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*F-M Simplified*

BY THE SAME AUTHOR

*U H F RADIO Simplified*

*TELEVISION Simplified*

## PREFACE

It is the purpose of this book to present not only the facts concerning F-M, but also to indicate and help bridge the gap between amplitude-modulation techniques and the newer frequency-modulation characteristics. The chapters in Section 2 analyze the types of circuit found in F-M receivers, together with the reasons for the form of each stage. In Section 5, the actual commercial applications are then studied with particular attention to each different design. Since every practical book is concerned with servicing procedure, such a chapter is included also. The similarity between A-M and F-M receiver servicing techniques is pointed out and an explanation is given of those principles required to cope fully with F-M. Section 3 deals with F-M transmitters and Section 4 describes alignment procedure for F-M receivers.

The desire to produce a perfect product is the aim of every progressive engineer. In radio, it has long been realized that the present system of communication lacks much to be considered perfect. Consequently, a constant search has been maintained for methods of improvement. One of the most obvious frailties of the system of amplitude modulation has been its susceptibility to noise and interference, both man-made and natural. It is common experience to find that clear reception from local stations other than those having 5,000 watts or more is often made impossible by the simultaneous appearance of signals from transmitting towers many miles distant. If the frequency of these interfering signals is sufficiently close to that of the desired signal, little other than increasing the power of the desired station will permit clear, undisturbed reception. Obviously, however, this is not the most desirable approach to the problem, for the same solution can be applied equally well to other stations, including those which cause the interference. A better solution is *frequency modulation*. How this new system of transmitting intelligence proves to be superior to amplitude modulation is discussed in considerable detail in Section 1.

The design of any receiver is governed not only by the type of signal to be received but, also, the frequency at which the signal is sent. F-M, arriving on the broadcast scene long after A-M was, of necessity, relegated to the high frequencies. In many respects this is advantageous since it has been found that there are fewer natural disturbances to be found at the higher frequencies. On the other hand, operation at the higher frequencies dictated the development of improved and modified techniques at all stages of the transmitter and receiver design. It requires, also, that the men who are to service and maintain this equipment extend their knowledge beyond the amount required to handle conventional A-M sets.

A set of questions is included at the end of the text for those who wish to gauge their progress through the book and their understanding of the principles involved. Instructors will also find the questions useful for classroom exercises. With the exception of the mathematical problems, no question can be properly answered with a single word or sentence.

In the compilation of the data and material required for a book of this type, the author wishes to extend his appreciation to the Stromberg-Carlson Co., the Philco Corp., Allen B. Dumont, Inc., U. S. Television, *Communications Magazine*, R E L, Z. Benin of Zenith, G. Wallin and C. W. Dymond of Motorola, and General Electric Corp. for their aid. Finally, there are the author's colleagues who, through their many suggestions, helped make this a more complete book.

M.S.K.

*Chicago, Illinois*  
*June, 1947*

# CONTENTS

CHAPTER	PAGE
SECTION I THE FUNDAMENTALS OF F-M	
1 INTRODUCTION TO FREQUENCY MODULATION . . . . .	3
2 F-M FROM PHASE MODULATION . . . . .	25
3 F-M AND INTERFERENCE . . . . .	38
SECTION II F-M RECEIVER PRINCIPLES	
4 HIGH FREQUENCY OSCILLATORS . . . . .	61
5 I.F. AMPLIFIERS . . . . .	78
6 LIMITERS . . . . .	97
7 F-M DETECTORS . . . . .	122
8 AUDIO AMPLIFIERS AND HIGH FIDELITY . . . . .	157
SECTION III F-M TRANSMITTERS	
9 COMMERCIAL F-M TRANSMITTERS — PART 1 . . . . .	181
10 COMMERCIAL F-M TRANSMITTERS — PART 2 . . . . .	211
SECTION IV RECEIVER ALIGNMENT	
11 F-M RECEIVER ALIGNMENT PROCEDURE . . . . .	245
SECTION V COMMERCIAL RECEIVERS	
12 R.F. TUNERS FOR F-M RECEIVERS . . . . .	271
13 COMMERCIAL F-M RECEIVERS . . . . .	291
14 SERVICING F-M RECEIVERS . . . . .	321
QUESTIONS AND PROBLEMS . . . . .	343
INDEX . . . . .	355





SECTION I

*The Fundamentals of F-M*



# CHAPTER 1

## INTRODUCTION TO FREQUENCY MODULATION

**The Old and the New.** For as long as radio has been commercially available, there has been a constant search for methods and means of improving the system — especially in two respects — fidelity of reproduction and interference. True, other aspects have not been neglected, but the greatest effort has been directed toward these two items. Of the two, good fidelity is simpler to achieve, because fidelity is the product of the circuits and systems evolved by engineers; what man can devise, man can generally improve. Troublesome interference is, for the most part, due to sources external to the receiver and therefore harder to cope with. Also, a large percentage of the worst interference comes from natural sources, generally beyond the limited power of man to control. Prior to the advent of frequency modulation, the best apparent solution was the employment of brute-force method high-powered radio signals to override the disturbing interference completely. Through the use of frequency modulation, the signal must still possess greater power at the receiver input than the undesired voltage, but the amount of signal required completely to obscure the noise now is considerably reduced. With amplitude modulation, the signal-to-noise ratio must be of the order of 100 to 1 or better to obscure the noise; with frequency modulation, this ratio need be only 2 to 1.

Frequency modulation, because of the ability 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. Compare how much farther away an A-M broadcasting station could be and still ruin reception: with A-M a voltage only 1/100 as strong as the desired carrier will provide sufficient inter-

ference 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 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 we eliminate 99 per cent of the disturbance. Now only stations in the immediate vicinity are concerned, and F.C.C. regulations do not permit simultaneous use of the same frequency in the same service area.

It is quite popularly supposed that F-M is capable of providing greater fidelity than A-M. Indirectly, this is true, since F-M eliminates more of the interference and thereby provides 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 vs 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 back to their original form, in which they become intelligible at the loud-speaker.

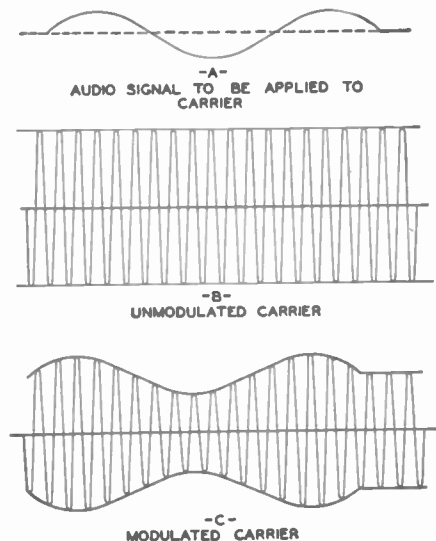


FIG. 1.1. The components of an amplitude-modulated wave.

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 dis-

tortion 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, undermodulation 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.

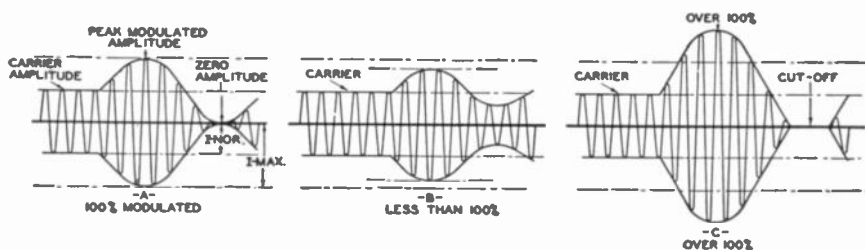


FIG. 1.2. Various degrees of amplitude modulation.

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, 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.

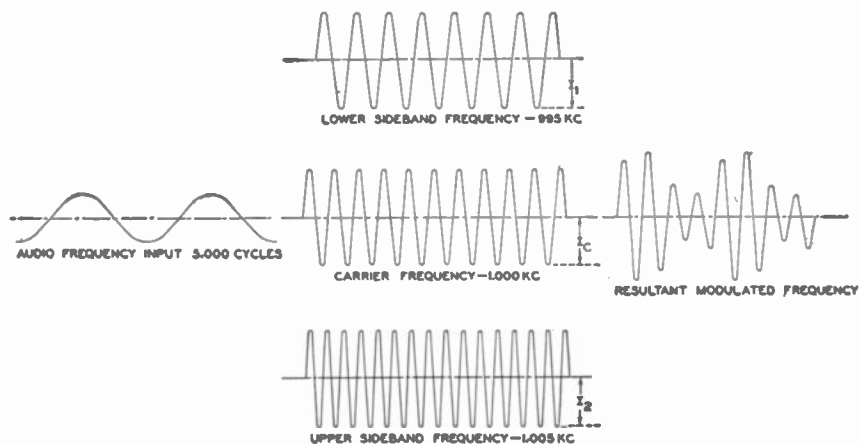


FIG. 1.3. The sidebands generated with amplitude modulation.

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  $\pm 5$  kc from the carrier are formed, making it necessary to receive a bandwidth of 10 kc ( $\pm 5$  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 is somewhat analogous to amplitude modulation, where the amplitude of the carrier is increased and decreased about the average, normal value.

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  $90^\circ$ , the frequency variation of the carrier from its mean or resting position also reaches a maximum. Between  $90^\circ$  and  $180^\circ$  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  $270^\circ$ , 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  $270^\circ$ , the carrier frequency is at its lowest point. From here it slowly follows the audio voltage back to normal. This occurs at  $360^\circ$ , 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,<sup>1</sup> 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

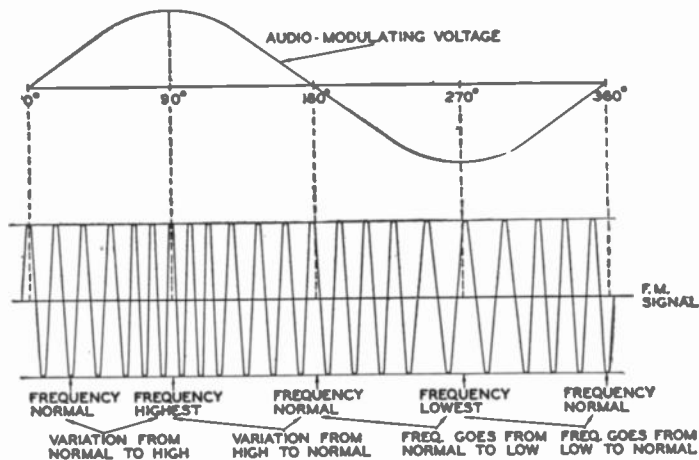


FIG. 1.4. Frequency variation in an F-M signal.

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 takes 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

<sup>1</sup> 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.

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 job is accomplished will be described in detail in Chap. 7. For the present of greater concern is what happens and this we have just learned.

The most frequently used name for the F-M detector is "discriminator," arising possibly as a result of its action in discriminating between the different frequencies that arrive at its input. Actually, if we wish to be consistent, the A-M detector is also a discriminator against the various amplitudes of the incoming signal. As used here and elsewhere, however, the word "discriminator" will apply only to F-M reception.

It is simple in F-M to deduce that 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 only shift the carrier frequency until it is brought to zero; anything less is obviously impossible since negative frequencies do not exist. Engineering limitations, of course, prevent us from introducing this large a shift, but even if it were possible, it would still be undesirable because of the limitation imposed on

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  $\pm 75$  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. With this limitation, the modulation does not produce distortion. Any degree of 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 ( $\pm 75$ ) kc is to be employed for the modulation and the remaining 50 ( $\pm 25$ ) 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. In other words, there are as many sideband frequencies above the carrier as below.

If we attempt to illustrate the sidebands with a diagram, we obtain the result shown in Fig. 1.5B. Fig. 1.5A represents the

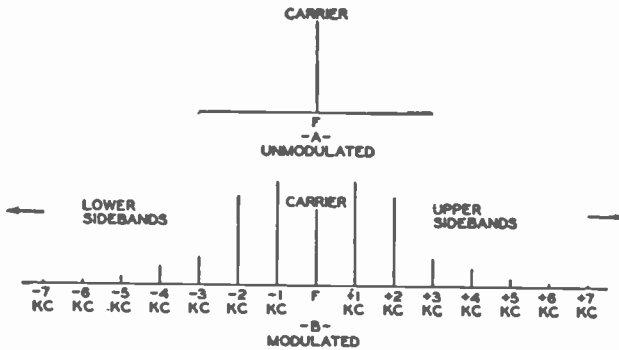


FIG. 1.5. An F-M signal with its sidebands developed as a result of audio modulation.

carrier before the modulation is applied. Fig. 1.5B 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. For, 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? Yet, we know from experience that this is entirely possible.

Upon examination of Fig. 1.5B, we can count seven sideband frequencies below the carrier and seven sideband frequencies above the carrier; all due, of course, 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 audio-modulating 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 exist an infinite number of these sidebands. Actually, both statements are reconcilable. For one condition represents the theoretical condition and the other the practical modification of this. 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 present are so weak beyond their band limits that, as 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 necessity 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 to actually 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 is a characteristic of frequency modulation.

The sum of the amplitudes of the sidebands plus 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, the first set of sidebands, at 90.001 mc, and 89.999 mc would both possess equal amplitudes but be of opposite polarity. The next or second set of sidebands would be at 90.002 mc and 89.998 mc, be equal in amplitude and have the same sign. Of course, although each set would have the same amplitude, different sets would, of course, differ from one another.

