Report on Single Sideband Techniques and Design Requirements

Bureau of Ships

Navy Department

INTRODUCTION

The forces afloat are entering a new era of communication techniques. In order to improve fleet communications the use of single sideband will be used extensively in the very near future.

The "Report on Single Sideband Techniques and Design Requirements" is published as a source of information pending the publishing of a complete reference manual.

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TABLE OF CONTENTS

TABLE OF CONTENTS, Cont.

TABLE OF CONTENTS, Cont.

LIST OF ILLUSTRATIONS

LIST OF TABLES

SECTION I

SINGLE SIDEBAND THEORY

1-1. Power Advantage of SSB Over AM

Single sideband is a very efficient method of voice communication by radio. The amount of radio frequency spectrum occupied is no greater than the frequency range of the audio or speech signal transmitted whereas all other forms of radio transmission require from twice to several times as much spectrum space. The RF power in the transmitted signal is directly proportional to the power in the original audio signal and no strong carrier is transmitted. Except for a weak pilot carrier in present commercial usage, there is no RF output when there is no audio input.

The power output rating of an SSB transmitter is given in terms of peak envelope power. Peak envelope power may be defined as the RMS power at the crest of the modulation envelope. The peak envelope power of a conventional AM signal is four times the carrier power when 100% modulated.

An excellent comparative evaluation of SSB versus AM is given by John F. Honey of the Stanford Research Institute. Some of the more important conclusions follow.

SSB is well suited for long-range communications in LF, MF, and HF frequency ranges because of its spectrum and power economy and because it is less susceptible to the effects of selective fading and interference than AM. The principle advantages of SSB arise from the elimination of the high-energy AM carrier and from further reduction in sideband power permitted by the improved performance of SSB under unfavorable propagation conditions.

Under ideal propagation conditions but in the presence of noise, SSB and AM perform equally well (same S/N ratio) if the total sideband power output of the two transmitters is the same. The AM signal is quite susceptible to degradation by multipath transmission and selective fading principally as a result of these effects on the carrier. This difficulty is essentially eliminated in SSB. The difference in performance of AM and SSB is most outstanding under very poor propagation conditions.

In the presence of narrow band man-made interference, the narrower bandwidth reduces the probability of destructive interference. A statistical study was made of the distribution of signals on the air versus signal strength and the analysis indicated that the probability of successful communication will be the same if the SSB power is equal to one-half the power of one of the two AM sidebands.

Thus SSB can give from 0 to 9 db improvement under various conditions when the total sideband power is equal in SSB and AM. In general, it may be assumed that 3 of the possible 9 db advantage will be realized on the average contact. In this case, the SSB power required for equal performance is only equal to the power in one of the AM sidebands. For example, this would rate a 100 watt SSB and a 400 watt (carrier) AM transmitter as having equal performance. It should be noted that in the above comparison it is assumed that the receiver bandwidth is just enough to accept the transmitted intelligence in each case.

It is believed that the above comparative evaluation by Honey is more realistic than comparisons based on other assumed conditions. To help evaluate other methods of comparison the following points should be considered. In conventional AM two sidebands are transmitted each having a peak envelope power equal to 1/4 carrier power. For example a 100 watt AM signal will have 25 watts peak envelope power in each sideband or a total of 50 watts. When the receiver detects this signal, the voltages of the two sidebands are added in the detector. This is called coherent detection of the two sidebands. Thus the detector output voltage is equivalent to that of a 100 watt single sideband signal. Grammar[2] uses this comparison to say that a 100 watt SSB transmitter is just equivalent to a 100 watt AM transmitter. This is valid only when the receiver bandwidth used for SSB is the same as that required for AM (e.g., 6 kc), there is no noise or interference other than broadband noise, and when the AM signal is not degraded by propagation. By using half the bandwidth for SSB reception (e.g., 3 kc) the noise is reduced 3 db so the 100 watt SSB signal becomes equivalent to a 200 watt carrier AM signal. Honey estimates that the AM signal will be degraded another 3 db on the average due to narrow band interference and poor propagation conditions which gives his 4 to 1 power ratio.

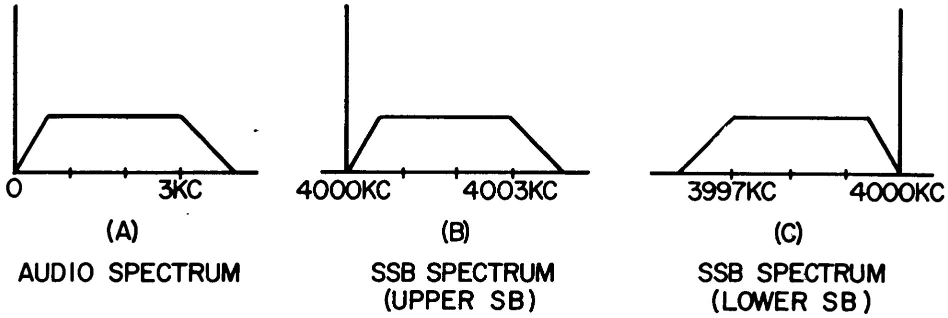
In this connection, it is pointed out that 3 db signal to noise ratio is lost when receiving only one sideband of an AM signal. The narrower receiving bandwidth reduces the noise by 3 db but the 6 db advantage of coherent detection is lost, leaving a net loss of 3 db. Poor propagation will degrade this one sideband reception less than double sideband reception, however. Also under severe narrow band interference conditions (e.g., an adjacent strong signal) the ability to reject all interference on one side of the carrier is a great advantage.

1-2. Nature of SSB Signal

The nature of a single sideband signal is most easily visualized by noting that the SSB signal components are exactly the same as the original audio signal components except that the frequency of each is raised by the frequency of the carrier. The relative amplitude of the various components remains

[2]Footnote references are made to the Bibliography which is found on page 77.

the same, however, (Actually the first statement is only true for the upper sideband since the lower sideband frequency components are the difference between the carrier and the original audio signal). Figure



(a) Audio Spectrum

(b) SSB Spectrum (Upper SB)

(c) SSB Spectrum (Lower SB)

Figure 1-1. Relationship of Audio and SSB Spectrums

1-1(a) and (b) shows how the audio spectrum is simply moved up into the radio spectrum to give the upper sideband. The lower sideband is the same except inverted as shown in Figure 1-1(c). Either sideband may be used. It is readily apparent that linear amplifiers must be used to amplify an SSB signal. It is also apparent that the carrier frequency of an SSB signal can only be changed by adding or subtracting to the original carrier frequency. This is done by heterodyning, using converter or mixer circuits similar to those in a superheterodyne receiver.

It is noted that a single sine wave tone input results in a single steady sine wave RF frequency output. Since it is difficult to measure the performance of a linear amplifier with a single tone, it has become standard practice to use two tones of equal amplitude for test purposes. The two RF frequencies resulting beat together to give the SSB envelope shown in Figure 1-2.

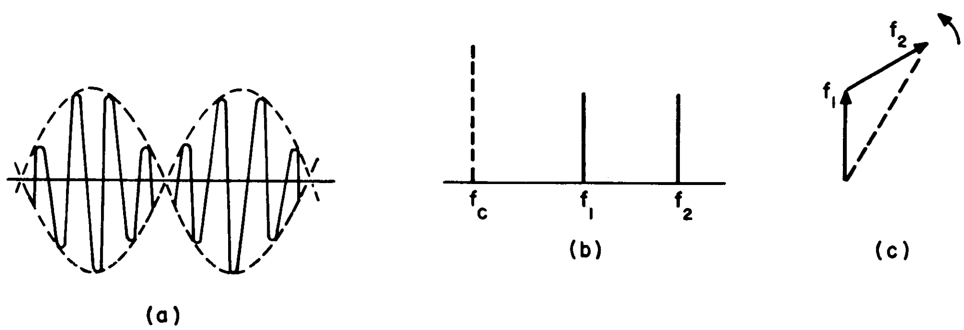
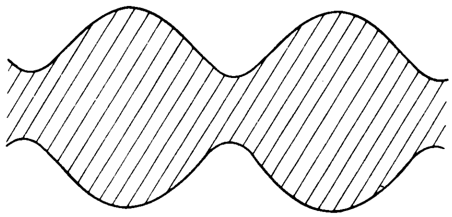


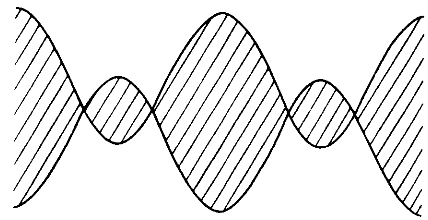
Figure 1-2. Two-Tone SSB & Frequency Components

It is noted that it has exactly the shape of half sinewaves and from one null to the next represents one full cycle of the difference frequency. How this envelope is generated is shown more clearly in Figure 1-2(b) and (c). f_1 and f_2 represent the two tone signals. When a vector representing the lower frequency tone is used as a reference, the other vector rotates around it as shown and this generates the SSB envelope. When the two vectors are exactly opposite in phase, the output is zero and this causes the null in the envelope. If one tone has twice the amplitude of the other, the envelope shape is shown in Figure 1-3(a). Figure 1-3(b) shows the SSB envelope of three equal tones of equal frequency spacings and at one particular phase relationship.



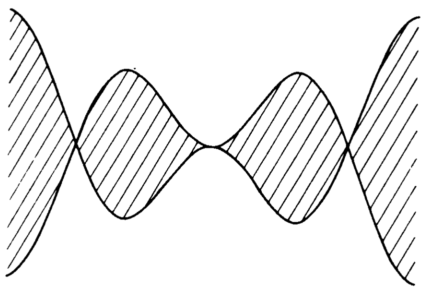
(a)

One tone twice the amplitude of other tone



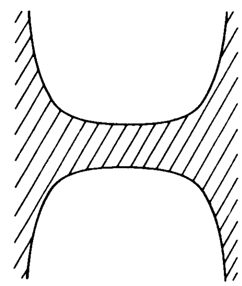
(b)

Three equal tones



(c)

Four equal tones



(d)

Square wave

Figure 1-3. Some SSB Envelopes

Fig. 1-3(c) shows the SSB envelope of four equal tones with equal frequency spacings and at one particular phase relationship. The particular phase relationships chosen are such that at some instant the vectors representing the several tones are all in phase. Figure 1-3(d) shows SSB envelope of a square wave. A pure square wave requires infinite bandwidth but its SSB envelope requires infinite amplitude also. It should be noted in particular that the SSB envelope shape is not the same as the original audio wave shape and usually bears no similarity to it. This is because the percentage difference between the RF frequencies is very small even though one audio tone may be several times the other in frequency.

1-3. Distortion Products Due to Nonlinearity of RF Amplifiers

When the SSB envelope of a voice signal is distorted, a great many new frequencies are generated. These represent all of the possible combinations of the sum and difference frequencies of all harmonics of the original frequencies. For purposes of analysis and measurement, two equal amplitude tones are used. Since the SSB RF amplifiers use tank circuits, all distortion products are filtered out except those which lie close to the desired frequencies. These are all odd order products; third order, fifth order, etc. The third order product frequencies are 2p - q and 2q - p where p and q represent the two SSB RF tone frequencies. The fifth order product frequencies are 3p - 2q and 3q - 2p. These and some higher order products are shown in Figure 1-4.

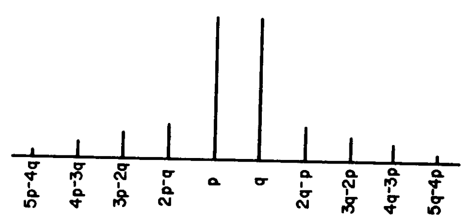


Figure 1-4. SSB Distortion Products

It is noted that the frequency spacings are always equal to the difference frequency of the two original tones. Thus when an SSB amplifier is badly overloaded, these spurious frequencies can extend far outside the original channel width and cause an unintelligible splatter type of interference in adjacent channels. This is usually of far more importance than the distortion of the original tones with regard to intelligibility or fidelity. To avoid interference in another channel, these distortion products should be down about 40 db below adjacent channel signal. Using a two-tone test, the distortion is given as the ratio of the amplitude of one test tone to the amplitude of a third order product. This is called the signal-to-distortion ratio and is usually given in db. The state of the art in building linear amplifiers has limited S/D ratios to the order of 25 to 30 db until recently. Within the last few years commercial performance on the order of 30 to 35 db has been achieved. The Collins development using a two-stage power amplifier with feedback makes 40 db possible and practical.

Two means are employed to keep the amplitude of these distortion products down to acceptable levels. One is to design the amplifier for excellent linearity over its amplitude or power range. The other is to use a means of limiting the amplitude of the SSB envelope to the capabilities of the amplifier. An automatic gain control, discussed in Section 3 and which is similar to AVC in a receiver, is used. It should be noted that the RF wave shapes are always sine waves because the tank circuits make them so. It is the change in gain with signal level in an amplifier that distorts the SSB envelope and generates these distortion products.

Speech clipping as used in AM is of no practical value in an SSB transmitter because the SSB RF envelopes are so different from the audio envelopes. A heavily clipped wave approaches a square wave and a square wave theoretically gives an SSB envelope with peaks of infinite amplitude as previously shown in Figure 1-3(d). Figure 1-5 shows the SSB envelope of a square wave with all harmonics above the 9th removed and assuming no nonlinear audio phase shift. The level at which clipping starts is indicated by the dashed line. The ineffectiveness of this type of peak amplitude control is obvious.

1-4. Addition of Distortion Products

The SSB signal is usually generated at an IF frequency and at a low signal level. Several stages are required to heterodyne this SSB signal to the desired carrier frequency and amplify it to the desired power level. Each amplifier and mixer stage generates distortion products. The distortion is kept very low in the low level stages by optimum operation at a low signal level. By this means the total distortion is kept down to 50 db or better up to the point where it feeds the last 2 or 3 power amplifier stages.

It should be noted that the distortion products generated in every stage have exactly the same frequencies. Also, they are all either in phase or 180° out of phase with each other unless the RF coupling circuits shift the phase of some frequencies more than others. In any case, the distortion products tend to add vectorially. Since it is unusual to have enough selectivity in the power amplifier stages where most of the distortion is generated to cause nonlinear phase shift across the pass band, the vector addition of the two third order components is practically identical and they are usually found to be of equal total amplitude.

1-5. Distortion of Audio Tones

When the signal-to-distortion ratio is good enough to avoid adjacent channel interference, the distortion of the original audio tones is so low as to be negligible normally. The nature of this distortion may be of academic interest, however. There are two basic sources of distortion. The common source is distortion of the SSB envelope wave shape. The other is due to phase distortion. The RF phase at the peak of the SSB envelope may be shifted by a change in resistance loading across a circuit (such as grid current) feeding back through a 1/8 wave line to cause a change in reactance which shifts the phase of the RF signals. When detected by beating against an independent reinserted carrier, this causes a distorted audio envelope because the audio envelope is advanced or retarded in phase during the interval of this RF phase shift.

Figure 1-6 shows two audio tones with a frequency ratio of 2 to 5, the resulting audio wave shape of their sum, and the SSB envelope wave shape. The shaded-in areas represent the distortion caused by flattening the SSB envelope peak. It is noted that the distortion causes depressions in the wave that may appear at any part of the audio cycle. In order to see these depressions on the detected tones on an oscilloscope, the frequency and phase relationship between the two tones must be locked in some manner. When peak flattening of an SSB envelope occurs all tones present are each reduced in amplitude proportional to the amount of peak flattening. Phase distortion would show up in a similar manner except that humps as well as depressions would occur.

1-6. Multi-Channel SSB

In commercial use, advantage is taken of the practicability of transmitting several voice or teletype channels over one transmitter channel. Since the transmitter is linear, this can be done. Two separate voice channels are easily obtained by using a twin SSB generator. Two identical SSB generators are used but one selects the upper sideband and the other selects the lower sideband. These two independent sidebands are combined and fed through the heterodyne and amplifier stages of the transmitter. Thus we have two independent voice channels as shown in Figure 1-7, and they occupy the same space as a single channel conventional AM signal. Present commercial practice goes one step further and places two voice channels in

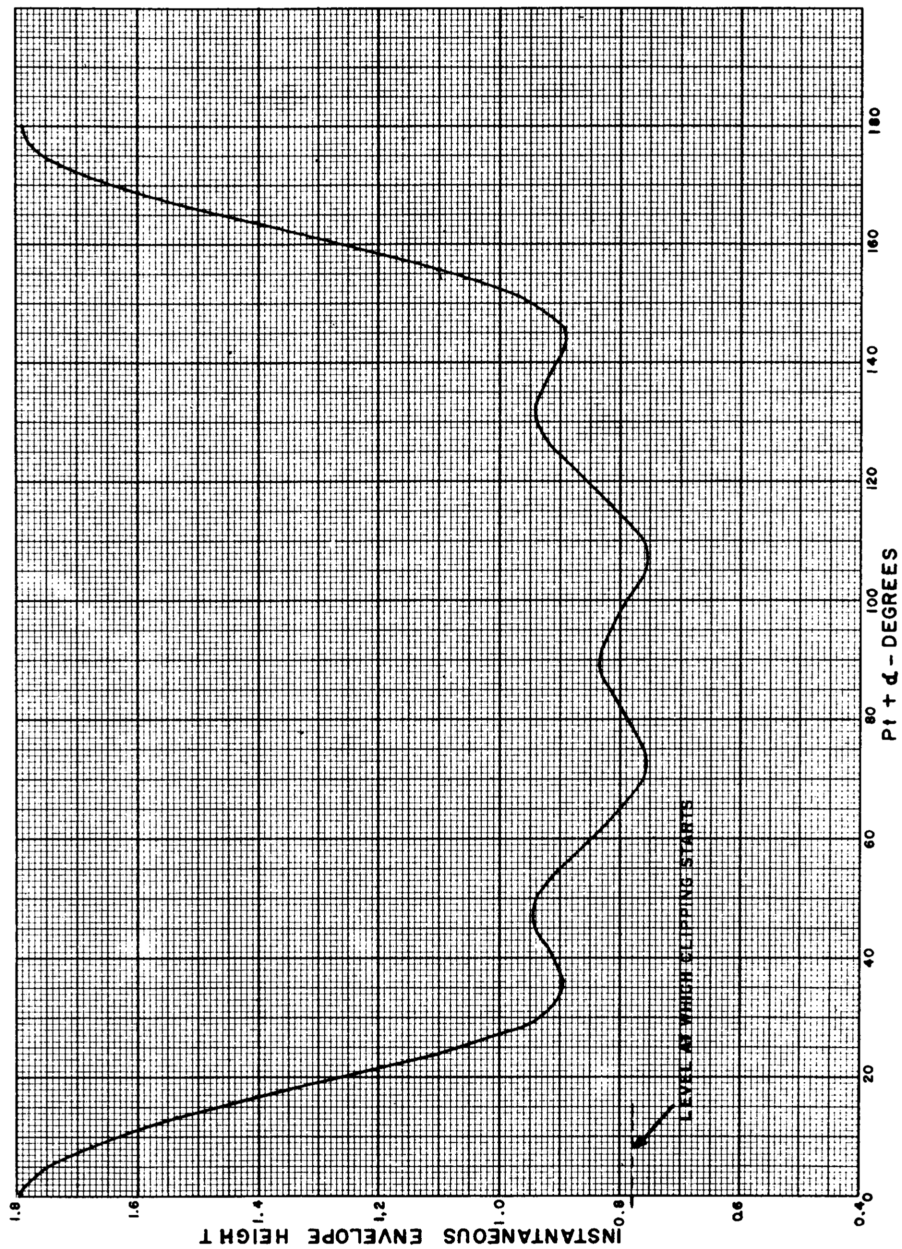


Figure 1-5. SSB Envelope of an Audio Square Wave with All Harmonic Components Above the 9th Removed

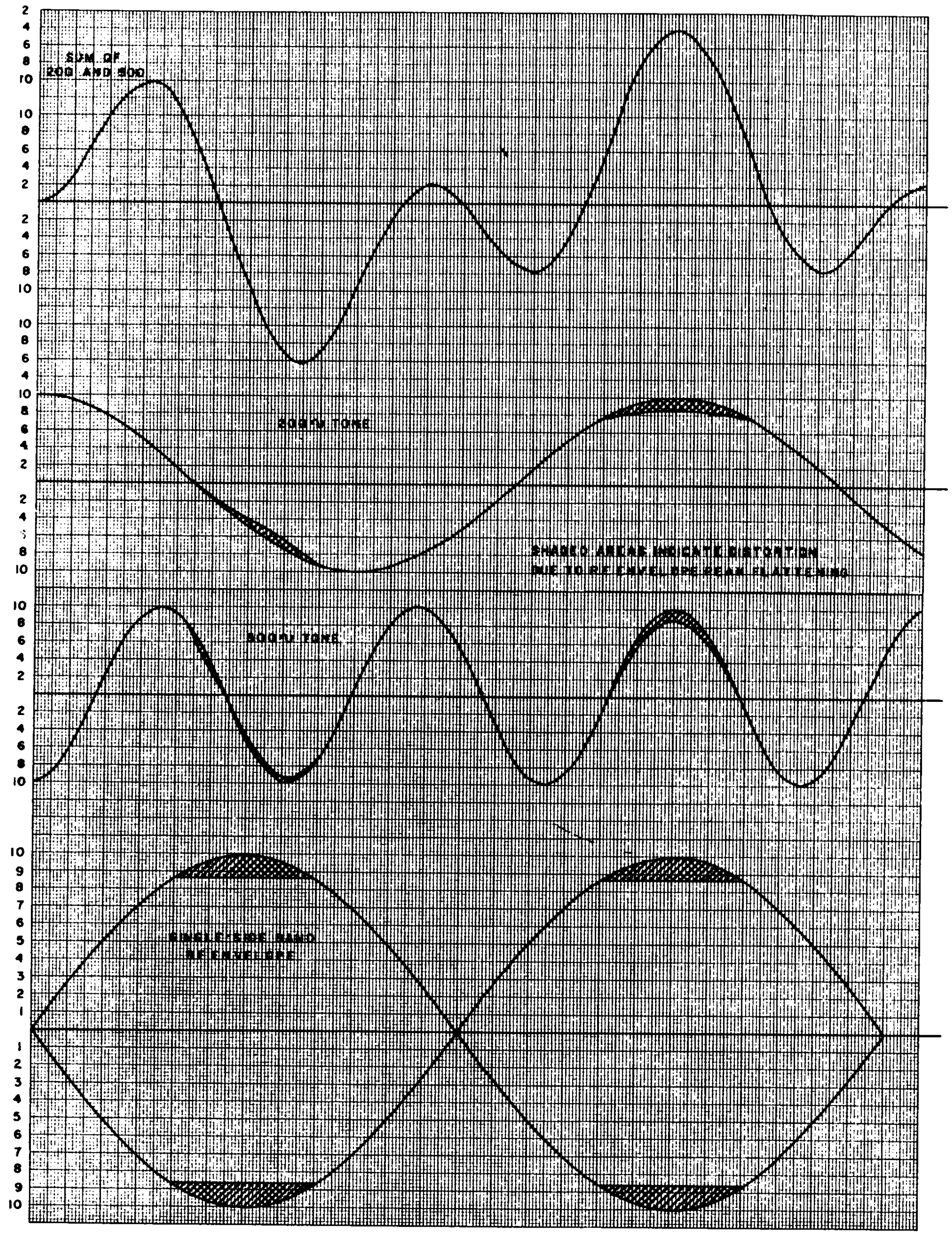


Figure 1-6. wave Shapes of Twin Channel SSB

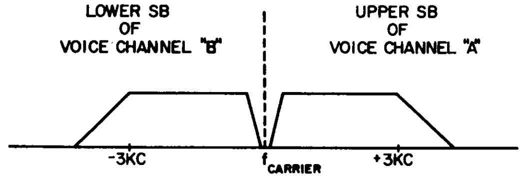


Figure 1-7. Spectrum of Twin Channel SSB

each sideband as shown in Figure 1-7, giving a total of four separate voice channels. The two outer channels are heterodyned or "sidestepped" out 3 kc in the terminal equipment.

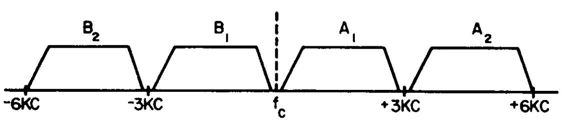


Figure 1-8. Spectrum of 4-Voice Channel SSB

As many as 16 channels of teletype are carried by one SSB transmitter. Frequency shift keying with frequency shifts as low as 17 cps are used. In addition, frequency diversity is obtained by transmitting the same signal in both sidebands separated by perhaps 6 kc. The receiving equipment uses the stronger of the two signals so if selective fading takes out one signal, the other usually remains and gives more reliable communication. It is noted again that many separate communication channels can be transmitted over one SSB transmitter. The channels can be very close together since the spacing between channels is determined by the terminal equipment. This economy of RF spectrum space is a very important advantage of SSB. Another is that one transmitter can handle several channels simultaneously so that several separate transmitters and frequency assignments are not required.

The effective transmitter power for each channel goes down as the square of the number of channels used. This is due to the voltage-power relationship. If each of two signals is allotted 1/2 of maximum voltage capability, each can have only 1/4 of maximum transmitting power. For 3 channels, the power for each is 1/9 and for 4 channels, it is 1/16 of peak transmitter power. Actually in voice communication, more effective power can be used for each channel because of the statistical probability that peaks will not occur in all channels at the same instant. The automatic load control discussed in Section 3 automatically keeps the utilization of available transmitter power capability near maximum. When many channels are used (e.g., 16) and statistical probability is considered, the power available for each channel becomes nearly a direct division between channels instead of square law. Thus using 32 channels instead of 16, would only reduce the effective power per channel one-half.

1-7. Carrier Frequency Stability Requirements.

Reception of an SSB signal is accomplished by simply heterodyning the carrier down to zero frequency. (The conversion frequency used in the last heterodyne step in the receiver is often called the reinserted carrier). If the SSB signal is not heterodyned down to exactly zero frequency, each frequency component of the detected audio signal will be high or low by the amount of this error. An error of 10 to 20 cps for speech signals is acceptable from an intelligibility standpoint but an error of 50 cps seriously degrades the intelligibility. An error of 20 cps is not acceptable for the transmission of music, however, because the harmonic relationship of the notes would be destroyed. For example, the harmonics of 220 cps are 440, 660, 880, etc. but a 10 cps error gives 230, 450, 670, 890, etc. or 210, 430, 650, 870, etc. if the error is on the other side. This would simply destroy the original sound of the tones and the harmony between tones. For voice and teletype communications, frequency errors of ±5 cps are acceptable in most applications.

Amateurs using SSB completely suppress the carrier so the combined frequency stabilities of all oscillators in both the transmitting and receiving equipment add together to give the frequency error in detection. They are willing to put up with a poorer quality signal than would a person using a commercial telephone service, however. Even so, frequency stability is a major problem with them.

Further development in the art of building very stable oscillators to the order of 1 cps in 30 mc will avoid the requirement of transmitting a pilot carrier in most applications. In order to overcome much of the frequency stability problem, it is common commercial practice to transmit a pilot carrier at a reduced amplitude. This is usually 20 db down from one tone of a two-tone test signal or 26 db below the peak envelope power rating of the transmitter. This pilot carrier is filtered out from the other signals at the receiver and either amplified and used for the reinserted carrier or used to control the frequency of a local oscillator. By this means, the frequency drift of the carrier is eliminated as an error in detection.

In SSB communication with high speed aircraft, the doppler effect must be considered. The frequency error is 1 cycle per mc of carrier frequency at an aircraft speed of 672 MPH toward or away from the station. The use of a pilot carrier can eliminate this difficulty.

1-8. Advantage of SSB with Selective Fading

On long distance communication transmission circuits using conventional AM, selective fading often causes severe distortion and at times make the signal unintelligible. When one sideband is weaker than the other, distortion results; but when the carrier becomes weak and the sidebands are strong, the distortion is extremely severe and the signal may sound like monkey chatter. This is because a carrier of at least twice the amplitude of either sideband is necessary to properly demodulate the signal. This can be overcome by using exalted carrier reception in which the carrier is amplified separately and then reinserted before the signal is demodulated or detected. This is a great help but the reinserted carrier must be very close to the same phase as the original carrier. For example, if the reinserted carrier were 90 degrees from the original source, the AM signal would be converted to phase modulation and the usual AM detector would deliver no output.

The phase of the reinserted carrier is of no importance in SSB reception and by using a strong reinserted carrier, exalted carrier reception is in effect realized. Selective fading within one sideband simply changes the amplitude and the frequency response of the system and very seldom causes the signal to become unintelligible. Thus the receiving techniques used with SSB are those which inherently greatly minimize distortion due to selective fading.

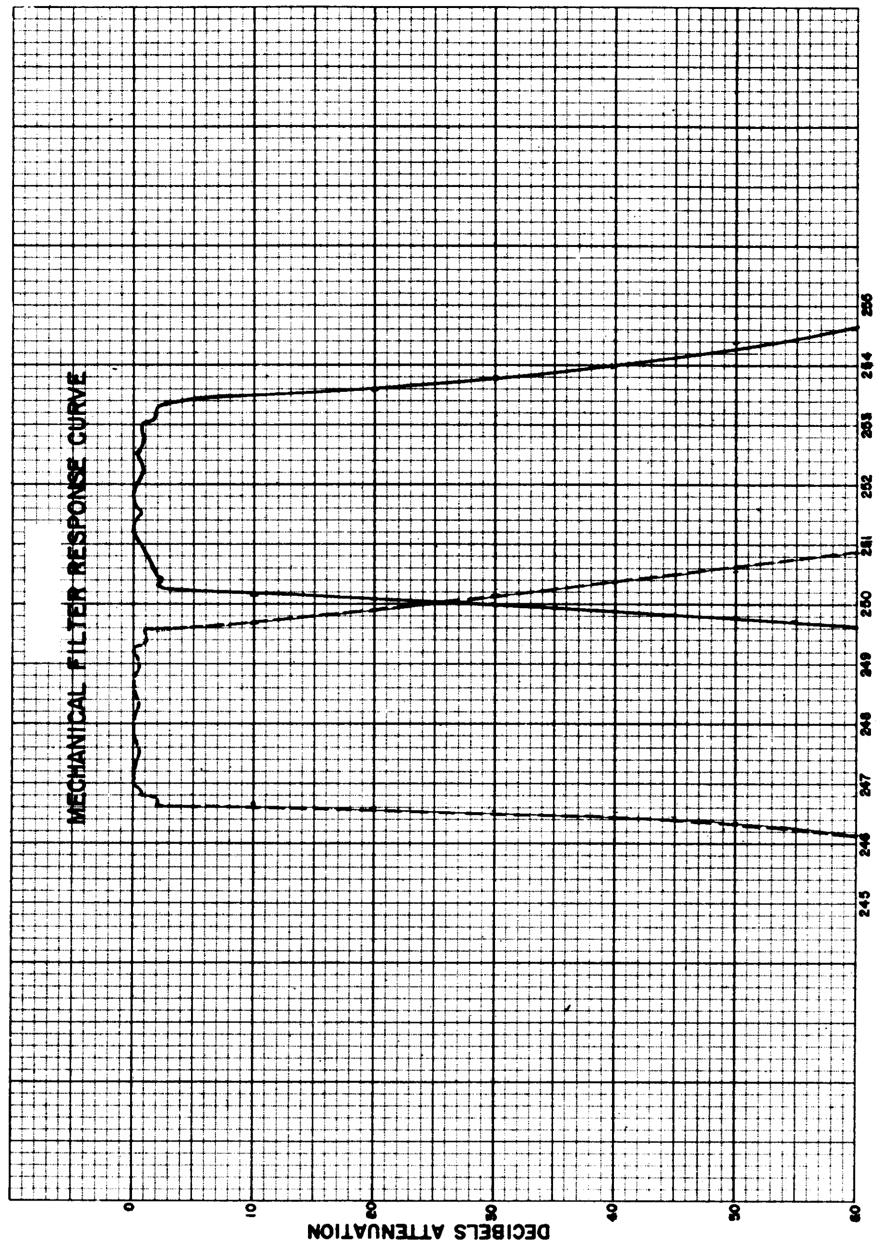


Figure 2-1. Mechanical Filter Response Curve With Collins Types F250Z-1 and F250Z-2 Mechanical Filters

Figure 2-2. Deleted

SECTION II

SINGLE SIDEBAND GENERATORS

2-1. General

Nature did not provide us with a means of generating one sideband without generating the other one at the same time. This is inherent in any modulation or mixing process. Thus SSB generators generate both sidebands by a modulation process and then use some means of selecting the desired sideband and rejecting the other. There are two principal methods of accomplishing this. One is the "filter" system and the other is the "phasing" system.

2-2. Phasing SSB Type Generator

The "phasing" system in effect uses two modulators with their outputs combined so that the desired sideband outputs add and the undesired sideband outputs balance out. To accomplish this, the carrier and also the entire audio frequency range of signal input must be shifted exactly 90° and without amplitude change over any part of the audio frequency range before going into one of the balanced modulators. For high grade performance where 40 db or more of undesired sideband rejection is required, the phase shifting networks and amplitude balancing adjustments become very critical. A variation of 1% in amplitude or 1 degree in phase in any of several places will reduce the rejection to 40 db. This inherent difficulty greatly overshadows the advantage of generating SSB directly at carrier frequency and at a relatively high power level.

2-3. Filter Type SSB Generators

The "filter" system simply passes the two sidebands from a balanced modulator through a narrow bandpass filter which passes the desired sideband and rejects the other. The degree of undesired sideband rejection is essentially limited to that of the filter characteristics. The early SSB transmitters generated SSB at a carrier frequency of 20 or 25 kc because this was the range where filters, using the techniques of that time, had the best characteristics. Later, crystal lattice filters were developed for commerical use using a carrier frequency of 100 kc. Filters of this type still set the standards for today. These high performance crystal filters are very difficult to manufacture and their cost approaches $1000 each. Presumably, crystal filters can be built for higher frequencies but they are not generally available. For reasons discussed in Section 5. 0, it is desirable to use SSB generator carrier frequencies in the 250 kc to 500 kc range for high frequency communications transmitters.

2-4. The Collins' Mechanical Filter

This recent development produced an excellent filter for SSB generators. A carrier frequency of 250 kc is about optimum for HF transmitters covering the 2. 0 to 30 mc frequency range. Filters have been developed for selecting either sideband with a 250 kc carrier frequency and they may be used in SSB receivers also. For amateur transmitters and receivers, filters with a 3 kc bandwidth are available at 455 kc. These filters are only about 1 x 1 x 3 inches in size and need no tuning adjustment. Their small size, light weight, ruggedness, stability and availability make them excellently suited for SSB generator filters.

The frequency response of Collins' Type No. F250Z-1 and F250Z-2 mechanical filters is shown in Figure 2-1. They have 3.2 bandwidth. Type No. F250Z-1 selects the upper sideband and Type F250Z-2 selects the lower sideband. These filters have an insertion loss of about 10 db in power and input and output impedances of 45,000 ohms. The voltage gain of 6 db is possible, however, when the input impedance and output impedance are about 850 ohms and 45,000 ohms respectively.

The carrier frequency is located about 20 db down on one skirt or the other to give the desired sideband.

2-5. Carrier Suppression

Normally SSB is generated with the carrier suppressed. If a pilot carrier is desired, it may be added later. Balanced modulators are used to suppress the carrier and are trimmed into accurate balance. The filter which selects the sideband also attenuates the carrier another 20 to 40 db. Since the carrier fed into the modulator is 20 to 30 db stronger than the audio signal, a great deal of carrier suppression is necessary.

2-6. Twin SSB Generators

A twin SSB generator generates two independent SSB signals from two separate signal sources. One SSB generator uses the upper sideband and the other uses the lower sideband. These two SSB outputs are combined and passed through the transmitter together. A block diagram is shown in Figure 2-3.

