

MILTON S. KIVER

F-M

Simplified

Third Edition

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D. VAN NOSTRAND COMPANY, INC.

Princeton, New Jersey

Toronto

New York

London

D. VAN NOSTRAND COMPANY, INC.
120 Alexander St., Princeton, New Jersey (*Principal office*)
24 West 40 St., New York 18, N. Y.

D. VAN NOSTRAND COMPANY, LTD.
358, Kensington High Street, London, W.14, England

D. VAN NOSTRAND COMPANY (Canada), LTD.
25 Hollinger Road, Toronto 16, Canada

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D. VAN NOSTRAND COMPANY, INC.

Published simultaneously in Canada by
D. VAN NOSTRAND COMPANY (Canada), LTD.

Library of Congress Catalogue Card No. 60-10289

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01612c40

PRINTED IN THE UNITED STATES OF AMERICA

F-M

Simplified

BY THE SAME AUTHOR

Television Simplified, 5th Ed.

Television Receiver Servicing, 4th Ed.

Introduction to UHF Circuits and Components

PREFACE

The tremendous vigor and vitality that frequency modulation broadcasting is exhibiting—in response to an increasing demand for high fidelity—has prompted a complete revision of this book. Every section of the text has been reworked to bring it completely into step with current practice. While the most extensive revision has occurred in those chapters dealing with F-M reception, considerable additional material has been added to the chapters dealing with antennas and transmitters. At the same time, circuits and practices no longer in vogue have been deleted.

The chapter sequence has remained unaltered. After the fundamentals of F-M have been presented in the first three chapters, a full chapter is devoted to the propagation, reception, and transmission of F-M signals. Then, in Chapters 5 through 10, F-M receivers are analyzed stage by stage and circuit by circuit, at all times adhering to direct, non-mathematical explanations. Chapter 11 deals with circuit alignment, first with the use of a signal generator and a VOM or VTVM, and then visually with an oscilloscope and a sweep generator. With these circuit explanations and techniques understood, the discussion turns to commercial F-M receivers (Chapter 12) and to their servicing and maintenance (Chapter 13).

The final three chapters are devoted to modern F-M transmitters. Equipment of major importance is included. Transmitters are grouped according to the process they utilize to generate the F-M signals, i.e., whether it is by the reactance method or by the phase-shift method.

A set of questions is placed at the end of each chapter for those who wish to gauge their progress through the book and their understanding of the principles involved. With the exception of the few mathematical problems, no question can be properly answered with a single word.

New to this edition is a bibliography which has been selected, in part, to direct the reader to those references which outline the foundations upon which F-M rests, its advantages and limitations. Often, these articles were

written by the men who pioneered this field and the reader who takes the time to study these papers in detail will come to appreciate first-hand the development steps that lead to the present state of the art. Other references, more recent in origin, show what is being done now in the field.

A final addition to this edition is a "Common Troubles Check List" which lists the defects most often encountered in F-M receivers, their possible causes and the section of the receiver in which they occur.

M.S.K.

Highland Park, Illinois
January, 1960

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Chapter 1

INTRODUCTION TO FREQUENCY MODULATION

Advantages of F-M. Although many improvements had been made in radio since the days of the first radio receiver, it was not until the advent of frequency modulation that noise interference was efficiently reduced in radio reception. Noise interference, for the most part, is due to sources, either man-made or natural, that are external to the receiver. Much of the worst interference comes from natural sources, generally beyond the power of man to control. A-M receiver circuits and systems designed by engineers were of limited effectiveness in their suppression of interference. The best solution seemed to be the employment of brute-force method, using high-powered radio signals to override the disturbing interference. In A-M sets the signal-to-noise ratio had to be of the order of 100 to 1 or more to obscure the noise. With the use of frequency modulation, however, it became possible to reduce considerably the strength of the signal required to override the noise. Although the signal at the receiver input must still possess greater power than the undesired voltage, the signal-to-noise ratio in F-M receivers need be only 2 to 1.¹

The ability of frequency modulation signals to override interference with low signal-to-noise ratios indirectly permits the use of many more stations within a given area than is possible with amplitude modulation. At the receiver, the desired signal must be only twice as strong as the undesired signal to hide the interference completely. Consider how much farther away an unwanted A-M broadcasting station can be and still ruin reception: with A-M a voltage only 1/100 as strong as the desired carrier will provide sufficient interference to be heard in the output of the receiver. Listen some night to an A-M receiver, especially at the high frequency end of the band. Signals, not only from nearby communities but from localities hundreds and

¹ In a properly designed F-M set it is possible for the signal-to-noise ratio to drop below 2 to 1 and still have the signal override any noise present. However, to be on the conservative side, the 2 to 1 ratio value will be retained.

thousands of miles distant, ride in to add their voice to a complete hodgepodge of whistles and signals, an intolerable situation for intelligible listening. Bring into play a 2 to 1 signal-to-noise ratio attainable with F-M, and with one quick sweep 99 per cent of the disturbance is eliminated. Now only signals from stations in the immediate vicinity are recognizable, and F.C.C. regulations do not permit simultaneous use of similar frequencies in the same service area.

It is quite popularly supposed that F-M is capable of providing greater fidelity of reproduction than A-M. This advantage is real, but it is an indirect benefit that results from F-M's ability to eliminate more of the interference and thereby provide a signal at the receiver that adheres more closely to the original signal at the studio. But greater fidelity is not an intrinsic, direct characteristic of the process of frequency modulation itself. Fidelity is a result of the correct design of the various stages that form either system of transmission and, since the same basic components are present in each system, both have fundamentally the same potentiality of good fidelity. Indirectly, we find other limitations, aside from the circuits themselves, that contribute to provide better fidelity with F-M. For example, with A-M only audio frequencies extending to 5000 cycles are transmitted, a limitation employed to prevent each station from occupying too great a frequency range within the limited A-M broadcast band and to permit additional stations to operate. F-M, arriving on the broadcast scene later, had the advantage of operating at the higher frequencies, where spectrum space is more plentiful. Add to this the already noted intrinsic tendency of F-M completely to override an interfering signal, and it is apparent why the misconception exists concerning fidelity. With the proper care in design, amplitude modulation can provide the listener with just as much fidelity as he obtains with F-M.

A-M versus F-M. In order to understand why F-M proves so effective in reducing interference, we must first make a thorough investigation of the properties of a frequency-modulated wave. Perhaps the easiest way of beginning is by a comparison of the effects on a carrier wave when it is amplitude-modulated and when it is frequency-modulated.

When a continuous, radio-frequency carrier wave is amplitude-modulated, the amplitude is varied in a manner determined by the audio-modulating voltage. Refer to Fig. 1.1 in which the high-frequency carrier and the low-frequency modulating wave are shown separately. By applying the carrier to the grid of a tube, and then varying the tube's plate voltage in accordance with the modulating voltage, we obtain at the output the amplitude-modulated wave shown in Fig. 1.1C. The variations imposed onto the carrier contain the intelligence of the audio wave. Now the A-M signal can be transmitted over the high-frequency circuits, through the air and to the receiver. At the receiver, the 2nd detector converts the amplitude variations

to their original form, in which they become intelligible at the loudspeaker.

A closer inspection of the modulated wave, Fig. 1.1C, reveals that there are certain limitations to the strength of the audio-modulating voltage. For example, it must not be too large. A small modulating voltage will affect slightly the amplitude of the carrier, and the resultant variations will be identical with those of the modulating wave. This means that no distortion, or loss of fidelity, has occurred. Increasing the strength of the modulating voltage causes correspondingly greater rises and indentations in the final signal waveform, but the essential form of the audio voltage is still preserved. However, when the applied modulating voltage becomes too great, the carrier becomes highly distorted. An example is shown in Fig. 1.2C. For the maximum modulation with no distortion the modulating voltage can only decrease the amplitude of the carrier to zero, momentarily, or raise it to twice the normal value at the peaks. We have now what is commonly known as 100 per cent modulation. Anything less is undermodulation; anything more is overmodulation. Of the two, un-

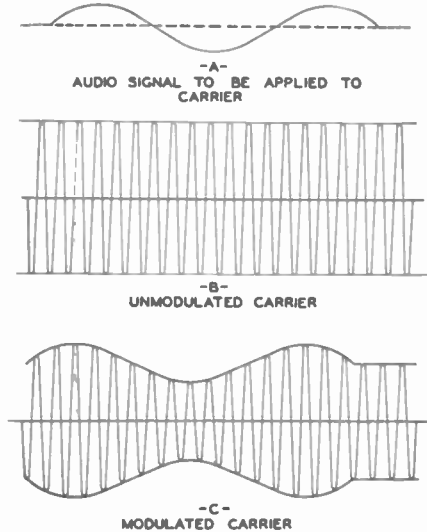


FIG. 1.1. The components of an amplitude-modulated wave.

dermodulation is to be preferred because, although it results in less power at the 2nd detector, it does not introduce any appreciable distortion. By distortion, we refer to the change in the shape of a wave from its original form.

The degree of modulation of a carrier wave is generally expressed as a percentage. It is the ratio of the peak amplitude of the modulating signal

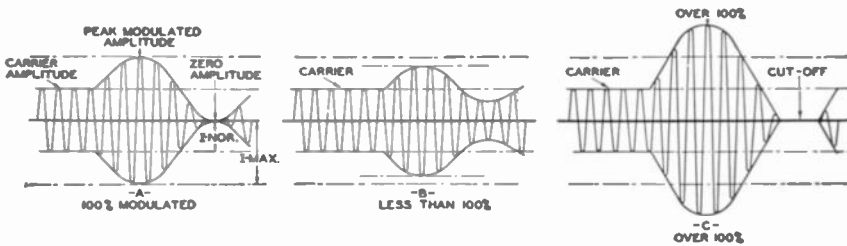


FIG. 1.2. Various degrees of amplitude modulation.

dermodulation is to be preferred because, although it results in less power at the 2nd detector, it does not introduce any appreciable distortion. By distortion, we refer to the change in the shape of a wave from its original form.

The degree of modulation of a carrier wave is generally expressed as a percentage. It is the ratio of the peak amplitude of the modulating signal

to the peak amplitude of the higher frequency carrier. The word "peak" is used because each of these waves are varying sinusoidally and one reference point is necessary. Mathematically, the ratio is expressed as

$$K = \frac{i_m}{I_m}$$

where i_m = peak value of the modulating signal
 I_m = peak value of the carrier

When this ratio is equal to 1, we have 100 per cent modulation wherein the amplitude of the modulated wave rises to twice the normal carrier value and decreases to zero when the two voltages act in opposition (see Fig. 1.2A).

The process of varying the amplitude of a continuous carrier wave causes the appearance of frequencies in addition to that of the carrier. Where a single, audio sine wave is used as the modulating voltage, two additional frequencies are generated, each separated from the carrier frequency by an amount equal to the audio-modulating frequency. As an illustration, consider the case of a 1000-kc carrier modulated by a 5-kc audio note. If an analysis is made of the amplitude-modulated wave at the transmitter output,

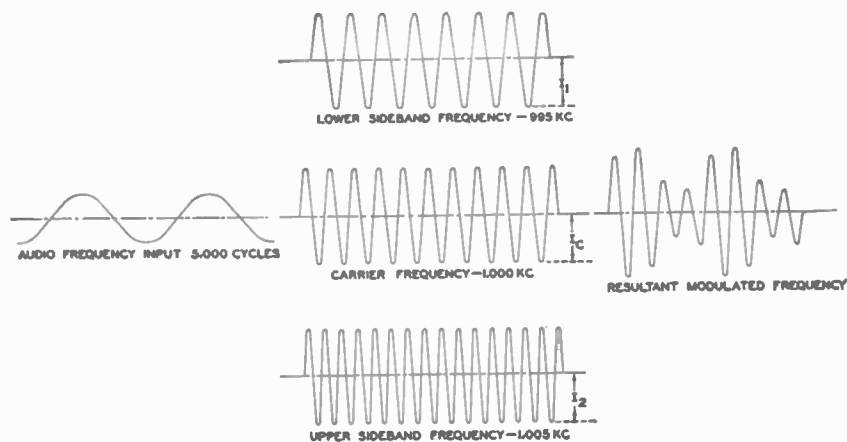


FIG. 1.3. The sidebands generated with amplitude modulation.

we will find that its components are the 1000-kc carrier, a frequency of 1005 kc, and one of 995 kc. Of the two additional frequencies produced, one is located above the carrier whereas the other is located the same frequency distance below the carrier. Together, these new frequencies are designated as sidebands. Graphically, they may be represented as shown in Fig. 1.3.

Now, obviously, the additional frequencies must in some way be related to the modulating voltage (which contains the desired intelligence), for

with the removal of this voltage we find that the sidebands are no longer present. Without modulation only the carrier remains. The moment we speak into a microphone, however, a modulating voltage appears and, coincidentally, the sidebands. A mathematical and experimental analysis reveals that all of the extra power contained in the modulating signal goes directly into the sidebands. The carrier itself is left untouched. Thus, although it may not be apparent from an inspection of Fig. 1.2, amplitude modulation gives rise to new frequencies, and these frequencies contain all the power and intelligence of the modulating voltage.

In all transmitting and receiving networks, provision must be made to include these sidebands. With a modulating frequency of 5000 cycles, or 5 kc, sidebands located ± 5 kc from the carrier are formed, making it necessary to receive a bandwidth of 10 kc (± 5 kc) wide. Anything less will partially or totally eliminate the sidebands, suppressing part or all of the intelligence. Although only one modulating frequency is mentioned, i.e., 5 kc, it is understood that any audio frequency may be used. Generally, the F.C.C. restricts the highest audio frequency to 5 kc, but this is only to limit the bandwidth of each station to 10 kc. Consequently, more stations per band are accommodated.

Frequency Modulation. In F-M we maintain the amplitude constant and alter the carrier frequency. In this instance, both the amplitude and phase of the carrier are fixed, while the frequency shifts back and forth about one central position in accordance with an audio-modulating voltage. This process is somewhat analogous to amplitude modulation, where the amplitude of the carrier is increased and decreased about the average, normal value.

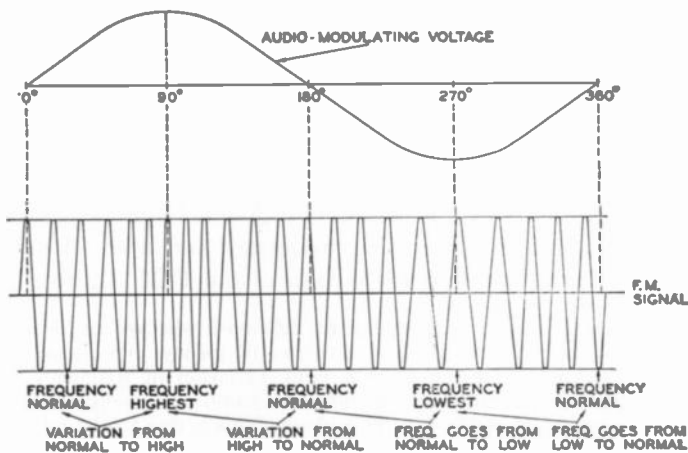


FIG. 1.4. Frequency variation in an F-M signal.

The physical appearance of a wave that has been frequency-modulated by a sine wave is shown in Fig. 1.4. At each instant throughout the application of the modulating signal, the frequency of the carrier will depend *directly* upon the amplitude of the audio signal. At the start of the audio cycle, the frequency of the carrier is slightly affected because within this region the modulating voltage is small. As the audio cycle approaches its positive peak, at 90° , the frequency variation of the carrier from its mean or resting position also reaches a maximum. Between 90° and 180° of the audio cycle, the voltage is returning to zero and the modulated carrier is shifting back, in step, to its resting position. Continuing on to 270° , we see that the audio voltage is again rising, although this time in a negative direction. The carrier frequency is likewise going in a negative direction, as indicated by a lower frequency. At 270° , the carrier frequency is at its lowest point. From here it slowly follows the audio voltage back to normal. This occurs at 360° , after which a new cycle begins.

To summarize, we see that the frequency variations of the carrier are in step, at all times, with the amplitude variations of the applied modulating voltage. As the amplitude of this latter voltage rises, the frequency of the carrier likewise increases,² whereas, when the audio signal reverses and moves to the opposite polarity, the carrier frequency decreases correspondingly.

The variation in carrier frequency is wholly dependent upon the *amplitude* of the acting audio voltage and not at all on its frequency. This is similar to A-M where the changes in carrier amplitude are influenced only by the amplitude of the audio signal. The frequency of the audio voltage, in both forms of modulation, determines the rapidity with which the amplitude or frequency changes take place but exerts no influence on the extent of these changes. A strong audio signal, at any frequency, will cause a wide frequency deviation from the center or resting frequency; a low or weak audio signal will cause less frequency variation. The frequency of each of these signals will determine only the speed with which the changes occur between points.

In the receiver the detector, or demodulator, must be capable of restoring the frequency variations to their corresponding audio-amplitude fluctuations. Demodulation, as its name suggests, is the reverse of modulation. In F-M receivers, some circuit is needed which will give a large output for a large frequency deviation from the resting frequency and a small output for a small frequency variation. This is the task of the F-M detector. How this

² As far as the ear is concerned, it makes no difference whether an increase in signal voltage results in an increase or decrease in carrier frequency. The important point is that a large modulating signal results in a large variation of frequency. At the receiver, the ear responds primarily to the different amplitude changes (and frequency changes) of the audio signal and not to their phase. This is due to the relative inability of the ear to respond to phase shifts.

job is accomplished will be described in detail in Chap. 9. For the present, of greater concern is what happens and this we have just learned.

It is simple to deduce that in F-M the signal will occupy a large bandwidth, since the primary purpose of the modulating audio signal is to vary the carrier frequency. In amplitude modulation, it will be recalled, the amplitude of the carrier was varied, and from this a frequency spread (sidebands) was created. A-M sidebands are not easily visualized and only by resort to mathematics or by direct experimentation can they be shown to exist. The maximum degree of amplitude modulation that can be applied to an A-M wave without producing distortion is limited to a value equal in amplitude to the carrier wave. This is defined as 100 per cent modulation. For the F-M system, the term "100 per cent modulation" is much more arbitrary and the limits to which a carrier can be frequency-modulated extends far beyond those presently utilized.

It is known, from theoretical considerations, that we can alter the frequency of a carrier only to the point where we do not, so to speak, obliterate it. Consider a carrier of 90 mc which is being modulated. We can shift the carrier frequency only until it is brought to zero; anything less is impossible since negative frequencies do not exist. Engineering limitations prevent us from introducing this large a shift, but even if it were possible, it would still be undesirable because of the limitation imposed by the number of commercial stations that could be licensed. Recognizing these difficulties and concerned with the problem of accommodating many stations in the available radio spectrum, the F.C.C. has decreed that the maximum deviation through which the carrier may be shifted shall be limited to ± 75 kc during modulation. This may be designated arbitrarily as 100 per cent modulation, although it must be remembered upon what limitations this value is founded.

As a comparison with A-M, it is to be noted that when a carrier is completely or 100 per cent modulated, its amplitude varies periodically between zero and twice the normal carrier value. Within this limitation, the modulation does not produce distortion. Any modulation greater than 100 per cent would definitely produce a distorted output, as shown in Fig. 1.2C. In F-M, we could, if we wished, shift the carrier far beyond the arbitrarily defined limit of 75 kc and still introduce no distortion. Again we see that the term "100 per cent modulation" does not apply to F-M in the same sense that it does to A-M. However, from certain considerations that will be developed later, the value of 75 kc is satisfactory. In addition, to diminish the effects of overlapping, an additional 25 kc is allotted as a means of protection between stations operating on adjacent frequencies. Each station, then, is assigned a bandwidth of 200 kc. Of this 200-kc bandwidth, 150 (± 75) kc is to be employed for the modulation and the remaining 50 (± 25) kc is to

function as a guardband. The very definite need for the guardbands will become more apparent after a closer study of the F-M sidebands.

F-M Sidebands. Before we approach the somewhat complex subject of the sideband frequencies generated in F-M, let us pause and quickly review the situation from the A-M angle. We have noted that sideband frequencies were not present until an audio signal was applied. The carrier wave independently has no sideband frequencies. Upon the application of a modulating voltage, say of one frequency, two sidebands were formed. Their frequencies were above and below the carrier by an amount equal to the modulating frequency. Throughout this process of modulation, the carrier power remained unchanged, the sidebands deriving their power from the audio signal.

Now let us look at the F-M situation. Upon the application of a sine-wave modulating signal, the carrier frequency is shifted back and forth, from a maximum position above the carrier frequency to a minimum position below. And in the process of shifting, additional sidebands are formed, with frequencies intermediate and beyond the maximum and minimum points reached by the carrier. In fact, theoretically, the sidebands are infinite, stretching away on either side of the carrier center position in equal number.

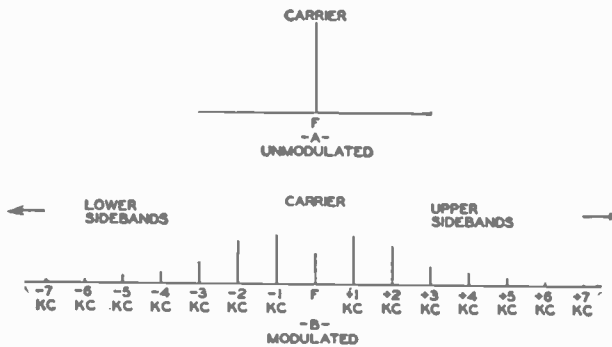


FIG. 15. An F-M signal with its sidebands developed as a result of audio modulation.

If we attempt to illustrate the sidebands with a diagram, we obtain the result shown in Fig. 15B. Fig. 15A represents the carrier before the modulation is applied. Fig. 15B reveals considerable information regarding F-M and helps to answer the confusing problem arising from the statement that infinite number of sidebands are present when a carrier is modulated by an audio signal. But, you may ask, if we have an infinite number of sidebands, how is it possible to have more than one station on the air, at any one time?

Upon examination of Fig. 15B, we can count seven sideband frequencies

below the carrier and seven sideband frequencies above the carrier, all due to the single modulating note. Each of these frequencies, it can be mathematically shown, is separated from its neighbor by an amount equal to the frequency of the modulating signal itself. Thus, if we whistle a 1000-cycle note into a microphone, a series of sidebands appear, above and below the central resting point, each 1 kc (1000 cycles) apart. With the carrier situated at 90 mc, there would be sidebands at 90.001 mc, 90.002 mc, 90.003 mc, etc.; also sidebands at 89.999 mc, 89,998 mc, 89.997 mc, etc. It is thus clearly indicated that the frequency of the modulating audio signal accomplishes two things:

1. Fixes the separation of the sidebands.
2. Determines the rapidity with which the sideband distribution changes.

Returning to Fig. 1.5B, we can count but fourteen sidebands in all: seven above the carrier, seven below. Apparently this is a misrepresentation, since a moment ago it was stated that there exists an infinite number of these sidebands. Actually, both statements are reconcilable; for one condition represents the theoretical condition, and the other the practical modification. With any strength of modulating voltage, we will find a certain number of sidebands generated which have sufficient power to be of value toward the reception of that signal. Beyond this, additional sidebands exist but contain *so little power* as to be of no practical importance in the formation of the signal at the receiver.

Here, then, is the reason why more than one station can operate at any time. The sidebands of the other stations are so weak beyond their band limits that, so far as interference is concerned, they are without effect. Included, too, is the answer for the establishment of guardbands about each station. Each station is limited to 75 kc on each side of its carrier frequency, but we know from theoretical considerations that additional sidebands exist. Hence, to be absolutely certain that no interference will be caused to adjacent stations, the 25-kc guardband (at each band end) is added. In A-M no such complications arise, and consequently there is no need for such precautions.

Sideband Power. Let us return again to a study of Fig. 1.5B. If we compare the amplitudes of the several sidebands with each other and with the carrier, we are at once struck with two facts: first, we find that the carrier power diminishes during modulation; second, it is possible for one or more sidebands actually to contain more power than the carrier.

The power that is taken from the carrier, during modulation, is distributed among the various sidebands. The louder the modulating signal, the greater the energy that will be taken from the carrier. In fact, it is perfectly possible for the carrier, during one of these modulation sweeps, to contain

no energy at all; the sidebands then possess it all. A moment's reflection should show that a transfer of some or all of the original carrier energy to the sidebands must occur, because the total frequency-modulated signal does not vary in amplitude. Thus, the only way to satisfy this condition, during modulation, is to transfer part (or all) of the carrier energy to the sidebands. This power transfer is a characteristic of frequency modulation.

The sum of the amplitudes of the sidebands plus that of the carrier gives the value of the unmodulated carrier, provided the addition is performed vectorially. This has special significance because not all the sideband amplitudes have the same polarity at one time. Counting from the carrier, in either direction, we find that all the odd-numbered sidebands have opposite polarity. The even-numbered ones have the same polarity. But, in any case, correspondingly placed sidebands (with respect to the carrier) have the same amplitude. A simple example will make this clear.

With a carrier situated at 90 mc and a modulating frequency of 1000 cycles (or 1 kc), we have sidebands at 90.001 mc, 90.002 mc, 90.003 mc, etc.; also sidebands at 89.999 mc, 89.998 mc, 89.997 mc, etc. Counting away from the carrier, in the first set of sidebands, at 90.001 mc and 89.999 mc, both frequencies would possess equal amplitudes but would be of opposite polarity. The next or second set of sidebands would be at 90.002 mc and 89.998 mc; these frequencies also would be equal in amplitude and would have the same sign. Of course, although each set would have the same amplitude, different sets would differ from one another.

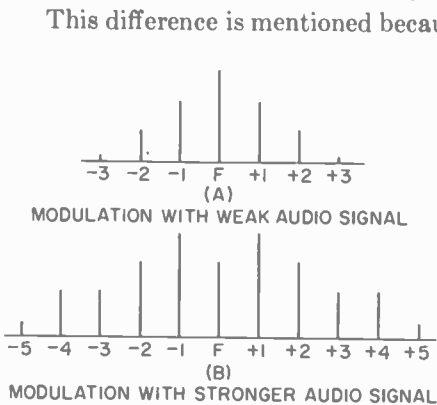


FIG. 1.6. A comparison of the frequency spread of two F-M signals modulated by audio voltages of differing intensity.

This difference is mentioned because the arithmetic addition of the amplitude of the different sidebands of Fig. 1.5B, together with that of the carrier, will not give the same value as the unmodulated carrier. If the diagram of Fig. 1.5B were drawn up with strict regard for polarity, some of the sidebands would have to be drawn below the reference line. That they were not is due to the fact that, as far as this discussion extends, it makes no essential difference. We are interested only in the frequencies generated by modulation and in their distribution.

Sideband Variation. It would be instructive to see the effect upon sideband distribution when we vary the amplitude of the modulating voltage. This is shown in Fig. 1.6. It is apparent that, as we increase the intensity of the audio signal, the number of significant sidebands increases. The sidebands of the low-intensity modu-

lating voltage that were considered negligible now gain enough power to make them significant. The energy is spreading out, creating more sidebands of appreciable power and thereby causing a greater overall frequency spread about the carrier. This is what we mean when we say that the louder the modulating signal, the greater our carrier will deviate from its mean position. At low modulating intensities, the important sidebands are clustered about the carrier. As we raise the volume, the energy spreads out and more sidebands must be included.

It is now possible to replace the elementary explanation regarding frequency spread given at first with the more accurate discussion given above. We see that the shifting of the energy of an F-M wave is directly related to the intensity of the applied audio signal. When a shift takes place, say away from the carrier, the former energy distribution is replaced by this newer arrangement, with every sideband and the carrier affected. Thus, energy is taken by some and given up by others. The *total* energy, however, under all conditions, remains constant.

Modulation Index. With an appreciation of the relationships between modulating signal strength and sideband formation, we can more easily approach the problem of designing networks to receive and pass the necessary band of frequencies. It is quite evident that if we limit the amplitude of the modulating signals we control the spread of significant sidebands and thus prevent any interference to adjacent channels. The limit set by the F.C.C. is plus and minus 75 kc about the frequency of the carrier. Steps must be taken, at the transmitter, to see that these limits are not exceeded. But is the strength of the audio signal the only determining factor in bandwidth spread? Or must we add another factor? Let us see.

It has been noted that, whenever sidebands are formed, they are spaced by an amount equal to the frequency of the modulating audio signal. Thus, for an audio voltage with a frequency of 100 cycles, successive sidebands are created 100 cycles apart, whereas, for a frequency of 15,000 cycles, the sidebands are spaced at intervals of 15 kc. However, if the two signals are causing 10 important sidebands about the carrier, the 100-cycle note would result in an F-M signal with a spread of 10×100 or 1000 cycles (1 kc) and a *total* bandwidth of 2 kc. With the 15-kc note, the total frequency spectrum would extend 300 kc. The obvious conclusion from this illustration is that the frequency bandwidth required by a frequency-modulated carrier depends upon two factors:

1. The intensity of the applied modulating voltage.
2. The frequency of this voltage.

Recognition of the importance of both these considerations has led to the formation of the modulation index, which is defined as:

deviation of F-M carrier

 audio frequency causing this deviation

where the deviation of the F-M carrier is a direct result of the intensity of the applied signal.

We note from this formula that, if the frequency of a signal is kept constant, the bandsread about the carrier will depend directly upon the strength of the modulating voltage; whereas if the sound amplitude is kept constant, the lower audio frequencies will produce more sidebands. In practice, however, neither one is kept constant and both vary simultaneously.

The spread of any F-M signal will depend upon the modulation index. At the present time, the F.C.C. has specified that the maximum shift of an F-M carrier should be limited to 75 kc. The highest audio modulating frequency in use is 15,000 cycles. The ratio of these two maximum quantities, according to the modulation index, is 75 kc/15 kc, or 5. This figure, besides being known as the modulation index, is also assigned the special name of deviation ratio. The reason for designating this one case as the deviation ratio is that any other combination of audio frequency to carrier shift less than the maximum values will always produce a narrower signal spread. This will be seen as we progress.

Index Table. Taken by itself, the index figure does not directly tell us anything. But, when applied to appropriate mathematical tables derived from Bessel functions, we can obtain the number of significant sidebands that will be formed at this particular value of modulation index.

A knowledge of Bessel functions is not specifically required in order to use the tables derived from these equations. Because of the complexity of Bessel equations, it is convenient merely to accept the tables without proof. A modified version sufficient for our discussion is given in Table I. Full inclusion of all possible values would fill many pages and add very little to what we can learn from the condensed version of Table I.

In Table I there is a list of the number of significant sidebands that would be obtained for some of the common values of the modulation index. Thus, with an index of 5, there are 8 important sidebands (on either side of the carrier) formed; with an index of 10, the number of sidebands increase to 14. For each value of index a certain number of significant sidebands are formed. In general, both vary directly. A large index results in a large number of sidebands; a small index, a smaller number of sidebands. It is interesting to note that when the index becomes very small, of the order of 0.4 or less, only 2 sidebands are formed, similar to A-M operation when one modulating frequency is employed.

To compute the bandwidth required for each modulation index, simply multiply the frequency of the audio-modulating signal by twice the number

TABLE I

Modulation Index	Number of Significant Sidebands		Bandwidth Required (f = frequency of audio signal)
	Above Carrier	Below Carrier	
0.1	1	1	$2f$
.02	1	1	$2f$
.03	1	1	$2f$
.04	1	1	$2f$
.05	1	1	$2f$
0.1	1	1	$2f$
.2	1	1	$2f$
.3	1	1	$2f$
.4	1	1	$2f$
.5	2	2	$4f$
1.0	3	3	$6f$
2.0	4	4	$8f$
3.0	6	6	$12f$
4.0	7	7	$14f$
5.0	8	8	$16f$
6.0	9	9	$18f$
7.0	10	10	$20f$
8.0	12	12	$24f$
9.0	13	13	$26f$
10.0	14	14	$28f$
11.0	16	16	$32f$
12.0	17	17	$34f$
13.0	18	18	$36f$
14.0	19	19	$38f$
15.0	20	20	$40f$

of sidebands given for the modulation index. Thus, suppose we have a modulation index of 5 when the audio frequency is 5000 cycles, or 5 kc. From Table I, a modulation index of 5 gives us 8 important significant sidebands, each, in this case, to be situated 5 kc apart; thus, $5 \text{ kc} \times 8 = 40 \text{ kc}$ on each side of the carrier, or a total bandsread of 80 kc.

It is always completely puzzling to anyone approaching F-M for the first time to learn that, although the carrier frequency in the F-M transmitter is never actually shifted beyond the $\pm 75 \text{ kc}$ limits specified by the F.C.C., sidebands do appear beyond these limits. As an illustration, consider the case when the modulation index is 5, obtained by shifting the carrier $\pm 75 \text{ kc}$ with a 15,000 cycle audio note. From Table I we see that 8 significant sidebands will be obtained. This, of course, means 8 above and 8 below the carrier position. Since each sideband is separated from its neighbor by 15,000 cycles, or 15 kc, we find that the total required bandwidth is

$$2 \times 15 \text{ kc} \times 8 = 240 \text{ kc}$$

Thus we obtain a signal extending over 240 kc (± 120 kc) when the carrier frequency is shifted only ± 75 kc in the transmitter during the process of modulation.

To explain this situation physically, we might use the analogy of a man moving his finger back and forth at the center of a small pool of water. We know from actual experience that, although the man may move his finger only slightly back and forth, water ripples will appear far beyond this little area. The greater the distance covered by the man's moving finger, the greater the spread of the water ripples. In F-M, the greater the carrier swing, the greater the number of sidebands obtained.

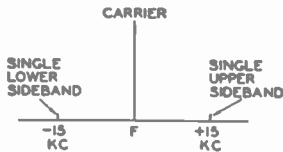


FIG. 1.7. The sideband distribution when the modulation index is small, 0.2 or less.

In commercial practice, it seldom happens that a 15-kc note will have sufficient volume to spread the transmitter carrier to the 75-kc limits. As the frequency of the signal is lowered, the number of sidebands that extend beyond the ± 75 -kc limits also decreases, until at 50 cycles a full carrier swing will just produce sidebands up to, but not beyond, the 75-kc limits. Here, then, is the reason for the additional ± 25 -kc guardbands. They tend to absorb any sidebands that

extend outside the 75-kc limits.

Many incorrect interpretations have been placed upon the F.C.C. regulation that, during modulation, the carrier is not to be shifted more than ± 75 kc. This does not mean, for example, that the end significant sideband must not extend beyond ± 75 kc. It merely means that during the process of modulation, when the carrier frequency is being altered by the applied audio voltage, the maximum shift be no greater than ± 75 kc. However, when the full 75-kc shift is utilized, it will generally be found that significant sidebands exist beyond the F.C.C. limits. This is perfectly permissible.

When the carrier swing is very small, 0.4 or less for the modulation index, then Table I reveals that only 2 sidebands (one above and one below the carrier) are obtained. If the audio-modulating signal is 15 kc, then a modulation index of 0.2 means that the carrier frequency is only shifted 3 kc back and forth.

$$0.2 = \frac{\Delta F}{f}$$

$$\Delta F = 15 \text{ kc} \times 0.2$$

$$\Delta F = 3 \text{ kc}$$

The frequency distribution will appear as shown in Fig. 1.7, with one carrier and two significant sidebands.

Note again that the sidebands will be spread farther from the center than the frequency shift that produced them. The sidebands are ± 15 kc from the center, whereas the carrier was shifted only ± 3 kc. This situation will be generally true of all F-M transmissions. However, if the reader remembers the water analogy, it may help him to understand this new idea.

We have, up to this point, a fairly complete picture of F-M, both before and during modulation. Of remaining interest is phase modulation, not only because it represents a very possible method of sending intelligence but, more important from our viewpoint, because phase modulation indirectly results in F-M. The latter aspect is all the more important because interference affects an F-M signal through its phase disturbances. Moreover, phase shifting is employed in commercial transmitters to produce F-M. To appreciate either of these two aspects, a brief discussion of phase modulation is necessary.

PROBLEMS

1. What effect does the choice of the high frequencies to transmit F-M signals have on the ability of F-M to override interference? What would be its effect if A-M was used at the same frequencies?

2. What is the significance of 100 per cent modulation when applied to A-M broadcasting? Explain its counterpart in F-M.

3. In an A-M wave, where is the intelligence contained? How does this differ from the conditions prevailing in frequency modulation?

4. What is the relationship between the sidebands produced in amplitude modulation and the audio-modulating frequency?

5. Is the carrier frequency actually shifted when a wave is frequency-modulated? What is the F.C.C. regulation in this regard?

6. What is a discriminator? What is the accepted application of this word?

7. What are guardbands and why are they required in F-M transmission?

8. Discuss the difference between the sidebands produced through amplitude modulation and those produced by frequency modulation.

9. What do we mean when we refer to a sideband as being "significant"?

10. What influence does the audio-modulating signal frequency have in the production of an F-M signal?

11. Where do the F-M sidebands obtain their power? Contrast this with the situation existing in an A-M signal.

12. Explain how the sidebands vary, both in power and number, during different portions of an audio-modulating cycle.

13. The F.C.C. specifically limits carrier shifting to ± 75 kc. Can sidebands extend beyond these limits? Explain, using a numerical example.

14. Explain the importance of the modulation index.

15. With a modulation index of 3 and a modulating frequency of 5 kc, compute the complete bandwidth required when a 90-mc carrier is employed.

16. Analysis of an F-M signal revealed that the carrier frequency was 94 mc, and the upper and lower sideband limits were at ± 98 kc. Further analysis indicated

that the modulating frequency was 14 kc. What modulation index was being employed at the transmitter?

17. Under what conditions are the bandwidths of F-M and A-M signals exactly equal to each other?

18. The F.C.C. specifies that the modulation index, with a 15,000-cycle modulating frequency, shall not exceed 5. Assuming that the amplitude of the modulating signal remains constant, what happens to the modulation index when the audio frequency is reduced? Does the bandwidth increase or decrease as we lower the audio frequency? Illustrate your answer by computing the bandwidth required for audio-modulating frequencies of 12 kc, 9.375 kc, and 7.5 kc.

19. A sideband is considered insignificant when it contains less than 1 per cent of the total carrier power. Express this in terms of db.

20. Does a signal with a large modulation index necessarily require a wider bandwidth than a second signal having a smaller index? Explain your answer.

Chapter 2

F-M FROM PHASE MODULATION

Phase Modulation. It has been demonstrated that intelligence can be imposed onto a carrier by variation of its amplitude or its frequency. The only remaining electrical property is the phase of the carrier. By suitably shifting the phase of the wave in response to an applied audio-modulating voltage, we can accomplish the same transfer of intelligence as we have seen occur in the previous two methods of modulation.

A comparison is perhaps the simplest way of approaching phase modulation. Consider two carriers that are being transmitted at the same time, both of equal frequency but differing in phase by 45° . The phase difference merely indicates that when the voltage of one wave reaches its peak, at any one instant, the other carrier is 45° from its peak value. Fig. 2.1 illustrates just such a situation. In all other respects the signals are equivalent. If nothing is done to alter this phase relationship between the two carriers, nothing will be heard at the output of a phase-modulation receiver.



FIG. 2.1. Two carrier waves differing by 45° .

But suppose the phase relationship is altered in a definite rhythm or pattern—perhaps in accordance with the amplitude and frequency of an applied sine wave. Let these audio variations be applied to the carrier that lagged by 45° . The other carrier will be kept fixed. Then the phase of the carrier to which the modulation is applied might swing, under the action of the audio signal, from 45° lag to 60° lag (see Fig. 2.2). The 60° lag would occur when the audio-modulating signal was at its positive peak. As the audio voltage decreased from this peak toward zero, the carrier would follow suit by slowly returning to 45° lag, its normal position. On the negative half cycle of the audio signal, the phase lag of the carrier shifts in the opposite direction, or toward a value less than 45° . This is opposite to its increase in

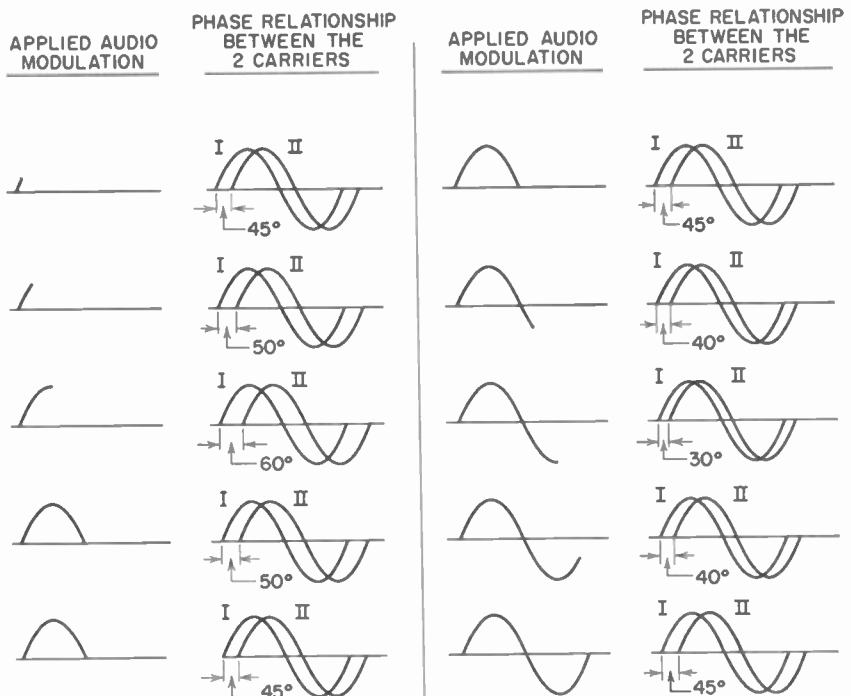


FIG. 2.2. A simplified illustration of phase modulation. Carrier I remains fixed while the phase angle between it and carrier II varies.

phase when the audio signal was on the positive half cycle. At the negative audio peak the phase of the carrier would be brought to 30° . Note that this point is just as far below the normal value of 45° as 60° is above. From the cycle's maximum negative peak back to zero the phase returns, in step, to 45° .

This phase shift represents one complete cycle for the audio-modulating voltage. It would occur over and over again as long as the modulating signal was active. At the proper receiver, the phase variations would be converted to the corresponding audio sound much the same way as an F-M receiver converts frequency modulation into sound. Instead of A-M or F-M we are dealing with P-M, or phase modulation.

Factors Affecting Phase Shift. The rapidity of phase shift from one value to another is determined by the modulating signal frequency. However, the number of degrees through which the carrier is shifted during modulation is dependent upon the strength of the audio signal. A weak signal, for example, might shift the phase only to 50° at its positive peak and 40° at its negative peak. A stronger signal at the microphone could, at its peak, cause a considerably larger phase shift. The receiver, working on phase dif-

ferences, would note the different shifts and respond accordingly. This relation of phase shift to strength of the audio signal is entirely analogous to F-M, where a strong audio signal is needed for a wide frequency shift and a weak modulating signal produces a small frequency shift. At the F-M receiver, the discriminator is sensitive to frequency differences and derives the appropriate audio variations from the carrier. In both F-M and P-M, the total power remains fixed throughout the entire sequence of modulation. There are other points of similarity between the two systems, with the same mathematical design formula applicable to both.

The foregoing illustration is useful in introducing the basic concepts of phase modulation, but it does not clearly indicate how frequency modulation results. For this, a vector approach can be better employed.

Elementary Vectors. A vector is a shorthand method of illustrating certain facts much as numbers represent a shorthand method of expression. When we write the number 15 to designate a group of objects, we mean 15 individual pieces or parts or units. The number 15 saves us the trouble of drawing 15 squares, or circles, or other geometrical figures to indicate the total of such units.

There are certain types of physical quantities which possess not only magnitude but also direction. The magnitude may be indicated by a number, but what about the direction? For this we require an arrow. Thus a force of 100 pounds produced by a tractor pulling a block of wood, for example, in one direction could be illustrated by a number and an arrow, as shown in Fig. 2.3A.

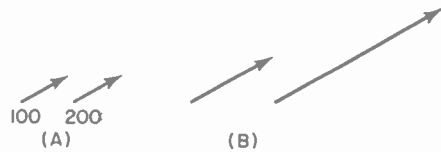


FIG. 2.3. Vectors represent definite forces and the directions in which they act.

A simpler method, however, is to have the length of the arrow indicate the

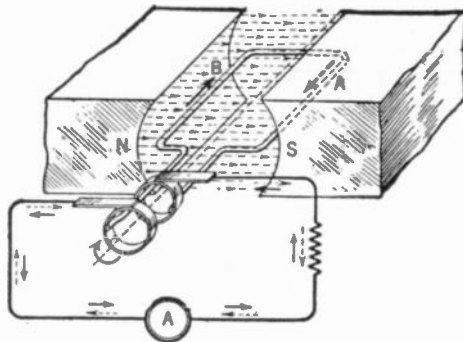


FIG. 2.4. The basic arrangement for the generation of an alternating voltage.

magnitude of the force, and the direction in which the arrow is drawn denote the direction in which the force is acting. If the arrow is made $\frac{1}{2}$ inch long for 100 pounds of force, then an arrow 1 inch long would represent a force of 200 pounds (see Fig. 2.3B).

Vectors Applied to Radio. Of specific interest is the application of vectors to radio or, more specifically, to electric waves. We know, from our first lessons in radio, that an alternating current is produced whenever a coil is rotated in the magnetic field between two magnets (see Fig. 2.4). If the field is uniform, the voltage produced across the output terminals of the coil is a sine wave. In Fig. 2.5 we see that the voltage generated depends upon

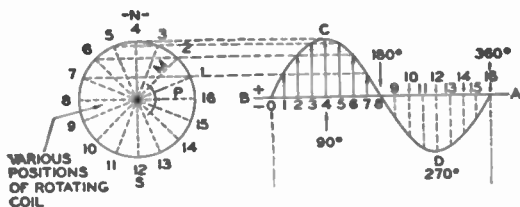


FIG. 2.5. The variation in a-c voltage with coil position.

the position of the coil in the magnetic field. This may be expressed mathematically by the formula

$$e = E_{\max.} \sin \theta \quad (1)$$

where

e = instantaneous value of the voltage developed when the coil is at some angle θ

$E_{\max.}$ = maximum value of the voltage

θ = angular position of the coil

Since the position of the coil determines the amount of voltage generated,

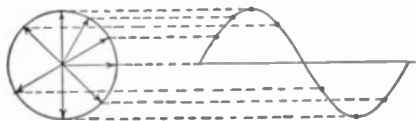


FIG. 2.6. Substitution of vectors in place of the rotating coil. The arrowed vectors indicate the coil position at various instants.

suppose we replace the coil by a vector, as shown in Fig. 2.6. The value of the voltage, at each instant, is then given by the vertical height of the vector, this vertical being proportional to the sine of the angle θ . When the vector is placed horizontally, the angle θ is zero. This makes $\sin \theta$ equal to zero (see Fig. 2.7A). From formula (1),

$e = 0$ when $\theta = 0$ and no voltage is generated. At some other angle, say 30° , the voltage generated is

$$e = E_{\max.} \sin 30^\circ.$$

Since $\sin 30^\circ = \frac{1}{2}$, then $e = \frac{1}{2}E_{\max}$. This is shown with the vector placed at an angle of 30° with the horizontal (Fig. 2.7B). A vertical line dropped

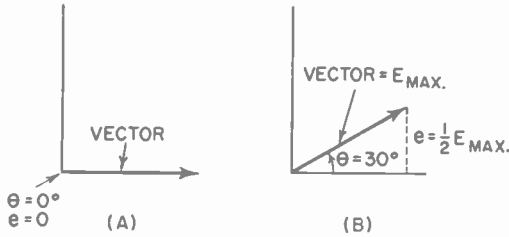


FIG. 2.7. The value of the generated voltage depends upon sine of the angle, θ .

from the head of the arrow will then be $\frac{1}{2}$ as long as the length of the vector. If the length of the vector is made equal to the value of E_{\max} , say 100 volts, then the length of the vertical will represent 50 volts. Thus, the vector is made equal to the peak voltage of the sine wave; a vertical line dropped from the head of the arrow to the horizontal axis represents the voltage generated when the coil is in the position of the vector.

The rapidity with which the vector rotates or makes one complete revolution is the same as the frequency of the a-c wave it represents. If the frequency of the wave is 10,000 cps, then the vector will rotate 10,000 times in one second.

Two waves acting at the same time can be represented by two vectors, as in Fig. 2.8. Any phase difference between the waves is indicated by the relative position of their vectors. Two vectors displaced by 45° represent two electrical waves 45° apart in phase. Here again we note the simpler notation made possible by the use of vectors.

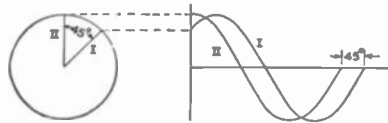


FIG. 2.8. A comparison between the vector and the conventional methods of illustrating a 45° phase difference between two waves.

$$\begin{array}{c} \vec{V}_1 \\ \text{PLUS} \\ \vec{V}_2 \end{array} = \vec{V}_1 + \vec{V}_2 = \vec{V}_1 + \vec{V}_2$$

FIG. 2.9. The addition of two vectors that are in phase, i.e., point in the same direction.

Many times, when two waves are present in the same circuit, we wish to ascertain the total effect and not merely the individual results. If the voltages are in phase, given by two parallel vectors, they may be added directly (see Fig. 2.9). However, voltages generally differ in phase and direct addi-

tion is no longer possible. In Fig. 2.10 two vectors, OA and OB , represent two voltages that differ in phase. The resultant of the two is obtained

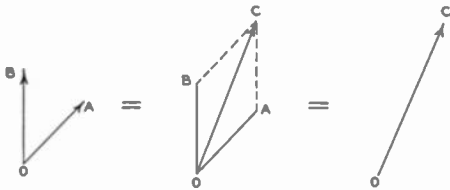


FIG. 2.10. The addition of two out-of-phase vectors.

by constructing a parallelogram, $OBCA$. Line BC is parallel and equal to OA ; line AC is parallel and equal to OB . The diagonal drawn between points O and C is the resultant of the vectors OA and OB .

Another interpretation of the resultant is to consider vector OA as forcing a particle to move from O to A , while vector OB forces it to move from A to C . Hence, if two forces, OA and OB , were applied simultaneously to the particle placed at point O , the particle would be forced to travel along line OC .

Vectors Applied to Modulation. The vector concept lends itself quite readily to the three types of modulation. For A-M, in Fig. 2.11, the carrier

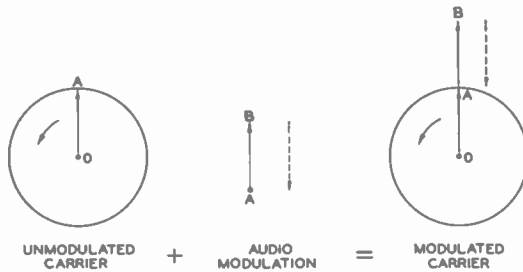


FIG. 2.11. Amplitude modulation demonstrated by vectors.

vector OA is rotating at a frequency determined by the oscillator in the transmitter. Let this be 1000 kc. In one second, therefore, there are 1,000,000 complete revolutions producing 1,000,000 cycles. When audio modulation is added, represented by vector AB , the length of the overall vector OB will depend on how AB varies from moment to moment. When the audio voltage is at its peak, AB attains its greatest length; for weaker modulating signals, AB becomes small.

The complete picture shows OA , the carrier vector, rotating rapidly at the radio-frequency rate while the length of $OA + AB$ varies slowly as the strength of the audio signal increases and decreases. This is evident, too, in Fig. 2.12, where one audio cycle covers many carrier cycles. Since in Fig. 2.11 AB is governed by the audio-modulating voltage it will either add to OA or be out of phase (oppose OA) by 180° . Again, reference to Fig. 2.12

will show the same thing to be true there. Note that AB can either add to OA or subtract from it, but nothing more, since this is amplitude modulation and only the amplitude of the vector OA can change.

In F-M, the length of the vector OA remains constant while the speed (frequency) at which it rotates varies from instant to instant according to a pattern set by the audio modulation. Thus, at one moment, it may be spinning around faster than normal; at the next, back to normal; a second later, slower than normal, and so on. These changes represent our frequency variation or modulation and can be detected by the proper receiver.

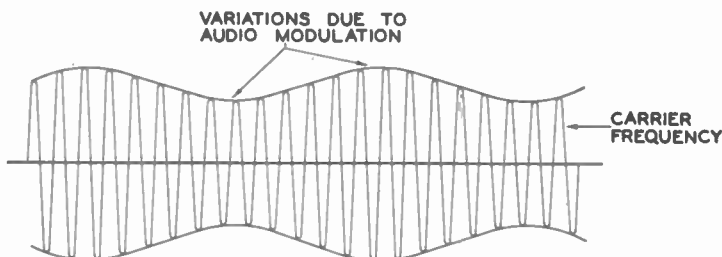


FIG. 2.12. The customary method of indicating an amplitude-modulated wave.

Now we come to phase modulation where we operate on the relative phase of a carrier, leaving the amplitude and frequency fixed. However, as we shall soon see, varying the phase of a wave results, indirectly, in frequency modulation and this can be detected by an F-M discriminator.

F-M from P-M. In the vector diagram illustrating phase modulation, Fig. 2.13, OA , as before, is the carrier to be phase modulated. It rotates at the frequency of the carrier. OB , another vector, is separated from OA by 45° . This vector is shown for reference only, and will aid in visualizing the phase changes of OA . Although it is not necessary to include OB in each diagram, it may help to understand the phase changes of OA by considering OB as always present.

The frequency and amplitude of OA are kept fixed. If now we vary the phase of OA (with respect to OB) then, as OA rotates, it will "wobble," or fluctuate back and forth, about its mean position.

Suppose that at the instant the audio modulation is applied, the vector OA is at point 1 in Fig. 2.13. Due to the acting modulation, its phase (or position) is shifted to point 2. From position 2, the audio signal brings the vector back to position 1. From position 1 the vector goes to position 3, due, this time, to the opposite half cycle of the audio wave. To complete the cycle, the vector goes back to position 1, the mean point. Thus, under the force of the applied audio voltage, the phase of OA changes from position 1 to position 2, back to 1, then to position 3, and finally back to the starting

point, position 1. This represents one complete cycle and recurs at the audio frequency.

Now let us see what is necessary when we shift the phase of a wave. To advance the wave position where it is slightly ahead of its normal position, such as shifting it from position 1 to position 2, we must momentarily cause it to speed up for if it had continued moving at its normal speed, it could

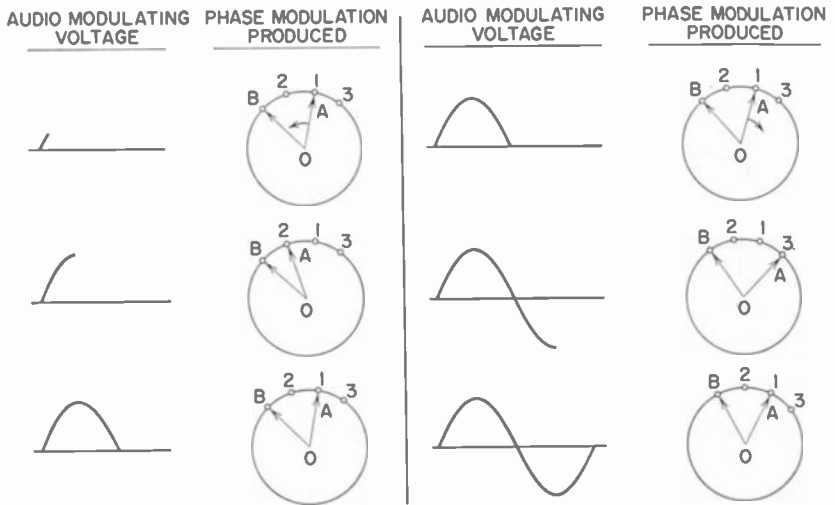


FIG. 2.13. The phase variation of a carrier that is being phase-modulated by an audio signal.

never have reached the advanced position. From point 2, back to point 1, necessitates a gradual lessening in the speed of rotation. From the normal position to position 3 means a still further decrease in speed on the part of the rotating vector. Finally, to complete the cycle, the vector is brought back to its original position at 1 by a corresponding speed increase.

For the sake of simplicity, the vector OA is stopped while the various positions assumed due to phase modulation are shown. Actually OA is revolving at a tremendous rate while the phase varies, relatively, very slowly. But the overall frequency change due to these phase variations still takes place. This is the important point. Each time the carrier vector wobbles back and forth to reach the new phase positions dictated by the audio modulation, we find the frequency also changes in order to have the vector reach the new positions.

In Fig. 2.13, points 2 and 3 were the maximum phase shifts of the carrier vector OA . If the amplitude of the driving audio voltage increases, for example, then the phase shift will increase and OA will have to travel through

a greater angle. Conversely, a smaller audio voltage results in a smaller shift. As we shall soon see, the size of the shift has a very direct bearing on the amount of indirect F-M generated.

The *average* frequency of the rotating vector OA , throughout the entire process, remains constant. This must be so, of course, since we are not, directly, varying the frequency of the wave. But at those instants when the phase of OA is changing, its instantaneous frequency is either increasing or decreasing. If this were not so, then how could OA ever reach the different phase positions rotating at its fixed rate? Obviously it could not. Note that what we are doing is adding sufficient change, either positive or negative, to a fixed frequency to permit the carrier to reach the desired phase position. Distinguish this from "pure" F-M where the carrier frequency itself is being directly affected and shifted in response to the modulating voltage. In the end, both may result in frequency modulation, but it is important to realize that a difference does exist.

Factors Affecting Indirect F-M. The amount of indirect F-M that is formed by phase modulation depends upon two factors. First, there is the extent of the phase shift. Quite naturally, the larger this phase shift, the more the carrier frequency must be altered to meet the new condition. The two are directly proportional to each other. Second, there is the frequency of the modulating audio signal. If the audio frequency is high, the shifting from one position to another will have to take place in less time, which means that, in any one instant, a greater carrier frequency change must occur. The result—more frequency modulation. On the other hand, if the same phase shift occurs at a lower audio frequency, the rapidity with which the carrier must change decreases. Consequently, the equivalent carrier-frequency change decreases. As with phase shift, the extent of the indirect F-M formed varies *directly* with the audio-modulating frequency.

Because of the importance of the concept of forming frequency modulation from phase changes, it may prove helpful if we consider the same problem from still another angle.

When the phase of a carrier is shifted, so that it differs from another constant carrier by more than it did previously, it means that each wave must go through its cycle just a trifle faster in order to attain the necessary shift within the allotted time. Consider the two waves, A and B , shown in Fig. 2.14. Wave A is untouched by the modulating voltage and completes its cycles with unvarying regularity. Wave B , however, is subjected to a modulating voltage which causes it to shift in phase, relative to wave A , by 30° . The full shift must be completed by the time position C is reached.

In order to attain this variation, each of the four cycles shown must contribute a portion of this shift, all contributions adding to the necessary 30° . But for each wave to assume an additional amount means that, besides the

regular 360° , each must add a little extra. This can only be accomplished if each wave swings through its cycle of variations a trifle faster than it normally would. By the time point C is reached each cycle has added sufficient phase variation to fulfill the required 30° . Henceforth, with no further variation, wave B will reach its maximum positions (and minimum positions) 30° ahead of wave A ; its frequency will continue at the same rate as that of wave A , exactly as it existed prior to the modulation. It is only *during* the shifting in phase that the frequency variations appear and, coincidentally, the indirect F-M. When the shifting ceases, indirect F-M disappears.

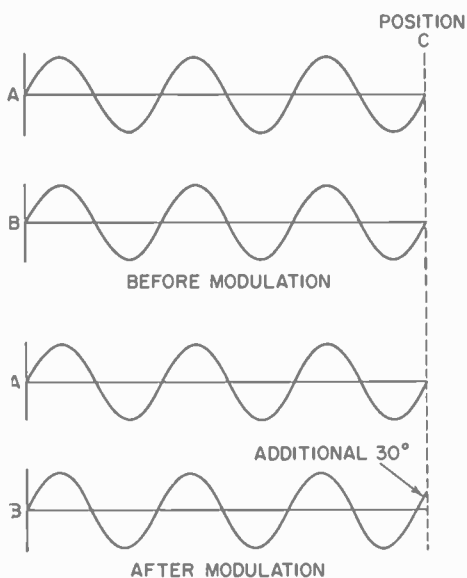


FIG. 2.14. Prior to modulation, carriers A and B are in phase. After modulation, B leads A by 30° .

To demonstrate how the degree of phase shift and the rapidity with which it occurs affect the indirect F-M, consider the following two illustrations. A carrier of 90 mc (90,000,000 cps) is to be shifted ahead in phase to a maximum of 50° , this to be accomplished in $1/1000$ of a second, as determined by the audio-modulating frequency. In $1/1000$ of a second, 90,000 cycles of the carrier will occur. To complete a 50° shift in this interval, each carrier cycle must contribute an additional $50/90,000$ degree, or approximately 0.0005° . To achieve this, the frequency of each wave must be increased slightly. This increase produces the indirect F-M. The same line of reasoning would apply if the wave were to drop 50° behind, except that now each wave loses 0.0005° from its cycle. The result—a decrease in frequency.

If we maintain the phase shift at 50° , but decrease the time in which this change must be attained, then we have the equivalent of an increase in audio-modulating frequency. Assume the completion of the phase shift is to be accomplished in $1/10,000$ of a second. In this time interval, the number of carrier cycles occurring will be reduced from the previous 90,000 to 9000. For the 50° shift, we calculate that $50/9000$ or approximately 0.005° are needed from each cycle compared to the 0.0005° given above. The larger the shift of each wave, the greater its frequency change.

Expressed in mathematical terminology, the extent of the indirect F-M formed through phase modulation is given by

$$\Delta F \text{ (frequency shift)} = f \times \Delta\theta \sin \omega t$$

where

f = the frequency of the audio-modulating voltage (or any other voltage) that is causing the phase shift

$\Delta\theta$ = the maximum angle through which the carrier is shifted

$\sin \omega t$ = the sinusoidal variation of the applied audio voltage

The inclusion of the term "sin ωt " indicates, mathematically, that the audio voltage is a sine wave and, hence, its force varies sinusoidally. As a result, the phase shift follows suit. Since all complex waves can be resolved into pure sine waves, the expression is applicable to all types of electrical waves employed in communications.

A comparison of the foregoing equation with true or direct F-M reveals the differences between them. In direct F-M the value of the carrier itself swings between its maximum limits, say 75 kc, about the central or resting position. The carrier is actually being shifted directly by the modulation. In indirect F-M, produced by the phase modulation, the carrier is not actually shifted directly by the modulation. Rather, the effect of the phase shifts is to either add or subtract frequency variations to the fixed carrier.

In the next chapter we will use these facts to demonstrate why F-M is so effective in reducing interference.

PROBLEMS

1. Differentiate between phase modulation, amplitude modulation, and frequency modulation.
2. Name and explain the factors which affect the phase shift of a phase-modulated wave.
3. Why are vectors useful in describing phase modulation? Could they be used to illustrate amplitude and frequency modulation?
4. Illustrate vector addition.
5. How do we obtain F-M from P-M?
6. Distinguish between indirect and direct frequency modulation.
7. List the factors which affect the amount of indirect F-M produced from a phase-modulated carrier.
8. What commercial system currently employs P-M to produce an F-M wave?
9. Two vectors are displaced from each other by an angle of 50° . One vector is 10 units long, the other is 7 units long. Draw these to scale and then determine graphically the length of the resultant vector. What angle does this resultant vector make with the 10-unit vector?
10. Two vectors are 180° out of phase. One vector is 12 units long, the other is 21 units long. Draw the resultant vector. What is the angle between the resultant vector and the longer vector?
11. A carrier is being phase-modulated by a 1000-cycle audio sine wave. At what points throughout the modulating cycle is no indirect F-M being produced? Why?
12. In the preceding example, at what time during the modulating cycle is the maximum amount of indirect F-M being generated? Why?

Chapter 3

F-M AND INTERFERENCE

Interference Suppression, F-M's Main Advantage. Frequency modulation was popularized because of its effectiveness in combating interference against which amplitude modulation operationally is seemingly ineffectual. The fact that a wider audio-frequency range may also be included without unduly increasing the bandwidth required by an F-M station is advantageous but not overly important. There is much evidence that the inclusion of audio frequencies above 5000 cycles is not greatly desired by 90 per cent of the listeners. Fidelity is another controversial issue. If fidelity is interpreted to mean the true reproduction of sound, than an A-M system is just as capable of good fidelity as any F-M network. By and large, interference suppression is the chief reason for the present popularity of F-M.

The word "interference" is applied to any voltage arriving at the input of a receiver or generated in the receiver itself which obstructs, to any noticeable degree, the satisfactory reception of the desired signal. Obviously, there are many sources that may produce interfering voltages, but only those that have been proved by practical experience to be the most important will be investigated. The others either occur too infrequently or else possess insufficient strength to warrant a separate discussion.

INTERFERENCE FROM OTHER STATIONS ON SAME FREQUENCY

One of the most prevalent annoyances in an amplitude-modulated receiver is the piercing beat-note whistle that is heard whenever two signals operating within the same channel are received simultaneously. The whistle is produced whenever the difference of two carrier frequencies is equal to a frequency in the audible range. Within the same channel, this difference seldom exceeds several hundred cycles. As they are close in frequency, the two signals pass through all the circuits together until the 2nd detector is reached. Here they are converted to their original audible form and the beat note ap-

pears. It is impossible to remove the beat note by any adjustment in the receiver because both signals are operating within the same band. A directive antenna favoring the desired signal may prove helpful but, in most cases, hardly feasible because other desired stations at other frequencies arrive from different directions. At night the interference becomes stronger because of improved propagation characteristics. In A-M, the interfering signal need be only 1/100 as strong as the desired signal to make itself heard. In other words, a noise-to-signal ratio only 1 to 100 will cause noticeable disturbance.

With F-M, a considerable improvement is possible and the desired signal need only be twice as strong as the interference to completely override it. To see why this is so, we shall examine closely the action of an F-M signal when in the presence of an interfering voltage of smaller amplitude.

Amplitude and Phase Modulation from Interference. In Fig. 3.1 we have two signals, separated slightly in frequency and differing in amplitude.

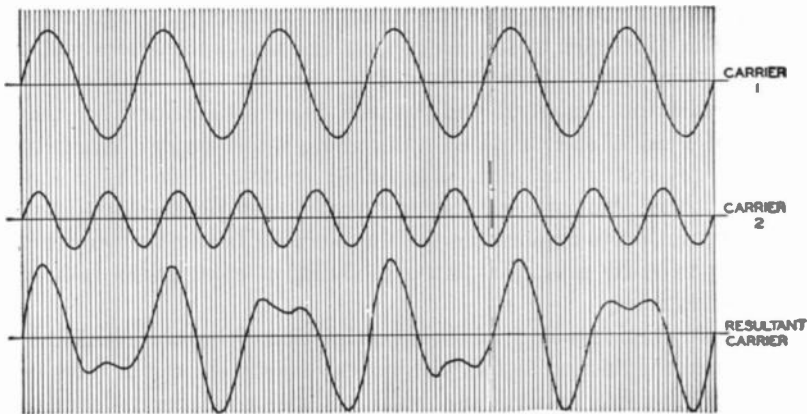


FIG. 3.1. The combination of two carriers to form a resultant which is amplitude- and phase-modulated.

Both are plotted separately to scale. Below these two waveforms is the resultant derived by combining the two waves at every point. If we examine the resultant closely and compare it with the desired signal, we note two differences. First, the amplitude of the resultant is not constant, but varies at a rate equal to the difference in frequency between the two received signals. Thus, if both voltages differ in frequency by 1000 cycles, the amplitude of the resultant will increase and decrease 1000 times per second. In an A-M receiver, this fluctuation would be "skimmed off" (detected by the 2nd detector) and heard as a beat-note whistle in the loudspeaker. What the effect is on an F-M receiver soon will become apparent.

We note, as a second effect of the mixing of two signals, that, although

the *average* frequency of the resultant is the same as the larger signal, the two waves (the larger or desired signal and the resultant) do not always have the same relative position. At times the resultant will lead the desired signal in phase, at other moments there will be no phase difference, and, finally, there will be a lag. In other words, with respect to the original desired signal, the resultant has become phase-modulated. From preceding paragraphs, we are aware that, indirectly, frequency modulation is also produced. The amount of the indirect F-M produced is directly proportional to the phase shift introduced into the resultant wave and to the frequency difference between the two carriers arriving at the receiver. This follows from the formula previously given.

The degree of phase difference caused by the two interacting waves depends wholly upon their amplitude ratio. When the ratio is 2 to 1, the maximum angle of phase shift produced is slightly under 30° . This means that the resultant wave leads and lags, alternately, the larger component by this amount. At larger ratios of desired signal to interference, the phase shift is even less. The difference in frequency between the two carriers will set the rate of variation of the phase shift.

Let us pause for a moment and consider these conclusions, for they hold the secret of F-M effectiveness in suppressing interference. Whenever two voltages are present at the input of an F-M receiver, then Fig. 3.1 demonstrates that their resultant will be amplitude- and phase-modulated. The amplitude variation is responsible for the piercing beat-note whistle that is so annoying in A-M sets; the phase modulation contributes indirect frequency modulation.

If we could eliminate both of these modulations from the resultant signal we would have the desired signal with no variations, since it is only in these two respects that the desired and resultant signals differ. Then, too, we should have effectively eliminated the disturbance of the interfering voltage, since it was this voltage that acted with the desired signal to cause these variations.

Eliminating Amplitude Variations. The simpler to remove is the amplitude modulation. This is accomplished by means of the limiter stage in the F-M receiver. Within limits, different input voltages to the limiter produce a constant output. By this device we accomplish the eradication of all amplitude modulation, leaving only the phase modulation. In this respect, then, we have made the resultant wave and the desired signal similar.

Minimizing Phase Modulation. There is no practical method currently available for removing the F-M produced by the phase variations and still receive the desired F-M signal. But, if we cannot eliminate the indirect F-M, we can, at least, reduce it to the point where it becomes negligible

when compared with the regular frequency variations of the F-M signal. It is for this reason (among others) that we resort to wide-band F-M.

The degree of the indirect F-M formed from phase modulation has been shown in Chapter 2 to depend upon $f\Delta\theta$, where f is given in cycles and $\Delta\theta$ is expressed in radians. A radian is equal to 57.3° , which means that there are 2π , or 2×3.1416 , radians in 360° . When the two reacting signals differ in amplitude by a ratio of 2 to 1, the maximum phase shift introduced in the resultant wave is approximately 30° or about $\frac{1}{2}$ a radian. By way of illustration, let us say that the frequency difference (f) between the two carriers is 1000 cycles, or 1 kc, and the signals differ in strength by 2 to 1. Substitution of these values in the foregoing formula indicates that a frequency shift of 1000×0.5 or 500 cycles occurs. The shift is, periodically, above and below the average frequency of the stronger signal. In this instance, the frequency variations shift at a rate of 1000 times a second, which is the value of f .

Now compare this variation with ordinary F-M. F.C.C. regulations permit a broadcast station's carrier to be shifted from its assigned position to a maximum of 75 kc, or 75,000 cycles. The interference, in the above example, indirectly produced a frequency shift of only 500 cycles. Certainly this is insignificant when compared to the audio intensity developed by a swing of 75,000 cycles! And, remember, for 500 cycles to be produced, the interfering voltage had to possess one half the amplitude of the desired carrier. In the more usual case, the ratio of the noise-to-signal voltage is much smaller, say 1 to 10 or 1 to 20. Under these conditions the suppression is better still. Thus, the use of wide-band F-M completely swamps the small F-M modulation developed indirectly from the interference. Herein lies the power of F-M. However, as we reduce the swing of the desired carrier, the effect of the interference becomes more and more important. For best reproduction, wide-band F-M is required.

Frequency Separation of Interfering Signals. There is another important fact to be learned from the equation $\Delta F = f\Delta\theta$. As we bring the desired carrier and the interfering signal closer in frequency, f becomes smaller. If the two signals are at the same frequency, their difference (f) becomes zero and no indirect F-M appears ($\Delta\theta$ times zero is zero). In other words, the interference can have no effect. The greater the frequency separation between the signals, the greater the indirect F-M produced. But even this has a limit. First, the farther the interfering signal frequency is from the center of the frequency selectivity of the tuned circuits, the more its intensity is decreased by the selectivity of the receiver circuits themselves. Second, the frequency difference f , if greater than 15,000 cycles, becomes inaudible to most people and cannot be heard at the output. There is still

another effect tending to decrease the interference when the signals are widely separated, but this is reserved for a later section. In summary, we see that when the interfering signal approaches the frequency of the desired signal, the $f\Delta\theta$ relation works against it whereas, when the two are widely separated, the selectivity of the circuits and our aural limitations impose other barriers.

In amplitude modulation the aural effect of bringing two frequencies together is to produce a lower beat note. This, however, does not affect the loudness of the beat-note whistle. From this viewpoint alone, the situation is more favorable for F-M. Add the fact that stations can differ in amplitude by a ratio of 2 to 1 for F-M and still come through clear and free and we see the great improvement that the newer system offers. It also makes possible the closer location of broadcast stations. Aiding the latter point is the use of high frequencies for F-M operation. At the high frequencies, radio waves travel essentially in straight paths from transmitter to receiver. The ionosphere cannot be used to reflect the waves because penetration is accomplished without appreciable bending. Hence, long distance interference from other stations is avoided. Admittedly, though, this is not a property of either F-M or A-M, but of the wavelengths employed. If we operated the amplitude-modulated transmitters in this upper range, the interference from a distant station would also be decreased.

As long as the desired signal is at least twice as powerful as the interfering signal, only the desired station will be heard in the loud-speaker. As the two signals approach each other in strength, the extent to which the interfering voltage makes itself felt increases sharply. When this signal finally reaches the level where it is stronger than the desired signal, a transition occurs and it assumes full control, completely drowning out the other carrier. The worst situation is in effect when both voltages are equal, for now no clear-cut tendency exists one way or other. However, the moment one signal becomes even a trifle stronger the response changes, with the stronger signal assuming noticeable control. The process is complete when the ratio reaches the 2 to 1 point.

Domination by the Stronger Signal. The reason for the sharp transition in control when one signal becomes even slightly stronger than the other can best be appreciated by vectors. Consider two signals (Fig. 3.2), both present at the discriminator of an F-M receiver. The larger vector, 1, represents the desired signal; the smaller vector, 2, is the interfering signal; and R is their resultant. At the output of the stage it is not 1 or 2, separately, that is heard, but the resultant. Our interest is chiefly in the indirect F-M produced by the interacting of 1 and 2. Any amplitude variations will be effectively eliminated by limiters preceding the discriminator. The signal, at the discriminator, can be made wholly frequency-modulated.

In Fig. 3.2A, vector 1 is operating at our assigned frequency of 90 mc. This means that it is rotating about point O at 90,000,000 times a second. Vector 2 is also rotating but at its own rate, say 90,005,000 times a second. Their difference is 5000. Vector R will also rotate about O , but its speed will depend upon whether 1 or 2 is greater in amplitude. This will become evident as we proceed.

Since signal 2 has a slightly higher speed (frequency) than 1, it means that 2 will gradually move farther and farther away from its relative position shown in Fig. 3.2A. At one point in time, it may have advanced from

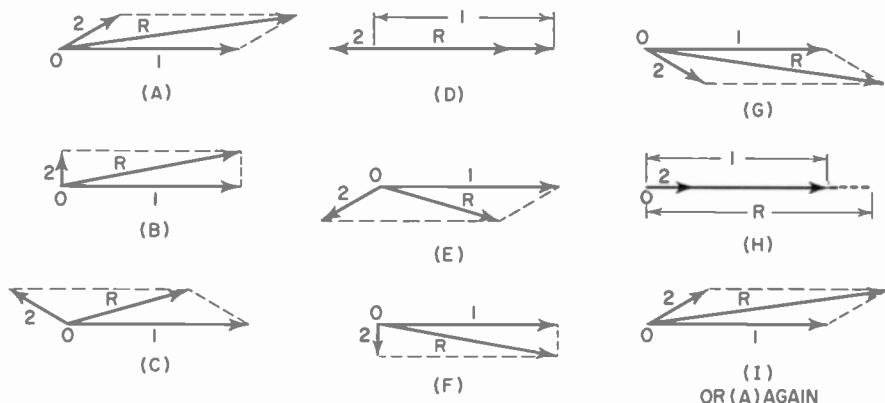


FIG. 3.2. The resultant wave (R) produced by the interaction of a strong signal (1) and a weak interfering signal (2).

vector 1, as shown in B ; at some later instant it would be as seen in C (if we momentarily stopped the motion), then D , etc., until gradually it would work its way back to the position shown in A . The sequence is repeated for as long as the rotation is allowed to continue. If vector 2 has a lower frequency than vector 1, it would lose instead of gain a little each instant and so appear to move in the opposite direction with respect to vector 1.

The reader will note the similarity between this action and that of two cars speeding around a large oval. The speedier car will gradually increase its lead over the slower car. If this is allowed to continue long enough, then, in time, the faster car will be one full lap ahead of the slower car. At this point both cars are again racing alongside each other.

If we stop at each of the instants shown in Fig. 3.2 and draw the resultant vector, we see that it fluctuates about vector 1. At C it has shifted to its maximum position ahead of vector 1 and at E it has fallen to its maximum position behind vector 1. At other times it assumes some intermediate position. These fluctuations represent the phase modulation of vector R . As R rotates rapidly about point O , it is also wobbling back and forth about the

mean positions of vector 1. In other words, vectors R and 1 possess the same average frequency, but R differs in amplitude and phase from 1. The amplitude fluctuations are suppressed by the limiter stage (or stages), whereas the phase variations will, as already noted, cause frequency modulation. The indirect F-M depends upon the maximum angular shift between 1 and R and the speed with which R fluctuates about 1. All this, it will be recognized, has been discovered before. That it appears here is partial evidence that we are on the same track.

In this first example, vector 1 was very much larger than vector 2. This was done purposely in order to show clearly the effect of a large desired signal. In the next case, let vector 1 remain larger than 2, but only slightly so. Now let us see what happens. The action is illustrated in Fig. 3.3.

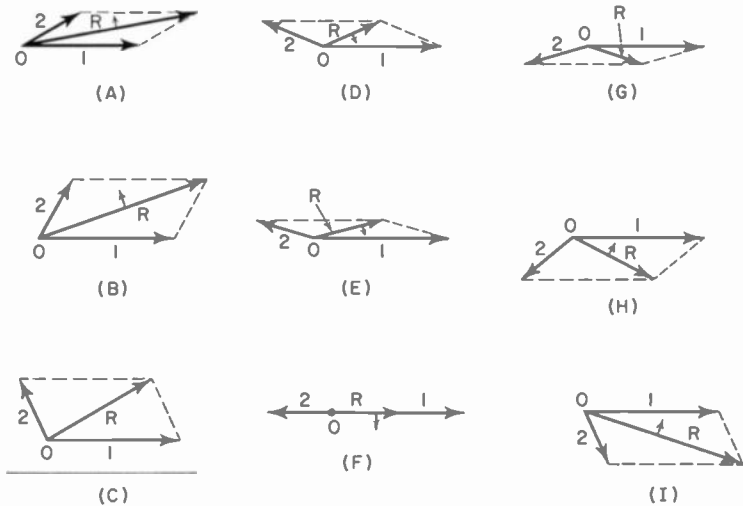


FIG. 3.3. The amplitude and phase variation of a resultant (R) carrier due to the interaction of two signals. The small arrows on R indicate whether its phase (with respect to the desired signal, 1) is going in a positive or negative direction.

The first evident conclusion to be drawn is that here the resultant fluctuates between wider limits. But since it still fluctuates about vector 1, its frequency is that of 1. If its frequency equaled that of 2, then it would follow this vector around, with 2 as its center position. This, however, does not occur. Hence, we conclude that by bringing the two signals closer in amplitude we only succeed in causing more phase modulation in the resultant vector R . Its *average* frequency is still the same as that of the larger signal. What we hear from the output of the receiver is signal 1, but with sufficient interference to cause distortion. This, it will be recalled, is another conclusion previously discovered.

Now, if we make signal 2 greater than signal 1, what will happen? We have merely to return to Fig. 3.2 and transpose the numbers 1 and 2 (vectors) to see the result. The resultant begins to fluctuate about vector 2 and the signal heard in the loud-speaker is determined primarily by 2, with a certain percentage of indirect F-M added, due to signal 1. The transition from vector 1 to 2 is sharp and demonstrates why the predominant signal assumes control in F-M systems. Once the ratio of the two signals reaches 2 to 1, the F-M effect of the smaller on the larger, with wide-band F-M modulation, becomes negligible. This is true whether the interfering signal is another station or just plain noise of any type.

ADJACENT CHANNEL INTERFERENCE

It is not at first apparent why any interference is caused by stations on the channels adjacent to the desired signal, but such interference is possible. Due to this possibility, it is the practice of the F.C.C. never to permit stations to broadcast on adjacent bands in the same community. The closest that two stations in the same region can transmit (in frequency) is on alternate channels. Thus, if the three frequencies 90.1 mc, 90.3 mc and 90.5 mc are available, only 90.1 mc and 90.5 mc would contain stations, at any one time, in any one service area. In some other region, however, it would be perfectly feasible to have a station assigned to the 90.3-mc frequency.

The interference between two stations transmitting on adjacent bands is a result of:

1. The type of selectivity curve designed for most commercial receivers.
2. The fact that sidebands of the two adjacent stations can interact to form audio frequencies at the output of the discriminator.

It was found in a previous section that F-M interference produced when two stations operate on the same band is due to the introduction of phase

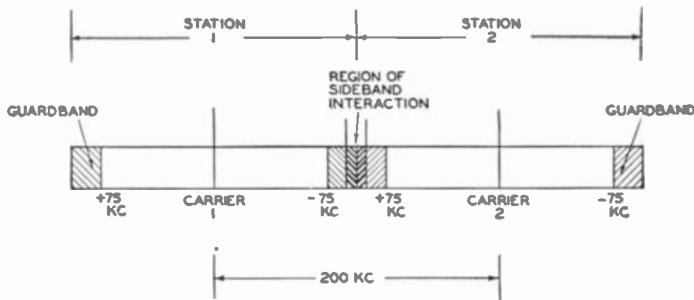


FIG. 3.4. In the darkened region between station 1 and station 2 it is possible for their sidebands to interact and produce interference.

variations. The frequency of the audio voltage developed from the phase modulation is equal to the frequency of separation of the two carriers. Obviously, if the interacting carriers differ by more than 15 kc, no audible interference appears at the loudspeaker. When one of the carriers arriving at the receiver is centered on an adjacent channel, the possibility of its creating audible interference in reception is non-existent because of the 200-kc frequency difference between the carriers. But consider the possibility of interference when the sidebands of each carrier react with each other. These would be capable of producing an audible note if both sets of sidebands extend into their respective guardbands. This can occur on strong modulation. The region that is sufficiently close to an adjacent guardband to cause audible interference is indicated in Fig. 3.4.

The selectivity characteristic of the tuning circuits of the receiver must also be a partner to the action, since a sharp cut-off at the edge of the desired channel would prevent reception of adjacent channel signals (see Fig. 3.5).

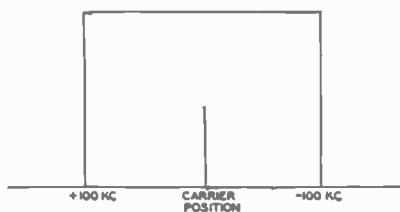


FIG. 3.5. Perfect response characteristic for receiver tuning circuits.

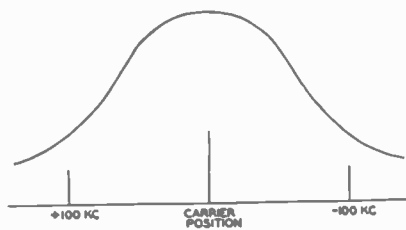


FIG. 3.6. Practical modification of Fig. 3.5.

In commercial receivers, the added expense necessary to include sharp cut-off resonant circuits is prohibitive. As a compromise we find the familiar tapered response shown in Fig. 3.6. With this form, it is quite possible to receive the sideband frequencies of adjacent channel stations. If, during any portion of the modulation, the adjacent signal is stronger at the discriminator than the desired frequencies, the desired signal will be ruined. On the other hand, if the desired signal has a greater intensity at the discriminator, no distortion will be present.

The solution of interchannel interference lies in good selectivity response of the receiver plus the prohibition of near-by adjacent channel stations. The latter has been accomplished already by existing F.C.C. regulations and the former is a matter of individual design. As it is, the situation using F-M is considerably better than the corresponding situation for A-M for several reasons.

1. The use of a guardband on F-M decreases the possibility of signals on adjacent bands interfering with each other.

2. The line-of-sight transmission characteristics of the higher radio frequencies limits the interference from stations in other areas. At the lower A-M frequencies, long distance propagation is always possible, especially at night when the attenuation is less.

3. The inherent characteristics of the F-M system where a 2 to 1 ratio causes the "practical" elimination of the weaker signal. Using A-M, this does not occur until the stronger signal is 100 times more powerful than the weaker one.

In neither system are stations in the same service area placed on adjacent channels. However, because of reason No. 2 above, the advantage of such a condition benefits F-M more than it does A-M.

STATIC

Perhaps second in importance as a disturbing influence in radio reception is static. This is especially true during the six warmer months of the year when thunderstorms are more prevalent. During a local thunderstorm, only the very loudest programs can be heard, and even these with difficulty. The strength of the electrical waves set up is strong enough either to completely drown out radio signals or else present sufficient interference to destroy the clarity of reception. This is the existing situation with A-M and its only solution, at the time of this writing, has consisted of the following:

1. The insertion of limiters in the audio section of the set. These stages limit the maximum amplitude of all signals and prevent thundering crashes of volume accompanying a strong outburst of lightning. The limiter, however, is quite ineffective when the incoming interference is of the continuous, crackling variety. As the name suggests, these devices only limit incoming bursts. The signal, during these moments, is still ruined.

2. Designing the receiver selectivity to include no more than the necessary commercial bandwidth, 10 kc wide. It has been discovered that the energy contained in a static outburst is spread over many frequencies, with the greatest concentration at the lower frequencies. Hence, the narrower we make the bandwidth to which the receiver will respond, the less static noise we will obtain.

Beyond these two refinements of an ordinary receiver, very little can be done to overcome the effects of static. With F-M, we not only have the advantage of the 2 to 1 ratio, but also the advantage of frequency. Since most of the energy of an outburst is located at the low frequencies, very little reaches the high-frequency bands allocated to F-M. In addition disturbances due to distant thunderstorms seldom, if ever, reach the receiver because of the limited transmission range of high frequencies.

THERMAL AGITATION AND TUBE HISS

Even in the complete absence of natural disturbances and interfering stations, there remains a practical limit to the weakest signal that can be received by any set. The limit is fixed at the first or second stages of a receiver and is due to two *main* sources, thermal agitation and tube hiss.

Thermal agitation arises from the random motion of electrons in any conductor. The movement of the electrons, in both directions, constitutes a current flow. Since there are usually a few more electrons moving in one direction than in the other, a voltage is set up across the conductor which is proportional to the net current flow and the value of the conductor resistance. The polarity of the voltage due to thermal agitation changes constantly, depending on the direction in which the maximum number of electrons are moving. Because of this, there is no definite pattern to the random voltage, or, for that matter, any one frequency at which the electrons move. It has been found that the energy of the disturbance is distributed uniformly throughout the entire frequency spectrum used for communications.

The amount of voltage that is developed by thermal agitation in conductors can be computed from the following relationship:

$$E^2 \text{ (rms)} = 4KTR \times (f_2 - f_1)$$

where

E = the rms value of the voltage generated across the resistance

K = a constant = 1.37×10^{-23} watt-second/degree

T = the temperature of the conductors (it is expressed in absolute degrees, Kelvin, which is equal to 273 plus the temperature in degrees centigrade)

R = the value of the resistance of the conductor, in ohms

$f_2 - f_1$ = the bandwidth of the receiver (for F-M, this would be equal to 200 kc).

To compute the voltage, merely substitute the known quantities in the equation, take the square root of the answer, and this will be the rms voltage.

An inspection of the preceding formula indicates that, with all other factors constant, the wider the bandwidth which the set is designed for, the larger the amount of thermal agitation voltage developed. Hence, so far as this one point is concerned, narrow-band F-M offers less intrinsic noise than wide-band F-M. However there are many other advantages to be gained by the use of a wide band and this is the form of present-day F-M receivers.

The other main source of internal noise is in the tubes. We obtain a series of many overlapping impulses due to the fact that the current flow from the cathode to the plate of a tube is not a continuous fluid but a moving congregation of separate particles, the electrons. This is known as the

“shot effect.” We obtain noise even when so-called steady current is flowing, because, at any single moment, more electrons are impinging on the plate than at some other moment. Over any time interval, the current is steady, but instantaneously it fluctuates quite rapidly because of its nonfluid nature. The instantaneous fluctuations represent the noise component. Examination has revealed that the energy of the noise is distributed evenly throughout the frequency spectrum. In this respect it resembles the noise arising from thermal agitation. For the purposes of this discussion, we can combine thermal agitation and tube hiss under the general heading of random noise as a single form of interference.

RANDOM NOISE

Random noise consists of many frequencies unrelated in phase. In the absence of a carrier, the different frequencies beat with each other to produce an audible noise when the random voltages are demodulated at the discriminator. The loud hiss received when tuning between stations is due to the random voltages. This is true of F-M and A-M sets. In fact, with F-M, the amount of interstation noise is greater because of the wider bandpass of F-M receivers. Many manufacturers eliminate interstation hiss by special silencer circuits that cut in whenever the station is tuned out. This makes for a quiet receiver.

A more important situation arises when a carrier is present. In this case we have interactions between each of the random noise voltages and the carrier plus the interactions of the random voltages among themselves. The first is by far the stronger of the two interactions, so that we can disregard the result of the action among the random voltages themselves.

The effect of the random pulses on the carrier is twofold:

1. Amplitude modulation of the carrier.
2. Phase modulation directly and, from this, frequency modulation indirectly.

If we are considering frequency modulation, then the limiter stages will remove all amplitude variations. The indirect F-M produced will depend, as before, upon the difference in frequency between the carrier and each random voltage plus the phase-angle variation of the resultant. The amount of frequency modulation is zero when the carrier frequency is the same as that of the random voltage. As the two separate, the F-M produced increases and the output from the discriminator follows directly. The graph in Fig. 3.7 illustrates the situation. This is a graphic representation that the interference becomes stronger as the frequency difference between the carrier and the noise pulses increases.

Whereas the double triangular plot indicates the extent of the noise gen-

erated, not all of it is effective at the loud-speaker. The ends of each triangle extend to 75 kc, whereas the audio amplifiers (and our ears) cut off at about 15 kc. Hence, for practical purposes, we can discount that section of the plot extending beyond 15 kc.

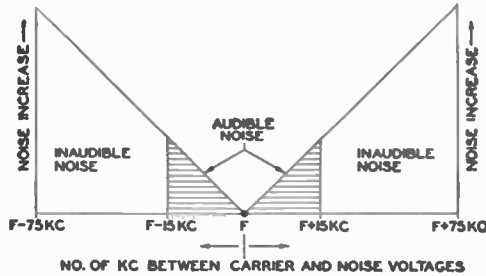


FIG. 3.7. Noise response of an F-M receiver. The noise intensity increases directly with frequency separation between carrier and interfering signal.

Now compare this situation with the corresponding A-M case. To place A-M and F-M on a comparable basis we will assume that the full 15,000 cycles can be passed by the A-M receiving networks. Each random voltage (of different frequency) will mix with the A-M carrier to form a beat note. The amplitude of each note is the same because all random voltages have approximately the same strength. Thus, in place of the triangular response of Fig. 3.7 we obtain the rectangular response shown in Fig. 3.8. This noise extends to 15 kc on either side of the carrier and then stops abruptly because of our aural limitations.

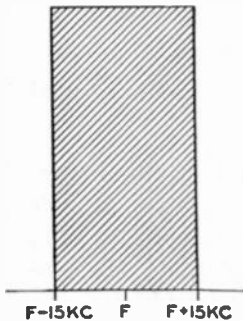


FIG. 3.8. The noise characteristic of an A-M receiver.

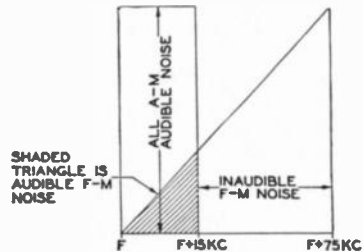


FIG. 3.9. A comparison between A-M and F-M noise characteristics. Note that all the A-M noise is audible, whereas only a small segment of the total possible F-M noise is effective.

In Fig. 3.9, half of each response curve is superimposed on the other to indicate the greater effectiveness of F-M over A-M. Only half the plots are

used because both halves of each curve are identical. A rough comparison readily indicates that the F-M system gives less noise in the output. Just how much the difference actually is can be shown mathematically to be 18.75 decibels or an equivalent signal-to-noise voltage ratio of 8.65. In other words, because of intrinsic characteristics, an F-M system is more effective against the ever-present noise interferences than the A-M method of transmission. All this, of course, in the presence of a carrier. Without the carrier, it has already been noted that the F-M system is noisier.

Noise and the Deviation Ratio. The deviation ratio of an F-M system has a direct bearing on its ability to suppress noise. By deviation ratio we mean the ratio of the maximum carrier swing to the highest audio frequency. For the figures quoted previously, a deviation ratio of 5 was used; that is, a maximum carrier shift of 75 kc and an audio frequency of 15 kc.

Suppose, however, that the F-M system is designed for a maximum frequency deviation of 60 kc. The highest audio frequency will be kept at 15 kc. The foregoing set of conditions limits the shifting of the carrier, under high level audio voltages, to 60 kc. At the receiver, with the *same* degree of noise present as before, we have less desired signal pushing through to override the noise. As a consequence, the signal-to-noise ratio is lower and the effectiveness of F-M has decreased. It is the ratio of the desired F-M swing to the indirect F-M swing produced by the interference that accounts for the efficacy of the system. For a deviation ratio of 4 ($60 \div 15$), the signal-to-noise ratio reduces to 6.92, or a decibel value of 16.75. A comparison between this F-M system and A-M is shown in Fig. 3.10. It is quite evident that the effectiveness of the F-M has diminished, as indicated by the larger proportion of the shaded triangle.

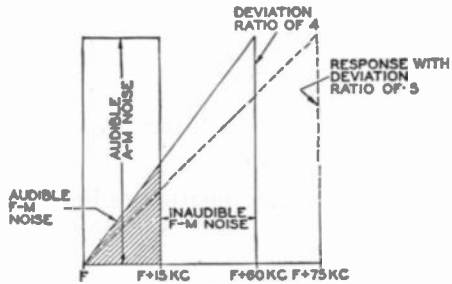


FIG. 3.10. The increase in noise in an F-M receiver with decrease in deviation ratio.

If we continue to limit the swing of the F-M carrier, the signal-to-noise improvement (over A-M) diminishes correspondingly. Comparison diagrams are shown in Fig. 3.11 for the instances where the F-M shift is reduced to 45 kc, then 30 kc, and finally 15 kc. Note that even in the worst situation, with a total frequency shift of only 15 kc, we still obtain a better signal-to-noise ratio than possible with A-M. The advantage is 4.1875 decibels.

The importance of obtaining the highest deviation ratio is thus evident. That the deviation ratio does not extend beyond 5 is due to the fact that the additional ether space required would permit less stations to be assigned

operating licenses. If, for example, we permitted a deviation ratio of 7, then for a maximum audio-modulating frequency of 15 kc, we would need 105 kc on either side of the carrier. Add to this a 25-kc guardband and we have a total bandwidth of 260 kc. Since a deviation ratio of 5 has proved satisfactory, it has been established as standard.

Pre-Emphasis and De-Emphasis. In the analysis of many broadcast programs, it was found that most of the energy is contained at the lower audio frequencies. In addition, it has further been brought to light that the

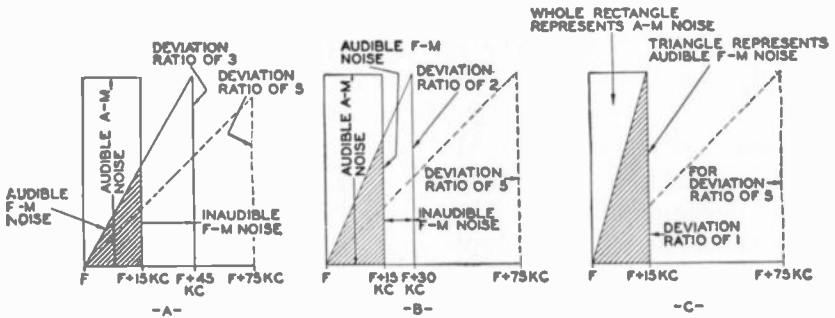


FIG. 3.11. Further comparisons between the noise in A-M and F-M systems with various F-M deviation ratios.

greatest irritating noise generated is located from 5000 cycles up. To reduce the effect of the noise, a pre-emphasis network is inserted in the audio section of the transmitter. The function of the circuit is to favor the frequencies above 1500 cycles. It accomplishes this by proportionately attenuating the lower frequencies more than the higher frequency components of the signals passing through the network.

Pre-emphasis is applied to the audio signals at the first audio amplifier. Beyond the pre-emphasis network, the audio voltages combine in the usual manner with whatever noise is present in the system. At the receiver there is a de-emphasis circuit with the reverse properties of the pre-emphasis circuit. The frequencies above 1500 cycles are reduced to their original values. At the same time a similar reduction in noise occurs. The overall effect is a return of the signal to its proper relative proportions, but with a considerable reduction in noise.

A typical circuit which will favor the higher frequencies is shown in Fig. 3.12A. It consists of a resistor, R_1 , and an inductance, L_1 . The incoming audio signal, coming from a microphone, record player, or tape recorder, is made to pass through this network before it is applied to the audio amplifier. R_1 and L_1 are in series and each applied voltage will divide between them in direct proportion to their impedances. L_1 , however, presents an impedance

that increases with frequency ($X_L = 2\pi fL$), so that, as the signal frequency rises, more of its voltage will appear across L_1 and less across R_1 . Since only the voltage that appears across L_1 reaches V_1 , we can see that the higher frequencies will receive favored treatment.

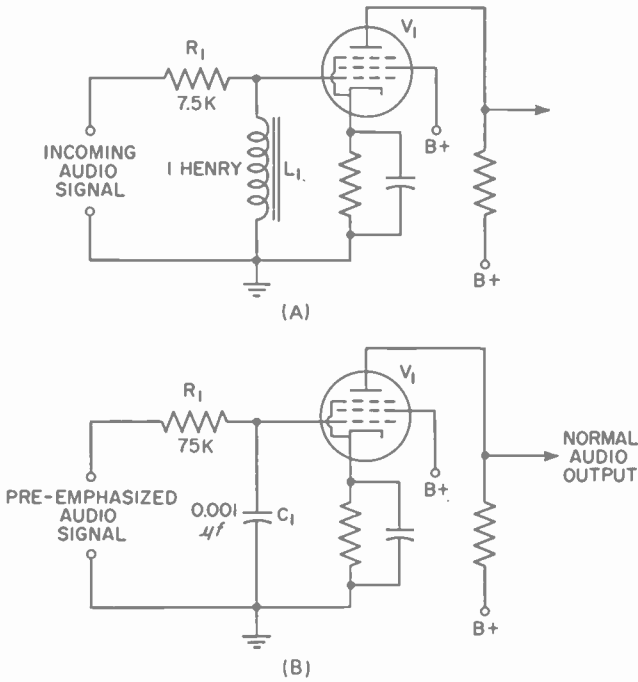


FIG. 3.12.

A de-emphasis circuit is shown in Fig. 3.12B. The audio signal to be de-emphasized is applied to the series combination of R_1 and C_1 . Again, this voltage will divide in direct proportion to the relative impedance of each component. C_1 , however, presents a decreasing impedance with frequency ($X_C = \frac{1}{2\pi fC}$), so that the higher frequencies will develop less voltage here than the lower frequencies. If the pre-emphasis networks are properly designed, their effects will counterbalance and the original audio signal will appear at the receiver output.

Another beneficial effect of de-emphasis is concerned with the noise that is produced by another station or the ever-present random noise. As we have seen, the greater the difference between our carrier frequency and the interference, the greater the indirect F-M formed (see Fig. 3.7). Through the use of the de-emphasis network, the triangular response of Fig. 3.7 is

modified to that of Fig. 3.13. The de-emphasis action, by reducing the level of all frequencies above 1500 cycles, slices off a considerable portion of the noise.

IMPULSE NOISE

Impulse noise, as distinguished from random noise, consists of short, sharp bursts of energy caused by agencies, man-made or natural, external to the receiver.

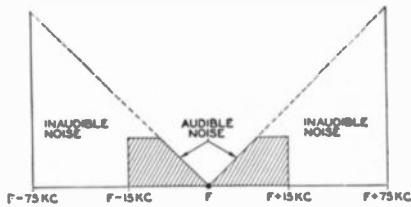


FIG. 3.13. Improvement in noise reduction due to pre-emphasis circuit in transmitter.

A very familiar example of such interference is the type of noise generated by auto-ignition systems or sparking in electrical machines. Although the average value of these bursts is quite low, the peak value may exceed the signal and hence appear noticeably in the loudspeaker.

As long as the peaks exceed the signal, F-M will be unable to override them. At best, all we can do is to minimize the effect of these peaks. For this purpose, the limiter stage can be very useful if designed to respond instantaneously to the rapidly recurrent pulses. How this is accomplished will be made evident in Chapter 8.

HUM

Hum, in a receiver, may affect the desired signal at two points, in the radio-frequency stages and in the audio amplifiers. Hum, in the majority of instances, arises from insufficient filtering in the power supply, although any unshielded transformers and chokes in the power circuit may just as readily be the source, even if the final output from the unit is pure d-c.

Regarding the effect on the F-M receiver output we find that the hum introduced by way of the audio amplifiers is the most annoying. Hum effect is negligible if arising in the R.F. or I.F. stages because the resulting amplitude modulation introduced is removed at the limiter and any phase modulation is generally exceedingly minute.

We must treat hum, present in the audio amplifiers, in exactly the same manner as we do in A-M receivers. Once the F-M signal has been converted to audio variations, it no longer has the protection of F-M. Hence, anything that will affect its amplitude will be carried through to the loudspeaker. Careful by-passing of all d-c connecting points is necessary, plus adequate shielding of transformers and chokes in the power supply. Filament or heater leads must be positioned as far away as possible from grid wires, plate leads and any other links in the audio channel that would affect the audio signal.

The methods are familiar to the radiomen who have encountered the same problem in A-M sets.

PROBLEMS

1. Name the most important types of interference that can affect radio reception and explain each briefly.
2. What is the cause of a beat-note whistle in an A-M receiver? How is this eliminated in F-M receivers?
3. Explain how phase and amplitude modulation arise from interference.
4. How does an F-M receiver effectively deal with the phase and amplitude modulation produced through interference?
5. Of the phase and amplitude modulation produced through interference, which is the most troublesome to an F-M receiver? Why?
6. In the expression, $\Delta F = f \times \Delta\theta$, what does each symbol represent? Indicate the proper units for each symbol.
7. How much indirect F-M is produced when two incoming signals differ in frequency by 5000 cycles and the maximum phase modulation produced is 60° ?
8. What are the advantages of wide-band F-M as compared to narrow-band F-M?
9. What effect does the frequency separation of the two interfering signals have on the amount of interference produced? Explain. For what frequency separation is the interference a minimum?
10. What is the significance of the 2 to 1 ratio in F-M reception?
11. What factors limit the effect of interference in an F-M receiver?
12. Can stations operating on adjacent channels interfere with each other? How?
13. What have the F.C.C. and the receiver manufacturers done to minimize interference arising from stations operating on near-by channels?
14. What is static? Why does it affect an F-M receiver less than an A-M receiver?
15. Under what general category can thermal agitation and tube hiss be placed? What is the frequency spectrum distribution for this type of noise?
16. "An F-M receiver is noisier than an A-M receiver." Explain under what conditions this statement is true.
17. Why is the noise characteristic of an F-M receiver triangular whereas in an A-M set it is rectangular?
18. What is the most difficult type of noise for an F-M receiver to cope with? Why?
19. What is the relationship between the deviation ratio of an F-M signal and its ability to suppress noise and interference?
20. When the deviation ratio of an F-M signal is 1, is the F-M system of modulation superior to A-M modulation? Explain.
21. How does poor filtering in the power supply affect an F-M receiver? How does this differ from its effect in an A-M set?
22. Explain pre-emphasis and de-emphasis.
23. Why are pre-emphasis and de-emphasis beneficial to an F-M receiver? Could the same networks be employed for A-M transmission? Explain.

Chapter 4

PROPAGATION, RECEPTION, AND TRANSMISSION OF F-M SIGNALS

Introduction. In this chapter we shall discuss the manner by which the generated signal at the broadcast station is radiated and the means by which this signal is intercepted at the receiver.

The present allocations established by the Federal Communications Commission set aside the frequencies extending from 88 to 108 mc for the transmission of F-M signals. Now, this band of frequencies is almost 70 times higher than the present standard A-M broadcast band, and just as differences exist between A-M and F-M receivers because of the difference in operating frequency, so differences exist between low frequency and high frequency antenna systems. In order to understand why these differences exist, a general review of radio wave propagation will be undertaken.

An antenna is simply a wire or group of wires which is placed out in the open as far as possible from surrounding objects. Energy is fed to the antenna from the transmitter with the result that currents are established in the wires, flowing first in one direction and then in the other. The currents, in turn, establish magnetic fields which expand and contract about the wires in step with the periodic rise and fall of the rapidly alternating currents. Now, if all the electromagnetic energy established by the antenna currents returned to the wires, no signal would be radiated. However, not all the energy is returned, and that portion which does not return represents the transmitted signal.

The purpose of the receiving antenna is to intercept some of this radiated energy, developing therefrom a voltage which is applied to the receiver terminals. Since the fields that the receiving antenna intercept are due to the transmitting antenna, we could quite logically state that the receiving antenna and the transmitting towers are coupled to each other. To be sure, the coupling is loose, because the two antenna systems are widely separated; but

coupling exists. If we think back to our own experience with coupled coils in radio circuits, we know that maximum voltage is induced in the secondary coil when it is held in the same position as the primary coil. Minimum voltage is induced when the two are held at right angles to each other. By the same token, if the transmitter antenna wires are placed in a horizontal position, then the receiving antenna should be placed in the same relative position. Signals from antennas held in this position are said to be horizontally polarized. If desired, the antenna wires may be mounted vertically, in which case the electric component is said to be vertically polarized. Polarization of an antenna is important, as we shall see in the following discussion.

WAVE PROPAGATION

Types of Radio Waves. Radio waves radiated by an antenna system may follow one of two general paths to reach the receiver and are classified either as ground waves or sky waves, depending on the path that they travel. Ground waves are those which either travel not far above the ground or else are actually guided by the ground. Sky waves are those which travel upward away from the earth for distances sometimes as much as 250 miles before they are bent sufficiently to enable them to return to earth.

Although there is essentially only one mode for sky wave transmission, ground waves may be subclassified into surface waves, direct waves, and ground-reflected waves. Each of these subdivisions enters into the propagation picture in a different way. This will become clearer as we progress.

Surface Waves. Perhaps the best point at which to begin is the surface wave because propagation at low frequencies is accomplished principally through this medium. A surface wave is a wave which travels in contact with the ground, actually being guided by it. The wave must be vertically polarized because, to low and medium frequency voltages, the earth is a fairly good conductor, and a horizontally polarized wave would have the electric field parallel to the ground. The conducting earth would short-circuit the field, preventing its propagation for any appreciable distance.

The surface wave, as it moves over the surface of the earth, induces charges in the earth. The periodic change in wave polarity causes the charge concentration to vary from point to point. These concentrations can vary only by having electrons flow from point to point, and this, in turn, constitutes a current. Since the earth possesses resistance, forcing current to flow through it results in a power loss, the power being supplied by the transmitted wave. As the frequency of the signal increases, the amount of power that is lost rises until, at about 2 mc, the coverage obtained by means of the surface wave is confined to very short distances. However, surface propagation in this manner is extremely useful at lower frequencies. Ship-to-shore installations operating below 500 kc use enormous antennas but are

capable of transmitting steady signals for distances of 1000 miles or more with unvarying regularity. At the A-M broadcast frequencies, 550 to 1600 kc, the primary and secondary service areas receive their signals almost entirely by the surface wave.

The other divisions of the ground wave, the direct ray and the earth reflected ray, do not become important until the signal frequencies rise above 30 mc. Long-distance communication between 2 mc and 30 mc is accomplished principally by the sky wave.

Sky Wave—Ionosphere. The energy radiated by an antenna which does not follow the contour of the earth travels upward through the air until

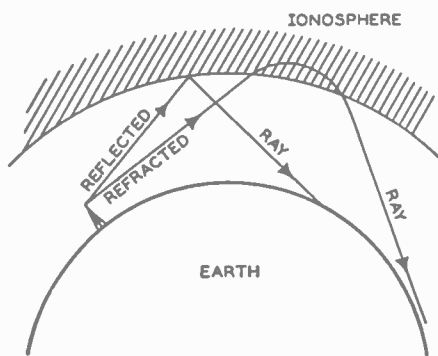


FIG. 4.1. Illustration of reflection and refraction of radio waves at the ionosphere. Reflection occurs less frequently than refraction.

it reaches a region where there exists a concentration of ions and electrons. This region is the ionosphere. For those who are unfamiliar with this terminology, ionization is the process of separating a molecule (in this instance a gas molecule) into a positive ion and one or more electrons. Molecules, of their own accord, seldom separate in this manner. However, by the application of sufficient energy from some outside source, separation can be achieved. In the ionosphere ultra-violet radiations from the sun are this source of energy. These rays, penetrating the gaseous layers surrounding the earth, supply enough energy to cause the electrons to separate from their molecules, the result being a concentration of negative electrons and positively charged ions. Radio waves traveling up from the antenna enter this ionized region and, by the interaction of the wave energy and the free electrons, are bent back toward earth. This process is essentially one of refraction, similar to the effect prisms have on light rays. However, reflection from the ionosphere also occurs, although to a considerably smaller extent. Both methods are illustrated in Fig. 4.1.

The layers of ions, or ionosphere as they are collectively called, are found at distances of from 70 to 250 miles above the surface of the earth. Analysis of this region has disclosed that, although there is ionization of the gas molecules throughout the entire area, there are distinct layers in which the ion concentrations rise to a maximum. The layers are distinguishable and have been given specific designations. The lowest layer, called the *E* layer, is found about 70 miles above the earth, and this height is fairly constant.