This is mentioned because the outright addition of the different sidebands of Fig. 1.5B, together with 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 the discussion extends, it makes no essential difference. We are only interested in the frequencies generated by modulation and 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 modulating 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, given up by others. The *total* energy however, under all conditions, remains constant.

Throughout each instant of the audio-modulating cycle, the nature of the F-M sideband distribution alters, always in response to changing instantaneous values occurring throughout the cycle. The picture, from moment to moment, might follow this form: at

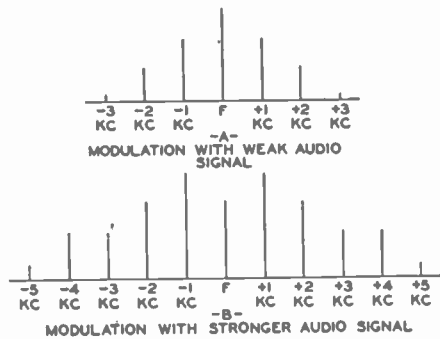


FIG. 1.6. A comparison of the frequency spread of two F-M signals modulated by audio voltage of differing intensity.

point *A*, Fig. 1.7, the modulating voltage is zero, and only the carrier frequency is being transmitted by the equipment. Gradually, as we move from point *A* to point *B*, the amplitude of the audio voltage increases and sidebands appear, absorbing power from the carrier. At the outset, say for points near *A*, only a few sidebands close to the carrier are important. But as we approach point *B*,

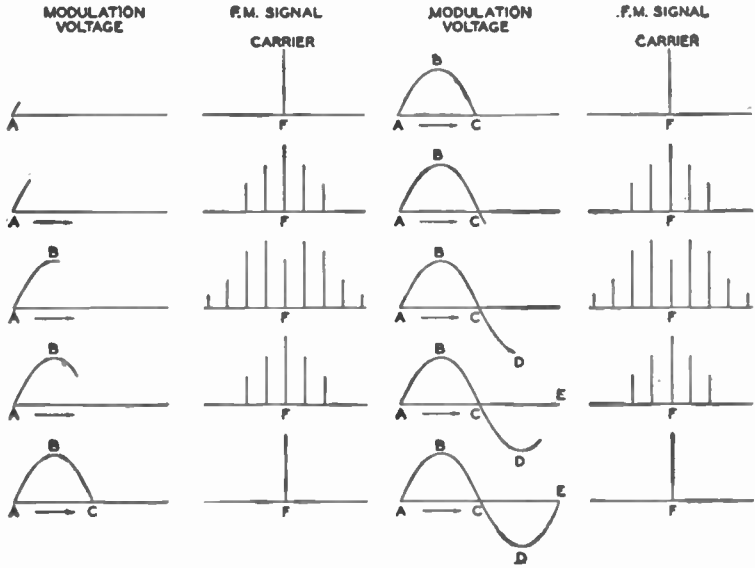


FIG. 1.7. A simplified illustration of the expansion and contraction of the sideband energy in an F-M during modulation.

the number of significant sidebands steadily increases, with the greater percentage of the total power shifting outward, towards the ends of the band.

Point *B* represents the maximum positive peak of the cycle and from here to point *C*, the contraction of the sidebands begins. The power starts shifting its weight back to the carrier and the distant sidebands, deprived of this power, disappear. At point *C*, we reach the zero line and for the moment nothing remains but the carrier. During the negative half of the audio cycle, we experience the same process, with the exception that the polarities of the various sidebands are now reversed. Those that were positive



previously become negative and those that were negative assume positive values. The expansion and contraction of energy, however, is identical. The end of the cycle is reached at point *E*, and we are ready to go through another cycle or as many more as the transmission requires. The length of time that each cycle lasts is determined, of course, by the frequency of the audio signal. A wave with the same frequency but of greater amplitude would merely extend the bandwidth on its peaks, but otherwise follow the variations outlined above.

The example of a single sine-wave modulating signal can be expanded to include voice and music waves actually encountered. However, a discussion of voice and music waves can become very complicated because of the interactions between the numerous frequencies composing each complex wave which tend to form sidebands that would not exist if each component of a complex wave were treated separately. Therefore, for our immediate purpose, all we need consider is the preceding explanation which provides us with a sufficiently accurate picture.

**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  $10 \times 100$  or 1000 cycles (1 kc) and a *total* bandwidth of 2 kc. On the other hand, 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:

$$\frac{\text{deviation of F-M carrier}}{\text{audio frequency causing this deviation}}$$

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. On the other hand, 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 due to the fact 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.

**Bessel Function Charts.** 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. In addition, we can also determine the amplitude of each sideband. Thus, using the modulation index for the F-M signal and the proper tables, we can tabulate all the information we desire regarding the specific frequency spectrum formed.

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

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 two sidebands are formed, similar to A-M operation when one modulating frequency is employed.

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

From the same Bessel tables we can obtain the amplitude of each of the significant sidebands for any modulation index. Table II contains a limited number of modulation indices and their corresponding sideband amplitude factors. Consider the modulation

index of 5. The relative values of each of the 8 (upper and lower) sidebands are given as follows:

<i>Component</i>	<i>Relative Amplitude</i>
carrier .....	17.76%
1st set .....	32.76%
2nd set .....	4.66%
3rd set .....	36.48%
4th set .....	39.12%
5th set .....	26.11%
6th set .....	13.10%
7th set .....	5.34%
8th set .....	1.84%

By the word "set" we mean the corresponding upper and lower sidebands, there always being one sideband just as far above the carrier as below. As noted previously, these would have the same amplitude, but not necessarily the same polarity.

The values given in Table II are relative. To find the absolute or actual value of each sideband, we must first know the value of the carrier before modulation. Suppose this is 90 amperes. Then multiply each component given above by 90 and divide by 100 to eliminate the percentage. Thus, the following values are derived:

carrier .....	15.98 amps.	5th set .....	23.50 amps.
1st set .....	29.48 amps.	6th set .....	11.79 amps.
2nd set .....	4.19 amps.	7th set .....	4.81 amps.
3rd set .....	32.83 amps.	8th set .....	1.66 amps.
4th set .....	35.21 amps.		

With a modulation index of 5, the carrier is reduced from its initial value of 90 amperes to 15.98 amperes during modulation. The rest of its energy has been transferred to the sidebands.

When all of these values are added directly, we do not obtain a resultant of 90 amperes. This is due to the fact that we are dealing with different frequencies and the addition has to be made geometrically. The correct method requires that we take the

TABLE I

Modulation Index	Number of Significant Side-Bands		Bandwidth Required ( $f$ = frequency of audio signal)
	Above Carrier	Below Carrier	
0.01	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	$40f$

square root of the sum of the squares of these values, or

$$I = \sqrt{(15.98)^2 + 2(29.48)^2 + 2(4.19)^2 + 2(32.83)^2 + 2(35.21)^2 + 2(23.50)^2 + 2(11.79)^2 + 2(4.81)^2 + 2(1.66)^2}$$

$I = 90$  amperes.

With each different modulation index, the entire complexion of the F-M spectrum changes.

It is always completely puzzling to anyone approaching F-M

TABLE II

Mod- ula- tion Index	Carrier Value	1st Set of Side- bands	2nd Set	3rd Set	4th Set	5th Set	6th Set	7th Set	8th Set	9th Set	10th Set	11th Set	12th Set	13th Set	14th Set	15th Set
0	1.000	....	....	....	....	....	....	....	....	....	....	....	....	....	....	....
0.01	1.000	.005	....	....	....	....	....	....	....	....	....	....	....	....	....	....
0.05	.9994	.025	....	....	....	....	....	....	....	....	....	....	....	....	....	....
0.1	.9975	.0499	....	....	....	....	....	....	....	....	....	....	....	....	....	....
0.2	.9900	.0995	....	....	....	....	....	....	....	....	....	....	....	....	....	....
1.0	.7652	.4401	.1149	.0020	....	....	....	....	....	....	....	....	....	....	....	....
2.0	.2239	.5767	.3528	.1289	.0341	....	....	....	....	....	....	....	....	....	....	....
3.0	-.2601	.3391	.4861	.3091	.1320	.0430	.0114	....	....	....	....	....	....	....	....	....
4.0	-.3971	-.0661	.3641	.4302	.2811	.1321	.0491	.0152	....	....	....	....	....	....	....	....
5.0	-.1776	-.3276	.0466	.3648	.3912	.2611	.1310	.0534	.0184	....	....	....	....	....	....	....
6.0	0.1506	-.2767	-.2429	.1148	.3576	.3621	.2458	.1296	.0565	.0212	....	....	....	....	....	....
7.0	.3001	-.0047	-.3014	-.1676	.1578	.3479	.3392	.2336	.1280	.0589	.0235	....	....	....	....	....
8.0	.1717	.2356	-.1130	-.2911	-.1054	.1858	.3376	.3206	.2235	.1263	.0608	.0256	.0100	....	....	....
9.0	-.0903	.2453	.1448	-.1809	-.2655	-.0551	.2043	.3275	.3051	.2149	.1247	.0622	.0274	.0108	....	....
10.0	-.2459	.0435	.2546	.0584	-.2196	-.2341	-.0145	.2167	.3179	.2919	.2075	.1231	.0634	.0290	.0120	....

NOTE: Where blank spaces are indicated the values of the sidebands are insignificant.

for the first time to learn that, although the carrier frequency in the F-M transmitter is never actually shifted beyond the  $\pm 75$  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  $\pm 75$  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 ( $\pm 120$  kc) when the carrier frequency is shifted only  $\pm 75$  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.

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  $\pm 75$ -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  $\pm 25$ -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  $\pm 75$  kc. This does not mean, for example, that the end significant sideband must not extend beyond  $\pm 75$  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  $\pm 75$  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. 1.8, with one carrier and two significant sidebands.

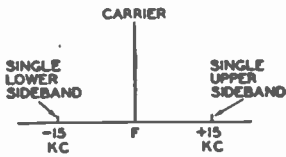


FIG. 1.8. The sideband distribution when the modulation index is small, 0.2 or less.

Note again that the sidebands will be spread farther from the center than the frequency shift that produced them. The sidebands are  $\pm 15$  kc from the center, whereas the carrier was shifted only  $\pm 3$  kc. This situation will be generally true of all F-M transmissions. However, if the reader remembers the water analogy, it may help 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. Again, Major Armstrong uses phase shifting to produce F-M in his type of F-M transmitter. To appreciate either of these two aspects, a brief discussion of phase modulation is necessary.



## 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  $45^\circ$ . The phase difference merely indicates that when the voltage of one wave reaches its peak, at any one instant, the other carrier is  $45^\circ$  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  $45^\circ$ .

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  $45^\circ$ . 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  $45^\circ$  lag to  $60^\circ$  lag (see Fig. 2.2). The  $60^\circ$  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  $45^\circ$  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 towards a value less than

$45^\circ$ . This is opposite to its increase in 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  $30^\circ$ . Note that this is

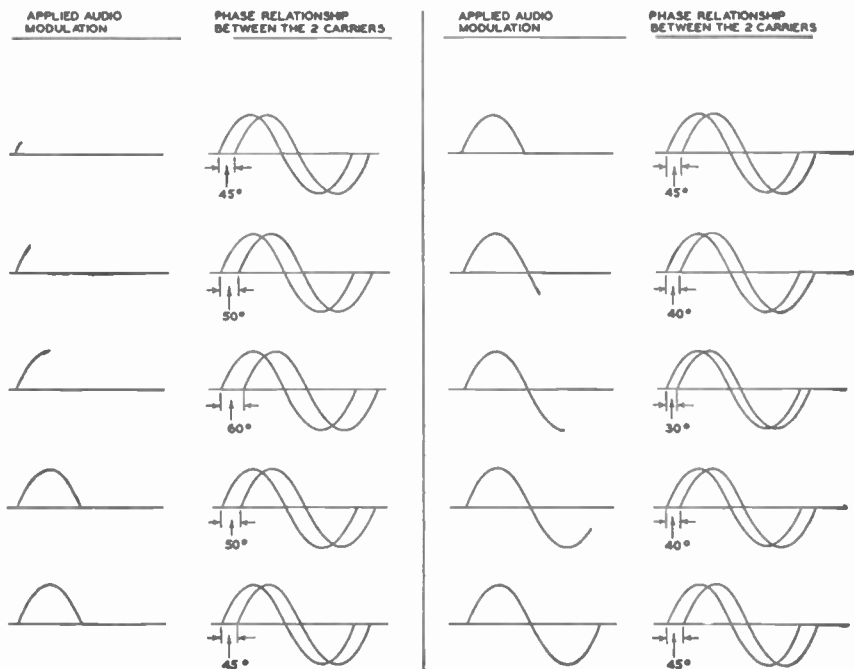


FIG. 2.2. A simplified illustration of phase modulation.

just as far below the normal value of  $45^\circ$  as  $60^\circ$  is above. From the maximum negative peak back to zero the phase returns, in step, to  $45^\circ$ .

This 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. On the other hand, 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 only shift the phase to  $50^\circ$  at its positive peak and  $40^\circ$  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 differences, would note the different shifts and respond accordingly. This 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 does not clearly indicate how frequency modulation is formed as a result. For this, a vector approach can be better employed.

**Elementary Vectors.** A vector is a shorthand method of illustrating certain facts much like 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. This is shown in Fig. 2.3A.

A simpler method, however, is to have the length of the arrow indicate the 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 1 inch long for 100 pounds of force, then an arrow 2 inches long would represent a force of 200 pounds (see Fig. 2.3B).