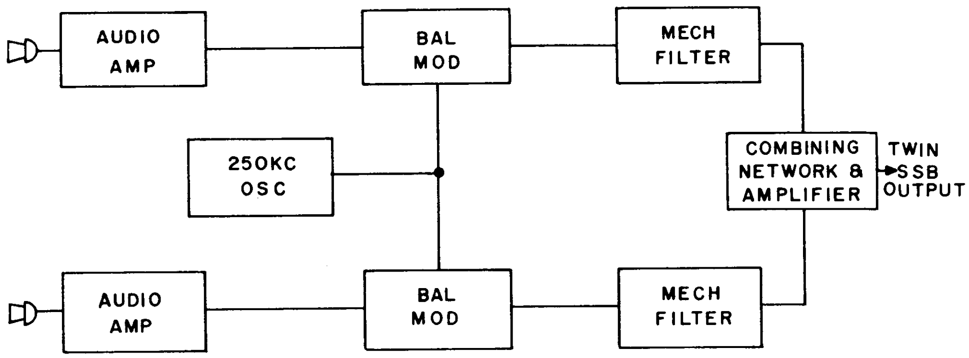


Figure 2-3. Twin SSB Generator

2-7. Balanced Modulator

A great many balanced modulator circuits are available and usable in this application. One is a simple ring modulator using a set of four matched germanium diodes. The circuit is shown in Figure 2-4. This circuit has the smallest number of modulation products in the output. Theoretically,

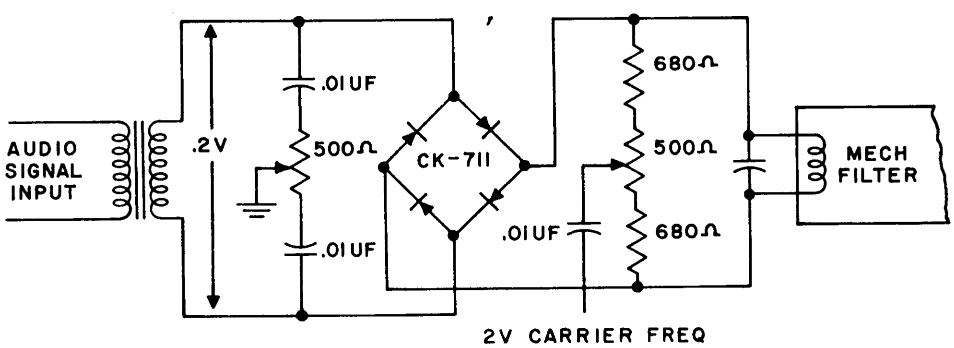


Figure 2-4. Ring Modulator Circuit

only the sum and difference of the odd harmonics of the carrier and signal frequencies appear in the output. Since the signal frequencies are at least 20 db below the carrier frequency and since the input levels are kept low to avoid distortion, the harmonics of the input signal in the output can be neglected. Thus for a 250 kc carrier frequency, the ring modulator output only contains pairs of sidebands at 250 kc, 750 kc and 1250 kc, etc. The mechanical filter, of course, does not pass any but the desired sideband at 250 kc. Only one carrier balancing potentiometer is required if the resistances on the other side are accurately center-tapped and the diodes are well balanced.

Another circuit which gives excellent performance and is very easy to adjust is shown in Figure 2-5. The adjustments for carrier balance are quickly made by adjusting the potentiometer for voltage balance and adjusting the small variable capacitor for exact phase balance of the balanced carrier voltage feeding the balanced modulator.

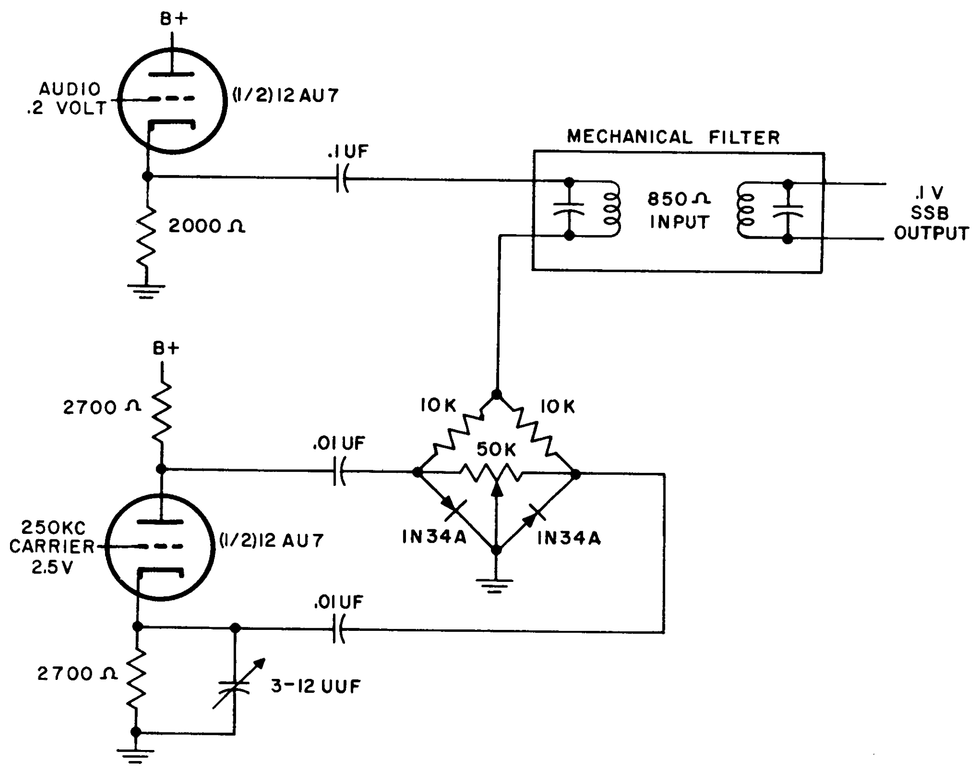


Figure 2-5. Balanced Modulator Circuit

SECTION III

AUTOMATIC LOAD CONTROL

3-1. Function of ALC

Automatic load control is a means of automatically keeping the signal level adjusted so the power amplifier is operating near its maximum power capability and also protecting it from being overdriven. In AM systems, it is common to use speech compressors and speech clipping to perform this function. These methods are not equally useful in SSB. The reason for this is that the SSB envelope is different from the audio envelope and the peaks do not necessarily correspond as explained in Section 1-5. For this reason a "compressor" located between the SSB generator and the power amplifier is most effective because it is controlled by SSB envelope peaks rather than audio peaks. This "SSB signal compressor" and the means of obtaining its control voltage is called the Automatic Load Control.

3-2. ALC Circuit Theory

A block diagram of an ALC circuit is shown in Figure 3-1. The compressor or gain control part of this circuit uses one or two stages of remote cutoff tubes such as a 6BA6 operating very similarly to the IF stages of a receiver with AVC.

The grid bias voltage which controls the gain of the tubes is obtained from a voltage detector circuit connected to the PA tube plate circuit. A large delay bias is used so that no gain reduction takes place until the signal is nearly up to the full power capability of the power amplifier. At this signal level, the rectified output overcomes the delay bias and the gain is reduced rapidly with increasing input signal so there is very little rise in output power above the threshold of gain control.

3-3. ALC Performance

3-3-1. Attack and Release Time

When a signal peak arrives that would overload the power amplifier, it is desirable that the gain of the ALC amplifier be reduced in a few milliseconds to a value where overloading the PA is overcome. After this signal peak passes, the gain should return to normal in about one-tenth second. These attack and release times are commonly used for general voice communications.

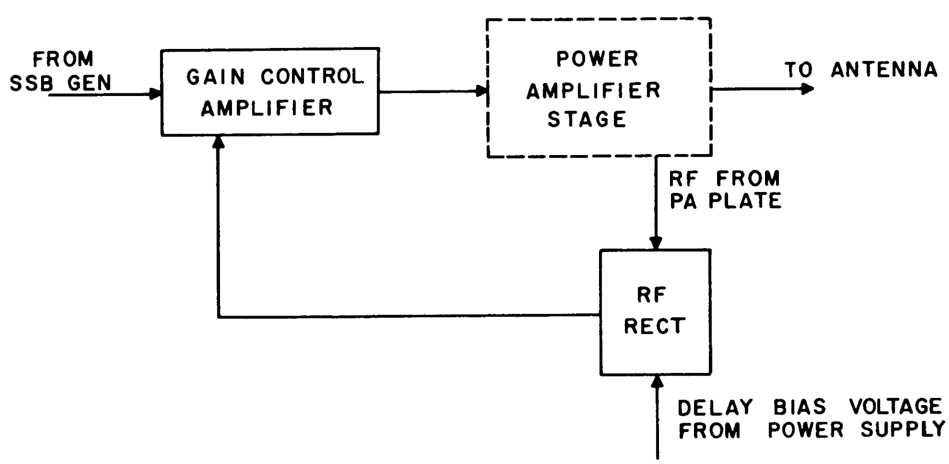


Figure 3-1. Block Diagram of Automatic Load Control

3-3-2. Range of Gain Control

The gain control range necessary depends upon the function it is to perform. In a four channel voice communication transmitter, it may be required to increase the gain when only one channel is in use at the moment to obtain best utilization of the transmitter power capability. An increase in gain of 8 db has been used for this purpose. If each channel never used more than 1/4 of the maximum voltage amplitude capability, a 12 db increase in gain would be required. Since the possibility of peaks occurring in all four channels at the same instant is quite low, the peak level of each channel can be somewhat more than 1/4 the maximum peak level. It is for this reason that only 8 db is used.

In single channel speech transmission the ALC may be required to perform the function of a speech compressor also, and a range of at least 10 db should be available for this. In general, a range of 12 db should be provided with very little distortion introduced by the ALC amplifier at any gain within this range. Input peaks as high as 20 db above the threshold of compression should not cause loss of control although some increase in distortion in the range between 12 and 20 db of compression can be tolerated because peaks in this range are infrequent. Another limitation is that the preceding SSB generator must be capable of passing signals above full power output by the amount of compression desired. Since the signal level through the SSB generator should be maintained within a limited range, it is unlikely that more than 12 db will be useful. If the input signal varies more than this, a speech compressor should be used to limit the range of the signal fed into the SSB generator.

Figure 3-2 shows the effecitiveness of the ALC in limiting the output signal to the capabilities of the power amplifier. An adjustment of the delay bias will place the threshold of compression at the desired power output.

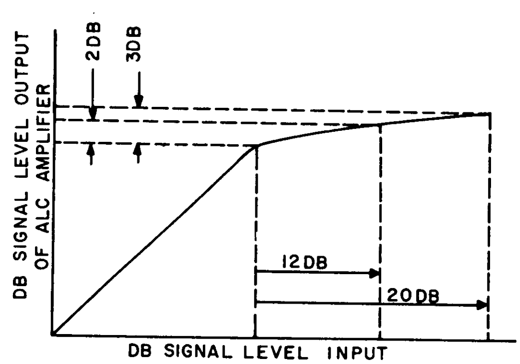


Figure 3-2. Performance Curve of ALC Circuit

3-3-3. Automatic Load Control Circuit

Figure 3-3 shows a simplified schematic of an ALC circuit. Two additional functions have been incorporated in the unit as developed and are included in this schematic. One is the isolating pads used when 2 or more SSB inputs are to be combined. These pads have at least 6 db of attenuation which provides 12 db isolation between SSB generator output. The other function is a means of carrier reinsertion for reduced carrier or SSB with carrier transmission.

This ALC uses two variable gain amplifier stages and the maximum over-all gain is only about 15 db. When twin SSB input is used, 6 db of this is lost in the input pads. Normally, the input level is about .1 volt and the output is adjusted to about .1 volt also.

A meter is incorporated which is calibrated in db of compression. This is useful in adjusting the gain for the desired amount of load control or "speech compression".

A capacity voltage divider is used to step down the RF voltage from the plate of the PA tube to about 50 volts for the ALC rectifier. The output of the ALC rectifier passes through RC networks for obtaining the desired attack and release times as well as isolating RF filter chokes and bypass capacitors. The 3300 ohm resistor and .1 mfd capacitor across the rectifier output stabilizes the gain around the ALC loop to prevent "motor-boating".

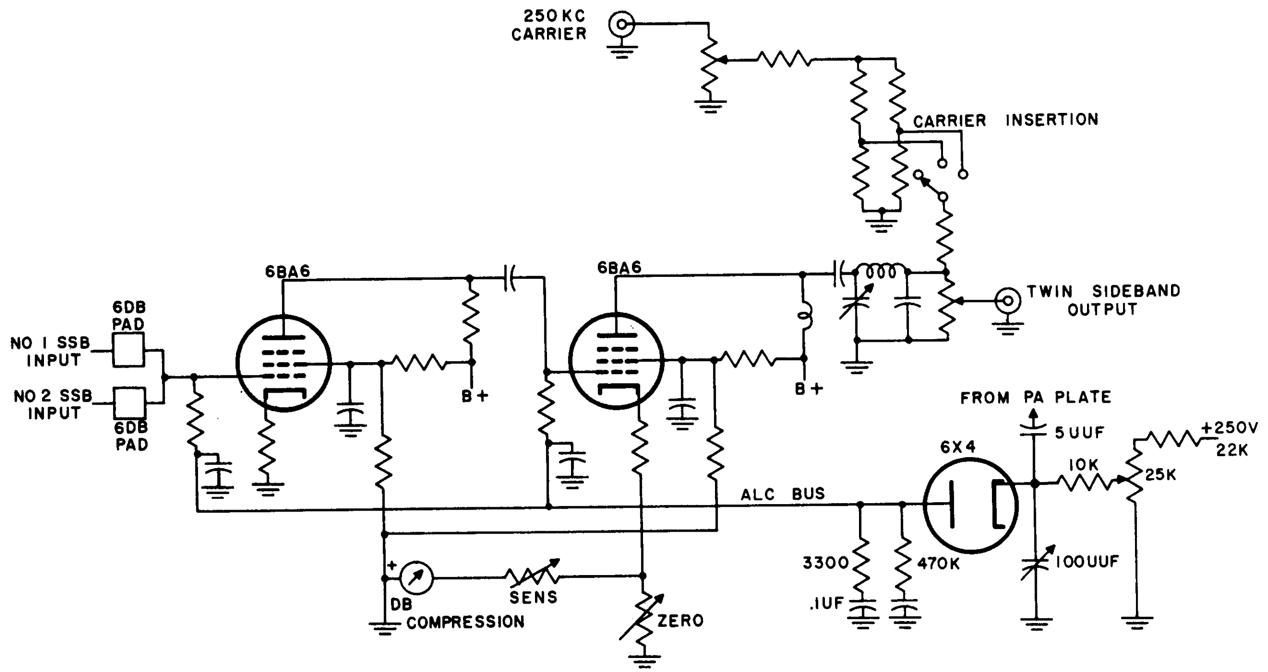


Figure 3-3. Simplified Schematic of Automatic Load Control

SECTION IV

SINGLE SIDEBAND CONVERSION FREQUENCY SCHEMES

4-1. General

Since we have selected the use of the filter method of SSB generation, the initial carrier frequency is determined by the frequency passband of the mechanical filter. Considering cost vs. performance of the mechanical filter, it is advantageous to use a frequency lower than the standard intermediate frequency of 455 kc. Using frequencies as low as 100 kc causes a difficult undesired signal rejection problem, however, so a compromise of 250 kc has been chosen for the SSB generator carrier frequency.

The band of sideband frequencies on either side of the carrier must be heterodyned up to the desired carrier frequency. In receivers the circuits which perform this function are called converters or mixers. They can also be considered as modulators. In this discussion, we will use the term "mixers".

4-2. Mixer Characteristics

One circuit which can be used for this purpose uses a receiving type mixer tube such as a 6BE6. The output signal from the SSB generator is fed into the No. 1 grid and the conversion frequency is fed into the No. 3 grid. This is opposite the usual grid connections but it gives about 10 db improvement in distortion. The plate circuit is tuned to select the desired output frequency product. Actually, the output of the mixer tube contains all harmonics of the two input signals and all possible combinations of the sum and difference frequencies of all the harmonics. In order to avoid distortion of the SSB signal, it is fed to the mixer at a low level such as .1 to .2 volts. The conversion frequency is fed in at a level about 20 db higher or about 2 volts. By this means, harmonics of the incoming SSB signal generated in the mixer tube will be very low. Usually the desired output frequency is either the sum or the difference of the SSB generator carrier frequency and the conversion frequency. For example, using an SSB generator carrier frequency of 250 kc and a conversion or injection frequency of 2000 kc as shown in Figure 4-1, the output may be tuned to select either 2250 kc or 1750 kc.

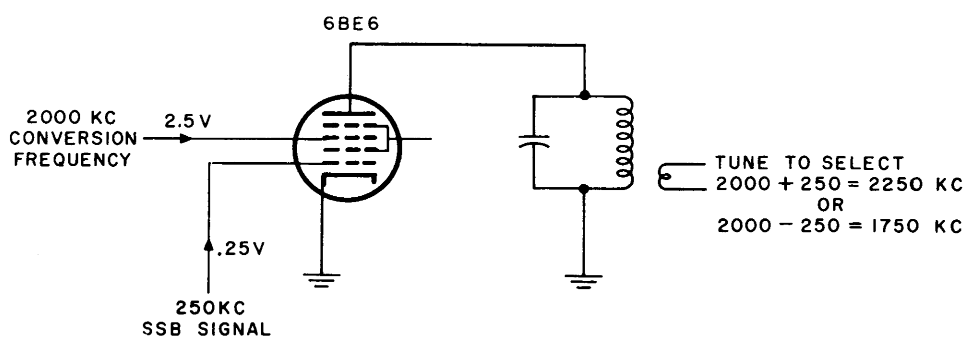


Figure 4-1. Mixer Circuit

Not only is it necessary to select the desired mixing product in the mixer output but also the undesired products must be highly attenuated to avoid having spurious output signals from the transmitter. The standards we have set up are that all spurious signals that appear within the assigned frequency channel must be at least 60 db below the desired signal and those appearing outside of the assigned frequency channel at least 80 db.

When mixing 250 kc with 2000 kc as in the above example, we may desire the 2250 kc product but the 2000 kc injection frequency will appear in the output about 20 db stronger than the desired signal. To reduce it to a level 80 db below the desired signal means that it must be attenuated 100 db.

4-2-1. Balanced Modulator Mixers

The principle advantage of using balanced modulator mixers is that the injection frequency theoretically does not appear in the output. There is a wide variety of balanced frequency that may be used. The principle by which the injection frequency is balanced out is shown by Figure 4-2. The injection frequency is fed to the two grids in phase, and

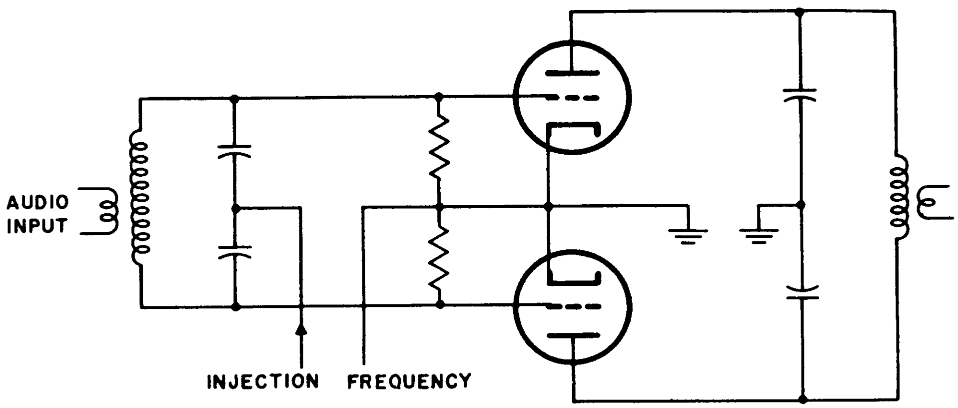


Figure 4-2. Balanced Modulator

will appear in phase at the two plates but since a push-pull circuit is used, the two injection frequency signals balance out. Another type of circuit uses push-pull input of the injection frequency and single ended signal input and output circuits. It is a common feature of all balanced modulator circuits that one of the input circuits must be a balanced push-pull circuit. This is a disadvantage when the circuits must be tunable across a wide frequency range since an accurate balance must be maintained. In practice when it is not practical to trim the push-pull circuits and the tubes into exact phase and amplitude balance, about 20 db of injection frequency cancellation is all that can be depended upon. With suitable trimming adjustments the cancellation can be 40 db, however, in fixed frequency circuits.

4-2-2. Twin Triode Mixer

The mixer circuit shown in Figure 4-3 was found to have about 10 db lower distortion than the conventional 6BE6 type converter tube. It has a

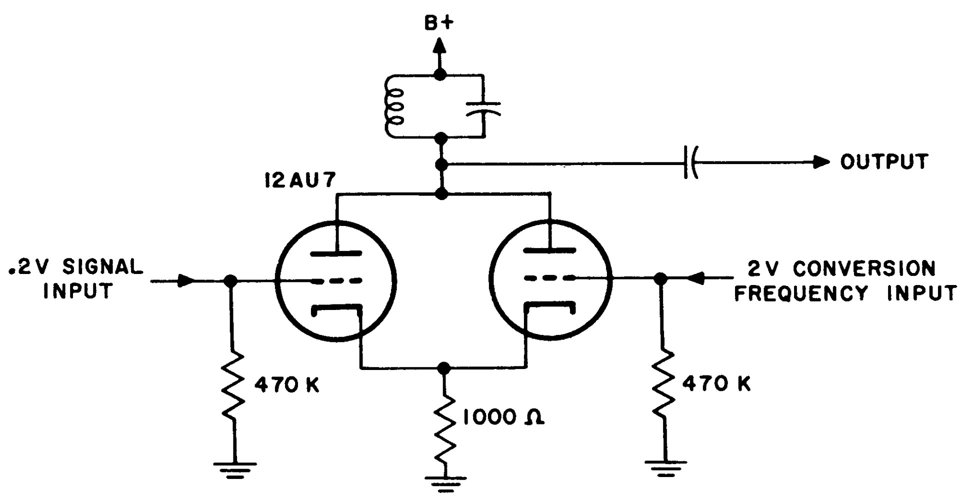


Figure 4-3. Twin Triode Mixer

lower voltage gain of about unity and a lower output impedance which loads the first tuned circuit and reduces its selectivity. In some applications the lower gain is of no consequence but the lower distortion is important enough to warrant its use in high performance equipment. The signal-to-distortion ratio of this mixer is on the order of 70 db compared to approximately 60 db for a 6BE6 mixer when the level of each of two tone signals is .5 volt. With stronger signals, the 6BE6 distortion increases very rapidly whereas the 12AU7 distortion is much better comparatively. The effective noise on the input grid is about twice as much (5 microvolts) in the triode mixer but this is low enough to be negligible in nearly all but receiver applications.

In a practical commercial equipment where the injection frequency is variable and trimming adjustments and tube selection cannot be used, it may be easier and more economical to obtain this extra 20 db of attenuation by using an extra tuned circuit in the output than by using a balanced modulator circuit.

4-3. Selective Tuned Circuits

The selectivity of the tuned circuits following a mixer stage often becomes quite severe. For example, using an input signal at 250 kc and a conversion injection frequency of 4000 kc the desired output may be 4250 kc. Passing the 4250 kc signal and the associated sidebands without attenuation and realizing 100 db of attenuation at 4000 kc which is only 250 kc away, is a practical example. Adding the requirement that this selective circuit must tune from 2250 kc to 4250 kc further complicates the basic requirement.

The best solution is to cascade the required number of tuned circuits. Since a large number of tuned circuits may be required, the most practical solution is to use permeability tuned circuits which are tracked and ganged together.

4-3-1. Number of Tuned Circuits Required

If an amplifier tube is placed between each tuned circuit, the over-all response will be the sum of their individual responses. Using identical tuned circuits, the over-all response is the response of one stage multiplied by n number of stages. Figure 4-4 is a chart which may be used to quickly determine the number of tuned circuits required. The Q of the circuits is assumed to be 50 which is normally realized in small permeability tuned coils. The number of tuned circuits with a Q of 50 required for providing 100 db attenuation of 4000 kc while passing 4250 kc is found as follows: Δf is 4250 - 4000 = 250 kc. f_r is the resonant frequency, 4250 kc. $\frac{\Delta f}{f_r} = \frac{250}{4250} =$ .059. The point on the chart where .059 intersects 100 db is between the curves for 6 and 7 tuned circuits so 7 tuned circuits are required. Another point which must b considered in practice is the tuning and tracking rror. For example, if the circuits were actually tuned to 4220 kc instead of 4250 kc, the $\frac{\Delta f}{f_r}$ would be $\frac{220}{4220}$ or 0522. Checking the curves shows that 7 circuits would just barely provide 100 db. This points up the need for very accurate tuning and tracking.

4-3-2. Coupled Tuned Circuits

When as many as 7 tuned circuits are required, it is not necessary to have the gain that the 6 isolating amplifier tubes would provide. Several vacuum tubes can be eliminated by using two or three coupled circuits between the amplifiers. With a coefficient of coupling between circuits .5 of critical coupling, the over-all response is very nearly the same as isolated circuits. The gain through a pair of circuits with .5 kc coupling is only eight-tenths that of two critically coupled circuits, however. If critical coupling is used between two tuned circuits, the nose of the response curve is broadened and about 6 db is lost on the skirts of each pair of critically coupled circuits. In some cases it may be necessary to use critically coupled circuits to broaden the nose of the response curve to avoid adversely affecting the frequency response of the desired passband. Another tuned circuit may be required to make up for the loss of attenuation on the skirts of critically coupled circuits. The type of coupling such as mutual inductance, top capacity coupling, etc., should be chosen to help maintain the desired response and constant gain characteristics over the tunable frequency range. The frequency response near resonance which effects the audio response of the sidebands can be determined by using the curves on pages 340 and 348 of Terman's "Radio Engineering".

4-4. Frequency Conversion Schemes

The example in the previous section points up the difficult selectivity problem encountered when strong undesired signals appear near the desired frequency. A high frequency SSB transmitter may be required to operate at any assigned carrier frequency in the range 1.75 mc to 30 mc. The problem is to find a practical and economical means of heterodyning the 250 kc SSB generator output up to any carrier frequency in this range. There are many modulation products in the output of the mixer and a frequency scheme must be found that will not have undesired output of appreciable amplitude at or near the desired signal. When tuning across a frequency range some products may "cross over" the desired frequency. These undesired crossover frequencies should be at least 60 db below the desired signal to meet our standards. The amplitude of the undesired products depends upon the particular characteristics of the mixer and the particular order of the product. In general, most products of the 7th order or higher will be at least 60 db down.[1] Thus any crossover frequencies lower than the 7th order must be avoided since there is no way of attenuating them if they appear within the desired pass band.

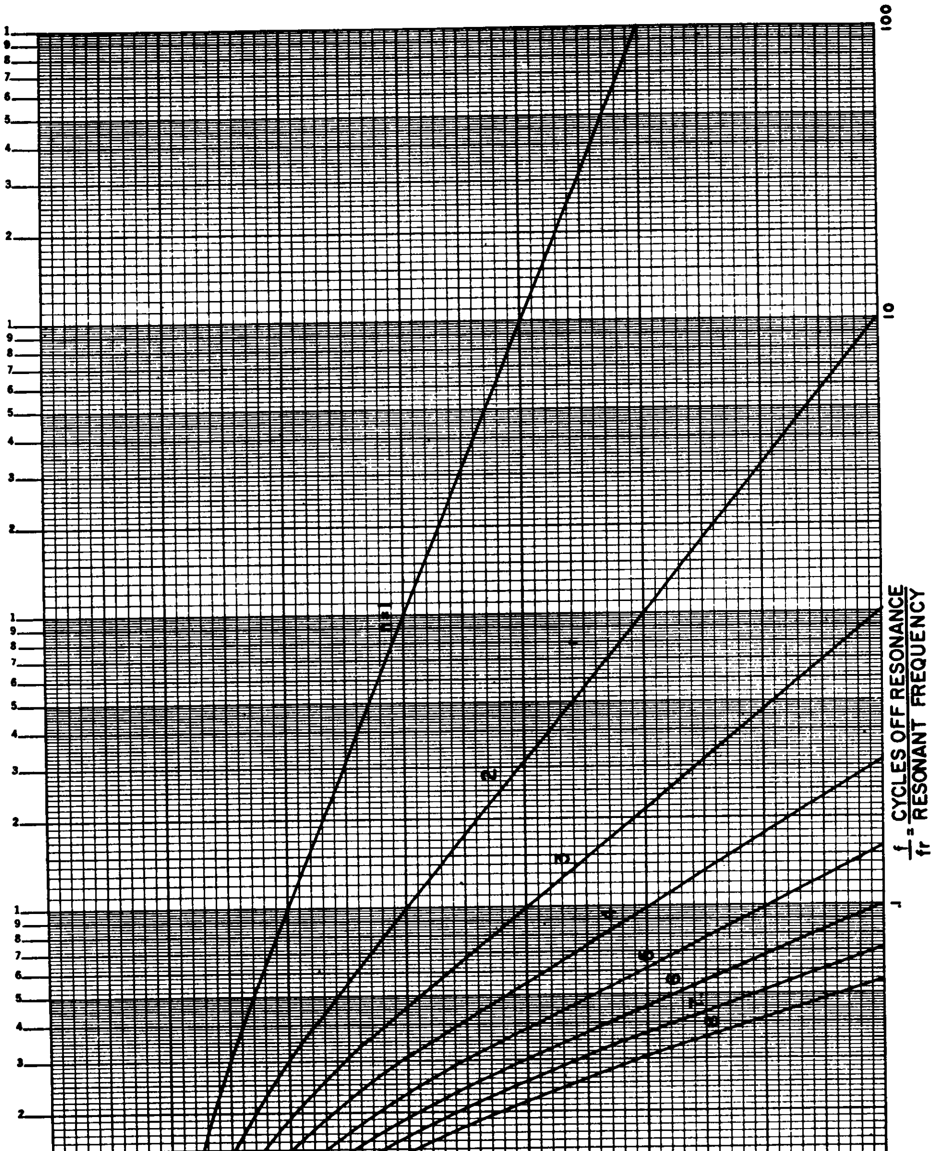


Figure 4-4. Response of "n" Number of Tuned Circuits Assuming Each Circuit Q is 50

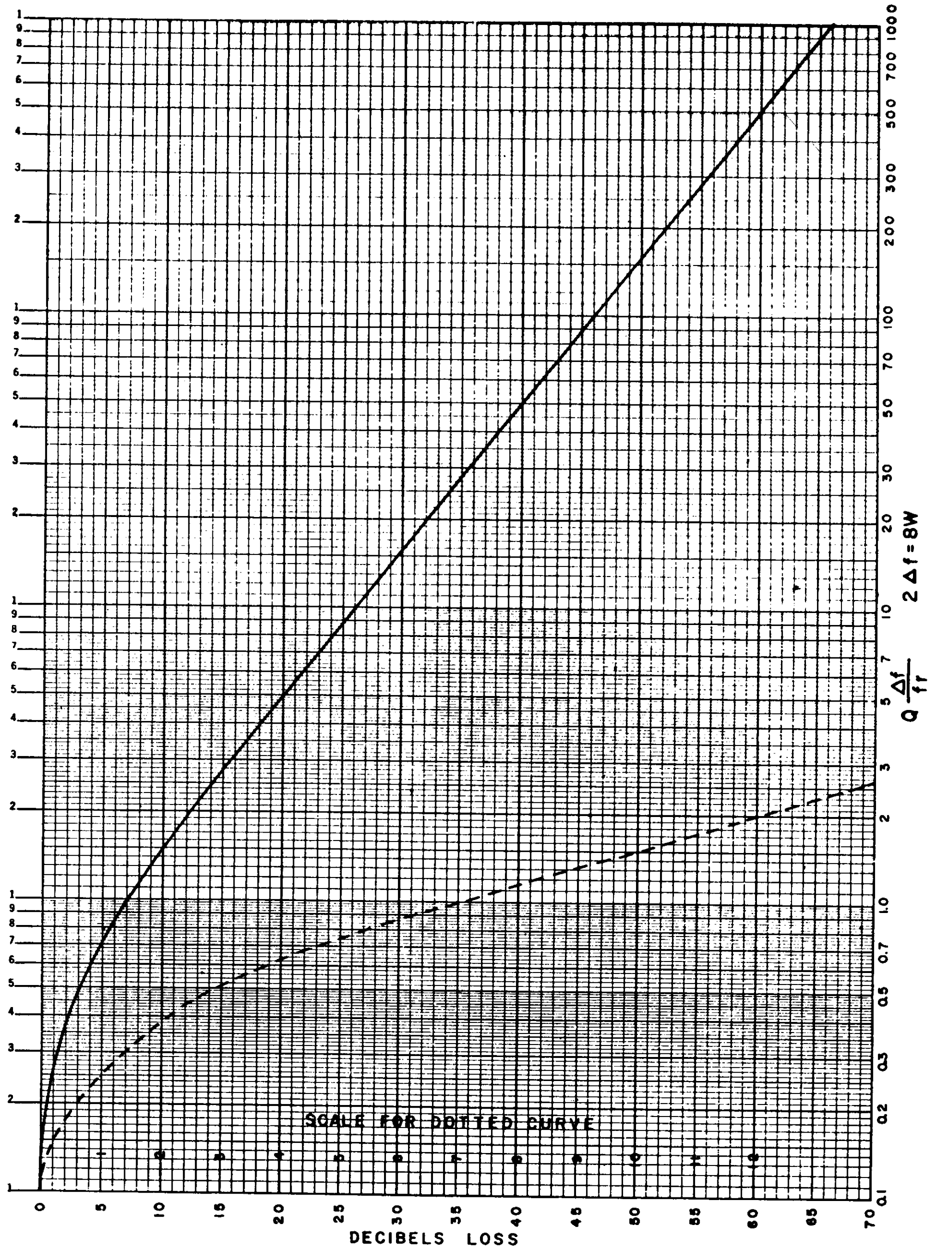


Figure 4-5. Generalized Selectivity Curve (Single Tuned Circuit)

4-4-1. Mixer Frequency Charts

There are several mixer frequency charts available to assist in working out a frequency scheme that offers the best compromise between the avoidance of undesired products and the selectivity requirements. Figures 4-6 and 4-7 show all products up to the 10th order. These figures show the products in graphical form so the crossover and nearby undesired frequencies can be readily observed. In order to work on an easier to read portion of the chart, the frequency scales may be multiplied or divided by the same number. For example, suppose the frequency range of 14 to 15 mc was to be covered using a 12 mc crystal oscillator and a 2 to 3 mc VFO. Find the point of 2 and 12 mc and also 3 and 12 mc and join them with a line on the chart. (Figure 4-7.) The undesired products over this frequency range can then be easily examined. The products due to sum mixing are the only ones of interest in this example. Starting with the VFO at 2.0 mc and increasing its frequency we find the following undesired products:

7th order at 2 + 12 = 14 mc.

6th order at 2.4 + 12 = 14.4 mc.

5th order at 3.0 + 12 = 15 mc.

Actually, there are two 5th order products appearing on the same frequency. One is 2(12) - 3(3) = 15 mc and the other is 5 times the oscillator frequency which will be much the stronger of the two. If the amplitude of this 5th order product is too high, some other frequency choice must be used to cover this band. The relative amplitude of the various mixing products is discussed in Section 4-4-2.

For a record and reference, it is suggested that lines be drawn on a mixer product chart when a frequency scheme is worked out. It will be found very useful in the development and testing of the equipment.

4-4-2. Relative Level of Mixer Products

The level of the mixer products can vary over very wide limits but some idea of the levels in our particular application will be quite useful. It has been found that the intelligence bearing signal should be at least 20 db below the conversion injection frequency input level. With levels of .1 volt and 1 to 2 volts, reasonably good conversion efficiency is realized with very low 3rd and 5th order distortion of the SSB signal. In this discussion we will assume a 10 to 1 or 20 db difference in mixer input levels.

The tube choice and its operating conditions have a pronounced effect on the amplitude of the various mixer products. For example, if the portion of the characteristic curve over which the tube is operated is a pure square function, some of the products would not be generated at all. Thus it is desirable to find a tube having an operating condition that is as near a square law characteristic as possible.

If it can be assumed that the lowest order term of the curvature appearing in the expression for the amplitude of a given product is the only one of significance, there will be a fixed relationship between the amplitude of the various products of a given order. In general, this is true but allowance must be made for perhaps 10 db error in some cases. In order to tabulate these relative levels, we must assume a fixed relative amplitude between the input signals. A ratio of $\frac{Q}{P} = \frac{10}{1}$ is used in the following table, where P and Q represent the amplitude of the two signals. If a greater ratio were used, the 3rd and higher order mixing products would be lower in amplitude than shown but the harmonics of the strong signal will be greater relative to a desired sum or difference frequency (Q + P) or (Q - P).

The relative amplitudes of each order are tabulated in Table 4-1 with the strongest product used as reference level in each order.

2nd Order

0 db	2Q
-14 db	P ± Q
-40 db	2P

3rd Order

0 db	3Q
-10 db	P ± 2Q
-30 db	2P ± Q
-60 db	3P

4th Order

0 db	4Q
-8 db	P ± 3Q
-24 db	2P ± 2Q
-48 db	3P ± Q
-80 db	4P

5th Order

0 db	5Q
-6 db	P ± 4Q
-20 db	2P ± 3Q
-40 db	3P ± 2Q
-66 db	4P ± Q
-120 db	5P

Making a further assumption relating the various order of products Table 4-1 was made. This assumption is that using the product Q ± P as reference, 2Q ± P is 1/2 or -6 db, 3Q ± P is 1/3 or -10 db and 4Q ± P is 1/4 or -12 db.

