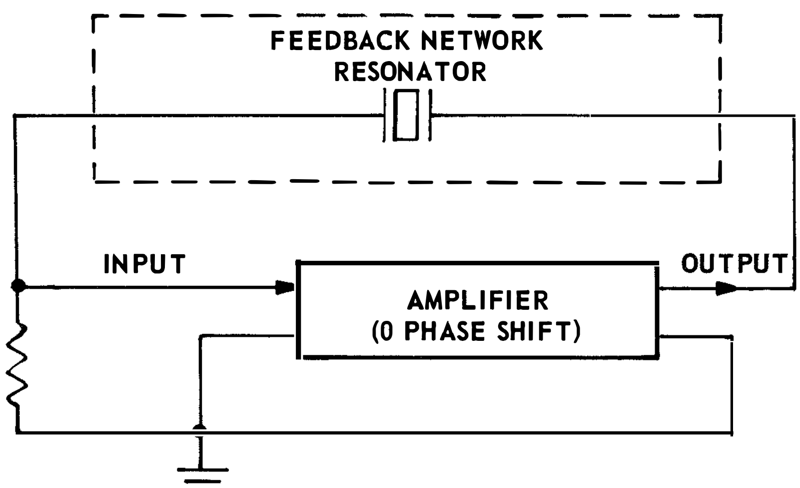


Figure 9-20. Basic Crystal Oscillator, Operating the Resonator at Series Resonance.

9-44. Figure 9-21 shows the basic Pierce oscillator circuit, (a) an equivalent circuit, (b) and a vector diagram, (c) showing phase relationships, neglecting circuit losses. In this oscillator, the feedback network operates at antiresonance, but the resonator operates at a point between its series resonant frequency and its antiresonant frequency where it is sufficiently inductive to resonate with C_p and C_g in series. The generator

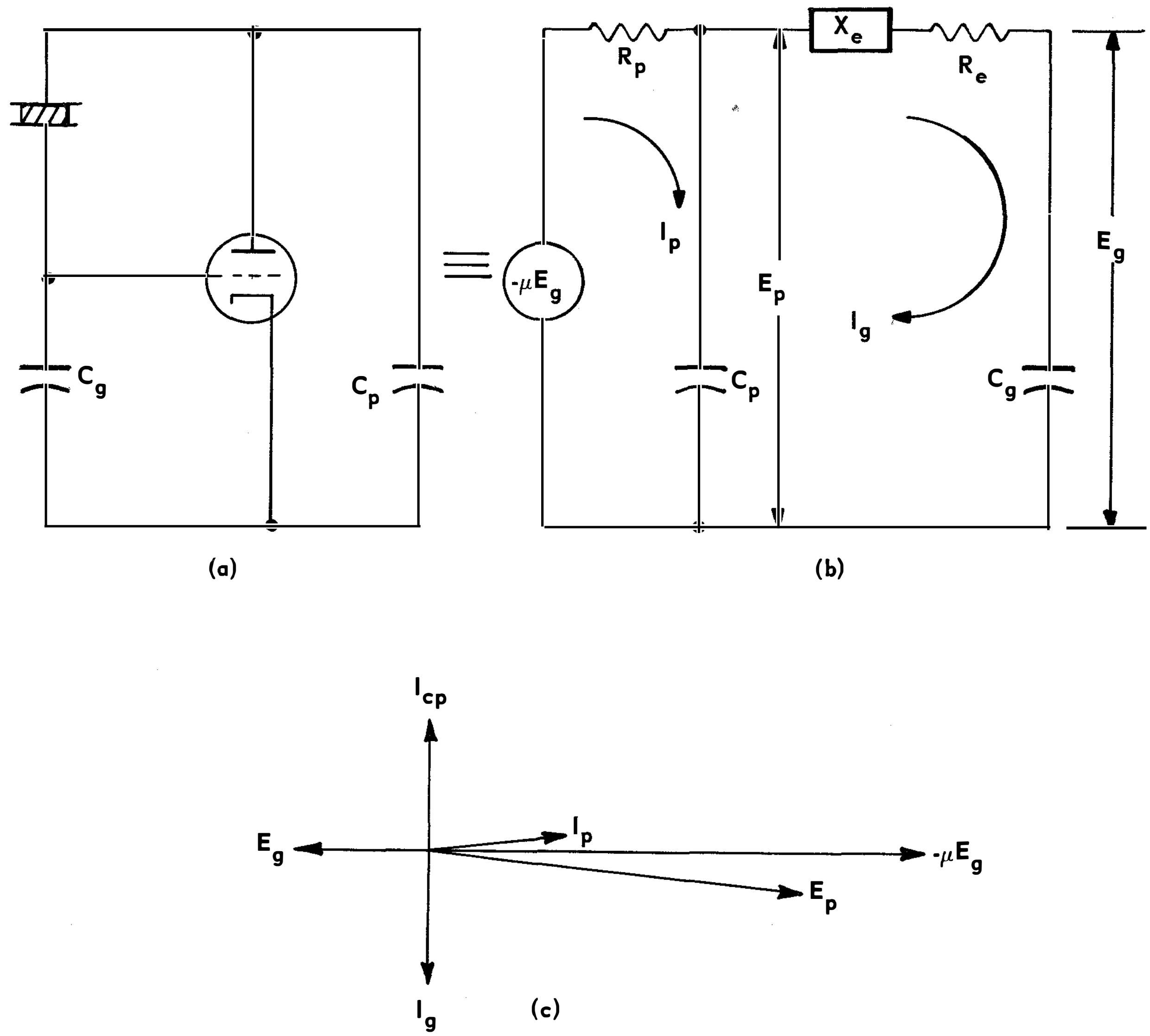


Figure 9-21. Basic Pierce Oscillator.

voltage $-\mu E_g$ is the grid voltage multiplied by the gain of the tube. Since the circuit representing the generator load is resonant, I_p will be in phase with $-\mu E_g$; and since the branch consisting of X_e, R_e, and C_g is inductive, I_g will lag $-\mu E_g$ by 90°. The voltage, E_g, developed across C_g will lag I_g by 90°. Therefore, E_g will lag $-\mu E_g$ by 180°, and since the tube introduces another 180° phase shift, the condition that there be zero phase shift around the loop is satisfied.

9-45. Precision Crystal-Controlled Oscillators.

9-46. In the design of precision oscillators, several precautions must be taken to minimize instabilities. The contruction of the quartz resonator itself was described in paragraph 9-28 of this section. Additional precautions to be observed in the use of quartz resonators in precision oscillators are:

(1) the components that make up the resonant circuit must be placed in a controlled environment;

(2) the amplitude of oscillation must be controlled to avoid instabilities caused by nonlinearity in the vibration pattern of the resonator at high amplitudes;

(3) phase instabilities in the active amplifying portion of the oscillator must be held to a minimum;

(4) external circuitry must be isolated from the oscillator so that reactive components are not reflected back into the resonant circuit to cause instability;

(5) the Q of the resonator should be high so that loop phase shifts can be compensated for with minimum change in resonator operating frequency. In addition, high Q makes possible low coupling between the resonator and the active amplifying portion of the oscillator, thus minimizing the effect of the active network on the resonator;

(6) nonlinearities in the active amplifying portion of the oscillator cause harmonic distortion. Adjacent harmonics are mixed together in the same or other nonlinear portion of the circuit after having passed around the feedback network. The fundamental frequency component thus produced is usually not phase stable and causes phase instability in the oscillator. Therefore, the amplifier must be operated on the linear portion of its characteristic.

9-47. Figure 9-22 illustrates the principle of operation of the Meacham oscillator. The resonator, Y1, offers a low resistance and zero phase shift to the frequency of oscillation. The opposite leg of the bridge circuit is an incandescent lamp which, when cold, also has a low resistance. Resistors R1 and R2 have about the same resistance as the effective resistance of the resonator at its series resonant frequency. This condition exists when oscillations start. The coupling between the amplifier and the feedback loop is relatively tight, and there is a large amount of positive feedback. As oscillations build up, lamp R3 is heated by the RF current and its resistance increases until it is almost equal to that of R1. As the lamp resistance increases, the bridge approaches balance; the positive feedback is reduced until the bridge is almost in balance and the residual positive feedback is just sufficient to sustain oscillation. This oscillator is usually used only at frequencies below 1 Mc because of the difficulties of obtaining transformers which do not cause phase instabilities at the higher frequencies.

9-48. Figure 9-23 is a schematic diagram of a typical Pierce oscillator used in high precision frequency standards. The 1-Mc quartz resonator, V1, is a fundamental AT cut crystal, sealed in an evacuated glass envelope. Its temperature coefficient is only several parts in 10^7 per degree centigrade. The components that make up the resonant

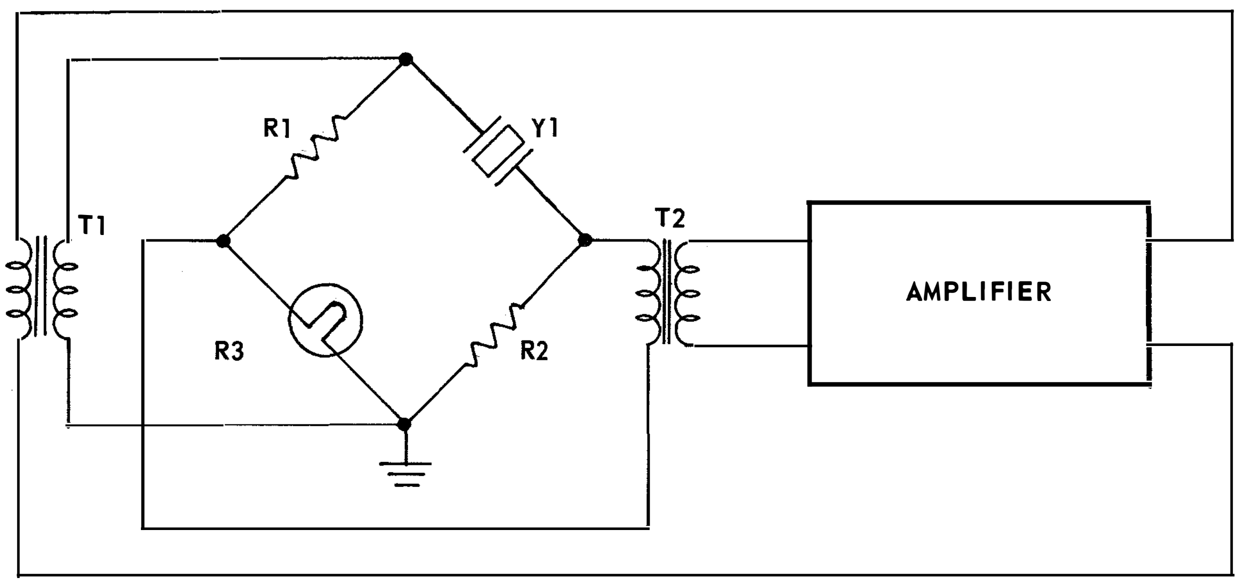


Figure 9-22. Meacham Oscillator.

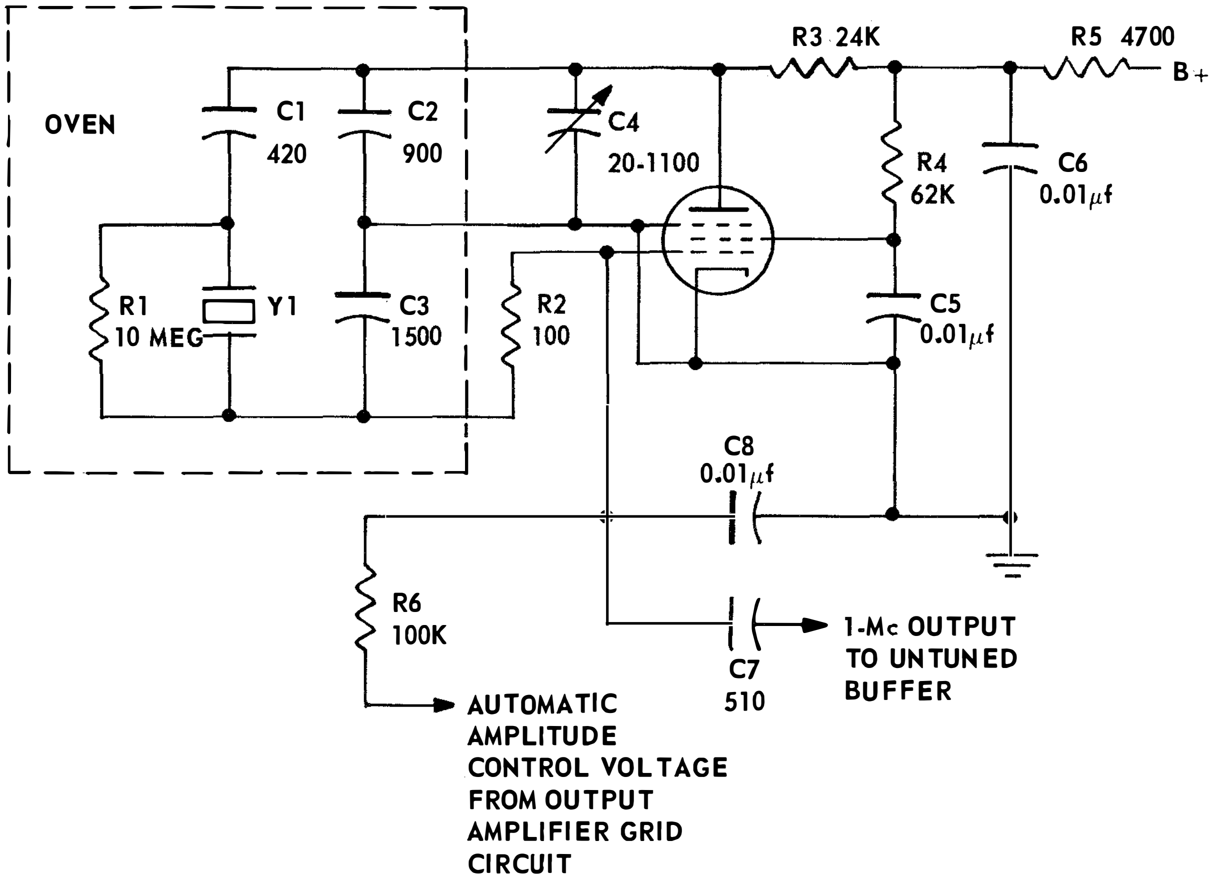


Figure 9-23. Precision Pierce Oscillator.

circuit, Y1, R1, C1, C2, and C3 are housed in an oven in which the temperature is held constant to better than 0.01°C. Capacitor C1 and resistor R1 hold the DC voltage impressed on the resonator, Y1, to a minimum. The resonator has a minimum Q of 1 million, thus capacitors C2 and C3 can be made large to bypass effectively the plate and grid of the tube to ground and reduce the coupling between the resonator and the active portion of the circuit to a low value. In addition, all frequency controlling components are isolated from other circuit elements by shielding. Capacitor C4 is a precision variable capacitor which provides a small range of adjustment of the resonant frequency of the circuit. The total range of adjustment is about 4 cps at the nominal operating frequency of 1 Mc. The output of the oscillator is coupled to an untuned buffer stage which isolates the oscillator from succeeding stages. Two stages of amplification follow the buffer stage and provide additional isolation. The amplitude of oscillation is controlled by negative voltage developed in the grid circuit of the last amplifier stage. Cathode bias on this tube delays development of negative voltage until the signal applied to the grid reaches a predetermined level. When this level is reached, the resultant negative voltage coupled to the grid of the oscillator tube through an RC filter increases the grid bias, thus reducing the tube gain and limiting the amplitude of oscillation to a low level. This automatic amplitude control system holds the operating power level in the quartz resonator to less than 0.1 microwatt. This type of oscillator has attained short-term stability of better than 1 part in 10^{10} and long-term stability of better than 1 part in 10^{9} per day. Oscillators are now being designed to use the 5th overtone AT cut crystal. These are expected to have even better stabilities than oscillators using the fundamental AT cut crystal.

9-49. Oven Theory.

9-50. General Theory.

9-51. Since all quartz resonators have some variation of frequency with temperature, the resonator must be kept at a constant temperature in order to achieve maximum stability. In most frequency standards this is accomplished by placing the resonator in an oven and maintaining the oven temperature at a level somewhat higher than the ambient temperature surrounding it. The six items listed below make up a typical oven.

(1) The resonator or device to be temperature controlled

(2) The oven heater

(3) A device for controlling the power delivered to the heater

(4) A temperature sensing element

(5) A heat sink (ambient temperature around oven)

(6) Thermal insulation or thermal resistance

The operation of the oven can be compared to the operation of an electrical bridge circuit as illustrated in figure 9-24. The arms of the bridge R1, R2, R3, and R4 represent thermal resistance, that is, resistance to heat flow. The temperatures T_h, T_r, T_a,

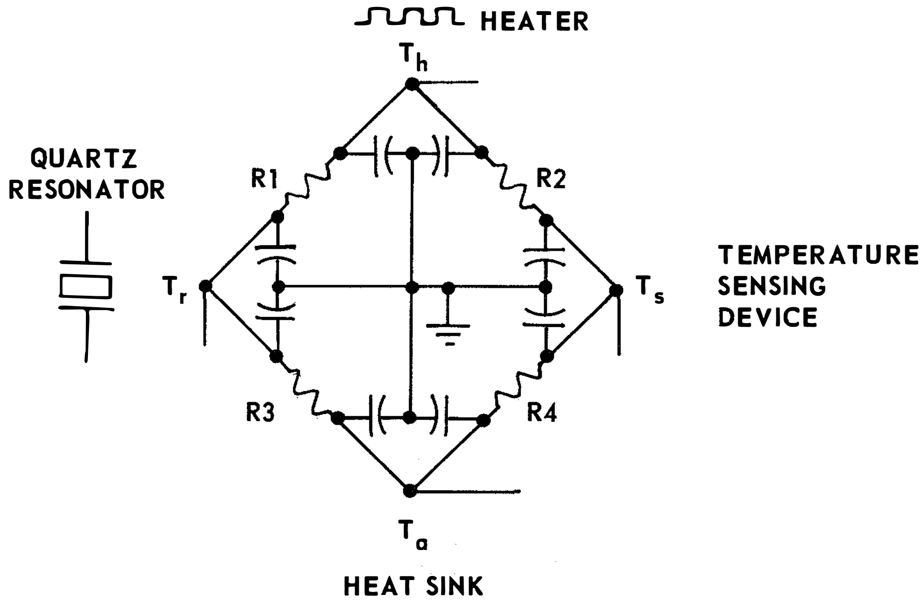


Figure 9-24. Oven Operation, Equivalent Electrical Circuit.

and T_s are analogous to electrical potentials at the points indicated. T_h is the heater temperature; T_r is the resonator temperature; T_a is the temperature surrounding the oven; and T_s is the temperature of the sensing element. The heat storage or thermal capacities of the materials in the heat flow path are analogous to electrical capacitance. The temperature of the heater T_h, is regulated by the sensing element through a servo system so that the temperature T_s remains constant. To maintain the temperature T_r of the resonator constant regardless of variations in T_a, T_r must equal T_s; that is, the bridge must be balanced, or R1/R3 must equal R2/R4. Conditions for balance are less critical if R1 and R2 are made very small as compared to R3 and R4, since T_h, T_s, and T_r will be more nearly equal. The time lag between T_h and T_s, caused by the time constant of the thermal resistance R2 and its associated thermal capacities, causes the servo system to hunt and this, in turn, causes the temperature T_s to cycle. Reducing the time constant of R2 and its associated capacities to a very low value eliminates this cause of hunting. However, another cause of hunting is the operating differential of thermostats. When these are used as temperature sensing devices, the time constant of R1 and its associated capacities must be made long in order to filter out variations in T_r caused by hunting in the servo system. If proportional control is used, the time lag due to operating differential in the sensing element is eliminated, and the time constant of R2 and its associated capacities can be made very low to eliminate hunting. If the servo system is free of hunting, then the time constant of R1 and its associated capacities can be made low, and if at the same time the time constants in the R3 leg and the R4 leg are made long, T_s and T_r will be on an isothermal line with T_h. Ovens using this system can maintain temperature within 0.01°C.

9-52. Typical Precision Oven.

9-53. Figure 9-25 illustrates the construction of a typical precision oven. In this oven, the resonator, the heater, and the temperature sensing element are all in an isothermal space. The resonator in its sealed envelope is housed in an aluminum cylinder

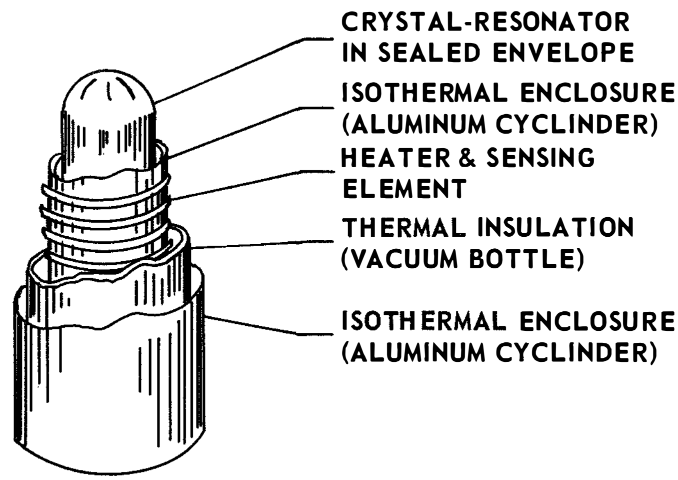


Figure 9-25. Typical Oven Construction.

upon which the heater is wound. Because of the high heat conductivity of aluminum and because the resonator is almost completely surrounded by aluminum, the temperature of the resonator is nearly identical to that of the aluminum enclosure. The heater is wound on this enclosure and tightly coupled to it thermally. Thus, the resistance R1 in figure 9-24 and the thermal time constant between T_h and T_r are nearly zero. The heater is constructed so that it is also the temperature sensing element, making R2 in figure 9-24 and the time constant between T_h and T_s nearly zero. The resistances R3 and R4 are made very large by housing this assembly in a vacuum bottle. The vacuum bottle is enclosed in a second aluminum cylinder, which makes T_a uniform on all sides of the oven.

9-54. Since the heater, the resonator, and the temperature sensing element have been placed in an isothermal space well isolated from the ambient temperature, the only remaining requirement is to maintain the heater at a constant temperature. The circuit of figure 9-26 satisfies this requirement. The bridge circuit, HR601, performs

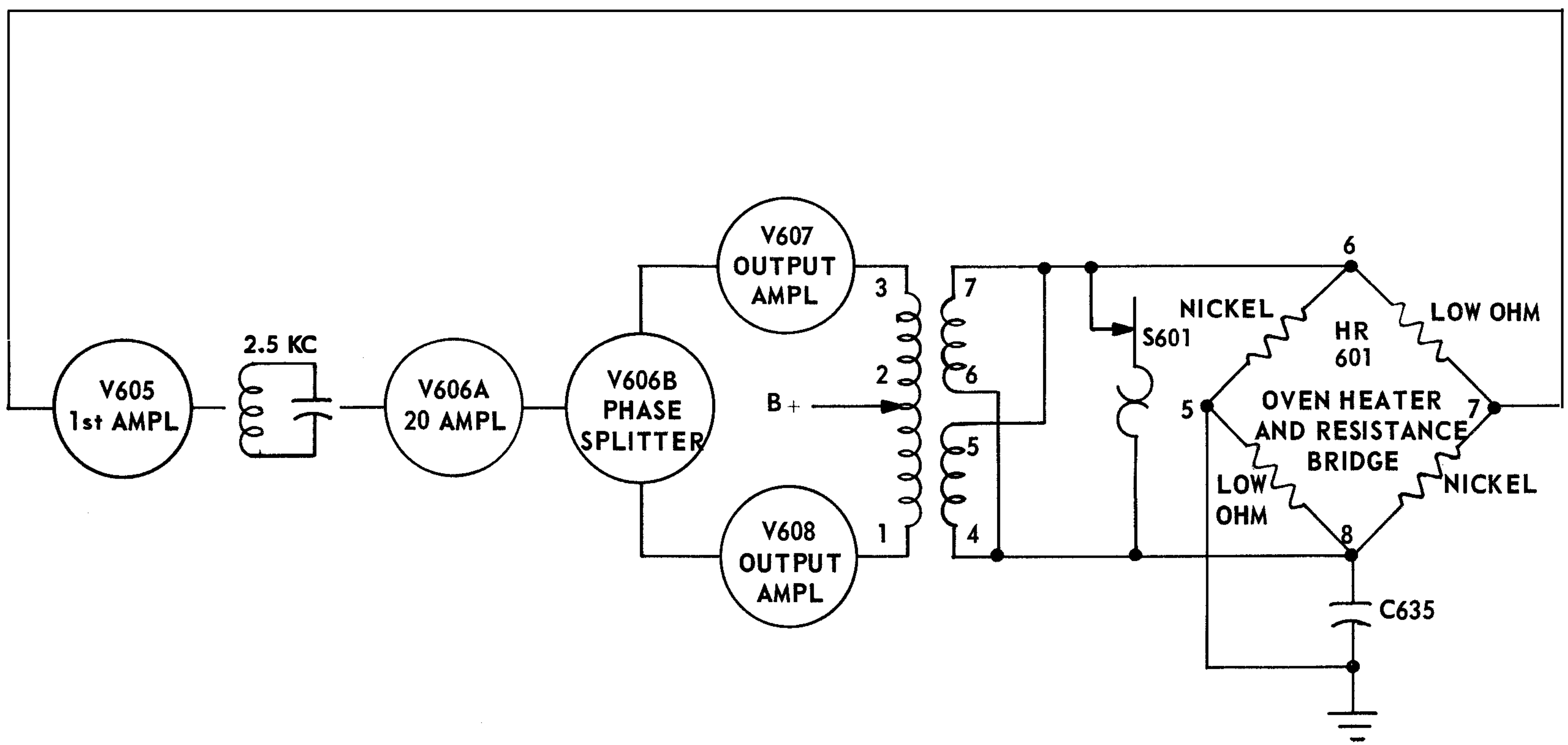


Figure 9-26. Oven-Control Oscillator.

two functions. It is the heating element for the oven and the control element for the oven oscillator. Two arms of the bridge are made of nickel wire, and the other two arms are made of low resistance wire. The arms are of selected lengths so that their resistances at the desired oven temperature are almost equal. When the oven temperature is low, the nickel wire has less resistance than the low resistance wire, and terminals 5 and 7 of the secondary winding of T601 see less resistance to ground than do terminals 4 and 6. At the same time, terminals 4 and 6 of T601 see less resistance to the feedback path than do terminals 5 and 7. As a result, the alternating current flowing through the bridge is applied as positive feedback to the first amplifier stage. Under these conditions, the circuit oscillates at an amplitude determined by the amount of bridge unbalance which, in turn, is controlled by the temperature. Thus, proportional control is provided, and the thermal lag inherent in thermostatic devices is eliminated. The power supplied to HR601 by the oscillator heats the oven. As the temperature approaches the desired level, the bridge approaches balance, reducing the amount of feedback until it is just sufficient to sustain oscillation. When the oven reaches this steady state condition, the oven-control oscillator supplies just enough power to the heater to replace the heat lost to the surrounding medium and maintains the oven temperature constant within 0.01°C. If for any reason the temperature of the oven rises above the desired level, the bridge becomes unbalanced in the opposite direction and resultant negative feedback prevents oscillation.

9-55. Frequency Divider Theory.

9-56. In order to obtain maximum stabilities, standard signals must be generated at higher frequencies than the lowest frequency required for use in the equipment. In order to obtain the lower frequencies, frequency dividers must be used, and if the divided frequency is to have the same stability as the original, these dividers must be under control of the standard. Figure 9-27 is a block diagram of a typical divider. The equipment contains two regenerative dividers which divide their input frequencies by 10. With a 1-Mc input, this circuit provides outputs at 1 Mc, 100 Kc, and 10 Kc. The principles of operation of the two dividers are identical except for the frequencies involved; therefore, only the 1-Mc to 100-Kc divider will be discussed. When the 1-Mc signal supplied to the injection grid of the 1 Mc - 900 Kc mixer is large enough to make the circuit sufficiently regenerative, noise energy at 900 Kc appearing at the signal grid mixes with the 1-Mc signal to produce sufficient 100-Kc signal to drive the 300 Kc multiplier. This circuit multiplies the 100-Kc signal by 3, producing a 300-Kc signal which drives the 900 Kc multiplier. The 300-Kc signal is again multiplied by 3 to produce a 900-Kc signal for mixing in the 1 Mc - 900 Kc mixer. The 100-Kc signal thus produced is under complete control of the 1-Mc signal. If the 1-Mc injection falls below the threshold level, the loop gain of the circuit falls below the level required to maintain operation, and no output is available. The second divider circuit operates from the 100-Kc signal in the same way to produce a 10-Kc signal. The cathode followers used to couple the signals to external circuits provide isolation.

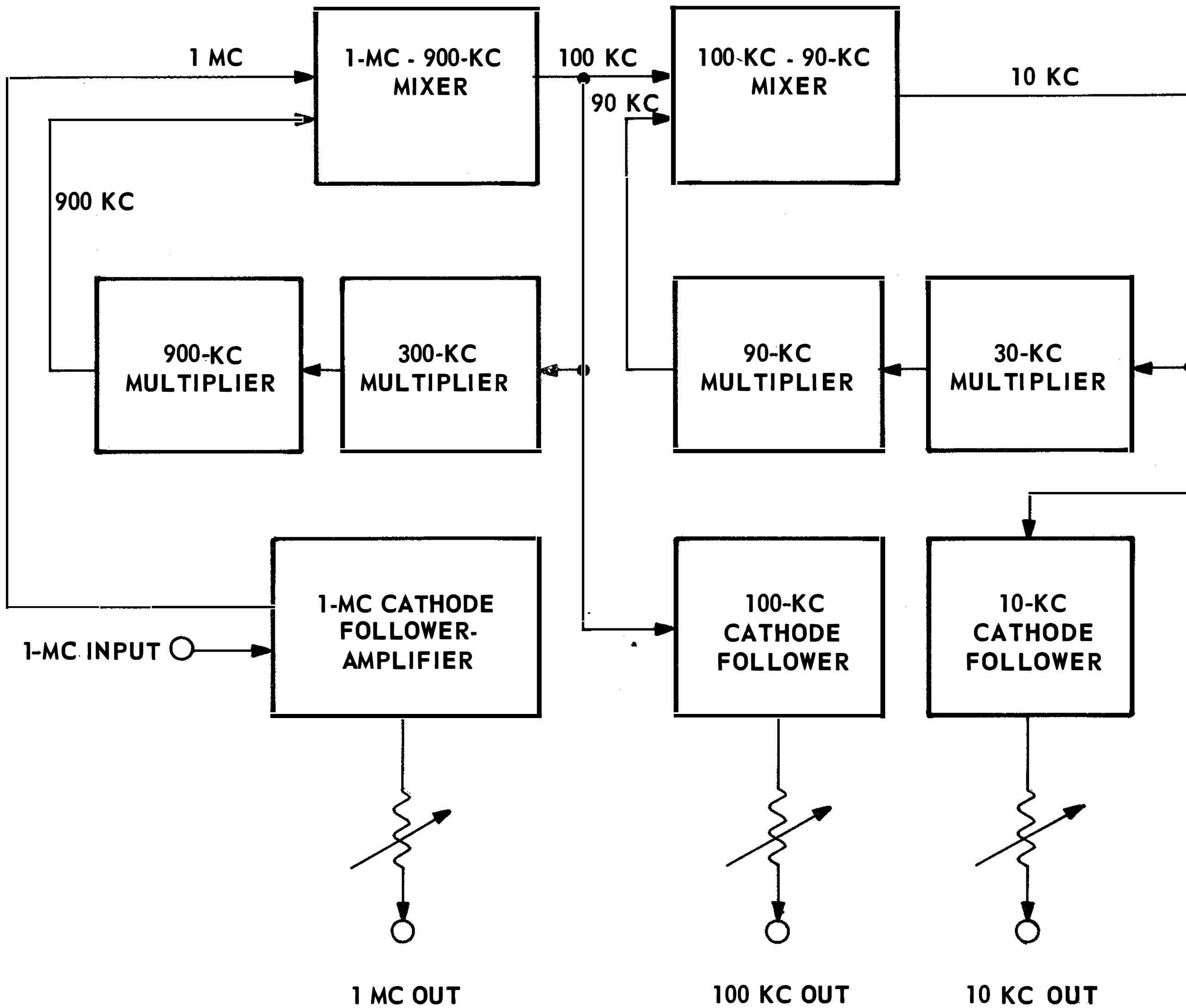


Figure 9-27. Block Diagram, Frequency Divider.

9-57. <u>System Considerations.</u>

9-58. In single sideband communications, the total frequency shift in the system, both transmitting and receiving, should not exceed 50 cps. At 20 Mc this requires a system stability of 2.5 parts in 10^6, including errors introduced by the propagating medium, Doppler shifts, and errors in terminal equipment. To assure total system stability of 2.5 parts in 10^6 over a period of months without readjustment, stabilities of 1 part in 10^8 per day are required in the frequency standards. Frequency, time, and phase stability requirements in other systems, such as Kineplex*, are even more severe. In the Kineplex system, 22-millisecond pulses are transmitted. Each pulse is a reference for the succeeding pulse. Short-term phase stability for this system must be within a few degrees over a 44-millisecond period. Frequency accuracy must be ±0.5 cps to prevent deterioration of the signal. Errors of ±1 cps cause noticeable distortion and ±3 cps is the practical limit of permissible frequency error. Time accuracy within ±1 millisecond would provide a signal with no noticeable deterioration, but ±5 milliseconds is the practical limit of time error. Thus, in these systems, total system frequency stabilities of 6 parts in 10^7 and total system time stabilities of 1 part in 10^8 are required. In mobile systems, Doppler shifts and time variations due to changing transmission path lengths must be compensated for either by automatic correction circuits or by manual readjustment of local equipment.

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* Registered in U. S. Patent Office

SECTION X

PRINCIPLES OF SERVOMECHANISMS

10-1. Definitions.

10-2. A servomechanism is most commonly defined as a feedback control system of which at least one element is mechanical in nature. Voltage regulators for power supplies, automatic volume and frequency control circuits used in radio equipment, and thermostats used to regulate temperature in home heating equipment and in various electrical applicances are examples of feedback control systems. Power steering, gun turret positioning devices, and airplane autopilots are examples of servomechanisms.

10-3. In all feedback control systems, the quantity to be controlled is measured in some manner. This measured value is compared to a desired, or reference, value to form an error signal, and the controlling action is governed by some function of the error signal. Feedback control devices may contain electrical, mechanical, pneumatic, hydraulic, and other types of elements. Frequently, a human operator is included in the feedback loop.

10-4. A Typical Position Servo System.

10-5. Figure 10-1 illustrates a typical position follow-up type of servomechanism. This type of device can be used to repeat the position of a shaft at a remote point. For example, in an airborne radio transmitter, it could make the shaft of a precision oscillator follow a dial in the pilot's control box. Follow-up servos are also used to repeat the shaft position of a delicate instrument at a shaft where a large amount of torque is needed. In this case, the purpose of the servo is to provide torque amplification.

10-6. In figure 10-1 reference input R is a shaft position. A voltage proportional to the shaft position is obtained from a linear potentiometer connected across a battery. This voltage is mixed with a voltage proportional to the controlled variable to form the error signal E, which is then amplified and applied to the winding of a motor. The motor shaft is coupled through a gear train to the load, which in most cases is a friction device, although it sometimes contains a significant amount of inertia. Coupled to the load shaft is another potentiometer which produces a voltage C, proportional to the position of the output shaft. The mixing circuit subtracts the controlled voltage C from the reference voltage R to obtain error signal E. The amplifier drives the motor in

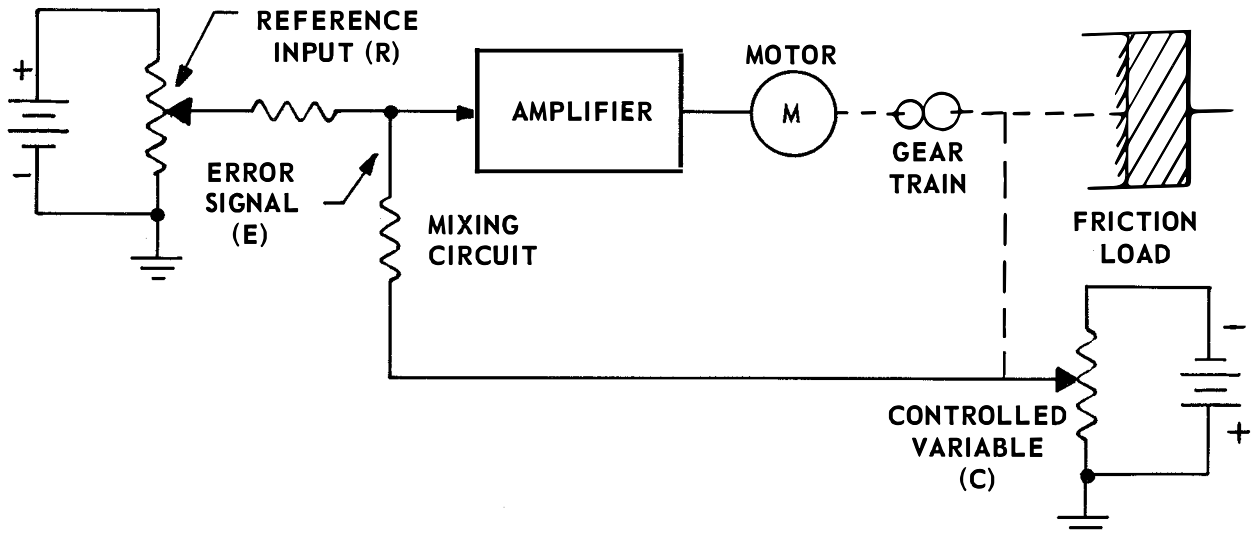

Figure 10-1. Position Follow-up Servomechanism.

such a way that if E is positive the motor turns in one direction, and if E is negative the motor reverses. When E is zero, the motor stops. Below the saturation level of the motor and amplifier, the motor speed is proportional to the error signal. A device of this kind is called a proportional controller.

10-7. Suppose that, when power is applied to the circuit, reference input R is greater than controlled variable C, so that E is positive. If the motor is connected so that a positive E causes the motor to turn in such a direction as to increase C, then as the motor turns, E will decrease and the motor will slow down until R and C become equal. Then E is zero and the motor stops. If the battery voltages feeding the reference input and the control variable potentiometers are equal, and the electrical angles of the two potentiometers are equal, this null condition will occur only when the load shaft is at the same angular position as the reference-input shaft. If some external force, such as vibration, displaces the output shaft so as to increase C, E will become negative, and the motor will apply torque to the load in a direction to decrease C. This torque is proportional to the displacement, and the result is similar to the effect of a spring. In a feedback device of the type described above, the motor will run as long as there is an error signal sufficient to overcome the load friction. Consequently, the amount of residual error in this type of servo depends only on the amount of friction in the load, and not upon the value of C.

10-8. In contrast to this type of behavior is a class of feedback control devices typified by the voltage regulator circuit (see figure 10-2). In this case, the controller is

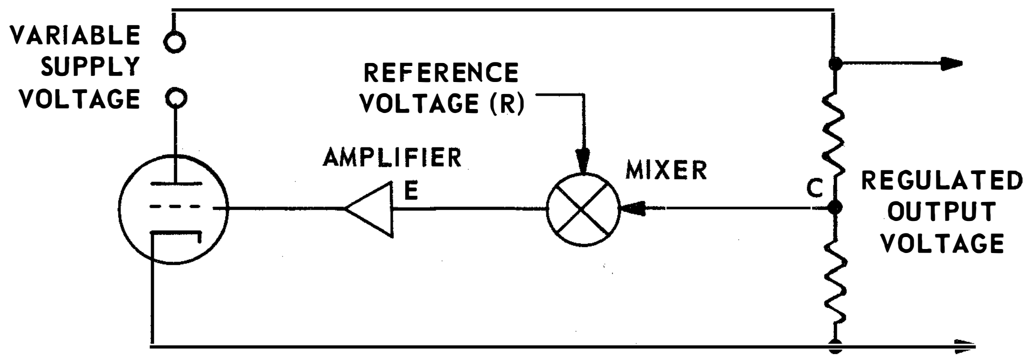

Figure 10-2. Voltage Regulator.

a tube whose plate-to-cathode resistance varies in proportion to its grid voltage. Feedback signal C, obtained from a voltage divider across the regulated output, is compared in a mixer circuit with a reference voltage obtained from a voltage reference tube. Resulting error signal E is amplified and applied to the grid of the controller. If the supply voltage suddenly drops, C will be reduced and E increased. The resulting change in the grid voltage of the controller will reduce its effective resistance and raise the output voltage. However, the output voltage cannot become quite as high as it was before the supply voltage dropped. If it did, E would return to its original value, and the resistance of the controller would become the same as it was before the decrease in the supply voltage. This condition could be corrected by replacing the controller with a motor-driven rheostat and connecting the motor to the amplifier. In this case, as soon as the supply voltage dropped, the resulting error voltage would be amplified and applied to the motor, causing it to drive the rheostat to reduce the resistance and increase the output voltage. The motor would continue to run until the error signal went to zero, at which time the output voltage would be up to its original value.

10-9. The accuracy with which a positioning servo can repeat shaft position R is in most cases limited by the amount of torque required to move the friction load. Figure 10-3 shows the torque-versus-error signal characteristic of a typical servo with the motor stalled. It shows that a certain error voltage must exist to produce enough torque to move the load shaft against the starting friction. The battery feeding the controlled variable potentiometer determines how many volts of error signal E will be produced per degree of displacement of the controlled variable shaft. Increasing this battery voltage will increase the stiffness of the system at the load shaft. Stiffness is defined as the reaction torque at the load shaft divided by displacement of the shaft. The greater the stiffness of the system, the smaller will be the residual error.

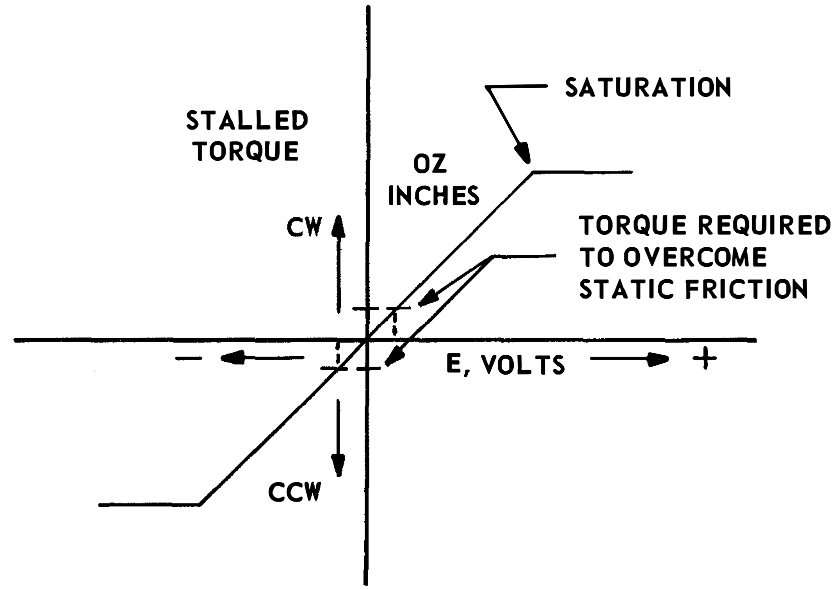

Figure 10-3. Idealized Torque vs. Voltage Curve.

10-10. If the controlled variable shaft is required to follow the reference shaft when it is moving at a constant speed, the motor must turn the output shaft at the same speed as the reference input shaft. Since a certain amount of power from the motor is required to overcome the running friction of the load, there must be a fixed difference between R and C sufficient to produce an error signal capable of driving the motor at the required speed. This difference between R and C indicates that the controlled variable shaft lags behind the reference input shaft by a certain number of degrees, although both are traveling at the same speed. This type of error, called a dynamic-tracking error, has a magnitude at any speed of reference input determined by the velocity constant of the servo. To obtain the velocity constant, the shaft of the controlled variable potentiometer is uncoupled from the gear train and displaced one degree from the null position, producing an error signal. The speed of the output shaft is measured. The ratio of this speed in degrees per second to the error in degrees required to produce it is the velocity constant of the servo system. In figure 10-1, increasing either the gain of the amplifier or the voltage of the battery across the controlled variable potentiometer will increase the velocity constant as well as the stiffness of the servo, so that increasing the overall gain of the system will decrease both the static friction error and the dynamic tracking error of the system.

10-11. Servo Stability Requirements.

10-12. A servo system, such as that shown in figure 10-1, may be represented by a block diagram such as figure 10-4. The box labeled KG represents the amplifier, the motor, and the gear train. K is the gain constant of the system, in this case the

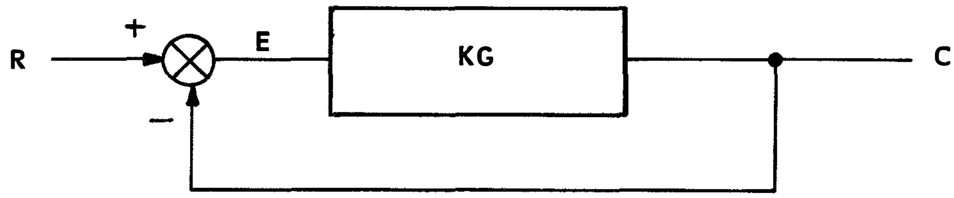

Figure 10-4. Servo Loop Block Diagram.

velocity constant of the entire loop. It includes the amplifier gain, the motor velocity constant, and the gain of the reference input and controlled variable potentiometers. The constant K is independent of the frequency of the applied signals, but G is an expression that describes the frequency response or time response of the amplifier and motor to error signals. From figure 10-4,

(1) $E = R - C$, and
(2) $C = KGE$

Solving (2) for E:

(3) $$E = \frac{C}{KG}$$

Substituting (3) at (1):

(4) $$\frac{C}{KG} = R - C$$

(5) $$C = RKG - CKG$$

(6) $$C(1 + KG) = RKG$$

Hence, the effective gain of the closed loop is:

(7) $$\frac{C}{R} = \frac{KG}{1 + KG}$$

Since the amplifier and motor must be built with physically realizable components, the function G represents a certain finite bandwidth. The inertia of the motor will tend to slow down its response to high-frequency error signals, and since a roll-off in frequency response is accompanied by a phase shift, there will be some frequency at which the controlled variable C will lag the error signal E by 180 degrees. At this frequency, the quantity KG becomes negative; if KG approaches -1, the denominator of equation 7 approaches 0 so that C/R approaches infinity. Physically, this can be interpreted to mean that an output C is obtained with no input R. This is the condition under which the loop will oscillate and is known as the Nyquist stability criterion. Because of the finite bandwidth of G, this condition for stability imposes a limitation on the value of K that may be used.

10-13. Now that the condition required for stability has been developed mathematically, the system shown in figure 10-1 may be examined to see what happens when a servomechanism is unstable. The higher the velocity constant, which includes amplifier gain and the voltage of the battery driving the controlled variable potentiometer, the greater will be the motor speed at any given value of error signal. Because the motor, gear train, and load possess inertia, the system in motion has kinetic energy equal to $J\omega^2$. In order to stop the motor, this energy must be dissipated. Because of inertia, the response of motor speed to a step of error signal voltage into the amplifier is as shown in figure 10-5. At the beginning of the step, a step of torque is applied to the rotor, producing acceleration. However, as the motor builds up speed, the friction in the load, motor bearings, and gear train dissipates an increasing amount of energy.