The next region of maximum ion concentration occurs at a distance of

it reaches a region where there exists a concentration of ions and electrons. This region is the ionosphere. For those who are unfamiliar with this terminology, ionization is the process of separating a molecule (in this instance a gas molecule) into a positive ion and one or more electrons. Molecules, of their own accord, seldom separate in this manner. However, by the application of sufficient energy from some outside source, separation can be achieved. In the ionosphere ultra-violet radiations from the sun are this source of energy. These rays, penetrating the gaseous

about 185 miles above the earth. This layer has been labeled the F layer, but it exists as one layer only at night, after sunset. During the day it separates into two distinct groups of ions, labeled F_1 and F_2 , respectively. The F_1 layer is usually found 140 miles above the ground while the F_2 layer forms at about 225 to 250 miles. At night, of course, the two regions combine again to form the single F ionized layer. The three are shown in Fig. 4.2 together with the lower E layer. The F_2 layer has the greatest degree of ionization and electron density. It is the layer that refracts the radio waves of higher frequency, the waves that other layers cannot turn back. If a wave does not receive sufficient bending in this final (F_2) layer, it continues into empty space.

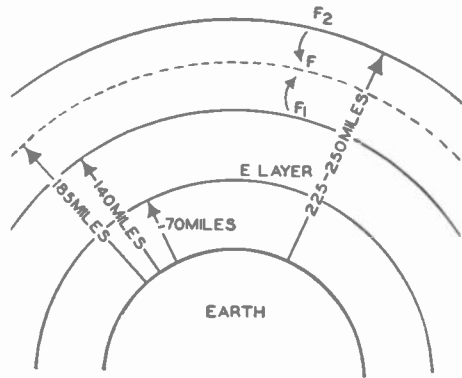


FIG. 4.2. The various layers of the ionosphere. The distances shown are only approximate values since the exact heights vary from day to day.

Wave Bending. When a radio wave enters the ionosphere, the electric field of the wave acts on the charged ions and electrons, causing them to

vibrate in accordance with the changes occurring in the electric field. The important vibrations are those of the electrons since they have a mass which is roughly 1800 times less than the ions, and therefore they are capable of executing greater vibrations than the ions. The frequency of these vibrations are the same as the frequency of the wave. Since moving electrons constitute an electrical current, we have, in effect, a large number of

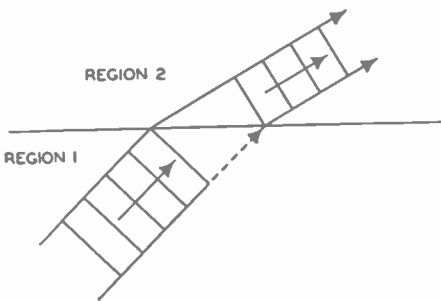


FIG. 4.3. Refraction of radio waves entering the ionosphere.

minute antennas, each radiating a signal having a frequency equal to the frequency of the arriving wave. At the start, when the electric field of the radio wave acts on an electron, the result is a vibratory motion which is parallel to the electric field. However, once the electron starts to move, it reacts with the earth's magnetic field, causing it to leave its straight line path and to travel a more or less spiral path. The energy that is radiated by the electron has now changed direction. In the conditions existing in the

ionosphere, this change in direction is such that the energy is reradiated partially or wholly toward the earth. We obtain the same effect as though the arriving radio wave had been bent around as in Fig. 4.3. The process of bending the radio wave is known as refraction.

The degree of bending that a wave receives depends primarily upon three factors: the angle at which the wave enters the ionosphere, the frequency of the wave, and the electron density of the layer. The importance of the angle may best be seen by reference to Fig. 4.4. The greater the angle ϕ , the more

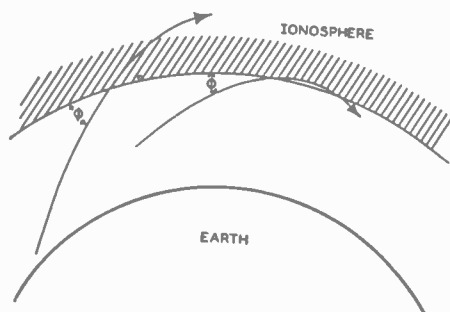


FIG. 4.4. As the angle ϕ which the radio wave makes with the ionosphere increases, more bending must occur before the radio wave is bent back to earth.

the wave has to be bent, or refracted, in order that it return to earth. Waves entering at very small angles require only a slight amount of bending to have their direction changed sufficiently to cause them to return to earth again.

The frequency of the wave is important because it determines the type (and length) of the path followed by the vibrating electron. When the frequency is high, the electron direction must reverse itself so rapidly that the electron never achieves a very high velocity.

Its interaction, then, with the earth's magnetic field is correspondingly slight, resulting in only a small change in electron path. Consequently, the energy reradiated by the electron does not differ appreciably in path direction from that followed by the original wave. However, as the frequency decreases, the velocity of the electron increases because the time between reversals is greater. Its path now is more truly spiral, and the electron direction is altered more by the earth's magnetic field. The reradiated energy thus shows a greater bending effect.

Finally, the density of the charged particles in the layers themselves will determine, in the final analysis, just how much the wave is bent. As the density increases, the amount of refraction increases for any one wave.

Critical Frequencies of Ionosphere. Radio waves may be sent upward at almost any desired angle, but the wave that will require the greatest amount of bending is one that is sent vertically upward or, what is the same thing, at right angles to the earth's surface. The frequency of the wave that is bent back after entering a certain layer at that vertical angle is known as the critical frequency for that particular layer. If a wave of high frequency enters this ionosphere layer at right angles, it will usually not be refracted enough to make it return to earth and will continue upward. How-

ever, this wave of higher frequency may enter the ionosphere at a smaller angle, say 50° , and have a better chance of being returned to earth. Of course, if the frequency is very high, not even a small angle of incidence will help, the wave in this case continuing out into space. At various times in the day and during the year the density of a given layer will change, whereupon its critical frequency changes. Should the density increase, the critical frequency value would rise. Should the density decrease, it would be necessary to lower the critical frequency. The critical frequency may thus be looked upon as an index of the ability of the ionosphere to return a wave to earth. All lower frequencies will be refracted to earth no matter what the angle used to transmit them upward, whereas higher frequencies may be returned only if the proper angle is used. The lowest angles achievable in practice are about 4° to 6° with the horizon. If a wave cannot be returned at this elevation, sky wave transmission cannot be used for this particular wave length.

Sporadic E. The discussion thus far has confined itself to what might be called normal conditions. Under these circumstances, frequencies as high as 30 to 35 mc can regularly be refracted by the F_2 layer. For frequencies beyond 35 mc, sky wave transmission becomes useless, the waves receiving insufficient bending to return to earth. However, every once in a while it is found that frequencies as high as 60 mc are refracted back to earth by the ionosphere. As this is an unusual occurrence, it is necessary to look for conditions in the ionosphere that are present only at these times and then disappear. Investigation has revealed that at these times portions of the lower E layer contain unusual concentrations of electrons, and it is these patches which refract the higher frequencies. The name given to these small, well-defined regions within the E layer is Sporadic E spots. These spots occur rather frequently, especially in summer, and are used to good advantage by the amateurs operating on the 56-mc band. They may last from a few minutes to many hours and may remain in one locality or travel about at high speeds. The time of the day or night does not appear to have too great a bearing on their appearance since long-distance communications have occurred at all times.

Another phenomenon that occurs frequently is the intense magnetic "storm" that disturbs long-distance communications for periods of from 3 to 5 days. During these storms the earth's magnetic field fluctuates widely in intensity and then gradually settles down to its normal value. The origin of these storms appears to be in the sun where certain types of solar disturbances cause the emission of clouds of electrons and other charged particles. These particles travel from the sun to the earth, penetrating through the F layers and reaching the E layer. The effect of these clouds of charged particles on our own ionosphere depends on their intensity and the extent

of the clouds. Under a severe invasion, the layers of the ionosphere lose their distinctiveness, varying widely in density, height, and refractive power. The F_2 layer may even disappear for a short time. Long-distance communication during these periods of turbulence is erratic.

The terms "reflection" and "refraction" have been previously illustrated (Fig. 4.1), and it is seen that when a wave is reflected from a surface its direction is sharply altered whereas when a wave is refracted, the change in direction occurs gradually. Most wave-bending in the ionosphere is achieved by refraction. However, when the frequency of the signal is sufficiently low, and the ionized layer sufficiently dense, the wave does not penetrate the layer to any appreciable extent, and change in direction occurs through reflection. The line of demarcation between reflection and refraction is not too distinct, and throughout the literature both terms are used interchangeably, especially when describing the phenomenon that occurs during periods of ionospheric disturbances.

H.F. Propagation. The discussion, thus far, has concerned itself exclusively with radio communication using frequencies up to approximately 30 to 35 mc. These are the highest frequencies that will be refracted by the F_2 layer under what may be termed normal conditions. For Sporadic E refractions it is possible to go as high as 60 mc although odd conditions are encountered. While Sporadic E refractions occur quite frequently, they cannot be depended upon for any regularly scheduled traffic. From here to the end of this chapter, frequencies in the very high region will be emphasized, extending beyond 30 mc.

Three Components of H.F. Propagation. High frequency wave propagation can be broken down into three categories, each of which is responsible for the reception of these signals by a different means. First is the so-called "line-of-sight" method which embodies both a direct ray and an earth reflected ray; second, there is refraction in the region of the troposphere; and last, there is diffraction around the curvature of the earth. These three methods are generally independent of each other. Although it may happen that all of these methods combine to lay down the same transmitted signal at the same point, it does not occur often. Usually the last two modes of propagation are responsible for signals appearing beyond the horizon, whereas the first method deals with the distances such that the receiving and transmitting antennas are in a direct line above the curvature of the earth.

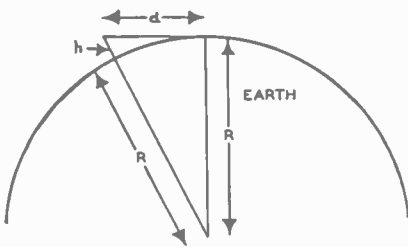


FIG. 4.5. How to calculate the line of sight distance.

Line-of-Sight Method. In this method of wave propagation, energy from the transmitting towers travels on a direct line to the horizon. Beyond the horizon, an extension of the straight line from the transmitter shows that the energy travels into the ionosphere and on into outer space. Unless this energy is intercepted by a receiving antenna before it travels away from the earth, communication by this method becomes impossible.

The following simple derivation will indicate how to determine the distance from the transmitting tower to the horizon. To simplify the mathematics, it will be assumed that the earth is flat over that small portion of its surface. This assumption results in a right triangle. (See Fig. 4.5.) From elementary geometry it is possible to write the following equation:

$$d^2 + R^2 = (R + h)^2 = R^2 + 2Rh + h^2$$

Since h is very small compared to the radius of the earth, the h^2 term can be neglected. This gives us

$$d^2 + R^2 = R^2 + 2Rh$$

or

$$d^2 = 2Rh$$

In this equation, R is most conveniently stated in miles, h is best given in feet. In order for d to be given in miles, h must be divided by 5280 to convert to miles and keep the units on both sides of the equation identical. The equation now becomes

$$d^2 = \frac{2Rh}{5280}$$

The radius of the earth is approximately 4000 miles. If we substitute this value in the above equation, we obtain

$$d = 1.23\sqrt{h}$$

where d is in miles and h is in feet. To show the relationship between d and h for various values of h , a graph of the equation has been drawn and is shown in Fig. 4.6.

It is assumed in the foregoing that the receiving point is at ground level, and the result indicates the distance from the transmitting antenna to the horizon. However, if the receiving antenna is not at ground level but is raised into the air, it should be evident that the direct-line distance between the two antennas can be made greater before the curvature of the earth again interferes with the direct line. (See Fig. 4.7.) By means of simple geometrical reasoning, the maximum distance between the two antennas is now increased to

$$d = 1.23(\sqrt{h_t} + \sqrt{h_r})$$

where d is the distance between antennas in miles, h_t is the transmitting antenna height in feet, and h_r is the receiving antenna height in feet. Because raising the antenna height increases the direct-line distance, trans-

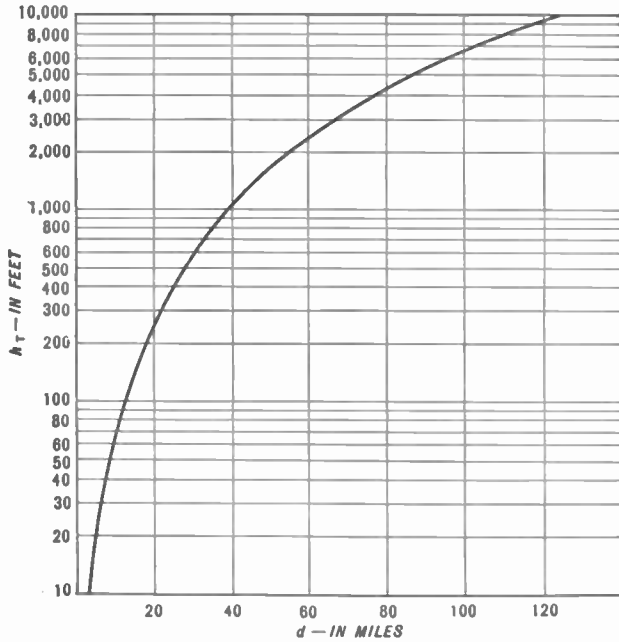


FIG. 4.6. The relationship between the height of the transmitting antenna (in feet) and the distance in miles from the antenna that the ray may be received.

mitting antennas for F-M and television stations are placed atop tall buildings or on high plateaus. There may be other obstacles in the path of the direct rays which would result in absorption of the signal energy and hence



FIG. 4.7. Increasing the line-of-sight distance from the receiving antenna to the transmitter by raising both structures as high as possible.

tend to weaken or distort the received sound or picture. These obstacles assume great importance for television stations where the interfering influences may result in distorted images on the cathode-ray tube screen.

The signal developed at the receiving antenna is due not only to the direct ray but, in addition, to a ground reflected ray. In Fig. 4.8, the waves reaching the receiving antenna may do so by one of two paths: by going directly to the receiving antenna; or by arriving there after reflection from the surface of the earth. Whether or not a good clear signal will be received depends upon the phase relationship of the combining signals. To illustrate this point further, note what happens to the reflected wave. At the point where the reflected ray impinges on the earth, a phase reversal of 180° and, in addition, an absorption of energy in an amount dependent upon the conductivity of the earth at this point have been found to take place. The phase shift produces a wave at the receiving antenna which acts in a manner opposite to the direct ray, lowering its signal level. Fortunately, two conditions act against this decrease by the reflected ray. One

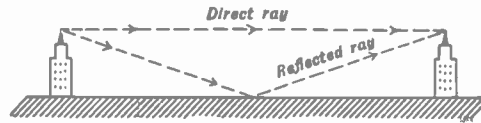


FIG. 4.8. The reflected radio wave, arriving at the receiving antenna after reflection from the earth, may lower the strength of the direct ray considerably.

is a weakening of the wave due to absorption at the point where the wave grazed the earth; the other results from an additional phase change (not the 180° just mentioned) arising from the fact that the length of the path of the reflected ray is longer than that of the direct ray. Thus, there is a total phase shift of 180° plus whatever else may have been added because of the longer path. As a result of these factors, the strength of the direct signal is not decreased as much by the reflected wave as one would at first expect. It has been found that the received signal strength increases with height of both antennas. At the same time, a decrease in noise pick-up occurs. For F-M and television signals this decrease is most important. By raising both antennas, we can increase the difference between the lengths of the paths that the direct and earth reflected rays travel and thus bring their phase difference closer to 360° . When a phase difference of 360° occurs between the signals, they add, producing a field strength which is approximately twice that obtainable from the direct wave alone. In general, the higher the antenna, the stronger the signal received.

The combination of a direct wave and an earth reflected ray is sometimes referred to in the literature as a space wave. It is not important in A-M broadcasting because the relatively long wave lengths used make the difference in paths between the two rays negligible. The waves are then nearly 180° out of phase because of the phase reversal in the earth reflected ray, and they effectively cancel each other completely. Hence, only the surface wave is important. However, as the signal frequency increases, and the wave length decreases, path differences between the direct and earth reflected ray

become more significant. At the same time, the effect of the surface wave diminishes.

Vertical Versus Horizontal Antennas. The relative merits of vertical versus horizontal polarization have been intensively studied, and it has been found that for antennas located close to the earth the vertically polarized rays yield a stronger signal. Upon raising the receiving antennas about one wave length above the ground, both types of polarization are equally good. Further increase up to several wave lengths has shown that the horizontally polarized waves find a more favorable signal-to-noise ratio and are therefore to be desired. Hence all F-M transmissions are done with horizontally polarized waves.

If signals were received only in accordance with the foregoing formulas and reasoning, the study of ultra-high frequency propagation would have ended there. But signals were consistently received at points beyond the horizon in usable quantity and evidently not due to any unusual phenomena. The reasons for these added field strengths are laid to two definite causes: refraction in the lower atmospheric regions and diffraction by the surface of the earth at the horizon.

These two effects tend to increase the range of an F-M signal for some distance beyond the horizon. A good approximation of the actual distance covered is to modify the line-of-sight equations by a factor of 1.41 instead of 1.23. Thus the previous equation of

$$d = 1.23\sqrt{h} \text{ and } d = 1.23(\sqrt{h_t} + \sqrt{h_r})$$

become

$$d = 1.41\sqrt{h} \text{ and } d = 1.41(\sqrt{h_t} + \sqrt{h_r})$$

With this understanding of the behavior of high frequency waves in the F-M band, we are in a position to appreciate better the reasons for the choice of receiving and transmitting systems.

Receiving Antennas. The primary purpose of an F-M receiving antenna is to develop as much voltage as possible of the desired signal. Now, the amount of voltage which the passing field induces in an antenna wire depends essentially upon two factors: (1) the manner in which the wire is held; and (2) the length of the wire. Technically, (1) is referred to as the polarization of the wire and (2) as its resonant frequency. The first factor, polarization, has already been considered, and it was seen that both receiving and transmitting wires should have similar polarization. In F-M reception, this means horizontal polarization.

The second factor that governs the amplitude of the induced voltage is the resonant length of the receiving antenna. Every length of wire contains resistance, inductance, and capacitance. The longer the wire, the greater

the inductance. The capacitance of an antenna is the capacitance formed between the various sections of the wire itself. (See Fig. 4.9.)

The resonant frequency of the antenna depends upon its inductance and capacitance just as it does in any conventional tank circuit. Since inductance (which predominates) increases with antenna length, increasing the length of the antenna decreases its resonant frequency. By the same token decreasing the antenna length increases its frequency. If the wave length of the received wave is known, it becomes a simple matter to compute the required length of the line. Thus, an antenna is similar to a resonant tank circuit, and its resonant frequency will vary with its length.

Tuned circuits are more sensitive to signals at their resonant frequency than at any other frequency. Hence, antennas designed for F-M are cut to specific lengths. The extent of the F-M band is 20 mc, and the antenna should be capable of receiving all frequencies within this band with equal response. However, whether a single frequency is to be received or a band of frequencies, a tuned antenna still provides the best results. In the case of a band of frequencies a compromise frequency is chosen generally at the center of the band and the antenna made resonant to it.

Wave Length. The distance covered by 360° of an electromagnetic wave

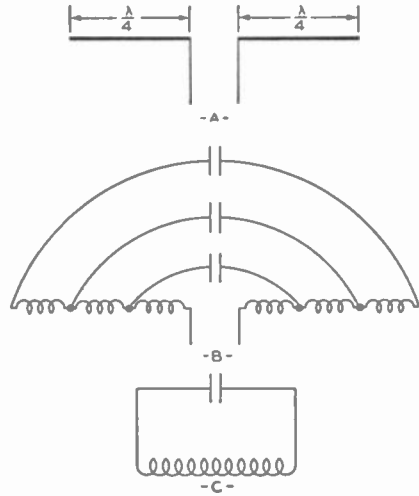


FIG. 4.9. (A) A half-wave dipole antenna. (B) The actual inductance and capacitance in the antenna wires. (C) The equivalent conventional L-C-circuit.

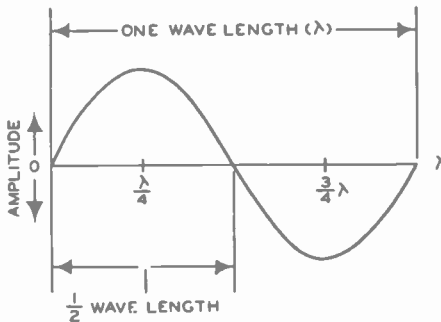


FIG. 4.10. The wave length of an electromagnetic wave.

is called one wave length and this is designated by the symbol λ (lambda). (See Fig. 4.10.) A wave length is equal to the velocity of travel of the wave divided by its frequency. Thus,

$$\lambda = \frac{V}{f}$$

where λ = the wave length in feet

V = the velocity in feet per second

f = the frequency of the wave in cycles per second

We can use 984,000,000 feet per second for the quantity V because the velocity of the electromagnetic wave is the same as light. Hence the foregoing formula becomes

$$\lambda = \frac{984,000,000}{f}$$

The use of the formula is illustrated by the following example:

EXAMPLE: What is the wave length of a 90-mc wave?

Answer:

$$\lambda = \frac{984,000,000}{90,000,000}$$

$$\lambda = 10.9 \text{ feet}$$

Half-wave Antennas. An ungrounded wire, cut to one-half the wave length of the signal to be received, represents the smallest length of wire that

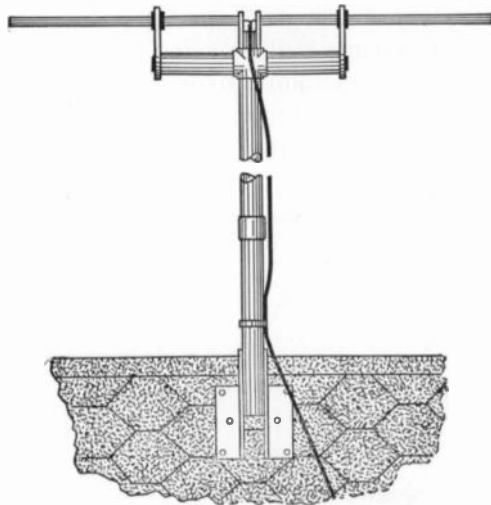


FIG. 4.11. Dipole antenna assembly used extensively for television receivers.

can be made to resonate at that frequency. The half-wave length antenna is most widely used because it is small, compact, and usually provides the receiver with sufficient signal. In troublesome areas it may be necessary to erect more elaborate arrays possessing greater gain and directivity than the simple half-wave antenna. They are, however, more costly and more difficult to install.

A simple half-wave antenna is erected and supported as indicated in Fig. 4.11. Metallic rods are used for the antenna itself, mounted on the supporting structure and placed in a horizontal position (parallel to the ground). Each of the rods is one-quarter of a wave length long, both together being equal to the necessary half wave length. In this arrangement, also known as a dipole antenna, the transmission lead-in wire is connected to the rods, one wire of the line to each rod. The line then extends to the receiver. Care must be taken to tape the line at several points to the supporting mast so that it does not interfere with the operation of the antenna. Taping also prevents the line from flapping back and forth in the wind. Any such motion could weaken the connections made at the rods.

When the properties of a dipole antenna are investigated, it is found that signals are received with greatest intensity when the rods are at right angles to the direction of the signal. This is illustrated in Fig. 4.12A. However, signals approaching the antenna from either end are very poorly received. To show how waves at any angle are received, the graph of Fig. 4.12B is commonly drawn. It is an overall response curve for a dipole antenna.

From the diagram, with the placement of the antenna as shown, the strongest signal would be received from direction A. As the angle made with

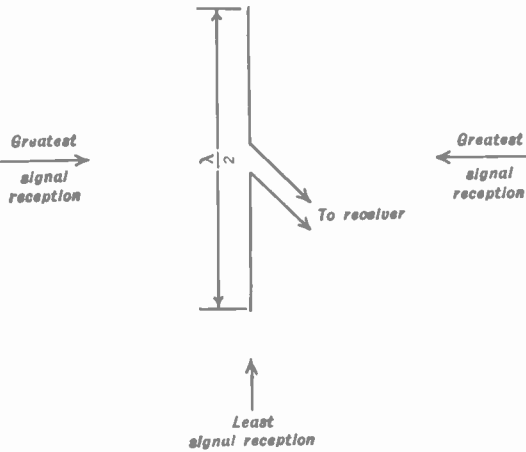


FIG. 4.12A. Dipole antenna, of the type shown, receives signals best from the directions indicated.

F-M SIMPLIFIED

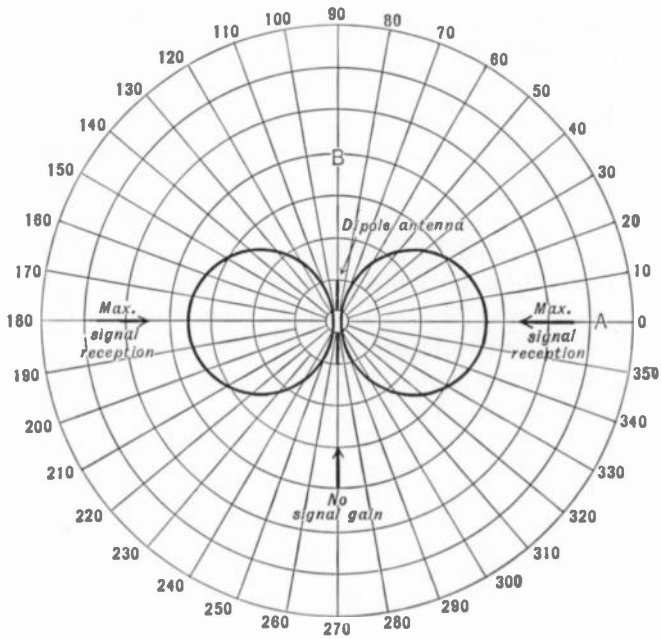


Fig. 4.12B. The directional response curve of a dipole antenna.

this point is increased, the strength of the signal voltage decreases, until at point B (90°) the received signal voltage is at a minimum (or zero). The reader can determine the reception for waves coming in at other angles by inspection of the graph. Notice that good signal strength is obtained from two directions and, because of this, the dipole response may be called bi-directional. Other systems can be devised that are uni-directional, non-directional, or having almost any desired properties. For each system, a response curve would quickly indicate its properties in any direction.

As stated, an antenna must be tuned in order to have the strongest signal develop along its length. Hence it becomes necessary to cut the wire (or rods) to a specific length. The length will vary with each different frequency, longer at the lower frequencies, and shorter at the higher frequencies. It might be supposed, then, that an F-M set, capable of receiving signals with frequencies ranging from 88 to 108 mc, would need several antennas. It is not necessary, however, to go to such extremes and, in nearly all instances, one antenna is sufficient, if tuned to a middle frequency.

Antenna Length Computations. With the F-M range of frequencies, a middle value of 98 mc might be chosen. Although an antenna cut to this frequency would not give optimum results at the extreme ends of the band, the reception would still be satisfactory.

To compute the length needed for the 98-mc frequency half-wave antenna, the following formula is used:

$$L_{\text{feet}} = \frac{468}{f_{\text{mc}}}$$

With f set equal to 98 mc, the length would be equal to $468/98$, or 4.7 feet. Practically, 5 feet might be cut, with each half of the half-wave antenna 2.5 feet long. For a full-wave length antenna, approximately 10 feet is needed. In congested areas, antenna length must be as short as possible, and only half-wave antenna systems are generally found. If longer lengths are desired, the equation should be modified by the proper factor. A full-wave length antenna requires a factor of 2; a wave length and a half requires a factor of 3, etc.

A little thought will show that a half-wave dipole, no matter in what position it is placed, will always radiate a maximum amount of energy at right angles to the wire axis. When the wire is held vertically, right angles to the wire is the horizontal plane and since this plane exists at all points around the wire, the radiation pattern is non-directional. However, when the wire is laid on its side or held horizontally, there are only two points in the horizontal plane that are at right angles to the wire. All other points in the horizontal plane make an angle of less than 90° with the wire, and the energy radiated in that direction varies accordingly.

If we were to illustrate the complete radiation pattern of a half-wave antenna in all directions, the radiation pattern would possess a doughnut shape. (See Fig. 4.13.)

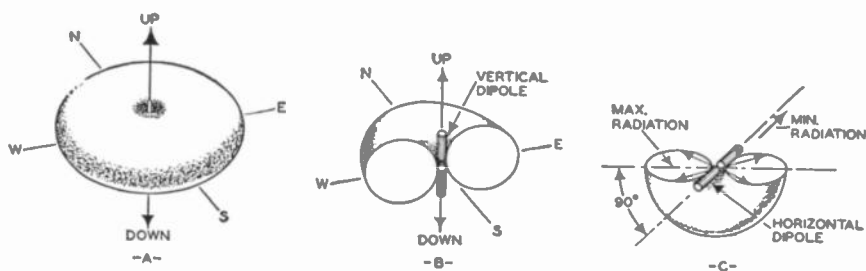


FIG. 4.13. (A) The doughnut-shaped pattern of a vertically placed dipole. (B) A cross section of the doughnut pattern. (C) A cross section when the antenna is horizontal. Compare (C) with Fig. 4.12B.

Connection to a half-wave dipole is made at its center terminals. The two conductors at one end of a transmission line connect here, as shown in Fig. 4.14, while the other end of the transmission line is connected to the

input terminals of the receiver. The impedance which is present across the two antenna terminals is 72 ohms and, for maximum transfer of energy from the antenna to the transmission line, the impedance of the line should also be 72 ohms. (The various types of transmission lines suitable for this purpose

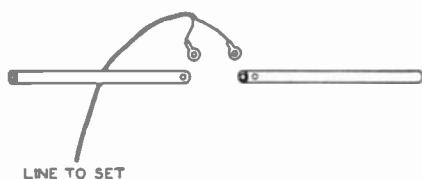


FIG. 4.14. How the transmission line is connected to the dipole antenna.

will be indicated presently. At that time, the significance of impedance matching will also be covered.)

A variation of the dipole is the "V" antenna shown in Fig. 4.15. Actually, this antenna is simply a half-wave dipole whose two arms are bent at an angle to each other. F-M transmitters emit horizontally polarized waves. However, near the receiving antenna, because of reflections from tall trees, buildings, and other obstructions, there may be a vertical component which, when combined with the horizontal component, will alter the angle of polarization above the horizontal plane. Under these conditions, dipole antennas



FIG. 4.15. The "V" antenna.

possessing the shape shown will receive the signals better than an antenna whose elements are perfectly horizontal. In the "V" antenna, the angle between the elements is adjustable over a wide range.

In strong signal areas, where roof antennas are either prohibited or impractical, an under-the-rug antenna of the type shown in Fig. 4.16 may prove

useful. The antenna is stretched out on the floor and its transmission line attached to the receiver. The unit is then moved about until a good clear signal is obtained. At this point, the antenna is placed underneath the floor rug.

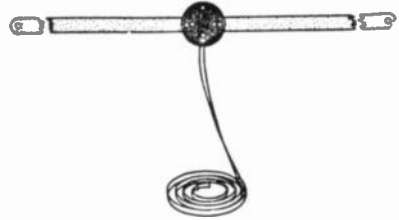


FIG. 4.16. An "under-the-rug" antenna.

Half-wave Dipole with Reflector.

The simple half-wave system provides satisfactory reception in most locations within reasonable distances of the transmitter. However, the signals reaching receivers situated in outlying areas are correspondingly weaker, and noise and interference have a greater distorting effect on the sound. For these locations more elaborate arrays must be constructed—systems that can develop more voltage and provide better discrimination against interference.

A simple yet effective antenna is shown in Fig. 4.17. The two rods are

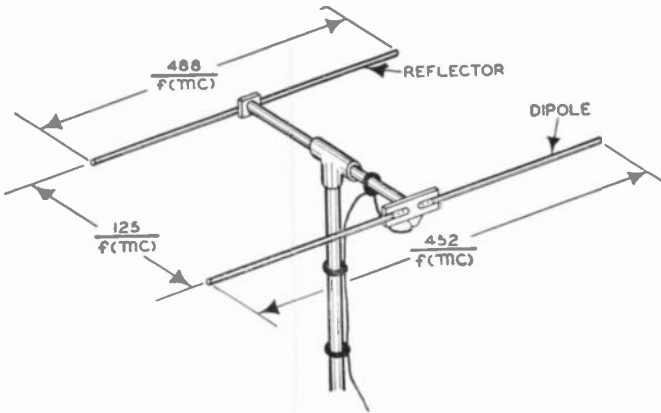


FIG. 4.17. A half-wave dipole and reflector. The length of each element and the spacing between them is given also.

mounted parallel to each other, cut to the dimensions indicated by the formula, and spaced as shown. The action of the second rod (the reflector), which is not connected to anything, is twofold. First, because of its position, it tends to concentrate signals reaching the front wire. Second, it acts to shield the front rod from waves coming from the rear. The signal voltage developed by this antenna is generally 5 db greater than that obtainable from a single half-wave antenna.

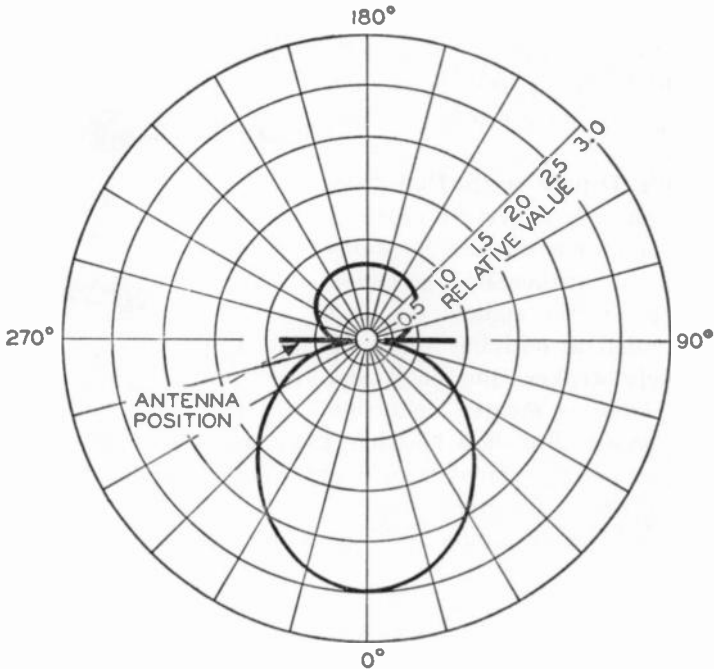


FIG. 4.18. The response characteristic of a dipole antenna with reflector. The reflector, not shown, would be placed behind the dipole, facing 180°.

The response curve of a half-wave dipole and reflector is shown in Fig. 4.18. While not all of the signals reaching the antenna from the rear are suppressed, they are diminished sufficiently to render the response of the array essentially uni-directional. This is advantageous in reducing the number of reflected rays that can affect the antenna. Finally, partial or complete discrimination is possible against interference, man-made or otherwise.

The method of erecting the antenna is similar to that of the half-wave dipole, although the adjustment of the position of the wires is more critical. A small displacement, one way or another, may alter the strength of the received signal appreciably. Many commercial antenna kits do not provide adjustment of the spacing distance between the two wires. However, if an adjustment is possible the spacing may be changed if experimentation indicates that it would result in better reception.

Gain. The gain of an array is a comparison of the signal power which this array would develop across its input terminals to the power which a standard or reference antenna would produce if placed at the same point in space as the directional array. The standard or reference antenna frequently chosen is the half-wave dipole.

Thus, when a certain array is said to possess a gain of 10, it means that this array will develop 10 times as much signal power as a half-wave dipole if it were positioned at the same point in space.

Often, the decibel is used as the unit of gain. In this case, since power is being employed, the formula is:

$$db = 10 \log_{10} \frac{P_1}{P_2}$$

A ratio of P_1 and P_2 of 100 would give the equivalent decibel rating of 20. This follows from that fact that the $\log_{10} 100 = 2$, and 2×10 equals 20 db.

Directivity. When speaking of the directivity of a directional antenna, we mean the sharpness with which the signal is confined or directed to a particular direction. It may, in a sense, be compared with the selectivity of a receiver in allowing one signal to pass through and rejecting all others. The more selective the set, the sharper and more peaked its tuning curve. In radio sets, sharp selectivity can usually be attained only when several tuning circuits are used in conjunction with each other. One coil and capacitor combination, by itself, would not be adequate. It is much the same with antennas. Ordinarily several radiators must be used before a highly directive pattern is obtained. Just one or two elements, by themselves, might show definite directive effects, but these would not be as clear-cut as those obtained when a greater number of antennas are used. Again if a system of wires possesses a marked directional pattern for sending, the same antenna will show similar directivity when receiving.

When speaking of antenna directivity, the term "beam angle" is often employed. Beam angle is the angle between the two points on the radiation curve at which the signal voltage is 0.707 (or 70.7%) of its maximum value. In Fig. 4.19, the angle between points A - B , or the angle θ , is the beam angle for this radiation pattern. At each of these points (A and B), the signal strength is 0.707 of its value at point C .

Points A and B are also known as the half-power points because the radiated power here is one-half of its value at point C . This is an equivalent expression because power is proportional to the square of the voltage and $(0.707)^2$ is approximately equal to 0.5.

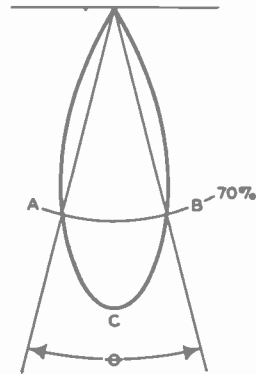
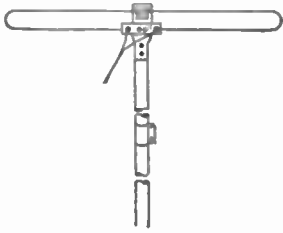
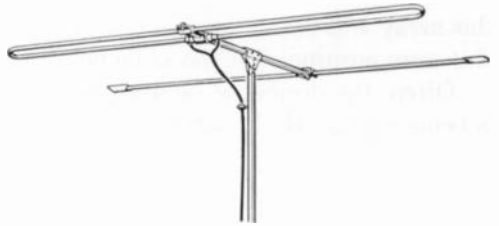


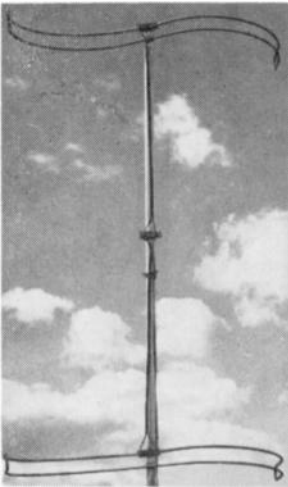
FIG. 4.19. The beam angle θ of an antenna is the angle between two points on its response curve at which the signal voltage is 70.7% of its maximum value.



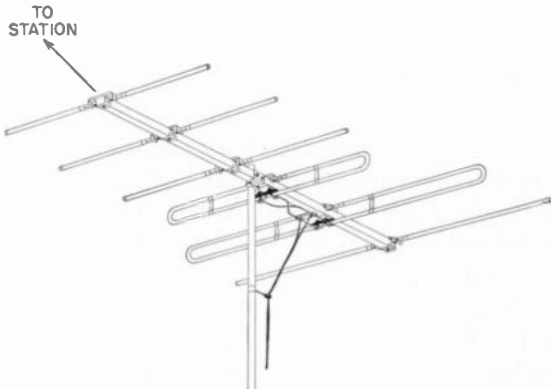
(A) Folded dipole



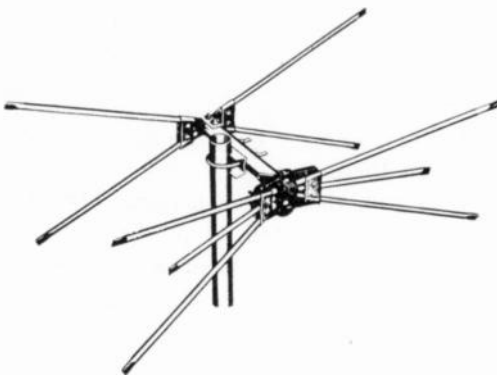
(B) Folded dipole with reflector



(C) "S" type folded dipole



(D) High-gain F-M antenna



(E) Conical antenna



(F) Cross-dipole antenna

FIG. 4.20. Other popular F-M antennas.

Other Receiving Antennas. The two antenna systems described are the most widely used. Other types are found, however, and the more popular of these are shown in Fig. 4.20.

1. *Folded Dipole*, Fig. 4.20A. This consists of two dipole antennas connected in parallel, each dipole being a half wave length long. The separation between the two sections is approximately 3 inches. The folded dipole has the same bi-directional pattern as the simple dipole. Its gain, however, is somewhat greater.

2. *Folded Dipole with Reflector*, Fig. 4.20B. The addition of a reflector has the same effect here as with the simple dipole. Reflector spacing and length are identical with the figures previously given for the simple dipole with reflector.

3. *"S" Type Folded Dipole*, Fig. 4.20C. This array uses the basic folded dipole, but by twisting the elements into the position shown, essentially omni-directional reception can be achieved. The input impedance and gain of this array is similar to that of the folded dipole.

4. *High-Gain Antenna*, Fig. 4.20D. There are two folded dipoles, one rear reflector and three directors in this array. These provide considerably more gain than a folded dipole. However, the antenna has a very narrow directivity pattern and, therefore, must be oriented accurately in the direction of the F-M stations.

One of the folded dipoles is tuned to the high end of the F-M band and one dipole is resonated to the low end. The combination, interconnected, provides a fairly even response over the entire band.

5. *The Conical Antenna*, Fig. 4.20E. Contains two or three rods on each side of the feed point. This serves to broaden the frequency response. Has basically the same directional pattern as the dipole. Each rod is one-quarter wave length long.

6. *The Cross-dipole Antenna*, Fig. 4.20F. When signals come from widely separated directions, it may be desirable to have a non-directional antenna. This is true of the unit shown in Fig. 4.20F and consists of two folded dipoles set at right angles to each other. A quarter-wave section (at the operating frequency of the antenna) of a twin-lead transmission line having a characteristic impedance of 300 ohms is then connected between the input terminals of the two dipoles. To transfer the signal from the dipoles to the receiver, a transmission line is connected from one of the dipoles to the F-M receiver.

After the particular antenna has been chosen, the following points should be kept in mind before installing the antenna.

1. The higher the antenna, the stronger the signal received.

2. The antenna should be set-tested with an actual connection to its receiver before the supports are fixed in place permanently.

Transmission Lines. With the antenna system in position the next problem is the transmission line that conducts the signal from the antenna to the receiver. Although many differently constructed transmission lines have been designed, only five types find any extensive use in F-M installations. These are the two parallel-wire types, the concentric or coaxial cable, and the twisted pair.

Probably the most popular transmission line is the parallel-wire transmission line shown in Fig. 4.21A. It is popular because its cost is relatively

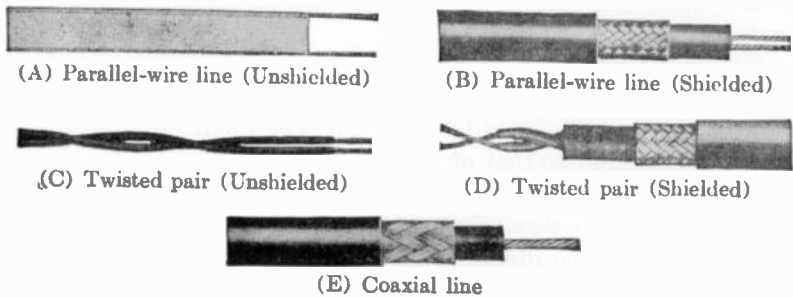


FIG. 4.21. Various types of popular transmission lines used for F-M and television installation.

low, it is easy to work with, and it possesses a low attenuation. Physically, the two wires of this transmission line are encased in a polyethylene strip which serves to hold the wires in place and act as a protective covering. It has excellent electrical and physical characteristics, being a flexible material that is not affected by sunlight, water, cold, or alkalis. At 50 mc, the line loss is less than 0.8 db per hundred feet of line. Its characteristic impedance ranges from 75 to 300 ohms and will match a folded dipole antenna. The line is balanced, which means that both wires possess the same average potential with respect to ground. It is, however, unshielded and therefore not recommended for use in locations where extraneous electrical signals are present.

A parallel-wire transmission line that is completely shielded is shown in Fig. 4.21B. The two wires are enclosed in a dielectric, possibly polyethylene, and then the entire unit is shielded by a copper braid covering. As a protection from the elements, an outer rubber covering is used. Grounding the copper braid converts it into a shield which prevents any stray interference from reaching either conductor. Furthermore, the line is balanced against ground. It is built with impedance values ranging from 95 to 300 ohms. The line loss is greater than the unshielded parallel pair, being on the order of 2.5 db per hundred feet at 50 mc.

The twisted pair of transmission line, Fig. 4.21C, is made by twisting wires about each other in the same manner as twisted lamp cord. It is an economical line, but it has the greatest loss and becomes impractical for lengths beyond 50 feet. The characteristic impedance ranges from 50 to 150 ohms and, at 50 mc, the db loss is 4 for each hundred feet of line. Unless this line is specially constructed, it will deteriorate in time under the ravages of the atmosphere. A shielded twisted pair line is shown in Fig. 4.21D. This line has more desirable characteristics than the unshielded twisted pair, but its cost is greater.

The transmission line shown in Fig. 4.21E is the coaxial or concentric cable. It consists of insulated center wire enclosed by a concentric metallic covering, which is, generally, flexible copper braid. The inner wire is held in position by a solid dielectric, which is chosen for its low-loss properties. The signal carried by the line is confined to the inner conductor, with the outer copper-braid conductor grounded so as to serve as a shield against stray magnetic fields. Because of this arrangement, the line is unbalanced and the input coil of the receiver must be connected accordingly. Coaxial cables are available in a range of impedances from 10 to 150 ohms.

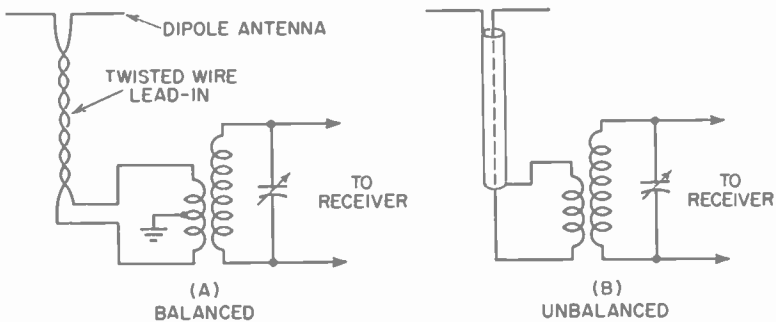


FIG. 4.22. Methods of connecting lead-in wires to the input coil of a receiver.

At the receiver, the connections for balanced and unbalanced line differ, as shown in Fig. 4.22. For a balanced line, the input coil is center-tapped and grounded at this tap. Stray fields, cutting across both wires of a balanced line, induce equal voltages in each line. The similar currents that flow because of the induced voltages are in the same direction on the two conductors of the line and they neutralize each other. In the unbalanced line, the outer conductor of the coaxial cable is grounded, thereby shielding the inner conductor from interference.

Impedance Matching. At two points in the antenna system, connections must be made to the transmission line. One point is at the antenna, and the other is at the receiver input. For the maximum transfer of power,

the antenna impedance should match the transmission line impedance, and a similar matching of impedances should occur at the receiver input.

The input impedance of most F-M receivers is about 300 ohms. To match this, a 300-ohm parallel-wire previously discussed would serve nicely. At the antenna, a match to the 300-ohm line could be made by using a folded-dipole antenna. Thus there is a matching of impedances at all connecting points in the antenna system, providing assurance of a maximum transfer of power.

Although this situation is fairly common, it does not represent, however, every type of installation that will be encountered. Suppose that, in the preceding example, the antenna is a dipole with an input impedance of 72 ohms. What would be the effect of connecting the 300-ohm line directly to the antenna? The answer to this question will depend upon the strength of the signals present in this particular area. If the signal intensity is of the order of 100 microvolts or more, then it is doubtful if the mismatch will have any noticeable effect upon the receiver output. True, a certain amount of the signal will be lost, but there is sufficient signal available so that the loss will not be noticed by the set listener.

However, if the signal level is appreciably below 100 microvolts it may readily happen that insufficient signal voltage will reach the set because of this mismatch. In this instance, use of a matching section between transmission line and antenna is highly desirable.

One method of achieving such a match is through the use of a special matching section called a "Q" section. A section of transmission line, electrically one-quarter wave length long, is inserted between the transmission line and the antenna. (See Fig. 4.23A.)

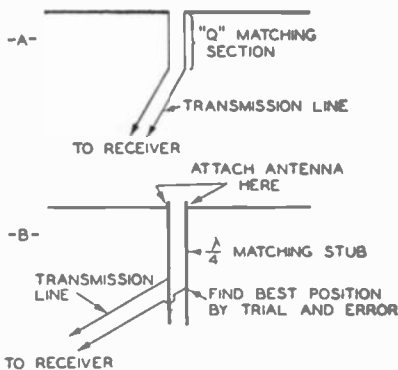


FIG. 4.23. Two methods of matching antennas to transmission lines.

The line is then connected to one end of this quarter-wave section, and the antenna is connected to the opposite end. In order to match the antenna to the line, the characteristic impedance of the "Q" section must be the geometric mean between the characteristic impedances of the antenna and the transmission line. It can be found by the formula:

$$Z_0 = \sqrt{Z_a \times Z_t}$$

where Z_a = impedance of the antenna

Z_t = impedance of the transmission line

Z_0 = impedance of the matching section

To illustrate the use of this formula, let us apply it to the problem on hand, namely, that of matching a 72-ohm antenna to the 300-ohm line. In this case,

$$Z_a = 72 \text{ ohms}$$

$$Z_t = 300 \text{ ohms}$$

$$Z_0 = ? \text{ (to be found)}$$

$$Z_0 = \sqrt{21,600} = 147 \text{ ohms}$$

147 ohms is an awkward value for a transmission line, but 150 ohms will do as well. Therefore, procure a 150-ohm parallel wire transmission line, cut it to a quarter wave length at the frequency used to design the antenna, and connect it between the line and the antenna, as shown in Fig. 4.23A.

Another solution is to use a quarter-wave matching stub. The stub consists of two rods or wires, their length being equal to a quarter wave length at the frequency used to design the antenna. In any such quarter-wave section, the impedance varies from a low of 60 to 70 ohms at one end to several thousand ohms at the other end. Clip the antenna terminals to one end of each cord (or wire) of the quarter-wave section, and then experimentally shift the ends of the transmission line back and forth along each rod (or wire) until maximum signal is reported at the receiver. Since the quarter-wave matching sections have a continuously varying impedance between 70 ohms to several thousand ohms, a point will be found where the impedance of the transmission line is matched. Fig. 4.23B illustrates the arrangement of the several components. Note that with this method any type of two-wire line may be used for the matching stub. Its characteristic impedance can be of any value.

When the receiver is situated in a vicinity where the signal level is high (close to the transmitting station), the simple and economical lead-in antenna of Fig. 4.24 can be used. The twin-lead transmission line is slit down the center and suspended from two insulators to form a dipole antenna. A ring clamp is used to prevent the transmission line from tearing farther.

To summarize:

1. The input system should be matched at each point, i.e., the antenna to the transmission line and the line to the receiver.

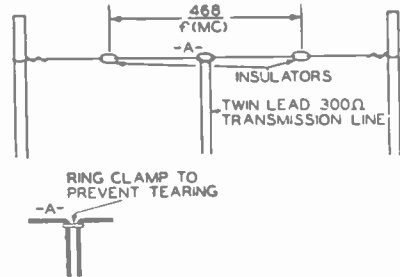


FIG. 4.24. A simple antenna constructed from a twin-lead transmission line. This can be successfully used in areas of high signal intensity.

2. For strong input signals, mismatching as much as 5 to 1 is permissible.
3. For nominal strength signals, a 2 to 1 mismatch is tolerable.
4. If a choice exists as to which point in the input system is to be mismatched, it is preferable that this occur at the antenna rather than at the receiver.

In noisy locations, where the transmission line must pass through areas where the noise generated by man-made machines is high, the use of a shielded lead-in is recommended. There is available shielded twin-lead transmission lines ranging in characteristic impedance values from 95 to 300 ohms. When these are used, care should be taken to see that the metallic braid which is wound around the outside of the line, and which acts as the shield, is carefully grounded at several points along its path from the antenna to the receiver. It is also important that this shield be grounded at each end. Failure to observe these precautions will reduce the effectiveness of the shield and permit external voltages to reach the inner line.

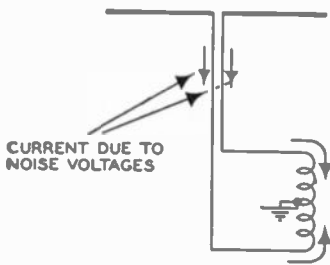


FIG. 4.25. With a parallel-wire transmission line, a balanced input coil is more effective in reducing the effect of noise voltages than an unbalanced coil.

Where the noise is only of mild intensity, its effects may be minimized if the 300-ohm unshielded line is twisted about every two feet on its way to the receiver. The purpose of this is to cause noise voltages to appear in equal strength in both conductors of the line. Since the currents developed by the noise voltages will flow in the same direction along both wires, cancellation at the receiver input will occur. (See Fig. 4.25.) Note that this will be true only when the receiver contains a balanced input, for then the two noise currents, made to flow through each half of the input in opposite directions, will effectively cancel each other. How well this cancellation

is achieved will depend upon how well the input of the receiver is balanced.

There are some F-M receivers which have an unbalanced input. For these sets the input impedance is generally 75 ohms, and a 72-ohm coaxial line is recommended. At the antenna, direct connection of the inner lead of the coaxial line to one rod, and the outer shield to the other antenna rod, will provide satisfactory results if the signal level is high. Note, however, that only half the antenna is being utilized: that section which is connected to the inner coaxial cable. The other half of the antenna is grounded out by its connection to the outer conductor of the cable, which itself is grounded.

When this system is to be used in a weak signal area, better results can be obtained by a converter at the antenna which will provide a match be-

tween the unbalanced 72-ohm coaxial cable, on the one hand, and the balanced impedance of the antenna, on the other. Such units are available at most of the large radio supply houses and consist of properly wound transformers.

F-M TRANSMITTING ANTENNAS

The design and shape of F-M transmitting antennas are governed (and, in the same sense, limited) by the following major considerations:

1. Horizontal polarization.
2. Location at center of service area.
3. Frequency allocations from 88 to 108 mc.

The F.C.C. established horizontal polarization as standard for all F-M broadcast signals, which requires, in general, that the radiating elements themselves be horizontally placed.

Antenna location at the center of the station service area is usually dictated by convenience, both with regard to studio placement and the relatively high expense of incorporating phasing networks to produce non-symmetrical radiational patterns. By placing the antenna at the center of its service area, a non-directional pattern is required.

Finally, there is the allocation of the F-M broadcasting services to the frequency band extending from 88 mc to 108 mc. As noted previously in the discussion of wave propagation, signals with frequencies extending above 35 mc are limited to the line-of-sight distance from the transmitting antenna to the horizon. This means that the extent of the area served by any one station depends upon the height of the antenna structure above the ground. The higher the antenna, the greater the area over which its signals can be received. Now, one of the simplest methods of achieving height is obtained by erecting the antenna atop a high building. To do this effectively, however, requires that the antenna structure be light in weight and fairly compact. Furthermore, precautions must be taken to prevent the antenna structure from becoming an obstruction to aircraft, from accumulating too much ice, from becoming targets for static discharges and lightning, and from buckling under the strain of high wind velocities.

While the height of the antenna determines the line-of-sight distance to the horizon, full use of that area is not possible unless sufficient power is radiated by the antenna. High frequency operation imposes certain limitations on the amount of power that can be economically generated in the transmitter. To overcome this limitation to some extent, it has become general practice to design antennas which can concentrate whatever power they are given only in the desired plane, thereby reducing to a minimum the amount of power which is radiated in undesired directions. In the latter

category would fall any power which is radiated vertically upward, since energy transmitted in this direction does not, at the present F-M frequencies, return to earth. To achieve this concentration of energy in the horizontal plane requires the vertical mounting (or stacking) of a number of units.

In the following pages, commercial F-M antennas will be analyzed to determine how the foregoing specifications are attained. Fortunately, in the range from 88 to 108 mc, such designs can be achieved with structures of moderate dimensions.

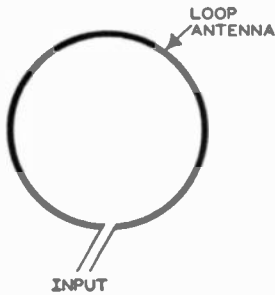


FIG. 4.26. A loop antenna.

The Cloverleaf Antenna. A very simple type of antenna which is capable of radiating a horizontally polarized wave to all points about the antenna in the horizontal plane is the loop antenna. (See Fig. 4.26.) The radiated field will be uniform if the current flowing around the loop is uniform, both in amplitude and phase.

To achieve this uniformity, the length of the loop (around its circumference) should not be greater than $\frac{1}{8} \lambda$ long. At the A-M standard broadcast frequencies this restriction can be observed quite readily, even when the loop contains more than one turn. To remain within this restriction at the F-M frequencies, however, requires that the loop diameter be extremely small, causing its input impedance to become quite low, and thereby making it difficult to feed power to the loop efficiently. Fortunately, the desirable radiation and polarization characteristics of the loop can be obtained without actually using a continuous loop. Instead, it is possible to employ a group of radiators which, when electrically combined, present the same electrical equivalent circuit as a loop. One such construction is the Cloverleaf antenna, shown in Fig. 4.27.

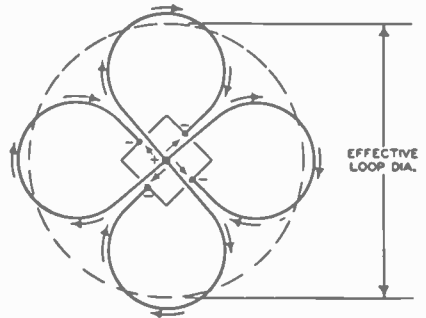


FIG. 4.27. Arrangement of four radiating elements constituting a radiating unit of the Cloverleaf antenna. Arrows indicate assumed instantaneous directions of current. (Courtesy I.R.E.)

The Cloverleaf antenna consists of four half-wave curved radiating elements. One end of each of the radiating elements is connected to a common central conductor of a coaxial line, while the other end of each of the four radiating elements is attached to a separate post of the tower structure which serves as the return conductor. With this arrangement, the currents coming up through the center conductor will, at point X, divide equally between

the four radiating elements in the manner shown in Fig. 4.27. If we consider only outer portions of each curved radiator, we note that the current flows in the same direction in each one, producing effectively a circular flow of current. Points on each radiator which extend in toward the center of the Cloverleaf have currents flowing in them which are in opposition to the currents flowing in the adjacent radiator, with the result that the fields due to this portion of the antenna current flow cancel each other. Thus, only the currents flowing in phase at the outermost portions of each half-wave antenna are effective in developing the radiational pattern for this array.

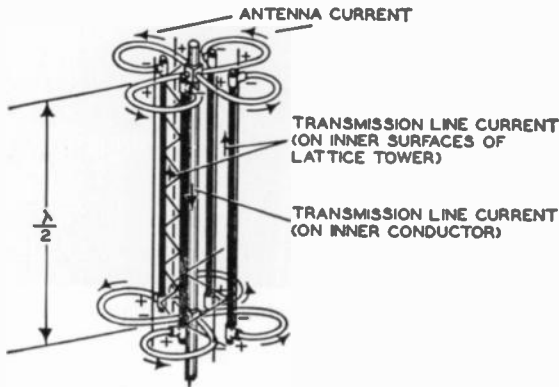


FIG. 4.28. A section of a Cloverleaf antenna showing two sections. Usually a commercial structure contains four or five sections.

To minimize the vertical radiation of the Cloverleaf, a series of four or five similar units are stacked one above the other, separated by distances of approximately one-half wave length. Two such units are shown in Fig. 4.28. Since there is a voltage phase reversal at half-wave intervals along a transmission line, the direction in which the individual radiating elements comprising each unit are curved is reversed. In this way the current flowing in all units all along the array do so in the same direction, and the field radiated by all sections is everywhere in phase.

Circular F-M Antenna. Another approach to a loop antenna is the circular antenna shown in Fig. 4.29. This unit may, in a sense, be considered as a circular folded dipole. In a folded dipole, shown in Fig. 4.30A, the current is maximum at the center, diminishing gradually to zero at the two ends of the dipole. (In contrast, the voltage is minimum at the center, increasing gradually toward the ends.) If, now, we place conducting plates at either end of the folded dipole, we accomplish two things. First, the overall length of the dipole decreases due to the capacitive effect of the end plates. Second, the current variation from the center of the folded dipole to either end

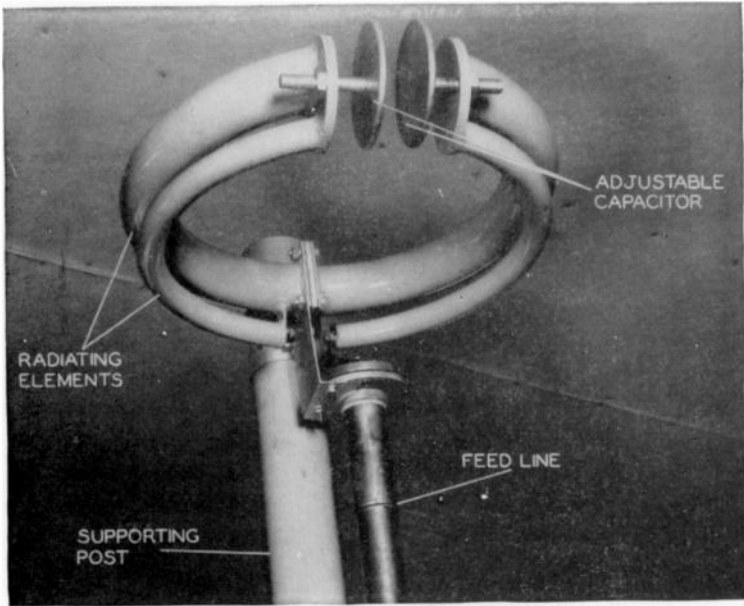


FIG. 4.29. A close-up view of a circular F-M antenna. (Courtesy G.E.)

is not as great as it is in the unloaded folded dipole. (See Fig. 4.30B.) Finally, by bending the dipole rods into circular form, a loop antenna is produced. The current distribution around the loop is now fairly uniform, producing a uniform field at all points around the antenna.

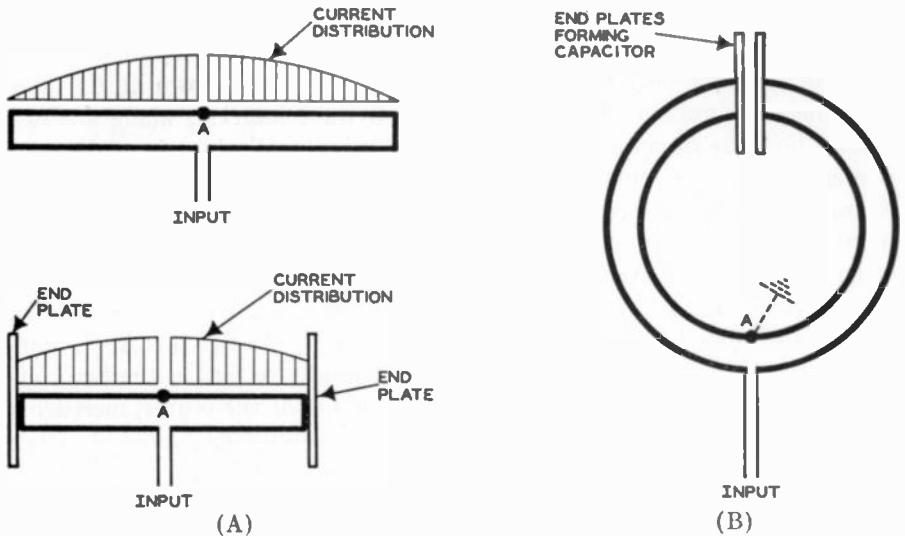


FIG. 4.30. Development of the circular F-M antenna from a folded dipole.

By adjusting the separation between the two end plates, which form a capacitor, the resonant frequency of the antenna can be varied over the frequency range for which the unit is designed. This can be done without any change in the main physical structure of the antenna loop.

Since the voltage at the center of the folded dipole (*point A* of Fig. 4.30B) is very low, connection of a ground here will not noticeably affect the operation of the unit. Such connection is made in actual installations, with *point A* attached directly to the supporting pole. A transmission line then brings the transmitter power to the input terminals.

To minimize vertical radiation from the array, four to six circular antennas are stacked vertically.

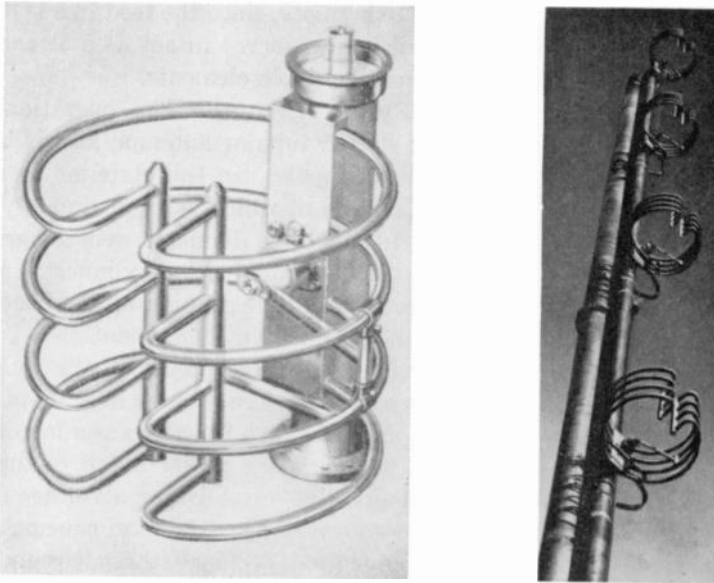


Fig. 4.31. Broadband dipole F-M broadcast antenna. *Left*: A single bay. *Right*: Four bays, vertically stacked. (Courtesy RCA)

RCA Broadband Antenna. The RCA broadband antenna, shown in Fig. 4.31, consists of four dipoles bent into the shape of a ring. Instead of connecting plates to the ends of the dipoles to form a resonating capacitor, the RCA unit achieves the same purpose by curving in the ends of the dipoles.

Of the four dipoles used for each section, only the center two are powered. The remaining two end elements are parasitic; their purpose is to broaden the response of the driven elements so that uniform transmission can be achieved over any channel in the 88-108 mc band. This is useful in multi-

plex broadcasting where two signals are transmitted simultaneously. Each shares the band normally utilized by one signal.

To insure uniform transmission of both signals, a wide-band antenna is needed, one that can provide close matching to the transmission line over the full 150-kc channel. When only a single signal is being transmitted, the matching requirements are less severe.

A 50-ohm coaxial transmission line feeds each broadband section. Both driven elements are fed in parallel, with the feedpoint located as shown in Fig. 4.31 (*left*) to obtain the proper impedance match. Furthermore, since the feed line is mechanically rigid, it also serves to act as a strengthening support for the dipole elements.

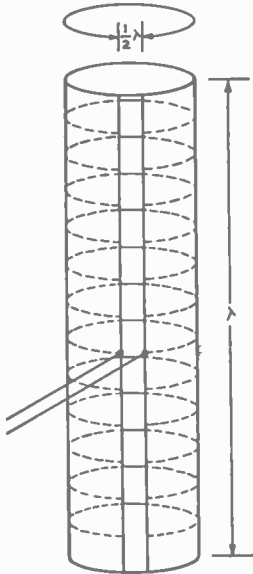


FIG. 4.32. The Pylon antenna may be considered as being made up of a large number of circular elements, each of which radiates.

The Pylon Antenna. The operation of the Pylon, or slotted tubular antenna, also is based on the loop. In appearance this antenna is a cylinder, each section of which is approximately 13 feet high and 19 inches in diameter with a narrow slot cut from top to bottom. The cylinder is fed by a single transmission line running up the inside of the cylinder to the midpoint of the slot.

To understand the operation of the Pylon, consider it to be composed of a large number of circular elements, each of which functions as a loop antenna.

(See Fig. 4.32.) When power is fed to the slot, it functions as an open wire transmission line, establishing a voltage distribution along the length of the slot as shown in Fig. 4.33, and causing currents to flow around the surface of the cylinder. Since these circumferential currents are very nearly in phase throughout the length of the cylinder, a field is radiated which is very nearly the same at all points in the horizontal plane. (See Fig. 4.34.)

Pylons may be stacked in one, two, four, or eight sections, the gain increasing with the number of sections. Furthermore, as the number of sections increase, the power radiated in the vertical direction decreases. It is important that each Pylon section is fed in phase, and this is accomplished by using suitable lengths of transmission lines.

Mechanically the Pylon cylinder is rolled from a single sheet of metal. It is capped on each end with a cast base to give the unit greater mechanical stability.

The Square-Loop Antenna. The circular radiation pattern of the loop can be obtained even when the shape of the loop is not circular or the cross section of the radiator is not round. In fact, the loop elements can be square, rectangular, or even triangular, without modifying the essential electrical behavior of the loop. Advantage is taken of this in the square loop F-M antenna. (See Fig. 4.35.)

To understand the electrical and physical construction of the square loop antenna, consider the two half-wave radiators, *A-B*, and *C-D*, in Fig. 4.36A. Each radiator is fed from a common point, *E*, and since the distance, *E-A*, is equal to the distance, *E-C*, the instantaneous voltage polarity at points *A* and *C* will always be the same. Hence, the currents flowing in *A-B* and *C-D* will, at any instant, be flowing in the same direction as indicated by the arrows, producing the equivalent of an in-phase current all around the circular path.

If the feeder line *AEC* is exposed, it, too, will radiate, tending to destroy the desired circular radiation pattern of the loop sections, *A-B* and *C-D*. To prevent this, the arrangement shown in Fig. 4.36B is employed. A coaxial transmission line is used for the feeder, and sections *A-B* and *C-D* are made rectangular in shape, with their center hollow. To feed section *A-B*, the inner conductor from point *E* goes to point *G*, where it enters at the center of section *C-D* and travels through the interior of this section until it reaches point *A*. At this point it makes electrical contact with section *A-B*.

To feed section *C-D*, another inner conductor extends from point *E* to point *H*, enters at the mid-point of section *A-B*, travels through the interior of *A-B* until point *C* is reached where electrical connection is made to section *C-D*. Thus, both feed lines from *E* are shielded until they actually reach the rectangular radiating sections, *A-B* and *C-D*.

The current distribution from end to end of each section (*A-B* or *C-D*) is maximum in the center, diminishing to a minimum at either end. By mak-

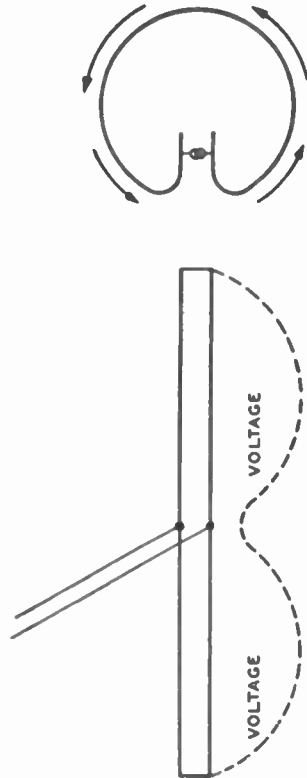


FIG. 4.33. Power is fed to the slot which functions as a transmission line. Currents flow around the circular path as shown.

ing the length of each section somewhat less than a half wave length and utilizing the end effect of each section, the difference between maximum and minimum current is only 12 per cent. This results in a radiation pattern which is very nearly circular.

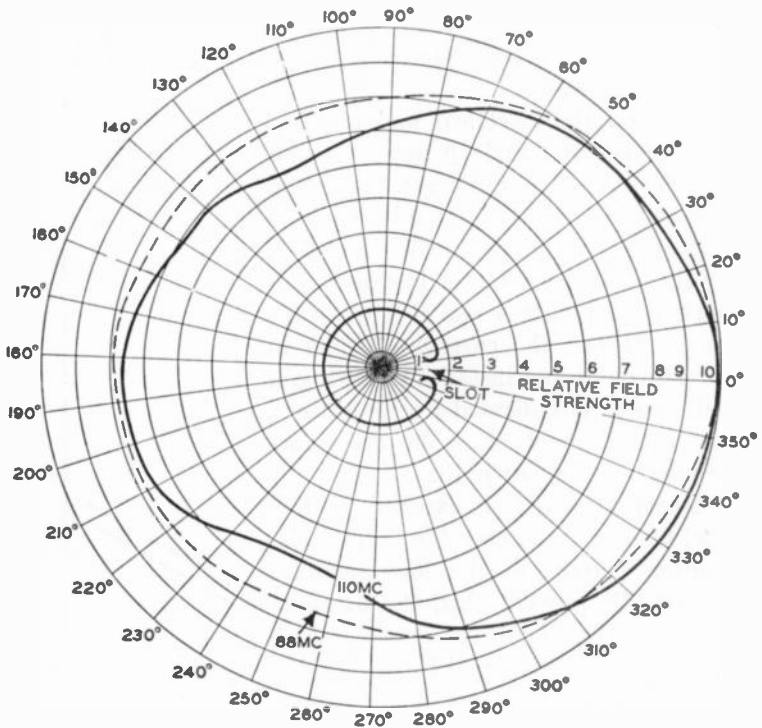


FIG. 4.34. The horizontal field pattern of the Pylon for frequencies at either end of the F-M band.

While the current at the center of each section is a maximum, the voltage here is very low, and actually grounding this point to the supporting structure does not noticeably affect the operation of the antenna. The outer conductor of the feeder coaxial cable also connects to the radiating section at this mid-point, while the inner conductor extends into the center of the radiating section.

In one commercial application of this F-M antenna, the number of sections is not two but four, permitting the circumference to be increased to approximately two wave lengths. The electrical transition from the two sections discussed above to four sections is shown in Fig. 4.37. As seen, the instantaneous current along the radiating members are always in the clockwise or counterclockwise direction.

The Turnstile Antenna. An F-M antenna which has been widely used is the turnstile antenna shown in Fig. 4.38. Essentially this antenna consists of two half-wave dipole antennas, positioned at right angles to each

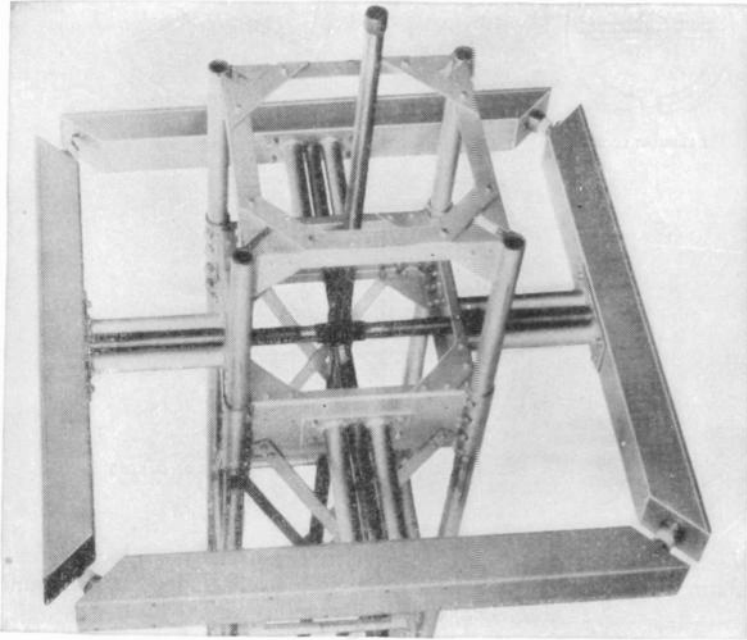


Fig. 4.35. A square-loop F-M antenna.

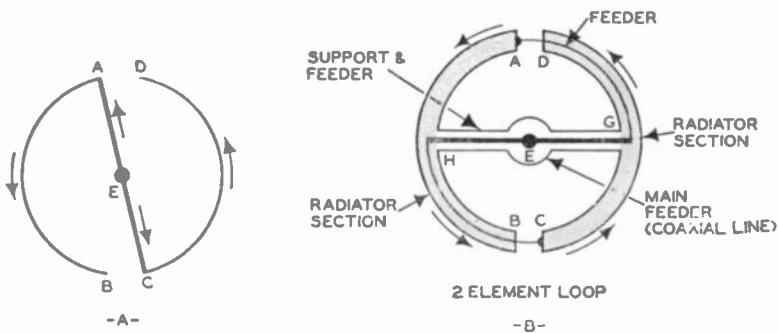


Fig. 4.36. Two steps in the development of the square-loop antenna.

other, and fed 90° out of phase. The radiation pattern from each dipole is the familiar figure-8 pattern. However, by positioning the two half-wave dipoles at right angles to each other and then feeding them as indicated, we

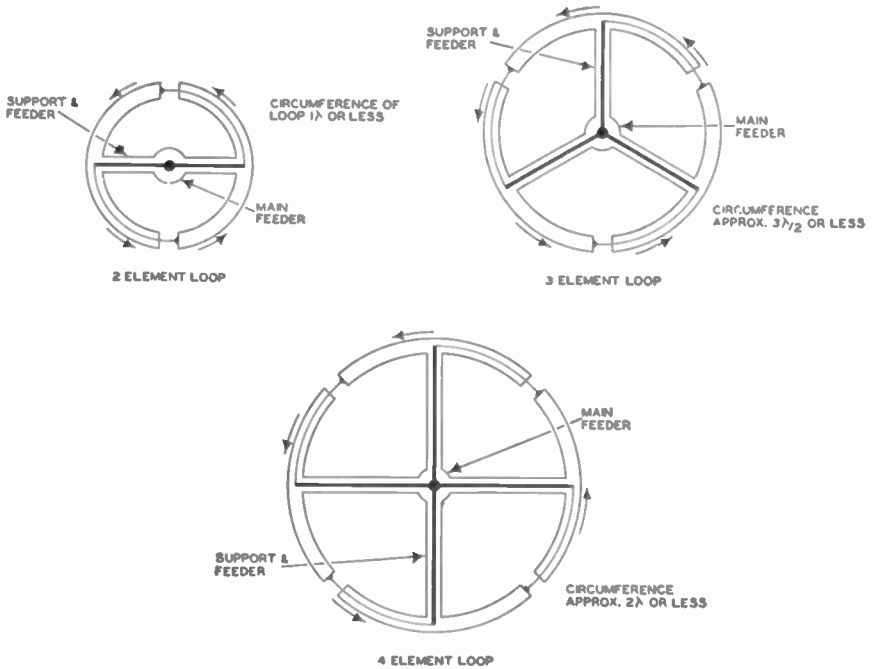


FIG. 4.37. Evolution from a two element to a four element loop. For ease in understanding, the loops are shown as circular, although in the F-M antenna of Fig. 4.35, the loop is rectangular.

obtain an overall pattern which is closely circular. Furthermore, the energy radiated by this system is horizontally polarized, which makes it suitable for F-M broadcasting. Finally, to minimize radiation in the vertical plane, a series of 3 to 5 turnstile units are mounted above each other, spaced by distances of approximately one-half length.

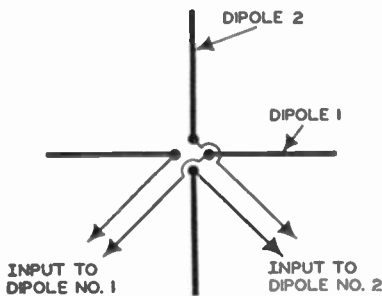


FIG. 4.38. The turnstile antenna. The input voltages to the two dipoles differ in phase from each other by 90° .

The turnstile, as originally designed, contained simple dipoles as the radiating elements. However, folded dipoles may be used as well and, in fact, have so been used. (See Fig. 4.39.) The folded dipole possesses a higher point input impedance than the simple dipole and for this reason is more desirable from the

standpoint of impedance matching and power transfer from the transmitter to the antenna system. Furthermore, the transmission line feeding system for the folded-dipole turnstile array can be designed so that the center of

the folded dipole may be grounded to the supporting mast, affording protection against lightning and static discharges. The radiation pattern of a folded dipole is the same as that of a plain dipole.

The super-turnstile antenna represents a further improvement in turnstile antenna design. In place of the simple dipole arms, solid radiating sheets approximately one-half wave length high are used. (See Fig. 4.40A.) In actual construction, the solid sheets are replaced by an open framework as shown in Fig. 4.40A and 4.40C because it was found that essentially the same results could be obtained.

The operation of the super-turnstile antenna and the reason for its peculiar shape may be understood by considering that a pair of super-turnstile radiators form a large plane surface containing a slot. In Fig. 4.41 such a slot is shown, the length of the slot being one-half wave length long, with the ends of the slot grounded. If, now, we connect a generator across the center of the slot, the voltage potential which will be established along the slot is indicated by the distance of the dotted lines from the sides of the slot. It is evident that the voltage potential will be greatest at the center of the rectangular slot, gradually diminishing to zero at the ends of the slot. Currents will flow in the plane surface with a magnitude and instantaneous direction as shown by the arrows. This same distribution of voltage and current will be obtained whether a solid metallic sheet is used for the plane surface or an open framework.

The radiators are attached to the supporting pole at the top and bottom. The R.F. currents in the side rods will radiate proportionately to the length of each rod and to the current in that rod. Hence, to obtain uniform radiation from each rod, the length is made inversely proportional to the amount of current flowing in the rod. Thus, at the center, where the current distribution is high, the rod length is made small; to obtain the same amount of radiation near the ends of the slot, where the current is small, the length of the radiators is increased. In this way, a very close approximation to the radiation pattern of two simple dipoles, spaced one-half wave length apart, is obtained.

To obtain a circular radiation pattern, each super-turnstile bay contains

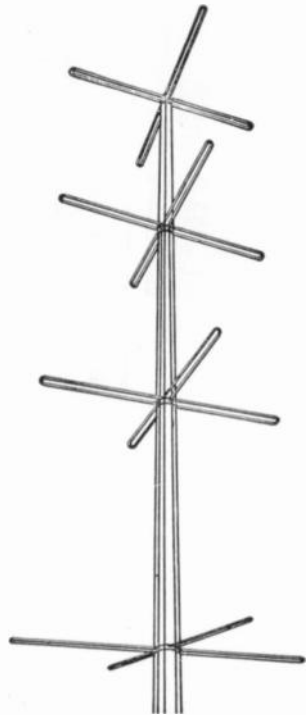
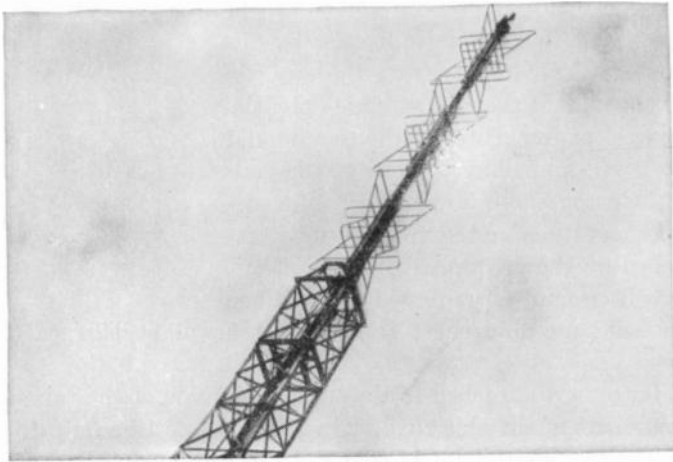
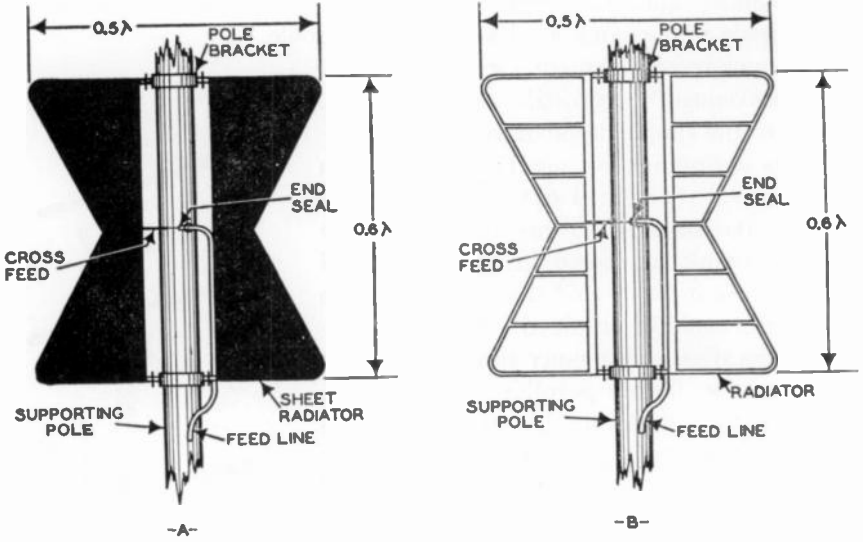


FIG. 4.39. A turnstile antenna array using folded dipoles.



-C-

FIG. 4.40. (A) The original form of the super-turnstile antenna using solid sheet radiations. (B) The present commercial appearance of the super-turnstile using an open framework. (C) An actual photograph of a super-turnstile antenna. (Courtesy RCA)

a pair of radiators facing north and south and a pair facing east and west. Energy concentration in the horizontal plane is achieved, as in previous ar-

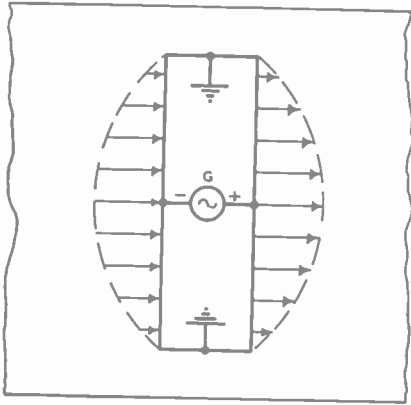


FIG. 4.41. Currents and voltages in the region of the half-wave slot in a large conducting surface.

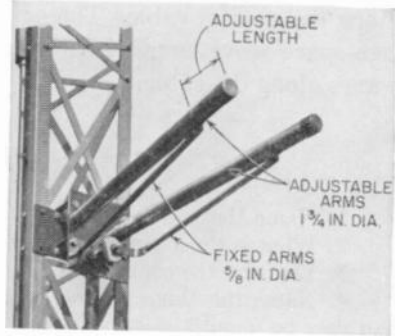


FIG. 4.42. The Multi-V, F-M broadcast antenna. (Courtesy Andrew Corporation)

rays, by mounting several bays above each other. Feeding of each pair of radiators is done at the center of the unit.

Multi-V Antenna. The Multi-V antenna is a lightweight transmitting array formed into the shape of a V. (See Figure 4.42.) This configuration gives a horizontal radiation pattern which is very nearly circular.

In order to achieve the desired input impedance, a folded dipole V is utilized. This gives rise to two sets of arms for each leg of the V. One of these arms is fixed and has a diameter of $\frac{5}{8}$ inches. The other arm of each set is adjustable, with a telescoping end section for frequency variation. Diameter of this arm is $1\frac{3}{4}$ inches. By this arrangement, a simple adjustment tunes the antenna to resonance anywhere in the F-M band without appreciably affecting the input impedance.

Another advantage of this tuning arrangement is that the two ends of the V are far enough apart to permit 5 kw or more of power to be fed into the array. Since this is basically a half-wave antenna, maximum R.F. voltage appears at the ends. In the V antenna, these points are separated sufficiently to prevent arc-over.

As commonly employed, several V sections are positioned one above the

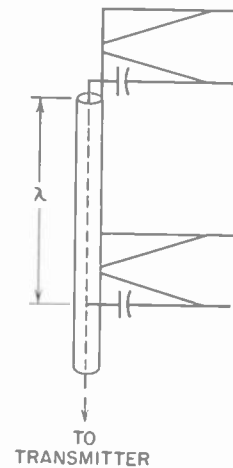


FIG. 4.43. Method of feeding the Multi-V antenna.

other. R.F. energy from the transmitter is then brought to each bay by means of a 51.5-ohm coaxial cable, as shown in Fig. 4.43. Three of the four elements of one V array connect to the outer coaxial conductor; the other conductor is capacitatively connected to the coaxial inner conductor. In order to feed the two bays in phase, they are separated by one wave length along the coaxial cable. The actual distance is somewhat less than one full free-space wave length because of the reduced velocity of electromagnetic waves along the cable.

PROBLEMS

1. Name the various categories into which radio waves can be divided.
2. What is the ionosphere and of what use is it?
3. Describe the convention regarding the polarization of radio waves.
4. Name the three general layers in the ionosphere. At what average heights can they be found?
5. Why is the ionosphere useful for long-distance communication? Which layer is responsible for the bending of 35-mc waves? Why?
6. What is the difference between reflection and refraction?
7. Name three conditions upon which ionospheric refraction depends.
8. What is meant by the critical frequency of an ionospheric layer?
9. What is the Sporadic E layer? What is its significance?
10. Does U.H.F. propagation differ from low-frequency radio wave travel? Explain.
11. Name the three methods whereby U.H.F. waves are propagated.
12. What is the range of frequencies that can be normally refracted by the ionosphere?
13. Discuss the line-of-sight method of propagating U.H.F. waves, deriving the equations applicable to this method.
14. What primary factors determine the amount of voltage developed in a receiving antenna?
15. What is the wave length of a 100-mc wave?
16. What would be the length of a half-wave dipole tuned to 106 mc?
17. A half-wave dipole antenna designed for 108 mc is to be used at 88 mc. By how much should its length be altered?
18. Name and sketch five different types of antennas that could be employed to receive F-M signals.
19. What precautions must be observed in choosing and installing a transmission line?
20. Define antenna gain and antenna directivity.
21. Illustrate a balanced and unbalanced input system.
22. Name and describe four types of transmission lines.
23. Discuss briefly the effects of impedance mismatch in the antenna system.
24. Describe two methods of matching impedances between the antenna and the transmission line.
25. What general factors determine the design and shape of F-M transmitting antennas? Discuss each item briefly.
26. Sketch the Multi-V antenna and describe its operation.

27. How is the amount of energy radiated vertically by an antenna minimized?
28. Explain the operation of the antenna in Fig. 4.29.
29. Compare the operation of the Pylon and square-loop antennas.
30. Explain the operation of the turnstile and super-turnstile antennas.

Chapter 5

R.F. TUNERS FOR F-M RECEIVERS

Introduction. In conventional sound broadcast receivers, it is current practice to tune the set by means of variable air capacitors or by changing the position of iron cores within certain inductances. These units are ganged together, enabling the use of one dial to tune several stages simultaneously. At the standard broadcast frequencies, 550 to 1700 kc, and even at the higher short-wave frequencies, the coils and capacitors used are quite substantial in appearance. As we increase the operating frequency of the receiver, however, the units decrease in size until, at the F-M band, 88 to 108 mc, the coils contain fewer than 10 turns and the capacitors have 3 or 4 small plates. Furthermore, extreme care must be taken in parts placement and in circuit wiring. Consider the latter, for example. Any length of wire, regardless of shape, contains a certain amount of inductance. This inductance is present at the lower frequencies just as concretely as it is at the higher frequencies. That it is important in one instance and not the other is merely due to its relationship to the rest of the circuit, particularly the tuning networks. At the low frequencies, the inductance introduced by the wiring is negligible when compared to the tuning coil's actual inductance. Hence, it may be (and is) disregarded. At the higher frequencies, however, this same inductance of the wiring forms a significant percentage of the tuned-circuit inductance and, as such, cannot be ignored. It is here that the difficulty arises. Not only must the connections be made as short as possible, but we find that there is considerable inductance present in the capacitor structure itself. Microphonic howls and inductive coupling between the various sections of the capacitor structure (and hence between the stages controlled by these sections) are two of the undesirable results that occur.

In present-day F-M receivers, tuning by means of suitable variable capacitors and fixed inductances is still widely used and, when properly constructed, will yield good results. In addition, there has been developed a

whole new group of tuners which are designed to meet the altered conditions imposed by the higher frequencies with greater efficiency and more gain. It is the purpose of this chapter to investigate these newer types of tuners so that the serviceman, when working with F-M receivers, will be able to service them intelligently. We are entering an era where service ability is more closely linked to circuit understanding than it has been previously.

Coaxial Tuning Capacitor.

Variable capacitors can take several forms, the most common of which is shown in Fig. 5.1. However, one F-M receiver manufacturer has developed a variable capacitor possessing a coaxial-type construction. (See Fig. 5.2.) One plate of the capacitor is formed by an outer cylinder which is heat shrunk onto a glass cylinder. This glass serves as the dielectric of the unit. The other plate of the capacitor is a movable cylinder whose diameter is slightly less than that of the glass

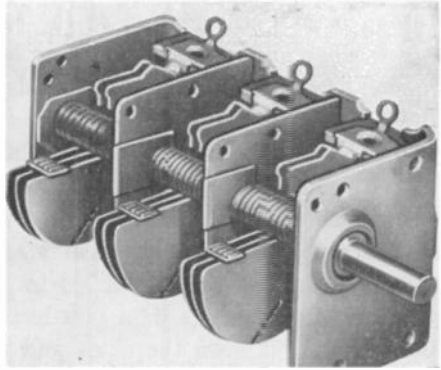


FIG. 5.1. A 3-gang variable capacitor employed in F-M receivers.

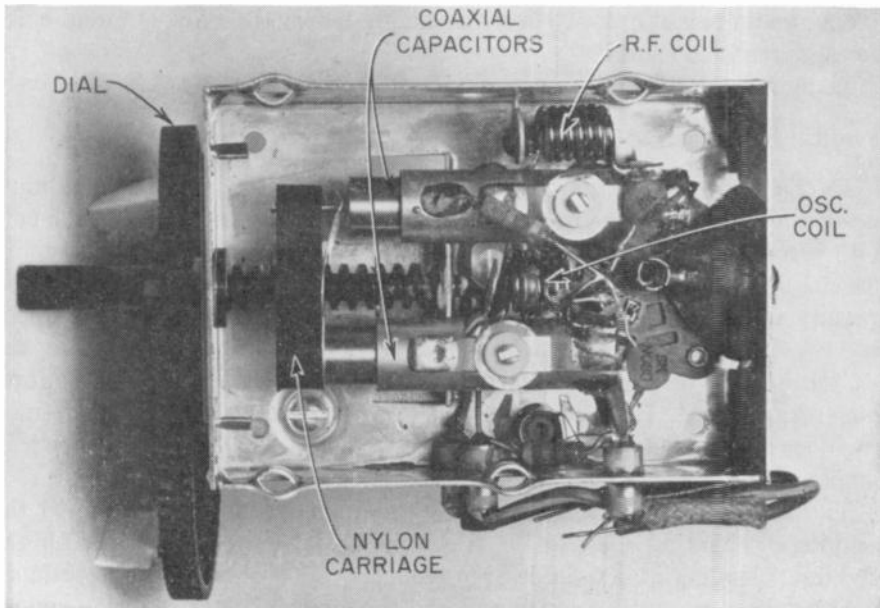


FIG. 5.2. A 2-section coaxially-constructed tuning capacitor. (Courtesy Granco)

cylinder. This enables it to be moved back and forth by a nylon lead screw to which it is attached.

One such capacitor is employed in the radio-frequency (R.F.) amplifier tuning circuit and another in the oscillator tuning circuit. Both capacitors are mechanically ganged together by a nylon carriage so that they can be varied in step across the F-M band.

Each variable tuning unit has a small compression-type variable trimmer placed across the gap formed by the two metal cylinder sections. This permits the circuit to be aligned at the high-frequency end of the band to achieve the proper tracking.

Permeability Tuning. There are several methods by which the inductance of the coils can be varied in order to tune the receiver. We can use brass or copper slugs or a powdered-iron, permeability-tuning element. A comparison between the slugs and the permeability element indicates that the inductance change, for a definite decrease in Q , is greater with the iron core than with the slug. Furthermore, a coil tuned with a slug possesses a wider bandwidth than one which has an iron core. The reason for this is due to the eddy current loss within the slug, which acts as a shorted turn.

Q of a tuning circuit is a measure of the sharpness of the response curve of that circuit. A tuned circuit whose Q is high will have a sharper (i.e., narrower) tuning curve than a circuit whose Q value is low. Thus, the selectivity of a high- Q circuit will be better. Also, high- Q tuning circuits provide a greater signal output (over the tuning band) than low- Q circuits, for the same received signal.

In most tuning circuits possessing coils and capacitors, Q is effectively governed by the coil. Hence, Q is given as $\frac{\omega L}{R}$, where $\omega = 2\pi f$.

In the design of a high-frequency permeability tuner, cognizance must be taken of the stray or circuit capacitance which is present across the coil. With careful design, the lowest value of the capacitance is close to 15 mmf. The small value, in conjunction with the relatively high value of inductance, presents a high L/C ratio. Since gain is directly proportional to the impedance presented by the resonant circuit and this, in turn, is governed by the L/C ratio of the network, we find that a high gain is available with the foregoing arrangement. However, when this is put to practical use in the receiver, it will be found that the R.F. selectivity is poor. The result, as shown in Chapter 7, is a large number of spurious responses.

The reason for poor R.F. selectivity at high frequencies is due to the damping or shunting effect of the R.F. tube input. A nominal value for the tube input loading at 100 mc is approximately 3500 ohms. This impedance is across the permeability-tuned coil and the stray 15-mmf shunting capacitance. At resonance, the impedance of the tuned circuit is given by

$$\frac{Q}{\omega C} \quad (14)$$

where Q = the Q of the coil
 C = 15 mmf

The coil Q is determined by the dimensions of the coil and is approximately 90. At 100 mc, the value of the foregoing expression, after substituting the figures listed, is 10,000 ohms. Since the tube shunts the tuned circuit, the total impedance of the two is 2600 ohms. Practically, then, the impedance of the tuned circuit, which is what the incoming signal "sees," is 2600 ohms. Hence, we can set

$$\frac{Q}{\omega C} = 2600 \text{ ohms}$$

Since ω and C have not changed, the lowered value (from 10,000 to 2600 ohms) must be due to the lowering of the effective Q of the network. Solving, we obtain an effective value of 23 for Q .

Now, let us increase the total capacitance from 15 mmf to 30 mmf by the addition of more capacitance. The Q of the coil remains approximately at 90 in spite of a decrease in the number of turns due to the added capacitance. Solving for $Q/\omega C$ now, we obtain a value of 5000 ohms. This value, in parallel with the tube (3500 ohms), produces a total input impedance of 2100 ohms. The effect on the overall Q is to bring it to an effective value of 38. Thus, although raising the shunt capacitance decreased the impedance (from 2600 to 2100 ohms), it did raise the effective Q of the circuit, in this instance by 65 per cent. The result is a slight loss in gain but a substantial increase in selectivity. Hence, in the receiver, a small capacitance is shunted across each coil.

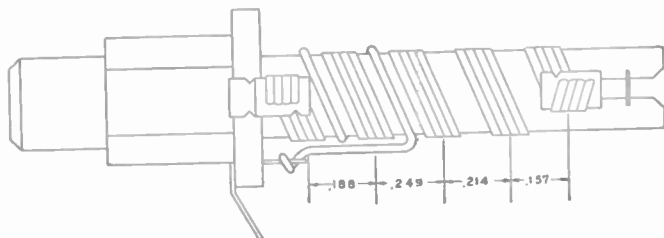


FIG. 5.3. A four-wire permeability-tuned coil. The variable pitch produces a linear characteristic.

In the particular permeability-tuned coils shown in Fig. 5.3, four-strand tinsel wire was used to retain wire flexibility in winding. The use of four parallel wires is due to the difficulty of obtaining the necessary inductance

change with the small number of turns on the coil. It was just pointed out that, for better selectivity, the shunt capacitance is increased and the coil inductance is lowered. Hence, the ratio of inductance change for a certain core travel decreases, and the frequency band is not covered by a complete travel of the core within the distance allotted to it on the dial. To overcome this difficulty, the width of each turn is increased. As the wire width increases, the tuning range is increased. A similar result may be had by winding the coil with parallel wires. This accounts for the four conductors shown in Fig. 5.3.

Coil Spacing: A serviceman, called upon to repair one of these coils, would be completely unable to return the set to its proper operating condition unless he wound the coils correctly. If the coils are closely inspected, it

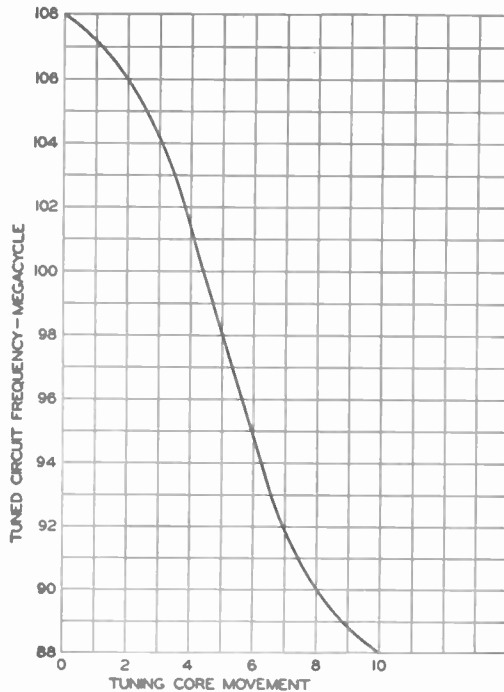


FIG. 5.4. The nonlinear shape of the tuning curve when the winding of the permeability-tuned coil is uniform in pitch.

will be found that the winding is not uniform in pitch. Fig. 5.4 illustrates the shape of the tuning curve for a permeability unit if the winding is made uniform in pitch. The nonlinear slope at each end occurs because the incremental inductance variation is maximum when the leading edge of the

core is in the center of the coil and minimum at the start or finish of the core travel.

To straighten out the curvature at the ends of the characteristic, the coil turns are bunched at the ends and spread out in the center. This increases the inductance variation (with core movement) at each end and decreases it in the middle. To insure further that an almost linear characteristic response is obtained, only part of the complete tuning curve is used. The coil and core lengths are made larger than necessary, and the core, over the band desired, is never out of the coil. The curve of Fig. 5.5 is the result.

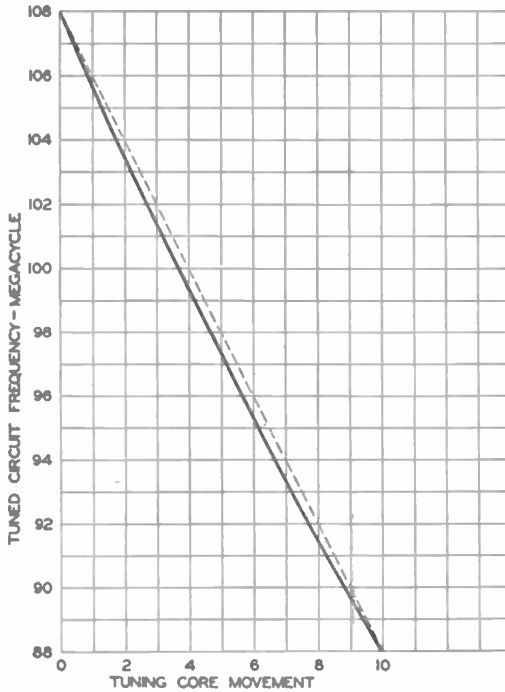


FIG. 5.5. The improved tuning curve (solid line) when the turn spacing is made logarithmic. The dotted line, which is exactly linear, is used for comparison.

In the mass production of permeability-tuned coils, it must be expected that a certain variation between units will occur. Since the R.F., mixer, and oscillator coils are tuned simultaneously by one mechanism, some method must be available which will permit a preliminary adjustment in order that all the coils track. The idea of using a variable trimmer capacitor is possible, but a variable capacitor is highly susceptible to change with temperature and would be highly detrimental to receiver stability. It was discovered, how-

ever, that, by means of a single adjustment of the initial core position with respect to the coil winding, it would be possible to compensate for small differences. Thus, when the coils are mounted and tested, any difference in tracking is compensated for by a single initial advancing or retarding of the iron core of each coil.

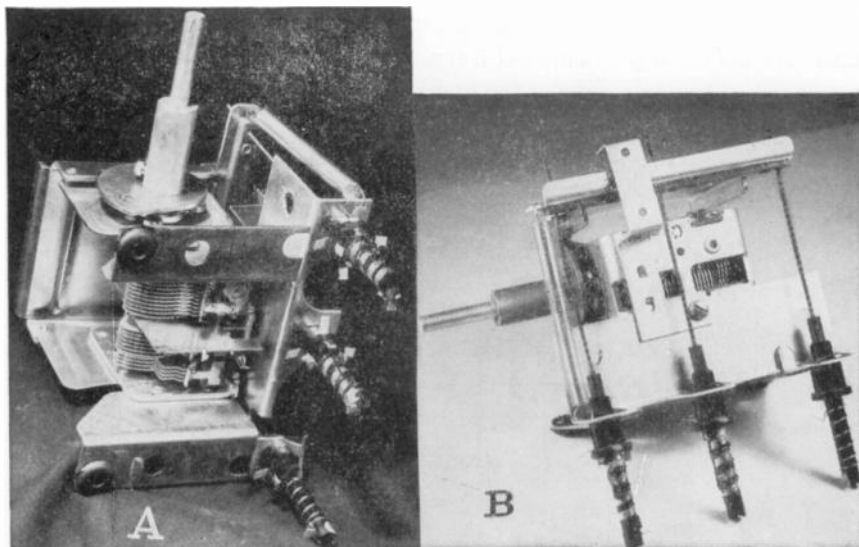


FIG. 5.6. (A) Receiver tuning assembly showing the mounting of the permeability coils. (B) Underside view.

Photographs of the entire tuning assembly are shown in Fig. 5.6. The F-M antenna, mixer, and oscillator coils are mounted on a bracket fastened to the side of a conventional variable capacitor. The latter is used, in this receiver, to tune the A-M broadcast and short-wave bands. A cam is mounted on the shaft of the variable capacitor and operates a rocker arm which moves the iron tuning cores in the three high frequency coils. The unit is very compact, reducing all stray capacitance and inductance to a minimum.

The front-end section of an A-M F-M receiver employing permeability tuning is shown in Fig. 5.7. The dotted lines represent the mechanical coupling which exists between the cores of the F-M tuning coils (L_2 and L_3) and the two variable capacitors in the A-M section. One variable capacitor tunes the A-M antenna coil (a rod antenna); the other tunes the A-M oscillator coil, L_7 .

In somewhat greater detail, the F-M R.F. amplifier is connected with the signal applied to the cathode and the grid placed at ground potential.

This is known as a grounded-grid amplifier and it possesses several advantageous characteristics for high-frequency operation. This is discussed more fully on page 103.

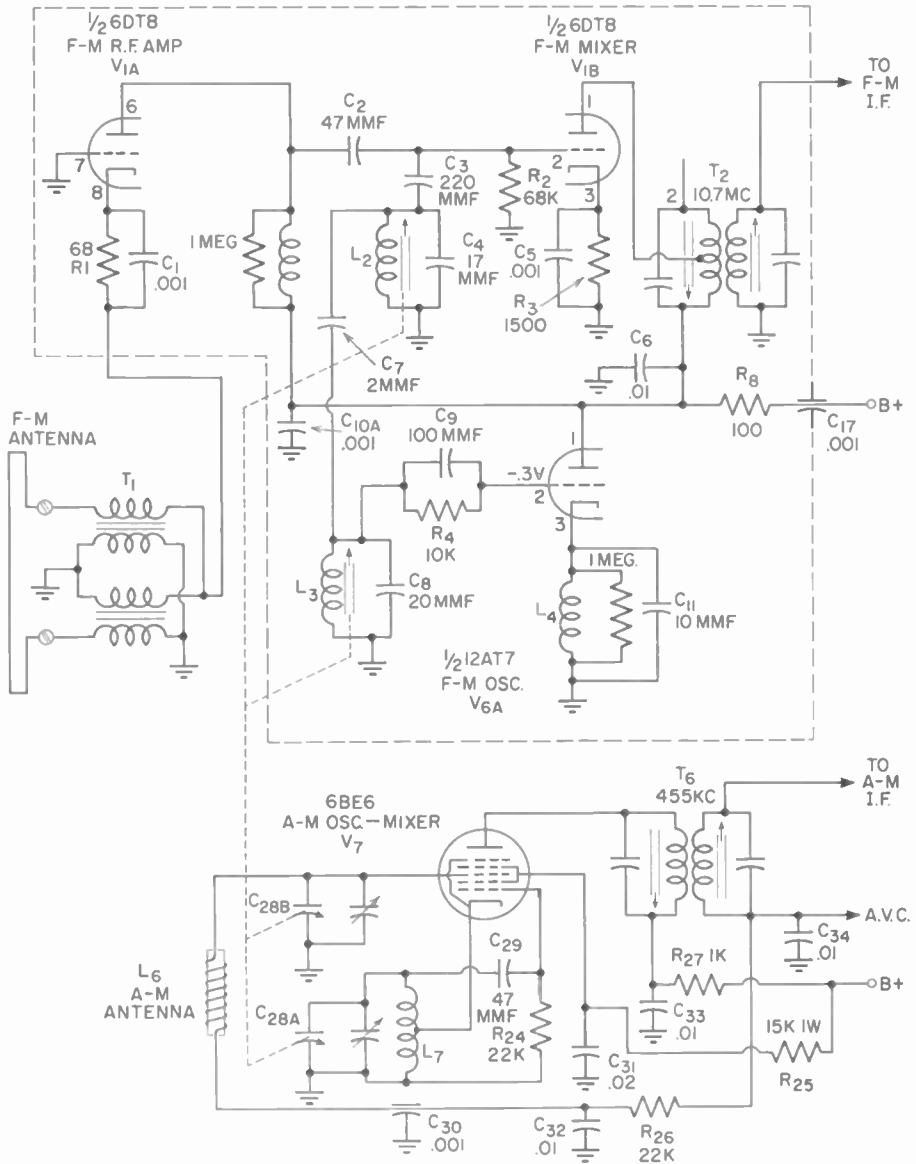


FIG. 5.7. The front-end stages of an A-M, F-M receiver employing inductance tuning on the F-M band.

The input circuit is untuned, possessing only a matching transformer to match the 300-ohm folded dipole, which is balanced, to the much lower unbalanced impedance of the R.F. amplifier cathode. R_1 establishes grid-to-cathode bias for the tube; C_1 bypasses the R.F. signal currents around R_1 .

The plate circuit of V_{1A} contains an untuned coil which acts as a wide-band load for the tube. The first signal tuning circuit, L_2 and C_4 , is in the grid circuit of the mixer, V_{1B} . L_2 is permeability-tuned and it, in conjunction with C_4 , establishes the input frequency to be received. The mixer grid circuit also receives a portion of the oscillator voltage by way of C_7 . The two signals combine in V_{1B} to produce the difference or I.F. signal (10.7 mc). This is then coupled to the I.F. amplifiers.