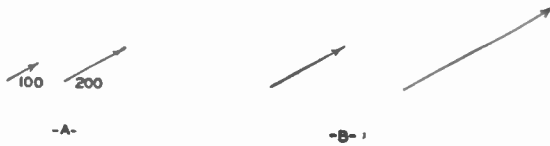


FIG. 2.3. Vectors represent definite forces and the directions in which they act.

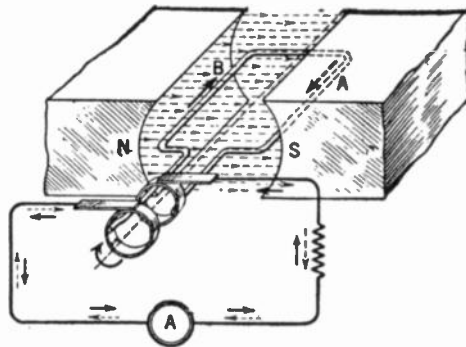


FIG. 2.4. The basic arrangement for the generation of an alternating voltage.

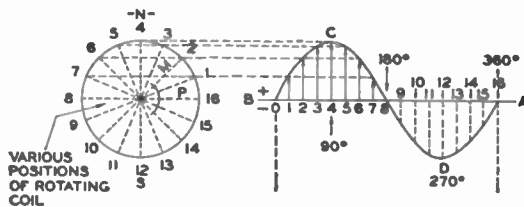


FIG. 2.5. The variation in A-C voltage with coil position.

**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 the position

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

$$e = E_{\max.} \sin \theta,$$

where

$e$  = instantaneous value of the voltage developed when the coil is at some angle  $\theta$ ,

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

$\theta$  = angular position of the coil.

Since the position of the coil determines the amount of voltage generated, 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  $\theta$ . When the vector is placed horizontally, the angle  $\theta$  is zero. This makes  $\sin \theta$  equal to zero (see Fig. 2.7A). From the formula,  $e = 0$  when  $\theta = 0$  and no voltage is generated. At some other angle, say  $30^\circ$ , the voltage generated is

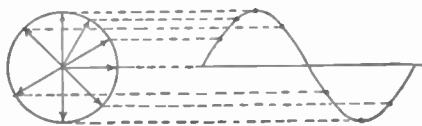


FIG. 2.6. Substitution of vectors in place of the rotating coil. The arrowed vectors indicate the coil position at various instants.

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

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  $30^\circ$  with the horizontal (Fig. 2.7B).

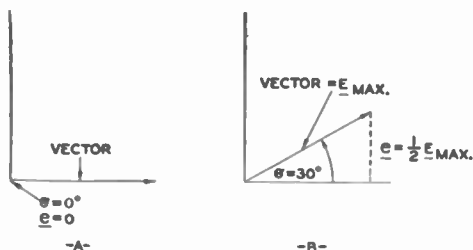


FIG. 2.7. The value of the generated voltage depends upon sine of the angle,  $\theta$ .

A vertical line dropped 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 cycles/sec, then the vector will rotate 10,000 times in one second.

FIG. 2.8. A comparison between the vector and the conventional methods of illustrating a  $45^\circ$  phase difference between two waves.

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  $45^\circ$  represent two electrical waves  $45^\circ$  apart in phase. Here again we note the simpler notation made possible by the use of vectors.

Many times, when two waves are present in the same circuit, we wish to ascertain the total effect and not merely the individual

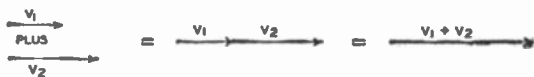


FIG. 2.9. The addition of two vectors that are in phase, i.e., point in the same direction.

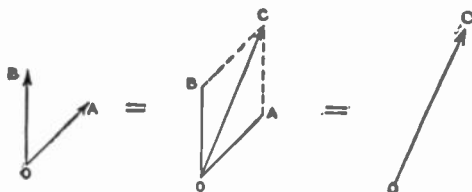


FIG. 2.10. The addition of two out-of-phase vectors.

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 addition 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 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.

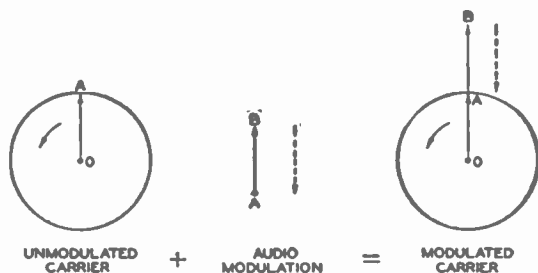


FIG. 2.11. Amplitude modulation demonstrated by vectors.

2.11, the carrier 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  $180^\circ$ . 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.

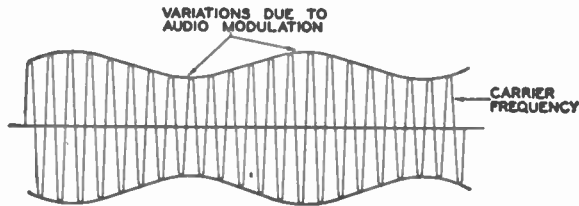


FIG. 2.12. The customary method of indicating an amplitude-modulated wave.

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 as shown in 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  $45^\circ$ . 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 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,

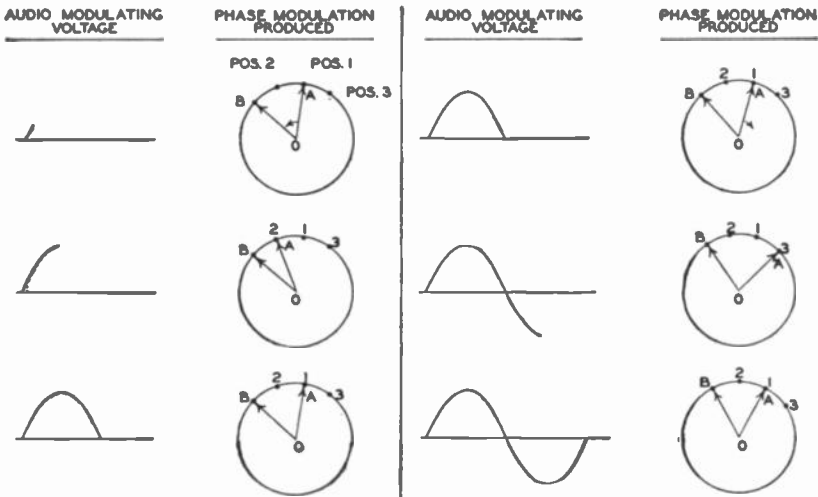


FIG. 2.13. The phase variation of a carrier that is being phase-modulated by an audio signal.

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 to a 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 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 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.

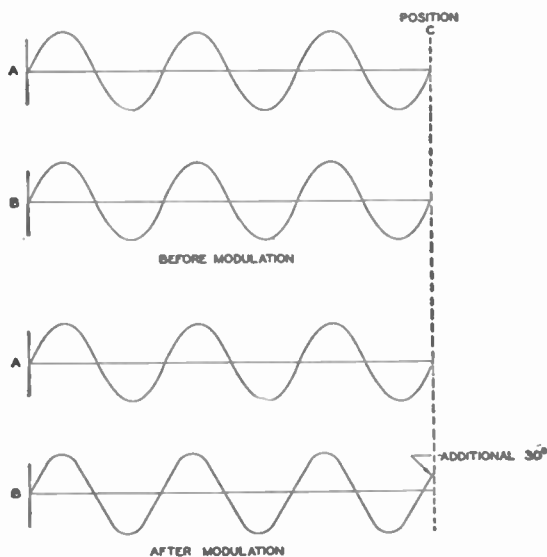


FIG. 2.14. Prior to modulation, carriers *A* and *B* are in phase. After modulation, *B* leads *A* by  $30^\circ$ .

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  $30^\circ$ . 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  $30^\circ$ . But for each wave to assume an additional amount means that, besides the regular  $360^\circ$ , 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  $30^\circ$ . Henceforth, with no further variation, wave  $B$  will reach its maximum positions (and minimum positions)  $30^\circ$  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.

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 cycles/sec) is to be shifted ahead in phase to a maximum of  $50^\circ$ , 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  $50^\circ$  shift in this interval, each carrier cycles must contribute an additional  $50/90,000$  degree, or approximately  $0.0005^\circ$ . To achieve this, the frequency of each wave must be increased slightly. This is the indirect F-M. The same line of reasoning would apply if the wave were to drop  $50^\circ$  behind, except that now each wave loses  $0.0005^\circ$  from its cycle. The result — a decrease in frequency.

If we maintain the phase shift at  $50^\circ$ , 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  $50^\circ$  shift, we calculate that  $50/9000$  or approximately  $0.005^\circ$  are

needed from each cycle compared to the  $0.0005^\circ$  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 “ $\sin \omega t$ ” 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.

## 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, then 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 appears. 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. Distant stations, ordinarily unheard during the day, chime in after sundown to add their transmissions to the existing confusion. In A-M, it requires only a noise-to-signal ratio of 1 to 100 to 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. 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

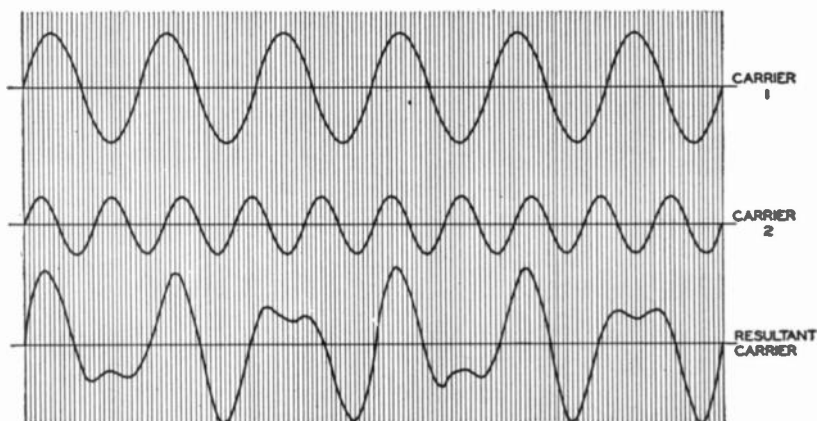


FIG. 3.1. The combination of two carriers to form a resultant which is amplitude and phase-modulated.

is 2 to 1, the maximum angle of phase shift produced is slightly under  $30^\circ$ . 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 to depend upon  $f\Delta\theta$ , where  $f$  is given in cycles and  $\Delta\theta$  is expressed in radians. A radian is equal to  $57.3^\circ$ , which means that there are  $2\pi$ , or  $2 \times 3.1416$ , radians in  $360^\circ$ . 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  $30^\circ$  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  $1000 \times 0.5$  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 voltages 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 loudspeaker. 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

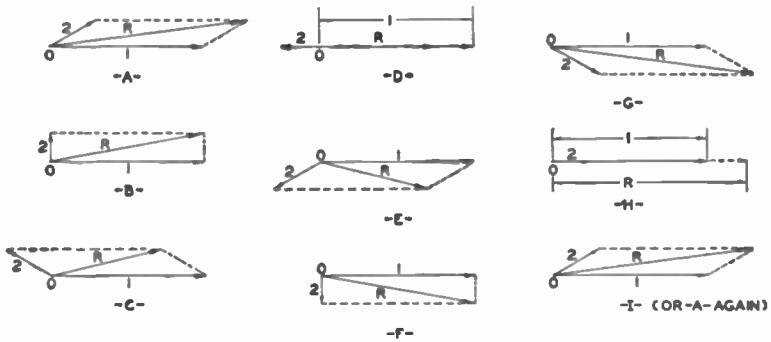


FIG. 3.2. The resultant wave ( $R$ ) produced by the interaction of a strong signal (1) and a weak interfering signal (2).

may have advanced from 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

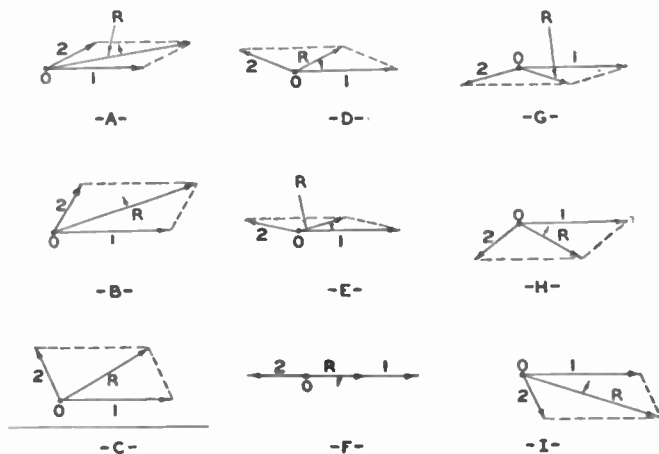


FIG. 3.3. The amplitude and phase variation of a resultant (*R*) carrier due to the interaction of two signals. The small arrows on *R* indicate whether its phase (with respect to the desired signal, 1) is going in a positive or negative direction.

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.