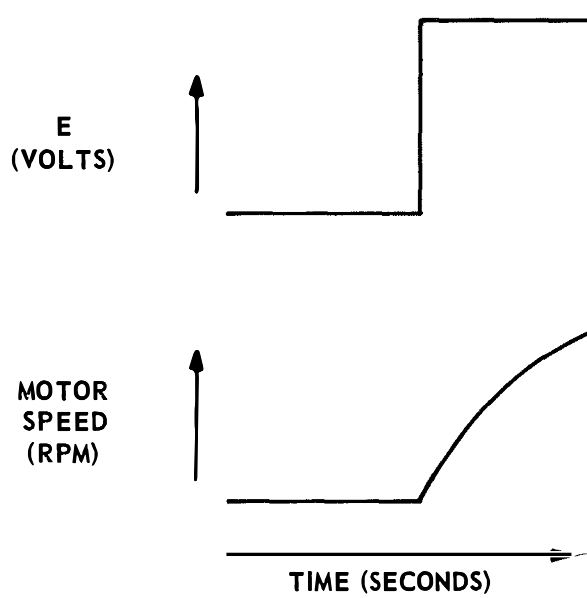

Figure 10-5. Response of Motor and Amplifier to Step Input.

Eventually the motor reaches a speed at which the amount of power supplied to the motor winding by the amplifier equals the total amount of power dissipated in the friction load, the motor and gear train bearing friction, and in the copper loss in the motor winding. At this point, the motor speed remains constant. If the system gain is low, the stiffness at the load will be quite small. As the output shaft approaches the null position, the motor torque drops off rapidly enough so that the friction can dissipate all the kinetic energy in the motor. Consequently, the response of the closed-loop system to a step input (10-6 a) will be as shown in figure 10-6 b. This condition of the servo loop is referred to as overdamped. If the gain is increased, increasing the stiffness, an oscillatory condition is reached in which the friction load cannot dissipate the inertial energy by the time the error signal gets to zero. In this case, the motor will overshoot the null position, and the position feedback signal will produce a reverse torque, causing the motor to overshoot in the opposite direction. This oscillation will continue with less energy imparted to the system on each oscillation, until a point is reached where the total system energy at the null is zero. This system is underdamped and has the response to an input step as shown in figure 10-6c. If the system gain is made sufficiently large, the stiffness is so great that the amplifier is able to add more kinetic energy to the motor in each successive cycle of oscillation than the friction load can dissipate. This condition results in divergent oscillations, as shown in figure 10-6d.

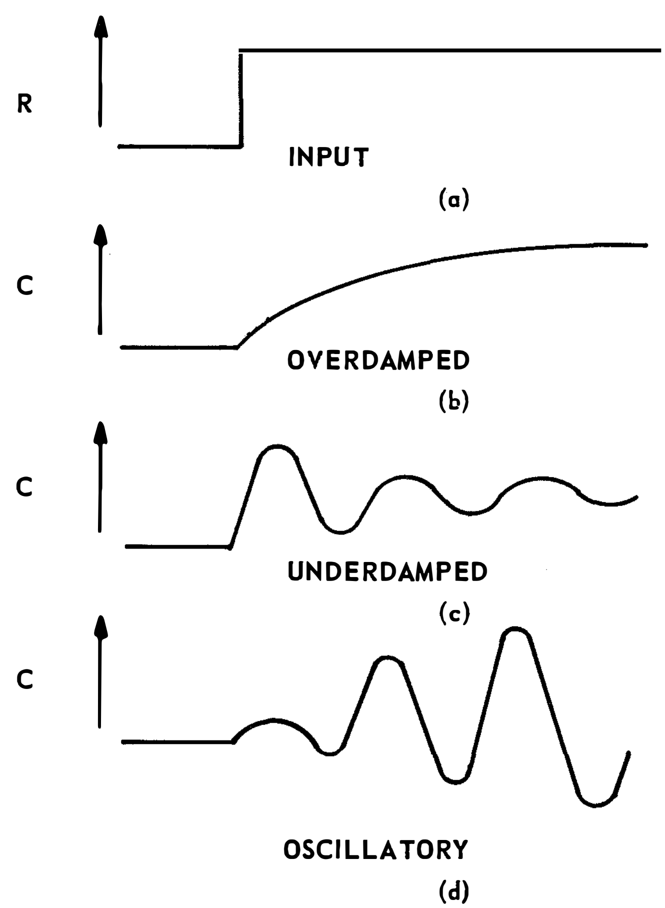

Figure 10-6. Response of Follow-Up Servo to Step Input.

10-14. Stabilizing Methods.

10-15. In many cases a servo may be satisfactorily damped by merely adding some sort of velocity-proportional friction device to absorb inertial energy. Automobile shock absorbers are an example of this type of damper, and small instrument servomotors are frequently equipped with a drag cup fastened to the motor shaft and turning in a fixed magnetic field. Such dampers produce a torque proportional to velocity. Friction dampers reduce the velocity constant of a system because the motor requires more voltage to run at a given speed. However, the damping effect allows the gain to be made up in the amplifier, so that for a given velocity constant a greater stiffness may be realized, and this reduces the static error in the system.

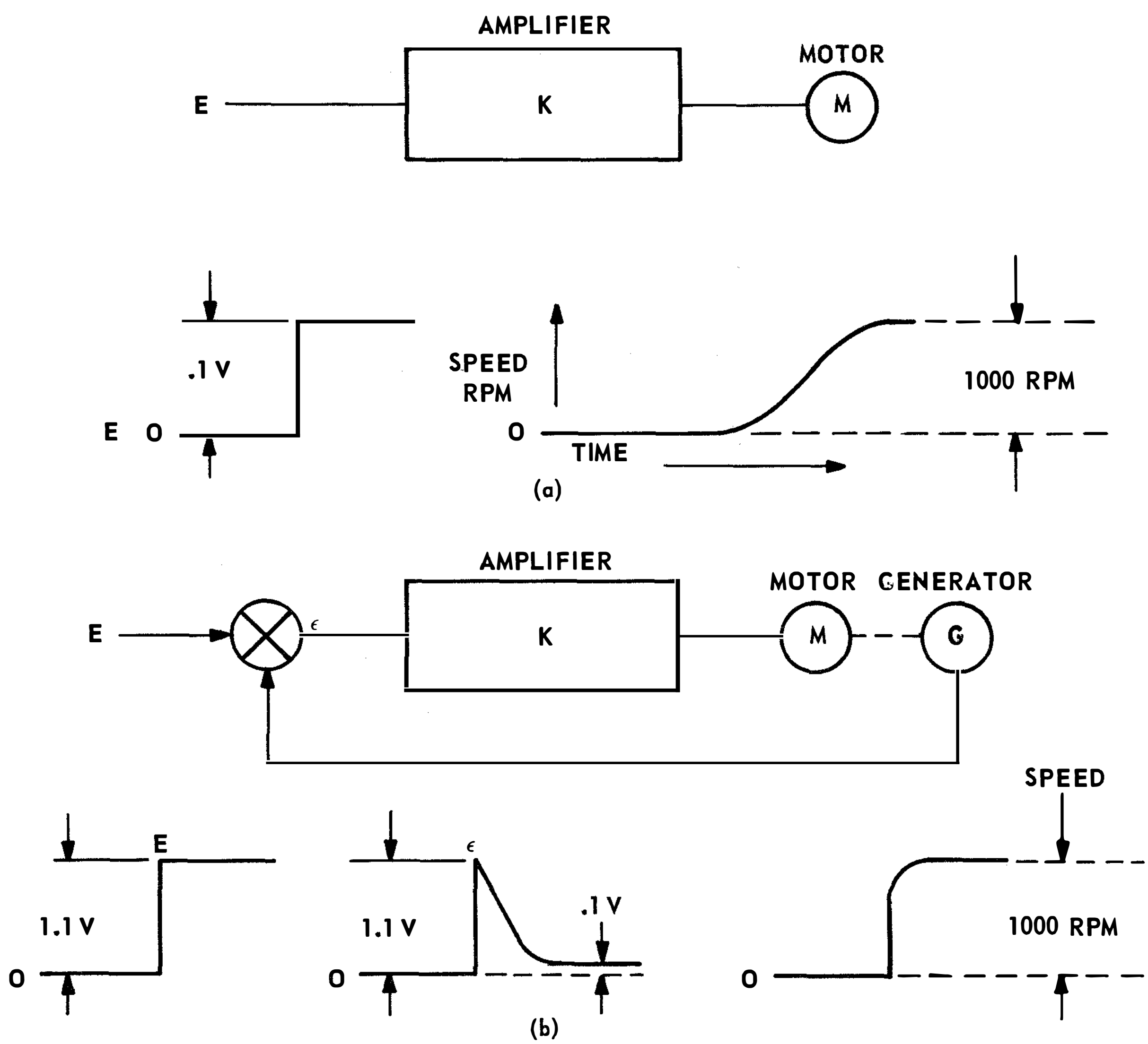

Figure 10-7. Effect of Rate Generator on Response Time.

10-16. Figure 10-7a shows the response of motor speed to a step of error voltage. For purposes of illustration, it is assumed that, in the steady-state condition, 0.1 volt is required to cause the motor to run at a speed of 1000 rpm. Because of the small error signal, very little torque is available to accelerate the motor initially; hence the rise time is excessive. A method of damping known as "rate feedback," in which a generator is coupled directly to the motor shaft, is shown in figure 10-7b. The output of the generator is a voltage directly proportional to the motor speed. If this voltage is fed back inversely to the amplifier, it results in a torque proportional to speed, similar to that obtained with a friction or drag cup damper. Since the subtracting is done at a low signal level, the amplifier is not required to supply any more power than when it is driving a motor with no load, the motor does not have to be so large, and there is no requirement for dissipating the motor's energy when it is running. The generator connected to the motor shaft puts out 1.0 volt per 1000 rpm. When excited with a step function, E, of 1.1 volt, error signal ϵ will be 1.1 volt at first, since the motor starts at zero speed and the initial generator output is zero. As the motor picks up speed the generator voltage, which is an exponential, will subtract from the 1.1-volt initial value. In the steady-state condition, when the motor reaches 1000 rpm, ϵ will be 1.1 volt minus the 1.0-volt generator output, or 0.1 volt. The voltage fed to the amplifier for the steady-state condition is the same as in figure 10-7a, but the 1.1-volt spike present in the amplifier input in figure 10-7b produces eleven times more acceleration torque at the motor, reducing the rise time. Because the amplifier gain, K, is the same in both cases and because the rate generator output is zero when the motor is stalled, the stalled torque for a given error signal is the same in both cases. Because of the much larger value of E required to obtain 1000 rpm with a rate generator, the velocity constant for this case will be 1/11 that of the motor alone. If, in figure 10-7b, the amplifier gain, K, were multiplied by 11, the two systems would have the same velocity constant but the system with the rate generator would have eleven times the stiffness (the positioning accuracy would be eleven times as great).

10-17. Another method of stabilizing a servo and allowing an increase in its stiffness is to use a lead network. Figure 10-8 shows a lead network and its transient response to a step input. If 1 volt is suddenly applied as E_{IN}, the entire 1 volt will appear as E_{OUT} because the voltage across C cannot be changed instantaneously. As the capacitor charges up. E_{OUT} will drop exponentially and approach the value it would have if C were not present, which is $E \dfrac{R_2}{R_1 + R_2}$. This transient response is similar to that of ϵ in figure 10-7b. Figure 10-9 shows how a lead network is used to accomplish a result similar to that obtained with a rate generator. If $\dfrac{R_2}{R_1 + R_2}$ is made equal to 1/11, a voltage of 1.1 volts applied at E will produce the wave form shown at ϵ, which will cause rapid acceleration of the motor to a speed of 1000 rpm. The capacitor in the lead network must be chosen to produce the same time constant in the signal at ϵ

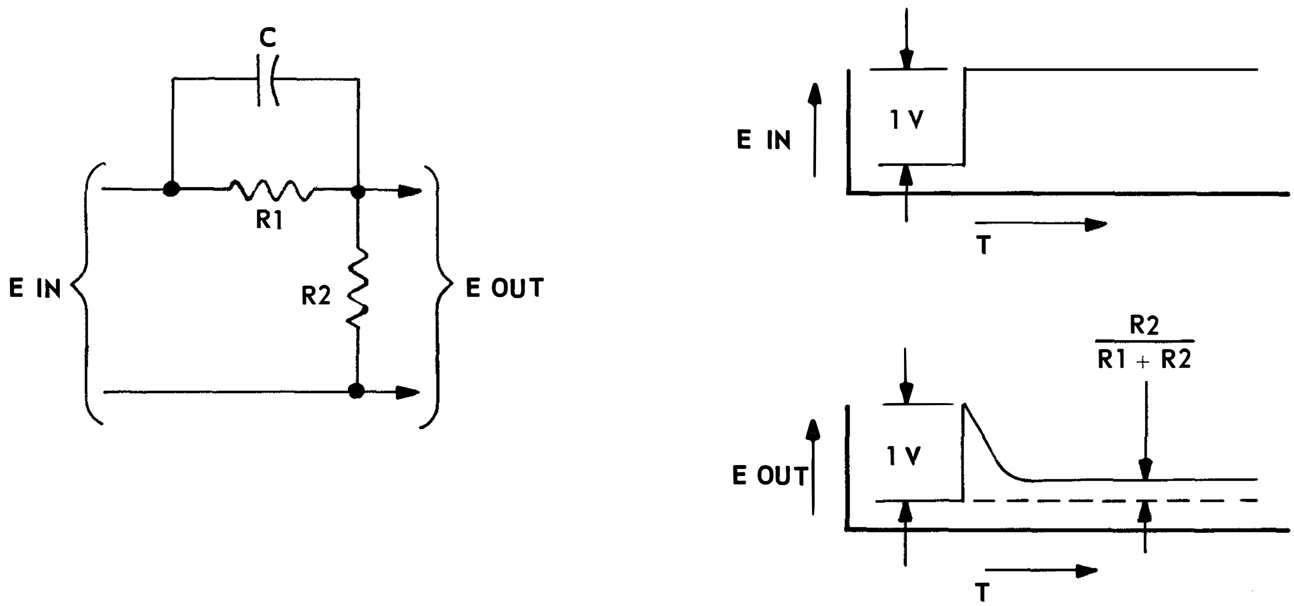

Figure 10-8. Lead Network.

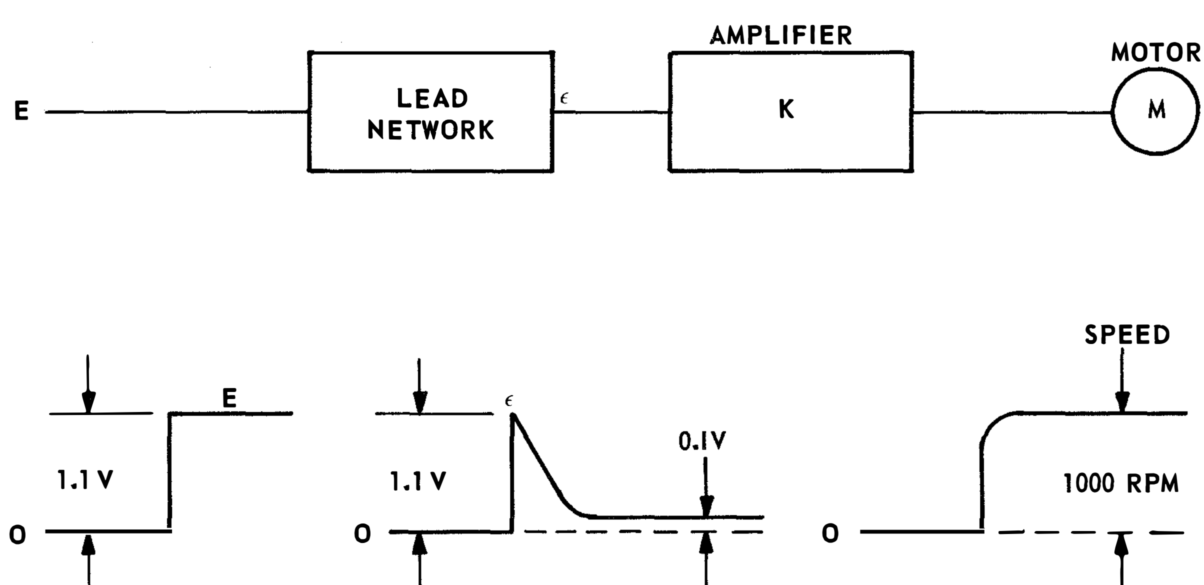

Figure 10-9. Effect of Lead Network on Response Time.

as that of figure 10-7b. The transfer function of the lead network of figure 10-8 may be derived by considering it a voltage divider with the parallel combination of R_1 and C in the top leg and R_2 in the bottom leg. Thus the output impedance is

$$Z_0 = R_2$$

The parallel impedance of R_1C is

$$\frac{R_1 \frac{1}{pC}}{R_1 + \frac{1}{pC}} = \frac{R_1}{1 + pR_1C}$$

and the total series impedance of the divider is

$$Z_t = R_2 + \frac{R_1}{1 + pR_1C}$$

Hence

$$\frac{E_0}{E_i} = \frac{Z_0}{Z_t} = \frac{R_2}{R_2 + \frac{R_1}{1 + pR_1C}} = \frac{R_2(1 + pR_1C)}{R_1 + R_2 + pR_1R_2C}$$

$$= \frac{R_2}{R_1 + R_2} \quad \frac{1 + pR_1C}{1 + p\frac{R_1R_2}{R_1 + R_2}C}$$

This expression may be rewritten as follows:

$$\frac{E_0}{E_i} = \frac{R_2}{R_1 + R_2} + \frac{R_2}{R_1 + R_2} \frac{\left(R_1 - \frac{R_1R_2}{R_1 + R_2}\right)Cp}{1 + \frac{R_1R_2}{R_1 + R_2}Cp}$$

$$\frac{E_0}{E_i} = \frac{R_2}{R_1 + R_2} + \frac{R_1^2 R_2 C}{\left(R_1 + R_2\right)^2} \frac{p}{1 + \frac{R_1R_2}{R_1 + R_2}p}$$

$$\frac{E_0}{E_i} = K_1 + K_2 \frac{Tp}{1 + Tp}$$

Therefore

$$E_0 = K_1 E_i + K_2 \frac{Tp}{1 + Tp} E_i$$

Thus, the lead network behaves like a straight feed, with a gain of K_1 plus a high-pass filter with a gain of K_2. The network differentiates low-frequency signals. When connected in a closed-position loop, as shown in figure 10-10, a lead network provides the sum of a position signal and a differentiated-position signal, so that the rate of change of ϵ is used, producing an effect similar to that of a rate generator, except that R is differentiated as well as C.

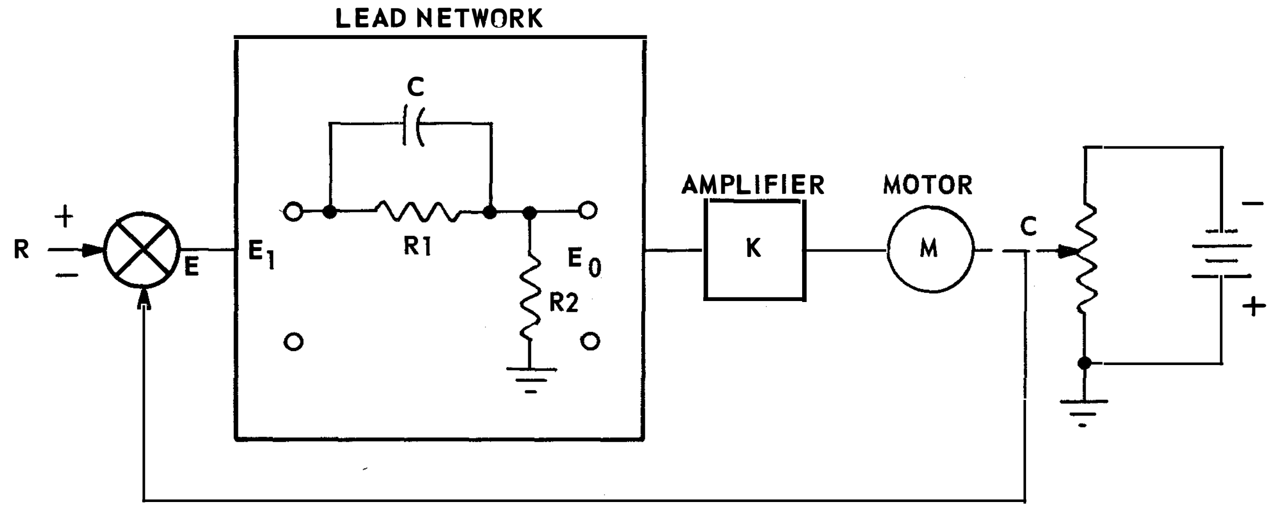

Figure 10-10. Position Follow-up Servo with Lead Network.

10-18. <u>Components and Circuits.</u>

10-19. In the preceding discussion of the position follow-up servomechanism of figure 10-1, it was assumed that the reference input and controlled variable voltages, the amplifier, and the motor were all direct current components. In practice, 400- or 60-cycle carrier systems are more commonly used for small, low-power servo systems in radio communications equipment. The use of an AC carrier system simplifies the design of the amplifier. Since an amplifier bandwidth of 40 or 50 cycles usually suffices, the DC operating point of the individual stages is of no consequence. Because of the low-frequency requirement, the junction transistor is well suited to servo work. Where 20 watts or more of amplifier output is required, a magnetic amplifier or saturable reactor driven by a transistor preamplifier may be used. Some servo amplifiers employ high-performance, magnetic amplifiers for all stages.

10-20. The most commonly used type of servomotor is the two-phase induction motor, and frequently a two-phase induction generator is built into the same case and coupled to the motor shaft. The schematic diagram for a typical motor generator appears in figure 10-11. The reference phase of the motor must be driven with a voltage 90° out of phase with the voltage on the control phase in order to obtain torque. If the control phase voltage leads the reference phase voltage, the motor will turn in one direction; and, if the control voltage lags the reference voltage, the motor will turn in the opposite direction. Thus, if the error signal source is a 400-cps signal, a phase reversal of the error signal produces a reversal of motor rotation, as required.

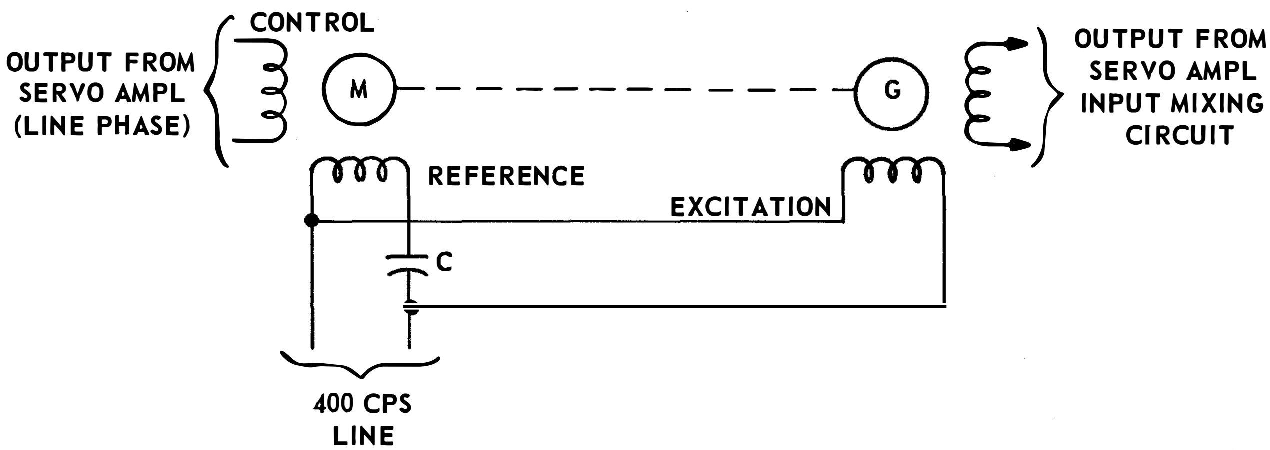

Figure 10-11. Motor-Generator Schematic Diagram.

10-21. In some cases, where the servo amplifier output is either in phase or 180 degrees out of phase with the line voltage, depending on the sense, the required quadrature relationship between control and reference winding is obtained by means of a phase-shift capacitor C, producing a quadrature voltage on the reference winding. Sometimes, however, it is more convenient to produce the required 90° phase shift inside the servo amplifier; in which case, the reference winding is connected directly to the 400-cps line. When its excitation winding is driven from the 400-cps line, the rate generator produces across its output terminals a voltage proportional to the speed of rotation and either in phase or out of phase with the excitation voltage, depending upon the direction of rotation.

10-22. If the purpose of the servomechanism is merely to repeat the position of a shaft or to produce an output shaft position proportional to a voltage used as a reference input, AC line voltages may be used across reference and controlled variable potentiometers in place of the DC voltages shown in figure 10-11. In this case, as the controlled variable voltage increases and becomes larger than the reference input, the phase of the error-signal voltage reverses and the direction of rotation of the two-phase servomotor is reversed. In this way, a carrier servo system may be constructed in which all variables are represented by 400-cps AC voltages.

10-23. In some transmitter tuning servos and antenna matching networks, it is desired to have the servomotor and gear train position a mechanical tuning element, such as a variable capacitor, so that the phase shift imposed upon an RF signal by the tuned circuit is zero. An RF phase discriminator circuit of the type used for detection of FM signals may be used to obtain a DC voltage proportional to the magnitude of the phase shift through the RF circuit, and of polarity determined by whether the output leads or lags the input. If this DC error information is to be fed into a carrier-type servomechanism, it may be converted to AC by means of an electromechanical chopper connected as in figure 10-12. The chopper consists of a vibrating reed and a pair of contacts which form a single-pole double-throw switch. The reed is excited by a magnetic coil which is driven by the 400-cps line. In most cases, the action of the reed contact is not in phase with the excitation voltage fed to the coil, so that a phase-shift network must be used on the coil to make the reed contact action either in phase with or in quadrature with the line, as required. In the chopper output waveform shown in figure 10-12, the peak-to-peak voltage of the square wave is the amplitude of the DC voltage connected across the contacts.

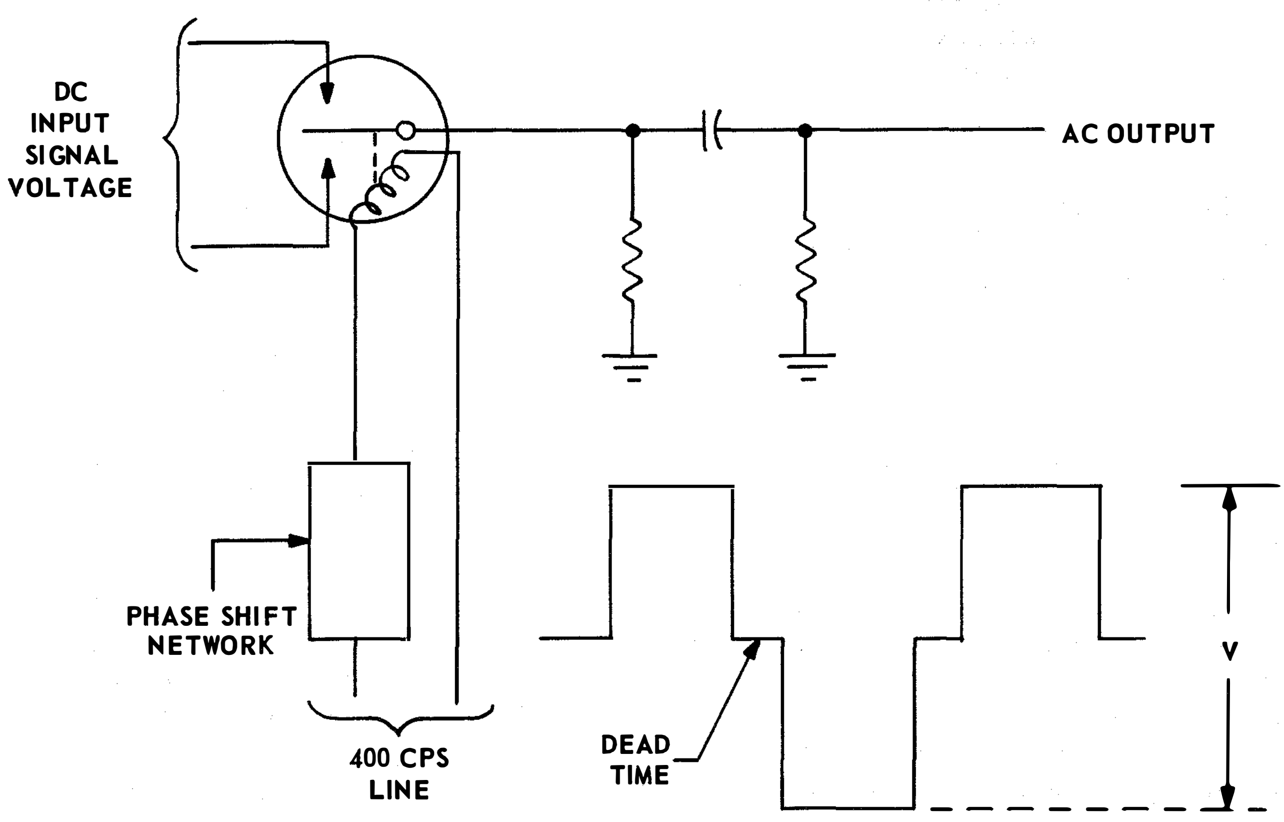

Figure 10-12. Electromechanical Chopper Circuit.

10-24. Another type of position transmitter frequently encountered is the synchro. Figure 10-13 illustrates a typical synchro error circuit. The synchro transmitter may be thought of as being a transformer with a single primary, a rotor winding, and three secondaries (the three stator windings). The stator windings are placed with their axes 120° apart. The rotor winding induces an AC voltage in each of the stator windings, proportional to the cosine of the angle between the rotor winding axis and the respective stator winding axis. The stator windings of the control transformer are connected directly across the stator windings of the transmitter. Therefore, the same voltages will exist across each of the control-transformer stator windings as are induced in the corresponding stator windings of the transmitter and the field pattern set up inside the core of the control transformer will be a replica of the field pattern in the transmitter. If the control transformer rotor winding axis is lined up with this field, the error-signal voltage developed across the rotor terminals will be maximum. As the control transformer rotor is turned, the error-signal voltage will decrease and become zero when the rotor axis makes an angle of 90° with the field pattern set up by the stator windings. Further rotation of the rotor will produce an increasing error voltage of the reversed phase. Thus, if the rotor of the transmitter is actuated by the reference input shaft of a position follow-up system (figure 10-1) and the control transformer rotor is coupled to the controlled variable shaft, the voltage developed across the rotor of the control transformer may be used as an error signal, E, to be fed into the servo amplifier, and the rotor of the control transformer will follow the transmitter rotor.

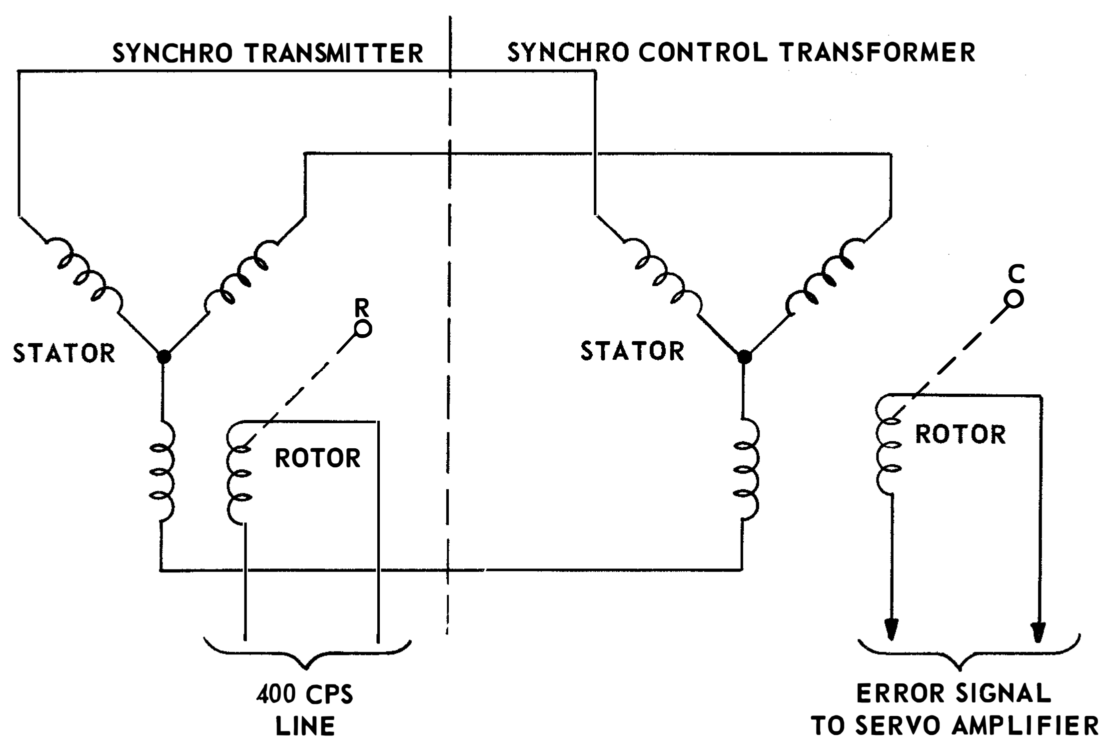

Figure 10-13. Synchro Error Circuit.

10-25. Frequency Setting of Stabilized Master Oscillator.

10-26. As an example of a servomechanism used in a single sideband exciter, let us consider the problem of setting the frequency of a stabilized master oscillator. Assume that you have a transmitter that has four output frequency bands as follows:

Band A - - 1.7 to 3.699 Mc, in 1 Kc steps, for a total of 2000 channels.

Band B - - 3.7 to 7.699 Mc, in 1 Kc steps, for a total of 4000 channels.

Band C - - 7.7 to 15.699 Mc, in 1 Kc steps, for a total of 8000 channels.

Band D - - 15.7 to 31.7 Mc, in 1 Kc steps, for a total of 16,000 channels.

Thus it can be seen that the master oscillator is required to supply a different number of frequency increments for each band. The master oscillator may be positioned servomechanically to 4000 increments of 0.5-Kc each. The 4000 frequency increments are selected by two remote-control dials called coarse and fine. The coarse selector has 160 selections or positions. The fine selector divides each coarse increment into 25 equal frequency ranges of 0.5-Kc. Positioning of the master oscillator is accomplished automatically when a frequency selection is made by setting the dials on the remote-control panel.

10-27. To assist in explaining operation of the modified two-speed servomechanism, discussion of a one-speed servomechanism is given. Figure 10-14 illustrates a basic one-speed servomechanism. Potentiometers R1 and R2 form a bridge circuit in which transformer T1 supplies excitation of 10-volts AC. The bridge output is taken across the potentiometer arms, and a servo amplifier serves as the error detecting device. In the condition shown in figure 10-14A, the potentiometer arms of R1 and R2 are positioned to points B. In this condition, the 10-volt AC excitation is applied to both R1 and R3, and voltage division occurs that causes 5 volts AC to appear between points B and C of each potentiometer. Therefore, no potential difference is present across the bridge, and no error is detected by the servo amplifier.

10-28. In figure 10-14B, the arm of R1 has been moved to a midrange position between points B and C. For the purpose of explanation, it is assumed that the voltage division across R1 now is 7.5 volts from point A to the potentiometer arm and 2.5 volts from point C to the potentiometer arm. Before correction, the arm of R2 remains at point B and the voltage division remains at 5 volts across each half of R2, as for figure 10-14A. Since the voltage from the potentiometer arm of R1 to point C is 2.5 volts and from the potentiometer arm of R2 to point C is 5 volts, a net error of 2.5 volts is detected by the servo amplifier. The servo amplifier converts the 2.5-volt error to servomotor power and applies it to the servomotor. This voltage will be either in phase or 180° out of phase with the R2 potentiometer arm, depending on the direction R1 is moved. This phase relationship will determine the direction of rotation of the

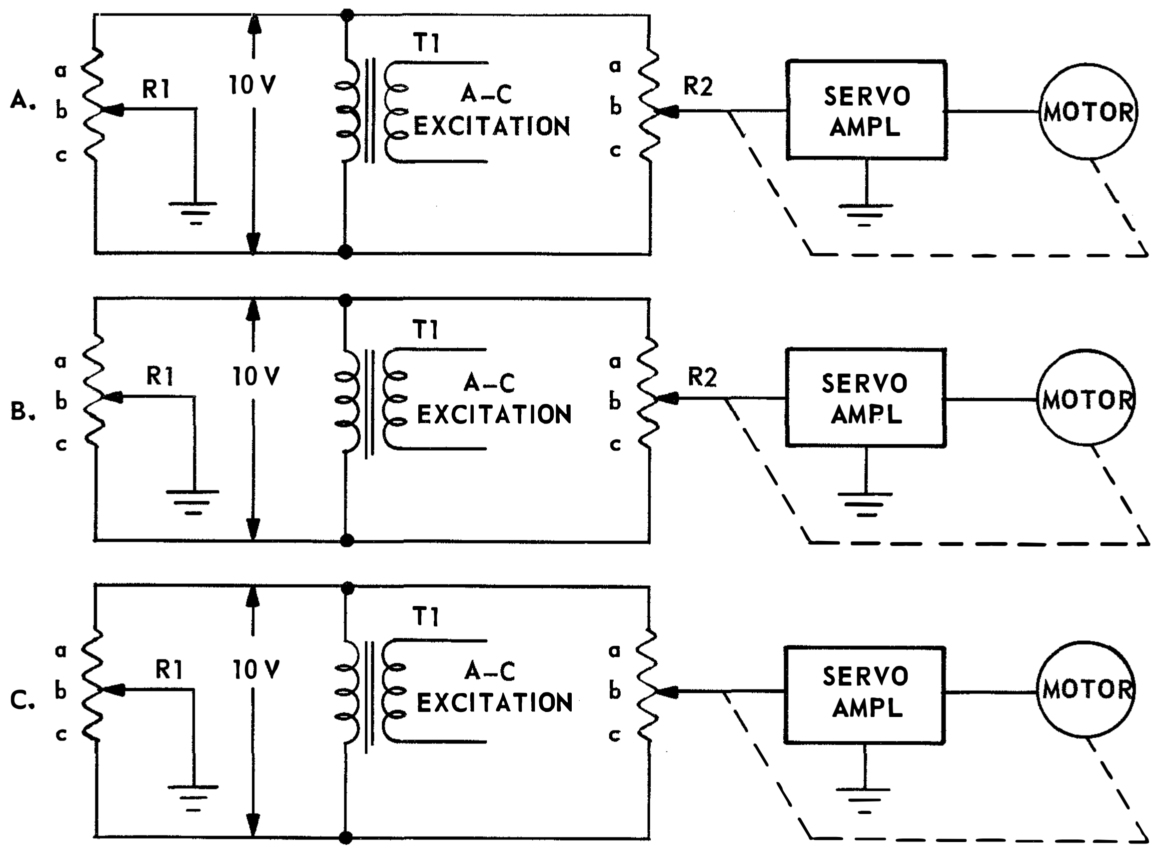

Figure 10-14. Simplified Schematic Diagram, One-Speed Servomechanism.

servomotor. The servomotor operates and drives the arm of R2 to a midrange position between points B and C, corresponding to the setting of R1. At this position, shown in figure 10-14C, the voltage potential across the bridge is reduced to a null, and the voltage between R2 potentiometer arm and point C has been adjusted to 2.5 volts. If the motor had attempted to drive the arm of R2 toward point A instead of point C, the bridge error would have increased to more than the original error. In the manner just described, a remote-control dial is used as the information source, corresponding to R1, to synthesize appropriate voltages which cause operation of a follow-up servomechanism corresponding to R2, the servo amplifier, and the motor. In turn, the servomechanism positions its associated circuits to a corresponding frequency setting.

10-29. A modified two-speed servomechanism is illustrated in figure 10-15. The frequency selector shafts are mechanically independent of each other, but the follow-up potentiometers associated with each are geared together in 16 to 1 ratio. The 25-step, continuous, follow-up system completes its range of each increment of the 160-step

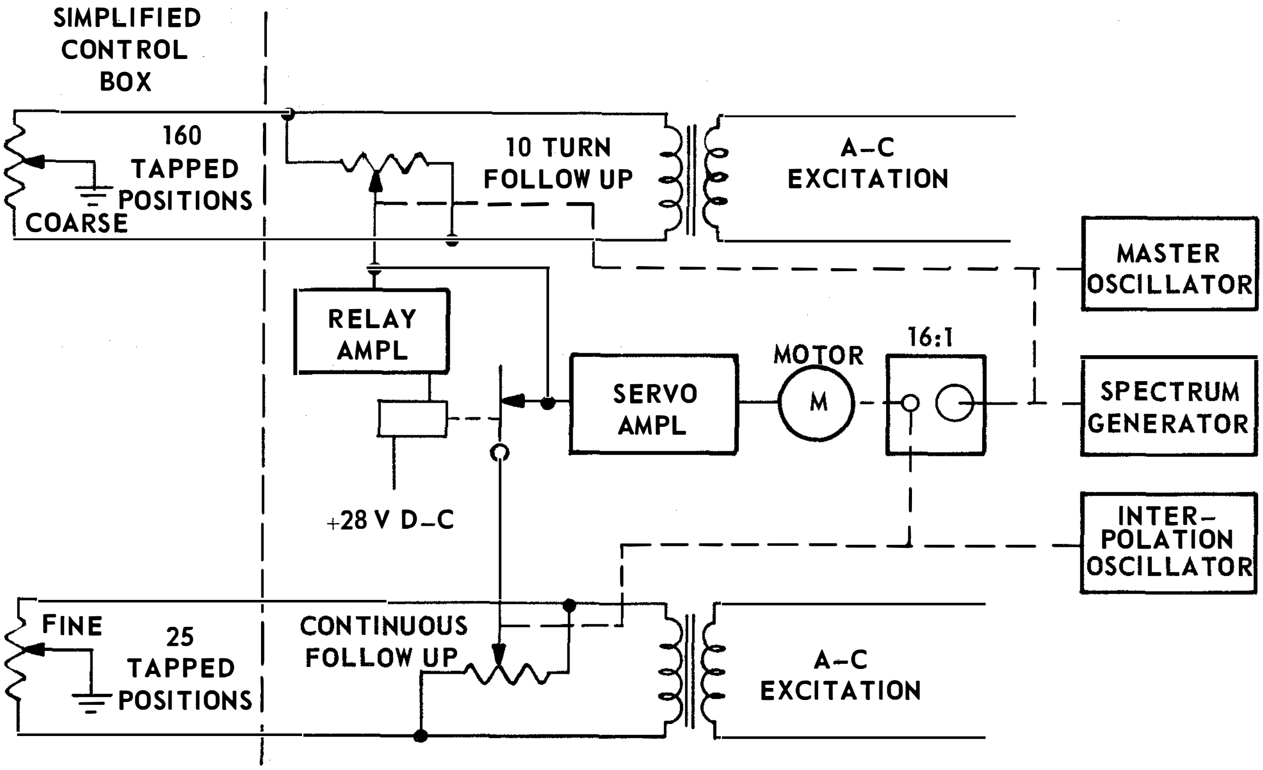

Figure 10-15. Schematic Diagram, Two-Speed Servomechanism Driving a Master Oscillator.

system. Therefore, the 160-step system is the coarse positioning system, which serves to eliminate ambiguity; while the 25-step system is the fine positioning system. The remote-control box is illustrated as a tapped information source only for explanation purposes and actually is a system for switching combinations of series and parallel resistors across the coarse and fine system control lines. The coarse system contains provisions for 160 discrete increments, and the fine system contains provisions for 25 discrete increments. The total capability is the product of the two, or 4000 increments. In the coarse follow-up system, a 10-turn potentiometer is used to resolve the 160 increments of the coarse positioning system. In the fine follow-up system, a continuous potentiometer is used to resolve the 25 increments of the fine system within each increment of the coarse system. The total capability of the follow-up system then corresponds to that of the information source. A relay circuit, operated by coarse-system errors, is used to remove the fine error from the servo positioning system until the coarse system has resolved the coarse error. When coarse positioning is completed, the fine error is used to select the proper increment of the fine system, within the selected coarse increments. The servo positioning system also causes proper setup of the injection signals for the triple-conversion circuits of a master oscillator stabilizing loop. The electrical stabilizing loop operates to provide correction of master oscillator frequency drift and serves to insert additional increments of master oscillator frequency, when necessary. The servomechanism just described is responsible for mechanical positioning of the master oscillator to obtain any of the 4000 increments of 0.5-Kc intervals within the frequency range of 2.0 to 4.0 Mc. Additional increments of 0.125 and 0.250 Kc, available from the master oscillator, are provided by the electrical stabilizing loop when necessary.

SECTION XI

LINEAR RF POWER AMPLIFIERS

11-1. Introduction.

11-2. The RF power amplifier of the SSB transmitter receives a low power-level, radio-frequency SSB signal from the exciter. The function of the power amplifier is to raise the power level of the input signal without changing the signal. That is, the envelope of the output signal must be a replica of the envelope of the input signal. A power amplifier which will perform this function is, by definition, a linear power amplifier.

11-3. Power Amplifier Classification.

11-4. Radio-frequency amplifiers are classified A, B, and C, according to the angle of plate current flow; that is, the number of degrees of plate current flow during a 360° RF cycle. Class A amplifiers have a continuous plate current flow and operate over a small portion of the plate current range of the tube, as shown in figure 11-1. This class of amplifier is used for amplification of small signals for low distortion. Its efficiency in converting DC plate power input into RF power output is quite low, usually less than 35 percent, but this is seldom of major importance where small signals are amplified.

11-5. Class B amplifiers have their grids biased to near plate current cutoff, so that plate current flows for approximately 180° of the RF cycle, as shown in figure 11-2. Amplifiers operated with appreciably more than 180° of plate current flow, but less than 360°, are called class AB amplifiers. High-power stages of RF linear amplifiers may be operated either class AB or class B to achieve maximum output power with low distortion. Plate efficiency depends upon the tube used and the operating conditions selected, with efficiencies in the range of 50 to 70 percent obtainable. The distinction between class B and class AB is somewhat arbitrary, since both operate over more than 180° but less than 360°. However, the class AB amplifier draws appreciably more static plate current than the class B amplifier, which draws only a small static plate current.