The F-M oscillator is of the ultraudion variety, with the internal capacitances of the tube playing a significant role in its operation. The tuning circuit, L_3 and C_8 , is essentially located between grid and plate. This can be seen by noting that the plate is at R.F. ground potential, placed there by C_{10A} . However, since the plate and cathode cannot both be at the same R.F. potential (and still have the circuit function), a choke coil is placed in the cathode lead.

L_3 is permeability-tuned. By mechanical coupling, its core moves in step with the core of L_2 so that both circuits will track over the dial.

The A-M section of the circuit uses a 6BE6 as a combination oscillator and mixer. The cathode, grid No. 1, and grids 2 and 4 (tied together internally) form the oscillator section of the converter. Grid No. 3 is the element to which the arriving signal is fed. The two voltages mix (by their effect on the same electron beam) and a difference frequency of 455 kc is produced. This is then transferred to the I.F. system. (The sum of these two frequencies is also developed, as well as various harmonics, but none of these are accepted by the I.F. system.)

Coaxial Tuners. A variety of transmission line tuners have been employed for F-M because of the high- Q values which they are able to develop. However, before we actually examine these units, let us briefly review the theory of transmission lines.

Transmission Lines: Every radio man is familiar with the conventional inductances, capacitances, and resistances, the types of which can be purchased in any radio shop. These units are said to be lumped because each is complete and distinct from the other.

Now, if two wires are mounted close together, using one to conduct the current away from the source of voltage and the other to bring the current back, it would be found that the source of voltage "sees" not just two wires with some resistance in them, but rather a complex impedance which, when broken down, will reveal inductance and capacitance, too. The last two com-

ponents are not visible upon a physical inspection of the wires, yet electrically they exist.

Of the two quantities, the presence of the capacitance can perhaps be more readily understood since, by definition, a capacitance is formed whenever two conductors are separated by a dielectric. In this instance the dielectric is air, although it may be any nonconductor. Since the dielectric is never a perfect nonconductor, it is represented schematically by placing a large resistance across the condenser, this denoting leakage.

The presence of the inductive reactance is much more difficult to explain. It is best to go back to the concept of magnetic lines of force and attack the problem from that angle. Inductance may be considered proportional to the number of magnetic lines of force per unit ampere that encircles a wire carrying the current. By starting with this idea, the inductance per unit length of any wire or system of wires can be developed.

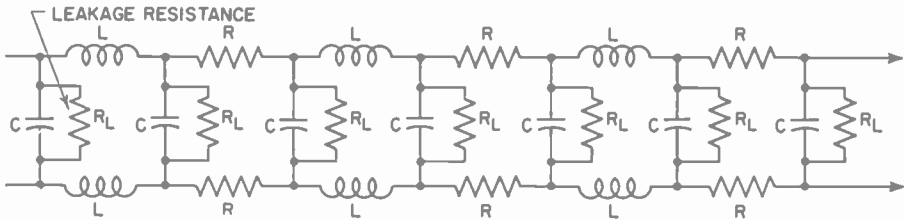


Fig. 5.8. The electrical components of a transmission line.

To represent the foregoing inductance, capacitance, and resistance in schematic form for analytical purposes, a diagram such as shown in Fig. 5.8 is often used in engineering books on transmission lines. Although the various components are shown separate and distinct from each other, they are actually distributed evenly along the line. It is only because of our inability to show these components as they really are that we resort to this method.

Since the transmission line contains exactly the same three components as any ordinary resonant circuit, it is reasonable to expect that the same results can be obtained from these as from a resonant circuit. Hence, the length of the transmission line determines the frequency at which it will operate.

In common use today are three types of transmission lines: the twisted-wire, the parallel-wire, and the coaxial transmission line. In the twisted-wire form, the two wires forming the line are twisted about each other, in the same manner as a lamp cord. In the next type, parallel-wire lines, the two wires are kept parallel and equidistant to each other, usually by means of spacers or by imbedding both wires in the dielectric material. The latter is

true of the polyethylene transmission line used as lead-in cables for F-M and television antennas. Finally, there is the coaxial cable, shown in Fig.



FIG. 5.9. A coaxial cable. D_1 is the inside diameter of the outer conductor; D_2 is the outside diameter of the inside conductor.

5.9, in which one conductor is placed inside and at the center of an outer conductor.

Transmission Line Tuners.

In choosing a transmission line, there were two factors to consider. First, the length of the line, in order to obtain the proper impedance at the operating frequencies, and, second, the best possible Q . It has

been determined that the maximum impedance results when the outer to inner conductor ratio is 9.2 to 1. Actually, a 10 to 1 ratio is used with very little decrease in impedance. For this particular line the expressions governing the length of the line in order to give the required inductance and the Q are as follows:

$$L = 0.0117l \text{ microhenries}$$

$$Q = 0.802D_1 \sqrt{F}$$

where L = the inductance presented by the length of line used

l = length of the coaxial line

D_1 = diameter of inner conductor

F = frequency at which the line is to be used

In common with conventional tuning circuits, one of the circuit components must be made variable in order that a definite range or band of frequencies be covered. It would be possible to resonate the inductance of the transmission line with a variable capacitor placed across the end of the line. However, since the magnitude of the inductance is low, the use of a gang capacitor—one section for each transmission line tuner—would introduce too much additional inductance to prove useful. We can, however, resonate the inductance of the transmission line with a fixed capacitor, and then vary the inductance of the line by means of a powdered-iron core. In other words, we would have permeability-tuning, but in a form which is substantially different from the previous unit.

The coaxial line, with an iron core, is shown in Fig. 5.10. The iron core is mounted on a threaded rod and its position in the coaxial line is adjusted by turning the head or top of the threaded rod. The Q of the tuner, with the iron core, at 100 mc is 335. The tube load reduces the effective Q of the circuit to 195 and this is approximately its final value, although the antenna loading lowers the Q somewhat. In the same diagram is the equivalent conventional circuit.

An inexpensive, high-quality, concentric capacitor was developed for this

particular unit. This is also shown in Fig. 5.10. The capacitor is of silver on mica construction. A highly desirable feature of the tuner is that the unit can be constructed as a mechanical assembly, without the necessity of electrical checking. With materials that are commercially available, the variation in inductance as calculated by the expressions previously given, is

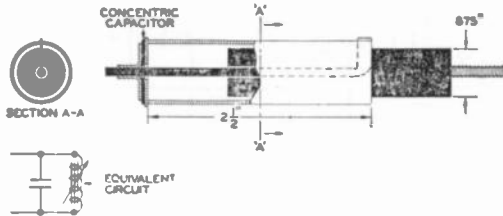


FIG. 5.10. The construction of the coaxial tuner. The equivalent coil and capacitor is also shown.

only ± 0.5 per cent. The coaxial line and capacitor type of construction forms a very compact unit. This reduces considerably the importance of the inductance present in leads connecting the tuned lines in the circuit. At 100 mc, the problem of confining the desired signal pick-up to the antenna terminals of the receiver is quite difficult. Chassis pick-up, for example, tends to degrade the image and adjacent channel attenuation. However, when the entire unit is confined within the coaxial cable, the undesired pick-up becomes negligible. Removing the antenna, in the presence of a strong signal, completely kills the receiver output.

One application of the tuner to a superheterodyne A-M F-M receiver is shown in Fig. 5.11. The same tubes, V_1 and V_2 , function for both A-M and F-M signals. However, by means of a two-position switch, different tuning circuits are brought in for A-M and F-M reception.

For F-M operation, switch S_1 is turned to the "F-M" position. F-M signals, entering the input terminals, encounter L_2 , the F-M antenna coil. This is the initial coaxial line tuner. The signals accepted by L_2 are applied to the control grid of V_1 . Here they are amplified and then developed across L_5 , a small R.F. choke.

The second coaxial tuning network, L_3 , obtains the signal from L_5 . L_3 is connected to grid No. 3 of V_2 , a 6BA7. This same tube also possesses an oscillator circuit between grid No. 1, the cathode, and grids 2 and 4. These two signals mix (through the electron stream) and develop a 10.7-mc difference frequency for the I.F. system. The F-M oscillator coil is a third coaxial line, L_4 .

Tuning across the F-M band, 88 to 108 mc, is achieved by mechanically coupling the iron core of each coaxial unit. The three units are adjusted to

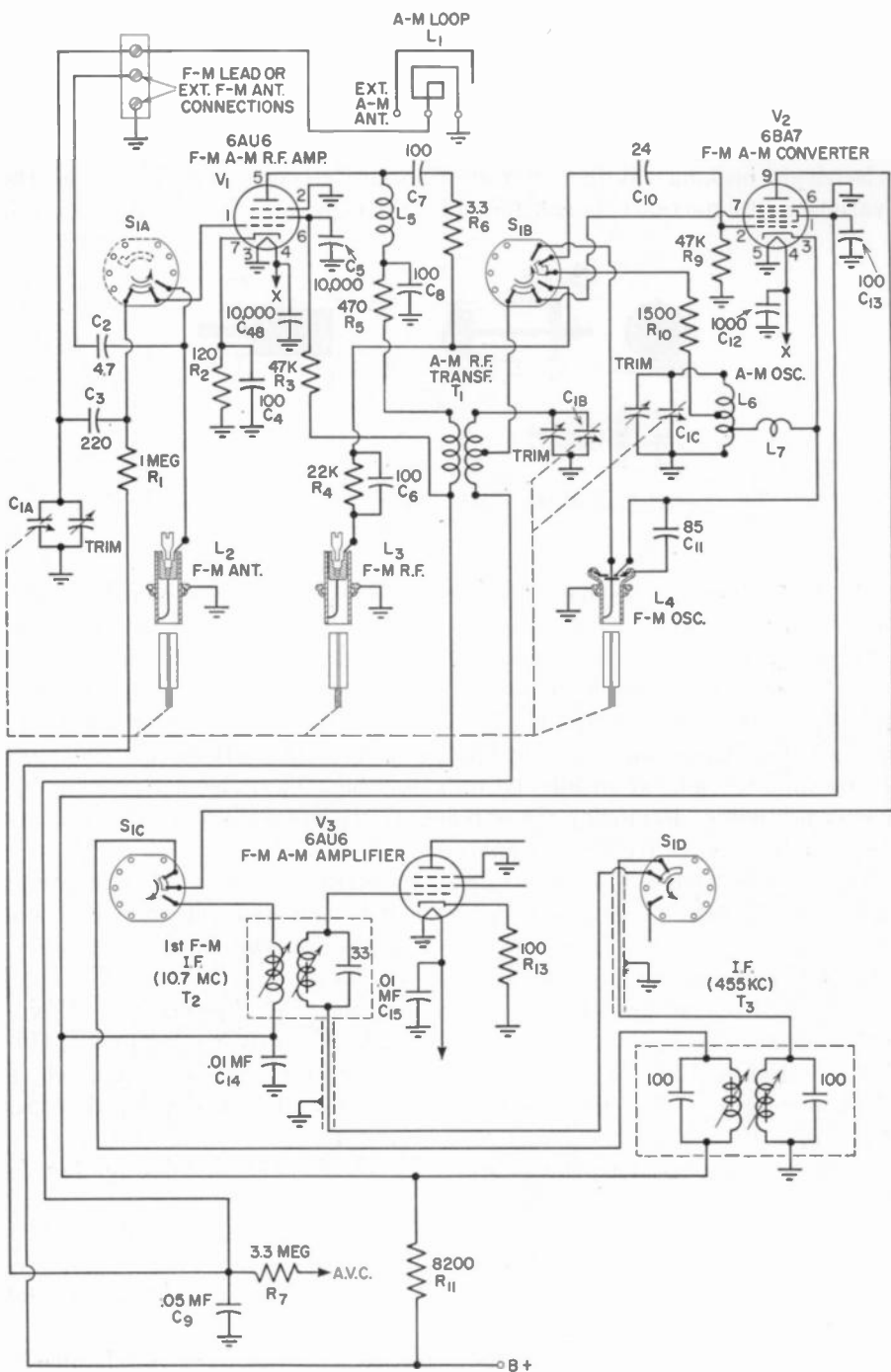


FIG. 5.11. The R.F. circuits of an A-M, F-M receiver using coaxial tuners for F-M reception. (Courtesy Motorola, Inc.)

track so that each will possess the proper frequency at each point on the tuning dial.

For A-M reception, switch S_1 is turned to the "A-M" position. The incoming signal is picked up by a small loop antenna and applied to grid No. 1 of V_1 . Tuning capacitor C_{1A} tunes the loop. The A-M output signal of V_1 appears now across T_1 . L_5 possesses too little inductance at the A-M R.F. frequencies to have any noticeable effect. C_8 , a 100-mmf bypass capacitor, in conjunction with resistor R_5 , do tend to reduce the amplitude of the A-M signal before it reaches T_1 . However, since the A-M signal is initially quite strong and since it is further amplified by V_1 , this decrease is useful in preventing V_2 from overloading.

From T_1 , the A-M signal is fed to grid No. 3 of V_2 where it combines with a local oscillator signal developed by L_6 . Again, V_2 serves as a converter, producing a 455-ke I.F. signal which is transferred to the following I.F. system.

In the output circuit of V_2 , the third section of S_1 (i.e., S_{1C}) brings in the desired I.F. transformer. When an F-M signal is being received, T_2 receives the 10.7-mc output of V_2 . When an A-M signal is being received, transformer T_3 is connected to the plate of V_2 .

The three A-M tuning circuits (loop antenna, T_1 , and L_6) are resonated to the proper frequency by a three-section variable capacitor. This unit is attached to the same mechanism employed to vary the iron core positions of L_2 , L_3 , and L_4 . In this way, only one tuning dial is needed for the entire receiver.

Parallel-Wire Transmission Line Tuners. Another approach to transmission line tuners, this time using parallel-wire lines, is shown in Fig. 5.12. The R.F. section consists of three sets of parallel-wire lines, each set having movable shorting bars which determine how much of the line is active. The shorting contacts are mounted on plastic bars and then attached to a common shaft. As the shaft is rotated counter-clockwise, the bars progressively short out more and more of the lines, raising their resonant frequency from 88 to 108 mc. (See Fig. 5.13.) To permit the three lines to track with each other, each line contains small end inductances and semi-fixed, temperature-compensated silver ceramic capacitors. At the high frequency end of the F-M band, the series capacitors are adjusted for

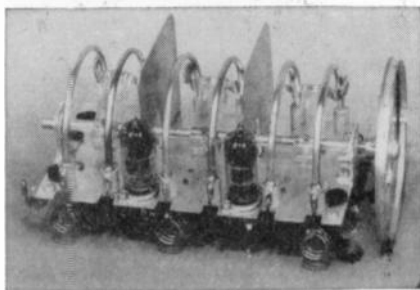


FIG. 5.12. A parallel resonant, line tuning assembly. (Courtesy Approved Electronic Instrument Corp.)

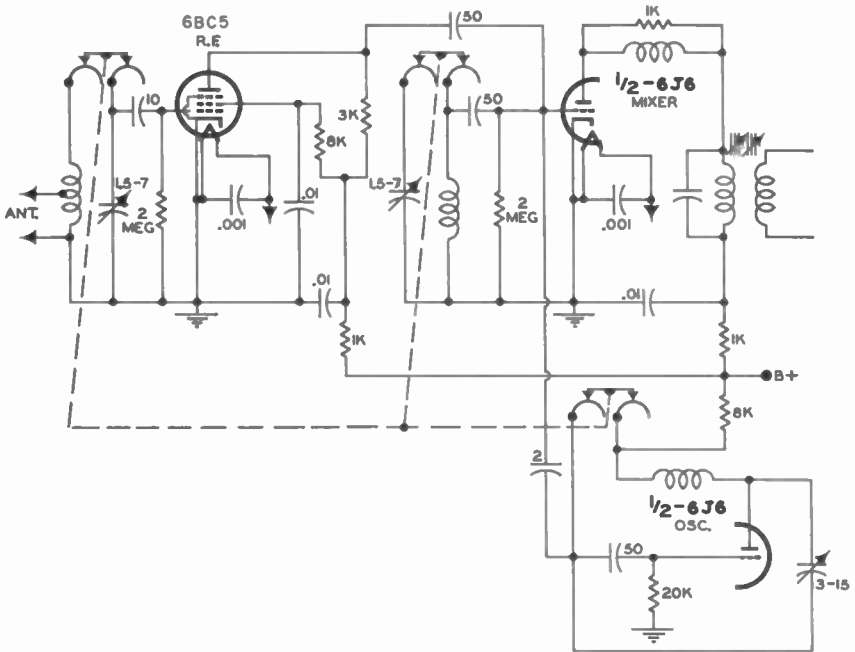


FIG. 5.13. Schematic diagram of the front-end section of an F-M receiver using parallel wire transmission line tuners.

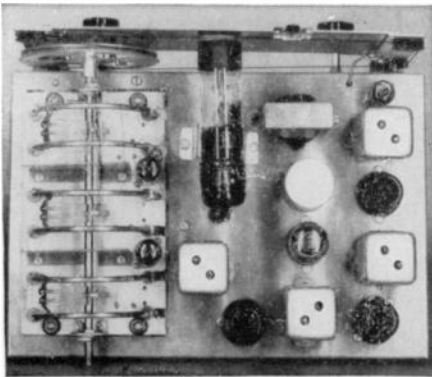


FIG. 5.14. The tuning assembly mounted on the receiver chassis. (Courtesy Approved Electronic Instrument Corp.)

maximum output. At the low end of the F-M band, 88 mc, the end inductance coil turns are either spread apart or squeezed together to achieve tracking here. The whole unit is rubber-mounted to give freedom from microphonics. Miniature tubes are used to provide excellent frequency stability and sensitivity. A 6AG5 is the R.F. amplifier, and a 6J6 double triode functions as a combination mixer-oscillator. Injection of the oscillator voltage into the mixer is accomplished with a 2-mmf coupling capacitor. The unit, mounted on the receiver chassis, is shown in Fig. 5.14.

Schematically, R.F. amplifiers used in F-M receivers do not differ from comparable R.F. amplifiers used in standard A-M broadcast receivers. (See

Fig. 5.15.) Structurally, of course, these amplifiers use smaller tubes and components as previously indicated.

Grounded-grid Amplifiers. In standard-broadcast A-M receivers, pentodes are used exclusively in the R.F. amplifier. Triodes have been ruled out because of their tendency to oscillate owing to the relatively large capacitance existing within the tube between plate and grid. Recently, however, an arrangement known as the grounded-grid amplifier has permitted the use of triode R.F. amplifiers with good results. The grounded-grid amplifier is contrasted with the conventional amplifier in Fig. 5.16. Note that the grid of the tube is at R.F. ground potential and that the signal is fed to the cathode. The tube still functions as an amplifier because the flow of the plate current is controlled by the grid-to-cathode potential. Instead of varying the grid potential and maintaining the cathode fixed, the grid is fixed and the cathode potential is

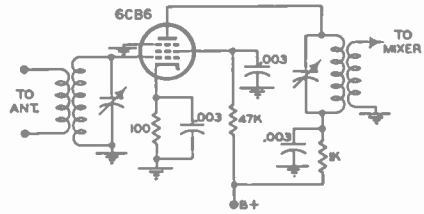


FIG. 5.15. Schematic diagram of an F-M, R.F. amplifier.

ground potential and that the signal is fed to the cathode. The tube still functions as an amplifier because the flow of the plate current is controlled by the grid-to-cathode potential. Instead of varying the grid potential and maintaining the cathode fixed, the grid is fixed and the cathode potential is

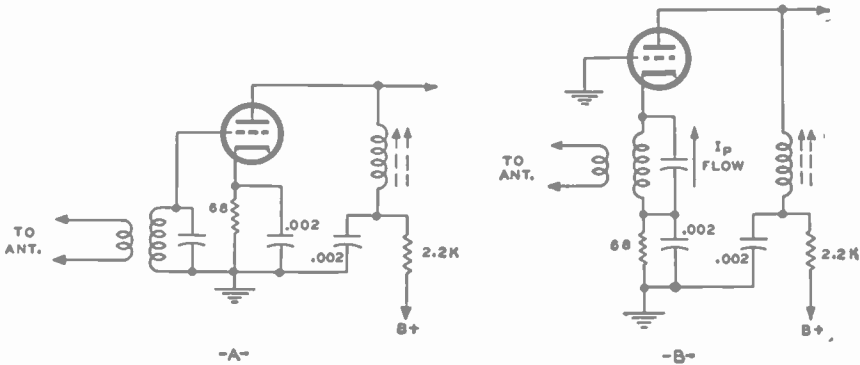


FIG. 5.16. (A) A conventional amplifier. (B) A grounded-grid amplifier.

varied. The net result is still the same. However, the grid, being grounded, acts as a shield between the input and output circuits, thereby preventing the feedback of energy which is so essential to the development of oscillations.

The desirability of using triodes in the R.F. stage of the high-frequency receiver is due to their low internal noise level. Tubes generate a small amount of noise voltage because the electrons moving from cathode to plate do so as separate units and not in a continuous stream. The amount of such noise voltage is seldom more than 10 to 15 microvolts and is ordinarily of

no importance. However, at the front end of a receiver, the signal level may approach this figure and, consequently, it is important that the tubes chosen generate as little noise voltage as possible. In this respect, the triode is considerably superior to the pentode. As a general rule, the greater the number of positive grids in a tube, the greater the internal noise generated.

In standard broadcast receivers, the signal reaching the set is seldom so weak that internal tube noise becomes an important factor. However, in high frequency reception, the ability of a set to receive weak signals is important and the minimum usable signal is determined by the amount of noise voltage which the set itself develops—hence the use of triodes in R.F. amplifiers.

Cascode Amplifiers. The ability of a receiver to amplify a signal is not limited by the amplification which can be obtained from vacuum tubes but by the noise which arises from the tubes and the associated receiver networks. Furthermore, the noise developed by the first stage (the R.F. amplifier) is actually the most important because, at this point in the system, the level of the incoming signal is more nearly on a par with the noise level than at any other point in the receiver. Whatever noise voltage appears at the grid of the R.F. amplifier is amplified along with the signal. Thus, for the best noise-free reception, particularly with weak signals, as much signal and as little noise as possible are desired at the front end of the set.

The best choice for low noise is a triode R.F. amplifier. Unfortunately, however, the gain of a triode is less than that obtainable from a well-constructed high-frequency pentode.

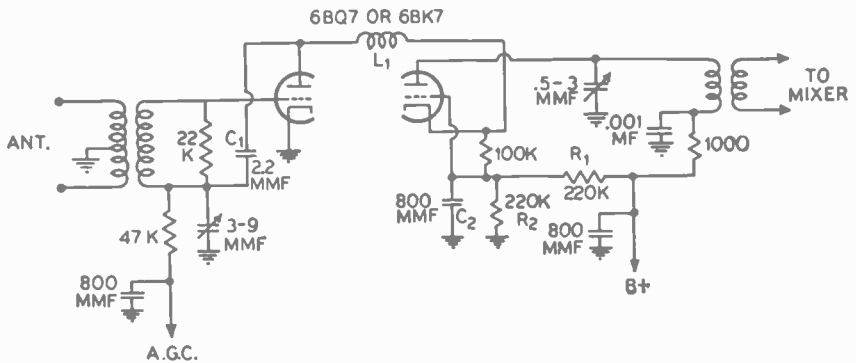


FIG. 5.17. A cascode R.F. amplifier.

A recent circuit development makes it possible to achieve low noise with good gain. The name of this circuit is "cascode" and its schematic is shown in Fig. 5.17. Basically, it consists of two triodes connected in series, that is,

the plate of the first section goes directly to the cathode of the second section. The same current flows through both tubes, and the amount of current is controlled by the bias on the first triode.

The input tuned circuit of this series amplifier connects to the control grid of the first triode; the output circuit is in the plate lead of the second triode. The first stage is operated as a conventional amplifier; the second stage is employed as a grounded-grid amplifier. The coil L_1 between both stages serves to neutralize the grid-to-plate capacitance of the first triode (with help from capacitor C_1) and it is designed to resonate with the grid-cathode capacity of the second section. While L_1 thus aids the stability of this combination, it is also largely responsible for the low noise qualities of the cascode circuit.

Direct coupling is used between the first triode plate and the second triode cathode. With cathode feed to the second triode, C_2 is used to place the grid at R.F. ground potential. Since the two triode sections are in series across a common plate supply, the cathode of the second triode is 125 volts positive with respect to chassis ground. A divider across the plate supply, consisting of R_1 and R_2 , places the grid of the second triode at a sufficiently positive potential (with respect to its cathode) for proper operating bias.

The cascode circuit is being widely used by many receiver manufacturers.

PROBLEMS

1. In what ways do high-frequency tuning circuits differ from low-frequency tuning circuits?
2. What ordinarily happens to the Q of a tuning circuit as its frequency is raised? Why?
3. What is slug tuning?
4. Why is a high L to C ratio desirable in a tuning circuit? When is it purposely lowered?
5. Why do we find four parallel wires used on some permeability-tuned coils? How could a single wire be made to produce the same effect?
6. What is the reason for the uneven turn spacing found on some coils?
7. What properties of a transmission line enable it to be used for tuning?
8. How can a coaxial transmission line be employed for tuning?
9. Describe the operation of a parallel-wire transmission line tuner.
10. Draw the circuit of a grounded-grid amplifier. What advantages does this amplifier offer over the amplifier of Fig. 5.16A for high-frequency operation?
11. What is a cascode amplifier? What advantages does it possess that make it suitable for use as an R.F. amplifier?
12. Draw the circuit diagram of a cascode amplifier.

Chapter 6

HIGH-FREQUENCY OSCILLATORS

Converters and Mixers. In order to take advantage of the benefits of the superheterodyne it is necessary to convert the incoming signal to the lower I.F. If no R.F. stage precedes the converter, then the incoming signal is fed directly to the converter (or mixer) input and reduced to the I.F. immediately. Otherwise, it is fed first through a stage of R.F. amplification and then applied to the converter.

The terms "frequency converter" and "frequency mixer" are used interchangeably by many writers, although a distinction does exist between them. A converter is a tube which produces the oscillator voltage and mixes this voltage with the incoming signal, all within the same tube. A mixer tube, on the other hand, is more limited in scope. It merely mixes or beats together a separately generated oscillator voltage with the incoming signal.

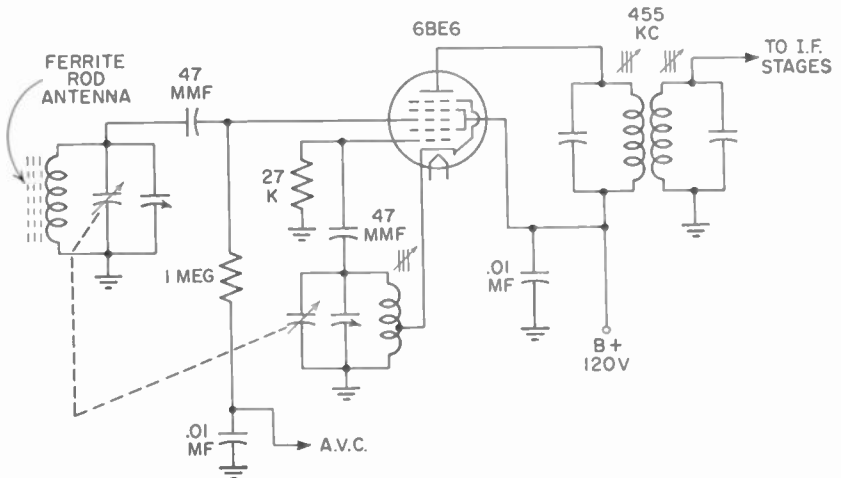


FIG. 6.1. A typical pentagrid converter circuit.

Its function is purely one of mixing. Whichever arrangement is employed depends to a great extent upon the frequencies involved. At the standard broadcast frequencies, a converter tube is more common because it gives satisfactory results and permits an economical arrangement where only an additional coil is needed to generate the oscillations. Fig. 6.1 is an example of a typical frequency converter.

As the signal frequency is raised above 50 mc, the operation of the oscillator becomes more critical. Instability in output voltage and interaction between the oscillator and the signal are more likely to occur with converters. To minimize these effects, the oscillator is separated from the mixing action, as shown in Fig. 6.2. A separate tube (usually a triode) is employed as the oscillator.

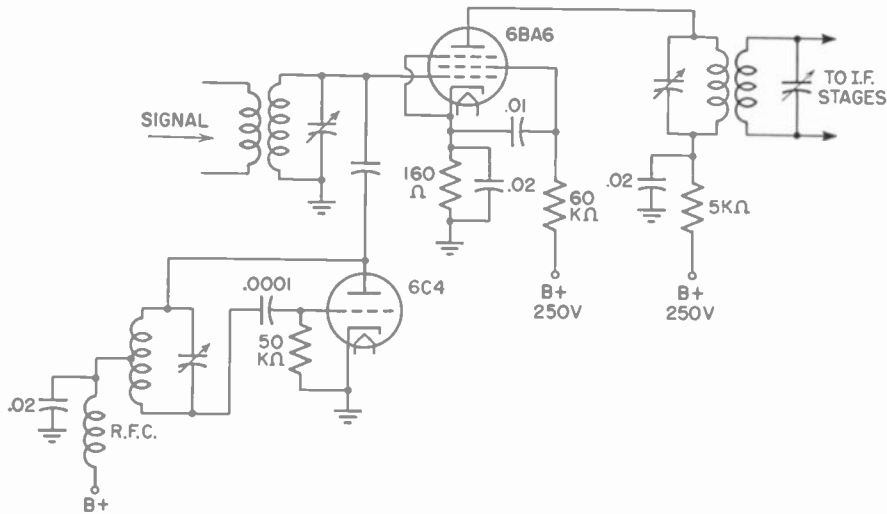


FIG. 6.2. A mixer (6BA6) with a separate oscillator.

The use of a separate oscillator increases the space and cost requirements of the set, but the stability is superior to the frequency converter circuit.¹

Oscillator Stability. Stability in the oscillator is one of the most important engineering aspects of high frequency receivers. Frequency conversion is based upon the fact that a fixed oscillator frequency, beating against an incoming signal, will result in a certain I.F. Little need be said about the constancy of the frequency of the incoming signal. Transmitter channel accuracy is held to within 0.002 per cent—safeguarded by many

¹This does not exclude the possibility of using a converter for F-M receivers in their present band. In fact, some manufacturers do incorporate such circuits using specially designed tubes. However, where good stability and low noise level are desired, separate oscillators are used.

elaborate electronic devices that report instantly any appreciable frequency deviation. At the receiver, the major limiting factor toward stability and the production of a constant I.F. is the stability of the oscillator.

When the oscillator drifts in frequency during receiver operation, the difference, or intermediate, frequency, produced as a result of the oscillator frequency mixing with the signal, changes. A relatively small oscillator drift shifts the signal partially or completely outside the range of the tuned I.F. stages. The receiver output then becomes distorted.

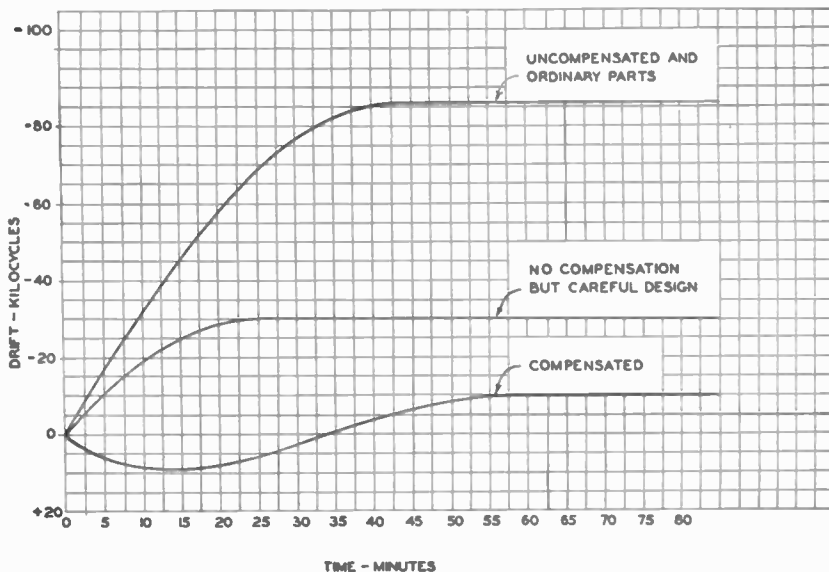


FIG. 6.3. The effect of careful design and compensation on oscillator frequency drift.

The desired stability in an oscillator has been found to be ± 0.01 per cent of the carrier or ± 10 kc at 100 mc. With ordinary precautions, such as the use of low-loss sockets, well-placed components permitting free circulation of air, and air-dielectric trimmer capacitors, the principal drift is due to the capacitance changes within the oscillator tube during the warm-up period. (Of consequence, too, but not quite as important, are the frequency changes due to the variations in oscillator tuning coil and capacitor. The total drift usually amounts to 0.03 per cent, or 30 kc at 100 mc. The drift is never completely developed the instant the set is placed in operation but is attained gradually, perhaps over a period of one-half hour to an hour. After the initial warm-up period, the oscillator frequency levels off to a fixed value unless a sudden change occurs in receiver operation. A typical drift curve is shown in Fig. 6.3. Note that the oscillator always drifts to a lower

frequency, a result of the increase in inductance and capacitance due to the increase in temperature.

In an inductance, increase in temperature—due to surrounding heat generated by the tubes, resistors, and transformers in the receiver—expands the copper wire of each turn. Consequently, the length of each winding increases, and with it the effective radius of the coil. Examination of the design formulas governing single layer coils reveals that increasing the length of a coil decreases the inductance by the first power. However, an increase in the winding diameter raises the inductance by the squared power. The net result, then, is an overall increase in inductance. With capacitors, the increase in capacitance is due to the linear expansion of the capacitor plates. The capacitance of a capacitor plate varies directly with its surface area.

Countermeasures. A simple countermeasure against the positive increase in inductance and capacitance is the use of a small, fixed ceramic capacitor having a negative temperature coefficient. Available in many sizes, the proper capacitor can be selected to counterbalance most positive temperature increases to the point where good stability is obtained. It is simple to install and is extensively used by many manufacturers. Other precautionary measures include regulation of the voltage applied to the oscillator plate, adequate electrical shielding of the oscillator plate, and the use of tubes with high mutual conductances and low input and output capacitances. A tube with a high g_m provides a strong output. Low input and output capacitances minimize the effect of the oscillator tube on the tuning coil and condensers in the circuit. During the heating-up period, the internal capacitance of a tube changes. This occurs, also, with changes in current through the tube. Hence, to minimize the effect of these changes on the output frequency of the oscillator, it is necessary that the internal capacitances form a negligible part of the total capacitance present across the tuning circuit.

Another method which has been used (although not recently) by some manufacturers is to operate the oscillator at some low frequency and utilize the second harmonic for mixing. The lower fundamental frequency permits the use of a high lump-circuit capacitance which, in effect, minimizes the stray and tube capacitances.

Harmonic operation is not feasible unless the oscillator voltage can be fed into the mixer tube by way of a separate grid. The capacitive and inductive injection methods shown in Fig. 6.4 (A and B) require a considerable oscillator voltage to be effective. The oscillator, in each instance, must transfer its voltage to the mixer grid tuned circuit. Since this network is resonant to the higher incoming signal (higher by the I.F. value), the impedance presented to the oscillator voltage will be low. Hence, consider-

ably more oscillator voltage must be available because of the mismatch. (In any mismatch there is inherently a large waste or loss of power and consequently a greater amount of power must be expended to develop the required smaller value.) Since the second harmonic output is lower than the fundamental, it means pushing this unit more. When the mixing voltage can be applied directly to a separate grid, with only a resistor as the impedance, then a closer match may be made with a resultant rise in efficiency.

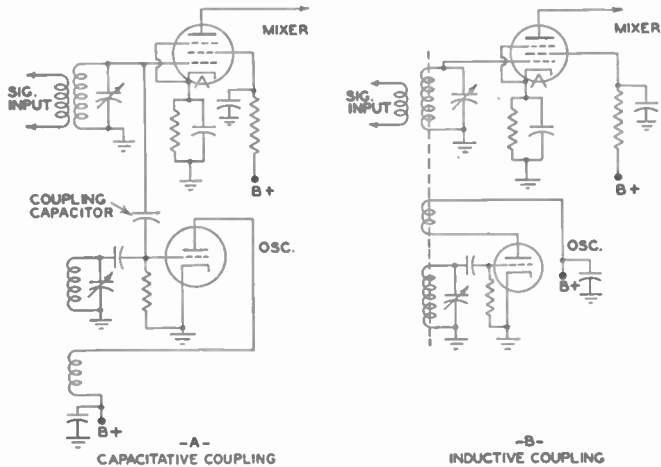


Fig. 6.4. Two methods of coupling energy from between the oscillator to the mixer.

High-Frequency Converter Operation. When F-M broadcasting in the 88-108 mc band was initiated, certain pentagrid converters which were then popular in A-M receivers could not be made to operate satisfactorily at the higher frequencies. The reason for this can perhaps best be understood by an examination of the action within such tubes. The 6A8 is typical of these converters and will serve to illustrate the points. The conventional schematic and the actual element placement are shown in Fig. 6.5. It is current practice to identify each grid by number, commencing with the grid nearest the cathode. The use of each grid is also indicated.

When the circuit is in operation, electrons leave the cathode area in pulses, controlled by the voltage variations on the oscillator grid, G_1 . As part of the oscillator, G_1 is negative during most of its cycles, becoming sufficiently positive (or less negative) for only a small portion of the time. It is during these short intervals that the electrons flow past G_1 , attracted by the positive potential on G_2 and G_3 . It is to be noted that G_2 , the oscillator anode, is not truly a grid, but only two posts or vertical rods without any of the usual grid turns of wire.

Some of the electrons that flow past the screen grid, G_3 , come to a halt in the region between G_3 and G_4 because of the negative potential on the signal grid, G_4 . This cloud of electrons constitutes a virtual cathode. The number of electrons reaching the plate of the tube is controlled by the variations of the potential on G_4 set up by the incoming signal. The overall action then is the control or modulation of the tube's electron flow by two sources, the oscillator grid variations and the input signal changes. The I.F. beat note is produced as a consequence of this interaction.

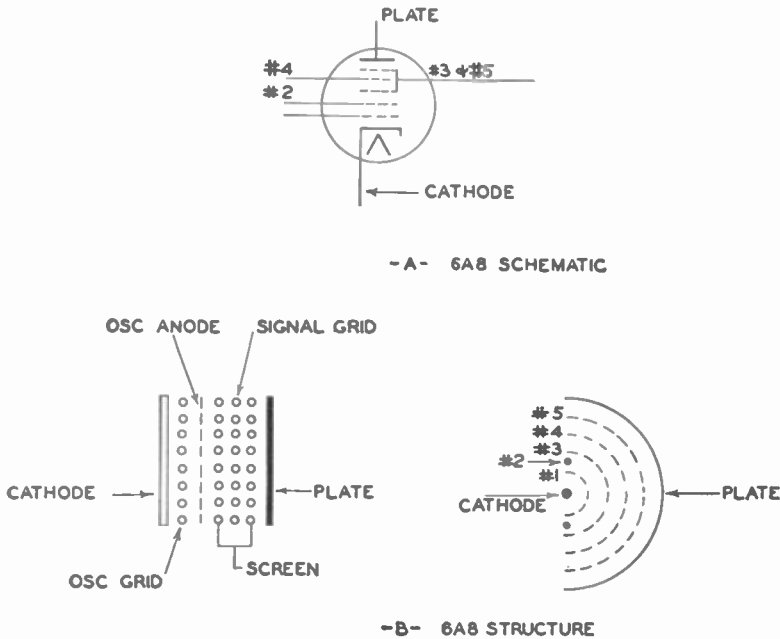


FIG. 6.5. The conventional schematic (A) and actual electrode placement (B) of a 6A8 converter tube.

It is seen from the above that the number of electrons in the space charge between G_3 and G_4 will vary in accordance with the voltage variations on the oscillator grid, G_1 . Because of the closeness of the space charge to G_4 , a voltage at the oscillator frequency will be induced into the circuit of the signal grid. The resultant current will not be in phase with the normal oscillator voltage and it will therefore act to buck or cancel out part of the effectiveness of the oscillator voltage. A reduction in gain results, or, in other words, the conversion transconductance is low.

A second and more serious reaction within the tube is the change produced in oscillator frequency by the signal grid. It is the signal grid, G_4 ,

that determines the number of electrons taken from the virtual cathode as constituted by the space charge and passed on to the plate. This, in turn, has been found to react on the oscillator grid, altering its capacitance with respect to the rest of its circuit and thereby forcing a change in frequency. With a separate oscillator, this latter effect cannot occur and the stability of the oscillator frequency from a separate source is not affected.

The use of a single tube to function as an oscillator and mixer is attractive economically and, under the pressure of this advantage, the tube companies have developed pentagrid converters which perform satisfactorily

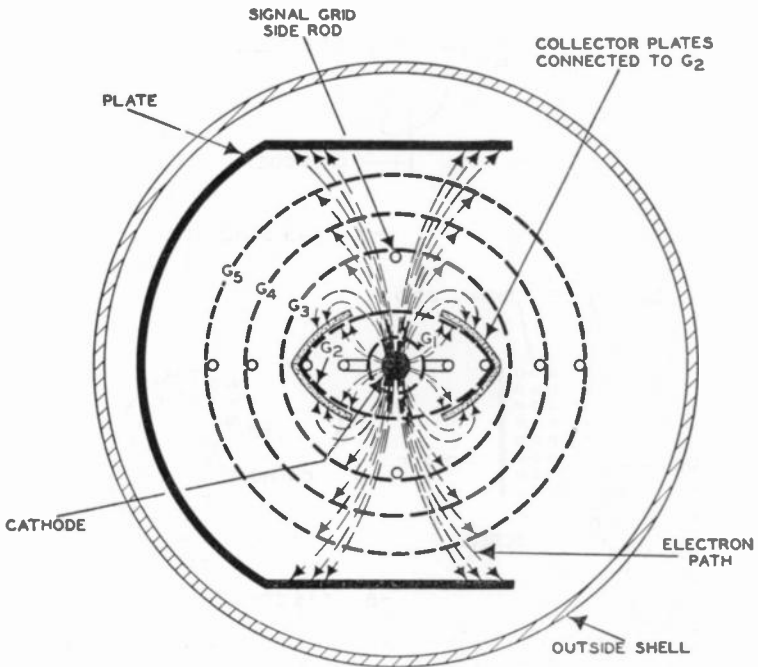


FIG. 6.6. The internal construction of the 6SB7Y tube.

at the F-M frequencies. Examples of such tubes are the 6SB7Y, the 6BE6, and the 6BA7.

An indication of the altered construction of these tubes can be seen from an examination of the internal structure of the 6SB7Y. (See Fig. 6.6.) At the center of the tube is the cathode, closely surrounded by G_1 which serves as the oscillator grid. Beyond G_1 is a second grid, G_2 , which functions as the oscillator plate and also serves to accelerate the electrons emitted by the cathode. Internally G_2 is connected to G_4 , and both elements are properly by-passed so that essentially only a d-c potential is present here. G_2

also has connected to it two side plates whose function it is to shield G_1 by intercepting any electrons that might be repelled by the negatively charged G_3 . In this way, interaction between the signal and the oscillator circuits is materially reduced. G_3 is located between G_2 and G_4 and receives the signal voltage. The two supporting rods of G_3 are placed in the center of the electron stream traveling from the cathode to the plate. Since the average potential of G_3 is negative, the effect of this rod placement is to split the electron stream into two diverging beams, as shown. Because of the particular structure of the elements and the distribution of the electric fields within this structure, electrons repelled back by G_3 are forced to strike the side plates of G_2 , effectively preventing them from reaching G_1 . G_5 is a suppressor grid and is externally connected to ground. Insertion of this suppressor grid raises the internal resistance of the tube and permits a higher conversion gain to be achieved.

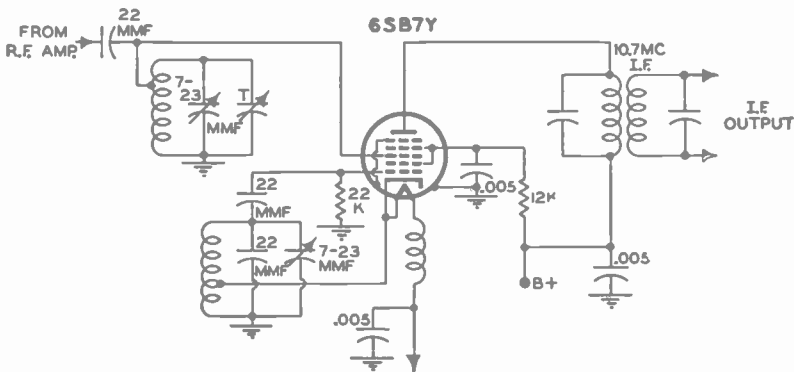


Fig. 6.7. A typical circuit application of the 6SB7Y tube.

Circuit connections for the 6SB7Y are shown in Fig. 6.7. The oscillator coil is connected between the cathode and G_1 . G_3 receives the signal voltage either from an antenna or an R.F. amplifier. The plate of the tube attaches to the primary of the first I.F. transformer.

In the choice between a frequency converter and a separate oscillator and mixer, there is one additional factor to consider—tube noise. As the number of grids or controlling elements within a tube increases, the amount of tube noise generated likewise increases. In fact, the amount of noise generated in a pentagrid converter, as compared with a straight R.F. amplifier, is in the ratio of approximately 4 to 1. The minimum signal reaching the grid of a pentagrid converter must be this much stronger in order to override tube noise. The noise will also limit the minimum signal that the receiver can amplify successfully without interference from the amplified tube noise.

Before we leave the subject of pentagrid converters, mention should be made of the gain that is obtained when the signal passes through the stage. Converter gain is defined as

$$\text{gain} = \frac{S_c r_p R_L}{r_p + R_L}$$

where

S_c = conversion transconductance, i.e., the effect an R.F. signal voltage has in producing a current in the plate circuit at the I.F. value

r_p = plate resistance of the tube

R_L = resistance of tuned plate circuit to the intermediate frequencies

In the equation, S_c , the conversion transconductance, is the controlling factor. S_c , in turn, depends upon the oscillator voltage. Increasing the gain of a converter lowers the minimum value of signal voltage needed at its grid to override the noise generated in the tube. Of triodes, pentodes, and pentagrid converters, the last possesses the smallest S_c . However, pentodes have a much higher internal plate resistance (r_p in the foregoing formula), and the overall gain, using the pentode, is higher.

When a separate oscillator is used, its voltage may be injected at the control grid, cathode, screen or suppressor grids of a pentode tube, or at the control grid or cathode if a triode is used as the mixer. The cathode is convenient, since control over the grid and plate circuits may thereby be effected simultaneously. Control grid injection is more critical than cathode injection because of the increased possibility of interaction between the signal and oscillator circuits. The screen and suppressor grids are seldom used for injection because a greater amount of oscillator driving voltage is needed to obtain a satisfactory degree of mixing.

Oscillator Coupling. Coupling to the control grid may be accomplished either inductively or capacitatively. Both methods are shown in Fig. 6.4. For cathode coupling, either method of energy transfer can also be employed. Loose coupling is necessary at the oscillator in order that the tuning circuit shall not be loaded down to any appreciable degree. With loose coupling, the oscillator is least affected by any undesirable feedback and the stability is much better.

Another reason for loose coupling is to keep the amount of oscillator energy that is fed to the mixer at the lowest point consistent with good operation. Every oscillator output contains a large number of harmonics. These harmonics may react with certain undesirable incoming voltages to produce a difference frequency equal to the I.F. The result is known as a spurious response and, if sufficiently strong, may easily interfere with the desired signal. By keeping the oscillator voltage fed to the mixer as low as possible

—consistent with good operation—we decrease the ability of these oscillator harmonics to produce spurious responses. It is well to keep this in mind when either building or repairing a receiver.

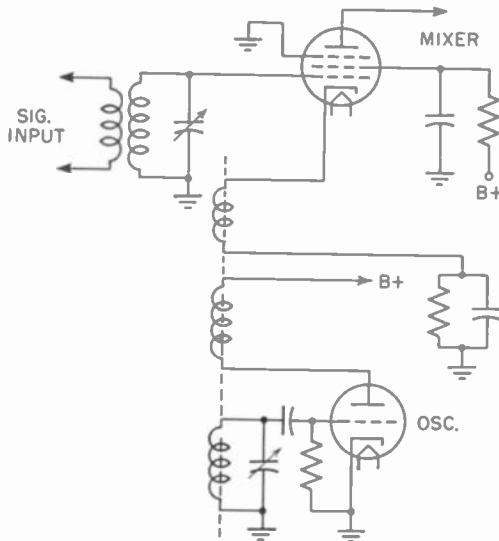


FIG. 6.8. Coupling oscillator voltage to the cathode of the mixer.

Previously the most important reason for not extensively using a triode for a mixer, despite its excellent signal-to-noise ratio and high conversion transconductance, was its relatively large grid-to-plate capacitance. A large C_{gp} permits a high percentage of feedback. Since the signal frequency is higher than the I.F., the I.F. tuning circuit in the plate circuit appears to be wholly capacitive to the signal. All frequencies higher than the resonant frequency of a tuned circuit find it much easier to pass through the capacitor than the coil. When this energy feeds back through C_{gp} , the Q of the input tuned-grid circuit is decreased. In a pentode feedback is negligible because of the extremely small grid-to-plate interelectrode capacitance.

Recently high frequency triodes have been developed in which the coupling between plate and grid is minimized so that undesirable feedback is not too serious. Some receiver manufacturers are using these tubes and the results appear satisfactory.

Types of Oscillators. Three oscillator circuits have been widely used. These are the conventional Hartley (A), the modified Hartley (B), and the ultraudion oscillator (C), the latter being a modification of the familiar Colpitts oscillator. (See Fig. 6.9.) Each unit, if properly constructed, oscillates readily and possesses good stability. Triode tubes are usually pre-

ferred because they require fewer components. Also, there are several high-frequency tubes available, such as the 12AT7 and 6U8, which contain two separate sections, one of which is a triode. This enables the designer to use the triode for the oscillator and the remaining section for the mixer.

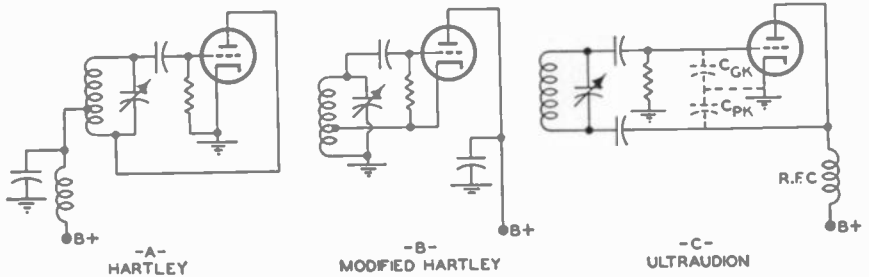


FIG. 6.9. Three popular oscillator circuits.

Crystal Oscillators. Great stress has been given to oscillator stability because of the commanding position that the oscillator holds in the superheterodyne circuit. This feature is especially important in F-M mobile receivers (used commercially) where the circuit must be designed to operate under conditions which are considerably less favorable to stability than is ever encountered in home receivers. Furthermore, most commercial operation is restricted to certain specific channels which have been allocated by the Federal Communications Commission for that purpose. Under these circumstances, preset circuits actuated by pressing a push button greatly simplifies receiver tuning and assures the user of a properly tuned receiver each time. Ordinary preset circuits, such as we find in standard-broadcast A-M home receivers are unsuitable for commercial usage because of their tendency to drift after a short period of operation. Preset circuits, to have any commercial value, must be able to retain their calibration under the rugged conditions encountered in such operations.

In order to obtain better frequency stability, quartz crystal oscillators are used. Crystals, on a production basis, can be obtained with a stability of ± 0.005 per cent, which represents 5 kc at 100 mc. Included in this figure are variations due to temperature and humidity. Desirable stability has been previously stated as ± 0.01 per cent, which is a lower degree of stability than the 0.005 per cent obtainable with a crystal. The need for specially constructed temperature-compensating components in crystal oscillators is also less. This does not mean that the crystal oscillator section need not be carefully constructed, but it does mean that greater leeway is possible in choosing components. Through the use of crystals, push-button tuning becomes perfectly feasible because consistent operation is achieved over long

periods of time. For the layman user of an F-M receiver, this form of selecting stations assures optimum results without much effort and eliminates entirely the need for any form of tuning indicator.

When crystal oscillators are used, however, reception is restricted only to those stations for which crystals are provided. In other words, each separate frequency to be received requires another crystal in the oscillator in order to develop the proper mixing frequency for the signal. Generally the number seldom runs beyond three or four, although the actual space required is small and a larger number of crystals could be accommodated without increasing the physical dimensions of the receiver appreciably.

Crystal Oscillator Circuits. In choosing a crystal oscillator circuit, we desire a circuit that will require nothing more than the insertion of the crystal to make it oscillate. For this reason, the tuned-grid, tuned-plate circuit using the crystal would be unsatisfactory. Here, besides the insertion of the crystal, we must tune the plate tank capacitor. A suitable circuit is

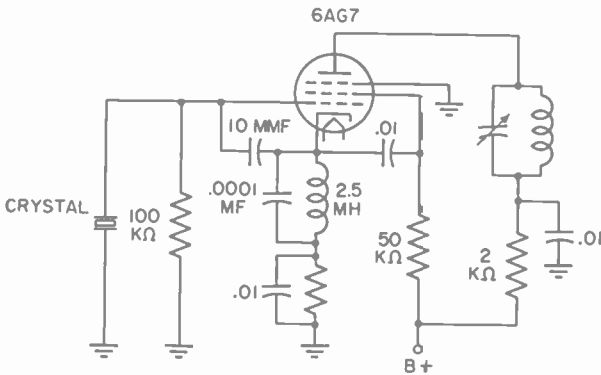


FIG. 6.10. A grid-plate crystal oscillator.

the grid-plate crystal oscillator shown in Fig. 6.10. The crystal is connected between the grid and ground. In the cathode tuned circuit, the radio frequency choke and the fixed capacitor, 0.0001 mf, tune to a frequency that is lower than the crystal frequency. Good output is obtained on the fundamental and on harmonics. Good output on the harmonics is especially desirable because, as we shall soon see, it is not the fundamental frequency that is used for mixing but a harmonic. In order to accentuate the desired harmonic, the plate tank circuit is tuned to that harmonic. The setting of the plate tank has no effect upon the functioning of the oscillator, but is inserted to insure that sufficient voltage is generated at the harmonic frequency.

For crystal control, the mixing oscillator frequency is generally below the signal input by the I.F. value, say, 10.7 mc. If the set is tuned to 152 mc

(one of the frequencies allocated to taxicabs), then the mixing oscillator must generate a signal which is 10.7 mc lower than this, or 141.3 mc. With a strong fundamental crystal frequency near 11.8 mc, multiplication by a factor of 12 is required to reach 141.3 mc. (The actual crystal frequency would be 11.775 mc.) This means that we must use the 12th harmonic of the fundamental crystal frequency. A well-designed crystal oscillator will develop an output which is rich in harmonics and possess sufficient voltage at the 12th harmonic to permit use of the simple oscillator circuit shown in

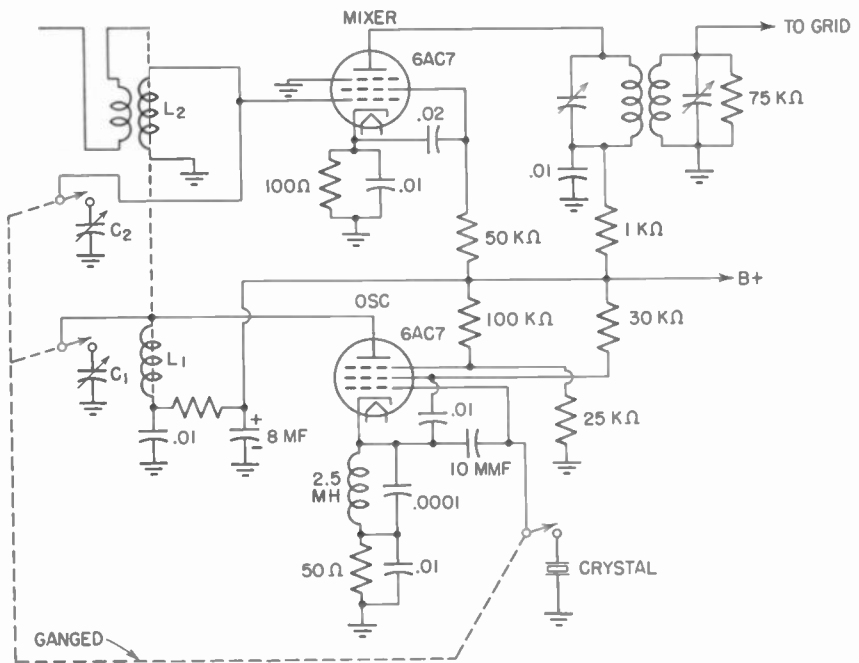


FIG. 6.11. The input stages of a crystal-controlled F-M receiver. The suppressor grid of the oscillator is made slightly positive to produce a strong output.

Fig. 6.11. For each frequency a separate crystal and two trimmer capacitors are brought in by a push button or a rotary switch. The trimmer capacitor C_1 across L_1 is peaked at the factory to resonate L_1 to the proper harmonic frequency. L_1 and its trimmer do not in any way influence the oscillations of the crystal unit. They merely serve to produce a strong output at the harmonic mixing frequency. The other trimmer capacitor, C_2 , tunes the antenna input coil, L_2 , to the signal frequency. The addition of an R.F. stage would add one more trimmer capacitor.

Instead of operating the crystal oscillator at the 12th harmonic of the crystal, many manufacturers prefer to use a crystal with a low fundamental

frequency and then use two quadrupler stages. The advantage of using a low fundamental frequency crystal lies in the fact that it is thicker than a higher frequency crystal. Thicker crystals are more easily manufactured and they can withstand greater current surges without cracking. The thinner a crystal is, the more easily it may break, either from jarring or excessive current flow in its circuit. Outweighing the advantage of a thicker crystal is the fact that a higher order harmonic must be used in order to obtain the proper mixing frequency. The higher the order of the harmonic, the lower the voltage developed in the oscillator. Too low a mixing voltage will not permit full advantage to be taken of the conversion transconductance of the mixer. The output on weak signals may drop so low as to make good reception impossible.

Where to Place Oscillator Frequency. The oscillator frequency, we know, must be separated from the incoming signal frequency by an amount equal to the intermediate frequency. The question then arises as to where this oscillator frequency is to be placed, whether above or below the signal. As far as the outcome is concerned, it makes absolutely no difference which point is chosen, as long as the difference between the point chosen and the signal results in the proper I.F. It is usual in most F-M high frequency sets to place the oscillator above the signal frequency. This is similar to the present practice at the standard broadcast frequencies, where the oscillator always operates above the signal. The reasons for the placement of each, however, are quite different. At the lower frequencies, the signal frequency range extends from 550 kc to 1700 kc. The usual I.F. is 455 kc. Suppose we designed the mixing oscillator so that it functioned 455 kc below the signal. Then it would have to cover a range from 95 kc to 1235 kc—or a range where the highest and lowest frequencies were in the ratio of 13 to 1. For a single coil and capacitor to cover a range this wide would present a practical problem that would be most difficult to solve. Most tuning combinations possess a tuning ratio of 2 or 3 to 1. Here we need 13 to 1. For this reason the oscillator frequency is made higher than the signal. Now let us see what happens. Being above the signal, the oscillator must generate frequencies from 1005 to 2155 kc—or a range with approximately a 2 to 1 ratio. This is easily obtained in practice with one capacitor and coil.

Now let us look at the higher F-M band, 88 to 108 mc. We will assume an I.F. of 10.7 mc. If the oscillator is placed above the carrier frequencies, its range will be from 98.7 to 118.7 mc. If placed below the carrier frequencies, its range will be from 77.3 to 97.3 mc. Design of tuning circuits does not enter into this discussion because both ranges are perfectly possible with one set of components. If we select the lower frequencies, we gain the advantages of greater ease in construction, more easily obtained stability, and

the use of less critical components. However, with the oscillator generating frequencies from 77.3 to 97.3 mc, any such signal which finds its way to the antenna will cause interference to nearby television receivers operating on channel 5 (76-82 mc) or channel 6 (82-88 mc). On the other hand, when the local oscillator operates above the F-M band, no interference results. It is principally for this reason that F-M oscillators operate 10.7 mc above the incoming signal.

PROBLEMS

1. Why are the I.F. amplifiers important in a superheterodyne receiver?
2. What is the difference between the terms "frequency converter" and "frequency mixer"? Which is found most frequently in standard broadcast A-M receivers? In F-M receivers?
3. Draw the circuit of a mixer with a triode Hartley oscillator.
4. Why is oscillator stability so important in a high-frequency receiver?
5. Discuss the various factors which govern oscillator stability.
6. What are some of the countermeasures employed to stabilize the oscillator frequency?
7. From the standpoint of noise and efficiency, what type of tube is most desirable? Why?
8. Draw representative circuits illustrating how the output of a separate oscillator is coupled capacitatively to the mixer. Illustrate, too, inductive coupling. Is there any other method of coupling possible between the oscillator and mixer? Explain.
9. Why is the frequency drift in an uncompensated oscillator always toward the lower frequencies?
10. In Fig. 6.3, the initial frequency drift of the oscillator frequency is positive. Explain why this differs from the situation in an uncompensated oscillator.
11. Are frequency converters ever used in F-M receivers? Draw the circuit of a frequency converter suitable for high-frequency use.
12. Which three oscillator circuits are commonly employed in F-M receivers? Draw one and couple its output capacitatively to a mixer.
13. Is it desirable for an oscillator to produce harmonics? What effect do these harmonics have on the receiver operation?
14. What desirable properties should an oscillator tube possess?
15. Why would a crystal oscillator be advantageous in an F-M receiver? What disadvantages would it introduce?
16. Are there any limitations on the placement of the oscillator frequency? Explain.
17. Draw the circuit diagram of a crystal-controlled oscillator and the associated mixer.
18. What type of crystal oscillator circuits are used?
19. What considerations govern the choice of a crystal?
20. Explain the operation of a mixer.
21. Define conversion transconductance. Contrast this with mutual conductance of a tube.
22. Why is loose coupling employed between an oscillator and a mixer?
23. Explain why the placement of the oscillator frequency in a standard broadcast receiver is restricted more than it is in a high-frequency F-M set.

Chapter 7

I.F. AMPLIFIERS

Design Factors of I.F. Amplifiers. The superiority of the superheterodyne lies in its I.F. amplifiers. The incoming signal frequency is purposely lowered to a level where a large degree of amplification is more readily, and efficiently, achieved, using conventional radio components. In addition, it permits the designing of fixed tuned circuits with a bandsread sufficiently wide to include all the sideband frequencies arising from modulation. An engineer, designing a tuned circuit for use at one frequency only, can better

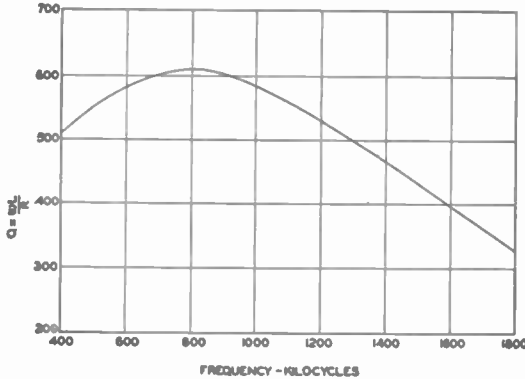


FIG. 7.1. The variation of Q (of a coil) with frequency.

obtain the maximum amplification than when he is faced with the prospect of having to adapt the circuit for uniform operation over a range of frequencies. A coil designed to function over many frequencies, such as is necessary in the input stages of a receiver, generally produces uneven amplification over the entire band. In Fig. 7.1 is shown a curve illustrating the variation in the Q of a coil ($\omega L/R$) with frequency over the standard broadcast band. Note that the Q has a high value at the beginning and then drops

fairly rapidly at the higher frequencies. This latter decrease is due to the fact that R increases more rapidly than ωL with the rise in frequency. R represents a combination of:

1. Losses in the form on which the coil is wound.
2. Skin effect, which confines the current to a narrow layer near the surface of the wire.
3. Eddy current losses in near-by metallic objects.

Through the use of a circuit tuned to one fixed frequency, the amplification and selectivity may be held constant—adding to the stability of the receiver.

In the design of an I.F. amplifier, or an I.F. system containing several stages, three basic factors must be considered:

1. Frequency of the I.F. stages.
2. Gain.
3. Selectivity.

These quantities are certainly interrelated, but the general procedure is to choose first the operating frequency and then to consider the problems of gain and selectivity together.

Choosing the Intermediate Frequency. The choice of an intermediate frequency may appear, at first, to be quite simple since we know that at the lower frequencies it is easier to arrange and design circuit components to obtain high gain. However, there is a limit to the low frequencies that can be used because of the stability of the oscillator and other circuit components located ahead of the I.F. stages in the receiver. When a set is first put into operation, it may require as much as an hour or so before the oscillator frequency becomes stable. When ordinary parts are used in the construction of a receiver, the oscillator may drift as much as 0.2 per cent in frequency. At 95 mc, this means a drift of 190 kc. With a low value of I.F., the station would soon drift outside of the I.F. range. Through compensation of an oscillator, such as the use of low-loss tube sockets, careful wiring, capacitors with negative temperature coefficients, etc., the amount of drift can be lowered to a figure of 0.05 per cent of the carrier frequency or 48 kc. This is a considerable improvement over the previous uncompensated figure. Further improvement can be achieved only at much greater cost, perhaps involving specially constructed components or the use of crystal-controlled oscillators.

A second limitation on the use of a low I.F. is the tendency of most oscillators (excepting, possibly, crystal oscillators) to lock-in with other more stable oscillators when they are quite close in frequency. With a low I.F., a strong signal could readily pull the oscillator out of its desired position into

the same frequency as the signal. As might be expected, under these conditions, no difference or intermediate frequency will be produced.

Still another consideration governing the choice of an I.F. value is the spread of the incoming signal. An F-M broadcast, fully modulated, occupies a frequency bandwidth of 200 kc. Obviously, it would be ridiculous to use any I.F. below this figure, for then how would it be possible to vary an I.F. of say 125 kc plus and minus 200 kc? There are no negative frequencies. Hence, an I.F. higher than 200 kc is necessary. But how much higher? 400 kc? 800 kc? 1 mc? or still higher? Before we definitely answer this question, let us investigate the different types of spurious responses capable of affecting a receiver and their influence on the choice of an intermediate frequency.

Spurious Responses. A spurious response is defined as the reception of a signal at a point other than desired on the dial. For example, a station may be received at more than one point on the dial, or two or more stations obtained at the same time when only one should be present. In using the phrase "spurious response," we refer only to interference caused by stations appearing at undesired points of the dial. Atmospheric and man-made noises are not included.

The most important spurious responses to which an F-M set is subjected are:

1. Image response.
2. Response of two stations separated in frequency by the I.F. value.
3. Direct I.F. response.
4. Combination of harmonics from oscillator and signal.

We may pause here and note that the appearance of any of these spurious responses is in no way connected with the noise-reducing qualities of frequency modulation discussed in Chapter 1. Any interfering signal, either arising from one of the sources listed above or from another station on the same channel, will be completely suppressed if the desired signal is at least twice as strong as the interference. However, the danger lies in the fact that this relationship may not exist. In fact, the undesired signal may be many more times stronger than our desired signal. In such an event, the desired signal could be either badly distorted or else completely obliterated. The choice of an I.F., coupled with proper design, will do much to minimize the interference from the four main sources of spurious responses.

Image Response. Image response is due to the mixing of an undesired signal with the oscillator voltage in the converter stage to produce a voltage at the intermediate frequency. Since a frequency equal to the intermediate frequency is produced, this signal will be accepted and passed by the I.F.

amplifiers. To understand how this may occur, suppose that a receiver is tuned to a station at 95.3 mc. With the oscillator set at 99.3 mc, the difference (4.0 mc) which is also the intermediate frequency will be produced in the frequency converter or mixer plate circuit and fed into the I.F. stages. Modulation would swing the carrier ± 75 kc about the center value.

Now suppose—and this may readily happen—that there is another powerful station in the vicinity operating on a frequency of 103.3 mc. If sufficient selectivity is not provided ahead of the converter stage—or an R.F. stage of amplification is not included in the set—then an appreciable amount of the signal voltage from the 103.3-mc station may appear at the converter grid. Subsequent mixing with the 99.3-mc voltage from the oscillator will produce a difference frequency of 4.0 mc which is also equal to the I.F. This second interfering signal is referred to as an image station.

It makes little difference—so far as the production of image response is concerned—whether the oscillator frequency is higher or lower than the desired station. In the preceding example, the oscillator was operating above the desired signal. It may also be located lower in frequency by an amount equal to the I.F. In fact, most manufacturers prefer the lower oscillator frequency because of the increased stability of oscillators at lower frequencies.

Using the last example, but with the oscillator below the desired signal, say at 91.3 mc, an image response will be obtained if the signal is below the oscillator by an amount equal to 4.0 mc. In this instance, this would mean a station at 87.3 mc. Under present allocations, the F-M band starts at 88 mc and no F-M station is located at 87.3 mc. However, if we tuned the receiver to another station, say at 97.3 mc, then with the oscillator at 93.3 mc and the image station at 89.3 mc we create a possibility for an image response. It will be noted that, whenever the oscillator frequency is higher than that of the desired station, the image is above the oscillator by an amount equal to the I.F. On the other hand, the image station is lower in frequency than the desired station when the oscillator is also lower. The image station and the desired signal are always separated from each other by an amount equal to twice the I.F. value. This is shown in Fig. 7.2.

One solution for minimizing interference due to images is increasing the selectivity of the tuned circuits preceding the converter stage. Not only is the number of tuned circuits important but also the attenuation they impose on any signal extending beyond the ± 100 -kc limits of the desired station. It is generally operationally difficult and most times economically impractical to develop in these high frequency input circuits a high degree of selectivity. At the low frequencies, one megacycle separation between signals represents a considerable percentage of the carrier frequency. Thus, for example, two stations separated by one megacycle at 15 mc could be resolved

readily, without interference, by a circuit possessing a much lower Q (and, hence, selectivity) than if the carrier frequency is raised to 90 mc. It is not the numerical frequency difference between stations that is important but the percentage difference. In the first case, the percentage separation is approximately 6.7 per cent; at 90 mc, a 1-mc separation represents a percentage separation of approximately 1.1 per cent. Two tuned circuits possessing the same selectivity could be more economically manufactured at the low frequencies than at the high frequencies.

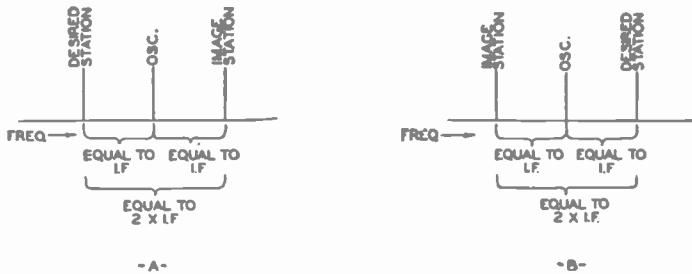


FIG. 7.2. The relationship between the desired and image stations when the oscillator frequency is above (A) and below (B) the signal frequency.

The foregoing illustration is one reason why superheterodynes have proved so popular. In a superheterodyne, the signal on reaching the receiver is converted or "stepped-down" to a lower frequency which is known as the intermediate frequency. Once the frequency is lowered, we can better apply selectivity and gain to the signal than if it had remained at 90 or 100 mc, its incoming value. No lowering occurs in tuned radio-frequency receivers—receivers which are practically out of production except in very cheap sets built to receive the standard broadcasting frequencies. Despite additions of R.F. amplifier stages to a receiver and despite any sharp selectivity of the input resonant circuits, it is still the I.F. amplifiers which are chiefly responsible for the selectivity or sensitivity obtainable from a superheterodyne.

The choice of an I.F. has a direct bearing on the amount of image response to which a set is subjected. If the I.F. is too low, say 1, 2, or 3 mc in value, then the possibility of interference from image stations is correspondingly increased because of the small percentage difference between the desired signal and an image signal. Raising the I.F. provides a greater separation between these signals and makes it easier for the preliminary or input circuits to weaken the interfering signal—weaken it probably to the point where the desired signal could normally override it. In practice, it is common to make the I.F. slightly greater than half the width of the entire F-M band. Thus, since the band extends for 20 mc—from 88 mc to 108 mc—an I.F. of 10.7 mc would eliminate interference from image stations entirely.

The reason is quite simple. We have just noted that the desired station and the image station are always separated by $2 \times$ I.F. value (see Fig. 7.2). But, if $2 \times$ I.F. value is greater than the width of the entire broadcast band, then it is obvious that the reception of image signals, within that particular band, is impossible.

Of course, we must not overlook the possibility of stations on adjacent bands causing trouble. Thus, for instance, adjacent to the F-M band, 88-108 mc, at the lower side is the television band, 82-88 mc. However, sufficient discrimination is provided by the input resonant circuits to make this possibility small.

Stations Separated by the Intermediate Frequency. The second source of spurious response listed is interference arising when two signals mix, both separated in frequency by an amount equal to the I.F. of the receiver. In this situation, one incoming signal acts as the mixing oscillator for the other signal, their difference frequency appearing at the output of the mixer or converter stage as the intermediate frequency. As an illustration of this type of spurious response, consider a receiver where the input selectivity is such that two signals, say 90.3 mc and 99.4 mc, are present at the converter grid in appreciable strength. When these mix with each other, in the converter circuit itself, a difference frequency of 9.1 mc (the I.F., say) will appear. The oscillator of the set itself does not enter into this action, and with poor input selectivity these two stations will be present no matter where the set is actually tuned. Hence, they will be heard all over the dial.

There are two solutions to the problem. One is to provide sufficient discrimination in the circuits preceding the mixer that they will reject two signals so widely separated in frequency. The other is to provide a high I.F., preferably an I.F. slightly greater than the entire bandwidth. In present-day operation, this would mean an I.F. above 20 mc. By so doing, we make it impossible for any two stations within the band to mix with each other at the converter (providing they reach this stage) to produce a difference frequency equal to the receiver I.F.

Direct I.F. Response. The third type of spurious response that may arise is due to direct reception of a frequency equal to the receiver I.F. itself. If some station or service is transmitting at the I.F. value used in the receiver, this signal could easily reach the I.F. amplifiers either through the usual input channel or else directly by the development of the signal voltage on the grid of the first I.F. amplifier when adequately shielding is not provided. That this can be a serious source of trouble has been demonstrated by many tests.

To forego the need of incorporating special filters, wave traps and shielding to prevent interference from the latter source, an I.F. is chosen whose frequency is seldom that of any commercial transmission. This accounts for

such seemingly odd figures as 4.3 or 10.7 mc for the value of an I.F. On many frequencies, the only traffic is ship-to-shore messages or perhaps traffic between mobile experimental equipment, etc. These frequencies have been set aside by the government for the particular services. Since the stations operating at these frequencies are not very powerful, the possibility of their interfering with our home receivers is fairly remote. Hence, these frequencies are chosen for the I.F. in the receiver.

Harmonic Spurious Responses. The fourth and final item on the list of spurious responses is the production of interference from harmonic relationships occurring between signal and oscillator voltages that could produce a voltage at the I.F. value. Perhaps the best way of demonstrating this source of spurious response is by an actual example. Suppose the oscillator is at 95.1 mc. Its second harmonic would be at 190.2 mc. If a strong signal of 89.7 mc overloads the input of the receiver, a very good chance exists that its second harmonic would be formed, the frequency being 2×89.7 mc, or 179.4 mc. The difference between 190.2 mc and 179.4 mc is 10.8 mc. This would reach the plate circuit of the converter stage and be accepted by an I.F. amplifier designed for 10.7 mc.

The probability of the foregoing situation occurring in a receiver depends to a great extent upon the care taken in the receiver design. Since the appearance of this form of spurious response depends upon the generation of secondary harmonics of both a signal and the oscillator, the obvious solution is to minimize the possibility of such harmonics. In the oscillator, it is almost impossible to prevent the appearance of harmonics because of the nonlinear operation of the circuit itself. In general, though, the strength of the harmonics decreases with the order of the harmonic. Thus, the fundamental frequency component is strongest, the second harmonic not quite as strong, the third harmonic weaker still, etc. The rapidity with which the strength of the harmonics decreases is a function of the Q of the tank circuit. The higher the Q , the greater the harmonic attenuation.

At the converter, best results are achieved at some value of oscillator voltage. This means that if the oscillator injection voltage is kept at or slightly below the optimum figure, the largest amount of I.F. voltage will be obtained at the converter output. Increasing the oscillator voltage does not improve the amount of desired I.F. produced, but it does strengthen the effectiveness of the harmonics of the oscillator in the conversion process. Hence, one way to minimize the effect of the oscillator harmonics at the converter is not to overload the converter with oscillator voltage.

To minimize the formation of any second harmonic components of signals appearing across the receiver input, the circuits should possess adequate selectivity. All R.F. tubes should be of the variable mu type, preferably controlled by some small amount of AVC. Although a sharp cut-off tube gen-

erally has a greater g_m (mutual conductance) and hence capable of greater gain, it is, at the same time, more easily overloaded, a condition that readily leads to the production of harmonics. By forfeiting a little gain and using a variable mu tube with a small amount of controlling AVC, we can also reduce the production of harmonics. The AVC voltage is obtained from the limiter, or, if necessary, from the discriminator.

The foregoing discussion has, by no means, exhausted the subject of spurious responses. However, the important sources have been covered and

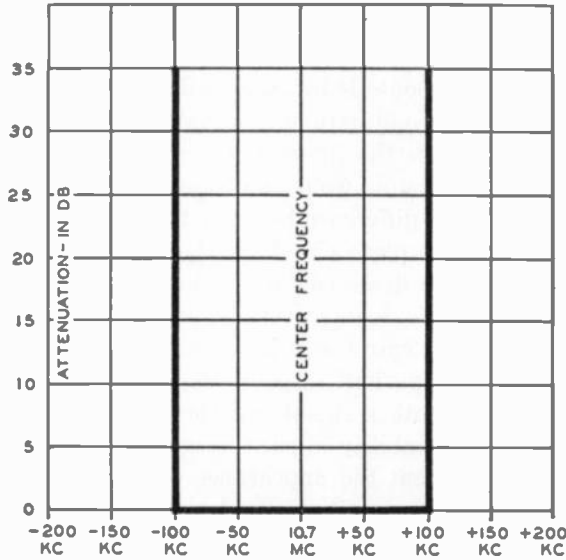


FIG. 7.3. An ideal selectivity response curve. All frequencies having values above $10.7 \text{ mc} + 100 \text{ kc}$ and below $10.7 \text{ mc} - 100 \text{ kc}$ are rejected by the circuit.

sufficient information obtained to understand their effect on receiver design and operation. It is quite evident that a high I.F. is desirable, and to a certain point, the higher the better. Against the use of a high I.F. are the disadvantages of reduced gain, necessity for greater care in choosing components to prevent excess losses, additional shielding and increased tendency toward feedback through the tubes and adjacent circuits. Also, we have the ever-present requirement of selectivity which rears its head at every turn and is certainly one of the most important factors in any consideration of operating frequency. All these factors must be weighed against each other and the trend of late is to use high intermediate frequencies. For the F-M band, the choice is 10.7 mc.

Gain and Selectivity. After the frequency of the I.F. system is chosen, we are faced next with the problems of gain and selectivity. So far as the

latter is concerned, the response should be uniform up to the sideband limits on either side of the carrier. A desired shape for the response curve is shown in Fig. 7.3. In general, though, the expense involved in attempting to construct a circuit possessing anywhere near such a response characteristic is prohibitive. The more readily achieved curve, shown in Fig. 7.4, is one practical compromise. At the plus and minus 75-kc points from the carrier, the response is down about 4 db. In terms of voltage attenuation, this represents a decrease in the ratio of 1.25 to 1. At 100 kc the attenuation has

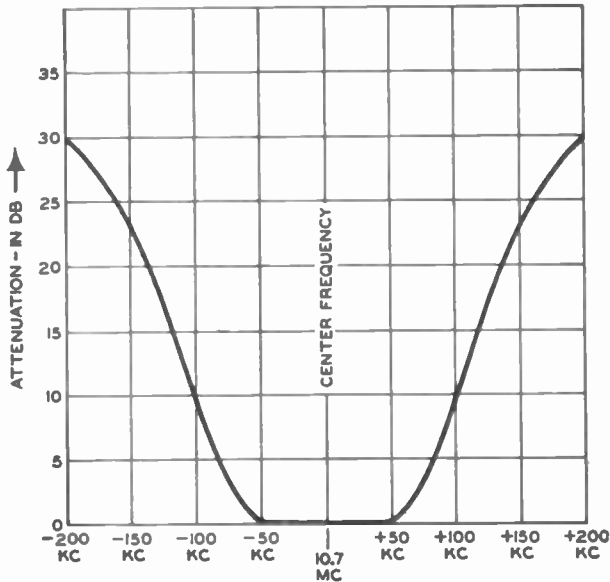


FIG. 7.4. The practical selectivity curve of an F-M receiver.

risen to 10 db (3.2 to 1), and at 200 kc from the carrier, it has reached 30 db (14 to 1). These represent figures which can be achieved in practice and which tend to provide the necessary discrimination against unwanted signals in adjacent channels. It must be understood that when we assign a figure such as a 30-db loss at 200 kc, we are referring to the total attenuation of the I.F. system and not to any single stage. Thus, suppose there are four tuned circuits in the I.F. system. Then each circuit contributes $\frac{1}{4}$ of 30db, or 7.5 db. The overall curve, though, would introduce this much loss to the particular frequency attempting to pass through the amplifiers.

When we speak of the selectivity of a tuned circuit, we refer actually to its Q . Q is defined as the ratio of the inductive reactance of the coil in the tuned circuit to the a-c resistance of that coil. When the Q is high, the impedance presented to any frequency (or narrow band of frequencies) is high,

resulting in a large voltage being developed across the coil (at that frequency). When circuits are inductively coupled this results in a large transfer of energy between them. At other frequencies, the opposition offered is not as high, and the transfer of energy is reduced. Here is where the loss, or attenuation, arises at these other frequencies.

The overall selectivity curve (Fig. 7.4) is the product of the Q of each tuned circuit in the system. Each set of coils must be properly wound, correctly spaced and, at the high frequencies, loaded down to produce the proper characteristics. Not only must the engineer determine the effect of Q on attenuation but also its effect on the gain of a stage, because the stage amplification is governed by the formula

$$A = gm \frac{\omega L Q}{2}$$

where g_m = mutual conductance of the tube and ωL = inductive reactance of the coil. This formula is based on a double tuned arrangement (Fig. 7.5) where $L_1 = L_2$ and $Q_1 = Q_2$. This is customary as it simplifies mass production.

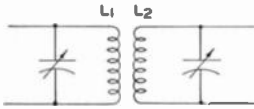


FIG. 7.5. A double tuned transformer. For ease in production, $L_1 = L_2$ and $Q_1 = Q_2$.

The special importance of gain in its relationship to limiter operation is discussed at a later point. In the tuned circuit, a high Q means a high L to C ratio. The higher L is, the more inductive

reactance (ωL) is presented to the tube and the greater the output voltage. But from the formula

$$f = \frac{1}{2\pi \sqrt{LC}}$$

where L = inductance in circuit

C = capacitance across coil

f = resonant frequency at which circuit functions

we see that, for any one frequency, large L requires a correspondingly small C . The extent to which C can be lowered depends upon:

1. The capacitance presented by the tube.
2. The stray wiring capacitance.

For a typical case, 6BA6, with a mutual conductance of 4400 micromhos, has an output capacitance of 5 mmf. Stray wiring capacitance may amount to an additional 5 to 10 mmf, depending upon how well the layout was accomplished. Thus, from these extraneous sources alone we have 15 mmf, and this will represent the minimum capacitance. If a coil is designed which utilizes only the stray capacitance, the operation of the amplifiers may be

come unstable because of the changes that occur in this capacitance with heat, length of operation, moisture and other factors. The I.F. tuned circuits would repeatedly shift frequency each time the set was operated. Good reception would be difficult. Experience has demonstrated that a physically inserted capacitor of at least 35 mmf should be included for stable operation. However, this limits the highest value of inductance for any one frequency and consequently places a ceiling on the gain possible for any stage.

The gain of a stage also varies directly with the mutual conductance of the tube, and a high g_m value is desirable. The particular usefulness of a high g_m is the large change in current flow caused by a relatively small change in grid voltage. In any plate load circuit, whether it be resistive or inductive, a large current variation gives rise to a large voltage drop and a consequent increased transfer of energy. After all, amplification within any circuit is merely a measure of the effect a small incoming voltage has in producing a large output. The greater this ratio, the higher the amplification factor. A high g_m is one way of achieving this goal.

To obtain greater values of mutual conductance, it is necessary that the grid exercise greater control over the electron current flow within the tube. Increased cathode emission, by an increase in the cathode area, would result in a greater transconductance. The interelectrode capacitance, however, would also increase by as great a factor, further limiting the L/C ratio of the tank circuit. One solution employed in the design of many miniature tubes is based upon the fact that by bringing the grid closer to the cathode, g_m would increase inversely as the square of the spacing, whereas C_{gk} would only rise inversely as the spacing. In other words, the g_m would rise more rapidly than the interelectrode capacitance. To a certain point this procedure is advisable but its limits—both in the nearness to which the grid can approach the cathode and the amount of interelectrode capacitance that can be tolerated—are quite evident.

Due to the importance of the interelectrode capacitances and the mutual conductance of a tube in determining the gain and stability of an amplifier, a "Figure of Merit" has been assigned to tubes on the basis of their ratio. Expressed mathematically,

$$\text{Figure of merit} = \frac{g_m}{C_{in} + C_{out} + C_{gp}(A + 1)}$$

- where C_{in} = the input capacitance of the tube
 C_{out} = the output capacitance of the tube
 C_{gp} = capacitance between grid and plate electrodes
 g_m = mutual conductance of tube
 A = amplification of the stage

The values for the three capacitances can be obtained from any tube man-

ual. The same is true for G_m . The value of the stage amplification, however, must be computed. The amplification is given by the equation

$$A = Z_L \times g_m$$

where Z_L is the load impedance expressed in ohms and G_m in mhos. For the tuned circuits used in the I.F. stages of an F-M receiver, we can assume a nominal value of 10,000 ohms for Z_L . With the value of A computed, we may substitute in the formula given above to determine the value of the "Figure of Merit." Table IV contains a list of current pentodes with their relative standing.

TABLE IV

Tube Type	Tube Capacitances			g_m (micromhos)	A	Figure of Merit g_m/C_{total}
	C_{gp} (mmf)	C_{in} (mmf)	C_{out} (mmf)			
6AK5	0.01	4.3	2.1	5,100	51	737
6AK6	0.12	3.6	4.2	2,300	23	216
6AU6	0.0035	5.5	5.0	4,450	44.5	418
6BA6	0.0035	5.5	5.0	4,400	44	417
6AQ5	0.35	8.3	8.2	4,100	41	131
6BC5	0.025	6.5	1.8	5,000	50	524
6BJ6	0.0035	4.5	5.5	3,600	36	355
6CB6	0.020	6.5	2.0	6,200	62	635
6CL6	0.12	11	5.5	11,000	110	370
6CU5	0.7	13.2	8.6	7,500	75	100
6SG7	0.003	8.5	7.0	4,700	47	300
6AG5	0.025	6.5	1.8	5,000	50	524

The procedure for choosing a tube for use in an amplifier is to tabulate a list of all suitable tubes and then pick the one that has the highest "Figure of Merit."

We have undertaken a fairly detailed inspection of the operation and design of high-frequency I.F. amplifiers because of the importance of these details to those who are going to service or build such amplifiers. Every single component, down to the exact layout of the wiring, must be taken into careful consideration in either repairing or building an F-M receiver. But, without an understanding of the reasons behind this necessity, repairing or construction of sets soon devolves into a mechanical procedure, easily forgotten—certainly never appreciated. It is only when one has a full appreciation of the task can the best results be obtained.

Commercial I.F. Amplifiers. Schematically, the I.F. amplifier of an F-M receiver (excluding the limiters for the present) does not differ in form from those amplifiers with which we are familiar from sound broadcast receivers. A typical diagram is shown in Fig. 7.6. The general tendency in

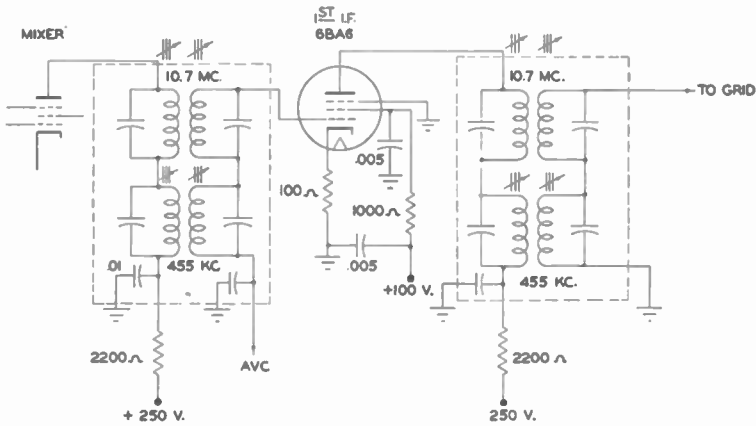


FIG. 7.6. An I.F. amplifier. The symbol above each coil indicates permeability tuning. (Courtesy RCA)

these higher frequency interstage coupling units is to vary the frequency of the tuned circuit by means of permeability tuning. The threaded core is some magnetic material, generally powdered iron. It is threaded and, as it is moved deeper into the center of the coil, it acts to increase the inductance of the coil thereby decreasing the resonant frequency. The advantage of this method of tuning over a variable capacitor is that it permits a higher inductance to be used and a wider tuning range to be covered. The capacitance associated with the inductance at a high I.F. is kept, as we have already noted, as small as practical. With a small variable capacitor only a limited tuning range is possible.

To cover a greater range, a large capacitance would be required. But to accommodate the larger capacitance, L would have to be lowered. With a decrease in the L/C ratio occasioned by such a procedure, a decrease in gain would occur—hence the advisability of permeability tuning. From time to time manufacturers employ complex coupling between stages in place of the usual two-coil arrangement shown in Fig. 7.6. Fig. 7.7 is indicative of this

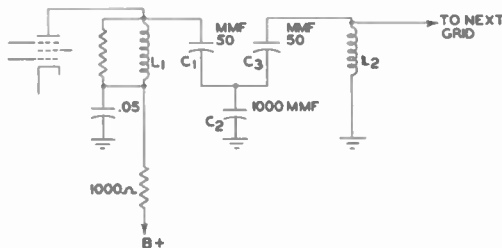


FIG. 7.7. A complex interstage coupling network.

type of a circuit. Here L_1 and L_2 are not inductively coupled to each other. Rather the transfer of energy takes place through the three capacitors C_1 , C_2 , and C_3 . C_2 is the common coupling capacitor, being common to both circuits. Since the transfer of energy takes place in this manner, the circuits are said to be capacitatively coupled. The advantages to be gained through such an arrangement are greater selectivity and a more uniform response throughout the desired range. Its disadvantage is increased cost.

There are many similar combinations possible and probably the reader has encountered some of them at one time or another. Their only difference from the double-tuned transformer is in the response they offer and the attenuation imposed on adjacent channel signals. They are widely used in television, where signal bandwidths extend to 6 mc. In such situations, simple coupling devices prove inadequate in maintaining the gain and at the same time presenting uniform response throughout the entire bandwidth. However, in F-M, 200 kc represents a relatively small percentage of the I.F. value, and conventionally coupled circuits are satisfactory.

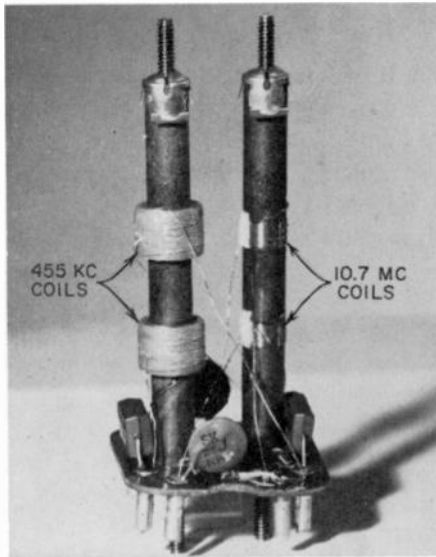


FIG. 7.8. Internal construction of an A-M, F-M, I.F. transformer. The capacitors at the bottom are connected across the coil windings.

In order to enhance the sales value of F-M receivers (due to its present restricted use), it is common practice to provide for the reception of standard broadcast stations. To keep the total cost of the receiver at a minimum, the same tubes are used for all frequencies, each band corresponding to a different position on the selector switch. One I.F. is employed for the F-M band and one set of I.F. coils for the standard broadcast band. To keep connecting wire lengths to a minimum and for space economy, both I.F. transformers are placed within the same can. In essentially all recent constructions, each set of I.F. coils is mounted on a separate hollow post. Inside each post there are two separate iron cores, one adjustable from the top of the can, the other from the bottom. In addition, the fixed capacitors

which resonate each coil winding are also placed within the housing can in order to form as compact an assembly as possible. (See Fig. 7.8.)

No difficulty is encountered by placing the transformers in series with

each other, as shown in Fig. 7.6. At any one time only one signal is being received and therefore only one set of I.F. transformers is in use. If the signal is A-M, the lower set of I.F. transformers in Fig. 7.6 is being used. The upper transformers which are in series with the A-M units do not affect the A-M signal because the inductance in the 10.7 mc coils is almost negligible at 455 kc.

When an F-M signal is being received, the upper set of I.F. transformers are used. The 455 kc units do not affect the F-M signal because the capacitor shunting the 455 kc I.F. windings acts as a short circuit to the higher 10.7 mc signal, effectively permitting the 10.7 mc signal to pass through the series 455 kc units with negligible loss.

PROBLEMS

1. What purpose do the I.F. amplifiers serve in a superheterodyne?
2. What advantages are obtained through the use of an I.F. amplifier?
3. What three basic factors must be considered in the design of an I.F. amplifier?
4. Define spurious responses.
5. Which spurious responses prove most troublesome to F-M reception?
6. What governs the choice of an intermediate frequency?
7. Explain image response. Is this phenomenon present in A-M receivers? Explain.
8. What effect does image response have on the choice of an intermediate frequency?
9. How can image response be minimized?
10. How can stations separated in frequency by an amount equal to the receiver I.F. cause interference?
11. What solutions can be employed against this type of interference?
12. What do we mean by the phrase "harmonic spurious responses"? What can be done to reduce their effect?
13. Define the phrase "selectivity of a receiver"? Why is this important?
14. Are the gain and the selectivity of a receiver related? How?
15. What is the difference between the ideal response of a tuning circuit and its actual response? Illustrate by means of graphs.
16. Can interference destroy the desired signal in an F-M receiver? Explain.
17. What relationship exists between the gain of a tuned circuit and the Q of the same circuit?
18. What tends to lower the Q of a tuned circuit?
19. What do we mean by the "Figure of Merit" of a tube? How does this factor influence the design of an I.F. amplifier?
20. Draw a tube symbol and indicate the various capacitances associated with the tube.
21. How is the "Figure of Merit" of a tube obtained?
22. Draw the schematic circuit of an I.F. amplifier.
23. How is one I.F. system used in receivers capable of receiving A-M and F-M signals?
24. What types of interstage coupling are found in the I.F. amplifiers of an F-M receiver?
25. How are most I.F. stages tuned? Why?

Chapter 8

LIMITERS

Why Limiters Are Necessary. The need for limiters in F-M sets arises not from any intrinsic property of an F-M wave but because of the limitations of some of the F-M detectors currently in use. These detectors are, in varying degrees, sensitive to amplitude variations in an F-M signal. As a result, the audio output contains voltages due to both the frequency modulation and the amplitude modulation. The insertion of a total or partial limiter substantially removes the A-M and presents to the detector a wave that is wholly F-M. Just how much limiting is required will depend upon the type of detector employed. The Foster-Sceley detector requires at least one stage and should preferably have two. The ratio detector frequently operates without any limiters, but some limiting is desirable. The 6BN6 and 6DT6 do their own limiting. However, all F-M detectors are sensitive in some degree to amplitude modulation; not one responds only to F-M.

Limiter Operation. A limiter performs its function of removing amplitude modulation by providing a constant amplitude output signal for a comparatively wide variation in input voltages. A simple illustration is shown in Fig. 8.1 where the forms of the input and output voltages of a limiter are

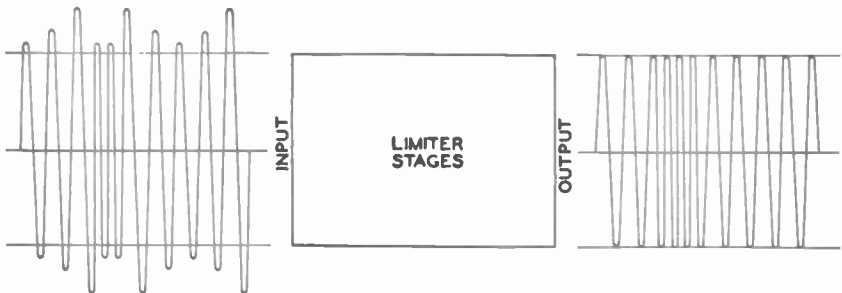


FIG. 8.1 The limiter removes the amplitude variations (modulation) from the input signal

indicated. A more exact representation of the ability of a limiter to remove A-M from a wave is given by the characteristic curve of Fig. 8.2. For all signals possessing more than a certain minimum input voltage at the antenna, the limiter produces a substantially constant output. In this region, starting at point *B* and extending to the right, the stage is purely a limiter in its action.

In the region from *A* to *B*, however, the stage functions as an amplifier because different input voltages produce different output voltages. Any signal too weak to drive the limiter beyond (or to the right of) point *B* will cause amplitude variations in the discriminator

output that are in no way part of the original F-M signal. This represents distortion. For proper operation, it is at all times necessary that sufficient amplification be given an incoming signal in order that it arrive at the limiters strong enough to drive the tube beyond point *B*.

Assuming that the F-M signal, as it leaves the transmitter, is wholly frequency-modulated with no amplitude variations, there are two general points in the path of the signal where amplitude variations may be inserted. First, there are atmospheric disturbances, completely random in nature and covering every frequency used for communication. The disturbances may come from natural sources, chiefly electrical storms, or they may come from man-made devices, such as the sparking in automobile ignition systems, electrical machines, etc. Again, there are other stations that may affect

the desired signal, although, for the most part, careful assignment of frequencies will minimize such interference. These are all sources external to the receiver and, largely, beyond the control of the set designer.

Within the receiver itself, the unequal response of the tuned circuits is the second important contributing factor toward amplitude variation within an F-M wave. The ideal response curve is a rectangle. This, however is seldom achieved with practical equipment. Economic considerations impose restrictions on the maximum cost of the tuned circuits. As a compromise, the sloping response characteristic is employed with most tuning circuits. An F-M signal applied to a tuned circuit having the response shown in Fig.

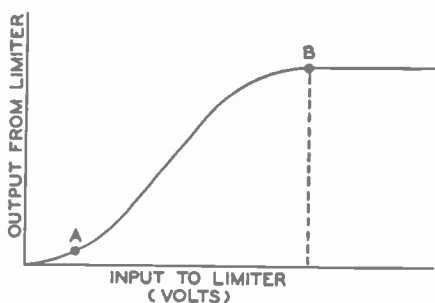


Fig. 8.2. For proper limiting action, the input voltage must be sufficiently strong to operate the limiter at point *B* or beyond.

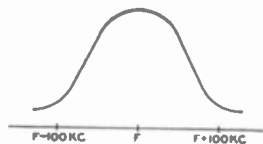


Fig. 8.3. Conventional selectivity characteristics.

8.3 will receive amplification dependent upon the frequency. The center frequency, to which the circuit is peaked, receives the greatest amplification. As the signal frequency moves closer to the ends of the band, the attenuation increases.

Influence of Limiters. In order for a limiter stage to be effective, it must function beyond the knee of its characteristic curve for all frequencies of the received F-M signal. In this respect the limiter has a very decided influence over the design and selectivity of the I.F. stages and the amplification that these stages must be capable of providing. Let us consider, for example, the single limiter characteristic shown in Fig. 8.4. According to the

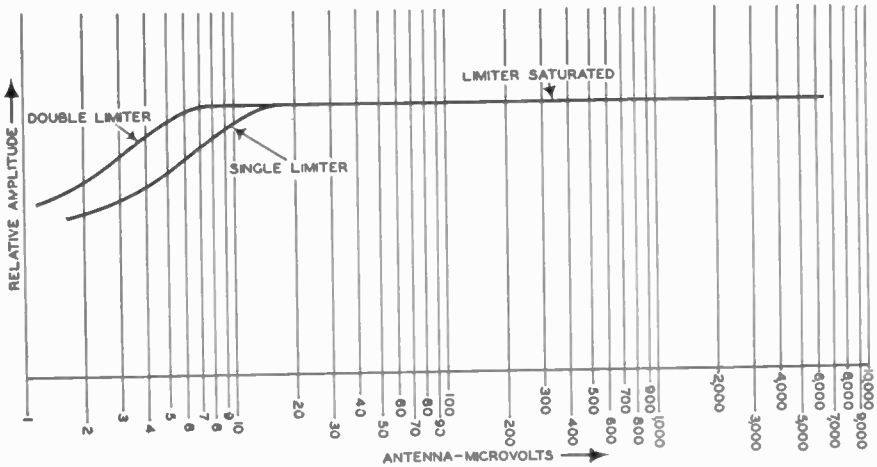


Fig. 8.4. A comparison of the input signal voltage required to produce constant output from single and double limiters.

graph, the knee of this limiter curve is reached for inputs of 15 microvolts at the antenna. As long as the incoming signal has this value, the limiter will function at saturation and no amplitude variations will be obtained at the output. Now consider the curve of Fig. 8.5. This shows the overall selectivity of a typical receiver from the antenna to the limiter. Examination indicates that the farther we move from the mid-frequency to which the receiver is tuned, the greater the attenuation to which the signal is subjected. Thus, at a point 65 kc from the carrier frequency, the signal is subjected to an attenuation rating of 1.25. Glancing back at Fig. 8.4, if 15 microvolts is required at the carrier frequency to give limiter saturation operation, then one and a quarter times this amount, or 19 microvolts, would be needed to produce the same result for that portion of the signal 65 kc off response. Frequencies 75 kc distant are attenuated about two times the mid-frequency value and, at 100 kc, the attenuation ratio is 7. Thus, if we compute the

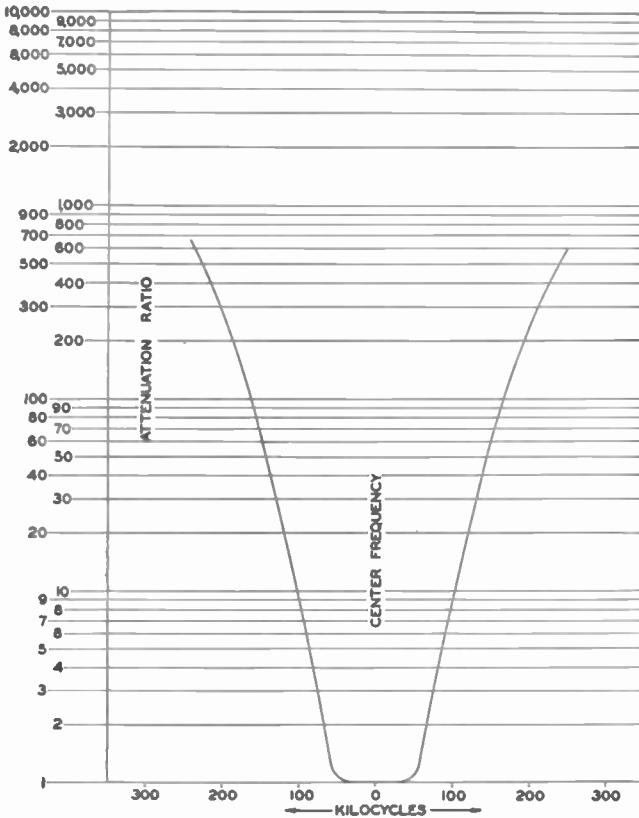


FIG. 8.5. Selectivity curve for an F-M receiver tested.

sensitivity of a receiver at points 100 kc off resonance, we find that we need signals of 105 microvolts for complete limiting action. All this is highly important in designing the characteristics of the intervening I.F. stages and the limiter. Beyond ± 100 kc, high attenuation of the signal is desirable as this minimizes adjacent station interference.

In practice, with current production limitations for the frequencies near 100 kc, the following gains can be obtained from the various stages of a well-designed F-M receiver.

Directional antenna	5x
Each R.F. stage	8x
Converter	15x
Each I.F. stage (except the one preceding the limiter)	50x
I.F. stage preceding limiter	40x
Single limiter	2.5x
Cascaded limiters	6x

For complete saturation of most limiter stages, an input voltage of 2 volts is necessary. With a 10-microvolt receiver sensitivity (meaning the receiver will reproduce clearly all signals of 10 microvolts or more) an overall gain of 2 volts/10 microvolts, or 200,000, must be available for the signal prior to its application to the grid of the final limiter stage. It then depends upon the designer what combination of R.F. and I.F. stages he will use. Note the relatively poor gain of the R.F. system as compared to each I.F. amplifier. The real usefulness of the R.F. stages lies in their partial discrimination of undesirable signals, and the boosting of weak signals to improve the signal-to-noise ratio. The front end of the receiver is particularly vulnerable to extraneous voltages and anything that will help keep these to a minimum is desirable. However, as far as gain is concerned, it is far more advantageous to add I.F. stages than R.F. stages.

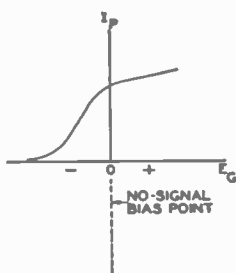


FIG. 8.6. The no-signal bias point for a limiter stage using grid-leak bias.

The lower gain obtained from the I.F. amplifier that just precedes the limiter is due to the loading effect of the limiter on the I.F. stage. Grid-current flow in the limiter (as we shall presently see) acts to lower the impedance across the I.F. output coil. The result—a decreased overall impedance with a corresponding lowering in gain. The relationship between amplifier gain and load impedance has been previously given.

Limiting Methods. There are two widely used methods for obtaining the desirable limiting action. First, we may use grid-leak bias. This limits the plate current on the positive and negative peaks of the incoming signal. Second, we can limit by employing low screen and plate voltages, producing what is known as plate-voltage limiting. Many designers combine both methods in one stage and obtain a better degree of limiting.

Grid-Leak Bias Limiting. The principle of operation of a grid-leak limiter is not difficult to understand. Initially there is no bias and the tube is operating at the zero bias point of its $E_g I_p$ curve. This is shown in Fig. 8.6. Upon the application of a signal, the grid is driven positive and current flows in the grid circuit charging up capacitor C_1 (see Fig. 8.7). It is easier for the current to charge C_1 than to attempt to force its way through the relatively high resistance of R_1 . The capacitor continues to charge throughout the entire portion of the positive half of the incoming signal.

During the negative portion of the input cycle, the accumulated electrons on the right-hand plate of C_1 flow through R_1 and the coil to the left-hand plate. The capacitor discharges because it represents a potential dif-

ference and, as long as a complete path is available, current must flow. The electrons passing through R_1 develop a potential difference with the polarity as indicated in Fig. 8.7. From the moment the input voltage departs from

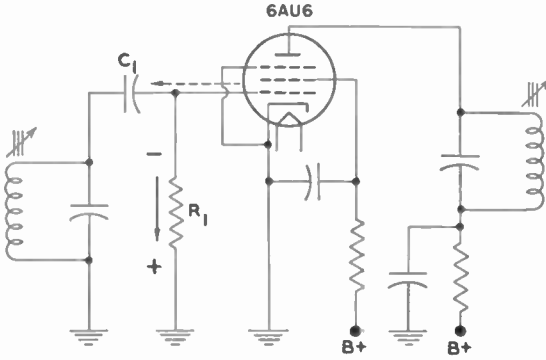


FIG. 8.7. A grid-leak bias limiter stage.

its positive values to the beginning of the next cycle, the discharge of C_1 continues.

At the start of the next cycle, there will be some charge remaining in C_1 . Hence, the grid does not draw current again until after the input voltage has risen to a positive value sufficient to overcome the residual negative voltage in C_1 . When the input voltage becomes sufficiently positive, current flows, recharging C_1 . This sequence will recur as long as an input voltage is present. Because of the fairly long time it takes for C_1 to discharge through R_1 , not all the charge on C_1 will disappear during any one cycle of input voltage. Hence, each succeeding cycle of input voltage will add a little to what remains from the previous cycle. After a few cycles a point of equilibrium is reached and the voltage across R_1 remains constant. This establishes the operating bias and the input voltage will fluctuate around this point. The diagram in Fig. 8.8 shows how the tube starts off with zero bias and, after a few cycles, reaches the equilibrium bias point.

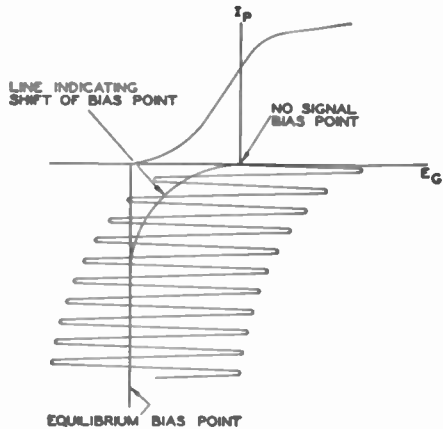


FIG. 8.8. The shift in bias potential during the first few cycles.

The equilibrium bias-point voltage is dependent upon a variety of fac-

tors. It depends upon the strength of the input voltage, the voltages on the tube elements, and the time constant of the grid capacitor and resistor. For the latter, the time constant is defined as the time, in seconds, for the capacitor to discharge 63 per cent of its voltage through R_1 . Mathematically, $t = RC$ where R = resistance in ohms, and C = capacitance in farads. Thus, if we assume values of 0.0001 mf for C and 50,000 ohms for R , we have

$$\begin{aligned} t &= R \times C \\ t &= 0.0001 \times 10^{-6} \text{ farad} \times 50,000 \text{ ohms} \\ t &= 5 \times 10^{-6} \text{ second} \end{aligned}$$

In other words, it requires 5×10^{-6} second for the charge on the capacitor—no matter what it is—to diminish to 37 per cent of its initial value.