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 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 loud-speaker. When one of the carriers arriving at

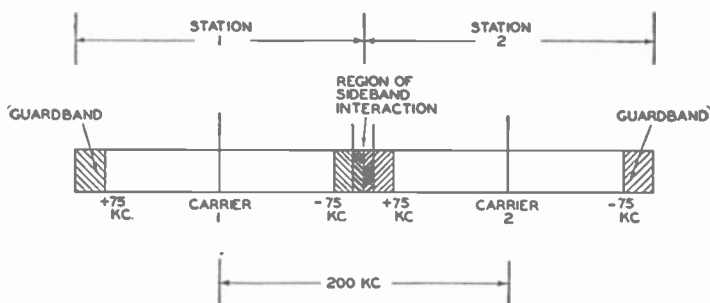


FIG. 3.4. In the darkened region between station 1 and station 2 it is possible for their sidebands to interact and produce interference.

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 guard bands. This can occur on strong modulation. The region that is sufficiently close to an adjacent guard band 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). 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

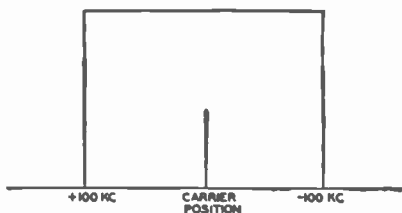


FIG. 3.5. Perfect response characteristic for receiver tuning circuits.

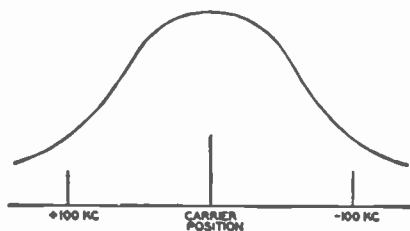


FIG. 3.6. Practical modification of Fig. 3.5.

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

$T$  = the temperature of the conductor (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

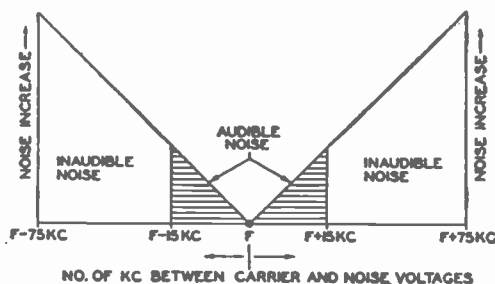


FIG. 3.7. Noise response of an F-M receiver. The noise intensity increases directly with frequency separation between carrier and interfering signal.

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 generated, 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.

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.

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

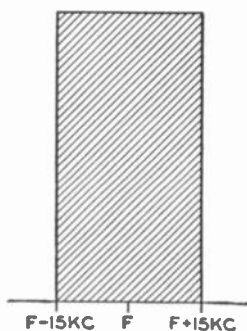


FIG. 3.8. The noise characteristic of an A-M receiver.

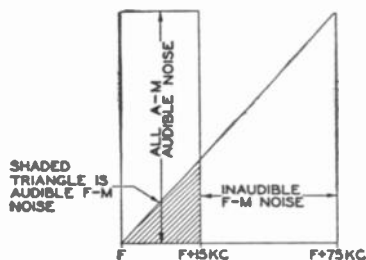


FIG. 3.9. A comparison between A-M and F-M noise characteristics. Note that all the A-M noise is audible, whereas only a small segment of the total possible F-M noise is effective.

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 in the last chapter, 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 fre-

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

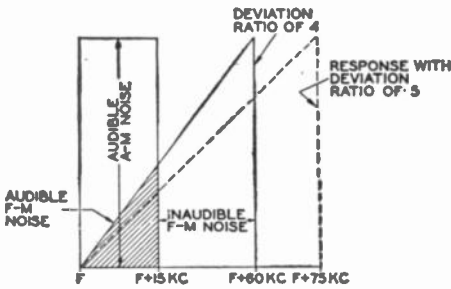


FIG. 3.10. The increase in noise in an F-M receiver with decrease in deviation ratio.

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.

If we continue to limit the swing of the F-M carrier, the signal-to-noise improvement (over A-M) diminishes correspondingly.

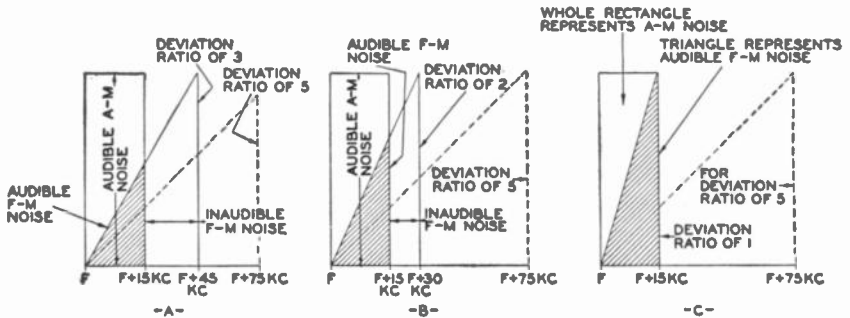


FIG. 3.11. Further comparisons between the noise in A-M and F-M systems with various F-M deviation ratios.

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 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.

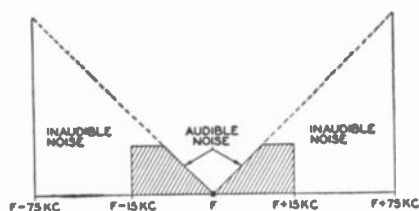


FIG. 3.12. Improvement in noise reduction due to pre-emphasis circuit in transmitter.

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. 3.12. The de-emphasis action, by reducing the level of all frequencies above 1500 cycles, slices off a considerable portion of the noise. The actual arrangement of the components of the de-emphasis circuit is shown in Chapter 7.

### 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. 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 loud-speaker. 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 6.

### 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 loud-speaker. 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.



*SECTION II*

*F-M Receiver Principles*



## CHAPTER 4

### 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 "frequent converter" and "frequency mixer" are used interchangeably by many writers, although a distinction does

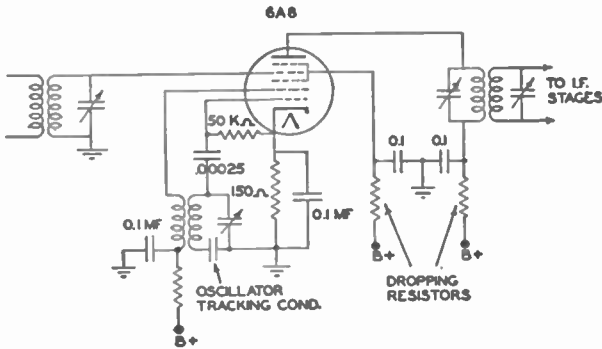


FIG. 4.1. A typical pentagrid converter circuit.

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. 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. 4.1 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. 4.2. A separate tube (usually a triode) is employed as the oscillator.

This increases the space and cost requirements of the set, but the stability is superior to the frequency converter circuit.<sup>1</sup>

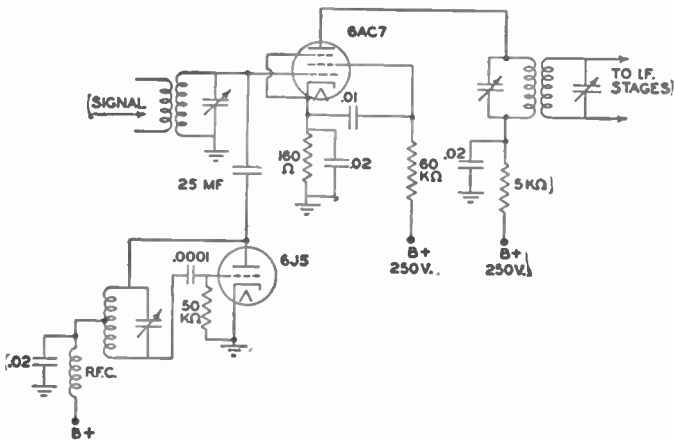


FIG. 4.2. A mixer (6AC7) with a separate oscillator (6J5).

**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 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 opera-

<sup>1</sup> The above statement does not exclude the possibility of using a converter for F-M receivers in their present band. In fact, two or three manufacturers do incorporate such circuits using specially designed tubes. However, most sets have separate oscillators for the reasons given.

tion, 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.

The desired stability in an oscillator has been found to be  $\pm 0.01$  per cent of the carrier or  $\pm 10$  kc at 100 mc. With ordinary precautions, such as the use of low-loss sockets, well-placed components

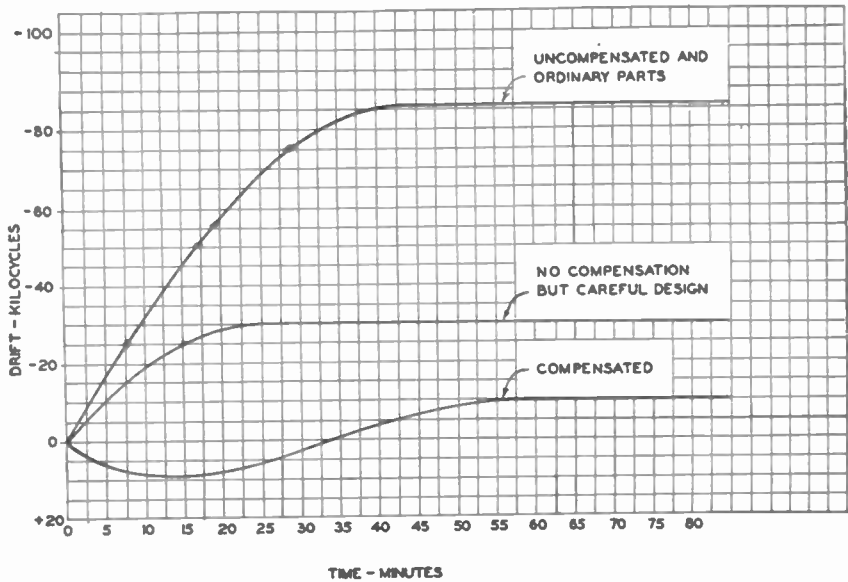


FIG. 4.3. The effect of careful design and compensation on oscillator frequency drift.

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 condenser. 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. 4.3. 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 condensers, the increase in capacitance is due to the linear expansion of the condenser plates. The capacitance of a condenser 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 mica condenser having a negative temperature coefficient. Available in many sizes, the proper condenser 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, 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 used 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. 4.4 (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 pre-

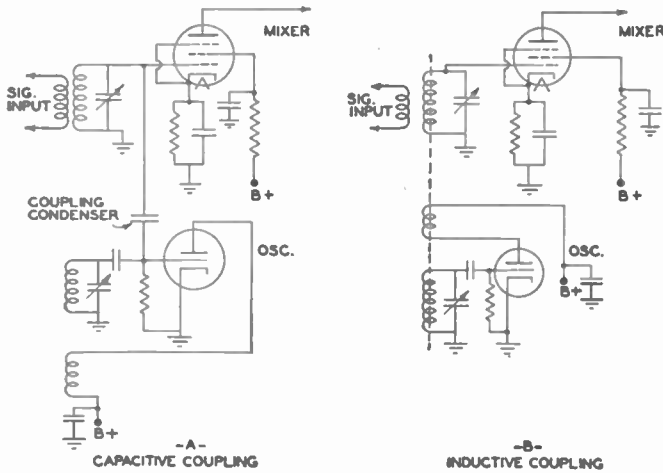


FIG. 4.4. Two widely used methods of coupling between the oscillator and mixer.

sented to the oscillator voltage will be low. Hence, considerably 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.

**Need for Separate Oscillator.** The inability of conventional pentagrid converters to function satisfactorily at higher frequencies 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 is shown in Fig. 4.5. 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  $G_1$ . As part

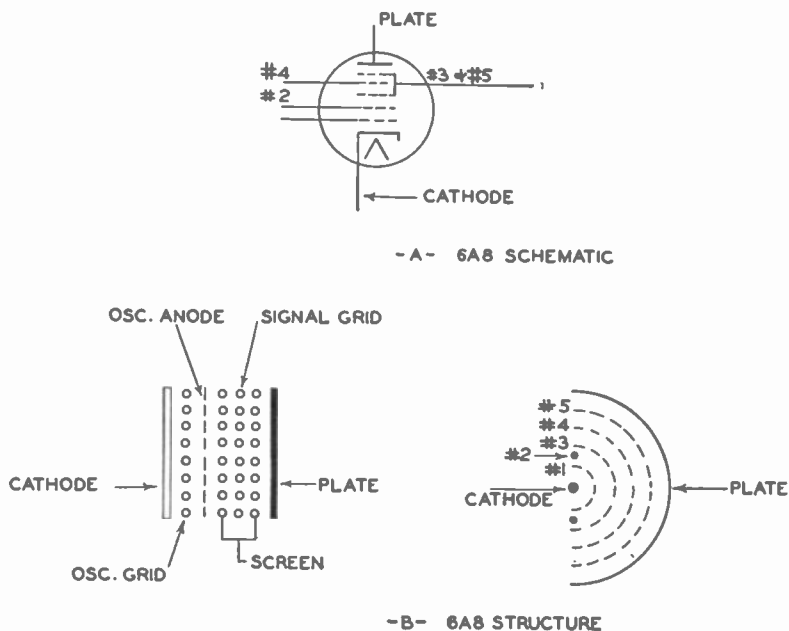


FIG. 4.5. The conventional schematic (A) and actual electrode placement (B) of a 6A8 converter tube.

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  $G_3$  come to a halt in the region between  $G_3$  and  $G_4$  because of the negative potential on

$G_4$ . 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.

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,  $G_1$ . Due to 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.