11-6. The class C amplifier is biased well beyond cutoff, so that plate current flows less than 180° of the RF cycle as shown in figure 11-3. The principle advantage of the

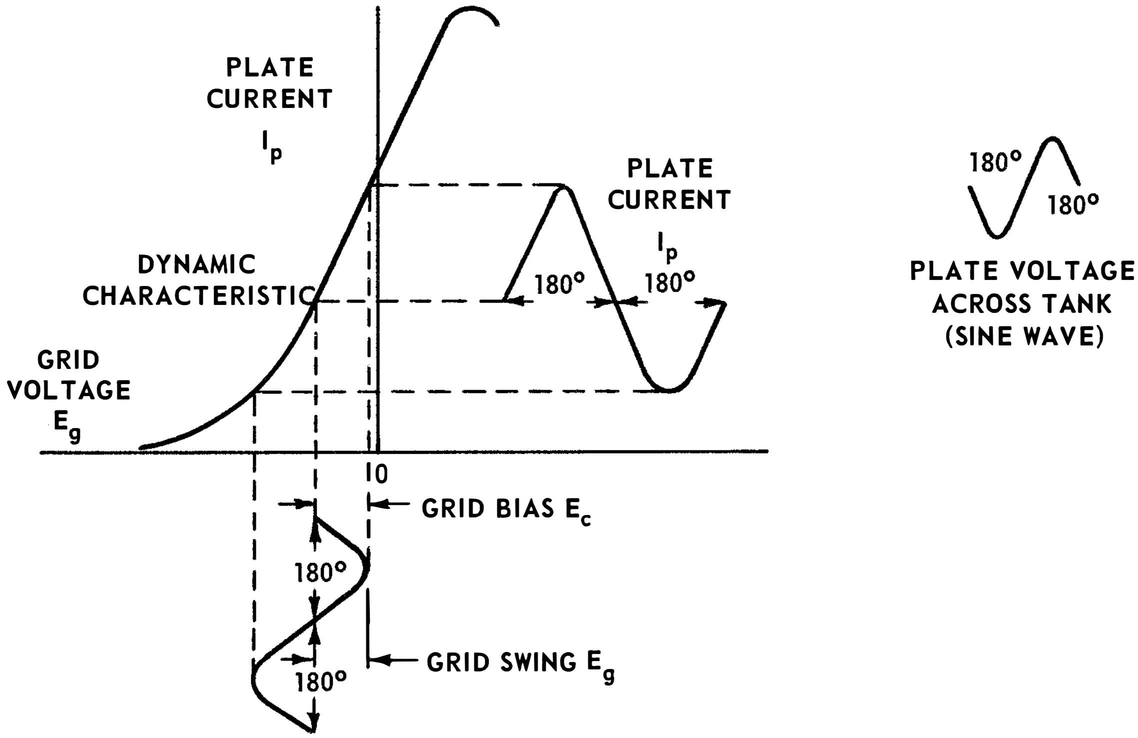

Figure 11-1. Class A Tube Operation.

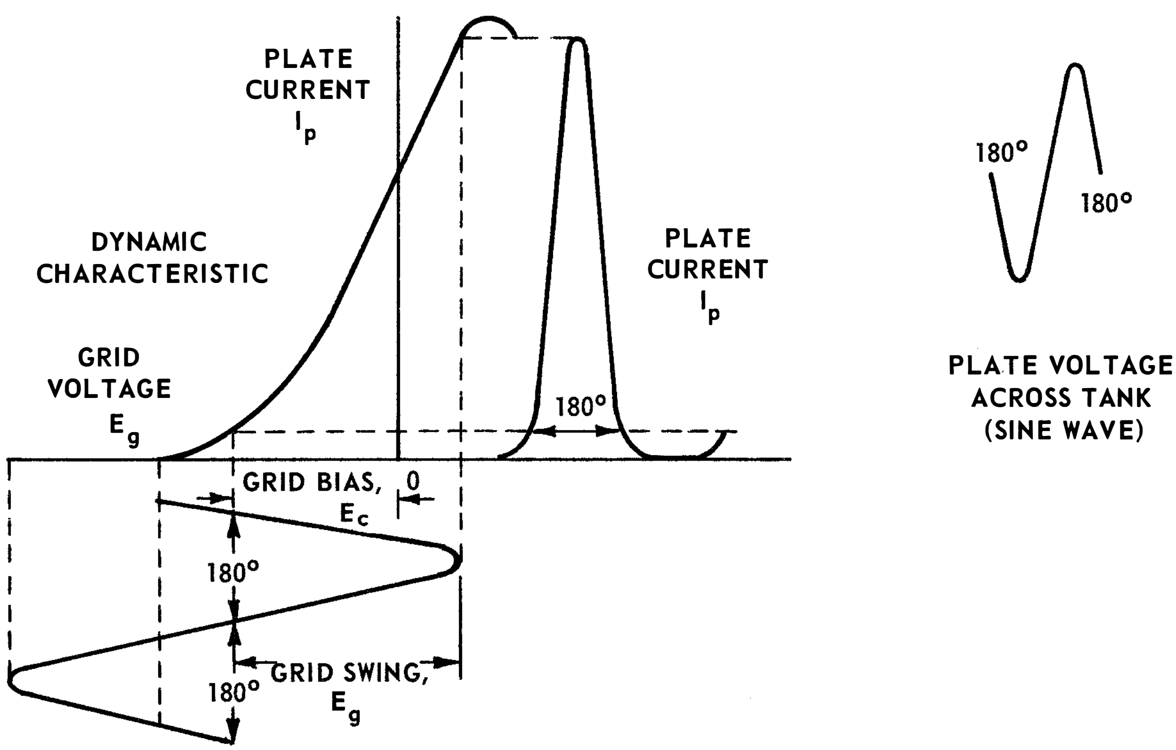

Figure 11-2. Class B Tube Operation.

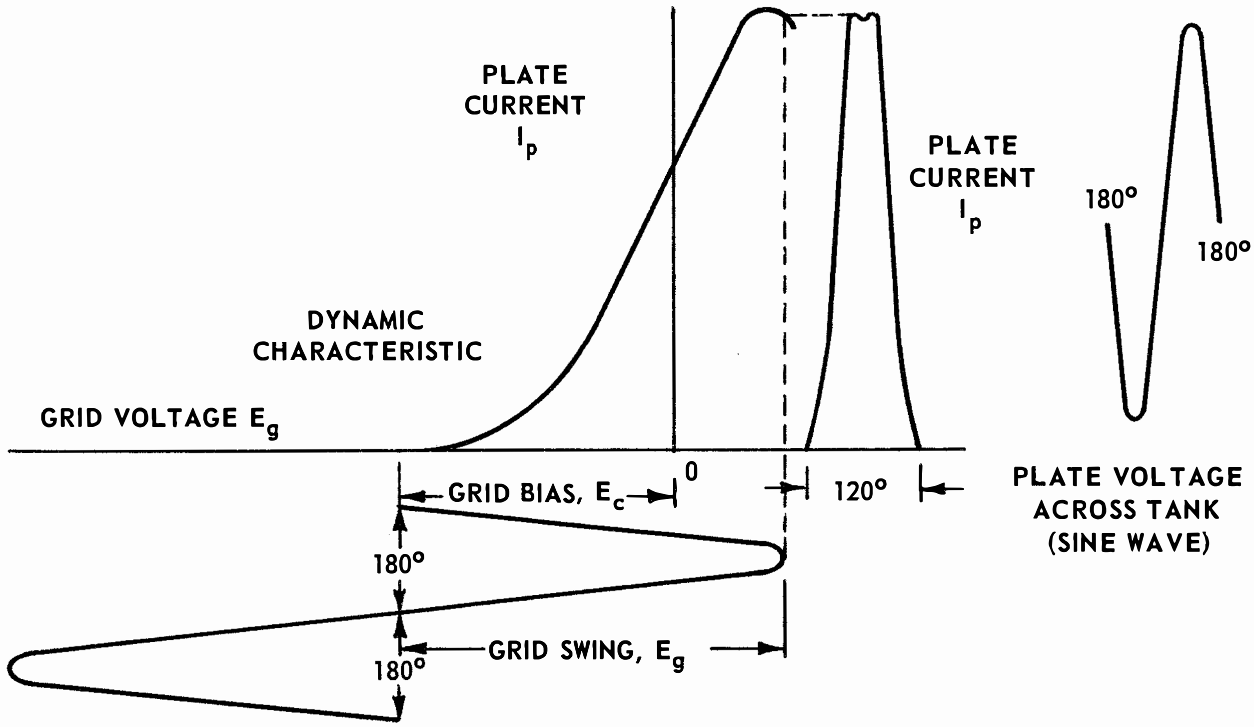

Figure 11-3. Class C Tube Operation.

class C amplifier is high plate efficiency, from 65 to 85 percent. Class C amplifiers are not suited for SSB use because they are not linear amplifiers and will not respond to low-level input signals.

11-7. A subscript number is commonly added to the amplifier class designator to indicate whether or not the tube is operated in the positive grid region over part of the cycle. For example, class AB_1 indicates that the grid never goes positive, so that no grid current is drawn. Class AB_2 indicates that the grid does go positive, so that grid current is drawn. Because class A amplifiers are always operated without grid current, and class C amplifiers are always operated with grid current, subscript designators are omitted.

11-8. RF Power Amplifier Tubes.

11-9. Conventional grid-controlled power amplifier tubes are classified according to the number of elements contained within the envelope. At one time, the triode, which has a control grid in addition to the filament and plate, was the only available transmitting tube in the medium- and high-power range. Tetrode tubes include a screen-grid placed between the control grid and the plate. The screen-grid furnishes an accelerating potential for the electron stream and provides an electrostatic shield between the control grid and the plate. The two grids of most transmitting-type tetrodes have a "beaming" influence on the electron stream which improves the tube characteristics. This beaming action tends to reduce the DC screen current and increase the effect of the control grid. Power pentodes have an additional grid, called the "suppressor grid," located between the screen grid and the plate. In some "beam-power" tubes, this element may consist of beam-forming plates which, in general, give an improved plate characteristic when the plate voltage swings below, or in the region of, the DC screen voltage.

11-10. Triode power amplifier tubes have the advantage of simplicity, low cost, and availability in all sizes. In general, they require a large amount of driving power. Also, since the grid is exposed directly to the plate, there may be considerable capacitive coupling between plate and grid within the tube. This plate-to-grid capacitance must be carefully and completely neutralized in RF amplifiers to prevent self-oscillation. The amplification factor of triode tubes is on the order of 5 for low-μ, 20 for medium-μ, and 50 for high-μ tubes. Generally, only low- and medium-μ triodes can be used in linear power amplifier circuits; therefore, a large grid-voltage swing is required to drive the amplifier to rated power output.

11-11. In tetrode power amplifier tubes, the screen grid acts as an electrostatic shield between the plate and the control grid, reducing the plate-to-grid capacitance. This can reduce the required neutralization capacity by a factor of one hundred. However, since the gain of the tetrode tube is so much higher than that of the triode, neutralization of the small residual plate-to-grid capacitance is still required for best, high-gain linear amplifier performance. Because of the high gain of the tetrode, the tube requires relatively low drive to obtain rated power output, and advantage which allows fewer amplifier stages to be used to obtain a given output.

11-12. Pentode construction is used in most small, receiving-type amplifier tubes and, in some cases, in transmitting tubes rated up to 1 kw output. Receiving-type pentodes provide good performance with low plate voltage; in larger tubes, pentodes give improved efficiency because the RF plate voltage swing can be increased somewhat. Disadvantages of the pentode are that it is more complex than the tetrode, is more expensive, and requires extra circuitry for the suppressor grid. These disadvantages have limited the development of the pentode as a high-power transmitting tube. At present, the pentode provides little advantage over a well-designed tetrode.

11-13. For linear RF power amplifier use, desirable amplifier tube features are:

a. high gain,

b. low plate-to-grid capacitance,

c. good efficiency,

d. linear characteristics which are maintained, without degradation, at all frequencies in the desired operating range.

11-14. The need for power amplifier tubes in the VHF and UHF ranges has spurred development of tubes suitable for efficient operation at those frequencies. This has resulted in tubes with better performance in the HF range (3 to 30 Mc). A comparison can be made between the older type 813 tube and the newer type 4X250B tube, both being in the same power class. The small, compact design of the 4X250B results in short lead lengths, better screening, closer element spacing, and improved performance, easily maintained over the HF frequency range. The ceramic construction (rather than glass) of an increasing number of new tubes promises to result in a more rugged and longer-lived tube. Ceramic-sealed tubes now available include the RCA 6118, which is smaller than the 4X150A; the Eimac 4CX300A, with characteristics similar to the 4X250B; an all-ceramic version of the 4X250B; the Eimac 4CX5000A, capable of 10 kw of RF output; and an RCA super-power, shielded-grid tube that will deliver 500 kw of RF output. Tube manufacturers are developing additional types of power amplifier tubes which promise still better performance in new equipments.

11-15. At the low signal levels encountered in SSB exciters, conventional receiving-type RF amplifier tubes are used. The type 6CL6, a miniature 9-pin tube, is commonly used in exciters for delivering about 0.1 w output, and is frequently used to excite the type 4X250B power amplifier tube. The 4X250B is used in small, compact equipment to deliver 500 w by paralleling two tubes, and to deliver 1 kw by paralleling four. The type 4CX5000A provides an output power level of 5 to 10 kw; by paralleling four of these tubes, power output up to 45 kw may be obtained.

11-16. Basic Linear Power Amplifier Circuits.

11-17. General.

11-18. For linear operation, RF power amplifiers may be operated class A or class AB. The amplifiers used are quite conventional, being either grid driven or cathode driven (grounded grid) amplifiers. However, the design considerations are extremely stringent to produce maximum linearity for a given tube in a given circuit. The tube operating point must be carefully chosen and precisely maintained, neutralization must be as effective as possible, and input and output impedances must be held as constant as possible. RF feedback circuits are often used to improve neutralization and linearity. Generally, class A pentode power amplifiers are employed in low-level power stages to preserve linearity in these stages while producing enough power to drive the higher level stages. Class AB_1 or AB_2 triode or tetrode power amplifiers are employed in the high-level power stages to obtain the desired power output.

11-19. Grid Driven Triode Power Amplifier.

11-20. Figure 11-4 is a simplified schematic of a typical grid-driven triode power amplifier. This amplifier, operating class AB_1, produces up to 2.5 kw, using the type 3X3000A-1 triode. The triode tube, having a large plate-to-grid interelectrode capacitance, always requires neutralization to prevent oscillation when used in the grid-driven circuit. The only types of triodes capable of class AB_1 operation are the low amplification factor types, such as the 3X3000A-1. Due to the low amplification factor, very high RF grid excitation voltage is required, on the order of 1000 v for the 3X3000A-1. A similar tube, suitable for class AB_2 operation, is the 3X2500A-3, which has an amplification factor of 20. This medium-μ triode requires less grid swing, but it requires grid driving power for class AB_2 operation. Neutralization, of course, is still required.

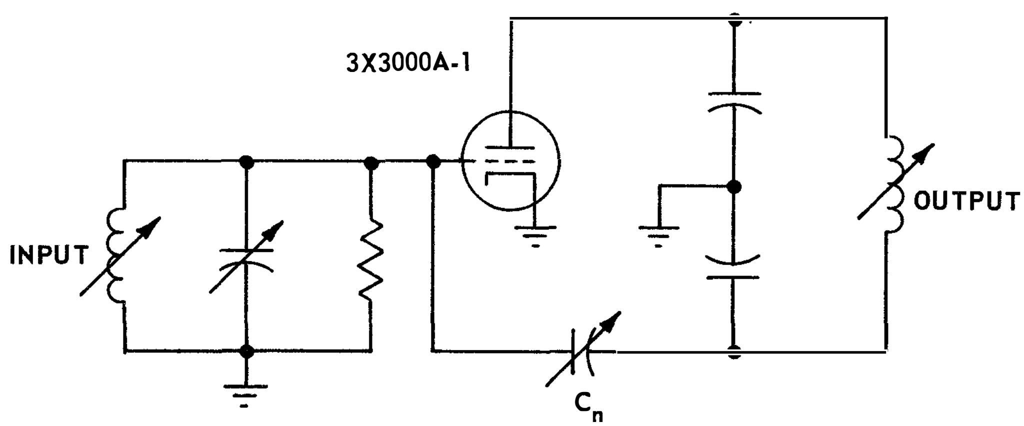

Figure 11-4. Grid-Driven, Plate-Neutralized Triode Power Amplifier.

11-21. A swamping resistor is used in the grid circuit for stability and to maintain a constant input impedance to the stage. When the stage is operated class AB_2, the grid current represents a varying load to the driving source. By adding the swamping resistor, the grid current drawn represents only a small portion of the total grid load, so that the driver load impedance is relatively constant. The swamping resistor increases the required driving power and improves stability by affording a low impedance to ground for regenerative feedback through the plate-to-grid capacitance.

11-22. Cathode-Driven Triode Power Amplifier.

11-23. Figure 11-5 is a simplified schematic of a typical cathode-driven (grounded grid) triode power amplifier. This amplifier, operating class AB_2, produces 4 to 5 kw, using the type 3X2500A triode. In the cathode-driven amplifier, the control grid is at RF ground and the signal is fed to the cathode. The main advantage of operating the triode

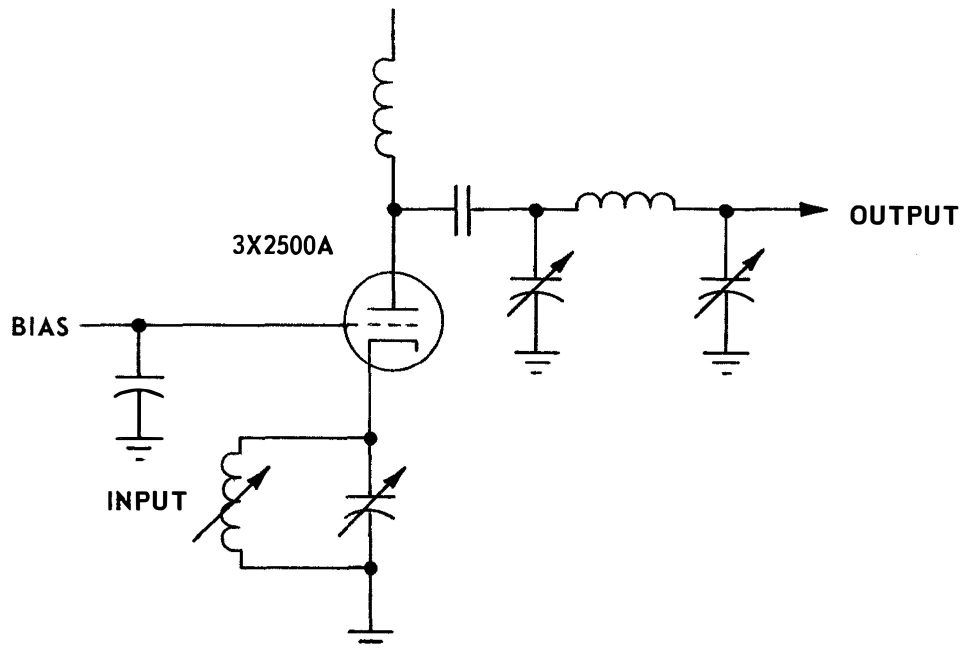

Figure 11-5. Cathode-Driven Triode Power Amplifier.

in this manner is that the control grid becomes an effective screen between the plate and the cathode, making neutralization seldom necessary. The small values of plate-to-cathode capacity have very little effect on the input signal, because the input circuit impedance is usually quite low. Since neutralization is not usually required, triodes with an amplification factor of 20, such as the 3X2500A, can be used. Another advantage of the cathode-driven power amplifier is that the feed-through power is an effective load across the input circuit, making swamping resistors unnecessary. The main disadvantages of this circuit are that a large driving power is required and only power gains of from six to ten can be realized. Most of the power required for driving, however, feeds through the stage and appears in the plate circuit so that it is not lost. The cathode-driven circuit is a convenient circuit to use when high power has already been developed and needs another step up.

11-24. Grid-Driven Tetrode Power Amplifier.

11-25. Figure 11-6 is a simplified schematic of a grid-driven, tetrode power amplifier. This amplifier, operating class AB_1, produces 250 watts per tube, using the type 4X250B tetrode. In general, the same design considerations exist for tetrode amplifiers as for triode amplifiers. That is, grid circuit swamping is required to hold the input impedance constant if the tetrode is driven into the grid current region; and neutralization is generally required if the tube is to operate over the entire high-frequency range. However, since the plate-to-grid capacitance is small in the tetrode, neutralization is much simpler. The tetrode amplifier, being a high-gain tube,

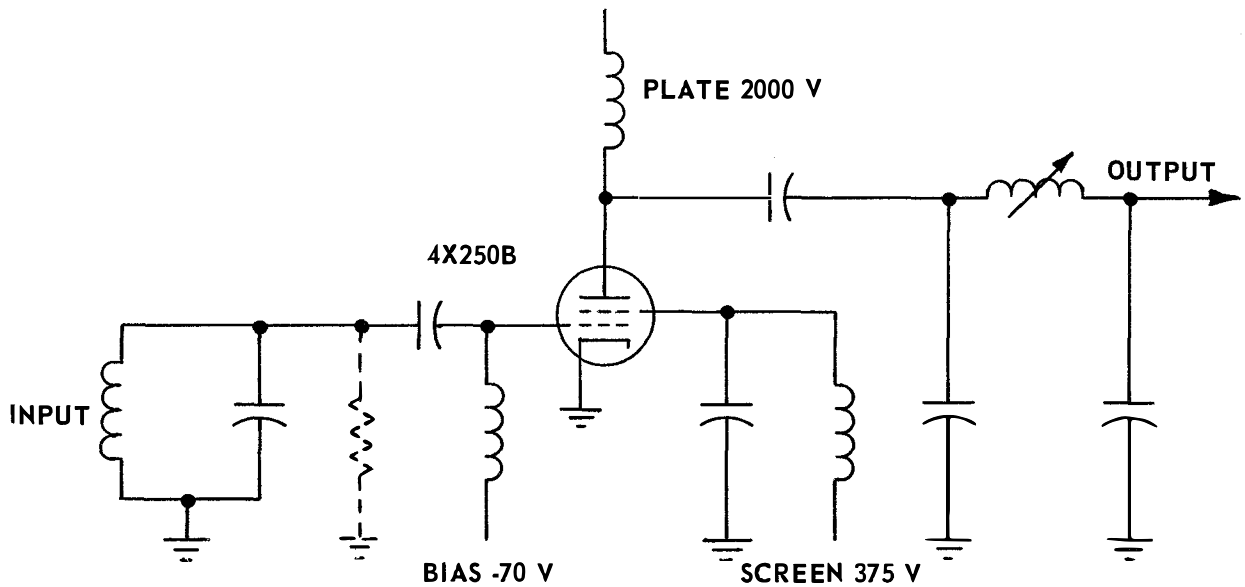

Figure 11-6. Grid-Driven Tetrode Power Amplifier.

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

11-26. Power Amplifier Output Networks.

11-27. Tank Circuit Considerations.

11-28. The plate tank circuit of an RF power amplifier must perform four basic functions.

a. Maintain a sine wave RF voltage at the plate of the tube.

b. Provide a low impedance path from plate to cathode for harmonic components of the plate current pulses.

c. Provide part or all of the necessary attenuation of harmonics and other spurious frequencies.

d. Provide part or all of the impedance matching from the tube plate to the antenna.

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

11-29. The Q of the plate circuit, of which the tank is a part, must be sufficient to keep the RF plate voltage close to a sine wave shape. This is often referred to as the "flywheel effect." If the plate circuit Q is insufficient, the RF waveform may be distorted, which will result in low plate efficiency. This loss of efficiency is seldom noticed unless the plate circuit Q is less than 5. A plate circuit Q of at least 10 is known to be sufficient for linear operation and is a recommended minimum.

11-30. A power amplifier operating either class AB, B, or C delivers power to the tank circuit by plate current pulses. The harmonic content of these pulses is determined primarily by the angle of plate current flow, the harmonics being greater with a smaller angle of plate current flow. In a linear power amplifier, the second harmonic component can be as great as 6 db below the fundamental at full peak envelope power. The higher order harmonic components drop off rapidly but their magnitude varies greatly, depending upon the pulse shape. These harmonics must be attenuated in the output network so that they are 50 db, 80 db, or more, below the fundamental component. The Pi-L network will attenuate the second harmonic to about 50 db below the fundamental, which is from 10 db to 15 db more attenuation than can be obtained from the simple Pi network. Where more attenuation is required, external filters of either the low-pass or band-rejection type are added. Increasing plate circuit Q increases harmonic attenuation, but since doubling the Q results in only about 6 db more second harmonic attenuation, Q's above 20 are seldom used below 30 Mc.

11-31. The Pi-L output network is ideally suited to matching a tube load to a 52-ohm coaxial transmission line. Loads with a standing wave ratio as high as 4 to 1 can be matched easily. This can be done with any value of tube load impedance, whereas the simple Pi network has difficulty matching to low load impedance when the tube plate load resistance is high. The Pi-L network has only four variable elements, and they can be ganged to have only a tuning control and a loading control, as shown in figure 11-7. Since in the Pi-L network C_2 and L_2 affect loading in the same direction, the extra capacity and inductance range of the elements required to extend the loading

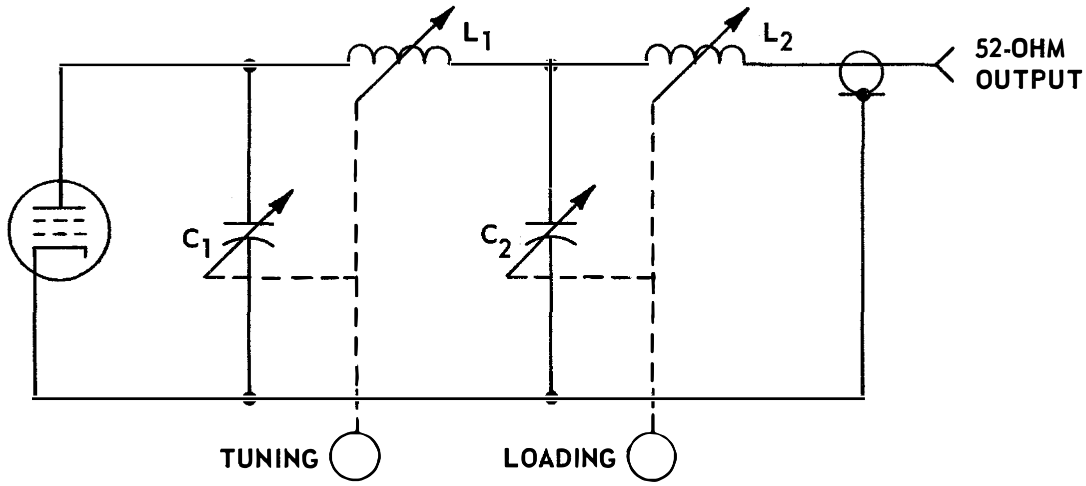

Figure 11-7. Tuning Controls for Pi-L Output Network.

range of the circuit is relatively small. For example, the loading control varies about ±25 percent to match a 52-ohm load with a 4:1 SWR. The tuning control varies about ±10 percent.

11-32. Circuit Losses.

11-33. Nearly all of the tank circuit loss occurs in the coils. These losses are closely related to the ratio of plate circuit Q to coil Q, but other design considerations enter in. These circuit losses are shown in figure 11-8 for a Pi-L network, which has lower

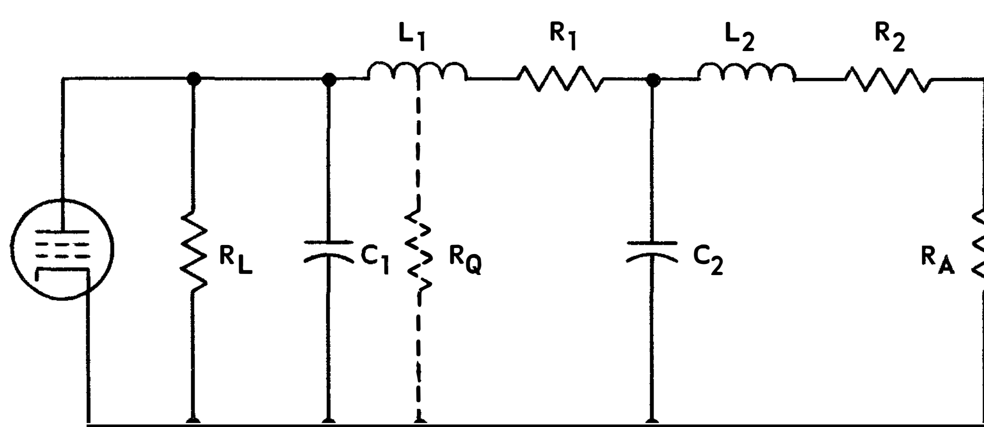

Figure 11-8. Circuit Losses in Pi-L Network.

losses than other networks for 50 db of second harmonic attenuation. Resistances R_1 and R_2 represent the equivalent series resistance of the coils determined from coil Q and reactance. Resistance R_q is the equivalent load resistance in series with L_1 and is determined from the relationship

$$R_q = \frac{R_L}{Q^2 + 1} = \frac{R_L}{(R_L/X_c)^2 + 1}$$

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

$$\text{Percent loss} = \left(\frac{R_1}{R_q + R_1} + \frac{R_2}{R_a + R_q} \right) \times 100$$

11-34. <u>Tank Coil and Capacitor Requirements.</u>

11-35. The frequency range and method of tuning are major factors in determining tank circuit components. Continuously variable coils and capacitors, which will cover the entire frequency range without any band switching, are the most desirable. However, this is not practical in Autotune transmitters because of the limited torque available to drive the tuning elements and the often short repositioning time specified. With these limitations, bandswitching is almost essential. Where instantaneous frequency change is specified, it is common to switch from one pretuned RF unit to another, and manually tuned circuits are suitable for this purpose. Servo control of the tuning elements permits incorporation of various automatic tuning or prepositioning circuits and is well suitable for driving continuously variable elements. A practical way to design a transmitter is to use continously variable elements which can be operated either manually or by an accessory servo system.

11-36. The use of continuously variable elements has the following advantages.

a. Circuit Q can be kept more uniform across the frequency range.

b. Circuit losses can be kept to a minimum.

c. The range of variable coils and capacitors can be less.

d. A maximum amount of harmonic attenuation is more easily maintained across the frequency range.

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

11-37. Neutralization.

11-38. Effects of Plate-to-Grid Capacitance.

11-39. The purpose of neutralization is to balance out the effect of plate-to-grid capacitive coupling in a tuned RF amplifier. In a conventional tuned RF amplifier using a tetrode, the effective input capacity of the tube is given by the following equation:

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

Where

C_{in} = tube input capacitance

C_{gp} = plate-to-grid capacitance

A = voltage amplification from grid to plate

θ = phase angle of plate load

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

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

This input resistance is in parallel with the grid current loading, grid tank circuit losses, and driving source impedance. When the plate circuit is tuned to the inductive side of resonance, energy is transferred from the plate to the grid circuit through the plate-to-grid capacitance (positive feedback). This introduces negative resistance in the grid circuit. When this shunt negative resistance across the grid circuit is lower than the equivalent positive resistance of the grid loading, circuit losses, and driving-source impedance, the amplifier will oscillate. As the plate circuit is tuned to the capacitive side of resonance, the input resistance becomes positive and power is transferred from the grid to the plate circuit. This is why the grid current in an unneutralized tetrode RF amplifier varies from a low value to a high value as the tank circuit is varied from below to above resonance. If the amplifier is over-neutralized, the effect reverses. This effect can be observed in a pentode or tetrode amplifier operating class A or AB_1 by placing an RF voltmeter across the grid circuit and tuning the plate circuit through resonance.

11-40. Neutralizing Circuits.

11-41. Most of the neutralizing circuits developed for use with triodes may be used equally successfully with tetrodes. However, those circuits which require balanced tank circuits for neutralizing purposes only are undesirable, because the trend in RF power amplifier design is toward single-ended stages. A conventional, grid-neutralized amplifier is shown in figure 11-9. Capacitor C_3 balances the grid-to-filament capacity to keep the grid circuit in balance. When $C_1 = C_2$ and $C_n = C_{gp}$, it is readily seen that a signal introduced into the grid circuit will not appear across the plate circuit because the coupling through C_n is equal and opposite to the coupling through C_{gp}. The relationship for no coupling from the grid circuit to the plate circuit is given by the relationship

$$\frac{C_1}{C_2} = \frac{C_{gp}}{C_n}$$

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

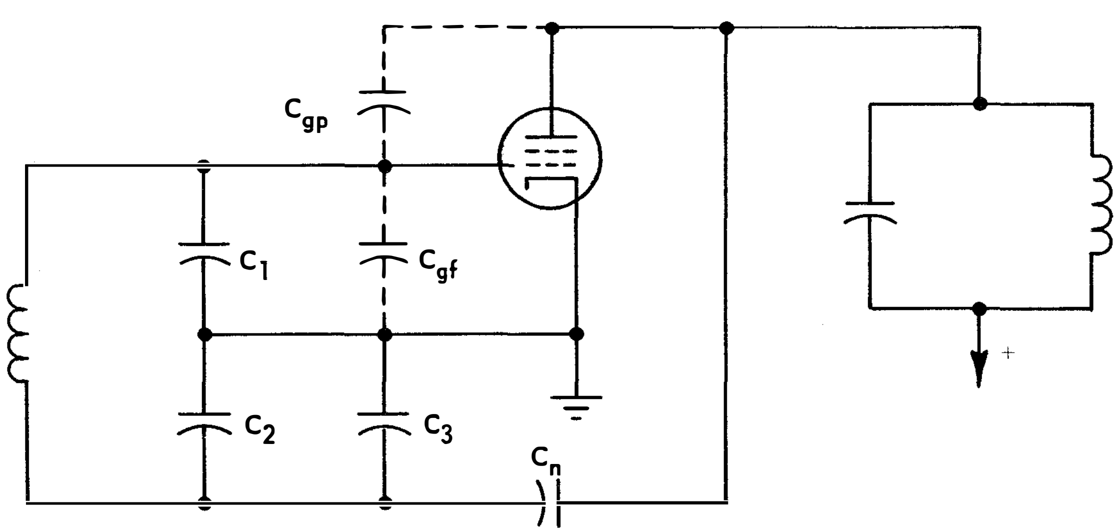

Figure 11-9. Conventional Grid-Neutralized Amplifier.

when C_2 is very much larger than C_1. By placing most of the grid tuning capacitance across the grid tank coil, using the bypass capacitor C from the bottom end of the grid tank circuit to ground for C_2, and using the grid-to-filament capacity for C_1, the modified grid-neutralized circuit shown in figure 11-10 results. The relationship for neutralization of this circuit is given by

$$\frac{C_n}{C} = \frac{C_{gp}}{C_{gf}}$$

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

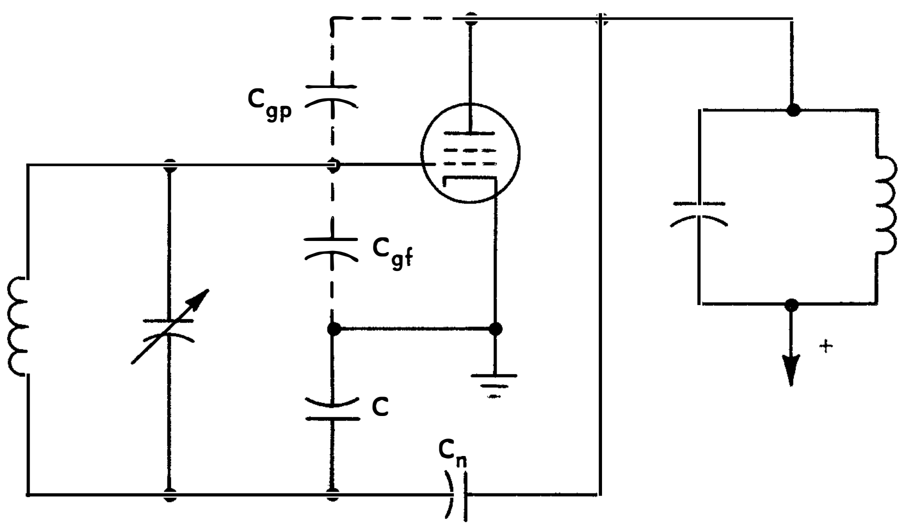

Figure 11-10. Modified Grid-Neutralized Amplifier.

11-42. Testing for Proper Neutralization.

11-43. When a power amplifier stage is properly neutralized, the power output peaks at the same time the plate current dips. An indication of this simultaneous peak and dip is often the most convenient way of testing for proper neutralization. To perform such a test, the DC cathode current, or plate current, of the neutralized stage is used to obtain an indication of plate current dip. The power output from the same stage or the grid drive to any succeeding stage is used to obtain an indication of power output. A power amplifier is usually checked for proper neutralization near the high-frequency end of its range where neutralization is more critical.

11-44. When the drive to a neutralized stage is so low that a plate current dip is not present, the best way to test for proper neutralization is by injecting a test signal into one circuit and checking for coupling of the signal into another circuit. In the modified grid-neutralized circuit shown in figure 11-10, proper neutralization balances out coupling between the input tank circuit and the output tank circuit, but it does not remove all coupling between the plate circuit and the grid-to-cathode circuit. Therefore, a test signal injected into the plate circuit will result in grid-to-cathode signal, even with proper neutralization. However, a test signal injected into the plate circuit will not result in a signal in the grid coil with proper neutralization. The presence of a signal in the grid coil can be detected by using an inductive coupling loop. This circuit can also be neutralized by inductively coupling an input signal into the input circuit and adjusting the neutralizing capacitor for minimum signal in the plate circuit.

11-45. RF Feedback Circuits.

11-46. Introduction.

11-47. RF feedback is a very effective means of reducing distortion in a linear power amplifier. Twelve db of RF feedback produces nearly 12 db of distortion reduction, and this distortion reduction is realized at all signal levels. However, voltage gain per stage is reduced by the amount of feedback employed; for example, with 12 db of feedback, the gain is reduced by one-quarter.

11-48. Feedback Around One Stage.

11-49. Figure 11-11 shows a negative feedback circuit around a one-stage RF amplifier. The voltage developed across C_4 is introduced in series with the voltage developed across the grid tank circuit and is in phase opposition to it. The feedback obtainable with this circuit can be varied between zero and 100 percent by properly

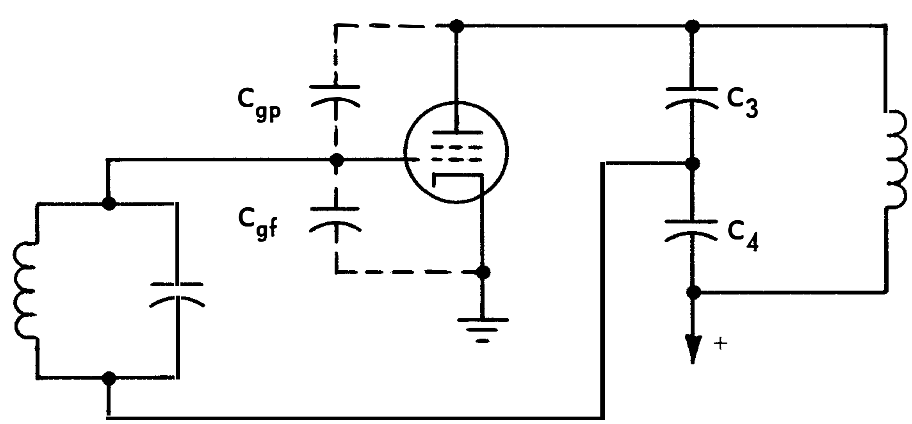

Figure 11-11. One-Stage Feedback With Neutralization.

choosing the values of C_3 and C_4. It is necessary to neutralize this feedback amplifier, the neutralization requirements being

$$\frac{C_{gp}}{C_{gf}} = \frac{C_3}{C_4}$$

To satisfy the neutralization requirement, it is usually necessary to add capacity from the plate to the grid. Using this circuit, a problem is encountered in coupling into the grid circuit. Inductive coupling is ideal, but the extra tank circuit complicates the tuning of the power amplifier if several cascaded amplifiers are used with feedback around each. The grid can be capacity coupled to a driver with a high source impedance, such as a tetrode or pentode. However, if this is done, feedback cannot be used in the driver because it would cause the source impedance to be low.

11-50. Feedback Around Two Stages.

11-51. Feedback around two RF stages has the advantage that more of the tube gain can be realized while nearly as much distortion reduction can be obtained. For instance, 12 db feedback around two stages provides about the same distortion reduction as 12 db around each of two stages separately. Figure 11-12 shows a negative feedback circuit around a two-stage amplifier, with each stage neutralized. The small feedback voltage required is obtained from the voltage divider C_6 and C_7. This feedback voltage is applied to the cathode of the first stage. The feedback divider can be left fixed for a wide frequency range since C_6 is only a few micromicrofarads. For example, if the

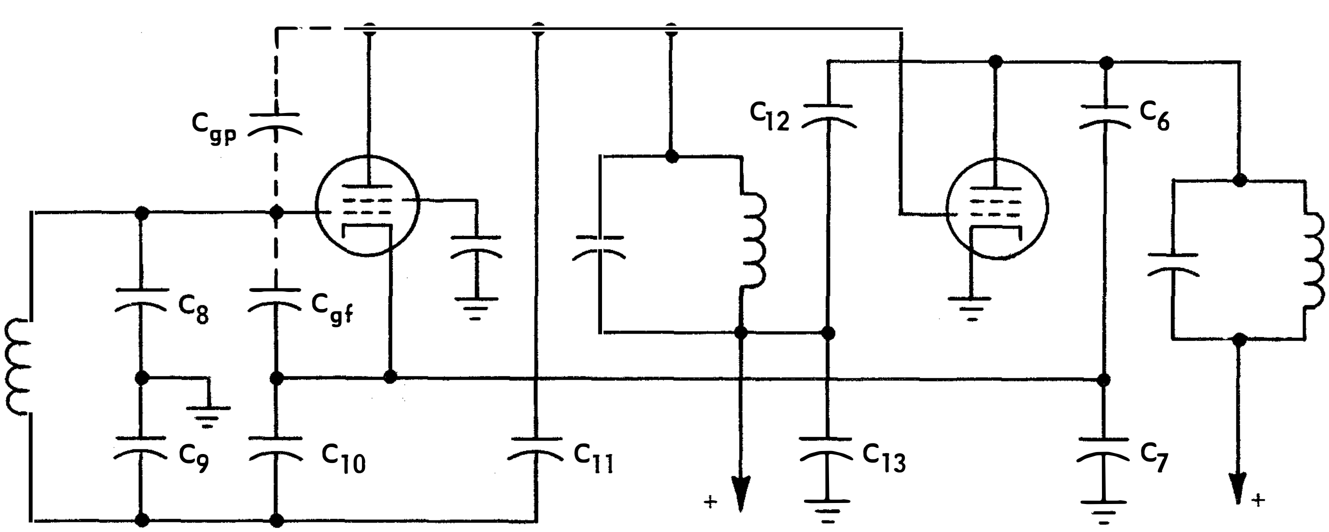

Figure 11-12. Two-Stage Feedback With Neutralization.

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

11-52. It is necessary to neutralize the cathode-to-grid capacity of the first tube in the two-stage feedback circuit to prevent undesirable feedback coupling to the input grid circuit. The relationship for the circuit which accomplishes this cathode-to-grid neutralization is

$$\frac{C_8}{C_9} = \frac{C_{gf}}{C_{10}}$$

To reduce the voltage across the input tank coil and minimize the power dissipated by the coil, the input circuit can be unbalanced by making C_9 up to five times C_8, as long as C_{10} is increased accordingly. The cathode-to-grid capacity of the first tube can be neutralized by injecting a test signal into the cathode of the tube. The neutralizing bridge is then adjusted for minimum signal, as indicated by a detector which is inductively coupled into the input coil.

11-53. Except for tubes with very small plate-to-grid capacity, it is necessary to neutralize C_{gp} in both tubes. This neutralization for the second tube is realized by choosing C_{12} and C_{13} so that the ratio C_{12}/C_{13} equals the ratio C_{gp}/C_{gf} in the second tube. If neutralization of C_{gp} is necessary for the first tube, it is obtained by satisfying the relationship

$$\frac{C_{gp}}{C_{11}} = \frac{C_{gf}}{C_{10}} = \frac{C_8}{C_9}$$

The screen and suppressor of the first stage should be grounded to keep the tank output capacity directly across the interstage circuit. This avoids common coupling between the feedback on the cathode and the interstage circuit.

11-54. In a two-stage feedback amplifier, the voltage fed back to the cathode of the first stage must be in phase with the grid input signal, measured from grid to ground. If the feedback voltage is not in phase with the grid input signal, the resultant grid-to-cathode voltage increases, as shown in figure 11-13. When the output circuit is properly tuned, the resulting grid-to-cathode voltage on the first tube is minimum, which will make the voltage across the interstage tank circuit minimum also.

11-55. Automatic Load Control.

11-56. Automatic load control is a means of keeping the signal level adjusted so that the power amplifier works near its maximum power capability without being overdriven

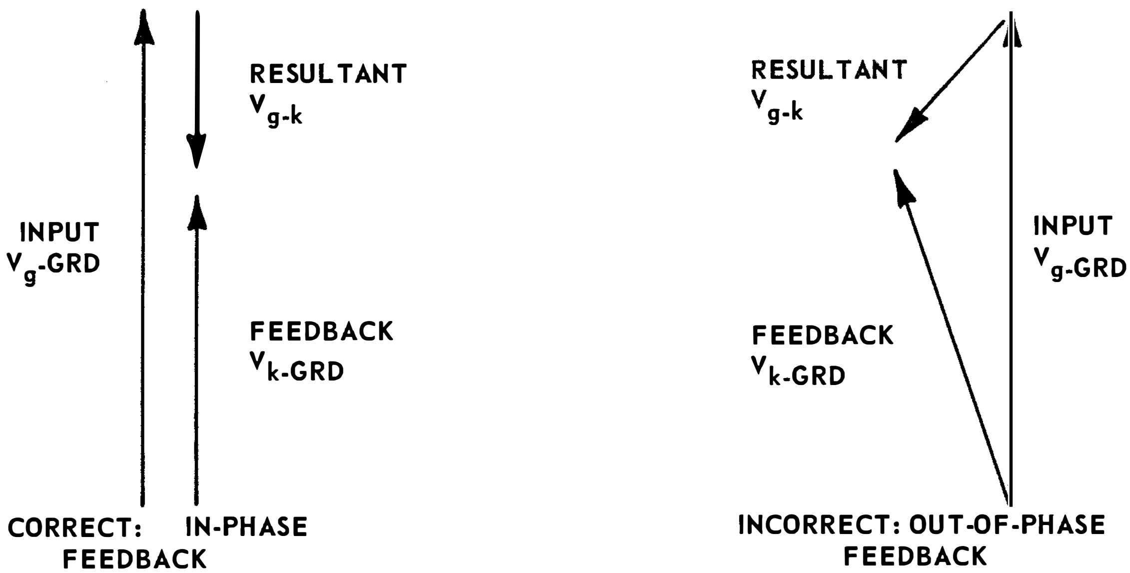

Figure 11-13. Vector Relationship of Voltages for Two-Stage Feedback.

on signal peaks. In AM systems, it is common to use speech compressors and speech clipping to perform this function. However, in an SSB system, these methods are not equally useful because the peaks of the SSB signal do not necessarily correspond with the peaks of the audio signal. Therefore, the most effective means of control is obtained by a circuit which receives its input from the envelope peaks in the power amplifier and uses its output to control the gain of the exciting signal. Such a circuit is an automatic load control (ALC) circuit.

11-57. Figure 11-14 is a simplified schematic of an ALC circuit. This circuit uses two variable-gain stages of remote cutoff tubes, such as the 6BA6, operating very similarly to the IF stages of a receiver with automatic volume control. The grid bias voltage of the variable-gain amplifiers is obtained from the ALC rectifier connected to the power-amplifier plate circuit. The capacity voltage divider steps down the RF voltage from the power amplifier plate to about 50 volts for the rectifier. A large delay bias is used on the rectifier so that no reduction of gain takes place until the signal level is nearly up to full power capability of the power amplifier. The output of the ALC rectifier passes through RC networks to obtain the desired attack and release times. Usually a fast-attack time, about 2 milliseconds, is used for voice signals, so that the gain is reduced rapidly to remove the overload from the power amplifier. After a signal peak passes, a release time of about 0.1 second returns the gain to normal. A meter calibrated in db of compression is used to adjust the gain for the desired amount of load control.