5×10^{-6} second, or 5 microseconds, may not appear to be a very long interval, but it is long if the input wave has a frequency of 100 mc. At 100 mc, each cycle requires but 1/100 of a microsecond and 500 input cycles will pass in 5 microseconds.

Time Constant of Limiter Grid-Circuit. The ability of a grid-bias arrangement to remove amplitude variations from the incoming wave depends upon its ability to change its bias as rapidly as the peak amplitude of the incoming wave changes.

For example, suppose the amplitude of the incoming signal suddenly increases. By increasing the bias developed across R_1 , the grid-leak resistor, correspondingly, it is possible to neutralize the rise in signal amplitude. The result is then the same as with a smaller signal and a smaller bias. On the positive peaks, the bias regulates the grid so that it is barely driven positive or just enough to keep the capacitor charged and E_g constant. On the negative peaks, clipping by current cut-off occurs. Thus, by the automatic bias regulation, we can force a constant output for varying inputs.

If a variation in amplitude occurs so rapidly that the bias is unable to change with it, and thus neutralize it, then this change will appear in the output. The ability of the bias to change is a function of the time constant of the grid-leak resistor and capacitor. As we noted previously, with a capacitor of 0.0001 mf value and a resistor of 50,000 ohms, the time constant is 5 microseconds. Compared to the time of one cycle of a 100-mc carrier, an interval of 5 microseconds was long. However, even though the frequency of the incoming signal is 100 mc, it rarely occurs in practice that each individual cycle will change in amplitude. Hence, we need not make the time constant this low. For most amplitude variations, time constants from 10 to 20 microseconds are more than adequate. The greatest difficulty arises when sharp impulses of the staccato variety, such as we obtain from the ignition system of a car, reach the receiver. If the bias is unable to follow the quick rise and fall of the impulse, a plate current variation will

occur and the noise is heard in the speaker. In practice, time constants as low as 1.25 microseconds have been used and the effect of most impulses has been minimized. It must be understood that where the strength of the impulse is greater than the signal, its effect will still be felt. However, the limiter tends to limit the intensity of the impulse and the output noise is not as strong as that of an unobstructed impulse.

Limiting with Low Tube Voltages. The saturation or limiting effect of low plate and screen voltages is best seen by reference to the graphs of Fig. 8.9. In Fig. 8.9A, we have the normal extent of the $E_g I_p$ curve of a

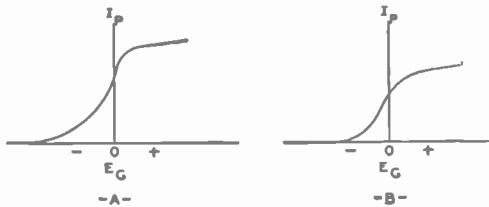


Fig. 8.9. (A) The normal characteristic curve for a sharp cut-off tube. (B) The modified curve obtained with lowered plate and screen voltage.

sharp cut-off pentode, say, a 6AU6. A sharp cut-off tube is necessary to remove fully the negative ends of the wave. It would be quite difficult to obtain a sharp cut-off in any extended cut-off tube and amplitude variations would be present on the negative peaks of the wave.

When we lower the plate and screen voltages, the extent of the characteristic curve is diminished. This is evident in Fig. 8.9B. It requires much less input signal now to drive the plate current into saturation. An F-M receiver having such a limiter would be capable of providing good limiter action with weaker input signals than if the tube were operating with full potentials.

We can, if we wish, operate a tube with lowered electrode potentials only and obtain satisfactory limiting. However, by adding grid-leak bias it is possible to raise somewhat the electrode voltages (thus providing higher gain) and still obtain good limiting. A typical circuit is in Fig. 8.10.

It was noted previously that, in general, there are two major types of noise which effect a radio wave. One is random or fluctuation noise; the other is the sharp, staccato impulse noise. For the latter type of interference, a low time-constant grid network is best suited to minimize—if not eliminate altogether—the effects of these quick-acting impulses. However, a low-valued capacitance and shunt resistance greatly damp the input tuned circuit and consequently lower the effective value of the signal reaching the grid. In order to develop a large signal voltage across a resonant circuit, a

high impedance should be presented to that signal. Shunting a low resistance across the circuit reduces the total impedance. In addition, large input signals cause sufficient grid-current flow in the input circuit actually to detune the resonant circuit. With larger time constants, better regulation of the stage is provided.

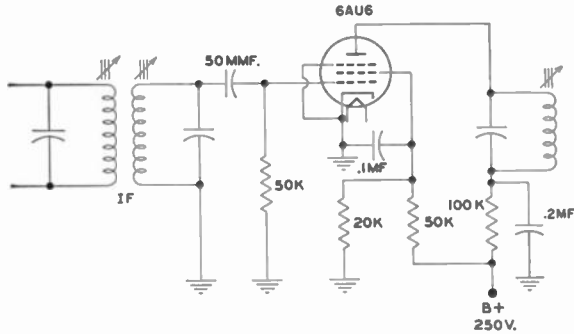


FIG. 8.10. A limiter stage functioning with lowered electrode voltages and grid-leak bias.

Cascaded Limiters. To obtain the advantages that low time constants offer in combating sharp impulses with the better regulation and higher gain of long time-constant networks, some manufacturers have used two limiters in cascade. This is shown in Fig. 8.11. The first limiter has a time constant

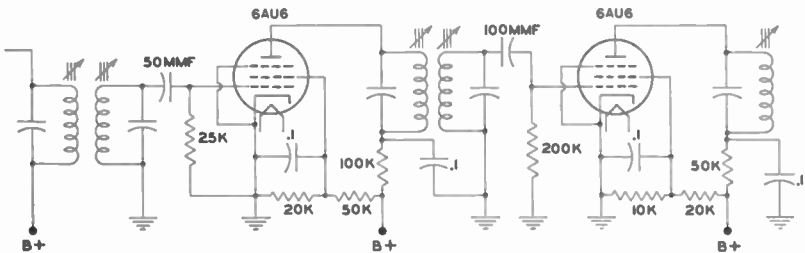


FIG. 8.11. A two-stage (or cascaded) limiter circuit.

of 1.25 microseconds. Because of the higher values possible in the second circuit, the gain of both stages averages about 6. With only one stage, values of 2.5 are usual. The increased gain permits the receiver to give full limiter action with weaker input signals.

The advantage can also be seen by reference to the graph of Fig. 8.4. With two limiters, the knee of the curve is reached by signals of 7 microvolts' strength at the receiver input. With one limiter, an input signal of 15 microvolts' strength is required. Any signal weaker than 15 microvolts with

one limiter does not give complete limiting action. When the limiter is not operated at saturation, it merely becomes an amplifier and all amplitude noise rides through as readily as the signal itself. This noise is plainly heard at the outer edges of practically all signals or when tuning between stages.

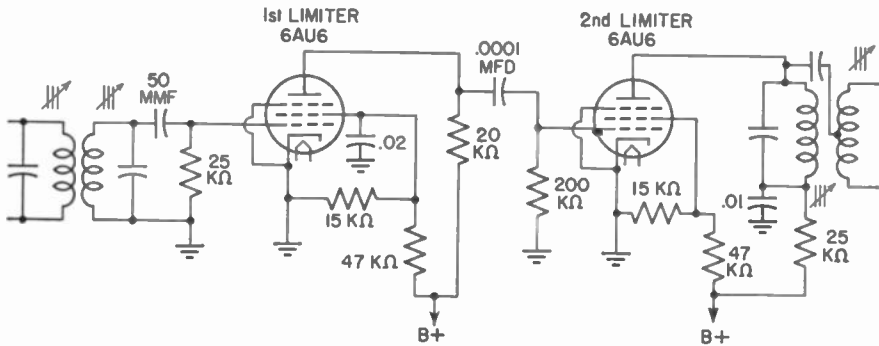


FIG. 8.12. A two-stage resistance-coupled limiter.

Commercial Limiter Circuits. There are two basic limiter circuits in commercial receivers. First, there is the single-stage unit similar to the circuit shown in Fig. 8.10. This functions satisfactorily in sets located near powerful stations where the input signal level is sufficiently strong to provide good limiting. On weak signals, however, noise will be noticeable un-

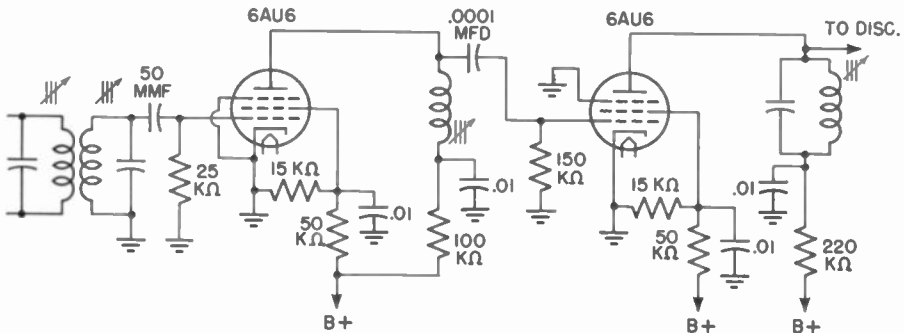


FIG. 8.13. Two limiter stages, impedance-coupled.

less a high degree of amplification is available in the stages preceding the limiter. Cascaded limiters may be either resistance-coupled, as shown in Fig. 8.12, impedance-coupled as in Fig. 8.13, or transformer-coupled. If properly shielded and constructed, the transformer-coupled unit possesses the advantage of an additional voltage gain between circuits, which is not

present in the other two circuits. The transformers must be well shielded to prevent feedback and regeneration.

The conventional discriminator is unresponsive to amplitude modulation when the signal is *exactly* at the center frequency and the circuits are symmetrical. This is true because the discriminator consists of two individual rectifiers which are balanced against each other and the output is the difference between their voltages. At the center frequency, a correctly balanced circuit yields zero output. To obtain full benefit from this feature, however, it is necessary that the resonance point of the discriminator coincide with the resonance points of all the preceding circuits. With strong signals this is difficult to obtain because the large current flow in the grid circuit of the limiter acts to detune the network, and the subsequent loading in the limiter plate circuit produces the same result in the discriminator. However, with strong signals, it is usually possible for the signal to override the noise, and any small distortion introduced because of detuning is generally negligible. With weak signals, however, the effect of noise is correspondingly greater. The advantage to be gained by having the circuits properly aligned at low signal strength is important. It is for this reason that servicing manuals urge alignment with weak generator voltages.

In the choice of a limiter tube, a tube with sharp cut-off characteristics is desirable. In addition, the mutual conductance (g_m) should be as high as possible. A large g_m will provide more current flow for any given strength of signal and permit the tube to cut-off sharply with smaller input voltages.

PROBLEMS

1. Why are limiters necessary in F-M receivers?
2. Do all F-M receivers require limiters? Explain.
3. In what section of the F-M receiver are limiters placed? Could they be placed elsewhere in the circuit? Explain.
4. Describe the operation of a limiter.
5. How does noise affect the F-M signal? Does it affect an A-M signal in a similar manner? Why?
6. What advantage do double limiters possess over single limiters?
7. Are limiters always effective in reducing interference? Explain.
8. How much gain can be normally expected from each stage preceding the F-M detector? What is the significance of this gain at the limiter? How is the total gain for a receiver computed?
9. Draw the circuit for a grid-leak bias limiter stage.
10. Explain the operation of a grid-leak bias limiter.
11. In what other circuits in radio do we find grid-leak bias?
12. What do we mean by "time constant"? Where is it used? How is it computed?
13. Why is the time constant of the limiter grid circuit important?
14. What effect does the use of lowered voltages have on tube operation?

15. Draw the schematic diagram of a limiter using grid-leak bias and lowered tube voltages. The grid circuit has a time constant of 50 microseconds.

16. Draw the schematic diagram of a two-stage, transformer-coupled limiter circuit. Assign values to all parts, using a 5-microsecond time constant in the grid circuit of the first limiter and a 25-microsecond time constant in the grid circuit of the second limiter.

17. What advantage would a 2-stage transformer-coupled limiter possess over a 2-stage resistance-coupled unit? What disadvantage?

18. Draw the schematic circuit of the I.F. and limiter section of an F-M receiver. Use two I.F. stages and two resistance-coupled limiter stages.

Chapter 9

F-M DETECTORS

Discriminator Action. The heart of the F-M receiver, where the frequency variations are reconverted to intelligible sound, is the discriminator. Since the intelligence is contained in the different instantaneous positions of the carrier frequency, the demodulating or detecting device must be such that its output varies with frequency. This is the bald, overall action of a discriminator. A closer inspection of the present discriminator will disclose that it is a balanced circuit, producing its frequency discriminations by properly combining the outputs of essentially two circuits. Whereas this form of discriminator does not readily indicate its mode of operation, its forerunner does, and we will investigate it first. The circuit is shown in Fig. 9.1.

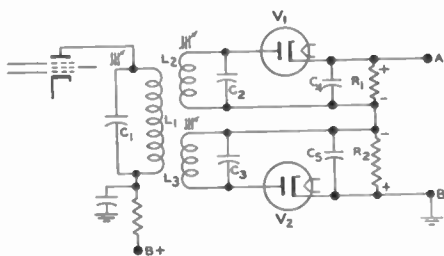


FIG. 9.1. An early form of discriminator. The secondary contains two separate windings.

L_1 and C_1 form the plate load of the last limiter stage preceding the discriminator. The tuning circuit is broadly tuned to the I.F. center value, broad enough to pass the 200 kc required by a frequency-modulated signal. L_1 , L_2 , and L_3 are all inductively coupled and the energy contained in L_1C_1 is transferred to L_2C_2 and L_3C_3 . Each of the two secondary circuits is a detector circuit in itself, complete with diode rectifier and load resistor.

They are placed end to end, as shown in Fig. 9.1, in order that their output may combine to provide both the negative and the positive half cycles of the audio wave.

To obtain the frequency discriminating action, L_2C_2 is peaked to a frequency approximately 75 to 100 kc below the center I.F. value, while C_3L_3

is peaked to a frequency the same number of kilocycles above the I.F. center point. It makes little difference which circuit is above or below, provided both are not peaked to the same point. In Fig. 9.2A, the frequency response curves of both tuned circuits are shown with respect to frequency. In Fig. 9.2B, they have been arranged—as the circuit wiring is arranged—to produce either a negative or positive output. If there is any doubt of this, trace the flow of current through V_1 and V_2 and through their respective load resistors, R_1 and R_2 . From the polarities obtained (Fig. 9.1) it can be seen that the polarity of the voltage obtained at the output terminals A and B will depend upon which resistor develops the larger voltage. If the voltage

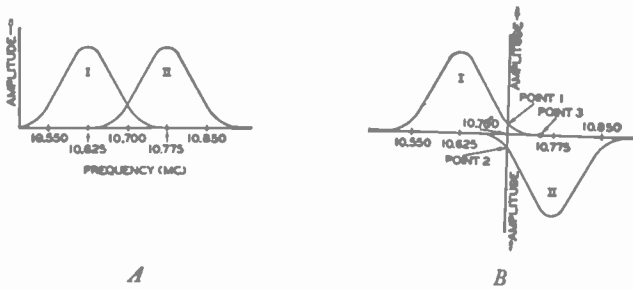


Fig. 9.2. (A) The secondary response curves. (B) The same curves as in (A) in a position corresponding to the voltage developed across the load resistor.

appearing across R_1 is larger than the voltage at R_2 , the output polarity will be positive. If, however, E_{R_2} is greater than E_{R_1} , the output voltage will be negative. Now we can discuss the discriminating action.

Each secondary resonant circuit is tuned to a different peak, and the amplitude of the voltage developed across R_1 or R_2 will depend upon the frequency of the signal present at that particular instant. For example, suppose that the incoming carrier is at the mid-point frequency, hence containing no modulation. From Fig. 9.2B, the amount of voltage developed across L_2C_2 , at this frequency, is shown by point 1 on its response curve. Assuming no loss of voltage in the tube, this same voltage will be developed across R_1 , with the top end of the resistor positive. L_3C_3 will also respond to this frequency, the amount of voltage indicated from the response curve by point 2. This, too, will be the amount of voltage present across R_2 , but of such a polarity as to oppose E_{R_1} . The output, because of the cancellation of voltages, will be zero. This is as it should be, for with no frequency modulation no output should be obtained.

When modulation is present and the frequency of the carrier begins shifting above and below the center frequency, then a definite output voltage will appear. Suppose, for example, that the carrier frequency shifts to

point 3, which is above the normal center frequency. The voltage across L_3C_3 will be greater than that present across L_2C_2 because the frequency of the carrier is now close to the resonant peak of L_3 and C_3 . Hence, while some voltage may be present across R_1 to cancel part of the voltage at R_2 , there will still be considerable voltage remaining at R_2 . This will appear across terminals A and B . As the frequency of the carrier shifts back and forth, at a rate determined by the frequency of the audio note that produced the modulation, the output will rise and fall, through positive and negative values, and the frequency variations will be converted into their corresponding audio variations.

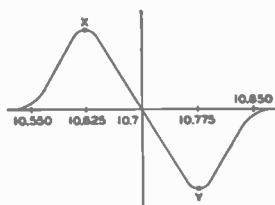


FIG. 9.3. The resultant curve derived by combining both separate curves of Fig. 9.2B.

Since the output voltage represents the difference between the potentials present across R_1 and R_2 , we can combine both response curves into one resultant curve. This has been done in Fig. 9.3. If properly designed, the response curve will be linear throughout the operating range X to Y and no distortion will be introduced when the F-M signal is converted to audio voltages. Should the extent of the linear operating section X - Y be so small that the signal variations extend into the curved sections beyond, then a good reproduction of the audio signal will not be obtained. To guard

against this, it is customary in practice to design a discriminator that is linear for a range greater than the required ± 75 kc. In most cases, the curve extends to ± 100 kc or more. The methods for aligning a discriminator are given in Chapter 11.

Capacitors C_4 and C_5 , Fig. 9.1, are for the purpose of by-passing the intermediate frequencies around the load resistors to prevent them from reaching the audio amplifiers. In choosing these capacitors, care must be taken to see that they are not made so large as to by-pass the higher audio frequencies. Usual values range around 0.0001 mf. This is large enough for the I.F., but too small for the highest audio notes. R_1 and R_2 may be about 100,000 ohms.

In this discriminator, we have essentially two separate actions in converting the F-M to audio signals. First, within the tuned circuits, the frequency modulation is changed to the corresponding amplitude modulation. If we were to picture the waveform change, it would look as shown in Fig. 9.4. As yet, however, we cannot hear any audio signals because the A-M wave is at the I.F. values. By inserting an amplitude detector—a diode—we reconvert the A-M to audio. Thus we change F-M to A-M and then obtain—with a conventional A-M detector—the audio output. It is for this reason that limiters are necessary before the discriminator. Since part of the

discriminator relies on amplitude demodulation and since this type of detector is sensitive to amplitude differences, we must use a limiter to present to the discriminator a pure F-M signal. Only in this way can we be sure that the A-M obtained is completely derived from the F-M and not from any extraneous, unwanted signals that may have found their way into the

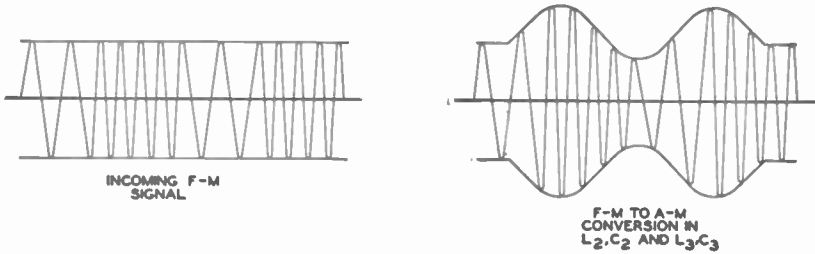


FIG. 9.4. The first step in the process of converting the F-M signal to the final audio voltage.

circuit. It will be shown later that with a ratio detector it is possible to reduce the effect of amplitude modulation with less limiting.

A Modified Discriminator. The foregoing discriminator contains one primary and two secondary capacitors. A more compact unit, one that requires only two adjustments, is the discriminator shown in Fig. 9.5. Both

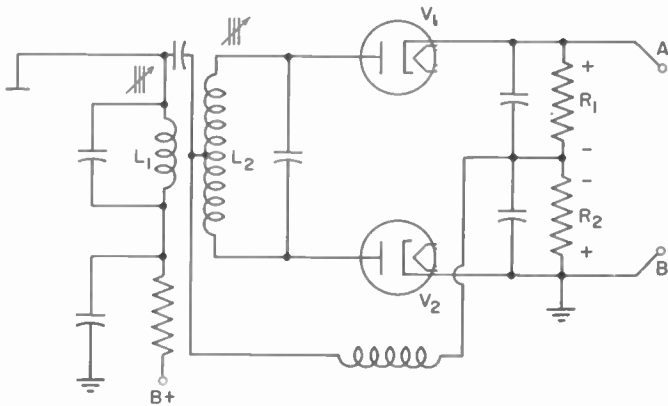


FIG. 9.5. A modified discriminator in widespread use today.

secondary windings are combined into one coil and a single capacitor is used to bring the secondary to the mid-frequency of the I.F. band. Both primary and secondary circuits are inductively tuned by means of movable cores, in accord with present practice. The discriminator output voltage does not depend now on the different response of two tuned circuits to an incoming

frequency; instead, the voltage applied to each diode depends upon the phase of the voltage in the secondary winding as compared to the phase of the primary voltage. In other words, by using a single secondary circuit, we must rely upon phase variations within that circuit to discriminate against different frequencies. In the previous discriminator, each secondary reacted separately to each frequency, their rectified outputs combining to give the audio signal. Now each different frequency alters the phase response of the secondary network and this, in turn, causes each diode to receive a different amount of voltage. As before, the rectified voltage across R_1 and R_2 gives the proper output.

Fundamentals of Coupled Circuits. Before we examine this circuit in greater detail, let us review some fundamental considerations of resonant circuits and the phase relationships of coupled circuits. In Fig. 9.5 we have two circuits that are inductively coupled. Energy from the left-hand coil is transferred to the right-hand coil through the alternating lines of force that cut across from one coil to the other. The energy in L_2 appears as a voltage within the coil and, if a complete path is available, current will flow because of the pressure of the induced voltage. To determine the phase of the current flowing in the circuit with respect to the induced voltage, we must know the frequency of the induced voltage, since a resonant circuit responds differently to each frequency. This, incidentally, is the reason resonant circuits are useful in receiving one station to the exclusion of all others.

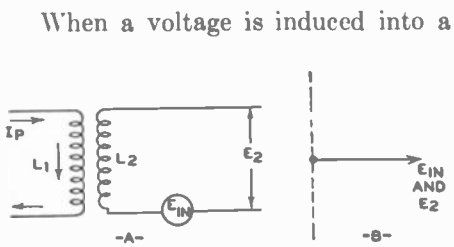


FIG. 9.6. The induced voltage in L_2 , due to I_p , is in phase with the voltage appearing across L_2 .

When a voltage is induced into a coil, such as in the resonant circuit shown in Fig. 9.5, the current that flows will develop across the coil a voltage that does not necessarily possess the same phase and amplitude as the induced voltage. This is very important and merits additional investigation. Consider, for example, two coils inductively coupled, as shown in Fig. 9.6. Because of the primary current in L_1 ,

a voltage appears across L_2 . If L_2 is an open circuit, there is no current flow, and the voltage appearing across L_2 is exactly the same voltage—in phase and amplitude—as the induced voltage. If now we connect a resistor across L_2 , as in Fig. 9.7A, permitting current to flow, then the voltage appearing across L_2 will differ from the induced voltage—both in phase and amplitude. The voltage that is induced in the secondary coil has a complete circuit consisting of the coil and the resistor. The opposition to the flow of current consists of the inductive reactance of L_2 and the resistance of R . If R is

small in comparison to the inductive reactance, the current flowing in the circuit will be governed almost completely by the inductive reactance and will lag the voltage by slightly less than 90° (this from elementary electrical theory). This is shown graphically by the two vectors, E_{in} and I_s , in Fig. 9.7B.

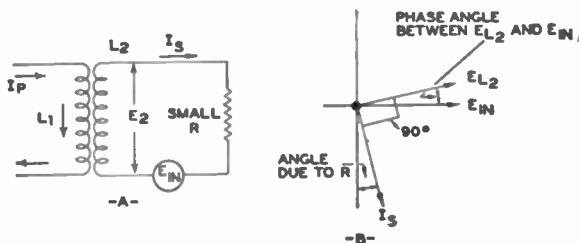


FIG. 9.7. The current and voltage phase relationship in the secondary circuit.

Now this current, in flowing around the circuit and through L_2 will develop a voltage drop across L_2 because of the coil's inductive reactance. As is true for every coil, the voltage will lead the current by 90° . On the vector graph it can be placed in the position marked E_{L_2} , 90° ahead of I_s . Thus we see the extent of the phase difference between E_{L_2} and the induced voltage E_{in} that produced it.

The two voltages also differ in amplitude because the voltage drop across the coil plus the voltage drop across the resistor must add up, vectorially, to equal E_{in} .

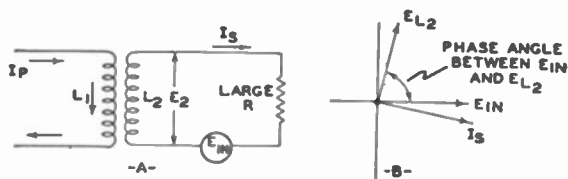


FIG. 9.8. The same circuit as shown in Fig. 9.7 except that R is now large. Note the difference in the phase angle between E_{in} and E_{L_2} .

Following the same line of reasoning, we get the vector graph of Fig. 9.8 when the inserted resistance is high in comparison to the inductive reactance. Note that, because of the high resistance, the current flow is more nearly in phase with E_{in} . However, no matter how close the voltage, E_{in} , and the circuit current approach each other in phase, the voltage drop across the inductance, L_2 (or any inductance), is always 90° out of phase with the current through it.

Secondary Resonant Circuit. The next step is to replace the resistor,

R , with a capacitor, C_2 (see Fig. 9.9). This is now a resonant circuit and the frequency of the induced voltage will determine the phase of the secondary current with respect to the voltage.

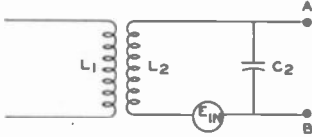


FIG. 9.9. A secondary resonant network.

As far as the induced voltage is concerned, the coil and capacitor form a series circuit. But if we consider the output terminals of the circuit, points A and B , then we have a parallel circuit. At the moment only the current *within* the circuit is of interest. Hence, we shall regard this as a series-resonant circuit.

The impedance offered to the circulating current will depend upon the frequency of the induced voltage. If the resonant frequency of the circuit is the same as the frequency of the incoming signal, a large current will be obtained. At the resonant frequency, the series impedance of the circuit is low because the inductive and capacitive reactances completely cancel each other. Above and below resonance, the reactances no longer cancel, and the resultant of the two, together with any coil and wiring resistance, presents a greater opposition to the current.

The nature of the current flow—for any one given induced voltage—is shown by the familiar curve of Fig. 9.10. At the resonant frequency, f_0 , the opposition is least and the current largest.

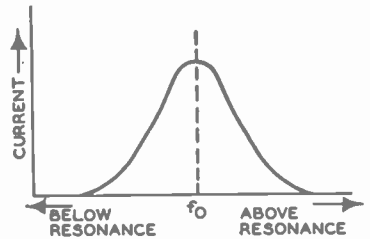


FIG. 9.10. The variation in current in a series resonant circuit.

On either side of the resonant point the current decreases because the circuit impedance increases.

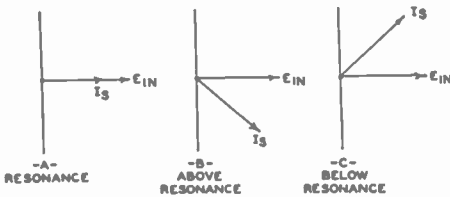


FIG. 9.11. The phase displacements between the induced voltage and the secondary current for different signal frequencies.

opposition to the current arises from the incidental resistance in the circuit. Hence, with pure resistance alone, I_s and E_{in} are in phase. Vectorially, this is shown in Fig. 9.11A. At frequencies *above* resonance, the opposition of the coil increases, because inductive reactance (X_L) is equal to $2\pi fL$. Capacitive reactance, on the other hand, decreases with frequency

according to the relation $1/2\pi fC$. Hence, *above resonance*, the X_L predominates and the current will *lag* E_{in} . This is illustrated by Fig. 9.11B. Below resonance, the opposite occurs, and the current *leads* E_{in} . See Fig. 9.11C.

The foregoing phase shifting with frequency is the basis of operation of the discriminator shown in Fig. 9.5. By properly utilizing these phase variations, we can convert F-M to audio, obtaining the same results as we did in the previous discriminator.

Discriminator Development. In order to utilize these phase shifts within the secondary resonant circuit, it is first necessary to establish a reference level. At this reference level, which is the mid I.F., no output should appear across the output terminals A and B . This is the same now as it was before, namely, no output with an unmodulated carrier. On either side of the reference point we must unbalance the voltages applied to V_1 and V_2 in order that a resultant voltage will appear across A and B . Above resonance, for example, we might have R_1 develop the greater voltage; below resonance, E_{R_2} would predominate. In this manner, swinging above and below the center frequency, we obtain positive and negative output audio voltages.

A simple resonant circuit, by itself, is unable to provide the balanced differential output we desire. Hence, some modifications must be made. First, by center-tapping the secondary coil and connecting a diode to each end, as shown in Fig. 9.12, we can obtain a balanced arrangement. This will

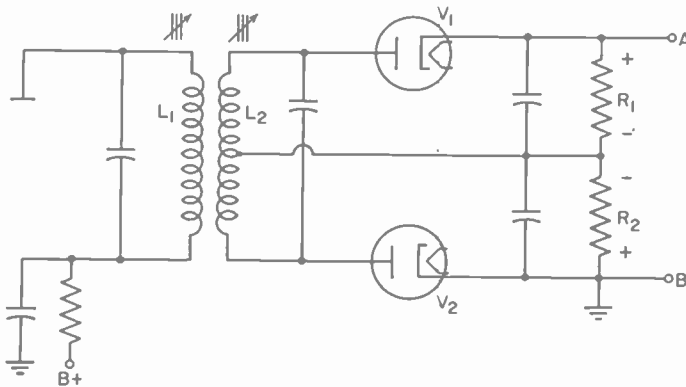


FIG. 9.12. Development of the modern discriminator. The inclusion of a center-tapped secondary and two separate rectifiers produce a balanced network.

permit a balancing of one tube against the other to give zero output at mid-frequency. The circuit, however, is still incomplete because there is nothing to unbalance the voltages applied to V_1 and V_2 above and below the center frequency. In the present circuit, V_1 and V_2 will have at *all* frequencies exactly the same applied voltage because L_2 is center-tapped and each tube

receives exactly half of any voltage developed across L_2 . As a result, the rectified voltages across R_1 and R_2 will be equal, and, because of their back-to-back placement, opposite in polarity. No resultant will ever appear at terminals A and B . To produce the proper unbalance, C_3 and L_4 are added, as shown in Fig. 9.13. If we compare this arrangement with the discriminator circuit of Fig. 9.5, we note they are identical.

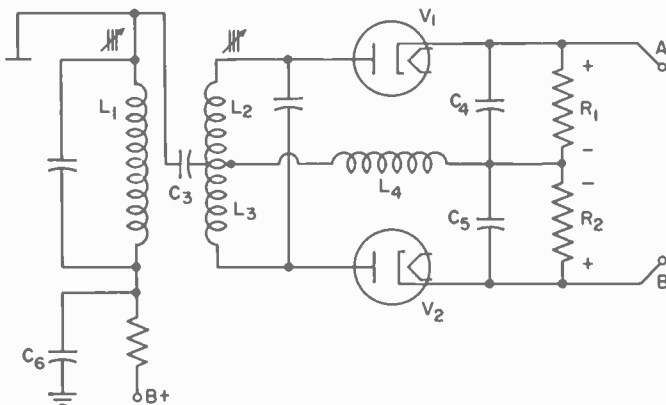


FIG. 9.13. The complete discriminator. This is known also as a Foster-Seeley discriminator.

Discriminator Operation. The circuit of V_1 includes L_2 , which is half of the secondary coil, choke L_4 , C_4 , and R_1 . The other half of the balanced secondary includes V_2 , L_3 , L_4 , C_5 , and R_2 . Note that L_4 is common to both V_1 and V_2 and is introduced to provide the reference voltage in order that an unbalance will occur at frequencies above and below the mid-frequency. One end of L_4 attaches to the top of L_1 through C_3 whereas the other end reaches ground through C_5 or, what is the same thing, the bottom of L_1 . L_4 then is in parallel with L_1 . C_3 , C_5 , and C_6 , which aid in placing the choke L_4 across L_1 , have negligible opposition at the high I.F. values. Hence, all of E_{L_1} appears across L_4 . Thus, the reference voltage for the discriminator is the voltage across the primary coil, L_1 . More on this in a moment.

The secondary coil is divided into two equal sections, L_2 and L_3 . Whenever current flows through the coil, we know that one end becomes positive with respect to the other end. The tap, being in between both ends, will be negative with respect to the positive end and positive with respect to the negative end. We could, if we wished, arbitrarily call this point *zero*. Both secondary circuits connect to the center terminal, and we can indicate the polarity of the voltage applied to each tube with respect to this center terminal. This is shown in Fig. 9.14. The arrow pointing up is the positive voltage, E_2 ; the arrow pointing down is the negative voltage, E_3 . Both are taken

with respect to the center tap. Now let us add L_4 with its voltage. We shall assume that a mid-frequency signal is being received. Hence, the current that flows through the secondary resonant circuit is in phase with the induced voltage. Later we shall extend this to include all conditions.

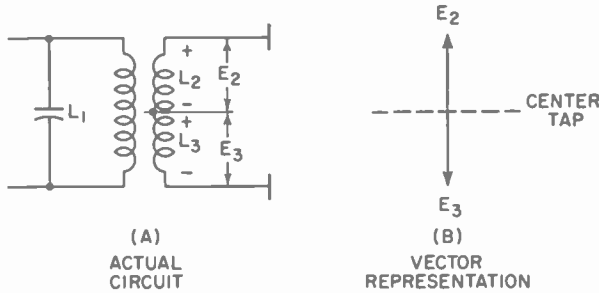


FIG. 9.14. Effect of center-tapping the secondary coil in the discriminator.

The incoming signal causes a current to flow through L_1 , the primary coil. The magnetic flux, due to the primary current, extends to the secondary winding and gives rise to the induced voltage, E_{in} . The greatest voltage is induced in the secondary when the primary flux is changing most rapidly. The flux changes most rapidly when the current does. In an a-c wave, this change occurs when the current is passing through zero, going from one

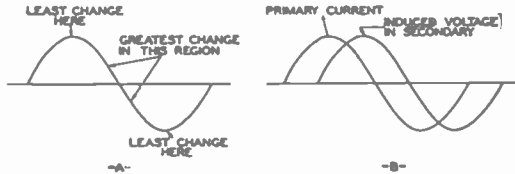


FIG. 9.15. (A) The regions of least and greatest change in an a-c wave. (B) The phase relationship between primary current and induced secondary voltage.

polarity to another (see Fig. 9.15A). However, the induced voltage is zero when the primary current is not changing, or at those times when the primary current has just reached the peak positive or negative points. If we plot these facts, we see that the current and induced voltage differ by 90° , with the current leading. This is shown in Fig. 9.15B.

The voltage, E_1 , developed across L_1 will always be 90° ahead of the current through L_1 . This, of course, is true of any inductance. If we put E_1 on a vector plot, as in Fig. 9.16A, then it must be placed 90° ahead of I_1 , the current through L_1 . E_{in} , on the same vector graph, must be placed 90° behind I_1 as determined above. Hence, we see that the primary voltage

and the induced secondary voltage differ by 180° . This is true at all frequencies. Remember that we are discussing E_1 and the *induced* secondary voltage. We have not, as yet, mentioned the voltage across the secondary coil. Most radio men are no doubt familiar with the results which we have just evolved. But, for the sake of clarity, this was methodically determined in order to show how these principles are used in the discriminator.

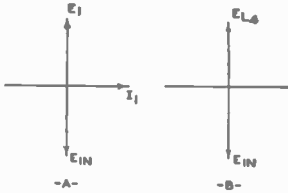


FIG. 9.16. (A) The phase differences between E_1 , E_{in} and I_1 . (B) E_1 replaced by its equivalent E_{L_4} .

We can replace E_1 by E_{L_4} because, as has been already shown, E_{L_4} is in parallel with E_1 and is the same voltage. Also, we can drop the primary current vector, it having already served its purpose. The result is merely E_{in} and E_{L_4} , shown in Fig. 9.16B. To the vectorial diagram, we must still add E_2 and E_3 and then combine them with E_{L_4} . Fig. 9.13 shows quite definitely that the voltage acting at V_1 , for example, is the combination of E_{L_4} and E_2 . Likewise, the voltage

at V_2 is E_{L_4} and E_3 . From their relative magnitudes we can determine which will predominate, and this will indicate the output voltage.

To cover fully the operation of the discriminator, frequencies above, below, and at the mid-point value must be considered. The first illustration will cover the operation with mid-frequency signals. This represents the case of no modulation. Secondly, we will use a frequency above the center point; lastly, a frequency below. Thus, the entire range of an F-M signal will be covered. The output voltage from the discriminator will give the circuit response to input frequency variations.

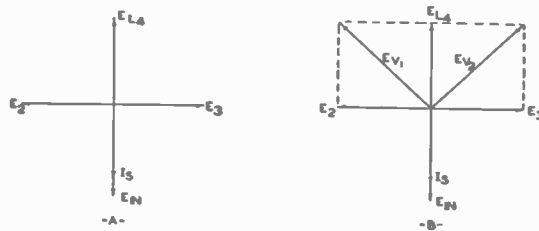


FIG. 9.17. (A) Vector diagram for an unmodulated signal. (B) The vector addition of E_2 with E_{L_4} and E_3 with E_{L_4} .

When the incoming frequency is at the I.F. mid-point, the induced voltage, E_{in} , produces a current in the secondary circuit which is in phase with E_{in} . At the resonant frequency, the impedance presented to any voltage is purely resistive. Hence, current and voltage are in phase. On the vector diagram, Fig. 9.17A, we can indicate the in-phase condition by drawing

E_{in} and I_s (I_s , the secondary current) along the same line. The length of each vector differs because of the resistance in the circuit.

The voltage developed in L_2 and L_3 , because of I_s , differs in phase from I_s by 90° . Again, this condition is true of any inductance. On the vector diagram, we can insert E_2 and E_3 , the voltages developed across L_2 and L_3 , both 90° from I_s . E_2 and E_3 are on opposite sides of I_s , because of the center tap on the secondary coil (see Fig. 9.14). E_2 and E_3 , taken with reference to the center tap, are 180° out of phase with each other. To show this graphically, in Fig. 9.17A, they are placed pointing in opposite directions.

We can now combine E_2 and E_{L_4} to arrive at E_{V_1} , the effective voltage applied to tube V_1 . Also E_3 and E_{L_4} to derive E_{V_2} , the voltage applied to V_2 . It is evident from Fig. 9.17B that E_{V_1} and E_{V_2} are both of equal amplitude. Hence, the same current will flow through each tube and the same voltage will appear across R_1 and R_2 . Since the voltages across these two resistors are in opposition, no output will be obtained. For an F-M signal containing no modulation, no output is derived.

Conditions with Higher Input Frequency. Consider next the circuit conditions when the modulated F-M signal swings to a higher frequency. At any frequency, the voltage induced in the secondary of the discriminator will be 180° out of phase with the voltage across the primary. Hence, to start the vector diagram, we can insert these two voltages (see Fig. 9.18A). Again, for E_1 we can substitute E_{L_4} , and thus far the conditions are the same as for the previous case.

A voltage having a frequency higher than the resonant frequency of the secondary circuit will encounter more opposition than a voltage at the resonant frequency. Also, because this is a series-resonant circuit (at least, as far as the induced voltage is concerned), the higher frequency will meet more opposition from the coil than the capacitor. Hence, in the present instance, the coil will cause the current to lag behind E_{in} (see Fig. 9.18A). Now we

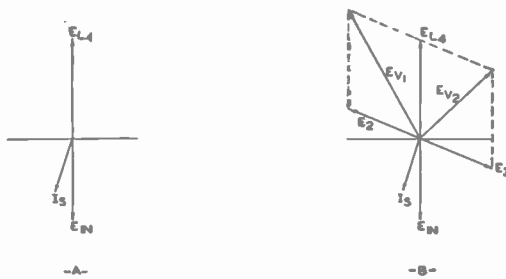


FIG. 9.18. Circuit phase conditions when the signal frequency is above the I.F. center value.

can draw in E_2 and E_3 , since they differ from the current (I_s) by 90° . I_s is responsible for E_2 and E_3 appearing across the secondary coil, and there is always a 90° phase difference between voltage and current in a coil.

Adding E_2 and E_{L_4} , E_3 and E_{L_4} , vectorially, we see that the resultant vectors are no longer equal. In this case, E_{V_1} is greater than E_{V_2} . Consequently, R_1 will develop a larger voltage than R_2 , and the output voltage

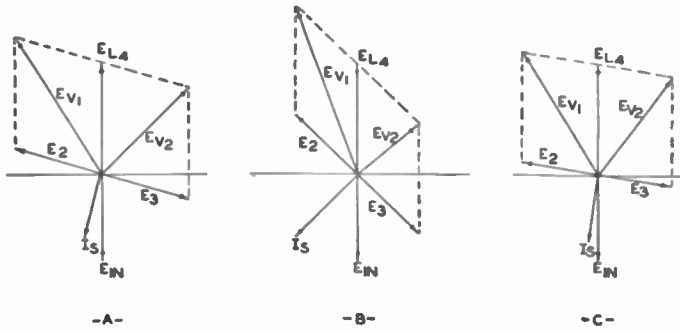


FIG. 9.19. Several additional off-frequency conditions. The input frequency is high.

will be positive with respect to ground. The amount by which I_s will differ in phase from E_{in} depends upon the incoming frequency. The farther this frequency is from the mid-frequency or resonant point of the secondary circuit, the more I_s and E_{in} will differ in phase. As the position of I_s changes, the positions of E_2 and E_3 will likewise change. Several off-frequency conditions are shown in Fig. 9.19, illustrating the phase shifting that occurs.

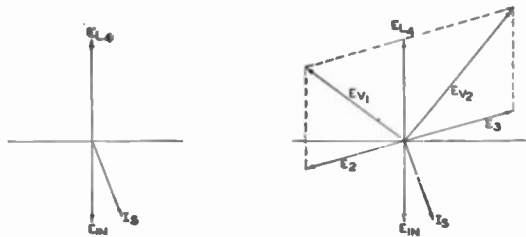


FIG. 9.20. The shift in vectors when the signal is below the I.F. mid-point.

Input Frequency Below Mid-Frequency. The phase analysis when the frequency swings below the mid-frequency value is just as readily accomplished. We can draw in immediately the vectors E_{in} and E_{L_4} . I_s follows next. Its position is ahead of E_{in} because, as the frequency drops below resonance, the capacitive reactance of the capacitor becomes greater than the impedance of the coil. In a capacitive circuit, current leads voltage.

This places I_a ahead of E_{in} . E_2 and E_3 are 90° out of phase from I_a , and these can be added to the vector diagram. Addition with E_{L_4} shows the unbalance of voltages, this time favoring V_2 (see Fig. 9.20). The output voltage from the discriminator now becomes negative.

As a summary of the overall picture, the following points are evident:

1. At the resonant frequency, which is the mid-frequency of the carrier, no output is obtained from the discriminator.
2. At frequencies off resonance, an unbalance is created in the circuit, and an output voltage is obtained.

The unbalance that arises from the shifting frequency is made linear with respect to frequency in order that the audio output will be a faithful reproduction of the modulation as effected at the transmitter. The same discriminator S-shaped curve derived for the previous discriminator (Fig. 9.3) is obtained again. The linear section of the curve X - Y generally extends for 200 kc, although the frequency shifts are restricted to a maximum of 75 kc on either side of the carrier. The extension is added protection to insure faithful reproduction.

Other Discriminators. Although the preceding discriminator is widely used, slight variations have been introduced by certain manufacturers. A modified discriminator employed in many sets is shown in Fig. 9.21. The

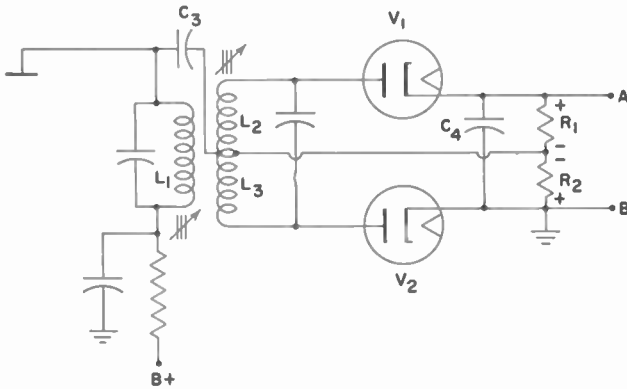


FIG. 9.21. A modified discriminator.

difference between this unit and the preceding circuit is to be found in the elimination of L_4 and the use of only one capacitor across the output instead of two. It may appear that E_{L_4} , the reference voltage so necessary for the proper operation of the discriminator in Fig. 9.13, has been eliminated. Actually this is not so. The reference voltage is still present, but in a slightly altered position.

If we trace the circuit from the top of L_1 through C_3 and R_2 to ground, we see that R_2 is in parallel with L_1 and hence E_1 appears across R_2 . If we travel around the network containing V_2 , we find that both E_3 and E_1 (from across R_2) act on V_2 . This is identical with the preceding case, except that the E_1 voltage was obtained from choke L_4 . Now we have eliminated L_4 and placed E_1 , the reference voltage, across R_2 . Thus, in this circuit, R_2 performs double duty. Not only does it develop the rectified voltage for V_2 , but it also serves to apply E_1 to V_2 .

In the V_1 network, R_1 is also found to be in parallel with L_1 . The circuit extends from the top of L_1 through C_3 to R_1 , to the top of R_1 , then through C_4 to ground. Both C_3 and C_4 offer slight opposition to any high I.F. Thus, R_1 is substantially in parallel with L_1 . Hence, R_1 is to V_1 what R_2 is to V_2 . No unwanted intermediate frequencies reach the following audio stages because of C_4 . Its low reactance to intermediate frequencies by-passes them from point A. However, being only 0.0001 mf or so, it does not provide an easy path for the rectified audio frequencies, and these reach the audio amplifiers. Aside from these changes, the operation of this circuit is identical with the preceding network.

The advantage of the circuit lies only in the fewer parts required. R_1 and R_2 must be kept as high as possible because they are in parallel across L_1 . A low value would decrease the impedance of L_1 with a subsequent decrease in gain. With the choke, R_1 and R_2 are isolated from L_1 , and their effect is not as marked. As a general rule, R_1 and R_2 may be higher in value in the unit of Fig. 9.21 than that of Fig. 9.13.

The circuit diagram of Fig. 9.21 may be drawn in many ways, one variation being shown in Fig. 9.22. At first glance there appears to be no relationship between the two circuits. On tracing out the wiring, however, it becomes evident that exactly the same discriminator is used. Each manufacturing firm draws schematics along its own peculiar lines and sometimes it requires a little patience to recognize the similarity of identical units.

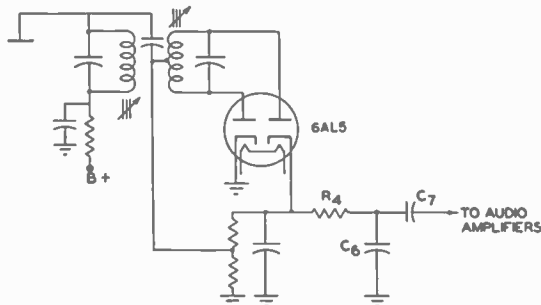


FIG. 9.22. The same discriminator as shown in Fig. 9.21, but slightly rearranged.

C_6 and R_4 of Fig. 9.22 form a de-emphasis network to return the signal to its correct values. It will be recalled that pre-accentuation or pre-emphasis is used at the transmitter in order to raise the relative intensity of the higher audio frequencies in an attempt to better the signal-to-noise ratio. At the receiver the reverse action must be instituted to bring the signal back to its proper form. C_7 is a coupling condenser.

F-M Ratio Detector. The inherent limitations of the preceding discriminators necessitated a limiter stage. The limiter removed all traces of amplitude modulation and presented a pure F-M wave to the discriminator. The disadvantage of employing a limiter is the excessive amplification that must be given to the signal before it reaches the limiter. This is necessary in order that the weakest desired signal be brought up to the point where it will drive the limiter to saturation.

If a detector could be built which would develop an output that was independent of amplitude variations in the incoming signal, then a great simplification in circuit design could be effected. It would mean that perhaps one, at most two, I.F. stages would be needed. Recently, the ratio detector was developed which does not respond as readily to amplitude variations as the Foster-Seeley circuit itself, without a limiter.

To see why a ratio detector is more immune to amplitude variations in the incoming signal, let us compare its operation with that of the Foster-Seeley discriminator.

In the discriminator circuit of Fig. 9.13, let the signal coming in develop equal voltages across R_1 and R_2 . This would occur, of course, when the incoming signal is at the center I.F. value. Suppose that each voltage across R_1 and R_2 is 4 volts. When modulation is applied, the voltage across each resistor changes, resulting in a net output voltage. Say that the voltage across R_1 increases to 6 volts and the voltage across R_2 decreases to 2 volts. The output voltage would then be equal to the difference between these two values, or 4 volts.

However, let us increase the strength of our carrier until we have 8 volts, each, across R_1 and R_2 , at mid-frequency. With the same frequency shift as above, but with this stronger carrier, the voltage across R_1 would rise to 12 volts and that across R_2 decrease to 4 volts. Their difference, or 8 volts, would now be obtained at the output of the discriminator in place of the previous 4 volts. Thus, the discriminator responds to both F-M and A-M. It is for this reason that limiters are used. The limiter clips off all amplitude modulation from the incoming signal, and an F-M signal of constant amplitude is applied to the discriminator.

When unmodulated, the carrier produced equal voltages across R_1 and R_2 . Let us call these voltages E_1 and E_2 , respectively. With the weaker carrier, on modulation, the ratio of E_1 to E_2 was 3 to 1 since E_1 became

6 volts and E_2 dropped to 2 volts. With the stronger carrier, on modulation, E_1 became 12 volts and E_2 dropped to 4 volts. The ratio was again 3 to 1, the same as with the previous weaker carrier. Thus, whereas the difference of voltage varied in each case, the ratio remained fixed. This demonstrates, in a very elementary manner, why a ratio detector could be unresponsive to carrier changes.

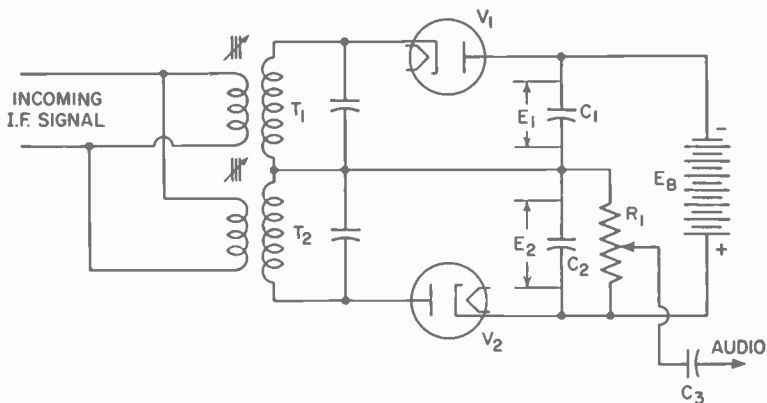


FIG. 9.23. Preliminary form of the ratio detector.

An elementary circuit of a ratio detector is shown in Fig. 9.23. In this form the detector is similar to the detector of Fig. 9.1 where each tube has a completely separate resonant circuit. One circuit is peaked slightly above the center I.F. value (say T_1); the other is peaked to a frequency below the center I.F. value (say T_2). The output voltage for V_1 will appear across C_1 , and the output voltage for V_2 will be present across C_2 . The battery, E_b , represents a fixed voltage. Since C_1 and C_2 are in series directly across the battery, the sum of their voltages must equal E_b . Also, due to the manner in which the battery is connected to V_1 and V_2 , no current can flow around the circuit until a signal is applied. Now, although $E_1 E_2$ can never exceed E_b , E_1 does not have to equal E_2 . In other words, the ratio of E_1 to E_2 may vary. The output voltage is obtained from a variable resistor connected across C_2 .

When the incoming signal is at the I.F. center point, E_1 and E_2 will be equal. This is similar to the situation in the previous discriminators, especially Fig. 9.1. However, when the incoming signal rises in frequency, it approaches the resonant point of T_1 , and the voltage across C_1 likewise rises. For the same frequency the response of T_2 produces a lower voltage. As a consequence, the voltage across C_2 decreases. However, $E_1 + E_2$ is still equal to E_b . In other words, a change in frequency does not alter the total voltage but merely the ratio of E_1 to E_2 . When the signal frequency

drops below the I.F. center point, E_2 exceeds E_1 . Again, however, $E_1 + E_2$ equals E_b . The audio variations are obtained from the change of voltage across C_2 . Capacitor C_3 prevents the rectified d-c voltage in the detector from reaching the grid of the audio amplifier. Only the audio variations are desired.

The purpose of E_b in this elementary explanatory circuit is to maintain an output audio voltage which is purely a result of the F-M signal. E_b keeps the total voltage ($E_1 + E_2$) constant, while it permits the ratio of E_1 to E_2 to vary. As long as this condition is maintained, we have seen that all amplitude variations in the input signal are without effect.

The problem of deciding upon a value for E_b is an important one. Consider, for example, that a weak signal is being received. If E_b is high, the weak signal would be lost because it would not possess sufficient strength to overcome the negative polarity placed by E_b on the tubes, V_1 and V_2 . The tubes, with a weak input voltage, could not pass current. If the value of E_b is lowered, then powerful stations are limited in the amount of audio voltage output from the discriminator. This is due to the fact that the voltage across either capacitor—either C_1 or C_2 —cannot exceed E_b . If E_b is small, only small audio output voltages are obtainable. To get around this restriction, it was decided to let the average value of each incoming carrier determine E_b . Momentary increases could be prevented from affecting E_b by a circuit with a relatively long time constant.

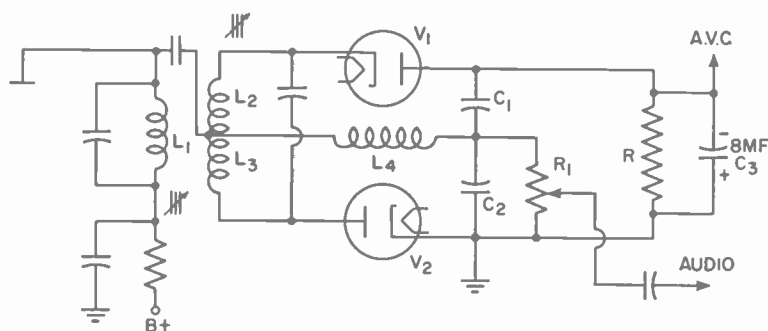


FIG. 9.24. Practical form of the ratio detector.

The practical form of the radio detector is shown in Fig. 9.24. It uses the phase-shifting properties of the discriminator of Fig. 9.13. R and C_3 take the place of E_b , and the voltage developed across R will be dependent upon the strength of the incoming carrier. Note that V_1 and V_2 form a series circuit with R (and C_3) and any current flowing through these tubes must flow through R . However, by shunting the 8-mf electrolytic capacitor across R we maintain a fairly constant voltage. Thus, momentary changes in car-

rier amplitude are merely absorbed by the capacitor. It is only when the *average* value of the carrier is altered that the voltage across R is changed. The output audio frequency voltage is still taken from across C_2 by means of the volume control.

Since the voltage across R is directly dependent upon the carrier strength, it may also be used for A.V.C. voltage. The polarity of the voltage is indicated in Fig. 9.26. With this detector, it becomes possible to design an F-M receiver containing only two I.F. stages. The reduction in costs is considerable and permits marketing F-M sets in a range comparable to the present A-M receivers. The detector possesses the disadvantage of being more difficult to align, and greater care must be taken to obtain a linear characteristic. Distortion tends to become more noticeable at high input voltages, although this can be minimized by special compensations of the circuit.

Ratio Detector Modifications. The ratio discriminator shown in Fig. 9.24 represents but one form of this circuit. In Fig. 9.25 there is a modifica-

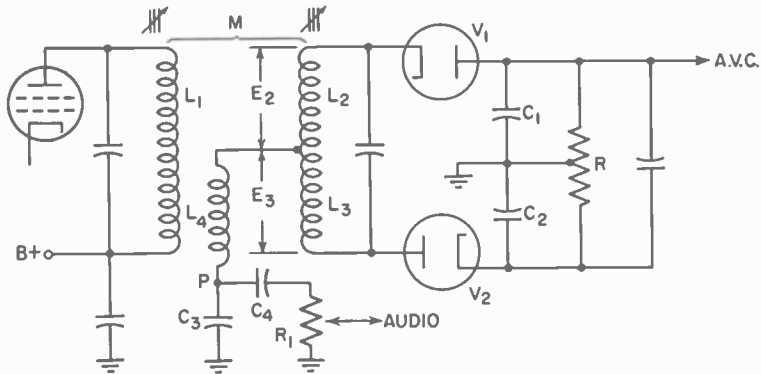


FIG. 9.25. A symmetrical ratio detector.

tion which possesses greater symmetry. Furthermore, it has the advantage of making available a voltage which can be used, in conjunction with a tuning eye, to indicate when the station is properly tuned in. L_1 , L_2 , and L_3 are inductively coupled, which is similar to the previous discriminators. However, in place of the former, direct capacitive connection between L_1 and the secondary coil center-tap, we now have L_4 . It will be remembered that it is due to the presence of E_1 in the secondary as a reference voltage that this circuit is able to function. In Fig. 9.25, we have, by using L_4 , substituted it for the previous method of obtaining this reference potential. L_4 consists of several turns of wire which are closely coupled to L_1 . Hence, the voltage induced in L_4 will remain constant so long as the primary voltage is constant. Since E_{L_4} depends upon E_1 and not upon the secondary circuit, it

can be used as the reference voltage in place of the previous arrangement. If E_{L_4} is varied, it would not affect the *relative* amplitudes or phases of any of the secondary voltages.

To understand better the other modifications, the circuit of Fig. 9.25 has been rearranged (see Fig. 9.26). The path from the center tap on the sec-

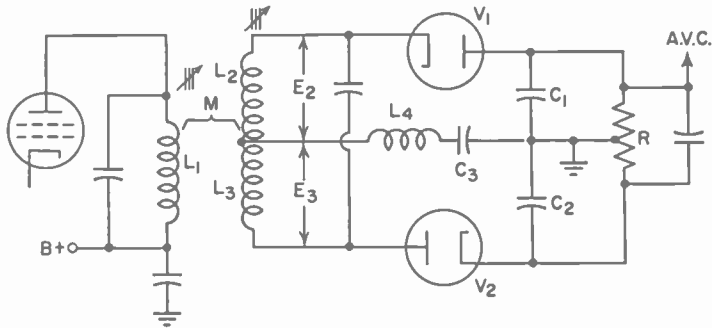


FIG. 9.26. The circuit of Fig. 9.25 rearranged to better indicate its mode of operation.

ondary coil to the connection between C_1 and C_2 contains L_4 and C_3 . Disregard C_4 and R_1 for the moment, as they are merely placed across C_3 to feed the audio output to the following amplifiers. The voltage which is applied to V_1 consists of E_{L_4} and E_2 . Across V_2 we have E_{L_4} and E_3 . At the center I.F. value, both tubes receive the same voltage, with C_1 and C_2 charging to the same potential. But what of the potential across C_3 ?

At this moment it is zero. The reason can be found if we trace the current paths for each tube. The current in V_2 flows up through L_3 , through L_4 to C_3 and C_2 . Current will flow in this path until the voltage across C_3 and C_2 equal the voltage across the tube. Then the flow ceases. The current from V_1 will flow from its plate to C_1 and C_3 , then L_4 and finally up through L_2 to V_1 again. As before, there will be a movement of electrons around this path until C_1 and C_3 charge to the potential at the tube. Note, however, that the current from V_2 flows through L_4 and C_3 in a direction *opposite* to the current produced by V_1 . Hence, what actually happens is that these two currents in the middle branch buck each other. At mid-frequency the resultant is zero. The voltage across C_3 , then, is also zero. Hence, the current in the circuit flows from V_2 , through L_3 and L_2 to V_1 and through C_1 and C_2 back to V_2 . At mid-frequency, point P , Fig. 9.25, is zero with respect to ground.

At other frequencies, the voltage applied to each tube will differ with the result that a net current flows through C_3 and L_4 . Which way the current flows will depend upon which tube receives the greater voltage. It is thus

evident that the potential variations across C_3 will vary directly with frequency and consequently represent the audio or demodulated voltage. By means of a coupling capacitor and a volume control, we can feed these audio variations to the succeeding amplifiers.

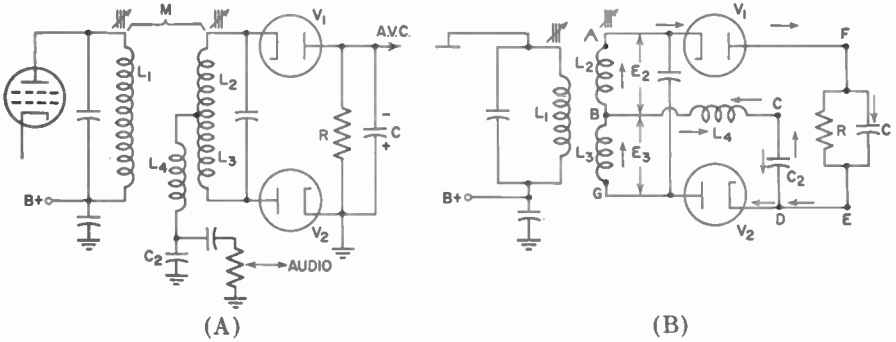


FIG. 9.27(A). Another form that the ratio detector can assume. (B) A rearrangement of Fig. (A).

Another arrangement of the ratio detector is shown in Fig. 9.27A. This is similar to the circuit of Fig. 9.24 except that capacitor C_2 has been retained and C_1 has been eliminated. The audio voltage is obtained from C_2 , the same as in Fig. 9.24. L_4 is coupled to L_1 and furnishes the reference voltage which causes the potentials applied to V_1 and V_2 to change with frequency.

In order to understand the operation of the circuit without C_1 , it must be kept in mind that the voltage across C_2 is determined by:

1. The potential of R and C . This, in turn, is fixed by the average amplitude of the incoming F-M signal.
2. The frequency of the incoming signal.
3. The *relative* currents flowing through V_1 and V_2 . This, of course, depends upon No. 2.

To demonstrate better the various current paths in the secondary network, the rearranged diagram of Fig. 9.27B will be used. Note that no connections have been altered, merely the relative placement of several components.

The voltage applied to V_2 is the vector sum of E_{L_4} and E_3 (see Figs. 9.18 and 9.20). Similarly, the voltage active across V_1 is composed of E_{L_4} and E_2 . As the frequency shifts in response to modulation, the total voltages at V_1 and V_2 will follow suit. This has already been noted in previous paragraphs.

Consider, now, the current paths for each of the tubes. V_1 is part of

the complete path $AFEDCBA$. Its current can also flow through path $AFEDGBA$. For V_2 , the two paths are: $GBAFEDG$ and $GBCDGB$. In other words, currents from each tube can flow around the outer path ($GBAFEDG$) or part of each can be diverted through L_4 and C_2 of the center path.

When the total voltages applied to each tube are equal (at mid-frequency), no current flows through L_4 and C_2 . This is true in all ratio detectors. At points other than mid-frequency, however, the current of one tube is greater and a portion of it does pass through L_4 and C_2 . Hence, the voltage across C_2 will be a function of frequency. Due to the fact that each tube is connected into the circuit in an opposite manner, their currents (from V_1 and V_2) flowing through L_4 and C_2 will likewise be opposite. Consequently, E_{C_2} will possess one polarity for frequencies above resonance and the opposite polarity for frequencies below resonance. By attaching a potentiometer across C_2 , we can obtain an audio voltage for the audio amplifiers.

It may not be amiss at this point to note again the difference between the ratio detector and its predecessor in Fig. 9.5. The latter unit operates on the difference of the output voltages of two diode detectors. The diode load resistors are connected with their voltages opposing each other. The resultant of these two then becomes the output audio voltage. The response of this discriminator to amplitude modulation requires that a limiter (at least one) always precede it. Consequently, a comparatively large amount of amplification must be available so that the signal can be in a position to drive the limiter to saturation. Economically, this requirement raises the price of the receiver.

In the ratio detector, the two diodes are connected in series, and a controlling voltage is established in the circuit which is dependent upon the average value of the incoming carrier. Because of the long time-constant of the R - C filter in the network, instantaneous changes in signal amplitude are prevented from affecting the audio output voltage. Furthermore, this control voltage sets the limit to the maximum audio voltage than can be obtained. The ratio detector is more immune to amplitude modulation than the Foster-Seeley discriminator if the latter is considered by itself, without any limiter. However, when a limiter is added, the Foster-Seeley circuit possesses a slight edge in performance. The ratio detector, however, is more desirable economically since the amplification required ahead of this stage is less and advantage can be taken of this fact to reduce the number of stages in the I.F. system. Many manufacturers attempt a compromise by having the I.F. amplifier preceding the ratio detector operate as a partial limiter. While this reduces the gain of the I.F. system, it does improve the performance of the detector for F-M signals possessing large amounts of amplitude modulation.

The 6BN6 Gated-Beam Tube. Another approach to a combined limiter-discriminator combination, one that differs considerably from any of the previous circuits, is provided by the 6BN6 gated-beam tube. This tube, designed by Dr. Robert Adler of the Zenith Radio Corporation, possesses a characteristic such that, when the grid voltage changes from negative to positive values, the plate current rises rapidly from zero to a sharply defined maximum level. This same maximum value of plate current remains, no matter how positive the grid voltage is made. Current cut-off is achieved when the grid voltage goes about 2 volts negative.

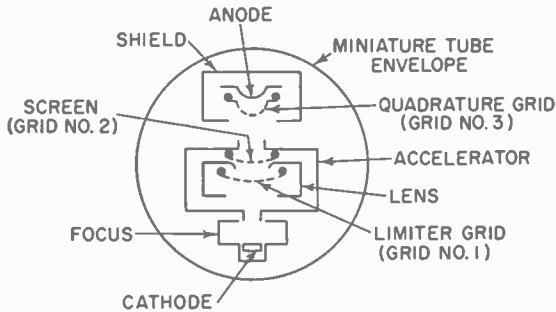


FIG. 9.28. The internal construction of the beam-gated tube.

The reason for this particular behavior of the tube stems from its construction. (See Fig. 9.28.) The focus-electrode, together with the first accelerator slot, forms an electron gun which projects a thin-sheet electron stream upon grid 1. The curved screen grid, together with the grounded lens slot and aided by the slight curvature of grid 1, refocuses the beam and projects it through the second accelerator slot upon the second control grid. This grid and the anode which follows are enclosed in a shield box. Internally, the focus, lens, and shield electrodes are connected to the cathode. The accelerator and the screen grid receive the same positive voltage because both are connected internally.

The foregoing design is such that the electrons approaching the first grid do so head-on. Hence, when grid 1 is at zero potential or slightly positive, all approaching electrons pass through the grid. Making the grid more positive, therefore, cannot increase the plate current further. When, however, grid 1 is made negative, those electrons that are stopped and repelled back toward the cathode do so along the same path followed in their approach to the grid. Because of the narrowness of the electron beam and its path of travel, electrons repelled by the grid form a sufficiently large space charge directly in the path of other approaching electrons, thus causing an immediate cessation of current flow throughout the tube. In conventionally con-

structed tubes, the spread of the electron beam traveling from cathode to grid is so wide that those electrons repelled by the grid return to the cathode without exerting much influence on other electrons which might possess greater energy and therefore be able to overcome the negative grid voltage. It is only when the control grid voltage is made so negative that no emitted electrons possess sufficient energy to overcome it that current through the tube ceases. These differences between tubes may be compared to the difference between the flow of traffic along narrow and along wide roads. Along the narrow road, failure of one car to move ahead can slow down traffic considerably; along the wide road, more room is available and the breakdown of one car has less effect.

The electron beam leaving the second slot of the accelerator approaches grid 3 also in the form of a thin sheet. Thus, this section of the tube may also serve as a gated-beam system. If this second grid is made strongly negative, the plate current of the tube is cut off no matter how positive grid 1 may be. Over a narrow range of potentials in the vicinity of zero, the third grid can control the maximum amount of current flowing through the tube. However, if the third grid is made strongly positive, it also loses control over the plate current, which can never rise beyond a predetermined maximum level.

So much for the operating characteristics of the tube. Now let us see how it can be made to function as a limiter-discriminator. A typical circuit is shown in Fig. 9.29.

It has been noted that when F-M signals reach the discriminator they contain amplitude variations. When the 6BN6 gated-beam tube is used, these signals are applied to control grid 1. If the signal has received sufficient prior amplification, it will have a peak-to-peak value of several volts. Upon application to grid 1, current through the tube will start to flow only during the positive part of the cycle and will remain essentially constant no matter how positive the signal may become or what amplitude variations it may contain. Thus, signal limiting is achieved in this section of the tube, the electron beam being passed during the positive half-periods of the applied signal and cut-off occurring during negative half-periods. The groups of electrons that are passed then travel through the second accelerator slot and form a periodically varying space charge in front of grid 3. By electrostatic induction, currents are made to flow in the grid wires. A resonant circuit is connected between this grid and ground, and a corresponding voltage of approximately 5 volts is developed at grid 3. The phase of this voltage is such that it will lag the input voltage on grid 1 by 90° , assuming that the resonant circuit is tuned to the intermediate frequency. (Because of this 90° difference between grid voltages, grid 3 is often referred to as the quadrature grid.)

The idea of electrostatic induction, while it has not been labeled as such, was encountered in the previous discussion of the 6A8 pentagrid converter. Whenever a group of electrons approach an element in a tube, electrons at that element will be repelled, resulting in a minute flow of current. By the

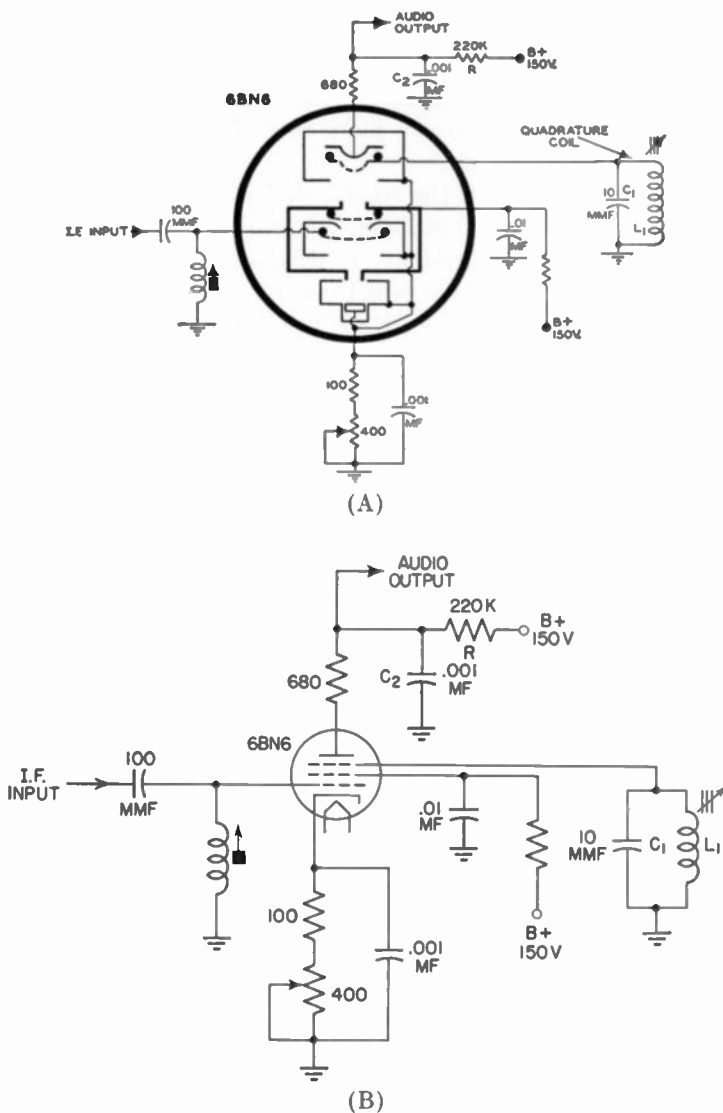


FIG. 9.29. The beam-gated tube connected as a limiter-discriminator. (A) Tube shown in pictorial form. (B) Tube drawn schematically.

same token, electrons receding from an element will permit the displaced electrons to return to their previous positions. Again, a minute flow of current results, this time in a direction opposite to that of the first flow. If sufficient charge periodically approaches and recedes from an element, the induced current can be made substantial. This is precisely what occurs at grid 3 in the 6BN6.

In the gated-beam tube, grids 1 and 3 represent electron gates. When both are open, current passes through the tube. When either one is closed, there is no current flow. In the present instance, the second gate lags behind the first. Plate-current flow starts with the delayed opening of the second gate and ends with the closing of the first gate. Now, when the incoming signal is unmodulated, and L_1C_1 of Fig. 9.29 is resonated at the I.F. frequency, the voltage on grid 3 will lag the voltage on grid 1 by 90° . However, when the incoming signal is varying in frequency, the phase lag between the two grid voltages will likewise vary. This, in turn, varies the length of the period during which plate current can flow. (See Fig. 9.30.)

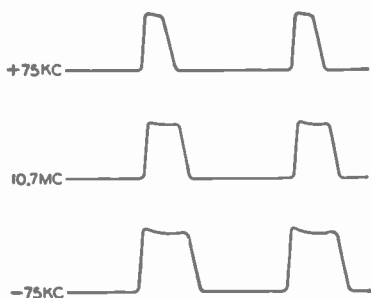


FIG. 9.30. The variation in current pulse duration in the beam-gated tube with frequency modulation.

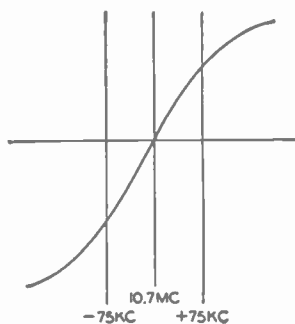


FIG. 9.31. Discriminator response of the beam-gated tube when connected as shown in Fig. 9.29.

Thus, plate current varies with frequency, and the circuit is designed so that the current varies in a linear manner. By placing a resistor in the plate lead, R of Fig. 9.29, we can obtain an audio voltage to feed the audio amplifiers that follow. A typical discriminator response for an F-M receiver with a 10.7-mc center frequency is shown in Fig. 9.31. Note that this curve does not possess any sharp bends (such as Fig. 9.3 does, for example) at frequencies beyond the range of normal signal deviations. This makes the receiver easier to tune.

In the circuit of Fig. 9.29, a 680-ohm resistor is inserted between the load, R_1 and the plate of the tube. By-passing of the I.F. voltage is accom-

plished by C_2 , but since this capacitor is placed beyond the 680-ohm resistor, a small I.F. voltage appears at the anode of the tube. Through the interelectrode capacitance that exists between the anode and grid 3, the I.F. voltage developed across the 680-ohm resistor is coupled into L_1C_1 . The phase relations existing in this circuit are such that this feedback voltage aids in driving the tuned circuit.

Bias for grids 1 and 3 is obtained by placing a resistor in the cathode leg of the tube. Since A-M rejection, especially at low input signals near the limiting level, is a function of the correct cathode bias, the cathode resistor is made variable. This permits adjustments to be made in the field in order to compensate for tube or component changes.

The 6DT6 Detector. The 6BN6 tube is, we have seen, of special construction. Recently, another tube has been similarly employed, although its internal structure is very much like an ordinary pentode. However, the only difference is that the control and suppressor grids are both capable of sharply cutting off the plate current. In this sense they resemble grids 1 and 3 of the 6BN6. The circuit of an F-M detector using a 6DT6 (or a 3DT6) is

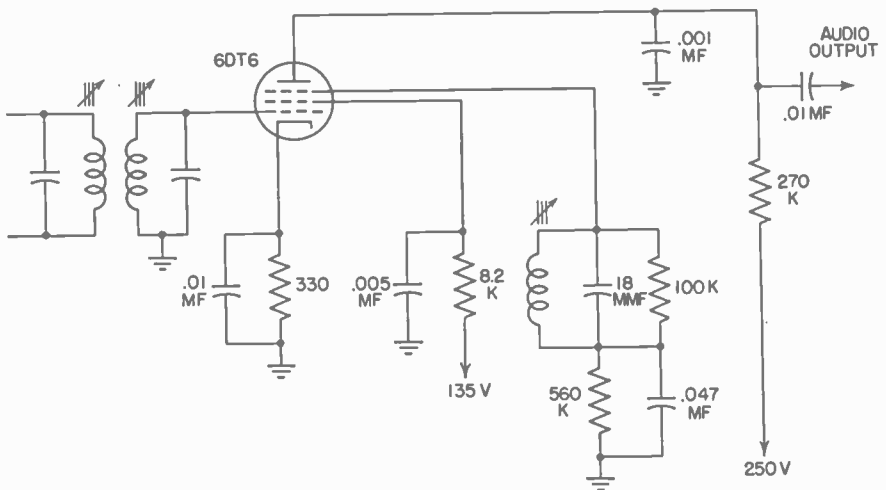


FIG. 9.32. An F-M detector circuit using a 6DT6.

similar to the 6BN6 circuit. (See Fig. 9.32.) So long as the incoming signal is moderate to strong, quadrature-grid detection takes place essentially as it does in the 6BN6 arrangement.

On weak signals, the 6DT6 circuit has a tendency to break into oscillation at the I.F. value. This serves to maintain the detected output signal constant in spite of the fact that weak signals tend to vary considerably in amplitude due to noise or fading. The oscillations arise because of

the feedback which takes place between the suppressor grid and the control grid within the tube. The incoming signal at grid No. 1 locks in with these oscillations and actually causes them to shift in frequency as the modulation moves the signal frequency back and forth. Normal quadrature-grid detection takes place in the oscillating detector. This oscillation boosts the sensitivity of this circuit to weak signals, causing it to deliver a clear output under extremely adverse receiving conditions.

When moderate or strong signals are received, the control grid draws grid current and this loads down the input tuned circuit. This not only kills any tendency to oscillate, but it also broadens the tuning response all of which tends to limit these signals, thereby providing a certain amount of limiter action. In the 6BN6, limiting is achieved within the tube itself.

Tuning in F-M Receivers. Proper reception of an F-M signal is obtained when the signal is perfectly centered in the R.F. and I.F. tuning circuits.

Since satisfactory location of this center point is necessary if the full advantages of F-M are to be enjoyed, it is best to incorporate some form of visual indicator into the circuit. To date, two types of indicators have been employed: tuning tubes and tuning meters. The

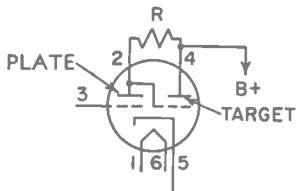


FIG. 9.33. The schematic symbol for a "Magic Eye" tube.

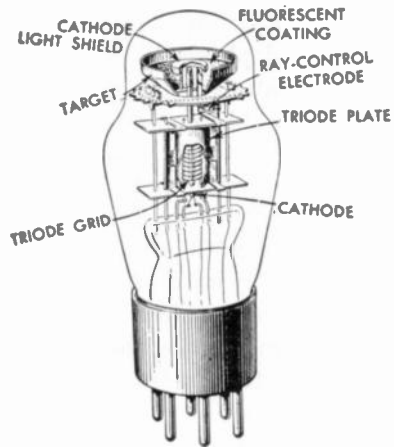


FIG. 9.34. The internal structure of the "Magic Eye" tube.

tuning tube (or tuning eye) is a vacuum tube possessing a small fluorescent screen. From the pattern produced on the screen, it is possible to tell when the signal is maximum and, sometimes, whether the receiver is exactly in the center of the channel. The same indications can be obtained with a tuning meter, this time from the movement of the meter needle.

For a long time, tuning tubes were the most popular form of indicator. However, recently, there has been an increase in the number of F-M receivers using tuning meters. Meters may not be more accurate than tuning tubes (although frequently they are), but their indications are easier for the set

user to interpret. On the other hand, meters are more costly, and the less expensive tuning tube often is still preferred.

The following discussion shows typical applications of both types of indicators.

Tuning Tubes. Indications with a tuning tube, such as the 6E5, are obtained visually by means of a fluorescent target. The 6E5 tube contains two main sections: (1) A triode which functions as a d-c amplifier and (2) a fluorescent screen.

In operation, a positive voltage is applied to both the fluorescent screen, known as the target, and the plate of the triode (see Fig. 9.33). Electrons from a cathode which is common to both sections of the tube bombard the target and cause a fluorescent glow. The breakdown illustration in Fig. 9.34 illustrates clearly the common cathode extending through the tube.

In order to have a sector of the 360° of the fluorescent screen serve as the indicator, a thin rod extends up from the triode plate (see Fig. 9.34) and projects into the region between the upper section of the cathode and the fluorescent screen. The rod is known as the ray-control electrode. Since it is attached physically and electrically to the triode plate, it assumes the same potentials as the plate. It is on this attachment that the entire operation of the tube depends.

Referring to Fig. 9.33, when the control grid is biased almost to cut-off, a small current flows to the plate and subsequently through resistor R . The voltage drop across R is very low, placing both the target and the triode plate at almost the same potential. This means, further, that the ray-control electrode projection of the triode plate has practically the same potential as the fluorescent screen.

Electrons leaving the upper section of the cathode are attracted by both the ray-control electrode and the fluorescent screen. Their path is essentially straight, from the cathode to the screen. Thus, the fluorescent screen emits light at all points except directly opposite the thin rod. The electrons are prevented from reaching the screen by the rod and a shadow is observed. This is shown in Fig. 9.35A.

When the negative voltage on the control grid is increased further, a slight overlapping appears on the fluorescent screen. This is due to the paths taken by the electrons in passing the ray-control electrode on their way to the fluorescent screen. The electrode is nearer the cathode than the fluorescent plate, and it exerts a greater influence over the electrons. Hence, when the electrode reaches the same potential as the screen, which occurs with no plate current in the triode, it attracts more electrons because it is closer to the cathode. Electrons then flow in the manner shown in Fig. 9.35B and produce an overlapped pattern on the fluorescent screen.

For the opposite set of conditions, when the voltage applied to the triode

grid becomes less negative, the current flow through the tube and resistance R increases. This raises the voltage drop across R and places the triode plate at a less positive—or more negative—potential than the fluorescent target. The ray-control electrode also becomes increasingly negative with respect to the fluorescent screen. Electrons leaving the cathode and traveling to the screen will shy away from the ray-control electrode. Thus, very few electrons will impinge on that section of the screen directly behind the ray control. The effect is a dark sector and the “eye” is said to open. The visual effect is indicated in Fig. 9.35C.

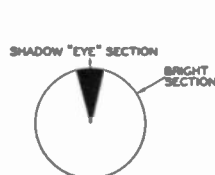


FIG. 9.35A. The width of the eye when the control grid of the tube is quite negative.

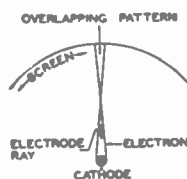


FIG. 9.35B. The electronic paths in the “Magic Eye” tube when an overlapped pattern is produced.



FIG. 9.35C. The eye opening when the grid becomes more positive.

In amplitude-modulated receivers, the A.V.C. line is attached directly to the grid and the station tuned in by observing the width of the dark sector on the screen—or, as it is commonly referred to, the shadow. The negative A.V.C. voltage reaches a maximum when the station is correctly tuned in. This is indicated at the electron-ray tube when the width of the shadow is narrowest. On very strong signals the two light ends of the dark sector not only approach each other but actually overlap.

Connecting the Tuning Eye in the F-M Set. In an F-M receiver, there are two points where the tuning indicator may be connected. One point is at the limiter grid circuit; the other is at the F-M detector.

A simple arrangement is to connect the tuning eye into the grid circuit of the limiter. When two limiters are employed, the connection is made in the grid circuit of the first limiter. Here, the negative voltage developed across the grid-leak resistor is directly proportional to the strength of the incoming signal.

A typical receiver circuit is shown in Fig. 9.36. The voltage across the 350,000 ohm resistor is applied to the grid of a 6E5 tuning-eye tube. Since the voltage at point A is negative with respect to ground, it is of a suitable polarity to be fed to the grid of the 6E5. The stronger the signal, the more negative the voltage at point A and the narrower the eye shadow. In tuning, the listener adjusts the dial until the eye shadow closes as far as it will go.

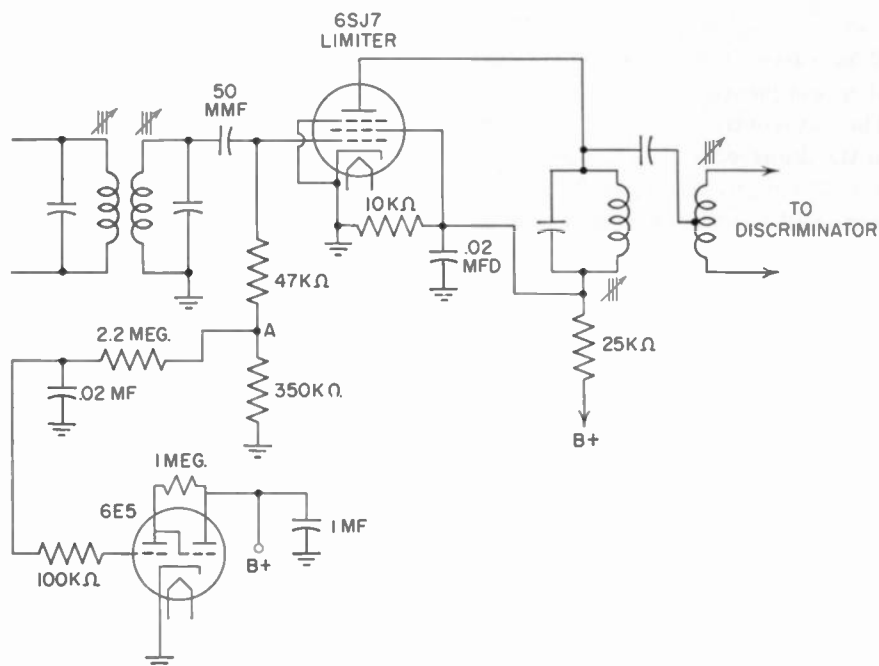


FIG. 9.36. The circuit for connecting a 6E5 tube to the grid of the limiter stage.

The 2.2-megohm and 100,000-ohm resistors, together with the 0.02-mf capacitor, form a decoupling unit to prevent the alternating components of the input signal from reaching the grid of the 6E5. The edges of the two sections of the fluorescent target become hazy when this occurs.

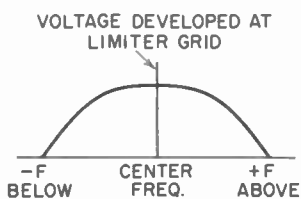


FIG. 9.37. The variation of limiter grid voltage with strong signals.

The disadvantage of the simple arrangement of obtaining the indicator voltage from the limiter grid circuit is due to the fact that, with strong signals, the voltage developed at the grid of the limiter does not permit exact positioning. Fig. 9.37 shows that there is a region about the resonance point where the voltage across the grid-leak resistor of the limiter does not change appreciably. Hence, it becomes possible to tune through a wide range and not obtain a noticeable change in the aperture of the tuning eye. In addition, despite the indications at the grid of the limiter, we still have no way of determining how the discriminator is functioning. Finally, there are receivers which use only partial limiter stages and some other means must be devised if tuning devices are to be used with these receivers.

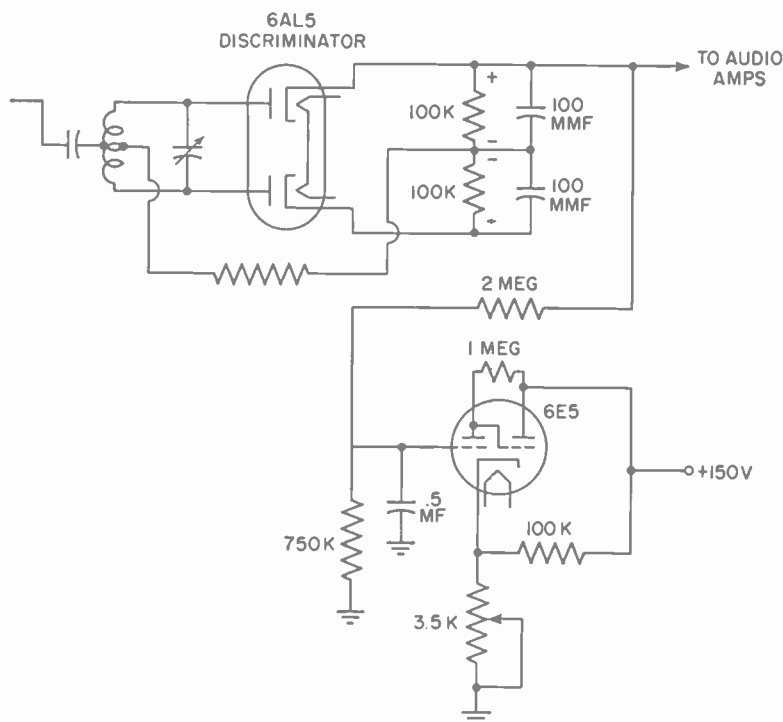


FIG. 9.38. A 6E5 tube connected to the discriminator output.

One method, found in some sets, employs a 6E5 tuning tube that obtains its input from the discriminator output. This is shown in Fig. 9.38. The 6E5 is adjusted to cut-off by variation of the 3500-ohm variable resistor in its cathode circuit. The arm of the 3500-ohm resistor is rotated until the shadow on the fluorescent screen of the 6E5 just closes when no voltage is applied to the grid of the tube. When the set is detuned in one direction, the output from the discriminator is, say, negative and the shadow overlaps. Detuning in the opposite direction produces a positive voltage and the shadow becomes wider. When the receiver is correctly set, the average voltage from the discriminator is zero and the shadows just close. This is the correct tuning point. The 2-megohm and 750,000-ohm resistors, together with the 0.5-mf capacitor, form a filter that eliminates the audio variations from the discriminator output to the 6E5 and present only an average voltage. The regular audio voltage is still taken from its usual point on the discriminator for the audio stages.

In F-M receivers employing ratio detectors, a limiter may not be used. In this case, a suitable indicating voltage for the tuning eye may be obtained from one of two points in the ratio detector circuit itself. One point, marked

AVC in Figs. 9.24, 9.25, and 9.26, will provide a negative voltage that reaches a negative peak when the station is correctly tuned in. The tuning indicator tube would be connected as shown in Fig. 9.36, except that the end of the 2.2-megohm resistor would go to the AVC point in the ratio detector instead of Point A.

If the tuning tube is to indicate the center-of-channel, as in Fig. 9.38, it

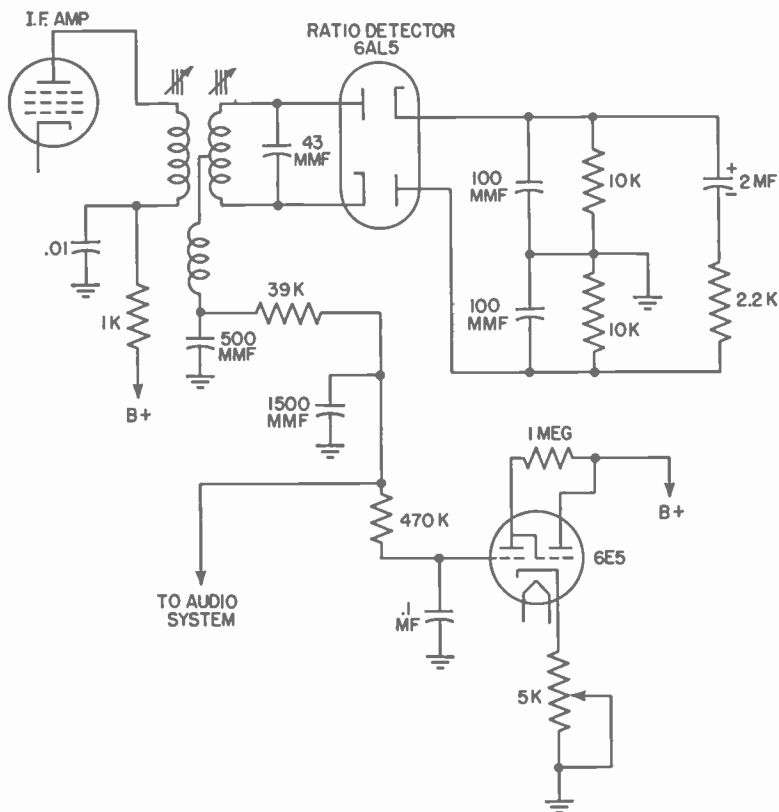


FIG. 9.39. A 6E5 tube connected to the output of a ratio detector.

would be connected to that point in the ratio detector where the audio output is obtained. A typical circuit is shown in Fig. 9.39. Operation of this circuit is identical with the tuning eye circuit in Fig. 9.38.

Other tuning tubes, similar in operation to the 6E5 but possessing differently shaped fluorescent areas, are also available. Thus, the indicating pattern of the EM84/6FG6 is a bar whose length varies with signal strength, becoming longer as the negative voltage on the grid increases. (See Fig.

9.40A.) In the DM70, Fig. 9.40C, the pattern is in the form of an exclamation point. Only the top segment of this pattern varies with signal, the length of the illuminated area decreasing with negative grid voltage. Still another arrangement appears in Fig. 9.40B, this being used in the EM80 tube. In spite of the differences in indication, all of these tubes would be employed in essentially the same manner.

An entirely different type of indicator tube is the 6AF6-G. This contains only an indicator section with two ray electrodes mounted on opposite sides of the cathode and connected to individual base pins. These provide two separate shadow angles in place of the one in the 6E5. Two symmetrically opposite shadow angles are obtained when the two ray-control electrodes are connected together; two unlike patterns are obtained when each ray-control electrode is separately connected to a different part of the circuit. In applying the 6AF6-G, it is necessary that each ray electrode receive a fairly good positive voltage, on the order of 60 volts or more. This means that the tube must be tied in to a section of the circuit capable of providing this voltage, as well as a fairly large signal swing. One application of the 6AF6-G is shown in Fig. 9.41. The

ray electrode at terminal 4 connects to the voltage dropping resistor in the plate circuit of the A-M, I.F. amplifier, while the ray electrode at pin No. 3 ties in to a similar point in the F-M limiter circuit. From this dual connection, it can be seen that each shadow angle is governed by a different signal, one by A-M, the other by F-M; therefore, only one of these will be activated when the set is tuned to a signal.

Operation of the circuit is fairly simple. When a strong signal is received in the A-M circuit, the AVC bias which is fed to the grid of the I.F. amplifier will be most negative. This will cause the plate current to be at its lowest value and the voltage at the top of resistor R_1 will be most positive. When the ray electrode is most positive, the shadow is at a minimum. This, then, is the indication of resonance. A similar action occurs in the F-M circuit, except that the limiter tube develops the greatest negative grid bias (by grid-leak action) when a signal is tuned in. This decreases the

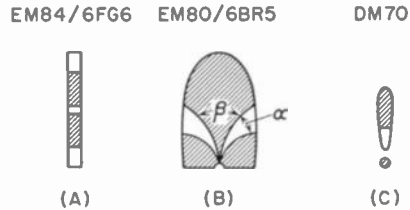


FIG. 9.40. Pattern indications of three tuning tubes. (A) Shaded rectangles expand and contract lengthwise as negative grid voltage varies. (B) Large shaded area on top and the two smaller shaded areas below vary in size as the grid voltage changes. Angle β becomes smaller as grid voltage becomes less negative. This causes angle α to increase. (C) Shaded area in elongated top segment varies in length as grid voltage is changed. In each tube, shaded areas (representing fluorescent screens) become maximum when grid voltage is most negative.

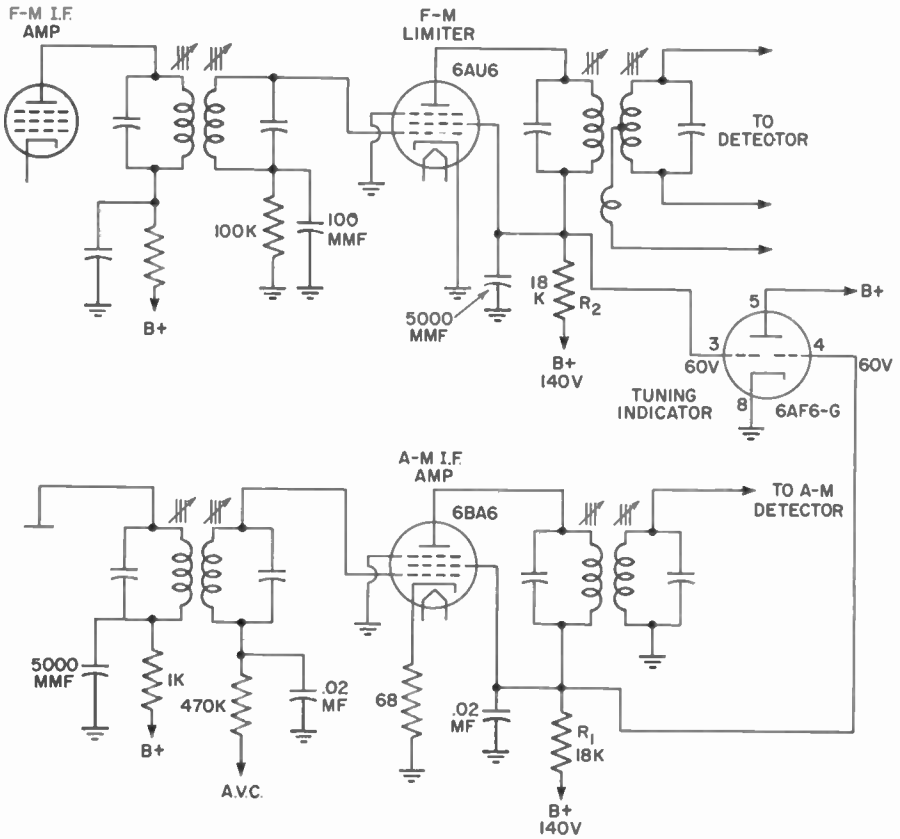


FIG. 9.41. Use of the 6AF6-G tuning indicator in an A-M, F-M combination receiver.

current flowing through the tube and raises the positive voltage at the top of the resistor R₂.

The 6AF6-G tube is, in part, equivalent to that portion of the 6E5 which contains the ray-control electrode. Missing from the 6AF6-G tube is the triode amplifier which is part of the 6E5. As a matter of fact, it will be found that the 6AF6-G is sometimes preceded by two triode amplifiers, one for each ray-control electrode. When this is done, negative grid voltages are required by the amplifiers for tuning indication. This, of course, is similar to the action of the 6E5.

The Electron-Ray Indicator. Another indicator, developed by General Electric, is the 6AL7-GT. This is an electron-ray tube containing a triode and three deflection plates and operating on a principle similar to the "Magic-Eye" tubes. In addition, the tube contains a control grid, located

in the triode section of the tube, which can be biased to regulate the brightness of the fluorescent target so that the electron beam can be prevented from reaching the screen when the set is off-station and permitted to reach the screen when the set is tuned to a station. Note that this differs from the action of the control grid in the "Magic-Eye" tubes.

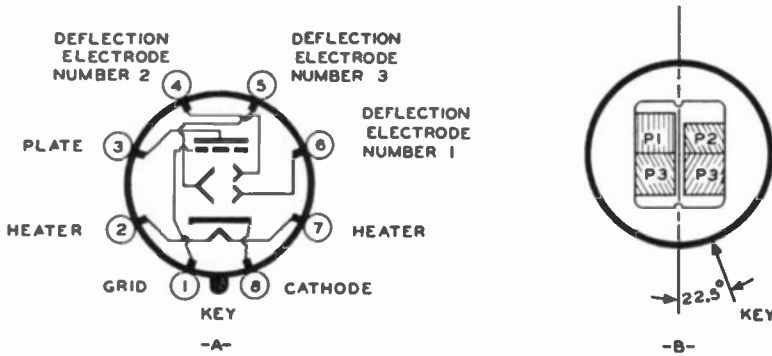


FIG. 9.42. (A) Schematic symbol for the 6AL7-GT tuning indicator. (B) The fluorescent screen used in this tube.

The fluorescent screen of the tube is rectangular-shaped and divided into quarters. (See Fig. 9.42.) Of the four quarters, however, the top two (shown as P1 and P2 in Fig. 9.42) connect to separate deflection electrodes or plates while the bottom two quarters connect to the same electrode. This arrangement means that, as far as independent variation is concerned, the top two quarters can be varied in size independently of each other whereas the bottom two quarters, being part of the same electrode, will vary in step with each other. The amount of negative bias applied to each electrode controls the size of that portion of the fluorescent screen. The amount of bias on the control grid varies the brightness (but not the size) of all four sections of the fluorescent screen.

The chart and circuits in Fig. 9.43 illustrate how the 6AL7-GT indicator can be connected into a circuit and the resulting variation of the fluorescent screen area for off-tune and on-tune conditions. The following explanation will indicate the operation of the tube in each of the circuits shown.

Circuits A and B of Fig. 9.43 show how the tube may be connected so that it operates from the discriminator voltage alone. In each of these two circuits deflection electrodes 1 and 3 are connected to ground, thereby giving them a fixed potential. Hence, none of the fluorescent areas controlled by these electrodes will vary in size, whether the set is tuned to a station or not. This means that P1 and the left-hand portion of P3 will act as a combined reference area. On the other half of the target, a variable P2 is combined

PATTERN SEQUENCE DURING TUNING

CONTROL VOLTAGE SOURCE	SIGNAL	CIRCUIT (SEE BELOW)	OFF CHANNEL (-)	ON CHANNEL OFF TUNE (-)	ON TUNE	ON CHANNEL OFF TUNE (+)	OFF CHANNEL (+)
DISCRIMINATOR	FM	A & B					
DISCRIMINATOR AND SQUELCH	FM	C					
DISCRIMINATOR AND LIMITER	FM	D					
AVC	AM	E					

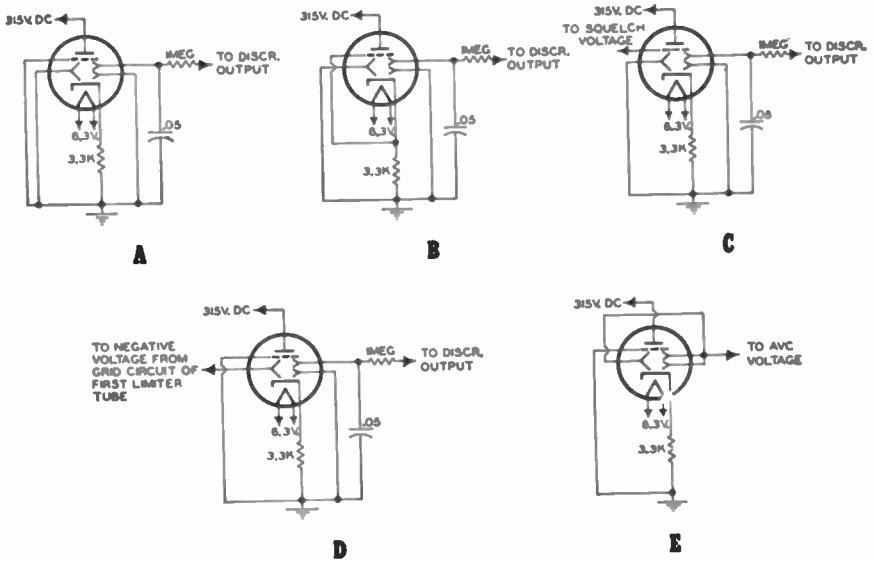


FIG. 9.43. Various applications of the 6AL7-GT tube and the patterns appearing on its fluorescent target area. (Courtesy G.E.)

with a fixed P3. P2 is connected to the discriminator and its size will vary according to the voltage it receives (either negative, zero, or positive) from this circuit.

When the station is completely off channel, the discriminator voltage is zero, placing P2 at the same potential as P1 and P3. Hence, both halves of

the fluorescent screen will be equal in size. This is true whether the set is off-channel above the desired station or below. Furthermore, it is also true when the set is tuned precisely on frequency. This similarity of indications may appear to be confusing, but actually it is not since music or speech will be heard when the set is correctly tuned to a station and only noise will be obtained for both off-channel positions.

When the set is on channel but off-tune, the average output voltage from the discriminator will be negative on one side of the station and positive on the other side. When the voltage is negative, P2 shrinks in size; when the voltage is positive, P2 becomes elongated.*

The different connections of the control grid in circuits A and B merely affect the brightness of the total fluorescent area.

In circuit C, we obtain essentially the same tube indications with the exception that a squelch voltage is applied to the control grid, blanking out the fluorescent screen when the receiver is completely off-channel.

In circuit D, P3 receives a negative voltage from the grid circuit of the first limiter tube (if two are used). When the set is completely off-channel, the voltages from the limiter and discriminator are zero, and all electrodes possess the same potential. Hence, the left- and right-hand sides of the target area become equal in size. When the set is on tune, the discriminator voltage is zero, and areas P1 and P2 are equal. At this point, however, the negative voltage from the limiter is quite high, and both quarters of P3 shrink down to practically nothing. On either side of the correct tuning point, the negative grid voltage from the limiter decreases, and the area of P3 increases. However, P2 either will become smaller or larger in size, depending upon which side of the station the set is tuned to.

In circuit E, all deflection electrodes are tied together and all receive the same A.V.C. voltage. Hence, both sides of the target area will vary in step. This same arrangement, with the three electrodes connecting to the grid of the first limiter (if two are used), can serve also as a signal strength meter for F-M receivers. If a limiter stage is not used, a practice common with ratio detectors, then the negative voltage may be obtained from the point marked "AVC" in Fig. 9.25 or Fig. 9.27A.

Note, however, that, when the 6AL7-GT is employed in this fashion in F-M receivers, its indications are not as useful as they are when employed for center-of-channel location.

Tuning Meters. A tuning meter is generally formed using a 1 ma D'Arsonval type movement. The meter is driven by a triode tube which

* In a Foster-Secley detector, the connection would be made to a point such as A in Fig. 9.13. In a balanced ratio detector, the connection would be made to a point such as P, Fig. 9.25. And in an unbalanced ratio detector, the 6AL7-GT would be attached to point C, Fig. 9.27B.

serves as an amplifier between the meter and the point where the indicating voltage is obtained. The choice of the latter depends on the manner in which the meter is to function. For example, if it is to serve simply to indicate signal strength, then the unit would be connected as shown in Fig. 9.44. The meter is inserted between R_3 and a network of two resistors, R_4 and R_5 . With no voltage applied to the control grid, R_4 is adjusted until the B+ voltage it applies to the meter equals the positive voltage developed across R_3 . With no voltage difference between the two meter terminals, no current flows through the meter and it indicates zero.

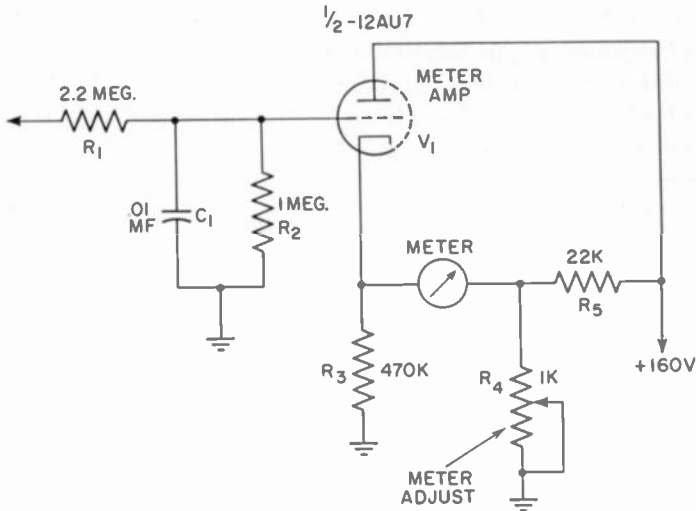


FIG. 9.44. A meter indicating network employed in A-M, F-M receivers.

The control grid of V_1 is now connected to a point in the circuit where the strength of the incoming signal is revealed by a negative voltage. In an A-M receiver, this would be the A.V.C. line. In an F-M receiver, this could be the grid of a limiter, or the point marked "A.V.C." in Fig. 9.25 or Fig. 9.27A. In any event, when the signal is correctly tuned in, the negative voltage applied to the grid of V_1 will be highest. This will reduce the current through V_1 , decreasing the positive voltage developed across R_3 . If the meter is properly connected, the needle will deflect up-scale.

The same arrangement, using a zero-center meter, will reveal when the center-of-channel of an F-M signal is reached. Now R_4 is adjusted, with no voltage on the control grid, until the meter needle is at its center zero mark. The control grid of V_1 must now be connected to the F-M detector where a zero d-c voltage is developed when the tuner is on-channel, a negative d-c voltage when it is tuned off-channel in one direction and a positive d-c volt-

age when it is tuned off-channel in the other direction. Foster-Seeley and ratio detectors can provide such a voltage.

C_1 and R_1 in Fig. 9.44 function as a d-c filter, removing all signal components from the voltage reaching the grid of the meter amplifier, V_1 .

Interstation Noise Suppression Systems. As long as a station is tuned in, no difficulty is encountered from background noise. However, as soon as the station is tuned out, the input voltage consists of random noise, too weak to operate the limiter at saturation or beyond. Under these conditions the limiter functions as an amplifier and the amplitude-modulated random voltages produce the familiar loud hiss at the speaker. This is the interstation hubbub, a distracting, hissing, sizzling noise. To eliminate this distraction, many manufacturers have included interstation noise suppression systems—generally some simple arrangement that automatically cuts off one or more audio amplifiers when the input signal decreases below a certain level.

The basic idea behind most of these systems involves the use of a tube that is maintained at cut-off as long as sufficient signal voltage is being received. As soon as the input signal decreases, because of a change of dial setting, the flow of electrons is resumed through the tube. This current is made to flow through the first audio amplifier cathode resistor and biases the tube to cut-off. This condition is maintained until the signal strength of the input of the receiver again increases.