Order of Product	Level	Product
2	0	Q ± P
2	+14	2Q
3	+ 4	3Q
4	- 2	4Q
3	- 6	P ± 2Q
5	- 6	5Q
4	-10	P ± 3Q
5	-12	P ± 4Q
2	-26	2P
3	-26	2P ± Q
4	-26	2P ± 2Q
5	-26	2P ± 3Q
5	-46	3P ± 2Q
4	-50	3P ± Q
3	-56	3P
5	-72	4P ± Q
4	-82	4P
5	-126	5P

Table 4-1. Approximate Relative Amplitude of Mixer Products

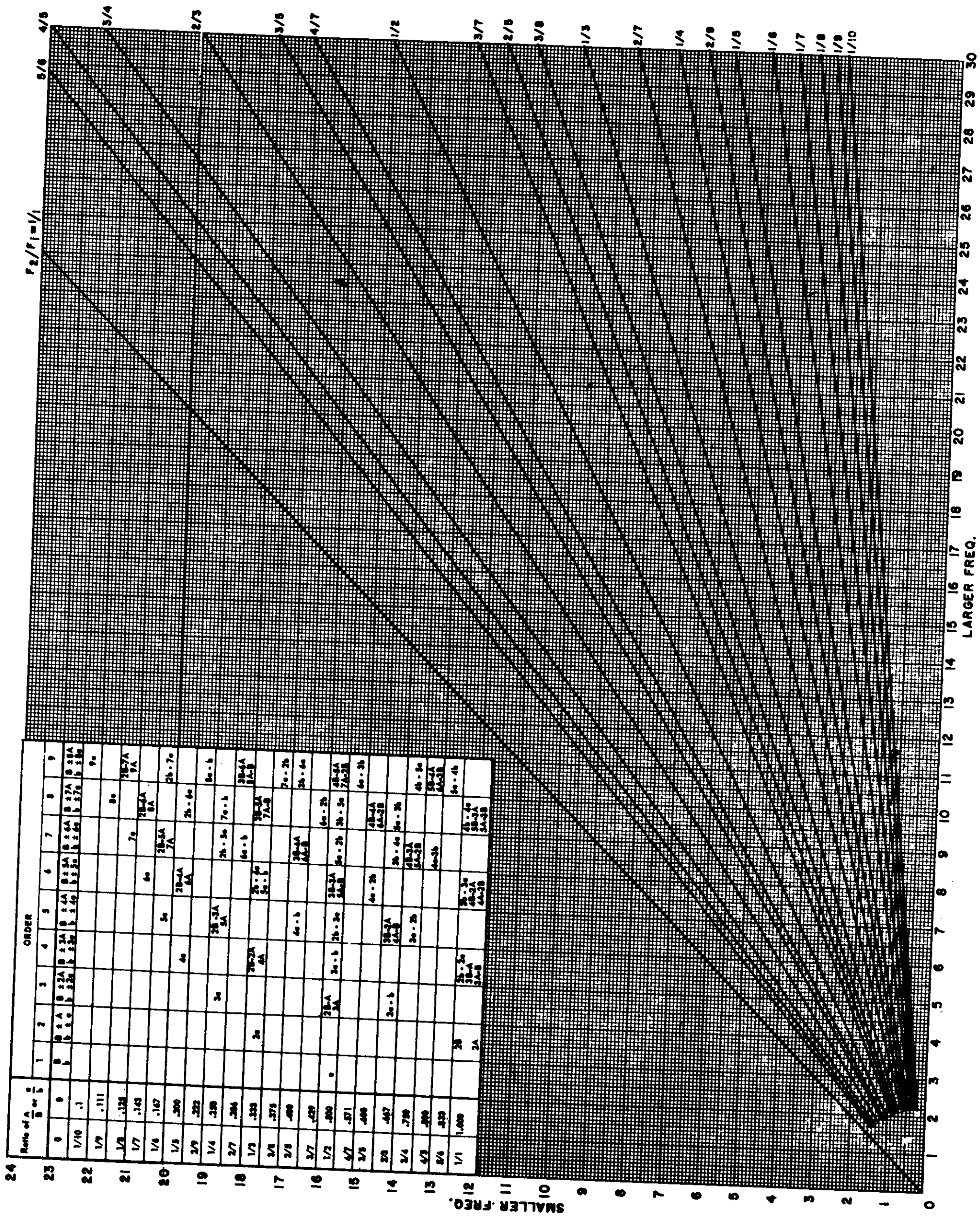


Figure 4-6. Mixer Frequency Chart

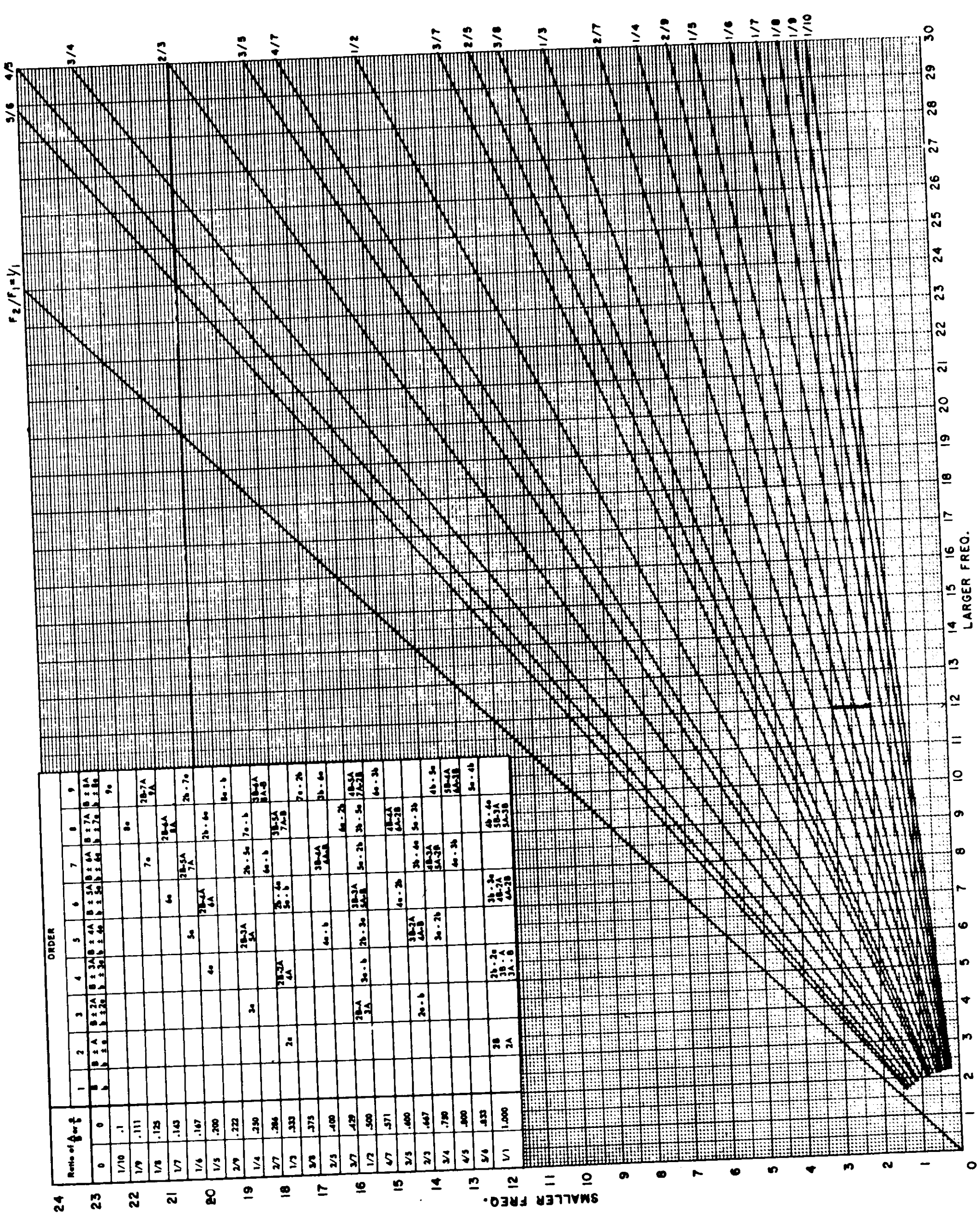


Figure 4-7. Mixer Frequency Chart Showing Undesired Products

This assumption relating the various orders probably is very pessimistic in that the higher order products may be down much more than shown as indicated by a few scattered measurements in the laboratory. An important point to note is that only products through the 5th order are listed and higher order products will exist between those shown below -6 db. The strong ones are harmonics of Q. It has been considered that any direct harmonic of the conversion frequency, Q, below the tenth must not cross over the pass band. As indicated in the table, only the 2nd and 3rd and perhaps the 4th harmonic of the signal frequency need be avoided as crossovers.

The purpose of this discussion is mainly to call attention to the relative amplitude of some of the mixing products. Knowing the amplitude of a nearby product that must be kept below a given level in the output is helpful in determining the selectivity requirements of the tuned circuits.

4-4-3. An Amateur Band Mixing Scheme

This is a typical example of the possibilities when only a few narrow bands of frequencies are to be covered. In this application it is desired to use a 250 kc SSB generator, a low frequency VFO and a high frequency crystal oscillator in order to obtain good frequency stability economically. A block diagram is shown in Figure 4-8. Even though the bands to be covered are narrow, it still is not possible to avoid all undesired products easily. Any scheme chosen is simply the best compromise in the opinion of the designer.

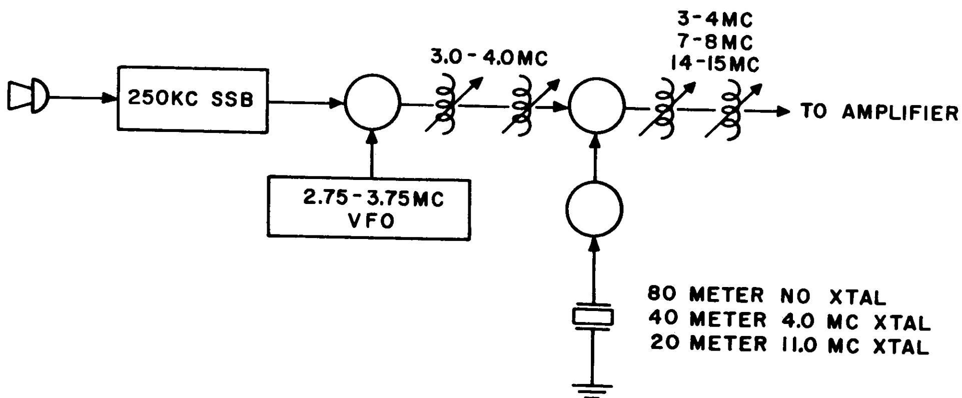


Figure 4-8. Amateur Band Frequency Scheme

4-4-4. A Continuous Coverage Mixing Scheme

When the range 1.75 to 30 mc must be covered by an SSB exciter, it becomes impractical if not impossible to accomplish this with a low frequency VFO and fixed frequency high frequency injection for each band. There is one possible solution, however, and this is to vary both input frequencies when covering a band. Figure 4-9 shows such a scheme.

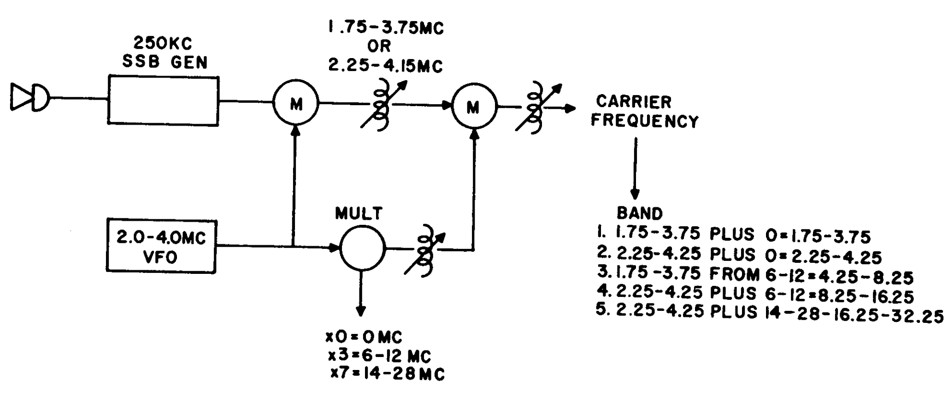


Figure 4-9. Continuous Coverage Frequency Scheme

When drawing a line between the end frequencies of a band on the mixer product chart, it is noted that it is more nearly parallel to the lines representing the mixer products. Thus a fairly wide range of frequencies can be covered in each band. Figure 4-10 shows each of the frequency bands of the scheme used in Figure 4-9 plotted on a mixer product chart. The location and order of the undesired products can be readily observed on this diagram. The nearness of strong undesired products can also be observed and this information is used to determine the selectivity requirements of the tuned circuits.

The frequency scheme for band 3 was chosen to avoid a very strong 2nd order product only 250 kc away from the output frequency which would occur if 2.25 - 4.25 plus 2.0 - 4.0 = 4.25 - 8.25 mc were used. This product is the 2nd harmonic of the VFO frequency, or 4- 8 mc and it is almost impossible to separate this from the desired product.

In general, for most applications when the intelligence bearing frequency is lower than the conversion frequency, it is desirable that the ratio of the two frequencies be between 5 to 1 and 10 to 1. This is a compromise between avoiding low order harmonics of the signal input appearing in the output and minimizing the selectivity requirements.

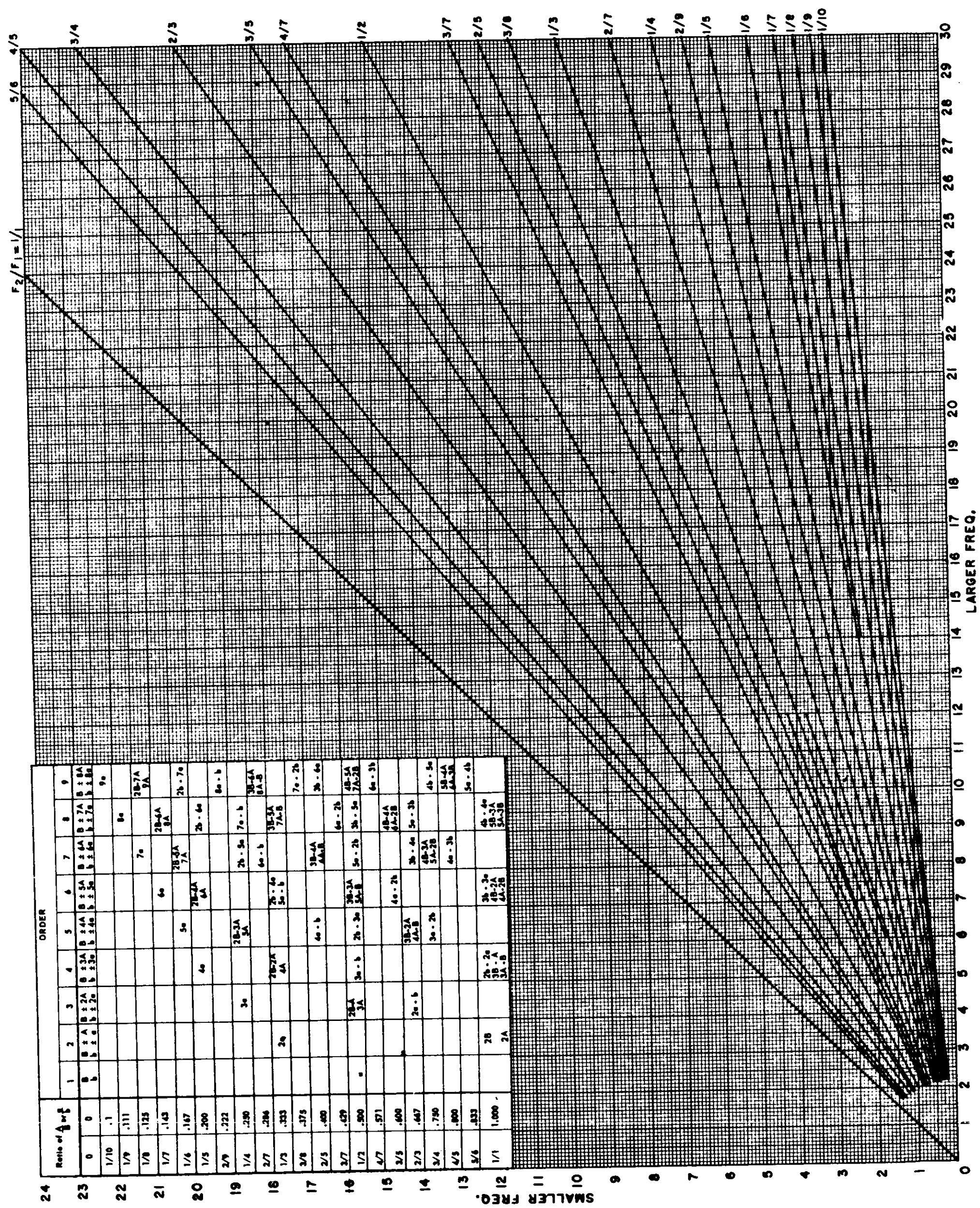


Figure 4-10. Mixer Product Chart

SECTION V

SINGLE SIDEBAND RF POWER AMPLIFIERS

5-1. Cathode Driven Linear RF Amplifiers (Grounded Grid)

5-1-1. General

The principal requirement of an SSB amplifier is that it must operate as a linear amplifier with very low distortion.

One of our first interests in this field was for an SSB power amplifier that would deliver 4 kw of peak envelope power. At that time, all tubes available for this application required that they be driven into grid current. To avoid distortion from envelope peak flattening due to grid loading of the input signal, it is necessary to devise a low RF source impedance. The other alternative is to choose a tube and its operating conditions so that the minimum possible grid driving power is required. In the past it had been the practice to simply use a grid swamping resistor to load the input circuit to 10 times the grid drive power required. This reduces the power gain per stage and is wasteful of power in the higher power stages.

There appeared to be no tetrodes suitable for this application and power level so after a brief search the field was limited to triodes as it was felt that triodes would have inherently less distortion and better efficiency. After comparing several tubes in the field, the 3X2500A-3 tube was selected.

A very objectionable feature of triode tubes is that they must be neutralized in conventional grid driven circuits. The neutralizing circuit almost doubles the input and output circuit capacities. This is undesirable in equipment operating over a wide frequency range and up to 30 mc.

5-1-2. Cathode Driven Amplifiers

The use of a cathode driven amplifier overcomes the main disadvantages just discussed. The grid acts as an excellent screen between the plate and cathode so neutralization is not required. The very small plate to cathode capacitance feeds the low input impedance of 150 to 200 ohms so very little plate to cathode coupling actually exists on frequencies below 30 mc. The fedthru power is an effective load across the input circuit so swamping resistors are not required and this fedthru power appears in the output circuit in addition to the power developed by the tubes. Also there were some not too clear ideas about the inherent feedback or degenerative nature of the grounded grid circuit helping to reduce any nonlinearity of the tubes. An amplifier using a pair of 3X2500A-3 tubes in a grounded grid circuit was built and the signal-to-distortion ratio at 4 kw output was in the range of 35 db as best as could be determined at that time.

5-1-3. 500 Watt SSB Amplifier Tests

Further interest led to means of developing a 500 watt power amplifier. Since there are no modern triode tubes with compact construction and short leads for 30 mc operation, the tetrode field was again examined. Tetrodes can also be operated as cathode driven amplifiers by bypassing both grids to ground and supplying normal DC grid bias and screen voltages. The tests and conclusions from the observations made are covered in the report made at that time and which are reviewed in the following section. Signal-to-distortion measurements were not made as the necessary equipment was not available at that time.

5-1-4. Class "B" RF Amplifier Linearity Tests on 4-400A Tetrode

An amplifier was set up for the purpose of determining the general characteristics of a 4-400A amplifier at a power level suitable for 500 watt single sideband transmitters. The amplifier was first constructed as a grounded grid tetrode amplifier upon which most of the testing was done and then it was converted to a conventional tetrode and a few comparative observations made.

In constructing the grounded grid tetrode amplifier, it was found necessary to use proper grid bypassing and short common impedance paths to avoid a parasite at 175 mc. With proper grounding this parasite was eliminated without suppressors of any kind. An outstanding characteristic of the grounded grid amplifier is apparently complete elimination of reaction of the plate circuit on the input circuit without neutralization. At our test frequency of 10 mc, it behaved like a perfectly neutralized amplifier with no trace of instability. This is attributed to the use of both grids as screens between the input and output circuits and the low input impedance of approximately 200 ohms due to the fedthru power.

For 500 watt transmitter output we chose tube operating conditions for delivering 700 watts to allow for tank circuit losses which would appear in a complete transmitter. It was thought that a pair of tubes would be required so we operated our one tube with a power to the load of 350 watts.

To get a picture presentation on an oscilloscope of the linearity of the amplifier, we modulated the RF input signal so that detectors could be placed across the input tank coil and the output circuit to give an audio output for feeding the horizontal and vertical inputs of the oscilloscope. The exciter for the amplifier consisted of a 6V6 crystal oscillator followed by a 6146 Class "C" amplifier with an adjustable plate and screen voltage for varying the output signal level for the continuous tone of measurements. The modulation necessary for the oscilloscope presentation was obtained by simply supplying variac controlled raw AG directly to the plate of the 6146 amplifier. Its RF output envelope is then in the form of half sine wave pulses similar to that of a half-wave rectifier. The shape of this modulation envelope, of course, is not important with regard to our linearity presentation.

The two diode detectors were made as nearly identical as possible and germanium diodes were selected to give matched outputs. One detector was connected directly to the input tank circuit through a capacity divider and the other detector was connected to a variable pickup link coupled to the plate tank coil. This link was varied in each case so the detector outputs were very nearly alike. By this means, the non-linearity within the detectors themselves is balanced out. This linearity tracer is more completely described in Section 6-1-1.

The most useful observation is the curve presented on the scope which will be a straight line when the amplifier is perfectly linear. The effects of varying the bias and screen voltages are observed directly so that the optimum voltages can be very quickly determined.

One purpose of our test was to find the optimum value of screen voltage. To do this, we chose screen voltages in 25 volt steps from 150 to 375 volts and made a set of measurements for each chosen value of screen voltage. The grid bias was adjusted in each case for best amplifier linearity and the loading was adjusted to the point which gave 350 watts output without getting into the knee at the upper end of the curve. It was soon found that a plate voltage of 3000 volts was much preferable to 2500 volts, so most of the data were taken with 3000 volts. Data on a few of the operating conditions are appended to this section. In general, the linearity curve of a Class "B" linear amplifier takes the form of an ess curve. The lower part of the ess is straightened out by using sufficient static plate current and the upper part of the ess is avoided by limiting the operating level to a point just below this portion of the curve.

A few tests were also made at a power level of 700 watts output from a single 4-400A amplifier. It was found necessary to go up to a plate voltage of 3500 volts for efficient operation and low driving power. A typical set of data for this class of operation is also appended. It is noted that the driving power for a single tube with 700 watts output is about identical with the driving power required for a pair of tubes delivering the same power. With a little heavier drive and loading and tolerating a little more non-linearity, an output of 1 kw could be readily obtained.

The following are some observations and general conclusions:

(a) For a triode tube operating Class "B" RF linear amplifier there is a certain value of static current necessary for the most linear operating condition. This value of plate current is determined by the tube characteristics themselves and is not appreciably affected by any operating conditions such as DC plate voltage, loading, etc.

(b) Higher DC plate voltages result in better efficiency and lower driving power. Lower driving power refers here to the power taken by the grids rather than the fedthru power although the fedthru power in a grounded grid amplifier does reduce slightly with higher DC plate voltages. The reduced fedthru power associated with higher DC plate voltages results in a slight increase in the power gain of a grounded grid amplifier.

(c) The maximum DC plate voltage is limited by the static plate dissipation of the tube since it is necessary to use a given static plate current for the reason given above. The DC plate voltage should be chosen so that the static plate dissipation will be well below the maximum allowable plate dissipation of the tube.

(d) In tetrode linear amplifier circuits, the screen voltage has a very pronounced effect on the tube characteristics. Lower screen voltages affect the tube characteristics so that a lower value of static current is required for optimum linearity. The static plate current required varies as the three-halves power of the screen voltage, which means that the optimum static plate current is more than doubled when changing the screen voltage from 300 volts to 500 volts DC. Thus lower screen voltages with their attendant low static plate currents are very advantageous from a static plate dissipation standpoint so that much higher intermittent peaks could be handled in an ordinary single sideband voice transmission. Lower screen voltages with their lower static plate currents will allow higher DC plate voltages then without exceeding a reasonable value of plate dissipation.

(e) Lower screen voltages also allow the use of greater values of RF plate voltage swing. To avoid operating up in the nonlinear knee portion of the curve, the minimum plate swing must be kept substantially above screen voltage. Lower screen voltages allow greater plate swing for this reason and it seems that the plate swing can be increased substantially more than the reduction in screen voltage for a given amount of linearity. This greater allowable RF swing, of course, results in a substantial improvement in the plate efficiency.

(f) The use of lower screen voltage has the adverse effect of increasing the grid driving power. In a grounded grid tetrode amplifier, the screen consumes driving power as well as the No. 1 or control grid. The choice of screen voltage affects the amount of the No. 1 grid current and the screen grid current but in general for a given power output when one goes up, the other one goes down to some extent.

(g) In a grounded grid amplifier, whether it be triode or tetrode, it is desirable to have a fairly large ratio of fedthru power to peak grid driving power. The fedthru power in effect acts as a swamping resistor across the circuit to reduce the effect of grid loading. It has been stated in the literature that it is desirable to have this ratio be at least 10 to 1. However, for the high degree of linearity we are striving for we probably should use ratios on the order of 20 to 1 if practical unless it is possible to obtain a very low driving source impedance.

(h) With a single 4-400A operating as a straight-up tetrode amplifier, it is possible to obtain 350 watts output in a Class "AB_1" operating condition. The screen voltage necessary is 500 volts and the plate voltage is 3000 volts. Neither the control grid nor the screen grid draw current until the point of 350 watts output is reached where they both start drawing current. The static plate current for best linearity is actually above 133 ma which gives 400 watts of static plate dissipation. This test was made using 130 ma of static plate current. Lower static plate currents, can of course, be used if the attendant nonlinearity can be tolerated.

(i) The high screen voltage required for Class "AB_1" operation also results in low RF plate swing for a good degree of linearity which results in quite low plate efficiency.

(j) A grounded grid tetrode amplifier could be operated under similar voltage and bias conditions as a conventional amplifier so that it too will operate without grid current or grid driving power. The fedthru power would not be affected greatly by this class of operation, however. It is doubtful whether this operating condition would serve advantageously in a power amplifier as explained in Section 5-1-5.

(k) In general, it is not believed that this degeneration in a grounded grid amplifier can be used to advantage because the effect of grid loading on the over-all linearity is much more important than the tube linearity itself. For this reason the tube operating conditions should be adjusted for best tube linearity and a low driving source impedance.

(l) It was noted that the optimum load on the tube usually occurred at approximately at the point of maximum power output for a fixed input RF voltage. This adjustment, of course, should be made at about one-half of maximum input voltage to avoid getting into a condition of plate saturation.

(m) The input impedance of a grounded grid amplifier is directly proportional to the plate load impedance except for the grid current loading.

(n) The proper loading on a conventional tetrode linear amplifier can be accomplished by adjusting the loading for the proper predetermined ratio of RF grid swing to RF plate swing. Although not checked, it is believed that this statement is also true for a grounded grid amplifier providing the adjustment is made well below the point of plate saturation or heavy grid current.

(o) For grounded grid operation of the 4-400A, a screen voltage of 300 volts (filament to screen) gives a reasonable compromise between low driving power, low static plate current and good plate efficiency.

(p) The high power gain of a "straight-up" tetrode Class "B" RF amplifier requires the virtual elimination of stray plate to grid coupling and near perfect neutralization.

It is pointed out again that the proper choice of static plate current straightens out the lower portion of the ess curve. The upper part is affected mostly by screen voltage and minimum plate swing so that for some conditions where substantial nonlinearity can be tolerated, the static plate current for best over-all linearity may be lower than that indicated only by the lower part of the curve.

The following sets of data are included to show a few sets of operating conditions. The loading, bias and excitation were juggled around for what looked like the most linear condition as observed on the oscilloscope and are not necessarily optimum. All data except the last set are for grounded grid tetrode operation. All voltages are with respect to the cathode. The zero signal grid and screen currents were always zero. The plate dissipation is calculated and really includes the tank circuit loss. The grid drive was calculated by multiplying the peak cathode swing by the sum of DC grid and screen currents. The fedthru power was calculated by the ratio of cathode swing to plate swing times the power output. This data represent a single tone test.

	Grounded Grid		Grounded Cathode
DC screen voltage	300	300	500 volts
DC plate voltage	3000	3500	3000 volts
Static plate current	60	60	130 ma
DC grid bias	-60	-59	-80 volts
Peak cathode swing	87	113	0 volts
Peak grid swing	0	0	82 volts
Min. plate voltage	660	500	600 volts
Max. Sig. grid current	3.6	10	0 ma
Max. Sig. screen current	4.1	20	.5 ma
Max. Sig. plate current	195	267	220 ma
Max. Sig. plate dissipation	235	235	310 watts
Static plate dissipation	180	210	400 watts
Grid driving power	.63	3.4	0 watts
Fedthru power	6.55	15.8	0 watts
Power output	350	700	350 watts

The following data is for a conventional neutralized tetrode amplifier. The operating condition was Class "AB_1" and neither grid nor screen drew current up to the point of 350 watts output where they both started drawing current. The linearity curve was not as good as for the previous grounded grid tests.

5-1-5. Cascade Class AB_1 Cathode Driven Amplifiers

As has been noted, the only way that advantage can be taken of the inherent feedback in a cathode driven amplifier is to drive it with a source of a high internal impedance. By operating the tubes without grid current this advantage can be realized. Several stages can be cascaded as shown in Figure 5-1.

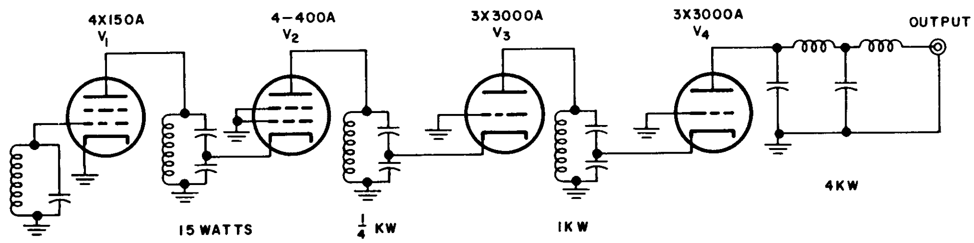


Figure 5-1. Cascade Class AB_1 Cathode Driven Amplifiers

It is noted that it is a characteristic of cathode driven amplifiers that a variation or change in load impedance on the stage output reflects back as a nearly proportional change on the input because the input and output circuits are essentially in series. Thus a change in load of V_4 reflects all the way back to the input of V^2 and the plate load of V_1.

Similarly, a change in gain (such as may be caused by nonlinearity of gain with signal level) of a cathode driven stage reflects back as a change of the input impedance of that stage and also back through the preceding cathode driven stages.

If the first cathode driven stage is fed by a source with high internal impedance such as a tetrode amplifier, changes in gain or nonlinearity in any cathode driven stage will be compensated for. For example, let the gain of V_4 decrease slightly. This causes the input impedance to V_4 to increase which reflects back through V_3 and V_2 so that the load resistance on V_1 increases. Since V_1 has high internal resistance the output voltage will rise when its load resistance increases.

This increased voltage will raise the output voltage of each GG stage so that the output is nearly up to where it should be even though the gain of V_4 is low. A change in gain or nonlinearity of any other preceding grounded grid stage will be compensated for also by the high source impedance of V_1.

It should be noted that the stages are directly coupled with very low phase shift between stages although 180° networks could be just used so the input impedance is directly proportional to the load impedance. This circuit can be shown in a simplified manner as follows:

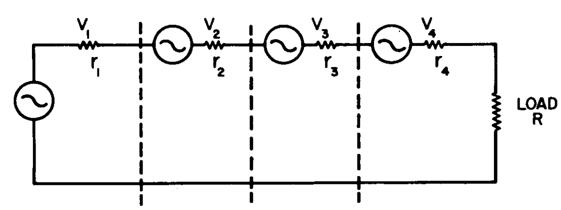


Figure 5-2. Equivalent Circuit of Figure 5-1.

r_1 has very high resistance and r_2, r_3, r_4 have very low resistance compared to the load R.

The only tube that can operate Class AB_1 and give much power with reasonable efficiency is the 3X3000A-1. This is a low mu triode and as a result, the stage gain will be about 4. Such low stage gains have the disadvantage of the requirement for more stages and tuned circuits. Another disadvantage is the large power dissipated in the interstage circuits because of the high driving voltage required and the high circuit capacitances across these tank circuits. Also, feeding the filaments with this much RF voltage on them presents a problem. For these reasons, this type of circuit was not tried.

5-1-6. Feeding Filaments of Cathode Driven Tubes

Cathode driven amplifiers in the high frequency range are usually operated with the grid at RF ground potential. The filament is then "hot" with RF. One method of feeding AC heating power to the filaments is to place a choke coil of large diameter tubing across the input circuit and run the filament conductors up through this choke coil. Both ends of the filament conductors are bypassed to the tubing choke coil. This is the method used in the present Collins 30 to 50 kw transmitters. Another method which has been used successfully is to wind the cathode tank circuit coil with large tubing and run the filament conductors up through it. The inductance of the coil can be varied by a sliding contact on it and this has no effect on feeding the filaments. These two methods have the disadvantage of requiring a physically large choke or tank coil and long small sized filament conductors. The voltage drop in these conductors must be made up by using a higher voltage filament transformer. If it is desired to meter the filament voltage at the tube, another pair of metering wires must be run up through the coil.

From the circuit standpoint, the simplest method is to use a special low capacitance filament transformer that can have RF between the filament winding and ground. A filament transformer was built for a 3X2500A-3 tube to examine the possibilities in this direction. A "hypersil" core with a cross section of 1-1/16" by 2-13/16" and a window area of 1-11/16" by 4-1/4" was used. The primary consisting of 275 turns of #18 wire was wound on one leg. The secondary or filament winding consisted of 11 turns of #10 copper bus wire supported on grooved mycalex insulating strips. The temperature rise of the primary winding was 20°C after two hours of operation. The #10 wire ran quite hot but it was found that the actual dissipation in this winding was less than that of an ordinary type winding. Also, it was anticipated that this transformer could be located in the plenum chamber for the tube so it could have adequate cooling. The one transformer built, operated at maximum flux density at 60 cps so the design would require more turns to keep the flux density down for 50 cps operation. The regulation from no load to full load was 10%. The capacitance of the filament winding to core was only 30 mmfd. The basic construction allows the use of shields if desired to confine the RF fields and currents to desired paths.

5-1-7. Conclusion Regarding Cathode Driven Operation

Since it appears impractical to operate available tubes Class AB_1, i. e., with no grid current and no screen current, the possibility of taking advantage of the inherent feedback in the circuits is lost because a low driving source impedance is then required. The power gain per stage is moderate with a gain of about 10 for a 3X2500A-3 triode stage and a gain of around 25 for a tetrode stage being about maximum. The linearity is the same as for conventional grid driven circuits with low source impedance and about 30 db signal-to-distortion ratio is about all that can be expected. The amplifiers are comparatively very stable and relatively free of parasites. They are the best choice in high power applications where the linearity requirements are not too high.

5-2. Evolution of PA With Feedback

Early testing with large receiving and small transmitting type tubes showed that it was very difficult to find tubes and operating conditions that would give an S/D ratio of 40 db or better. The power obtainable at this S/D figure was very low and on the order of 1 watt. The S/D ratio for most tubes operating somewhere near their full output was usually about 30 db or less. This made it apparent that tubes had to be run at very low outputs to yield a transmitter capable of even a 30 db S/D ratio. This led to ideas for reducing the distortion in the tubes.

5-2-1. Cathode Follower

The first attempt was to use tubes in a cathode follower circuit to take advantage of the inherent degeneration. This met with some success in regard to obtaining low distortion but the immediate difficulty was the low gain per stage. The voltage gain is less than unity through the tube so all gain had to be achieved by a voltage step-up in the tank circuits. This was limited by the dissipation of the tank coils since the circuit capacity across the coil was appreciable. The resulting gain per stage was very low and the tuning was sharp because of the high Q circuits.

While looking for ways to overcome the difficulty of the feeding of the filament of high power cathode follower stages, it was found that cathode follower performance of the tube could be retained by moving the ground point of the circuit from the plate to the cathode as shown in Figure 5-3(a) and (b).

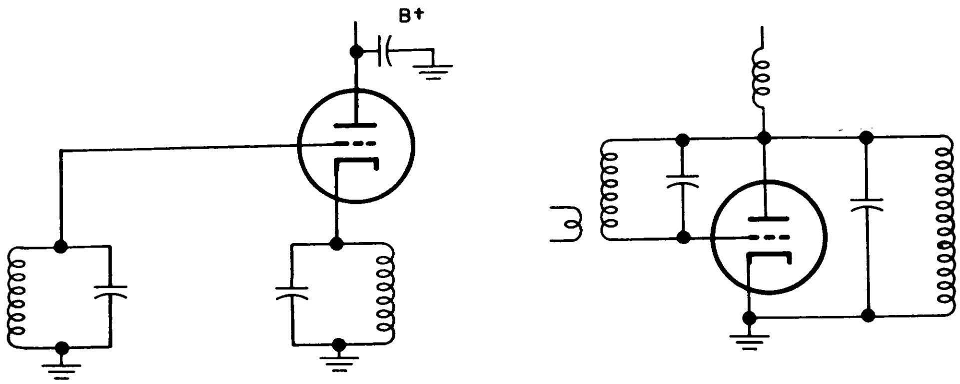


Figure 5-3. Cathode Follower Circuits with Different Ground Points

In Figure 5-3(b) it is noted that both ends of the input tank circuit are "hot" so inductive coupling from the previous stage is almost necessary.