There is still one additional disadvantage associated with a pentagrid converter — 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 a 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 and the incoming signal voltage, since both voltages are needed to produce the mixing action. The greater the effect that an input signal has in producing an output I.F. voltage, the higher  $S_c$  becomes. The oscillator voltage is usually of little concern since it can be made as strong as desired. It is the input signal that is of primary concern, for the variations in this signal contain the intelligence. 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 above formula) and the overall gain, using the pentode, is higher.

The oscillator voltage may be injected at the control grid, cathode, screen or suppressor grids of a pentode tube, or at the control grid and cathode if a triode is used as the mixer. The most convenient place is the cathode, since a control over the grid and plate circuits may thereby be effected simultaneously. Injection at the control grid 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 capacitively. Both methods are

shown in Fig. 4.4. For cathode coupling, only an inductive transfer of energy proves satisfactory (see Fig. 4.6). 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.

Possibly the most important reason for not extensively using a triode for a mixer, despite its excellent signal-to-noise ratio and high conversion transconductance, is its relatively large grid-to-plate capacity. 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 condenser than the coil. When this energy feeds back through  $C_{gp}$ , the  $Q$  of the input tuned-grid circuit is decreased. In a pentode this effect 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

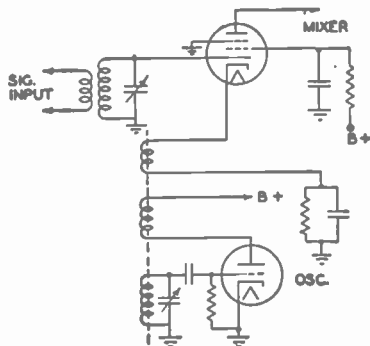


FIG. 4.6. Coupling oscillator voltage to the cathode of the mixer.

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 in the majority of sets. These are the familiar Hartley, the tuned-grid tickler oscillator, and the electron-coupled Hartley, which is a modification of the regular Hartley. They are shown in Fig. 4.7. Each unit, if properly constructed, oscillates quickly and possesses good stability. Triode tubes are usually preferred because they require but one regulated voltage; pentodes

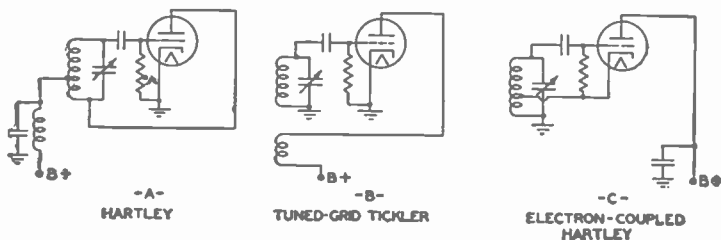


FIG. 4.7. Three popular oscillator circuits.

need two. A popular triode is the 6J5, although, of late, other high frequency triodes have been made available. These have:

1. Low internal shunting capacities.
2. Special construction enabling the tube to function readily at higher frequencies.
3. A high value of mutual conductance to provide a strong output voltage.

**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 emphasized when push-button tuning is utilized, which is true of many F-M and television receivers. Manual F-M tuning, at best, requires a sensitive "magic-eye" tube to aid the listener center the station correctly. It is especially important in F-M receivers that the station be correctly tuned in, else many of the benefits possible with F-M are lost. Obviously, preset circuits operated by pressing a push-button would greatly simplify receiver tuning and assure the listener of a properly centered signal each time. Preset circuits have long been regarded as representing the best solution to F-M

tuning problems if some method could be evolved which would prevent or minimize oscillator drift. It has been the common (and sad) experience of numerous owners of push-button receivers that, after a short period of operation, resort had to be made to the manual control either for complete tuning or else as a vernier adjustment for the push-button itself.

The one way out of this difficulty is by the use of quartz crystal oscillators. Crystals, on a production basis, can be obtained with a stability of  $\pm 0.005$  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  $\pm 0.01$  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.

One limitation of crystals is the restriction of reception to only those stations for which crystals are provided. In other words, each separate station requires another crystal in the oscillator in order to produce the proper intermediate frequency. If there are ten stations in a community, ten separate crystals are necessary. These could be bought with the receiver and installed (a very simple operation) by the radio dealer. Difficulties would arise only if a new station was erected in the vicinity, requiring the installation of an additional crystal, or if the owner moved to another city where stations utilized different frequencies. These are the only two instances, however, where the use of crystals imposes added difficulties. Since most families remain within one community for long periods of time, the limitation is not too serious. In any event, each of the foregoing limitations can be corrected for a nominal charge by any radio serviceman.

**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

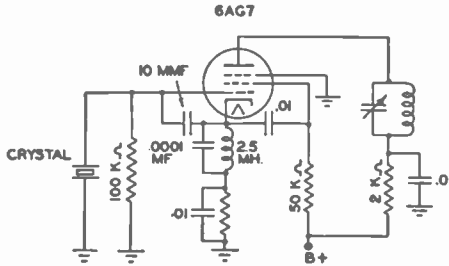


FIG. 4.8. A grid-plate crystal oscillator.

plate tank condenser. A suitable circuit is the grid-plate crystal oscillator shown in Fig. 4.8. The crystal is connected between the grid and ground. In the cathode tuned circuit, the ratio frequency choke and the fixed condenser, 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 below the signal input by 10.7 mc, which is the recommended intermediate frequency. The range, then, that the mixing oscillator must cover is from 77.3–97.3 mc. With a fundamental crystal frequency between 6 and 12 mc it becomes necessary to obtain strong outputs from the 8th to the 12th harmonic. For the 12-mc crystal, we need only go as high as the 8th harmonic; for the 6-mc crystal, the 12th harmonic is needed. 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. As a compromise, most manufacturers use a crystal whose frequency may be used for the fundamental and the harmonic needed as given in Table III.

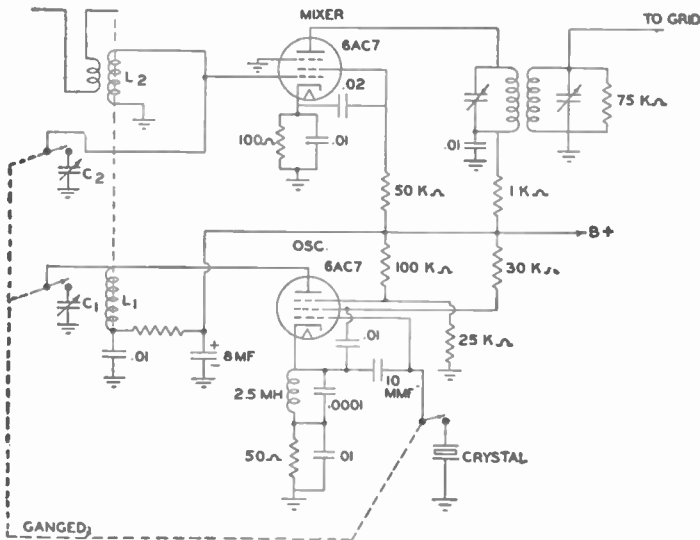


FIG. 4.9. The input stages of a crystal-controlled F-M receiver. The suppressor grid of the oscillator is made slightly positive to produce a strong output.

The input portion of an F-M receiver, using a crystal oscillator, is shown in Fig. 4.9. For each station, a separate crystal and two trimmer condensers are brought in by a push-button or a rotary switch. The trimmer condenser  $C_1$  across  $L_1$  is peaked by the manufacturer 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 are merely inserted to produce a strong output at the harmonic mixing frequency. The other trimmer condenser,  $C_2$ , tunes the antenna input coil  $L_2$  to the signal frequency. The addition of an R.F. stage adds one more trimmer

condenser that must be changed each time another station is received.

TABLE III. The range of fundamental crystal frequencies that may be used to produce the 77.3 mc to 97.3 mc mixing frequencies needed.

Range of Crystal Frequencies	Harmonic Used	Range of Harmonic Frequencies
9.6625 mc to 12.1625 mc	8th	77.3 mc to 97.3 mc
8.58888 mc to 10.8111 mc	9th	77.3 mc to 97.3 mc
7.7300 mc to 9.7300 mc	10th	77.3 mc to 97.3 mc
7.02727 mc to 8.8455 mc	11th	77.3 mc to 97.3 mc
6.44166 mc to 8.1083 mc	12th	77.3 mc to 97.3 mc

**Choice of Crystals.** In choosing a crystal for the oscillator, special attention must be given to its frequency change with tem-

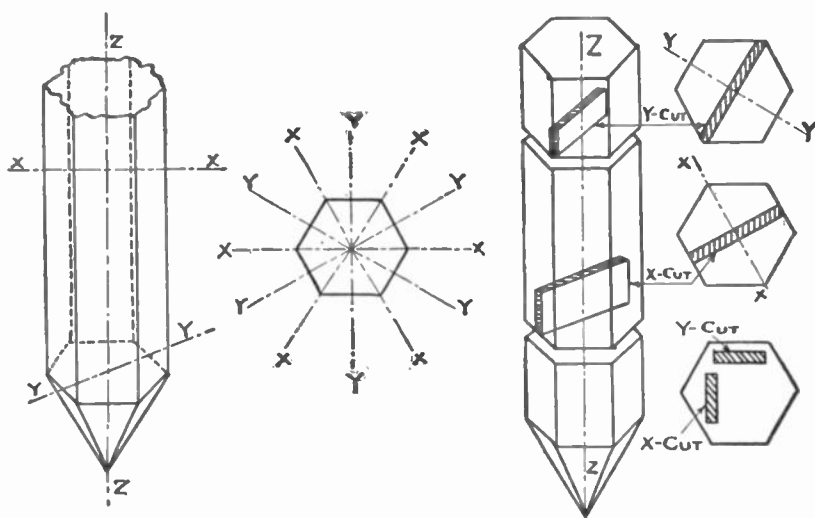


FIG. 4.10. The three axes of a crystal. Cuts other than those shown also possible.

perature. This characteristic of the crystal depends upon the manner in which it is cut from the original crystal piece. A crystal has three major axes, designated by the letters  $X$ ,  $Y$ , and  $Z$ . The  $X$  axis is known as the electric axis, the  $Y$  axis is the mechanical axis and the  $Z$  axis is the optic axis (see Fig. 4.10). Crystal slabs cut

from each of these axes have different temperature coefficients. For example, X-cut crystals have negative coefficients between 15 to 25 cycles/mc/° C. This means that for each degree change (Centigrade side) the resonant frequency of the crystal decreases 15 to 25 cycles/mc. If the crystal has a fundamental frequency of 5 mc, and the temperature changes 1° C, then the resonant frequency will decrease to 4.9999 mc. We arrive at this figure by assuming that the crystal will decrease 20 cycles for each megacycle or a total of 100 cycles. Y-cut crystals generally run from -20 cycles/mc/° C to 100 cycles/mc/° C.

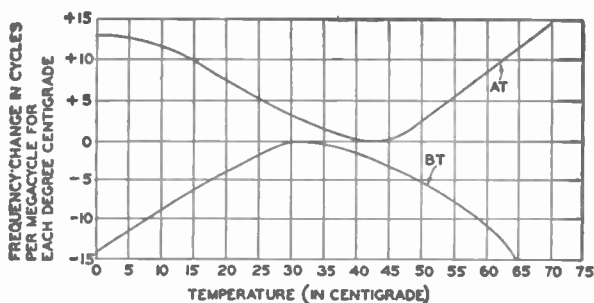


FIG. 4.11. Variations in frequency for AT- and BT-cut crystals.

In addition to these basic cuts, it is possible to obtain cuts possessing properties intermediate between two of these fundamental units. This makes it possible to offset a positive increase in frequency with temperature by a negative increase in temperature.

The AT-cut is an example of such a special cut. It has a very low temperature coefficient and is used for frequencies up to 5 mc. Above this, to 10 mc, the BT-cut is useful. The variation in frequency with temperature for both these cuts is shown in Fig. 4.11. Note how the BT-cut reaches a zero coefficient of temperature around 25° C which, translated into the more common Fahrenheit, is 77°. This is close to the average operating temperature inside the radio chassis.

**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 many F-M high-frequency sets to place the oscillator below the signal frequency.\* This is quite contrary 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 simply explained. 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 condenser 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 1005–2155 kc — or a range with approximately a 2 to 1 ratio. This is easily obtained in practice with one condenser and coil.

Now let us look at the higher F-M band, 88–108 mc. We will assume an I.F. of 15 mc. If the oscillator is placed above the carrier frequencies, its range will be from 103–123 mc. If placed below the carrier frequencies, its range will be from 73–93 mc. Design of tuning circuits does not enter into this discussion because both ranges are perfectly possible with one set of components. But in choosing the lower frequencies, we gain the advantages of greater ease in construction, more easily obtained stability, and the use of less critical components. The advantages

\* While the reasons outlined stress placement of the oscillator on the low-frequency side of the signal, there are sets manufactured in which the oscillator frequency is above the signal. In this position any radiated oscillator power will not cause interference in nearby television receivers which do receive signals in the 76–82 mc and 82–88 mc video channels. Interference may arise when the oscillator frequency is below the F-M (88–108 mc) signal frequency.

of lower frequency oscillators become even greater when crystal control is desired because here we are using a fairly high harmonic of the crystal and the strength of the harmonic voltages decreases rapidly. The lower the crystal harmonic used, the simpler the receiver design becomes. Hence, the oscillators in many F-M receivers will be in the range from 77.3–97.3 mc, this being 10.7 mc below the signal range, 88–108 mc.