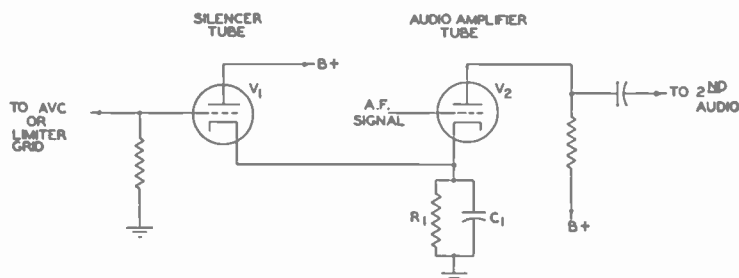


FIG. 9.45. A simple silencer arrangement.

A simple circuit illustration of this principle is shown in Fig. 9.45. V_1 is the silencer, or squelch tube. V_2 is the first audio amplifier. Both tubes have a common cathode resistor and capacitor, R_1 and C_1 . As long as V_1 is maintained at cut-off by a negative voltage on its grid, V_2 functions normally. Removal of the negative grid voltage at V_1 , or at least a decrease to permit an appreciable current to flow, quickly brings V_2 to cut-off because of the excessive voltage appearing across R_1 . This holding action continues until V_1 is again cut off by the application of a negative voltage. Tubes capable of high current flow are chosen for the squelch tube, as this permits decisive

control with relatively small negative grid voltages. High- μ triodes are commonly used. Quite possible, too, is any one of a number of pentodes, although, in their case, provision must be made for screen voltage.

In an F-M receiver, there are two points where the required negative biasing voltage to control the squelch tube may be obtained. One such point is in the grid circuit of the first limiter stage; the other is at the output of the discriminator. In the circuit shown in Fig. 9.46, the grid of the silencer

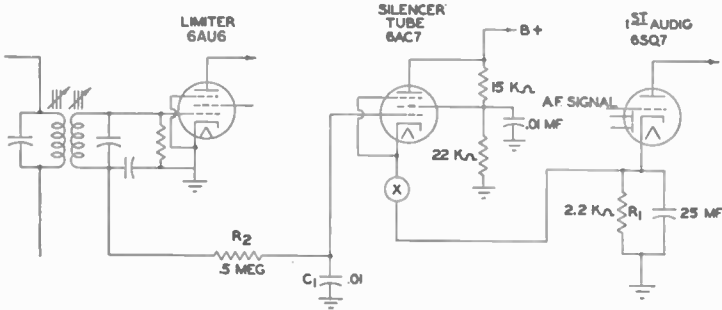


FIG. 9.46. Another possible silencer circuit.

tube is directly connected into the grid circuit of the 6AU6 limiter tube. The negative voltage developed here is sufficient, with signal reception, to keep the 6AC7 at or beyond cut-off. During these periods, the triode section of the 6SQ7 operates normally as the first audio amplifier. As soon as the negative voltage is removed from the grid of the 6AC7, a high current flows through R_1 and biases the 6SQ7 to cut-off. The 6AC7 has a mutual conductance of 9000 micromhos and is capable of large currents with relatively small negative voltage changes on its grid. R_2 and C_1 form a filter to prevent I.F. currents from reaching the 6AC7. Only the average rectified, or d-c, voltage present on the limiter grid is desired. A switch is provided in the cathode circuit of the 6AC7 to permit the listener to cut out the silencing action if desired. This cutting-out action, for example, might be needed on weak signals when the presence of the silencer could prevent reception.

In Fig. 9.47, the d-c voltage available at the discriminator is utilized for control. If we examine the individual polarity across each of the 100,000-ohm discriminator resistors, we see that point *A* will always be negative with respect to point *B* as long as a signal is being received. Note that we are utilizing only the voltage appearing across the lower resistor and not across both resistors as is common for the audio output. The voltage obtained from point *A* consists of an audio component due to the shifting of the frequency of the signal and a d-c component arising from the amplitude of the incoming carrier. As long as the signal possesses sufficient strength to oper-

ate the limiters at saturation, the negative voltage present at point *A* will be strong enough to keep the silencer tube, a 6J5, at cut-off. This will permit the 6SF5, first audio-amplifier, to function normally. However, when the signal is tuned out, the voltage at point *A* drops, the 6J5 begins to conduct, and the 6SF5 is prevented from operating. R_1 is the resistor common to both triodes. A filter is inserted between point *A* and the grid of the 6J5 to eliminate the audio component of the voltage at point *A* and present only a d-c voltage to the silencer.

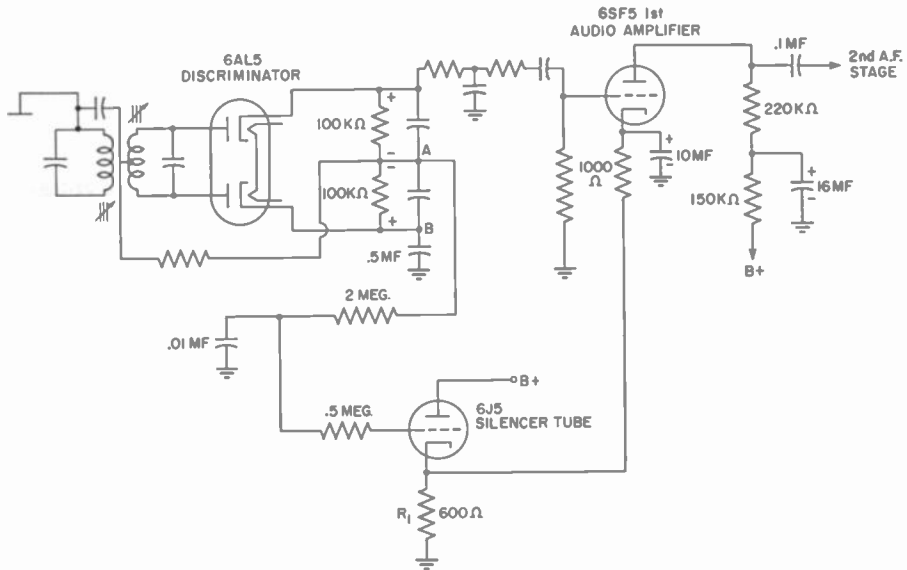


Fig. 9.47. A silencer functioning on the voltage obtained from a discriminator.

Automatic Volume Control (A.V.C.). Automatic volume control is a common feature in the standard broadcast sound receiver. It helps maintain a fairly constant level of output volume and largely counteracts normal changes in signal intensity due to external sources. Within the receiver, A.V.C. also helps to prevent over-loading of the I.F. stages. In F-M receivers, the need for A.V.C. is not quite so urgent. As long as the signal is sufficiently strong, it will operate the limiters at saturation, and any change in signal strength beyond this point will have no effect on the output audio fidelity or strength. Also, the I.F. stages preceding the limiter are operated at maximum gain in order to insure saturation at the limiter of all desired signals. Any signal stronger than the average will suffer only amplitude limitation before it reaches the limiter, but this will not affect its frequency modulation. Since the I.F. amplifiers are operated at maximum gain, signals that are too weak to saturate the limiter are being received below the rated

sensitivity of the receiver, and the inclusion of A.V.C. would not prove helpful. As long as the amplitude of the signal being received is of no consequence beyond a certain point, there is little need for A.V.C.

In the ratio detector, sensitivity to amplitude modulation is low if the circuit is aligned properly. In addition, many circuit designers incorporate partial or full limiting ahead of the detector as a precautionary measure. Sometimes, because the ratio detector does possess a source of A.V.C. voltage, it is used. In detectors using 6BN6 or 6DT6 tubes, limiting occurs within the tube and the resulting sensitivity to A-M is low enough to provide satisfactory reception on all but very weak signals.

PROBLEMS

1. Contrast the purpose of an A-M detector with that of an F-M detector. State clearly why each should differ.
2. Draw the schematic diagram of an early type of discriminator which employed two secondary windings.
3. Explain the operation of the circuit drawn for Question 2.
4. Would the foregoing discriminator function if one of the diodes became inoperative? Give the reasons for your answer.
5. Draw the response curve for this early type of discriminator. What would be the effect on the curve if the two tuned secondary circuits were brought closer in frequency? How would this affect the reception of a fully-modulated F-M signal?
6. Draw the circuit of a modified discriminator (Foster-Seeley type) widely used today.
7. Describe the operation of the modified discriminator circuit briefly.
8. Indicate the exact function of each component in the discriminator.
9. Are there any modifications that can be made in the circuit? Explain.
10. What limitations of the Foster-Seeley type of discriminator resulted in the development of a ratio detector?
11. Draw the circuit of a ratio detector.
12. Explain briefly the operation of a ratio detector.
13. Are special tuning devices necessary in F-M receivers for proper reception of signals? Explain.
14. What types of tuning devices are used? How do they function?
15. Draw the circuit of a simple tuning device connected to the limiter.
16. Explain how the circuit drawn in Question 15 operates.
17. What limitations does this method of connecting a tuning device possess?
18. Draw the schematic diagram of a tuning device connected to the discriminator output.
19. Explain the operation of the circuit drawn for Question 18.
20. Why are interstation noise suppression systems used? What is the basic idea behind most of these systems?
21. Draw the schematic diagram of a simple silencer circuit and explain its operation.
22. What type of tube would be most desirable in silencer networks? Why?
23. Does the silencer network function when a loud burst of interference is being received? Explain your answer.
24. Explain briefly the operation of a beam-gated limiter detector circuit.

Chapter 10

AUDIO AMPLIFIERS AND HIGH FIDELITY

AUDIO AMPLIFIERS

High Fidelity. To many people, both in and out of the radio profession, the great advantage of F-M lies in its ability to reproduce the full range of audio frequencies. For sales appeal, the term "high fidelity" was popularized and everything desirable was associated with it. And yet, like everything else that has been oversimplified, unwarranted assumptions have been made. High fidelity represents not one but a host of ideas, and what may appear to be high fidelity to one person may not be so to another. To a large extent, high fidelity involves the listener, and thus any complete consideration of the meaning of this phrase must not only include the technical details of the electrical system but the relationship of the listener himself to the system. It is because of this latter aspect that we find any one precise definition of high fidelity impossible. Nonetheless, we may still investigate the contributing factors which add up to provide good reproduction of a broadcast. Briefly, these factors include the electrical system (receiver and transmitter), the loudspeaker, the surrounding acoustics of the room where the reproduction occurs, and, finally, the aural capabilities of the listener.

In considering the listener, we find that the human ear has a frequency response characteristic that varies not only with each individual but also with age and even slightly with the length of time that the ear is subject to the music or speech of any one continuous listening. As an example of the average frequency characteristics of the ears, consider Fig. 10.1. Here we have the different intensities which each frequency must possess in order to sound the same to our ears as the 1000-cycle note. An example will perhaps make this clearer. Dealing with the bottom curve, we note that a 50-cycle note must be 50 db stronger than a 1000-cycle signal in order for both sounds to *seem* equally intense to us. Each curve represents the response at different sound levels. Thus, the bottom curve is for very soft sounds, almost at the

threshold of audibility. The top curve represents situations where the sound is so loud or intense that hearing involves pain. The other curves are for intermediate degrees of intensity. In each curve, the 1000-cycle frequency is taken as the level of reference. From these curves, several important general conclusions may be drawn.

1. The best response occurs around 3000 to 4000 cycles.
2. The low frequency response is poorest at low intensity levels. Thus, when the volume is turned down low the low frequency notes suffer the most. Many manufacturers incorporate compensation for this in their tone controls.
3. The overall ear response is flatter as the volume increases. Note that the curves near the top of Fig. 10.1 tend to be fairly flat.

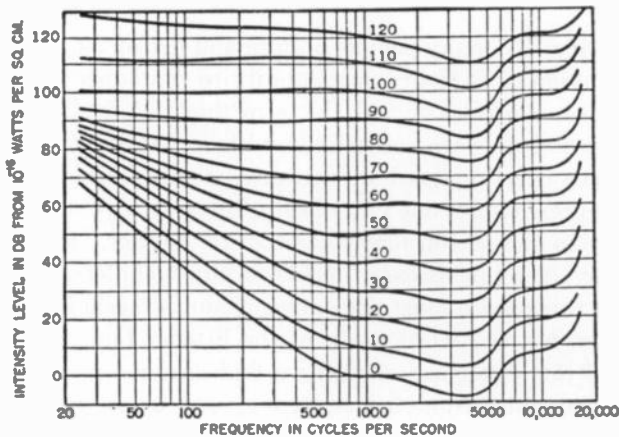


FIG. 10.1. Frequency and loudness characteristic curves of the ears, as obtained by Fletcher-Munson.

Although not indicated in the diagram, individual hearing ability decreases with age. As one grows older, the ability to hear any of the audio frequencies deteriorates, with the higher frequencies subject to the greatest loss.

To consider only one of the foregoing conclusions, provision must be made in the radio receiver to compensate for the change in hearing response at different volume levels. As mentioned, most manufacturers incorporate low frequency compensation on volume controls at the lower level. Beyond this, however, very little has been done and the listener is obliged to resort to his tone control for further adjustments.

The dependence between volume and the frequency response is indicated

more clearly in Fig. 10.2. Two curves are shown, one for the average listener and one for the more critical listener. The data for the average-listener curve were obtained by using people of all ages and both sexes. We see that, when the sounds are very loud, the frequency range for the average person extends from 20 to 15,000 cycles. As this volume is decreased, both ends of the curve drop and the frequency response becomes narrower. As before, the ear is most sensitive to frequencies between 3000 and 4000 cycles.

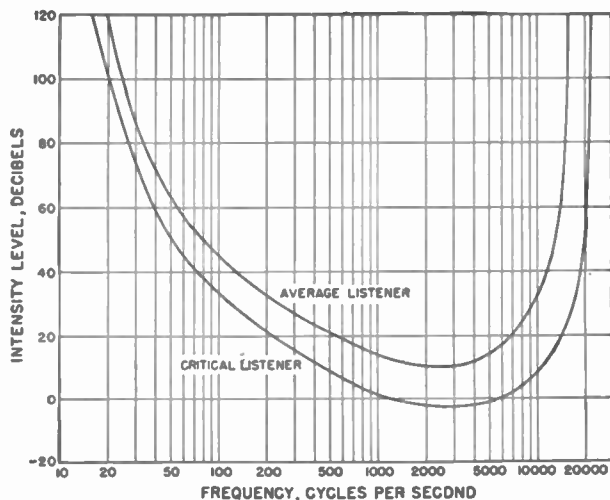


FIG. 10.2. A comparison between the hearing ability of average and critical listeners.

In the discussion of these curves, no mention was made of noise, which is present to a certain degree in all homes. The surrounding, or (as it is more technically known) ambient, noise will be most noticeable when the volume is turned down low. It will tend, also, to mask the low and high frequencies because, as we have seen from the charts, these require greater intensity to produce the same aural sensation. It has been discovered that the average residential noise level is 43 db. In many homes, with children romping about, the noise is considerably higher and the ability of a person to hear is that much more impaired.

The Loudspeaker. As a converter of electrical currents into acoustic sounds, the speaker is as much an integral part of this system as the listener. In midget sets the small diameter cones are capable of a restricted range with best response at the frequencies between 4000 and 5000 cycles. Since it requires a large cone area to reproduce effectively the low frequencies, high fidelity reproduction from these small sets is impossible. In console receivers a popular solution to the problem of efficient reproduction is through the use

F-M SIMPLIFIED

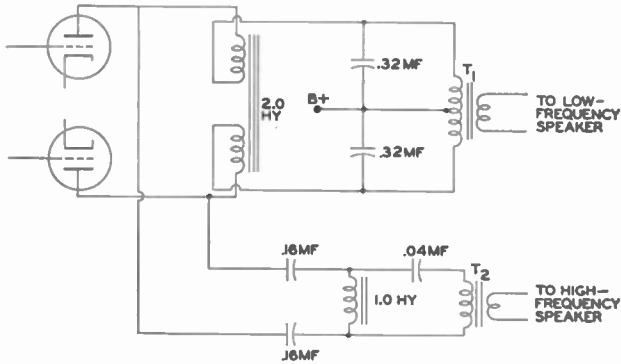


FIG. 10.3. A frequency-dividing network designed to feed two speakers.

of two speakers. A frequency-dividing network, shown in Fig. 10.3, separates the frequencies as they come from the output transformer and thus permits each speaker to respond only to frequencies within its designed range. Another approach is with a coaxial speaker, where one small cone is positioned at the center of a much larger cone (see Fig. 10.4). Again, frequency-dividing networks provide each speaker with its proper band of frequencies. The name "coaxial" is derived from the similarity of this arrangement to the coaxial cable where one conductor is located at the center of an outer hollow conductor.

Balance. It is common belief that everything less than the full frequency



FIG. 10.4. A coaxial type of speaker. (Courtesy of Jensen.)

coverage of 30 to 15,000 cycles leaves us short of our goal of high fidelity. Extensive tests have indicated, however, that of greater importance than trying to establish the maximum range is the balance achieved between the highs and the lows. By balance, we refer to the low and high frequency limits of our system. If we take a system and extend the high frequency response without providing a corresponding extension at the low frequencies, we find that the results do not tend to be as pleasing despite the fact that, from a true fidelity viewpoint, we are now in a position to receive more of the audible frequency range. The aural effect of extending the high frequency end, for example, without similar compensation at the low frequency end is to give the speaker response a shrill tone. Small radios possess speakers that accentuate the higher frequencies. This requires frequent use of the bass portion of the tone control to produce a compensating mellow effect.

The concept of balance has not been reduced to a definite mathematical formula, but it has been suggested that a good, empirical test is to have the product of the high and low frequency limits of the band fall approximately between 500,000 and 600,000. Thus, as an illustration, a band that starts at 90 cycles at the low frequency end should extend up to 6000 cycles at the high end in order to preserve the balance between the upper and lower limits. The chart in Fig. 10.5 shows how the limits vary from a small table

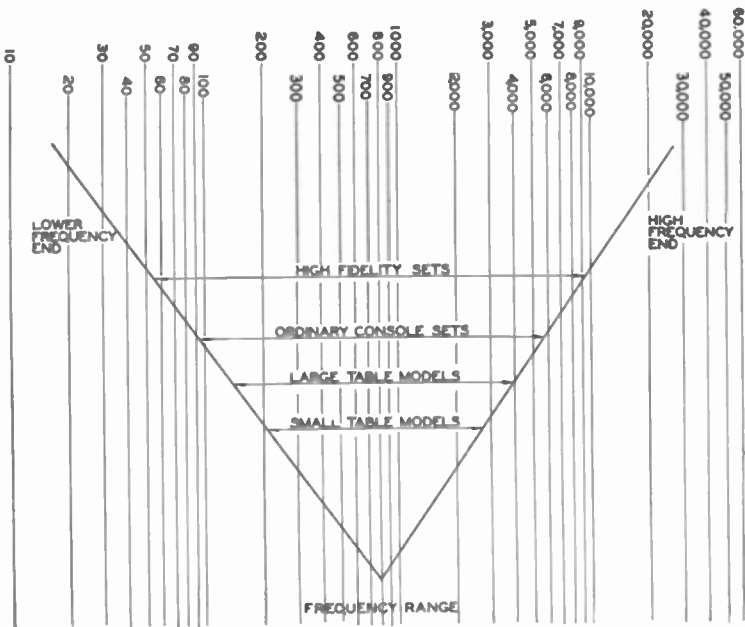


FIG. 10.5. Desirable frequency ranges for various types of receivers in order to obtain the proper tonal balance.

model to a large console radio with a proper balance maintained in all instances.

Bandwidth Limits. It has been assumed throughout the foregoing discussion that the receiver networks are fully capable of passing all the signal sidebands. Although this may readily be accomplished if the manufacturer takes the proper precautions in design, the simple fact remains that this is not so in many instances. The proper bandwidth response must be maintained in the R.F. and I.F. transformers, in the audio amplifiers, the output coupling transformer and, finally, in the loudspeaker itself. In the tuned coupling transformers of the R.F. and I.F. stages, the usual arrangements result in the familiar, rounded response. With this response, the lower audio frequencies which are closest to the center of the carrier (hence closest to the center of the curve) receive the greatest amplification. The higher frequencies, situated farther away, receive correspondingly less amplification. Hence, even if the signal is properly balanced at the set input, by the time it reaches the second detector the balance is distorted because of unequal amplification accorded the various frequencies. These are the present conditions in A-M receivers. With an F-M receiver, unequal response in the R.F. and I.F. circuits does not greatly affect the signal as long as its strength is sufficient at the limiter to cause saturation. However, in an F-M receiver, the bandpass of each tuning circuit must be wide enough to permit reception of the full 200 kc, and the discriminator must be capable of converting F-M to A-M linearly. Careful design and construction are needed to balance all these factors—a fact that frequently varies in direct proportion to the price of the receiver. The latter statement is not meant to be an indictment of present practice but merely a recognition of some of the commercial considerations that often waylay high fidelity.

We have touched briefly on the subject of high fidelity to indicate some of the highly personal factors that combine to form this concept. It is because individual tastes differ that we find an extensive use of tone controls in most F-M receivers. Some of the common methods of achieving this tone variation will be taken up in a later section.

Phase Inverters and Push-Pull Amplifiers. In order to be able to produce the full range of audio frequencies at full volume without undue overloading, push-pull output amplifiers are required. Advantages of push-pull amplifiers are (1) the cancellation within the stage of second harmonic distortion, (2) greater permissible drive at the input producing a stronger output with less odd-harmonic distortion, and (3) smaller output transformers and less supply voltage filtering. The last advantage reduces hum and permits a greater dynamic range from the loudspeaker. At low volumes, hum can be very disturbing, and the current practice of accentuating the bass response of a receiver by special cabinet construction tends to ac-

centuate hum further. In a balanced push-pull amplifier, a-c ripple from the power supply is eliminated in the plate transformer.

If we use a push-pull stage, such as the typical unit shown in Fig. 10.6, then the input voltages to each tube must be 180° out of phase with each

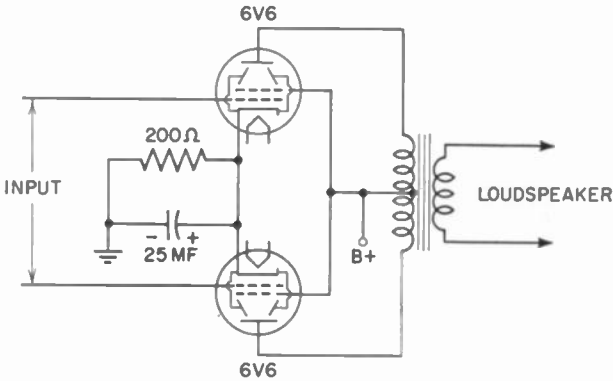


FIG. 10.6. A push-pull amplifier.

other. The simplest, although not the cheapest, means of accomplishing this is by using input transformers. Ordinary transformers, however, have a poor frequency response, and compensated transformers are quite costly. The best solution is the use of a phase inverter whereby a portion of the signal is taken and inverted so that it becomes 180° out of phase with its input signal.

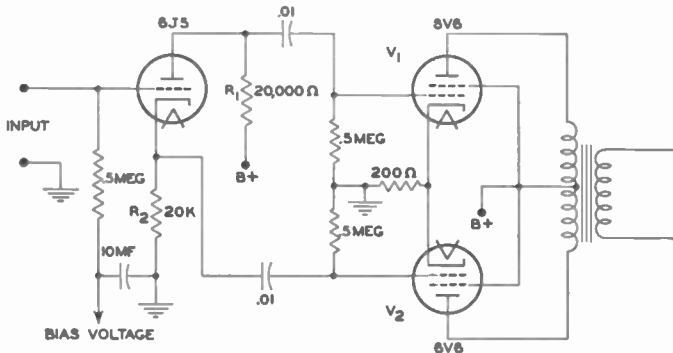


FIG. 10.7. Phase inversion supplied by a single triode.

This makes available two properly phased signals for the push-pull input.

Present phase inverters fall into several categories. The simplest arrangement is shown in Fig. 10.7, where a single triode (6J5, 6C5 etc.) supplies both out-of-phase voltages. One push-pull amplifier receives its

excitation from the plate of the phase inverter; the other push-pull tube receives its signal from the inverter cathode. That these two signals are 180° out of phase can be noted by tracing the current through the circuit. Electrons reaching the plate of the 6J5 must pass through R_1 and R_2 in series, with the top end of R_1 always opposite in sign with respect to the cathode end of R_2 . Perhaps a better way of visualizing the inverting action is by considering the path for the a-c component in the circuit. In Fig. 10.8 the

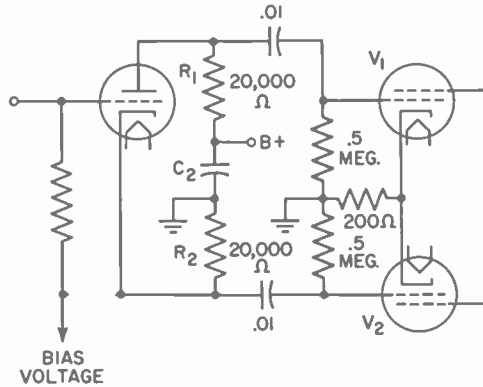


Fig. 10.8. Fig. 10.7 rearranged to emphasize the balance of voltage across R_1 and R_2 .

circuit is redrawn slightly to emphasize this aspect. Now we see that, as far as the signal is concerned, R_1 and R_2 are electrically connected at their common intersection to ground (through C_2). Each signal voltage is then taken from the remaining two ends and fed to each grid of the push-pull amplifier. Thus we have a balanced arrangement against ground.

In this inverter, R_1 and R_2 must always be equal in order that equal voltages are fed to V_1 and V_2 . Because R_2 is in the cathode leg of the 6J5 and unby-passed, the tube functions as a degenerate amplifier. The a-c voltage developed across R_2 opposes the input signal, decreasing its effectiveness at the grid. As a result, the gain of the stage, grid to grid at the push-pull amplifier, is approximately 1.8 to 2 times the input signal to the phase inverter. Consequently, if a large voltage is needed to drive V_1 and V_2 , this arrangement is not very feasible. It does, however, possess the advantages of simplicity, low distortion due to the degenerative feedback, freedom from changes in tube emission, and the ability of always developing equal signals for the push-pull amplifier.

Two variations of the circuit are given in Fig. 10.9. In Fig. 10.9A the degeneration is eliminated by feeding the input signal between R_2 and R_3 . R_2 is fully by-passed by C_1 to provide the proper bias for the tube. Hence,

no audio voltages appear across R_2 . R_g , the grid input resistor, is connected to the top of R_3 and, since no degeneration appears between this point and cathode, full input voltage is effective in varying the current through the phase inverter. The gain of this network is higher than in the previous circuit, but its fidelity is not as good because of the removal of the degeneration.

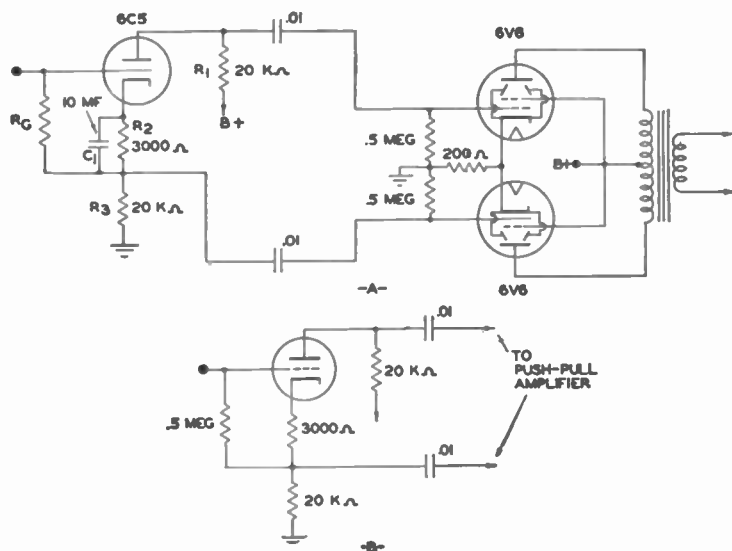


FIG. 10.9. Two alternate phase-inverter circuits.

In Fig. 10.9B, the smaller cathode resistor is left unby-passed to provide a small amount of negative feedback. This aids the stability of the phase inverter and provides a better high and low frequency response.

Another phase inverter circuit is shown in Fig. 10.10. Here, either V_1 or V_2 may be separate tubes or else each may be part of a duo-triode. The output of the first triode, V_1 , is fed to one grid of the push-pull amplifier. A portion of this voltage, however, is also applied to the grid of the second triode (V_2) and the output from this tube applied to the grid of the other push-pull amplifier tube. Since the input and output voltages of V_2 differ by 180° , we have the output voltage of V_2 differing from the output voltage of V_1 by 180° . In this way each grid of the push-pull amplifier receives its voltage in proper phase relationship. Since V_2 is used for the sole purpose of taking a portion of the output voltage of V_1 and reversing it in phase, it is actually the phase inverter.

The amount of voltage that is fed to the grid of V_2 depends upon the ratio of R_4 to $R_3 + R_4$ and the amplification of V_2 . Thus, for the voltage

amplification at V_2 to be 40, the voltage tapped off at R_4 should be $1/40$ of the total voltage across $R_3 + R_4$. With this method, each grid of the push-pull amplifier tubes receives the same input voltage, and the circuit is balanced. In the present circuit, the ratio of R_4 to $R_3 + R_4$ should equal $1/40$, and substitution of the given values for R_3 and R_4 , $10,000/410,000$, shows this to be approximately true.

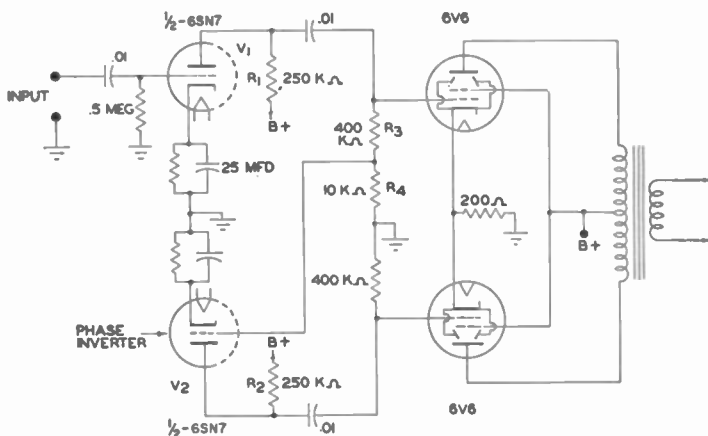


FIG. 10.10. Phase inversion using a separate tube. This arrangement is sometimes called a paraphase circuit.

Unlike the previous phase inverters, this circuit is capable of fairly high gain. Its greatest disadvantage, however, is the unbalance that occurs whenever the amplification of either triode changes. Both triodes are unrelated and any change that occurs in one section is not automatically compensated for by the other triode. Hence, unless care is taken to measure the characteristics of each tube from time to time (and making compensations therefore), some unbalance will always be present.

Partial compensation for unbalance is obtained by the circuit of Fig. 10.11. A change has been introduced in the grid circuit of the push-pull amplifiers by the insertion of the common resistor, R_6 . Since R_6 is also in the input circuit of the phase-inverter tube, V_2 , any changes here will also affect this tube's operation. V_2 receives its input voltage from R_3 and R_6 , whereas its output voltage appears across R_4 , R_5 , and R_6 . For V_1 , the output signal is developed across R_2 , R_3 , and R_6 .

To understand how the compensation is accomplished, let us suppose that both V_1 and V_2 are functioning normally. The voltage for tube V_3 appears between point A and ground; for tube V_4 , between point C and ground. The voltage present at R_6 , under normal conditions, is zero because the out-

put voltages from V_1 and V_2 are out of phase and hence cancel across this common resistor. But, now, let us suppose that the amplification of one tube decreases, say, V_1 . Thus, if point A went positive, say, 8 volts at one instant, point C (with respect to ground) might be 10 volts negative because amplification of V_2 . Under these circumstances, the net voltage appearing across R_6 would not be zero but some small negative value. Since the voltage

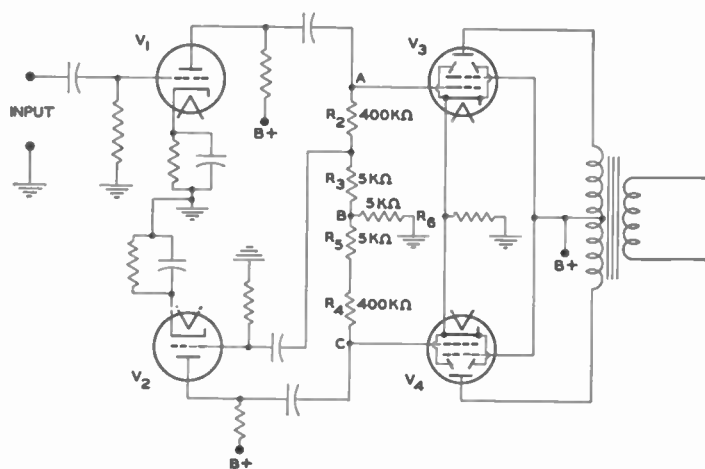


FIG. 10.11. A modified phase inverter to produce automatic balancing.

applied to the grid of V_2 is composed (at the moment) of the positive voltage of R_3 and the small negative voltage of R_6 , the net voltage acting at the grid of V_2 is slightly less than it ordinarily would be. As a result, the output voltage of V_2 is decreased, counterbalancing the increased amplification of this tube.

Conversely, if V_2 should decrease in amplification, a greater voltage would be applied to its grid, giving a correspondingly larger output. With limits, the common resistor R_6 tends to maintain a balanced circuit, although at some sacrifice in overall gain.

Another popular inverter type is the cathode-coupled inverter shown in Fig. 10.12. The first triode, V_{1A} , receives the audio signal from the previous stage across its grid resistor, R_1 . This signal is amplified and forwarded through C_1 to the grid of one push-pull amplifier tube. The load resistor for V_{1A} is R_2 . Since the cathode resistor, R_4 , is unby-passed, the signal variations appear in full across it as well as across R_2 . The grid of V_{1B} is placed at a-c ground potential by C . Thus, the signal variations appearing across R_4 are also instrumental in varying the current through V_{1B} and producing an amplified version across plate load resistor R_3 . From here, the signal is coupled to the grid of the second push-pull amplifier tube.

The two output signals are 180° out-of-phase with each other. A positive increase in signal at the grid of V_{1A} will cause a corresponding increase in voltage at the cathode end of R_4 . Since the a-c potential of the grid of V_{1B} is fixed (at ground), the positive rise in its cathode voltage is equivalent to a negative increase in grid voltage. Thus, for the same applied voltage, the grid of V_{1A} is driven in the positive direction while the grid of V_{1B} goes in the negative direction.

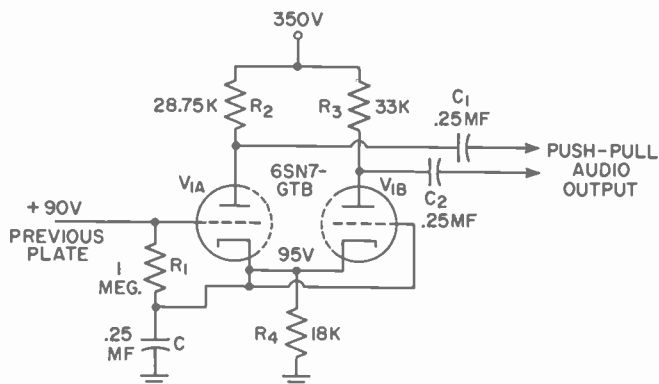


FIG. 10.12. A cathode-coupled phase inverter. (Courtesy Radio & TV News)

The plate resistors have different values in order to balance the two circuits. Since the cathodes have a fairly high positive voltage, because of the large value of R_4 , a fairly high positive voltage is required at each grid. 95 volts, positive, is indicated for each cathode; therefore, each grid should be given a positive voltage of 90 volts. This is obtained by direct connection to the plate of the previous stage. The grid of V_{1B} connects directly to the bottom end of R_1 in order to obtain the same B+ as the grid of V_{1A} . However, by adding C at the bottom of R_1 we are able to keep the grid of V_{1B} at a-c ground.

The cathode-coupled inverter possesses good balance at both high and low frequencies. It is also exceedingly stable because of the large amount of degeneration used. It does not, however, provide any gain and, in this sense, is not as good as either of the previous inverters. The addition of an extra amplifier ahead of the stage will compensate for this.

Negative Feedback. The beneficial effects of feeding back some voltage which is out of phase with the input voltage was mentioned briefly above in connection with phase inverters. Actually in the case of Fig. 10.7, the degeneration (or negative feedback, as it is known) was unintentional. It was necessary to leave R_2 unby-passed in order that the alternating voltage

developed across it could be transferred to the next stage. However, in many amplifiers, the negative feedback is purposely introduced. Its advantages are:

1. Improvement of the frequency response of the amplifier.
2. Decrease in distortion in an amount depending upon the percentage of voltage fed back.
3. Increased stability, with less tendency for regeneration (whistles or howls) to appear.

It is true, of course, that these advantages are not obtained without a loss—in this case, gain. For, as the degree of negative feedback increases, the gain of an amplifier decreases. However, for most amplifiers, especially those used in radio receivers and small amplifiers, the amount of available gain exceeds the need, and a slight loss for the advantages noted above is voltage well spent.

In negative feedback, a certain proportion of the output voltage is fed back to a previous stage in opposition to the input voltage that is present at this point. The circuit in Fig. 10.13 illustrates the principle. R_1 and R_2 form a voltage divider across the output of V_1 . Through the connection from point A to point B , a portion of the audio voltage that is developed at R_L is fed back to the grid circuit of V_1 . Now let us determine whether the voltage fed back is in phase opposition to the incoming voltage appearing across R_2 .

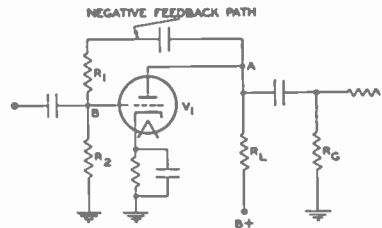


FIG. 10.13. A simple negative feedback circuit. R_1 and R_2 form a voltage-divider network across R_L .

If the incoming signal to V_1 at any instant is going in the positive direction, then the voltage at point A is going negative. This action is due to the increased current through the tube, resulting in a greater voltage drop across R_L , leaving the top end (point A) more negative (or less positive because of the increased voltage drop in R_L) than it was before the application of the signal. By connecting points A and B together, the voltage from R_L (E_L) will cancel part of the positive incoming audio voltage at point B . This cancellation is the essential principle of negative feedback. When the incoming signal is strong, more voltage is fed back from A to B and a greater cancellation occurs. With a weak signal, the opposition of the feedback voltage is correspondingly reduced. Thus, the negative feedback voltage acts to maintain a constant output. Again, if distortion is introduced between points B and A , the voltage transferred back will contain this dis-

tortion. However, because the voltage fed back from A to point B is opposite in phase to the signal passing through V_1 , the feedback voltage will introduce this distortion in a manner opposite to that produced by the tube. When the signal, with this reverse distortion, passes through the tube, the impact of the tube distortion is partially nullified. The result—a decrease in the overall distortion as seen in the output.

The amount of feedback in the circuit of Fig. 10.13 is determined by the ratio of R_2 to $R_1 + R_2$. ($R_1 + R_2$ parallel R_L and the output is developed across them.) This voltage is divided between them in direct proportion to their relative resistances, with the larger resistor obtaining a larger percentage of the voltage. If R_2 is large, a greater percentage of voltage will be transferred to point B . This, in turn, will lower the gain considerably. Generally a compromise is reached, where the gain is still appreciable but sufficient feedback is developed to afford good stability with low distortion.

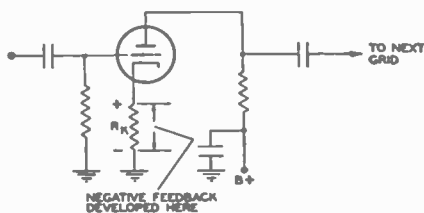


FIG. 10.14. Negative feedback obtained from an unby-passed cathode resistor.

One very simple form of negative feedback is obtained by removing the cathode by-pass capacitor (see Fig. 10.14). The a-c portion of the plate current must now flow through the cathode resistor R_k instead of being by-passed. As a result, an alternating voltage is developed across R_k which opposes the input signal. To illustrate, suppose the grid of V_1 is made positive by an incoming signal. The plate

current through V_1 will increase, increasing the voltage drop across R_k with the polarity as noted. However, the voltage at the grid that caused the increased flow of current through the tube is the grid-to-cathode voltage. Since, in this instance, the cathode becomes more positive, part of the positive signal voltage rise at the grid is counterbalanced by the positive voltage rise at the cathode.

There are many other possible arrangements and several of the more common methods are shown in Fig. 10.15. In each instance the opposition of the voltage being fed back may be noted by following the procedure outlined above. Feedback may occur from the plate of one stage to its grid, or it may extend over several stages. Whatever the method, care must be taken to see that the feedback is degenerative, not regenerative. Regenerative feedback means voltage fed back in phase, not out of phase, and under these conditions oscillations will occur.

Output Stages. Mention has already been made at the start of this chapter of the advantages of push-pull operation and, for this reason, most high fidelity amplifiers use this form of circuit. However, before we ex-

amine some of the most popular push-pull circuit arrangements, it should be noted that single-ended stages are used also, particularly in table-model receivers where the output power requirements are quite low. For example, the audio amplifier section of one such F-M receiver is shown in Fig. 10.16. The sound output of the preceding ratio detector is fed through a volume

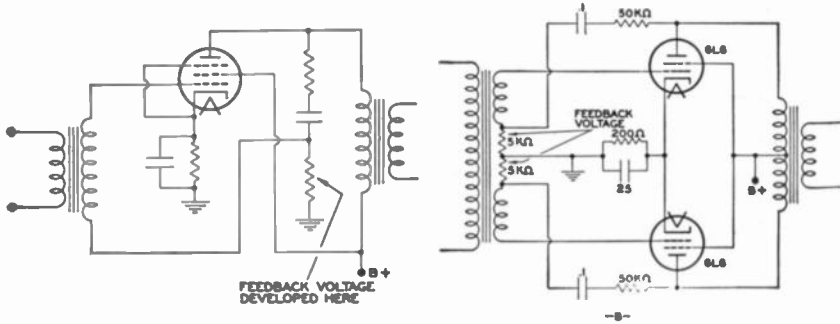


FIG. 10.15. Two additional methods for obtaining negative feedback.

control to one triode section of a 12AX7. This is a voltage amplifier. The signal next is applied to a 6AK6 pentode output stage and, from here, to an output transformer and a 5-inch speaker. In essentially all respects, this amplifier section is equivalent to those found in most A-M radio receivers.

Push-pull output circuits fall into two general categories, those using

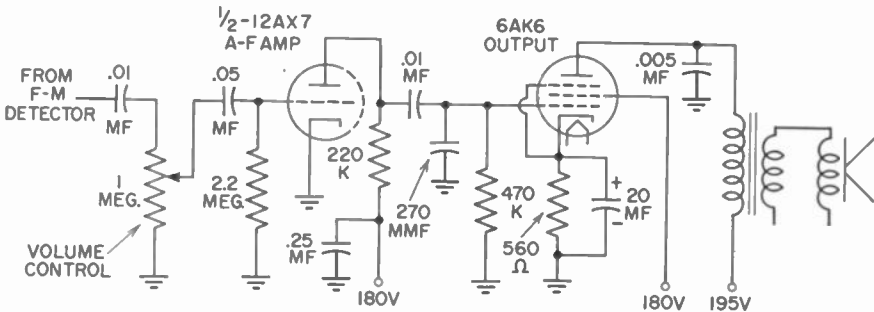


FIG. 10.16. The audio amplifier section of a table-model F-M receiver.

triodes and those using pentodes or tetrodes. (The latter two tube types are grouped together because they function in essentially the same manner and provide similar results.) In the early days of high fidelity, triodes or pentodes and tetrodes connected as triodes were used most extensively because they provided the best results from a distortion standpoint. Then, to reduce whatever distortion did appear, negative feedback was inserted.

The famous Williamson amplifier used such an arrangement, as shown in Fig. 10.17. The input signal is amplified by one triode section (V_{1A}) of a 6SN7GT. The signal is then directly-coupled to the second triode, V_{1B} , where two oppositely-phased voltages are developed to drive V_2 . Thus V_{1B} is a split-load phase inverter. V_2 serves as a driver, to strengthen the signal it receives to drive output tubes V_3 and V_4 . These two tubes are beam power tetrodes connected as triodes. The signal is then fed to the output transformer and, from here, to the loudspeaker.

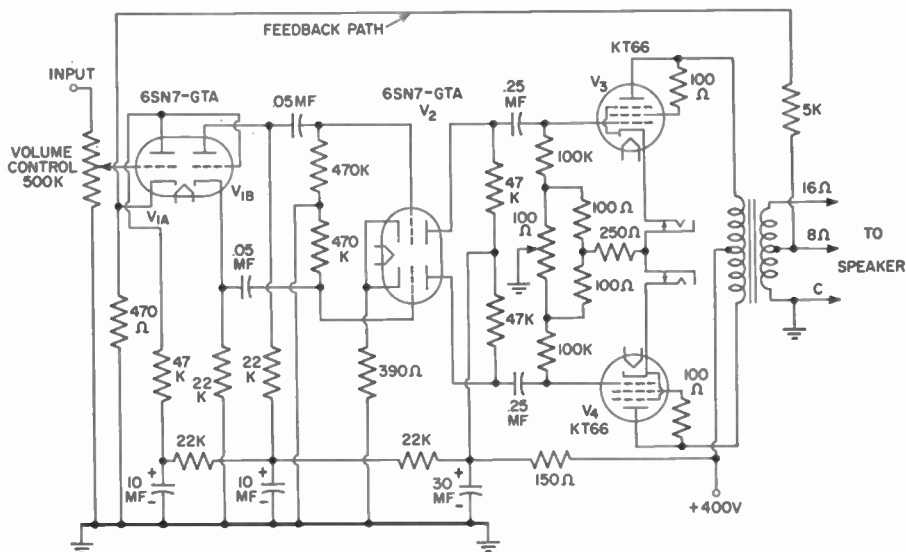


FIG. 10.17. The Williamson amplifier in which the two beam-power tetrode output tubes, V_3 and V_4 , are connected as triodes.

Negative feedback, to stabilize the system and reduce distortion, is employed between the secondary of the output transformer and the cathode of V_{1A} . As a further aid to the minimization of distortion, the two output amplifier circuits are balanced by a 100-ohm potentiometer. This is in the grid circuits of V_3 and V_4 ; it is adjusted correctly when both tubes draw the same cathode current with no applied signal.

A tetrode or pentode, connected as a triode, requires less driving power for a specified output than does a triode. However, a tetrode or pentode does introduce a greater amount of distortion. A configuration which combines high power sensitivity with low distortion is the Ultra-Linear arrangement shown in Fig. 10.18. The plates of the output tubes are connected to opposite ends of the primary winding of the output transformer, as before. However, special taps are available on the winding for the screen grid of each

tube. This has the effect of providing some negative feedback to the screen-grid circuit which straightens out the characteristic curves of these tetrode (or pentode) tubes. The result is very low distortion with good power sensitivity.

Still another output power amplifier circuit that has received attention is the cathode feedback circuit shown in Fig. 10.19. A special winding is added to the output transformer and the cathode of each tube is connected to one end of this winding. This arrangement has excellent fidelity characteristics (i.e., low distortion), but it requires a considerable amount of driv-

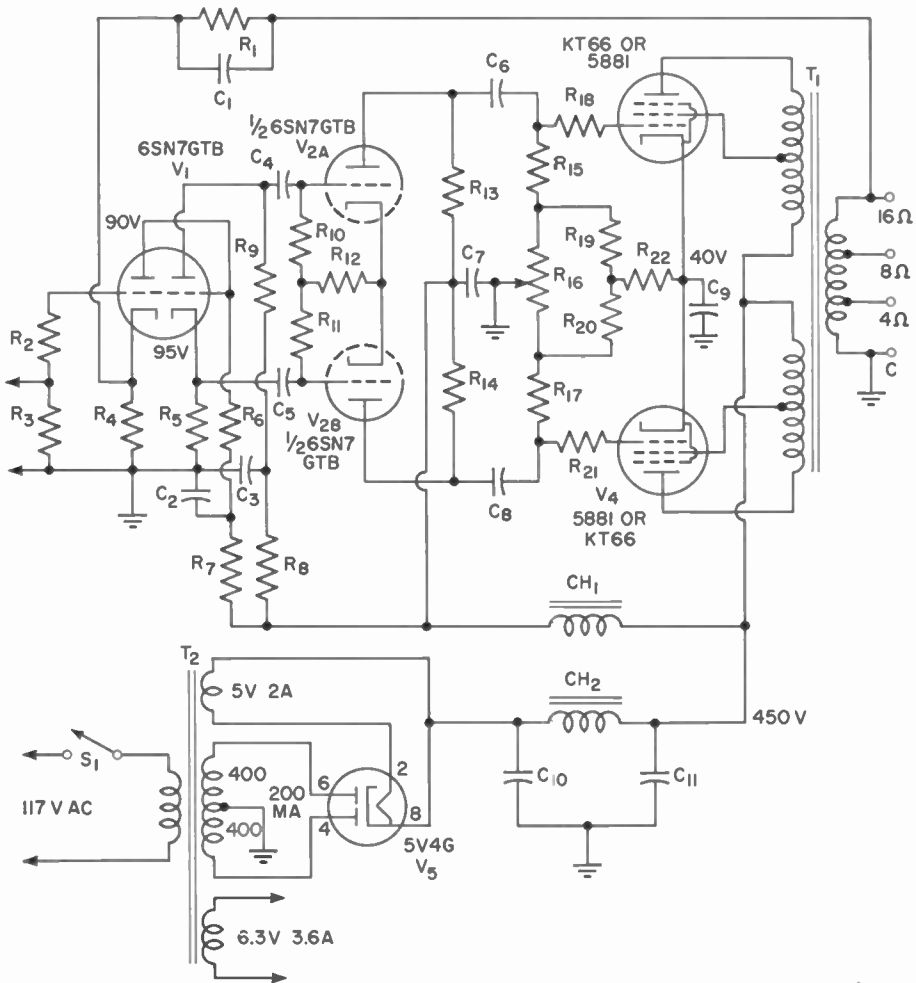


FIG. 10.18. A power amplifier in which the output tubes (V_3 and V_4) are employed in an ultra-linear arrangement.

ing voltage because of the cathode degeneration. In this sense, it is quite similar to the cathode follower, where the gain is low.

Where extremely high output power is desired, on the order of 60 watts or more, it is common practice to parallel the output amplifiers. Generally, two tubes in parallel for each half of the output amplifier suffice for most high power requirements. However, if more power is needed, additional tubes can be added.

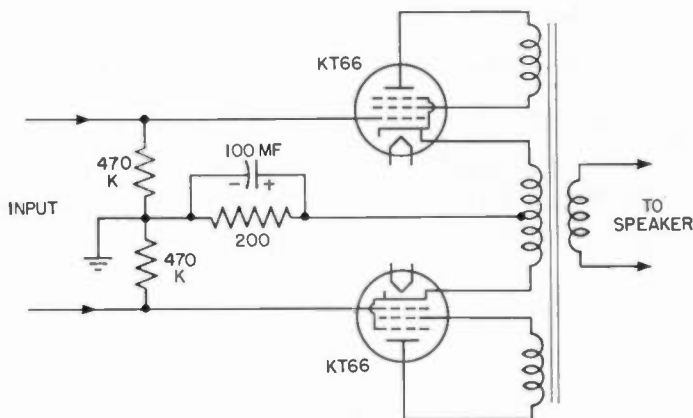


FIG. 10.19. A push-pull output amplifier using cathode feedback.

Most push-pull amplifiers are conservatively operated as Class A or Class AB₁. This results in low distortion. True Class-B operation requires more driving power and introduces stricter design requirements that can raise the cost of the unit. Consequently, Class-B operation is not used as often.

Cathode Followers. It is common practice in modern high-fidelity receivers to have the tuner separate from the audio amplifier. The tuner, whether it be a combination A-M, F-M unit or designed for A-M or F-M alone, receives the incoming signal, amplifies it, and detects it. After perhaps one stage of audio amplification, the signal is ready to be applied to the audio amplifier where it will be strengthened sufficiently to drive the loudspeakers.

In order to transfer the signal from the tuner to the amplifier, a shielded coaxial cable is employed. This cable has a fairly low impedance, of the order of 600 ohms, and to shunt the normal plate load resistor of an amplifier with so low an impedance would not only reduce the output voltage to a very small value, but it would also cause considerable signal distortion.

Much more useful in matching the cable's impedance is the special am-

plifier stage known as a cathode follower. The basic form of this circuit is shown in Fig. 10.20. The load resistor is removed from the plate circuit and shifted to the cathode branch. No by-pass capacitor is used and any applied audio voltages between grid and cathode will appear across R_k . The output signal can then be taken from this point.

Since the cathode resistor is unby-passed, considerable degeneration will occur. Furthermore, the phase of the voltage across R_k is the same as the signal phase at the grid. For example, if a positive-going signal is applied to the grid, the rise in plate current will produce a greater voltage drop across R_k , making the cathode more positive. Likewise, a negative-going voltage at the grid will reduce the plate current, decreasing the voltage drop across R_k . This will make the cathode less positive or, what is the same thing, more negative.

Thus, the voltage across the cathode resistor "follows" the grid and this is the reason why this circuit is called a "cathode follower."

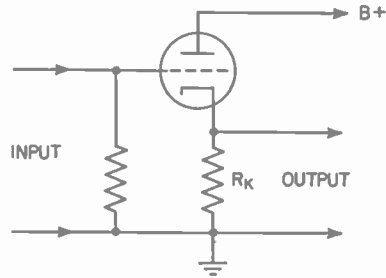


FIG. 10.20. The basic cathode follower.

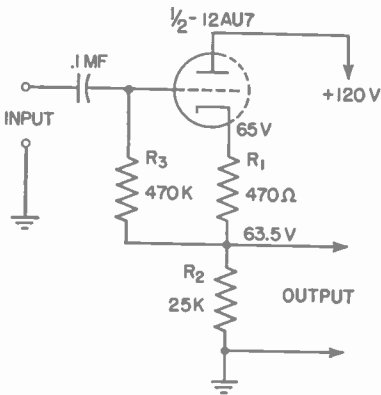


FIG. 10.21. A modification of the cathode follower of Fig. 10.20. The voltages shown are with respect to ground.

The gain of a cathode follower is normally on the order of 0.9. This means there is a slight loss, due to the large amount of degeneration or negative feedback. The degeneration is also responsible for the low output impedance of the stage. It is this low output impedance which makes it possible to use a coaxial cable between the cathode follower and the power amplifier.

A modification of the cathode follower, which is used more frequently than the basic arrangement of Fig. 10.20, is shown in Fig. 10.21. This possesses a separate bias resistor R_1 which establishes the operating voltage between grid and cathode. The grid input resistor, R_3 , is then connected between the grid and the bottom end of R_1 . With a large resistor in the cathode circuit, the cathode voltage will be correspondingly high—here on the order of -65 volts. If R_3 is returned to ground, the grid would see a negative bias of -65 volts, sufficient to cut the tube off. (Actually, the tube could not

cut itself off, for then the cathode voltage would disappear. However, enough current would flow to bring the tube close to cutoff.)

All this is avoided by returning R_3 to the bottom end of R_1 . Now the grid-to-cathode voltage is -1.5 volts and the d-c voltage drop across R_2 does not affect this bias. However, since R_2 is unby-passed, it still produces the cathode follower action.

Preamplifiers. Before we leave the subject of audio amplifiers, mention should be made of preamplifiers. These are *specially* designed amplifiers which take the exceedingly small voltages developed by modern magnetic cartridges such as are used in record players and amplify them so that they can be fed into the input of a power amplifier.¹ The average voltage obtained from a magnetic cartridge is on the order of 15 to 20 millivolts. This voltage is too small to drive a power amplifier.

The word "special" is emphasized in connection with preamplifiers because of the precautions necessary to minimize hum and resistor and tube noise (described earlier on page 38). With the desired signal possessing a very small amplitude, any hum or noise which, in ordinary amplifiers, could be ignored, must be carefully avoided in preamplifiers. Otherwise, the amplification accorded the signal will serve equally to strengthen the hum or noise and provide an input signal to the main amplifier having a strong hum or noise background.

To reduce tube hum, the filament has a spiral construction in which the magnetic field ordinarily set up by the current heating the filament is reduced to as low a value as possible. Additional precautions are also taken to prevent current leakage from filament to cathode, another source of hum trouble. So important is the need to minimize hum in the preamplifier that many designers use d-c to heat the filaments instead of a-c. The required current can be obtained from a separate power supply or by utilizing the cathode currents flowing through the power output amplifier.

Noise arises from the current flowing in the preamplifier tubes and from the grid and plate resistors used with these tubes. Special construction will also reduce tube internal noise. Resistor noise is reduced by using deposited film resistors and noninductive wire-wound resistors. These produce considerably less noise than composition resistors of the same rating.

The circuit diagram of a two-stage preamplifier is shown in Fig. 10.22. It is similar in design to a conventional audio voltage amplifier, incorporating, however, the low hum and low noise features mentioned previously. The 12AX7, for example, is specially designed to possess low noise and low hum. Also, the resistors are carefully selected for their low noise qualities.

Most preamplifiers also incorporate an equalization network, either

¹ By power amplifier, we refer here not solely to the final stage of an audio amplifier system, but the entire system, such as that shown in Fig. 10.17.

between the two stages or between the last stage and the output of the section. The same preamplifier (Fig. 10.22), with a number of equalizing circuits, is shown in Fig. 10.23. Five equalizing networks are shown, and the desired one can be selected by a special 5-position switch.

Equalizing circuits are needed to counterbalance the inverse process which is applied to records when they are manufactured. For example, when a record groove is cut for a low frequency sound, the applied voltage from the microphone (or its amplifiers) is reduced in order to avoid cutting too wide a groove. This is called de-emphasizing the low frequencies. On the

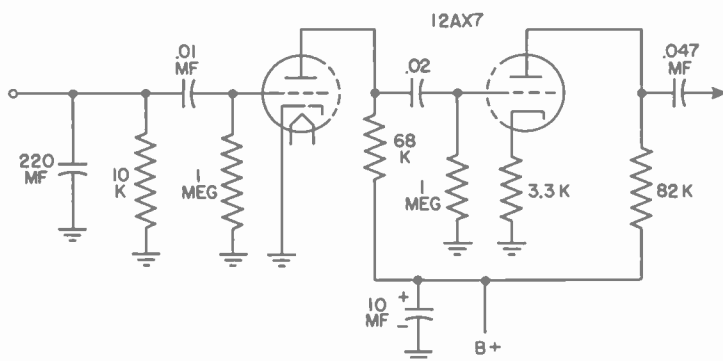


FIG. 10.22. A record player preamplifier.

other hand, frequencies above 1000 cycles are given additional amplification—i.e., pre-emphasized—in order to overcome the surface noise which is present in the record. At the amplifier, on playback, these signals must be returned to their proper relative amplitudes. This is the purpose of the equalizing networks. They are designed to de-emphasize the high frequencies and emphasize the low frequencies. When properly carried out, the net result is an exact reproduction of the original audio signals fed to the recording system.

Five different equalizing networks are used in the preamplifier in Fig. 10.23. Prior to 1953, there was no general agreement on the amount of de-emphasis and pre-emphasis to employ when cutting a record. Each record manufacturer set his own standard. Thus, Columbia Records had one recording and playback curve, RCA utilized another, etc. To develop the proper sounds from these records, each had to be played back with the correct equalization network. In 1953, however, agreement was reached by the major record manufacturers on a standard curve and, as this was done under the auspices of the Record Industry Association of America, Inc., the curve is now known as the RIAA curve.

In the preamplifier of Fig. 10.23, RIAA equalization is obtained with the selector switch in the third position. Since many of the older records are still in circulation, equalizing networks for the more popular of these are also included. These are: LP, NAB, and AES. One position, too, is provided for European records, as these require still a different equalization network.

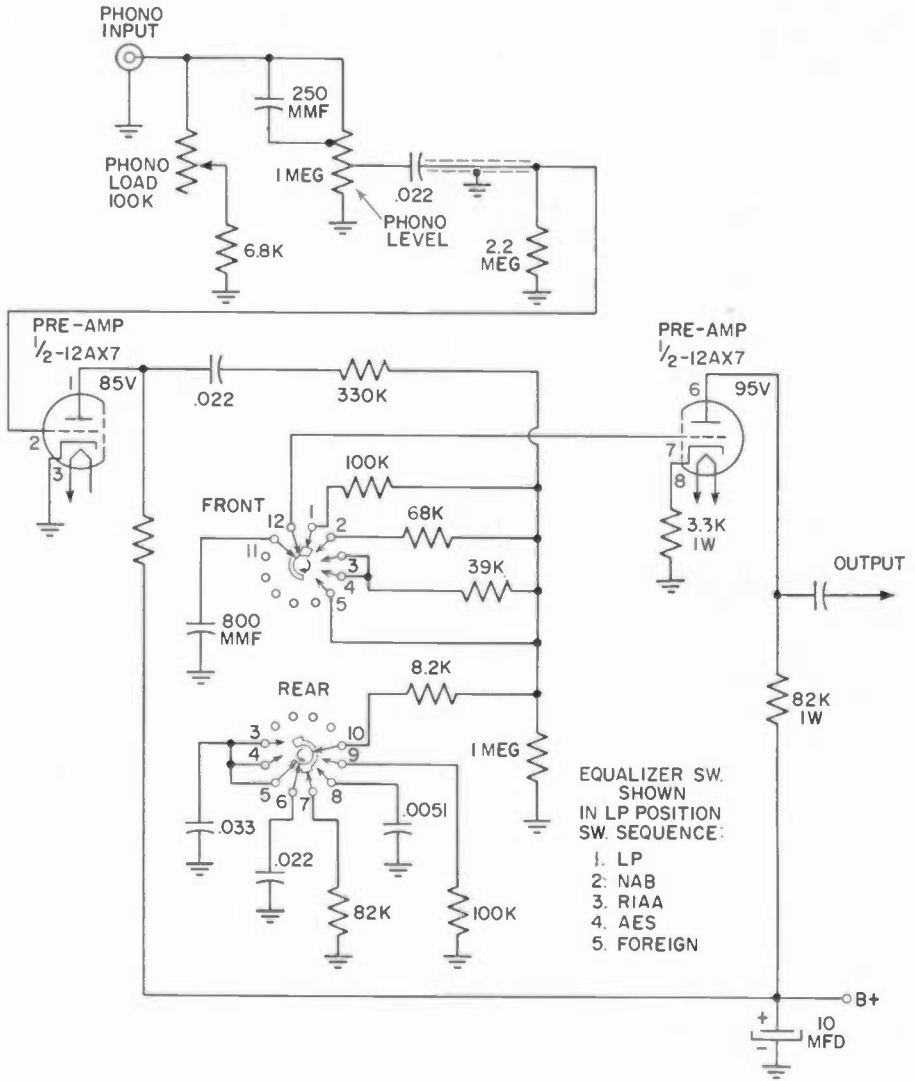


FIG. 10.23 A commercial preamplifier with an equalizing network. (Courtesy H. W. Sams & Co.)

A comparison of these different equalization curves are shown in Fig. 10.24. These are the curves of the playback amplifier equalization networks where the low frequencies are emphasized and the high frequencies are de-emphasized. In the record-cutting equipment, the inverse equalization characteristic would be applied. That is, the low frequencies would be de-emphasized and the high frequencies emphasized.

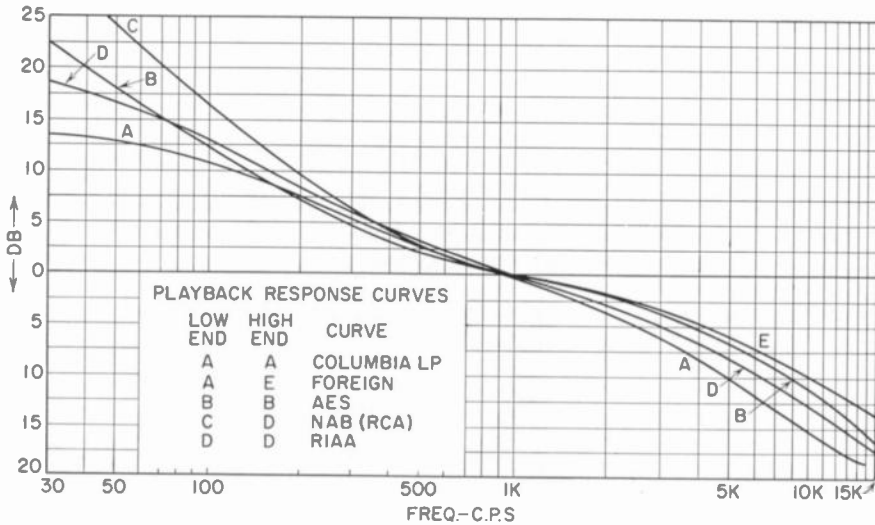


FIG. 10.24. Five equalization curves that have been used by record manufacturers. The standard curve, at present, is the RIAA (curve D). (Courtesy *Radio Electronics*)

Tone Control Circuits. It is due to a recognition of the variation in individual tastes plus the fact that radios are used in widely differing locations that tone controls are incorporated into a large number of sets. Tone control permits the listener to vary the amplification applied to a range of frequencies and in this way cause a definite accentuation or boosting within this range. For the reception of music, for example, many listeners prefer to turn their tone control to the point where the bass frequencies predominate. For speech, the adjustment is made toward the higher frequencies. Whatever the preference, we are more interested in this discussion as to how tone control is electrically achieved rather than why a certain tone is preferred.

Tone-control systems are essentially of two types. In one system, the apparent accentuation of one range of frequencies is achieved not by an actual boost within this range but by decreasing the strength of the other frequencies present. For example, to raise the level of bass frequencies, many manufacturers provide an adjustment which will accomplish this boost in

a relative manner, by decreasing the intensity of the treble frequencies. The effect of the high frequency decrease upon the listener is as though there had been an accentuation of the bass tones. The second system involves an actual boosting of the frequencies that we desire to accentuate.

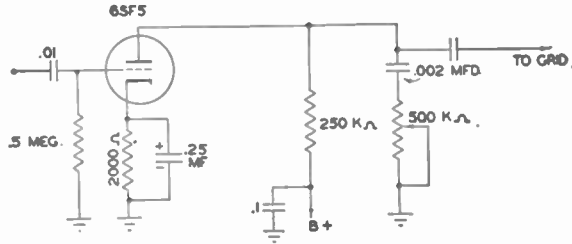


FIG. 10.25. A simple bass-boost tone control. The variation of control is accomplished with the 500,000-ohm potentiometer, but the entire tone control circuit includes also the 0.002-mf capacitor.

Simple Bass-Boost Control. A simple tone control that is found in many sets is shown in Fig. 10.25. It is placed in the plate circuit of one of the audio amplifiers and functions through its ability to decrease the relative strength of the high frequencies that are permitted to reach the speaker. When the

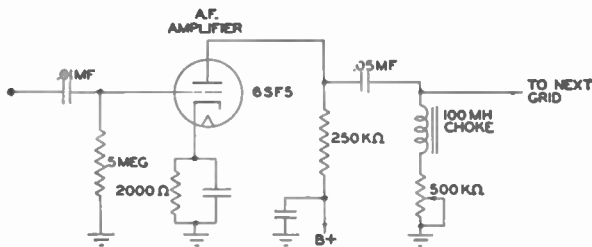


FIG. 10.26. A simple treble tone control.

center arm of the resistor is at the end nearest the capacitor, the decrease of the high frequencies is maximum because only the capacitor is opposing their passage around the plate circuit. At these times the bass output will seem strongest; moving the center arm to the opposite end of the resistor permits a greater percentage of the highs to reach the output because the by-passing path's resistance has been increased.

Treble Control. The opposite effect can be achieved by replacing the series capacitor of Fig. 10.25 with a series choke coil, as shown in Fig. 10.26. Now we have a path for the low frequencies to be shunted away from the output. When the center arm is at the end of the resistor nearest the coil, the maximum shunting effect is imposed on the low frequencies. The posi-

tion would correspond to greatest treble output from the set. When the center arm is at the opposite end of the resistor, the opposition to the lows has increased, providing more frequencies from this range for the output. If we wish to combine both circuits, obtaining a more flexible control, we have the arrangement of Fig. 10.27. This unit is particularly useful at low

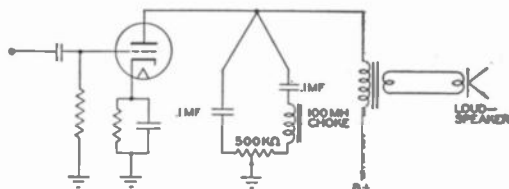


FIG. 10.27. A combination "treble-bass" tone control.

volume levels because it tends to provide a more uniform response over the entire audio range. It was noted from the response curves of the ear in Fig. 10.1 that, as the volume is decreased, the high and low frequencies decrease more rapidly than the middle-range frequencies. In the control of Fig. 10.27, the middle frequencies are attenuated more than those at either end of the audio range. The overall result is a more uniform distribution.

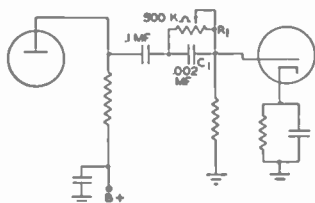


FIG. 10.28A. A series treble tone control.

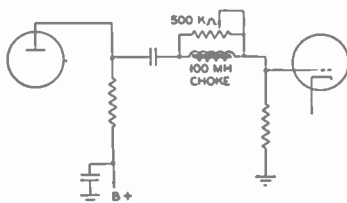


FIG. 10.28B. A series bass-boost tone control.

Instead of placing the tone control in shunt across the circuit, it is also possible to place it in a series branch. For low frequency attenuation, the circuit of Fig. 10.28A is possible. The low frequencies lose more voltage across C_1 than the higher frequencies. Hence, when R_1 is completely in the circuit, the greatest amount of loss occurs at the low frequencies. By gradually shunting out a part of R_1 , we can raise the intensity of the lows at the output. As a treble attenuator, the capacitor can be replaced by a choke (see Fig. 10.28B).

Tone Control by Inverse Feedback. Tone variation can be also accomplished by inverse or negative feedback methods. Two popular methods

are shown in Fig. 10.29. In Fig. 10.29A, negative feedback effect is incorporated in the cathode leg of the tube. R_1 and C_1 , in series, provide the degenerative effect. The value of C_1 determines the frequencies at which the effect first becomes noticeable. If C_1 is small in value, the decrease in output occurs at the middle and low frequencies; if C_1 is large, the output decreases only for the very low frequencies. The whole operation of this network depends upon the fact that if the cathode bias resistor is not adequately by-passed, the voltage across it will vary and tend to counteract the effect of the input voltage. When C_1 is small, only the higher frequen-

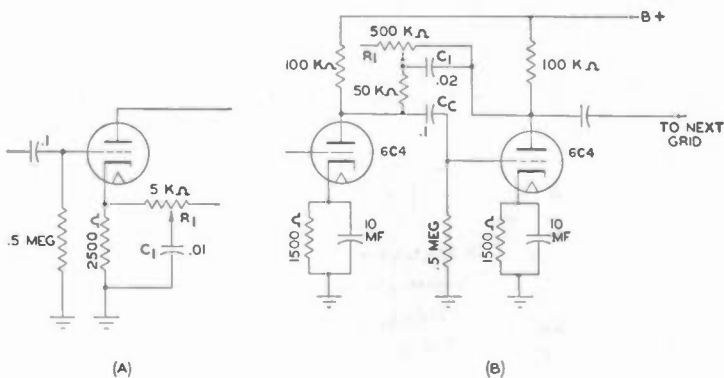


FIG. 10.29. Two common methods of applying negative feedback to obtain tone control.

cies are readily by-passed, the middle and lower frequencies producing a variable drop across C_1 because of the substantial impedance offered them by the capacitor. The variable cathode bias (for the frequencies mentioned) will partly counteract input voltages of the same frequencies. The result is that the higher frequencies remain untouched and appear more strongly at the loudspeaker. As C_1 is made larger in value, its opposition to even the lower frequencies decreases. Hence, these, too, pass through undiminished. The inclusion of the additional frequencies lessens the aural effect of the highs, and the tone becomes deeper. R_1 is inserted to provide a variable adjustment.

Another approach to the problem of negative feedback tone control is the plate-to-plate connection illustrated in Fig. 10.29B. Actually, what we do here is to feed back a voltage from the plate of the second 6C4 to the grid of the same tube. We use the plate-to-plate connection to keep the positive B+ voltage of the 6C4 from reaching its grid. C_c isolates the voltage from the grid. In the circuit, C_1 and R_1 combine to regulate the high frequency voltage that is fed back from the plate to the grid. Since the high

frequencies return out of phase, they decrease other incoming high frequencies, permitting the lows to dominate at the output. The position of the center arm of R_1 determines how much high frequency degeneration occurs.

In all of these circuits, it is to be understood that no network permits one range of frequencies to pass, excluding all others. It is merely that one range of frequencies is offered less opposition than other frequencies and, hence, a proportionately greater amount of voltage from this range passes through.

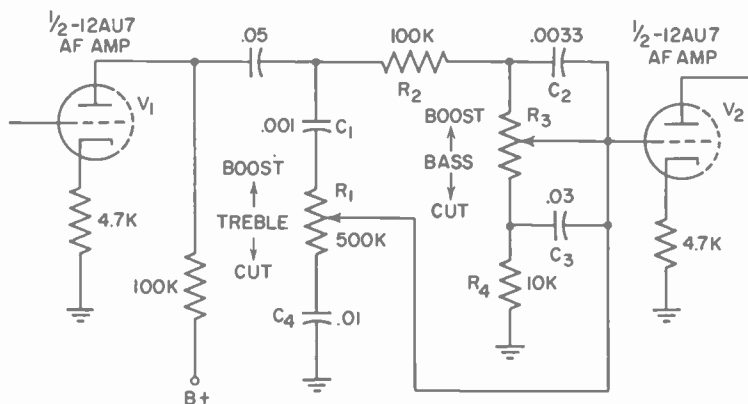


FIG. 10.30. A continuously variable tone control.

Continuously Variable Tone Controls. All of the preceding tone-control circuits have been so designed that, in order to obtain either a high or low frequency output sound, the opposite range of frequencies is attenuated. The circuits have the advantages of economy in construction and simplicity of design. Much more complex are those arrangements whereby a boost or increase is given to the section of frequencies we desire to accentuate. A typical circuit is shown in Fig. 10.30. It is capable of independently boosting or attenuating both the high and low frequencies. This is graphically illustrated by the curves shown in Fig. 10.31. At the low-frequency end of the graph, the upper curve demonstrates the effect of bass boost. This is 16 db at 30 cycles. The lower curve at this end is the bass cut and it reaches a value of 15 db. At the right-hand side of the graph, the extent of the treble boost is revealed by the upper curve and is seen to reach a value of 16 db at 20,000 cycles. Treble cut, at this frequency, is 17 db. The action at intermediate frequencies, between 30 cycles and 20,000 cycles, is indicated by these curves.

In the circuit, R_2 , R_3 , R_4 , C_2 , and C_3 form the bass boost and cut circuit. R_3 is the variable front-panel adjustment. In its maximum clock-

wise position, the bass output is greatest; in its maximum counter-clockwise position, the bass output is lowest. The taper on the control is such that, in its mid-position, the resistance between the arm and the lower end of the control represents only about 10% of the total resistance of the potentiometer. This means that any medium frequency signal passing through will be reduced in the ratio of 10 to 1. This is also true at the high frequencies, when C_2 and C_3 effectively short out R_3 , because then R_2 and R_4 possess the same 10 to 1 ratio.

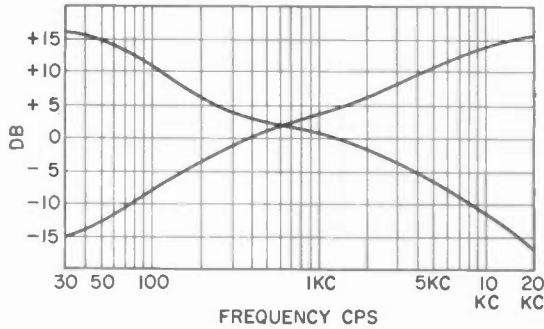


FIG. 10.31. Curves showing the compensating effect of the bass and treble tone controls of Fig. 10.30.

Note that the capacitance values of C_3 and C_2 are also proportioned in the 10 to 1 ratio to insure that this division is maintained at all frequencies with the arm of R_3 in the center position.

Consider, now, what happens when the arm of R_3 is moved up. This is the bass boost position. C_2 is shorted out. For the high audio frequencies, C_3 presents a very low impedance, and any signal coming from the previous stage divides between R_2 and R_4 . Since R_4 has a resistance one-tenth that of R_2 , the next stage (V_2) receives one-tenth of the total signal.

As the signal frequency decreases, the impedance of C_3 rises, applying more and more voltage to the grid of V_2 . At 30 cycles, the impedance of C_3 is so high that V_2 receives nearly 9/10 of the applied signal. This represents a boost over the center setting of the bass control.

When the arm of R_3 is turned to its maximum counter-clockwise position, C_3 is shorted out. High audio frequencies pass through C_2 readily and divide between R_2 and R_4 in the ratio of 10 to 1. As the frequency decreases, the impedance of C_2 rises and this places a greater impedance in the path of the signal *before* it reaches R_4 . Hence, less voltage develops across R_4 , and less reaches V_2 . At 30 cycles, the voltage reaching V_2 is reduced 15 db below its value when the arm of R_3 is in mid-position.

The treble control circuit consists of C_1 , C_4 , and R_1 . The potentiometer,

R_1 , has the same type of taper as R_3 so that, when it is in mid-position, only 10% of the 500,000 ohms is between the arm and C_4 . In this position, only 10% of all the high frequencies applied to this network reach V_2 .

When the arm is turned to the top of R_1 , considerably more of the high-frequency signal reaches V_2 . By the same token, when the arm is turned to the bottom of R_1 , very little of this signal reaches V_2 . This is the treble cut position.

There are other versions of this control, but with the help of the foregoing explanation, there should be little difficulty in understanding their operation.

Automatic Tone Compensation. Before we leave the subject of tone controls, there is one commonly used network that provides a fixed amount of tone compensation at the volume control. In Fig. 10.32 a series resistor and capacitor are tapped across the lower section of the volume control. When the volume is turned up high, the tone compensation network has no appreciable effect. At low volume levels, however, when the center arm of R_1 is near the lower end of the volume control, R_2 and C_1 become effective. Their purpose is to attenuate somewhat the amount of middle and high frequencies that reach the succeeding amplifiers. The need for the compensation is due to the fact that at low volume levels the human ear is least sensitive to the low frequencies. Partially to offset this limitation, some of the higher frequency voltage is by-passed by C_1 and R_2 . The aural effect is an apparent increase in low frequencies at low volume.

Many high-fidelity amplifiers extend the range of tone compensation by inserting several compensation networks, each designed for a different level of sound. The circuits are brought into the system by a special selector switch, usually labeled a "Loudness" control. This is discussed more fully in Chapter 12.

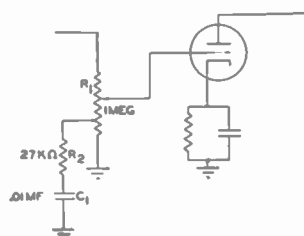


FIG. 10.32. A fixed-tone compensation circuit.

PROBLEMS

1. State briefly what factors must be taken into consideration in any discussion of high fidelity.
2. How do the properties of the human ear affect high fidelity? Describe some of these properties.
3. What have radio manufacturers done in recognition of the frequency and loudness characteristics of the human ear?
4. Describe what is meant by "frequency balance."
5. Draw the diagram of a frequency-dividing network designed to feed signals to special high and low frequency speakers.
6. What are the advantages of push-pull amplifiers over single-ended amplifiers?

7. Must push-pull amplifiers always be transformer-coupled to the previous amplifiers? Explain.
8. Draw the circuit diagram of a push-pull amplifier.
9. What is a phase inverter? Why is it useful?
10. Draw the circuit diagram of a single phase inverter tube driving a push-pull amplifier.
11. Why can a single tube be used to provide two out-of-phase voltages?
12. What are the advantages and disadvantages of using a single tube to drive a push-pull amplifier?
13. Draw the circuit schematic of a single amplifier and a separate phase inverter driving a push-pull amplifier.
14. What is negative feedback?
15. What are the advantages of negative feedback?
16. Describe several methods for achieving negative feedback.
17. What precautions must be observed when using negative feedback?
18. What type of tone control is used on most radio receivers? Explain how it functions.
19. Draw the diagram of an audio amplifier containing a simple bass-boost tone control.
20. Illustrate the differences between bass and treble tone controls.
21. Explain how negative feedback can be used to obtain tone control.
22. Draw the diagram of a circuit using negative feedback to obtain tone control.
23. Explain how the circuit of Fig. 10.30 operates.
24. What is automatic tone compensation? Why is it used? How is it achieved?
25. What is a cathode follower? Where would it be found in audio equipment.
26. Illustrate, with simple diagrams, three different types of audio output circuits.
27. Explain the advantages and disadvantages of the following output amplifiers: triodes; tetrodes connected as triodes; tetrodes with cathode feedback.
28. When are preamplifiers required? What precautions must be taken when preamplifiers are designed?
29. What is meant by equalization? Where is it generally applied in audio equipment?

Chapter 11

F-M RECEIVER ALIGNMENT

The ability of F-M to reduce interference depends to a great extent upon the proper centering of the signal within the bandpass channel of the receiver's tuned circuits. This, in turn, requires that the tuned circuits of the receiver be correctly aligned. When a station is tuned in manually on an F-M receiver, the speaker output gradually increases in volume, accompanied by much more noise than is customarily heard with A-M receivers. However, if the dial rotation is continued, a point is reached where the station program comes through loud and clear with all background noise completely absent. Interstation noise, unless removed by a special noise-limiter circuit, is much greater with F-M than A-M receivers. The reasons for this have been discussed previously.

Two methods of aligning F-M receivers are currently in use. The most popular method, and probably the faster one, uses a signal generator and a VTVM. The second method requires more extensive equipment, namely, an F-M signal generator and an oscilloscope. The latter approach, because of the oscilloscope, is referred to frequently as the visual method. Since both methods are likely to find wide usage, both will be considered in detail. The reader can then choose the one best suited to his purpose.

Commercial F-M receivers can be divided into four categories: first, the familiar limiter and discriminator combination; second, the ratio detector; third, receivers that use the 6BN6 (or 3BN6) gated-beam tube; and, fourth, the 3DT6 or 6DT6 F-M detector. The two latter types have been used, thus far, to a greater extent in the F-M sound section of television receivers than in F-M broadcast receivers. However, since they are F-M detectors, they will be discussed here.

To systematize the alignment presentation for the various types of detectors, we will first examine the alignment of those receivers containing the limiter and discriminator type of detector. Furthermore, since the align-

ment can be achieved by either of the methods mentioned, we will begin with the method requiring the least amount of equipment and then progress to the more extensive visual procedure.

LIMITER-DISCRIMINATOR RECEIVERS

Signal Generator—VTVM Method. For the purpose of alignment, the receiver can be divided into three main sections: the R.F. stages, the I.F. and limiter stages, and the discriminator. As a general rule, the I.F. and limiter tuned circuits are peaked first, and then the discriminator is adjusted. Finally, the R.F. stages are aligned. Although some change in order may occur, such as adjusting the discriminator before the I.F. and limiter stages, this sequence will generally prove most satisfactory.

I.F. System Alignment. To adjust the I.F. stages, the signal generator is placed at the grid of the mixer tube. The VTVM is connected into the grid circuit of the limiter. If two limiter stages are employed and both are tuned, the VTVM is used initially at the grid of the first limiter. On the other hand, if there is no tuned circuit between the two limiters, then the VTVM is placed initially in the grid circuit of the second limiter.

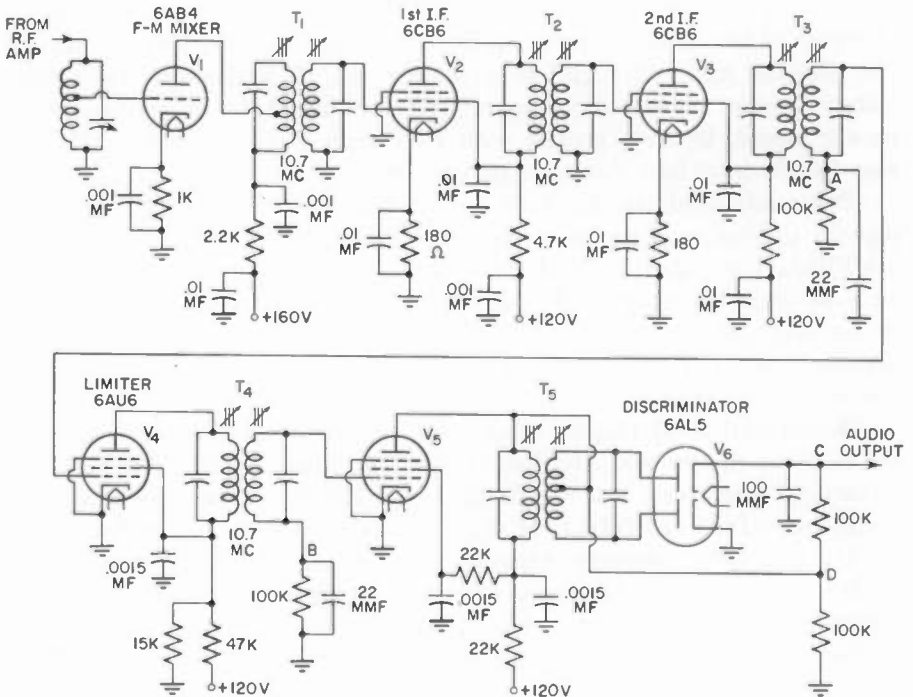


FIG. 11.1. A typical I.F. and discriminator system of an F-M receiver.

The I.F. system of a typical F-M receiver having two limiters is shown in Fig. 11.1. The signal generator is connected to the control grid of the 6AB4 mixer through a 0.01-mf capacitor. The ground lead of the generator connects to the receiver chassis. The VTVM is set to a low d-c scale and the d-c probe is placed at point *A*. The meter common lead goes to the chassis.

With the signal generator accurately set to deliver a 10.7-mc unmodulated signal, and the receiver in operation, the cores of transformers T_1 , T_2 , and T_3 are each adjusted for maximum indication on the VTVM. These adjustments should be made with as low an input signal as possible, consistent with a usable indication on the meter. The stages will thus be aligned for maximum response at low signals. Upon reception of strong signals, a certain shifting in the peak of the limiter will result, producing some distortion. However, strong signals can deal more effectively with interference and distortion, and the effect of this peak shifting is less noticeable than it would be if the misalignment occurred on weak signals.

After the foregoing peaking has been carried out, the d-c probe of the VTVM is shifted to point *B*. Everything else remains the same. Now, both the primary and secondary cores of T_4 are adjusted for a maximum indication on the VTVM. This completes the alignment of all the I.F. stages and the two limiters.

Note that the signal generator is placed at one end of the I.F. system and the VTVM is positioned at the other. Occasionally, when the system is badly out of tune or one of the tuning circuits is defective, it may be necessary to conduct this procedure on a stage-by-stage basis. The VTVM is initially placed at point *A*. The signal generator is then connected to the control grid of V_3 . The generator ground lead still goes to the receiver chassis. With the generator set to 10.7 mc (unmodulated), adjust both cores of T_3 for maximum indication on the VTVM. Next, the signal generator is shifted to the control grid of V_2 , and the cores of T_2 are adjusted in a similar manner. Then, the signal generator is moved to the grid of V_1 , and the cores of T_1 are adjusted.

One further point before we discuss discriminator alignment. When the signal generator is connected to the grid of the mixer tube, its output may be severely reduced because of the presence of the mixer grid tuned circuit. This circuit is resonant to a much higher input signal, and its impedance at 10.7 mc is low. Consequently, a very high setting of the signal generator amplitude control may be required to inject a small amount of signal into the I.F. system.

If this occurs, unsolder the lead to the grid of the mixer tube and substitute a 100,000-ohm resistor from the mixer grid to ground. The signal generator voltage is applied across this resistor.

A question often asked is whether the high-frequency oscillator should be disconnected—or prevented from functioning—during this period of alignment. For clear-cut results, it is better to stop the oscillator. We can remove the oscillator tube, if a separate one is used, or short the plates of its tuning capacitor if the B+ is not thereby grounded out. In crystal oscillators, removal of the crystal itself will prevent operation of the oscillator.

If only one limiter is present in the receiver, the VTVM is connected into its grid circuit. If two limiters are employed and the coupling between them is untuned, the d-c probe of the VTVM goes to the grid of the *second* limiter and remains there throughout the entire alignment of the I.F. system.

Discriminator Alignment. The next step, after the I.F. and limiter adjustment, is alignment of the discriminator. The signal generator *remains* at the mixer grid. However, the d-c probe of the VTVM is moved to point *D*, Fig. 11.1. Common lead of the instrument connects to the receiver chassis.

Adjust the primary core of T_5 for maximum meter indication. Then, move the meter probe to point *C* and rotate the secondary core until a zero reading is obtained. If the core is rotated in one direction from this zero point, the VTVM will indicate a positive reading; if the core is turned in the opposite direction, the meter will indicate a negative voltage. It is important that the zero reading occur between a positive and negative reading

(as the core is rotated). This is the only correct setting for this secondary core.

Since most VTVM's have the zero marking at the extreme left-hand side of the scale, the reading of negative values is not too convenient. For such units, an artificial zero mark can be achieved by rotating its "Zero Adjust" knob until the needle is brought some distance up scale. Let us say it is brought to a marking of 3. This now becomes the new zero. Needle deflection to the right of this value represents a positive movement of the needle deflection to the left of 3 represents negative voltage values.

Meters with a center zero scale can be employed directly.

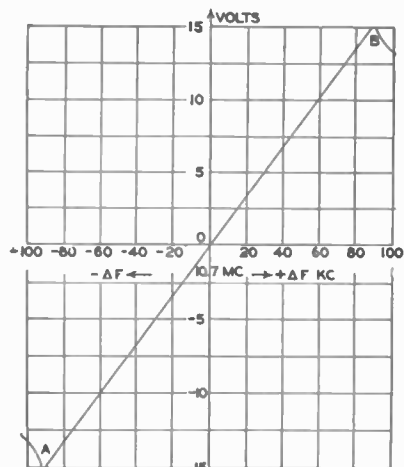


FIG. 11.2. The proper output characteristic of a discriminator.

The foregoing adjustments should cause the discriminator transformer to produce the response characteristic shown in Fig. 11.2. However, a test should be made of the discriminator linearity, at least up to ± 75 kc. Most

manufacturers design the discriminator transformer to be linear up to ± 100 kc, but ± 75 kc will suffice. Vary the frequency of the signal generator to obtain several points such as ± 20 kc, ± 40 kc, ± 60 kc, and ± 75 kc, about the chosen carrier frequency, 10.7 mc. The meter across the load resistors should indicate equal readings (although of opposite polarity) for each set of frequencies. Thus 10.7 mc + 20 kc should produce the same deflection as 10.7 mc - 20 kc. If these tests indicate that the response is not linear, then

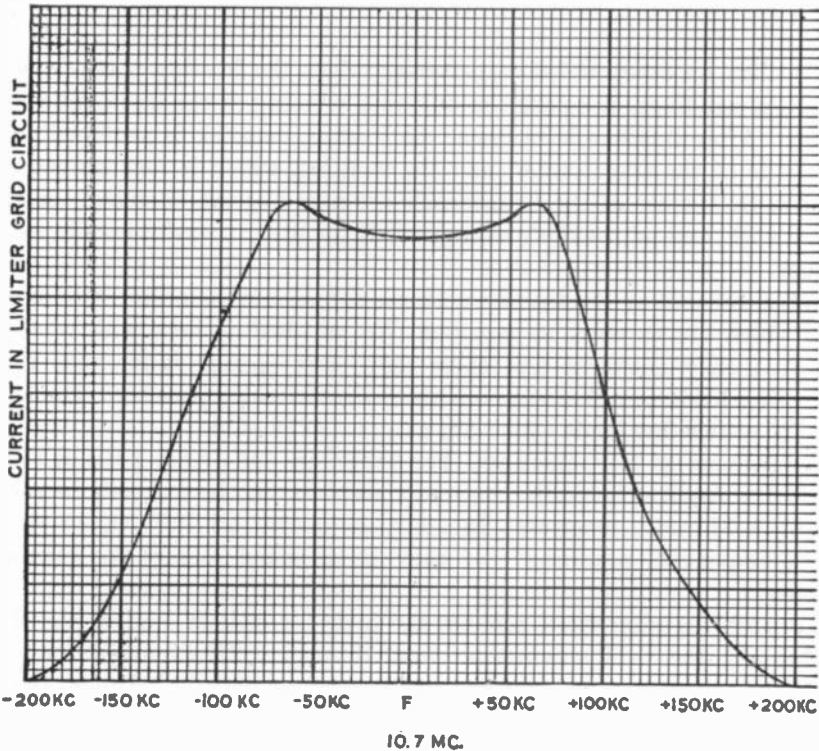


FIG. 11.3. A double-peak response curve for an F-M, I.F. system.

the various cores throughout the I.F. system should be touched up until nearly equal meter deflections are obtained.

Overcoupled I.F. Transformers. There are in use some I.F. transformers which possess a double peak, due to overcoupling, in order to obtain a wide-band characteristic. (See Fig. 11.3.) For these, a slightly modified approach is necessary. When an overcoupled stage is to be aligned with an A-M signal generator, it is necessary to use loading networks. The network loads the circuit so that the transformer is effectively below critical

coupling. In this condition, the transformer can be peaked. The loading network to be used will be specified by the manufacturer and may consist of a single resistor or a resistor and capacitor in series. When aligning a loaded stage, a greater signal is required from the signal generator and the stage will tune broadly. If it is found that the generator does not possess sufficient strength, it will be necessary to increase the size of the loading resistor in order to reduce the effect of the loading.

Where the manufacturer's service data are not available, or the size of the loading resistor is not known or specified, a resistor of 680 ohms may be used.

To align an overcoupled transformer, place the loading network across one of the windings, and peak the unloaded winding at the intermediate frequency. After this has been done, the loading network is switched to the other side of the transformer, and the winding, which is now unloaded, is peaked. Both sides of the transformer are now aligned. Remove the loading network, and adjust the windings of all the other transformers in similar manner.

When connecting the loading network across a transformer winding, use leads which are as short as possible. We are dealing here with relatively high frequency networks and there may exist enough inductance in the connecting leads to affect appreciably proper circuit operation. Hence the need for connecting leads which are as short as possible.

Alignment of R.F. Section. The procedure for aligning the R.F. portion of the receiver is similar, in many respects, to the methods in use for present A-M radio receivers. In addition, nearly all R.F. tuned circuits are single-peaked, requiring that they be adjusted merely for maximum response.

To start the alignment, connect the high (or signal) lead of the signal generator to the F-M antenna terminal through a 270-ohm resistor.¹ The low side of the generator connects to the receiver chassis. (Unless the circuits are completely out of alignment, it is perfectly feasible to adjust the R.F. circuits and the oscillator at the same time.) The indicating meter, a VTVM, is placed in the grid circuit of the limiter. Set the signal generator to some frequency at the high end of the band, say, 106 mc, for the circuit shown in Fig. 11.4.² Adjust the front dial of the receiver to read exactly the same frequency. With the signal generator and receiver both in operation, adjust trimmer capacitors C_1 , C_2 , and C_3 for maximum deflection of the meter. Do this carefully and use as low a signal as is consistent with readable meter indications.

¹ If a balanced input is employed, the signal generator is connected to the antenna terminals with a 150-ohm resistor in each lead.

² This frequency will vary with the receiver.

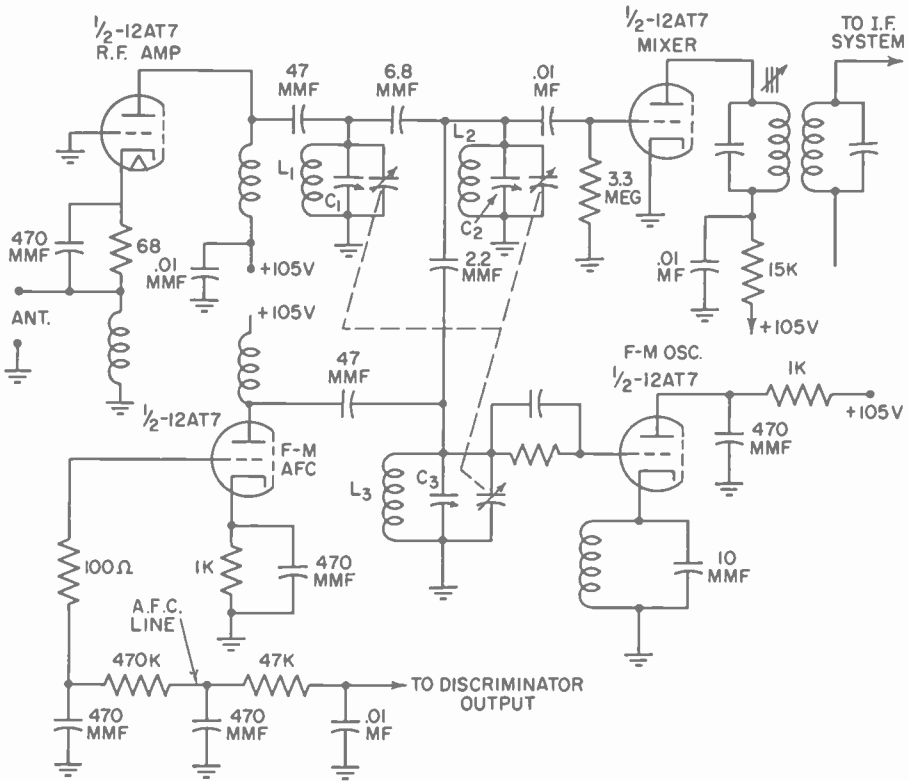


FIG. 11.4. The R.F. section of an F-M receiver.

The circuit in Fig. 11.4 possesses an automatic frequency control (A.F.C.) circuit. This is designed to counteract any frequency drift that may develop in the F-M oscillator while a station is being received. In the present instance, we wish to adjust the oscillator's frequency without any control action by the A.F.C. network. Therefore, this stage is disabled. Generally, receivers which possess A.F.C. circuits have a front panel disabling switch and the cut-out is accomplished this way. If, by chance, this is not true, then the stage may be disabled by simply grounding the A.F.C. line which regulates the operation of this stage. See Fig. 11.4.

The next step is to reduce the signal generator frequency to 90 mc (for the circuit in Fig. 11.4) and to set the receiver dial to the same value. Now, L_1 , L_2 , and L_3 are adjusted for maximum meter deflection. Since these coils have no slugs, their inductance is altered by compressing or expanding the coil turns until the desired maximum reading is obtained. After this has been done, the entire R.F. alignment is rechecked to make certain optimum peaking has been achieved.

If coils L_1 , L_2 , and L_3 do possess internal movable cores, then these would be adjusted in the foregoing step instead of the physical compression and expansion.

The R.F. adjustment frequencies of 106 mc and 90 mc are peculiar to the circuit of Fig. 11.4. Other receivers may use other frequencies and, for each, the manufacturer's instructions should be followed.

This now completes the R.F. alignment of the receiver.

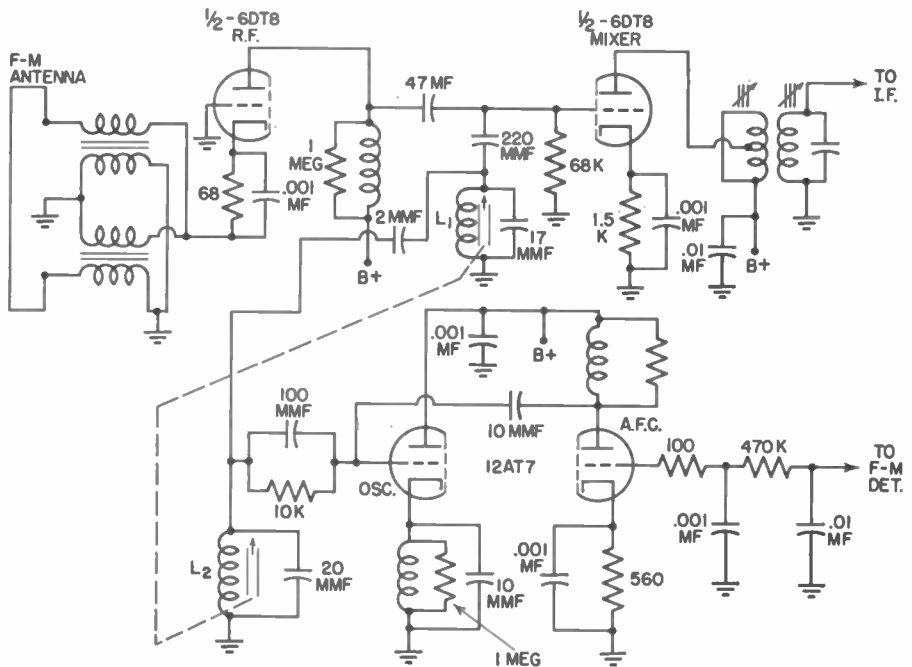


FIG. 11.5. An R.F. front end section in which inductive tuning is employed.

When the R.F. stages employ inductive tuning, as they do in Fig. 11.5, the adjustments to be made are slightly modified. There are no trimmers to adjust, the only movable items are the tuning cores (in L_1 and L_2) themselves. Therefore, adjustment is made on these cores. The signal generator, connected to the antenna terminals, is set to 98 mc. This is the mid-frequency of the F-M band. The receiver dial is also set to 98 mc. For indication, a VTVM is connected to the grid of the limiter. Then, the tuning cores are individually adjusted for maximum reading on the VTVM. Provision exists for this to be done. The adjustments are performed several times to make certain the best position is obtained. Then the cores are again secured to the mechanical tuning mechanism so that they can be varied in unison.

Visual Alignment. In the visual method of circuit alignment, the response curve of the circuit under test is produced directly on the screen of an oscilloscope. This is achieved by feeding the output of an F-M signal generator into the receiver. This signal sweeps back and forth across a band of frequencies 0.5 to 1 mc wide. The center of this band is placed at 10.7 mc, which means that the generator output varies from 10.2 mc to 11.2 mc. As the signal passes through the tuned circuits, it is modified in amplitude by the selectivity characteristics of the tuned circuits. At the limiter, the grid bias will vary in accordance with these amplitude variations, and it is this pattern which is presented on the oscilloscope screen.

By this method of alignment, results of any adjustments may be viewed immediately. In this sense, this method is superior to the previous one where we really do not know exactly what the final response curve looks like.

The Oscilloscope. The oscilloscope, shown in Fig. 11.6, is a voltage indicator. Hence, in order to obtain any indications on the screen, the plates (or the input terminals) must be placed across two points in a circuit where a difference of potential exists. If we desire to determine the form of the current wave, it becomes necessary to pass the current through a resistor and then to obtain the waveform of the voltage across the resistor. Only in this way is it possible to determine the current variations present.

For alignment of an F-M receiver, we still use the points indicated previously. In the grid circuit of the limiter, we obtain the necessary voltage for the oscilloscope from across the grid resistor. At the output of the discriminator, the other indicating point used, the substitution of the oscilloscope introduces no problem.

In connecting the oscilloscope, the vertical input and ground leads are placed across the two points whose voltage waveform is to be viewed. At the points mentioned, one end is usually at ground potential, and it is best to connect the ground terminal of the scope to the grounded point. This



FIG. 11.6. An oscilloscope suitable for aligning and servicing F-M circuits. (Courtesy Simpson Electric Co.)

places the scope ground at the same potential as the set ground and avoids shock. If both points at which the voltage is to be measured are above ground, care will have to be exercised that one does not come in contact with the case of the oscilloscope, this usually being at oscilloscope ground. Remember that the so-called "ground" of one system does not necessarily have to be at the "ground" of another system, unless both are directly connected with a conductor.

In order to use the oscilloscope to its fullest extent, it is essential to have a good understanding of the type of indications that will appear. Unless the proper interpretations are made from the observed results, very little benefit will be derived from its use.

We know, from the previous discussion, that an S-shaped curve represents the characteristic response of the discriminator. The usable region, between points *A* and *B* (Fig. 11.2), must be made linear if amplitude distortion is to be avoided during the F-M demodulation process. Beyond the two points the characteristic may curve considerably without affecting the response in any way.

With an F-M signal generator sweeping over the entire band, the full S-shaped curve will be obtained on the screen of the scope. If the entire I.F. system is properly aligned, the signal generator may be placed at the grid of the mixer. If the discriminator is aligned before the other stages, the signal generator is generally placed at the grid of the limiter preceding the discriminator. The vertical input terminals of the scope, for discriminator alignment, connect between point *C* and ground (Fig. 11.1).

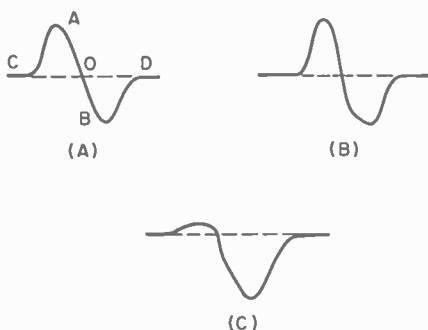


FIG. 11.7. Discriminator response curves. (A) Desired form; (B) and (C) distorted form.

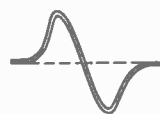


FIG. 11.8. A double pattern.

With the F-M signal generator in operation, the curve that should be seen on the oscilloscope screen is shown in Fig. 11.7A. The extent of this curve is greater than that shown in Fig. 11.2 and is obtained by using a

half-megacycle sweeping range. This is done purposely so that the two end sections, *C, D*, will serve to form a zero line. If we then mentally connect sections *C* and *D* by a straight line, we can tell more easily when section *A-O* is equal in length to section *O-B*.

If the secondary of the discriminator input transformer is properly aligned, but not the primary, the linearity of the curve between points *A* and *B* will be distorted, perhaps producing the result shown in Fig. 11.7B. The primary winding controls the linearity of the response, and it should be adjusted until the operating portion of the curve is linear.

Misalignment of the secondary, with the primary winding properly adjusted, shifts the cross-over point (point *O*, in Fig. 11.7A) from its normal position (see Fig. 11.7C). Adjustment is required until both sections of the curve are symmetrical with respect to the center point.

Although we require but one S-shaped curve, we sometimes get two. (See Fig. 11.8.) To understand the reason for this, it is necessary to know how the F-M sweep signal is produced and how the oscilloscope operates in conjunction with the generator. In an F-M signal generator, of the type likely to be employed for the present purpose, the 60-cycle sine-wave input from the power line is used to frequency-modulate an oscillator. It may do this through a reactance tube (to be described later) or it may do it by some other means. As a result, the frequency variation that appears at the output of the generator varies back and forth at a 60-cycle rate.

At the same time, a portion of this sine-wave voltage is brought to an outlet at the signal generator panel. From this point it is connected to the horizontal input terminals at the oscilloscope by means of a cable. In other words, we replace the sawtooth sweep of the indicating scope with this 60-cycle sine-wave voltage.

A sawtooth voltage, shown in Fig. 11.9, when applied to the horizontal

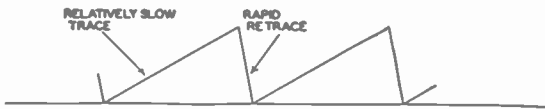


Fig. 11.9. The application of a sawtooth wave to the deflecting plates of a scope causes a relatively slow left-to-right motion and rapid retrace.

deflection plates of an oscilloscope, will cause the electron beam to move slowly from left to right and then, at the retrace, to swing rapidly back to the left-hand side of the screen. With a 60-cycle sine-wave deflecting voltage, such as that obtained from the foregoing F-M signal generator, the action of the beam is entirely different.

Sine-Wave Deflection. In Fig. 11.10 we have the familiar sine wave, with letters from *A* to *E* inserted at convenient points to aid in the explanation. At point *A*, the voltage is zero and, if applied to the horizontal deflection plates, would not affect the beam. Hence, when the sine wave of Fig. 11.10 is applied to the deflecting plates, and the voltage is zero, the electron beam will pass through the center of the deflecting system unaffected and strike the fluorescent screen at its center.

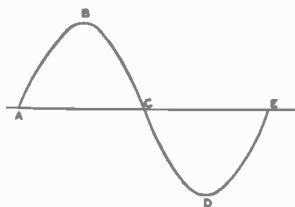


FIG. 11.10. A sine wave.

From *A* to *B*, the voltage is increasing, and the beam is deflected to one side, say, the right. From *B* to *C*, the deflecting voltage is decreasing, and the beam is gradually returning to the center of the screen. It arrives at this position when point *C* is reached.

From *C* to *D*, the voltage has reversed itself, and the beam is now shifting to the opposite side of the screen. At *D*, the maximum left-hand position has been reached. From *D* to *E*, the beam returns to the center of the screen again.

Now let us apply the output from the signal generator to the grid of the last limiter tube. The generator output is varied by the same 60-cycle sine wave. This means that the output frequency starts from its center position (which is indicated by the dial on the front panel), shifts to a higher frequency, reverses and returns to its center position, drops to a lower frequency, and finally returns to the carrier or resting position to end the cycle.

When the F-M signal is applied to the discriminator circuit, you will find, by combining the two actions described above, that one S-shaped curve is traced out on the left-to-right motion of the beam, and a similar S-shaped curve is traced out as the beam travels from right to left on the screen. Remember that with this signal generator we are using a sine-wave sweep; the sawtooth sweep of the oscilloscope itself is in the off position and without effect. However, although the voltages for the beam deflection and the sweep generator oscillator are obtained from the same source, it does not necessarily follow that these voltages are still in phase with each other by the time they reach the beam or the modulating circuit. To bring them into phase, a special control, the phase control, is incorporated into the sweep generator. By varying this control, the two portions of Fig. 11.8 can be blended into one.

Sometimes, the two curves can be brought together at all points except perhaps at the extreme ends. This is caused by some circuit unbalance and can be ignored.

The Howard W. Sams organization recommends a double "S" pattern in

their alignment instructions. (See Fig. 11.11.) This is obtained by using a 60-cycle sine-wave sweep in the F-M signal generator and a 120-cycle sawtooth voltage for deflection in the horizontal system of the oscilloscope.

To see just how this double pattern is obtained, let us follow the output frequency of the F-M signal generator and the associated motion of the electron beam in the oscilloscope. This is shown in Fig. 11.12. At time *A*, when the beam is at the extreme left-hand side of the scope screen, the output frequency of the signal generator will be 10.8 mc. (We are assuming here a frequency sweep of +100 kc and -100 kc about 10.7 mc.) Since the output of the discriminator at 10.8 mc will be maximum, by Fig. 11.2, the beam will be shifted to point *A* in Fig. 11.12C. This represents the response of the discriminator to 10.8 mc.