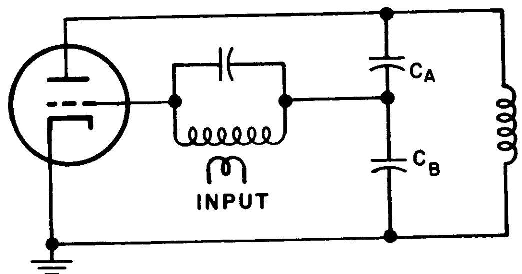


Figure 5-4. Single Stage with Feedback

5-2-2. Single Stage with Feedback

Inspection of Figure 5-3(b) revealed that by moving the top end of the input tank down on a voltage divider tap across the plate tank circuit, the feedback could be reduced from 100%, as in the cathode follower, down to any value desired. This circuit is shown in Figure 5-4. This made a more practical circuit since the losses in the input tank were greatly reduced. (The neutralization required is discussed in Section 5-3-2.) Twelve db of feedback was chosen as a good compromise between distortion and stage gain. The voltage developed across C_b is then 3 times the grid-to-cathode voltage. Experiments with this circuit were encouraging but the requirement of inductive coupling between stages was very undesirable. To overcome this, the circuit in Figure 5-5.was developed by simply moving the ground point to the point common to both tank circuits. The advantages of direct coupling between stages far outweighed the disadvantages of having the RF feedback voltage appear on the cathode.

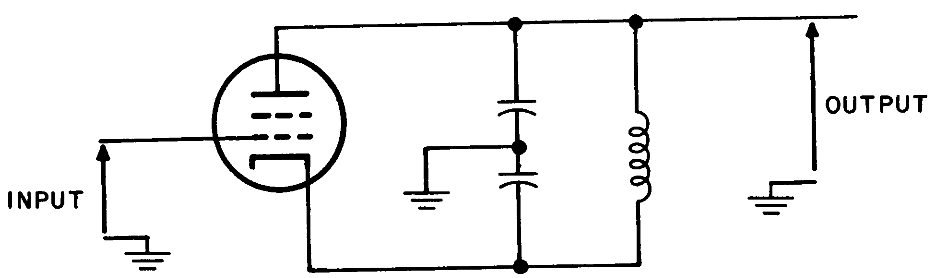


Figure 5-5. Single Stage with Feedback and Unbalanced Input and Output Connections

Separate stages using a 6AG7, 6L6, 4X150A and a 4-1000A were built, using this basic circuit. In the low power stages a plate load impedance of 5000 ohms was used for stability reasons and because the coil loss can be expected to give such a value at 30 mc. High transconductance pentodes and tetrodes are very desirable to maintain a good voltage gain per stage because the tube gain is reduced by the 12 db of feedback used. Tubes such as the 6AG7, 6CL6 and the 4X150A yielded voltage gains of 10 to 12 whereas the 6L6 only yielded a gain of 5.

The PA stage required a means of matching to 52 ohms and Figure 5-6 shows the circuit used.

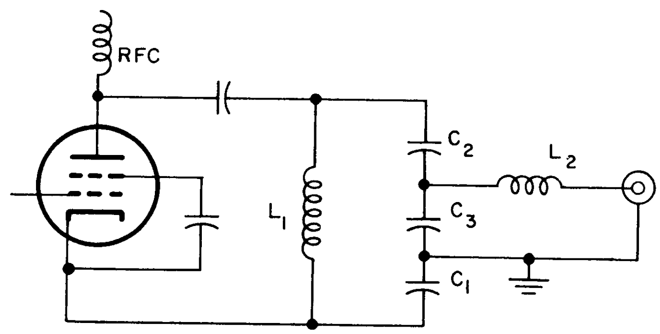


Figure 5-6. Amplifier with Feedback and Impedance Matching Output Network

Tuning this circuit presented quite a problem, however. The ratio of X_{LI} to X_{CI} determines the feedback so it is necessary to vary them together when changing frequency. The tuning and loading functions are then accomplished by varying C_2 and C_3. L_2 can also be varied to adjust the loading. Although this circuit had satisfactory performance, the tuning requirements were very undesirable. Also feeding AC to the filament with RF on it was objectionable. It should be noted, however, that with proper neutralization no difficulty was had with the feedback and there was no tendency toward instability. Theoretically, it should be stable because aside from some stray lead inductances and capacitances the phase shift within the feedback loop is limited to that of the single tank circuit which does not exceed 90°

5-2-3. Feedback Around a Two-Stage Amplifier

The maximum phase shift possible by two simple tuned circuits does not exceed 180° so feedback around two stages should also be stable. A gain-phase plot was made assuming tank circuits with identical values of loaded Q. This plot is shown by the outer curve in Figure 5-7. The inner curve is the gain-phase plot of a single tuned circuit. If the Q of the two tuned circuits are not equal, their gain-phase curve will lie somewhere between these two curves. It is apparent that stability should be easily achieved if additional phase shift due to stray inductance and capacitance can be kept small and if less than 20 db of feedback is used.

Another major point of interest is that 12 db of feedback around two stages is essentially just about as effective in reducing distortion as the use of 12 db around each of the two stages individually. Thus by using feedback around two stages an advantage of 12 db voltage gain over each pair of stages is realized.

The basic circuit of the two stage feedback amplifier is shown in Figure 5-16 of Section 5-3. It is noted that voltage is fed from a voltage divider across the output circuit directly to the cathode of the first tube. The input tank circuit thus is outside the feedback loop. The neutralizing and tuning requirements are discussed in Section 5-3.

This circuit also overcomes two other objectionable features of the single stage feedback circuit. One is that the filament of the output stage can now be operated at RF ground potential. (Incidentally, this also overcame a difficult low frequency parasite problem discussed in Section 5-5). The other important advantage is that any conventional output network may be used.

Two such two stage amplifiers were built. One for low level using a 6AG7 and a 4X150A and another for 1 kw output using a 4X150A and 4-1000A. They operated successfully on the test frequency of 4 mc and the distortion reduction was approximately the same as the amount of feedback used. Operating the low power amplifier at a level sufficient to drive the second one yielded S/D ratios on the order of 50 db. The output amplifier stages yielded S/D ratios on the order of 40 db.

The amplifiers were then rebuilt to operate at 30 mc. The feedback circuits were found to operate successfully. The miscellaneous difficulties encountered in the amplifier development were never found to be caused directly by the feedback.

Quite a bit of effort was spent working out the neutralizing requirements. The feedback and neutralizing circuits can be integrated together where practical. This is covered in detail in Section 5-3.

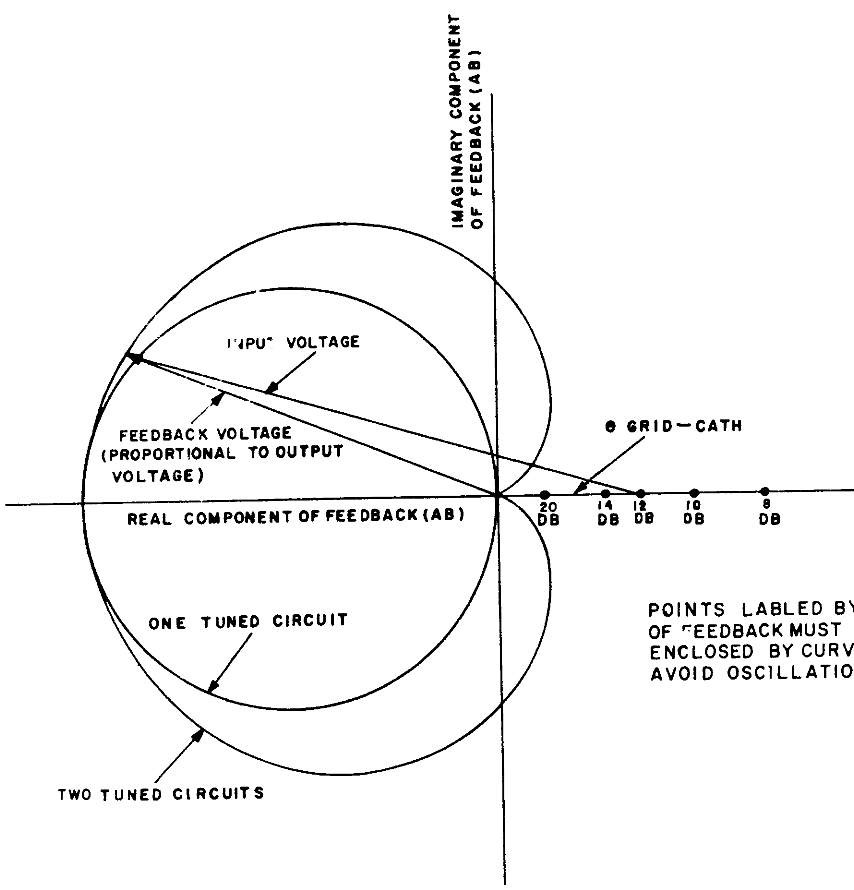


Figure 5-7. Phase Gain Plot of One and Two Tuned Circuits

The interstage circuit required for overcoming a low frequency parasite is covered in Section 5-5.

The 4X150A amplifier was developed further for a 100 watt final amplifier. A 6CL6 driver stage was used and it had to be neutralized and loaded for good stability. A pair of 6AU6's parallel connected were tried without success. The capacitance between the grid and plate leads connecting the two sockets together was more than could be tolerated. One 6AU6 was tried but it was necessary to operate it far above its plate dissipation rating to drive the 4X150A. It performed well and did not require neutralization however. After a search for a better tube, the 6CL6 was used again. The 6CL6 and 6AG7 tubes were designed for video use and operate fairly well in our application. It would help a great deal if a similar tube were available but designed for high frequency RF use and have much lower grid-to-plate capacitance.

The 4X150A delivered 100 watts operating Class AB_1. Up to 150 watts output could be obtained operating Class AB_2 but the distortion was down to about 25 db. A design was frozen and several amplifier units were built. The schematic is shown in Figure 5-8 and the following table shows typical operating conditions.

	6CL6	4X150A
D. C. plate voltage	250	1250 volts
D. C. screen voltage	150	250 volts
D. C. grid voltage	-3	-45 volts
D. C. plate current (static)	45 ma	50 ma
D. C. plate current (2-tone)	45	120 ma
D. C. plate current (1-tone)	45	150 ma
D. C. grid current	0	0 ma
Peak RF grid voltage	3	45 volts
Plate power input (2 tone)	--	150 watts
Peak envelope power output	--	100 watts (into 52 load)
Signal-to-distortion-ratio at 100 watts P. E. P.	--	40 db

Table 5-1. Typical Operating Condition of Amplifier Units

The operating condition of this amplifier can be charged appreciably by simply varying the loading of the power amplifier. An optimum valve of plate load resistance can be found for each set of D. C. supply voltages. Figure 5-9 shows curves that were calculated for a 4X150A power amplifier tube. It is noted that about 5000 ohms plate resistance is optimum for the plate and screen voltages chosen. The distortion is dependent upon the operating condition and increases rapidly when the plate voltage swings below the screen voltage. These curves were calculated for a single tone test condition.

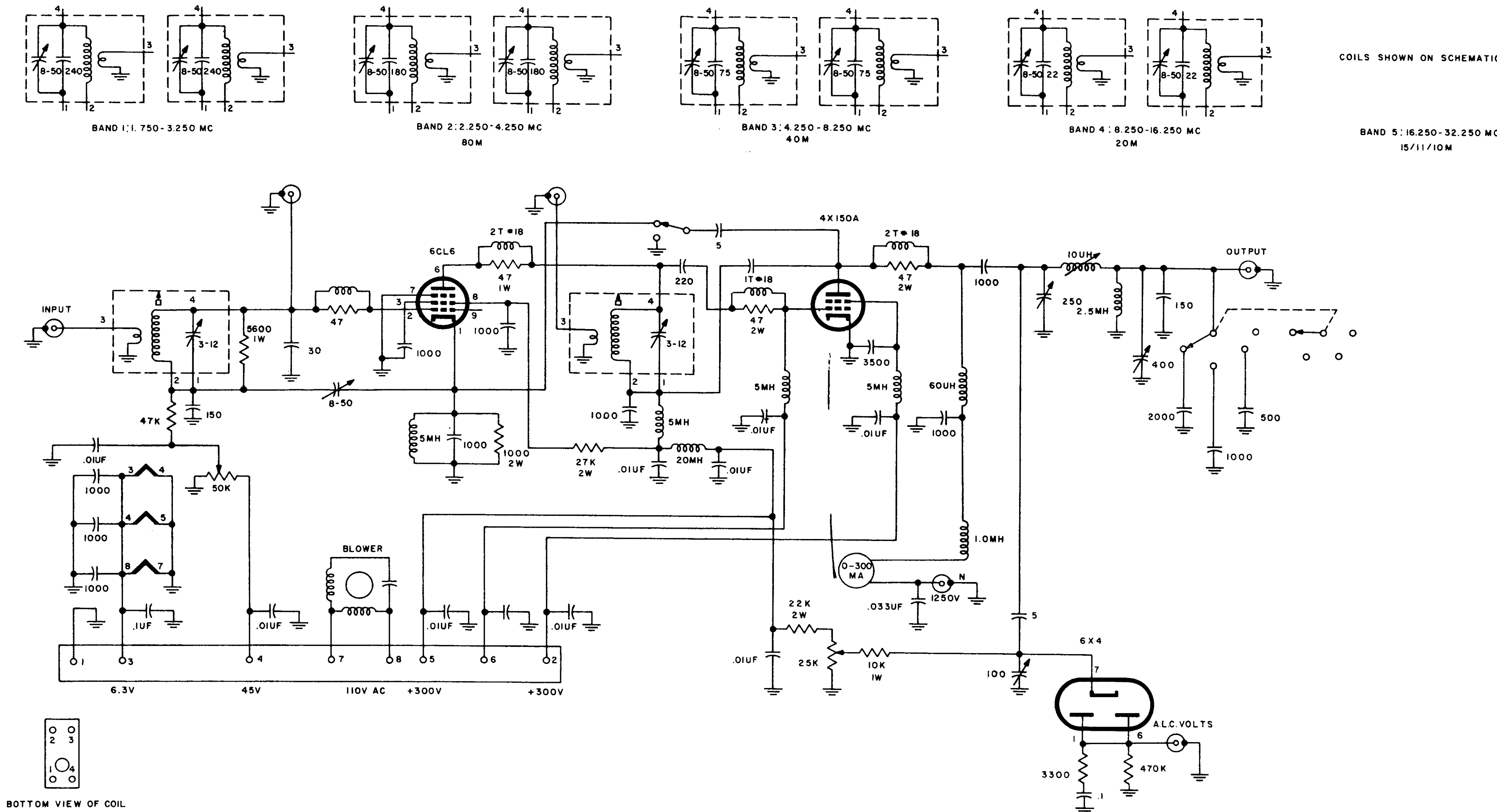


Figure 5-8. Schematic of 100 Watt Class AB, Amplifier

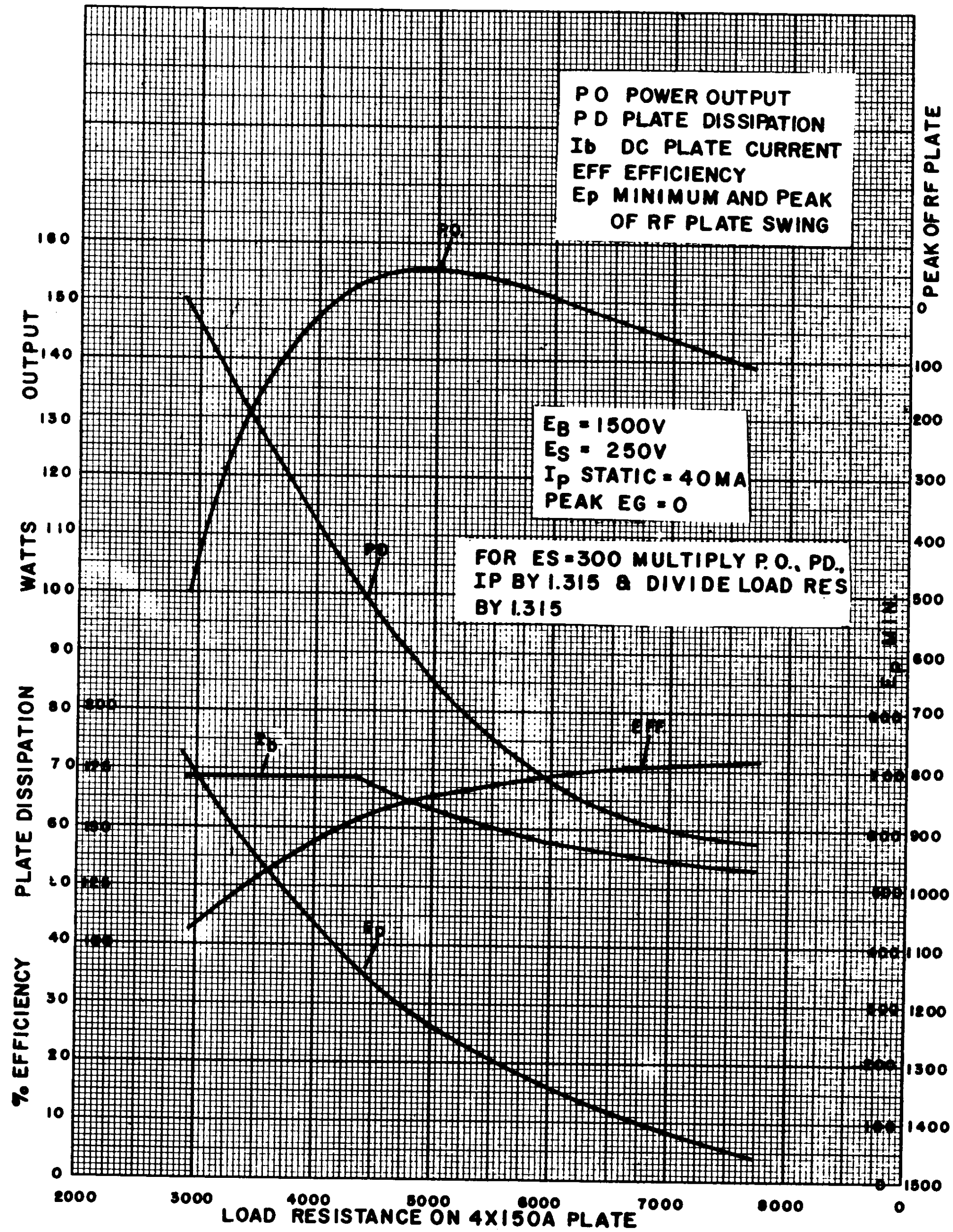


Figure 5-9. Calculated Curves for a 4X150A Power Amplifier Tube

An amplifier using a pair of 4-1000A tubes in parallel driven by a single 4X150A was built and operated successfully. This was followed by another amplifier using three 4-1000A tubes in parallel as the prototype for a 4 kw envelope power amplifier. In each of these amplifiers improvements in the control of lead inductance, stray capacity, parasitic circuits, screen and filament bypassing and mechanical layout were made.

At first, it was planned to use two cascade two stage feedback amplifiers. The first would use a 6CL6 and a 4X150A and the second would use a 4X150A and three 4-1000A tubes. The first one had to be loaded fairly well to keep noise from being troublesome. Also, in a practical equipment, some means of keeping the SWR low on the coaxial cable between the amplifiers is required. The input circuit to the 4X150A in the high powered amplifier was changed by removing the tuned input circuit and replacing it with a resistor of about 300 ohms and a pair of balun coils to step the impedance up by a 4 to 1 ratio from the 52 ohm input. Not only did this remove a tuned circuit but because of the low input resistance, it was no longer necessary to neutralize the grid-to-plate capacity of the 4X150A input stage.

Some experimenting was done to determine the effect of varying the swamping resistor across the interstage tank circuit of the 4 kw feedback amplifier. Since changing this resistor also affected the feedback and gain, the effect could not be determined directly but it seemed to have little effect. A value of 5000 ohms was chosen as being a good compromise between circuit stability and a good margin of reserve gain from the first stage so the feedback would be effective.

Some curves of S/D ratio vs. signal level were run on a 4X150A. It was noted that the S/D ratio was down in the range of 40 to 50 db when operating at 25% or less of maximum signal level. Since this is the only suitable tube available for exciting the 4X150A in the 4 kw feedback amplifier, it was thought that this tube could give enough gain at low distortion to replace the low power feedback amplifier. A new 5 kw amplifier using a pair of AX9907R tubes was built up using this idea. For some reason, the distortion out of this first 4X150A gets about 6 db worse when exciting the second 4X150A. This is true even when feedback around the 4X150A and AX9907R's is removed.

The reason for this has not been determined. This effect is undesirable because the distortion in the first stage approaches that of the last two stages and results in degraded over-all performance.

5-2-4. Advantages of Two Stage Feedback Amplifier

The main advantage is that it makes possible and practical S/D ratios higher than can otherwise be achieved. Many other advantages appeared as the development progressed. These advantages are listed below:

(1) Makes S/D ratios of 40 db practical.

(2) Makes use of high gain tetrodes so the number of stages required are very few. (.1 watt to 5 kw in three stages).

(3) Allows tubes to be run at higher efficiency and closer to maximum output.

(4) Allows less static plate current to be used in the PA tubes.

(5) Fewer tuned circuits because of fewer stages.

(6) Internal impedance and regulation requirements of power supplies is less severe.

(7) Keeps amplifier gain constant.

5-3. Neutralization and RF Feedback

It is usually the purpose of neutralization to balance out the effects of the plate-to-grid capacitance coupling in a tuned rf amplifier.

5-3-1. Effects of Plate-to-Grid Capacitance

In a conventional tuned rf amplifier using a tetrode tube, the effective input capacity of the tube is given by:

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

where C_{in} = tube input capacitance.

C_{gp} = grid to plate capacitance.

A = voltage amplification from grid to plate.

θ = phase angle of plate load.

In an unneutralized 4-1000A amplifier with a gain of 33, a calculation shows that the input capacitance of the tube with the plate circuit in resonance is increased by 8.1 mmf due to the unneutralized C_{gp}. This is not particularly important in amplifiers where the gain (A) remains constant but if the tube gain does vary, serious detuning and rf phase shift may result. A grid or screen modulated rf amplifier is an example where the gain varies from a maximum down to zero. The gain of a tetrode or pentode rf amplifier operating below plate saturation varies with loading so that if it drives a following stage into grid current the loading increases and the gain falls off. The possibility of undesirable effects from unneutralized or inaccurately neutralized C_{gp} may be made by estimating the change in gain, using equation 1, then determining the effect this change in input capacity has on a grid circuit with a known or estimated Q and capacitance.

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

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

This resistance is in shunt with the grid current loading, grid tank circuit losses and driving source impedance. When the plate circuit is inductive there is energy transferred from the plate to the grid circuit through the plate-to-grid capacitance (positive feedback) which 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.

When the plate circuit is in resonance ($\theta = o$) the input resistance due to the plate-to-grid capacitance becomes infinite. As the plate circuit is tuned to the capacitive side of resonance, the input resistance becomes positive and power is actually 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 with the plate circuit on the low frequency side of resonance to a high value on the high frequency side of resonance. Grid current, of course, is a common indication of rf voltage on the grid but it is the voltage change which is of primary concern. In a pentode or tetrode amplifier operating Class A or AB_1, the effect of plate-to-grid feedback can be observed by placing an rf voltmeter across the grid circuit and observing the change as the plate circuit is tuned through resonance.

If the amplifier is over-neutralized, the effects reverse so that with the plate on the low frequency side of resonance the grid voltage is high and on the high frequency side, it is low. One useful "rule of thumb" method of checking the neutralization of a stage (assuming it is nearly correct to start with) is to tune both grid and plate circuits to resonance. Then observing the rf grid voltmeter, tune the plate circuit to the high frequency (low C) side of resonance. If the indication goes up, more neutralization is required and if it goes down, less is required. This indication is very sensitive in a neutralized triode Class C amplifier and correct neutralization exists when the grid voltage (or current) peaks at the plate current dip. In tetrode power amplifiers this indication is less pronounced. Sometimes in a supposedly neutralized tetrode amplifier, there is practically no change in grid voltage as the plate circuit is tuned through resonance and in some amplifiers it is unchanged on one side of resonance and drops slightly on the other. Another observation sometimes made is a small dip in the center of a broad peak of rf voltage. These various effects are probably caused by coupling from the plate to grid circuit through other paths which are not balanced out by the neutralizing circuit used.

5-3-2. Neutralizing Circuits.

There are many neutralizing circuits available and most of those developed for use with triodes may also be used with tetrodes. The trend in rf power amplifier design is toward single ended stages and those circuits which require balanced tank circuits for neutralizing purposes only are undesirable for many reasons. One circuit which is being used extensively for neutralizing tetrode power amplifiers is a form of grid neutralization. This is shown in Figure 5-10. Its similarity to a conventional grid neutralized amplifier is shown by comparing this circuit with Figure 5-11.

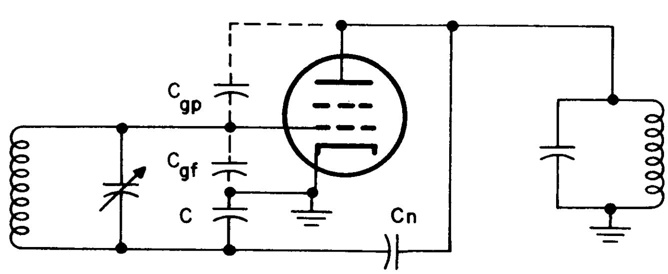


Figure 5-10. Bruene Neutralizing Circuit

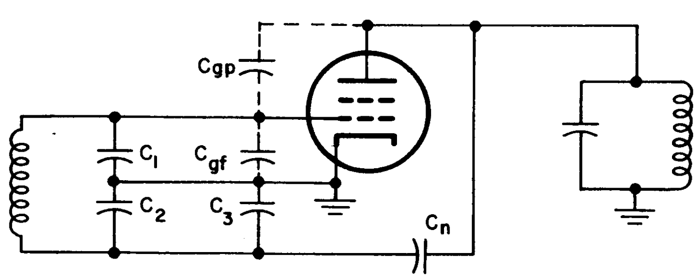


Figure 5-11. Conventional Grid Neutralized Amplifier

C_3 is used to balance the grid to filament tube capacitance to keep the grid circuit balanced. When $C_1 = C_2$ and $C_n = C_{gp}$, it is readily observed that when a signal is introduced into the grid circuit, none will appear across the plate circuit since the coupling through C_n is equal and opposite to the coupling through C_{gp}. Figure 5-12 makes this more clear.

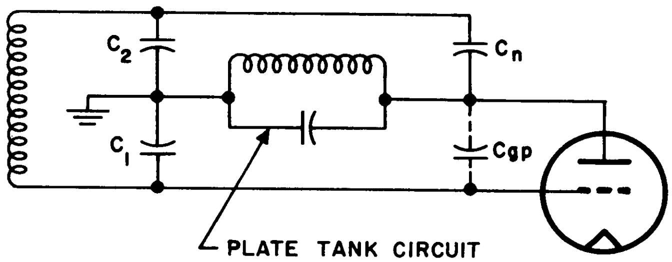


Figure 5-12. Simplified Schematic of Grid Neutralizating Circuit

The relationship for no coupling from one circuit to the other when a signal is introduced into one circuit, is given by the relationship

$$\frac{C_1}{C_2} = \frac{C_{gp}}{C_n} \tag{3}$$

This indicates that the grid tank circuit need not be balanced. If C_2 is made larger than C_n must be correspondingly larger. In a tetrode amplifier C_{gp} is very small (approximately .1 mmf) so that practical values (5 mmf) can be used for C_n when C_2 is very much larger than C_1 (approximately 50 times). The remaining change to make Figure 5-11 like Figure 5-10 is to place most of the grid tuning capacitance across the grid tank coil and let the voltage division be made by using the grid to filament capacity (both inside and outside of the tube) for C_1 and the bypass condenser from the bottom end of the grid tank circuit to ground for C_2. The relationship for neutralization in Figure 5-10 is

$$\frac{C_n}{C} = \frac{C_{gp}}{C_{gf}} \tag{4}$$

This relationship assumes perfect screen and filament bypassing and negligible effects from stray inductances and capacitances.

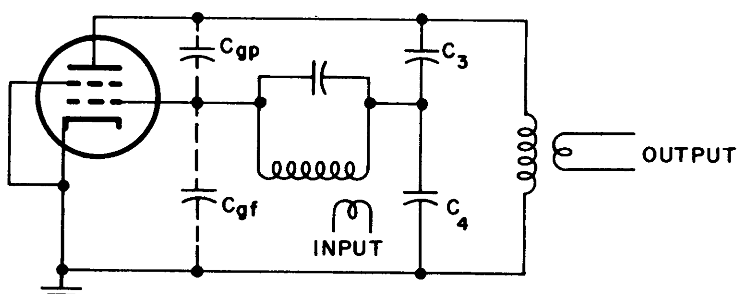


Figure 5-13. RF Amplifier with Feedback

5-3-3. Feedback Around a One-stage RF Amplifier

Figure 5-13 shows an rf amplifier with negative feedback. The voltage developed across C_4 due to the voltage divider action of C_3 and C_4 is introduced in series with the voltage developed across the grid tank circuit and is in phase opposition to it. The feedback can be made any value from zero to 100% by properly choosing the values of C_3 and C_4.

It is necessary to neutralize this amplifier and the relationship for neutralization is:

$$\frac{C_{gp}}{C_{gf}} = \frac{C_3}{C_4} \tag{5}$$

It is usually necessary to add capacity from plate to grid to satisfy this relationship.

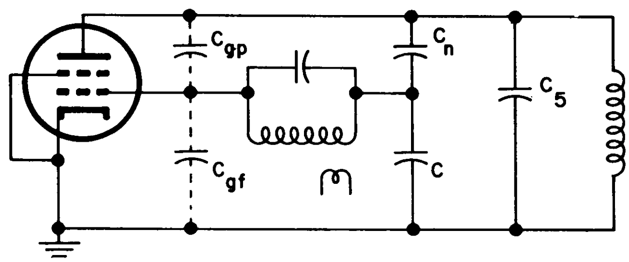


Figure 5-14. Neutralized Amplifier and Inherent Feedback Circuit

Figure 5-14 is identical to Figure 1 except it is drawn to show the feedback inherent in this neutralizing circuit more clearly. Since Cgp is very small in tetrode power amplifiers, the resulting feedback is usually quite small and is on the order of 1 db with 2 db being about maximum. This feedback has sometimes been referred to as degeneration in grid neutralizing circuits. Figure 5-14 is also the same as Figure 5-13 except that C_n and C replace C_3 and C_4 and the main plate tank circuit capacity is C_5.

Using the circuit of Figure 5-13 presents a problem in coupling into the grid circuit. Inductive coupling is ideal but the extra tank circuits complicates the tuning of a transmitter which uses several cascaded amplifiers with feedback around each. The grid could be capacity coupled to a high source impedance driver such as a tetrode or pentode plate but the driver then cannot use feedback because this would cause the source impedance to be low. One possibility is to move the circuit ground point from the cathode to the bottom end of the grid tank circuit. The feedback voltage then appears between each cathode and ground. Redrawing Figure 5-13 to change the ground point gives the circuit shown in Figure 5-15. The input can be capacity coupled and the plate can be capacity coupled to the next stage. This is a satisfactory circuit for the low powered stages for this reason. Also cathode type tubes are available so the heaters can remain at ground potential. The output voltage available with capacity coupling, of course, is less than the plate-to-cathode rf voltage developed by the amount of feedback voltage across C_4.

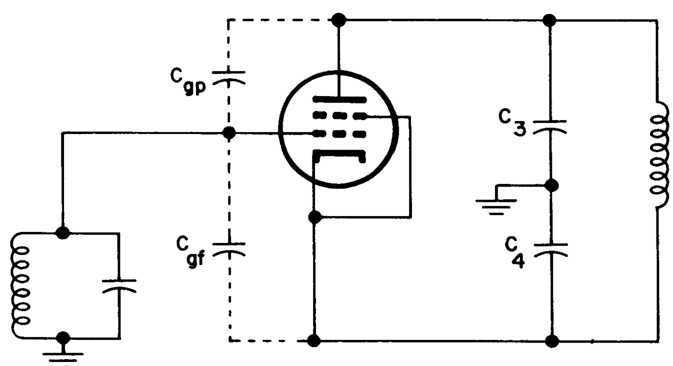


Figure 5-15. Unbalanced Input to RF Amplifier with Feedback

Amplifier stages using 6AG7, 6L6 and 4X150A tubes were built using this circuit with satisfactory performance. No difficulty with the feedback was experienced and each amplifier was stable at operating frequency using 12 db of feedback. Low frequency parasitic oscillations were encountered in each amplifier due to the cathode, screen and plate feed rf chokes. They were eliminated by the use of series plate feed, large coupling and blocking condensers and shunt resistance loading of the screen or cathode chokes. A stage using a 4-1000A was built using this feedback circuit but the low frequency parasites were particularly troublesome since the value of inductance in the chokes is necessarily low. Also the filament was "hot" so the entire filament transformer was floated at rf with bypassing around it and rf chokes placed in the transformer primary. This circuit was abandoned before the parasites were completely eliminated.

One fact worth noting is that the voltage gain per stage is reduced by the amount of feedback employed so that with 12 db of feedback the gain is reduced to one-quarter. To realize a good gain per stage such high trans-conductance tubes as the SAG7 and 4X150A are required. These tubes will provide a net gain of 10 to 15 times whereas a 6L6 only provides a net gain of 5.

5-3-4. Feedback Around Two RF Stages

Feedback around two rf stages has the advantage that more of the tube gain can be realized and nearly as much distortion reduction can be realized using 12 db around both stages as is realized using 12 db around each of two stages separately.

The phase shift in one tuned circuit cannot exceed 90 degrees so two circuits can only cause a phase shift approaching 180 degrees. At 180 degrees the gain should be zero so an amplifier using two tuned circuits within the feedback loop should be stable with a good margin for stray phase shifts due to coupling condensers, lead lengths, etcc

Figure 5-16 shows the basic circuit of a two-stage amplifier with feedback. Inductive input coupling is used and any form of output coupling may be used. The small feedback voltage required is obtained from the voltage divider C_6 and C_7 and is

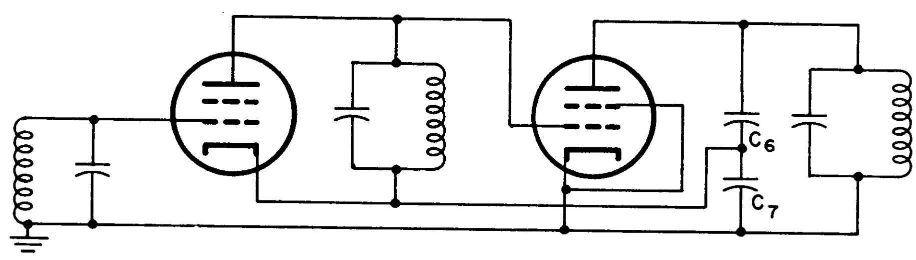


Figure 5-16. Basic Circuit of Two-Stage Amplifier with Feedback

applied to the cathode of the first tube. C_6 is only a few mmfd so this feedback divider may be left fixed for a wide frequency range. If the combined tube gain is 160 times and 12 db of feedback is desired, the ratio of C_7 of C_6 is 40 which may be 400 mmf to 2.5 mmf for example. A complication is introduced into this simplified circuit by the cathode to grid capacity of the first tube which causes an undesired coupling to the input grid circuit. It is necessary to balance or neutralize out this capacitance coupling.

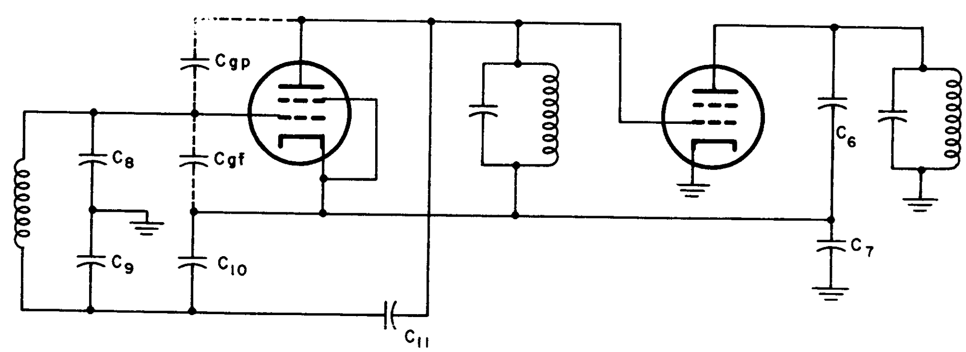


Figure 5-17. Two-Stage Amplifier with Feedback Including Circuit for Neutralizing the Cathode to Grid Capacity of the First Tube

A circuit to accomplish this is shown in Figure 5-17. The relationship of the capacitors is

$$\frac{C_8}{C_9} = \frac{C_{gf}}{C_{10}} \tag{6}$$

The input circuit may be unbalanced by making C_9 perhaps five times C_8 to reduce the voltage across the input tank coil and minimize the power dissipated by the coil. Then C_{10} of course, must also be five times C_{gf}.