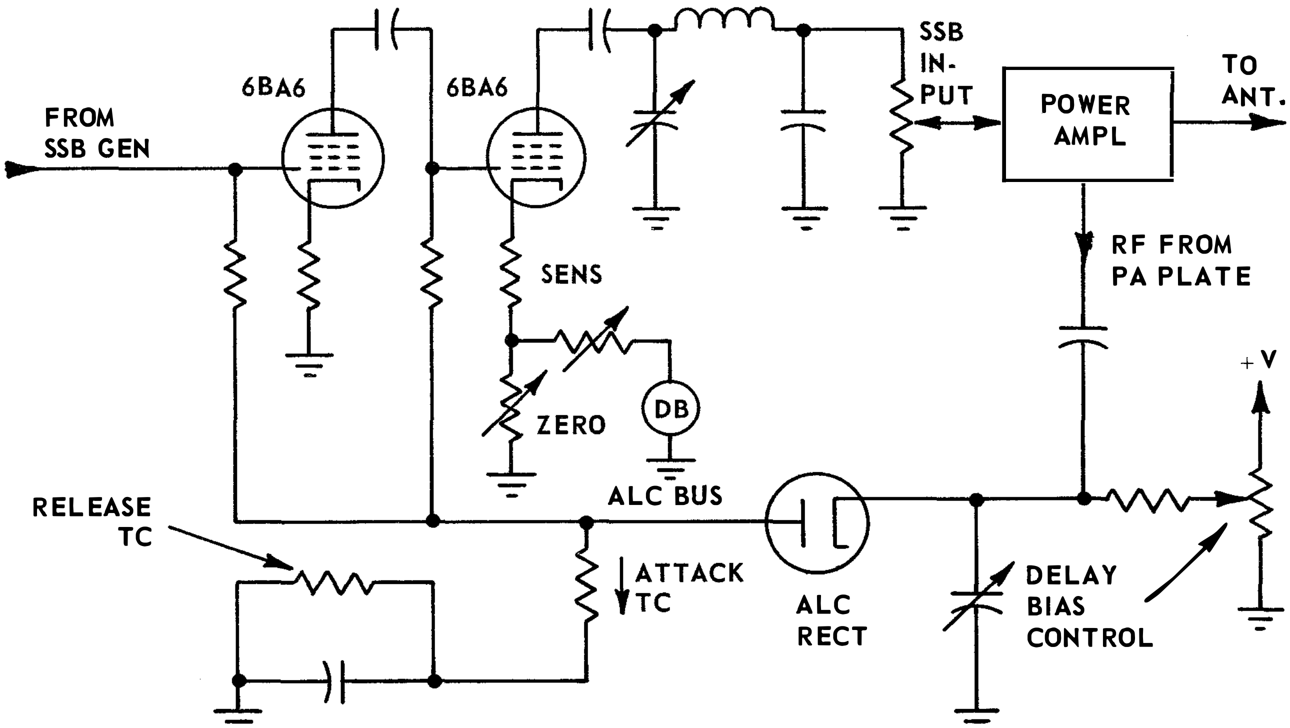

Figure 11-14. Automatic Load Control Circuit.

11-58. In single-channel speech transmission, the ALC circuit performs the function of a speech compressor. To do this, a range of 12-db is usually provided, with control maintained on input peaks as high as 20 db above the threshold of compression. Since the signal level should be fairly constant through the preceding SSB generator, it is unlikely that more than a 12-db range of ALC would be useful. If the signal level varies more than 12 db for the SSB generator, a speech compressor in the input audio amplifier is usually used to limit the range of the signal fed into the SSB generator. Figure 11-15 shows the effectiveness of the ALC circuit in limiting the output signal to the capabilities of the linear power amplifier. An adjustment of the delay bias will put the threshold of compression at the desired level.

11-59. Linear Power Amplifier Tuning.

11-60. Introduction.

11-61. When a power amplifier is operated class C, a pronounced plate current dip and grid current peak are fairly accurate indications of proper tuning. In a linear power amplifier, the use of these indications is limited. For instance, in a class A amplifier there will be no plate current dip; therefore, the class A amplifier output circuit must be tuned for an indication of maximum input to the next stage. In class AB amplifiers, the plate current dip is not always readily detected. This does not mean that conventional tuning procedures will not properly tune a linear amplifier,

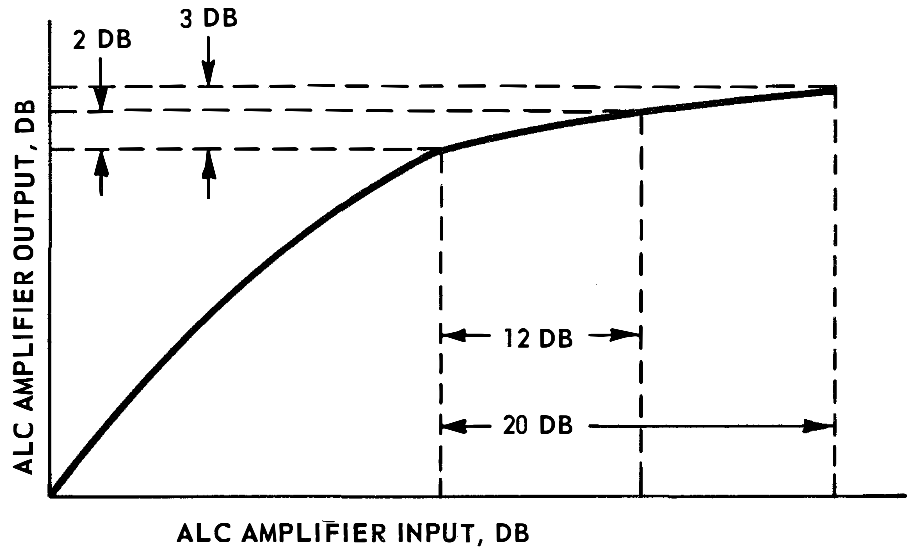

Figure 11-15. Automatic Load Control Performance Curve.

but tuning a linear amplifier with conventional procedures is much more exacting. One procedure commonly used is to increase the drive to a stage in order to obtain a good plate current dip indication.

11-62. In low Q tank circuits, the point of plate current dip is not a true indication of exact resonance, because the plate current dip occurs at maximum impedance rather than when the tank circuit appears as a pure resistance. This is especially true for Pi networks and Pi-L networks. For instance, in a network with a Q of ten, the phase angle at maximum impedance is about 17° from unity. Tuning this far from resonance in a linear amplifier with RF feedback can be much more serious than in a class C amplifier, because the phase angle of the feedback voltage is critical.

11-63. Phase Comparison Tuning.

11-64. Use of a phase comparator circuit to compare the phase of the input signal to the phase of the output signal affords the most sensitive means of tuning a linear power amplifier stage. This circuit employs a phase discriminator, such as shown in figure 11-16, for phase comparison. A balanced, push-pull voltage is obtained through a 90° phase-shift network to provide the voltage $E_a + E_b$. In the figure shown, $E_a + E_b$ is in phase with the current in the inductive branch of the grid tank circuit. Since the current in the inductive branch is 90° out of phase with the voltage across the tank circuit, the induced voltage $E_a + E_b$ is also 90° out of phase with the voltage across the tank circuit. From the output of the stage, E_c is obtained. When E_c is exactly 90° out of phase with E_a and E_b, the voltages across the two crystals, CR1 and CR2, are equal in magnitude. Then, the DC currents in the diode loads are equal and flowing in opposite directions, producing zero output. When E_c is not exactly 90° out of phase with E_a and E_b, the voltages across the two crystals are

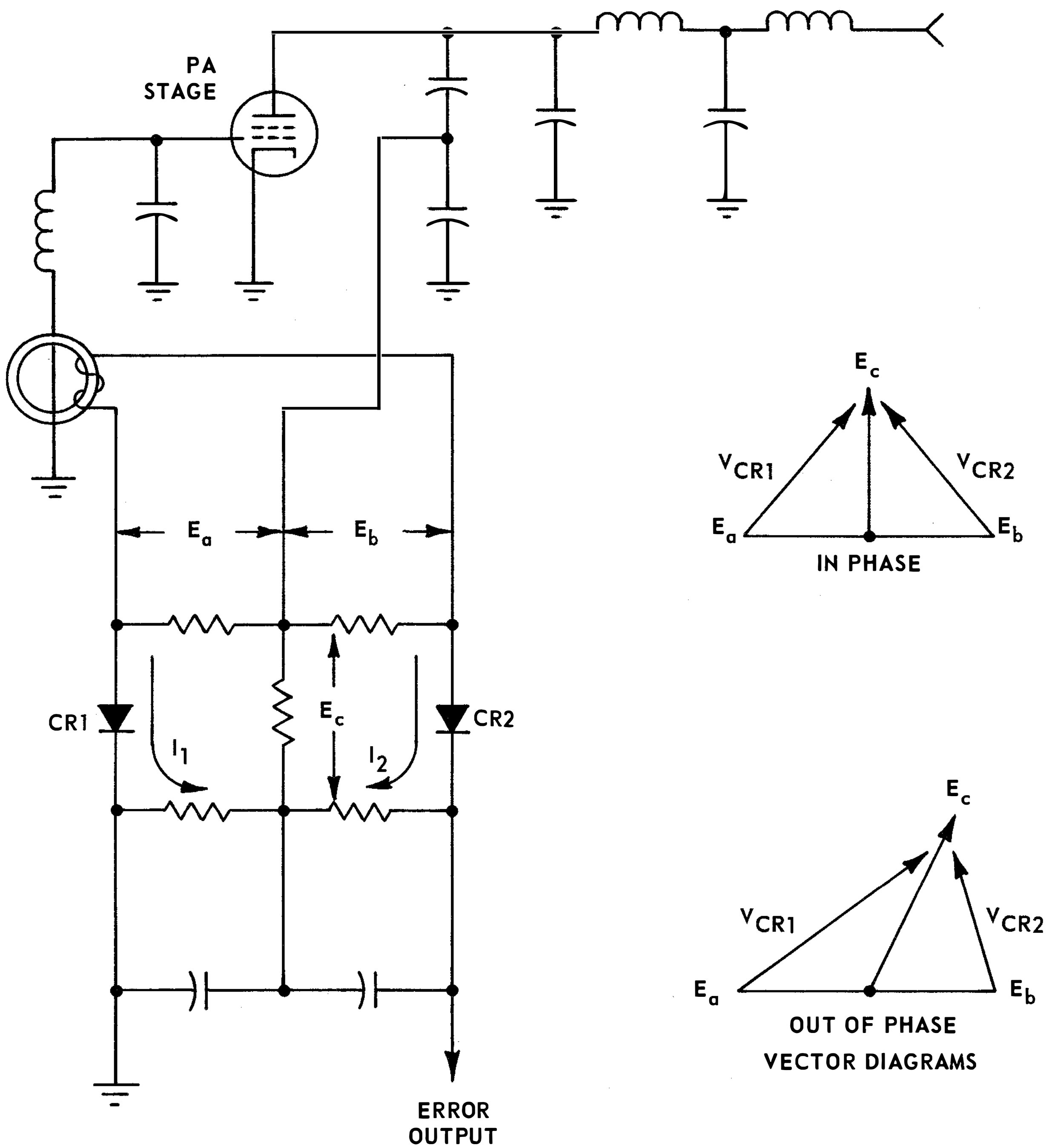

Figure 11-16. Phase Discriminator for Power Amplifier Tuning.

unequal in magnitude. This will cause the DC currents in the diode loads to be unequal, which will produce an output. The error signal derived from this circuit can be used to operate a zero-center meter for manually tuning the output circuit. When tuned for zero meter indication, the output voltage is exactly 180° out of phase with the input voltage, the condition for true resonance.

11-65. The phase discriminator can also be used to obtain an error signal for servo-tuning the stage. However, for servo tuning, coarse positioning information is necessary because the phase discriminator responds to harmonic tuning points and because there is insufficient output from the phase discriminator over much of the frequency range. This coarse positioning information can be provided with a coarse follow-up potentiometer which receives information from the exciter frequency control circuits. Such a system requires that the master potentiometer track the tuning curves of the amplifier tank circuits and that sequencing controls be used to initiate and halt coarse positioning at the proper times. Pretuning information can also be derived from the exciter RF output signal by using a coarse discriminator circuit, such as is shown in figure 11-17. This circuit is a series RC network fed with RF voltage from the exciter.

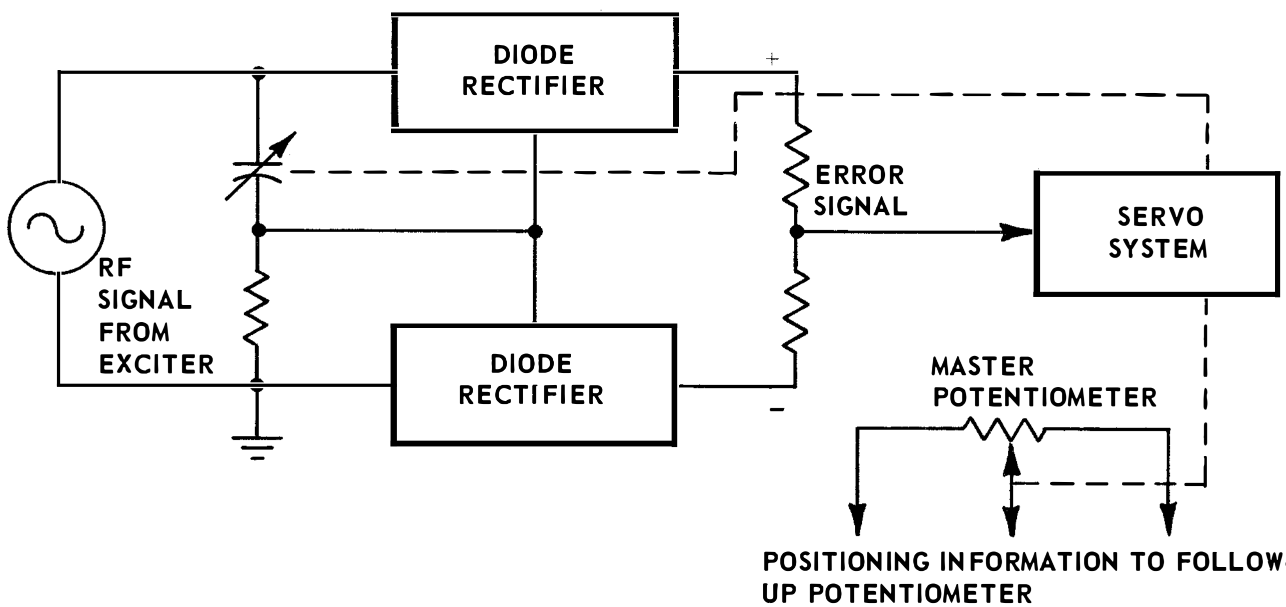

Figure 11-17. Coarse Discriminator for PA Tuning.

A servo system then drives the capacitor in the RC bridge to produce zero error signal at the same time it positions a master potentiometer. A second tuning servo then drives a follow-up potentiometer which is wound to cause the tuning servo to track the tuning curve of the amplifier tank circuit. To tune the amplifier automatically, the error signals from the phase discriminator and the coarse discriminator can be combined to operate a single servo. The servo system will then operate over the whole frequency range and have a precise zero error signal position, as shown in figure 11-18.

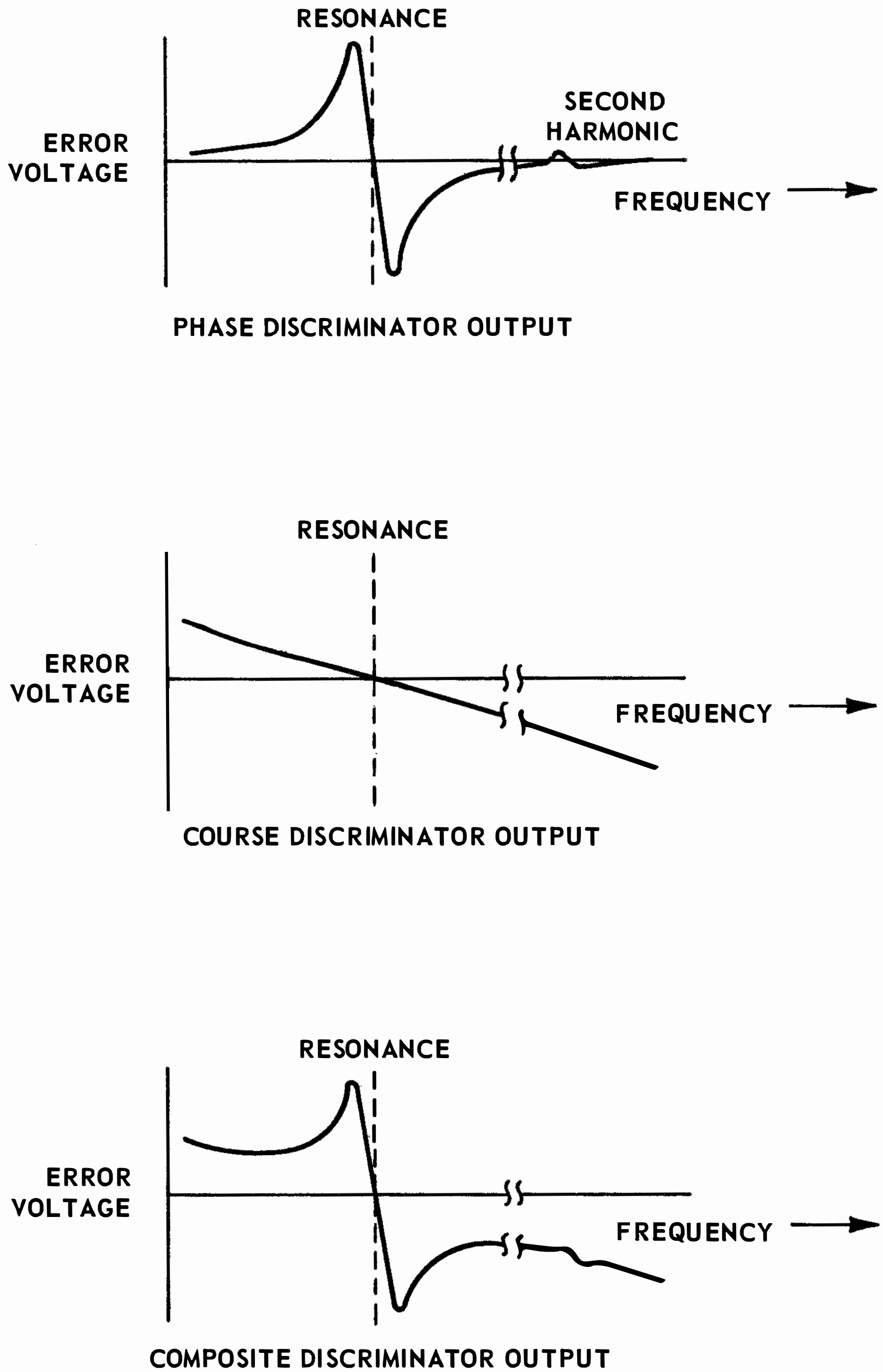

Figure 11-18. Discriminator Output Curves for Power Amplifier Tuning.

11-66. Loading Comparator Circuit.

11-67. Since the voltage gain of a tube is dependent upon the load resistance, a loading comparator circuit, as shown in figure 11-19, can be used to determine proper loading. The loading comparator is designed so that a predetermined ratio between positively rectified grid voltage and negatively rectified plate voltage produces zero-error signal output. The power amplifier is then manually or automatically loaded until the error-signal output goes to zero. The clamping diode is required so that the circuit will maintain control under light load when the amplifier is driven into plate saturation. In plate-saturated operation, the rectified grid voltage will continue to rise with reduced loading, while the rectified plate voltage remains relatively constant. This will cause the circuit to lose its sense of direction and result in reducing the load even further. To maintain the sense of direction under this condition, the clamping diode prevents the rectified grid voltage from exceeding a voltage which is proportional to plate current. Therefore, in plate-saturated operation, which is similar to class C operation, loading is determined by the ratio of plate current to RF plate voltage. Proper compromise of the magnitude of the plate, grid, and clamping signal voltages results in a loading comparator that produces proper loading information regardless of the operating conditions, provided the plate circuit is held at resonance.

11-68. Antenna Tuning and Loading.

11-69. The output network of a variable frequency transmitter must be capable of tuning and loading into a transmission line which presents different impedances at different frequencies. This requires output networks which will match a wide range of load impedances with the power amplifier output. In fixed-station equipment, the

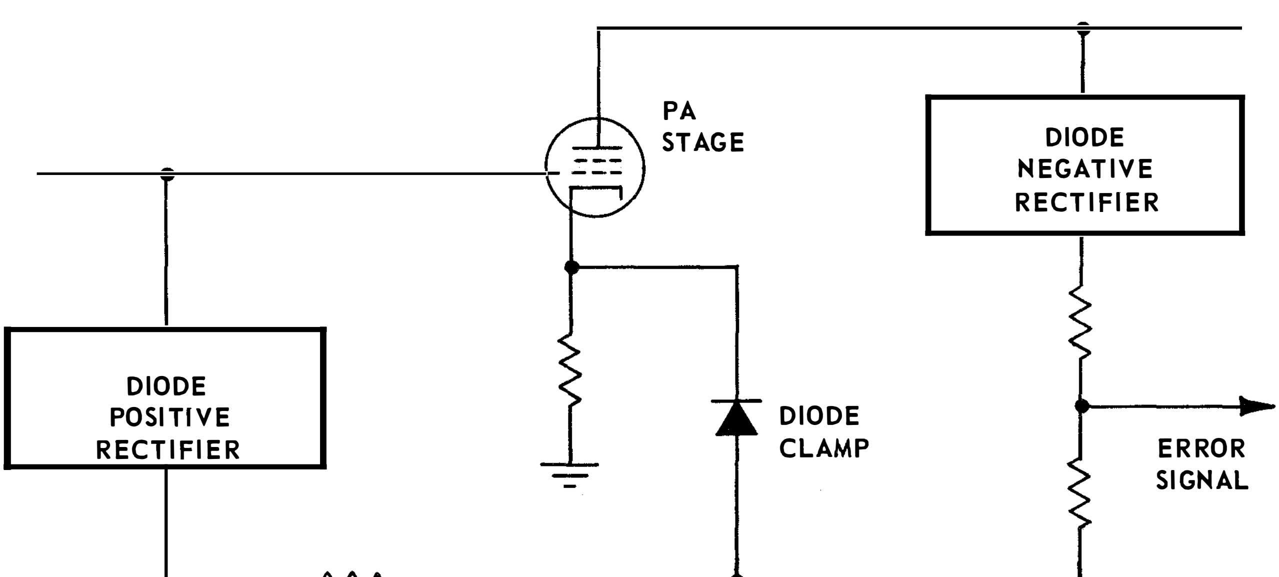

Figure 11-19. Loading Comparator for PA Loading.

power amplifier usually works into a transmission line and antenna designed so that the load impedance presented to the amplifier varies over only a limited range. In this case, the output network is designed to match the load impedance directly. In mobile and airborne equipment, the power amplifier usually works into a coaxial transmission line terminated with any one of a wide variety of antennas which present unwieldy terminating impedances. In such cases, an antenna coupler is used. For mobile transmitters, the antenna coupler is commonly located near or in the transmitter to provide proper coupling between the transmitter output networks and a transmission line which is terminated with a mismatched antenna. For airborne transmitters, the antenna coupler is placed near the antenna to terminate the transmission line properly and provide coupling for maximum power transfer to the antenna.

11-70. Two power amplifier control functions are required to match properly the load impedances presented to the power amplifier. One is a phasing control, or tuning control, which will balance out undesirable reactance and make the load resistive or as nearly resistive as possible. The other is a load control which will provide the proper terminating impedance. Figure 11-20 shows several ways that the output network components can be ganged to provide tuning and loading with two controls. The tuning control is adjusted to produce a plate current dip, which indicates maximum impedance. For more precise tuning and automatic tuning, the phase discriminator circuit is used. The loading control is adjusted to produce a pre-established value of grid voltage and plate current or, in some cases, a pre-established value of screen current and plate current. For more precise loading, or automatic loading, the loading comparator circuit is used. The loading and tuning circuits must be so designed that the controls will not lose sense of direction under any circumstances. This is absolutely essential for automatic loading and tuning, and is highly desirable for manual loading and tuning.

11-71. Power Supplies for RF Power Amplifiers.

11-72. Fixed transmitters up to 1 kw usually use a single-phase AC power source. Larger fixed transmitters usually use a three-phase AC power source. Mobile equipment may operate from 6- to 28-volt DC power sources, using dynamotors or vibrator power supplies to obtain the required high voltages. Airborne equipment usually uses the 400-cycle power source of the aircraft.

11-73. In addition to supplying the required DC voltage and output current, the power supply must have adequate DC regulation, good dynamic regulation, and low ripple or noise output. Most high-voltage power amplifiers have a varying load characteristic so that good DC regulation is essential. To reduce ripple and noise, filters are used between the rectifier and the power supply load. The filter chokes place a high impedance between the rectifier and the load, making large capacitors necessary in the output side of the filter to minimize voltage fluctuations caused by rapid variations in load current. This is particularly necessary in high-voltage power supplies for linear power amplifier stages.

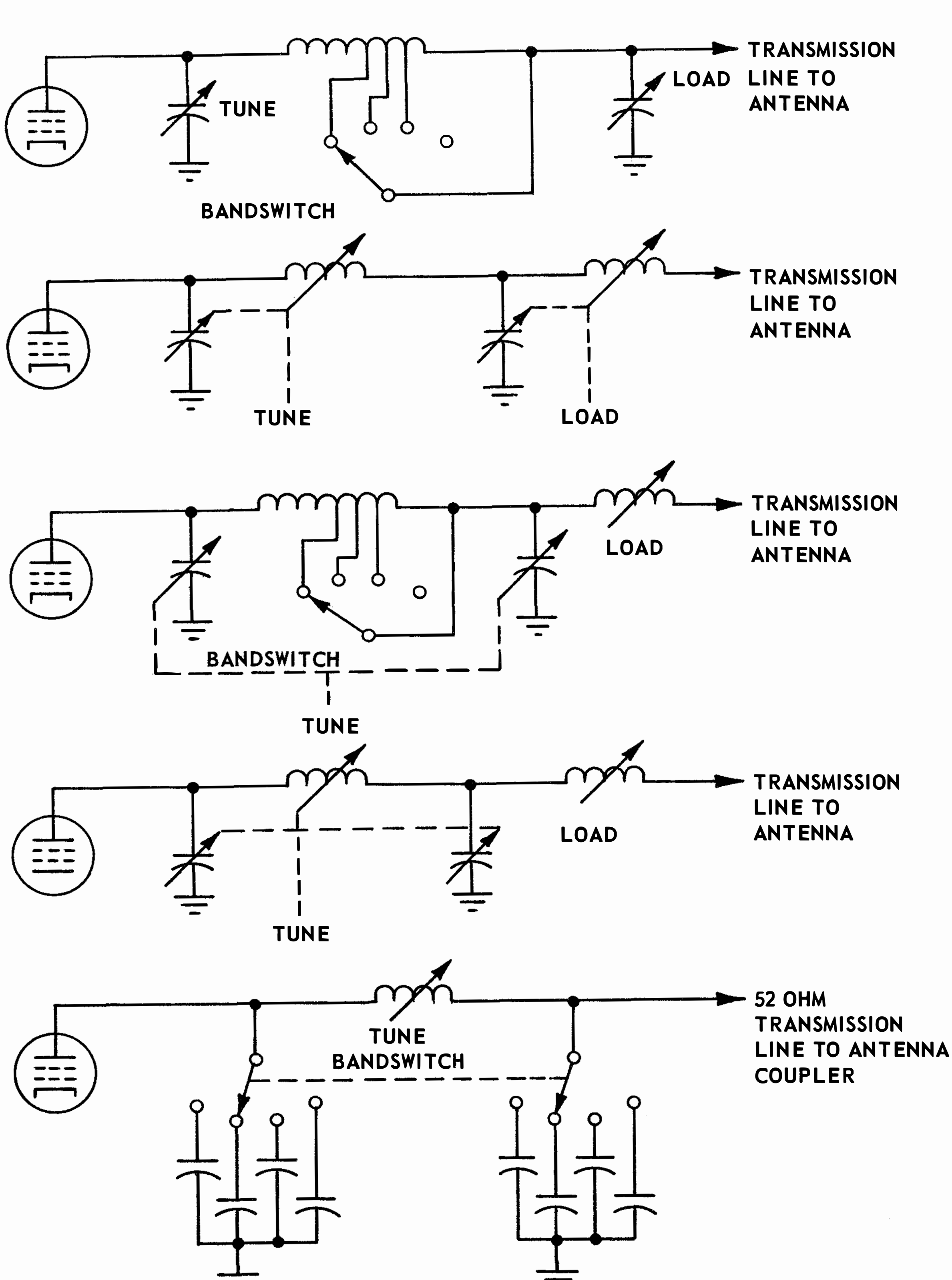

Figure 11-20. Output Network Tuning and Loading Controls.

11-74. Vacuum rectifiers can be used for small, low-voltage power supplies which have relatively constant load. Gas-type rectifiers are required where better regulation is necessary. The mercury-vapor rectifier is the most common gas-type rectifier, because it has long life when properly operated. Operating a mercury-vapor rectifier above or below its rated temperature changes the vapor pressure in the tube and reduces its peak inverse voltage capability, making the rectifier more susceptible to arc-back. Equipment which is subject to wide ambient temperature variations, such as military equipment, uses inert gas rectifiers, such as the 3B28 and 4B32. These tubes can be operated in ambient temperatures from -75°C to +90°C, which is frequently a necessary feature. The tube life of the inert gas rectifier, however, is only about one-third of the tube life of an equivalent mercury-vapor rectifier. Metallic rectifiers, such as selenium and copper oxide, are frequently used in power supplies delivering less than 100 volts.

11-75. Rectifier tube life is increased by operating the filaments 90° out of phase with the plate voltage. This minimizes the difference in voltage from each end of the filament to the plate and allows a more uniform emission over the entire filament. A 60° phase difference between the filament and the plate voltage is often used when it is more easily obtained, because almost the full advantage of quadrature operation is realized. Tube ratings of some of the larger rectifier tubes are increased for quadrature operation.

11-76. Transient voltages and currents, which far exceed the steady state values, occur in power supplies when the supply is energized. If these transient peaks exceed the peak inverse voltage rating of the tube, an arc-back may result. For this reason, rectifier tubes are often operated so that the normal peak inverse voltage does not exceed one-half of the rated peak inverse voltage. If this is not possible, a step-start circuit is used, which starts the transformer with resistors in series with the primary. After a short time delay these resistors are shorted out. Some high-voltage rectifiers are started with a resistor in series with the filter capacitor, the resistor being shorted out after a short time delay. This prevents a transient due to the charging current required to bring the voltage up on the filter capacitor. The added resistance in the circuit prevents excessive current in the rectifier.

11-77. Control Circuits.

11-78. Power amplifier control circuits must perform three functions: provide circuit control, equipment protection, and personnel protection. In small transmitters, the control circuits may consist of nothing more than an on-off switch to supply heater power and a push-to-talk button to apply plate voltage and put the transmitter on the air. In larger equipment, pushbuttons are usually used to initiate a certain sequence of relay operations which complete a function in the proper manner. Many transmitters, particularly those suitable for remote control, are capable of complete energization from a single pushbutton control. The filament on-off switch, or pushbutton, initiates a sequence of functions that applies power to the filaments, starts the cooling system, and energizes time-delay circuits which make the power amplifier ready for the application of plate power. In the off position, the power amplifier is shut down.

11-79. Filaments of high-power amplifier tubes are energized separately and, in the case of mercury-vapor tubes, a time delay allows for warmup. Because the life and reliability of many components is greatly dependent upon operating-temperature control, the blower is actuated at the beginning of the starting sequence. Air interlocks prevent the application of power to high-power tubes before cooling air is present and a blower-off delay maintains cooling air after shutdown. In various power amplifier stages, it is essential that bias voltage be applied before plate or screen voltage is applied. This requires sequencing the application of the bias voltage and the plate voltage as well as interlocks between the two, so that the loss of bias voltage will result in removing the plate voltage.

11-80. Power amplifier control circuits are sequenced and interlocked so that everything else must be on and functioning before the high voltage plate transformer is energized. Certain power tetrodes require that screen voltage be applied simultaneously with plate voltage to prevent excessive screen dissipation. To prevent high current and high voltage transients, plate voltage is often applied through step-start circuits, which place resistors for a short time in the power supply circuit.

11-81. Medium and high power tubes are nearly always protected from excessive plate current by overload relays. These relays remove the high-voltage primary power if the plate current exceeds a preset value. Many overloads which occur during normal operation will clear themselves when the high voltage is removed. For this reason, large power amplifiers are usually provided with an overload recycle circuit. This circuit brings the power amplifier back on after an overload. If the overload recurs, the power amplifier will again shut down. The number of recycles before permanent shutdown can generally be preset with a recycle counting switch.

11-82. Tube Operating Conditions for RF Linear Power Amplifiers.

11-83. General.

11-84. SSB amplifiers provide linear amplification and operating conditions similar to those of audio amplifiers. There is one fundamental difference, however, between audio and RF linear amplifiers. This is that the input and output voltages of a tuned RF amplifier are always sine waves because the tuned circuits, if they have adequate Q, make them so. Therefore, any distortion in an RF amplifier results in distortion of the SSB modulation envelope and not in the shape of the RF sine wave. Restated, distortion in an RF linear amplifier causes a change in gain of the amplifier when the signal level is varied. The greatest difference between an audio amplifier and an RF linear amplifier is in the grid driving power requirements when driving the amplifier into grid current. In the audio amplifier, the driver must supply all of the instantaneous power required by the grid at the peak of the grid swing. To deliver this peak power, the audio driver must be capable of delivering average sine wave power equal to one-half of the peak power. In an RF linear amplifier, the tank circuit averages the power of the RF cycle due to its "flywheel" effect, so that the driver need only be capable of delivering the actual average power required, and not the peak. With these reservations in mind, examination of the audio or modulator data of a tube will give a good idea of its RF linear power-amplifier operating conditions.

11-85. Class A RF Linear Amplifiers.

11-86. In low-level amplifiers, where the output signal voltage is less than 10 volts, small, receiving-type tubes, such as the 6AU6, are very satisfactory for class A service. For voltage levels above 10 volts, the 4X150A is the best choice for class A operation, because it has short leads, low plate-to-grid capacitance, and high transconductance. Class A amplifier tubes should be operated in as linear a portion of the plate characteristic curves as is practical. This can be done by inspecting the plate characteristic curves of the tube. Usually the static plate current which results in near maximum plate dissipation is the best. The maximum output voltage should be kept to about one-tenth of the DC plate voltage, or less, to obtain signal-to-distortion ratios of 50 db or better. The DC plate voltage regulation for class A operation is seldom of importance, and cathode bias and screen dropping resistors are commonly used. Even with tubes such as the 6AU6 and the 4X150A, which have short leads and low grid-to-plate capacitance, it is desirable to load the input and output circuits to 5000 ohms when operating up to 30 Mc.

11-87. Class AB RF Linear Amplifiers.

11-88. In the power range from 2 watts to 500 watts, class AB_1 is normally used. This class of operation is very desirable because distortion due to grid current loading is avoided and because high power gain can be obtained. At present, tubes are not available which will give low distortion with good plate efficiency, operating class AB_1 at power levels above 500 watts. Therefore, for higher power levels class AB_2 operation is used.

11-89. For class AB operating conditions with a given screen voltage and given plate load, there is one value of static plate current which will give minimum distortion. The optimum value of static plate current for minimum distortion is determined by the sharpness of cutoff of the plate current characteristic. Grid bias is then set to produce the optimum static plate current. This optimum point is determined from the load line on a set of constant-current plate current curves. Values obtained from this curve are then plotted to obtain the plate current vs grid voltage curve shown in figure 11-21. This curve is the dynamic characteristic of the tube. By projecting the most linear portion of the curve to intersect with the zero plate current line, the grid bias is determined. This point of intersection is often referred to as the projected cutoff. The static plate current which will flow with this grid bias is the proper static plate current for minimum distortion. This procedure is used in audio amplifier design

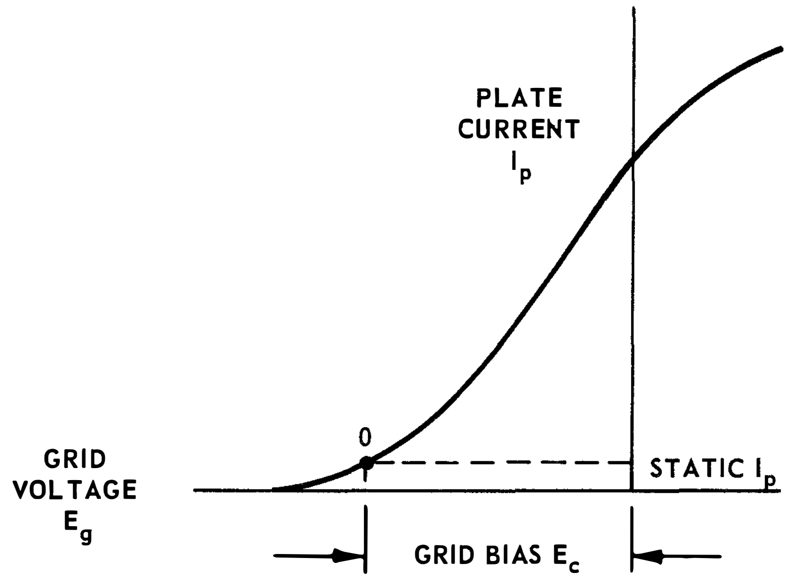

Figure 11-21. Optimum Static Plate Current for Linear Operation.

and is nearly correct for RF linear amplifier design. Perhaps a more accurate procedure for determining the proper bias for RF amplifiers is to choose the point Q so that the slope of the curve at Q is one-half of the slope of the major linear portion. This will allow the amplifier to operate class A with small signals, and deliver power over both halves of the cycle. With a large signal, the tube delivers power over essentially one-half of the cycle. Then, the change in plate current relative to plate voltage swing over half the cycle will be half as much for small signals as it is for large signals, and linear operation is obtained at all signal levels.

11-90. The screen voltage of a tetrode tube has a very pronounced effect on the optimum static plate current, because the plate current of a tube varies approximately as the three-halves power of the screen voltage. For example, raising the screen voltage from 300 to 500 volts will double the plate current. The shape of the dynamic characteristic will stay nearly the same, however, so that the optimum static plate current for minimum distortion is also doubled. A practical limit is reached because high static plate current causes excessive static plate dissipation. In practice, it is found that the static plate current determined by the above method is so high that plate dissipation is near or beyond the maximum rating of the tube when using desired DC plate voltage. For example, one of the better, medium-power triodes for linear amplifier service, the 3X2500A3, requires approximately 0.5 ampere of plate current for minimum distortion. Using a desirable plate voltage of 5000 volts, static plate dissipation is 2500 watts, which is the maximum, rated, plate dissipation for the tube.

For this reason, it is often necessary to operate the tube below the optimum static plate current, which can be done without causing appreciable distortion. In tetrodes, the optimum static plate current is a function of screen voltage, and the high screen voltages required for class AB_1 operation usually require an excessive amount of plate current for minimum distortion. A choice must then be made between operating the tube at lower than optimum static plate current or using a lower screen voltage and driving the tube into the grid current region, a second principal cause of distortion.

11-91. Estimating Tube Operating Conditions.

11-92. The operating conditions of a tube operating class AB in an RF linear power amplifier can be estimated from the load line on a set of constant plate current curves for the tube, as shown in figure 11-22. From the end point of the load line, the instantaneous peak plate current, I_p, and the peak plate voltage swing, E_p, can be established.

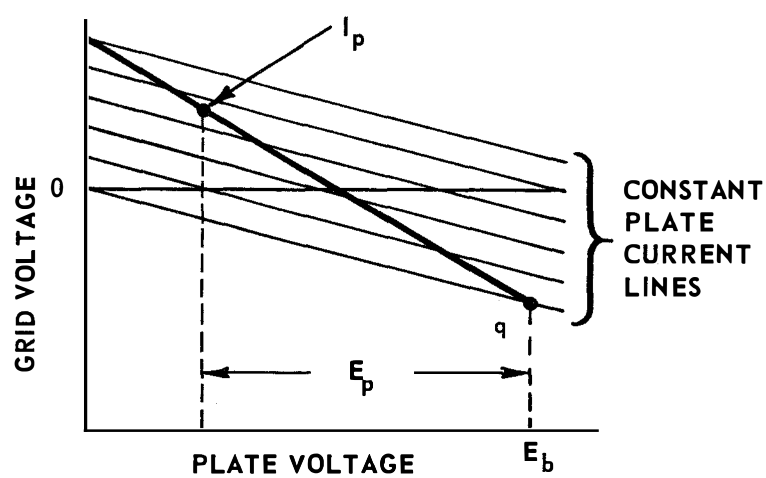

Figure 11-22. Graphic Determination of Tube Operating Conditions.

From these two values, the principal plate characteristics can be estimated by using the following relationships for a single-frequency test signal:

$$\text{DC plate current, } I_B = \frac{I_p}{\pi}$$

$$\text{plate input watts, } P_{in} = \frac{I_p E_B}{\pi}$$

$$\text{average output watts and PEP, } P_o = \frac{I_p E_p}{4}$$

$$\text{plate efficiency, } E_{ff} = \frac{\pi E_p}{4E_B}$$

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

$$\text{DC plate current, } I_B = \frac{2I_p}{\pi 2}$$

$$\text{plate input watts, } P_{in} = \frac{2I_p E_B}{\pi 2}$$

$$\text{average output watts, } P_o = \frac{I_p E_p}{8}$$

$$\text{PEP watts, } P_o = \frac{I_p E_p}{4}$$

$$\text{plate efficiency, } E_{ff} = \left(\frac{\pi}{4}\right)^2 \frac{E_p}{E_B}$$

An actual tube, with moderate static plate dissipation, will have operating characteristics similar to those shown in figure 11-23 for the single-tone and two-tone signals. Plate dissipation and efficiency at maximum signal level are affected but little by even rather high values of static plate dissipation. In practice, the peak plate swing is limited to something less than the DC plate voltage, in order to avoid excessive grid drive, excessive screen current, or operation in the nonlinear plate current region. Most tubes operate with an efficiency in the region of 55 to 70 percent, at peak signal level.

11-93. Distortion.

11-94. Causes of RF Linear Power Amplifier Distortion.

11-95. The principal causes of distortion are nonlinearities of grid current loading and of the amplifier tube's plate current characteristic. In order to confine distortion

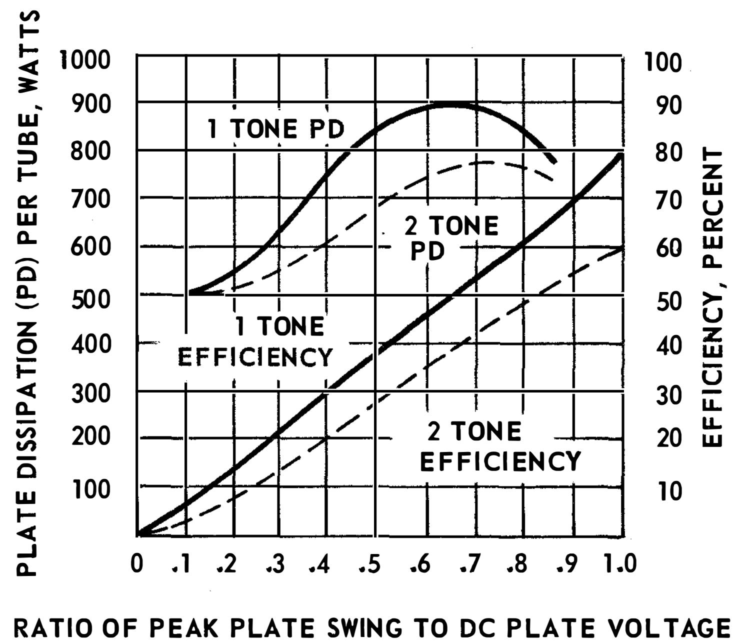

Figure 11-23. Efficiency and Plate Dissipation for Class AB Operation.

generation to the last stage or two in a linear power amplifier, all previous stages are operated Class A.

11-96. The generation of distortion products due to the nonlinear characteristics of the amplifier tube can be derived from the transfer characteristic of the tube, also called the dynamic characteristic, as shown in figure 11-24. The shape of this curve and the choice of the zero-signal operating point, Q, determine the distortion which will be produced by the tube. A power series expressing this curve, written around the zero-signal operating point, contains the coefficients of each order of curvature, as shown in the following expression:

$$I_p = K_O + K_1E_g + K_2E_g^2 + K_3E_g^3 + K_4E_g^4 + K_5E_g^5 + \ldots K_NE_G^N$$

In this expression I_p represents instantaneous plate current, K_1, K_2, etc., the coefficients of their respective terms, and E_g the input grid signal voltage. The values for the coefficients are different for every power series written around different zero-signal operating points. By making input signal E_g consist of two equal-amplitude

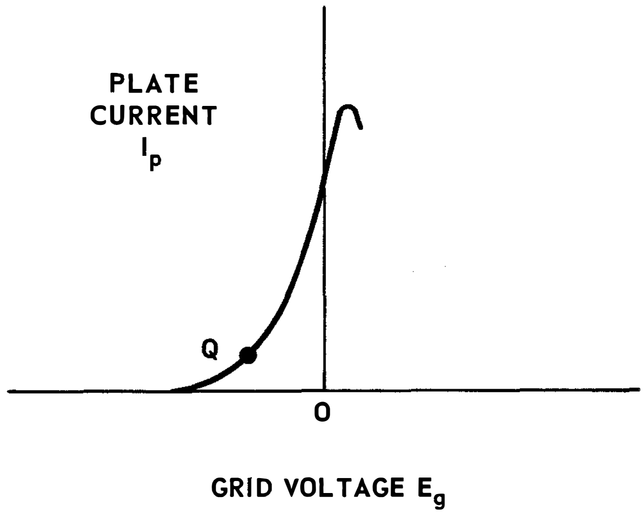

Figure 11-24. Typical PA Tube Transfer Characteristic Curve.

frequencies with a small frequency separation, the distortion products of concern in linear amplifiers can be obtained. Figure 11-25 shows the spectrum distribution of

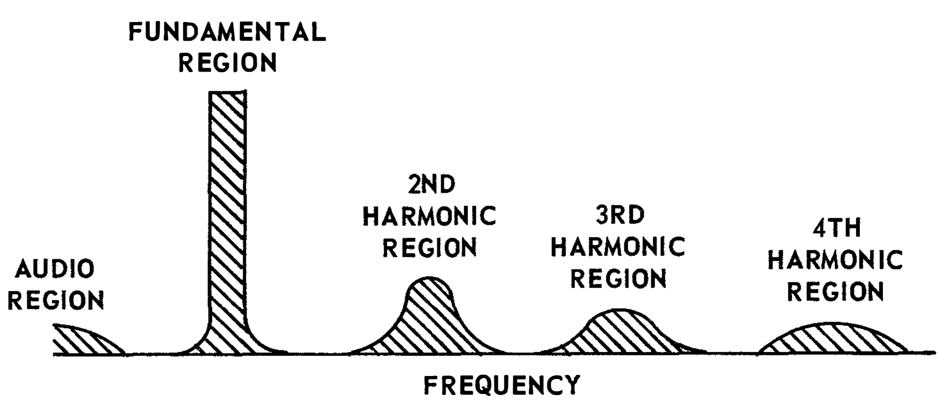

Figure 11-25. Spectrum Distribution of
Products Generated in PA Stage.

the stronger plate current components. It is seen that tuned circuits can filter out all products except those which are near the fundamental input frequencies. This removes all of the even-order intermodulation products and the harmonic products. The odd-order intermodulation products fall close to the original frequencies and cannot be removed by selective circuits. Figure 11-26 shows these odd-order products, which fall within the passband of selective circuits. The inside pair of intermodulation distortion products are third-order, the next fifth-order, seventh-order, etc. The first

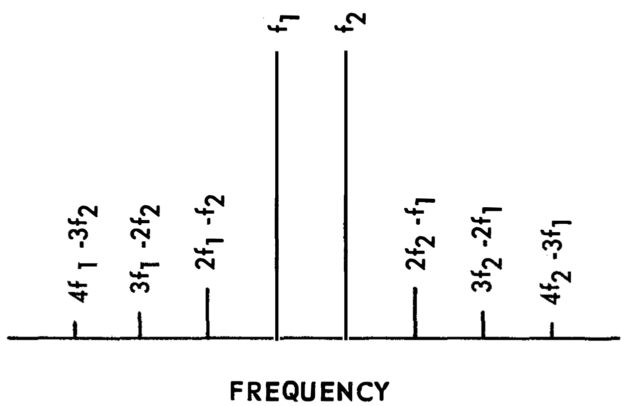

Figure 11-26. Odd-Order Intermodulation Products Causing SSB Distortion.

and most important means of reducing distortion, then, is to choose a tube with a good plate characteristic and choose the operating condition for low, odd-order curvature.