As the beam moves across the screen from left to right, the frequency output of the F-M signal generator decreases until it becomes 10.7 mc when the beam is in the center of the screen. This is time *B* and the section of the response curve traced out is *A-B* of Fig. 11.12C. (The dotted arrows show the manner in which the curve is traced out.)

From time *B* to time *C*, the frequency output of the generator decreases to 10.6 mc and section *B-C* of the response curve is traced out. Point *C* (or time *C*) occurs when the beam is at the extreme right-hand edge of the screen. If we assume that the beam loses no time in snapping back to the left-hand side of the screen (i.e., retrace), then point *C* also appears at the left, as shown in Fig. 11.12C. Actually, some small interval is required for this action to occur, but we will ignore it here.

From time *C* to time *D*, the beam is again traveling from the left-hand side of the screen toward the center and the frequency is changing from 10.6 mc to 10.7 mc. During this interval, section *C-D* of the response curve is traced out.

The final section, *D-E*, is then traced out from time *D* to time *E*.

The advantage of this type of presentation is its symmetry. If the reader adjusts the F-M discriminator until he obtains a linear "X" pattern, as shown, he then knows that the stage is properly adjusted. The crossover point must lie in the center of the pattern.

The difference between the patterns of Fig. 11.11 and Fig. 11.12C is the greater sweep used to obtain the pattern of Fig. 11.11. For Fig. 11.12C, a sweep of ± 100 kc was indicated. For Fig. 11.11, a sweep of ± 225 kc was used. This additional sweep range provides the flat end sections where the circuit response is zero because the signal is outside the range of the tuned circuit. These end sections further aid in lining up the center of the "X."

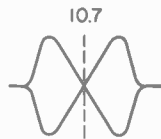


Fig. 11.11. The F-M detector response curves obtained by using a 60-cycle sine-wave sweep in the F-M signal generator and a 120-cycle sawtooth voltage for beam deflection in the oscilloscope.

If any reader wonders how it is possible to cause the output of the sweep generator to be at the maximum end when the beam is at the extreme left-hand side of the screen, the answer is with the phase control on the generator. This is adjusted until the pattern shown is obtained.

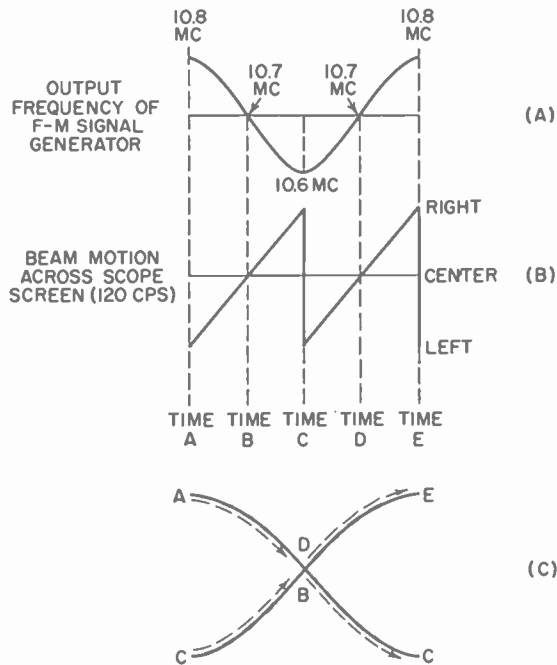


FIG. 11.12. How the pattern of Fig. 11.11 is produced.

Visual Discriminator Alignment. To align the discriminator input transformer visually, connect the output terminals of the signal generator between grid and ground of the last limiter tube. A 0.01-mf capacitor should be inserted between the signal lead of the generator and the grid of the limiter to prevent the generator from upsetting the d-c voltage at the limiter grid. Attach a wire from the "Vertical" post of the oscilloscope to terminal C of Fig. 11.1. The other end of the discriminator load, usually ground, is connected to the ground terminal of the oscilloscope. In addition, an outlet is provided on all F-M signal generators from which a portion of the 60-cycle modulating voltage can be obtained. The voltage is applied to the "Horizontal" input terminal of the oscilloscope. The ground terminal of the generator is connected to the oscilloscope ground post. The internal sawtooth sweep of the scope is turned off.

If a pattern such as that shown in Fig. 11.11 is desired, then no connec-

tion is made from the signal generator to the oscilloscope. Instead, the latter is set to produce a horizontal sawtooth sweep of 120 cycles per second.

Set the generator to the I.F. value, say, 10.7 mc, and adjust the sweep control so that the output frequency sweeps for about 450 kc about 10.7 mc. This will produce the complete response pattern on the screen plus a good indication of the circuit response beyond the desired spread about 10.7 mc. The primary and secondary cores of the discriminator coil are adjusted now until the desired curve is obtained. This curve is shown in Fig. 11.7A. When this curve has been obtained, the discriminator circuit is aligned.

Visual Alignment of I.F. Stages. For the visual alignment of the I.F. system, the vertical input leads of the oscilloscope are placed across the grid-leak resistor in the grid circuit of the first limiter stage (if two are used). The signal generator is connected between control grid and ground of the mixer tube. Again, use the 0.01-mf capacitor in the signal lead of the generator. Adjust each of the cores of the I.F. transformers (T_1 , T_2 and T_3 in Fig. 11.1) for maximum amplitude and symmetry of the curve shown in Fig. 11.13. Then, move the oscilloscope to the grid of the second limiter (point *B* in Fig. 11.1) and adjust the two cores of T_4 for maximum amplitude and symmetry of the response curve. The level of the input signal is kept as low as possible in order not to drive the limiter beyond saturation.

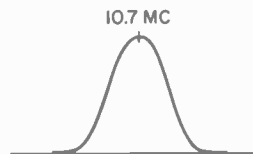


FIG. 11.13. The normal response of an F-M I.F. system.

The foregoing completes the visual alignment of the I.F. and discriminator stages.

R.F. System Alignment. Alignment of the R.F. stages is achieved as follows: Attach the signal generator to the antenna terminals of the receiver, leaving the oscilloscope connected across the grid-leak resistor of the limiter stage. Just how to connect the signal generator leads to the receiver will generally be specified by the manufacturer. Philco, for example, states that two simple dipole aeri-als should be constructed using 30-inch lengths of rubber-covered wire. One dipole is connected to the receiver antenna terminals and one dipole is connected to the output terminals of the signal generator. The two dipoles are then spaced several feet apart and the signal transfer affected in this manner. Other manufacturers specify that the signal generator output lead should connect to the receiver antenna terminal through series resistor. The value frequently chosen is 270 ohms. The other generator lead goes to the receiver chassis.

The sweeping range of the generator is set for approximately 1 mc and the center output frequency adjusted to 108 mc. Set the dial of the receiver to the same frequency. Now adjust the oscillator and input trimmer capacitors (as indicated by the manufacturer) until the response curve seen at the os-

cilloscope is symmetrical and at a maximum. Note that, since the oscilloscope is placed in the limiter grid circuit, the signal generated by the signal generator will first have to pass through the I.F. system before it reaches the oscilloscope. Hence, the response pattern seen on the oscilloscope will be that of the I.F. system. (See Fig. 11.14A.) However, if the R.F. stages

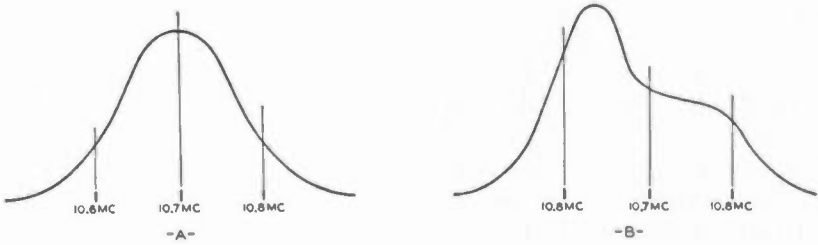


FIG. 11.14. (A) The response characteristic of a correctly aligned I.F. system. (B) A distorted response indicating circuit misalignment.

are not properly adjusted, the response pattern will become distorted, one possible shape shown in Fig. 11.14B. The R.F. stages are correctly adjusted when the I.F. response curve observed on the oscilloscope screen possesses its proper form with a maximum amplitude.

The same procedure is repeated at 88 mc. Practically all R.F. coils are of single-peak design, and the foregoing method of adjusting for one maximum response at each test point is sufficient.

Marker Frequencies. In using the oscilloscope, it would aid considerably if the frequency at specific points on the curve seen on the screen were identified in some manner by special markers. Such indicators are known as "marker frequencies" or "marker points."

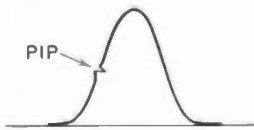


FIG. 11.15. The marker pip as it appears on the response curve.

Some sweep generators have built-in marker generators and it is possible to inject the marker signal into the sweep signal directly in the instrument. The marker dial is calibrated and can be set to the desired frequency. Let us say that it is 10.6 mc. Then, wherever the marker pip appears on the response curve represents 10.6 mc. (See Fig. 11.15.)

For visual alignment of the R.F. circuits, the F-M signal generator is connected to the antenna terminals. Assume that the initial alignment is to be made at 108 mc. To produce a 10.7-mc marker on the response curve (obtained at the grid of the limiter), the marker generator must inject a 108-mc signal into the sweep signal. After this passes through the mixer, the 108-mc signal is converted down to 10.7 mc.

There is still another method whereby we may obtain a marker indication at any point, but it requires the use of an A-M signal generator. This second generator may be connected in parallel with the sweep (or F-M) generator or it may be placed somewhere along the path through which the signal from the sweep generator must pass. The second generator is then set at the frequency at which we wish the marker indication to appear. The result on the screen of the oscilloscope will be a small pip or wiggle at the proper point in the characteristic curve.

As an illustration, suppose the I.F. system of the receiver is being aligned. Connect the F-M signal generator across the grid circuit of the mixer stage and, at the same time, connect the generator that will supply the marker frequency in parallel with the F-M generator. (Place an isolating resistor of 270 ohms in series with the signal lead from the A-M signal generator. This will keep interaction between the two generators to a minimum. Also keep the output of the A-M generator as low as possible while still being able to observe a wiggle in the response curve.) Let the F-M generator sweep over the band in the usual manner. Set the A-M generator—which develops a single signal at each dial setting—to 10.7 mc. The oscilloscope, connected into the grid circuit of the limiter stage, will indicate a wiggle at the point where 10.7 mc is located. The response curve should peak at this point. If it does not, the several tuning circuits are adjusted until it does.

The marker indication is also good when aligning the discriminator. It may be used to indicate the two end points of the S-shaped discriminator curve plus the cross-over or central point. These are just a few of the uses of the marker points. The serviceman, in dealing with the equipment, will undoubtedly find many more.

ALIGNMENT OF F-M CIRCUITS CONTAINING THE RATIO DETECTOR

To align F-M receivers that do not use the Foster-Seeley discriminator but have another type of detector, such as the ratio detector, for example, it is necessary to modify the process somewhat. Furthermore, ratio detectors fall into two categories, balanced and unbalanced, and this introduces slight variations in the alignment procedure. We will start first with the unbalanced ratio detector, using an A-M signal generator and a vacuum-tube voltmeter.

An unbalanced ratio detector is shown in Fig. 11.16. Connect the output lead from the A-M signal generator through a 0.01-mf capacitor to the control grid of the I.F. amplifier tube just preceding the detector. Set the signal generator to the I.F. value. This signal should be unmodulated. Connect the vacuum-tube voltmeter between point A and ground. This connects the meter across the 25-mf capacitor and the 15,000-ohm resistor.

With the equipment thus set up, adjust the primary core of T_1 until

quencies which are equally above and below 10.7 mc. The procedure is identical to the one discussed with Foster-Seeley discriminators.

After this has been done, the two 100,000-ohm resistors are removed from the circuit.

To align the I.F. system shift the A-M signal generator to the grid of the converter or mixer tube. Set the generator to the I.F. frequency. The vacuum-tube voltmeter is connected between point *A* and ground of the ratio detector. Now adjust the primary point secondary cores of all I.F. transformers for maximum meter indication. If the transformers are overcoupled, use a loading network as previously outlined.

It is possible to align the I.F. system first and then the ratio detector. In this case, the signal generator is left at the grid of the mixer for the entire operation. There is no need to shift it to the control grid of the I.F. amplifier preceding the detector for the latter's alignment.

To align the R.F. sections of the receiver, shift the signal generator to the antenna terminals of the set (using an appropriate matching network as indicated by the manufacturer). The vacuum-tube voltmeter remains connected between point *A* and ground. Set the generator and the receiver dial to 108 mc and adjust the proper R.F. trimmers or cores for maximum meter reading. Next, change the generator frequency and the receiver dial to 88 mc (or thereabouts) and adjust the proper trimmers or cores (as indicated in the manufacturer's service manual) for maximum indication on the meter. The set alignment is now complete.

In reading through the service literature that is put out by manufacturers, variations of the foregoing method will be found. Thus, some manufacturers suggest that an output meter be placed across the speaker voice coil instead of the vacuum-tube voltmeter mentioned previously. Since the signal must now pass through the audio amplifiers in order to reach the output meter, the I.F. signal from the generator must be amplitude-modulated. Practically all A-M signal generators contain provisions for modulating their outputs with a 400-cycle audio note, hence this stipulation presents no difficulty. However, since a well-aligned ratio detector does not readily pass an amplitude-modulated signal, complete alignment of the detector is left for the last. Here is the sequence of alignment.

Connect an output meter across the voice coil of the speaker and connect the signal generator to the grid of the tube just preceding the detector. Set the generator to the I.F. value (10.7 mc) and switch on the modulation. Now, with the secondary trimmer or core of the discriminator coil tuned as far out as it will go, adjust the transformer primary for maximum meter reading. This brings the primary winding into alignment. The secondary, however, is still far out of alignment and this is done purposely in order to permit the detector to respond to A-M signals.

The I.F. and then the R.F. sections of the receiver are aligned next, at all times modulating the carrier signal with the 400-cycle (or whatever other low frequency note the generator contains). All adjustments in these two sections of the set are made for maximum meter indication.

After these circuits have been adjusted, the signal generator is moved back to the control grid of the tube preceding the ratio detector and set at the I.F. value. The secondary of the detector transformer is now adjusted for minimum reading on the output meter. At this point the A-M rejection properties of the detector are best, and the circuit is in alignment.

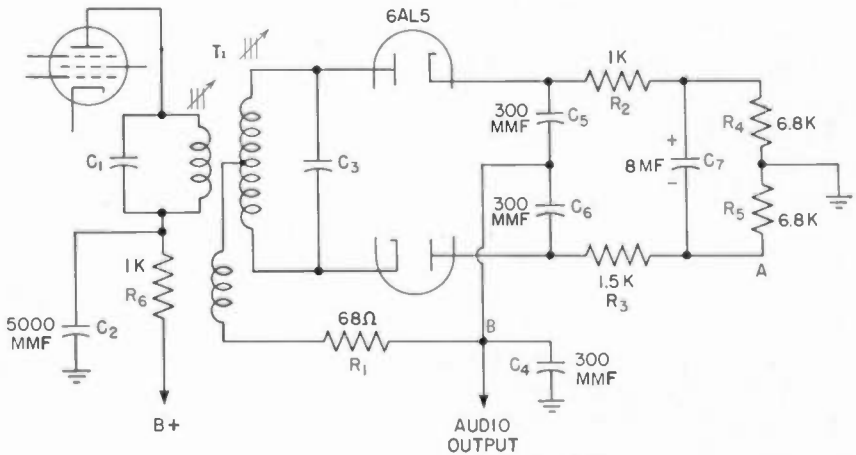


FIG. 11.18. A balanced ratio detector.

Balanced Ratio Detectors. When a balanced ratio detector (Fig. 11.18) is encountered, the alignment procedure is modified only when the detector circuit itself is being adjusted. To adjust the detector, we proceed as follows: Connect the A-M signal generator to the control grid of the last I.F. amplifier. Set it to the I.F. value (say, 10.7 mc). Connect input lead from a vacuum-tube voltmeter to point A, Fig. 11.18. Attach the common lead to the receiver chassis. Now, adjust the primary core of T_1 for maximum meter deflection. Next, move the input voltmeter lead to point B and adjust the secondary core of T_1 for zero meter reading. Check the linearity of the detector response as discussed with the unbalanced ratio detector.

Visual Alignment. Visual alignment of receivers employing the ratio detector is best accomplished by first adjusting the ratio detector and then adjusting the I.F. and R.F. circuits. The ungrounded vertical input terminal of the oscilloscope is connected through a 10,000-ohm resistor, to point B. (Point B, Fig. 11.19, represents the audio output terminal of the detec-

The purpose in removing the 5-mf electrolytic capacitor, as indicated in the foregoing paragraph, is to permit the voltage variations caused by the F-M signal passing through the I.F. system to appear at point A. When viewed on the oscilloscope screen, these variations represent the response curve of the I.F. system.

Sweep alignment of the R.F. stages of an F-M receiver is seldom required. Simple peaking using an A-M generator, as previously described, is the method generally recommended by receiver manufacturers. However, if sweep alignment of the R.F. stages is desired, it is carried out as follows:

Connect the sweep generator to the antenna terminals of the receiver. The oscilloscope is connected to point A, Fig. 11.19. The sweeping range of the generator is set for approximately 1 mc, and the center output frequency is adjusted to 108 mc. Set the dial of the receiver to the same frequency. Now adjust the oscillator and input trimmer capacitors (as specified by the manufacturer) until the I.F. response curve possesses maximum amplitude and is symmetrical about 10.7 mc. However, if the R.F. stages are not properly adjusted, the response pattern will become distorted. The R.F. stages are correctly adjusted when the response curve possesses its proper form with a maximum amplitude.

The same procedure is repeated at 88 mc, completing the front-end alignment of the set.

Alignment Procedure for Beam-Gated Tubes. The alignment procedure for F-M receivers using the 6BN6 (or 3BN6) beam-gated tube differs in many respects from the alignment procedure employed for sets having one of the other detectors. The first difference occurs in the alignment of the I.F. stages themselves. Since the 6BN6 removes amplitude variations, and therefore cannot be transformed into an A-M detector, it becomes impossible to place either a vacuum-tube voltmeter or an oscilloscope at

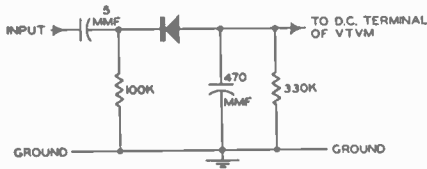


FIG. 11.20. Circuit of the probe required in the alignment of F-M receivers employing the beam-gated tube.

any point in this circuit and obtain the proper indications for peaking the I.F. coils. Hence, to align the I.F. system, a special probe detector must be constructed. A suggested circuit of a probe is shown in Fig. 11.20, using a germanium crystal. When constructing this probe, be careful to keep the length of all connecting leads between components as short as possible.

To align the I.F. system, connect the input lead of the probe to grid 1 of the 6BN6. (See Fig. 11.21.) The output lead connects to the d-c scale of a vacuum-tube voltmeter, while the other output lead connects to the ground or common terminal of the meter. This same lead also connects to the re-

ceiver chassis. An A-M signal is connected, in turn, to each of the control grids of the I.F. amplifier tubes, and the transformer associated with that stage is adjusted for maximum meter indication. This is done for each I.F. amplifier in exactly the same manner that it is done in sets using the ratio or Foster-Seeley detectors, always using as low a signal as possible.

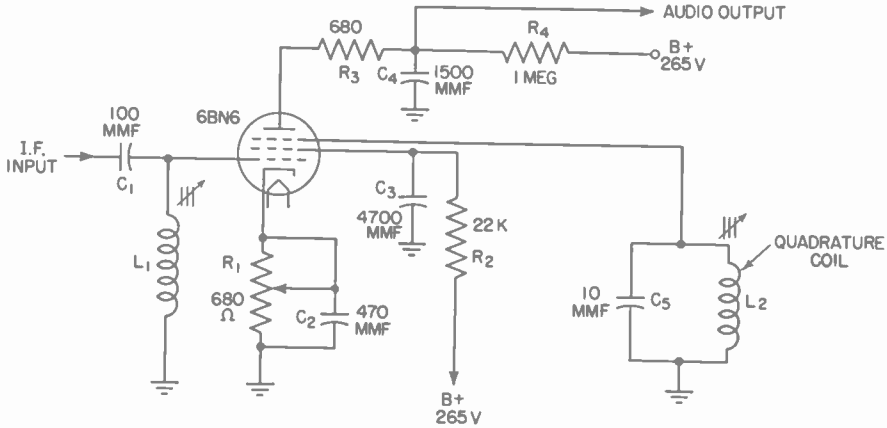


Fig. 11.21. The beam-gated tube connected as a limiter-discriminator.

After the I.F. system is aligned, the signal generator is moved to the input terminals of the receiver, and the R.F. section is adjusted for maximum indication on the probe meter. If visual alignment is desired for either the R.F. or I.F. sections of the receiver, the output of the probe detector connects to the vertical input terminal of the oscilloscope, and an F-M sweep generator is used.

To adjust the discriminator, the probe detector with its meter is left connected to grid 1 of the 6BN6. The signal generator, however, is removed, and the regular set antenna is now connected to the receiver and a station tuned in. The set should be tuned for maximum deflection on the probe meter, thereby indicating that the station is properly tuned in. As an added precaution, the dial should be rocked back and forth until the meter deflection is maximum. The volume control should be turned completely counter-clockwise so that no sound is heard from the speaker.

When the station is correctly tuned in, the volume control is advanced until speech or music is heard, and the quadrature coil of the 6BN6 is adjusted for best sound from the speaker.

There still remains the adjustment of the variable resistor in the cathode leg of the tube. The function of this resistor, it will be recalled, is to provide the proper grid bias for the limiter, in order that the best limiting action (for A-M rejection) be obtained. A method of adjusting this control which is

both simple and satisfactory is to place the set in operation just as it normally would be and then tune in a signal of medium strength. Insert a resistive attenuator pad between the antenna and the receiver and reduce the received signal to a level where hiss is heard with the sound. Now adjust the cathode control of the 6BN6 for clearest sound and the least amount of hiss. During the adjustment, the hiss may disappear. If this occurs, the input signal must be further reduced so that the hiss never disappears during alignment.

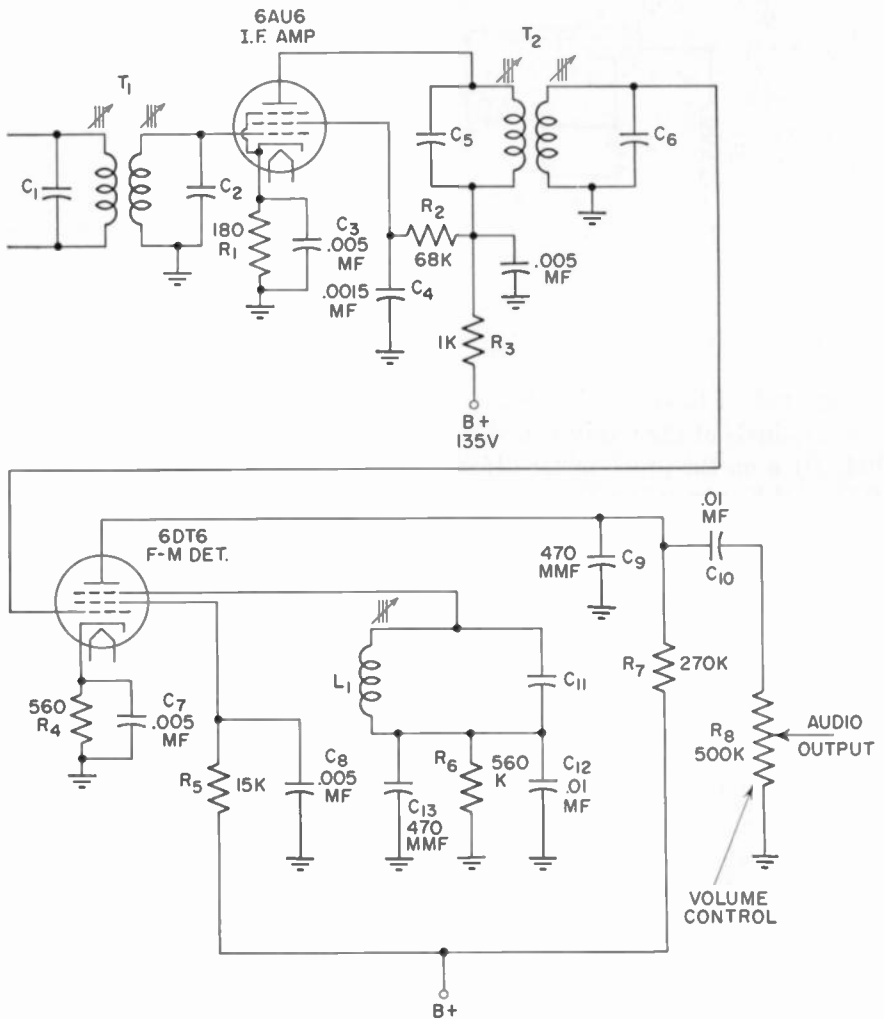


FIG. 11.22. An I.F. amplifier and 6DT6 F-M detector circuit.

It is possible to align a 6BN6 F-M system without test instruments if the tuned circuits are not too far out of alignment. The procedure is as follows. Tune in a station and then use a resistive antenna attenuator to decrease the incoming signal until a hiss is heard in the sound. Now, adjust the I.F. and R.F. tuned circuits, including L_2 and R_1 (Fig. 11.21) for maximum clear sound and minimum hiss. If the hiss disappears during alignment, increase alignment until it reappears.

6DT6 F-M Detector. Although this type of F-M detector circuit (Fig. 11.22) is similar in many aspects of its operation to gated-beam tube detectors, there are certain differences, arising chiefly from the extreme sensitivity of the circuit. Here is a typical procedure, as recommended by Motorola for the above circuit.

Best alignment is achieved using a received station signal instead of a generated signal. Since the detector is extremely sensitive, a relatively small input signal will cause grid current to flow in both the I.F. amplifier and the detector stages. Grid current through the tuned coils will load them down, making the adjustment quite broad and alignment impossible. For this reason it is necessary to use a very weak signal when aligning the driver and the detector input coils. Actually, the signal should be well down into the noise level for proper tuning action.

Procedure (for Strong Signal Areas):

1. Connect the negative lead of the VTVM to the junction of R_6 (560K) and the bottom of L_1 . Connect the positive meter lead to chassis ground.
2. Tune in a station.
3. Set the volume control for average usable sound amplification.
4. Adjust the quadrature coil (L_1) for maximum negative reading on the VTVM.

There are two points of tuning for the quadrature coil—one of which is incorrect. The correct tuning point will produce approximately $2\frac{1}{2}$ volts. The incorrect tuning point will produce approximately $1\frac{1}{2}$ volts. Severe misalignment of the sound I.F. and F-M detector grid coils will reduce the value of this tuned voltage. If this occurs, tune for maximum negative reading on the VTVM; later adjustment of the input coils will produce the $2\frac{1}{2}$ volts.

After the correct tuning point has been established, make the final adjustment of the quadrature coil based on minimum sound distortion. Make no further adjustments of the quadrature coil during the remainder of the alignment.

5. Reduce the signal input at the antenna (disconnect one or both leads and separate from the receptacle or insert resistors) until the sound has been weakened considerably (hiss should be heard).

6. Adjust the primary and secondary of T_2 for best signal-to-noise ratio as determined by listening to the sound. If the signal is too strong, exact tuning will be difficult. (Cores of transformer must be turned as far from each other as possible so that the cores are just entering the coils.)

7. Adjust transformer T_1 for best signal-to-noise ratio as determined by listening to the sound output. If the signal is too strong, exact tuning will be difficult. If there are any other I.F. transformers, adjust their cores in similar fashion.

8. Readjust for best possible signal-to-noise condition.

9. If considerable alignment was required to complete the foregoing procedure, it would be advisable to recheck the tuning of the quadrature coil using a strong signal as in Step 4. However, if the quadrature coil is realigned, it will be necessary to repeat Steps 5, 6, and 7 for tuning of T_1 and T_2 , using a weak signal.

Procedure (for Weak Signal Areas).

1. Connect the negative lead of the VTVM to the top of R_6 . Connect the positive meter lead to chassis ground.

2. Using maximum available signal input, roughly align the cores of T_1 , T_2 , and L_1 for maximum quadrature grid bias (meter reading) of $1\frac{1}{2}$ volts. (See note under Step 4 of the procedure for strong signal areas.)

3. Using maximum available signal, align the quadrature coil (L_1) for minimum sound distortion.

4. Using the weakest signal possible, adjust the primary and secondary of T_1 for best signal-to-noise conditions. (Turn cores of transformer as far from each other as possible so cores are just entering the coils.)

5. Using a weak signal, adjust transformer T_1 for best signal-to-noise ratio. (Turn core as close to chassis metal as possible.)

6. Repeat the procedure several times, if required, until the optimum adjustment is obtained. Keep in mind that the I.F. amplifier and detector input coils must always be readjusted on a weak signal if the quadrature coil setting is changed.

PROBLEMS

1. What two methods can be employed to align F-M receivers? Can the same methods be used with A-M receivers? Explain.

2. What similarities exist between the basic alignment procedure for A-M and F-M receivers?

3. How is a Foster-Seeley discriminator aligned, using an A-M signal generator and a voltmeter?

4. Describe the alignment of a two-stage transformer-coupled limiter. What precautions must be observed when limiters are aligned?

5. Can improper alignment of the I.F. and limiter stages affect the operation of the discriminator? Explain.

6. What effect does the type of F-M detector used have on the method of alignment?
7. What effect will a strong signal have on the alignment of the I.F. and limiter stages? Assume the indicating meter is placed in the grid circuit of the limiter.
8. What is the purpose of an overall alignment check on the I.F. and detector stages? Describe how this check is carried out.
9. How are over-coupled I.F. transformers aligned?
10. Why is the I.F. system aligned before the R.F. section of the receiver? Would it be possible to reverse the process? Explain.
11. Describe the alignment of the R.F. section of an F-M receiver.
12. What equipment is needed to run a visual check on a receiver? Explain how each piece of equipment is used.
13. Explain the difference between sine-wave deflection and sawtooth deflection of the electron beam in the oscilloscope.
14. What would happen if a sweep generator using sine-wave modulation was used in conjunction with an oscilloscope in which the electron beam was deflected by a sawtooth voltage? How would the response pattern appear on the scope screen?
15. Why is a phase control incorporated in sweep signal generators?
16. What are marker frequencies? Why are they used?
17. Describe the alignment of the I.F. system and the discriminator by the visual method.
18. How are marker signals developed?
19. What differences are there between the alignment of balanced and unbalanced ratio detectors?
20. Describe the alignment procedure for F-M systems using beam-gated detectors.

Chapter 12

COMMERCIAL F-M RECEIVERS

We have studied the various sections of an F-M receiver in considerable detail. Now we shall observe how these principles are applied to a complete receiver. It will be found, after examination of the many receivers which are available, that they differ markedly in essentially two respects—type of R.F. tuner employed and the circuitry of the F-M detector or discriminator. The remaining sections of the receivers, which include the I.F. stages, the audio amplifiers, and the power supply, are designed along closely similar lines. Although few of these stages are exact duplicates of each other, complete understanding of the circuit in one receiver will lead readily to comprehension of corresponding circuits in other receivers.

With regard to discriminators, all of the presently available commercial F-M receivers employ either a Foster-Seeley discriminator or some form of ratio detector. The other two types of F-M detectors which have been described have appeared only in the sound section of television receivers. Their quality has apparently not been regarded as good enough for a commercial F-M receiver. Hence, we will not encounter these detectors in our examination of commercial F-M receivers here.

We will examine in this chapter four receivers which are typical. The first is a low-cost receiver capable of receiving either A-M or F-M signals. The second is an F-M tuner designed to receive F-M signals only, amplify these signals, and then demodulate them, and equipped with an external amplifier system, to drive a loudspeaker. The third is an A-M, F-M tuner performing the same functions as the foregoing F-M tuner, except that it covers both commercial broadcast bands. In addition, it possesses other features which add to its versatility. The fourth is a complete A-M, F-M receiver, containing both a tuner and a power amplifier. Only an external speaker is required to provide a complete high-fidelity system. The unit also possesses an internal preamplifier, enabling records to be handled as

well. Equalization, loudness correction and a continuously-variable bass-treble tone control are also available in this receiver. With these four receivers as a background, the reader will be able to analyze almost any type of F-M or A-M, F-M combination receiver he is likely to encounter.

A Low-Cost A-M, F-M Receiver. A low-cost A-M, F-M receiver, housed in a small table model cabinet, is shown in Fig. 12.1. A total of seven tubes perform all the functions needed to deal with both A-M and F-M signals. A 4 x 6 inch speaker reproduces the sound output. No power transformer is employed, the d-c voltage being developed by a small selenium rectifier. This permits the receiver to operate from either 120 volt a-c or d-c power line. A slide-rule type of tuning dial carries one set of markings for the F-M channel and another set for the A-M channel. The band to be received is selected by a two-position switch marked "F-M" and "A-M."

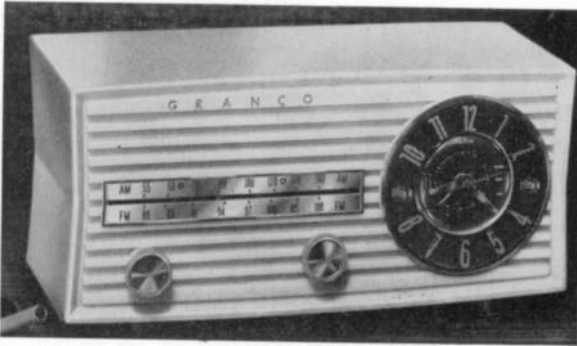


Fig. 12.1. The Granco Model 770 A-M, F-M receiver.

The F-M Section. The tubes in this section are: An F-M, R.F. amplifier (12BA6), an F-M mixer (using one-half of a 12AT7), an F-M oscillator (using the other half of the 12AT7), two I.F. amplifiers each using a 12BA6, a ratio detector (19T8), an audio frequency amplifier using the triode section of the 19T8, and a 35C5 output amplifier. Of these tubes, the two I.F. amplifiers, the A.F. amplifier, and the audio output serve both F-M and A-M channels. The remaining ones deal solely with F-M. Incoming F-M signals are developed across L_1 , a broadly tuned air-core coil, and applied to grid No. 1 of the F-M, R.F. amplifier, V_1 . Here they are amplified, appearing across the output load inductance, L_2 . Since the R.F. amplifier is untuned, all of the signals of the F-M band are received equally well. However, the signal reaching the mixer, V_{2A} , is determined by the R.F. tuning circuit, L_{10} , C_7 , and the small trimmer capacitor (T) shunting C_7 .

C_7 is mechanically ganged to C_{10} , the F-M oscillator tuning capacitor, and to C_8 and C_9 . The latter two capacitors tune the A-M ferrite rod an-

tenna (L_3) and the A-M oscillator. In actual construction C_7 and C_{10} differ markedly from C_8 and C_9 . The first two capacitors are coaxially shaped units in which one cylinder moves within another cylinder, the two being separated by a glass dielectric. (This type of variable capacitor was previously discussed in Chapter 5.) C_8 and C_9 , on the other hand, are conventionally-shaped capacitors in which each has a variable set of plates which mesh with a fixed set of plates. All four units, however, are mechanically ganged together so that one front panel knob controls them.

Feeding into the F-M mixer is another signal developed by an ultrasonic oscillator. This oscillator operates above the incoming signal, combining with it in the mixer to form the 10.7-mc I.F. voltage. This signal is developed across the primary of T_1 and inductively coupled to the secondary where it is applied to the control grid of the first I.F. amplifier, V_4 . It will be noted that the secondary of the 455-kc A-M, I.F. transformer T_2 is in series with the secondary of T_1 . No difficulty is caused by this arrangement because this winding of the 10.7-mc transformer possesses negligible inductance at 455 kc. By the same token, the by-pass capacitor of the 455-kc transformer permits the high frequency signals to pass around this winding.

V_4 amplifies the F-M signal and then transfers it by means of T_3 to the grid of the second F-M, I.F. amplifier, V_5 . Additional amplification occurs here, after which the signal is forwarded to T_5 and, from here, to an unbalanced ratio detector. The stabilizing network in the detector is formed by C_2 and R_{14} . C_{19} is placed in parallel with C_2 to offset the appreciable inductive reactance that is common to electrolytic capacitors such as C_2 .

The audio output of V_{6A} travels through the de-emphasis network, R_{12} and C_{21} , down to position 1 of switch S_1 . This is the previously mentioned two-position front panel switch which enables the set user to select either F-M or A-M. The switch, as shown, is in the F-M position; therefore, the signal received from the ratio detector will pass from contact 1 to contact 2 and from here to the volume control and the A.F. amplifier. The triode section of the 19T8 serves as an audio voltage amplifier, amplifying the voltage received from the movable arm of the volume control and then forwarding it to V_7 , the audio output stage. Here sufficient power is developed to drive a 4 x 6 inch loudspeaker.

A small amount of negative feedback is obtained from the secondary of the output transformer and applied to the bottom of the volume control. This stabilizes the audio frequency system and tends to reduce the distortion produced here.

When switch S_1 is in the F-M position, contacts 4 and 5 are electrically connected and the 80 volts applied to contact 5 reach the F-M oscillator, F-M mixer, and second F-M, I.F. amplifier. These stages serve only the

F-M signals. When switch S_1 is in the A-M position, the B+ voltage is removed from these stages, inactivating them.

The power supply, serving all sections of the receiver, consists of a selenium rectifier and a fairly simple filter network. Output voltages of 110 and 80 volts are developed. The filaments of each of the seven tubes are series-connected in the order shown. Appropriate dropping resistors are inserted in the circuit to provide the proper voltage distribution within the filament network; in addition, by-pass capacitors are shunted across various points in this chain to make certain that R.F. or I.F. signals do not travel from one stage to another through the filament circuit. If this is permitted, regeneration may result.

The A-M Section. A-M signals are picked up by a ferrite rod antenna, L_3 , and applied to the signal grid (No. 3) of the 12BE6 A-M converter. At the same time, the control grid (No. 1) and cathode of this tube serve as an oscillator, using the tuning circuit formed by L_4 , C_9 , C_{13} , and a trimmer capacitor (T) to develop the necessary oscillations for the mixing operation. An output I.F. signal of 455 kc is developed in the plate circuit of V_3 , across the primary winding of T_2 . From here, it is coupled to the secondary winding and the control grid of V_4 . After the signal is amplified, it is placed across the primary winding of T_4 , inductively coupled to the secondary winding, and then rectified by the cathode and control grid of V_5 . Since we are now operating on A-M, switch S_1 has been rotated one position counter-clockwise and the B+ voltage which was previously applied to the plate of V_5 has now been removed. The same is true of the screen grid. The A-M detector load resistor is R_{10} and the audio signal which appears here is brought to position 3 of S_1 . From position 3, it travels to position 2 and from here to the volume control. From this point, the audio signal is amplified by V_6 , then by V_7 , and finally applied to the loudspeaker.

Resistor R_9 and by-pass capacitor C_{14} filter out the audio variations and permit the average d-c voltage developed across R_{10} to serve as an A.V.C. voltage for V_4 and the signal grid of V_3 . This arrangement, it will be seen, is a conventional A.V.C. network and functions in the same manner as it does in any A-M receiver. In the A-M position, contacts 5 and 6, switch S_1 , are electrically connected and this enables the 80 volts from the power supply to power both the screen grid and plate of V_3 . At the same time, V_1 , V_{2A} , V_{2B} , and V_5 are rendered inactive.

From this analysis, it can be seen that both the A-M and the F-M circuits are straightforward and neither contains anything that has not been previously discussed. One interesting feature is the power line antenna. The power line, in addition to its regular function, is employed also as an antenna and F-M signals which it captures are capacitance-coupled to a

short section of wire which is connected to the antenna input terminals. This provides satisfactory F-M reception for all moderate and strong signals. If it is desired to extend the reception range and pick up weak signals, an outside antenna is employed. In this case, the power line antenna is simply set aside.

Alignment Procedure. The alignment procedure for both A-M and F-M sections of this receiver follows a standard pattern. Allow the set to warm up 15 to 20 minutes before beginning the alignment. Always align the receiver using the lowest signal from the signal generator which will give a usable reading on the indicating device. The volume control should be set to the maximum position. The manufacturer recommends alignment of the A-M circuits first. Since this is a transformerless receiver, either an isolation transformer should be employed between the receiver and the power line or a .1-mf capacitor should be connected in series with the ground side of the signal generator and the B- point of the receiver. These precautions should be carefully observed; otherwise the danger exists of short-circuiting either the receiver or the signal generator with disastrous results.

A-M Alignment. Connect the signal lead of the signal generator to pin 7 of the 12BE6, V_3 , through a series .1-mf capacitor. The low side of the signal generator connects to the receiver chassis. Set the signal generator to a frequency of 455 kc, modulated by a 400-cycle signal. The band switch of the receiver is turned to the A-M position and the tuning capacitor is fully opened. Also, an output meter is connected across the voice coil of the speaker.

With equipment and receiver in operation, adjust the primary and secondary cores of transformer T_2 and T_4 for maximum indication on the output meter. Next, fashion a loop of five to ten turns of wire and connect the open ends of this loop to the signal generator. Position the loop so that it is close to the ferrite antenna rod of the receiver. Set the signal generator frequency to 1650 kc. Some indication of signal should appear on the output meter. If none is visible, bring the loop closer to the receiver until an indication is obtained. Then adjust the trimmer capacitors of C_8 and C_9 for maximum output meter reading. Next, reduce the signal generator frequency to 610 kc and tune the front dial to the same frequency. Adjust the core of L_4 for maximum output on the meter. Do this while slightly rocking the tuning capacitor back and forth to make certain that the peaking is done at the proper point.

This completes the adjustment of the A-M section of the receiver. Various signals should be tuned in now to make certain they can be received properly.

F-M, I.F. Alignment. As a first step in the F-M, I.F. alignment, connect two matched 100,000 ohm resistors (plus or minus 1%) in series from point

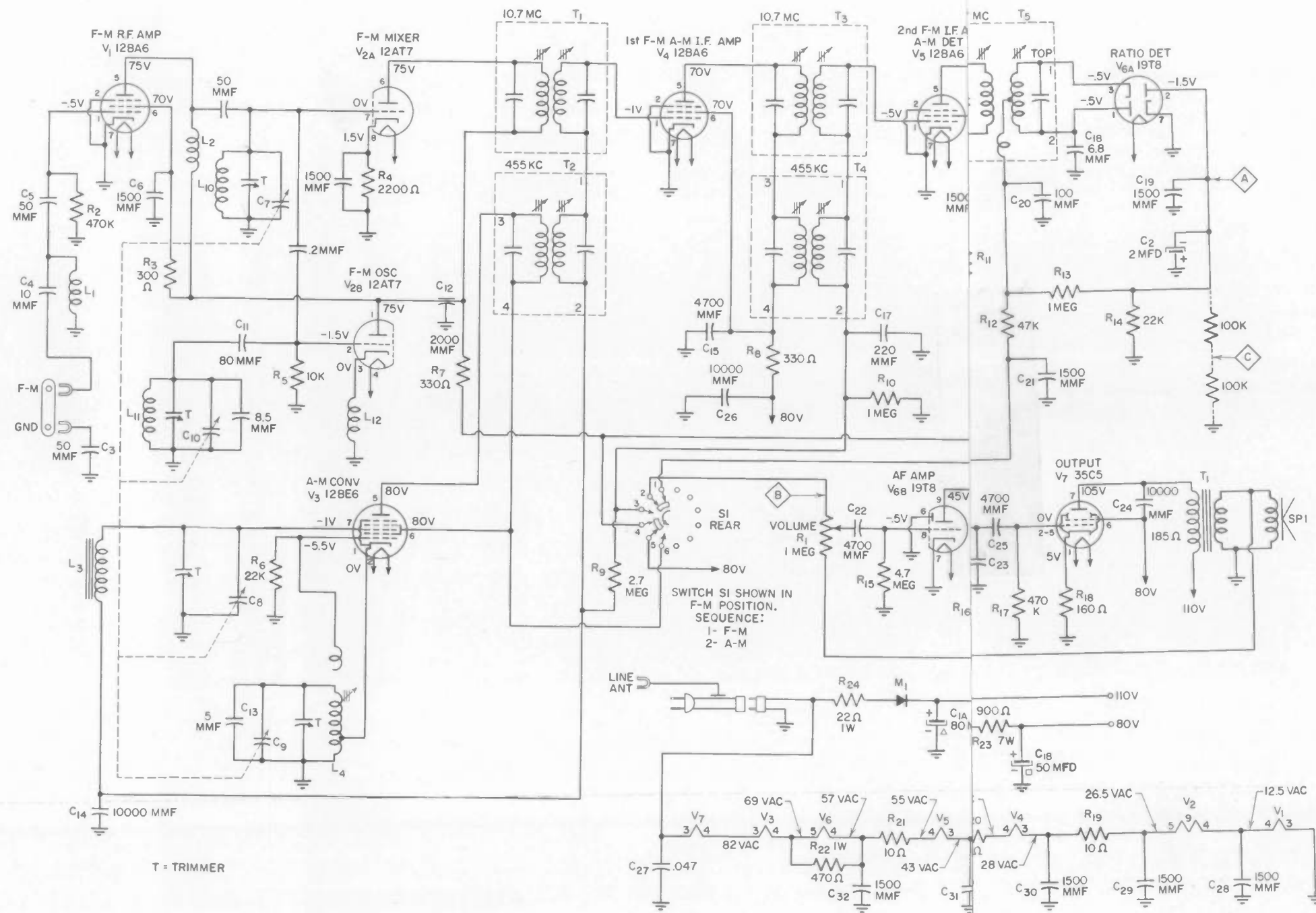


FIG. 12.2. Schematic diagram of a low-cost AM-FM receiver. Courtesy Howard W. Sams and Company, Inc.

A (Fig. 12.2) to the chassis. The junction of these two resistors is alignment point C. The signal generator has its signal lead connected to pin 7 of V_{2A} . This is done through a .01-mf capacitor. The low side of the signal generator connects to the chassis. Set the signal generator to a frequency of 10.7 mc, unmodulated. Turn the band switch to the F-M position. A VTVM is used as the indicator, with its d-c probe going to point A. The common lead of the VTVM connects to the chassis. As a final preparatory step, the front dial is set to a point where no interference is heard.

Now, adjust the following cores for maximum indication on the meter: the primary core of T_5 , the primary and secondary cores of T_3 , and the primary and secondary cores of T_1 . These should be turned carefully to make certain that maximum indication is obtained. Next, the d-c probe is moved to point B and the common lead of the VTVM is shifted to point C. Rotate the secondary core adjustment of T_5 until a zero reading is obtained on the meter. This zero reading should be between a positive reading when the core is rotated in one direction and a negative reading when the core is rotated in the opposite direction. This is similar to the method previously discussed.

This completes the complete alignment of the F-M, I.F. system and ratio detector.

F-M, R.F. Alignment. The signal generator is brought now to the input antenna terminals and each lead is connected to an antenna terminal through a 150-ohm resistor. The signal generator is initially set for a frequency of 108 mc. The front panel dial is turned to the same frequency. The d-c probe of the VTVM is connected to point A and the common lead of the instrument is connected to the chassis. Then, the trimmer capacitor across C_7 and the trimmer capacitor across C_{10} are each adjusted for a maximum reading of the meter.

The signal generator frequency is turned next to 88 mc and the front panel dial is changed to suit. L_{10} and L_{11} are adjusted by either compressing or expanding the coil turns until maximum deflection is obtained on the VTVM. This should be done very carefully because not much positional change is needed to produce a marked change at the meter.

The alignment of the entire F-M section of the receiver is completed now and stations should be tuned in at various points on the dial to make certain that every adjustment has been properly made. Note that nothing is done to coils, L_1 and L_2 . These serve not as tuning coils, but as loads for their respective circuits— L_1 for the antenna and L_2 for V_1 .

F-M Tuner. A tuner designed to operate only in the F-M band, 88 to 108 mc, is shown in Fig. 12.3. Seven tubes are employed, six in the signal circuits and one in the a-c power supply. Two outputs are provided: one at low level which is not affected by the volume control on the tuner; and the

other at high level which is controllable. Overall sensitivity of the tuner is high enough to permit operation with an indoor antenna made up of 300-ohm twin lead, if reasonable signal strength is prevalent in the area. If the signal level is low, an outdoor antenna is recommended.



FIG. 12.3. The Heath F-M tuner, Model F-M 3A. (Courtesy Heath Co.)

A schematic diagram of the tuner is shown in Fig. 12.4. The sequence of stages is conventional, starting with an R.F. amplifier. However, this amplifier is of the cascode type to provide high R.F. gain and to reduce oscillator leakage to the antenna. A 6BQ7A twin triode is employed in this circuit. The incoming signal is received by an antenna transformer whose purpose is to match the input impedance of the first triode to the 300-ohm twin lead transmission line. (Input impedance of nearly all commercial F-M receivers and tuners is currently established at 300 ohms.) The antenna coils are made broadly resonant with the 10-mmf capacitor and 6800-ohm resistor so that any signal in the F-M band can be received. Automatic gain control (A.G.C. or A.V.C.) is applied to the control grid of V_{1A} . This makes it necessary to feed the signal from the antenna coil to the grid through a 47-mmf capacitor. The capacitor will pass the R.F. signal, but block the d-c voltage from the antenna coil and ground.

The first half of the 6BQ7A tube acts as a conventional triode voltage amplifier. Its plate load is made up of the plate resistance of the second half of the tube which is in series with the first half and the 10,000-ohm resistor, R_1 . Voltage amplified by the first half of the tube is conducted through a neutralizing choke to the cathode of the second half, causing this element to swing by approximately the same amount. The neutralizing choke resonates with the circuit and tube capacitance in the middle of the F-M band, providing added gain to the stage and preventing oscillation. In the second section of the 6BQ7A, the grid is at R.F. ground potential, while the cathode potential is varied. Loading for the second half of the 6BQ7A is the 10,000-ohm resistor, R_1 , and the R.F. tuning circuit that follows it. As

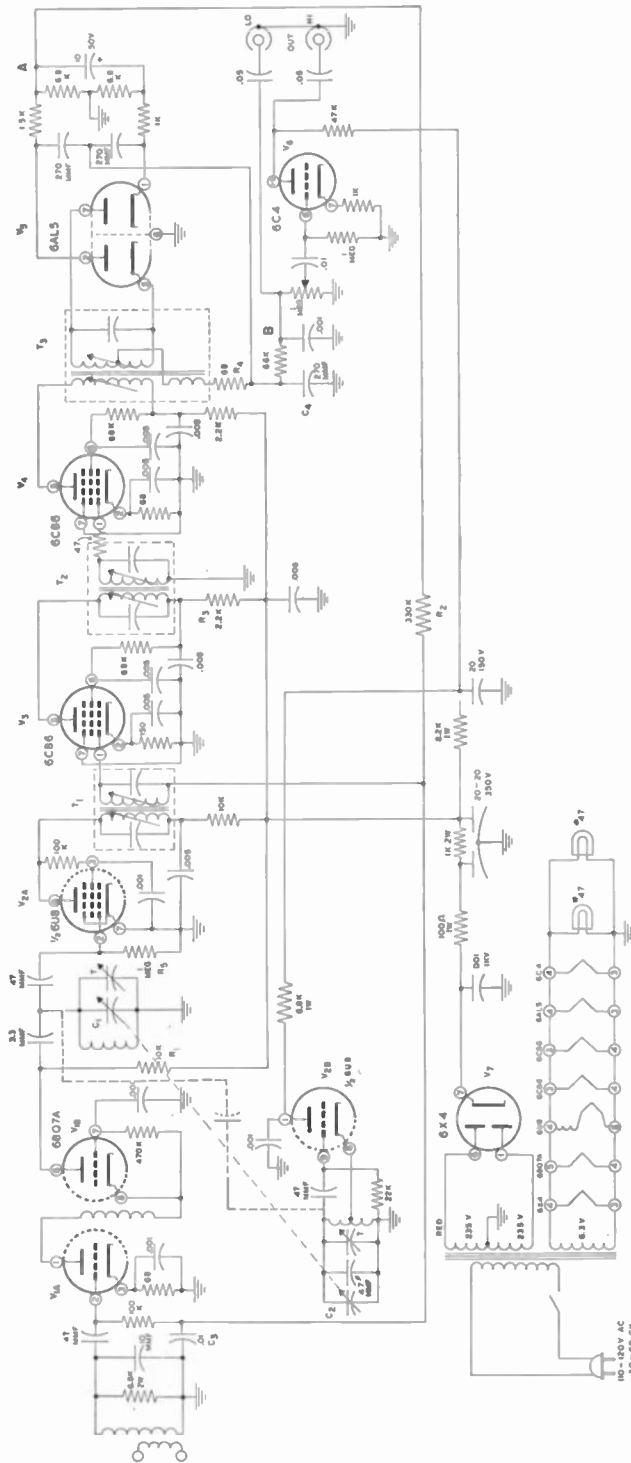


Fig. 12.4. Schematic diagram of the Heath F-M tuner shown in Fig. 12.3.

indicated earlier, the main advantage of the cascode circuit is low noise and a gain equivalent to that of a pentode.

The mixer which follows V_1 uses the pentode section of a 6U8. The triode section of this tube is connected as a Hartley oscillator, and the frequency of the oscillator is 10.7 mc higher than the incoming frequency. Energy from the oscillator is shown in Fig. 12.4 as being capacitatively coupled to the mixer. This is true; however, an actual coupling capacitor (as the diagram appears to indicate) is not used. Instead, after the circuit is wired, the 47-mmF capacitor in the oscillator grid circuit is physically pressed close to the 47-mmF capacitor connected to the control grid of V_{2A} . This close positioning creates sufficient capacitance between the oscillator and mixer circuits to enable the proper amount of energy to pass from one to the other.

It is interesting to note that V_{2A} uses contact bias. If the cathode of a tube is tied directly to circuit ground and the grid is returned to ground through a high resistance (here, 1 meg) a very small amount of current will be drawn by the grid. This current will be limited by the resistor, however, and a slight negative voltage will appear at the grid. Biasing in this manner is useful where the cathode impedance must be kept low and the signal level is also low.

Two stages of I.F. amplification are employed. The 6CB6's are high-gain tubes, providing the signal with considerable amplification. The I.F. transformers, of which there are two, each have adjustable iron cores for peaking the primary and secondary windings to 10.7 mc.

A balanced ratio detector follows the second I.F. The 10-mf capacitor with the two 6800-ohm resistors to which it is attached serve as the amplitude stabilizing elements which are chiefly responsible for the A-M rejection properties of this detector. The 1500-ohm and 1000-ohm resistors in series with the diode sections of the 6AL5 help to maintain current balance in the circuit. Audio, from the tap on the detector transformer secondary, is passed through a 68-ohm resistor and a 270-mmF capacitor network which by-passes all remaining I.F. signal to ground, leaving only the audio signal. This is next passed through a 68,000-ohm resistor and a .001-mf capacitor de-emphasis network. The audio signal appears across the 1-megohm volume control. The low-level output signal is taken from the fixed resistance terminals of the control; hence, the control has no effect on this output. Output from the variable tap is connected through a .01-mf capacitor to the grid of the 6C4 audio amplifier.

The 6C4 stage is a conventional audio amplifier. Since the cathode resistor is unby-passed, generation or negative feedback results. This reduces stage gain. However, it also reduces noise and distortion and, in this sense, is useful. A fairly low value of plate load resistance keeps the output impedance low. This also reduces hum pickup in the interconnecting audio

cable and minimizes high-frequency loss. However, since this output impedance is higher than that provided by a cathode follower, the length of the cable should be kept as short as possible.

A.V.C. bias is obtained from the stabilizing network in the ratio detector and fed to the control grids of V_3 and V_{1A} . The system functions in exactly the same manner as the A.V.C. network in an A-M radio receiver.

Note, however, that, because V_{1A} and V_{1B} are in series with each other, varying the grid bias on V_{1A} varies the current through both sections of V_1 . In essence, then, we are controlling the gain of both triodes.

The power supply has a 6X4 full-wave rectifier and a three-section R-C filter. Output B+ voltages of 175 and 100 volts are produced.

F-M Receiver Sensitivity. In recent years, F-M set manufacturers have established a method of indicating receiver sensitivity by giving the number of microvolts required at the input terminals to produce a certain amount of output quieting. For the present receiver, the manufacturer states that 5 microvolts are required for 20-db quieting. Eight microvolts are needed for 30-db quieting.

This method of designating receiver sensitivity serves to indicate the relative freedom of a receiver from objectionable internal noise during pauses in modulation when receiver noise is least likely to be masked by the modulation of the broadcast. Internal noise in the receiver consists largely of the noise generated in the R.F. amplifier and mixer and in the resistors associated with these stages. For those readers who may wonder how the quieting figure is obtained, here is the procedure recommended by the Institute of Radio Engineers. An F-M signal generator is connected to the receiver and the output level is set initially for about .001 volt (1000 microvolts). The generator is adjusted for a deviation of 30% of 75 kc. This would be about 22.5 kc and the frequency will change back and forth at the rate of 60 cps. At the output of a receiver tuned to the signal, the 60-cps note would be heard. The volume control of the receiver is adjusted to some convenient level well below the audio overloading point.

The modulation switch is turned off and on alternately while the signal output of the generator is reduced gradually until a point is reached at which there is a 30-db voltage difference (voltage ratio of 31.62) in receiver output between the time when the signal is modulated and when it is unmodulated. This value of input signal is the quieting figure. Obviously, the smaller the input signal, for a specified quieting figure, the more desirable the receiver, all other things being equal. This means that 5 microvolts for 30-db quieting is better than 10 microvolts for 30-db quieting; or it is even better than 5 microvolts for 20 db quieting. Note again that, if the signal level at a particular location is high, very little tangible value is gained by having a receiver with an extremely low quieting figure be-

cause there always will be enough signal coming in, modulated or unmodulated, to override the noise developed by the set.

Circuit Alignment. The alignment of this tuner can be carried out with a signal generator and a VTVM. The signal lead of the signal generator is connected to the control grid of V_{2A} through a .01-mf capacitor. The generator ground lead connects to the receiver chassis. The d-c probe of a VTVM (or a 20,000-ohm VOM) goes to point *A* in the ratio detector. The ground lead of the instrument attaches to the receiver chassis. Set the meter to read -DC, for the output voltage is negative at this point. Generator frequency is 10.7 mc.

With the instruments and the tuner in operation, adjust the primary winding of T_3 and the primary and secondary windings of T_2 and T_1 for maximum meter indication. Always use as low a signal from the generator as possible, reducing the generator output as alignment proceeds.

Next, move the d-c probe of the VTVM or VOM to point *B*. All other connections remain unchanged. Now adjust the secondary of T_3 for a zero reading on the meter. This zero point should be between a positive indication on one hand and a negative meter indication on the other. This completes the alignment of the I.F. stages and the ratio detector.

To align the R.F. circuits, connect the signal generator to the two antenna terminals. Since the output impedance of most generators is of the order of 50 ohms, insert a resistor of 125 to 150 ohms in each lead. This will match the generator to the receiver.

The VTVM or VOM is connected between point *A* and chassis ground. Set the generator to 106 mc and set the receiver dial to the same frequency. Carefully adjust the R.F. trimmer (located on C_1) for maximum meter reading. Rock the dial back and forth while adjusting the trimmer in order to find the highest peak response. Adjust the trimmer on the R.F. oscillator tuning capacitor, C_2 , for the highest meter reading. Go back and check the R.F. tuning trimmer (on C_1) to make certain it is still at its peak.

This completes the alignment of the front-end section of this unit. No adjustment is recommended at the low end of the F-M band by this manufacturer. In other sets, this may be done.

A-M, F-M Tuner. The circuit diagram of an extensive A-M, F-M tuner employing 12 tubes is shown in Fig. 12.5. The unit itself is pictured in Fig. 12.6. The 12 tubes perform 14 functions, which includes an automatic frequency control (A.F.C.) for the F-M oscillator, a special tuning meter amplifier, and a cathode follower output. The audio signal is obtained from this latter stage and made available to a separate audio amplifier system which would amplify the signal and then feed it to a loudspeaker. While the overall circuit is extensive, it follows closely the principles of the receiver previously outlined.

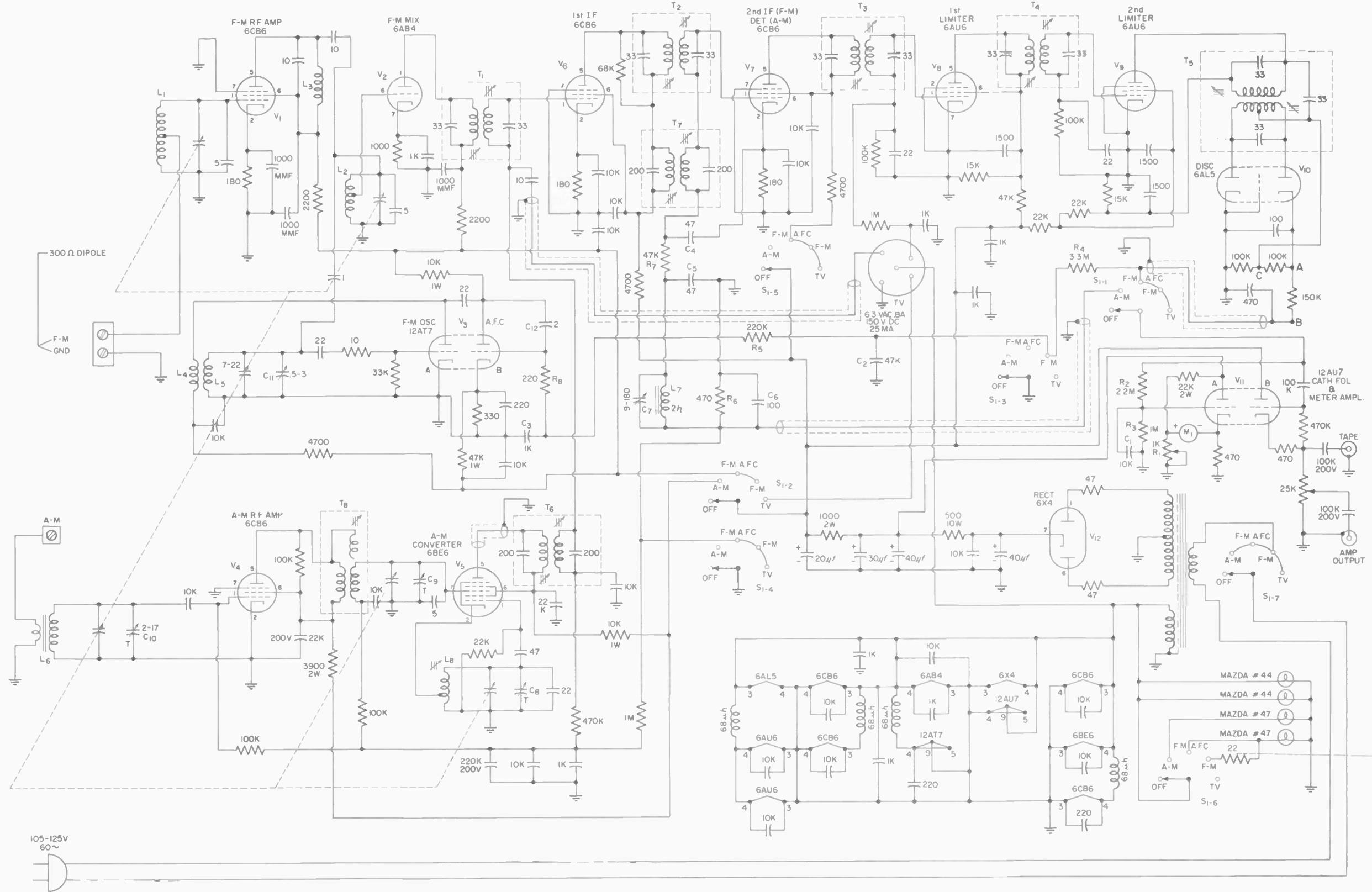


FIG. 12.5 Schematic diagram of the Allied Model KN 110 AM-FM tuner. Capacitance values are given in micromicrofarads unless otherwise specified. The letter K in a capacitance value stands for 1,000.

F-M Section. The F-M section consists of an R.F. amplifier, a mixer, an oscillator, A.F.C., two I.F. stages, two limiters, and a Foster-Seeley discriminator. The output from the discriminator is applied to a cathode follower and from here the signal can be transferred to an external audio power amplifier. The R.F. amplifier employs a high-frequency pentode tube operated conventionally. The antenna is tapped down on L_1 in order to achieve the proper impedance match. The signal reaching the grid of V_1 is amplified and then transferred to the second R.F.-tuned circuit in the grid circuit of V_2 , the F-M mixer. L_3 , at the plate of V_1 , serves simply as a broadband load for the R.F. amplifier.

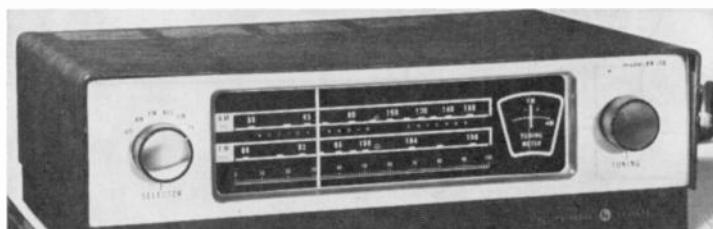


FIG. 12.6. The Allied Model KN110 A-M, F-M tuner.

The high-frequency oscillator, V_{3A} , is a tuned-grid circuit employing inductive feedback between the plate and the grid. The oscillating voltage developed across L_5 is brought by a 1-mmf capacitor to the grid of V_2 . The incoming signal and the oscillator signal mix in V_2 , and a 10.7-mc I.F. is developed across the primary winding of T_1 .

At the frequency at which F-M oscillators function, any slight tendency to drift would tend to distort the sound output of the receiver and require frequent returning of the front dial. This can be minimized by shunting an automatic frequency control circuit across the oscillator. The purpose of this auxiliary circuit is to maintain the oscillator frequency constant at whatever position it is set by the tuning control. This is achieved by having the A.F.C. tube function as a reactance tube. The circuit is so connected that V_{3B} appears to V_{3A} as a small variable capacitor. While the functioning of reactance tubes is discussed in detail in Chapter 14, it is not too difficult to see how V_{3B} can appear as a variable capacitor to L_4 . The oscillator and A.F.C. circuit of Fig. 12.5 is shown separately in Fig. 12.7. Note how the plate of V_{3B} is coupled to L_4 by the 22-mmf capacitor. This not only brings the oscillator voltage directly to the plate of V_{3B} , but it also means that any change in R.F. current flowing through V_{3B} will be coupled to L_4 . This latter point is important.

The same L_4 voltage is also brought to the control grid circuit of V_{3B} by capacitor C_{12} and resistor R_8 , causing a current to flow through these

components. The capacitive reactance of C_{12} is several times greater than the resistance of R_1 , so that the current through R_8C_{12} will lead the applied voltage by essentially 90° . The voltage developed across R_8 by the current will be in phase with the current, and it is this voltage which acts as the grid signal. Hence, the grid signal will lead the R.F. voltage from L_4 by 90° .

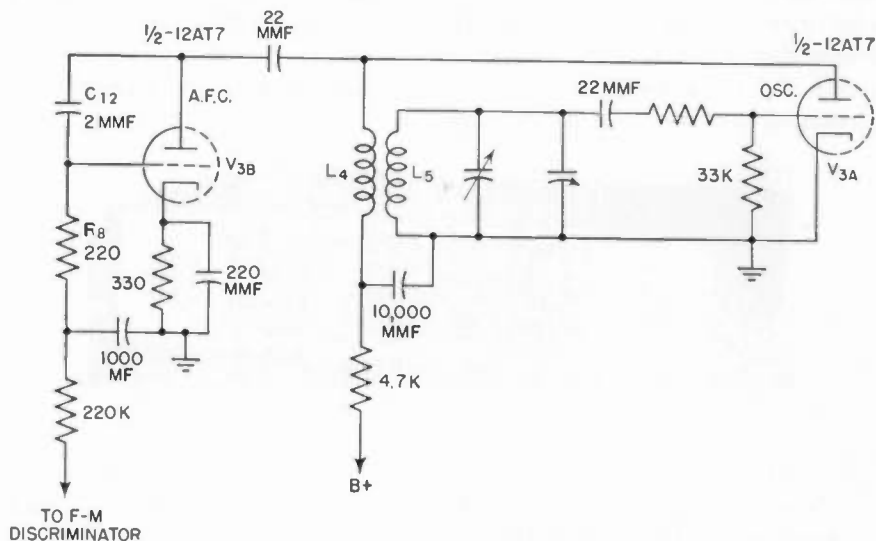


FIG. 12.7. Simplified diagram of the A.F.C. and R.F. oscillator stages of the receiver shown in Fig. 12.5.

Next, let us consider the plate current of V_{3B} . It is governed by the grid voltage and is in step with this grid signal. Thus, the plate current of V_{3B} leads the voltage across L_4 by 90° . Since this R.F. plate current flows through L_4 , V_{3B} will appear as a capacitor shunting L_4 . This is so, because, in any capacitor connected across L_4 , the voltage and current relationships just described would be present. The value of this capacitor is governed by the amount of plate current flowing which, in turn, depends upon the grid bias of V_{3B} . This voltage is obtained from the output of the discriminator. When a station is properly tuned in, the average d-c voltage across the discriminator load resistors is zero. However, if the station drifts to one side or the other, a negative or positive voltage will develop. This voltage, properly filtered to remove the audio variations, is fed to the grid of the A.F.C. tube where it will vary the plate current. This will have the same effect as varying the value of the capacitance which the reactance tube (V_{3B}) shunts across the oscillator circuit. Thus, if the oscillator tends to increase its frequency, the capacitance presented by V_{3B} will increase, causing the

frequency of the oscillator to drop. By the same token, if the oscillator frequency decreases, the A.F.C. capacitance will similarly decrease and offset the change in oscillator frequency.

Sometimes V_{3B} is shown with no capacitance bridged from plate to grid as C_1 in Fig. 12.7. In such instances, the internal capacitance between plate and grid of the tube is being used to bring the R.F. voltage to the grid circuit.

Beyond the F-M mixer, the 10.7-mc I.F. signal is amplified by V_6 and V_7 , then further amplified and limited by V_8 and V_9 . The latter two tubes operate with grid leak bias and low plate and screen voltages, thereby effecting signal saturation by two methods. Next, the signal is applied to the 6AL5 discriminator (V_{10}) where the audio variations are re-obtained. The signal is then brought to terminal 3, section 1, of switch S_1 (labeled $S_{1.1}$ in Fig. 12.5). From here, with the switch in the proper position, it is transferred to the grid of the cathode follower. The output from this stage is obtained from the cathode circuit.

Switch S_1 is a five-position switch with the following positions: OFF; AM; FM-AFC; FM, and TV. In the OFF position, the power is completely removed from the receiver. In the AM position, the A-M stages are activated and those stages which operate solely on F-M have their B+ voltage removed. In the third position, FM-AFC, the F-M section of the receiver is in operation and the A.F.C. stage is functioning. For the fourth position, FM, the entire F-M section is functioning as in the previous switch position, but the A.F.C. circuit is inactivated. This is done to permit the tuning in of weak signals; when these signals are properly brought in, the A.F.C. control can be re-established. Without this cutout, the stronger stations would tend to dominate, and any weak signal near a strong signal would be obscured. In the final position, TV, the filaments to the various tubes are powered, but all of the B+ is removed. This position is made available should this unit be used in conjunction with a TV receiver. By keeping the filaments lit, the user can always bring in the A-M or F-M sections of this receiver whenever he desires simply by rotating switch S_1 .

Switch S_1 has 7 sections. That is, it possesses 7 separate sections, all mechanically mounted on the same actuating shaft. This means that all sections are in position 1 at the same time, or position 2 at the same time, etc. Each section is electrically independent of all the others and each deals with a different segment of the circuit. Section 1 (labeled $S_{1.1}$ in Fig. 12.5) is concerned with bringing the proper audio voltage to the cathode follower, V_{11B} . Section 2 ($S_{1.2}$) provides B+ voltage to the F-M, R.F. amplifier, F-M mixer, F-M oscillator and A.F.C. tube when the F-M section of the system is in operation. Section 3 ($S_{1.3}$) inactivates the F-M, A.F.C. line when this is desired. Section 4 ($S_{1.4}$) grounds the A.V.C. line when the A-M

section of the receiver is not in use. Section 5 ($S_{1.5}$) brings B+ voltage to V_7 when the set is operating on F-M. It removes this voltage from the tube when the receiver is set for A-M. For the latter condition, the tube functions as a diode to detect the A-M signal. Section 6 ($S_{1.6}$) lights the appropriate pilot bulb for the service desired from the receiver. Finally, section 7 ($S_{1.7}$) serves as the power on-off switch for the entire receiver.

It is suggested that the reader become fully acquainted with these switch sections because they are so vital to every portion of the receiver. Without them, the receiver front panel would be hopelessly filled with switches.

The meter amplifier, V_{11A} , functions whenever an audio signal is fed to the cathode follower. This amplifier uses the d-c voltage output of the F-M discriminator to establish the reading of the meter. The latter is connected between the cathode of the tube and R_1 . The meter is adjusted initially so that it reads mid-scale when no signal is received. For F-M reception, the d-c voltage developed at the output of the discriminator is zero when a station is properly tuned in. Therefore, the meter amplifier will receive a zero d-c voltage and the meter needle will remain stationed at mid-scale. If the receiver is tuned to one side of the station, the voltage from the discriminator will assume a negative or positive polarity. This will cause the meter needle to swing away from center. If the receiver is tuned in the opposite direction, the polarity of the d-c voltage at the discriminator output will likewise reverse and the meter needle will swing in the opposite direction. Proper tuning is indicated when the meter needle is precisely at the central point. This is the best type of F-M tuning indicator since it reveals when the station is properly centered in the band pass of the tuned circuit of the receiver. Resistors R_2 and R_3 and capacitor C_1 serve to filter the signal received from the discriminator, removing the audio variations and permitting only the d-c voltage from the latter stage to reach V_{11A} . Resistors R_4 and R_5 and capacitors C_2 and C_3 perform the same function for the voltage which is applied to the grid of the A.F.C. tube, V_{3B} . Here, too, only the average d-c output voltage of the Foster-Seeley discriminator is desired.

A-M Section. The A-M section consists of a separate 6CB6 R.F. amplifier (V_4), a separate 6BE6 A-M converter (V_5), one I.F. stage (V_6), and an A-M detector (V_7). After this, the signal is applied through section 1 of S_1 to the input of the cathode follower. In detail, the antenna for the A-M, R.F. amplifier is a ferrite rod, L_6 . Its signal is fed to V_4 , amplified and then transferred to V_5 . Here the signal is combined with a locally generated oscillator voltage and the resultant 455-ke I.F. voltage is developed across the primary of T_6 . The signal is then inductively coupled to the secondary and applied to the control grid of the first I.F. amplifier, V_6 .

The next stage, V_7 , functions as an A-M diode detector because, during A-M operation of the receiver, the screen grid and plate elements are not given any B+ voltage. The load resistor for the detector is R_6 while C_4 , C_5 , C_6 , and R_7 serve as R.F. filters. The audio output developed across R_6 is transferred to section 1 of switch S_1 , and when this switch is turned to the A-M position, the signal is brought both to the meter amplifier and the cathode follower. At the same time, the d-c voltage developed across R_6 is employed as an A.V.C. voltage and is used to regulate the gain of V_6 , V_5 , and V_4 . Thus, the first three stages of this A-M section are subject to A.V.C. control. This is somewhat more extensive than is normally employed.

The d-c voltage which the meter amplifier receives is derived from the A.V.C. line; therefore, it possesses a negative polarity. When no station is received, the d-c voltage at the grid of V_{11A} is zero, and the meter needle is in the center of the range. When a signal is received, a negative voltage will develop, this voltage reaching its highest value when the station is properly tuned in. Therefore, for A-M reception, optimum setting of the tuning control is indicated when the meter needle is as far to the right as it will go. Note that this differs from the meter performance during F-M reception.

Power for all sections of the receiver is provided by a 6X4 full-wave rectifier. Voltages of 120 and 160 volts are available. As is common practice, B+ voltage is applied only to those sections of the receiver in use for a specific setting of the selector switch. All other portions of the receiver have their B+ removed. This is done to avoid interference which would certainly arise if all sections received B+ at the same time.

Note the absence of any volume control on the front panel of the receiver. Actually, the output obtained from the cathode follower can be regulated, but this is done not by a front panel control but by a rear panel screwdriver control. It is generally adjusted once and then left at this position. Further volume regulation would be made by a suitable front panel control in the audio power amplifier receiving the tuner signal.

One additional feature of this receiver is a 10-kc filter inserted between the output of the A-M detector and the cathode follower. The components of this filter, C_7 and L_7 , are designed to remove the 10-kc whistle which frequently develops when A-M stations are close enough to beat with each other. The beating or mixing of the two signals occurs in the A-M detector, and the resultant 10-kc note then passes through the audio system and is heard in the loudspeaker. To adjust L_7 and C_7 , several methods are available. When the A-M section of the receiver is aligned, the 455-kc I.F. signal can be modulated with a 10-kc note instead of the normal 400-cycle note.

TABLE I. ALIGNMENT PROCEDURE FOR AM-FM TUNER OF FIG. 12.5. (Courtesy H. W. Sams & Co.)

Initial steps.

Volume control should be at maximum position. Output of signal generator should be no higher than necessary to obtain an output reading. Use an insulated alignment screwdriver for adjusting.

A-M ALIGNMENT

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	CONNECT VTVM	ADJUST	REMARKS
0.1 mf	High side to pin 7 (grid) of 6BE6 (V_6). Low side to chassis.	455 kc (400 ~ Mod.)	AM	Point of non-interference.	AC probe to amplifier output jack. Common to chassis.	Primary and secondary cores of T_1 and T_2 .	Adjust for maximum deflection.
0.1 mf	High side to pin 7 (grid) of 6BE6 (V_6). Low side to chassis.	455 kc (10 kc Mod.)	AM	Point of non-interference.	AC probe to amplifier output jack. Common to chassis.	C_7	Adjust for MINIMUM deflection.
	Loop	1600 kc (400 ~ Mod.)	AM	1600 kc	AC probe to amplifier output jack. Common to chassis.	C_8 (Osc. Trimmer)	Fashion loop of several turns of wire and radiate signal into loop of receiver. Adjust for maximum output.
	Loop	600 kc	AM	Tune for 600-ke signal.	AC probe to amplifier output jack. Common to chassis.	Core of L_2	Fashion loop of several turns of wire and radiate signal into loop of receiver. Adjust for maximum output.
	Loop	1500 kc	AM	1500 kc	AC probe to amplifier output jack. Common to chassis.	C_9 and C_{10}	Fashion loop of several turns of wire and radiate signal into loop of receiver. Adjust for maximum output.
	Loop	600kc	AM	600 kc	AC probe to amplifier output jack. Common to chassis.	Core of T_3	Fashion loop of several turns of wire and radiate signal into loop of receiver. Adjust for maximum output. Repeat steps 3, 4, 5, and 6.

F-M I.F. ALIGNMENT USING A-M SIGNAL GENERATOR AND VTVM

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	CONNECT VTVM	ADJUST	REMARKS
0.01 mf	High side to pin 1 (grid) of 6AU6 (V_6). Low side to chassis.	10.7 mc (Unmod.)	FM	Point of non-interference.	DC probe to point C. Common to chassis.	Primary core of T_4 , primary and secondary cores of T_1 .	Adjust for maximum deflection.
0.01 mf	High side to pin 1 (grid) of 6AU6 (V_6). Low side to chassis.	10.7 mc (Unmod.)	FM	Point of non-interference.	DC probe to point B. Common to chassis.	Secondary core of T_4 .	Adjust for zero reading. A positive and negative reading will be obtained on either side of the correct setting.
0.01 mf	High side to F-M R.F. stator lug of tuning gang. Low side to chassis.	10.7 mc (Unmod.)	FM	Point of non-interference.	DC probe to point C. Common to chassis.	Primary and secondary cores of T_4 , T_1 , and T_1 .	Adjust for maximum deflection.

F-M I.F. ALIGNMENT USING F-M SIGNAL GENERATOR AND OSCILLOSCOPE

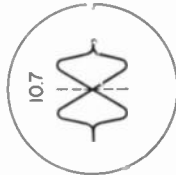
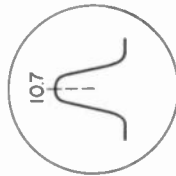
Use frequency-modulated signal with 60 ~ modulation and 450 kc sweep. Use 120 ~ sawtooth voltage in scope for horizontal deflection.

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	RAND SWITCH POS.	RADIO DIAL SETTING	CONNECT SCOPE	ADJUST	REMARKS
0.01 mf	High side to pin 1 (grid) of 6AU6 (V _g). Low side to chassis.	10.7 mc (450 kc sweep)	FM	Point of non-interference.	Vert. Amp. thru 10K to point C. Low side to chassis.	Primary core of T ₁ , primary and secondary cores of T ₂ .	Adjust for curve of maximum amplitude and symmetry similar to Fig. A.
0.01 mf	High side to pin 1 (grid) of 6AU6 (V _g). Low side to chassis.	10.7 mc (450 kc sweep)	FM	Point of non-interference.	Vert. Amp. to point A. Low side to chassis.	Secondary core of T ₁ .	Adjust so that 10.7 mc occurs at center of crossover lines similar to Fig. B. Slightly retouch primary of T ₁ for maximum amplitude and straightness of crossover lines.
0.01 mf	High side to F-M R.F. stator lug of tuning gang. Low side to chassis.	10.7 mc (450 kc sweep)	FM	Point of non-interference.	Vert. Amp. thru 10K to point C. Low side to chassis.	Primary and secondary cores of T ₁ , T ₂ , and T ₁ .	Adjust for curve of maximum amplitude and symmetry similar to Fig. A.

F-M R.F. ALIGNMENT

Last 2 steps are not necessary unless the receiver fails to track properly.

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	RAND SWITCH POS.	RADIO DIAL SETTING	CONNECT VTVM	ADJUST	REMARKS
270 carbon resistor	High side thru 270 ohms to F-M Ant. terminal. Low side to chassis.	108 mc	FM	108 mc	DC probe to point C. Common to chassis.	C ₁₁	Adjust for maximum deflection.
270 carbon resistor	High side thru 270 ohms to F-M Ant. terminal. Low side to chassis.	106 mc	FM	106 mc	DC probe to point C. Common to chassis.		Adjust for maximum deflection by bending F-M R.F. rotor plates of tuning gang.
270 carbon resistor	High side thru 270 ohms to F-M Ant. terminal. Low side to chassis.	88 mc	FM	88 mc	DC probe to point C. Common to chassis.	L ₁ and L ₂	Adjust by compressing or expanding coil turns for maximum deflection while rocking tuning gang.



While most audio signal generators only provide 400-cycle modulation, nearly all have provision for external modulation. A suitable 10-kc signal can be obtained from an audio oscillator for this purpose.

Another approach is simply to apply the 10-kc signal from an audio oscillator at the junction of R_7 and C_4 . The 10-kc note will then be heard in the loudspeaker, and C_7 can be adjusted for minimum sound.

Receiver Alignment. The alignment procedure for this receiver is given in chart form in Table I. Included are instructions for a visual alignment of the I.F. section and the discriminator.



FIG. 12.8. The Fisher Model 500 A-M, F-M receiver (Courtesy Fisher Radio Corp.)

An Extensive A-M, F-M Receiver. The receiver shown in Fig. 12.8 employs 14 tubes not only to provide reception of A-M and F-M signals and reproduce them on an output loudspeaker, but also to enable records and audio inputs from other sources to play through the same audio system. Thus, we have combined on one chassis a complete A-M receiver, a complete F-M receiver, and a phonograph playback system. (See Fig. 12.9.) The only additional equipment required with this unit, a Fisher Model 500 receiver, are a record changer and a suitable record cartridge.

F-M Receiver Section. Most of the system is similar to the F-M sections of previously described receivers. However, there are several circuits which warrant special attention. In the R.F. amplifier, the two triode sections of the 6BC8 form a cascode arrangement. The first section, V_{1A} , is connected as a conventional grounded cathode amplifier, with the incoming signal received at the control grid and the output signal developed in the plate circuit. This signal is then transferred by C_{10} to the cathode of the

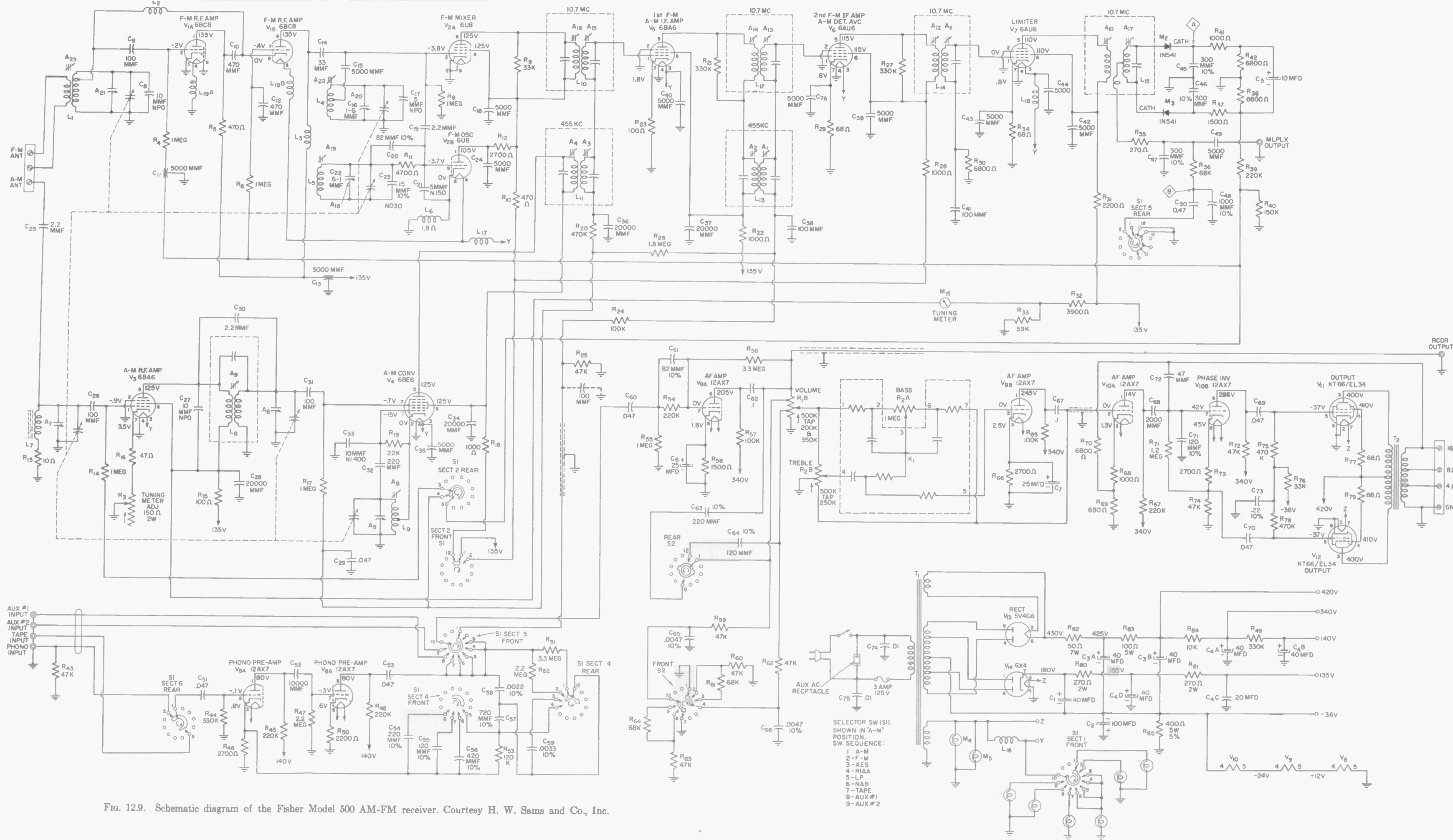


FIG. 12.9. Schematic diagram of the Fisher Model 500 AM-FM receiver. Courtesy H. W. Sams and Co., Inc.

second triode, V_{1B} . The grid of this section is placed at R.F. ground potential by C_{12} ; thus, this section is operating as a grounded-grid amplifier. Output from V_{1B} is also obtained from its plate circuit and fed by C_{14} to the grid of mixer, V_{2A} .

Both control grids of V_{1A} and V_{1B} have applied to them a negative control voltage obtained from resistor R_{38} in the ratio detector. This resistor has developed across it a negative voltage whose amplitude is a measure of the strength of the incoming signal. This voltage, properly filtered to remove any audio signal, is employed as an A.V.C. voltage for F-M reception.

The pentode section of a 6U8 serves as the F-M mixer, the triode section of this tube provides the necessary oscillator voltage. The tube is constructed so that each section is electrically independent of the other and an external 2.2-mmF capacitor must be connected between the two control grids in order to bring the oscillator signal to the mixer. No F-M A.F.C. circuit is used here, the oscillator tuning circuit relying on two temperature compensating capacitors, C_{23} and C_{21} , to furnish the necessary frequency stabilization. These capacitors have a negative temperature characteristic, so that, as the circuit warms up, the capacitance value of each unit decreases. This counterbalances the decrease in frequency which the same warm-up produces in the tuning and trimmer capacitors and in the coils of these resonant circuits.

For F-M signals, there are two I.F. amplifiers and one limiter before the signal reaches the balanced ratio detector. In place of vacuum-tube diodes, the ratio detector employs two matched germanium diodes. Additional balancing is achieved by resistors R_{37} and R_{41} . The audio signal developed by this stage is transferred to switch S_1 , section 5 through a deemphasis circuit R_{36} and C_{48} . The connection at this switch is made at terminal 2. This switch, which has nine different positions, is shown in position 1. This is the A-M position, for the reception of A-M signals. The remaining positions are:

2 — F-M	6 — NAB
3 — AES	7 — TAPE
4 — RIAA	8 — AUX #1
5 — LP	9 — AUX #2

Positions 3, 4, 5, and 6 refer to record equalization, while 7, 8, and 9 enable other audio signals, obtained from auxiliary equipment (including a tape recorder), to be passed through the audio system. A "Phono" input, also available, is active during the four equalization positions.

We might digress here for a moment and examine the notation employed for the selector switch. In a previous receiver, Fig. 12.6, one wa:

of illustrating complex selector switches was demonstrated. There, the various terminals of each section were shown in numbered sequence and the terminal to which contact was being made was indicated by the position of a central arm with an arrowhead at one end. While this demonstrated the action of the switch quite clearly, it did not depict the switch as it is physically, and anyone not familiar with these units could easily be lead astray when forced to work between the schematic diagram and the actual receiver.

A more accurate representation of the selector switch is rendered by the notation employed in Fig. 12.9. Each section of the switch is identified by a different number and, if there are front and rear plates on the section, they are also shown.

Each section of the switch contains a movable, circular metallic strip which, as it rotates, contacts certain switch terminals. Some terminals are blank, indicating that no connection is made to them. Where a connection does exist, a small arrow (denoting a contact arm) extends from the terminal in toward the movable contacting plate. If the arrow does not touch the plate, then no contact is made between the two. However, in some position of the switch, contact will occur.

Some arrows are longer than others and these generally remain in contact with the movable plate (as it is rotated) longer than the shorter arrows. If the plate forms a full ring, then the longer arrow will contact the ring in every position of the switch. At other times, the contact will exist for several switch positions and be broken for other positions.

Three pieces of information are usually given concerning a selector switch. First, its present position is indicated. In Fig. 12.9, switch S_1 is in the A-M position. Second, the other positions are also designated as to function. Finally, the direction in which each section is to turn from its present position, in order to follow the sequency, is indicated, too. This is done in Fig. 12.9. Note that the front plate of a section will turn in one direction, while the rear plate will turn in the opposite direction.

Let us examine some of the sections of switch S_1 to illustrate some of the foregoing points. In section 5, front, the arrow from terminal No. 10 extends far enough out to contact the movable plate in all positions of the switch. The wire from terminal No. 10 goes to the grid of the first A.F. amplifier, and all signals pass through this section of the circuit. In the position shown, the movable plate is touching the arrow extending inward from terminal No. 1. Thus, the signal at terminal No. 1 makes electrical contact with terminal No. 10 and this brings the output from the A-M detector to V_9 .

In the next position of the switch, the plate is moved one position clockwise. Contact now is no longer made to terminal No. 1, but to terminal

No. 2. While this terminal has an arrowhead, it does not show any wire coming to it from the circuit. Thus, it would appear as if, in the F-M position of the switch, nothing from the F-M section of the receiver will reach V_9 . However, if you examine section 5, rear, of switch S_1 (near the F-M detector), you will note that the output of the ratio detector does reach terminal No. 2. Since the section has only one set of terminals, the connection at terminal No. 2 of section 5, rear, is the same as a connection at terminal No. 2 of section 5, front. (Most of the time, the front plate contacts terminals that the rear plate does not, and vice versa. Occasionally, both plates will have terminals in common, such as No. 2 here.)

Note, too, that section 5 is so set up that those contacts not in use are grounded by the rear plate. This is useful in preventing any signal leakage into the active channel from an unused circuit.

Some of the sections have a large number of contacts, whereas some have very few. Section 2, front, has only three terminals to which wires go. Remember, too, that there is no direct electrical contact between the front and rear movable plates of a section. Any contact that may exist (and this is not done too often) is achieved through a terminal that both plates are contacting at the same time. No section 3 of switch S_1 will be found in Fig. 12.9. Such a section is physically present in the receiver, but it serves simply to provide tie points for some of the components in the switch circuitry.

To return to the circuit analysis, the output of the ratio detector is applied to V_{9A} through the selector switch when the latter is in the F-M position. This stage is an audio voltage amplifier, with a small amount of negative feedback developed by R_{56} , the 3.3-megohm resistor connected between plate and grid. The combination of C_{61} and R_{54} just in front of R_{56} is designed to improve the high-frequency response of the stage. It does this by permitting the higher audio frequencies to reach the grid of V_{9A} with less attenuation than lower audio signals. The latter are faced with a higher impedance from C_{61} and R_{54} .

The output of V_{9A} is transferred to the volume control. In addition to its normal function of varying the volume of the output audio sound, this control also has associated with it a four-position "Loudness" switch, S_2 .

The purpose of the "Loudness" control stems from the way the human ear responds to tones of different frequencies. This was previously discussed in connection with Fig. 10.1 where it was noted that the overall ear response becomes flatter as the volume increases. When the volume is turned down, the ear hears less of the high and low audio frequencies. If we listened to this music as it was being played by an orchestra, the sound level would be high and both high and low audio frequencies would be heard as well as

the medium frequencies. However, when the selection is played back at a much lower level (simply because most home playback systems are not capable of the sound intensity of the original), the low and high audio frequencies suffer a proportionately greater loss due to the aforementioned characteristics of our ears. It is the purpose of the loudness control to compensate for this loss by increasing the amplitude of the high and low audio frequencies.

Note that, since the ear becomes less sensitive to bass frequencies than to high frequencies as the volume decreases, greater boost is given the bass frequencies by these loudness controls. Sometimes high-frequency compensation is ignored completely because it is felt that the additional boost does not warrant the expense required to achieve it.

The loudness control is set according to the level of the sound. If a high level of sound is desired, less loudness correction is required; if the sound is normally kept now, a greater amount of correction is added.

Some readers may wonder why a special loudness control is needed since the base and treble controls can provide the same emphasis to the two ends of the audio frequency range. First, the loudness control is designed according to the frequency and loudness characteristics of the ears, as revealed by the curves of Fig. 10.1. Thus, the control introduces a fixed amount of compensation at each position.

Second, if we employed the normal bass and treble controls to compensate for the loss in hearing, we would not have enough additional leeway in these controls to permit adjustment for the individual tastes of the listener. Because of this, two sets of controls are provided.

Following the volume control, there are separate bass and treble controls of the continuous variety. These function in the same way as the controls discussed in Chapter 10. The bass control provides a boost of 17 db when tuned fully clockwise and a reduction of 15 db when tuned fully counter-clockwise. The treble control has approximately the same range.

V_{9A} , V_{9B} and V_{10A} are single-ended audio voltage amplifiers. V_{10B} is a split-load phase inverter, feeding equal and oppositely phased signals to the push-pull output stage. Beam power tetrodes are employed here. The signal is then transferred to an output transformer with secondary impedances of 4, 8, and 16 ohms. The amplifier is rated at 30 volts with a uniform frequency response from 16 to 32,000 cycles.

A small amount of negative feedback is provided from the secondary of T_2 to the cathode of V_{10A} . In addition, C_{72} and C_{71} provide special high-frequency feedback to offset a shift in phase that occurs for these frequencies in transformer T_2 . (It will be noted that the phase of the high frequencies fed back to the grid of the phase inverter is the same as the

phase of the signals reaching this point through the other feedback path. This is true because the signal fed back through R_{70} is applied to the cathode of V_{10A} and not its grid.)

A-M Section. The A-M section of the circuit is straightforward, employing an R.F. amplifier, V_3 , a converter, V_4 , one I.F. stage, V_5 , which is shared with F-M, and a detector, V_6 . The latter tube, although a pentode, functions as a diode on A-M because section 2 of switch S_1 removes its plate and screen-grid voltages at this time. The A-M detector load resistor is R_{25} and, from this point, audio and negative A.V.C. voltages are obtained. The audio signal is fed to the grid of V_{9A} through section 5, front, of switch 1, while the A.V.C. voltage is brought to section 2, rear, switch S_1 by resistor R_{26} . The switch, in the A-M position, transfers this A.V.C. voltage to the control grid of V_3 . A.V.C. voltage is also applied to V_4 and V_5 , although this is done directly, without any switches.

A tuning meter is employed in conjunction with the A-M section. One side of this meter is connected to the screen grid and plate of the A-M R.F. amplifier. The other side of the meter is connected to 135 volts through a resistive voltage-dividing network, R_{32} and R_{33} . Thus, this side of the meter receives a fixed voltage.

The voltage drop across R_{15} (in the screen grid and plate circuit of V_3) will vary with the strength of the signal passing through V_3 . This is because the gain (and, consequently, the current) of V_3 is controlled by the A.V.C. voltage.

A variable resistor in the cathode leg of V_3 permits adjustment of the needle swing so that the needle does not go off scale with the strongest received signal.

Interestingly enough, the same tuning meter also serves as a tuning indicator on F-M. When the selector switch is turned to the F-M position, section 2, rear, of S_1 feeds the negative d-c voltage developed by C_5 (of the ratio detector) to the control grid of V_3 . This voltage will regulate the amount of current flowing through V_3 and, consequently, the voltage present at its screen grid. Thus, even though V_3 amplifies only A-M signals—and, hence, operates only when the set is turned to A-M—it serves as d-c amplifier on F-M for the tuning meter. In recognition of this latter function, V_3 , always has its plate and screen grid voltages, whereas these are removed from V_4 in the F-M position of S_1 .

Preamplifier. The phono preamplifier employs a single 12AX7, V_8 . This added amplification is provided for magnetic cartridges only, since the output of a crystal cartridge would overload the audio system with this additional pre-amplification. The signal from the cartridge is passed through the two amplifiers and an equalizing circuit before it is transferred by sec-

TABLE II. ALIGNMENT PROCEDURE FOR THE A-M, F-M RECEIVER OF FIG. 12.8. (Courtesy H. W. Sams & Co., Inc.)

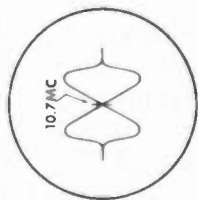
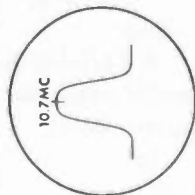
Pre-alignment instructions
Volume control should be at maximum position. Output of signal generator should be no higher than necessary to obtain an output reading. Use an insulated alignment screwdriver for adjusting.

A-M ALIGNMENT

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	OUTPUT METER	ADJUST	REMARKS
1. 0.01 mf	High side to pin 7 (grid) of 6BE6 (V ₁). Low side to chassis.	455 kc	AM	Tuning gang fully open	Across voice coil	A ₁ , A ₂ , A ₃ , A ₄	Adjust for maximum output.
2. 200 mmf	High side to AM antenna terminal. Low side to chassis.	1100 kc	AM	1400 kc	Across voice coil	A ₅ , A ₆ , A ₇	Adjust for maximum output.
3. 200 mmf	High side to AM antenna terminal. Low side to chassis.	600 kc	AM	600 kc	Across voice coil	A ₈ , A ₉	Adjust for maximum output. Repeat steps 2 and 3.

F-M I.F. ALIGNMENT USING A-M SIGNAL GENERATOR AND VTVM

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	CONNECT VTVM	ADJUST	REMARKS
4. Direct	High side to ungrounded tube shield on 6U8 (V ₂). Low side to chassis.	10.7 mc (Unmod.)	FM	Point of non-interference	DC probe to point A. Common to chassis.	A ₁₀ , A ₁₁ , A ₁₂ , A ₁₃ , A ₁₄ , A ₁₅ , A ₁₆ , A ₁₈	Adjust for maximum deflection.
5. Direct	High side to ungrounded tube shield on 6U8 (V ₂). Low side to chassis.	10.7 mc (Unmod.)	FM	Point of non-interference	DC probe to point B. Common to chassis.	A ₁₇	Adjust for zero reading. A positive and negative reading will be obtained on either side of the correct setting.



F-M I.F. ALIGNMENT USING F-M SIGNAL GENERATOR AND OSCILLOSCOPE

Use frequency modulated signal with 60 ~ modulation and 450KC sweep. Use 120 ~ sawtooth voltage in scope for horizontal deflection.

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	CONNECT SCOPE	ADJUST	REMARKS
4. Direct	High side to ungrounded tube shield on 6U8 (1 ₂). Low side to chassis.	10.7 mc (450 kc Sweep)	FM	Point of non-interference	Vert. Amp. to point A. Low side to chassis.	A ₁₀ , A ₁₁ , A ₁₂ , A ₁₃ , A ₁₄ , A ₁₅ , A ₁₆	Disconnect stabilizing capacitor (C5). Adjust for curve of maximum amplitude and asymmetry as in Fig. 1.
5. Direct	High side to ungrounded tube shield on 6U8 (1 ₂). Low side to chassis.	10.7 mc (450 kc Sweep)	FM	Point of non-interference	Vert. Amp. to point B. Low side to chassis.	A ₁₇	Reconnect capacitor (C5). Adjust A ₁₇ so that 10.7MC occurs at center of crossover lines as in Fig. 2. SLIGHTLY retouch A ₁₀ for maximum amplitude and straightness of crossover lines.

F-M R.F. ALIGNMENT

DUMMY ANTENNA	SIGNAL GENERATOR COUPLING	SIGNAL GENERATOR FREQUENCY	BAND SWITCH POS.	RADIO DIAL SETTING	CONNECT SCOPE	ADJUST	REMARKS
6. 270-ohm Carbon Resistor	High side to F-M antenna terminal. Low side to chassis.	106 mc (45 kc Sweep)	FM	106 mc	DC probe to point A. Common to chassis.	A ₁₈	Adjust for maximum deflection.
7. 270-ohm Carbon Resistor	High side to F-M antenna terminal. Low side to chassis.	90 mc	FM	90 mc	DC probe to point A. Common to chassis.	A ₁₉	Adjust for maximum deflection.
8. 270-ohm Carbon Resistor	High side to F-M antenna terminal. Low side to chassis.	106 mc	FM	106 mc	DC probe to point A. Common to chassis.	A ₂₀ , A ₂₁	Adjust for maximum deflection.
9. 270-ohm Carbon Resistor	High side to F-M antenna terminal. Low side to chassis.	90 mc	FM	90 mc	DC probe to point A. Common to chassis.	A ₂₂ , A ₂₃	Adjust for maximum deflection. Repeat steps 6, 7, 8 and 9 until no further improvement is obtained.

tion 5, front, of switch 1 to V_{9A} . Four types of equalization are provided: AES, RIAA, LP, and NAB. This is the conventional line-up, although frequently a fifth network entitled "Foreign" is incorporated also.

The 12AX7 tube, as we have already noted in Chapter 10, is specifically designed for low-level audio amplification, having low-noise and low-hum characteristics. In the present circuit, interference from the latter source is avoided completely by powering the filament of V_8 (and V_9 and V_{10} as well) with d-c. This voltage is obtained from the power supply.

Power Supply. Two separate full-wave rectifiers in the power supply develop four different positive voltages (420, 340, 140, and 135 volts) and one negative voltage (-36 volts). The latter potential is produced by having all of the circuit current pass through R_{85} . Its purpose is to heat the filaments of the three 12AX7 tubes, V_8 , V_9 , and V_{10} .

Extensive filtering is employed in both full-wave circuits. It is particularly important in a high-fidelity system to keep the level of a-c hum far below the lowest usable level of the audio signal.

The front part of section 1, switch S_1 , brings in a different pilot bulb for each different function of the receiver. Thus, there is one bulb for the A-M position, one for F-M, one bulb when the phonograph is being played, and one bulb each for the positions of tape, auxiliary input No. 1, and auxiliary input No. 2. These bulbs are physically positioned one above the other behind the front panel and serve to remind the set user of the position in which the receiver is set.

The complete alignment procedure for this receiver is given in Table II. The various steps are straightforward and follow the general alignment sequence previously given. Adjustments are labeled by the capital letter A and extend from A_1 to A_{23} .

In the F-M I.F. alignment instructions, the table indicates that the output of the signal generator is connected to the ungrounded tube shield on V_2 (6U8). To unground this shield, it is lifted and then made to rest on the glass envelope of the tube in such a way that no contact with the bottom grounded sleeve occurs. In this position, the shield is floating and when it receives the generator signal, it capacitatively couples this to the tube elements through the glass envelope. This is a simple way to inject the signal into a circuit and it works exceptionally well.

PROBLEMS

1. In what respects is an A-M F-M tuner different from a complete A-M F-M receiver?
2. Which stages in Fig. 12.2 serve both A-M and F-M signals? Which stages serve one signal only? For the latter question, indicate which signal.

3. If resistor R_{11} in Fig. 12.2 opened up, what effect would this have on receiver operation?
4. What test instruments are needed to align the receiver in Fig. 12.1?
5. How are the F-M R.F. tuning circuits adjusted in Fig. 12.2?
6. Describe in detail the operation of the R.F. amplifier in Fig. 12.4.
7. What features exist in the Heath F-M tuner that are not found in the Granco receiver?
8. Describe the complete alignment procedure for the Heath tuner.
9. What purpose is served by the following components in Fig. 12.4: R_2 , C_3 ; R_3 ; R_4 , C_4 ; R_5 .
10. What stages in Fig. 12.4 are controlled by the A.V.C. voltage? Where is the A.V.C. voltage obtained?
11. How is F-M receiver sensitivity usually stated? Explain the meaning of this notation.
12. Draw a block diagram of the tuner shown in Fig. 12.5.
13. What is the purpose of the A.F.C. stage in Fig. 12.5? Where does this stage obtain its regulating voltage? Trace out the path.
14. Explain how the tuning meter in Fig. 12.5 operates on A-M and F-M.
15. Indicate the function served by each section of selector switch S_1 in Fig. 12.5.
16. Describe the F-M alignment procedure for the tuner in Fig. 12.5.
17. What is the purpose of C_7 , L_7 in Fig. 12.5? How does it achieve this?
18. Describe the operation of the F-M detector used in Fig. 12.5.
19. What form of A.V.C. is employed in the A-M section? F-M section?
20. Identify all of the R.F. and I.F. tuning circuits associated with the F-M section of the receiver in Fig. 12.5.
21. What can the receiver in Fig. 12.8 do that none of the tuners could? What stages enable it to do these additional functions?
22. Identify the loudness control circuitry in Fig. 12.8. What purpose does this control serve?
23. Explain how the treble and bass controls operate in the receiver of Fig. 12.8.
24. Where is the record equalizing network found in the circuit of Fig. 12.8? Identify the components of this network.
25. Identify the following, using Fig. 12.8.
 - (a) The section of switch S_1 which determines whether B+ is applied to the F-M section or the A-M section.
 - (b) The components of the de-emphasis network at the output of the F-M detector.
 - (c) All of the circuit components of the A-M local oscillator.
 - (d) Every tube in the receiver which has audio current flowing through it.
 - (e) The stabilizing components in the F-M detector.
26. Indicate how you would visually align the F-M I.F. and detector stages in Fig. 12.8.

Chapter 13

SERVICING F-M RECEIVERS

Extension of A-M Servicing Methods. The servicing of F-M receivers conforms to the same basic procedures currently in use for A-M sets but modified sufficiently to meet the differing characteristics of the F-M receiver. A comparison of corresponding stages in both types of receivers shows that, except for the second detector and limiters (if any), there is no *functional* difference between them. Remember that we are concerned at the moment only with the function of each stage and not its design. From the serviceman's point of view, function is all-important and design is secondary. The serviceman, called upon to repair a set, is interested only in what each stage does in order that he may properly apply his servicing instruments.

The F.C.C. has assigned F-M to the higher frequencies, 88 to 108 mc. When the serviceman, who has long been accustomed to working with receivers at the considerably lower A-M broadcast frequencies, is confronted with an F-M receiver, he will find himself considerably handicapped, unless he is familiar with the modified operation of radio components at frequencies of 100 mc. Let us, therefore, examine the behavior of the most common radio components as the signal frequency rises to the region of 100 mc.