Except for tubes with very small plate to grid capacity, it is still necessary to properly neutralize both tubes. If the ratio of C_6 to C_7 is chosen to be equal to the ratio of C_{gp} to C_{gf} in the second tube, this tube will be neutralized. Tubes such a 6AU6 or a 4X150A have very low C_{gp} and probably will not need to be neutralized when used in the first stage. If neutralization is necessary, condenser C_{11} is added for this purpose and the proper value is given by the following relationship:

$$\frac{C_{gp}}{C_{11}} = \frac{C_{gf}}{C_{10}} = \frac{C_8}{C_9} \tag{7}$$

If neither tube requires neutralization, the bottom end of the interstage tank circuit may be grounded. The screen and suppressor of the first tube should also be grounded then to keep the tank output capacity directly across this interstage circuit and to avoid common coupling between the feedback on the cathode and the interstage circuit. A slight amount of degeneration occurs in the first stage since the tube also acts as a grounded grid amplifier with the screen as the grounded grid. The mu of the screen is much lower than that of the control grid so that this effect may be unnoticed and would only require slightly more feedback from the output stage to overcome this.

5-3-5. Tests for Neutralization

Neutralizing the circuit of Figure 5-10 balances out coupling between the input tank circuit and the output tank circuit but it does not remove all coupling from the plate circuit to the grid-to-cathode tube input. This latter coupling is degeneration or feedback, so applying a signal to the plate circuit will cause a signal to appear between grid and cathode even though the stage is neutralized. A bench test for neutralization is to apply a signal to the plate of the tube and detect the presence of a signal in the grid coil by inductively coupling to it. No signal will be present when the stage is neutralized. Of course, a signal could be inductively coupled to the input and neutralization accomplished by adjusting one branch of the neutralizing circuit bridge (C_n for example) for minimum signal on the plate circuit.

Neutralizing the cathode to grid capacity of the first stage of Figure 5-17 may be accomplished by applying a signal to the tubes cathode and asjuting the bridge balance for minimum signal on a detector inductively coupled into the input coil.

5-3-6. Tuning a Two-Stage Feedback Amplifier

Tuning the two-stage feedback amplifier of Figure 5-17 is accomplished in an unconventional way because the output circuit cannot be tuned for maximum output. This is because the output circuit must be tuned so the feedback voltage to the cathode is in phase with the input signal applied to the first grid. When it is not in phase, the resultant grid to cathode voltage increases as shown in Figure 5-18(a).

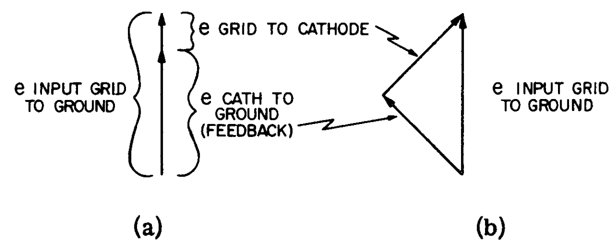


Figure 5-18. Vector Relationship of Feedback Voltage

When the output circuit is correctly tuned, the resultant grid to cathode voltage on the first tube will be at a minimum. Thus the voltage on the interstage tank circuit will be at a minimum also. Experimentally, it was found the two stage amplifier with feedback may be easily tuned by placing a voltmeter across the interstage tank circuit (hot side to ground is O.K.) and tuning the input and interstage circuits for maximum reading and tuning the output tank circuit for minimum. If the second tube is driven into grid current, the grid current may be used as the indicator. Also it has been noted that on high powered stages where operation is well into Class AB that the plate current dip of the output tube indicates resonance also.

5-4. Parasites

5-4-1. LF Parasite in Feedback Amplifier

Quite often low frequency parasites were encountered in the interstage circuit of the two-stage feedback amplifier. Oscillation occurs in the first stage due to LF feedback to its cathode. RF chokes, coupling and bypass condensers of course provide the LF resonant tank circuit. When the feedback and second stage neutralizing circuits were combined, it was necessary to use the circuit shown in Figure 5-19. This circuit has the advantage that only one

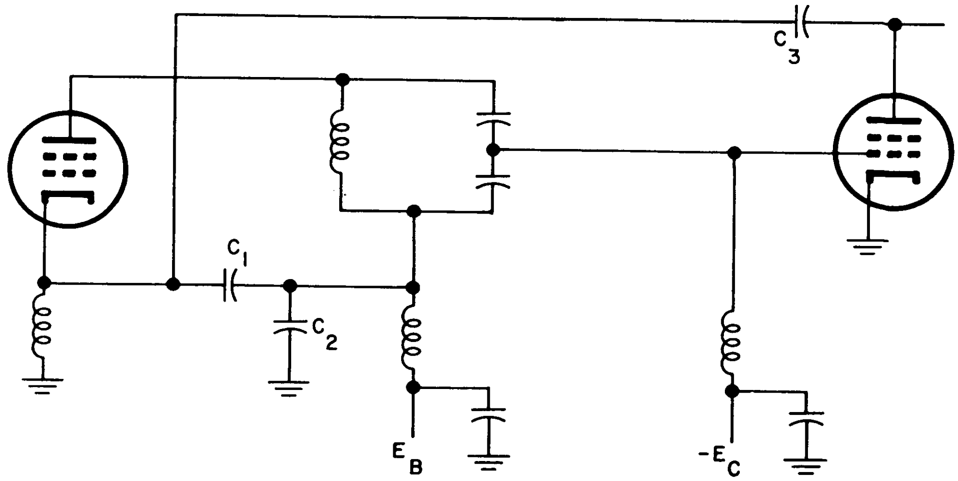


Figure 5-19. Interstage Circuit When Combining Neutralization and Feedback

capacitor, C_3, is required from the plate of the output tube thus keeping the added capacity across the output tank at a minimum. It is very convenient, however, to separate these circuits so neutralization and feedback can be adjusted independently. Also it may be desirable to be able to switch the feedback in and out.

For these reasons, the circuit shown in Figure 5-20 is used. The switch shown switches the feedback in and out. A small coupling capacitor from the top of the tank circuit to the second grid may be used if an impedance stepdown is not required.

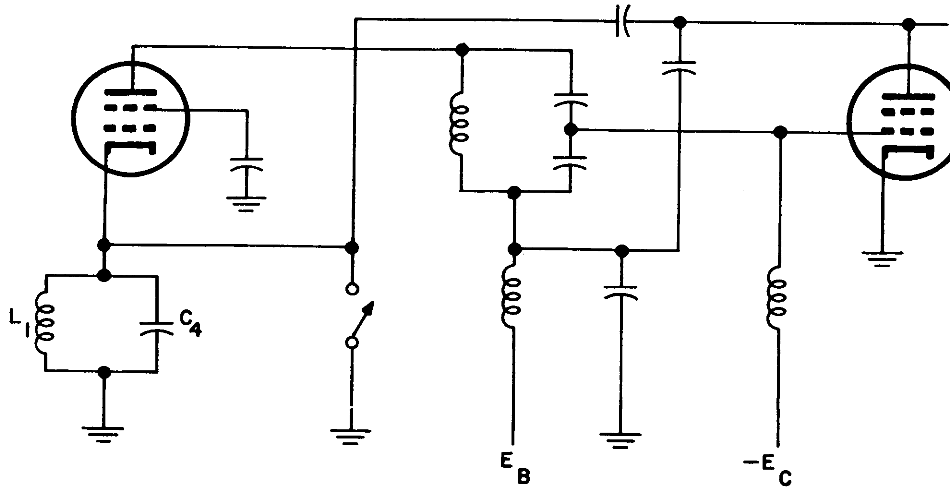


Figure 5-20. Interstage Circuit with Separate Neutralizing and Feedback Circuits

A slight tendency for a low frequency parasite still may exist with this circuit. The value of L_1 should have as little inductance as possible without upsetting the feedback. If it is too low, it "tunes" out part of the feedback capacitor, C_4, and causes the feedback to increase at low RF frequencies. In some cases, a swamping resistor may be necessary across L_1. Its value should be high compared to the reactance of C_4 to avoid a phase shift of the RF feedback.

5-4-2. VHF Parasites in Tetrode Amplifiers

VHF parasitic circuits are inherently present in HF tetrode amplifier circuits. The main parasitic resonant circuit is formed by the tube output capacity resonating with the lead inductance from the plate through the tank condenser and back to the cathode. See Figure 5-21. Usually its frequency is in the range of 50 to 200 mc.

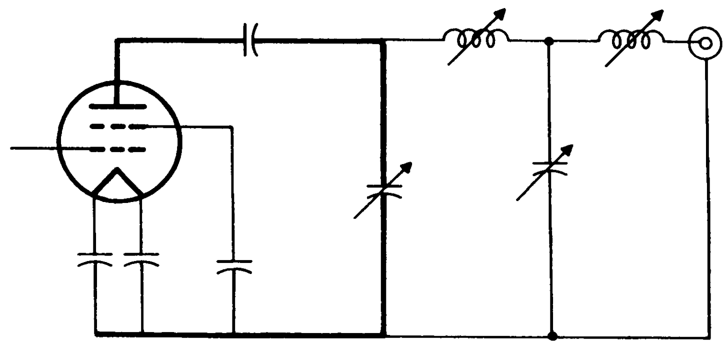


Figure 5-21. VHF Parasitic Circuit in Tetrode Amplifier

There are several ways that energy can be fed back from this plate circuit to cause oscillation. A similar parasitic circuit may exist in the grid circuit which often causes a tuned-plate-tuned grid oscillation. In this case, energy is fed back through the plate to grid tube capacitance. Oscillation cannot take place, theoretically, if the grid circuit parasitic resonant frequency is lower than the plate. Usually the grid parasitic frequency is higher because the components are smaller and lead lengths shorter. A coil in series with the grid will lower the grid parasitic frequency. Another way which has been successful is to raise it to a very high frequency. In one application, this was done by using fixed ceramic capacitors with low lead inductance in the grid tank. The lead inductance from the tube pin to the capacitor was kept very low by using 1/2 inch wide strap spaced from the chassis with 1/8 inch thick teflon insulation. This increased the minimum circuit capacitance some but not enough to be serious. In some cases, it will be necessary to damp the parasitic circuits with suppressors. A suppressor in series with the grid is effective for tuned grid-tuned plate type of parasites.

Quite often a parasite will persist when the grid is shorted to ground. In this case energy is being fed back to either the screen or the cathode. The screen and filament lead lengths through their bypass capacitors are involved along with the tube interelectrode capacitances. Best results are obtained by using low inductance bypass capacitors with as short leads from tube to ground as possible. A 4-1000A has appreciable screen and filament lead inductance and the bypassing arrangement makes a lot of difference when operating at 30 mc. Special tube sockets which made connection near the tube base and very low inductance sheet teflon capacitors made a vast improvement over the use of conventional ceramic capacitors of the Centralab 850 type. Next best were some special flat disc type capacitors as shown in Figure 5-22.

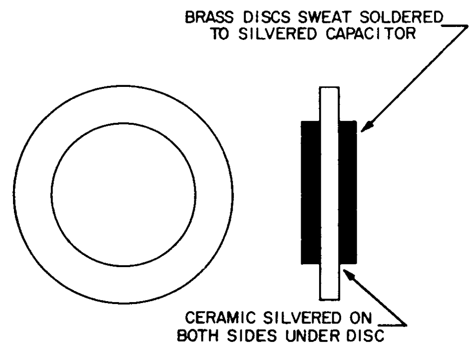


Figure 5-22. Special Low Inductance Bypass Capacitors

The screen lead inductance affects the neutralizing of a tetrode so it is desirable to keep it low for this reason also.

It is desirable to keep the frequency of parasitic resonance in the plate circuit as high as possible. When operating on a carrier frequency that is a subharmonic of the parasitic frequency, a high voltage may be developed which will result in high harmonic output at the parasitic frequency. Also, the parasitic resonance must be at least three times the highest operating frequency for parasitic suppressors to be effective without dissipating a lot of power.

Parasitic suppressors are usually required even after the above measures have been taken, but the amount of suppression required will be much less. Parasitic suppressors load the resonant circuits to such an extent that the tube gain is less than unity. Since the plate circuit is common to all of the common parasitic circuits, it is the most effective location for a parasitic suppressor.

A low value resistor shunted by a coil is an effective suppressor. Figure 5-23 shows such a suppressor in the plate circuit. The inductance across R adds in series with the original inductance L_2 to lower the parasitic frequency somewhat. R must appear across a reasonable part of the total parastic

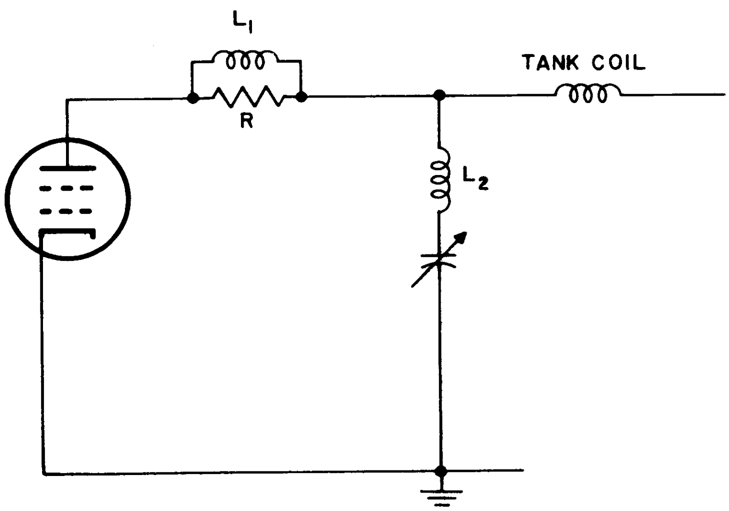


Figure 5-23. Parasitic Suppressor in the Plate Circuit

circuit inductance to be effective, however. The suppressor inductance L_1 should be low enough so that at the highest operating frequency it shunts R_1 enough to keep it from dissipating appreciable power at carrier frequency. Resistances on the order of 10 to 50 ohms are commonly used for R_1. The optimum value of R and L_1 must be found experimentally for each application.

5-4-3. Measuring Parasitic Tendency

A method of measuring the db margin below oscillation using commonly available laboratory equipment will be described. The theory and method of measurement presented is believed sound but it has not been checked in the laboratory. Basically, this method simply measures the change in Q of the parasitic resonant circuit with the amplifier turned off and then on. If the Q rises, the circuit is regenerative and the ratio of the two Q's can be used to determine the db margin below oscillation. Consider a VHF parasitic circuit as shown in Figure 5-24. The inductance from the tube plate back to the cathode resonates with the tube output capacitance at the parasitic frequency. Some loss appears in the circuit and this can be represented by an equivalent parallel resistance R_c. When all filament and DC supply voltages are applied to the tube, additional resistance

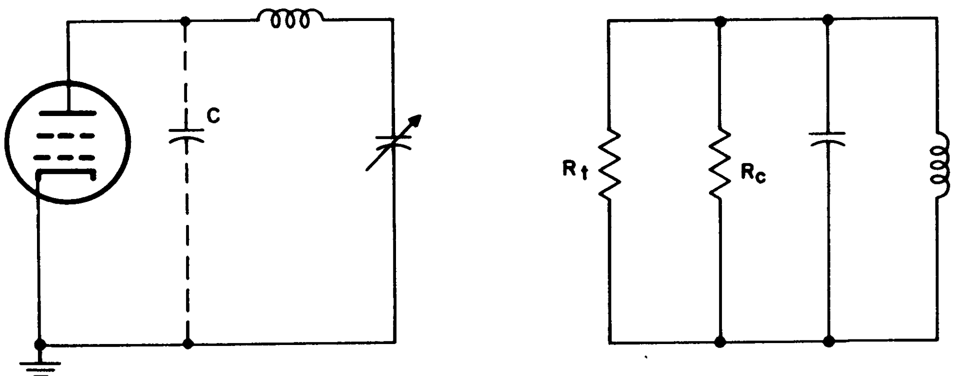


Figure 5-24. VHF Parasitic Circuits

appears across the parasitic circuit. One is the static plate resistance of the tube which is normally on the order of 100,000 ohms for tetrodes. This is very high compared to the circuit resistance so it will not be considered further. If the amplifier circuit is regenerative at the parasitic frequency, the tube will tend to supply power to the parasitic circuit and thus appear as a negative resistance R_t. Increased regeneration causes $-R_t$ to become lower and when $-R_t = R_c$, oscillation takes place. The effective Q of the circuit is increased if regeneration is present and becomes infinite at the point of oscillation. The equivalent parallel resistance of R_t and R_c determines the equivalent Q. The ratio of the equivalent Q to the original circuit Q is given by the expression.

$$\frac{Q_{eq}}{Q} = \frac{R_t}{R_t + R}$$

The sign of R_t is negative if the circuit is regenerative.

$$\text{Let } r = \frac{R_t}{R} \text{ then } \frac{Q_{eq}}{Q} = \frac{r}{r + 1}$$

This points out that a ratio of two readings can be used rather than the absolute values and this simplifies the measuring techniques.

It is noted that a similar expression can be derived using feedback equations which give the gain of the amplifier.

$$\frac{G_{\text{with feedback}}}{G_{\text{without feedback}}} = \frac{1}{1 + AB},$$

$$\text{which also equals } \frac{r}{r + 1} = \frac{1}{1 + \frac{1}{r}}$$

When r = 1 oscillation takes place so the margin below oscillation is measured from this point and may be expressed in db since only ratios are involved.

One method of obtaining the ratio r is shown in Figure 5-25. The procedure is as follows:

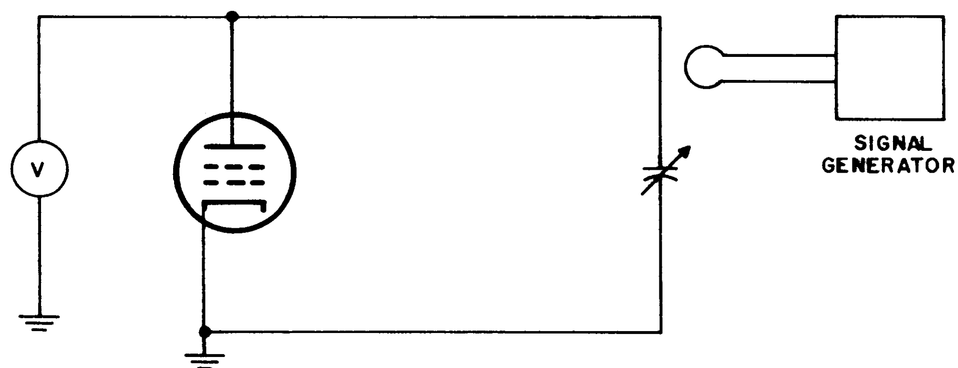


Figure 5-25. Measurement of Parasitic Tendency

1. A sensitive RF voltage detector is connected to the plate of the tube.

2. A signal generator is inductively coupled loosely into the parasitic circuit. The signal level, coupling and voltage detector sensitivity are adjusted until a scale reading of about 1/4 full scale is obtained.

3. Filament and all DC supply voltages are applied to the tube. Normal operating screen voltage should be used and the plate voltage should be lowered to avoid excessive plate dissipation. Lower grid bias until full rated plate current flows.

4. Read the meter again.

5. If the second reading is higher than the first, there is a tendency to oscillate at this frequency. Calculate the ratio of the two meter readings and find the db of margin below oscillation using the chart in Figure 5-26.

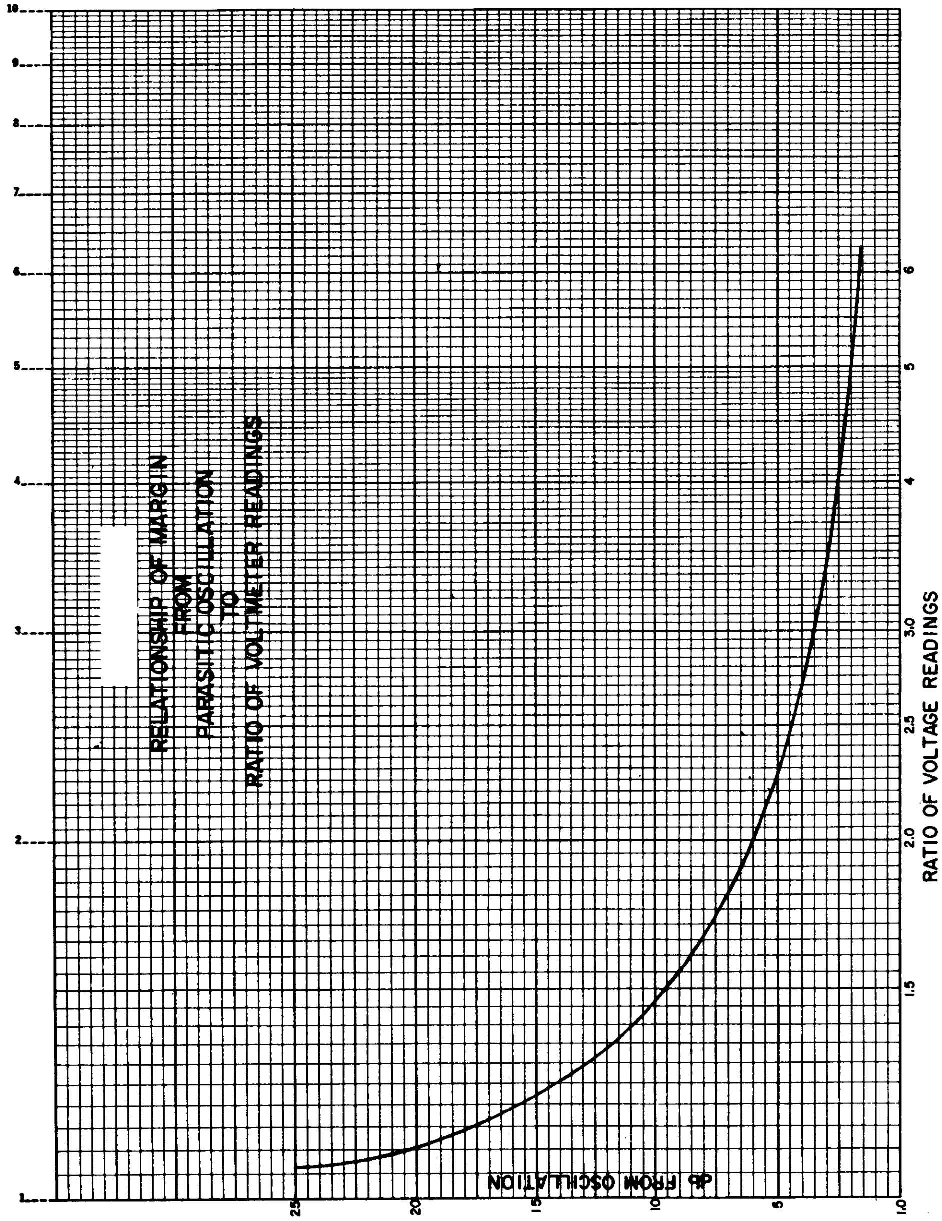


Figure 5-26. Relationship of Margin from Parasitic Oscillation to Ratio of Voltmeter Readings

This measurement only applies to the output frequency of the signal generator. The signal generator should be set at the frequency where the parasitic circuit is at resonance. This can be found by sweeping across the parasitic frequency range and watching for a peak in the voltmeter reading. A grid-dip oscillator can also be used to determine the parasitic resonance very quickly. In fact, it may be used as the signal generator.

Another method of coupling the voltage detector to the circuit is shown in Figure 5-27. Care must be used to avoid direct coupling from the signal generator to the voltage detector. A field strength meter or a receiver can be used as a very sensitive voltage detector.

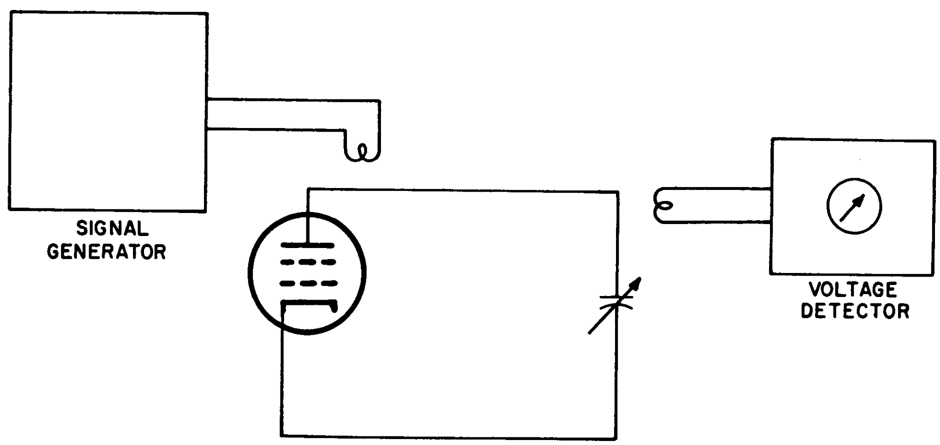


Figure 5-27. Inductive Coupling to Detector

Other variations of this basic method may be more suited to some applications. It is only necessary to avoid appreciable loading of the parasitic circuit with either the source or the detector. Direct coupling may be necessary when measuring low frequency parasitic circuits.

5-5. Some Tank Circuit Considerations

The plate tank circuit of an RF power amplifier performs four basic functions:

(a) It must maintain a sine wave RF voltage on the plate of the tube.

(b) It must provide a low impedance path from plate to cathode for the harmonic components of the plate current pulses.

(c) It must provide part or all of the necessary attenuation of harmonics and other spurious frequencies.

(d) It is required to perform part or all of the impedance matching from the tube plate to the antenna.

Additional requirements of the output network directed by company policy are that it be single ended and designed for 52 ohms nominal output impedance for feeding coaxial cable transmission lines. A direct coupled network such as the Pi-L network is most suitable for these requirements. Also, it is the most satisfactory circuit for meeting the various other requirements as discussed in the following section.

5-5-1. Circuit Q

Sufficient plate circuit Q is required to keep the RF plate voltage close to a sine wave shape. This is often referred to as the "flywheel effect". If the Q is insufficient, the RF waveform may be distorted unfavorably, resulting in lowered plate efficiency. This loss of efficiency is seldom noticed unless the Q is less than 5. It is suspected that low circuit Q will increase distortion in linear amplifiers but this limit has not been investigated. A circuit Q of 10 or more is recommended as it is known to be sufficient for this purpose.

5-5-2. Harmonic Attenuation

RF power amplifiers operating either Class AB or Class C deliver power to the tank circuit in pulses. The harmonic content of the plate current pulses is determined principally by the angle of plate current flow. The second harmonic component is on the order of 6 db below the fundamental in linear Class "B" RF amplifiers at full peak envelope power. At lower signal levels, it will be less. The higher order harmonic components drop off rapidly with increasing order but their magnitude varies greatly, depending upon the pulse shape. These harmonics must be attenuated so they are 50 db, 80 db or often still further below the fundamental. A Pi-L network will attenuate the 2nd harmonic to about 50 db below the fundamental. Higher order harmonics are attenuated more as long as the network elements maintain the low frequency values of inductance and capacitance. If more attenuation is required in special cases, external filters of either the low-pass or band rejection type may be added.

At some VHF frequency the lead inductance from the tube plate through the plate tank condenser and back to ground will resonate with the tube output capacity. Harmonics falling at this frequency may d velop appreciable voltage and power at the plate which must be attenuated in the remainder of the tank circuit. It is important that both capacitors in a Pi-L network have very low lead inductance so they will look like shunt capacitances instead of inductances at these VHF frequencies. The distributed capacity of the coils makes them look something like a transmission line in that they first go parallel resonant, then series resonant, etc., as frequency is increased. At frequencies near the "series resonance" the coils are poor from a harmonic attenuation standpoint. Little importance has been given to these considerations in older equipment designs but they should be seriously considered in new design as spurious harmonic output in the VHF region must be kept to a minimum to avoid interference with other services. Vacuum variable capacitors are very desirable in high power amplifiers as they have low series inductance. It may not be necessary to go to extremes but the tank circuit elements should be effective to about twice the highest carrier frequency. A low-pass filter in the output line can attenuate all higher frequencies.

The harmonic attenuation of an output network can be calculated by assuming the tube to be a constant current source with estimated or calculated ratios of fundamental to harmonic current. A Pi-L network has 10 to 15 db more 2nd harmonic attenuation than a simple Pi network.

Higher plate circuit Q increases the harmonic attenuation but doubling the Q only gives about a 6 db reduction of the second harmonic so Q's above 20 are seldom used for this reason below 30 mc.

5-5-3. Circuit Losses

Nearly all of the tank circuit loss occurs in the coils. It has been found that for about 50 db of second harmonic attenuation, the Pi-L network has lower losses than other networks. The losses are closely related to the ratio of circuit Q to coil Q but other factors in the design of the network enter in also. Figure 5-28 shows a Pi-L network with r_1 and r_2 representing the equivalent series resistance of the coils as determined from Coil Q and reactance. The

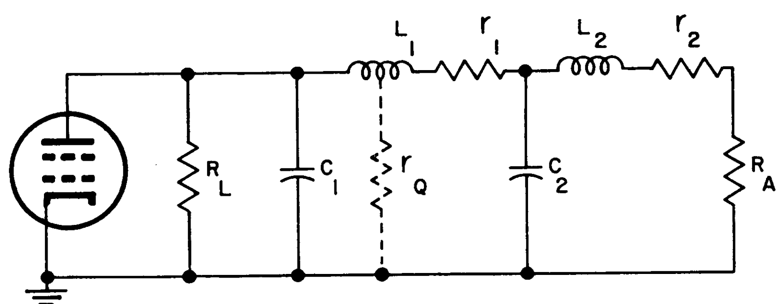


Figure 5-28. Nomenclature for Computing Coil Losses in Pi-L Network

value of r_q is the equivalent load resistance in series with L_1. It may be found from the relationship

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

The network loss is then given by the equation:

$$\text{Percent loss} = \left(\frac{r_1}{r_q + r_1} + \frac{r_2}{R_A + r_2}\right) 100.$$

R_A is the series resistive component of the load.

It should be noted that operating with a high equivalent impedance across C_2 increases the network losses. For this reason, the equivalent resistance across C_2 is chosen to have a nominal value of 300 to 500 ohms depending on other requirements. Lower values will decrease losses more but at the expense of rapidly reduced harmonic attenuation. Of course, the most effective means of reducing losses is to use coils with high values of Q in circuits with low values of operating Q.

5-5-4. Impedance Matching

The Pi-L network is ideally suited to matching a tube load to a 50 ohm transmission line. Loads with a standing wave ratio of up to 4 to 1 can be matched easily. Also, this can be done with any value of tube load impedance whereas a simple pi has difficulty matching to low load impedance when the tube plate load resistance is high.

5-5-5. Tank Coil and Condenser Requirements

The frequency range to be covered and the method of tuning are major considerations when choosing tank circuit components. The trend is toward the use of continuously variable coils and capacitors that will cover the entire frequency range required without switching of any kind. This is not practical in Autotune transmitters because of the limited torque available to drive the tuning elements and the short repositioning time allowed. In the latter application, bandswitching is almost essential. Most applications, however, require either instantaneous frequency change or the change time not of any great importance and one minute is acceptable. For instantaneous frequency changes, it is common practice to switch from one pretuned RF unit to another and manually tuned circuits are suitable for this purpose. Servo control of the tuning elements allows the incorporation of various types of automatic tuning or automatic prepositioning and is well adapted to driving continuously variable elements. Thus a practical way to design a transmitter is to use continuously variable elements designed so they can be operated either manually or by an accessory servo system.

The use of continuously variable elements has the advantages that:

(a) The circuit Q can be kept more uniform across the frequency range.

(b) The circuit losses can be kept at a minimum.

(c) The range of inductances and capacity variation of the elements can be less.

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

Vacuum variable capacitors are becoming widely used in transmitters with power levels of 1 kw and higher. They are expensive but they have a minimum to maximum capacity range that is not obtainable by any other type. Additional advantages are their small size and low series inductance, particularly in high voltage applications. Air variable capacitors should not be forgotten, however, as they can give satisfactory performance at much lower cost in some applications. They should be considered where:

(a) Plate modulation is not required.

(b) Plate voltage is not over 2500 volts.

(c) Frequency range is 5 to 1 or less.

They are easier to tune and require no turn-counting dials. Care should be used in selecting the capacitor and mounting it with relation to the amplifier tube and other components in order to control parasitic resonances and series inductance.

Variable inductance 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 as shown in Figure 5-29, to keep high voltages from developing in them. The "series" self-resonance of

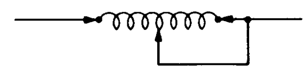


Figure 5-29. Variable Inductance Coil

the shorted out section must not be near the operating frequency as high circulating currents will develop and cause appreciable power dissipation. Usually the self-resonance must be kept higher than the highest operating frequency. If the coil must have so much inductance that this is not possible, additional shorts can be added to raise the self resonances in the unused section. Care must be used, however, to avoid a discontinuity in the inductance variation as new shorts are added.

5-5-6. Two-control Pi-L Network

It is desirable to have a minimum number of tuning controls on the output network since this simplifies the tuning procedure and minimizes the number of servo actuating units required. The Pi-L network only has four variable elements and it has been found that they can be ganged to have only a tuning and loading control as shown in Figure 5-30. Fortunately, C_2 and L_2 affect loading in the same direction so the extra capacity and inductance range of the elements required to cover various load impedances is relatively small[1]. For example, to match a 50 ohm load with a 2:1 SWR, the loading control varies about ±15% and the tuning control only about ±5%.

[1] "Pi-L Network Load Matching" report written by R. L. Uhrik, Nov. 15, 1953.

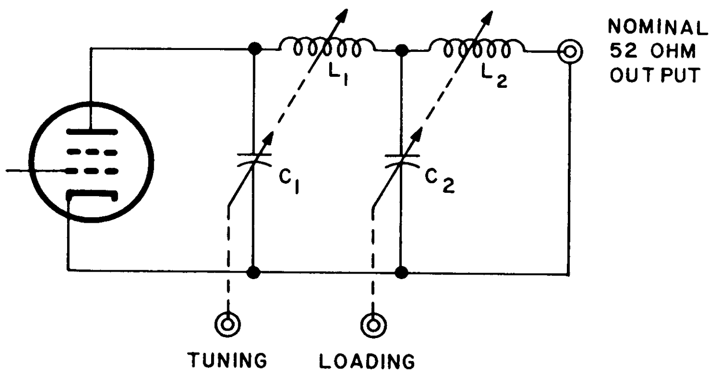


Figure 5-30. Two Control Pi-L Network

5-6. Choosing Tube Operating Conditions for Linear RF Amplifiers

5-6-1. General Considerations

SSB amplifiers must provide linear amplification and operating conditions similar to those of audio amplifiers. Although there is a close similarity between audio and linear RF amplifiers, there is one fundamental difference which is sometimes overlooked when applying audio operating conditions to an RF amplifier. The input and output voltages of a tuned RF amplifier are always sine waves because the tuned circuits with adequate Q make them so. Thus distortion in an RF amplifier results in distortion of the SSB modulation envelope (not the shape of the RF sine wave). Another way of putting it is that distortion in linear RF amplifier causes a change in gain of the amplifier when the signal level is varied. For RF service, the plate load line is a straight line on constant current curve data whereas it is a straight line on $e_p i_p$ curves for audio. Since these load lines are not quite the same, particularly in Class AB service, the power output, plate current and plate efficiency are about the same for similar audio and RF operating conditions, but not exactly the same.

The greatest difference between audio and RF is in the grid driving power requirements when driving into grid current. In audio, the driver must be capable of supplying all of the instantaneous power required by the grid at the peak of grid swing. This peak power is equal to the peak grid voltage swing mutliplied by the peak grid current. To deliver this peak power the driver must be capable of delivering an average sine wave power of one-half this peak value. The actual average driving power over an entire audio cycle may be much less than this because the duration of the grid current pulse is usually short. In RF, the tank circuit averages the power over the RF cycle due to its "flywheel" effect so the driver need only be capable of delivering the actual average power required, and not the peak. With these reservations in mind, one can get a good idea of the SSB amplifier operating conditions by examining the audio or modulator data on the tube. The following sections will discuss the choice of operating conditions and methods of calculation.

In general, all but the last two or three stages of a transmitter are operated Class A in order to confine the significant distortion generation to the power stages. The power amplifiers normally operate Class AB for higher efficiency.

5-6-2. Class "A" RF Amplifiers

In low level amplifiers, where the output signal voltage is less than 10 volts, small receiving type tubes are used such as those in receiver RF and IF amplifiers. The 6AU6 is a typical tube type since it has very low grid to plate capacitance and is very suitable for this class of service. To keep the distortion low, Class A amplifiers must operate well below their maximum output voltage capabilities. For voltage levels above 10 volts, the 4X150A is the best choice electrically because of its short leads, low plate-to-grid capacitance and high G_m. Other tubes such as the 807 can be used but their input and output circuits must be heavily loaded to avoid instability and this results in low stage gain.