11-97. Nonlinearity caused by grid current loading is a function of the regulation of the grid driving source. In general, this regulation, with varying load, is poor in linear amplifiers. It is common practice to use swamping resistors in parallel with a varying grid load so that the resistance absorbs about ten times the power that the grid of the tube requires. This provides a low, constant driving-source impedance and improves linearity at the expense of increased driving power.

11-98. The instantaneous plate current of all tubes drops off and causes distortion when the instantaneous plate voltage is low. The main reason for this drop is that current taken by the grid and screen is robbed from the plate. In all but a few transmitting tetrodes, the plate can swing well below the screen voltage before plate saturation occurs. However, when the plate swings into this region, the instantaneous plate current drops considerably. If distortion requirements are not too high, increased plate efficiency can be realized by using large plate swings. However, to minimize distortion, the allowable plate swing may have to be reduced.

11-99. Distortion Reduction.

11-100. There is a need for reduced levels of intermodulation distortion from RF linear power amplifiers used in SSB systems. This need exists, not because the distortion noticeably reduces the intelligibility of the SSB signal, but because distortion products outside of the channel width necessary for transmission of intelligence interferes with adjacent channel transmission. The distortion of some of the early SSB power amplifiers was so great that voice channels were placed a full channel width apart to avoid adjacent channel interference. Recent power amplifier developments permit adjacent channel operation using power amplifiers with signal-to-distortion ratios of from 35 db to 40 db. However, power amplifiers with signal-to-distortion ratios of from 45 db to 50 db would further increase the utility of single sideband.

11-101. There are two basic means of reducing distortion to levels better than is obtainable from available tubes. These are RF feedback and envelope distortion canceling.[1] RF feedback is very effective and quite easy to obtain. Ten db of RF feedback will produce nearly 10 db of distortion reduction, which is realized at all signal levels. Envelope distortion canceling has an inherent weakness, because it depends upon envelope detection for its feedback signal. This means that distortion canceling must be instantaneous to be perfect. Since some delay is inherent in the envelope detector and feedback loop, the effectiveness of this circuit depends upon how short the time delay can be made. Development work indicates that a combination of RF feedback and envelope distortion canceling will provide more distortion canceling than either method separately. Using 10 db of RF feedback around all three stages of a 20-kw PEP power amplifier, and signal synthesized from the input envelope to grid modulate the first stage, a better than 50 db signal-to-distortion ratio has been obtained for all distortion products at any signal level up to the 20-kw PEP.

11-102. Linearity Tracer.

11-103. The linearity tracer consists of two SSB envelope detectors, the outputs of which connect to the horizontal and vertical inputs of an oscilloscope, as shown in figure 11-27. A two-tone test signal is normally used to supply an SSB modulation envelope, but any modulating signal that provides from zero to full amplitude can be used. Even speech modulation gives a satisfactory trace. This instrument is particularly useful for monitoring the signal level and clearly shows when the amplifier is overloaded. It can also serve as a voltage indicator, which can be useful for making tuning adjustments. The linearity trace will be a straight line regardless of the envelope shape, if the amplifier has no distortion. Overloading, inadequate static plate current, and poor grid circuit regulation are easily detected with the linearity tracer. The instrument can be connected around any number of power amplifier stages, or it can be connected from the output of the SSB generator to the power amplifier output to indicate the overall distortion of the entire RF circuit. A circuit diagram of an evelope detector is shown in figure 11-28. Any type of germanium diode may be used for the detector, but the diodes in each of the two required envelope detectors must be fairly well matched. The use of matched diodes cancels the effect on the oscilloscope of their nonlinearity at low signal levels. A diode load of from 5000 to

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[1] W. B. Bruene, "Distortion Reduction Means for Single Sideband Transmitters," IRE Proceedings, Dec. 1956.

10,000 ohms minimizes the effect of diode differences. Operation of both detectors at approximately the same signal level is important so that diode differences will cancel more exactly. It is desirable to operate the envelope detectors with a minimum of 1 volt input, to further minimize diode differences. It is a convenience to build the detector in a small shielded enclosure with coaxial input and output connectors. Voltage dividers can be similarly constructed so that it is easy to patch in the desired voltage step-down from the voltage sources. A pickup coil on the end of a short coaxial cable can be used instead of voltage dividers to obtain the RF input signal.

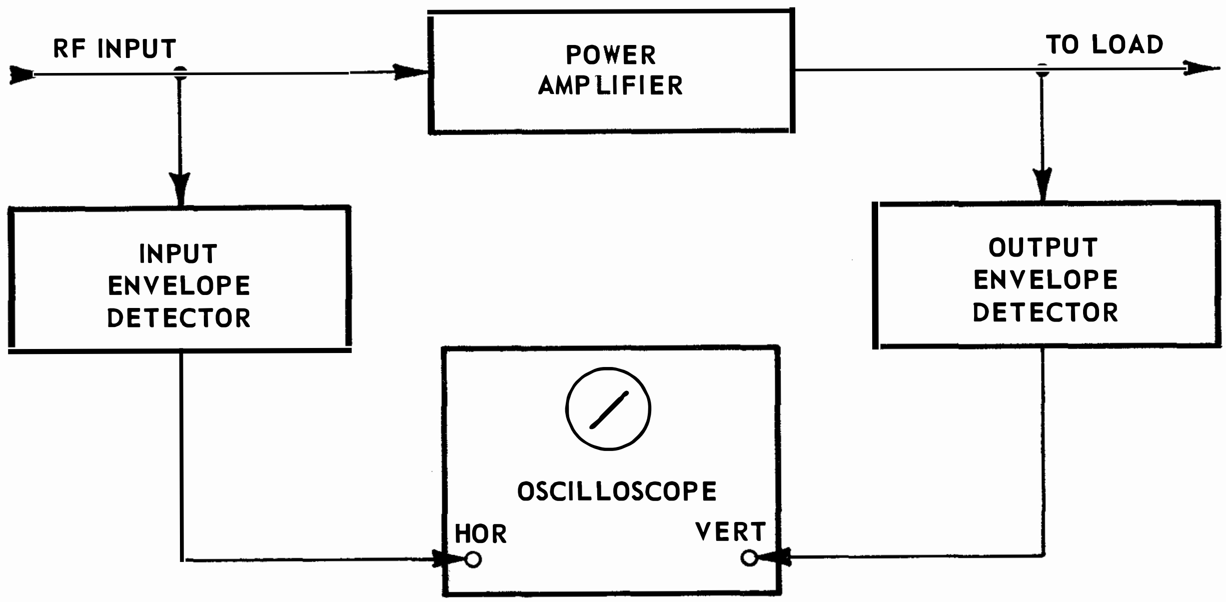


Figure 11-27. Block Diagram, Linearity Tracer.

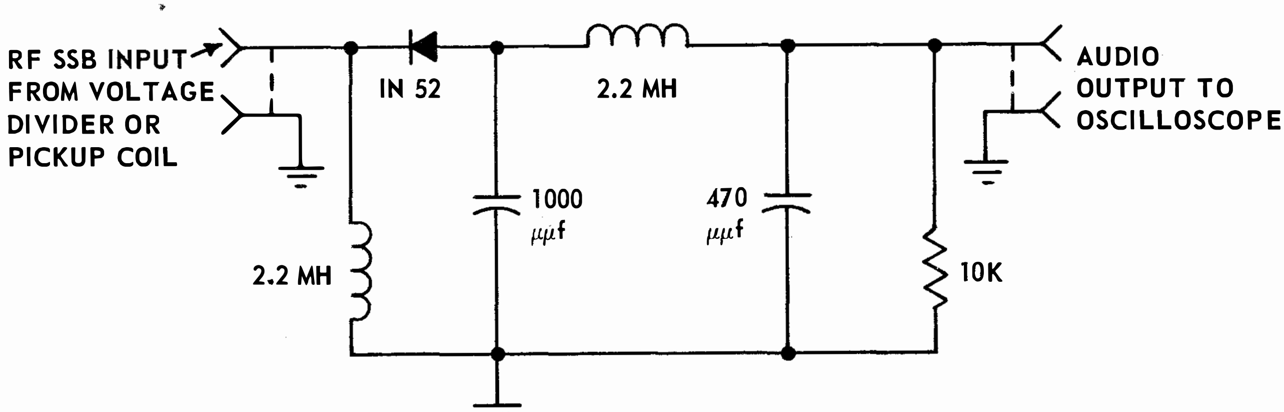


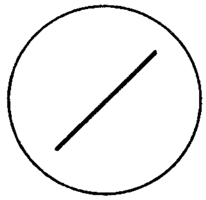
Figure 11-28. Schematic Diagram, Envelope Detector.

11-104. The frequency response and phase shift characteristics of the oscilloscope vertical and horizontal amplifiers should be the same and should be flat to at least twenty times the frequency difference of the two test tones. Excellent high-frequency characteristics are necessary because the rectified SSB envelope contains harmonics to the limit of the envelope detector's ability to detect them. Inadequate frequency response of the vertical amplifier may cause a little foot to appear at the lower end of the trace. If the foot is small, it may be safely neglected. Another effect, which may be encountered, is a double trace; but this can usually be corrected with an RC network between one detector and the oscilloscope. The best way to test the linearity tracer is to connect the inputs of the envelope detectors in parallel. A perfectly straight, diagonal trace on the oscilloscope will result if everything is working properly. One of the detectors is then connected to the other source through a voltage divider. This will not require an appreciable change in the setting of the oscilloscope gain controls.

11-105. Figure 11-29 shows some typical linearity traces which might be observed in linear power amplifier operation. Figure 11-29A indicates proper linear operation. Inadequate static plate current in class A amplifiers, class AB amplifiers, or mixers will result in the trace shown in figure 11-29B. This condition can be remedied by reducing the grid bias, raising the screen voltage, or lowering the signal level through mixers and class A amplifiers. The trace shown in figure 11-29C is caused by poor grid circuit regulation when grid current is drawn, or by nonlinear plate characteristics of the tube at large plate swings. This can be remedied by using more grid swamping or lowering the grid drive. The trace shown in figure 11-29D is a combination of the traces shown in B and C. The trace shown in figure 11-29E is caused by overloading the amplifier. It can be remedied by lowering the signal level.

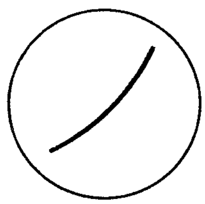
11-106. Phase Discriminator.

11-107. Tuning an amplifier plate circuit tank to resonance consists of tuning for a 180 degree phase shift between grid and plate. By using a phase discriminator, it is possible to compare accurately the phase relationships of the grid and plate signals. This circuit has the advantage of an output which can be used in automatic control or tuning circuits. Figure 11-30 shows a simplified phase discriminator circuit. Notice that the diode, resistor, and capacitor network is essentially symmetrical. The input from the grid circuit is coupled from the inductive branch of the grid tank. Current through L_g is 90 degrees out of phase with the voltage on the grid. (In a resonant circuit, the current through the inductor is 90 degrees out of phase with the voltage across it.) The voltage induced in the discriminator circuit is applied across R_a and R_b, which are equal resistances. The voltage across R_a and R_b is divided equally, so that CR_1 and CR_2 have an equal voltage applied to them. During the half cycle, when point A is positive, CR_1 conducts, and current flows through R_1, C_1, R_c, and R_b, charging C_1 such that the grounded end is positive with respect to point C. During the half cycle when point B is positive, CR_2 conducts and current flows through R_2, C_2,



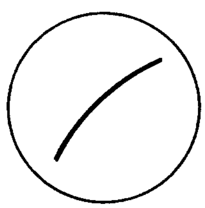
A

PROPER OPERATION



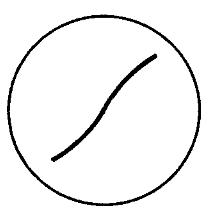
B

INADEQUATE STATIC PLATE CURRENT



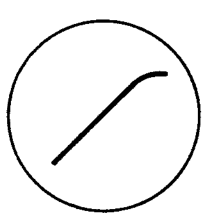
C

POOR GRID CURRENT REGULATION OR NONLINEAR PLATE CHARACTERISTIC



D

INADEQUATE STATIC PLATE CURRENT AND POOR GRID CURRENT REGULATION OR NONLINEAR PLATE CHARACTERISTIC



E

OVERLOADING

Figure 11-29. Typical Linearity Traces.

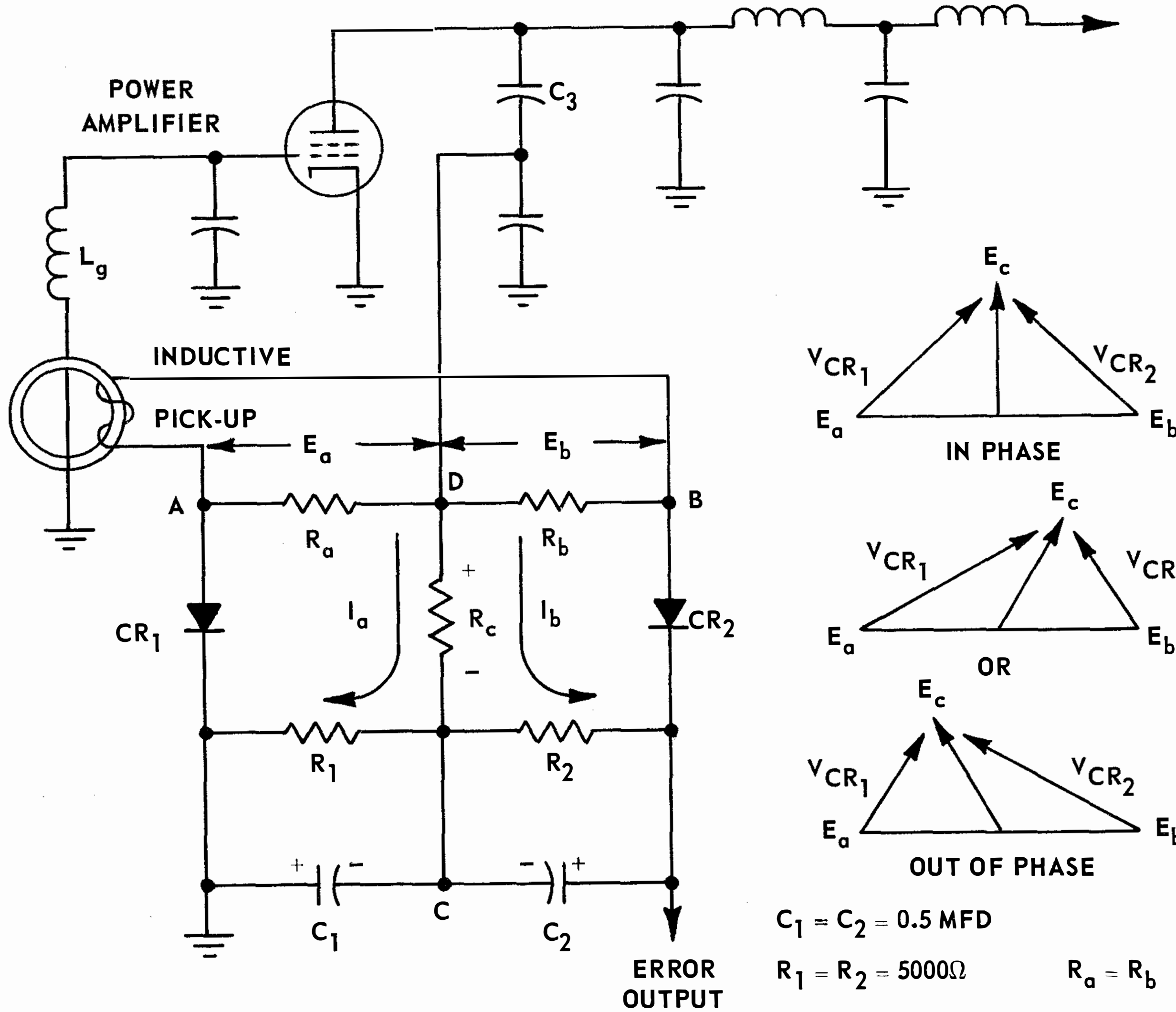


Figure 11-30. Simplified Phase Discriminator.

R_c, and R_a, charging C_2 so that the output end is positive with respect to point C. If the time constants in the charging paths are small with respect to the discharge time constants, capacitors C_1 and C_2 will remain charged between half cycles. Notice that the charging paths of both capacitors is through the resistor R_c, and that during both half cycles the current flow through R_c is in the same direction. This means that the RF signal appearing across R_c is rectified by the diodes so that during each half cycle point D is negative with respect to point C. A graphic representation of this current is shown in figure 11-31. Now let us couple part of the power amplifier plate signal back to point D in the discriminator network. The RF signal is an alternating current flowing between the plate and ground, through C_3, R_c, and C_1, or C2. This signal is

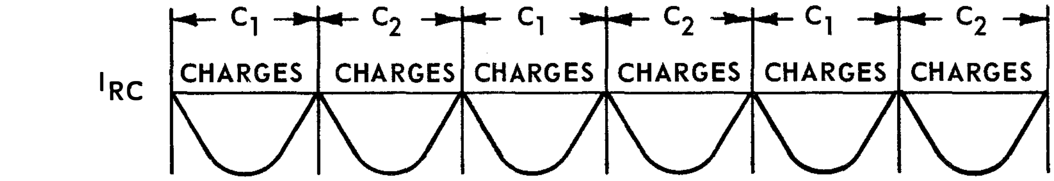


Figure 11-31. Charge Current Graph, Zero Plate Signal.

shown superimposed on the R_c current waveshape in Figure 11-32. If the plate signal is fed back so that it is in phase with C_1 charging current, it will add to the charge on C_1, allowing C_1 to charge higher and C_2 lower. Conversely, if the plate signal is 180 degrees out of phase with the C_1 charging current, C_1 will not charge as high, but C_2 will charge more. Now, if the plate signal is 90 degrees out of phase with both the C_1 and C_2 charging current, both capacitors will charge equally.

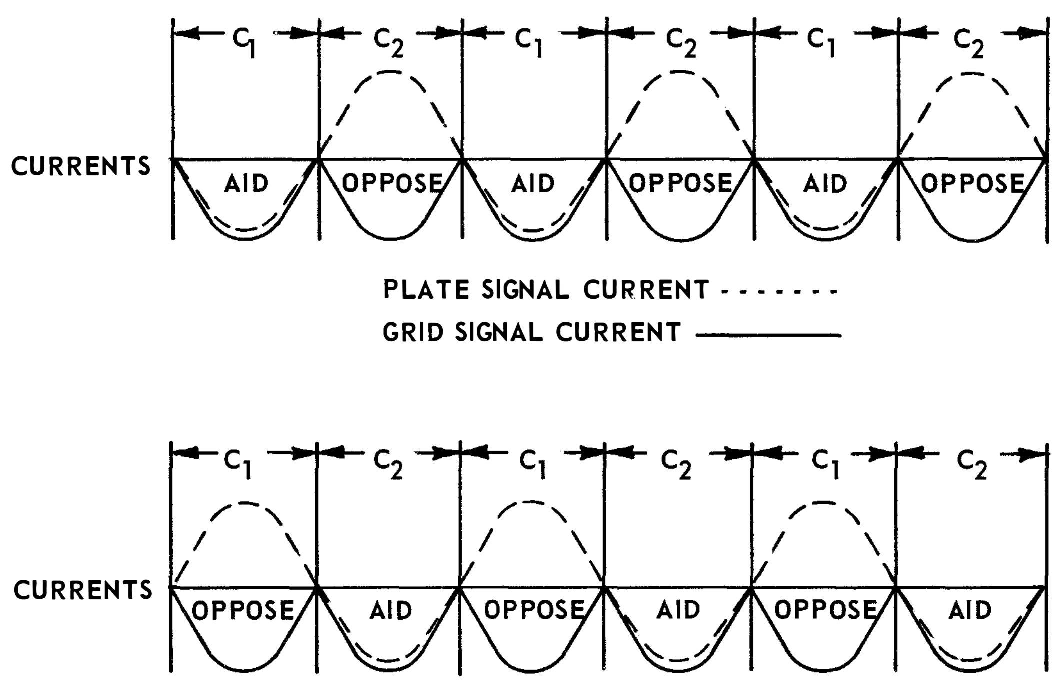
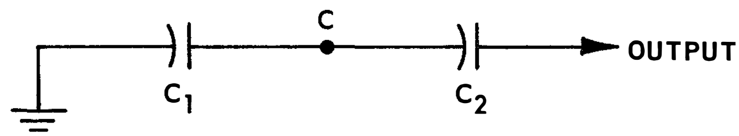


Figure 11-32. Charging Current with Plate Signal.

11-108. The discriminator output is taken off across C_1 and C_2 in series, as shown in figure 11-33.



EXAMPLE (A)	−4V	+4V	= 0V OUTPUT SIGNAL
EXAMPLE (B)	−5V	+3V	= −2V OUTPUT SIGNAL
EXAMPLE (C)	−3V	+5V	= +2V OUTPUT SIGNAL

Figure 11-33. Output Signal.

Example (a) shows the condition of no feedback, resulting in an output voltage of zero. Example (b) assumes that a plate signal is fed back in phase with the C_1 charging time, resulting in a -2-volt output. Example (c) assumes that the plate signal fed back is in phase with C_2 charging time, resulting in +2 volts output. The phase discriminator develops an output DC voltage which is zero when the plate tank oscillations are 180 degrees out of phase with the grid tank voltage. If the plate tank is detuned on either side of 180 degrees, the output will be a DC voltage, either positive or negative, depending on whether it leads or lags the grid signal. In actual power amplifier circuits, this DC voltage can be used to operate a device to correct the tuning automatically. This device is called an automatic tuning control.

SECTION XII

GENERAL MAINTENANCE TECHNIQUES

12-1. Transmitter Maintenance Techniques.

12-2. For proper operation of a communications 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 overall system. Periodic checks and adjustments of the equipment, by qualified personnel will ensure that the overall 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 conditions. 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 determined by the equipment designer, and cannot be controlled by the maintenance technician. However, by employing the correct maintenance techniques (alignment, system measurements, parts replacements, etc.) the maintenance technician can ensure that the performance of the equipment will be optimum, consistent with equipment design.

12-3. A majority of maintenance personnel are familiar with the importance of accurate frequency adjustments and output monitoring of existing AM communications transmitters. Transmitter output monitoring and accurate transmitter frequency adjustments are even more important in single sideband than in AM applications. Not only must the single sideband 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.

12-4. The method employed in testing a single sideband transmitter will vary slightly for different equipment applications. For example, some permanent ground installations have special test units incorporated in the overall system configuration. These test units, often referred to as distortion measuring equipment, are usually designed for the specific application 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.

12-5. No attempt will be made in this text to present a detailed step-by-step alignment procedure for any particular single sideband transmitter because such information can be obtained, and should be obtained, from the equipment manufacturer's publications. Instead, the information presented will cover some of the maintenance techniques peculiar to single sideband transmitters in general.

12-6. RF Carrier Oscillator.

12-7. The relationship of 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 contains none of the message information, it is still required at the receiver for demodulation of the signal. In single sideband suppressed carrier systems a locally generated carrier is inserted 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 the single sideband system, the 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 RF oscillator in the transmitter cannot be stressed too strongly.

12-8. The stability of the 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 regulated power supplies, is the most common method used to obtain good frequency control in 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 ensure that the strict requirements established by the circuit designer are maintained in actual equipment operations.

12-9. The frequency of the 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 or 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 oscillator. Test instruments specifically designed for single sideband applications should be used whenever such instruments are available; sufficient warmup time should be allowed for all test instruments to ensure stable operation of such equipment. If a check using the proper test equipment and procedures reveals that the frequency of the 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 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 adjusted. Usually, a small trimmer capacitor is incorporated in the oscillator circuit designed 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, adjustment 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 RF carrier oscillator.

12-10. Carrier Balance in Balanced Modulators.

12-11. In the transmitter theory portion of this course it was stated that the primary purpose of any balanced modulator circuit is to produce the sidebands of an amplitude modulated RF 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 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 45 db is desirable in single sideband systems, separate bias supplies and RC balance adjustments are usually provided in the modulator to ensure 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 reinsertion circuits and that the amount of carrier due to unbalance of the balanced modulator be kept to a minimum. Attainment of this condition ensures more accurate control of the pilot carrier level by the carrier level control.

12-12. A certain degree of balance is built into a balanced modulator circuit by the equipment designer in the choice of tubes, components and circuit arrangements. To compensate for changes in value of components due to environmental and operational conditions, separate balance controls are usually incorporated in the circuit design. The maintenance technician is responsible for the proper adjustment of these controls to ensure that the proper degree of balance is maintained in the modulator circuit under operational conditions.

12-13. The first step in adjusting a balanced modulator is to check the circuit operating voltages and make any necessary corrections (parts replacement, bias supply adjustments, etc.) for any errors found. When a change of parts is required, caution should be observed to ensure that the replacement parts comply with the values and tolerances 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.

12-14. The carrier balance can be adjusted by applying a test tone of constant amplitude and frequency (usually around 1000 cycles per second) to the input of the single sideband exciter and coupling a suitable AM receiver to the output of the exciter as shown in figure 12-1, or at 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 for controlling the frequency and amplitude of the output signal. The only requirements of the AM receiver are that it have suitable response characteristics in the proper frequency range to make the test.

12-15. After all connections are made, sufficient time should be allowed for warmup of the equipment. The carrier level control should be turned to its minimum position; that is, it should be adjusted so that no carrier is applied to the single sideband signal through the carrier reinsertion circuits, if one is used. 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 modulator exists. The carrier balance control is then adjusted for minimum tone output from the

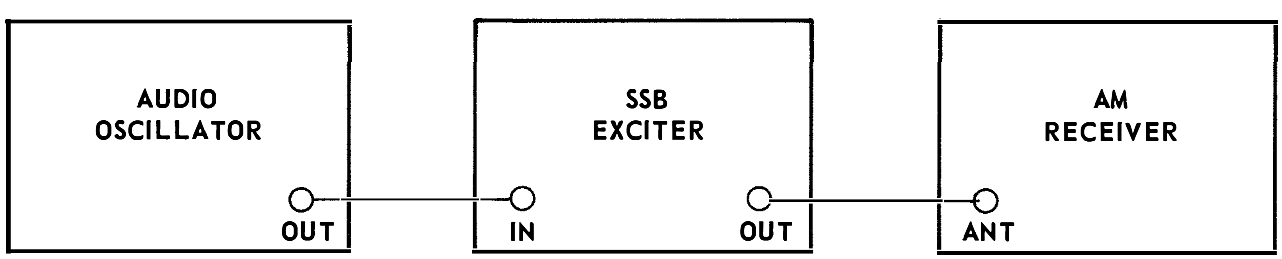


Figure 12-1. Equipment Set Up for Carrier Balance Adjustment.

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 setting 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 the technician allow sufficient amplitude of tone signal to indicate a definite null point when the carrier balance control is being adjusted for minimum signal level.

12-16. The reason for using an AM receiver instead of a single sideband receiver for making this adjustment is that a transmitted 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 to be noted in the received signal. If a single sideband 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 single sideband receiver. In this case, the carrier balance adjustment would have no effect on the received test tone; therefore, such a carrier balance adjustment would be meaningless.

12-17. In the diode rectifier balanced modulator (balanced bridge ring- or lattice-type modulators), an important factor that must be considered is the proper ratio of the RF 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 RF voltage should be on the order of 3 to 6 volts, which is at least 8 to 10 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.

12-18. Sideband Filters.

12-19. The purpose of the sideband filters in a single sideband 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 band of frequencies with minimum loss and distortion, but it is equally important that all frequencies outside these desired bands be attenuated sufficiently so that they do not appear in the output of the transmitter. This fact is most important in dual-channel single sideband systems where information from two unrelated sources is transmitted on opposite sides of a suppressed or reduced carrier.

12-20. As a general rule, the maintenance technician should not attempt to adjust sideband filters. The necessary adjustments are made and sealed during the construction of the filter unit. This fact is especially true of mechanical filters, because some types of filters are hermetically sealed and no external adjustments are possible. Attempts to adjust mechanical filters should not be undertaken under any conditions.

12-21. 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 the 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.

12-22. The importance of the sideband filters in regard to the overall 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.

12-23. Sideband Suppression.

12-24. By using an oscilloscope and an audio oscillator it is possible to closely approximate the amount of undesired sideband suppression. In this procedure (see figure 12-2) a single-tone signal (usually 1000 cycles) from the audio oscillator is applied to the single sideband 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 (see figure 12-3A). If the single sideband 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 as in figure 12-3B. The shape of the multiple-frequency envelope will depend upon the number, amplitude, and wave shape of the separate frequencies present.

12-25. Assuming that the observed signal has the appearance of figure 12-3B, the suppression ratio between the desired and undesired sidebands can be expressed as:

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

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 RF carrier must also be sufficiently suppressed to prevent it from appearing in the

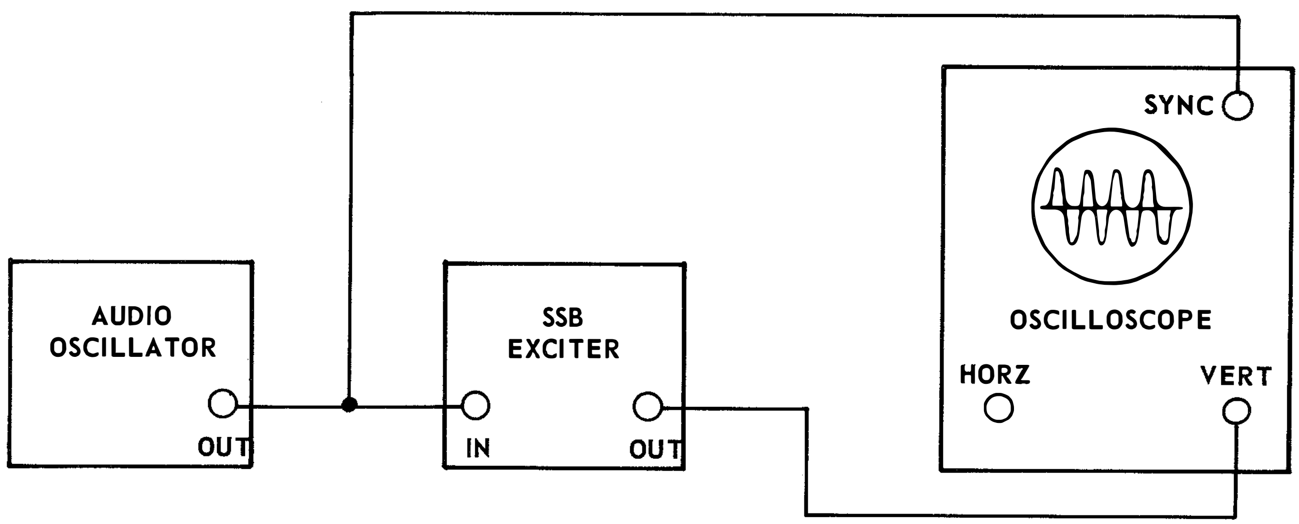


Figure 12-2. Equipment Setup for Measurement of Sideband Suppression.

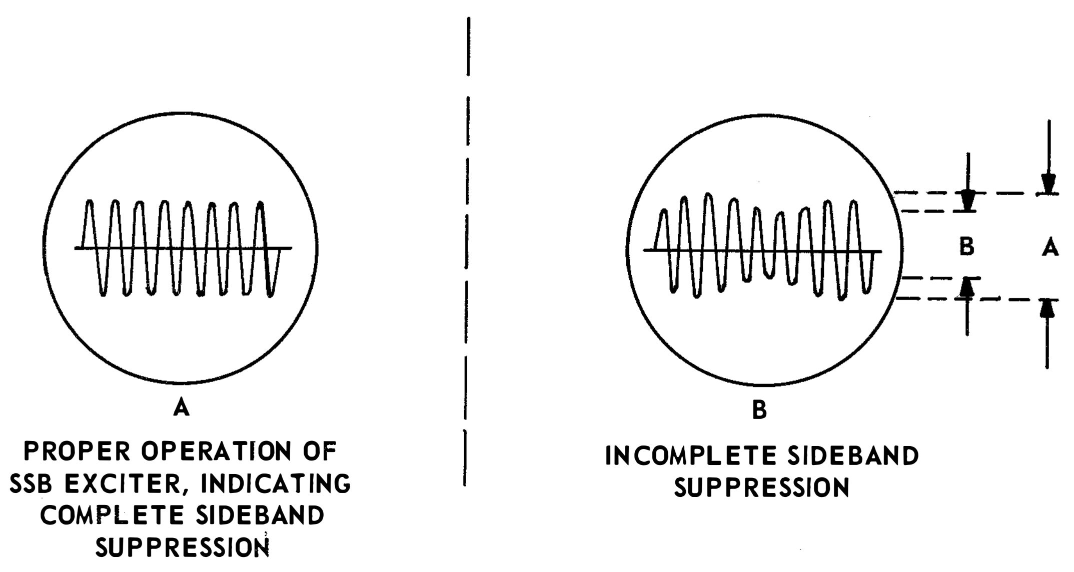


Figure 12-3. Waveforms Observed During Measurements of Sideband Suppression.

output and further complicating the output waveform. The amount of undesired sideband suppression should be approximately 35 to 45 db for proper operation of the system.

12-26. Gain of Low-Frequency Circuits.

12-27. Generally speaking, testing the gain of the low-frequency section of a single sideband transmitter provides a means for determining the gain of the circuits between the input to the transmitter and the medium-frequency circuits. The exact procedure used in making this test will vary for different equipment designs and applications. However, the overall procedure used is basically the same for all types of equipment.

12-28. In order to test the 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 medium-frequency or the high-frequency 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 the tube from its socket. After one of the oscillators has been disabled and the equipment has warmed up sufficiently, the operating voltages of the low-frequency circuits should be checked with an accurate and reliable voltmeter, preferably a vacuum-tube voltmeter. The next step is to apply a test tone (usually 1000 cycles) 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 low-frequency circuits and, in particular, the level of the RF carrier applied to the low-frequency 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 RF 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 medium-frequency and the high-frequency mixers, the gain must be controlled so that the signal to oscillator-injection voltage ratio is of the proper value to ensure linear mixer operation, low noise levels, and minimum spurious products.

12-29. Some procedures use voltmeters to measure the signal voltage, while others use meters to measure grid and plate current in the low-frequency 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 that was removed from the equipment at the beginning of the test.

12-30. Distortion in Single Sideband Transmitters.

12-31. An important requirement of single sideband 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 (second, fourth, etc.) are sufficiently removed from the desired signal so that normal tuning will eliminate them. Most of the intermodulation distortion in linear amplifiers is caused by the odd-order products (third, fifth, etc.) which fall in or near the desired frequencies. Of equal importance are the distortion products which fall outside the single sideband channel, because these products may cause interference in the reception of weak signals in equipment operating on the adjacent channels.

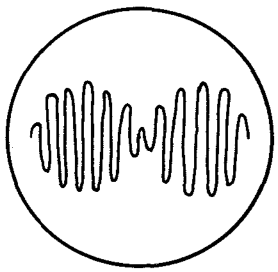
12-32. A form of distortion peculiar to phase-shift single sideband transmitters is called post-phasing distortion. This form of distortion is caused by harmonics generated in the audio amplifiers following the audio phase difference networks and by improper adjustments of the balanced modulators. When 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 produce distortion in the desired sideband. 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.

12-33. Proper choice of tube characteristics and operating conditions by the circuit designer can keep distortion products to a minimum. Although the maintenance technician has no control over these factors, he can ensure that the distortion is kept low by observing proper maintenance procedures. These procedures include proper adjustment of element voltages, balance controls, neutralizing controls, and proper replacement of shields, shielded leads, and feed-through capacitors.

12-34. Two-Tone Test Signal.

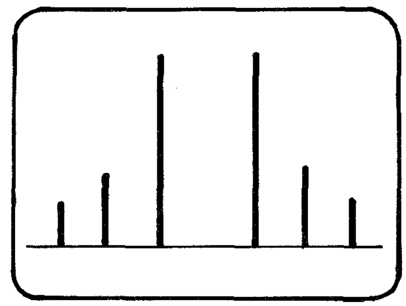
12-35. The most widely used method of testing single sideband transmitters is the two-tone test. This test involves the application of two separate tone signals to the input of a system or circuit and observing the results on an oscilloscope, spectrum analyzer, or some other indicating device. The two tones should be equal in amplitude and have a difference in frequency of about 1000 cycles in order to achieve the results desired from the test. Typical examples of two-tone test waveforms are illustrated in figure 12-4.

12-36. The sources of the test tone can be either RF signal generators or audio oscillators, whichever are applicable to the test being performed. In filter system transmitters, the two-tone test signal can sometimes be obtained by applying a



A

TWO-TONE TEST
OBSERVED ON
OSCILLOSCOPE



B

TWO-TONE TEST
OBSERVED ON
SPECTRUM ANALYZER

Figure 12-4. Examples of Ideal Two-Tone Test Waveforms.

1000-cycle audio tone to the transmitter input and slightly unbalancing the low-frequency 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 single sideband 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 ensure that the amplitude of the two tones is maintained constant. Upon completion of the two-tone test using these methods, the carrier balance should be reset and checked to ensure that it is at the proper point for the operation of the equipment.

12-37. Signal-to-Distortion Ratio.

12-38. The signal-to-distortion ratio, 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 the two-tone test method. Although present designers of linear amplifiers can produce signal-to-distortion ratios of approximately 35 to 40 db, lower overall system distortion values can be obtained by using some form of distortion cancellation, such as RF feedback. Since the principle causes of distortion in linear amplifiers are nonlinearity and grid current loading, care must be exercised to ensure that the linear amplifiers are not overdriven in a single sideband system.

12-39. The signal-to-distortion measurement can be made by applying the two tones at the same level to the input of the transmitter, as shown in figure 12-5, and measuring both direct output level and 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.

12-40. Signal-to-Noise Ratio.

12-41. The signal-to-noise ratio, usually expressed in db, is defined as the level of the desired signal compared to the noise level in a particular application, and is generally a function of the bandwidth of the overall system. In cases where impulse noise is

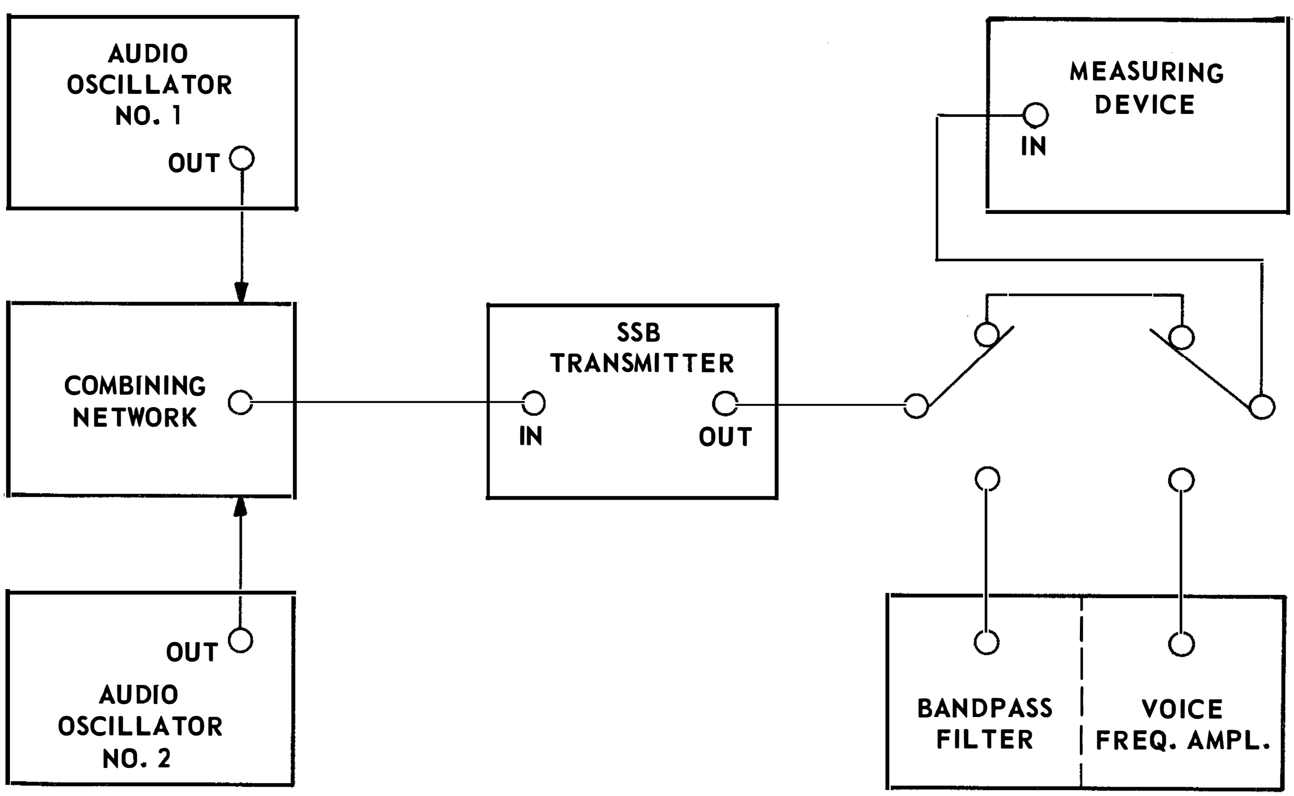


Figure 12-5. Equipment Setup for Measurement of Signal-to-Distortion Ratio, Using Two-Tone Test.

to be measured, the ratio is usually expressed in terms of peak values; that is, the ratio of peak signal to peak noise. In random noise measurements, rms values are generally used; that is, the ratio of rms signal voltage to rms noise voltage.

12-42. 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 consulted for the exact procedure used in checking a particular system.

12-43. Audio-Frequency Response.

12-44. Frequency response is defined as the range of frequencies passed with a specified allowable loss by a system or circuit. In single sideband transmitters, the frequency response is primarily determined by the insertion loss and frequency characteristics of the sideband filters associated with the low-frequency balanced modulators. Proper impedance matching of the sideband filter input and output to the associated circuits is also an important factor in determining the overall system frequency response.

12-45. The methods of measuring the audio frequency response of the single sideband system 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.

12-46. Transmitter Power Output.

12-47. There is a direct relationship between 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 the specific amount of distortion that can be tolerated. If no value of distortion is specified, it is understood that distortion will be kept within the limits considered to be acceptable for the system. The desired relationship, as well as the importance of this relationship, between transmitter power and permissible distortion should not be overlooked when considering the power output rating of a transmitter.

12-48. The power output of a single sideband transmitter is usually given in terms of peak envelope power (PEP), with this term being defined as the rms power developed during the peak RF cycle. Peak envelope power is equal to the sum of the amplitudes of the sideband components and the pilot carrier if one exists. The measurement of PEP is usually made using the two-tone test procedure. With the carrier turned on during the two-tone test, maximum PEP occurs when the peaks of the two tones are in coincidence. It should be noted that a carrier, when used as part of the sideband signal, must be inserted. In a single sideband, suppressed-carrier transmitter, of course, no carrier is used in making the PEP measurements.

12-49. One method of measuring the peak envelope power of a single sideband transmitter is shown in figure 12-6. In this method, the peak envelope voltage (E_p) of the

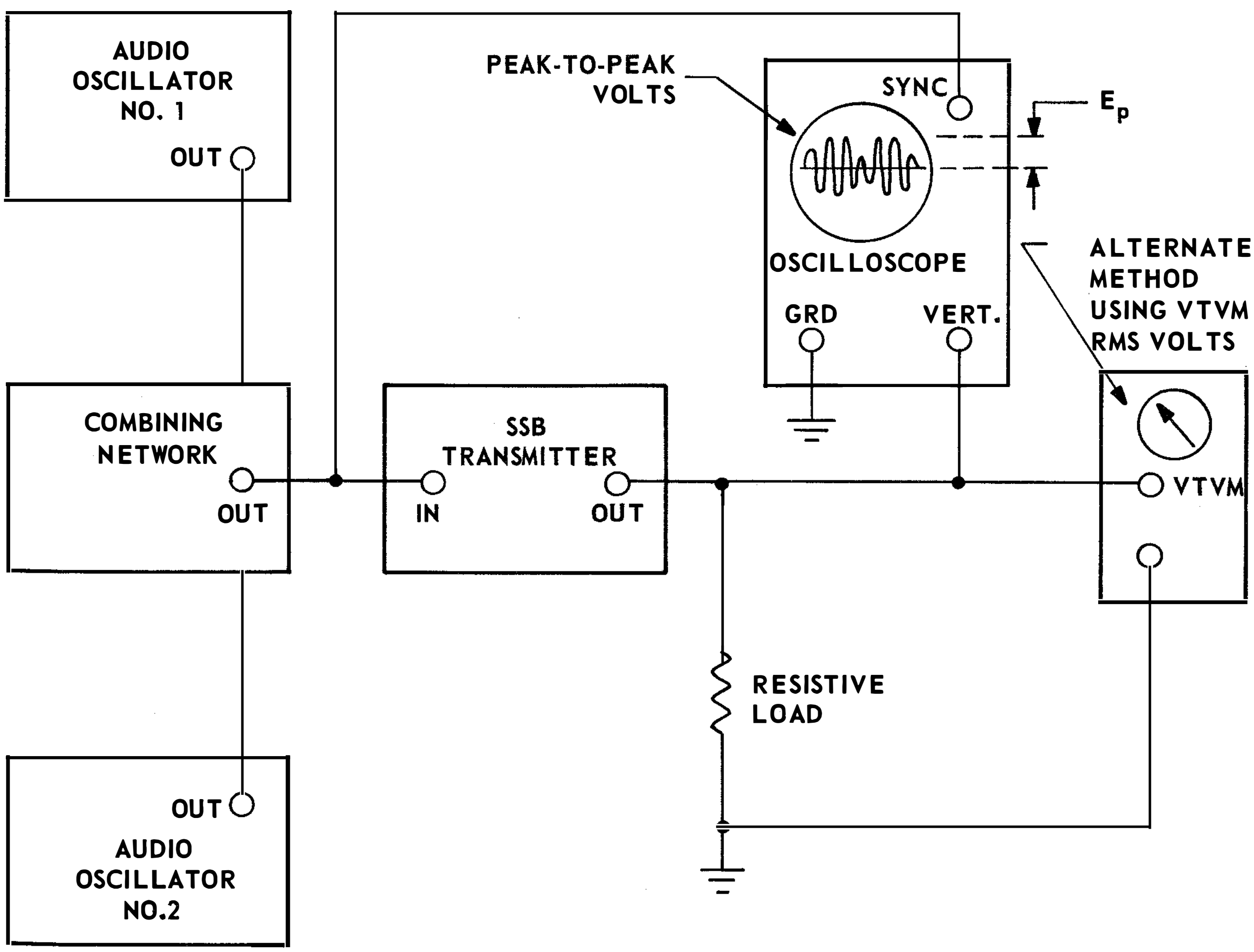


Figure 12-6. Equipment Setup for Measuring Peak Envelope Power, Using Two-Tone Test.