## CHAPTER 5

### 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,

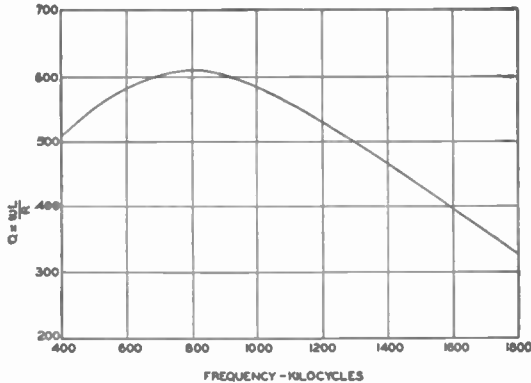


FIG. 5.1. The variation of  $Q$  (of a coil) with frequency.

designing a tuned circuit for use at one frequency only, can better obtain the maximum amplification than if he were 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. 5.1 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  $\omega L$  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, condensers 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 listed above.

**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  $\pm 75$  kc about the center value.

Now suppose — and this may readily happen — 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

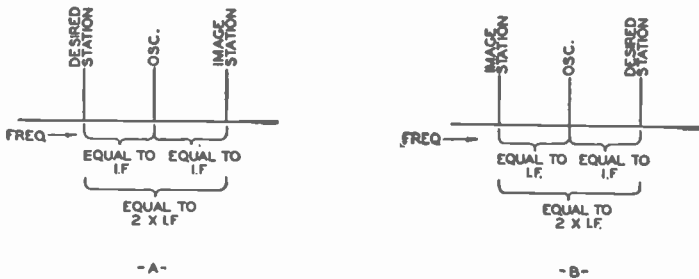


FIG. 5.2. The relationship between the desired and image stations when the oscillator frequency is above (A) and below (B) the signal frequency.

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. 5.2.

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  $\pm 100$ -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.

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. 5.2). 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 90.5 mc overloads the input of the receiver, a very good chance exists that its second harmonic would be formed, the frequency being  $2 \times 90.5$  mc, or 181.1 mc. The difference between 190.2 mc and 181.1 mc is 9.1 mc. This would reach the plate circuit of the converter stage and be accepted by an I.F. amplifier designed for 9.1 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 second 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 generally 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 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. This not only is advantageous in reducing the probability of spurious responses, but also in permitting a lower oscillator frequency to be used. When the I.F. is high, the oscillator can be operated that much lower than the signal. An oscillator operating at a lower frequency can be more easily designed for greater stability and efficiency than an oscillator functioning at the much higher frequencies. 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 of 10.7 mc has been recommended. It would not be at all surprising to see it rise — and settle — at some higher frequency, possibly between 20 and 30 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. 5.3. 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. 5.4, 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 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 30 db 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

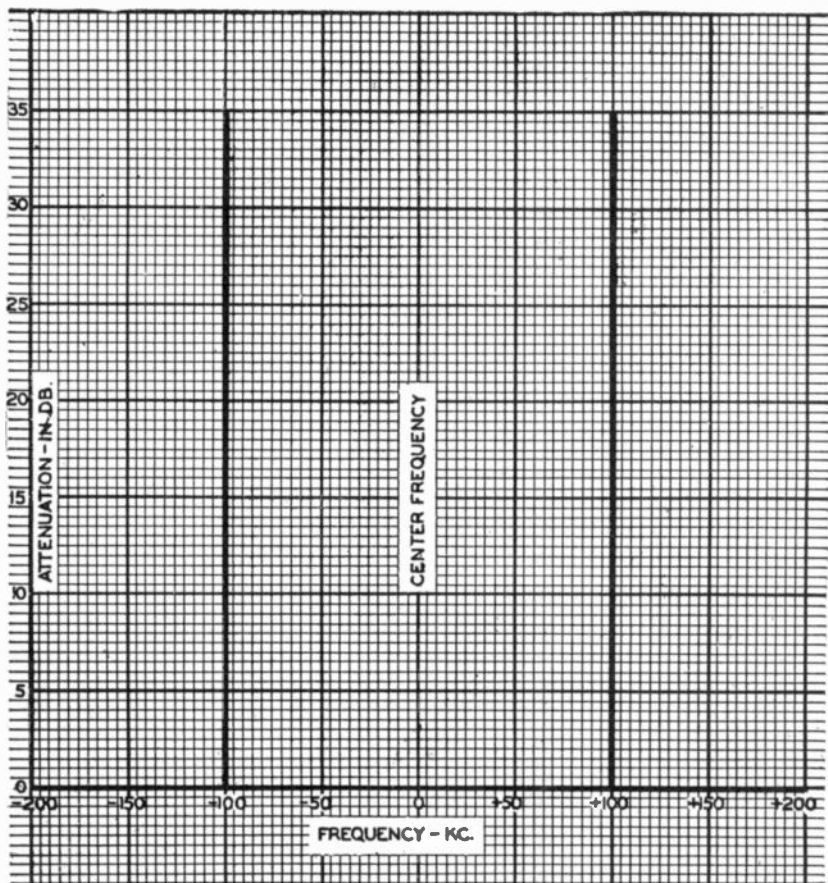


FIG. 5.3. An ideal selectivity response.

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. 5.4) 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,



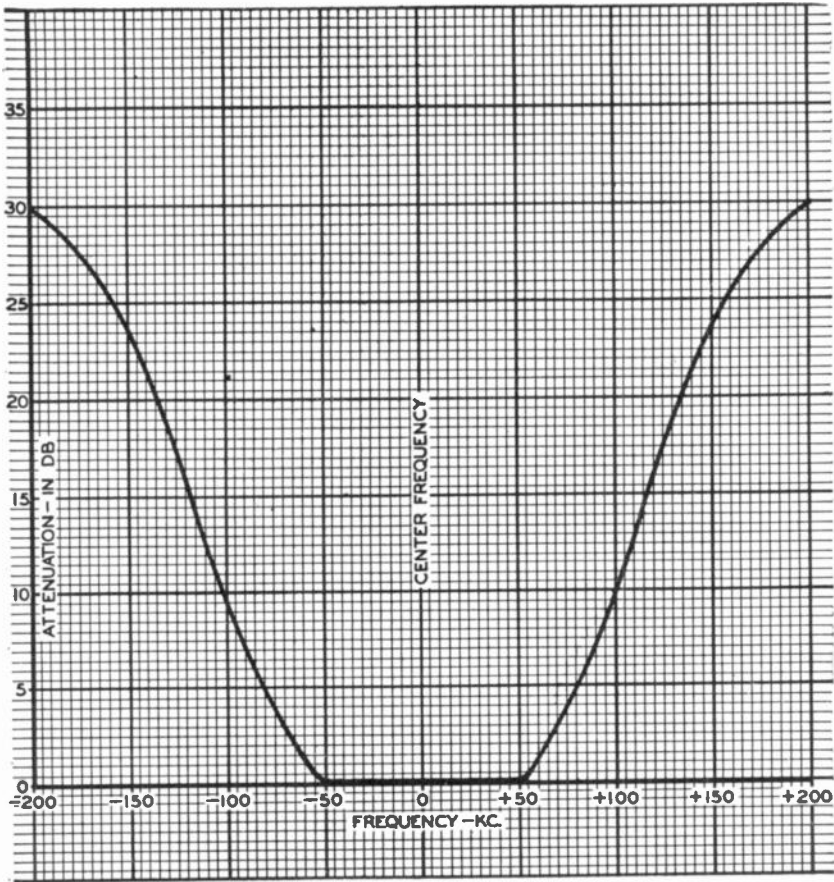


FIG. 5.4. The practical desired selectivity for an F-M receiver.

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

$$A = g_m \frac{\omega L Q}{2},$$

where  $g_m$  = mutual conductance of the tube and  $\omega L$  = inductive reactance of the coil. This formula is based on a double tuned

arrangement (Fig. 5.5) where  $L_1 = L_2$  and  $Q_1 = Q_2$ . This is customary as it simplifies mass production.

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 ( $\omega L$ ) is presented to the tube and the greater the output voltage. But from the formula

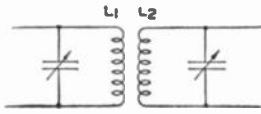


FIG. 5.5. A double tuned transformer. For ease in production,  $L_1 = L_2$  and  $Q_1 = Q_2$ .

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

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, 6AC7, with a mutual conductance of 9000 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 would be unstable because of the changes that take place 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 impossible. 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 con-

ductance 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. The solution employed in the construction of the 1851-2-3 tube series was 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  $C_{pk}$  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 capacities 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 manual. 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			$G_m$ (micro-mhos)	$A$	Figure of Merit $G_m/C_{total}$
	$C_{gp}$ (mmf)	$C_{in}$ (mmf)	$C_{out}$ (mmf)			
6AB7	0.015	8	5	5,000	50	363
6AC7	0.015	11	5	9,000	90	519
6AG5	0.025	6.5	1.8	5,000	50	524
6AG7	0.06	13	7.5	11,000	110	405
6AK5	0.01	4.3	2.1	5,100	51	737
6AK6	0.12	3.6	4.2	2,300	23	216
6C6	0.007	5.0	6.5	1,226	12.3	110
6D6	0.007	4.7	6.5	1,600	16	141
6R6	0.007	4.5	11	1,450	14.5	93
7V7	0.004	9.5	6.5	5,800	58	358
7W7	0.0025	9.5	7.0	5,800	58	348
6AU6	0.0035	5.5	5.0	4,450	44.5	418
6BA6	0.0035	5.5	5.0	4,400	44	417
6SH7	0.003	8.5	7.0	4,900	49	313

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.

**I.F. Amplifier Adjustment.** Schematically, the I.F. amplifier of an F-M receiver (excluding the limiters for the present) does not

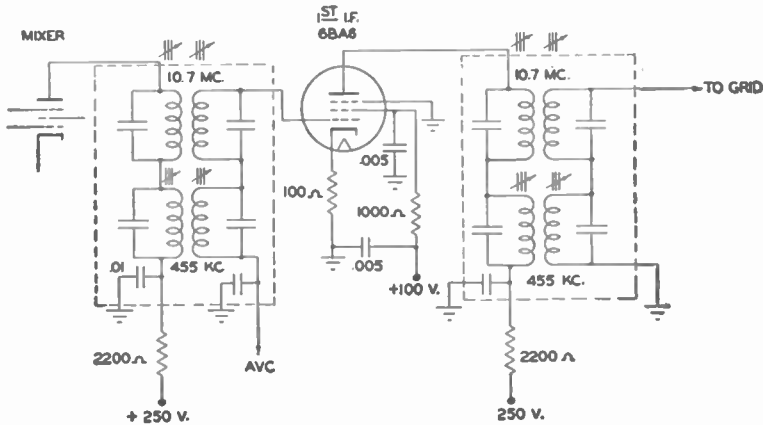


FIG. 5.6. An I.F. amplifier. The symbol above each coil indicates permeability turning (courtesy RCA).

differ in form from those amplifiers with which we are familiar from sound broadcast receivers. A typical diagram is shown in Fig. 5.6. The general tendency in 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 condenser 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. 5.6. Fig. 5.7 is indicative of this type of a circuit and is employed in some Scott receivers. 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 condenser, being common to both circuits. Since the transfer of energy takes place in this manner, the circuits

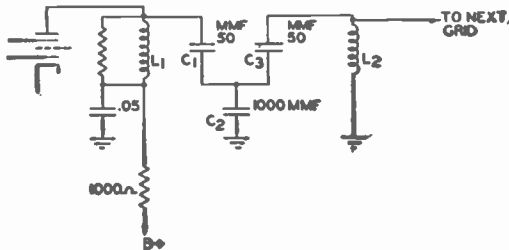


FIG. 5.7. A complex interstage coupling network used in Scott receivers.

are said to be capacitively 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.

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 and short-wave 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 and short-wave frequencies. To keep connecting wire lengths to a minimum and for space economy, both I.F. transformers are placed within the same can. In some constructions, all windings are placed on the

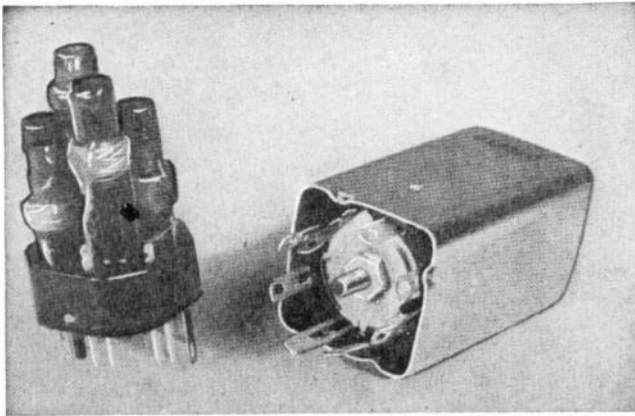


FIG. 5.8. The compact A-M, F-M, I.F. transformer showing the method of coil arrangement.

same post. In others, like the one illustrated in Fig. 5.8, there are several posts, each for an individual primary and secondary winding of the tuned circuits. In the composite I.F. transformer shown, an indented area at the underside of the unit contains a stack of silvered mica plates. These are the condensers which resonate the coils to the approximate frequency. The threaded iron core, which is mounted in the hollow center section of each post, then permits adjustment for an accurate alignment. The two primary condensers are silver printed upon one mica disc; the secondary condensers utilize another mica disc. Contact to each silvered surface is made with double-ended metal strips. With two ends,

separate lead connections for the chassis wiring and coil terminals may be made. To make the transformer as compact as possible all by-pass condensers associated with that particular section of the circuit are contained within the unit by means of additional mica discs. In this way frequency variations due to temperature changes are limited approximately to a value between 10 and 20 kc. Separate openings at the top of the metallic shield provide access to each iron core.