Resistance: At the low frequencies, the resistance of a conductor is given by

$$R = \rho \frac{l}{A}$$

where R = resistance of wire, in ohms

ρ = specific resistivity

l = length of conductor

A = cross-sectional area of conductor

As the frequency of the current through the wire increases, it will be found that the resistance offered by the same length of wire will also in-

crease. To understand the reason for the increase, let us consider what occurs within a length of wire when a current flows through it. It is known that current flow has associated with it a magnetic field or, what is the same thing in this instance, circular magnetic lines of flux. These are everywhere encircling the current. The definition of inductance depends upon these flux linkages and is expressed by the formula

$$\text{Inductance (henries)} = \frac{\text{Flux linkages encircling conductor}}{\text{Current producing these linkages (in amps)}} \times 10^{-8}$$

Consider now the end view of a small, round section of wire that has a current flowing through it (see Fig. 13.1). Each small section of current flowing

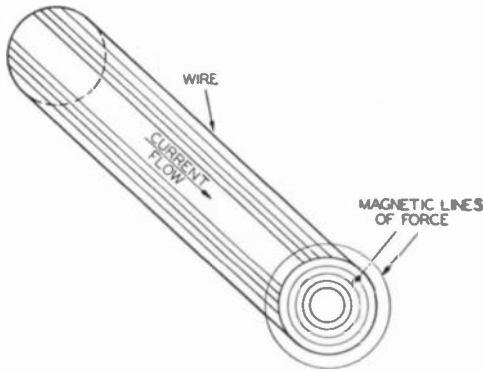


FIG. 13.1. Current flow through a wire and the magnetic lines of force encircling the wire.

through the wire has magnetic lines of flux encircling it, but the sections of current at the outer edges of the wire have fewer lines of flux around them than the currents at the center of the wire. This is because the flux produced inside the wire by the central currents does not encircle the outer currents and hence does not influence their flow. The flux produced by the currents flowing at the surface of the wire does, however, encircle the currents at the center and, consequently, exerts an influence upon them. From the definition just noted for inductance, it is seen that, since more flux linkages encircle the center of the wire, the inductance of the wire at the center will be greater than that at the surface. As the frequency of the currents increases, the inductance at the center of the wire offers more opposition (reactance) than the outer sections of the wire where there is less inductance. Hence, the current, seeking the path of "least resistance," will tend to concentrate more at the surface (or skin) of the conductor as the frequency rises. This concentration has, in effect, reduced the useful cross-sectional

area of the conductor. The resistance, due to this decrease in effective area, will rise, a phenomenon generally known as "skin effect."

Another fact of interest to the serviceman is that, at the high frequencies, the equivalent circuit of so-called "pure" resistances becomes as shown in Fig. 13.2. The resistive and inductive effects of resistor are due, of course, to the previous explanation. The capacitance arises from the capacitance between the terminals and the capacitance between parts of the resistor itself.

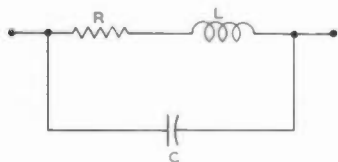


FIG. 13.2. The equivalent circuit of an ordinary resistor at the higher frequencies.

Inductance: Inductance changes slightly from low to high frequencies, due again to the effect of the flux linkages and the currents. As the frequency goes up, the distribution of the currents in the wire will change and, because inductance depends upon this distribution, it too will change. The result is a slight decrease.

At low frequencies, those employed for A-M sound broadcasting, the inductance of the tuning coils is comparatively high. Hence, a wire several inches long that might be used to connect the coil to a tube or capacitor would not add any significant amount of inductance. So long as the coil inductance is in the henry or millihenry range, no ordinary length of connecting wire will add sufficient inductance to alter the coil characteristics appreciably. When the signal frequencies reach 100 mc, however, the inductance of the tuning coils approaches the point where even the small connecting link of wire can become important.

Note, then, that connecting wires become important at the higher frequencies, not because their inductance changes to any appreciable extent, but because the important components in the circuit become smaller. Even such things as the plates of capacitors introduce inductance into the circuit, and we find that trouble arises because of the interaction of the magnetic fields set up between ganged capacitors connecting to different circuits. It is common practice to shunt paper and electrolytic condensers with small mica capacitors in order to neutralize any inductance existing in these units.

Capacitance: The increased importance of stray capacitance in a high frequency circuit can be attributed to the same reasons as the importance of connecting wires. The stray capacitance does not increase, but rather the desired capacitance decreases, hence, assigning a more prominent role to any stray capacitance. In addition, there is leakage conductance which must be considered together with capacitance. Capacitors and insulators may, because of certain conditions, allow small amounts of currents to flow through them. They act then as high resistances, or poor conductors. The

conductance is low and is labeled "leakage conductance." This has a direct influence on the type of insulating material which can be used successfully in high frequencies circuits, especially for tube sockets. Moisture and other atmospheric conditions may reduce considerably the effectiveness of these insulators at the higher frequencies and the result is poor operation of the set.

Trouble Shooting Procedures. The serviceman, when faced with the problem of trouble shooting or repairing an inoperative F-M set, will find that all his experience gained from servicing A-M receivers is applicable here, too. Thus, for example, signal tracing, starting from the speaker and working forward, is the best method of servicing an F-M receiver just as it is for any A-M circuit. In addition, despite the fact that an F-M receiver is designed for the reception of F-M signals, A-M signal generators of the proper frequency range will serve for the signal tracing. The alignment procedures for the commercial models described earlier included alignment with an A-M signal generator. If a set can be aligned with an A-M signal generator, it most certainly can be tested with one.

Following recognized procedure, after the defect has been localized in a particular stage, voltage and—with the power off—resistance and continuity checks will lead the serviceman to the faulty component. The only real difficulty which the serviceman will encounter and one which is due to lack of experience is the localizing of defects in the R.F. stages at the front end of the set. Component failure, here, especially if one of the coils or other R.F. components are faulty, will prove especially troublesome. However, if the serviceman is familiar with the basic operation of high frequency circuits, as outlined in the preceding section, and uses all special instructions issued by the manufacturer, he should be able to surmount successfully any difficulties that may arise. At all times, in replacing components and/or connecting wires, duplicate exactly what has been taken out. It is a common practice among present-day servicemen to replace faulty components with whatever part happens to be on hand. Although this procedure may work in A-M receivers, it will be less successful in F-M sets. This is also true of television receivers, which function at the same approximate band of frequencies.

Preliminary Tests: In receiving a set which is inoperative, there is seldom any reason for the serviceman to suspect that the tuned circuits are out of alignment. The more complicated receivers become—and the trend is in that direction—the less likely it is that someone owning a receiver will venture to meddle with it to the extent of adjusting the circuit alignments. Possibly the only meddling that is done nowadays is to remove the tubes to have them tested at a near-by radio store. Here, the possibility of improper replacement is quite likely and it may save the serviceman hours of work if he first checks the tubes and the sockets they are in.

As a routine measure, the serviceman should check all tubes in the set.

Tube failure is one of the major reasons for sets becoming inoperative. It is also one of the easiest checks to make.

Many, if not all, of the present F-M receivers are combination A-M and F-M sets. This gives the serviceman another means of localizing trouble in the receiver. If it is found, for example, that the set is perfectly normal on the A-M bands, but inoperative when the switch is turned to F-M, then the trouble must be contained in some part of the receiver which is exclusively used for the F-M. It indicates, for example, that the audio amplifiers and the power supply may be eliminated immediately. Again, since most sets (for the sake of economy and compactness) have but one set of I.F. amplifiers, the chances are good that these are also in working order. It may be, of course, that one of the F-M-I.F. coils is shorted, but this is not a common occurrence and may, at first, be discounted. Thus, through the simple process of testing the set on A-M, we have eliminated anywhere from four to six tubes and as many stages. A glance at the circuit diagram of the set will show which circuits are common to both signals and which are not. Combination receivers are becoming increasingly numerous because of the many popular types of services available. The additional circuits may make the overall circuit more complex and yet, on the other hand, they can make the task of trouble shooting easier. A common defect will show up in all circuits; a special defect will be confined to one particular path.

Phonograph attachments are common in modern F-M receivers. These phonographs generally feed through the entire audio section of the set. As a simple test of the audio system, the serviceman has only to touch the phonograph needle and note whether any sound is heard from the loudspeaker.

Visual Check. An important step in the servicing of any electronic circuit is a careful visual inspection of the unit. Do this first with the power off and, if nothing wrong is found, with the power on. Here are some of the things to look for:

1. Are all the tubes lit or is one or more unlit? If all are unlit, then a fuse may be blown or the various filaments may be connected in series and one set of filaments is open.
2. Are there any charred components, particularly resistors?
3. With the power on, do any of the components overheat or smoke?
4. What effect does rotation of the various front panel controls have (with the power on)?
5. If printed circuitry is employed, are there any visible breaks in the wiring?

These are but a few of the questions that a serviceman might ask himself as he visually examines a receiver. Additional questions will undoubt-

edly suggest themselves. Make as thorough an investigation as you can; you will find that it will frequently assist you materially.

General Servicing Procedure. For the purposes of testing a receiver to determine the cause and location of any defect, it is helpful to divide the set roughly into the following six categories:

1. Power supply.
2. Audio amplifiers.
3. F-M detector or discriminator.
4. I.F. system
5. H.F. oscillator and mixer.
6. R.F. section of receiver.

The foregoing classification is true for all superheterodyne receivers. Since no TRF types of commercial F-M sets are being manufactured at present, the analysis will apply to all receivers that the serviceman is called on to repair.

Power Supply: Trouble in the power supply produces either a completely inoperative set or else low volume in the receiver output. Under these conditions, the first place to test is the B+ terminal of the power supply. A partial short-circuit in any portion of the receiver may result in an excessive current drain on the power supply and lower the B+ voltage considerably. The audio output becomes weak and distorted. A total short-circuit will burn out the rectifier tube and cause any protective fuses to blow. Never replace a blown fuse until the cause of the damage has been determined. A hoarse, raspy audio output is another indication of trouble in the power supply, this time a faulty, electrolytic filter capacitor. This should not be confused with a scratchy output, which is a result of an incorrectly centered speaker cone.

Audio Amplifier System: The simplest and most rapid method of testing the audio amplifiers is by applying an audio signal and noting whether an output is obtained. With the same strength signal, the output should become louder as the test voltage is moved back away from the speaker. The output will be loudest when the signal is applied to the grid of the first audio amplifier. A more sensitive indicator is an output meter connected across the voice coil leads.

F-M Detector or Discriminator: The various types of commercial F-M detectors currently in use were analyzed in Chapter 9. To trace an A-M signal through the detector, we must connect the indicator (a voltmeter or ammeter) in such a way that response to an A-M signal is possible. In the diagram of a Foster-Seeley discriminator, Fig. 13.3, connection of the meter across either load resistor will result in a deflection if the A-M signal voltage

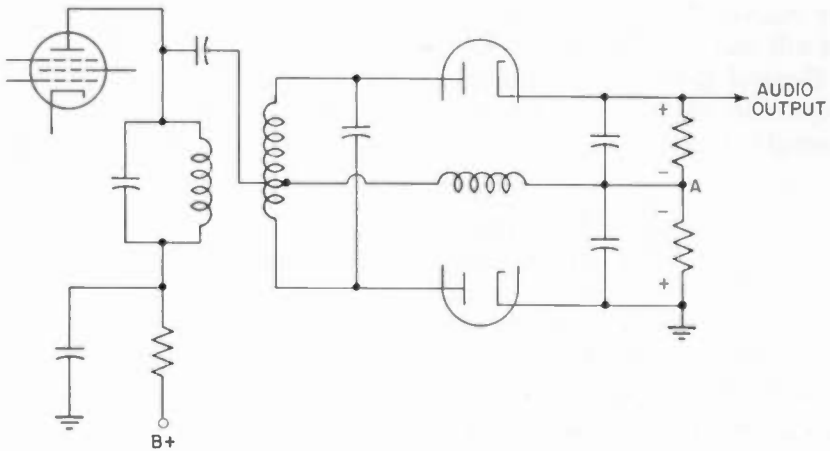


FIG. 13.3. A Foster-Seeley F-M discriminator.

is being passed through the circuit. Each half of the discriminator is responsive to A-M and, when the circuit is functioning properly, the d-c voltmeter connected across either resistor will show a reading. It is usually most convenient to connect one end of the voltmeter to ground and the other end to the junction of the two load resistors, such as point A, Fig. 13.3. The ground terminal of the signal generator is connected to the receiver chassis. The output lead is connected through a 0.05-mf capacitor to the plate or grid of the preceding limiter. If the frequency of the signal is varied, by rotating the knob of the signal generator, the meter deflections will increase and decrease.

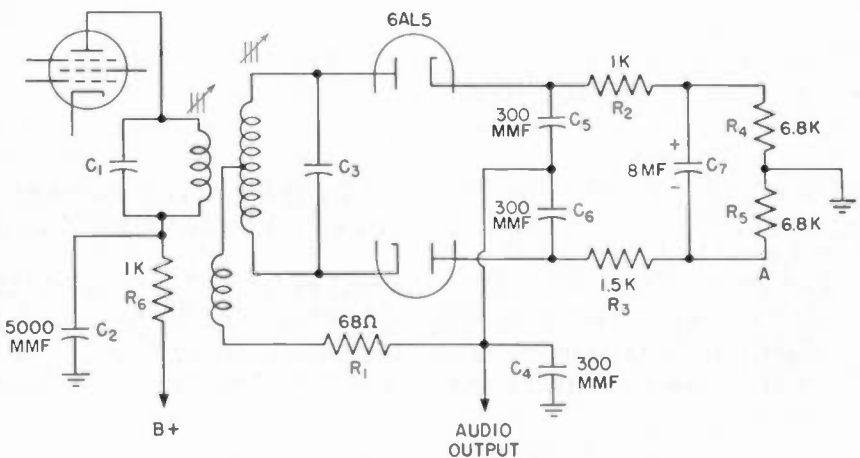


FIG. 13.4. A balanced ratio detector.

The linearity of the detector to F-M signals can be determined by connecting the voltmeter across both load resistors and observing whether equal (and opposite polarity) deflections are obtained for frequencies located equal distances above and below the I.F. mid-value. The linear response should extend ± 75 kc from this I.F. point.

In the ratio detector, reproduced in Fig. 13.4, the voltmeter is connected between the point marked A and ground. Since point A is negative, the negative lead of the d-c voltmeter is placed here. The signal generator is then attached between the grid and ground of the last I.F. amplifier. If the circuit is operating normally, the meter will show a deflection. As the amplitude of the signal is slowly varied, the voltmeter will follow the changes in step. In addition, if the signal generator is set at the mid I.F. value and then varied from 75 kc below this point to 75 kc above, the indicator will show only a small change. This is true, providing the circuit is correctly aligned and the voltage output of the signal generator does not change with frequency. If either diode is inoperative, there will be no meter deflection.

It is well to remember that if the secondary of the F-M detector (or discriminator) is detuned, distortion will result. If the primary is detuned, the signal output will be weakened, usually accompanied by some distortion, also.

F-M Detector Checking with Sweep Generator. For those service shops that possess sweep generators, the F-M detector may be checked by observing its response or S-curve. Connect a sweep signal generator to the grid of the tube preceding the F-M detector. Set it to sweep 450 kc above and below the I.F. Connect an oscilloscope across the output terminals of the detector. The appearance of the S-curve on the oscilloscope screen will tell the complete story concerning the operating condition of the detector. When the transformer coupling network connecting the F-M detector to the previous stage is not properly tuned or centered, the S-curve will appear either as shown in Fig. 13.5A or B. In either case, the sound output will be

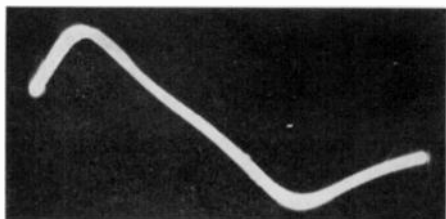


FIG. 13.5A. Secondary tuning core adjustment of discriminator transformer improperly set.

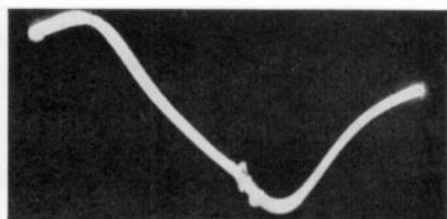


FIG. 13.5B. Primary tuning core adjustment of discriminator transformer improperly set.

distorted. To correct this condition, adjust the iron cores of the F-M detector primary and secondary windings. The primary controls the linearity of the S-curve, and the secondary adjustment governs the centering of the S-curve.

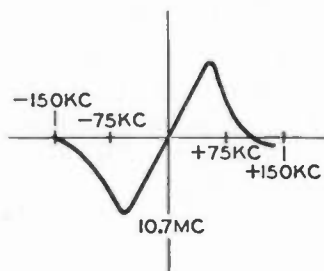


FIG. 13.6. Bandpass of discriminator too narrow.

Another defect made evident by the S-curve occurs when the bandwidth of the tuning circuit is too narrow. (See Fig. 13.6.) When this happens, the output becomes distorted when the incoming signal is fully modulated. The sounds become "hissy" or "raspy." The solution for this, assuming that the circuits were not originally designed this way, is to realign them.

Whenever the complete response curves are viewed on the oscilloscope screen, marker signals obtained from a separate A-M signal generator should be used to identify the mid I.F. and the end frequencies of the linear section of the curve.

In the Foster-Seeley discriminator the two diodes should be matched fairly closely. If one discriminator diode has low emission, the S-shaped curve will now assume the appearance shown in Fig. 13.7. This condition results in poor noise rejection at low signal inputs and sound distortion at high signal inputs. When both discriminator diodes have low emission, the sensitivity of the receiver is poor.



FIG. 13.7.

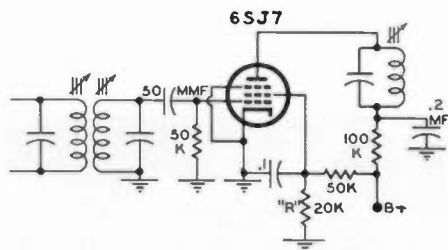


FIG. 13.8. If "R" should open up, the limiter may become noisy with moderate signals.

Checking Limiters. When a receiver that has been operating normally suddenly becomes noisy, the discriminator tubes and alignment should be checked. If the receiver contains a limiter stage, this too should be checked. Limiters become noisy when their grid-leak circuits open, or the voltages applied to the electrodes rise above normal. In many circuits, the low voltages for the screen and plate are obtained by the voltage divider arrangement shown in Fig. 13.8. If *R* should open, the voltage applied to the screen will rise. Under these conditions it will require considerably

more adjustment to bring the receiver back to normal.

stronger signals to bring the tube to saturation, and the limiting action of the tube is impaired. Ordinarily, most limiter tubes will saturate with signal voltages of 2 volts or more. The best method of testing the limiter is to measure its grid, screen, and plate voltages and compare the values obtained with those specified by the manufacturer.

A noisy ratio detector can usually be traced to an open electrolytic capacitor. This capacitor, together with its shunting resistor, stabilizes the ratio detector against amplitude modulation of the received carrier. If the capacitor becomes defective, noise will ride through and be heard. The serviceman will find that sets using the ratio detector will receive weaker signals with less background noise than the same signal in a set possessing the Foster-Seeley discriminator. In the latter receiver, noise disappears only when the preceding limiter is driven to saturation, but no such "threshold" effect is apparent in the ratio detector. However, in all instances, when the noise or interference is stronger than the signal, the noise will prevail.

I.F. System: The placement of the indicator when the I.F. system is tested will depend upon the type of F-M detector used. For the conventional discriminator, which has one or two limiters preceding it, it is possible to place the indicating meter in the grid circuit of the last limiter, if there are two, and test the I.F. system. The signal generator is connected to the grid of the mixer tube, and the frequency of the signal set at the I.F. If all the intervening stages between the mixer and the limiter (where the meter is located) are in proper operating condition, the meter will deflect. If no indication is obtained, the signal generator will have to be brought back to the grid of the first tube preceding the limiter. In this way, only one stage is under test. If the meter deflects, indicating that the stage under test is functioning properly, the generator is moved back one stage and the signal applied again. In this manner, we gradually work back to the point where the trouble exists in the I.F. system. However, as a preliminary test, the signal is fed through the entire I.F. system, since if no trouble exists in this portion of the receiver, valuable time is not spent testing each stage.

While testing the I.F. system, it is possible also to determine whether the circuits are in alignment. Maximum deflection of the meter should be obtained when the test signal is at the I.F. value. Shifting the frequency an equal number of kilocycles above and below this point should produce equal readings on the meter.

Note that the foregoing procedure is followed only for F-M sets employing conventional discriminators with their attendant limiters. In receivers possessing the ratio detector and in which no limiter is employed, the voltmeter is connected to the point where the A.V.C. voltage is obtained from the detector. The positive lead of the meter is attached to the chassis. Any signal fed into a properly operating I.F. system will result in a voltage at this point in the detector.

The I.F. system of present F-M sets is generally a combination of A-M and F-M transformers with a single tube for each stage serving both A-M and F-M. With the exception of the I.F. switch and the F-M, I.F. transformers, parts giving normal indications on A-M will test normal on F-M. Hence, if the I.F. system is found to be in operating condition on A-M but not on F-M, it would be best to check these particular components first.

R.F. Section: To test the R.F. stages of the receiver, we still retain the A-M signal generator, provided its range extends to the necessary frequencies. Again, as before, the position of the indicating meter will depend upon the type of F-M detector in the circuit. The explanation given above will apply equally here. For the signal tests, use an R.F. signal generator with an amplitude-modulated output. The ground lead of the signal generator is connected to the chassis ground; the output lead is connected through a 0.05 mf-capacitor to each of the test points, as explained below.

The input of the R.F. section of the receiver consists of an R.F. amplifier, a mixer, and an oscillator. In some sets, no R.F. stage is included; in others, such a stage is employed only for the F-M signals.

In the testing of an F-M receiver, the input circuits are tested last. Hence, if the foregoing step-by-step procedure has indicated that all of the other sections of the receiver are functioning properly, then, in the testing of the R.F. circuits, it is best to start at the oscillator. The reason is that the oscillator is the one stage, in the front end of the receiver, which is most critical in operation and most likely to be at fault. By measuring the voltage between the control grid and cathode of the oscillator, we can determine whether the stage is functioning. Measure this voltage with a high-resistance voltmeter, otherwise the value read can be misleading. The voltage at the control grid (with respect to the cathode) may have any value between -2 volts and -10 volts. Consult the service manual issued by the manufacturer for the proper value. Too low a value may be due to a bad tube or lowered operating potentials. Measure the plate and screen voltages and, if these are within the correct range, change the tube. The oscillator grid voltage should be measured on all bands since it is very possible that the oscillator will function on some frequencies and not on others. Reasons for this may be a defect in the switching system or the plain fact that a tube will oscillate readily at the low frequencies and yet fail to do so at the higher frequencies.

If the oscillator is operating satisfactorily, place the signal generator at the signal grid of the mixer tube. Adjust the front dial of the receiver and the signal generator for the same frequency setting. If the set, at this point, is operating satisfactorily, the indicator should show a deflection. If no response is obtained, change mixer or converter tubes, measure voltages at the electrodes and/or make resistance checks to locate the defective component.

Once the defect has been traced to a single stage, a routine analysis will, in almost all instances, reveal the trouble.

In probing around the underside of the chassis, the serviceman must be careful not to rearrange any components—especially small coils. It is impossible to emphasize too strongly the criticalness of these high frequency circuits, a fact which is usually fully appreciated only after such a circuit has been built or rearranged and it becomes inoperative. Many of these circuits, especially the oscillator, contain compensating capacitors. Their purpose is to compensate for changes in the electrical characteristics of other capacitors and/or the coils. If one of these special capacitors becomes defective, i.e., open, the circuit is detuned. If the serviceman is unaware of this situation, he may conclude that the circuits merely require alignment and rectify the situation by readjusting the circuits. For the moment, the set will function. However, after a short while it will be found that the set drifts during each warm-up period, and this drift may easily be sufficient to shift the circuits out of range of the signal. The station can be brought in by manually retuning the set—but this will have to be done at least once each time the set is turned on and possibly more.

No special mention has been given to the testing of any R.F. amplifiers since the procedure is identical with what has already been given. Generally, instead of feeding the signal directly at the grid of the R.F. amplifier, the signal is fed in at the antenna terminals. The method for connecting the signal generator, especially with F-M receivers possessing balanced input, is identical with the methods outlined for the alignment of the receivers.

A set, in transit from the factory to the retailer and from the retailer to the customer, is often subjected to many jarring blows and knocks that can cause wires to short together, poorly soldered connections to break, trimmer capacitor plates and iron core slugs to bend and stick, plus a host of other damaging mishaps which will render a set inoperative. Sometimes these happen to a set when it is moved about in the home, although generally sets are handled fairly carefully by their owners.

It may be instructive, before we end this chapter, to examine some typical troubles encountered in F-M receivers which often baffle the inexperienced serviceman.

Case 1. An A-M, F-M receiver was operating satisfactorily on A-M and phono, but was dead in the F-M position. This set used the same tubes for both A-M and F-M, and consequently the tubes, as a source of trouble, could be eliminated. The audio system was similarly eliminated since the phonograph fed its signal through this system. Checking the oscillator grid voltage disclosed that it was zero in the F-M position. Obviously, the F-M oscillator was not functioning. In checking the various components in this cir-

cuit (coils, resistors, and capacitors) it was found that the plates of the trimmer capacitor used in this circuit were shorting together through a cracked mica sheet which served as the capacitor dielectric.

By replacing the trimmer capacitor and realigning the oscillator, the set returned to normal operation on F-M.

Case 2. It should be fairly evident by this time that high frequency circuits such as we find in the front end and in the I.F. stages of an F-M receiver are more critical than low frequency circuits. Rough handling in shipment have been the cause of poor sensitivity of many an F-M set. If the output of a set is weak, check the alignment of the circuits. With a signal generator and a VTVM, this job should not take more than 15 to 20 minutes and it may save hours of needless search for a nonexistent defective component.

Case 3. A receiver operated normally on A-M and F-M, but was intermittent or hummed constantly on phono. In A-M or F-M sets, the phono is usually connected into the circuit just ahead of the volume control. This control, in turn, attaches to the control grid of the first audio amplifier. Since the hum appeared only in the phono position of the selector switch, the trouble must have existed between the phono input and the volume control.

In the present instance it was found that the trouble was due to a poorly soldered ground connection on the phone jack. With the ground lead off, the phono was floating, picking up some of the stray 60-cycle voltage present in nearly all sets. These stray 60-cycle fields are due to poorly shielded power transformers or filament wires that extend into every section of a set wherever a tube is located.

Poorly soldered ground connections cause servicemen no end of grief. Whenever a set is troubled with intermittent operation, hum pickup, or noisy operation, and the defect is difficult to locate, take 10 minutes or so to go over many of the soldered connections in the set with a hot soldering iron. Check, too, at the same time to see whether any bare wires are accidentally touching each other or the sides of the chassis when they are not supposed to be grounded. The time thus spent will be well worth your while.

To illustrate the foregoing remarks, here is a case that is periodically encountered. The set becomes noisy, especially at the high frequency end of the dial and when the volume is turned up high. The noise is somewhat like the sizzling sound obtained by frying butter in a pan.

The source of this noise can usually be traced to an ungrounded speaker when the latter unit is not attached directly to the receiver chassis. To remedy this, run a wire from the speaker casing to the chassis, soldering the wires at both ends to insure proper electrical connection.

Case 4. Many manufacturers of A-M, F-M sets place both I.F. coils within the same can. Where space is limited, bare wires running from the various transformer windings to the connecting lugs may touch each other, shorting out one or both I.F. windings. When this happens, reception on either band (or both) will depend upon which winding is shorted out. Thus, if the A-M windings are shorted, there will be reception only on F-M; no A-M signals will be heard. When the short occurs across the F-M winding, reception on this band may be either totally absent or weak. In spite of the short, some reception may be obtained because the shorting wire may contain enough inductance at 10.7 mc to permit some signal to develop across it and be passed onto the next stage. At the A-M, I.F. frequencies, however, the short effectively cuts out all signals.

Case 5. A certain A-M, F-M receiver was found to develop an intermittent bad hum and sometimes the set became completely inoperative. In this set, all filament and ground returns in the R.F. section are made through a single ground strap that connects the R.F. section to the chassis. This is frequently broken in shipment or in moving the set around. To remedy the condition, replace the strap with flexible copper braid.

Case 6. Some F-M discriminators have a small capacitor connected between the top of the primary winding of the discriminator transformer and the center tap on the secondary. This was noted in the discussion on discriminators. If this capacitor should become shorted, B+ will be placed on the plates of the discriminator diodes. The result may be either a completely dead set or the appearance of "motorboating." Since the capacitor is generally not readily accessible, the only positive cure is replacement of the entire transformer.

Case 7. In almost every receiver, small 100- or 220-mmf by-pass capacitors are connected from grid to ground or plate to ground of the first audio amplifier. The purpose of these units is to prevent any I.F. voltage from entering the audio amplifier system. When one of these capacitors shorts out, no audio output will be obtained. When they open up, the set frequently becomes noisy.

To localize this noise on a set just received, proceed as follows. Ground the grid of the final output amplifier. This will eliminate the noise. Next, ground the grid of the previous (generally the first audio) amplifier stage. The noise now will be heard, indicating that it is arising in the stage under test.

Case 8. Low volume and poor or fuzzy tone can usually be traced to insufficient voltage being applied to the audio amplifiers. Electrolytic capacitors that have excessive leakage, rectifier tubes with poor emission, or weak selenium rectifiers will produce these symptoms. Especially sensitive

to voltage drops is the local oscillator. Make certain when checking a dead set that the voltages are within 10 per cent of those specified by the manufacturer. Remember, too, that when an oscillator is operating properly, it will develop negative biasing voltages from 2 to 10 volts, depending upon the circuit. Check for this voltage.

Decrease in voltage has been noted quite frequently in sets using selenium rectifiers, and the best solution to this trouble is replacement of the unit.

In some districts, the a-c voltage of the local power lines is subject to fairly wide fluctuations during the 24-hour period of a day. If a set is sensitive to these variations, especially voltage reductions, it can (and will) happen that the oscillator within the set will cease to function when the line voltage is low. A complaint of set operation only during certain hours of the day can be checked for this cause in the shop by connecting a Variac between the set and the power line. Gradually lower the line voltage to the set and observe at what input voltage the set ceases to operate. Usually, most sets will continue to function down to line voltages of 100 volts. Some, however, will die out before this value is reached, and the foregoing complaints are heard with these latter sets. Larger filter capacitors, lower-valued filter resistors, or a change in rectifier tubes are generally helpful in such cases.

Conditions will sometimes arise, when checking the various voltages in a set, that lead to a false conclusion regarding the component at fault. A case that actually occurred concerned a certain receiver that developed considerable distortion when in operation 5 minutes or longer. A voltage check of the power tube showed a positive voltage at the control grid. The first thought was a leaky or shorted coupling capacitor between the plate of the preceding tube and the grid of the power tube. However, replacement of this capacitor did not correct the condition. The fault was found to lie with the tube itself and replacing the tube cleared up the trouble.

A similar instance involved a set which possessed a weak, distorted output followed, after several minutes, by a completely dead output amplifier. Checking the audio amplifier system revealed that the output amplifier did not possess any cathode voltage, indicating that the tube itself was not conducting current. A further test showed continuity in the cathode circuit. Voltages on all the other elements were in order, yet the tube did not seem to be drawing current. It was found that the 500,000-ohm volume control across the control grid of this stage showed continuity only at certain positions of the movable arm.

What was happening here was this: With the grid circuit open, electrons collected on the grid and, having no path through which to escape, soon biased the tube to cut-off. When the set was turned off, the movable arm

on the control moved down the resistor and, since the lower portion of the volume potentiometer did show continuity, the electrons which had collected on the grid were able to leak off. When the set was turned on again, some audio output was usually able to pass through the tube until it once again became blocked by the accumulation of electrons on the grid when the volume control arm was turned up.

Case 9. A series of troubles are encountered periodically that are due to the development of leakage paths between components that should normally be isolated from each other. These leakage paths permit voltages from one circuit to reach other circuits from which they would normally be excluded. The results in several instances are given below:

(A). A set was noisy, with the signals weak and fuzzy. Trouble was traced to leakage between the primary and secondary fixed capacitors of an I.F. transformer, which were imbedded in wax, and enough dirt and moisture had collected to permit a leakage path to be formed between the two units. The faulty condition was remedied by cleaning off the wax.

(B). Receiver had a loud hum when a station was tuned in. The control grid of the R.F. amplifier was connected to a 3.9-megohm resistor through which it received A.V.C. voltage. Apparent corrosion between the wafers of the tube socket allowed a 500,000-ohm to one-megohm leakage resistance to form between the various pins and the center terminal which is grounded.

Cleaning off the corrosion may help although, if the corrosive action is well advanced, the only solution will be to change the socket. It is well to remember that this same thing can happen in almost any wafer socket.

(C). A set exhibited a severe a-c hum after it had been in operation for several months. Bridging the existing filter capacitors with units known to be good had very little effect. Grounding the control grid of the audio output tube removed the hum; grounding the grid of the previous audio amplifier did not affect the hum. The trouble, then, was in this stage. It was discovered that leakage between the cathode by-pass capacitor used here and one of the power-supply filter capacitors was the cause of the trouble. Both units were contained in the same case.

(D). Hum due to electrical leakage between the heater (or filament) and cathode of a tube will be encountered occasionally. The minute current that flows between the heater and cathode develops enough voltage across the cathode resistor to introduce a 60-cycle variation in the tube's current flow.

A-c, d-c sets are particularly susceptible to this source of hum because of the series connections of the filaments.

The solution to hum developed in this manner is replacement of the faulty tube. Hum can also be reduced by operating the tubes at or slightly below their normal filament voltages. Operating a tube with more than rated

voltage aggravates the problem. In high-gain audio systems, place the input tube as close as possible to the ground end of the filament chain. Also, if cathode self-biasing circuits are used, keep the value of the cathode resistor as low as possible. Keep the capacitance of the cathode by-pass capacitor as large as possible because this reduces the impedance of the cathode circuit to 60-cycle voltages. In a well-designed amplifier, fixed bias is frequently used in the first audio amplifier, permitting the cathode to be grounded directly and thereby avoiding trouble from this source of annoyance.

PROBLEMS

1. What effect does increase in frequency have on the ordinary resistor? Why is this important in F-M circuits?
2. Why is stray capacitance more important at the high frequencies than at the low frequencies?
3. How can stray capacitance be kept at a minimum?
4. What test instruments would be considered as absolutely essential for F-M receiver servicing? Explain the reason for your choice.
5. What do we mean by a visual check? What sort of things can be checked in this manner?
6. What preliminary tests should be made before any actual work of servicing is begun?
7. How is the audio system of a receiver tested?
8. Outline the method you would employ to check the Foster-Seeley discriminator and the ratio detector.
9. Would a-c hum in the power supply affect an F-M receiver? Explain.
10. How does testing of the I.F. system depend upon the type of F-M detector used?
11. Must the indicating meter always be placed in the output of the F-M detector when the I.F. system is being tested? Explain.
12. List some of the defects which could possibly affect an I.F. system.
13. How is the R.F. section of an F-M receiver tested?
14. What indication do we obtain from a normally functioning oscillator?
15. In tuning over the F-M band, a hiss is heard in the speaker. At no point, however, are any stations heard although stations are known to be operating. What is probably wrong with the set?
16. Microphonic tubes have what effect on a set?

Chapter 14

COMMERCIAL F-M TRANSMITTERS

(PART 1)

To generate F-M signals, two general methods are in use. One is the reactance method; the other, the phase-shift method. In the reactance method, the frequency of an oscillator is made to vary with modulation through the use of a vacuum tube. In the phase-shift method, frequency modulation is achieved by varying the phase of a signal obtained generally from a crystal oscillator. Chapters 14 and 15 describe commercial F-M transmitters employing reactance tubes, whereas Chapter 16 is concerned with transmitters utilizing the phase-shift system.

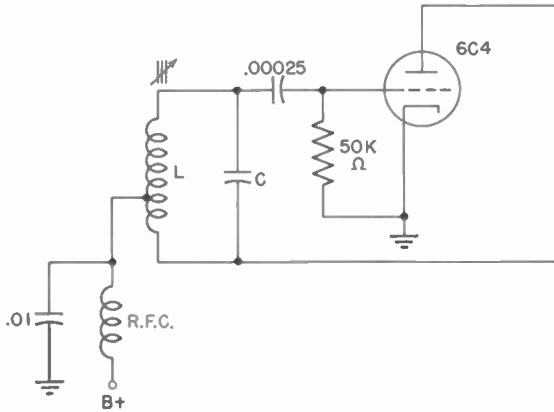


Fig. 14.1. A Hartley oscillator.

Reactance-Tube Circuit. Probably the most direct method of producing a frequency-modulated wave is through the actual variation of one of the frequency-determining components of an ordinary oscillator. Consider the Hartley oscillator shown in Fig. 14.1. In this circuit, the frequency which

the oscillator will generate is determined primarily by L and C according to the familiar formula,

$$f = \frac{1}{2\pi\sqrt{LC}}$$

By varying either or both of these components, it is possible to alter the resonant frequency. For the production of a frequency-modulated signal, we would have to have either L or C vary in accordance with the spoken word or a musical selection.

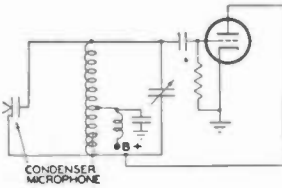


FIG. 14.2. An elementary method of frequency modulating an oscillator.

One crude and elementary method of obtaining a variation in capacitance is by the attachment of a condenser microphone in parallel with the tuning capacitor C . Such an arrangement is shown in Fig. 14.2. As long as no sound reaches the microphone, the total capacitance across the coil remains constant. The frequency due to this combination of L and C would represent the central or resting frequency. C , in this case, is the sum of the average capacitance of the microphone plus the tuning capacitance.

When the microphone is actuated by sound, the movable capacitor plate is forced to vibrate back and forth about its rest or center position. At some time, under pressure of the moving air, the capacitor plate approaches the fixed plate and the total capacitance increases. (Capacitance varies inversely with plate spacing.) As a consequence, the frequency of the oscillator decreases. Upon being released from this position, the movable capacitor plate swings past its center position and comes to rest at a point farther away from the fixed plate. Under these circumstances, the total capacitance has decreased below its average value, increasing the oscillator frequency. Thus the frequency of the oscillator varies in step with the impressed audio variations and a frequency-modulated signal is developed. True, the linearity of this converter is not very good, and the results would not be commercially satisfactory. However, it does illustrate the production of a frequency-modulated wave.

The attachment of a capacitor microphone across the tank circuit is impractical, but we can obtain a variation in either inductance or capacitance by means of a vacuum tube. As will be seen, the tube does not actually "add" or "subtract" a capacitor or coil across the tank circuit. However, because of its presence in the circuit, the effect produced is as though a variable inductance or capacitance were actually in the circuit. Consequently, the frequency of the oscillator changes. If the circuit is designed so that the frequency variations are governed directly by the audio sounds

reaching the microphone, we have the desired frequency-modulated signal.

Reactive Impedance. Before we actually examine the circuit of a reactance tube, let us review briefly the ideas of resistive and reactive impedance. Consider pure resistance first. Any physics book will demonstrate that what is called resistance is merely the ratio of voltage across a conductor to the current flowing through it. We express this by Ohm's Law, $R = E/I$. If a given voltage produces a high current flow, then we say that the conductor possesses a small resistance or opposition. If a small current is produced, for the same voltage, then we say that the conductor has a high resistance. In any event, the resistance is never measured as such, but solely as the ratio of voltage to current. Recall any resistance measurements or experiments that you have performed or seen performed. Were any of these ever made without the use of voltage and current? The answer is no. In a limited sense, then, we may consider resistance as existing only when current is flowing through the circuit. Without current and voltage, resistance loses its significance.

What has all this to do with reactance-modulator tubes? We shall see in a moment. Now let us turn our attention to inductive and capacitive reactance. We learn early in our study of radio that for many common frequencies the opposition that a resistor offers is constant. But when we insert a capacitor or an inductance into the circuit, we note that the opposition depends upon the frequency of the applied voltage. For the capacitor, the opposition decreases with rise in frequency; for the inductance, it rises directly. Here, then, more so than in the case of the resistor, we see that the concept of opposition cannot be separated from the applied voltage and its ratio to the current flowing in the circuit.

From these few facts we can draw one simple, yet highly important, conclusion. A circuit may appear (to some other circuit, for example) either capacitive, inductive, or resistive depending upon the voltage and current conditions in that circuit and *not* upon whether capacitance, inductance, or resistance is actually present. In other words, if we take a tube and so connect it that the voltage between its plate and cathode is 90° lagging the plate current, then it will appear as a capacitor to any circuit connected to this tube. Why? Simply because it is the voltage-to-current ratio and their relative phase that are the determining factors and not the mere fact that certain components have been placed in the circuit. Remember these points, for they will greatly simplify the understanding of reactance tubes.

A Reactance-tube Modulator. The fundamental circuit of a reactance-tube modulator is shown in Fig. 14.3. Basically, its function is to vary the frequency of the oscillator in step with an applied audio voltage. For the moment, the audio voltage has been omitted. However, as soon as we understand how a tube can cause the frequency of an oscillator to vary, we will incorporate an audio signal.

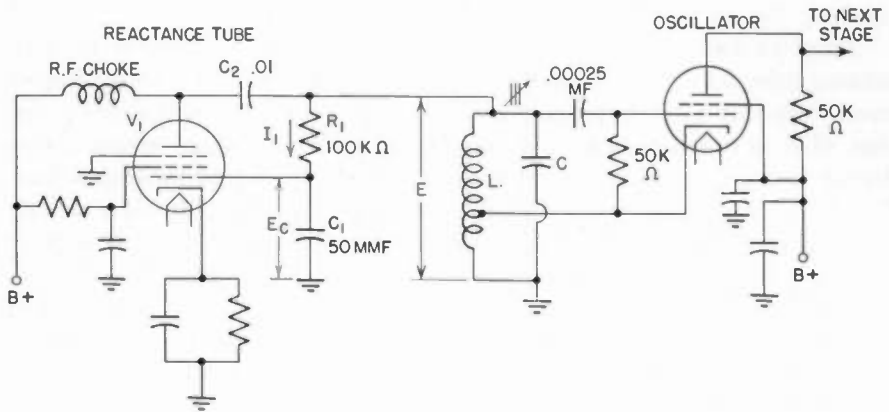


FIG. 14.3. A reactance-tube modulator.

To begin, we note that, since the reactance tube is connected across the tank circuit of the oscillator, any voltage here will also be present across R_1 and C_1 (see Fig. 14.3). If we make the resistance of R_1 much higher than the opposition of C_1 (to this voltage E), then the current that flows through the R_1C_1 branch will be governed by the resistor and be practically in phase with the applied voltage E . Using vectors, we could show this condition as in Fig. 14.4A.

Examine the voltage that is developed across capacitor C_1 . We know that with every capacitor the current and the voltage (across that capacitor) are out of phase (see Fig. 14.4B). In this instance, since the current through the series branch is practically in phase with the total applied voltage E , and, at the capacitor, the current and E_c (the capacitor voltage) are 90° out of phase, we can add E_c to the vector diagram of Fig. 14.4A to obtain the result shown in Fig. 14.4C. E_c lags the current and is placed behind it.

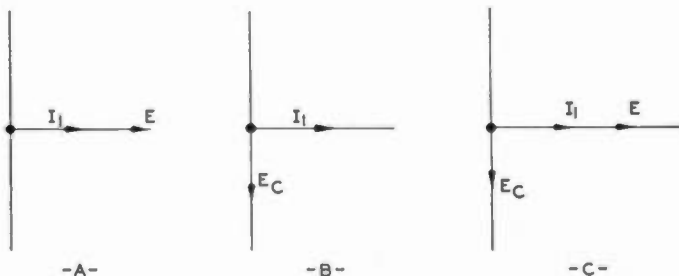


FIG. 14.4. (A) Phase relationship between oscillator tank voltage and current through R_1 and C_1 . (B) Current and voltage phase for a capacitor. (C) The phase of E_c with respect to E and I_1 .

Returning to Fig. 14.3, let us investigate the position of E_c in the circuit. By its attachment between the grid and ground, it constitutes the grid voltage. Hence, whatever phase relationship E_c might have to the rest of the circuit, the same relationship will exist for the grid voltage of V_1 . And, since the grid voltage directly controls the plate current, we can derive a relationship between E_c , E_p (the grid voltage of V_1) and the plate current. C_2 serves only to prevent the $B+$ voltage of V_1 from being grounded by L . It is a relatively large capacitance and its impedance at the operating frequency is low.

Whenever the grid becomes positive, the plate current increases; whenever it goes negative, the current decreases. Both vary directly and both are in phase. Adding the plate current

(I_p) to the vector diagram that has been developed already, we obtain the diagram of Fig. 14.5A. The plate current, being in phase with E_c , is placed 90° behind voltage E . If we look carefully at this phase relationship between E and I_p , we are immediately aware of one fact. Because of the phase relationship, I_p acts as if it is coming from an inductance. Since I_p comes from the tube,

then as far as voltage E is concerned the tube possesses inductance and presents inductive reactance to the outside circuit. Thus, through the peculiar combination of R_1C_1 and the tube, we are able by a simple procedure to have the plate current of the tube lag the oscillator tank voltage E and appear inductive.

To progress one step further. Voltage E is developed across L and C of the oscillator tank circuit. If we plotted the relationship between the voltage to the current flowing through the inductance of the tank circuit, it would appear as shown by the vectors in Fig. 14.5B. E leads the inductive current of the tank circuit in exactly the same manner as it leads the plate current of V_1 . Again the similarity of the tube to an inductance is evident.

Because of the placement of V_1 across the oscillator tank circuit, I_p must flow through the oscillator coil and add to the inductive current already flowing through the coil L . That it will add is due to the voltage E . This voltage is present across both the tube and the tank coil. Since both inductive currents arise from the same voltage, they will be in phase. The result, due to the presence of the tube, is the same as though another inductance had been placed across the tank circuit. Actually, the tube represents this added inductance, arising from the plate-current phase relationship to the tank voltage. An equivalent circuit is shown in Fig. 14.6.

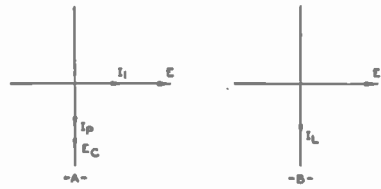


FIG. 14.5. (A) The phase angle between E and I_p of V_1 . Its resemblance to the phase relationship between E and I_L in the oscillator tank circuit (B) is evident.

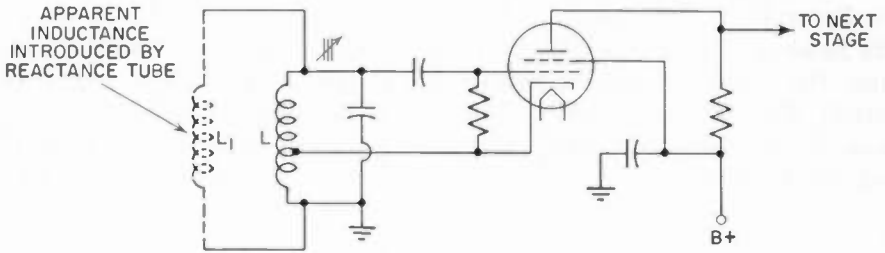


FIG. 14.6. Equivalent circuit of Fig. 14.3.

When coils are placed in parallel, their total inductance becomes less than each individual inductance. From the formula for frequency

$$f = \frac{1}{2\pi\sqrt{LC}}$$

it is evident that decreasing L will produce a higher frequency. Hence, by the addition of a tube across the tank circuit we have succeeded in shifting the oscillator frequency. The new frequency now represents the new resting point of the oscillator. No other change will occur as long as the operating voltages remain steady.

Producing F-M. Thus far we have succeeded only in moving the frequency of our oscillator from one point to another. There has been no frequency modulation. This is to be expected since no audio voltages have been applied. This is done in Fig. 14.7, which is essentially the same as the

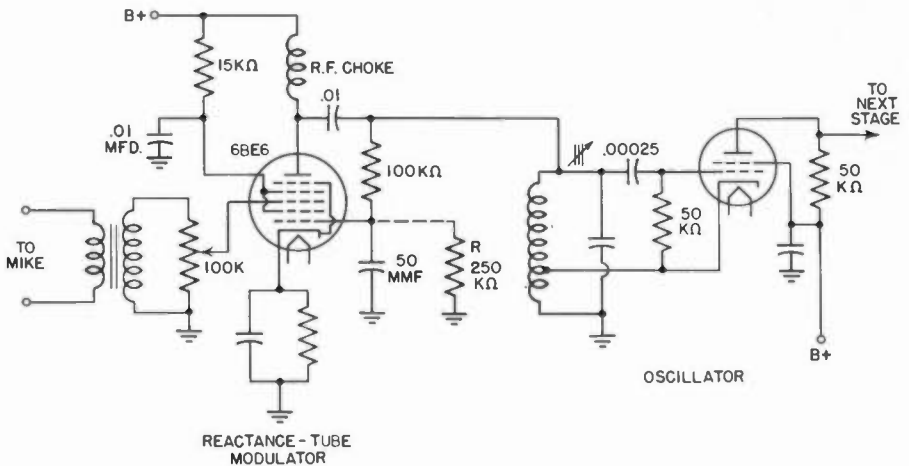


FIG. 14.7. A complete reactance-tube modulator. R may be connected across the 50-mmf capacitor to provide a d-c path to ground for any stray electrons trapped on the grid.

diagram of Fig. 14.3. However, by employing a pentagrid tube we obtain a separate grid (grid 3) for the insertion of the audio voltage. Grid 1 causes the plate current to appear inductive to the oscillator tank voltage E .

When no audio voltage is being impressed on the tube, the plate current is steady. Upon the application of an audio voltage, the plate current is varied in step with the voltage. On positive half cycles, grid 3 becomes positive, allowing more electrons to flow. On the negative audio voltages, the plate current decreases below its normal value. The rapidity with which the current rises and falls is entirely governed by the frequency of the audio voltages. People with high-pitched voices will produce faster variations in the 6BE6 plate current than persons having low-pitched voices.

At the oscillator, these variations in plate current (I_p) will have the effect of changing the amount of inductive current flowing through the coil L . We know that the oscillator tank voltage remains constant. Nothing has been done to change it. Therefore, with the voltage constant and a variation appearing in the inductive current of the tank circuit, we can come to only one conclusion. The inductance must be fluctuating. By changing the inductance, we can alter its opposition to the current flow, and this could

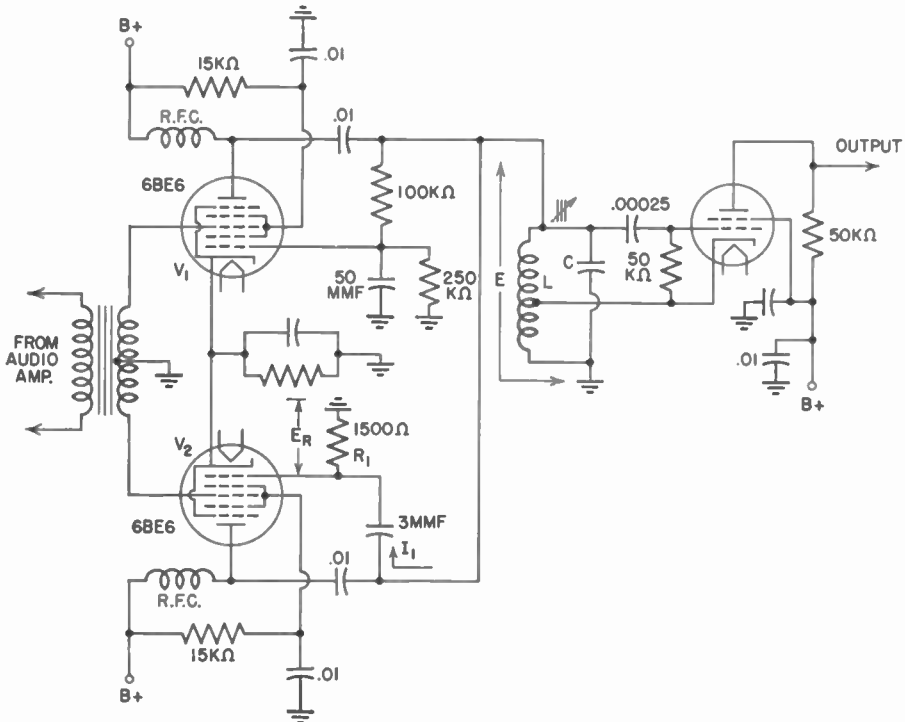


FIG. 14.8. A balanced modulator.

account satisfactorily for the various inductive currents. A moment's reflection will indicate that a variation in inductance must be followed by a variation in frequency. Consequently, the audio voltages will produce frequency modulation.

In passing, we may note that when the audio voltage is at its positive peak, the maximum inductive current will flow from the tube through coil L . At these moments, the overall or apparent value of L will be smallest (presenting the lowest inductive reactance and hence permitting the largest flow of inductive current). The frequency of the oscillator will be at its highest point. (See previous formula.) During the opposite half of the cycle, when the audio voltage is negative, the frequency is at its lowest point. Remember again that it is the varying effect of the tube across the tank circuit that is responsible for all this. The actual coil and capacitor of the circuit never change.

We have been concerned thus far with a reactance tube that produced a varying inductive current. It is possible to achieve the same results by having the tube appear as a capacitance. This will be done in the next section when a balanced, reactance-tube frequency modulator is investigated.

Balanced Modulators. The degree of frequency shift can be increased and the linearity improved through the use of a balanced modulator, such as illustrated in Fig. 14.8. Two tubes act as reactance units, their grids connected across opposite ends of the audio-signal transformer, whereas their plates are attached to one common point. The purpose of this arrangement is to produce opposite effects in the tubes which will aid each other. Let us see how this is accomplished.

One tube, V_1 , is connected as in the previous circuit. From this we know that the plate current produced by the tube will lag the tank voltage E and appear inductive. For V_2 , however, we find a slightly different arrangement. We still have a capacitor (3 mmf) in series with a resistor, but their impedance values are reversed. The capacitance, because of its very small capacitance, presents a much greater opposition to the flow of current through this branch than the 1500-ohm resistor. Hence, the current will be almost entirely capacitive and will lead the applied voltage E by practically 90° . This is shown vectorially in Fig. 14.9A. The voltage that is developed across R_1 is in phase with the current because, in all resistors, current and voltage are in phase. We may add E_R to the vector diagram, as in Fig. 14.9B. Since the grid obtains its input voltage from the resistor, we see that E_g and E_R are one and the same thing. In the vector diagram of Fig. 14.9B we can place E_g at the same point as E_R .

Proceeding with the analysis, we note that I_p of V_2 , being governed directly by E_g , is also in phase with it. Hence, we arrive at a situation

where, for V_2 , the plate current is 90° ahead of the applied oscillator voltage E (see Fig. 14.9C). Thus, by rearranging the capacitor-and-resistor combination and altering their relative values we obtain a plate current that leads instead of lags the voltage E . Thus, V_2 appears capacitive.

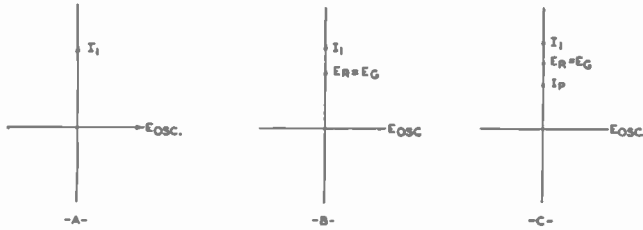


Fig. 14.9. The current and voltage phase relationship in the grid network of V_2 (Fig. 14.8).

Within the tank circuit, this capacitive current combines in phase with the capacitive current flowing through the tuning capacitor C . Whether the effective capacitive effect across the entire circuit increases or decreases depends upon the variations in the plate current coming from V_2 . If the capacitive current increases, this would have the same effect as a decrease in capacitive reactance. The reason is simple. The tank voltage E remains constant. Hence, the only way the current through C (or L) can change is through a change in the reactance in that branch. For an increase in capacitive current, we must have a corresponding decrease in capacitive reactance. In terms of capacitance this means a larger capacitor because

$$X_c = \frac{1}{2\pi fC}$$

If X_c goes down, C must increase. The opposite current variation would mean a decrease in C .

Now let us apply an audio voltage to this reactance-tube modulator and determine how the frequency changes.

The grids of the 6BE6's are connected across opposite ends of the audio input transformer. When one grid goes positive, the other grid is driven negative. Suppose that the audio voltage applied to V_1 is positive, and that to V_2 is negative. Then the plate current I_{p1} will increase and I_{p2} will decrease. Across the oscillator tank circuit this will appear as an increased inductive current and a decreased capacitive current. We have seen that an increasing I_{p1} means a lower effective inductance across the tank circuit. We have also determined that a decreasing capacitive current has the

apparent effect of a decrease in C . Hence, as a result of the application of the audio voltage, both the effective values of L and C have decreased. From

$$f = \frac{1}{2\pi\sqrt{LC}}$$

we note that the frequency rises. But since both L and C are effected, the overall frequency change is much greater than we could have obtained with the single-tube arrangement of Fig. 14.7.

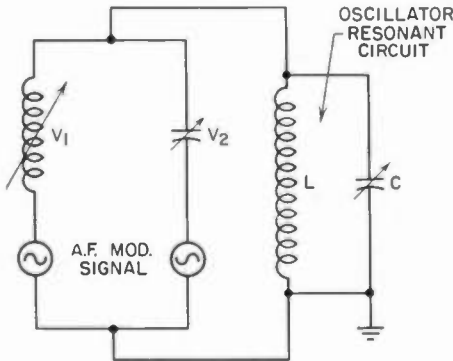


FIG. 14.10. An approximately equivalent circuit of the balanced modulator of Fig. 14.8. V_1 is represented by a variable inductance while V_2 by a variable capacitance.

We could go through the same analysis for the opposite situation when the audio input voltage reversed. However, from what has already been given there is no need for this. The frequency variation is now in the opposite direction, toward the lower frequencies. Between the positive and negative audio voltage peaks, intermediate frequency shifts take place at each point directly proportional to the amplitude of the applied voltage.

An approximately equivalent circuit of the balanced modulator of Fig. 14.8 is shown in Fig. 14.10.

The Complete Transmitter. Our progress, to this point, in the forming of an F-M signal has consisted solely in examining the action in a reactance tube and the associated oscillator. In practice, the oscillator frequency is somewhere in the neighborhood of 5 mc. Through the application of an audio signal to the reactance tube, we can vary the oscillator frequency about 4 kc. The output, then, from the oscillator will be $5 \text{ mc} \pm 4 \text{ kc}$. It is not practical to attempt to obtain a greater frequency deviation because of the non-linearity that rises. Hence, if we desire to reach the $\pm 75 \text{ kc}$ authorized by the F.C.C., frequency multiplication must be achieved in the stages that follow the oscillator.

A block diagram of a complete reactance-tube F-M transmitter is shown in Fig. 14.11. We are already familiar with the reactance tube and the oscillator. Frequency multipliers are needed to develop the final, signal-frequency variation $\pm 75 \text{ kc}$. The power amplifier, of course, is used to raise the signal level to the rated power of the unit, be it 5 kw, 10 kw or even 50 kw. Note that until we reach the power amplifier stages we have relatively low power stages. This permits the use of small, inexpensive receiving

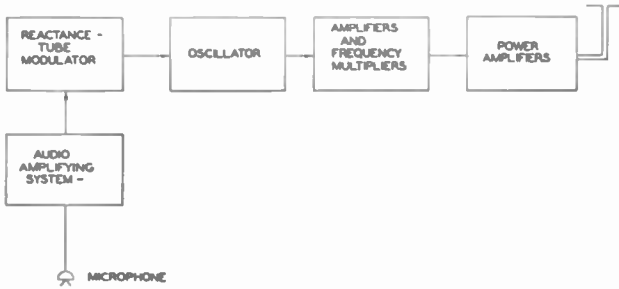


Fig. 14.11. The block diagram of an F-M transmitter using a reactance-tube modulator.

type tubes, with a substantial saving in cost. In A-M transmitters, the modulation occurs at or near the final power amplifier. This necessitates large modulating voltages, with their attendant high-power tubes. The cost of such an arrangement is usually considerable.

A second addition to each reactance-tube transmitter is a frequency-centering system. The fundamental frequency that is received from the oscillator is the result of the normal resonant frequency of the oscillator tank circuit plus or minus whatever effect the reactance tube may cause by virtue of its attachment across the oscillator. It is well to keep in mind that the simple attachment of the reactance tube (even when no audio signal is present) produces a frequency shift in the oscillator. Of course, the change is fixed, not varying as when the audio is active. As an example, the oscillator may have a frequency of 4.995 mc. By attaching the reactance tube, we may raise the frequency to 5.000 mc. This now represents the normal oscillator output frequency and it is around this value that the audio signal will swing the frequency.

Once the normal or center oscillator frequency has been established, it is extremely important that it remain fixed. Every station is assigned to one frequency and thereafter it becomes their responsibility to maintain the center or normal frequency at this point. Failure to do so will not only

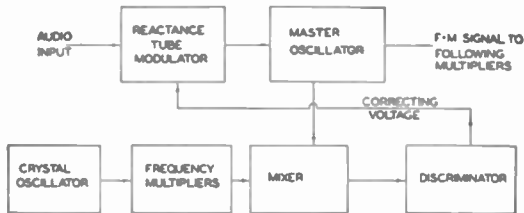


Fig. 14.12. A block diagram of a discriminator control system which keeps the master oscillator at its assigned center frequency.

cause interference to stations on other bands, but undoubtedly result in revocation of their operating license by the F.C.C.

A simple method that has been successfully used to keep an oscillator on frequency is the system shown in Fig. 14.12. This network is arranged to produce an output voltage from the discriminator if the oscillator frequency shifts. By applying this voltage to the reactance tube, we can vary its plate current and, through this, its effect on the oscillator. If the discriminator voltage is properly applied, it will counteract any tendency on the part of either the oscillator or the reactance tube to change the center transmitter frequency. Note that the correction is applied only to maintain a fixed center frequency. It does not, in any way, interfere with the frequency-modulation excursions above and below the resting frequency.

Frequency Multipliers. The output from the modulated oscillator is usually within the frequency range of 4.7 to 6 mc. The frequency variation, due to the modulation, is generally near 4.2 kc. This is the initial maximum swing, destined to become ± 75 kc at the output of the transmitter. Under present frequency allocations, the output carrier is confined to the range 88 to 108 mc. Hence, we must increase the relatively low oscillator frequency to some value between 88 and 108 mc. Suppose, as an illustration, we wish to broadcast on 90 mc. The most common arrangement is to utilize two triplers and a doubler, thus providing the minimum number of frequency multipliers with the proper amount of frequency multiplication. The total amount, then, by which the oscillator frequency is increased is 18 times. Since we require a 90-mc output, we find that an oscillator operating at 5 mc will do nicely. Again, with an eighteenfold multiplication, an initial frequency swing of 4.1667 kc must be obtained for the loudest audio signal. Thus, ± 4.167 kc, increased 18 times, gives the maximum ± 75 kc permitted by the F.C.C.

A frequency multiplier is essentially nothing more than an ordinary amplifier with the output circuit tuned to a harmonic of the input frequency. In the circuit of Fig. 14.13, the resonant tank, L and C in the output of V_1 , would be tuned to three times the input frequency. Any voltage that is developed across the tank and transferred to the next stage would be the third harmonic voltage. All other harmonics in the circuit would develop very little voltage across the tank impedance and only a negligible amount would reach the following stage. By successively choosing the desired harmonic, we can raise the original oscillator frequency to any desired value.

Grid-Leak Bias and Distortion. In order to appreciate fully the operation of a frequency multiplier and its function in the F-M transmitter, there are several facts we must know. First, in any class C amplifier, such as we find in transmitters, the large grid-leak bias acting at the input of the tube produces a distorted wave in the plate circuit. Grid-leak bias is actu-

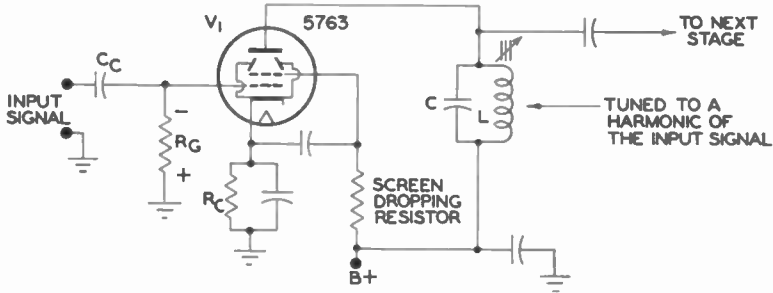


FIG. 14.13. A frequency multiplier. The main grid bias is developed by R_g ; R_c is merely for protective bias.

ally self-bias, where the amount of voltage developed across the grid-leak resistor, R_g , is directly dependent upon the strength of the incoming signal.

Initially, when no input voltage is applied, no bias is present on the grid. Upon the application of a signal voltage, the grid is driven positive on the positive half cycles and current flows in the grid circuit. The coupling capacitor becomes charged. During the negative portions of the signal the capacitor discharges, tending to maintain a voltage across R_g with the polarity as shown. As long as the strength of the incoming voltage is constant, the bias will keep the tube fixed at one operating point.

The amount of bias voltage developed across R_g will depend upon the size of this resistor (for any one value of input signal). In practice, the resistor is chosen to give a bias equal to approximately 2 or 3 times the cut-off voltage of the tube. If the latter voltage is given as 15 volts by the manufacturer, then the grid-leak bias will be between 30 and 45 volts. Its position is indicated in Fig. 14.14. With this as the operating point, the incoming voltage will vary above and below this value. At the positive peaks, the grid will be driven sufficiently positive to maintain the bias voltage across R_g fixed.

Plate current flows only for that portion of the input voltage when the

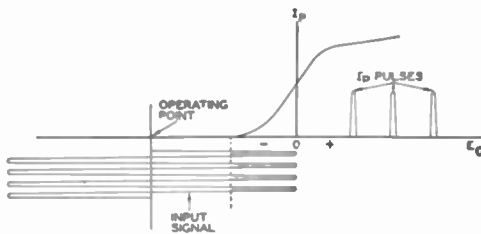


FIG. 14.14. The operating characteristics of a frequency multiplier. Only the shaded portion of the input signal is effective in producing plate current.

total grid voltage is more positive than the plate-current cut-off value. This region is shown shaded in Fig. 14.14. The plate current, during those moments when it is permitted to flow, will do so in pulses.

The fact that the plate-current form, produced as a result of the action of the grid bias, is not an exact duplicate of the input signal immediately indicates that distortion has been introduced. It can be demonstrated mathematically that when a wave is distorted, odd and even harmonics of that wave are produced. If the plate tank circuit is tuned to one of these harmonic frequencies, then the major portion of the voltage developed across the coil will be at this harmonic. Transfer to a succeeding stage, also tuned to the same harmonic, will practically eliminate all the other frequencies. Thus, in Fig. 14.13, tuning L and C to the third harmonic of the input will produce this component across the circuit.

It will be found that multiplication seldom exceeds the third harmonic in any one stage. The reason is simply due to the fact that in any frequency multiplier, the voltage that can be obtained through the use of a harmonic is proportionately lower than if we used the fundamental. For the third harmonic, the output voltage developed across L and C would be usually less than $\frac{1}{2}$ of the fundamental if L and C were tuned to the fundamental frequency. Higher harmonics will give progressively less output.

Although the pulses that are fed into L and C do not resemble sine waves, the fly-wheel action of the tank redevelops them. The pulses keep the current circulating between the capacitor and the coil and this action forms the necessary sine waves at whichever frequency the circuit is tuned.

The plate tank circuit of the frequency multiplier is tuned broadly enough to include any frequency variations of the carrier about its resting point. Thus, if the input frequency to the first tripler is 5 mc and is accompanied by a ± 4.167 -kc variation due to audio modulation, it will vary between the limits of 5.004167 mc and 4.995833 mc (4.167 kc has been changed to 0.004167 mc and added and subtracted from 5 mc). At each point in this range, the third harmonic will be produced and developed across the tank circuit. Hence, the output will fluctuate between 15.012501 mc and 14.987499 mc. This can be expressed as 15 mc \pm 12.501 kc. Thus we see that both the carrier and the sidebands each receive the same frequency multiplication.

With one more tripler, the foregoing values would be converted to 45 mc \pm 37.503 kc. Add a doubler, and we obtain the desired frequency, 90 mc, with ± 75 -kc deviation.

After these multipliers, power amplifiers are added in sufficient number until the desired output power is attained.

RCA F-M Transmitter. The present RCA F-M transmitter incorporates many of the preceding circuits and ideas in its design. It uses a slightly modified pair of push-pull modulators, a Hartley oscillator, two triplers, a

buffer amplifier, a doubler, and several power amplifiers. A block diagram is shown in Fig. 14.15.

The balanced modulator differs from the previous balanced modulator in that the grids receive their energy from the oscillator tank circuit by means of a link-coupling arrangement. Previously, it will be remembered, we obtained the R.F. voltage for each grid by means of a resistance-capaci-

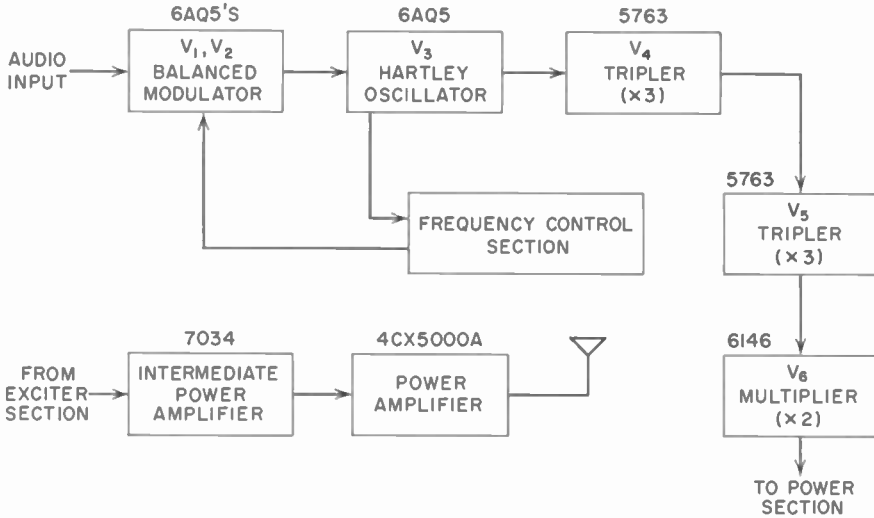


Fig. 14.15. The outline of a modern RCA F-M transmitter. The notation above each block designates the tubes used in that stage.

tance circuit coupled directly to the oscillator tank. Through this R-C arrangement, each grid received voltages that made them lead or lag the oscillator tank current by 90° . If the plate current of one modulator tube was 90° leading, the current of the other tube was 90° lagging.

In the RCA transmitter, the R.F. voltage from the oscillator is transferred to L_1 inductively. (See Fig. 14.16.) The voltage developed across L_1 produces a current flow in the link-coupling circuit, establishing a similar voltage across L_2 at the other end. The voltage from L_2 is then coupled to L_3 and L_4 inductively.

If we examine the grid circuits of the two modulator tubes, we note that they form essentially a push-pull circuit. The two far ends of L_3 and L_4 are placed at the same R.F. potential by C_3 . Thus we may consider L_3 and L_4 as being directly connected. C_1 and C_2 resonate L_3 and L_4 to the same frequency as the oscillator tank circuit. Since the grids are so connected, each receives R.F. voltages that are 180° out of phase with re-

spect to each other. If one grid is, at one instant, positive, the other grid is negative.

To obtain the required reactance from each modulator, the designers have arranged the link-coupling network, L_1 and L_2 , in conjunction with the other coils— L_3 , L_4 —so that the R.F. voltage transferred from the oscillator is shifted 90° . Thus, the voltage is first shifted 90° and then applied to L_3 and L_4 to make each grid 180° out of phase with the other. As a result of this arrangement, each tube presents to the oscillator tank a dif-

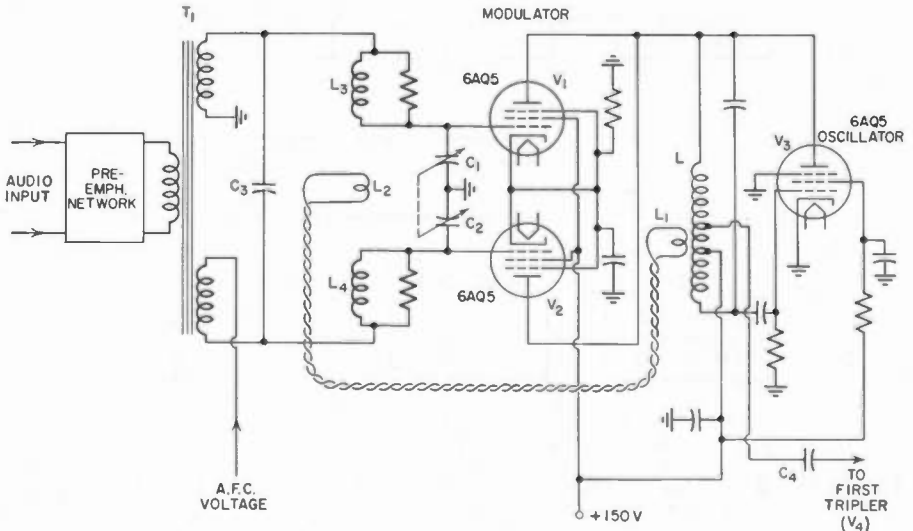


Fig. 14.16. A simplified diagram of the balanced modulator and oscillator of the RCA F-M transmitter.

ferent form of reactance. However, with no audio input signal at V_1 and V_2 , the balanced connection of each modulator acts to cancel the reactance produced by each tube at the oscillator tank.

The audio voltage is applied through transformer T_1 to each grid of the modulator. Since the grids are connected to opposite ends of the transformer, the audio voltage will be applied 180° out of phase to each grid. Application of the modulating voltage will upset the balance of V_1 and V_2 , causing one tube to draw more current and the other tube less current. Across the oscillator tank circuit, the result is the same as if we had varied the reactance (inductive or capacitive), thereby producing a frequency shift. The two tubes, thus, produce a greater frequency shift than if one tube alone had been used in the modulator.

From L , the frequency variations are fed to V_4 , the first tripler, by means

of capacitive coupling through C_4 . The output circuit of V_4 is tuned to the third harmonic of the input or oscillator frequency. This, as we have already seen, will triple the input frequency and also the *frequency variation*, which is the F-M. V_5 is another tripler, and V_6 is a doubler. The layout, to this point, represents the exciter unit and is completely contained within one

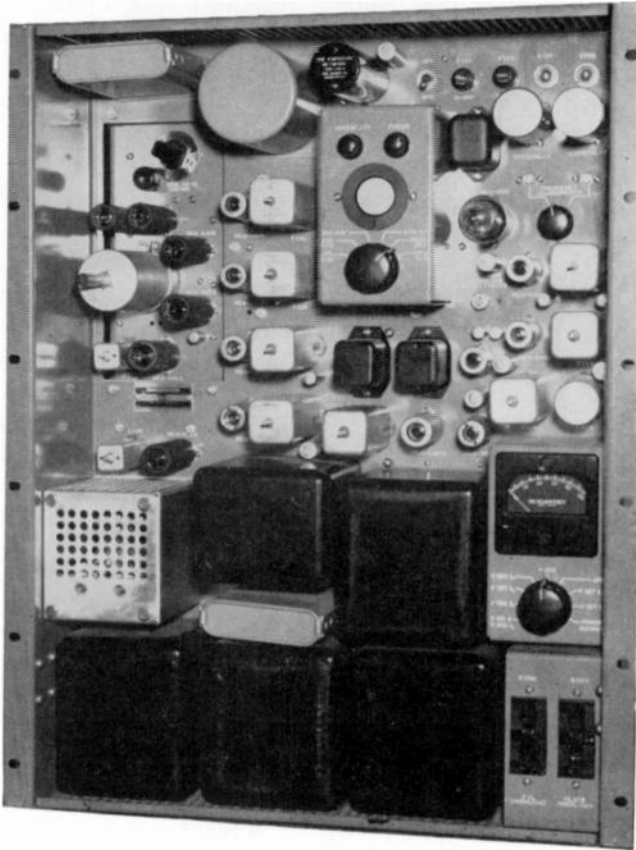


FIG. 14.17. Exciter and driver sections of RCA F-M transmitter Model BTE-10B. (Courtesy RCA)

cabinet (see Fig. 14.17). Included also is a power supply and a frequency control unit.

The center frequency, at the output of the exciter, is somewhere between 88 and 108 mc. The signal level at this point is on the order of 10 watts. This is now fed to an intermediate power amplifier where it is raised to 250 watts. Then the signal is applied to a 4CX5000A power tube, and the out-

put from this stage is the desired 5 kw. The signal then goes to the F-M antenna.

Frequency-Control Circuits. The F.C.C. specifies that each F-M broadcast station must be maintained within 2 kc of the assigned center frequency. To insure that this condition is maintained at all times, it has become customary to employ automatic frequency control. This is especially necessary in transmitters that employ reactance-tube modulators because in these units the main oscillator is usually some form of Hartley oscillator. Unless definite precautions are taken, frequency drift will occur because of the effect of heat, humidity, and aging of tubes and circuit components. With a properly designed, frequency-correcting network, the carrier will be confined to the regulation limits at all times.

The frequency control system which is employed in the RCA F-M transmitter of Fig. 14.15 is based upon a comparison of two oscillators—a crystal oscillator and the master frequency-modulated oscillator—to obtain a mean center-frequency control. The frequencies from both units are each divided to a common frequency and then combined in a balanced phase detector. When the two oscillators are exactly in step, the two inputs to the phase detector are 90° out of phase and the detector output is zero. If the mean frequency of the master oscillator tends to vary from the frequency of the crystal oscillator, the 90° phase relationship is no longer obtained, and the phase detector develops a d-c output which is used as a correction voltage to offset the tendency of the master oscillator to vary. The correction voltage is applied to the control grid of a modulator tube where it directly affects the reactance shunted by this tube across the master oscillator tuning circuit.

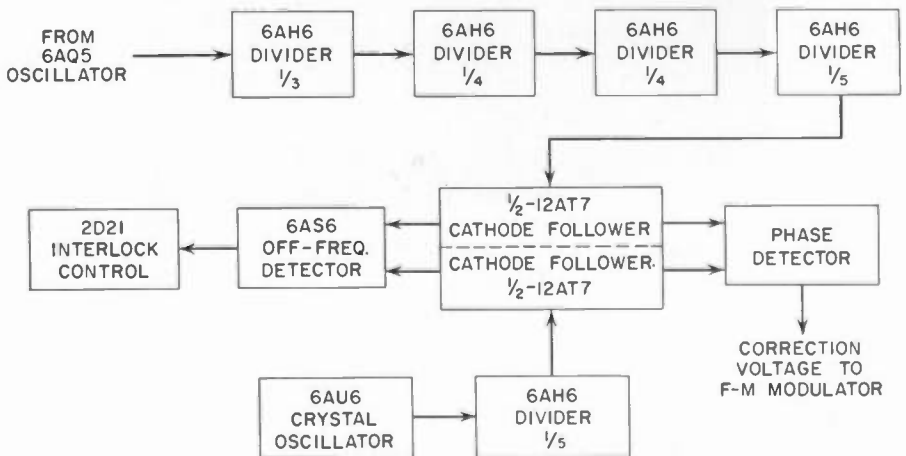


FIG. 14.18. The automatic frequency control section of the F-M transmitter of Fig. 14.14.

A block diagram of the frequency control system is shown in Fig. 14.18. A portion of the output from the master oscillator is fed to a series of frequency dividers where the input frequency is divided down by a factor of $3 \times 4 \times 4 \times 5$, or 240. Since the output from the master oscillator falls somewhere between 4.5 and 6.0 mc, the output of the frequency divider chain will be 1/240 of these frequencies, or 18.75 kc to 25 kc. (It is understood that only *one* center frequency will be obtained from the master oscillator and consequently only one frequency will appear at the end of the divider chain.)

At the same time, a crystal oscillator, operating at some frequency between 93.75 kc and 125 kc, has its signal pass through a divider where the initial frequency is lowered by a factor of 5. This will produce a frequency of 18.75 kc and 25 kc, or the same frequency that the master oscillator should produce at the end of its frequency chain if it is exactly on frequency. The two divided signals are then passed through separate triode cathode followers and applied to the phase detector.

The reason for the extensive frequency division which must be employed stems from the limitations of the phase detector. The crystal oscillator frequency is quite stable in value. However, the instantaneous frequency deviations of the carrier may extend for as much as 75 kc which, in terms of equivalent phase variations, will amount to thousands of degrees. The phase detector, on the other hand, operates over a range of only 180° . Hence, to employ this type of detector, the frequency deviation must be reduced, and this is achieved by the frequency-divider chain. Of course, not only are the frequency deviations reduced, but the carrier frequency as well. In fact, by the time the signal deviations from the master oscillator are reduced by a factor of 240, the instantaneous carrier variation at 100 per cent modulation produces a phase variation of approximately 33° when the modulating frequency is 30 cycles.

There is still another reason for using frequency division, and this concerns the varying amplitude of the center-frequency component. It was shown in Chapter 1 that, when a carrier is frequency-modulated, power for the sidebands is obtained from the carrier. At times, therefore, the carrier amplitude will decrease to zero, making accurate synchronization impossible. The frequency division, however, acts to concentrate the carrier power into a smaller and smaller bandwidth, resulting at the end of the frequency division chain in a signal of substantially constant amplitude (containing no less than 96 per cent of the power of the unmodulated carrier) for synchronization purposes.

Frequency Division. Before we analyze the phase detector, let us examine a frequency divider stage. This is most commonly a combination mixer and oscillator. (See Fig. 14.19.) As an indication of how it functions

as a frequency divider, consider the oscillator section first. The oscillator, consisting of L_1 and C_1 in the plate circuit and L_2 in the grid circuit, is tuned to the desired output frequency. L_1 and C_1 determine the fundamental frequency at which the circuit will oscillate. As is usual in such oscillators, the tube operates on the nonlinear portion of its characteristic curve and the plate current contains many harmonics. Although harmonics are generally not desirable, their presence is required here for the mixing action.

The frequency to be divided is fed to the oscillator through the capacitor, C_2 . Let us say that its value is 6.0 mc. An output one-fourth of this is desired, so that the oscillator functions at 1.5 mc. At the tube the 6.0 mc

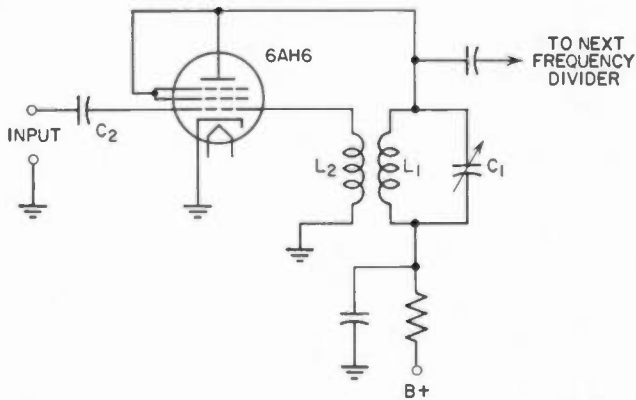


FIG. 14.19. A frequency divider similar to those used in the RCA F-M transmitter.

will mix with either the third harmonic (4.5 mc) or the fifth harmonic (7.5 mc) of the oscillator to produce a difference frequency of 1.5 mc. Because of the harmonic relationship between the incoming signal and the oscillator voltage, a lock-in will occur, with the incoming signal acting to keep the oscillator at some sub-harmonic value. If the incoming frequency should shift, then the lock-in will still remain, and the oscillator frequency will be shifted or "dragged along." However, the incoming signal will be able to maintain control only within certain limits. If these are exceeded, the oscillator breaks loose and returns to its natural frequency. However, in this circuit, the frequency shift is seldom so great that the lock-in is broken. The oscillator circuit is designed to follow the incoming frequency over fairly wide limits.

To recapitulate, the oscillator in the frequency divider is set at a sub-harmonic of the correct incoming frequency. Then, one of the harmonics of the oscillator mixes with the incoming signal to produce a difference frequency which is equal to the fundamental oscillator frequency. Because of

the harmonic relationship of the incoming signal and the oscillator frequency, the two lock-in. Once locked-in, any small shift in the signal will produce a shift in the oscillator and thus indicate to the following frequency dividers that the main F-M oscillator is off the center frequency.

Returning to the block diagram of Fig. 14.18, we see that the system contains four frequency dividers. Multiplying each of the frequency divisions together, we obtain a total of $3 \times 4 \times 4 \times 5 = 240$. Thus, if the main oscillator is at a frequency of 6.0 mc, the output of the final divider is 6.0 mc/240, or 0.025 mc, or 25 kc.

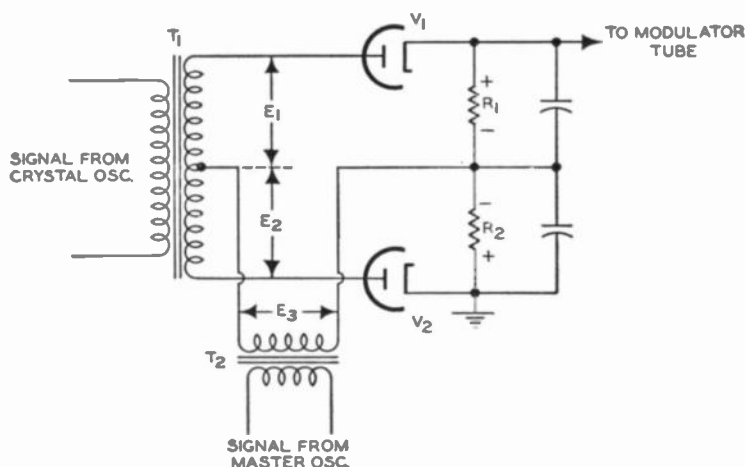


FIG. 14.20. A phase detector.

The overall purpose of the frequency divider is to bring to the phase detector immediate indication of any change in the main oscillator frequency. However, the sole fact that the frequency has shifted is not significant unless it is compared with a standard frequency, one that remains fixed. For this we employ a crystal oscillator. After the output of the crystal oscillator is decreased by one-fifth, its frequency has the same value that the main oscillator frequency should possess after it had been divided 240 times. Hence, if the main oscillator frequency should shift even the slightest amount, this becomes immediately apparent by comparison.

Phase Detector. The phase detector compares the phase between the frequency-divided signals from the master oscillator and the crystal oscillator. When the master oscillator is on frequency, the two signals at the detector will possess a phase difference of 90° . The reason for this particular phase difference is that, under these conditions, the output of the phase detector is zero. The latter condition should occur when the master oscilla-

tor center frequency is correct, since at this moment no correction voltage is needed.

To demonstrate that the foregoing conditions exist, consider the circuit of the phase detector shown in Fig. 14.20. The frequency-divided signal from the crystal oscillator is applied to the transformer, T_1 . Since the secondary is center-tapped, equal voltages appear across each half of the secondary winding. These are labeled E_1 and E_2 . Since the voltage polarities at either end of the secondary windings are opposite to each other with respect to the center tap, E_1 and E_2 can be represented (by vectors) as shown in Fig. 14.21A. E_3 differs from E_1 and E_2 by 90° , as specified above.

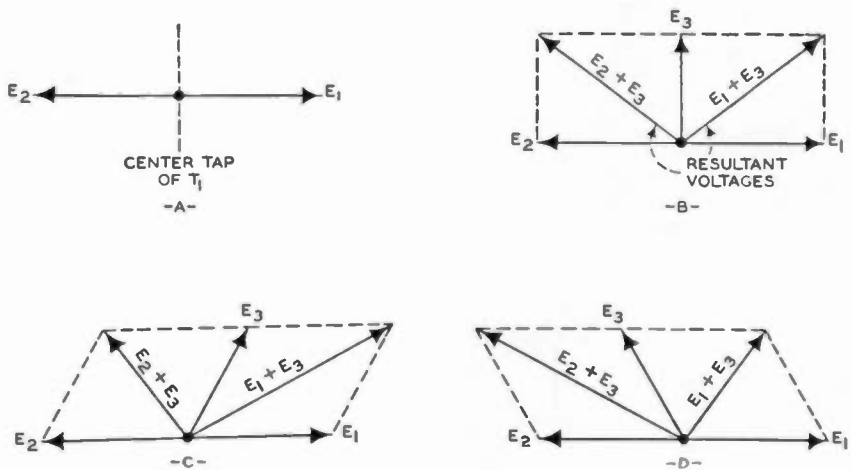


FIG. 14.21. Phase relationships between E_1 , E_2 , and E_3 of Figure 14.20.

If now, E_1 and E_3 are added vectorially, and the same is done to E_2 and E_3 , we obtain the resultant vectors or voltages as shown in Fig. 14.21B. $E_1 + E_3$ is applied to tube V_1 , while $E_2 + E_3$ is applied to V_2 . Since both resultant vectors are equal in length (i.e., amplitude), the same currents will flow through V_1 , V_2 , R_1 , R_2 , producing equal voltage drops across the two load resistors. Their combined voltage, which is fed to the control grid of the modulator tube, is zero because of the back-to-back placement of R_1 and R_2 .

Consider, now, what happens when the frequency (and, consequently, the phase) of the master oscillator changes from the desired center value. E_3 will no longer be exactly 90° from E_1 and E_2 . When the master oscillator frequency changes in one direction, E_3 will approach closer in phase to E_1 and hence cause $E_1 + E_3$ to be greater than $E_2 + E_3$. (See Fig. 14.21C.) V_1 will now conduct more strongly than V_2 , developing a greater voltage drop across R_1 and feeding a positive voltage to the modulator tube. How-

ever, when the master oscillator drifts in the other direction, the conditions shown in Fig. 14.21D obtain. Now V_2 will conduct more strongly than V_1 , giving rise to a greater voltage across R_2 and feeding a negative voltage to the modulator. In each instance, the modulator will tend to return the master oscillator to its proper center frequency.

A 10-cycle low-pass filter is inserted in the path of the correction voltage from phase detector to modulator. Since the lowest modulating frequency used is 30 cycles, the filter removes any residual modulation that may be left. The only changes that are permitted to reach the modulator are those caused by the drifting of the master oscillator, and these occur at a frequency less than 10 cps.

Off-Frequency Detector. In the same A.F.C. circuit, Fig. 14.18, there is a 6AS6 off-frequency detector which also receives the stepped-down master oscillator and crystal oscillator signals. (See Fig. 14.22.) The 6AS6 tube

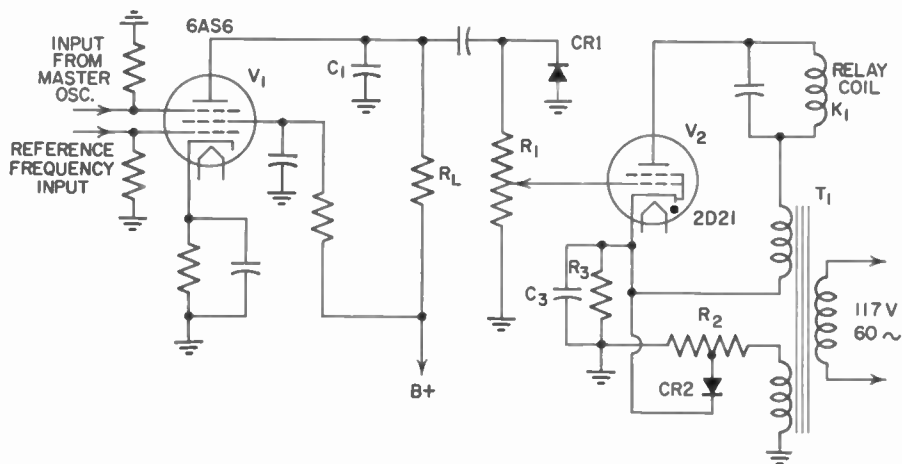


FIG. 14.22. Schematic diagram of the off-frequency detector and interlock control circuit.

acts as a mixer, receiving each signal at a different grid. When the master oscillator is on frequency, both signals possess the same frequency and zero-beat together, producing no a-c output. However, if a difference does appear between the two signals, sum and difference frequencies will be developed. C_1 is designed to offer a low impedance to the sum frequencies, effectively by-passing them away from load resistor, R_L . The much lower difference frequency does reach R_L and is rectified by $CR1$. A positive d-c voltage will then appear across R_1 . If this voltage exceeds the fixed cathode bias applied to the thyratron, V_2 , the tube will conduct and the relay K_1 will be

energized. This opens up the relay contacts (not shown) and prevents plate power from being applied to stages following the exciter unless the A.F.C. circuit is locked in. Thus, the transmitter is prevented from radiating a signal unless the proper frequency is being generated.

The foregoing automatic frequency control system represents one approach to this problem. Several other methods have been employed and it may be instructive to see how they operate. One interesting system employs a 2-phase motor; another utilizes a discriminator circuit. Both are described below.

Motor Control. An outline of an F-M transmitter using motor control in the A.F.C. network is shown in Fig. 14.23. Capacitors C_1 and C_2 , which

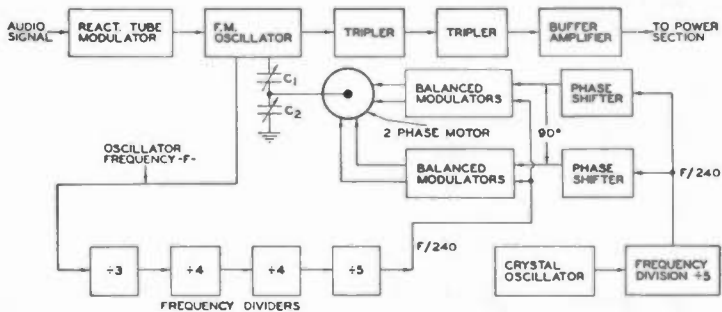


FIG. 14.23. An outline illustrating the several circuits of the frequency control section.

are mechanically linked to the 2-phase motor, are electrically connected across the master oscillator tuning coil. Any change in the settings of C_1 and C_2 will alter the frequency of the oscillator and, with this, the transmitted carrier frequency. A portion of the oscillator voltage is fed to a series of frequency-divider stages, where a total frequency division of 240 times occurs. Thus, with the main oscillator frequency somewhere between 4.5 and 6.0 mc, the output of the final frequency divider stage will decrease to the 18.75-25 kc range. (It should be understood that, for any one station, the main oscillator has but one frequency. However, to cover the entire F-M band, 88 to 108 mc, it is necessary to provide the values of 4.5 to 6.0 mc for the oscillator.) The divider output is then fed to a balanced modulator.

The other input to the modulators is obtained from a highly accurate crystal oscillator. The crystal oscillator frequency is lowered by $\frac{1}{5}$, at which value it will be exactly $\frac{1}{240}$ th of the correct main oscillator frequency. This frequency is then compared in the balanced modulator with the frequency obtained from the main oscillator. If any difference exists, the motor

is actuated and the tuning capacitors rotated until the oscillator is again centered at its assigned spot.

The balanced modulator of Fig. 14.24 differs from the balanced modulator of Fig. 14.8. The latter circuit modulates the frequency of an oscillator, whereas the present modulator mixes two frequencies. Both qualify for the designation of modulator, although their functions differ.

Balanced Modulators. In a balanced modulator we can feed two signals into the grids of the tubes and obtain only one of the input signals plus the sum and difference frequencies of the two signals at the output. The other input frequency does not appear at the output. In the circuit of Fig. 14.24,

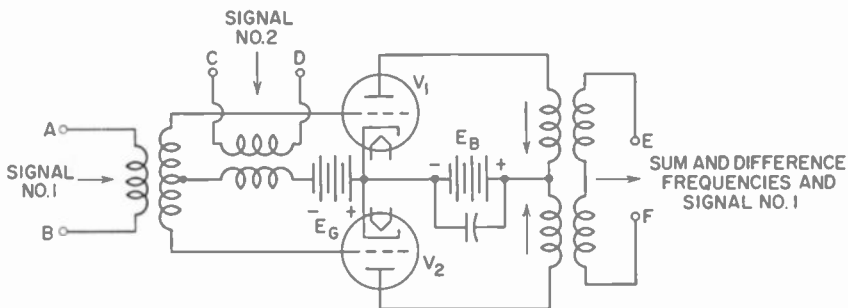


FIG. 14.24. A balanced modulator.

one signal is applied to terminals *A* and *B*, the other signal to terminals *C* and *D*, and the sum and difference frequencies appear across points *E* and *F*. It is customary to apply the higher frequency at terminals *C* and *D* and the lower frequency at *A* and *B*. However, in this case, both incoming signals have approximately the same frequency.

The reason the signal at terminals *C* and *D* does not appear across *E* and *F* is due to its method of application. Both tubes, V_1 and V_2 , are connected in push-pull, which means that their respective plate currents flow through the output coil or transformer in opposite directions. For V_1 , the plate current flows down through its winding, whereas for V_2 the flow is up through the coil to the center point. As long as the currents are equal, their opposite effect will cancel.

At the input circuit, any voltage across terminals *C* and *D* will be applied to both grids in like measure. The voltage will drive both tubes positive or negative at the same time. Both plate currents will thus be in phase, completely cancelling their magnetic effects in the output.

When we apply another signal at terminals *A* and *B*, we upset the equal voltages that the signal at *C* and *D* is applying to the grids. When *A* is positive, *B* is negative and these voltages, added to those already existing

on the grids due to the signal at C and D , produce an unbalance. First, one tube will conduct more current, then the other. The result of this mixing of frequencies in V_1 and V_2 is the production of the sum and difference frequencies across terminals E, F . We also find in the output, the frequency of the signal that is applied to terminals A, B .

Now let us see how these facts are useful in the F-M transmitter. Two balanced modulators are needed, the output of each connected to the windings of the 2-phase motor. The output from the fourth frequency divider (bringing the signal from the main oscillator) is applied to each balanced modulator. We could, for example, feed the signal into terminals, A, B of each modulator. The output of the frequency divider following the crystal oscillator is also fed to each modulator, say terminals C, D . However, there is introduced (through the phase-shift network) a phase difference between the two crystal oscillator voltages of 90° . The need for this is due to the use of a 2-phase motor. In order for the motor to rotate, the two currents flowing in its windings must differ by 90° —hence the phase-shift network.

In the modulators, the two signals mix, producing sum and difference frequencies. Of particular interest in this instance are the difference frequencies. When these currents flow through the two windings of the 2-phase motor, a torque is produced and the motor rotates the attached capacitors. If both signals have the same frequency, the difference frequency is zero, or d-c. Since the motor requires alternating currents, no torque is produced and no rotation occurs. When signals entering the modulators have identical frequencies, the main oscillator is obviously at the proper operating point.

It may be wondered why the motor does not respond to the sum frequencies, or the frequency of the signal that is applied across terminals A, B . These are also present in the output. Sum frequencies are produced as long as two signals are present, whereas difference frequencies occur only when the two signals differ in frequency. The reason the motor responds only to difference frequencies is due to its operational characteristics. It will not respond to frequencies beyond 1000 cycles. Since either input signal or the sum frequencies (produced through the mixing of the signals) are always above 1000 cycles, they cause no motor rotation.

Discriminator Control. Another interesting method of frequency control relies on the operation of a discriminator. In Fig. 14.12, a block diagram illustrates the essential features of the system. As before, a crystal oscillator is the standard against which the main oscillator frequency is compared. As soon as deviations occur in the oscillator center frequency, an output (d-c) voltage is obtained from the discriminator. If the oscillator frequency drifts to a higher value, the output voltage of the discriminator is of one polarity. Should the frequency drift below its normal value, then the voltage from the discriminator reverses. The application of the d-c voltage from the dis-

criminator to the control grid of the reactance tube controls the effect of the reactance tube on the main oscillator and, through this, the center frequency. The corrective voltage applied to the reactance tube is always of such polarity (either positive or negative) as to bring the main oscillator center frequency back to its assigned position.

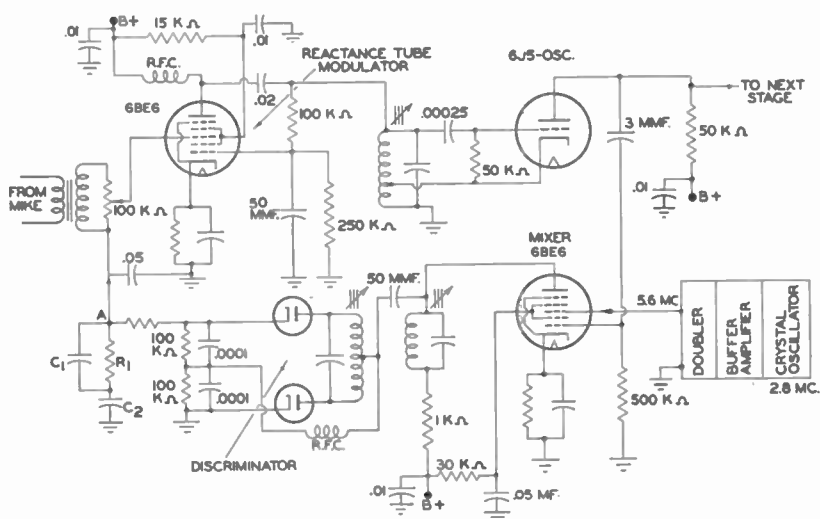


FIG. 14.25. A discriminator frequency-stabilization circuit.

A frequency stabilization circuit is shown in Fig. 14.25. The resonant frequency of the crystal oscillator is at 2.8 mc. This voltage is passed through a buffer amplifier and then doubled in the frequency multiplier to 5.6 mc. From the output of the doubler, the 5.6-mc voltage is applied to grid 3 of a 6BE6 tube, connected as a mixer.

The 5.0-mc oscillator voltage is picked off by means of a small capacitor and brought to grid 1 of the 6BE6. The two signals beat together in the tube, forming sum- and difference-frequency voltages across the plate tank circuit. Of particular interest is the difference frequency, 600 kc, since this is the center resonant point of the following discriminator. So long as this frequency is the difference frequency produced in the mixer, the output voltage from the discriminator is zero. The reason was shown previously in Chapter 9.

However, when a change occurs in the main oscillator frequency, the mixing will produce some other difference value. Consequently, an output voltage will appear at point A. If the input frequency to the discriminator is above 600 kc, the output voltage will be of one polarity. If it is below, the polarity will be opposite. From point A the voltage is fed to the grid of the

reactance tube. At the tube it affects the plate-current flow and, with this, the shift in oscillator frequency caused by the modulator connection.

The drift of an oscillator is seldom a rapid or sudden affair. Rather, the shift is gradual. To produce a large controlling bias voltage from point *A*, for only small changes in frequency, the bandwidth response of the discriminator is made fairly narrow. As the selectivity of the discriminator tuning circuit becomes sharper, more output voltage is obtained for any given frequency shift. With additional grid bias, the reactance modulator and the entire system become very sensitive to slight frequency changes. Of course, if the main oscillator frequency should suddenly shift by a considerable amount, we would find that a narrow-band discriminator is useless. However, this almost never occurs.

There is one precaution that must be observed when using this network. The signal fed from the oscillator to grid 1 of the mixer contains frequency modulation. If the correcting voltage developed at the output of the discriminator is permitted to follow these frequency variations, all the frequency modulation produced by the audio signal would be nullified. What we desire to prevent are changes in the main oscillator frequency due to uncontrolled causes, such as heat, humidity or aging of the tubes. We do not want to prevent frequency shifting due to applied audio voltages. The oscillator must be kept at its proper resting point, or central frequency. From this position it will swing back and forth under the influence of the audio modulation.

A low-pass filter is used to eliminate all the frequency modulation from the correcting voltage obtained at the output of the discriminator. In Fig. 14.25, this filter consists of C_1 , R_1 , and C_2 . It cuts off at approximately 10 cycles and effectively by-passes or eliminates all audio variations (caused by the reactance tube in modulating the oscillator) and permits only the very slow frequency drifts of the oscillator to be effective. The center frequency of the oscillator drifts very slowly, well within 10 cps.

PROBLEMS

1. What two general methods are employed to develop F-M signals?
2. What is a reactance tube? Does this differ in physical characteristics from any other tube? Explain.
3. Draw the schematic circuit for a simple reactance-tube modulator.
4. Explain the operation of the modulator drawn in Question 3.
5. A simple series circuit containing a resistor and capacitor is connected to an R.F. generator. When will the voltage developed across the resistor be nearly in phase with the applied R.F. voltage? When will this resistor voltage be almost 90° out of phase with the R.F. voltage?
6. To the circuit of a Hartley oscillator add a vacuum tube which appears as a pure resistor to the oscillator tank circuit.
7. Explain what causes a tube to function as a resistor, capacitor, or inductor.

8. How does an applied audio voltage cause the reactance tube to alter the frequency of an oscillator?

9. What are the advantages of a balanced reactance-tube modulator?

10. Explain briefly the difference between balanced and single-ended reactance-tube modulators.

11. Is the F-M signal fully developed at the output of the modulated oscillator? Explain.

12. The signal obtained from the output of an oscillator had a frequency of 5.6 mc with an F-M swing of 4.0 kc. How could we obtain a carrier frequency of 100.8 mc? How much frequency swing would we produce in raising the original 5.6 mc to 100.8 mc?

13. What precautions can be observed to insure that the frequency of a reactance-tube modulated oscillator is kept on-frequency?

14. Explain the operation of a frequency multiplier.

15. Explain briefly the operation of the frequency control employed in the RCA F-M transmitter.

16. Explain briefly the operation of the frequency-control circuit in Fig. 14.18. Illustrate by means of a block diagram.

17. How does a phase detector operate? Explain in detail.

18. How does a frequency divider function? What is its purpose in the RCA F-M transmitter?

19. Contrast a discriminator frequency control network with a motor-controlled circuit.

20. Draw the block diagram of a discriminator frequency-control network and explain how it functions.

21. Draw the block diagram of a complete F-M transmitter containing the following components.

1. Reactance-tube modulator.
2. A 3-stage audio-amplifier system.
3. Two triplers and a doubler.
4. An intermediate power amplifier.
5. A power amplifier.
6. A discriminator type of frequency control circuit.

Chapter 15

COMMERCIAL F-M TRANSMITTERS

(PART 2)

In this chapter we shall examine two additional methods* which have been employed to achieve frequency modulation directly by means of a reactance modulator. In one method, the modulator is a diode. In the second system, advantage is taken of the fact that the input capacitance of a tube is dependent on its mutual conductance. Each approach, in its way, is distinctive and demonstrates the versatility of the direct method.

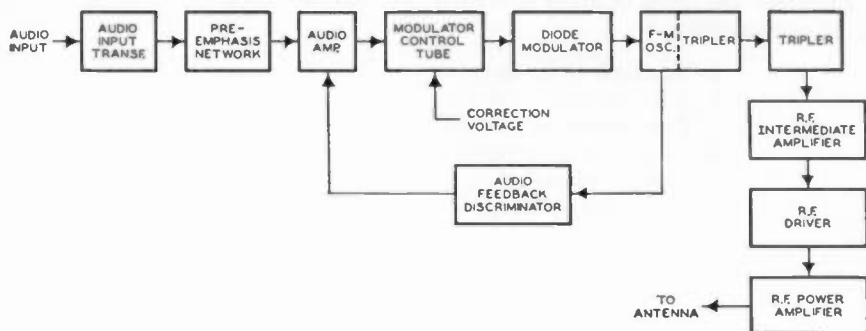


FIG. 15.1. A block diagram of an F-M transmitter in which the frequency of the master oscillator is varied by a diode modulator. The frequency stabilization system which supplies the above correction voltage is shown in Fig. 15.5.

Direct F-M with Diode Modulator. A block diagram of a transmitter using a diode modulator is shown in Fig. 15.1. The audio circuits, to the point where their signal is applied to the modulator control tube, are conventional. An audio feedback discriminator, which is peculiar to this cir-

* Commercial F-M transmitters using both modulation methods to be described are in operation today.

cuit, tends to stabilize the audio stages by applying a small amount of inverse feedback voltage. The operation of this circuit is simple and will be analyzed after the operation of the modulating system has been investigated.

The oscillator, using a 1614 beam-power tetrode, is connected as an electron-coupled Colpitts oscillator and tripler circuit. (See Fig. 15.2.) The

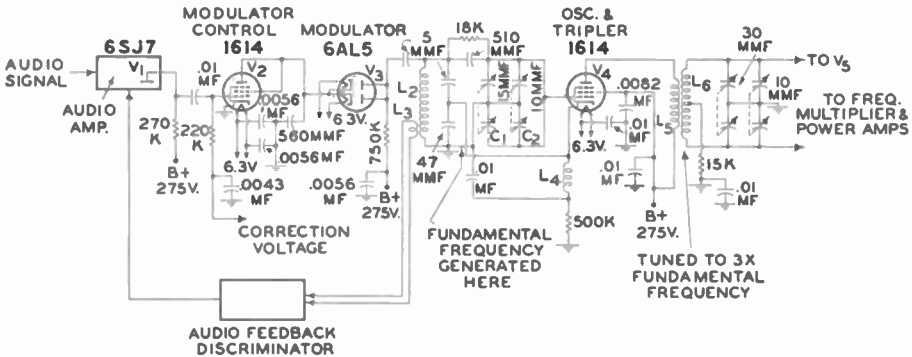


FIG. 15.2. The modulator and oscillator circuits of the transmitter shown in Fig. 15.1.

grid circuit generates a frequency $\frac{1}{9}$ that of the transmitted carrier frequency. This frequency is tripled in the plate circuit of the oscillator tube to $\frac{1}{3}$ the output frequency. The grid circuit of V_4 is tuned by the grid inductance, L_2 , by the coarse and fine variable capacitors, C_1 and C_2 , and by the capacitive reactance of the modulator tube, V_3 . In the plate circuit, L_5 in conjunction with L_6 forms a tuned coupling circuit having a very large coefficient of coupling. Tuning here is accomplished by the grid tuning capacitors of the following R.F. tripler stage. The final frequency multiplication (3 times) is achieved in the plate circuit of the following tube, V_5 (not shown). The frequency, at this point, is equal to the carrier frequency authorized to that station.

Beyond V_5 , intermediate and power amplifiers increase the power of the signal until it has attained the full value at which the station operates.

Modulating Circuits. The modulator circuit consists of two tubes, a 1614 beam-power tetrode and a 6AL5 duo-diode. The 1614 is connected as a triode and operates as a class A audio frequency amplifier. The 6AL5, with both sections connected in parallel, is placed in series with the plate of the 1614. Hence the instantaneous current through the diode depends upon the current flowing through the 1614. Since the 1614 is operating as a class A amplifier, its plate current and that of the diode will always be proportional to the applied audio signal.

The diode, in addition, is connected across the grid inductance of the oscillator. (See Fig. 15.3A.) This means that the interelectrode capacitance existing between cathode and plate of the diode and the tube's internal plate resistance are also shunted across the oscillator tank circuit. (See Fig. 15.3B.) By varying the value of this plate resistance, we change the overall

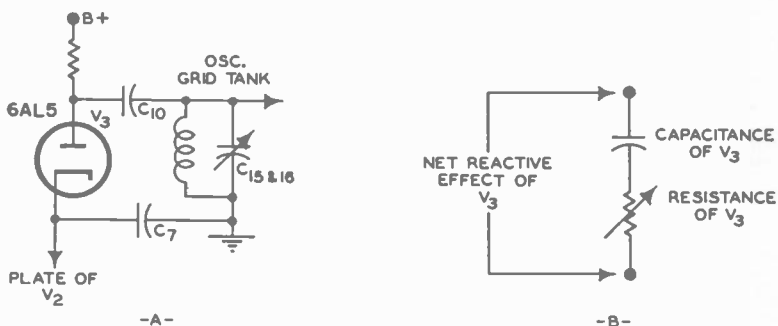


FIG. 15.3. (A) The diode modulator is connected across the grid inductance of the oscillator. (B) Equivalent circuit of diode showing its electrical effect on oscillator tank circuit.

reactance of the network, which, in turn, causes the frequency of the oscillator to change. Thus, the audio voltage at the grid of V_2 , by varying the current flowing through this tube and the diode, will also vary the reactive effect of the diode on the oscillator.