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 peak envelope power can be calculated in the following manner:

$$PEP = (0.707E_p)^2 \div R_l$$

If the voltage (E_l) developed across the load resistor (R_l) is measured on a VTVM calibrated in rms volts, the PEP can be calculated as follows:

$$PEP = (E_l)^2 \div R_l$$

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

12-50. Another rating given to single sideband transmitters is called peak sideband power. Peak sideband power is similar to peak envelope power, except that the measurements of this power term are made with the carrier off (if a carrier is used with the transmission); that is, all the transmitter power is applied to the sidebands and none is applied to the carrier. This also is referred to as "talk power." Talk power is a term often used in reference to single sideband transmitters. Talk 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 single sideband systems, talk power, peak sideband power and peak envelope power of such systems are all synonymous.

12-51. Monitoring and Testing of SSB Transmitters.

12-52. Transmitter monitoring simply means checking 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 samples voltages or currents at different points in the transmitter to ascertain that the 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.

12-53. 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 single sideband transmitters. The main difference is in the type of waveforms observed on the oscilloscope and in the interpretation of these waveforms. In all cases, however, the procedure recommended by the equipment manufacturer should be closely followed where it is available.

12-54. Since the linear amplifiers in single sideband 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 12-7. For proper indications using this method, care should be taken to ensure that the connections are made properly and that sufficient deflection voltage is applied to the deflection plates. To prevent damage to test equipment, it is important to observe all precautions pertaining to the use and operation of oscilloscopes in circuits where high voltages are present.

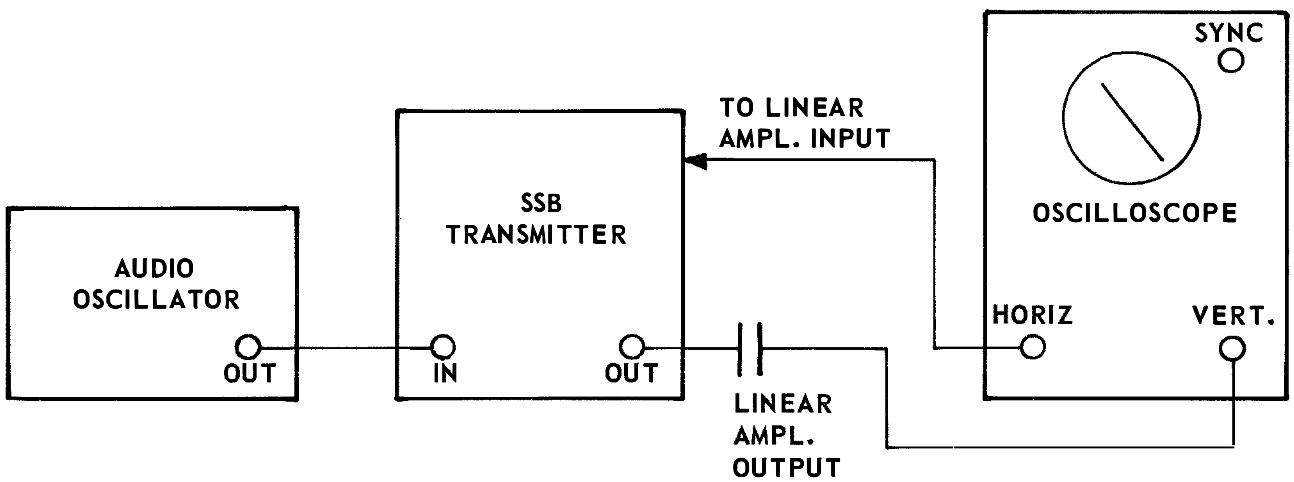


Figure 12-7. Equipment Setup for Monitoring a Linear Amplifier with an Oscilloscope.

12-55. The waveforms illustrated in figure 12-8 are typical examples of the indications which may be expected when a linear amplifier is monitored using an oscilloscope. The waveforms indicated are those which can be expected when using a single tone (including the inserted carrier). Figure 12-8A shows the waveform present when perfect linearity exists between the input and output of the linear amplifier. Figure 12-8B 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 the initial adjustments of the oscilloscope and may not necessarily be the fault of the transmitter. In this case, different RC combinations across one set of the oscilloscope plates should be tried until the desired waveform is obtained. Figure 12-8C indicates peak limiting caused by too much grid drive or not enough amplifier loading. Figure 12-8D indicates too much grid bias or operation of the amplifier on an improper portion of its operation curve. Figure 12-8E indicates a combination of the effects noted in figure 12-8C and 12-8D.

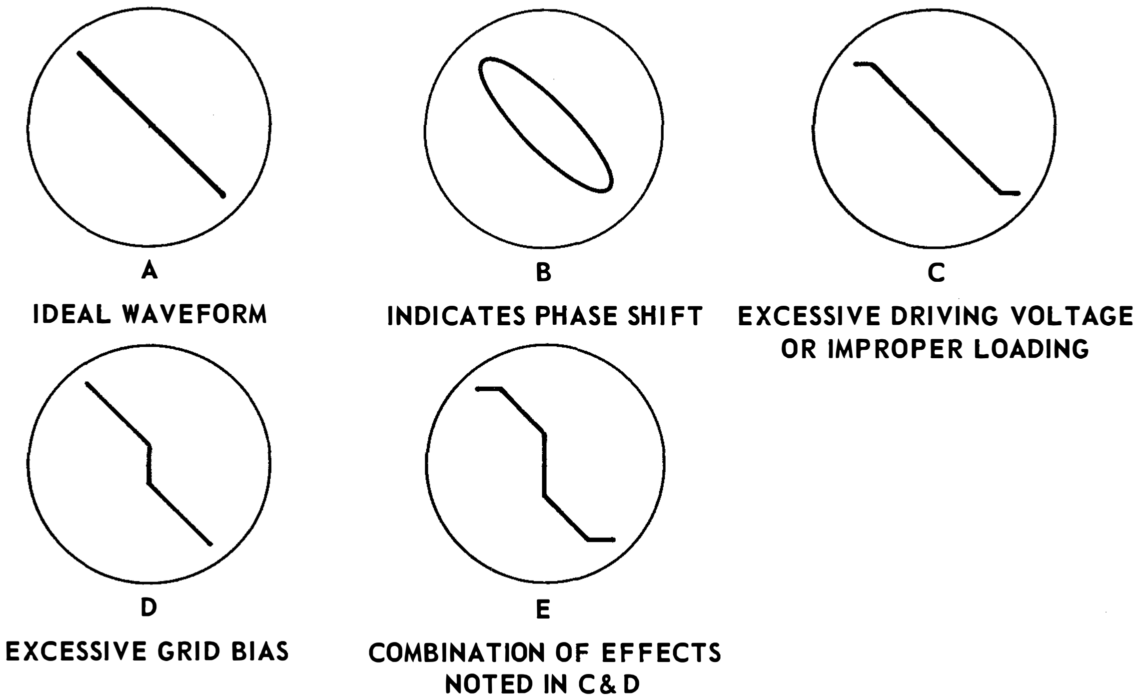


Figure 12-8. Typical Waveforms Observed When Monitoring a Linear Amplifier With an Oscilloscope.

12-56. When the linear power amplifier is monitored or tested using the two-tone test method, perfect linearity is indicated by the waveform in figure 12-9A. Some typical distorted waveforms and their causes are illustrated in the remaining portions of figure 12-9. The ideal waveform is illustrated by the dotted lines in these examples.

12-57. A form of test pattern peculiar only to phase-shift single sideband 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 12-10. 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 MF or HF 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. The double trapezoidal pattern indicates some typical troubles in phase-shift transmitters (see figure 12-11).

12-58. 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

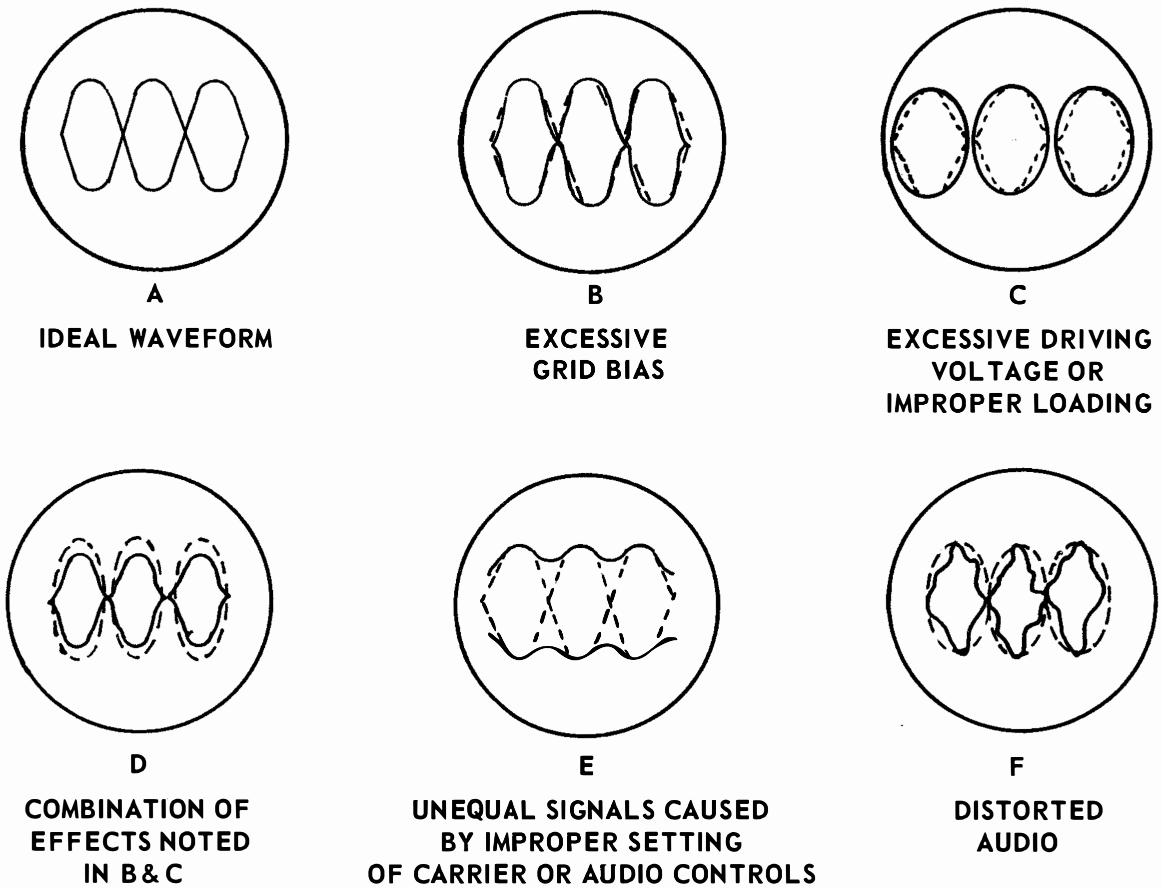


Figure 12-9. Waveforms Observed on an Oscilloscope When Monitoring a Linear Amplifier Using Two-Tone Test.

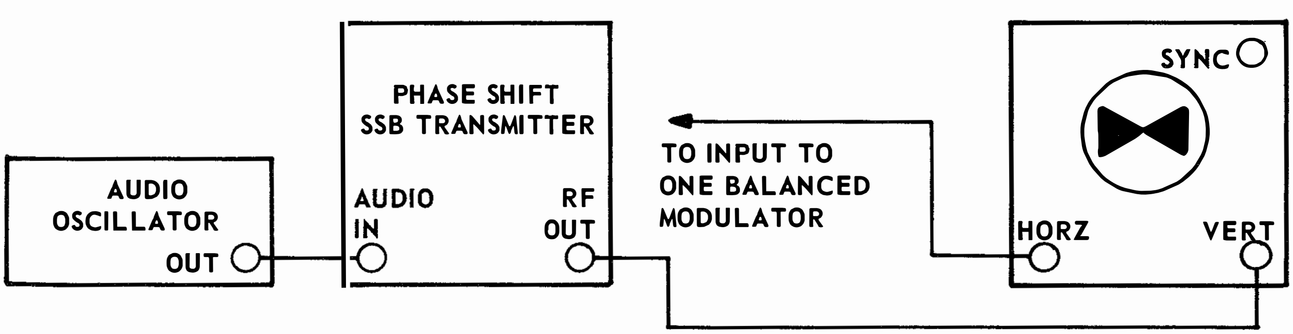


Figure 12-10. Method of Obtaining Double Trapezoid Pattern With the Phase System.

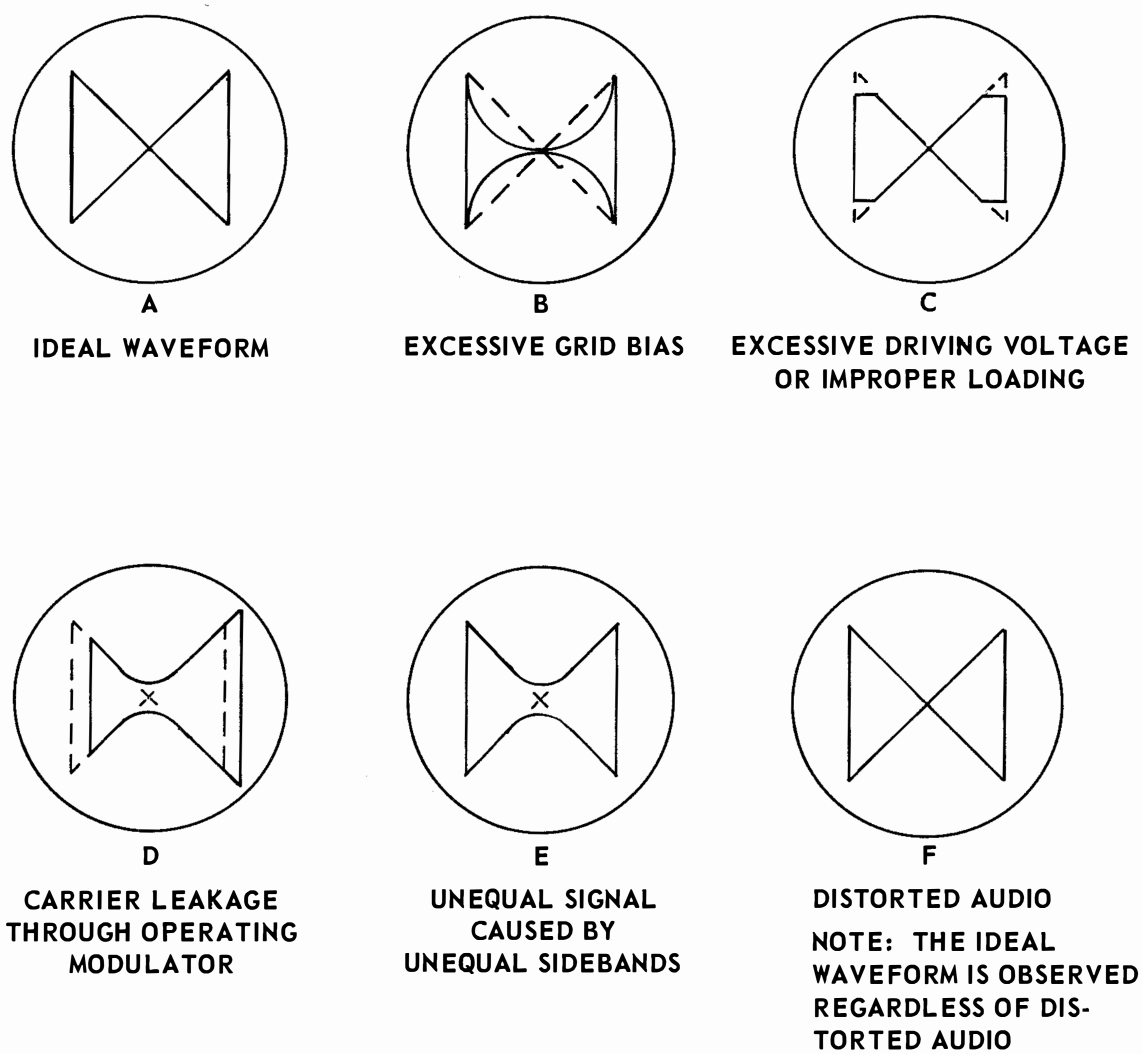


Figure 12-11. Double Trapezoid Waveforms Observed During Phase Shift SSB Transmitter Test.

maintenance of 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.

12-59. Receiver Maintenance Techniques.

12-60. General.

12-61. 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 ensure the design performance of the equipment.

12-62. 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 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.

12-63. Receiver Alignment.

12-64. 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, discriminator tuning becomes somewhat critical; and, in phase-modulated 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 the alignment.

12-65. In equipment employing crystals as either reference generators, oscillators, or filters, the alignment must center around the crystals, since the frequency of the crystals is not variable.

12-66. All of these factors must be considered in 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 the 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 variations in circuits and their layout, actual procedures and alignment specifications must be obtained from the manufacturer's data for the particular equipment involved.

12-67. Test Equipment Requirements.

12-68. Two very important factors must be borne in mind in the choice of test equipment for maintenance or adjustment of single sideband receivers: the close tolerance to which the oscillator frequencies must be held, and the sharp edges of bandpass which must not be defeated.

12-69. Signal generators designed for single sideband applications are presently available. Such a generator should produce a single sideband output signal with provisions 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 which 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. Because of their normal, acceptable frequency error, signal generators should never be used as reference for oscillator adjustments in single sideband transmitters or receivers. Frequency meters or standard reference oscillators, with a known accuracy, must be used for each adjustment. The reference used must have an accuracy 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 frequency error of the reference oscillator should be less than the allowable percentage of error of the oscillator to be adjusted by a ratio of at least ten to one.

12-70. Low-Frequency Reference Oscillator.

12-71. 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 low-frequency oscillator frequency is indicated. However, tube replacements 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.

12-72. The normally accepted standard frequency for low-frequency reference oscillators or carrier insertion oscillators in single sideband 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 already 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 standard frequency signals of WWV prior to use. It must be remembered, however, that the reception of WWV signals on the single sideband receiver, in itself, does not constitute a check of the receiver oscillators, because manual tuning or automatic frequency control action will correct the receiver tuning to compensate for low-frequency oscillator error. The error will then be reflected to appear as a high-frequency oscillator error, or possibly as an AFC error. Therefore, either an external receiver or a high-gain, wide-band oscilloscope should be used to compare the signal of the oscillator with the frequency standard.

12-73. 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.

12-74. In receivers designed for pilot carrier reception employing crystal carrier filters and comparator-type AFC circuits, a different approach should be used. In these receivers, the reference- or carrier-insertion oscillator frequency may 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 AFC 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 warmup time must be allowed for both the receiver oscillators and test equipment before any oscillator adjustment is made, and drift should be checked over a period of at least several hours before a final adjustment is made.

12-75. Heterodyne Oscillators.

12-76. 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 AFC circuits and pilot-carrier reception. Adjustment of the high- and medium-frequency oscillators may be satisfactorily accomplished by following normal AM alignment techniques. It must be remembered, however, that the use of crystal or mechanical filters in single sideband 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 bandpass.

12-77. Filters and IF Stages.

12-78. No attempt should be made to adjust the filters in single sideband equipment except by highly skilled personnel having a thorough understanding of filters. Such attempts should be made only when proper facilities and specialized test equipment are available.

12-79. Credit must be given to radio amateurs who are constantly seeking to improve radio circuits and equipment while operating under the handicaps of limited facilities and equipment and, usually, with very limited funds. Many filters and filter designs may be credited to these economy-minded amateurs. Such filters, however, are not typical of the filters found in military or commercial sideband equipment. Military or commercial filters are usually hermetically sealed or potted and should not be tampered with. After all attempts at proper alignment or circuit tests and adjustment have failed, replacement of a filter may be deemed necessary. Some lattice- or half-lattice 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 conditions.

12-80. 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 micromicrofarads. These often consist of a pair of leads given a slight wrap or twist or perhaps 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.

12-81. 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 completely rejected. Perhaps of equal importance in single sideband receivers is the shielding, especially in the proximity of 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. Decoupling resistors, capacitors, and RF chokes should be inspected when feed-through is evident. The elimination of feed-through is of utmost importance in dual-channel systems to prevent crosstalk or cross modulation.

12-82. IF alignment in single sideband receivers may follow a pattern similar to that in the IF alignment of standard AM receivers, with the exception that the bandpass is determined by the characteristics of the filters involved. IF 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 ensures maximum signal at the output of the receiver or, in the case of the carrier amplifier, maximum AVC voltage.

12-83. Automatic Frequency Control (AFC) Systems.

12-84. AFC systems should be checked for proper control or 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 sticking or binding. Cleaning and lubrication should be performed according to the manufacturer's specifications. Range and control or overall performance of the automatic frequency control system can be checked by manually detuning the receiver suddenly and checking the AFC response. Some mechanical systems employ a damping device, such as an oil-filter dashpot or flywheel to prevent oscillations of the mechanical drive and provide a hold feature to prevent sudden changes in frequency or deep carrier fading from affecting the AFC. These components should be inspected if they are included in the system.

12-85. The frequency limits of the carrier filter almost exclusively determine the range of the automatic frequency control system in pilot-carrier systems. Carrier systems must, of necessity, have a relatively narrow bandpass, usually less than 100 cycles. When the carrier in the IF amplifier falls outside the frequency range of the carrier filter, the carrier will be attenuated far below the level required for AFC operation and the system will lose reference to the carrier. It should, however, maintain control within the range of the carrier filter.

12-86. In the event of improper AFC operation in comparator-type circuits, the frequency of the low-frequency reference oscillator should be checked. This adjustment should be made so that the AFC system will align on the midband noise passed by the carrier filter in the absence of a signal. If the AFC control drifts during this check, the frequency of the reference oscillator may be suspected as being incorrect, after it is ascertained that the AFC balanced modulator or discriminator is properly balanced and adjusted.

12-87. Balanced Detectors (Demodulators).

12-88. 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 balance is not as important as that of balanced modulators in transmitters where the carrier suppression depends on the degree of balance in these modulators. Balanced detectors are used, rather, to ensure 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 RF 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 level 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 frequencies in the carrier circuit 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 tubes or diodes are replaced. 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.

12-89. Balanced modulators in comparator-type AFC systems should be balanced to produce minimum output voltage in the absence of either input signal. Product detectors or multigrid detectors are balanced in much the same manner as balanced detectors. Each signal grid should be operated in the linear portion of the characteristic E_gI_p curve. Balance may be checked by removing or disabling the signal or carrier from the opposite grid and adjusting the element voltage for minimum signal, then repeating the procedure for the opposite grid.

12-90. Carrier Circuits.

12-91. In exalted carrier and single sideband 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 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 thermal and other noise, and results in a lower signal-to-noise ratio in the output of the demodulator.

12-92. Most carrier circuits employ limiting stages combined with several modes of AVC 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 AFC 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 may take control of the AFC 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 AVC and AFC voltages should be checked for amplitude and range.

12-93. Gain and Distortion of Radio Frequency Stages.

12-94. The gain of the RF stages must be kept low in single sideband receivers to prevent overloading the receiver. The primary functions of the RF 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 credited to the RF stages and their malfunctions. These are listed as follows:

(a) 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 RF or mixer stages and producing undesired products at the receiver response frequencies. It is caused by nonlinearity in the RF circuits or tubes, or by overloading.

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

(c) Receiver desensitization or blocking. This condition is caused by the overloading of RF 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.

(d) Receiver spurious response. A spurious response results when frequencies other than the desired signal are converted to the IF frequency. Such a response is due to insufficient RF 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 RF stages or enter the IF stages through improper shielding, etc.

2. Image response. This response is the undesired IF signal, which is produced by heterodyning 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 RF system is good image rejection.

3. Harmonic functions. These functions include responses at submultiple harmonics, 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.

12-95. Most of these effects can be minimized by preventing overload to the RF stages. All of these effects must be eliminated in the RF stages because they cannot be eliminated after introduction in the IF stages. An attenuator is used in many single sideband receivers in the input or antenna circuit. Proper setting of the attenuator is an important aid in preventing overload in most cases. Proper AVC time constants and voltages also help prevent overload.

12-96. Moisture, dirt, corrosion, and poorly soldered joints all affect the selectivity of RF and IF circuits. These conditions should be checked during maintenance. Proper tube element voltages are necessary to prevent nonlinearity in these circuits.

12-97. Elimination of these forms of distortion is important in AM receivers and is even more important in single sideband receivers where improved reception is the intended objective. Overload is especially serious in single sideband receivers because the carrier insertion signal level is held constant and excessive sideband signal level at the detector will result in overmodulation distortion.

12-98. Intermediate Frequency (IF) Stages.

12-99. As in AM superheterodyne receivers, most of the gain and selectivity of single sideband receivers is provided by the IF stages. Also, as in AM receivers, overload and nonlinearity must be prevented in the IF stages. Overload in either the RF or IF stages of a single sideband receiver produces large amounts of intermodulation distortion, which is considered to be the most serious form of distortion affecting single sideband systems. Intermodulation in IF 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 IF frequency range due to the aforementioned RF malfunctions.

12-100. Harmonic distortion, also produced by overload or nonlinearity, is produced in greater amounts in full-carrier systems. It is more pronounced than intermodulation distortion in full-carrier 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 intermodulation distortion becomes more serious. 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 single sideband receivers under overload conditions.

12-101. Other malfunctions, such as motorboating or oscillation, feedback or feed-through, and improper coupling or decoupling, are common to RF stages in all types of receivers and are corrected or prevented in single sideband receivers by conventional methods.

12-102. Audio and Output Circuits.

12-103. 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. Single sideband radio equipment used for transmission or reception of pulse or multiplex signals should be tested using appropriate pulse test equipment. The output of receiving equipment handling such services is usually fed into auxiliary equipment; a thorough knowledge of this equipment is essential in the maintenance of such systems.

12-104. Dual-Channel Systems.

12-105. All of the preceding information also applies to dual-channel systems or double sideband receivers. It is important in such systems that both sideband channels have approximately the same gain. Tube transconductance should be checked in the medium- and low-frequency stages, and both channels of dual-channel systems should be balanced during maintenance.

12-106. Cross talk is perhaps the most serious trouble in dual-channel or multiplex systems. This can be produced by direct feed-through or coupling between channels because of improper shielding or poor isolation of circuits. Shielding should always be replaced, if removed, and bonding or grounding should be checked if feed-through is evident or suspected. Cross talk can also be produced by cross modulation. When the offending channel is the lower sideband of a carrier, the upper sideband of the same transmission is the affected channel, or vice versa. With voice signals, cross talk will be intelligible and will appear as a conversation in the background of the desired signals. Privacy of channels is thus impaired. The same holds true of two transmissions on or very near the same frequency. In this case, four combinations of cross talk, involving both sidebands of both transmissions, 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."

12-107. A form of adjacent channel interference occurs when two transmissions are so close together in frequency that the lower sideband of one transmission falls within the upper sideband of the other transmission. 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; therefore, it 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 position relative to the carrier) with respect to its own (or a different) carrier is demodulated in that form.

SECTION XIII

TEST EQUIPMENT AND TECHNIQUES

13-1. General.

13-2. When a transmitter is operated, three characteristics of the output signal are of prime importance. These are carrier frequency, signal level, and undesired power output. The carrier frequency is that frequency which designates the position in the spectrum occupied by the band of frequencies required to transmit the intelligence. Desired signal level is the RF power confined to those discreet frequencies within this band that are required to transmit the intelligence. Undesired power is the RF power at frequencies, both within this band and outside it, that are not necessary to the transmission of the intelligence and, therefore, interfere with the transmitted signal as well as with other communications channels.

13-3. Frequency Measurement.

13-4. Two commonly used methods of frequency measurement are frequency counter and converter, as well as receiver and frequency standard. A typical frequency counter for measurements up to 100 Mc is the Hewlett-Packard Model 524B with the Model 525A converter installed. Higher frequency measurements can be made with other converters which can be installed in place of the Model 525A. The frequency resolution of this instrument is variable by means of a switch on the front panel to as low as 0.1 cps but, in normal use, a resolution of ±1 cps is sufficient. The period of the count is the reciprocal of the resolution. For example, for a resolution of 0.1 cps the count period is 10 seconds and for a resolution of 1 cps the count period is 1 second. The readout accuracy of the counter is ±1 in the last digit so that with a resolution of 0.1 cps the accuracy of the reading displayed is ±0.1 cps and with a resolution of 1 cps the accuracy of the reading is ±1 cps. If the frequency to be counted is 10 Mc and the resolution is 0.1 cps, then the readout accuracy is ±1 part in 10^8. The overall accuracy of the measurement depends upon the accuracy of the time base standard. For example, if the oscillator producing the time base has an accuracy of 1 part in 10^8 then, with a readout accuracy of 1 part in 10^8, the overall accuracy of measurement is 2 parts in 10^8. The frequency counter is designed to count the frequency of a single sine wave of an amplitude of about 1 volt. It will not give a true indication of frequency in the presence of a complex wave since the digitizingof the data obtained from the incoming signal is accomplished by counting the number of times the instantaneous voltage crosses a given absolute value. Short term errors existing during the period of the count are integrated by the counter, and the average error is included in the frequency display at the end of each count. The internal frequency standard which provides the time base for the count in the Hewlett-Packard Model 525A has a stability of approximately 1 part in 10^6 per day, but provision is made for connection of an external standard if higher accuracy is desired.

13-5. A second method of frequency measurement using a receiver and associated frequency standard is less accurate than the frequency counter method. If a receiver such as the Collins Model 51J is used, the internal frequency standard can be calibrated to a standard of known accuracy, and by using the BFO, the kilocycle dial can be calibrated at adjacent, integral 100 Kc points on each side of the frequency at which the measurement is to be made. The accuracy of the measurement made with this receiver will be in the order of ±250 cps.

13-6. A highly accurate adjustment between two frequency standards can be made by feeding the outputs of the two standards into a receiver and adjusting one standard to agree with the other by observing the beat frequency on the receiver S meter. Where such comparison is made between two frequency standards, the 100 Kc output of one standard can be amplified to about 100 volts and applied to the calibrator crystal socket in the receiver to obtain strong 100 Kc points throughout the range of the receiver. A transmitter whose frequency standard is to be trimmed is tuned to an integral 100 Kc point; the receiver is tuned to that frequency, and the beat between the two is observed on the receiver S meter. The level of input to the receiver from the transmitter is adjusted for maximum swing on the S meter and then the transmitter standard is adjusted for zero beat. When using frequency standards with stabilities of 1 part in 10^8 per day or better, beats having periods of approximately 20 seconds can be obtained when comparing signals at 30 Mc.

13-7. Since the frequency counter will not give an accurate measurement of frequency in the presence of a complex wave and, since measurements of frequency by means of a receiver are confused in the presence of additional signals, it is necessary to disconnect modulation from the transmitter and make all frequency measurements on the reinserted carrier, or on a single tone of known frequency, with stability equal to that of the equipment under test.

13-8. Signal Level.

13-9. Most measurements of power output presume that the level of undesired power output is small compared to that of the desired power output. Normally, the rms sum of undesired power output will be in the region of 35 to 40 db below desired power output, or about 1 percent of desired power. If a suitable resistive load is available to terminate the RF output circuit, a vacuum-tube voltmeter such as the Hewlett-Packard Model 410B will give a reading of voltage across the load which is reasonably accurate for power computation. This meter is a negative-peak-reading meter calibrated in terms of rms. It has a very high-impedance AC probe which will not change the reactive characteristics of a 50-ohm line, provided the meter probe is connected across the line with a minimum of additional shunt capacity or series inductance. When using this method for measuring power, the accuracy of voltage measurement is important since the effect of any error is squared when computing $P = E^2/R$. The most accurate measurement available with the Model 410B is made with a single tone, although experience has shown that measurement of two equal tones on the RF output line will give a voltage indication for computing peak envelope power to an accuracy varying between 5 and 10 percent, the higher accuracy being obtained on the higher ranges of the meter. Table 13-1 shows comparative measurements made with a VU meter which reads slightly above the average value of the applied signal, a Ballantine Model 310A which reads the average value, a Ballantine Model 320 which reads true rms, a Hewlett-Packard Model 410B which reads negative peaks but is calibrated in rms, and an oscilloscope which was used to obtain the peak-to-peak voltage of the signal. Measurements were made on a signal containing from 1 to 16 equal audio tones. In table 13-1 the output level of the amplifier was reset for each reading so that the maximum reading on the VU meter was -3.0 VU. Since beats between tones began to affect the meter readings after more than 2 tones were combined, the minimum and maximum readings were taken each time. Where average readings are given they are the value indicated by the meter for a major portion of the period during which each reading was taken. A signal composed of 16 tones of equal amplitude was adjusted to indicate full scale maximum on the VU meter. Then the tones were dropped one at a time and the maximum readings were taken without further level adjustment. All readings were converted to dbt (decibels with 0 db equal to the indication for a single tone on each type of indicator). The theoretical peak was computed on the basis of a 6-db increase each time the number of tones was doubled and is indicative of theoretical peak envelope power. The true rms indication increases 3 db each time the number of tones is doubled and is indicative of true "heat" power. Statistically, the possibility of the practical peak indication approaching theoretical peak during a given time interval may be computed. The Hewlett-Packard 410B and the oscilloscope agree within a few tenths of a db when read for maximum peak indication over a one-minute interval, yielding the practical peak curve of table 13-1 which is indicative of practical peak envelope power. Note the double inflection in the VU and average curves at the two-tone and three-tone points, the divergence of the VU and average curves from true rms above three tones, and the fact that the VU indication remains almost exactly midway between average and rms throughout the chart.

13-10. The Ballantine Model 310A is an average reading vacuum-tube voltmeter calibrated in terms of rms but does not have as high an input impedance as the Hewlett-Packard Model 410B and has a cutoff frequency of about 2 Mc. However, since its sensitivity extends to less than 100 microvolts, it is useful for measuring low signal levels at IF frequencies. Dummy loads, such as those manufactured by the Bird Electronic Corporation, exhibit a resistance error of the order of ±0.5 ohm and present very little reactance to the signal source. Most of the loads are good for measurements into the hundreds of megacycles. If a resistive dummy load is not available,

Table 13-1. Comparative Meter Readings, 1 to 16 Equal Tones, with Amplifier Output Held Constant on VU Meter

Meter	VU		310A		320	410B			Scope
Reads	Average +		Average		True RMS	Neg Peak			P-P
Calibrated	VU		dbv		dbv	Volts rms			Volts
No. of Tones	Min	Max	Min	Max		Min	Avg	Max	Max
1		-3.0		-1.5	-1.5			0.85	2.6
2		-3.0		-1.7	-0.7			0.23	3.8
3		-3.0		-1.7	-0.8	1.3	1.4	1.5	4.5
4	-4.0	-3.0	-3.0	-1.7	-1.3	1.0	1.4	1.6	5.0
5	-4.0	-3.0	-2.7	-1.8	-1.2	1.1	1.4	1.8	5.5
6	-4.0	-3.0	-3.1	-1.7	-1.0	1.0	1.4	1.9	5.3
7	-4.0	-3.0	-3.0	-1.6	-0.95	1.1	1.5	2.1	6.0
8	-4.0	-3.0	-2.7	-1.8	-1.0	1.1	1.5	2.2	6.2
9	-4.0	-3.0	-3.0	-1.7	-1.0	1.1	1.5	2.3	7.0
10	-3.7	-3.0	-3.0	-1.8	-1.0	1.1	1.6	2.3	7.0
11	-3.7	-3.0	-3.0	-1.5	-1.0	1.05	1.5	2.4	7.0
12	-4.0	-3.0	-2.7	-1.8	-1.0	1.1	1.6	2.3	7.0
13	-3.9	-3.0	-2.7	-1.7	-1.0	1.1	1.6	2.2	7.0
14	-4.0	-3.0	-2.8	-1.8	-1.0	1.1	1.6	2.5	7.2
15	-4.0	-3.0	-3.0	-1.7	-1.0	1.1	1.6	2.5	7.2
16	-4.0	-3.0	-2.8	-1.7	-1.0	1.1	1.6	2.4	7.2

the measurement of voltage or current will not give an accurate indication of power. In these circumstances the calorimetric method of measuring power will give a more accurate indication. The flow meter and thermometers may be calibrated separately to high accuracies and errors in reading of the flow meter or thermometers do not appear in a squared term in the calculation of power as do the errors in voltage or current readings. The accuracy of the temperature readings, however, is affected by the temperature at which the calorimeter is operated with respect to the ambient temperature.

13-11. Another device for measuring power, which does not require a special load and which can be used while feeding an antenna, is a directional wattmeter.

$$\text{SWR} \quad \frac{1+\sqrt{\dfrac{\text{REFLECTED POWER}}{\text{FORWARD POWER}}}}{1-\sqrt{\dfrac{\text{REFLECTED POWER}}{\text{FORWARD POWER}}}}$$

The wattmeter will measure average power integrated as affected by the time constant of the metering network and, if both forward and reflected power are measured simultaneously, the formula will hold for multiple tones. An envelope-detecting voltmeter will read 45 percent of peak voltage when a two-tone RF signal is applied, assuming that the voltmeter reads the average level of the RF and the rms level of the envelope. Thus, 0.707 of 0.637 is 0.45 or 45 percent of peak envelope voltage. A modification of directional wattmeters to approach a measurement of peak envelope power is useful for speech and other complex single sideband signals, since these quantities are of greater importance in single sideband equipment than average power. With extra capacity to lengthen the time constant of the voltmeter circuit, it is possible to make the meter read approximately 0.8 of the peak envelope power so that it will follow speech peaks more closely.

13-12. Undesired Power Output.

13-13. Two types of undesired power may be present in the output of a transmitter. These are spurious response outside the passband; and intermodulation together with incidental amplitude and angle modulation products in or near the passband.

13-14. Spurious responses outside the passband consist mainly of harmonics of the desired output frequencies, products of frequency synthesis, and broadband noise from lower level stages amplified by the power amplifiers. The most direct method of measurement of this type of undesired response is the receiver/signal generator substitution method shown in figure 13-1. A portion of the transmitter output is sampled to provide approximately 1 or 2 RF volts at desired signal frequencies

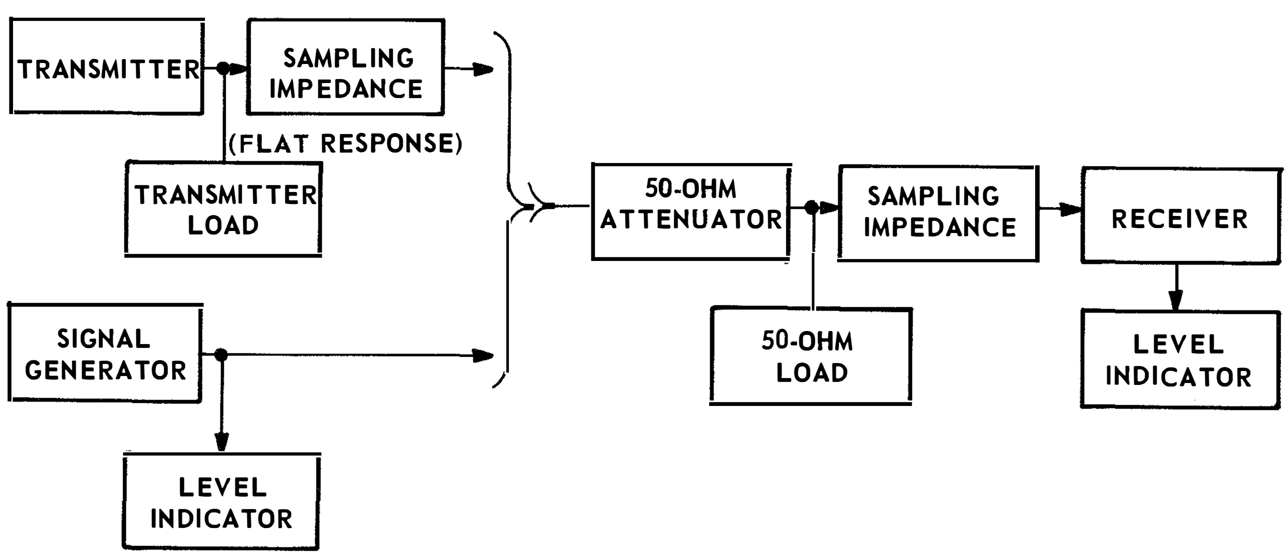


Figure 13-1. Receiver/Signal Generator Substitution.

through a 50-ohm attenuator to a 50-ohm load. When measurements are to be made, the transmitter is operated to provide carrier only or one sideband of modulation by a single tone of known frequency. The receiver is tuned to the transmitter output frequency and the 50-ohm attenuator is adjusted to obtain a convenient reference point on the receiver level indicator. The signal generator is then substituted for the transmitter and tuned for maximum indication on the receiver level indicator. Then the signal generator output, the 50-ohm attenuator, or both, are adjusted to obtain the same indication on the receiver level indicator as was obtained with the transmitter. The reading on the signal generator level indicator, corrected to compensate for the attenuator setting, is the amplitude of the signal across the 50-ohm load which was equivalent to that obtained from the transmitter, and is the reference or zero db indication for measurements of spurious products. The transmitter is then reconnected to the attenuator and operated exactly as before, but the receiver is tuned to a known point in its frequency range where a harmonic would appear, or a search for a spurious response is made. After the receiver is tuned for maximum level indication at a spurious response, the attenuator is adjusted to obtain a convenient reference point on the receiver level indicator. Once again the signal generator is substituted for the transmitter and tuned for maximum indication on the receiver level indicator at the frequency where the spurious response was found. The signal generator and attenuator are adjusted to obtain the same level indication at the receiver as produced by the spurious response and the voltage output of the signal generator, corrected to include the attenuator setting, provides a second reading of amplitude across the 50-ohm load. Comparison of the two readings gives the relative amplitude of the undesired response with respect to the desired signal. The accuracy of these measurements is determined by the accuracy with which the amplitude indication at the receiver level indicator is duplicated when the signal generator is substituted for the transmitter. If the receiver and signal generator are suitably isolated from any interconnecting paths such as the power line, are suitably shielded from each other, and the transmitter is shut down completely when measurements are made with the signal generator, it is possible to obtain reliable and repeatable measurements about 120 db below the desired output of the transmitter.

13-15. Since it is most practical to generate single sideband by generating the sideband frequencies at an intermediate frequency and then heterodyning these signals to the desired RF output frequency, products of output frequency synthesis may appear at the output of the transmitter. These products may be the actual heterodyning frequencies used to translate the intermediate-frequency single sideband signal to the RF output frequency or they may be mixer products of this process. Normal equipment specifications for these products are that they shall be from 70 to 80 db below the desired output of the transmitter. Low order harmonics of the output frequency are often specified at 50 to 60 db below the desired output.

13-16. Since the Q of the output circuits of a reasonably efficient RF power amplifier will be in the vicinity of 10 to 12, broadband noise generated in or amplified by the RF power amplifier stages will not be affected appreciably by the selectivity of the output tank circuits. In a linear power amplifier with three or four stages, this noise will be due primarily to thermal and shot noise in the lower level stages. The effects of this noise may be observed on a spectrum analyzer or on a highly selective receiver. It is not normal for this noise to interfere with communication on the channel of the transmitter causing the noise; however, if the noise is particularly severe, its broadband nature may cause it to interfere with adjacent channels several channel widths removed.

13-17. The second type of undesired power output includes those spurious responses inside or very near the passbands of frequencies including the intelligence to be transmitted. These in-band spurious products are caused by intermodulation distortion resulting from operation of mixers or amplifiers beyond their capabilities; amplitude or angle modulation resulting from imperfect stabilization of the oscillator or synthesizer from which translating frequencies are derived. In addition, the following characteristics are important to a good single sideband signal: a. suppression of opposite sideband; b. suppression of carrier; c. minimum compression of desired signal due to power amplifier loading.

13-18. Two equal-amplitude audio tones have been a standard test signal for distortion measurements because: a. one signal is insufficient to produce intermodulation; b. more than two signals result in so many intermodulation products that analysis is impractical; c. tones of equal amplitude place more demanding requirements on the system than it is likely to encounter in normal use. Any two tones will serve for this test but with many frequency relationships, intermodulation products and harmonics tend to merge, making identification of these products impossible. A 3-to-5 frequency ratio will alleviate this problem. Tones having a more complex ratio may produce products with frequency relationships more suitable to certain tests, but these products will be more difficult to identify than those of the more simple ratio. The following chart shows the relationships between products produced by distortion of an upper sideband of 300 Kc modulated by 3-Kc and 5-Kc tones. (See table 13-2).

13-19. The idealized spectrum analyzer pattern for a two-tone single sideband signal will consist of three discreet frequencies as illustrated in figure 13-2.

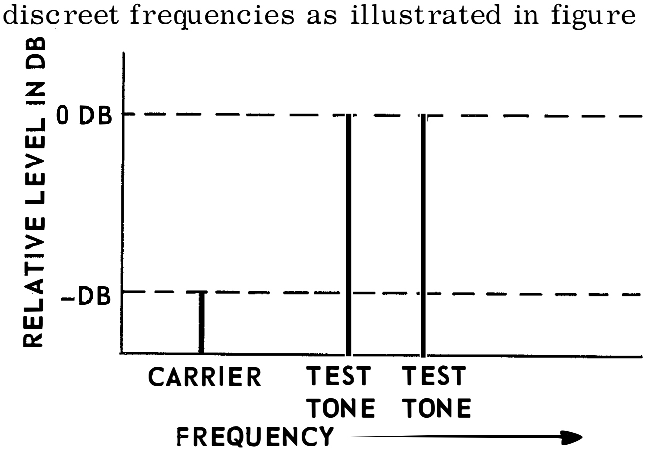


Figure 13-2. Idealized Spectrum Analyzer Pattern.

Table 13-2. Single Sideband Distortion Products

Frequency (Kc)	Order of Products							
		7	6	5	4	3	2	1
295								OSB
296								
297		IM						OSB
298								
299				IM				
300								CAR
301				AIM		IM		
302			AIM				AD	
303		AIM						3 Kc DT
304					AIM			
305		AIM						5 Kc DT
306			AIM				AH	
307				AIM		IM		
308							AS	
309				IM		AH		
310			AIM				AH	
311		IM						
312					AIM AH			
313		AIM						
314			AIM					
315				AH		AH		
316					AIM			

LEGEND:

A = Audio
AD = Audio Difference
AH = Audio Harmonic
AIM = Audio Intermodulation
AS = Audio Sum
CAR = Carrier
DT = Desired Tone
IM = Intermodulation
OSB = Opposite Sideband

These are the frequencies of each of the two audio test tones, translated to the desired RF output frequency, and the carrier (which should be suppressed to the required level). The amplitudes of all undesired products and the carrier are measured in terms of db below either of the two equal amplitude test tones. Practical circuits always have some degree of intermodulation distortion which appears in the form of new, discreet frequencies above and below the two test tones, as illustrated in figure 13-3. The spacing between each tone and the adjacent intermodulation products and the spacing between subsequent intermodulation products is equal to the spacing, F_1, between the two tones. The intermodulation products first adjacent to the desired tones are the third-order intermodulation distortion products; the next pair of products are the fifth-order intermodulation distortion products spaced equally outside the third-order intermodulation products; the next pair are the seventh, then the ninth and so on. The order of a distortion product is the sum of the coefficients in the frequency expression. For example, the third-order intermodulation products will be two times the frequency of one desired tone minus the frequency of one tone minus two times the frequency of the other. The odd-order products fall in or near the desired transmission band and are, therefore, the most objectionable because, once generated, they cannot be eliminated by either the transmitter or receiver. The signal-to-distortion ratio is the ratio of either of the two desired test tones to the largest undesired product expressed in db. A signal-to-distortion ratio of 40 db is usually acceptable for high-frequency communication systems when the equipment is tested on a two-tone basis. Unless unusual cancellation exists in the power amplifier, the third-order intermodulation products will be largest and the higher order products will be progressively smaller.