Class A amplifier tubes should be operated in as linear a portion of the plate characteristic curves as is practical. This can usually be determined by inspection of a set of plate characteristic curves of the tube. Usually static plate currents which result in near maximum plate dissipation are best. Typical operating conditions for the tube in either RF or AF amplifier service may be used as a guide but the maximum output voltage must be reduced to something like one-tenth the DC plate voltage or even less if signal to distortion ratios of 50 db or better are to be realized from the stage. The DC supply voltage regulation for this class of operation is seldom of importance and cathode bias and screen dropping resistors are commonly used.

Even with tubes such as the 6AU6 and 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.

5-6-3. Class AB_1 RF Amplifiers

Class AB_1 operation is normally used over the power range of 2 watts to 500 watts output. This class of operation is very desirable since distortion due to grid current loading is avoided and also because very high power gain can be achieved. At present, tubes are not available which will give low distortion with good plate efficiency operating AB_1 at power levels above 500 watts. For 2 watts output, the 6CL6 or 6AG7 tubes appear best. They are fairly linear since they were designed for video amplifier service. They have high G_m and a low grid swing requirement. The 4X150A is suitable for any power level up to 150 watts and perhaps even 300 watts per tube at higher voltage SSB ratings which may be forthcoming. The 6146 is suitable for Class AB_1 operation but it does not have the excellent screen shielding of the 4X150A and probably has substantially more distortion. The only other tubes that appear practical for Class AB_1 service are the 4-400A and 4X500A which can deliver up to 350 watts per tube. Each of these tubes must be neutralized for operation at 30 mc not only for circuit stability but to keep the distortion generation at a minimum.

5-6-4. Estimating Power Output of Class "AB" RF Amplifiers

The power output of a tube to be operated Class AB as a linear power amplifier may be quickly and quite accurately estimated as follows. Choose a load line or plate swing end point on a set of tube curves. From the peak plate voltage swing and peak plate current thus established, use the following formula where P_o is peak envelope power output.

$$P_o = \frac{i_p e_p}{4}$$

This formula is accurate if the tube operation is perfectly linear and will be within 10% even with quite bad distortion.

A general note or two regarding the selection of the end point of plate swing follows although more detailed information will be found in later sections. For best efficiency a large plate swing should be used. It is noted that constant current plate current curves depart from a straight line at low values of plate voltage. For low distortion, the plate swing should be limited to where the plate current deviates about 10% from a straight line. For Class AB_1 operation, the end point will also be at zero grid voltage for maximum power output. When operating Class AB_2, the end point of plate swing should be chosen so that the peak grid current is not over 10% of the peak plate current. These limits are very approximate and should be used only as a rough guide. It should also be noted that the above formula for power output is independent of static plate current.

5-6-5. Static Plate Current

For operating conditions with a given screen voltage, plate voltage and plate load, there is one value of static plate current that will give minimum distortion.

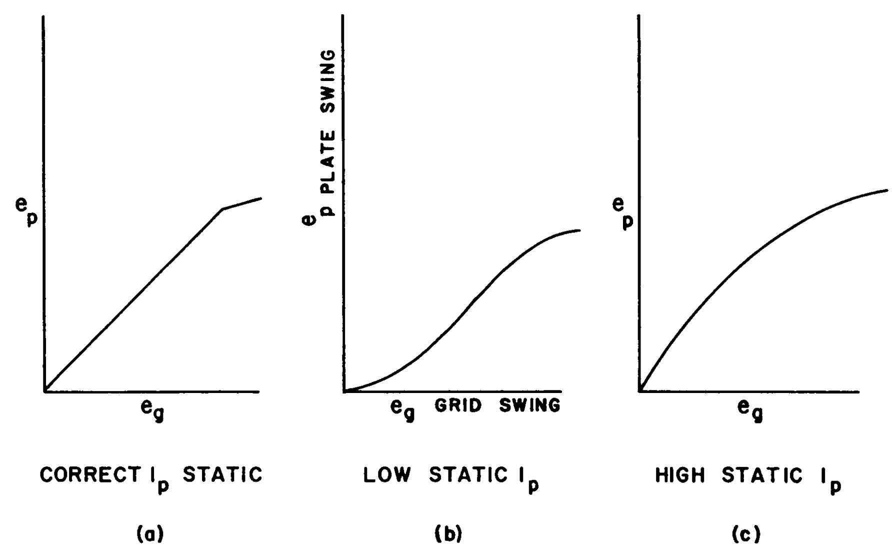


Figure 5-31. Effect of Static Plate Current on Linearity

Figure 5-31 shows the effect of static plate current on linearity. The grid bias determines the static plate current when the plate and screen voltages are given. The optimum value of static plate current for minimum distortion is determined by the sharpness of cutoff of the plate current characteristic. This point can be found analytically by choosing a load line on a set of constant current curves for the tube and plotting a curve of plate current vs. grid voltage from points along this line. Such a curve is shown in Figure 5-32. Draw a straight line along the most linear part of the curve and extend it to

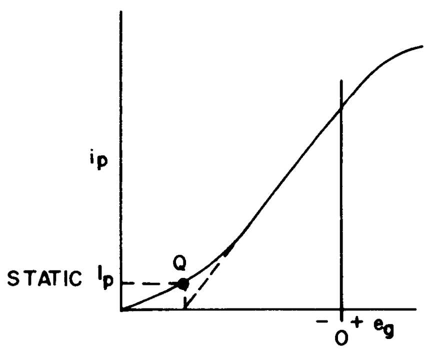


Figure 5-32. $e_g i_p$ Curve on Plate Load Line

zero plate current. This is often referred to as projected cutoff and the grid voltage at this point is the proper value of grid bias voltage, E_c. The plate current which will flow at this bias is the proper static plate current. This is the procedure used for audio amplifiers and is very nearly correct for tuned RF also. Perhaps a more correct method for tuned RF amplifiers is to choose the point of static plate current on the curve at point Q such that the slope of the curve at this point is one-half the slope of the major linear portion. The reason for this is readily seen by comparing operation with a very small signal to that of a large signal. With a small signal, the tube operates Class A and the tube delivers power over both halves of the cycle. With a large signal, the tube delivers power over essentially one-half the cycle. Thus the change in plate current relative to plate voltage swing over one-half the cycle, must be half as much for small signals as for large signals. Still another way of looking at it is that the current from the left of point Q should just match the amount of departure from linearity from the right of point Q.

The screen voltage applied to a tetrode tube has a very pronounced effect on the optimum static plate current. The plate current of a tube varies approximately as the three-halves power of the screen voltage. Thus for example, raising the screen voltage from 300 to 500 volts will double the plate current. The shape of the curve of Figure 5-32 will stay nearly the same, however, so the result is that the optimum static plate current for minimum distortion is also doubled. A practical limit is soon reached because high static plate current causes excessive static plate dissipation.

In practice, it has been found that operation at the exact point as determined above is not necessary and somewhat lower static plate current can be used before the distortion rises appreciably. RF feedback around an amplifier will reduce distortion from this cause and values of plate current about one-half of optimum may be used although it depends on how much distortion can be accepted.

5-6-6. Calculation of Tube Operating Conditions (Single Tone)

The operating conditions for a power amplifier can be calculated quite accurately by selecting a load line and using the 11-point analysis originated by Chaffee.[1] Sarbacher devised a mechanical aid for the Chaffee method and published an article on its use.[2] More recently, Eimac made available another aid and a well written procedure of its use.[3] Both of the latter references were written for Class C amplifiers, but a slight extension allows their use on Class AB operation also.

Basically, this method assumes sine wave voltages 180 degrees out of phase on the grid and plate. Points along the load line at each 15° over the cycle are found with the aid of the mechanical devices which simply have lines spaced according to the sine wave function. The currents read at these points along the load line are put into the equations given to determine the DC component of current and the fundamental AC component. Knowing these values of the plate, grid and screen currents all values of input, output and dissipation are readily computed.

The method of calcuation will now be given in more detail. Calculations are made on a selected load line and if it turns out that the desired operating conditions are not realized, it is necessary to try other load lines until either the desired or optimum conditions are found.

Figure 5-33 shows a load line drawn on a set of constant current curves. Below is a sine wave representing the RF plate voltage. Projecting upward from the 15° points on the sine wave establishes points A, B, C, etc. on the load line. Each of these points except point A, is used to represent the average plate current over a 15° interval extending 7-1/2° each side of the point. Since only half of the cycle is analyzed, point A represents only half a 15° interval or 7-1/2°. The average DC plate current is found by simply averaging these readings. This is done by using the equation:

$$I_{AV} = \frac{1}{12}\left(\frac{A}{2} + B + C + D + E + F + Q + F' + E' + D' + C' + B' + \frac{A'}{2}\right)$$

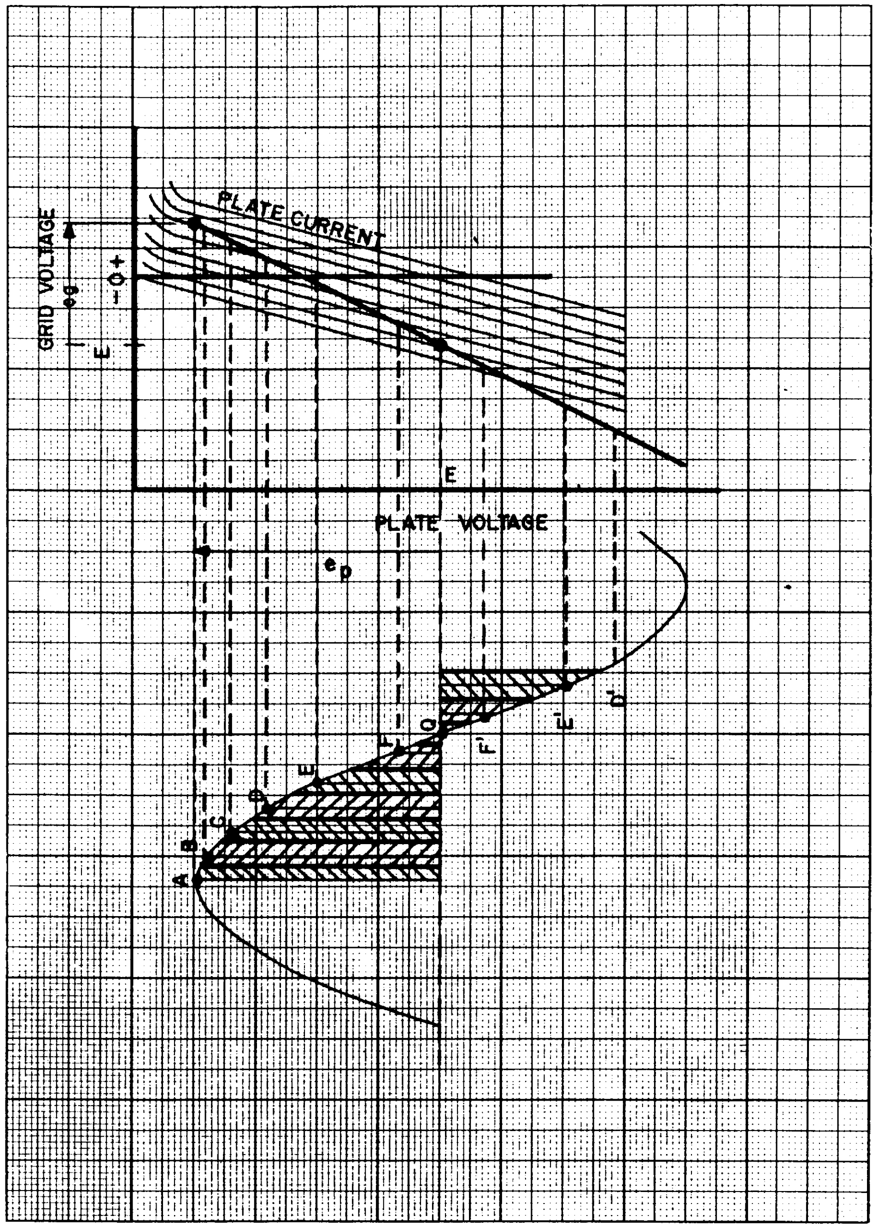


Figure 5-33. Principle of Graphical Analysis

The 1/12 term comes from using only one-half the cycle since it is symmetrical and there are twelve 15° intervals over one-half a cycle. This equation includes all the terms for Class A operation but normally for Class AB operation, those beyond F' or E' are zero. The references cited[2,3] are only concerned with Class C operation and they have dropped the Q and all following terms since they are always zero in Class C operation.

The fundamental AC component of current is obtained by using the equation:

$$I_1 = \frac{1}{12}\Big[(A-A') + 1.93(B-B') + 1.73(C-C') + 1.41(D-D') + (E-E') + .52(F-F')\Big]$$

It is noted that the coefficients of the various terms are twice the cosine of the angle except the (A-A') term where the coefficient is one-half twice the cosine of 0° or unity.

If the 2nd, 3rd or 4th harmonic components of current are required, the following equations are used:

$$I_2 = \frac{1}{12}\left[(A+A' + C+C' - E-E') + 1.93(B+B' - F-F') - Q\right]$$

$$I_3 = \frac{1}{12}\left[(A-A') + 1.41(B+D'+F'-B'-D-F) - 2(E-E')\right]$$

$$I_4 = \frac{1}{12}\left[(A+B+F+F'+B'+A'-C-E-E'-C') + 2(Q-D-D')\right]$$

These equations for harmonic current are only approximate and apply only to tetrode and pentode tubes.

This method of computing the DC and AC components applies to grid and screen current also. All values of DC and fundamental AC components of the plate, grid and screen, if any, should be calculated except the AC component of screen current is only of interest in grounded grid tetrode amplifiers. The AC components calculated using the above equations are peak values and not RMS values.

The following equations are used to calculate the various operating conditions:

$I_p(av)$ = DC plate current

$P_{in} = E_B I_p(av)$ watts input

$P.O. = \frac{I_{1p} e_p}{2}$ watts output

$Eff = \frac{P_0}{P_{in}} 100$ percent plate efficiency

$P_{pd} = P_{in} - P_0$ = plate dissipation

$I_g(av)$ = DC grid current

$P_{dr} = \frac{I_{1g} e_g}{2}$ = watts grid driving power

$P_c = I_g(av)\, E_c$ = drive consumed by bias supply

$P_{gd} = P_{dr} - P_c$ = watts grid dissipation

$I_s(av)$ = DC screen current

$P_{sd} = I_s(av)\, E_s$ = Screen dissipation

$R_L = \frac{e_p}{I_{1g}}$ = RF plate load resistance

The above calculations are for conventional grid driven circuits. Of course, when using triodes, the screen current is absent but all other calculations are the same for triodes and tetrodes.

5-6-7. Calculation of Cathode Driven Amplifiers

5-6-7-1. Using Constant Current Characteristics for Cathode-Driven Operation

Cathode driven (or grounded grid) amplifiers operate a little differently than conventional grid driven amplifiers. This discussion will cover tetrode tubes but the considerations for triodes are identical except that operations relating to the screen are omitted.

The calculations are minimized if a set of constant current curves are available for this type of operation which refers all voltages to the No. 1 grid instead of to the cathode. These curves can often be obtained from the tube manufacturer if they do not appear on the regular data sheet. The method of using conventional curves for cathode driven service will be discussed in the following section.

To calculate the operating conditions, draw a selected load line on the curves for cathode driven service. Remember that all voltages are referred to No. 1 grid but the actual DC supply voltages are normally measured to the cathode. In practice, the cathode and the negative side of the DC plate and screen supplies are at DC ground. The point, Q, on the load line must be at a plate-to-G_1 voltage on the curve that is the sum of the DC plate and DC grid bias voltages. For example, 3000 volts of E_B and 100 volts of E_C gives 3100 volts of plate-to-No. 1 grid voltage on the curves. The other coordinate of Q is at a cathode-to-grid No. 1 voltage equal to the grid bias voltage but of opposite sign. The plate current pulses occur when the cathode is driven in the negative direction. A typical load line on a set of curves is shown in Figure 5-34. The DC and fundamental AC components are calculated using the method explained in the previous section. The operating conditions are then calculated as follows.

DC Plate Input

$$P_{in} = E_B I_p(av)$$

Note that E_B is the DC plate voltage from plate to cathode and not plate-to-G_1.

RF Power Output

$$P_O = \frac{E_p I_{IP}}{2} \text{ watts.}$$

Fed-Thru Power

$$P_{ft} = \frac{e_k I_{IP}}{2} \text{ watts}$$

Plate Dissipation

$$P_d = P_{in} - (P_O - P_{ft})$$

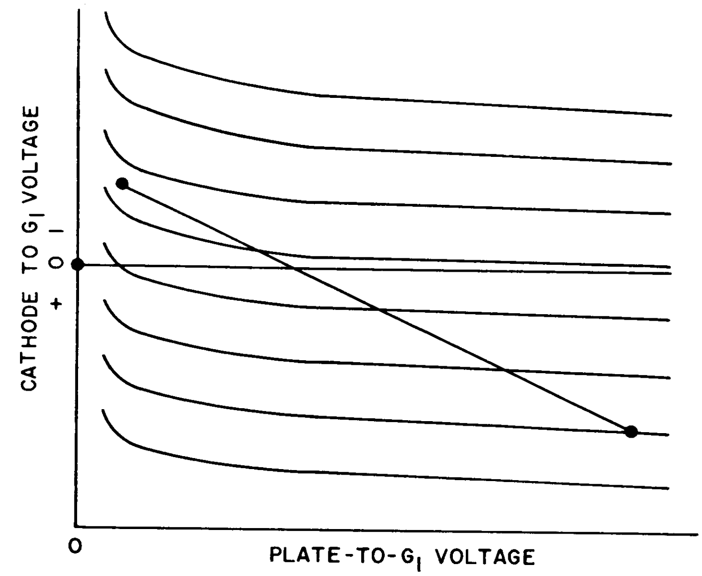


Figure 5-34. Load Line on Constant Current Curves for Cathode Drive Operation

Driving Power

This is composed of three components; the fed-thru power, the grid driving power and the screen grid driving power.

$$P_{dr} = P_{ft} + P_c + P_s$$

$$= \frac{e_k I_{1P}}{2} + \frac{e_k I_{1g}}{2} + \frac{e_k I_{1s}}{2} \quad \frac{e_k I_p}{2} \text{ etc}$$

Grid Dissipation

$$P_{gd} = \frac{I_{1g} e_k}{2} - I_g(av)E_C$$

Screen Dissipation

$$P_{sd} = \frac{I_{1s} e_k}{2} - I_g\,(av)E_s$$

Input Impedance

$$R_{in} = \frac{e_k}{I_{1p} + I_{1g} + I_{1s}}$$

This varies a little with signal level in a Class AB linear amplifier because the grid and screen currents do not vary linearly. At very low signal levels, the grid current is nearly always zero and in some cases the screen current may be zero also. In most SSB applications the cathode input impedance is very nearly that due to the fedthru power alone.

Other unknowns of interest may be calculated as discussed in the previous section.

5-6-7-2. Using Conventional Constant Current Curves for Grid Driven Operation

Triodes

These curves can be readily used for triode cathode driven amplifiers. A load line is selected as for grid driven service and all calculations are made. The fed-thru power must be added to the power output of the tube to obtain the total power output. Other than this, all the appropriate previous equations apply.

Tetrodes

Accurate calculations cannot be made easily using these curves because the screen to cathode voltage varies over the cycle and the curves only represent the condition when the screen to cathode voltage is fixed. Reasonably good estimates of power output and fed-thru power can be made if all pertinent considerations are kept in mind, however.

5-6-8. Different Screen Voltages

Very often it is desirable to operate tetrodes with screen voltages other than those for which the curves are available. Since the currents vary approximately as the three-halves power of the screen voltage, it is not too difficult to assign new values to the grid and plate voltage coordinates and to the constant current lines. For example, if it is desired to operate with a screen voltage of 500 volts and the curves are for 300 volts, all voltage values on the curves must be multiplied by the voltage factor of $\frac{500}{300}$ or 1.67. The current values are all multiplied by the three-halves power of the voltage factor of $(1.67)^{3/2} = 2.15$ which is the current factor. Using these new values on the curves, the operating conditions can be calculated as in the previous section.

5-6-9. Two-Tone Operating Conditions

An input signal of two tones of equal amplitude is used for making performance measurements on SSB transmitters and power amplifiers. A two-tone signal with an amplitude that will drive the amplifier to its rated peak envelope power is also considered as representative average operating conditions. The SSB envelope is the shape of half sine waves as shown in Figure 1-3(a). In some applications, tests are required with a pilot carrier 20 db below the level of either tone. This causes small sine wave ripples on the SSB envelope. Due to the complexity introduced by a carrier or third tone, only the two-tone condition will be discussed.

5-6-9-1. Theoretical Two-Tone Operating Condition

In order to show the nature of operating conditions using a two-tone signal a theoretical condition using an "ideal" tube will be presented. The "ideal" tube will be considered as having linear plate current variation with plate voltage from plate current cutoff down to zero plate voltage. Figure 5-35 shows a load line on a set of "ideal" tube curves. The theoretical power output and plate efficiency given by the following equations.

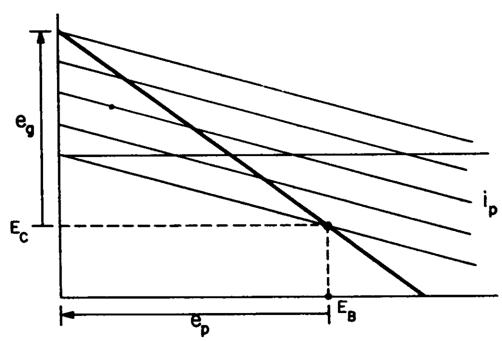


Figure 5-35. Load Line on "Ideal" Class "B" Tube Curve

Single Tone

$$I_B = \frac{i_p}{\pi} \quad \text{DC plate current}$$

$$P_{in} = \frac{i_p E_B}{\pi} \quad \text{watts plate input}$$

$$P_o = \frac{i_p e_p}{4} \quad \text{watts power output}$$

$$P_d = i_p \left(\frac{E_B}{\pi} - \frac{e_p}{4} \right)$$

$$\text{Eff} = \frac{\pi}{4} \left(\frac{e_p}{E_B} \right)$$

where i_p is peak plate current and e_p is peak plate voltage swing. It is noted that maximum efficiency is $\frac{\pi}{4}$ or 78.5%.

With a two-tone signal, the currents and voltages are averaged over the SSB envelope, which is the shape of half sine waves, giving the following equations:

Two Tone

$$I_B = \frac{2}{\pi^2} i_p$$

$$P_{in} = \frac{2}{\pi^2} i_p E_B$$

$$P_o = \frac{i_p e_p}{8}$$

$$P_d = i_p \left[\frac{2E_B}{\pi^2} - \frac{E_p}{8} \right]$$

$$\text{Eff} = \left(\frac{\pi}{4} \right)^2 \frac{e_p}{E_B}$$

It is noted that DC plate current, power input and efficiency vary linearly with signal level. The power output varies as the square of the signal level since i_p is proportional to e_p. The maximum theoretical plate efficiency for a two-tone signal is $\left(\frac{\pi}{4}\right)^2$ or 61.7%.

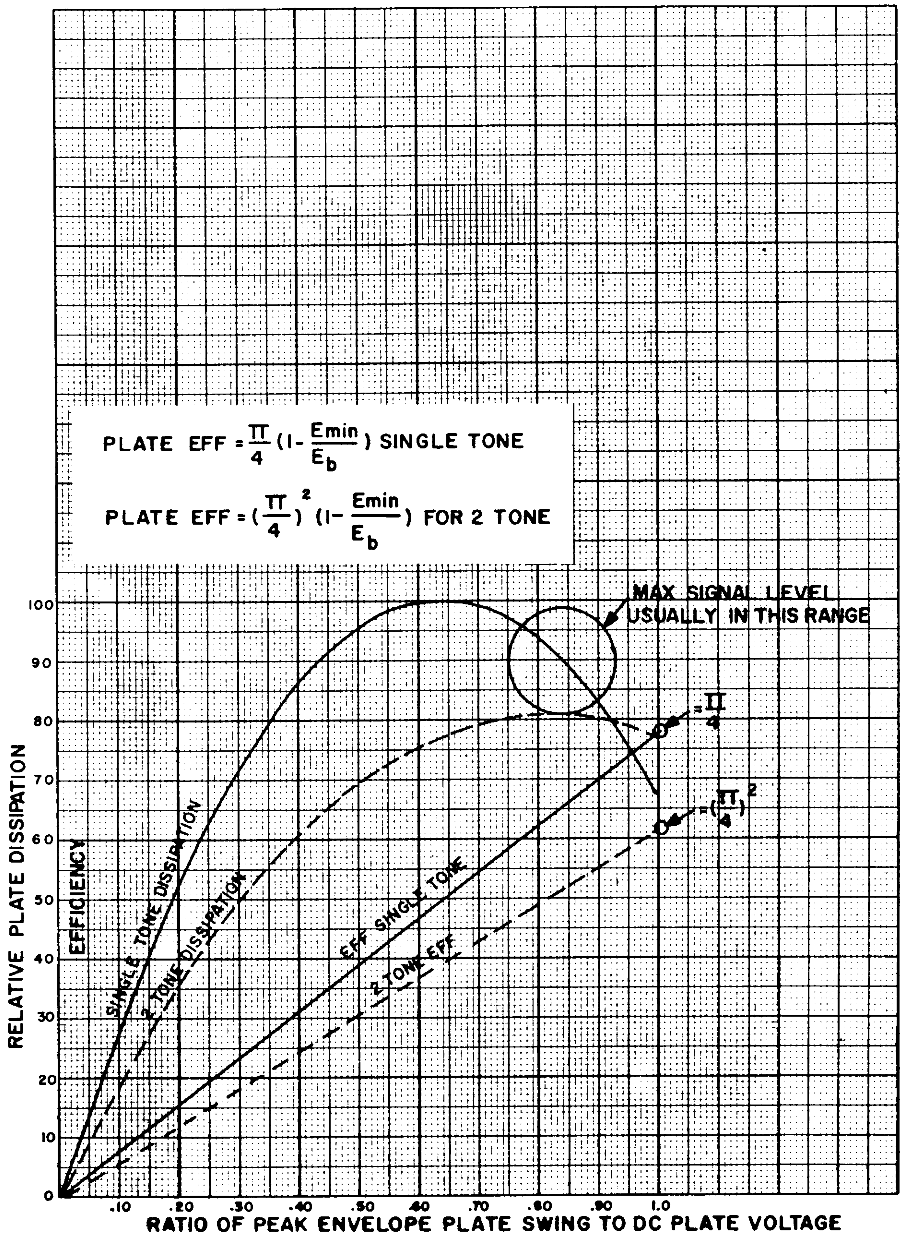


Figure 5-36. Plate Dissipation Versus Plate Voltage Swing of Theoretical Class B Tube

A plot of plate dissipation versus signal level is shown in Figure 5-36 for both the single tone and the two-tone cases. In the single tone case, the dissipation has a maximum when the peak plate swing is $\frac{2}{\pi} E_B$ of $.617E_B$. In the two-tone case, the maximum P_d occurs when e_p at the SSB envelope crest is $\frac{8}{\pi^2} E_B$ or $.81E_B$.

5-6-9-2. Actual Two-Tone Operating Conditions

Practical conditions differ from the simple theoretical case given in the previous section because tubes are not linear and require a certain amount of static plate current for low distortion. The static plate current causes static plate dissipation and the maximum plate dissipation is increased to some extent. Figures 5-39, 5-40 and 5-41 show some curves that indicate how various amounts of static plate dissipation affect the plate dissipation.

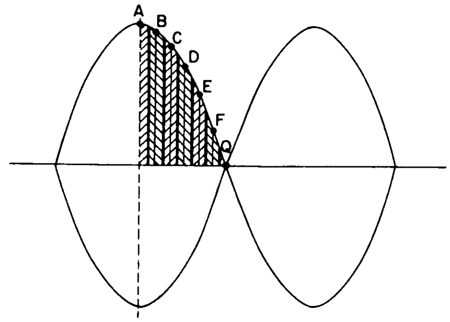


Figure 5-37. Graphical Two-Tone Analysis

5-6-9-3. Calculating Two-Tone Operating Conditions

This is accomplished by making single tone calculations at "15°" points along the SSB envelope. See Figure 5-37. Point "A" represents the peak envelope power point. The peak plate swing to use for calculating the conditions at points B, C, D, E and F is the peak plate swing of "A" multiplied by the cosine of 15, 30, 45, 60 and 75 degrees respectively.

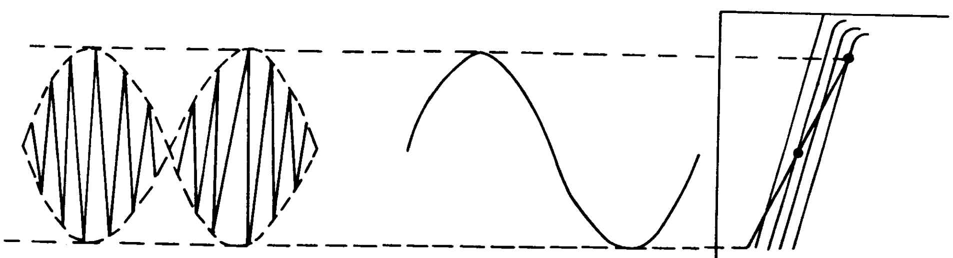


Figure 5-38. SSB Envelope and RF Sine Wave Relation

The calculating devices which locate the points for single tone calculations can be used to locate the peak plate swing for each of these calculations also. The operating conditions of interest must be calculated for each of these points plus the static condition for the seventh. The average value of all these conditions is calculated by using the following equation:

$$\text{Average two-tone value} = \frac{1}{12}\left(\frac{A}{2} + B + C + D + E + F + Q\right)$$

Plate current, plate dissipation or any other variable of interest may thus be calculated for the two-tone condition.

If the two-tone operating conditions are desired at another peak envelope power level, it is necessary to repeat the entire procedure. If a curve of plate dissipation versus signal level is desired as shown in Figure 5-36, it is suggested that the peak plate swing be chosen at the even multiples of 10% E_B and plotted in this manner. This makes it possible to compare the characteristics of various tubes more easily.

A computer would be a great time saver if a great number of two-tone calculations had to be made comparing various tube types or operating conditions. The writer believes that a fairly simple analog type computer could be designed to perform the above calculations.

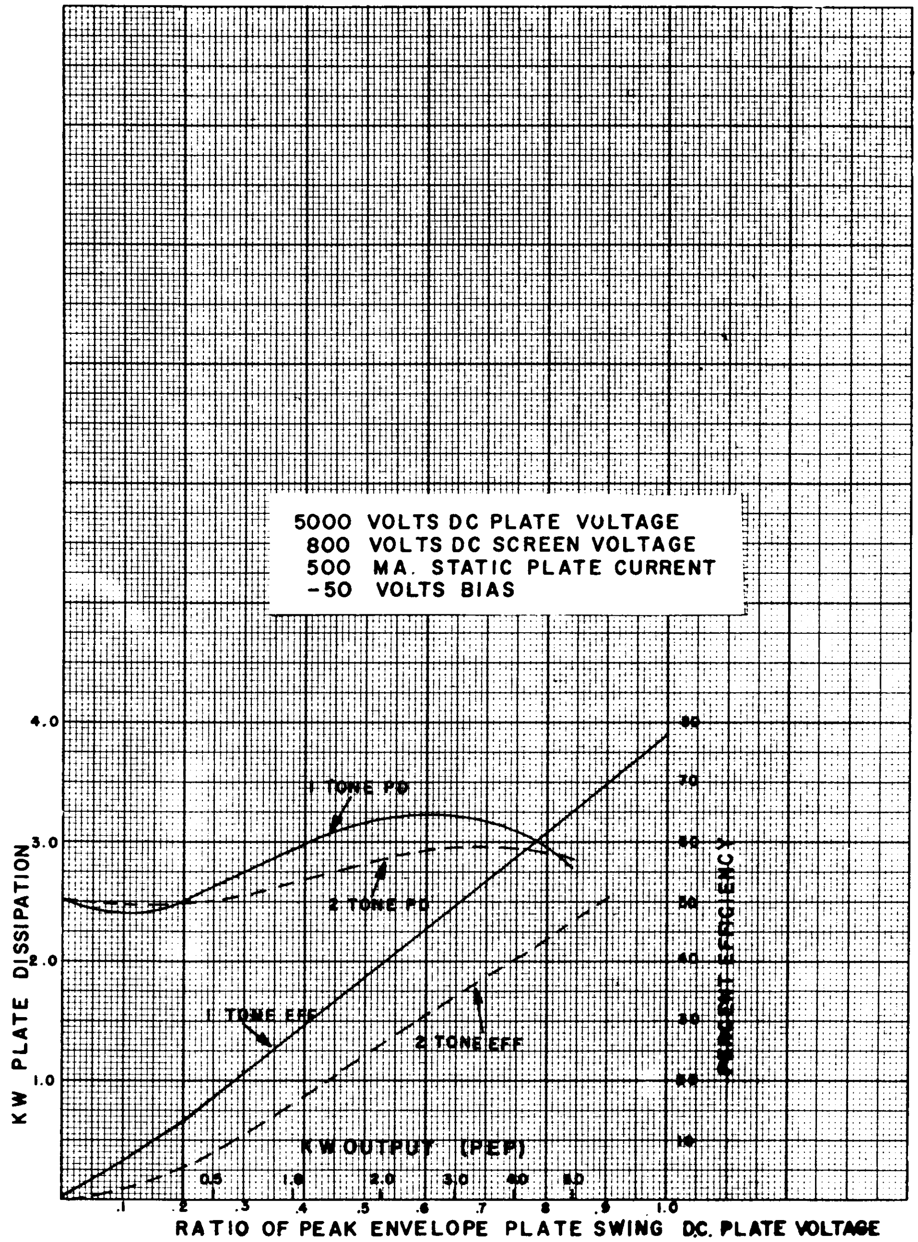


Figure 5-39. Calculated Plate Dissipation and Efficiency of 6166 SSB Linear Amplifier Versus Signal Level

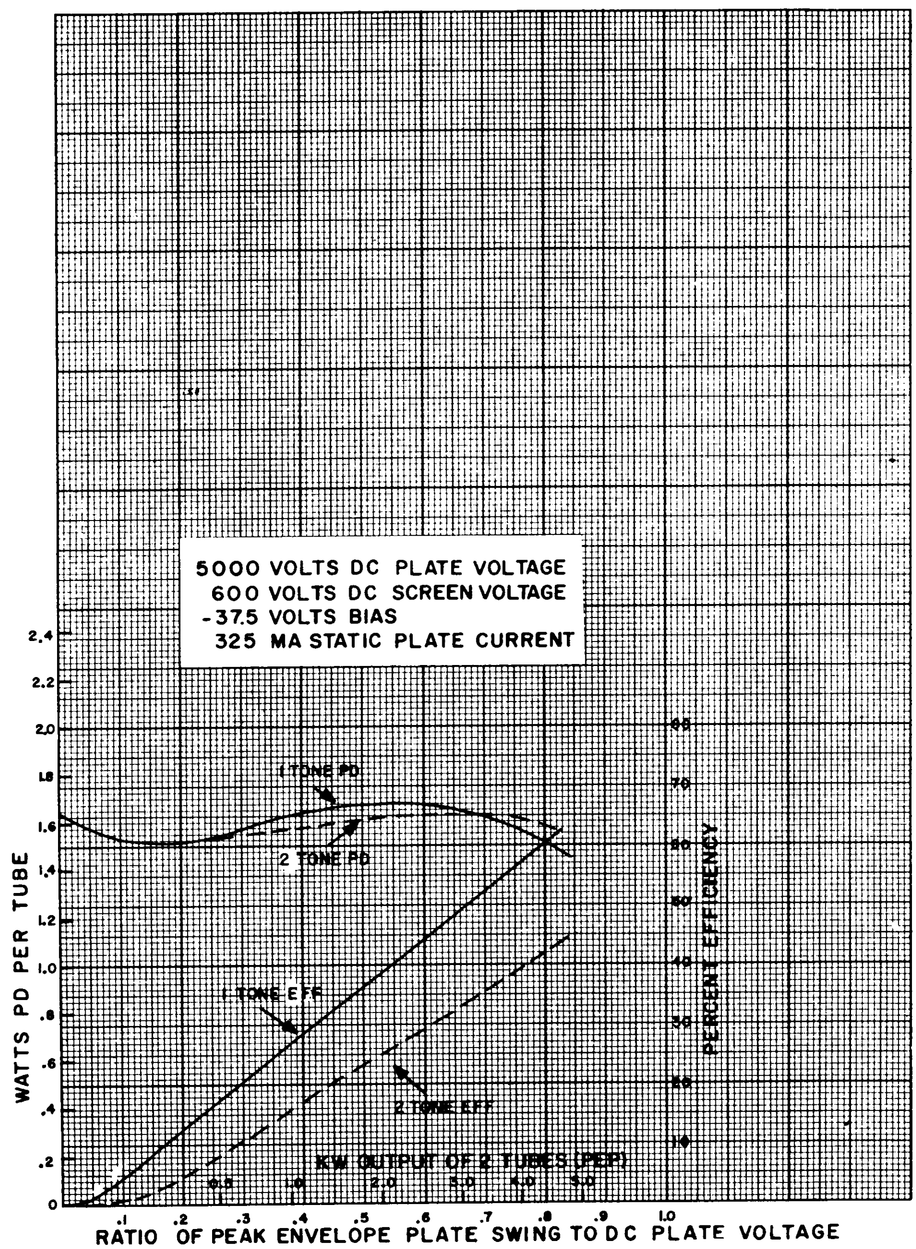


Figure 5-40. Calculated Plate Dissipation and Efficiency of AX9907R SSB Linear Amplifier Versus Signal Level

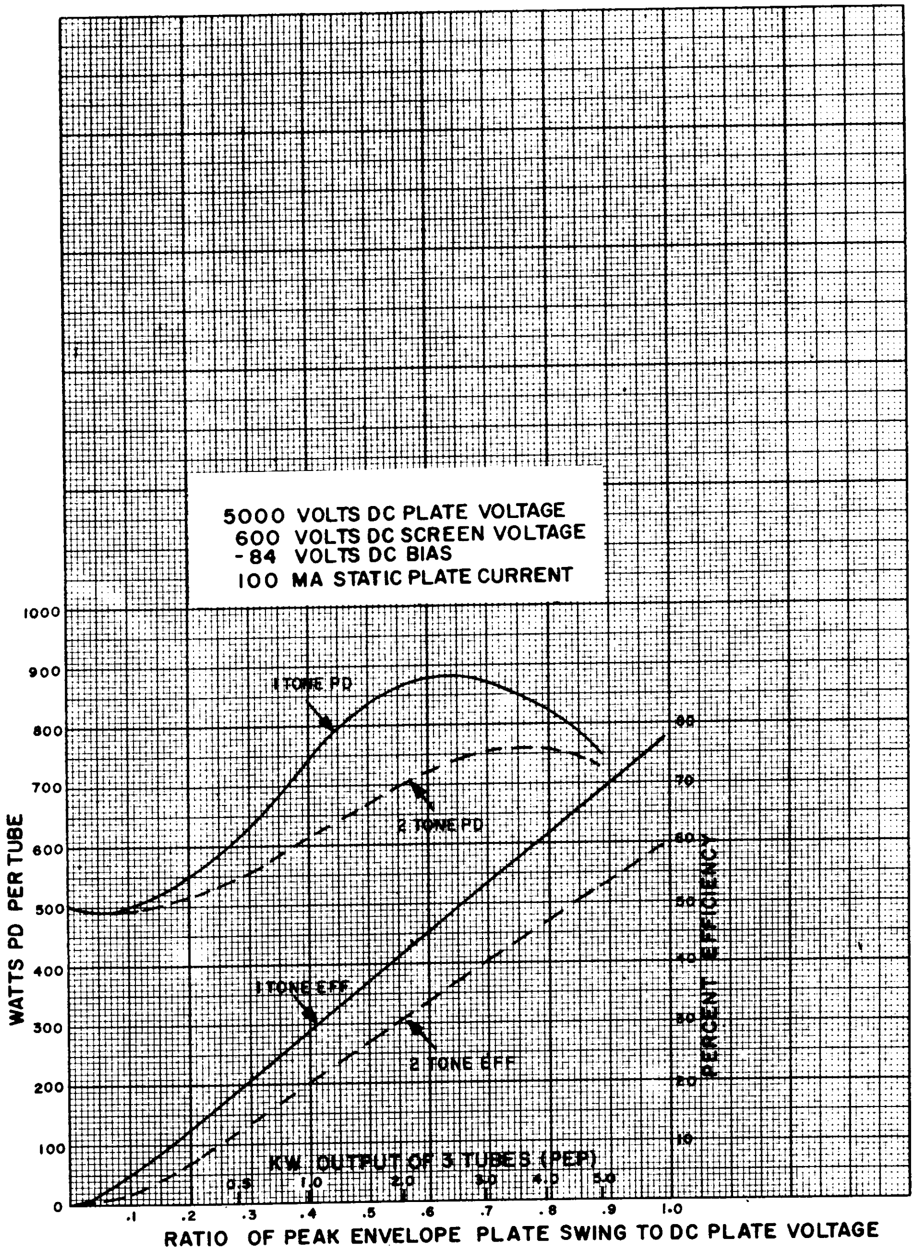


Figure 5-41. Calculated Plate Dissipation and Efficiency for 4-1000A SSB Linear Amplifier Versus Signal Level

5-7. Distortion

The nature of distortion generated in SSB power amplifiers is discussed in Sections 2. 3, 2. 4 and 2. 5. The requirements of the application will determine the signal-to-distortion ratio requirements of the power amplifier.[1] The reference cited discusses the relationship between the tone-to-distortion ratio of the amplifier and the speech-to-distortion noise ratio in adjacent voice channels on a multichannel equipment. In general, with speech modulation the noise or "splatter" appearing in an adjacent channel is appreciably less than the amplitude of the 3rd order product as measured in a two-tone test. How much less apparently depends on many factors but it seems that the adjacent channel noise is usually at least 6 db below or lower than the measured S/D ratio.