## CHAPTER 6

### 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 only because of the limitations of conventional F-M detectors, or, as they are more commonly known, discriminators. The conventional F-M discriminator is 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 limiter substantially eradicates the A-M and presents to the discriminator a wave that is wholly F-M. We mention conventional F-M detectors because, as we shall see later, there are discriminators that are insensitive to amplitude variations in the incoming F-M signals. For these, limiters are unnecessary. This results in a much simpler construction. Undoubtedly, the discriminator that requires the use of limiters will eventually be displaced by units that function without them. However, many sets still employ the conventional discriminator, and limiters are used extensively.

**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. 6.1 where the forms of the input and output voltages of a limiter are 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. 6.2. 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

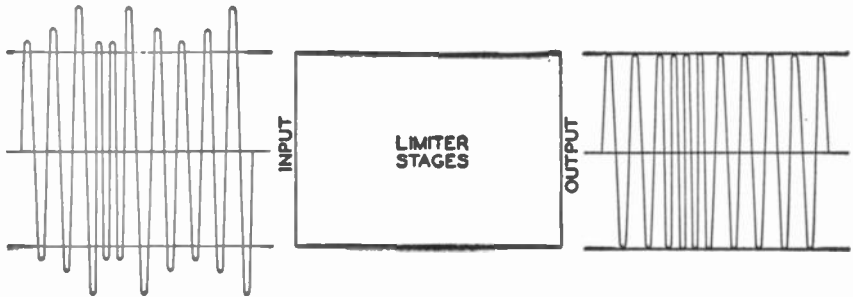


FIG. 6.1. The limiter removes the amplitude variations (modulation) from the input signal.

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, trolley lines, 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.

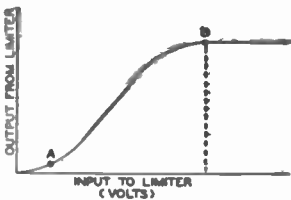


FIG. 6.2. For proper limiting action, the input voltage must be sufficiently strong to operate the limiter at point *B* or beyond.

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. 6.3 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.

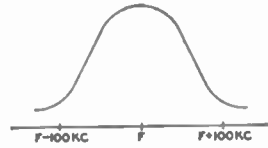


FIG. 6.3. Conventional selectivity characteristics.

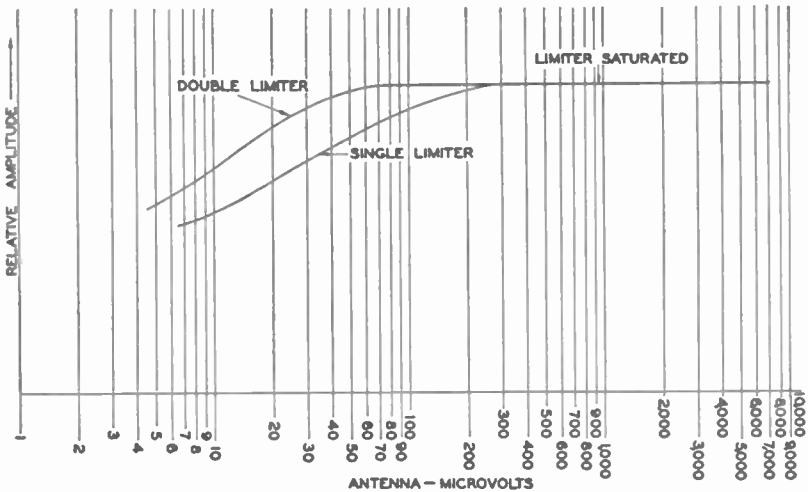


FIG. 6.4. A comparison of the input signal voltage required to produce constant output from single and double limiters.

**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 characteristics shown in Fig. 6.4. According to the graph, the knee of this limiter curve is reached for inputs of 25 microvolts at the antenna. As long as the incoming signal has this value, the limiter will function at saturation and no amplitude variations will

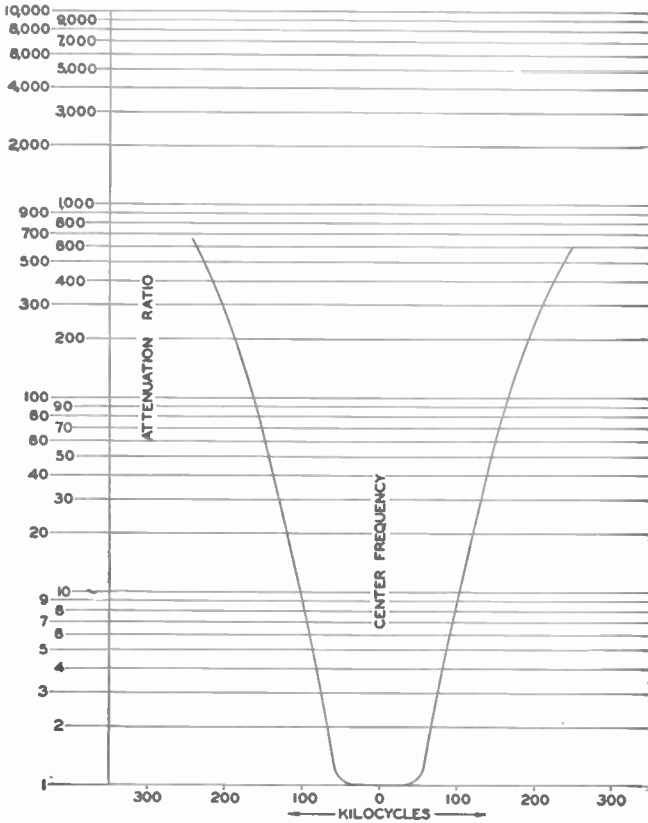


FIG. 6.5. Selectivity curve for an F-M receiver tested.

be obtained at the output. Now consider the curve of Fig. 6.5. 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. 6.4, if

25 microvolts is required at the carrier frequency to give limiter saturation operation, then one and a quarter times this amount, or 31 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 to one and a half times the mid-frequency value and at 100 kc, the attenuation ratio is 7. Thus, if we compute the sensitivity of a receiver at points 100 kc off resonance, we find that we need signals of 165 microvolts for complete limiting action. All this is highly important in designing the characteristics of the intervening I.F. stages and the limiter. Beyond  $\pm 100$  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 .....	10x	
Each I.F. stage (except the one preceding the limiter) .....	50x	
I.F. stage preceding limiter .....	40x	
Single limiter .....	2.5x	2
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.

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.

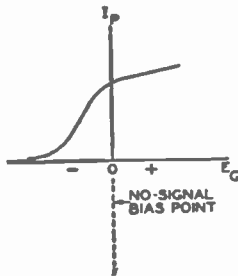


FIG. 6.6. The no-signal bias point for a limiter stage using grid-leak bias.

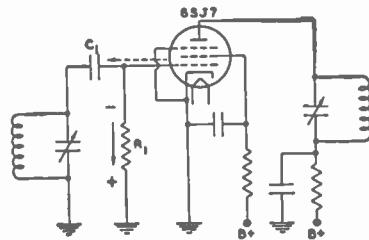


FIG. 6.7. A grid-leak bias limiter stage.

**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. 6.6. Upon the application of a signal, the grid is driven positive and current flows in the grid circuit charging up condenser  $C_1$  (see Fig. 6.7). 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 condenser continues to charge throughout the entire portion of the positive half of the incoming signal.

During the negative portion of the input cycle, the charge on condenser  $C_1$  begins to force its way through  $R_1$  back to ground. The condenser must discharge because it represents a potential difference and as long as a complete path is available, current must flow. This is fundamental to all electrical circuits. The electrons passing through  $R_1$  develop a potential difference with the polarity as indicated in Fig. 6.7. From the moment the input voltage departs from 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 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. Due to 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. 6.8 shows how the tube starts off with zero bias and, after a few cycles, reaches the equilibrium bias point.

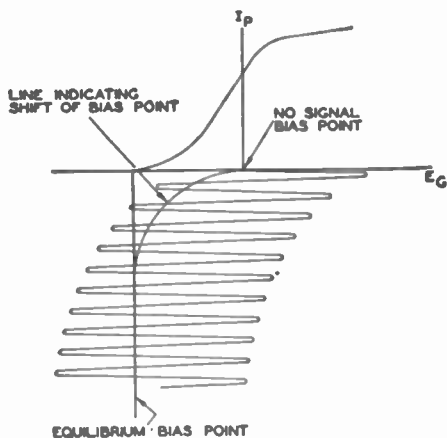


FIG. 6.8. The shift in bias potential during the first few cycles.

The equilibrium bias-point voltage is dependent upon a variety of factors. It depends upon the strength of the input voltage, the

voltages on the tube elements and the time constant of the grid condenser and resistor. For the latter, the time constant is defined as the time, in seconds, for the condenser 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

$$\begin{aligned} t &= R \times C \\ t &= 0.0001 \times 10^{-6} \text{ farads} \times 50,000 \text{ ohms} \\ t &= 5 \times 10^{-6} \text{ seconds} \end{aligned}$$

In other words, it requires  $5 \times 10^{-6}$  seconds for the charge on the condenser — no matter what it is — to diminish to 37 per cent of its initial value.

Five  $\times 10^{-6}$  seconds, 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 condenser 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 condenser. As we noted previously, with a condenser of 0.0001 mf value and a resistor of 50,000 ohms, the time constant is 5 microseconds. Com-



pared to the time of one cycle of a 100-mc carrier, 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

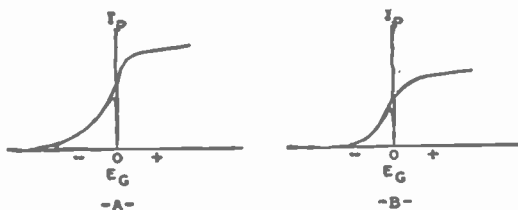


FIG. 6.9. (A) The normal characteristic curve for a sharp cut-off tube. (B) The modified curve obtained with lowered plate and screen voltage.

have been used and the effect of most impulses 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. 6.9. In Fig. 6.9A, we have the normal extent of the  $E_g I_p$  curve of a sharp cut-off pentode, say a 6SJ7 or 6SH7. 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. 6.9B. 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. 6.10.

It was noted previously that, in general, there are two major types of noise which affect a radio wave. One is random or fluctu-

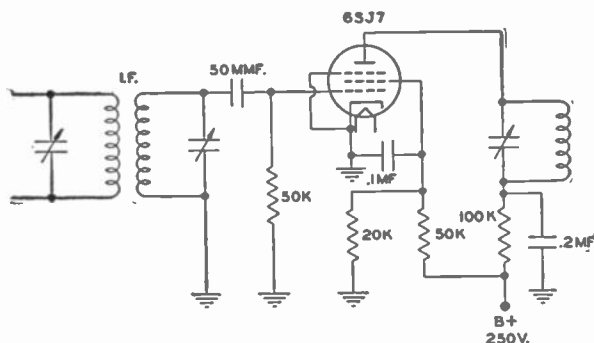


FIG. 6.10. A limiter stage functioning with lowered electrode voltages and grid-leak bias.

ation 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 damps the input tuned circuit and consequently lowers 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.

**Cascaded Limiters.** To obtain the advantages that low time constants offer in combating sharp impulses with the better regula-

tion and higher gain of long time-constant networks, many manufacturers have used two limiters in cascade. This is shown in Fig. 6.11. The first limiter has a time constant of 1.25 microseconds. Due to the higher values possible in the second circuit, the gain of both stages averages around 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. 6.4. With two limiters, the knee of the curve is reached by signals of 10 microvolts strength at the receiver input. With one

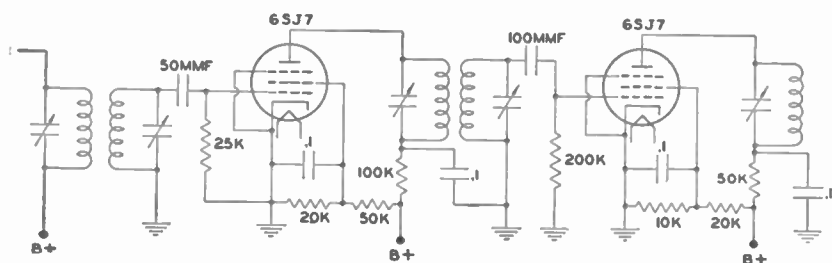


FIG. 6.11. A two stage (or cascaded) limiter circuit.

limiter, an input signal of 25 microvolts strength is required. Any signal weaker than 25 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.

**Commercial Limiter Circuits.** There are two basic limiter circuits in present commercial receivers. First, there is the single-stage unit similar to the circuit shown in Fig. 6.10. 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 unless a high degree of amplification is available in the stages preceding the limiter. Cascaded limiters may be either resistance-coupled, as shown in Fig. 6.12, impedance-coupled as in Fig. 6.13, 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

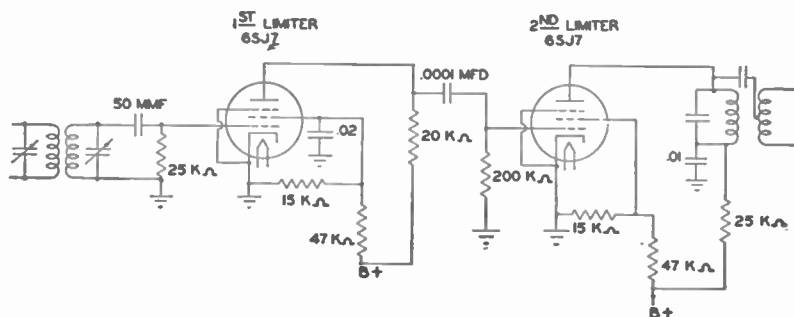


FIG. 6.12. A two-stage resistance-coupled limiter.