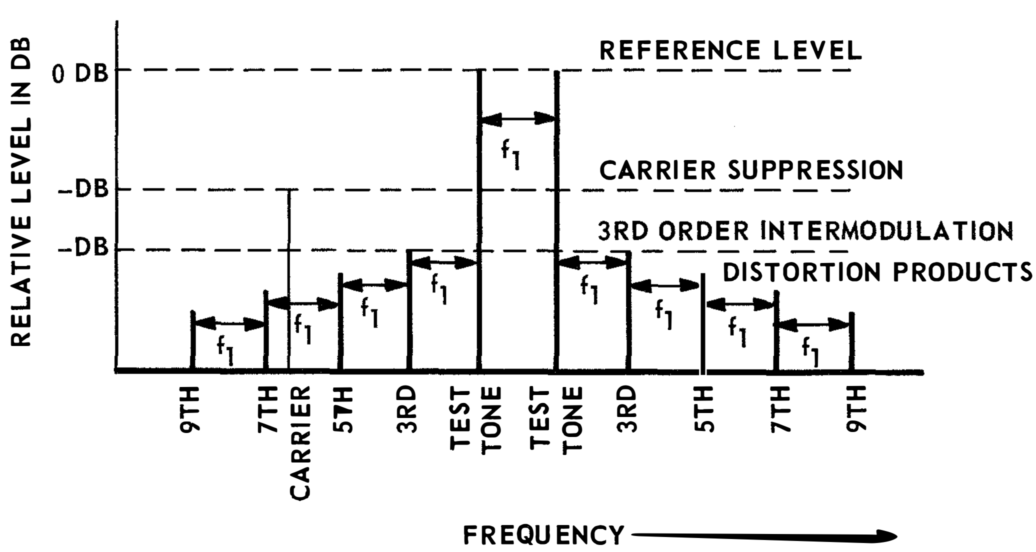


Figure 13-3. Practical Spectrum Analyzer Pattern.

13-20. Overall distortion resulting from several cascaded stages of amplifiers, modulators or mixers may be computed if the distortion of each stage is known. It is useful to note that each stage may be a "black box" actually composed of multiple stages, and one of the black boxes may be the distortion analysis equipment itself. To obtain the overall distortion in a system composed of several cascaded stages the following formula applies:

$$db_t = 10 \left\{ 10 - \log \left[\log^{-1} \left(10 - \frac{db_1}{10} \right) + \log^{-1} \left(10 - \frac{db_2}{10} \right) + \ldots\ldots + \log^{-1} \left(10 - \frac{db_n}{10} \right) \right] \right\}$$

where

db_t = Total or overall signal-to-distortion ratio in db

db_1 = Signal-to-distortion ratio of first stage in db

db_2 = Signal-to-distortion ratio of second stage in db

db_n = Signal-to-distortion ratio of nth stage in db

log = Logarithm with a base of 10

$\log^{-1}$ = Antilogarithm with a base of 10

This equation yields the following typical results from two stages:

Difference between signal-to-distortion ratio between the two stages (in db)	Amount by which the overall signal-to-distortion ratio is degraded beyond that of the poorer stage (in db)
0	3.0
3	1.8
6	1.0
10	0.4

13-21. Any of the oscillators in a transmitter or receiver, or in an analyzer, may have amplitude or angle modulation caused by such deficiences as power supply ripple, alternating currents in tube heaters, mechanical vibration, or strong electric or magnetic fields in the vicinity of the oscillators or their control devices. This incidental modulation causes new sidebands to be produced by the transmitter. These may be observed on the spectrum analyzer display as responses symmetrically located on either side of all desired tones. Each distortion product will also exhibit these sidebands. When the oscillator used in the frequency scheme of a transmitter is modulated, the sidebands produced thereby are often unequal in amplitude because of simultaneous angle and amplitude modulation. An analysis of this phenomenon is summarized in an article entitled "Linearity Testing Techniques for Sideband Equipment" by Icenbice and Fellhauer in The Proceedings of the IRE, December, 1956. Phase modulation distorts the amplitude symmetry of the two sidebands produced by a single sine wave by simultaneously angle- and amplitude-modulating a carrier, or other desired output signal, by subtracting a component from one sideband and adding it to the other.

13-22. Spectrum Analyzer Model 478R-1.

13-23. The most informative and universal method of measuring in-band spurious response is by means of a spectrum analyzer such as the Collins Model 478R-1. In this analyzer, a complete picture of the spectrum in the vicinity of the intelligence passband is plotted directly on an oscilloscope screen or may be recorded by means of a two-axis recorder. The problem of construction of this equipment was primarily to reduce intermodulation within the analyzer to a level appreciably below that to be measured. Although the analyzer is large and complex, the signal under test passes through only two tubes before detection by the narrow-band selective amplifier. These two tubes are both mixers, and much of the remainder of the equipment is devoted to ensuring that the injection of translating frequencies applied to these mixers is sufficiently free from noise, distortion, and incidental angle modulation.

13-24. The basic circuit of the spectrum analyzer is that of a wave analyzer for measuring frequencies generated by signals passing through an amplifier or mixer, or other system, with an unknown amplitude transfer characteristic. Figure 13-4 is a block diagram of the Model 478R-1. Additional selectivity, variable sweep width, and other features permit accurate and simultaneous measurements of level in db versus frequency of distortion, hum, noise, and other spurious products in a direct plot on the analyzer screen, or on a two-axis recorder. Included in the analyzer is a two-tone audio generator, consisting of two audio oscillators and filter-mixer panel. This portion of the analyzer generates a two-tone, audio test signal for use as an audio input intermodulation distortion measurement of the equipment under test.

13-25. The two-tone mixer panel satisfies several requirements for minimizing harmonic and intermodulation products in the two-tone audio test signal. One such requirement is sufficient isolation between the two audio signal generators to reduce intermodulation distortion in the output tubes of one generator because of coupling to the output of the other. This isolation is provided by pads in the output of each generator ahead of the mixing circuit. Second and higher order harmonics in the

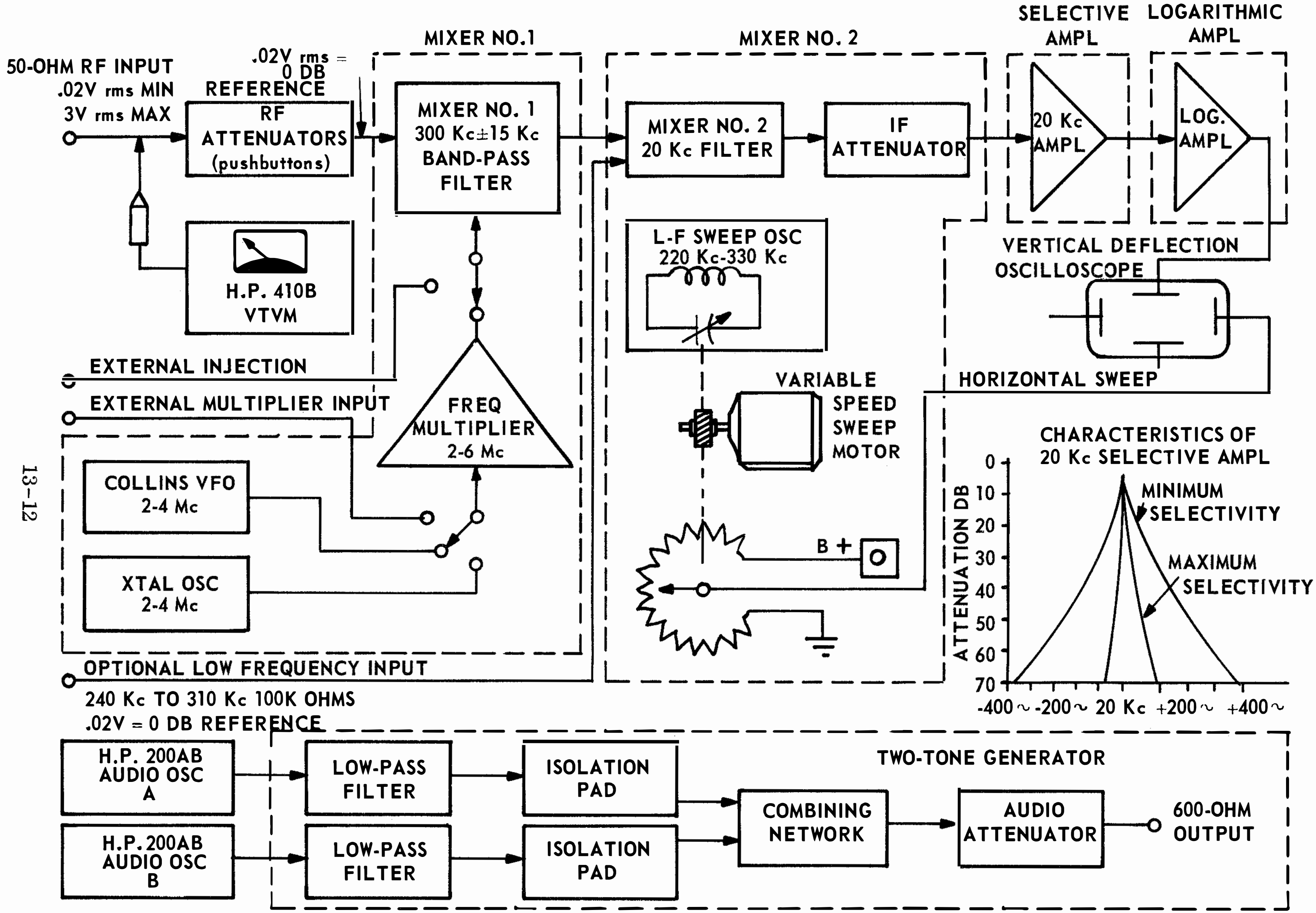
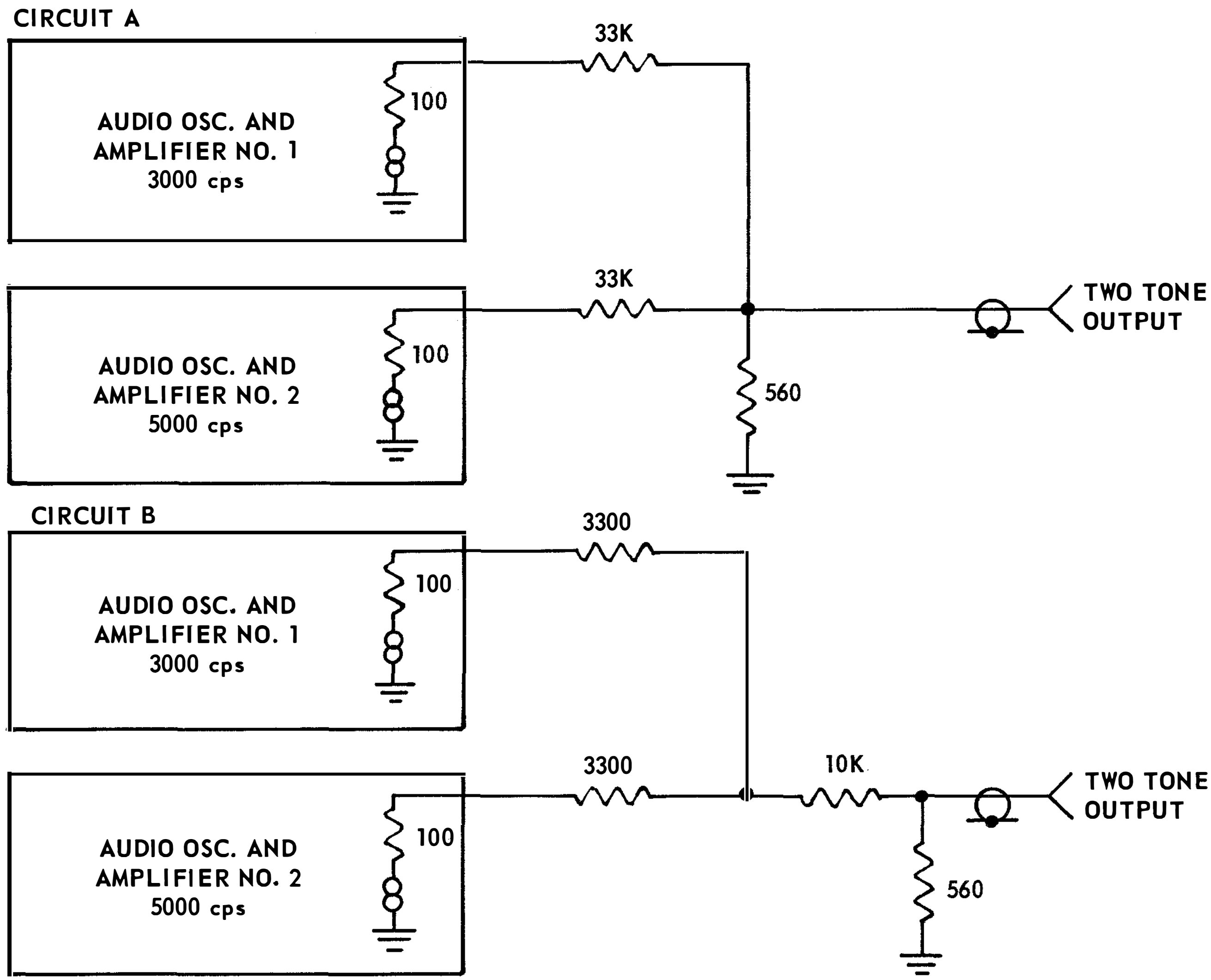


Figure 13-4. Block Diagram, Spectrum Analyzer Model 478R-1.

output of the generators are attenuated by plug-in, low-pass filters selected to have cutoff frequencies between the fundamental frequencies and the second harmonic frequencies of their respective generators. Second-order intermodulation distortion which appears to be third-order intermodulation results from direct mixing of the second harmonic output of one audio generator with the fundamental of the order. This effect is minimized by the low-pass filters. The difficulty of mixing the output of two signal generators so that they do not modulate each other is illustrated in figure 13-5. Both circuits A and B have the same output level but circuit A has approximately 50 db more isolation between the oscillators. The level of third-order intermodulation products is approximately 20 db higher in circuit B than circuit A. While the attenuation of the audio output signal from oscillator No. 1 is affected primarily by the output series resistor and the 560-ohm load, the attenuation between audio oscillator No. 1 and No. 2 is affected by the output series resistor of oscillator No. 2 and the internal generator impedance of oscillator No. 2 as well as the attenuation between oscillator No. 1 and the output.

13-26. During spectrum analysis of a transmitter, products observed by the analyzer can be more readily identified by removing one or the other of the audio tones and observing the effect on the intermodulation products of interest. The ON-OFF switches in the two-tone audio generator are provided with dummy loads in the OFF position to preserve the impedance termination on the audio mixing circuit and thereby prevent a change in the amplitude of the remaining tone when one tone is switched off. The audio filters may be switched out of the circuit to allow the use of audio frequencies beyond the range of the filters, and a set of decade attenuators is provided to enable rapid and accurate testing of equipment with different amplitudes of audio input.

13-27. The dynamic range of the analyzer is 70 to 80 db, displayed on one scale to an accuracy of ±1 db. Continuous metering circuits are provided on the front panel to ensure correct mixer injection level. The analyzer will accept a frequency spectrum with a center frequency from 1.7 Mc to 64.3 Mc and from 240 Kc to 310 Kc without additional coils or test equipment. The spectrum display is on a 17-inch cathode-ray tube. The signal to be analyzed is fed into the precision attenuator panel where it may be monitored by the internal vacuum-tube voltmeter, Hewlett-Packard Model 410B. The attenuator is adjusted to the proper level and the attenuated signal is applied to mixer No. 1 which converts the signal to the 300 Kc IF. The output of mixer No. 1 passes through a 300 Kc bandpass filter to mixer No. 2 where it is converted to 20 Kc. Mixer No. 2 can accept directly any frequency from equipment under test between 240 Kc and 310 Kc. The tuning capacitor of the injection oscillator for mixer No. 2 is rotated by a variable speed motor to sweep the frequency of this oscillator through the required range. Sweep widths of 4, 8, and 16 Kc are available. The output of mixer No. 2 is fed through a precision attenuator with 0.1 db steps to a narrow-band 20-Kc selective amplifier. The half bandwidth of the selective amplifier at 40 db below maximum response can be varied from 30 to 145 cps by a control on the front panel. The output of the 20-Kc selective amplifier is fed to a logarithmic amplifier. The output of



CIRCUIT	FREQ (cps)	3RD ORDER LEVEL (DB)
A	1000	75
	7000	75
B	1000	52
	7000	58

Figure 13-5. Two-Tone Generator, Source Isolation.

this amplifier is a DC voltage that is a logarithmic function of the input over a 70 db dynamic range. The calibrated attenuator between mixer No. 2 and the 20-Kc selective amplifier provides for checking the linearity of the logarithmic amplifier and the oscilloscope to ensure an accuracy of ±1 db throughout the 70-db dynamic range of the analyzer. The varying DC output from the logarithmic amplifier is applied directly to the vertical deflection amplifier of the oscilloscope or to an external recorder. Synchronized horizontal sweep voltage is provided by a potentiometer ganged to the oscillator sweep tuning capacitor.

13-28. The signal path in the analyzer includes only three nonlinear devices ahead of the 20-Kc selective amplifier, after which nonlinearity causes no further intermodulation. The first nonlinear device in the signal path is the Hewlett-Packard Model 410B VTVM probe, but the loading of this high-impedance probe on the 50-ohm circuit is so slight that negligible intermodulation distortion results. Mixer No. 1 and mixer No. 2 consist of only one tube each, and their operating characteristics have been very carefully selected to minimize intermodulation distortion. Microammeters on the mixer front panels provide continuous monitoring of injection grid current to ensure that the mixers are always operated under optimum injection-level conditions.

13-29. An ideal panoramic display of a constant carrier with no modulation would appear as a single line at right angles to the frequency axis. However, in practical equipment this display is a single plot of the selectivity of the analyzer. If the selectivity of the analyzer is changed, the displayed shape of the same carrier under test will change to the new shape of the selectivity curve of the test equipment. Signals under test which have sidebands or intermodulation products to be observed by the analyzer will produce individual responses corresponding to each of these sidebands, or products, together with the desired responses themselves and each response will be basically the shape of the analyzer selectivity curve, each with its own maximum amplitude. The maximum response from each discreet frequency is the required measurement. When these discreet frequencies or sidebands are spaced only a few cps apart, as may be encountered with hum modulation, their corresponding responses on the analyzer screen tend to merge into each other. The responses, for example, of hum modulation will appear on the skirt of the responses to the carrier for that modulation. The ability to separate such discreet frequencies is known as the resolving power of the analyzer. Maximum resolving power is attained when the equipment is operated with minimum sweep width, minimum sweep speed, and maximum selectivity. Since this mode of operation reduces the speed with which data may be obtained, provision is made for varying all three parameters so that data requiring less resolving power may be obtained more rapidly. With maximum selectivity, the speed of the sweep and the sweep width may be adjusted so that the frequency is swept through the response frequency of the analyzer so rapidly that the effective Q or selectivity of the analyzer will not allow the signal to build up to its peak amplitude before the sweep has passed this frequency. This error is always present with any reasonable amount of selectivity, but the effect will be negligible if the sweep width and sweep speed used are commensurate with the selectivity to which the analyzer is adjusted. The easiest check to ensure a safe sweep speed is merely to reduce speed about one-half and note whether the amplitude increases. If the peak amplitude increases more than 1 db, the original speed was too fast.

13-30. When single sideband suppressed-carrier equipment is under test, it is helpful to take the intermodulation distortion test data in steps of about 3 db. This is accomplished by inserting 3 db of attenuation in the output of the two-tone mixer by means of the audio attenuators and removing 3 db of attenuation from the RF input to the analyzer by means of the RF attenuators. This preserves the same amplitude of desired tones on the analyzer, thus allowing observation to be concentrated on the changes in the intermodulation products. In this way, the point on the distortion-versus-signal curve, at which the equipment under test should be operated, can be established rapidly. The intermodulation products will increase relative to the desired output signals at an ever increasing rate, as the rate signal level of balanced modulators, mixers, or amplifier stages in the RF equipment is approached. Near the overload point, intermodulation products commonly increase at a rate 2 or 3 db faster than the desired output signals. Beyond this point the rate of increase of the ratio of intermodulation distortion products to desired output signals becomes much more rapid. If the audio input signal level is reduced well below the normal level, the rate of increase of the amplitude of the intermodulation distortion products will be reduced until, at very low levels, the intermodulation distortion will change imperceptibly with respect to the desired output signals.

13-31. Angle Modulation Measurements.

13-32. In order to make use of the resolution available from the selectivity of the Model 478R-1 spectrum analyzer in measurements of hum-phase modulation sidebands, special precautions were taken to minimize hum-phase modulation on the injection oscillators and in the signal path through the analyzer. Electronically regulated power supplies were used to reduce hum ripple on the plate voltages to less than 1 millivolt; wherever possible, cathodes were operated at ground potential; and plate circuits were either shunt fed, or a pair of tuned circuits were coupled, to ensure that the grids of the following stages were grounded with respect to hum frequencies. In critical circuits, such as in the variable frequency oscillator, where the preceding methods were not applicable, filaments were supplied with direct current, circuits and components were selected to minimize microphonic pickup, and the regulating transformer was housed in a special case to reduce magnetic fields at harmonics of the 60-cps line frequency.

13-33. When angle modulation is analyzed on a spectrum basis, only a simple amplitude detecting rectifier circuit is required in addition to the analyzer selectivity. Since the selectivity of the analyzer can slope-detect angle modulation, it is necessary to integrate the output of the detector rectifier because the slope detection is a function of the analyzer selectivity and will not truly represent the spectrum of the equipment under test. On the Model 478R-1 spectrum analyzer, an external plug-in capacitor of several microfarads may be placed across the oscilloscope deflection input terminals to perform this integration and eliminate the slope detection.

13-34. A sinusoidally modulated FM wave has a spectrum which contains not just two side frequencies as in AM, but an infinite number of side frequencies spaced equally from the carrier by intervals equal to the modulating frequency. When the angle modulation level is very low, the amplitudes of the higher order sidebands drop very rapidly. When the modulation level is high, the amplitude of the carrier may be lower than some of the sidebands, and the sidebands will extend over a much large band of frequencies. A qualitative check of the effect of low-level incidental angle modulation on a carrier may be obtained by slope detection in a receiver which has a relatively high degree of selectivity, such as the Model 51J or Model R390. The receiver is tuned to one side of the carrier signal so that the S meter indicates one-half, or less, of the maximum deflection obtained when peaked exactly on the signal. If no hum or other tone is heard in the receiver as the AVC allows the sensitivity of the receiver to increase and the noise to rise, it may be assumed that closely spaced, discreet angle modulation spectra are below the noise level.

13-35. Compression Measurement.

13-36. An additional characteristic of power amplifiers, known as compression, is often measured as an indication of capability of the power amplifier and its power supply. Compression of the output signal may result from less than optimum DC regulation of the power supply for the power amplifier plate, screen and bias voltages and may serve as an indication of intermodulation in the power amplifier when subjected to close spaced tones. Since the power amplifiers are normally operated in class AB_1, or some other mode of class B operation with respect to plate current, the load on the power supply varies with the instantaneous amplitude of the signal envelope. Compression results when, in the presence of one signal which does not utilize full peak envelope power capability, a second signal is applied which approaches full peak envelope power. The amplitude of the first signal is compressed by an amount which is a function of the variation in the power supply output voltage as a result of the additional loading demand by the second signal. Measurement of compression must be conducted with selective equipment which is capable of observing the amplitude of one continuous signal as a second signal is varied in amplitude. Such a measurement is usually obtained with spectrum analysis equipment by observation of a continuous desired signal 10 to 20 db below peak envelope power while on and off. The effect on the fixed amplitude tone is plotted in terms of db versus the number of db by which peak envelope power is approached or exceeded. The Model 478R-1 analyzer is particularly adaptable to this type of measurement in that its db scale may be expanded 10-to-1 so that each inch of oscilloscope scale equals 1 db. The accuracy of the measurement is then ±0.1 db. Such measurements may be conducted using the same tones as are used with the standard two-tone test signal, stopping the sweep motor so that the amplitude of one tone is continuously monitored while the other tone is switched on and off. Intermodulation distortion may be produced by the RF power amplifier when operating with close spaced tones if the low-frequency AC impedance of the power supply is too high. The screen voltage supplies for pentode power amplifiers often present stringent requirements because such amplifiers are sensitive to screen voltage changes.

13-37. Intermodulation Measurements with Built-in Monitor.

13-38. Intermodulation distortion in a transmitter may be tested by means of a monitor which uses the same frequency scheme as the transmitter, but operating in reverse to translate the RF output signal back to audio or some convenient fixed intermediate frequency. In this manner, spectrum analysis can be made and compared with analysis of the original audio signals applied to the transmitter. The monitor itself must be carefully designed so that its intermodulation is lower than that of the transmitter to be tested. For example, if the level of intermodulation distortion products are 40 db below desired output signals, and the intermodulation within the test equipment is 46 db below desired output, the resulting intermodulation distortion measurement will be in error by approximately 1 db. Therefore, the result of the measurement will indicate intermodulation products 39 db below desired output rather than the actual 40 db figure. These relationships represent normal conditions but do not guarantee this result for every situation.

13-39. Measurements made by means of such a monitor, however, will not show incidental angle modulation, since use of the same translating frequency sources for both the transmitter under test and the monitor will tend to cancel the effect of such incidental angle modulation. Separate translating frequencies for the test equipment must be used to measure spurious sidebands caused by incidental angle modulation.

13-40. Linearity Measurement with Noise Loading.

13-41. Intermodulation distortion measuring equipment using two tones is very versatile for identification of linearity characteristics. However, measuring equipment using noise as the input signal has the advantage that the test signal more nearly simulates the complex signal typical of voice or multiple tone modulation. If band-limited noise is introduced into a system under test, linearity of the system may be partially described in terms of the noise outside the original band limits. If the output of a random noise generator is fed into a band-pass filter which passes equally all noise frequencies in the passband to be tested, except for a small portion at the upper and lower extremes, the noise loads all but a few cycles of the transmission band to any degree of modulation desired. At the receiving end of the system three band-pass filters with equal bandwidth and insertion loss are used for measurement purposes. One such filter is selected near the center of the transmitted noise passband and a true rms noise voltage from this filter is used as a reference signal level. The other two filters pass the distortion sidebands just outside the intended noise passband. The output of these two filters is measured separately with the true rms voltmeter and these levels, in db below the reference voltage, represent the intermodulation distortion generated at the low- and high-frequency ends of the loaded passband. Noise generators for this purpose are commercially available and the bandpass filters may be selected to satisfy the requirements of the equipment under test.

13-42. As indicated in figure 13-6, a transmitter loaded with noise signals over a discreet bandwidth, B, will have all third-order intermodulation products appearing in a band equal to three times the desired bandwidth and with the same center frequency. All fifth-order intermodulation products will fall inside a bandwidth five times the width of the desired band and having the same center frequency. All seventh-order products will fall inside a band having seven times the width of the desired band and having the same center frequency, and so on.

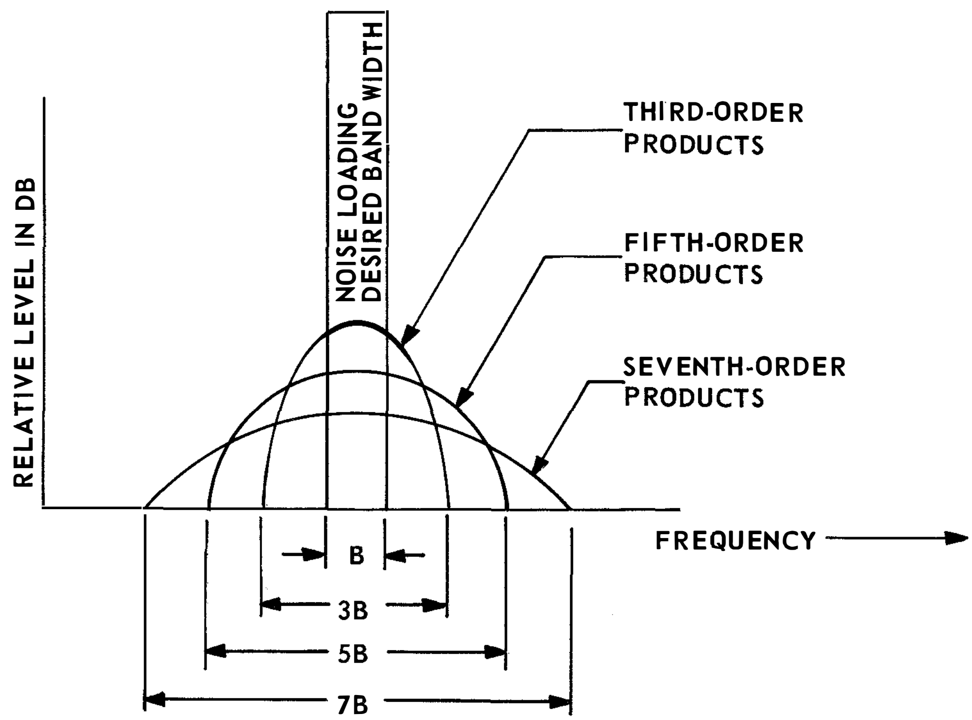


Figure 13-6. Intermodulation Products in a Noise-Loaded System.

Since the amplitudes of the intermodulation distortion products are usually in approximate inverse proportion to the order of the product, the shape of the curves describing the amplitudes of the products may be predicted. If a two-tone test signal is employed, the discreet frequency relationships between the desired signals and the intermodulation products will be known or may be computed. When these are superimposed upon the curves of the intermodulation products for a noise-loaded system (figure 13-5), an approximate plot of the entire spectrum resulting from the two-tone test signal may be predicted.

13-43. The Collins Model 478R-2 Baseband Spectrum Analyzer may be used in the same manner to test intermodulation distortion in a noise-loaded system or in one employing multiple discreet frequencies. Only one filter is necessary at the receiving end of the system. A sweeping frequency scheme is employed to allow a panoramic plot on an oscilloscope, or on a recorder, of the responses throughout and beyond the system bandwidth. This equipment permits simultaneous observation of a range of audio and video signals from 3 Kc to 2 Mc. The plot on the oscilloscope or recorder is in terms of db plotted on the Y axis against frequency in kilocycles on the X axis, as in the Model 478R-1 Spectrum Analyzer. The sweep width is variable from 0 to 70 Kc, and the main tuning dial is detented to position the center frequency at 50 Kc intervals over the 3-Kc to 2-Mc range, so that the complete frequency range may be examined accurately and rapidly in 50-Kc increments. This equipment is particularly suited for portable and field use when used with a portable two-axis recorder.

13-44. Delay Distortion.

13-45. A transmitter which has delay distortion but negligible nonlinear distortion will not cause the production of new output frequencies. The amplitude of the output components is not affected by delay distortion. The existence of delay distortion within the transmitter will not influence the results of measuring nonlinear distortion by either multi-signal loading or noise loading methods if the delay distortion does not vary with time. Phase always varies with frequency in a reactive network, but phase distortion is not necessarily produced. Instruments for the direct measurement of the phase of audio frequencies are commercially available. When a plot of the phase measurements against frequency is differentiated with respect to frequency, the derivative is the envelope delay. Phase delay is defined as the ratio of the phase with respect to frequency and approaches the value of the envelope delay expression, namely, the derivative of phase with respect to frequency, when the phase-frequency plot approaches a straight line. The perfect system is never available in practice, so the phase delay is never exactly equal to the envelope delay. Delay distortion measurements may be obtained by passing a modulated signal through a network and measuring the resultant modulation envelope phase shift caused by the network under test. Time delay may then be computed by using $T_d = \theta/360f_m$, where T_d is the delay in seconds, θ is the phase shift of the modulation envelope in degrees, and f_m the frequency of modulation in cps. Systems in which delay distortion can seriously affect or completely destroy the useful characteristics of the desired signal must be tested using equipment designed for this particular purpose, the common method being the measurement of envelope phase shift.

13-46. Field Test Set for Intermodulation Distortion Measurements.

13-47. A smaller portable spectrum analyzer would be useful for field test purposes. Such a unit could serve as a transmitter monitor as well as an intermodulation distortion analyzer. The following measurements could be provided by this equipment.

a. Linearity or intermodulation distortion in a transmitter, receiver, or audio amplifier.

b. Carrier leak or suppression.

c. Alignment checks.

d. Low-level RF voltage measurements.

e. Transmitter monitoring.

The basic technique of such a distortion analysis field test set would be to translate the RF signal through low-distortion mixers to audio and to separate the signals and distortion products with audio filters. The relative amplitudes of the intermodulation products with respect to desired signals would be obtained from attenuator readings and would be limited to about a 50-db range. When measuring intermodulation distortion in a transmitter or in a receiver, a study of relative magnitudes of the intermodulation distortion products, combined with familiarity with the frequency scheme, will usually isolate a malfunction with respect to the RF, IF or audio section of the equipment under test. Such an analyzer would not indicate RF signal levels directly in volts; however, it would indicate whether the same readings prevail that existed during a previous test. This function would be especially useful for trimming tuned circuits which operate at levels below normal VTVM sensitivities. The monitor portion of the test unit would allow aural checks with the equipment in operation to determine if speech or multiple-tone data circuits sound normal. When the equipment is used for monitoring purposes, either or both sidebands, as transmitted, would appear in the audio output without separation.

The field analyzer would consist of three basic units:

a. An audio two-tone-test signal generator.

b. An audio distortion analysis filter.

c. An RF to audio converter.

For monitoring purposes, a fourth unit consisting of an audio amplifier with provisions for headphones or loudspeaker could be used.

13-48. Audio Two-Tone Signal Generator.

13-49. The audio two-tone signal generator (figure 13-7) would consist of two oscillators whose frequencies would be controlled by audio resonators such as are provided in Kineplex* equipment. Each oscillator would be provided with a level control, its own amplifier, and a low-pass filter with the cutoff between the fundamental and second harmonic of the oscillator frequency. On-off switches would be provided for each tone to enable identification of intermodulation distortion products. Isolation pads in each signal path plus the additional isolation afforded by the resistive adding network

* Registered in U. S. Patent Office

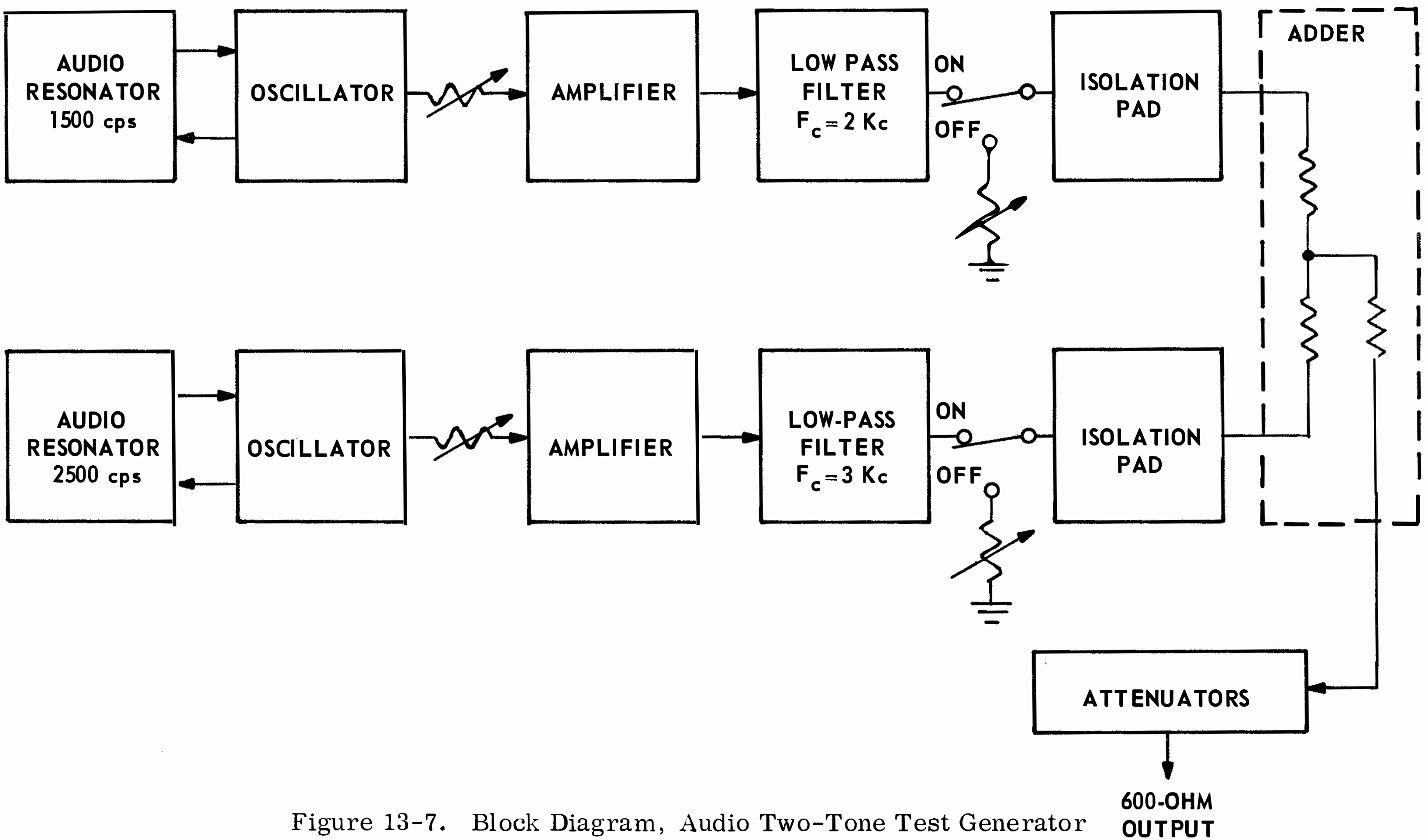


Figure 13-7. Block Diagram, Audio Two-Tone Test Generator

and isolating due to the presence of the low-pass filters would reduce intermodulation between the two audio amplifiers to a practical level. One- and ten-db-per-step attenuators in the combined output signal path would provide for intermodulation distortion measurements yielding a curve of intermodulation distortion versus audio input amplitude. The output impedance of the test generator would be 600 ohms, with the output signal level variable by means of the attenuators from approximately 3-volts-per-tone to 100 db below this level (approximately 30 microvolts).

13-50. Distortion Analysis Filter.

13-51. Although audio analysis may be performed with commercial wave analyzers, the process is slow and commercial units for this purpose are expensive and impractical for field use. The audio distortion analysis filter (figure 13-8) for this field test

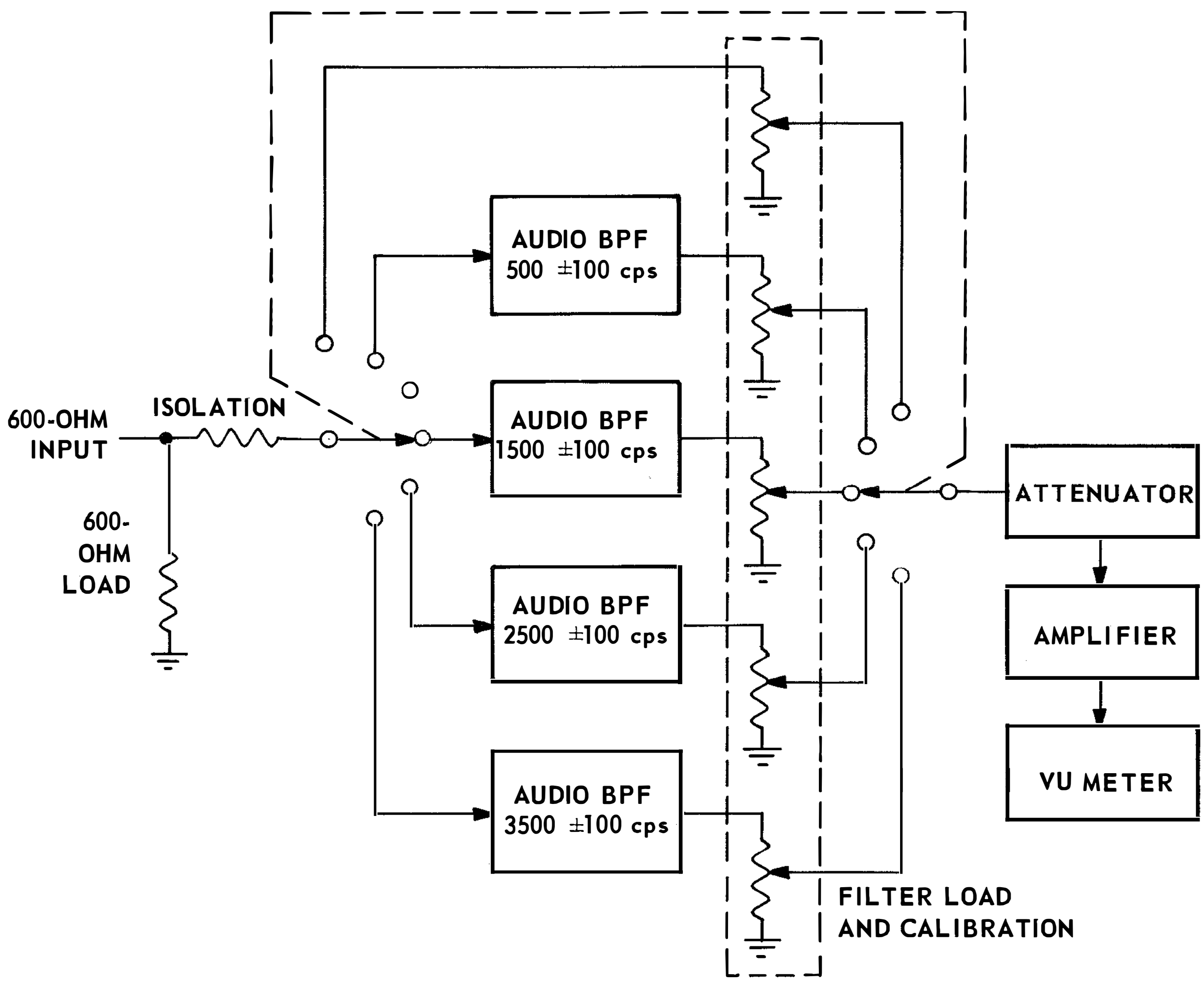


Figure 13-8. Block Diagram, Audio Distortion Analysis Filter.

set would have a 600-ohm input termination, suitable isolation of the internal audio band-pass filters from the 600-ohm input, and a selector switch for inserting any one of four audio band-pass filters or an unfiltered circuit in the path to the indicating meter. Each band-pass filter would be wide enough to pass only one of the discreet audio frequencies of interest, allowing for approximately 50-cps error in tuning of the vfo in the RF-to-audio converter. Two of the filters would pass the desired two-tone test signals, and the other two filters would pass the third-order intermodulation distortion products resulting from the two desired test tones. The unfiltered circuit would provide measurement of signal levels under normal operating conditions of the equipment monitored. Fifth-order products in a transmitter could be measured by tuning the RF-to-audio converter to one side so that the desired two tones fall on the frequencies of the first and second filters. Then the third-order intermodulation distortion product would appear in the third filter and the fifth-order intermodulation product in the fourth filter. Seventh-order products could be measured by side stepping the tuning of the RF-to-audio converter one additional slot so that the higher of the two desired test tones falls on the frequency of the first filter. Then the third-, fifth-, and seventh-order products would fall in the second, third, and fourth filters, respectively. Provision would be made for compensating for the insertion loss of the various filters individually, and a second gang on the selector switch would connect the attenuator, amplifier, and level indicating meter to the filter circuit to which the input was switched. With a 10-db-per-step attenuator and a level-indicating meter calibrated in db over a 1 to 10 db range, signals could be measured with an accuracy of ±1 db, while the dynamic range requirements on the meter amplifier would be only 10 db. Since it would pass only one frequency at a time, its distortion requirements would be negligible.

13-52. RF-to-Audio Converter.

13-53. The RF-to-audio converter (figure 13-9) would consist of three mixers. Only fixed-tuned low-pass and band-pass filters would be used in the signal path of the converter to eliminate the necessity for tracking tuned circuits. The first mixer input might be switched to a high-impedance input through a potentiometer or to a 50-ohm calibrated RF attenuator. A limited range of input level control would be obtained thereby to allow setting input signals at optimum mixer operating levels. There would be no internal gain controls. If required, additional attenuation could be obtained before the signal is applied to the converter by means of sampling impedances in the external isolating circuits. Transmitters and other equipment under test should have suitable test points provided for monitor pickup.