5-7-1. Causes of Distortion and Methods of Reduction

The principle causes of distortion are non-linear characteristics of the amplifier tubes and grid current loading. In order to confine the distortion generation substantially to the last stage or two, all other stages are usually operated Class A. The plate current curve of Class A amplifier tubes in general can be represented by a simple exponential curve as shown in Figure 5-4-2(a).

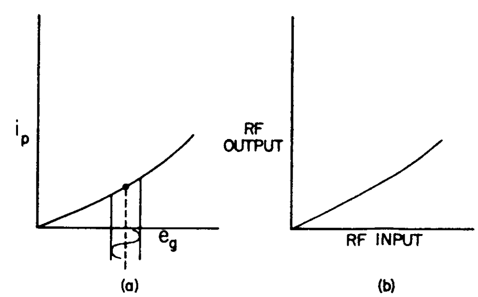


Figure 5-42. Effect of Nonlinear Plate Characteristics

The distortion is kept low by operating the tube in the most linear portion of its plate current characteristic and by keeping the signal level lcw. Figure 5-42(b) shows the nature of the linearity curve of a typical Class A amplifier. The curvature is gr atly exaggerated since for S/D ratios on the order of 50 db, it cannot be detected visibly.

Class AB amplifiers usually have a very similar curvature. When the curvature of the linearity characteristics are similar for a series of cascaded amplifiers, the distortion products generated add together in phase.

When amplifier tubes are driven into the grid current region, the resulting grid circuit loading causes the linearity curve to drop at large signal levels as shown in Figure 5-43. The distortion products from this type of curvature are 180° out of phase with those previously discussed. When both types of curvature exist, the distortion products tend

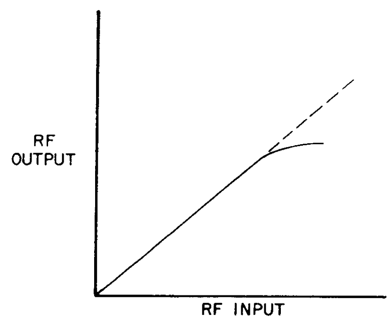


Figure 5-43. Effect of Grid Loading on Linearity

to cancel as shown in Figure 5-44. When this happens, the 5th order product is usually stronger than the resulting 3rd in the region of cancellation. For this reason, the value of distortion cancellation is not as great as it seems.

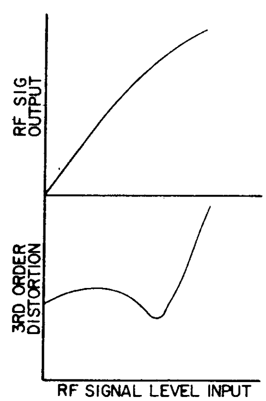


Figure 5-44. Distortion Cancellation

The nonlinearity caused by grid current loading is a function of the regulation of the grid driving source. The regulation of linear amplifiers with a varying load is poor in general. It is common practice to use a swamping resistor in parallel with the varying grid load. It is usually necessary to absorb about 10 times as much power in this swamping resistor as the grid consumes in order to obtain satisfactory regulation. Another way of providing a low driving impedance is to use a very high resistance tube such as a tetrode or pentode and an impedance inverting network.[1] The impedance inverting network can be a quarter wave or 90° network coupling the driver plate and PA grid tank circuits. Inductively coupled tank circuits also have this property. Figure 5-45 shows these two circuits. The disadvantage is that it is difficult to maintain proper coupling without special adjustment and it is seldom used in general frequency coverage transmitters for this reason.

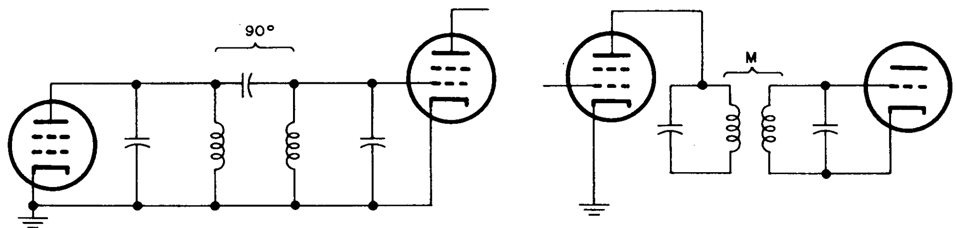


Figure 5-45. High Resistance Driver and Impedance Inverting Network

It is apparent that it is best to choose tubes and operating conditions for low grid driving power. Tubes are available that will operate Class AB_1 at power levels up to 500 watts and their use greatly simplifies the driver and bias regulation requirements.

In cathode driven amplifiers the total grid and screen driving power should not exceed 10% of the fedthru power at maximum signal level. For S/D ratios better than 30 db, it should be correspondingly less.

The plate current of all tubes drops off when the instantaneous plate voltage is low. Figure 5-46 shows a typical plate current curve taken along a straight load line on constant current curves. The grid and screen currents are also shown. Two effects seem to cause the drop in plate current. The main reason for the drop is that current taken by the grid

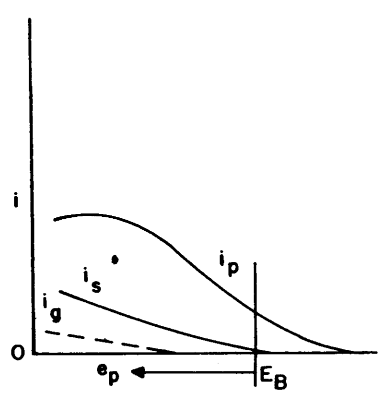


Figure 5-46. Instantaneous Tube Current Characteristic

and screen is "robbed" from the plate. It can be noted on tube curves that the plate current lines depart from straight lines by approximately the amount of grid and screen current. The amount of screen current and drop off of plate current also depends upon the tube geometry. In all but a few transmitting tubes such as the Amperex AX9907 the plate can swing well below the screen voltage before plate saturation takes place. When the plate swings down in this region, the plate current drops off quite a bit. If the distortion requirements are not too high, the high plate efficiency realized by using large plate swings can be realized. Figure 5-47 shows a typical linearity curve of a tetrode linear amplifier. At point "A" the plate is swinging down to the screen voltage. At point "B" it is swinging well below the screen and is approaching the grid to the point where saturation or plate current limiting takes place.

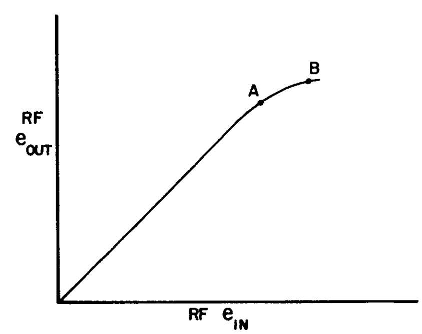


Figure 5-47. Linearity Curve of Typical Tetrode Amplifier

5-7-2. Estimating Distortion

A means of estimating distortion early in the design of a power amplifier is quite useful. This can be done working from a linearity curve drawn from calculated or from estimated operating conditions. I. Gerks derived an equation considering only the 1st and 3rd order components that will give the approximate signal-to-distortion ratio of a two-tone test signal, operating on a given linearity curve. Figure 5-48 shows a linearity curve with two ordinates

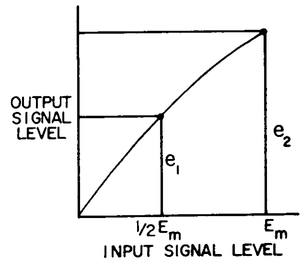


Figure 5-48. Ordinates on Linearity Curve for 3rd Order Distortion Equation

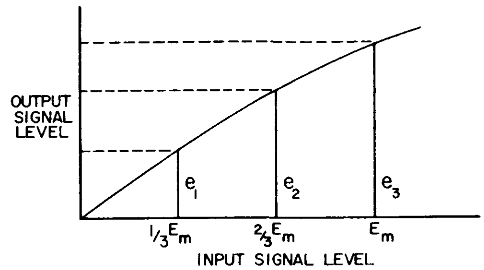


Figure 5-49. Ordinates for 3rd & 5th Order Distortion Equation

erected at half and full peak input signal level. The length of the ordinates e_1 and e_2 is scaled and used in the following equation:

$$\text{S/D ratio in db} = 20 \log \left(\frac{8e_1 - e_2}{2e_1 - e_2} \right) \tag{1}$$

K. Rigoni extended the work of Gerks to include the 5th order component of the linearity curve and also included the effect of 3rd and 5th order curvature on the level of the original tones.

Figure 5-50 shows the method of erecting ordinates for use in the following equations for 3rd and 5th order S/D ratio calculation:

$$S/D_{3rd} = 20 \log \left[\frac{2(-45e_1 + 252e_2 + 167e_3)}{45(-7e_1 - 4e_2 + 5e_3)} \right] \text{db} \tag{2}$$

$$S/D_{5th} = 20 \log \left[\frac{2(-45e_1 + 252e_2 + 167e_3)}{81(5e_1 - 4e_2 + e_3)} \right] \text{db} \tag{3}$$

The accuracy of the results obtained by using this method is limited by the accuracy to which the values of e_1, e_2 and e_3 can be determined. With reasonable diligence it is expected that accuracies on the order of ±2 db at 30 db S/D level can be obtained. If the distortion of a theoretical curve is to be calculated accuracies of ±1 db at 50 db S/D ratio can be realized because the values of e_1, e_2 and e_3 are then exact.

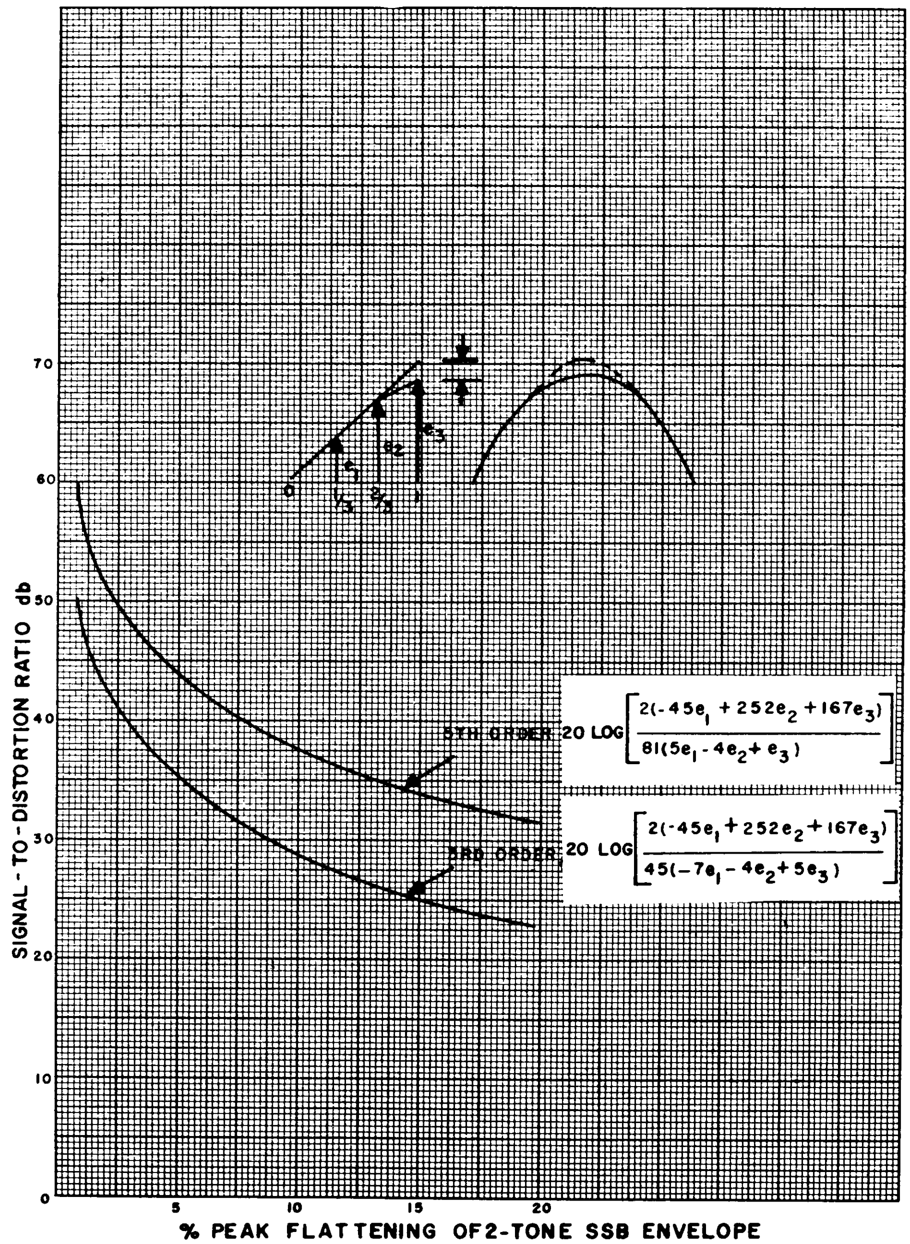


Figure 5-50. Relationship of Peak Flattening of Two-Tone SSB Envelope to Distortion

Envelope peak flattening which is due to grid current loading and plate current nonlinearity at large plate swings is often the major cause of distortion. The amount of envelope peak flattening due to grid current loading may be easily calculated. See Figure 5-50. The equivalent grid load resistance R_g in Figure 5-51 is calculated from the grid driving power and the RF grid swing.

$$R_g = \frac{e_g \text{ peak}}{2(P_g)}$$

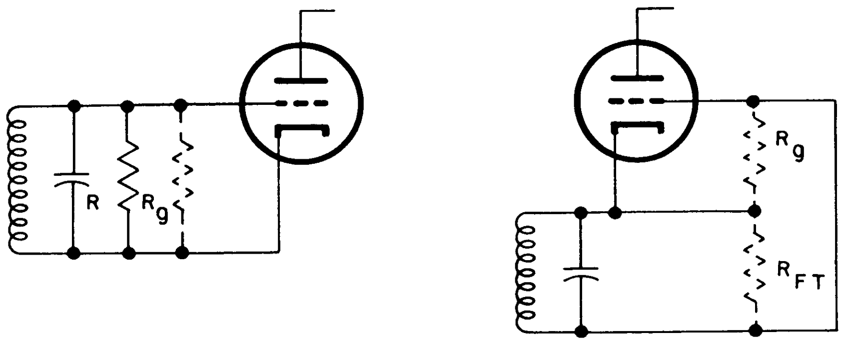


Figure 5-51. Grid Loading

The resistance of the swamping resistor, R, is known or can be chosen for the calculation. The equivalent resistance of R and R_g in parallel is then calculated by:

$$R_{eq} = \frac{R \quad R_g}{R + R_g}$$

If the source impedance looking back at the driver stage is very high compared to R, it will contribute little to improving the driving voltage regulation. In this case, the grid voltage will be reduced on the envelope peak by the amount of reduction from R to R_{eq}.

$$\text{Peak flattening} = \frac{R - R_{eq}}{R} 100 \text{ in percent.}$$

The calculation is made in a similar manner for cathode driven amplifiers. Use the equivalent resistance, R_{ft}, of the fedthru power at the cathode in place of R in the above equations. In tetrode cathode driven amplifiers the grid and screen driving power should both be considered in calculating R_g.

Figure 5-50 shows curves of the relationship of 3rd and 5th order S/D ratio to percent of peak flattening in a theoretical case. A sharp break in the linearity such as might be due to plate voltage swing saturation, Figure 5-52 will contain more 5th and higher order components than if it were a smooth curve from the ordinates e_2 and e_3. This type of nonlinearity is particularly objectionable because of the wide band over which the distortion products appear.

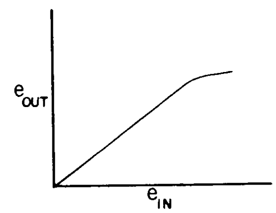


Figure 5-52. Linearity Curve with Plate Saturation

The other principal type of nonlinearity is due to the exponential plate current characteristic of tubes. Figure 5-53 shows such a curve. This type of curve is obtained with Class A amplifiers as well as Class AB. The curvature is kept at a minimum by proper tube choice, operating it at low signal level and over the most linear portion of its characteristic. in Class AB amplifiers, the use of the optimum value of static plate current will do most toward reducing this type of nonlinearity. A smooth curve of this type usually contains mostly 3rd order distortion products. Even though the 3rd order products may be high, the bandwidth over which significant higher order products appear may be relatively narrow.

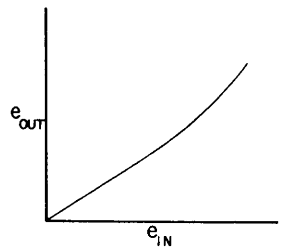


Figure 5-53. Nonlinearity Due to Exponential Plate Current Characteristic

A theoretical curve containing only 1st and 3rd order curvature is shown in Figure 5-54.

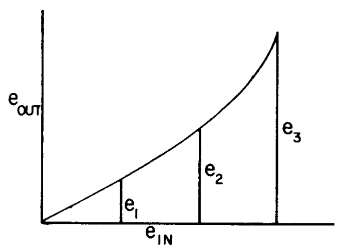


Figure 5-54. Linearity Curve with Only 1st and 3rd Order Components

Compound curves such as shown in Figure 5-55 have relatively stronger 5th and higher order distortion components because the 3rd tends to be cancelled as previously shown in Figure 5-50.

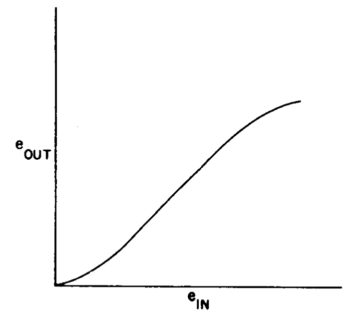


Figure 5-55. Linearity Curve with Compound Curvature

SECTION VI

SINGLE SIDEBAND TRANSMITTER PERFORMANCE MEASUREMENTS

6-1. General

Measurements of power output and distortion are of particular importance in SSB measurements. They are related to the extent that the distortion rises rapidly when the power amplifier is overloaded and the power output is often defined as the maximum peak envelope power output obtainable with a specified signal to distortion ratio. A plot of the S/D ratio vs. peak envelope power is an excellent way of showing a transmitters distortion and power capabilities. A typical curve is shown in Figure 6-1.

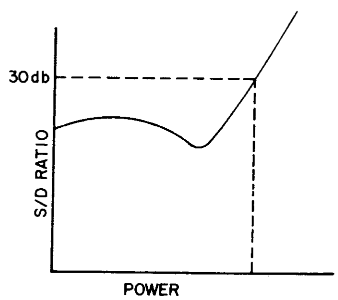


Figure 6-1. S/D vs. Power

Two tones of equal amplitude are used for nearly all measurements in order to provide a "modulation envelope". Some specifications require measurements to be made with a "pilot" carrier at a fixed level of 26 db below the rated peak envelope power output but it serves no useful purpose in aiding performance measurements.

6-2. Distortion Measurements

There are three different methods of indicating or measuring distortion and each has a separate field of usefulness. The "Linearity Tracer" is useful for quick observation of amplifier operation as the effect of various adjustments can be instantly observed. The "Pandapter" shows the relative amplitude of the distortion products to the test tones. It also provides a quick observation of the effect of adjustments and provides a fairly accurate measurement of distortion up to at least a 30 db S/D ratio. The "Sideband Spectrum Analyzer" is used for accurate measurement of the relative amplitudes the various distortion products and pilot carrier to the test tones.

6-2-1. Linearity Tracer

This instrument consists of two SSB envelope detectors the outputs of which connect to the horizontal and vertical inputs of an oscilloscope respectively. Figure 6-2 shows a block diagram of this instrument connected to a power amplifier. A two-tone test signal is normally used to supply an SSB

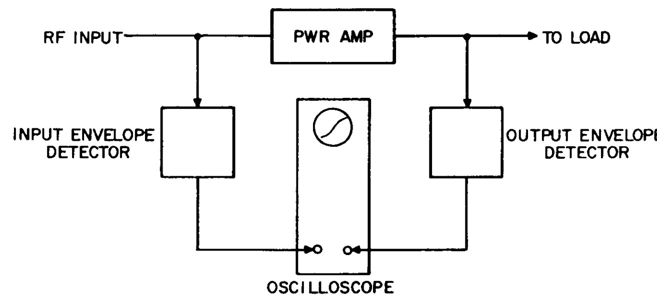


Figure 6-2. Block Diagram of Linearity Tracer

modulation envelope but any modulating signal that provides an envelope which varies from zero to full amplitude can be used. Even speech modulation gives a satisfactory trace so this instrument is unique in that it is an excellent visual monitor. It is particularly useful for monitoring the signal level and clearly shows when the amplifier under observation is overloaded. The linearity trace will be a straight line regardless of the envelope shape if the amplifier has no distortion. Overloading causes a sharp break in the linearity curve. Distortion due to too much bias is also easily observed and the adjustment for low distortion can easily be made.

Another unique feature is that the distortion of each individual stage can be observed. This is helpful in troubleshooting. By connecting the input envelope detector to the output of the SSB generator, the over-all distortion of the entire RF circuit beyond this point is observed which may include mixer stages.

It can also serve as a voltage indicator which is useful in making tuning adjustments.

Figure 6-3 shows the circuit of an envelope detector. Germanium diodes are used as detectors. Any type can be used but they must be fairly well matched in the two envelope detectors. The detectors are not linear at low signal levels but if the nonlinearity of the two detectors is matched, the effect of their nonlinearity on the scope trace is cancelled. The effect of diode differences is minimized by using a diode load of 5000 to 10,000 ohms as shown on the schematic. It is important that both detectors be operated at approximately the same signal level so their differences will cancel more exactly. Although they will operate well on

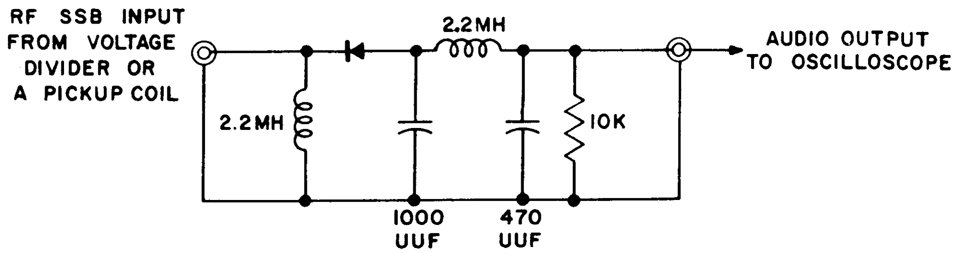


Figure 6-3. Envelope Detector

RF voltages below .1 volt it is desirable to operate them on voltages above 1 volt to further minimize this difference.

It is convenient to build the detector in a small shielded enclosure such as an IF transformer can fitted with coax input and output connectors. Voltage dividers can be similarly constructed so that it is easy to patch in the desired amount of voltage stepdown from the voltage sources. In some cases it is more convenient to use a pickup loop on the end of a short length of coaxial cable.

The frequency response and phase shift characteristics of the amplifiers in the oscilloscope should be the same and flat out to at least 20 times the frequency difference of the two test tones. An oscilloscope such as the DuMont Type 304H is excellent for this purpose. Excellent high frequency characteristics are necessary because the rectified SSB envelope contains harmonics extending 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 as shown in Figure 6-4.

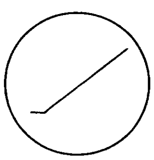


Figure 6-4. Effect of Inadequate Response of Vertical Amplifier

If it is small, it may be safely neglected.

Another effect often encountered is a double trace as shown in Figure 6-5.

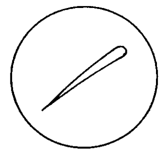
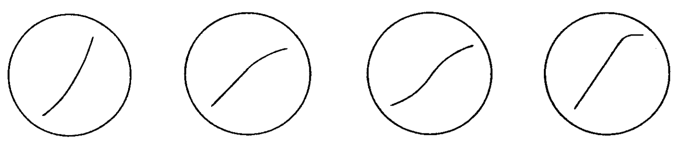


Figure 6-5. Double Trace Due to Phase Shift

This can usually be corrected with an RC network between one detector and the oscilloscope. These effects are usually easily remedied and an accurate linearity trace is not difficult to obtain. The best method of testing the setup is to connect the input of the envelope detectors in parallel. A perfectly straight line trace will result when everything is working properly. One detector is then connected to the other source through a voltage divider which will not require appreciable change in the setting of the oscilloscope amplifier gain controls. Figure 6-6 shows some typical linearity traces. The probable causes and remedies follow:



(a) (b) (c) (d)

Figure 6-6. Typical Linearity Traces

Figure 6-6(a). Inadequate static plate current in Class A or Class B amplifiers or a mixer. Reduce the grid bias, raise the screen voltage or lower the signal level through mixers and Class A amplifiers.

Figure 6-6(b). Caused by poor grid circuit regulation when grid current is drawn or by nonlinear plate characteristics of the tube at large plate swings. Use more grid swamping or lower the grid drive.

Figure 6-6(c). Effect of (a) and (b) combined.

Figure 6-6(d). Overloading the amplifier. Lower the signal level.

6-2-2. The "Pandapter"

This is an instrument which gives a pictorial presentation of the signal spectrum. Figure 6-7 shows a typical trace of two tones with 3rd and 5th order distortion products and a pilot carrier. This instrument sweeps the RF spectrum occupied by the signal and clearly shows the relative frequency and amplitude of each frequency component. A logarithmic amplifier is incorporated in it so the vertical scale can be graduated directly in db. By setting the input level so the two tones touch the top 0 db line, the relative amplitude of the distortion products can be read giving the signal-to-distortion ratio directly. Fairly accurate readings can be made to 30 db S/D ratio and up to 40 db can be estimated after experience is had comparing the trace with other more accurate measurements.

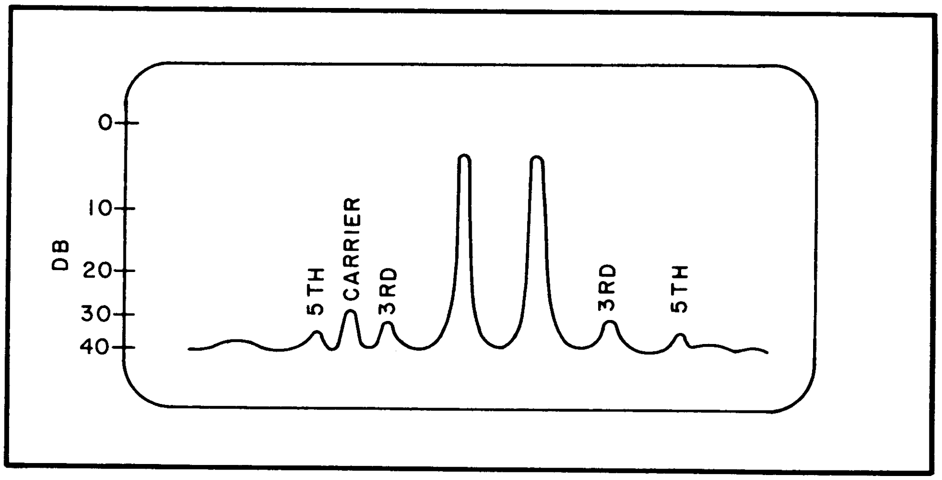


Figure 6-7. Typical "Panadapter" Presentation

6-2-3. Sideband Spectrum Analyzer.

This provides a method of accurately measuring the amplitude of the various frequency components including the distortion products in the sideband RF spectrum. A block diagram of the setup is shown in Figure 6-8.

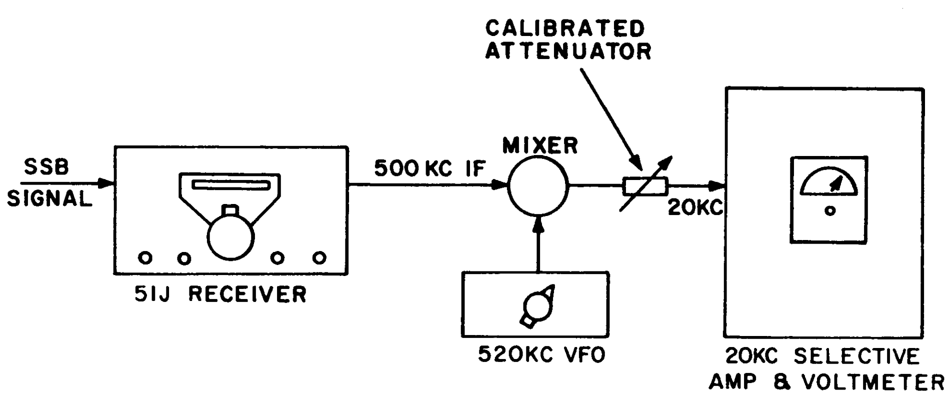


Figure 6-8. Block Diagram of Sideband Spectrum Analyzer

IF output is taken from the first IF stage in the receiver and mixed with the 520 kc VFO output which has a vernier dial for tuning over a range of several kc. The 20 kc selective amplifier has a bandwidth on the order of 100 cps so frequencies very close together can be separated.

In practice, the receiver is tuned to place the SSB signal in the center of its passband. The 520 kc VFO is then tuned to heterodyne one of the test tones to exactly 20 kc as indicated by a peak in the voltmeter reading of the selective amplifier. The voltmeter reading and the attenuator setting are noted. A distortion product is then tuned in with the VFO and the attenuator setting required to give the same meter reading is noted. The db difference of the two attenuator settings is the S/D ratio of that particular distortion product.

Since the response curve of the selective amplifier is very sharp, it is necessary that the transmitter, receiver and 520 kc VFO have excellent short time frequency stability. If the transmitter has three crystal oscillators as does the LD-T2, there is a total of six oscillator stabilities involved. When the selective amplifier bandpass is narrowed so that frequencies 100 cps apart and 40 db different in amplitude are to be measured, the frequency stability of all these oscillators becomes the limiting factor in making measurements.

A method of eliminating the effects of frequency drift is to use a monitor receiver that uses the same frequency scheme as the transmitter and the same conversion oscillators. In this manner, the HF oscillator frequency drifts cancel out. Only the difference in the drift between the 250 kc oscillator and the 270 kc VFO will cause the 20 kc signal to drift. These two oscillators can easily be made stable enough for this application.

The monitor mixers must be designed and operated so their distortion is much less than that of the transmitter under measurement such as on the order of 60 db. By accurately calibrating the 270 kc VFO dial the frequency of each frequency component can also be read.

Figure 6-9. Deleted

6-2-4. Deleted

6-3. Power Output

SSB transmitters are rated in terms of peak envelope power output but the power is usually measured with a two-tone test signal. When a test signal of two tones with equal amplitude is used and no carrier is present, the actual watts dissipated by the load is one-half the peak envelope power. Figure 6-10 shows the relationship of the average power to the peak envelope power for (a) a two-tone

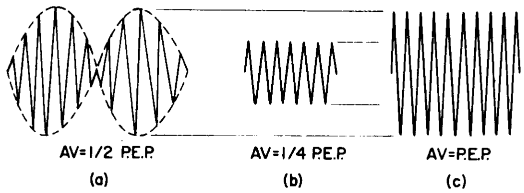


Figure 6-10. Relationship of Average Power to Peak Envelope Power

signal, (b) same as (a) but with one tone removed, and (c) when the amplitude of one test tone is doubled in amplitude. If three test tones of equal amplitude were used, the power in each tone in 1/9 of peak envelope power and the total output is 3(1/9) of 1/3 peak envelope power. These relationships apply if there is no distortion of the SSB envelope. The average to peak envelope power of a two-tone signal will increase a little if peak flattening takes place. This is usually neglected since it is small if the distortion is low.

When a pilot carrier 20 db below one of the two test tones is present, the peak envelope power is 2.2 times the average power. It is noted that when the carrier is left constant and the tones are reduced, this ratio increases.