To prevent amplitude modulation as well as frequency modulation, the

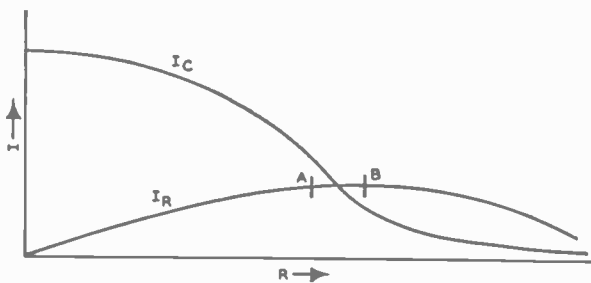


FIG. 15.4. Resistive and reactive current in diode modulator.

6AL5 is operated at a point where the change in resistive current through the tube is very small in proportion to the change in capacitive current. A plot of the current through the 6AL5 tube is shown in Fig. 15.4. It will be noted that, between the points A and B, a very small change in resistive current, I_R , occurs for a relatively large change in the resistive effect of the

tube. At the same time, between these two points, the change in capacitive current, I_c , is quite large. Hence, by operating the diode between points A and B , we secure frequency modulation accompanied by only a negligible amount of amplitude modulation.

Audio Feedback Discriminator. The purpose of the audio feedback discriminator is to provide an inverse feedback voltage for the audio amplifier, V_1 , in order to stabilize its operation and to reduce distortion. The discriminator circuit is, in essence, a Foster-Seeley F-M detector. The primary and secondary windings of the discriminator transformer are tuned to the frequency of the master oscillator, V_4 . A small coupling link, L_3 , transfers the signal from the oscillator grid coil to the discriminator transformer. When the oscillator is unmodulated, the output of the discriminator will be zero. This action is in accordance with the operation of a discriminator circuit. However, when an audio voltage is being applied to the circuit, the oscillator signal becomes frequency-modulated, and a similar audio voltage is obtained from the discriminator. This voltage possesses the proper phase to be applied to the grid of V_1 as an inverse feedback voltage.

Frequency Stabilization Circuits. We come now to the frequency stabilization circuits, these being necessary in all reactance-modulated F-M transmitters. In the previous chapter, one system using motor control, one system employing a discriminator, and one system using a phase detector were described. The present system represents still another approach, using a series of pulses and a pulse-counting circuit to keep the transmitter

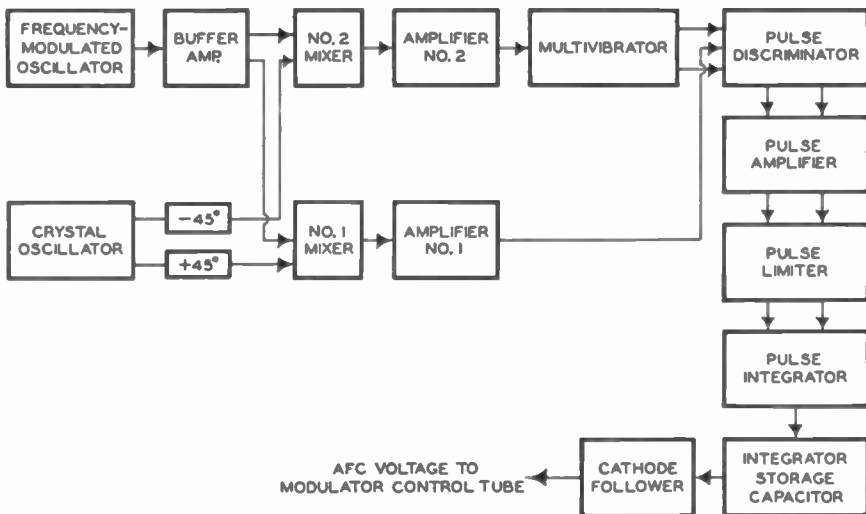


FIG. 15.5. A block diagram of the frequency stabilization system employed in the F-M transmitter of Fig. 15.1.

on frequency. When the master oscillator drifts to one side of its assigned frequency, the output pulses of the circuit have one polarity; when the drift is to the other side of the assigned frequency, the pulse polarity reverses.

A block diagram of the entire frequency stabilizer section is shown in Fig. 15.5. The master oscillator feeds a portion of its signal to two separate mixers through a buffer amplifier. The same amount of voltage is applied, in phase, to each mixer. At the same time a crystal oscillator, operating at the exact frequency to which the master oscillator should be set, also feeds its voltage to each of the mixers. However, the crystal oscillator output must first pass through two resistance-capacitance phase shift networks so designed that the crystal oscillator voltage reaching mixer No. 1 is advanced 45° in phase while the voltage reaching mixer No. 2 is retarded by 45° . The two voltages, then, are actually 90° out of phase.

Within each mixer, the signals from the crystal and master oscillators mix, and if the two are not equal in frequency, the difference and sum of these frequencies will be generated. Of interest in this discussion are the difference frequencies since the sum frequencies are by-passed to ground.

Any output frequencies developed in the mixers will be equal, but the phase of the voltage from mixer No. 2 will differ from that of mixer No. 1 by 90° because of the phase difference in the applied crystal oscillator voltages.

Let us examine this latter point in greater detail. When the master oscillator is operating at its proper center frequency, then its frequency and that of the crystal oscillator will be equal. Under these conditions, no difference frequency voltage will be obtained at the output of either mixer. However, when the two oscillator frequencies do not agree, the following will happen.

1. When the oscillator frequency is higher than the crystal reference frequency, the output of mixer No. 2 will lead the output of mixer No. 1 by 90° .
2. When the oscillator frequency is lower than the crystal reference frequency, the output of mixer No. 2 will lag the output of mixer No. 1 by 90° .

Fig. 15.6 illustrates these phase relationships, with the output of mixer No. 1 kept constant while the output of mixer No. 2 lags or leads by 90° . During modulation of the master oscillator, the applied audio-modulating voltage will swing the oscillator frequency above and below its operating point. This means that during the positive half of an audio frequency cycle, when the carrier shifts (in this circuit) in the low frequency direction, the output of mixer No. 1 will lead the voltage output of No. 2. During the negative half of the audio cycle, the output of mixer No. 1 will lag the out-

put of mixer No. 2. If the center operating frequency of the master oscillator is properly set, no correction voltage will be developed by the frequency stabilization network. However, if the master oscillator is not properly centered and has instead drifted to one side, then a difference frequency will appear in the mixer plate circuits even when no audio voltage is being applied to the circuit. When, now, an audio voltage is applied, the outputs of the mixers will not lead and lag for an equal length of time as they did when the master oscillator was on frequency. This will result in a correction voltage being developed by the stabilization system which will act to return the oscillator to its proper center frequency.

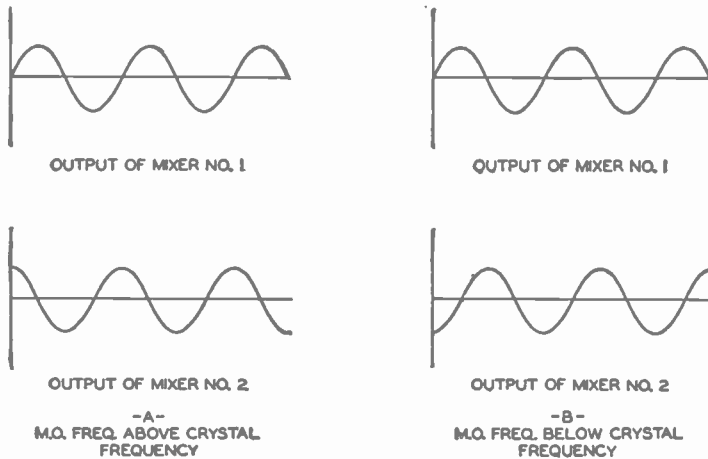


FIG. 15.6. The phase relationship between the outputs of mixers 1 and 2 when the master oscillator is not operating at its assigned center frequency.

The output of each mixer is fed to a separate amplifier. The output of mixer No. 1 is fed to amplifier No. 1. This tube is biased practically to cut-off and acts to square off the applied sine wave, giving it the shape shown in Fig. 15.7. This wave is then applied to the center of the following pulse discriminator.

The output of mixer No. 2 is passed through amplifier No. 2 and then used to trigger a multivibrator. Although it is beyond the scope of this text to go into the operation of multivibrators, it can be said that this circuit, containing two tubes, transforms the sine-wave output of amplifier No. 2 into a series of square waves. These square waves are then passed through a resistance-capacitance network which converts them into a series of positive and negative pulses. It is these pulses which are received by the same pulse discriminator to which the amplified output of mixer No. 1 is applied.

Note that the pulse discriminator receives equal and oppositely phased

pulses from the multivibrator. This is required because of the manner in which these voltages are applied to the pulse discriminator.

Pulse Discriminator. The purpose of the pulse discriminator is to combine the output of the multivibrator with the amplified output from mixer No. 1 in such a manner that voltage pulses appear on one side of its output circuit when the master oscillator is above its correct center operating frequency and on the other side of the output circuit when the oscillator frequency is below its correct operating frequency. If the master oscillator is on frequency, then during audio modulation, the number of pulses appear-

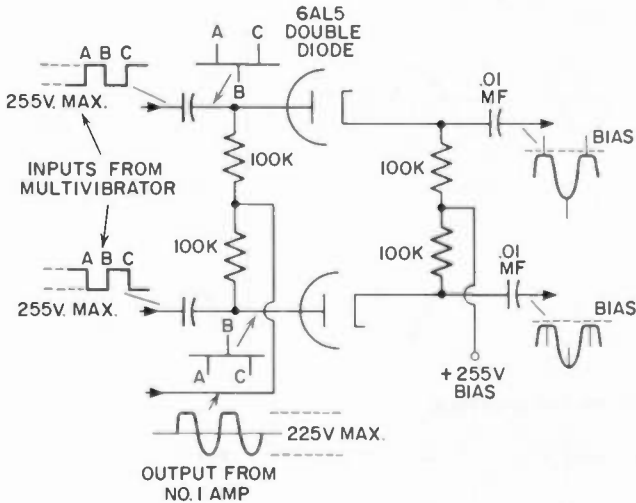


FIG. 15.7. The pulse discriminator circuit.

ing on one side of the discriminator during one half cycle of audio voltage will be exactly equal to the number of pulses appearing on the other side of the discriminator, during the next half cycle. However, this will not be true if the master oscillator has drifted.

The pulse discriminator consists of two diodes, the cathode of each being biased by a positive 255 volts. The maximum possible signal voltage available from amplifier No. 1 is also +255 volts, and therefore the voltage due solely to amplifier No. 1 cannot cause the discriminator to pass current. The same is true of the voltage pulses obtained from the multivibrator. However, when both voltages appear at the same time and add in the positive direction, current will flow. The voltage will add together on one side of the discriminator when one mixer voltage leads, and on the other side of the discriminator when the voltage from the other mixer leads.

Note that, whenever one of the diodes conduct, only the voltage pips

riding atop the wave from amplifier No. 1 appear in the output. If the voltages oppose each other, nothing appears at that side of the discriminator.

The positive pulses appearing on either side of the discriminator are kept separate from each other and are passed through separate pulse amplifiers and a pulse limiter. These stages bring all pulses to the same amplitude, removing any variations that might possibly exist between them. The pulses are then applied to a pulse integrator circuit in such fashion that half of the integrator responds to the pulses appearing on one side of the circuit and half of the integrator responds to the pulses appearing on the other side of the circuit. The currents flowing through the first half of the integrator charge a capacitor in one direction while the currents flowing through the second half of the integrator charge the same capacitor in the opposite direction. The voltage developed across the capacitor is then applied through a cathode follower to the modulator control tube where it can alter the center or resting frequency of the modulated oscillator.

When the master oscillator is on frequency, the average voltage developed across the capacitor over one audio cycle is zero. For all other conditions either a positive or negative resultant voltage will appear across the capacitor which, when applied to the modulator control circuit, will act to bring the master oscillator back on frequency.

In spite of the apparent complexity of this circuit, it is simple to operate because it requires no adjustments once it has been placed in operation. There are no tuned circuits, frequency dividers, or locked-in oscillators. Most tubes operate from cut-off to saturation, reducing to a minimum the effect of any variation in tube characteristics. Actually, under these conditions, the only positive manner in which a tube can interfere with the proper circuit operation is by complete failure, and generally tubes are removed from the circuit, during periodic inspections, long before this occurs.

F-M BY TUBE CAPACITANCE VARIATION

In still another approach to direct frequency modulation, advantage is taken of the fact that the input capacitance of a tube is dependent upon its mutual conductance. The mutual conductance, in turn, will vary as the plate current varies and, going one step farther, this will depend upon the grid voltage. Thus, tying all these facts together, we have a direct connection between the grid voltage of the tube (in this instance, the modulator tube) and the input capacitance of the tube. If, therefore, we apply an audio voltage to the modulator grid, we can cause the input capacitance of the tube to vary at an audio rate. By connecting the tuned circuit of an oscillator across the grid of the modulator tube, we effectively shunt this input capacitance across the oscillator and cause the generated frequency to vary in step with the applied audio voltage. The result is frequency modulation.

The basic and actual circuits of a suitable modulator and oscillator are shown in Fig. 15.8. The oscillator uses a 12J5GT triode arranged to operate as a conventional Hartley. Connected across part of the oscillator tank in-

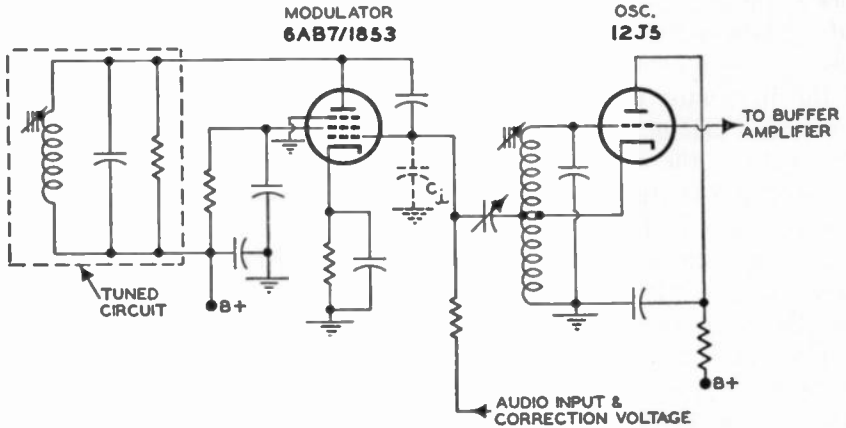


FIG. 15.8A. A simplified diagram of the modulator and oscillator used in a transmitter where F-M is obtained by varying the input capacitance.

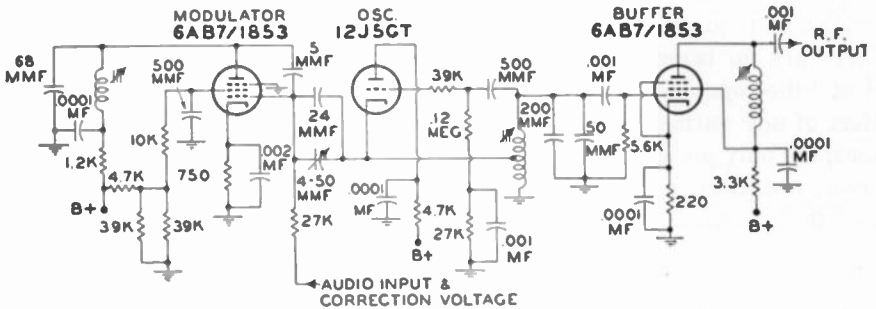


FIG. 15.8B. The complete schematic of the modulator and oscillator. Shown, too, is the buffer which follows the oscillator.

ductance is the modulator tube, a 6AB7/1853 pentode. This particular tube was chosen because it possesses a high g_m value and because this g_m can be made to vary linearly with grid voltage. The latter qualification is necessary to insure that the frequency modulation produced is directly proportional to the applied audio-modulating signal.

It can be shown by a rather lengthy mathematical analysis that the input capacitance of an amplifier tube is given by

$$C_i = C_{pk} + C_{gp}(1 + A)$$

where C_i = input capacitance of amplifier tube

C_{gk} = grid-to-cathode capacitance of tube

C_{gp} = grid-to-plate capacitance of tube

A = gain of the tube

For all practical purposes, C_{gk} and C_{gp} are constant and will not change appreciably with tube current. The gain, on the other hand, will vary with tube current because, for a pentode, the gain, A , can be shown to be equal to $g_m Z_L$, where g_m is the mutual conductance of the tube and Z_L is the load impedance into which the tube works. Hence, the foregoing equation can be rewritten in the form

$$C_i = C_{gk} + C_{gp}(1 + g_m Z_L)$$

Note that C_i is not only affected by the g_m of the modulator tube, but also by Z_L , the load impedance of this tube. In order to have C_i affected only by g_m , the plate load is made resistive by using a resonant circuit, which is resistive at its resonant frequency. A resistor is shunted across the resonant circuit in order to have it tune broadly and therefore remain in resonance over the range of frequencies covered by the oscillator when it is frequency-modulated. With this circuit modification, the equation for the input capacitance of the modulator becomes:

$$C_i = C_{gk} + C_{gp}(1 + g_m R_L)$$

where R_L now represents the load impedance.

Since the 6AB7/1853 modulator tube is pentode, the C_{gp} of this tube is very small, on the order of 0.015 mmf, and its effect, in comparison to C_{gk} (which is 15 mmf) would be almost negligible. To overcome this, a 5-mmf capacitance is shunted across the tube, from plate to grid providing sufficient input capacitance variation with changes in audio signal. Another capacitor, this one variable, connects the grid of the modulator tube to the tuning inductance of the oscillator. Its purpose is to provide a fixed frequency swing for a given change in audio voltage. To prevent too much R.F. signal from appearing on the modulator grid, this variable capacitance is connected to a tap on the tank inductance.

The audio voltage is applied to the grid of the modulator tube, and as it varies the current through the tube the g_m will change and with it the value of the input capacitance. The center frequency of the oscillator is set at some value between 3.66 and 4.5 mc, so that with a frequency multiplication of 24 (3 doublers and a tripler) the transmitted carrier frequency ends up between 88 and 108 mc. The 75-ke deviation for 100 per cent modulation is 75/24 or 3.12 ke at the oscillator frequency.

The sequence of low-power stages beyond the modulated oscillator consists of a series of three doublers and a tripler, at which point the signal

F-M SIMPLIFIED

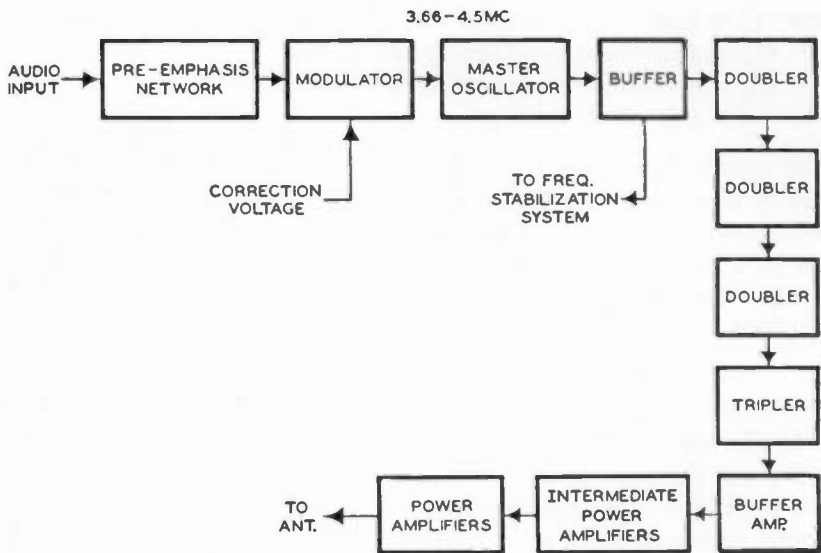


FIG. 15.9. A block diagram of the R.F. stages of an F-M transmitter employing the modulating circuit of Fig. 15.8.

frequency is between 88 to 108 mc. (See Fig. 15.9.) The signal is now transferred to a buffer and an intermediate power amplifier where its power is increased to 250 watts. Beyond this point, power amplifiers raise the signal power to 1, 3, 10, 20, or 50 kw, depending upon the authorized power for the particular station.

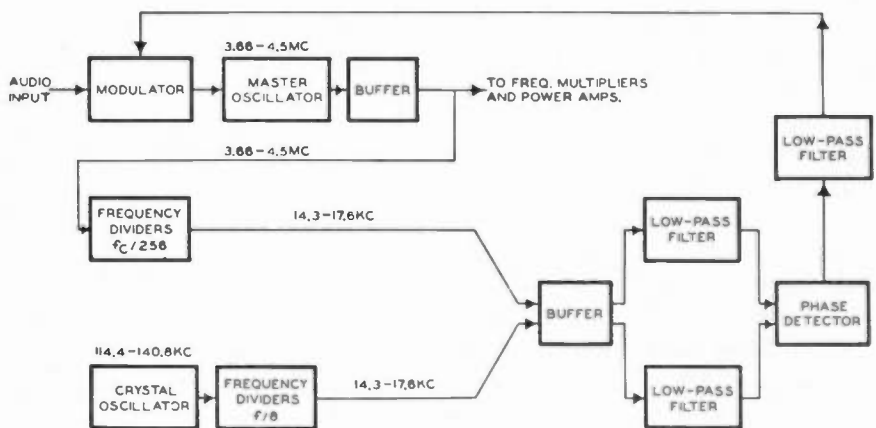


FIG. 15.10. A block diagram of the frequency stabilization system employed in the F-M transmitter of Fig. 15.9.

Frequency Stabilization System. To maintain the master oscillator within 2000 cycles of its assigned frequency (as required by the F.C.C.) a frequency stabilization system is incorporated into this transmitter. A block diagram of this system is shown in Fig. 15.10. The frequency output from a buffer immediately following the frequency-modulated oscillator is fed to a frequency-divider chain where the 3.66 – 4.5 mc frequency is divided 256 times to a frequency between 14.3 – 17.6 kc. (It is understood that only one center frequency will be obtained from the master oscillator and consequently only one frequency will appear at the end of the divider chain.)

At the same time, a crystal oscillator, operating at some frequency between 114.4 kc and 140.8 kc, has its signal pass through two frequency-divider stages where the initial frequency is lowered by a factor of 8 (2×4). This will produce a frequency between 14.3 and 17.6 kc, or the same frequency that the master oscillator should produce at the end of its frequency chain if it is exactly on frequency. The two divided signals are then passed through separate buffer amplifiers and low-pass filters and applied to the phase detector. The latter stage is identical with the phase detector described in Fig. 14.20.

PROBLEMS

1. Explain the function of each block in Fig. 15.1.
2. How is modulation achieved in this F-M transmitter? Describe in detail, using simple schematic circuits.
3. What is the purpose of the audio feedback discriminator in the transmitter of Fig. 15.1?
4. Draw a block diagram of the frequency stabilization system used in this transmitter.
5. Explain briefly the operation of the frequency stabilization system.
6. In what way does the F-M transmitter of Fig. 15.9 make use of the fact that the input capacitance of a tube is dependent upon the mutual conductance of that tube? Explain fully.
7. Draw a simple schematic circuit of the modulator and master oscillator of this transmitter (Fig. 15.9).
8. What type of frequency stabilization system does the F-M transmitter of Fig. 15.9 employ? Illustrate with a block diagram.
9. Would it be possible to interchange the frequency stabilization system employed in the transmitter of Fig. 15.1 for the unit used in the transmitter of Fig. 15.9? Explain fully, indicating what changes might be required in the stabilization circuit frequencies or components, if any.
10. What factors govern the input capacitance of an amplifier tube? Give the mathematical relationship that pertains.

Chapter 16

COMMERCIAL F-M TRANSMITTERS

(PART 3)

F-M Through Phase Modulation. It was previously seen that, indirectly, frequency modulation is produced when a carrier is phase-modulated. The development of F-M by this method is extensively employed because it is possible to have a crystal oscillator produce the transmitter frequency. The crystal oscillator not only provides a high degree of frequency stability and accuracy, but it eliminates the need for any frequency-controlling network, such as required in transmitters producing F-M directly. In the latter system, it will be recalled, it is not possible to use a crystal in the main oscillator because the F-M is produced here. In phase-modulated systems, the F-M is developed beyond the oscillator and crystal control is feasible.

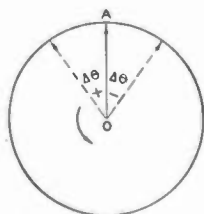


FIG. 16.1. Phase modulation represented by vectors. $\Delta\theta$ is the maximum phase shift of the carrier.

Before we investigate the operation of any phase modulator, let us briefly review the production of F-M by phase shifting or phase modulation. Phase modulation is produced when the carrier wave, as represented by vector OA in Fig. 16.1, is made to shift back and forth as it spins around at its fixed frequency. As a result of this shifting, we obtain the same effect as if the frequency were instantaneously being varied. The variation, superimposed on the regular, fixed frequency of the carrier, represents the frequency modulation.

The indirect F-M produced depends (1) upon the maximum angle that the carrier wave is shifted, as shown in Fig. 16.1, and (2) on the frequency at which the shifts take place. Mathematically, we can say that

$$\Delta F \text{ (the frequency swing)} = f \cdot \Delta\theta$$

where f = frequency at which the carrier wobbles or shifts back and forth
 $\Delta\theta$ = maximum angular shift of carrier in radians

This relationship should be kept in mind throughout the ensuing discussion. We will have recourse to it again later in the chapter.

Phase Variation. A change in the phase of a signal can be produced by passing the signal through a network containing resistance and reactance. When a voltage is applied to a capacitor and a resistor in series, for example, the current leads the applied voltage by an amount dependent on the relative values of resistance and capacitive reactance. This current develops a voltage across the resistor which leads the applied voltage. If the series combination is considered to be the input, and the output voltage is taken from across the resistor, a definite amount of phase shift is introduced. If the fixed-frequency signal from a crystal oscillator is passed through this network, its phase at the output is shifted by an amount depending on the ratio of the reactance to the resistance. If the resistor can be varied, the phase angle of the network changes to correspond with the newly established ratio of reactance to resistance. When the resistance is varied with an applied audio signal, the phase angle of the output changes in direct proportion to the audio-signal amplitude and produces a phase-modulated signal.

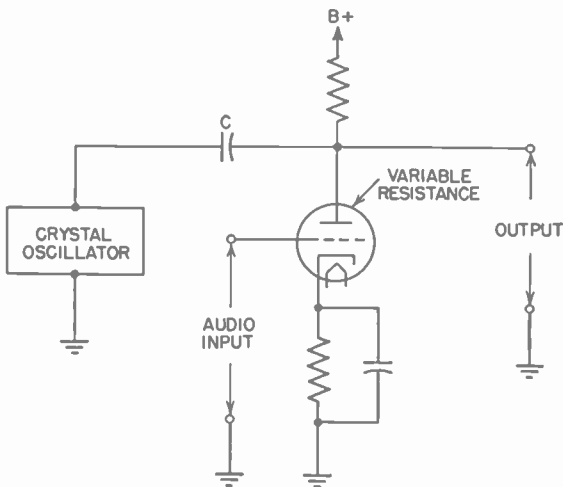


FIG. 16.2. A simple phase modulator.

P-M to F-M. The basic circuit of a phase modulator based on this approach is shown in Fig. 16.2. Here, the variable plate resistance of a triode replaces the resistor mentioned above. The plate resistance of the triode

changes with grid voltage and therefore serves as the variable resistor. Since the plate resistance of the triode varies with the audio signal applied to the grid circuit, the phase between the input to the circuit and the output changes with the audio signal. As the grid swings positive, the plate resistance drops and the phase angle of the output increases; when the grid swings negative, the plate resistance rises and the phase angle decreases. The change of plate resistance with various values of grid voltage is exactly proportional to grid voltage over a small range. If the phase angle of the network is changed between wide limits, the *amplitude* of the output changes. This means that the modulator can produce only a limited phase deviation without distortion. In general, it is reasonably good only over a range of less than 25° of phase shift.

Constant-Impedance Phase-Shift Modulator. In order to overcome to some extent the change in output signal amplitude as the phase varies, a constant-impedance network is employed. A phase modulator having such a network is shown in Fig. 16.3. The cathode resistor, R_k , is connected es-

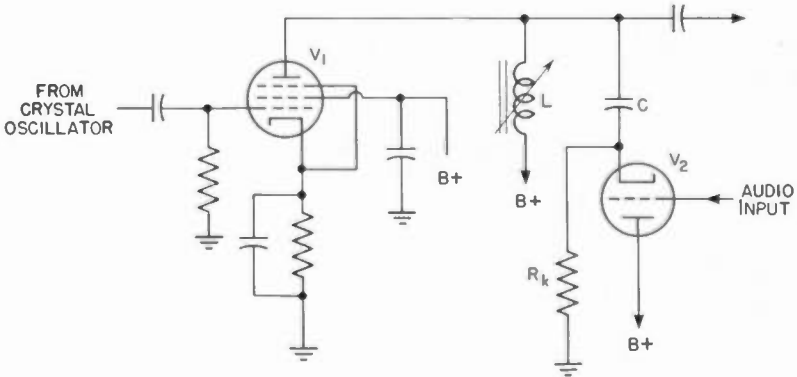
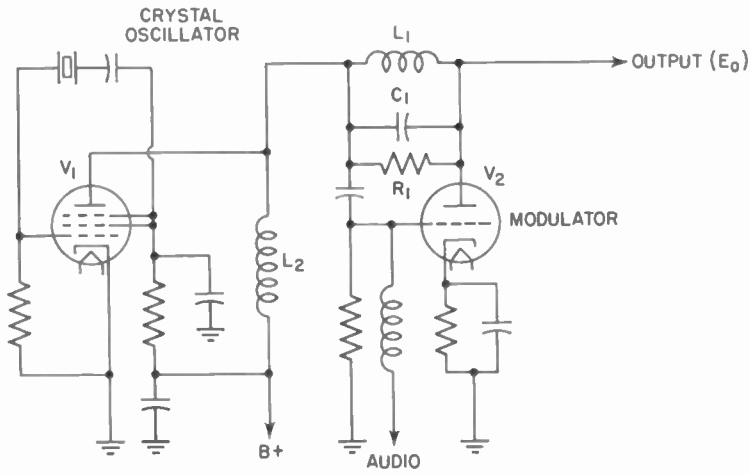


Fig. 16.3. Constant-impedance phase-shift modulator.

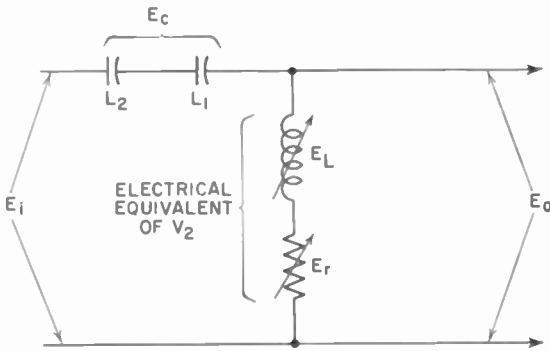
entially in parallel with the plate resistance of the tube. R_k remains fixed, but the tube plate resistance varies with the audio signal. Since the two are in parallel, the total resistance will vary with the audio signal. However, since only one resistor changes value, the total change will be less than if one resistor alone is employed, as in the previous circuit. It will also be found that the resistance variation with signal level will be more linear.

The inductor, L , in Fig. 16.3 serves to prevent any change in the total impedance, keeping the amplitude of the output constant. For any change in frequency, a change in capacitive reactance is canceled by an opposite change in inductive reactance.

G_m Phase Modulator. Another approach to phase modulation which depends on variations in transconductance is shown in Fig. 16.4A. The output of a crystal oscillator is fed to a modulator, V_2 , in which feedback exists between plate and control grid. This feedback network, consisting of



(A)



(B)

FIG. 16.4. (A) Actual circuit of 6M phase modulator. (B) Equivalent circuit.

L_1 , R_1 , and C_1 , is tuned somewhat above the operating frequency of the oscillator. Thus, this network presents a capacitive reactance to the crystal signal. Coil L_2 , the output load of the crystal oscillator, also is selected to present a capacitive reactance at the oscillator frequency. That is, its resonant frequency is above the oscillator frequency. Under these condi-

tions, it can be shown that the tube itself appears as a series combination of a coil and a variable resistance. (See Fig. 16.4B.) (The latter action, of course, is due to the signal feedback between plate and grid of V_2 .)

The values or magnitudes of the internal resistance and inductance of V_2 depend upon the mutual conductance of the tube. The latter, in turn, is governed by the audio modulating voltage applied to the control grid of V_2 .

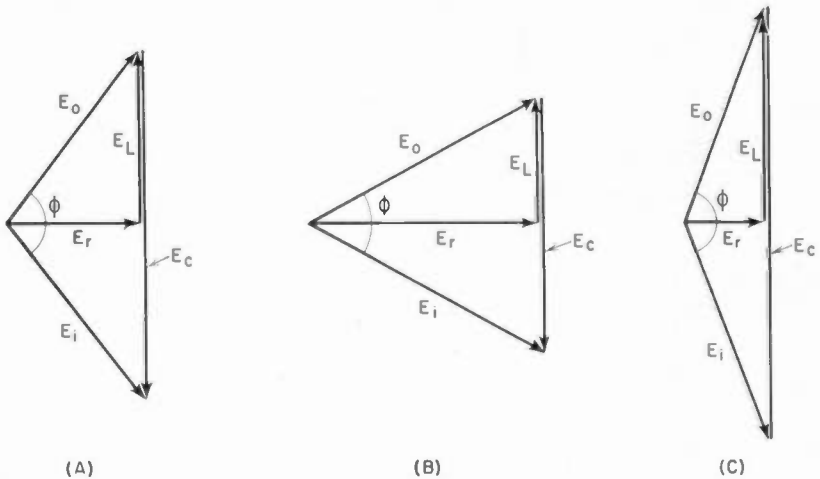


Fig. 16.5. How the voltages vary (with modulation) in the circuit of Fig. 16.4.

The vector diagrams of Fig. 16.5 illustrate what happens in this circuit. E_i is the input voltage to the modulator from the oscillator. E_o is the output voltage from the modulator. E_r is the voltage developed across the resistive portion of the internal impedance of V_2 , while E_L is the voltage developed across the inductive segment of V_2 . Finally, E_c represents the sum of the voltages developed across L_2 and across L_1 , R_1 , and C_1 . When no modulating voltage is present, the vector condition represented by Fig. 16.5A holds. When the grid is driven more negative, E_r increases and the vectors shift to the condition shown in Fig. 16.5B. Finally, when the grid is driven more positive, E_r decreases (indicating that the internal impedance of the tube decreases), and the various voltages assume the relationship shown in Fig. 16.5C.

Note the changing phase relationship between E_o and E_i in Fig. 16.5. This can vary between 0 and 180°. Finally, observe that, even with this wide a phase change, E_o remains fairly constant in amplitude. This is a major advantage of this method of phase modulation and considerable use is made of this circuit in mobile communications equipment. Distortion is very low, less than 2%.

Reactance-Tube Phase Modulator. It is possible to produce phase modulation by connecting a variable reactance across the resonant load circuit of an R.F. amplifier (Fig. 16.6). The variation in the phase angle

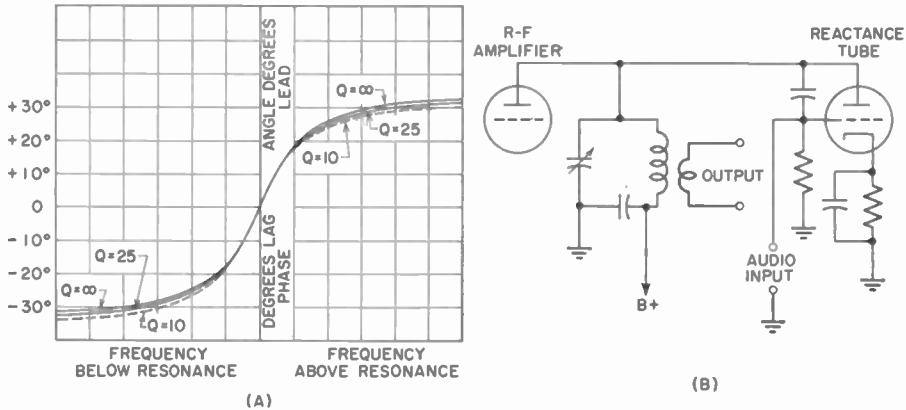


FIG. 16.6. Reactance-tube phase modulation.

of a parallel resonant circuit shown in (A) is plotted in respect to the frequency. As the frequency of the applied voltage increases, the capacitance in the circuit begins to be predominant, and the resultant total current leads the applied voltage. When the frequency of the applied voltage is lower than the natural resonant frequency of the tuned circuit, the inductance has a lower reactance than the capacitor and, consequently, draws the major share of the current. Therefore, the total current lags the applied voltage.

If the frequency of the resonant circuit is varied and the applied voltage kept constant, the same variation of phase angle is produced. The curve in (A) also applies, except that the center frequency is that of the applied voltage, whereas the frequencies on either side are those to which the resonant circuit is tuned. The circuit can be tuned by changing the value of either the inductance or the capacitance. Changing either at an audio rate with a reactance modulator produces phase modulation. The shape of the phase variation curve depends on the Q of the tuned circuit. This, in turn, depends almost entirely on the construction of the inductor. Therefore, injection of capacitance usually is employed to avoid changes in the shape of the curve.

The reactance modulator, in (B), is designed to inject a variable capacitance across the resonant-circuit load of an R.F. amplifier. Note that plate voltages for the modulator and the amplifier are supplied in common. The injection capacitance changes the tuning of the tank circuit in response

to the variations in audio signal. This, in turn, changes the phase angle of the current drawn by the tank. Consequently, the phase angle of the output from the tank circuit also varies. The curve in (A) shows that the change in phase is proportional to the detuning over a small range approximately between the limits of -25° and $+25^\circ$. This circuit, therefore, cannot produce much phase deviation. It has the advantage of permitting the use of a reactance tube with a crystal oscillator-amplifier combination. However, the wide deviations associated with the reactance modulator when used with a self-excited oscillator cannot be obtained.

Nonlinear Coil Modulator. A coil can be constructed which will have the property, when both radio and audio frequencies are passed through it, of introducing phase modulation into a carrier. Such a coil is called a "non-linear coil modulator" (Fig. 16.7A). The output from the R.F. amplifier

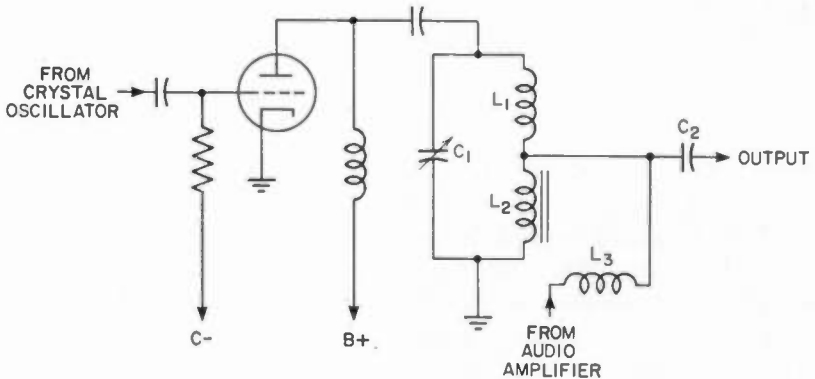


FIG. 16.7A. Nonlinear coil modulator.

is passed through a plate load consisting of the resonant circuit C_1 , and L_1 , and the special nonlinear coil, L_2 . L_2 is wound on a special permalloy core together with winding L_3 which carries the audio signal from the audio system. The rest of the circuit is a conventional R.F. amplifier. The nonlinear coil circuit produces a frequency deviation of nearly 1 kc at the output. This phase modulator circuit is relatively efficient in terms of the amount of initial phase deviation.

Normally, when a current is passed through an air-core coil the current flow has the same wave shape as the applied voltage. However, if a magnetic core is inserted into the coil, the situation changes. When a magnetic field exists in a magnetic material, there is a definite magnetizing force corresponding to that field. As the field increases in strength, the material becomes magnetized until a point is reached where the increase in the magnetizing force produces no increase in the magnetic field set up. When the

material is fully magnetized and the magnetic flux cannot increase, a state called *saturation* is reached.

When current flows in a coil, it sets up a magnetic field that magnetizes the core material placed in the immediate vicinity. As the current increases, the corresponding magnetic field increases, as does the magnetization of the core. Because of saturation, however, there is a definite point beyond which any additional current causes no additional magnetization. Special alloys, such as permalloy, when used as cores reach the saturation point at very low values of the magnetizing field. A coil wound around a permalloy core reaches saturation with a very small amount of applied current. When a sine-wave voltage is applied to such a coil, the magnetic flux increases rapidly until the core saturates; after this, the flux becomes relatively constant.

The relation between the applied magnetizing current and the voltage developed across the coil is shown in Fig. 16.7B. When the current begins to increase toward its maximum value, the magnetization of the core rises rapidly, with a rapid increase in flux. While the current flow is above saturation (A to B), there is no change in the flux, since the core is saturated. The same situation occurs on the negative half-cycle between points C and D.

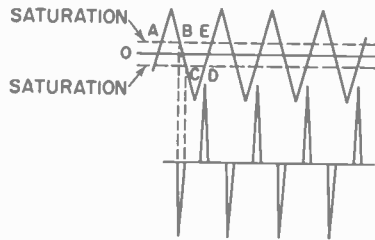


FIG. 16.7B. Voltage pulses developed across nonlinear coil.

The induced voltage depends on the rate of change of the flux. When the flux is not changing, no voltage is induced. In a coil wound on a permalloy core, the flux is changing rapidly during part of the cycle and, during that time, large voltages are induced in it. These voltage pulses of high amplitude occur *only* during the periods of rapidly changing flux. At other times, when the flux is nearly constant, little or no voltage is induced across the coil. It is shown in the illustration that these voltage changes take place exactly 90° after the current peaks. The polarity of the pulses depends on the direction of the magnetic flux. Therefore, on opposite half-cycles of magnetizing current, the pulses are of opposite polarity. This 90° difference is constant in respect to the magnetizing current, and, since this current is supplied by an R.F. oscillator, the pulse is constant in frequency.

Assume, however, that, in addition to the R.F. energy, audio signals are simultaneously applied to the nonlinear coil. They have the same magnetizing effect on the core material and they combine with the R.F. current to produce the current wave, as shown in the first three lines of Fig. 16.8. Curve A represents the current in the coil, caused by the carrier R.F. Curve B is the current produced by the modulating signal, assumed to be

sinusoidal, for simplicity of analysis. The combined waveform is shown in line C. It is clear that the resultant current no longer goes through the zero axis in the same time interval as before, and, therefore, the region of maximum rate of change of flux is different for each cycle and depends on the audio voltage. These combined currents produced voltage pulses across the coil at different instants during the audio cycle, as in D. The variation

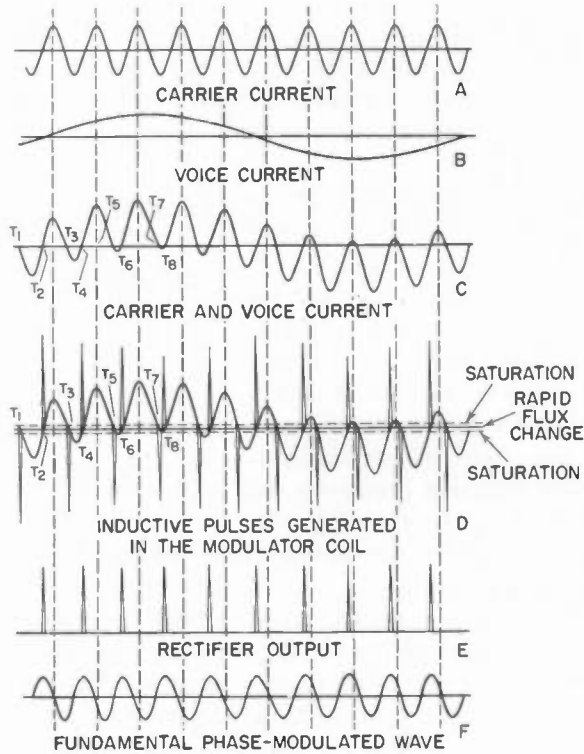


FIG. 16.8. Waveforms developed in nonlinear-coil modulator.

in the level of the R.F. current at different points of the R.F. cycle causes this effect. Sometimes, the pulse is produced at the normal interval of the unmodulated carrier. At other times, the pulses are spaced more or less than 360° apart. These variations in the spacing of the pulses, with different values of modulation voltages, are obviously equivalent to displacements in the relative phase of the pulses. In other words, a change in the A.F. voltage shifts the phases of the pulses in respect to the phase of those produced by the unmodulated carrier. Therefore, these pulses are effectively phase-modulated.

If the phase-modulated pulses of voltage derived from the nonlinear coil are applied to a rectifier and limiting amplifier, only the pulses of one polarity will be passed. Furthermore, the limiting action will reduce the slight variations in the amplitude of the pulses that appear in D. The action of this rectifier and limiter is shown in E. When pulses of sharp amplitude pass through a resonant circuit, they set it into oscillation at its natural resonant

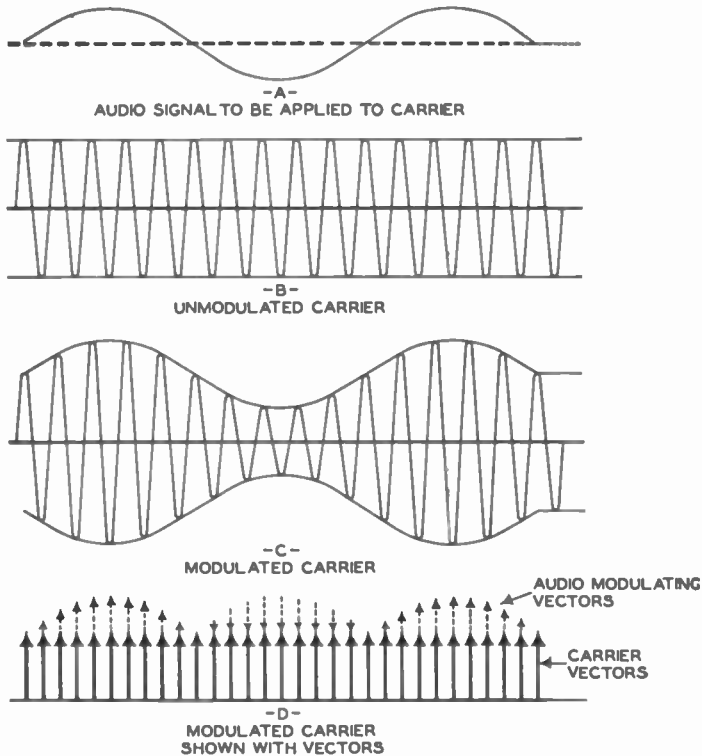


FIG. 16.9.

frequency. If these pulses from the output of the rectifier pass through a resonant circuit, which is tuned to their repetition frequency, a sine wave is produced. Since the pulses are phase-modulated by the audio voltage, the resultant sine wave also will be phase-modulated.

Balanced Modulator. A system of phase modulation that received considerable attention from the broadcast industry employs a balanced modulator developed by Major Armstrong. In order fully to appreciate how this system functions, let us vectorially compare A-M and P-M.

Figure 16.9 illustrates the conventional and the vectorial representation of amplitude modulation. In conventional notation, the low frequency audio

signal at (A) varies slowly from 0° to 360° . Directly beneath it, in (B), we have the unmodulated carrier. When the audio signal voltage is applied to the carrier, the result is a modulated signal containing all the intelligence (whatever it may be) of the audio voltage. This is shown in (C). In (D), we have the vector equivalent of the modulated carrier. Note how the amplitude changes at each point, as a result of the addition or subtraction of the audio voltage. Its frequency remains untouched, although, as we have seen, the sidebands formed in this process differ in frequency from the carrier by an amount equal to the audio-modulating voltage.

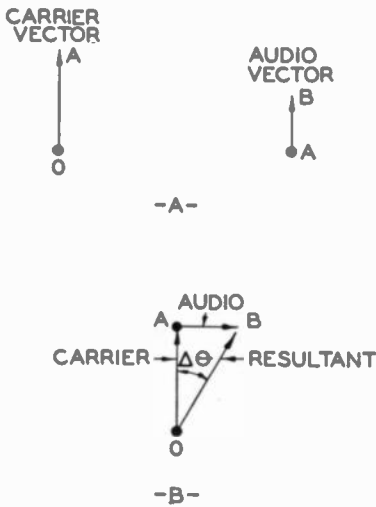


FIG. 16.10. (A) The separate audio and carrier vectors (OA and AB) and their vector addition to produce P-M. (B) $\Delta\theta$ is the phase difference between the unmodulated carrier and the modulated resultant.

Let us examine this critically. In Fig. 16.10A, the unmodulated carrier and the audio-modulating voltage are shown separately. If we combine them 90° out of phase as in Fig. 16.10B, then a resultant vector is formed which is displaced from the original position of the unmodulated carrier by some small angle. In other words, by combining the two voltages as shown, we have produced a resultant carrier which is displaced slightly from OA.

As the audio signal goes through one complete cycle, its amplitude will vary and, with it, the angular shift of the resultant carrier vector from the position it would occupy if no modulation voltage were present. The variation of the angular shift or displacement at various points in the audio-

In contrast to this, we have the formation of a phase-modulated wave. At each instant during modulation, we must apply the audio-modulating voltage so that it causes the phase of the carrier to advance or retard. It is this shifting, first in one direction, then in another, that is illustrated by the vector in Fig. 16.1. Now, it was noted in amplitude modulation, that the audio voltage increased or decreased the amplitude of the carrier vector. This meant that the audio voltage was being applied either in phase or 180° out of phase with the carrier voltage.

In phase modulation, we do not wish to alter the amplitude of the carrier, merely its relative phase from moment to moment. This phase change can be accomplished by adding the audio voltage vectorially at right angles to the carrier voltage or, electrically, by combining them 90° out of phase with each other.

modulating voltage cycle is shown in Fig. 16.11. Note that the resultant vector (which is the phase-modulated wave) first occupies a position to one side of the unmodulated carrier; then it moves to the other side. In this way it fluctuates back and forth, just as it was pictured in previous discussions. From the wobbling or shifting, an equivalent frequency modulation arises.

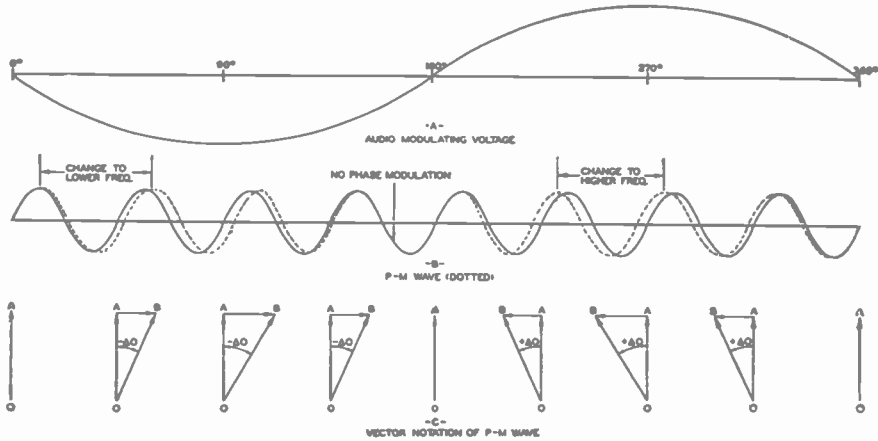


Fig. 16.11. Illustrations of phase modulation.

The formation of frequency modulation, or the variation in frequency in the phase-modulated wave, is clearly indicated in Fig. 16.11. The solid sine wave drawn at the center represents the unmodulated carrier. The other wave, drawn with dashes, is the resultant vector (or the modulated wave) at various instants during the application of the audio signal. At the left-hand side of the diagram the resultant wave is shifted behind the unmodulated carrier's position. This means, in effect, a decrease in frequency. Therefore, the dotted wave, at this point, spreads out (lower frequency) and the maximum peaks are farther apart than those of the unmodulated carrier (shown by the solid lines).

When, immediately after this, the audio-modulating voltage drops to zero, the dotted line blends into the solid line, indicating both are of the same frequency. With no audio voltage, there is no modulation.

On the positive half of the modulating cycle, the resultant, or phase-modulated carrier, is shifted to the opposite side, indicating that the phase of the carrier has advanced. This is equivalent to a slight increase in frequency, demonstrated by a comparison of the dotted and the solid curves. We note that the modulated carrier peaks move closer together, effectively increasing the frequency. Finally, at the zero point (360°), the modulation disappears and the phase-modulated carrier is again at its central value.

Armstrong System of F-M. In the Armstrong system of F-M we do not directly apply the audio voltage to the carrier as shown in Fig. 16.11. However, we obtain the same results. We start with a fixed source of high frequency voltage, a crystal oscillator (see Fig. 16.12). The output from the

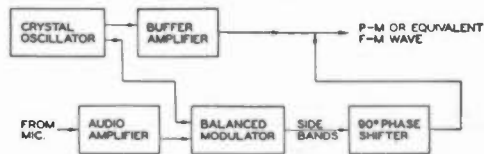


FIG. 16.12. The basic Armstrong system for generating F-M waves.

oscillator is then directed into two different paths. One path contains a buffer amplifier; the other path is through a balanced modulator where the audio-modulating voltage is applied to the carrier signal. At the modulator output, only the sidebands appear. The carrier voltage is suppressed because of the operation of the balanced modulator. The sidebands are then shifted in phase by 90° and combined with the unmodulated carrier present at the output of the buffer amplifier. The final result is a phase-modulated wave and, indirectly, also frequency-modulated.

The sideband frequencies obtained from the balanced modulator contain the intelligence of the audio voltage, much the same as though we had amplitude-modulated a carrier. The sidebands are then shifted in phase by 90° and combined with the carrier from the buffer amplifier.

Let us pause for a moment and compare the Armstrong method with the system in Fig. 16.10 for obtaining phase modulation. In one instance we apply the audio-modulating voltage directly to the carrier (maintaining 90° phase relationship) to obtain phase modulation whereas in Major Armstrong's system we use the sidebands produced by amplitude modulation and combine these with the carrier. In both methods, the results are seemingly the same and yet, obviously, both methods are not exactly alike. How does this happen?

The answer lies in the fact that Major Armstrong's system is definitely limited in application. First, we can produce phase modulation by applying the audio voltage directly to the carrier. This is entirely analogous to amplitude modulation, except for the manner in which the audio voltage and the carrier are combined. Here, the audio voltage is *directly* shifting the carrier's relative phase position and, if the system is properly designed, there will be no distortion produced for fairly wide angles of phase shift.

Major Armstrong's method will produce the desired phase and frequency modulation from the combination of a carrier and two sidebands *only* when the resulting carrier has a maximum phase shift 30° or less. Ex-

pressed in radians (one radian is equal to 57.3°) 30° is approximately 0.5 radians. If these voltages are combined so as to give a greater phase shift to the resultant, then along with the phase modulation we have amplitude modulation. This represents distortion and a waste of useful power. But, by keeping the phase shift within the above limits, pure phase modulation is obtained.

It can be shown mathematically that, when a carrier is phase-modulated at angles 30° or less, the resulting modulated wave will contain a total of only two sidebands, one above and one below the carrier. This is shown in Fig. 16.13. Furthermore, each sideband will be 90° out of phase with the carrier. Hence, in the Armstrong system, we take a carrier and separate it into two parts. One part is modulated by the audio signal to produce the necessary sidebands. These sidebands are then shifted 90° and combined with the remaining half of the carrier (which was fed into the buffer amplifier, Fig. 16.12) to produce the phase or frequency modulation.

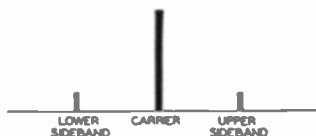


FIG. 16.13. A low value of phase shift produces only two sidebands.

To recapitulate, the Armstrong method can be used because:

1. A carrier, phase-modulated at angles of 30° or less, will produce only two sidebands.
2. Each sideband is 90° displaced from the carrier.

In the Armstrong system we take (a) one carrier and (b) two sidebands and combine them 90° out of phase to obtain a phase-modulated wave. Note, however, that this method is valid only for small phase shifts.

The actual Armstrong unit has the block arrangement shown in Fig. 16.12. The equivalent schematic is shown in Fig. 16.14. The quartz-crystal oscillator operates near 200 kc (depending upon the assigned frequency of the transmitter), and its output is applied to a buffer amplifier and a balanced modulator. Since the control grids of the balanced modulator tubes are connected in parallel, both grids become positive and negative in step with each other. The modulator plates, however, are connected in push-pull, with the plate current from each tube passing through the load circuit coils (L_1 and L_2) in opposite directions. Thus, in the absence of an audio voltage, each field cancels the effect of the other and no voltage is obtained across L_3 .

Upon the application of an audio-modulating voltage at the input terminals of T_1 , a varying potential is applied to the screen grids of V_3 and V_4 . Suppose, for example, that the screen grid of V_3 is going more positive. At the same time, the screen grid of V_4 is less positive since it is attached to

the opposite side of the transformer secondary winding. Because of the increased positive voltage, the current in V_3 will increase while the current in V_4 is lowered by a corresponding amount. In a push-pull transformer arrangement, such as we have for L_1 , L_2 , and L_3 , a decrease in one coil, say L_2 , and an increase in L_1 combine to aid each other and to produce a fairly large voltage across L_3 .

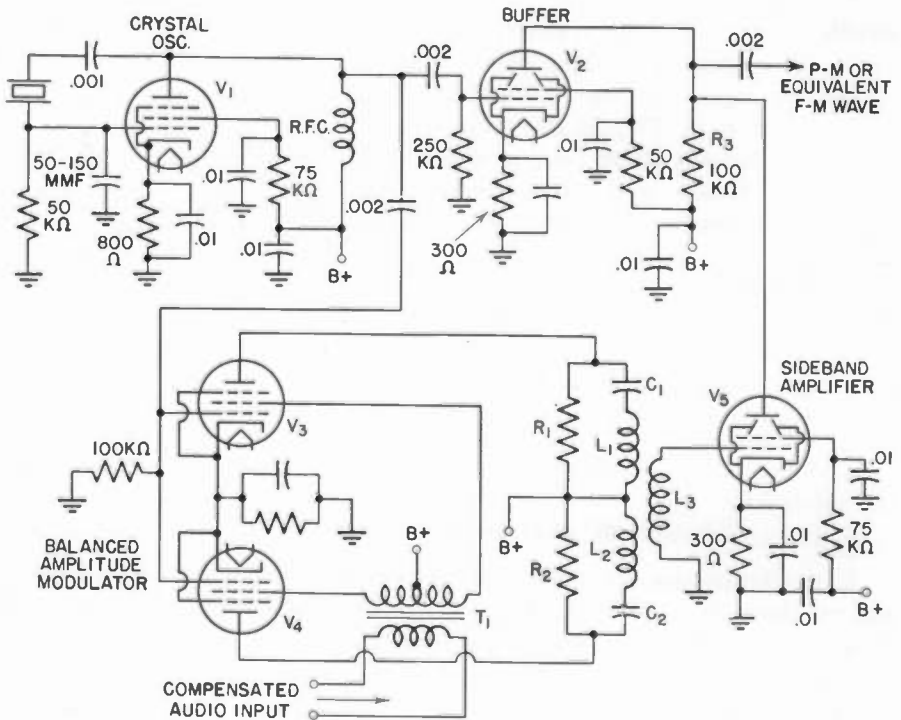


FIG. 16.14. The schematic diagram of the basic Armstrong modulation.

On the next audio half cycle, when the screen grid of V_3 becomes less positive and the screen grid of V_4 is driven more positive, the opposite set of conditions prevails. The voltage across L_3 is also of opposite polarity. Thus, until the audio voltage is active, nothing appears across L_3 . The effect of the audio signal is to unbalance the circuit and produce the sideband voltages across L_3 .

The arrangement of components in L_1 and L_2 also produces a 90° phase shift, which is accomplished through the insertion of capacitors C_1 and C_2 in series with L_1 and L_2 . These capacitors neutralize the inductive reactance of their respective coils (C_1 for L_1 , C_2 for L_2) and present to the tube a purely resistive load. Under these conditions, the grid voltages of each

modulator tube is in phase with its plate current. The plate current, in flowing through the load, will develop a magnetic field about each coil. The variation of a magnetic field is directly dependent upon the current that produced it. Hence, the control grid voltage of each tube, the plate current of that tube, and the magnetic field set up about the coil are all in phase.

The magnetic field, in cutting across L_3 , will induce a voltage. From elementary a-c theory we know that an induced voltage is maximum when the magnetic flux lines are changing most rapidly. This rapid change occurs when the field is going from positive to negative (or negative to positive) values. However, the induced voltage is minimum or zero when the flux is hardly changing at all. This absence of change occurs at the positive and negative peaks (see Fig. 16.15). If we plot the relationships of induced

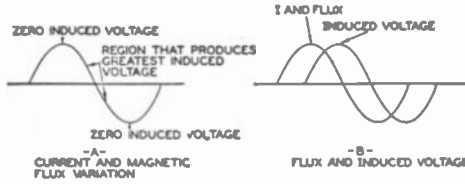


FIG. 16.15.

voltage to magnetic flux variations, we immediately see that a phase difference of 90° is introduced. In this manner we produce the necessary sidebands and shift them 90° with respect to the crystal oscillator carrier.

Tube V_5 amplifies the sidebands and combines them with the carrier. Resistor R_3 is the common load for V_2 and V_5 , and it is here that the combination of carrier and sidebands takes place.

Frequency Swing Compensation. Thus far, we have been concerned solely with the stage or stages where the phase modulation occurs. Once this is achieved, however, the next step is frequency multiplication, not only to raise the carrier frequency to the desired value, but also to increase the bandwidth occupied by the signal. However, before frequency multipliers are considered, it is first necessary to determine how much frequency shift is produced by a certain phase change. For this, we refer again to the equation

$$\Delta F = f \cdot \Delta\theta$$

where f = frequency of audio modulating voltage

$\Delta\theta$ = maximum phase shift in radians.

Actually, in the foregoing equation, $\Delta\theta$ should be $\Delta\theta \sin 2\pi ft$ because the angle varies from instant to instant. However, we are interested only in the maximum value of the phase swing; hence we use $\Delta\theta$. The maximum value of $\sin 2\pi ft$ is 1, and $\Delta\theta \times 1$ is $\Delta\theta$. $\Delta\theta$ should be expressed in radians, which

is, as we have seen, equal to the angle in degrees divided by 57.3; f is expressed in cycles.

Let us assume a maximum phase swing of 30° is produced in the phase modulator. At the highest audio frequency, utilizing the full phase swing of 30° (0.524 radian), the equivalent frequency swing or variation is:

$$\Delta F = 15,000 \times 0.524$$

$$\Delta F = 7860 \text{ cycles}$$

$$\Delta F = 7.86 \text{ kc}$$

Now, let us see what happens when low frequency audio signals are applied. The lowest audio frequency desirable is generally 50 cycles. Substituting this in the equation,

$$\Delta F = 50 \times 0.524$$

$$\Delta F = 26.2 \text{ cycles}$$

Note that both signals gave the same *phase* swing, yet the low frequency voltage produced a frequency variation of only 26.2 cycles whereas the highest audio frequency produced a frequency variation of 7.86 kc.

Suppose we transmitted the carrier with these frequency shifts. At your receiver, which would give the greatest output? The 7.86 kc, of course, since all F-M receivers develop a greater output for a wider frequency swing. But this does not represent the conditions at the transmitter. Here, both signals were equally strong. Something, then, must be done to place both signals, or all frequencies for that matter, on an equal footing. No matter what the frequency, signals of equal amplitude should produce equal frequency shifts.

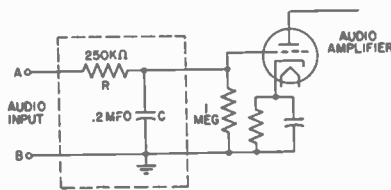


FIG. 16.16. A predistorter network.

To achieve this, a predistorter circuit is inserted in the audio amplifier stages (see Fig. 16.16). The resistor R is made very large in comparison to the reactance of capacitor C at all audio frequencies. This means that for any voltage impressed across terminals A, B , the current will be determined by the resistor R . The effect of

C is negligible. Thus, no matter what the audio frequency, the same voltage will produce the same current through R and C .

Across C , the voltage will depend upon the frequency. A higher frequency will develop less voltage across C than a lower frequency. This is because the capacitor offers less opposition to higher frequencies. Hence, we obtain the desired inverse effect to counterbalance the rising f effect of phase modulation. The predistorter components of R and C apply less

high frequency voltage to the grid of the tube. (In the phase modulators previously shown and discussed in this chapter, no such network was shown, but in each case one would be included.)

Since the low frequencies produce the smallest equivalent F-M, the audio-correction network reduces the effect of all frequencies higher than the lowest audio frequency—say, 50 cycles—so that, in the end, all the frequencies of the same strength will produce the same frequency modulation. The amount of equivalent F-M, then, that a 50-cycle voltage produces will determine how much multiplication is to be applied to obtain the final F-M swing.

Let us return to the circuit of Fig. 16.14. Utilizing the maximum phase shift of $\pm 30^\circ$, we discovered that, at 50 cycles, the carrier has a frequency variation of ± 26.2 cycles. The carrier frequency, at this point in the circuit, is the crystal oscillator frequency. For the circuit of Fig. 16.14, as employed in commercial F-M broadcasting, this frequency value would be around 200 kc. Hence, our F-M signal is 200 kc with a frequency swing of ± 26.2 cycles. What we desire, at the antenna, is the assigned station frequency, say, 90 mc ± 75 kc. The problem is to increase the 200 kc ± 26.2 cycles to 90 mc ± 75 kc. In order to deal with round figures, let us change the 26.2 cycles to 25 cycles.

Frequency Multiplication. If we divide 75,000 cycles (the desired frequency swing) by 25 cycles (the swing we now have) we see that a frequency multiplication of 3000 ($75,000/25$) is necessary. Although it may be a trifle difficult to have exactly a 3000 increase using triplers and doublers, we can get very close, say, 2916. This frequency multiplication is possible if we use the following assortment of doublers and triplers: 2, 3, 3, 3, 3, 3, 3, 2. The signal is passed through successive stages until the total multiplication of 2916 is obtained.

The frequency multiplication is applied not only to the frequency swing (the sidebands) but also to the carrier. This means that the original 200 kc of the crystal oscillator will be multiplied 2916 times—a final output frequency of 583,200 kc or 583.2 mc—resulting in a value considerably above the desired 90 mc. Something must be done to prevent the carrier from rising to a frequency higher than its assigned frequency, in this case, 90 mc, and, at the same time, to obtain the maximum frequency deviation of ± 75 kc for the sidebands.

A method to accomplish this is shown in Fig. 16.17. The entire transmitter is arranged in block diagram, with each unit numbered for easy identification purposes. At the input to the amplifier in block No. 6, the equivalent frequency modulation has been achieved. Thereafter, the carrier and its sidebands are sent through a series of frequency multipliers until, by the time the mixer is reached (block No. 13), the carrier has been increased to

32.4 mc and the frequency variation to ± 4.05 kc. Note that at the input to the mixer the carrier is 32.4 mc whereas at the output it has been reduced to 5 mc. The reason is the mixing action within the tube.

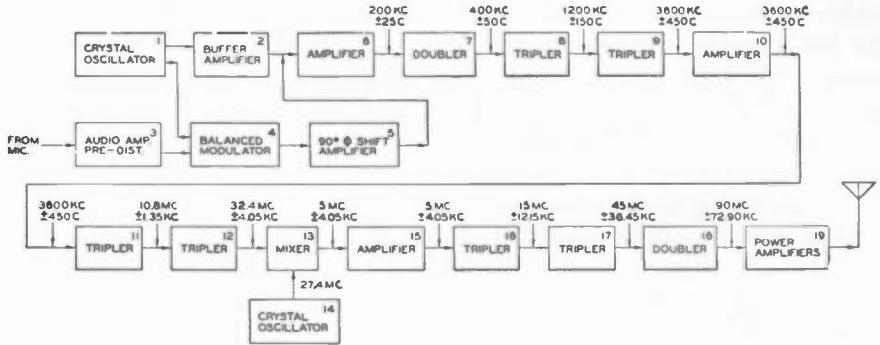


Fig. 16.17. A complete block diagram of the basic Armstrong F-M transmitter.

Whenever we mix two signals, the output consists of the original two mixing frequencies and the sum and difference frequencies. Of primary interest is the difference frequency, 5 mc. To accentuate this frequency and minimize or eliminate the others, we place a resonant circuit at the output of the mixer which is peaked at 5 mc. With this simple arrangement it is possible to reduce the original carrier to the point where it can be further increased to the desired 90-mc values.

But what of the sidebands during this process? The incoming signal to the mixer has a frequency variation of ± 4.05 kc. The range, then, of the input signal is from 32.39595 mc ($32.4 \text{ mc} - 4.05 \text{ kc}$) to 32.40405 mc ($32.4 \text{ mc} + 4.05 \text{ kc}$). We change 4.05 kc to 0.00405 mc and then add and subtract this figure with 32.4 mc to arrive at the foregoing values. When the 27.4-mc mixing frequency is added, the following output difference frequencies are derived:

$$\begin{array}{r}
 32.39595 \\
 -27.40000 \\
 \hline
 4.99595
 \end{array}
 \qquad
 \begin{array}{r}
 32.40405 \\
 -27.40000 \\
 \hline
 5.00405
 \end{array}$$

or, what is the same thing, $5 \text{ mc} \pm 4.05 \text{ kc}$. Look at this carefully for it demonstrates clearly that by mixing it is possible to lower the carrier frequency and, at the same time, retain the original frequency variation.

Hereafter, the tripling and doubling sequence is straightforward until we arrive at the 90 mc ± 72.9 kc. By assuming a starting frequency variation of 25 cycles, we obtain only a ± 72.9 -kc frequency shift at the output. However, with the 26.2 cycles proposed originally, we would come much

closer to ± 75 kc. In any event, we do not have to hit it exactly on the head. Anything slightly less is still quite satisfactory.

There is one more fact to be noted concerning the Armstrong transmitter. In addition to the audio predistorer, or audio-correction network, recently discussed, we also require a pre-accentuator. The pre-accentuator is used here as in the reactance-tube F-M transmitter, namely, to raise the level of the higher audio tones in order that they may be in a better position to combat the effect of noise at these frequencies. At the receiver, a de-accentuator or a deemphasis circuit returns the higher frequencies to their proper level.

Do not confuse the purpose of the two circuits. The audio correction network is needed to counteract the tendency of the higher frequencies to produce more equivalent frequency modulation than the lower frequencies, assuming both signals are equal in strength. This circuit is required *only* in phase-modulation systems.

The pre-emphasis network is placed in *all* F-M transmitters to produce a better signal-to-noise ratio at the receiver.

In F-M transmitters to be employed for purposes other than commercial broadcasting, other carrier frequencies would be employed. Further, the F-M bandpass would also be different, generally less. Hence, other multiplier arrangements would be utilized. The method (and the purpose) of multiplication, however, would remain the same as just discussed.

The Dual Channel Armstrong System. A modification of the preceding system which provides improved frequency stability and lower distortion has been achieved through a dual channel system of phase modulation. Essentially, phase modulation is still used to produce frequency modulation. The sidebands are generated in a balanced modulator and then combined 90° out of phase with the original carrier. The method of combination and the subsequent process of mixing are, however, quite different from the preceding modulator.

The basic circuit of the Armstrong Dual Channel Modulator is shown in Fig. 16.18. A crystal oscillator functions at a frequency near 200 kc, the same frequency as in the previous modulator. A buffer stage follows the crystal to serve as an isolation amplifier, improving circuit stability. If desired, the crystal oscillator can be placed in a temperature-controlled oven, although this is not necessary in this system. As we shall see presently, any variations in this oscillator have no effect on the final transmitter frequency.

The output of the buffer amplifier is coupled to the modulator input coil, L_1 (see Fig. 16.18B). The voltage appearing across L_1 is applied to the balanced modulator—consisting of tubes V_1 and V_2 —and to points C , D . In the modulator stages, the carrier signal is shifted 90° and then audio-modulated to produce the upper and lower sidebands. After this, the sidebands are recombined with the carrier at points C , D . It is here that the

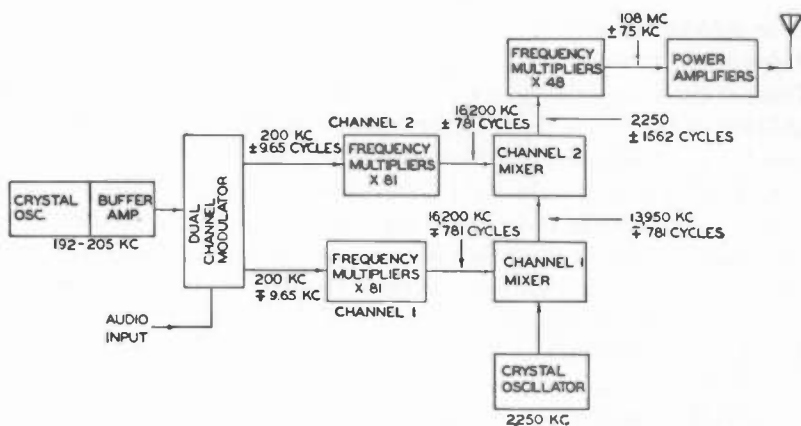


FIG. 16.18A. A block diagram of the dual channel modulator circuit.

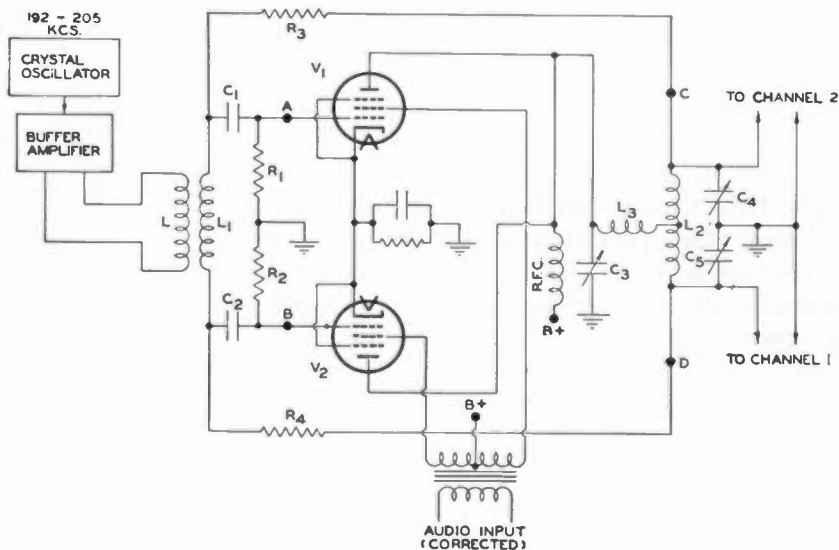


FIG. 16.18B. The modulating circuit of the dual channel F-M transmitter.

phase modulation (and, with it, of course, the frequency modulation) appears.

Modulator Operation. At the balanced modulator, the signal at L_1 is applied to the phase-shifting network, C_1R_1 and C_2R_2 . Both capacitors are equal in value, and the same is true of the resistors. At the frequency used by the carrier, the reactances of C_1 and C_2 are much greater than the resistance of R_1 or R_2 . Hence, the current through each branch will be prac-

tically 90° ahead in phase, with respect to the voltage across L_1 . Since the voltages across R_1 and R_2 are in phase with the current through the resistors, we note that the voltage received by each grid will lead the carrier voltage at L_1 by 90° . The required 90° phase shift has thus been acquired. Now to the production of the sidebands.

The grids of V_1 and V_2 are excited 180° out of phase, whereas their plates are connected across a common load. During those intervals when no audio-modulating voltage is present at the screen grids, no output is produced. Upon the application of an audio voltage, one tube draws more current than the other and sideband voltages appear. These are coupled through C_3 and L_3 to L_2 where the carrier voltage from the modulator input coil is directly applied.

To the carrier voltage arriving from L_1 , points C and D appear as the terminals of a purely resistive network. L_2 in conjunction with C_4 and C_5 forms a parallel resonant circuit at the carrier frequency of 200 kc. Because of the resistive impedance between points C , D , the carrier voltage here is in phase with the voltage at L_1 . Resistors R_3 and R_4 are merely inserted for the purpose of preventing undesirable interaction between the modulator and L_2 , C_4 , and C_5 .

The ground connection between C_4 and C_5 produces equal and opposite carrier frequency voltages for the two separate channels of amplifiers. The equality exists because C_4 and C_5 are equal in value.

In the circuit, to this point, the 90° -shifted sidebands and the direct carrier (or, at least a portion of it) are at points C and D . To understand how phase modulation is achieved it will be necessary to examine in greater detail the electrical action of L_2 , C_3 , L_3 , C_4 , and C_5 . These components of Fig. 16.18B have been rearranged as shown in Fig. 16.19. In the rearrangement, L_2 has been separated into two mutually coupled coils of equal inductance, L_{2A} and L_{2B} . Their total inductance is equal to L_2 .

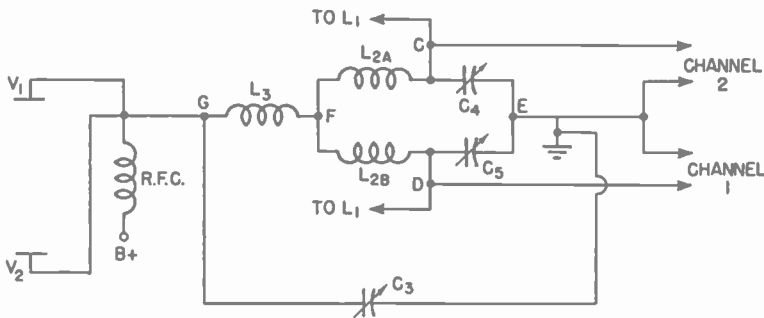


FIG. 16.19. The electrical network wherein the sideband and carrier voltage combine.

The carrier voltage is brought to points C , D . These two points also lead to the two tripler channels, as can be seen from Fig. 16.18B. Because L_{2A} plus L_{2B} equal L_2 , and $L_2C_4C_5$ is resonant to the carrier frequency, then the entire series branch shown in Fig. 16.19, around points C , E , D , and F , is resonant to the carrier frequency.

To the carrier voltage, this network offers maximum impedance because it is connected as a parallel resonant circuit. It is also to be noted that, since both arms of the network (L_{2A} and C_4 or L_{2B} and C_5) are equivalent, each is resonant to the carrier frequency. This leads to the following result: Any voltage placed between points C and D "sees" a parallel resonant circuit. However, any voltage applied between F and E "sees" two series resonant circuits. This latter fact should be kept in mind while we consider the introduction of the sidebands between points F and E .

The sideband voltages are applied from the R.F. choke in the plate circuit of V_1 and V_2 to L_3C_3 and the network within terminals $CDEF$. L_3 in conjunction with C_3 resonates at the carrier frequency. The presence of L_{2A} , L_{2B} , C_4 and C_5 in the resonant network of L_3C_3 does not produce any interference because a voltage at the carrier frequency applied between F and E "sees" only resistance. Thus, whether we are considering the circuit in Fig. 16.19 from the terminals leading to points C and D or from the terminal leading from the balanced modulator, we have a completely resonant circuit presented to the carrier frequency. This design is purposely effected in order to obtain the correct phase relationships when the sideband voltages are applied.

The sideband voltages are applied to this circuit between point G and ground. In the inductive branch of the resonant circuit (the branch containing L_3 and the network from point F to ground) the current lags the sideband voltages by 90° . This is true of all inductive branches. The inductive current flows down through L_3 , and, at point F , divides evenly between $L_{2A}C_4$ and $L_{2B}C_5$. Each of these two branches appears resistive to the sideband current, since the sideband frequencies are so very close to the carrier frequency.

The sideband currents, in flowing down each branch, develop voltages across C_4 and C_5 . However, across a capacitor, the voltage lags the current by 90° . Hence, the voltage that appears across C_4 and C_5 is 90° (owing to its flow through L_3) plus an additional 90° (owing to the voltage and current relationships across a capacitor) out of phase with respect to the sideband voltages applied by the modulator to terminal G and ground. And, owing to the modulator input network, C_1 , C_2 , R_1 , and R_2 , another 90° shift was introduced. The overall result is this: the sideband voltages appearing across C_4 and C_5 are $180^\circ + 90^\circ$, or 270° , out of phase with the carrier voltages at C_4 and C_5 . But 270° in one direction (say, clockwise)

is equal to 90° in the opposite (or counterclockwise) direction. Thus, the sideband frequencies and the carrier are combined 90° out of phase with each other. This is the resired result.

To recapitulate, we have the sideband frequencies formed at V_1 and V_2 and shifted 90° (by $C_1, C_2, R_1,$ and R_2) from the carrier. Then, in the network composed of $C_3, L_3, L_2, C_4,$ and C_5 , the sideband frequencies are shifted an additional 180° . Since $180^\circ + 90^\circ$ is equal to 270° , and 270° clockwise is the same as 90° counterclockwise, we find that the sidebands and the carrier are combined 90° out of phase with each other. This produces phase modulation and, with it, equivalent frequency modulation.

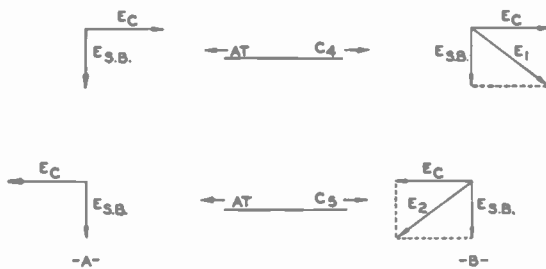


FIG. 16.20. The vectorial addition of the sideband and carrier voltage.

The sideband voltages at C_4 and C_5 are equal to each other, both in amplitude and phase. We can indicate this fact in Fig. 16.20A by making their vectors equal to each other and pointing in the same direction. To add the carrier voltages to these vectors, we note that across C_4 and C_5 they are equal but *opposite* in polarity. The carrier voltages at C_4 and C_5 are opposite in polarity because C and D (from whence they are applied) are at opposite ends of the input coil, L_1 (see Fig. 16.18B). They differ, in other words, by 180° . This is shown in Fig. 16.20A. Adding the vectors for the sidebands and the carrier voltages, we obtain the result shown in Fig. 16.20B. One resultant vector, E_1 , lags its carrier voltage; the other resultant vector, E_2 , leads its carrier voltage. In terms of equivalent frequency variations, this means that if, in channel No. 1, the frequency swing is $+9.65$ cycles, then, in channel No. 2, the corresponding frequency swing is -9.65 cycles. When one frequency swing is positive, the other frequency swing is negative.

Each channel contains four tripler stages, multiplying each signal frequency by $3 \times 3 \times 3 \times 3$, or 81. Hence, the input 200 kc, with the associated 9.65-cycle frequency swing, becomes at each channel output 16,200 kc. Each has a frequency variation of 781 cycles. This is shown in Fig. 16.18A.

The output of each channel is fed into a separate mixer. Feeding into the first mixer is the voltage from a 2250-ke crystal oscillator and the 16,200

kc \mp 781 cycles derived from channel No. 1. As a result of the mixing process, a difference frequency of $13,950 \text{ kc} \pm 781 \text{ cycles}$ is produced. This is fed to a second mixer, where it meets the incoming signal from channel No. 2. We obtain a carrier output from the second mixer of $16,200 - 13,950 \text{ kc}$, or 2250 kc . For the sidebands we have $\pm 781 - \mp 781 \text{ cycles}$, or $\pm 781 + \pm 781 \text{ cycles}$, or $\pm 1562 \text{ cycles}$. Note what has been done here. With the negative or minus sign, we have converted $\mp 781 \text{ cycles}$ to $\pm 781 \text{ cycles}$. Arithmetically and electrically, this is what actually occurs. Hence, we obtain a signal from the second mixer which has twice the frequency swing of either channel. From here, four doublers and a tripler (providing a multiplication of $2 \times 2 \times 2 \times 2 \times 3 = 48$) increases $2250 \text{ kc} \pm 1562 \text{ cycles}$ to $108 \text{ mc} \pm 75 \text{ kc}$. One or more power amplifiers then boost the signal power to its final output value.

In the Armstrong Dual Channel Modulator, the crystal control oscillator feeding the first mixer regulates the stability (and the value) of the final carrier value. In the foregoing illustration, if we desire an output carrier of 90 mc , the crystal control oscillator would have to be lowered to 1875 kc . Any other carrier frequency in the band 88 to 108 mc can be obtained accordingly.

Frequency Stability. The stability of the carrier frequency may be seen indicated by the following example. Assume that the crystal oscillator feeding the modulator unit shifts in frequency by 1 kc , say, from 200 kc to 199 kc . In each channel the incoming 199 kc is multiplied 81 times to a final value of $16,119 \text{ kc}$. If the control oscillator is set at 2250 kc , then the difference frequency output from the first mixer is $13,869 \text{ kc}$. At the second mixer, $13,869 \text{ kc}$ and $16,119 \text{ kc}$ mix to produce 2250 kc , which is exactly what we wish. The shift from 200 kc to 199 kc had no effect upon the frequency modulation. Hence, no matter to what value the first crystal oscillator changes, the output frequency from the second mixer will be entirely determined by the crystal control oscillator. Actually, all that the two channels and the balanced modulator do in this system is to produce the frequency modulation. The frequency variation is increased 81 times in the channels and then transferred to the control crystal oscillator carrier.

Although it makes no difference what change occurs in the frequency of the first oscillator, too much shifting would tend only to reduce the amplitude of the signal from each channel because of attenuation in the tuned circuits. Hence, it is customary to use a crystal oscillator, although any of the other types of familiar oscillators (Hartley, Colpitts, T.G.T.P., etc.) would be satisfactory.

The phase-shift method of securing frequency modulation produces a signal with very little distortion. It is claimed that the dual-channel principle has even decreased the previously small amount of distortion present.

The measured rms harmonic distortion is less than $1\frac{1}{2}$ per cent for all audio frequencies between 50 and 15,000 cycles at 25 per cent, 50 per cent, and 100 per cent modulation. This is well within the standards of "Good Engineering Practice" of the F.C.C. as applied to broadcast stations.

The Serrasoid F-M Modulator. An F-M modulator which is capable of securing a relatively large initial phase shift with only four tubes was

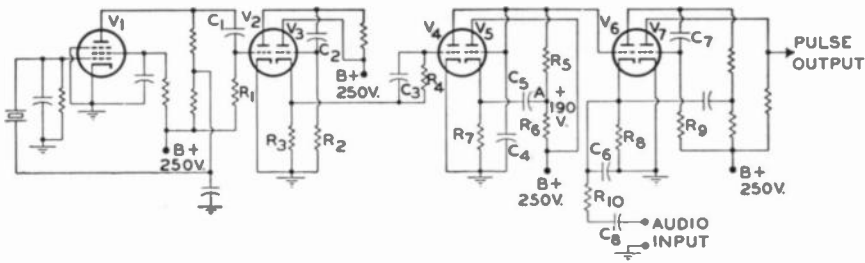


FIG. 16.21. The Serrasoid F-M modulator.

developed by J. R. Day. The name of this new unit is the Serrasoid F-M Modulator. For a modulating frequency of 50 cycles, peak deviations close to ± 100 cycles can be obtained. Since an F-M carrier is permitted a maximum frequency deviation of ± 75 kc, or 75,000 cycles, a fairly high multiplication is required. The figure generally falls between 800 and 1,000, depending on the deviation achieved in the oscillator.

The schematic circuit of the Serrasoid modulator is shown in Fig. 16.21. V_1 is a crystal oscillator which generates a frequency having a value of $1/972$ of the final carrier frequency.* By placing the crystal in an oven, the frequency generated can be held within ± 0.0002 per cent of its assigned value, thereby insuring that the final carrier stability is of the same order.

The oscillator circuit is so designed that the crystal current is very small, and the tube conducts for only a small fraction of time. The result is that narrow negative-going pulses are produced at the plate of the tube. (See Fig. 16.22A.) These pulses are applied to the grid of V_2 through a differentiating network composed of C_1 and R_1 which tend to narrow down the pulses still further. At the grid of V_2 the pulses plunge the tube into cut-off producing positive-going pulses

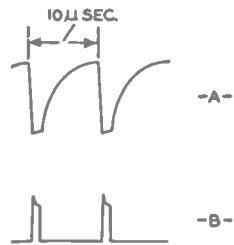


FIG. 16.22. (A) Voltage waveform at plate of V_1 Fig. 16.21. (B) Voltage pulses across R_3 of Fig. 16.21.

* This is for the circuit as developed by J. R. Day. In the commercial unit shown in Fig. 16.24, it would be $1/864$ of the final carrier frequency.

with flat tops at the plate. These pulses are applied through C_2 to the grid of V_3 , a cathode follower. This tube is biased beyond cut-off by the combination of cathode and grid-leak bias, the latter being developed by the combination of C_2 and R_2 and the positive pulses which are fed to the tube by V_2 . Because of its method of biasing, V_3 clips the bottom of the pulses, producing a pulse across R_3 which has steep sides and is fairly square. (See Fig. 16.22B.)

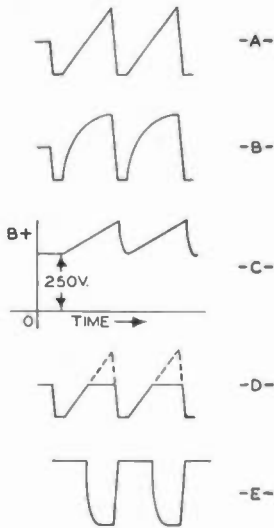


Fig. 16.23. Additional waveforms of Fig. 16.21. (See text for explanation.)

The pulses now are fed to V_4 which is a non-oscillating saw-tooth wave generator. Here is how it operates. The grid initially possesses no bias. Upon the arrival of a positive-going pulse from R_3 , grid current flows from V_4 into C_3 , charging this capacitor to essentially the peak value of the applied pulse. During the interval between pulses, C_3 discharges slowly through its shunting resistor, R_4 , but because of the high value of R_4 , the voltage across C_3 is sufficiently high to keep V_4 cut-off except when the pulses are active.

A capacitor, C_4 , is connected across the output circuit of V_4 . When the tube is cut-off, the voltage across C_4 rises because one end of this capacitor is connected through R_5 and R_6 to the B+ power supply while the other end connects to ground. C_4 continues to charge until a pulse arrives at the grid of V_4 , at which time V_4 is driven sharply into conduction. The pulse reduces the internal resistance of V_4 to a low value, and, since V_4 is directly across C_4 , the capacitor discharges rapidly through this low resistance. At the end of the pulse, V_4

again returns to cut-off, and C_4 again starts its charging.

Now, it is most important that the voltage across C_4 rise as linearly as possible (Fig. 16.23A) because the linearity of the modulation process depends upon it. Normally, however, the charging of a condenser is not linear but exponential, as shown in Fig. 16.23B. This is so because, as the voltage across a charging capacitor rises, it tends to buck or counteract the applied B+ voltage, reducing the ability of this latter voltage to keep the charging current constant. To insure that the rise will be linear, a "bootstrap" amplifier comprising V_5 , C_5 , and R_7 is added. The grid of V_5 is connected directly across C_4 , so that whatever voltage is present across C_4 becomes the grid voltage for V_5 .

During the charging period of C_4 , the grid voltage of V_5 is rising, causing the current through this tube to increase. This results in a rising voltage

across R_7 . Since C_5 is connected between the top of R_7 and R_5 , the voltage rise across R_7 is transferred, via C_5 , to the B+ voltage present at point A. Thus the B+ charging voltage for C_4 has superimposed on it the rising voltage from R_7 . (See Fig. 16.23C.) This increase in B+ voltage offsets the voltage rising across C_4 , keeping the current flowing into C_4 constant and producing a linear rise in voltage across this capacitor.

C_4 is also connected to the grid of V_6 , a cathode-biased amplifier. The bias on this tube is so adjusted that conduction begins when the saw-tooth voltage across C_4 has attained only half of its maximum value. (See Fig.

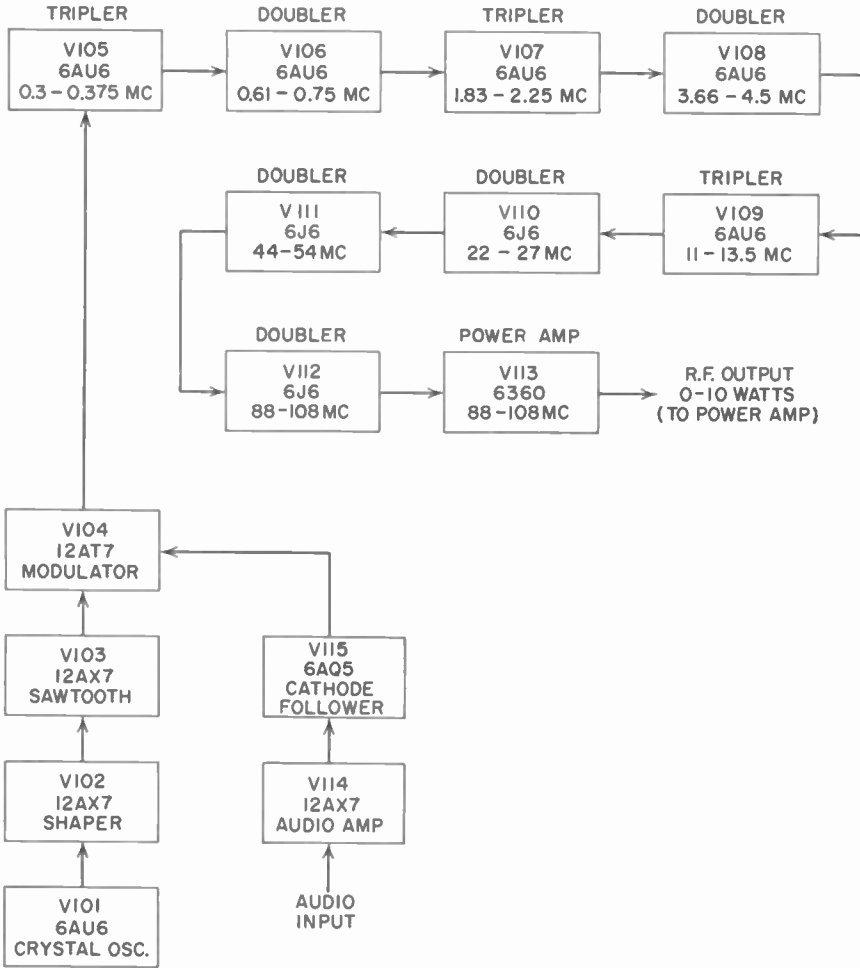


FIG. 16.24. A block diagram of a commercial F-M transmitter using the Serrasoid system. (Courtesy: Gates Radio Co.)

16.23D). 0.25 microsecond after V_6 starts conducting, grid current is drawn, stopping the charging of C_4 and maintaining its voltage at this value until discharge occurs, after which the process is repeated. The voltage at the plate of V_6 drops abruptly from 250 volts to that corresponding to a fairly low value and remains there until C_4 discharges, at which time it rises steeply back to 250 volts because the tube is cut-off. The plate waveform is shown in Fig. 16.23E.

Feeding into the cathode of V_6 is the audio-modulating voltage. As this voltage varies, it changes the bias on the tube and, therefore, the time when conduction will begin. When the audio voltage is positive, the start of tube conduction is delayed because a more positive cathode is equivalent to a more negative grid. Conversely, when the audio voltage is negative, tube conduction will commence sooner. This periodic advance and delay of the start of tube conduction causes the leading edge of the plate voltage pulses at the output of V_6 to become phase-modulated. (The end of tube conduction always occurs at the same time since this is controlled by the pulses coming from V_3 .)

Before the audio signal is applied to V_6 , it is acted on by a correction circuit (composed of C_6 and R_{10}) which counteracts the tendency of the higher audio frequencies to produce more equivalent frequency modulation than the lower frequencies, assuming both signals are equal in strength. This circuit, it will be recalled, is required when frequency modulation is derived from phase modulation, as it is here.

The pulses at the plate of V_6 are made narrower by passage through a differentiating network consisting of C_7 and R_9 and then applied through V_7 to a string of doublers and triplers. In the resonant circuits of these multipliers, the pulses are converted to sinusoidal waves possessing the same amount of phase or frequency modulation.

The block and schematic diagrams of a commercial F-M transmitter utilizing the Serrasoid method of modulation are shown in Figs. 16.24 and 16.25. After the initial modulated signal is formed by V_{104} , three triplers and five doublers multiply the signal 864 times to a final carrier value between 88 and 108 mc with 100% modulation (± 75 kc). The signal, at the output of V_{113} , has a maximum average power of 10 watts. This is then raised to the desired final level (250 watts, 100 watts, 5000 watts) by additional power stages.

In spite of the names which this manufacturer has given stages V_{101} through V_{104} , they function in essentially the same manner as V_1 through V_7 of Fig. 16.21. R_{168} and C_{112} serve the same purpose here as C_6 and R_{10} of Fig. 16.21. That is, they counteract the tendency of the higher audio frequencies to produce more equivalent F-M than the lower audio frequencies.

Approximately 30 volts rms of audio is required at the output of the

audio amplifier stage, as measured at TP_{118} , to modulate the F-M carrier 100%. Since the audio is applied to the cathode of V_{104} , the output impedance of the audio system should be low. For this reason, V_{115} is a cathode follower. V_{114} and V_{115} work together as a unit since they operate within a feedback loop. The negative feedback is provided by C_{163} and R_{158}, R_{159} in series. The audio input stage, V_{114} , is driven by audio input transformer, T_{102} . A suitable pre-emphasis network would be inserted between T_{102} and the audio input source.

In the frequency multiplier portion of the schematic diagram, V_{105} through V_{112} , the circuitry is fairly conventional except in one or two places. For example, between V_{105} and V_{106} there are three tuned circuits, L_{101} , L_{102} , and L_{103} . The purpose of these three circuits is to present to V_{106} a sine wave in which every cycle possesses the same amplitude. V_{105} receives a series of sharp spikes or pulses from V_{104} . The pulses are being applied at a frequency rate of 116 kc.* The plate circuit of V_{105} , consisting of L_{101} in parallel with a resistor and capacitor, is tuned to three times 116 kc or 348 kc. As each of the sharp spikes occurring at 116 kc arrives at the grid of V_{105} , the tube suddenly conducts and shocks the plate circuit of V_{105} into oscillation at its resonant frequency of 348 kc. If an oscilloscope were connected directly to the plate of V_{105} , these oscillations would appear as shown in Fig. 16.26A.

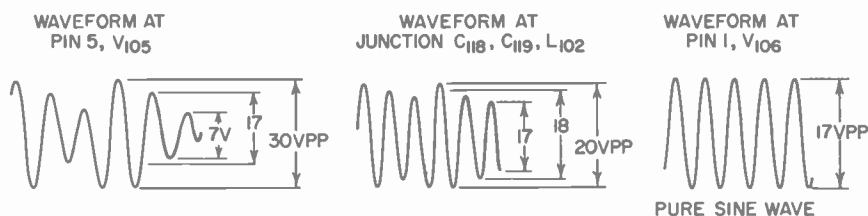


FIG. 16.26. Waveforms in the circuits between V_{105} and V_{106} of Fig. 16.25.

These oscillations, it will be noted, are damped. This indicates that the Q of the circuit is fairly low which, in the present instance, is due to the resistor shunted across L_{101} . If we removed the resistor, the three sine-wave cycles would more closely possess the same amplitude. However, the signal reaching V_{105} possess a frequency modulation of ± 116 cycles. At the output of this tube, the modulation has been increased to ± 348 cycles. It is the purpose of the resistor across L_{101} to broaden the circuit response so that the frequency can be shifted from 348 kc *plus* 348 cycles to 348 kc *minus* 348 cycles. Without the loading imposed by the resistor, the Q of the tuned

* This is a representative figure. The actual value would be governed by the assigned carrier frequency of the station.

circuit would be high and the flywheel effect of the plate circuit of V_{105} would prevent the circuit from following the frequency excursions faithfully and distortion would result.

The tuned combination of L_{102} and its shunting capacitor together with L_{103} and its shunting capacitor form a bandpass filter that removes the amplitude variation of 116 kc from the 348-kc signal coming from the plate of V_{105} . The grid of V_{106} is thereby fed with a pure sine wave of 348 kc. Bandwidth and coupling are determined by the values used for the two coupling capacitors, C_{118} and C_{119} . If amplitude variations of 116 kc are not removed from the 348-kc signal, the carrier will tend to become phase-modulated at the driving frequency of 116 kc and cause spurious frequencies to appear at the exciter output. The necessary bandwidth of circuitry between V_{105} and V_{106} is determined by the highest modulating frequency and the frequency deviation. For the circuits of V_{105} and V_{106} , this turns out to be 35 kc.

Similar reasoning is behind the selection of the tuning circuits in the remaining stages. In V_{106} through V_{112} , two coupled tuning circuits provide the necessary bandpass, because, as we go higher in frequency, the necessary bandspread is achieved more readily. That is, if the same Q is maintained, increase in operating frequency results in a proportionately greater bandpass.

It will be noted that a short coaxial cable jumper connects V_{109} and V_{110} . The combination of L_{110} , C_{137} , and C_{138} tunes the plate circuit of V_{109} to the proper resonant frequency. C_{137} and C_{138} are impedance-transforming capacitors that change the high impedance output of V_{109} to about 51 ohms at J_{101} . This makes it possible to carry the output of V_{109} over a considerable length of coaxial cable to another amplifier stage without serious attenuation. From this external amplifier (not shown), another length of coaxial cable may be brought back to J_{102} . The capacitor combination of C_{139} , C_{140} , and C_{141} then transfers the 51-ohm impedance back to a high impedance in addition to resonating L_{111} to the operating frequency.

J_{101} and J_{102} are used specifically for the purpose of inserting multiplex sub-carriers on the main carrier, should this be desired. If not, J_{101} connects directly to J_{102} .

V_{105} through V_{109} are single-ended stages and use 6AU6 pentode tubes. V_{110} through V_{112} are push-push stages and all use 6J6 twin triode tubes. The grids of these push-push stages are connected in normal push-pull fashion, but the plates are connected in parallel. With this type of connection, the plate circuit receives two pulses for every complete cycle of R.F. drive in the grid circuit. This makes it naturally suited to frequency doubling service since the circuit will not amplify a fundamental frequency and will

not triple. It will only double or quadruple. The quadruple frequency, however, will be out of the tuning range of the stage.

Before we leave the subject of F-M through phase modulation, brief mention should be made of a tube, the Phasitron, which can accomplish this directly. The tube, developed by Dr. Robert Adler of the Zenith Corporation in 1946, utilizes a series of 36 deflection grids, a 3-phase oscillator signal and an external modulating coil to produce an F-M signal through phase modulation. The grids, in conjunction with the 3-phase signal, develop the rotating electron disc shown in Fig. 16.27. The external modulating coil, through which the audio currents pass,

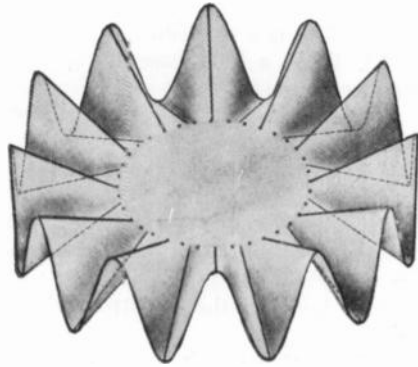


FIG. 16.27. The electron disc developed in the Phasitron tube.

causes the rotating disc to shift back and forth in accordance with the modulating signal. In essence, this is exactly the same action that occurs in the vector diagram used originally to explain phase modulation in Chapter 2.

A small number of F-M transmitters were built by General Electric using the Phasitron, but no extensive application ever developed.*

PROBLEMS

1. Name 4 methods of producing phase modulation.
2. Describe how the circuit of Fig. 16.2 produces phase modulation.
3. Draw a diagram of the G_m phase modulator and explain how it operates.
4. Describe the operation of the reactance-tube modulator. In what way does this differ from a reactance-tube modulator which produces F-M directly?
5. What is the difference between a linear and a nonlinear inductance? What causes the nonlinearity?
6. How does the circuit of Fig. 16.7A produce phase modulation?
7. What is the Armstrong system of generating frequency modulation? Why is it useful?
8. What factors govern the amount of F-M produced through phase modulation? How are these factors combined in a formula?
9. Draw a block diagram of the basic Armstrong system. Explain the function of each stage.

* Those readers desiring more information about this tube are referred to:

1. Robert Adler, "A New System of Frequency Modulation," *Proc. of the Institute of Radio Engineers*, January, 1947.
2. F. M. Bailey and H. P. Thomas, "Phasitron F-M Transmitter," *Electronics*, October, 1946.

10. What limitations must be observed when employing phase modulation to produce F-M? How does the Armstrong system observe these limitations?

11. What is a predistorter network? Draw a diagram of such a network and connect it to an amplifier.

12. Why is a predistorter network required in the Armstrong modulator?

13. When a full phase swing of 30° is utilized, how much equivalent F-M is produced when the audio signal frequency is 1000 cycles?

14. Differentiate between predistorter networks and accentuator circuits. Where is each used?

15. Why is predistorter circuit design based upon the lowest audio frequency used in the modulator?

16. How much frequency multiplication is required in the Armstrong F-M transmitter in order to obtain a fully modulated wave? How does this affect the carrier frequency and how is a final carrier value between 88-108 mc achieved?

17. How does the Armstrong Dual Channel Modulator differ from the basic circuit?

18. What stage determines the accuracy of the final carrier frequency in the Dual Channel Modulator Transmitter? Explain.

19. Explain by means of a block diagram the operation of the Dual Channel Transmitter.

20. What advantages does the Armstrong Dual Channel Transmitter possess over the basic Armstrong circuit?

21. Discuss the differences between A-M and F-M transmitters of comparable power. Limit the discussion to audio modulating power required, carrier frequency stability, types of tubes required and overall economy.

22. What does an F-M receiver contain to counteract the effect of the predistorter and accentuator networks in the transmitter? Explain your answer in detail.

23. Explain briefly the operation of the Serrasoid F-M modulator.

LOCATING COMMON TROUBLES

The purpose of the following check list is to point out those components or areas in a receiver where the defect for the various types of troubles indicated is most likely to be located.

No Sound

1. Defective tube.
2. A-c line voltage not reaching the receiver. Check fuse.
3. Selenium or semiconductor rectifiers defective.
4. Defective power supply component.
5. Try signal injection, starting with stage closest to loudspeaker.

A-M Reception, No F-M Reception

1. Defective tube. (Check those dealing with F-M only.)
2. Check A-M, F-M switch.
3. F-M oscillator defective.
4. Measure voltages on tubes dealing with F-M only.
5. Try signal injection in F-M section.

F-M Reception, No A-M Reception

1. Defective tube. (Check those dealing with A-M only.)
2. Loop antenna open.
3. Malfunctioning A-M oscillator.
4. Check A-M, F-M switch.
5. Measure voltages on tubes dealing with A-M only.
6. Try signal injection in A-M section.

Weak Sound—All Signals

1. Defective tube in power supply on audio amplifiers.
2. Defective volume control.

3. Low voltage in power supply, audio amplifiers, or I.F. stage.
4. Use signal injection in audio amplifier section to locate trouble.
5. Defective loudspeaker.

Weak Sound—A-M Only

1. Defective tube in A-M section.
2. Defect in A-M detector circuit.
3. Open loop antenna.
4. A-M oscillator not functioning strongly.
5. Check A-M, F-M switch.
6. Misalignment in A-M circuits.
7. Low voltage in A-M stage.

Weak Sound—F-M Only

1. Defective tube in F-M section.
2. Misalignment in F-M circuits.
3. Check F-M antenna.
4. F-M oscillator not functioning strongly.
5. Low voltage in F-M stage.
6. Check A-M, F-M switch.
7. Defective component in output network of F-M detector.

Hum—Always Present

1. Cathode-to-heater leakage in audio amplifier tube.
2. Defective filter in power supply.
3. Poor ground connection in audio amplifier circuits.
4. Leakage in multiple-section filter capacitor when one of the sections is in the audio system.

Hum—Present Only When Signal Is Being Received

1. Leakage across R.F. amplifier, oscillator, or mixer sockets.
2. Cathode-to-heater leakage in R.F. amplifier, oscillator, or mixer tube.
3. Poor ground connection in R.F. stages. (This includes R.F. amplifier, oscillator, and mixer.)

Hum—Phono Operation Only

1. Poor ground connection in phono assembly or in connecting cable.
2. Poor electrical connection at phono input jack.
3. Leakage from phono motor to cartridge head in pickup.

Fuse Blows

1. Defective tube, particularly in power supply.

2. Shorted filter capacitor in power supply.
3. Defective selenium or semiconductor rectifier in power supply.
4. Short circuit across B+ line (generally a shorted bypass capacitor).
5. Shorted winding in power transformer.
6. Shorted filament in parallel-wired filament circuits.
7. Try disconnecting all B+ lines, then reconnecting them one at a time.

Defective Operation During Certain Hours Only

1. Check power line voltage during these times.

Distorted Sound

1. Defective tube.
2. Low B+ voltage.
3. Misalignment of F-M detector.
4. Open filter capacitor.
5. Leaky audio coupling capacitor.

Signal Drift

1. Defective AFC circuit.
2. Defective oscillator and AFC tubes.
3. Open or shorted components in AFC line.
4. Change in component value in oscillator circuit.

Noisy Reception

1. Defective limiter.
2. Misalignment—all circuits.
3. Poor ground connections.
4. Defective bypass capacitor in audio amplifiers.
5. Dirt or grime on tuning capacitor plates.
6. Defective tube.
7. Loose connections.
8. Misalignment of F-M detector circuit.

Signals Only Over Portion of Dial

1. Misalignment of R.F. circuits.
2. Defective oscillator tube.
3. Tuning capacitor plates shorting.
4. Defective oscillator component.

Tuning Dial Markings Off

1. Misalignment of R.F. circuits.

Intermittent Operation

1. Defective tube.
2. Poor ground connection.
3. Fluctuating line voltage.
4. Defective selenium or semiconductor rectifiers.
5. Defective capacitor.
6. Defective resistor.
7. Loose connection.

Motorboating

1. Open filter capacitor.
2. Open bypass capacitor.
3. Poor lead dress.
4. Poor ground connection on shield.
5. Open grid resistor.

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