13-54. The first mixer of the converter would obtain translating signals from a multiple-crystal oscillator and multiplier to heterodyne the RF input to a range of 1.7 to 4.3 Mc. Since the output of the vfo need not be multiplied for this purpose, the stability and ease of tuning in the converter would be improved. The multiple-crystal oscillator and multipliers would be provided with suitable tuned circuits for rejecting the undesired multiples where necessary. A level adjustment would be required for each

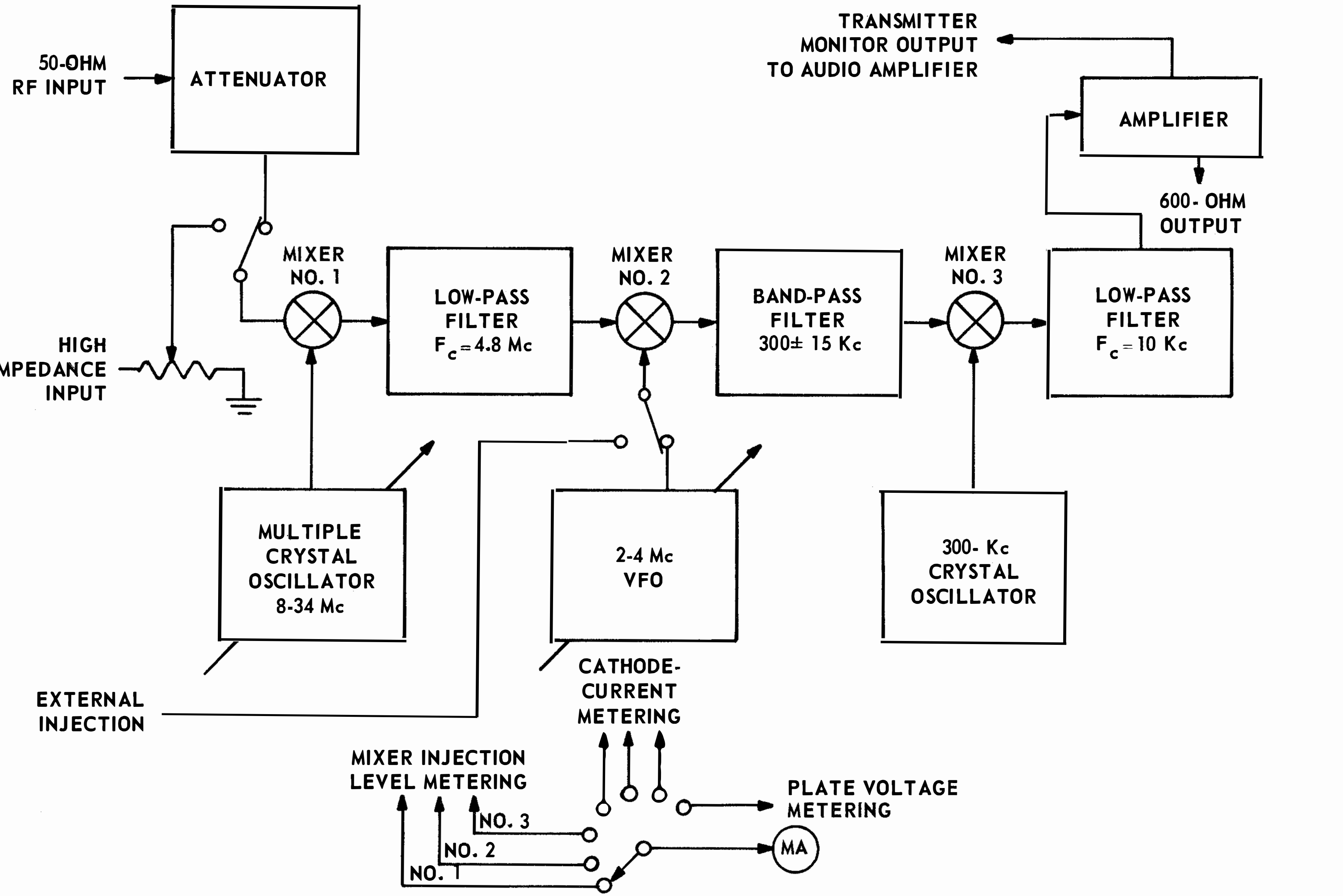


Figure 13-9. Block Diagram, RF-to-Audio Converter.

range. All crystals, coils, and level controls would be selected by a range position switch. Since the output of mixer No. 1 would be filtered by a low-pass filter, signals in the range 1.7 to 4.3 Mc could be applied through mixer No. 1, as an amplifier, directly to mixer No. 2 without heterodyning. The second mixer would obtain a heterodyning signal from a 2- to 4-Mc, ten-turn, variable-frequency oscillator, which could have a vernier control to facilitate fine tuning. If desired, the injection for mixer No. 2 could be applied externally to allow detection of signals at frequencies below 1.7 Mc. The output of mixer No. 2 would be filtered by a band-pass filter approximately 30 Kc wide and centered at 300 Kc. Mixer No. 3 would heterodyne the signal to audio with injection obtained from a 300-Kc crystal oscillator which could be provided with a trimmer for very fine tuning. The output of mixer No. 3 would be filtered to pass only the audio range, after which the signal would be amplified by a low-distortion audio amplifier and made available in the form of a 600-ohm output capable of driving the audio distortion analysis filter. This 600-ohm output could be bridged with a high-impedance input audio amplifier driving headphones or a loudspeaker. If a monitor audio amplifier and speaker were used, and if the low-frequency response were adequate, the monitor would indicate the presence of hum in the transmitter. The unfiltered circuit in the distortion-analysis filter could be used as an aid in rough-tuning the RF-to-audio converter in the absence of an audio monitor, or for measurement of the amplitudes of single signals which do not fall in the range of the filters but can be identified by audio monitoring. The latter application could facilitate frequency response measurements. Metering of the mixer injection levels would be provided to insure optimum mixer conversion efficiency with minimum intermodulation distortion. In addition, metering of mixer and audio amplifier cathode current and plate voltage would be provided. As in the Model 478R-1 Spectrum Analyzer, no preselectivity would be required and tuning of only the 2- to 4-Mc VFO and selection of the proper crystal oscillator circuit for mixer No. 1 would suffice to translate the RF signal to audio for these measurements.

13-55. The block diagrams of figure 13-10 through 13-12 demonstrate the use of the several sections of an analyzer of this type for tests of intermodulation distortion in a transmitter, a receiver or an audio amplifier. An entire communications system could be tested or the equipment could be used as a transmitter monitor as indicated in figures 13-13 and 13-14.

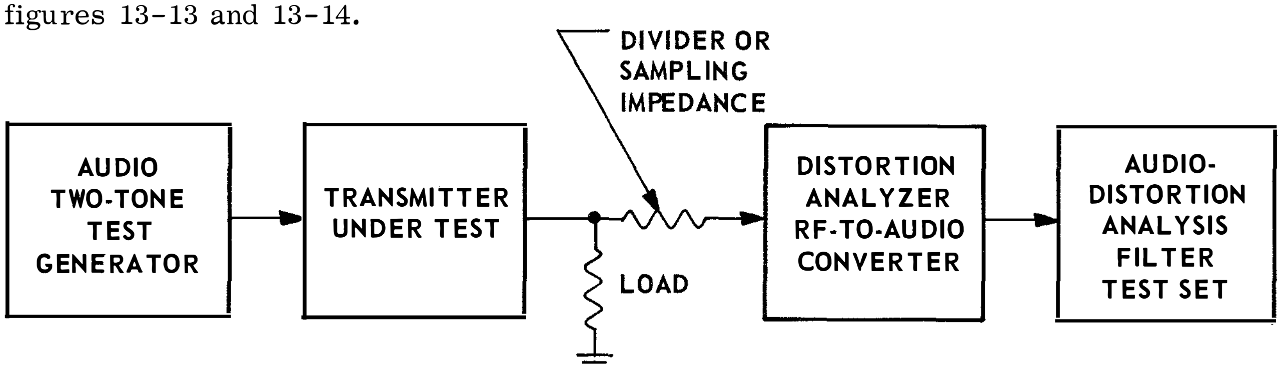


Figure 13-10. Intermodulation Test in Transmitter.

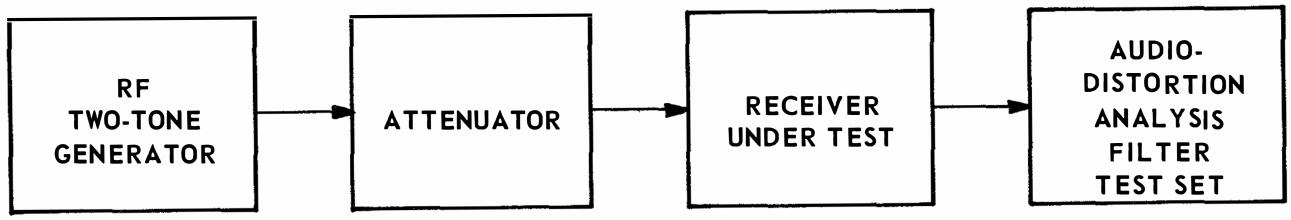


Figure 13-11. Intermodulation Test in Receiver.

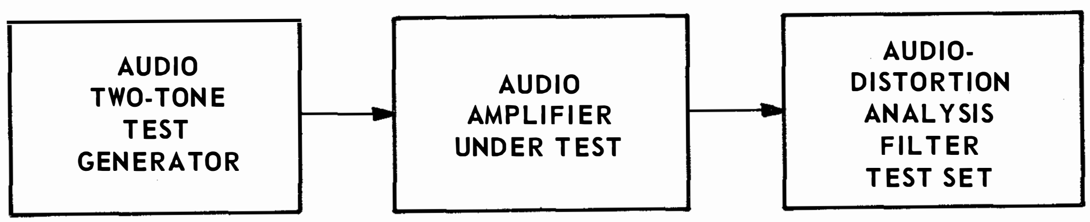


Figure 13-12. Intermodulation Test in Audio Amplifier.

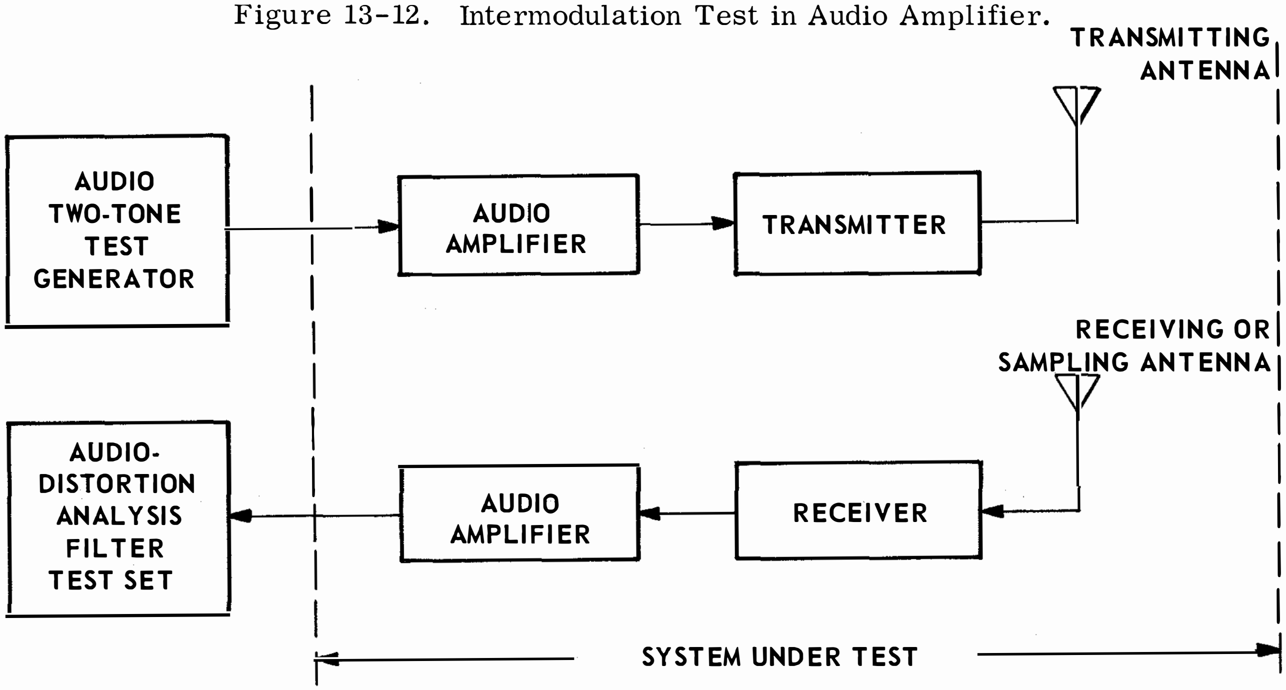


Figure 13-13. System Test for Intermodulation Distortion.

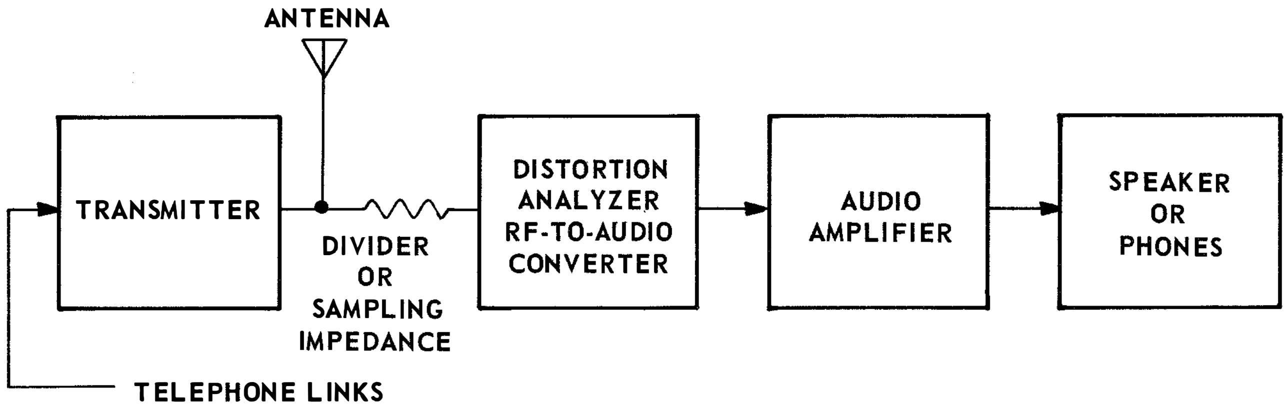


Figure 13-14. Block Diagram, Transmitter Monitor.

The analyzer would be designed so that it could be used to check itself as shown in figures 13-15 and13-16. This system would be capable of measuring intermodulation distortion products 6 db or more, higher in amplitude than the indicated measurements of these products when the analyzer is used to test itself. This condition would provide an accuracy of approximately 1 db in tests of each equipment.

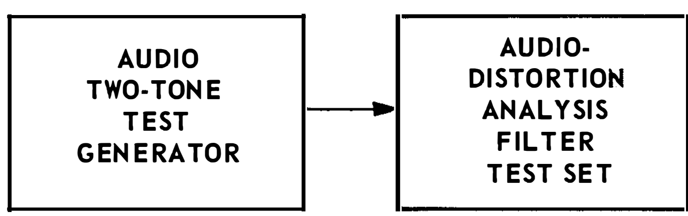


Figure 13-15. Test of Two-Tone Audio Generator.

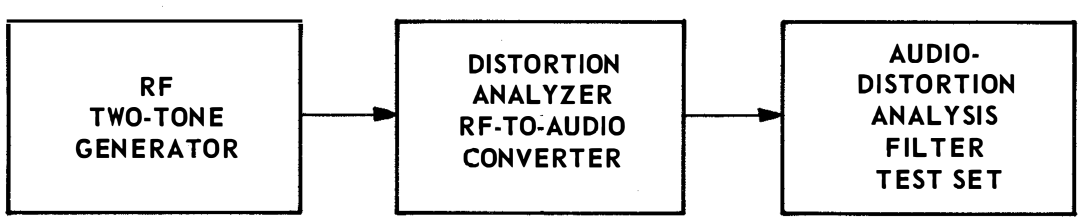


Figure 13-16. Intermodulation Test in RF-to-Audio Converter.

13-56. Types of Test Equipment.

13-57. Much of the test equipment used to service an AM system can also be used to maintain an SSB system. Most field personnel are familiar with such equipments as

AF signal generators, volt-ohm-ammeters, AC VTVM's, RF signal generators, frequency meters, oscilloscopes, dummy loads and RF sampling devices. These test equipments are sufficient to maintain most SSB receivers and transmitters. A spectrum analyzer can also be a valuable aid when a thorough study of an SSB signal is essential. The spectrum analyzer and its various functions are described in previous paragraphs.

13-58. Two-Tone Test Patterns.

13-59. To a great extent the proper operation of an SSB transmitter can be determined by studying a two-tone pattern. Two AF generators (properly isolated from each other), an oscilloscope, a dummy load, and an RF sampling device are required to generate and analyze a two-tone test pattern. The oscilloscope is used to display this pattern, obtained by injecting two AF signals of equal amplitude, but separated in frequency by approximately 1000 cps, into the AF input of the transmitter. Figure 13-17 shows the AF generator output connections that will reduce the interaction between the two generators; the 33,000-ohm resistor in series with each generator output reduces intermodulation products. The two AF signals are then terminated in a 560-ohm resistor. The two AF tones (1500 cps and 2500 cps) are adjusted to equal amplitude and to a value sufficient to develop normal transmitter output power. The SSB transmitter output is

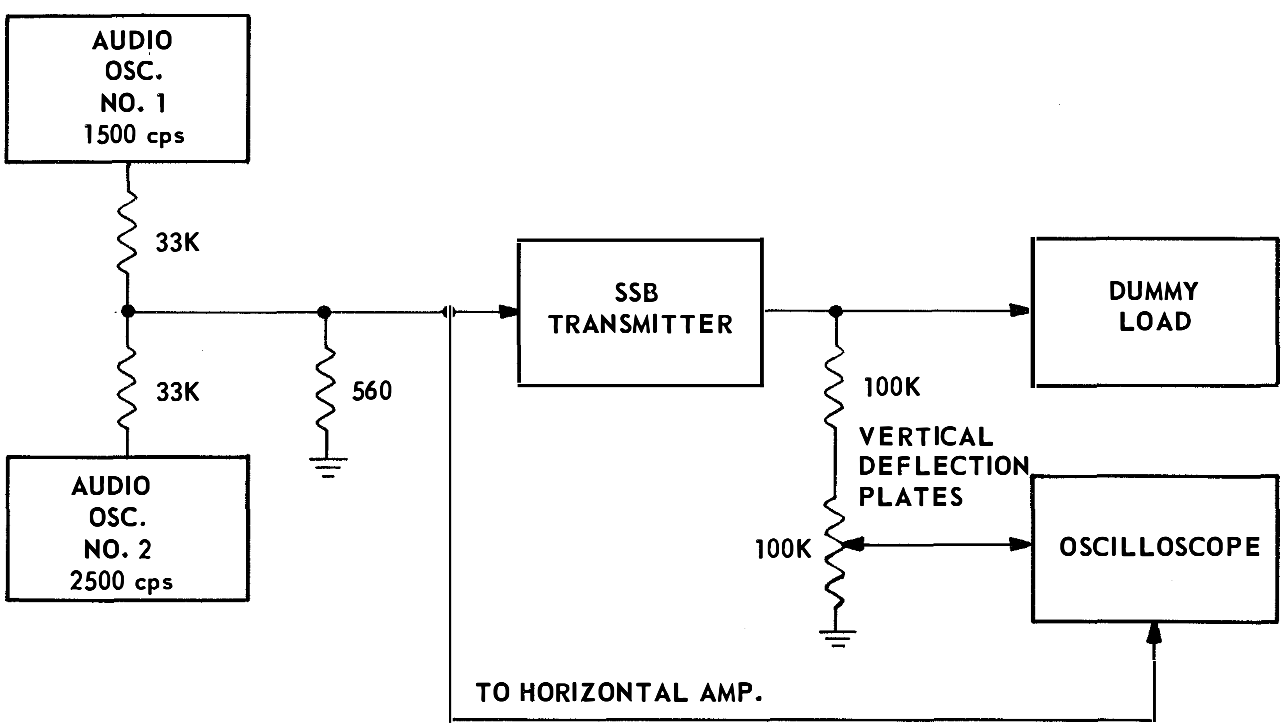


Figure 13-17. Circuit for Obtaining Two-Tone Test Pattern.

terminated in a 50-ohm resistive load capable of the necessary power dissipation, with the output signal fed to the dummy load by two sections of coaxial cable joined by a coaxial tee connector. The signal is also fed to the vertical deflection plates of the oscilloscope through the coaxial tee junction and variable 100,000-ohm carbon resistor. If a wideband oscilloscope (such as a Tektronix 545) is used, the potentiometer can be omitted; the signal can be fed through the 100,000-ohm resistor to the vertical amplifier input of the oscilloscope. All cables should be as short as possible.

13-60. By proper adjustment of both vertical gain and horizontal sweep speed, a two-tone test pattern will be observed on the oscilloscope (figure 13-18). By applying the two-tone AF signal to the horizontal amplifier input of the oscilloscope (as shown in figure 13-17), a double-trapezoid test pattern should be obtained. The test patterns obtained should be compared with those shown in figures 13-18, 13-19, and 13-20.

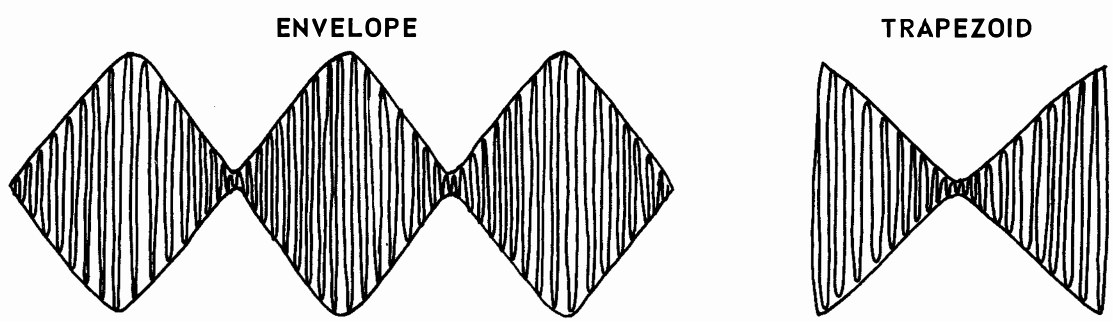


Figure 13-18. Two-Tone Test Pattern Indicating Proper Operation of SSB Transmitter.

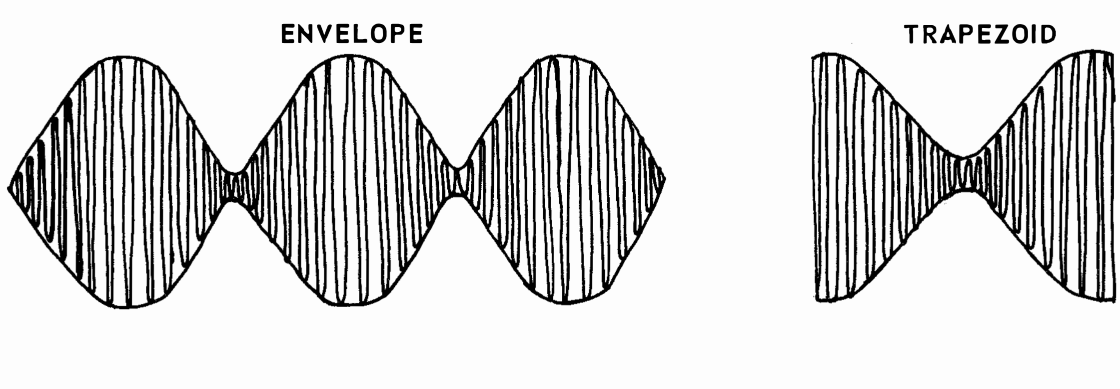


Figure 13-19. Two-Tone Test Pattern, Overdriven SSB Exciter or Power Amplifier.

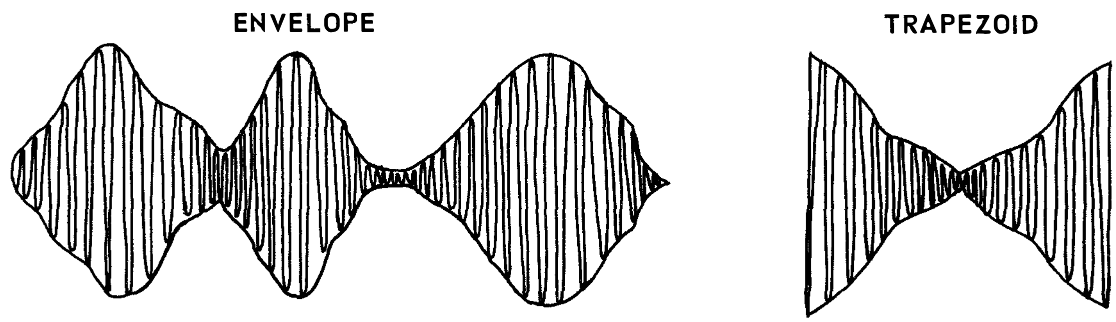


Figure 13-20. Two-Tone Test Pattern for Distorted SSB Audio Amplification

13-61. When an SSB transmitter is properly adjusted, the two-tone test pattern appears as shown in figure 13-18. Whenever possible the double-trapezoid test pattern should be analyzed in preference to the envelope. Each triangle of the trapezoid is subject to the same analysis as that obtained in an AM system. The sloping sides of the trapezoid patterns are straight when proper operation is obtained from the SSB transmitter. Since it is much easier to judge the straightness of a line than it is to judge the correctness of a sine curve, the double-trapezoid has the advantage of being more easily evaluated than the envelope. If considerable nonlinear amplification should exist in either the SSB exciter or final amplifier stages, the two-tone test pattern obtained would depart from that shown in figure 13-18. When the final amplifier is overdriven, the peaks of both the envelope and trapezoid test pattern will decrease or "flatten" as shown in figure 13-19. This condition results in increased amplitude of unwanted intermodulation products. These unwanted products consume power and cause considerable interference. Figure 13-20 shows the result of distorted audio amplification. Note that successive waves are not identical when viewing the envelope pattern. The trapezoid pattern conveys this defect by the sloping sides of the pattern. These are but a few of the variations which can be expected from the two-tone test pattern. The various indications will become more familiar with use.

13-62. Application and Specifications.

13-63. Each piece of test equipment is designed with certain capabilities, limitations, and specifications. Before using test equipment the technician should become familiar with the manufacturer's instructions. If an RF voltage is to be measured, use a voltmeter having the proper frequency response. If the voltage is developed from a high-impedance source, measure this voltage with a voltmeter of considerably higher input impedance; otherwise, the meter will load the circuit being measured. Loading of the circuit results in a lower voltage being developed across the RF voltage source and an erroneous voltage indication on the meter.

13-64. When measurements of AC signals are to be made, especially RF signals, stray shunt capacitance of the meters and their leads must be considered. For example, consider that the amplitude of a 30-Mc signal is to be measured at the grid of an RF amplifier using a Hewlett-Packard Model 410B, VTVM. This meter has an input resistance of 10 megohms, shunted by a 1.5 micromicrofarad capacitance. The capacitance reactance (Xc) $= 1/2\pi fc = 0.159/fc = 0.159/30 \times 1.5 \times 10^{-6} = 3,533$ ohms (f = frequency, c = capacitance). Since the 3,533-ohm reactance is in parallel with the 10-megohm resistance and much lower in value, the total input impedance of the voltmeter will be approximately 3,533 ohms. If the voltmeter is being fed at 30 Mc from a low-impedance source, such as 50 ohms, the meter will not load the circuit appreciably. If the 30-Mc source impedance is high, say 25,000 ohms, the meter will load the circuit and a false voltage indication will be obtained. The greater the shunt capacitance or the frequency being measured, the greater will be the error in the voltage indication.

13-65. The accuracy of the test equipment used must be better than the tolerance of the signal being measured. If a voltage to be measured must be 300 volts, ±6 volts, a meter with a tolerance of ±1 percent or better should be used. The accuracy of most meters is stated as plus or minus a certain percent of full scale. Greatest meter accuracy is obtained when the reading is at the middle of the scale or more; therefore always use a meter range that places the reading close to full scale. With the exception of vacuum-tube voltmeters, greater accuracy is possible when a meter is used in the horizontal position. It is essential that the test equipment used be properly calibrated, properly terminated, and in good condition.

13-66. Summary.

13-67. The demand for more communication channels in the high-frequency band and the large volume of information that must be carried on each channel has led to a search for a more effective and efficient method of communication. Technological advances in frequency control have made possible the use of single sideband techniques which eliminate dependence upon a carrier for automatic frequency control at the receiver. However, to assure full utilization of the advantages of these techniques, test equipment must be provided which is capable of making measurements with a much higher degree of resolution than has been customary. Precision frequency counters provide a convenient method of measuring frequency accuracy, and spectrum analyzers measure the degree of linearity and frequency stability directly in terms of bandwidth requirements. Laboratory test equipment capable of the required resolution has been built and used in the development of SSB equipment. The design and construction of precision test equipment for field use is receiving added attention as the use of SSB equipment becomes more widespread.

APPENDIX A

GLOSSARY OF TECHNICAL TERMS

1. Abscissa -- The horizontal distance from a point on a graph to the zero reference line. The units of this distance are indicated on a scale at the bottom or top of the graph.

2. Absolute Value -- The numerical value of a number, without relation to sign. Vertical lines on each side of a symbol specify that its absolute value is intended. The absolute value of z is $|z|$.

3. Active Network -- A network which receives power from both the input signal and other external sources.

4. AFC -- Automatic Frequency Control. A circuit which detects a change of frequency and feeds back an error signal to correct the change.

5. AGC -- Automatic Gain Control. A circuit which automatically maintains the gain of an amplifier so that the output is constant in amplitude (see 10., AVC).

6. ALC -- Automatic Load Control. A device which keeps the signal level adjusted so that a power amplifier works near its maximum power capability without being overdriven on signal peaks.

7. Antilogarithm -- Abbreviated antilog, the number corresponding to a given logarithm. For example, Log 563.2 is 2.7066; then 563.2 is the antilog of 2.7066.

8. AT-cut Crystal -- A piezoelectric crystal, commonly quartz, which is cut at an angle of 35° 15' from the Z axis. This crystal is normally used at high frequencies, in the range of 500 Kc to 120 Mc.

9. Attack Time -- The time necessary for a circuit to reach a steady state condition after a sudden rise in input level. The attack time will vary directly with the time constant of the circuit.

10. AVC -- Automatic Volume Control. A device which will automatically provide a substantially constant output volume with a variable input level.

11. Balanced Detector -- 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.

12. 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.

13. Balanced Modulator -- A modulator so connected that the circuit is in a state of balance with respect to one or both of 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.

14. Bandpass -- A band of frequencies passed with little or no attenuation.

15. Bandswitching -- The process of selecting any one of the bands in which a receiver or transmitter is designed to operate. Normally, two or more sections are switched simultaneously so that all circuits are tuned together. Switching may be done manually, or by means of a servomechanism.

16. Bandwidth -- The number of cycles per second expressing the difference between the limiting frequencies of a frequency band.

17. Beating Oscillator (Heterodyne Oscillator) -- An oscillator used for frequency translation or conversion in transmitters and receivers.

18. Bi-lateral Circuit -- A circuit that will properly function in either direction of signal path.

19. Bi-mode -- Having two modes of operation.

20. 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.

21. Birdie -- Sometimes called a tweet. A high pitched whistle sometimes heard while tuning a radio receiver. It is due to beating between two frequencies differing by about 10,000 cps.

22. BPF -- Bandpass Filter. A device which will allow a selected band of frequencies to pass through while rejecting all other frequencies.

23. 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.

24. 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.

25. 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 low frequency or reference oscillator.

26. Carrier Reinsertion -- The process used in SSB or DSB transmitters for reinserting 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.

27. Channel -- A limited band of frequencies, or a separate circuit, for conveying a signal isolated from other signals in the same system or frequency range.

28. Characteristic -- The integral (whole number) part of a logarithm. This integer indicates the position of the decimal point in the antilog of the logarithm.

29. Coherent Detector -- A form of synchronous or phase-locked detector used in double sideband suppressed carrier receivers.

30. Chronoscope -- An electronic instrument used for measuring extremely short intervals of time.

31. Clipper Circuit -- A circuit used to limit or clip off the peaks of a signal which are of too great an amplitude for desired signal level.

32. Cofunction -- The trigonometric function of a given angle is the cofunction of the complement of that angle. If $A + B = 90°$ then sine A = cosine B.

33. 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 such systems.

34. Combining Network -- A network for vectorially combining the outputs from two or more separate circuits.

35. Comparator -- A type of AFC 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.

36. Compatibility -- The characteristic of a transmitted signal which enables it to be received by more than one specific type of receiver, or the design capability of a transmitter to generate such a signal, or the design capability of a receiver to receive more than one type of transmission. This is a term generally used in reference to capability for operating SSB or DSB equipment in conjunction with conventional AM equipment.

37. 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.

38. Complementary Angles -- Two angles whose sum is 90°.

39. Components of a Vector -- Two directed segments of a given angle with the vector resultant. These components are determined by projecting the vector resultant on the vertical and horizontal axes.

40. Compression Circuit -- A circuit used to reduce the volume range of an audio signal. Weak signals are made stronger and strong signals are reduced in level. This action is achieved by making the effective gain vary automatically as a function of signal magnitude.

41. 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.

42. 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.

43. Corona Breakdown -- The breakdown or ionization of the air, causing a luminous discharge surrounding a conductor when the voltage gradient around the conductor exceeds a certain critical value.

44. Coupling Rods -- In a mechanical filter, the metal rods which connect the metal disks. By varying the length of these rods, the bandwidth of the filter is varied.

45. Cross Modulation -- A type of distortion produced by modulation of the desired carrier (or components of its sidebands) by the sidebands of an adjacent or other undesired carrier or signal.

46. Crossover -- A tweet or birdie. A spurious response in an IF amplifier caused by poorly chosen mixer frequencies.

47. Crystal Lattice Filter -- A wave or bandpass filter employing piezoelectric (quartz) crystals in a lattice network arrangement.

48. Damping -- Any action or influence that extracts energy from a vibrating system in order to suppress the vibration or oscillation.

49. db -- Abbreviation for decibel; refer to decibel, item 51.

50. dbm -- Abbreviation for decibels with reference to 1 milliwatt; a unit used in specifying input levels.

51. Decibel (db) -- A dimensionless unit for expressing the ratio of two values of power, the number of decibels being 10 times the logarithm to the base 10 of the power ratio. With P_1 and P_2 designating two values of power and n the number of decibels denoting their ratio:

$$n = 10 \log_{10} \frac{P_1}{P_2}$$

52. Doppler Shift -- The change in frequency of a signal due to 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.

53. 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.

54. Double Sideband (DSB) -- The transmission of both sidebands. Normally accepted as being a suppressed carrier transmission with the same information on both sidebands.

55. 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.

56. Electromechanical Chopper -- A device such as a vibrator or buzzer which interrupts the flow of direct current, breaking it up into short intervals. Such chopped DC can be used as AC for amplification through a transformer.

57. Electromechanical Filter -- A filter in which the electrical energy to be filtered is changed into mechanical energy, and the filtering process performed by a mechanical device, such as a specially designed rod, disk, or plate, all or portions of which vibrate at the specific frequency, or band of frequencies, to be coupled into another electrical, or output circuit.

58. Electromagnetic 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.

59. Envelope Detection -- Demodulation by rectification of a complete wave envelope (carrier and sidebands).

60. Electrostriction -- The mechanical deformation caused in a material by an applied electrical field.

61. Envelope Detector -- A filtered rectifier-type detector with which an RF signal is rectified, leaving only the varying voltage envelope as an output.

62. Envelope Distortion Cancelling -- A device which detects an intermodulation distortion envelope and feeds back the envelope voltage to the amplifier in order to decrease the distortion.

63. 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.

64. 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 sidebands. It is used to decrease the effects of

selective fading and detector distortion in detectors of AM, PM, and SSB receivers.

65. Exponent -- A symbol written above another symbol and on the right, denoting how many times the latter is repeated as factor. For example, $A^2 = A \times A$.

66. Fading -- Variations in amplitude of an RF signal at a receiver antenna due to in-phase 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.

67. Flywheel Effect -- The characteristic of a resonant circuit by which it sustains oscillation, producing a sine wave of essentially constant frequency, when fed with short pulses of energy at constant frequency and phase.

68. Forward Acting AVC -- An Automatic Volume Control, which acts on circuits (amplifiers) subsequent to the point of input to the AVC circuit in the receiver. Usually used in combination with conventional AVC as additional compensation.

69. 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.

70. Frequency Discrimination -- Detecting changes, or variations, in frequency.

71. Frequency 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.

72. 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.

73. Frequency Synthesis -- Producing a signal frequency by heterodyning and otherwise combining frequencies not necessarily harmonically related to each other or the frequency produced.

74. Frequency Synthesizer -- A multiple-crystal frequency generator in which the output frequencies of several crystal oscillators are mixed together to produce the desired output frequencies, all being controlled by one standard.

75. Frequency Translation (Conversion) -- Moving a signal or channel to another portion of the frequency spectrum by heterodyning. The sum or difference of the original signal and the heterodyning signal will be the desired channel at another point in the spectrum.

76. Frequency Translator -- Any device that will change a signal of one frequency to a similar signal of a different frequency. In single sideband, the device is usually an oscillator-mixer circuit.

77. Full-Carrier SSB -- See 37, Compatible Single Sideband.

78. Guard Band -- A frequency band left vacant between two neighboring channels to give a margin of safety against mutual interference.

79. Harmonic Distortion -- Impairment of fidelity caused by generation of new frequencies that are harmonics of the frequencies contained in the applied signal.

80. Heterodyning -- The process of combining, in a nonlinear device, two signals having a different frequency. Frequencies called "beats," equal to the sum and difference of the combining frequencies, are produced.

81. High Level Modulation -- Modulation produced at a point in a system where the power level approximates that at the output of the system.

82. Instantaneous Peak Plate Current -- The maximum instantaneous value of current that flows in the plate circuit of a vacuum tube.

83. Intelligence -- The message to be transmitted, such as speech, music, data, etc., or its code equivalent.

84. 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.

85. Interpolate -- To insert intermediate terms, as a series according to the law of the series. Interpolation is based on the assumption that the increase in the logarithm or trigonometric function is proportional to the increase in the number.

86. Interpolation Oscillator -- An oscillator, usually crystal controlled, used in an interpolation circuit. An interpolation circuit generates small increments of frequencies to be used in fine-tuning a receiver or transmitter automatically.

87. Kineplex -- A trade name (Collins Radio Company) of the process of kinematic multiplexing in data transmission.

88. 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.

89. Linearity Trace -- The trace on the cathode ray tube of an oscilloscope used to check the linearity of a signal.

90. Loading Comparator -- A circuit used to determine exact loading of a power amplifier. The circuit compares rectified grid and plate voltages so that when the two voltages are in an exact predetermined ratio the output of the comparator is zero.

91. Logarithm -- Abbreviated log, the power to which a number, called the base, must be raised in order to equal the number. Common logarithms use the base 10, and the natural logarithms use the base 2.71828 designated ϵ or by the Greek letter epsilon. The natural logarithms are also known as hyperbolic or Napierian logarithms.

92. Lower Sideband (LSB) -- A modulation signal spectrum which is the difference in frequency between the carrier and the modulating signal, displaced by an amount equal to the carrier frequency. The lower sideband is located below the carrier in frequency and is inverted; its bandwidth is the same as that of the original modulating signal.

93. Low Level Modulation -- Modulation produced at a point in the system where the power level is low compared with the power level at the output of the system.

94. 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. Conversely, this mechanical action produces an electrical signal at the transducer output. This is a distributed-constant system ideally suited for use as a mechanical filter in SSB applications.

95. Mantissa -- The decimal part of a logarithm.

96. Meacham Oscillator -- A form of crystal oscillator using a bridge circuit in which one arm of the bridge is a piezoelectric crystal and an opposite arm is a thermistor. This is an extremely stable oscillator used as a frequency standard.

97. Multichannel -- More than one branch or path over which signals may be transmitted.

98. Multiple Conversion -- More than one conversion step in a transmitter or superheterodyne receiver.

99. Multiplexing -- (multiplex transmission) the simultaneous transmission of two or more channels of intelligence over a single path, or the preparation of the intelligence for such transmission.

100. Negative Resistance -- The resistance of a device whose voltage-current characteristic has a negative slope; that is, when the voltage at a given point on the characteristic curve is increased, the current decreases. (This is just the opposite of Ohm's law, so the phenomenon is called negative resistance.)

101. Oblique Triangle -- Any triangle that does not contain a 90° angle.

102. 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 signal, or separate components of the same signals to an odd order, such as third order, fifth order, etc.

103. Ordinate -- The value that specifies distance in a vertical direction on a graph.

104. Passband -- In relation to a frequency response curve, the width of the frequency spectrum that occurs at a level 6 db below the peak of a curve. For example, a passband of 5 Kc.

105. Passband Nose -- That portion of a frequency response curve that occurs above a level 6 db below the peak of the curve. The nose normally extends on both sides of the resonant frequency point.

106. Passive Network -- A network which receives all of its operating power from the input signal.

107. Peak Envelope Power -- The rms power developed during the peak RF cycle occurring in the transmitter. Equal to the sum of the power amplitudes of the sideband components and the carrier.

108. Peak Envelope Voltage -- The peak voltage in a modulation envelope.

109. Peak Plate Voltage Swing -- In a vacuum tube circuit, the difference between the highest and lowest voltage values appearing on the plate of the tube during operation.

110. Peak Sideband Power -- The rms power developed during the peak RF cycle occurring in the transmitter without (minus) the carrier. Equal to the sum of the amplitudes of the sideband components only.

111. Peak-to-Valley Ratio -- In a rectifier, the difference between the peak and the minimum or zero voltages. In regard to a frequency response curve, the variation in amplitude within the nose of the curve.

112. 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.

113. Phase Discriminator -- Detects changes or variations in phase.

114. 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.

115. Phase Distortion -- Impairment of fidelity due to nonlinear phase characteristics which cause various frequencies of an applied waveform to be delayed disproportionately.

116. Phase-Locked Receiver -- A receiver in which the local oscillator is synchronized and maintained in phase with the received carrier (or its equivalent in case of suppressed-carrier reception).

117. Phase Modulation -- Variation of phase of an RF signal in accordance with intelligence to be transmitted. A form of angle modulation.

118. Phase Shift Network -- A network in which an applied signal is shifted in phase by a predetermined angle.

119. Piezoelectric Crystal -- A crystalline dielectric, commonly quartz, which exhibits a voltage generating effect when mechanically strained, and conversely, bends or vibrates when a voltage is applied to it.

120. Pi-L Network -- A filter network made up of a Pi section followed by an L section.

121. Pilot Carrier -- A reduced carrier, the amplitude of which is maintained at a level of from 10 to 20 db below that of the peak sideband. A carrier often used in SSB transmission as a reference for AVC and AFC systems in receivers.

122. Plate Current Dip -- The decrease in plate current resulting when the plate tank is tuned to resonance.

123. Post-Phasing Distortion -- A form of distortion due to harmonics generated in the audio amplifiers following the audio phase-difference networks, or to improper adjustment of the balanced modulators in phase-shift SSB transmitters.

124. Product Detector (Demodulator) -- A detector in which the amplitude of the output (audio) is proportional to the mathematical product of the amplitudes of both of its two inputs (carrier and sidebands).

125. Product Modulator -- A modulator in which the amplitude of the output 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.

126. Projection -- The projection of an oblique line on a horizontal is the length of the segment on the horizontal between the perpendiculars dropped from the ends of the oblique line.

127. Pulse Modulation -- Either the modulation of an RF 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 duration (PDM), pulse position (PPM), or pulse code (PCM).

128. Quadrature Input -- The use of two separate input signals, one of which is shifted in phase 90 degrees with respect to the other.

129. Quadrature Modulation -- The phase-shift method of modulation using two input signals in quadrature phase relationship with each other. Normally two balanced modulators are used in SSB applications.

130. Radius Vector -- The distance from any point to the origin (center) in a plane of rectangular coordinates. The radius vector is always positive when not zero.

131. 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.

132. 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.

133. Reference Angle -- The acute angle between the horizontal axis and the terminal side of a given angle. Any function of the given angle and the reference angle have the same absolute value.

134. Reference Oscillator -- An oscillator used for automatic frequency control reference in SSB transmitters and receivers. It is usually a low-frequency oscillator and ordinarily serves the dual function of frequency control and carrier insertion.

135. Release Time -- The time necessary for a circuit to reach a steady state condition after a sudden drop in input level. The release time will vary directly with the time constant of the circuit.

136. Resultant -- A force that combines the effects of two or more forces acting on an object, as the resultant of two vectors.

137. rms Calibrated VTVM -- A vacuum tube voltmeter calibrated to read rms values.

138. Roofing Filter -- Refer to item 20, bridging filter.

139. Scalar -- A quantity fully described by a number and having no direction; as opposed to a vector, which has both quantity and direction.

140. Selective Fading -- Fading that affects unequally the different frequencies of the received radio signal within a specified band.

141. Servomechanism -- A feedback control system in which one or more of the system signals causes mechanical motion to take place.

142. Shape Factor -- In reference to a frequency response curve, the ratio of the passband 60 db below peak divided by the passband 6 db below peak. For example, a shape factor of 2.2 represents a fairly sharp cutoff mechanical filter. Also see skirt selectivity.

143. 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 the difference (lower sideband) frequencies produced by modulation of the carrier with the modulating signal.

144. 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.

145. Sideband Filter -- A filter designed to pass the desired sideband frequencies with little or no attenuation and to reject all other frequencies, especially those

adjacent to the sideband frequencies.

146. 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.

147. Sidestep Oscillator -- In a stabilized master oscillator, a circuit which provides frequency variations that are used in the mixer-oscillator to cause generation of 1-kilocycle channels in the output.

148. Signal Clipping -- Preventing the peak amplitude of an electrical signal from exceeding a predetermined value.

149. Signal Compression -- A process in which the effective gain applied to a signal is varied as a function of the signal magnitude, the effective gain being greater for small than for large signals.

150. Signal-to-Distortion Ratio -- The ratio of the amplitude of the desired signal to that of the distortion products, usually expressed in db.

151. Single-Sideband Exciter -- A unit containing all the frequency-generation 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.

152. Single-Sideband Modulation -- A form of amplitude modulation in which one sideband and the carrier are suppressed.

153. Single-Sideband Suppressed Carrier (SSSC) -- Refer to single sideband modulation, item 152.

154. Skirt Selectivity -- The selectivity of a resonant circuit or filter expressed as the shape factor of the frequency response curve. The skirt is that portion of the curve which extends from the base line to the nose of the curve. The sharper the tuned circuit, the steeper the slope of the skirt portion of the curve.

155. Slaved Oscillator -- An oscillator controlled by, or synchronized with, another oscillator or oscillatory circuit.

156. S Meter -- A signal meter. A meter sometimes used in communication receivers to indicate relative signal strength of received signals.

157. Spectrum Analyzer -- A test instrument which is used to show the distribution of the energy contained in a spectrum of frequencies.

158. Splatter -- A term used to define the distortion products of a transmitter that fall outside of the frequency limits of the desired transmission.

159. Spurious Radiation, Mixer Products, Signals, Etc. -- In any system, descriptive of undesired response similar to the desired response, but resulting from malfunction or interference.

160. Square-Law Detection -- Detection which produces an output proportional to the square of the input, or inputs.

161. Square-Law Modulator -- A modulator whose output signal is proportional to the square of the input signal. With such a modulator the carrier and modulating signal are added together in the input to obtain the modulated carrier in the output.

162. Stabilized Master Oscillator -- A variable frequency oscillator, in a special circuit employing crystal frequency synthesis, providing a multiple number of stable channel frequencies.

163. Stability -- The degree to which a specified frequency may be maintained. Usually expressed as a tolerance in percent.

164. Standing Wave Ratio -- The ratio of the amplitude of a standing wave at an antinode to the amplitude at a node.

165. Swamping Resistor -- A resistor used in the grid circuit of an amplifier to maintain a constant input impedance to the stage and for stability.

166. SWR -- Abbreviation for standing wave ratio.

167. Synchronous Reception -- Reception of a signal by the use of a phase-locked oscillator in a special receiver. Ideally suited for DSB reception, or synchronized AM reception (SAM).

168. Thermal Agitation Noise -- Random frequencies generated by the heating effect at the cathode of a vacuum tube, or by heating any part of a circuit.

169. Time-Division Multiplex -- The 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.

170. Tweet -- Sometimes called a birdie. A high pitched whistle sometimes heard while tuning a radio receiver. It is due to beating between two frequencies differing by about 10,000 cps.

171. Two-Tone Test -- A test involving the application of two separate tone signals, equal in amplitude and differing in frequency by approximately 1000 cps, to the input of a system or circuit, and the observation of results on an oscilloscope, spectrum analyzer, or other indicating device.

172. Unbalanced Detector -- A detector in which no component of the input signals is cancelled, or in which the circuit is unbalanced.

173. Upper Sideband (USB) -- A modulation signal spectrum which is the sum of the frequencies of the carrier and the modulating signal, displaced by an amount equal to the carrier frequency. The upper sideband is located above the carrier in frequency and is not inverted; its bandwidth is the same as that of the original modulating signal.

174. Vector -- A quantity having both magnitude and direction, represented by a line terminated by an arrowhead. Often used to portray amplitude and phase of a sinusoidal signal; also called vector quantity.

175. Vestigial Sideband -- A form of SSB 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.

176. VU Meter -- An instrument for indicating the volume of a complex electric wave such as that corresponding to speech or music. The reading in volume units is equal to the number of decibels above a reference level.

177. Wide-Band Phase-Shift Network -- A network designed to provide a uniform phase shift over a wide band of frequencies.

178. WWV -- U. S. National Bureau of Standards radio station for broadcasting standard frequencies and time signals, located at Beltsville, Maryland.

179. Zero Level -- The reference level used for comparing sound or signal intensities. In audio frequency work, a power of 0.006 watt is generally used as zero level. In sound, the threshold of hearing is generally assumed as the zero level.