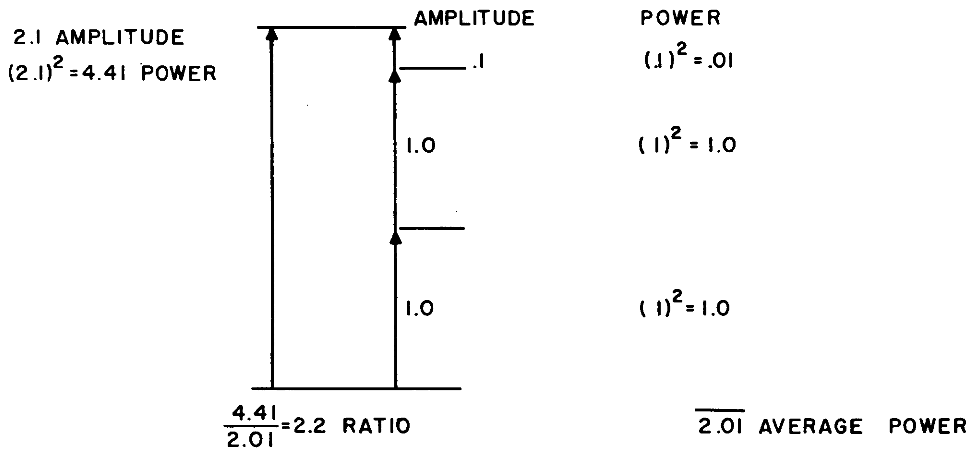


Figure 6-11. Method of Computing P. E. P. to Average Power

The power output can be measured by conventional means provided the measuring instrument characteristics are known. The calorimic method will show average power. Most RF ammeters read average current so if the equivalent series load resistance is known, the power can be computed by the I^2R relationship. Peak reading RF voltmeters can also be used across the load if the equivalent parallel resistance of the load is known. A Jones "micromatch" is not suitable when a two-tone signal is used because the diode rectifier characteristics and time constants cause a reading in between the average and peak envelope power. It can be calibrated for a given two-tone test condition, however, or a single tone test that is 6 db greater than one of the two-tones can be used.

6-4. Carrier Compression

This test applies only when a pilot or reduced carrier is used. It is a measure of the amount the carrier amplitude reduces when a two-tone test signal is applied that will drive the transmitter to rated peak envelope power output. This reduction is expressed in db and specification limits are usually 1 db. This can also be interpreted to be the change in steady stage gain from a small signal to the maximum rated signal level. This change is caused by distortion (particularly peak envelope flattening) and the effects of power supply regulation.

6-5. Noise Level

The hum and "unweighted" noise output is usually specified as 50 db (e.g.) below the level of one of two equal test tones. It could also be specified as 56 db (e.g.) below the peak envelope power.

Various means have been devised to make this measurement. Essentially, they require a linear diode rectifier that is flat within 1 db over the transmitter bandwidth. A carrier of some type is required for demodulating the reference tone and the noise. The noise reading is compared to a tone of a known amplitude and from this the noise level is calculated.

For example, the method used by Bell Labs and called for in some specifications is as follows. Apply a carrier 20 db below peak envelope power and a 1000 cps tone 35 db below peak envelope power. Read the noise meter rectified output which corresponds to the level of the one tone 35 db below peak envelope power.

Remove the tone and read the noise meter again. The difference between these two readings in db is added to 29 db to give the noise level referred to one tone of a two-tone signal. If it were added to 35 db it would give the noise level below the peak envelope power output. Figure 6-12 may help to show these relationships more clearly. Some specifications require that the amount of carrier compression be subtracted from the noise level thus obtained but this only reduces the measured value by 1 db or less.

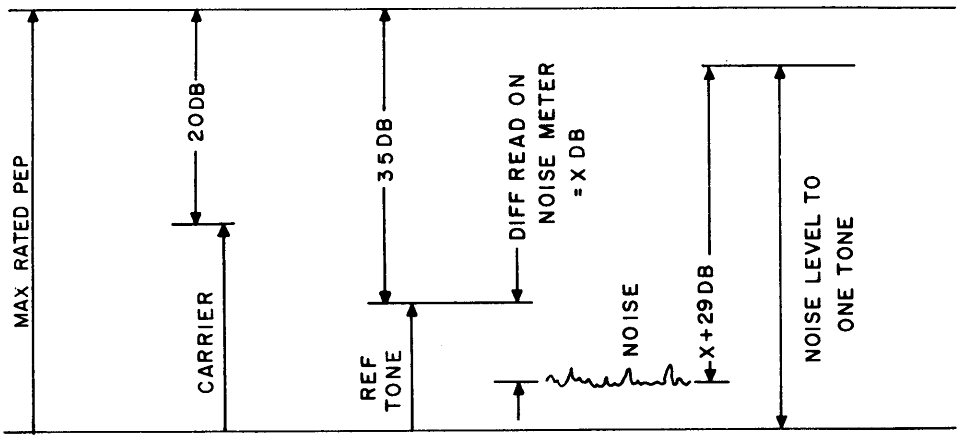


Figure 6-12. Relative Levels in Noise Measurement

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3. "The Care and Feeding of Power Tetrodes", Eitel-McCullough, Inc., page 17.

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5. "A Simplified Harmonic Analysis", E. L. Chaffee, Review Scientific Instruments, October 1936, volume 17, page 387.

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APPENDIX 1

TEST FOR PARASITIC TENDENCY

It is a common experience to develop an engineering model of a new equipment that is apparently free of parasites and then find troublesome parasites showing up in production units. Sometimes the parasitic oscillations may not show up until after the equipment has been in operation for some time. The reason, of course, is that the equipment has a parasitic tendency that remained below the verge of oscillation until some change in a component, tube gain or operating condition raised the gain of the parasitic circuit enough to start oscillation. In some cases, the parasitic tendency may have been unknown and in other cases when parasites were found in development they were not suppressed sufficiently. A reliable equipment must have a wide margin of stability at all parasitic resonances.

Some tests can be made to find the frequencies where parasitic tendencies occur. In most high frequency transmitters there are a great many resonances in the circuits at frequencies other than the desired operating frequency. Some are at low frequencies which usually involve RF chokes, coupling capacitors and bypass capacitors. Others appear above the operating frequency and often a dozen or more can be found using a grid-dip m ter. Also, occasionally a parasitic resonance near the desired op rating frequency can be found. Most of these "parasitic" resonant circuits are not coupled to a tube in such a manner as to be regenerative and have no significant tendency to oscillate. Even though not r generative, a high Q parasitic resonance on a harmonic of the operating frequency may build up a voltage creating an undesired amount of harmonic output and in some cases can cause voltage flashovers in high powered transmitters.

Usually there are a few parasitic resonant circuits that are coupled to a tube in some form of an oscillator circuit. If the regeneration is great enough, oscillation at the parasitic frequency results. These parasites, of course, are suppressed during development but those just below oscillation must also be found and suppressed to a safe level.

One test is to feed a signal into the grid of one stage and measure the signal level on the plate of this and one or two succeeding stages as shown in figure 1. The test is made with all operating voltages applied to the tubes. Class "C" stages should have bias reduced so a reasonable amount of static plate current flows. The sensitivity of the RF voltmeters shown in figure 1 must be sufficient to obtain some indication with a signal on the first grid at other than the operating frequency. The signal generator is tuned over a range of at least 100 kc to 200 mc for transmitters designed to cover the 2 mc to 30 mc frequency range, for example. If high inductance chokes or high capacitance capacitors are used that can resonate below 100 kc, tests should be mad of these frequencies also. As the signal generator is tuned across this frequency range, the relative levels of the RF voltmeters are watched and the frequencies at which peaks occur are noted. Each significant peak in voltage gain over either stage must be investigated further. This test only locates potential troublesome frequencies and does not measure how far below oscillation any parasite may be. This can be done if desired following the procedure given on pages 46 and 47.

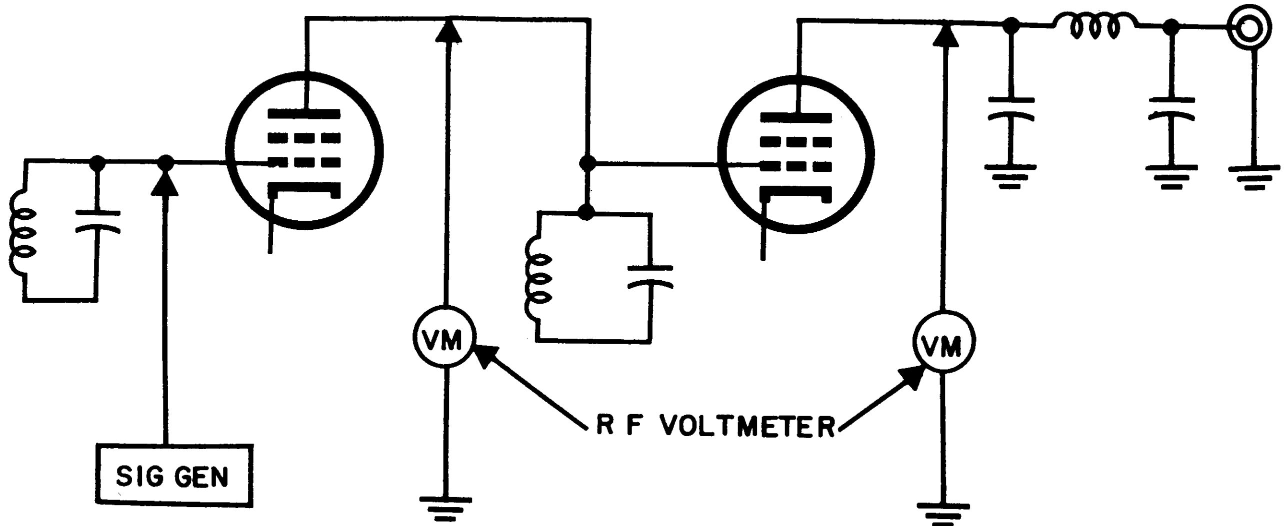

Figure 1. Measuring Parasitic Gain

An alternative is to change the circuit or add suppression to r duce all significant peaks by 10 db or more. It should be noted that peaks found in the above test do not necessarily indicate an oscillation tendency. It does weed out most of the unimportant resonances that can be found with a grid-dip meter, however.

Parasitic tendencies around the feedback loop in a two-stage RF feedback amplifier can be tested as shown in figure 2. The signal generator is connected to the high voltage feedback capacitor which

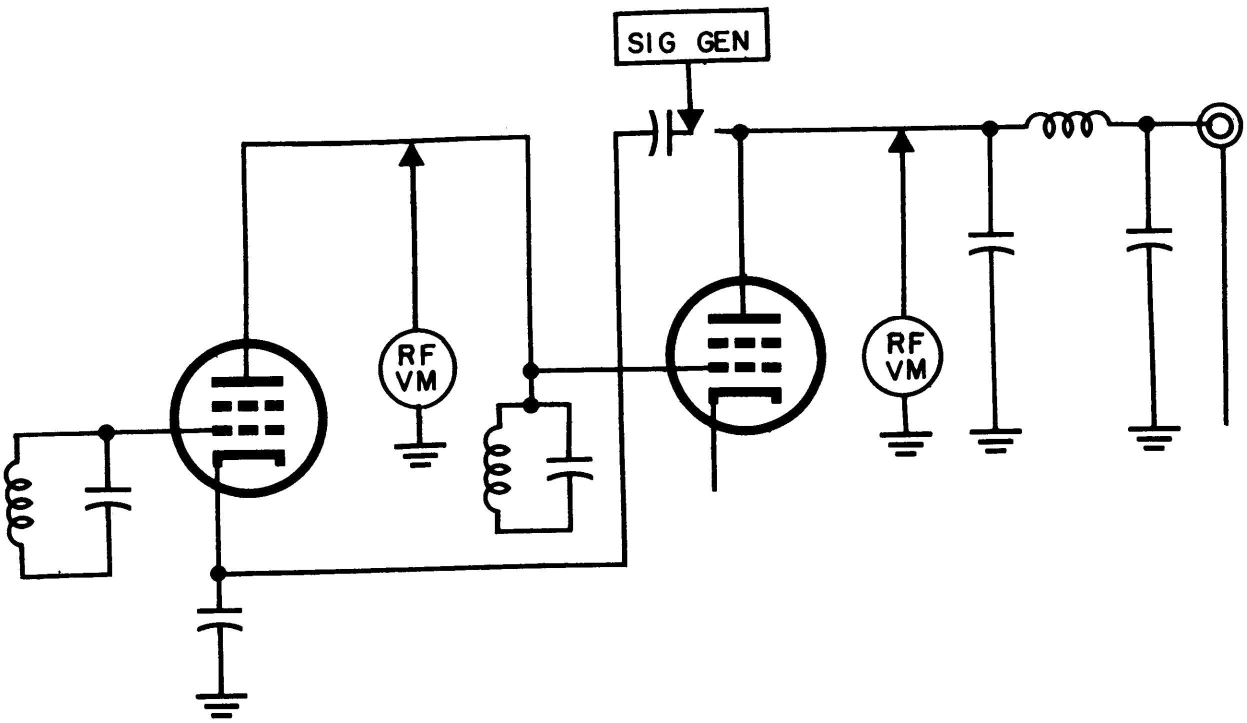

Figure 2. Measuring Parasitic Gain Around Feedback Loop

is disconnected from the PA plate. It is necessary to connect in this manner rather than directly across the cathode capacitor because the low output impedance of the signal generator may load parasitic circuits in the cathode circuit so that they might not be observed.

The gain around the feedback loop is measured in this test by taking the ratio of the output plate RF voltage to the signal generator output voltage. A gain of unity or more will cause oscillation if phase is correct. High gain at parasitic frequencies are undesirable regardless of phase and should be attenuated to at least 10 db below unity gain around the loop and around any stage.

The purpose of this discussion is to point out that every power amplifier can, and should be, tested for parasitic tendencies. Some thought and ingenuity may turn up better and easier tests. The end result is to develop equipment that is more trouble-free and more reliable.

APPENDIX 2

APPLICATION NOTES ON

TWO STAGE RF FEEDBACK AMPLIFIER

Neutralizing Procedure

Experience with several equipments using feedback around two RF stages has brought out several different methods of neutralizing. An important observation is that when all three neutralizing adjustments are correctly made the peaks or dips of the various tuning meters all coincide at the point of resonance. For example, the coincident indications when the various tank circuits are tuned through resonance and with feedback operating are:

PA plate circuit tuned through resonance

- PA plate current dip
- Power Output Peak
- PA RF grid voltage dip
- PA grid current dip (Note: PA grid current peaks when feedback is disabled and tube is heavily driven.)

PA grid circuit tuned through resonance

- Driver plate current dip
- PA RF grid voltage peak
- PA grid current peak
- PA power output peak

Driver grid circuit tuned through resonance

- Driver RF grid voltage peak
- Driver plate current peak
- PA RF grid voltage peak
- PA RF grid current peak
- PA plate current peak
- PA power output peak

Provisions for metering all of the above voltages and current in high powered transmitters is justified but low powered equipments usually do not have enough instruments to adequately check neutralization by this means. The most useful instruments are pow r output, PA plate current, PA RF grid voltage and driver plate current. The power output indicator can be an RF ammeter, RF voltmeter or directional coupler. If the PA is driven into grid current a PA grid current meter will serve as well as a PA RF grid voltmeter. The meters should be arranged so the following pairs of readings are displayed on m t rs located close together for ease of observation of cóincident peaks and dips:

- PA plate current and power output
- PA RF grid voltage and PA plate current
- PA RF grid voltage and power output
- Driver plate current and PA grid RF voltage

The third pair listed above may not be necessary if the PA plate current dip is pronounced but often the dip is small and in this case the power output indication is a much more sensitive indication. It is a good double check, also.

When the above instrumentation is provid d the neutralizing procedure is as follows:

1. Remove the RF feedback
2. Neutralize C_{gp} of driver stag (Driver neutralization)
3. Neutralize C_{gp} of PA stage (PA neutralization)
4. Apply RF feedback
5. Neutralize driver grid-cathode capacity (Feedback neutralization)

These steps will be explained in more detail in the following paragraphs:

Step 1: The removal of RF feedback through the feedback circuit must be complete. The switch shown in the feedback circuit on the schematic, figure 1, is one satisfactory method. Since C_6 is effectiv ly across the PA plate tank circuit it is desirable to keep it across the circuit when feedback is removed to avoid appreciable detuning of the plat tank circuit. Another method that can be used if properly don is to ground the junction of C_6 and C_7. Grounding through a switch or relay is not good enough b cause of common coupling through the grounding lead length. The grounding method shown in figure 2 is satisfactory This has been used in high powered transmitters and is an adequate and convenient means of removing feedback manually for neutralizing tests.

Step 2: Plate power and an excitation signal is applied. The driver grid tank circuit is resonated by tuning for a peak in driver RF grid voltage or driver plate current. The PA grid tank circuit is then resonated and adjusted for a dip in driver plate current. The driver neutralization is adjusted until the PA RF grid voltage (or PA grid current) peaks exactly at the driver plate current dip.

A Rule of Thumb for adjusting grid plat n utralizing of a tube without feedback is: With all circuits in resonance detune the plate circuit to the high frequency side of resonance. If grid current to next stage (or power output) increases, more neutralizing capacity is required and vice-versa.

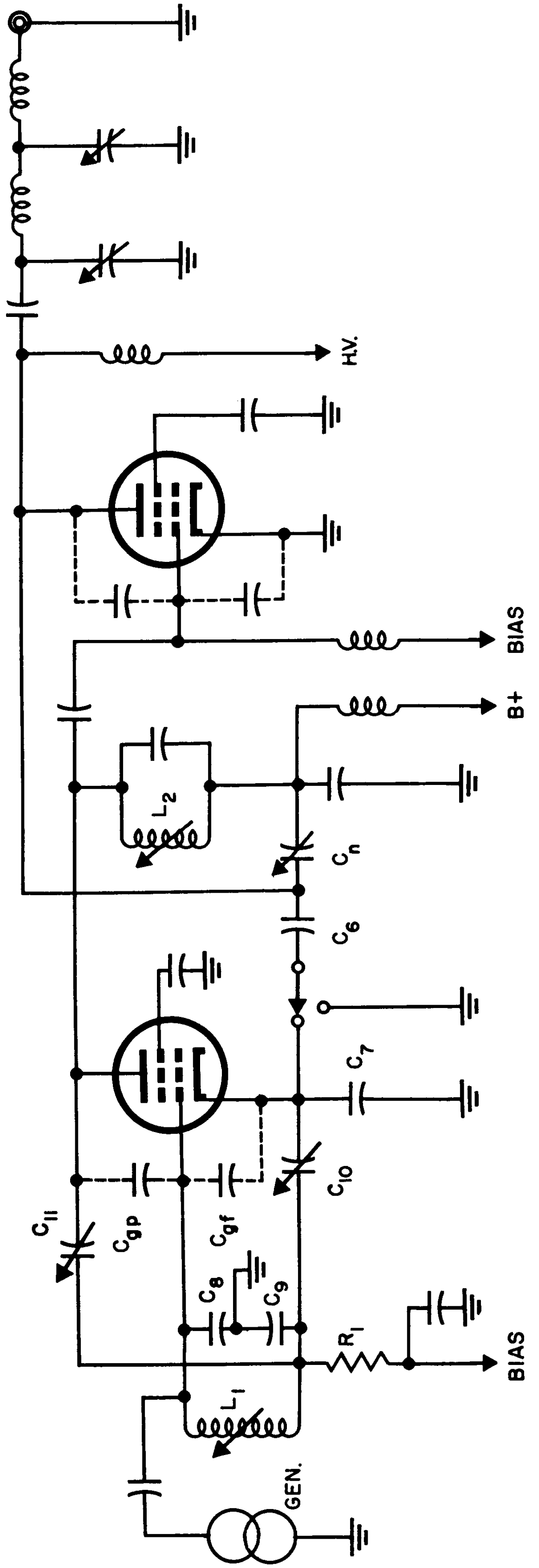

Figure 1. Two Stage Amplifier With Feedback

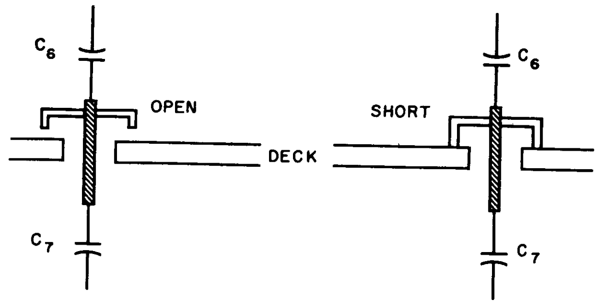

Figure 2. Feedback Shorting Device

If the driver tube operates Class A so that a plate current dip cannot be observed, a different neutralizing procedure is necessary which will be given in a subsequent section.

Step 3: Same as Step 2 except applied to the PA stage. Adjust neutralization for a peak in power output at the plate current dip.

Step 4: Reverse of Step 1.

Step 5: Apply plate power and an RF signal that will drive to nearly full power output. Adjust the feedback neutralization for a peak in PA power output at the PA plate current dip. The rule of thumb here is to decrease the feedback neutralizing capacity if power output rises as a tank circuit is tuned to high frequency side of resonance.

The procedure in the above steps applies when the neutralizing adjustments are approximately correct to start with. If they are far off some cut and try may be necessary. Also, the driver may oscillate if the feedback neutralizing is not near its correct setting.

In the above procedure it is assumed that a single test tone is used for the RF signal and that all tank circuits are at resonance except the one being detuned to make the observation. There is some interaction between the driver neutralization and the feedback neutralization so if an appreciable change is made in any adjustment the others should be rechecked. It is important that the grid-plate neutralization be accomplished first when using the above procedure, otherwise the feedback neutralization will be off a little since it partially compensates for that error.

Test Bench Neutralization

Section 5-3-5

gives a method where a sensitive RF detector is inductively coupled to a grid tank coil. This has been found difficult to apply in some equipment because of the mechanical construction of some variable inductance coils and because of undesired coupling. A shielded pickup link is necessary to eliminate capacitive coupling completely.

Another method for observing neutralization has been found that appears to be more accurate in practice. A sensitive RF detector such as a receiver is capacity coupled loosely to the grid of the stage being neutralized. About 1 mmfd coupling capacitance is usually enough. It must be small to avoid upsettin the neutralization when it is removed because the total grid-to-ground capacitance is one leg of the bridge. A sensitive vacuum tube RF voltmeter can be used if its capacity loading is taken care of in som manner. Figure 3 shows the feedback neutralization circuit. A signal generator is connected at point "S"

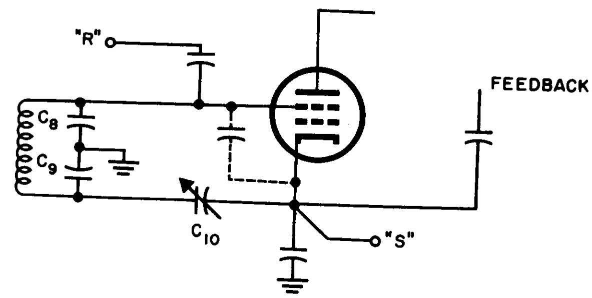

Figure 3. Feedback Neutralizing Circuit

and the receiver (or other sensitive detector) at poin "R". If C_{10} is not properly adjusted the S-meter on the receiver will either kick up or down as the grid tank circuit is tuned through resonance. C_{10} is simply adjusted for minimum deflection of the S-mete reading as the circuit is tuned through resonance.

The grid-plate capacity of the tube is then neutralized by connecting the signal generator to the plate of the tube and adjusting C_{11} in figure 1 for minimum deflection again as the grid tank is tuned. The PA stage can be neutralized in the same manner by coupling a receiver loosely to its grid and connect ing a signal generator to the plate of the tube. The signal can be fed into the PA output terminal if desired.

Some precautions are necessary when using this method, however. One is that the input capacitance of some tubes used as drivers (e. g. 6CL6) have appreciably more effective input capacitance when in operation and conducting plate current. This increas may be 3 or 4 mmfd and since this is part of the neutralizing bridge circuit it must be considered. This means that the neutralizing adjustment of such tubes must be made when they are conducting normal plate current. Stray coupling must be avoided and it may be helpful to remove filament power from the preceding stage or disable its input circuits in some manner.

It should be noted that in each of the above adjustments that minimum reaction on the grid is desired, not minimum voltage. Some voltage is inherent on the grid using this neutralizing circuit. For example the feedback neutralizing circuit of figure 3 can be redrawn to show this more clearly, as shown in figur 4.

C_8 and C_{gk} form a capacitive voltage divider so when a signal is applied at "S" a signal at "R" can be expected. Actually, the circuit design should call or C_8 to be several times the tube input capacitance because some of the feedback applied to the cathode is lost by this voltage division and it depends upon a relatively large value of C_8 to keep this loss down to 1 db or so.

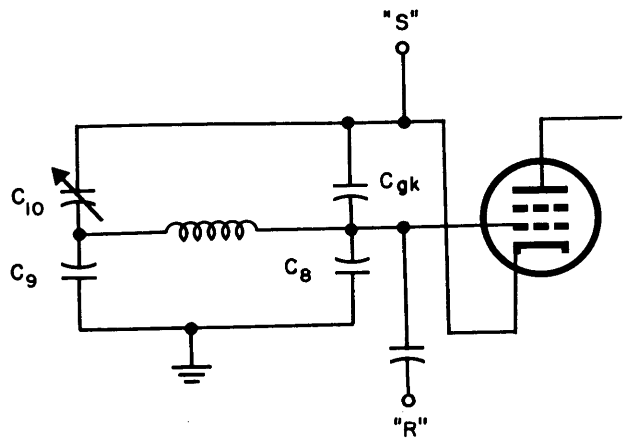

Figure 4. Feedback Capacity Division

Effects of Series Cathode Impedances

An impedance in series with the cathode or filament of a tube can cause the stage to be regenerative or degenerative depending upon the sign of the reactance. Text books give the input resistance of a pentode as

$$R = \frac{1}{w^2 g_m L_k C_{gk}} \quad \text{where}$$

$w = 2\pi f$

g_m = transconductance of tube

L_k = series cathode inductance

C_{gk} = grid-cathode capacitance

This can also be written as

$$R = \frac{X_{C_{gk}}}{g_m X_k}$$

where X_k is the series cathode reactance.

When tubes are operated Class AB the effective g_m can be given as

$$gm = \frac{i_p}{e_g}$$

where i_p is the peak of the fundamental ac component of plate current and e_g is the peak RF grid voltage. These two values are always determined when making a calculation of tube operating conditions. The input resistance can then be written as

$$R = \frac{X_{C_{gk}} e_g}{X_k i_p}$$

It is noted that the resistance decreases as the square of frequency when the series cathode reactance is inductive. This can cause serious loading of the input circuit at frequencies on the order of 30 mc and higher. This should be kept in mind when tube types are selected, the socketing arrangement chosen and when the cathode grounding or bypassing connections are made. In general the inductance in the cathode-to-ground lead should be as low as possible.

This input resistance becomes quite low with a 10 KW transmitting tube such as the type 6166. Even with reasonable precaution in filament bypassing this input loading consumes nearly 100 watts of drive power per tube at 40 mc, which is several times the actual grid driving power alone.

This power is not lost because it is fed through the tube and increases the output. One way to visualize this is to consider that the tube is being cathode driven (grounded-grid) by the magnitude of RF voltage that appears across the cathode inductance, L_k. This peak voltage $e_k = i_p X_{L_k}$ and it is in phase with the current because when the grid circuit is tuned the cathode inductance is tuned out. Input power is fed through the tube just like a cathode driven tube and the magnitude of this power is $P_f = \frac{e_k i_p X_{L_k}}{2}$.

An interesting observation is that this loading on the input circuit is degenerative and is a form of feedback if a high source impedance driver such as a tetrode is used. For example, if the gain of the PA drops a little its RF plate current will also drop, which causes the input resistance to rise. This effect would be useful except that PA screen current and grid current often are an appreciable part of the cathode current and they are not linear so this type of grid circuit loading may be more harmful than beneficial. A fixed value of resistance loading is usually better.

One method of eliminating this input loading is to series resonate the cathode lead inductance by proper choice of bypass capacitors. This is practical for fixed frequency operation, in particular.

Another important point to notice is that capacitive reactance in series with the cathode causes the input resistance to be negative. Inspection of this circuit, figure 5, shows its similarity to a Colpitts oscillator. The circuit is inherently regenerative and will oscillate if the reactance C_k is increased to the point where the negative input resistance equals the resistive load of the grid tank circuit. This explains why proper cathode and filament bypassing is so important from a stability standpoint, not only at carrier frequency but also at frequencies of both low and high frequency parasitic resonances in the circuit. Note, also, that the negative input resistance is not a function of

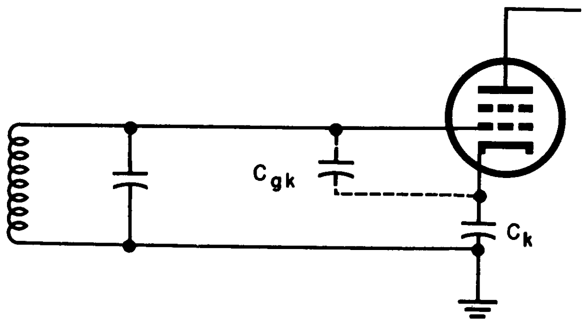

Figure 5. Capacitive Reactance in Cathode

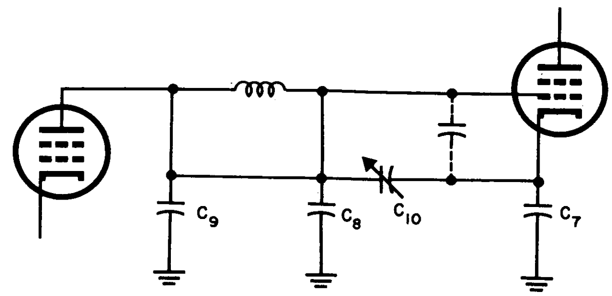

Figure 6. Feedback Neutralization With Pi-Network Input

frequency and can be written as

$$R = \frac{-C_k}{C_{gk}}\left(\frac{e_g}{i_p}\right) \quad \text{or} \quad \frac{-C_k}{C_{gk}\, g_m}$$

For example, if the QX or equivalent parallel resistance of the grid current is 5000 ohms, g_m is 10,000 umhos, C_{gk} is 10 mmfd, the circuit will start oscillating when C_k is reduced to 500 mmfd.

$$R = \frac{500}{10\ (10{,}000)\ \text{x}10^{-6}} = -5000 \text{ ohms}$$

Neutralization of Grid-Cathode Capacitance

In the preceding section it is noted that the input resistance depends upon the value of grid-cathode capacitance. If this capacitance is neutralized, the input resistance will be infinite and will not be affected by impedance in series with the cathode. Actually, this is the function of "Feedback Neutralization" in the two stage feedback amplifier, that is, to eliminate regeneration or degeneration due to the capacitor in series with the driver cathode. Several different neutralizing circuits can be used but the one shown in figure 1 and figure 3 is generally used because it is of the broadband type in that it needs no adjustment over a wide frequency range. These so-called broadband neutralizing circuits only neutralize the circuit at the resonant frequency of the tuned circuit. At other than resonant frequencies the gain of the amplifier stage is very low because the tuned circuit appears as a low reactance and causes the stage gain to be so low it doesn't oscillate even though not actually neutralized.

Another very similar neutralizing circuit is shown in figure 6. The capacitors are designated so the equations given previously apply to this circuit also.

Coil neutralization is a different but excellent means of neutralization but it must be readjusted for each carrier frequency. Figure 7 shows this circuit and L_3 is adjusted to resonate the grid-cathode capacitance. This adjustment cannot be made with a grid-dip oscillator because other capacitances, such as the grid-screen capacitance, cannot be eliminated from the neutralizing circuit. The usual reaction indications

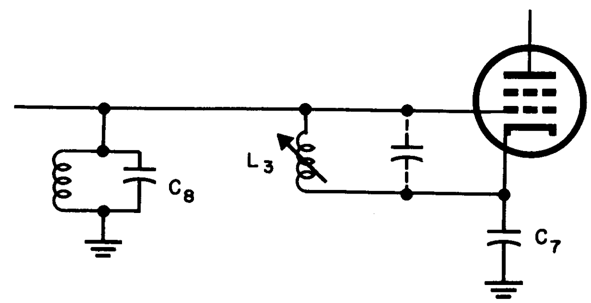

Figure 7. Coil Neutralization

outlined previously are used to adjust this circuit also. Since the setting of L_3 affects the grid circuit resonance the grid tank must be retuned after each adjustment of L_3. This circuit is useful on VHF frequencies where L_3 is not so large. It has the advantage that no feedback is lost due to capacity division as in other circuits.

Neutralizing the grid-cathode capacitance also has the effect of removing C_7 from the input circuit. The size of C_7 should be kept as large as practical, however, because it will make the neutralizing less critical and the driver will have less tendency to oscillate when out of neutralization.

Phase Angle Error of Feedback

Another point often considered but apparently of secondary importance is that the RF plate current through this cathode capacitor is in quadrature with the feedback current. This causes a phase shift in the feedback voltage on the driver cathode. The equivalent resistance across C_7 caused by the plate current is

$$Re = \frac{e_k}{i_p}$$

where e_k is the peak RF voltage across the cathode capacitor, C_7, divided by the fundamental component of driver plate current. Another way of estimating Re is to consider that the feedback drives the cathode of the tube and that the cathode driven input resistance is raised by the feedback as in the following equation.

$$Re = \frac{1}{gm} \frac{e_k}{e_{gk}}$$

where e_{gk} is the peak grid-to-cathode RF voltage. For example, let g_m = 10,000 umho, C_7 = 1000 uuf, frequency = 2 mc and use 12 db of feedback.

$$Re = \frac{1}{.010,000} \frac{1}{3} = 330 \text{ ohms}$$

X_{C_7} at 2 mc = 80 ohms

The phase shift of the feedback is then the angle whose tangent is $\frac{80}{330}$ or 13.6 degrees. This is a rather large error but it decreases directly with frequency. An increase in feedback will reduce the phase angle error but the error than becomes more important.

Driver Screen Voltage

The driver tubes are often operated Class AB_1 so the power they can deliver to the PA grid circuit depends upon the effective screen voltage. The actual screen-to-cathode voltage at the positive peak of RF grid voltage is the DC screen voltage minus the peak RF voltage fed back to the driver cathode. For example, if calculations of the driver stage indicate that 350 volts effective screen voltage is required and the feedback voltage is 150 volts peak, it will be necessary to use a DC screen supply voltage of 350 + 150, or 500 volts. High DC screen voltages call for high static plate current for good linearity which causes high static plate dissipation. All of these factors must be considered to work out the best compromise for the driver of a high powered feedback amplifier.

Parasites

Two kinds of parasites associated with this feedback circuit have been found. One is a VHF parasite which requires a high Q parasitic resonance in the PA plate circuit and a phase reversal between the driver plate and PA grid. Figure 8 shows the two

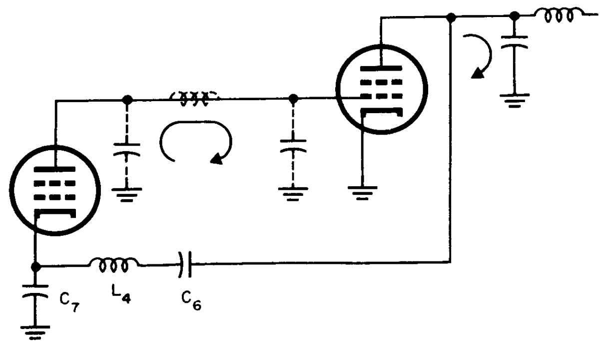

Figure 8. VHF Parasitic Circuit

conditions necessary. The feedback is positive due to the phase reversal caused by the Pi-Network type of parasitic resonance between the driver plate and PA grid. Very little coupling from plate to grid is required because the two tetrodes have very high gain. The usual parasitic suppression techniques can be used but a very effective one is to reverse the phase of feedback at the parasitic frequency. This is accomplished by adding a small inductance, L_4, in series with the feedback capacitor, C_6. These two components should go through series resonance just below the parasitic frequency to suppress it. This amount of inductance will increase the feedback slightly at the highest normal operating frequency.

A somewhat similar low frequency parasite can be encountered that is due to a phase reversal in the feedback voltage divider circuit. An RF choke is required across C_7 to provide a DC path. The phase of the feedback is reversed below the resonant frequency of this choke and C_7. This type of parasite can be controlled by proper choice of RF chokes, blocking capacitors and bypass capacitors or by loading them with resistance damping.

As mentioned previously, the driver grid-cathode capacitance is neutralized only at the operating frequency. Regeneration due to the cathode capacitor can cause parasitic oscillation if suitable parasitic resonances occur in the driver grid circuit.

GPO 925187

HEADQUARTERS
AIRWAYS AND AIR COMMUNICATIONS SERVICE
MILITARY AIR TRANSPORT SERVICE
Andrews Air Force Base
Washington 25, D. C.

22 July 1957

SUBJECT: ERRATA NOTICE

TO: All Concerned

1. The Collins Radio Company has advised that there are errors in the material within this publication. You are requested to make these changes:

a. Page 22 under paragraph 4-4-2 entitled "Relative Level of Mixer Products." Only the first two paragraphs are valid. The information contained in the balance of paragraph 4-4-2 should be disregarded.

b. Page 56. Correct the formulas shown to read as follows:

$$P.O. = \frac{I_p e_p}{2} \text{ watts output}$$

$$Pdr = \frac{I_g e_g}{2} \text{ watts grid driving power}$$

$$R_L = \frac{e_p}{I_g} = \text{RF plate load resistance}$$

c. Page 57. Correct the formulas shown by deleting the capital sub "I" in all formulas given.

H. B. MCCULLOUGH
Colonel, USAF
DCS/Personnel