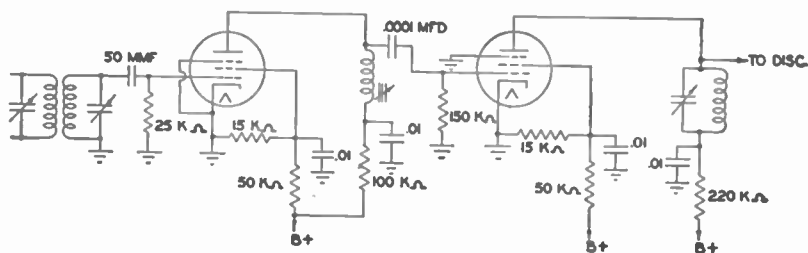


FIG. 6.13. Two limiter stages, impedance-coupled.

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.

Note that it is only when the incoming signal is exactly at the center frequency that the output of the discriminator is zero. Thus, no matter what type of signal is received at that one particular frequency, no output is obtained. However, when the incoming frequency shifts the slightest amount from this midpoint, a limiter becomes necessary. Since the incoming signal is off-center practically all of the time, we find that the limiter is always needed. Exception to the latter statement occurs only with discriminators that do not respond to amplitude variations.

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.

**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. Meters with a zero center scale are useful, but meters are no longer considered in vogue. However, some form of tuning eye or "magic eye" is satisfactory and this is the present trend among manufacturers.

Indications with an electron-ray tube are obtained visually by means of a fluorescent target. The more familiar of these tubes are the 6U5/6G5 and the 6E5. Both contain 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. 6.14). 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. 6.16 illustrates clearly the common cathode extending through the tube.

In order to have a sector of the  $360^\circ$  of the fluorescent screen serve as the indicator, a thin rod extends up from the triode plate (see Fig. 6.15) 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

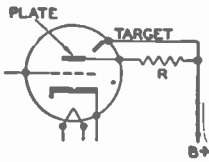


FIG. 6.14. The schematic symbol for a "Magic Eye" tube.

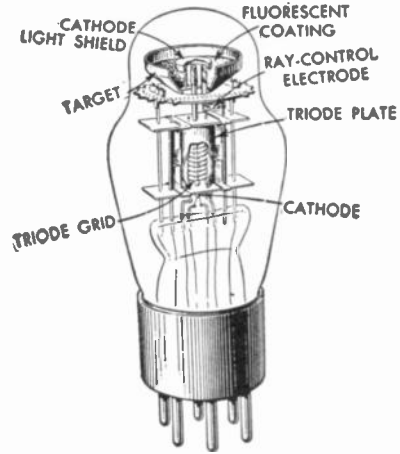


FIG. 6.15. The internal structure of the "Magic Eye" tube.

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. 6.14, 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. 6.16A.

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

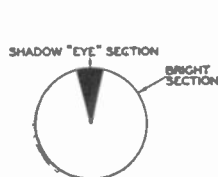


FIG. 6.16A. The width of the eye when the control grid of the tube is quite negative.

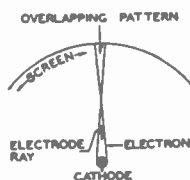


FIG. 6.16B. The electronic paths in the "Magic Eye" tube when an overlapped pattern is produced.



FIG. 6.16C. The eye opening when the grid becomes more positive.

is closer to the cathode. Electrons then flow in the manner shown in Fig. 6.16B 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. 6.16C.

In amplitude-modulated receivers, the A.V.C. line is attached





across the 350,000 ohm resistor is applied to the grid of a 6U5 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 6U5. The stronger the signal, the more negative

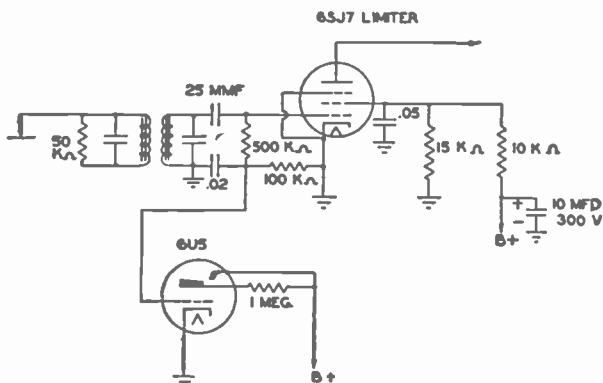


FIG. 6.18. The turning indicator network in the Howard 718 F-M receiver.

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. The 2.2 megohm and 100,000 ohm resistors, together with the 0.02 mf condenser, form a decoupling unit to prevent the alternating components of the input signal from reaching the grid of the 6U5. The edges of the two sections of the fluorescent target become hazy when this occurs. In Fig. 6.18 the circuit in the Howard F-M 718 receiver is practically the same, as above, with a simplification in the decoupling components. Here only the 0.02 mf condenser is used.

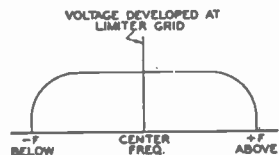


FIG. 6.19. The variation of limiter grid voltage with strong signals.

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. 6.19 shows that there is a wide 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 discriminators which do not require the use of limiter stages and some other means must be devised if tuning devices are to be used with these receivers.

One method, found in some sets, employs a 6E5 tuning tube that obtains its input from the discriminator output. This is shown

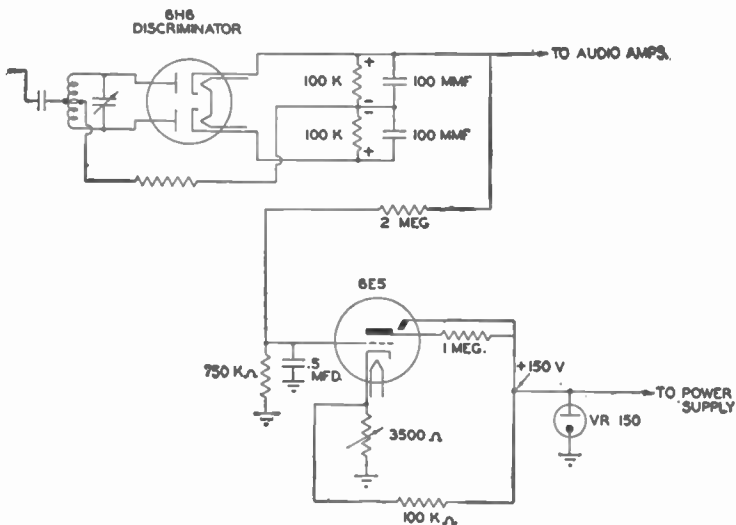


FIG. 6.20. A 6E5 tube connected to the discriminator output.

in Fig. 6.20. The 6E5 is adjusted to cut-off by variation of the 3500 ohm variable resistor in its cathode circuit. The VR-150 voltage regulator tube maintains a constant potential of 150 volts. 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 condenser, 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 another arrangement, the tuning indicator is attached somewhat differently to the output of the discriminator. The indicator functions on a combination of voltages taken from both the limiter and the discriminator. The result is a very sharp indication of resonance. The circuit is shown in Fig. 6.21. As far as the audio frequencies are concerned, the circuit consisting of a 6SJ7 limiter and a 6H6 discriminator is entirely conventional. The

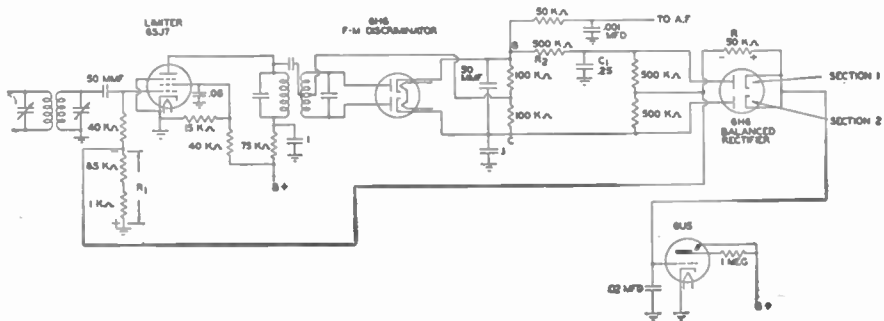


FIG. 6.21. A sharply indicating tuning eye utilizing voltage from the limiter and the discriminator. (Courtesy of Philco.)

tuning indicator, in this network, requires the use of a 6H6 balanced rectifier and the tuning eye tube, a 6U5. Connection into this circuit is also made from point *A* in the limiter grid circuit.

If we examine the schematic, we find that the grid of the 6U5 receives its potential from essentially two points. First, there is the voltage across resistor *R*, which is in the cathode leg of the balanced rectifier. A voltage develops across *R* only when current flows in the balanced rectifier. Second, the grid of the 6U5 receives voltage from the potential across *R*<sub>1</sub>, the 8500 and 1000 ohm resistors in the limiter grid network. Two resistors are used to permit attachment of a test meter. The division, however, has nothing to do with the operation of the tuning indicator. We may disregard the separation and label both as *R*<sub>1</sub>.

When a signal is tuned in, grid current flows in the limiter and a

negative voltage is developed on  $R_1$  at the point where the 6U5 grid is connected, point  $A$ . The signal continues to the discriminator and if the station is not correctly tuned in, the *average* voltage across  $B$  and  $C$  will not be zero. We stress the term "average" because audio voltage is also developed at these points. However, if the station is properly centered, the average voltage developed here will be zero. The output from the discriminator, besides feeding the audio stages through a de-emphasis circuit, is also applied to each plate of the balanced rectifier. Let us see what happens when the station is not properly tuned in.

A signal passing through the set, whether centered or not, will cause a negative voltage to appear at point  $A$ . However, if the station is off-center, a resultant voltage develops at point  $B$  which will be either positive or negative with respect to point  $C$ , depending upon whether the station is detuned to one side of the center resonant point or the other. It makes little difference to which side the station is detuned. The important point is the off-centered position.

If point  $B$  is positive with respect to point  $C$ , then section 1 of the balanced rectifier will conduct, causing a voltage to appear across resistor  $R$ . This positive voltage will counterbalance the negative voltage presented by resistor  $R_1$ , and the eye of the tuning indicator is kept open. If point  $B$  is negative with respect to point  $C$ , then section 2 of the balanced rectifier conducts, resulting in the same voltage across  $R$ . Thus, no matter to which side the set is detuned, the positive voltage developed across  $R$  counterbalances the negative voltage across  $R_1$  and prevents full closing of the tuning indicator eye.

When the set is adjusted so that the station is centered, neither section of the balanced rectifier conducts, no counterbalancing voltage appears across  $R$  and the full negative voltage from point  $A$  is present at the grid of the 6U5. The V-shaped shadow of the fluorescent screen becomes smaller and, with large signals, may even overlap. Tuning is sharp and correctly indicated when the shadow is smallest or overlap occurs.

A variation of the above circuit suitable for receivers not employing a limiter is shown in Fig. 6.22. The 6E5 tuning indi-

cator is biased approximately to cut-off by the voltage drop across  $R_1$ . This keeps the shadow on the eye very narrow, almost to the point where the edges on each side of the shadow overlap. The control grid of the indicator is attached to the cathode end of  $R_2$ , a resistor that develops a positive voltage each time one of the balanced rectifier diodes conduct. The diodes will conduct only when a station is not properly tuned in, for only under these circumstances will the average voltage appearing across points  $A$  and  $B$  be either positive or negative. Application of positive voltage from  $R_2$  to the grid of the 6E5 will counterbalance some of the

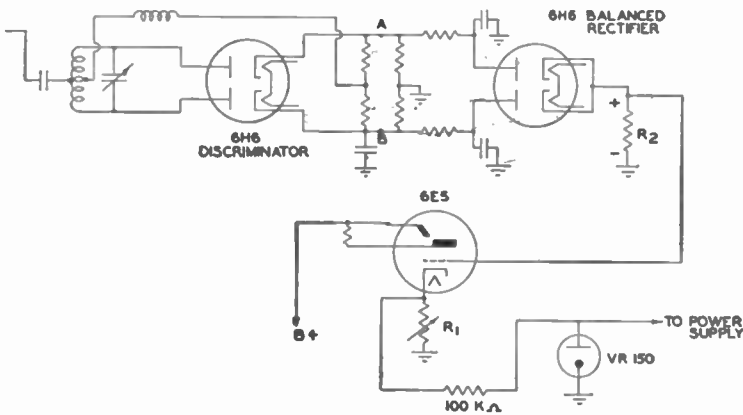


FIG. 6.22. A circuit similar to Fig. 6.21 but requiring only the discriminator voltage.

negative voltage from  $R_1$  and cause the eye to open. Correct tuning is indicated when the voltage across  $R_2$  is zero and the eye is practically closed.

The 6E5 is preferable to the 6U5 in this circuit because of the sharp cut-off characteristic properties of the 6E5 triode. Sharper indications with less voltage are possible.

**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, due to 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

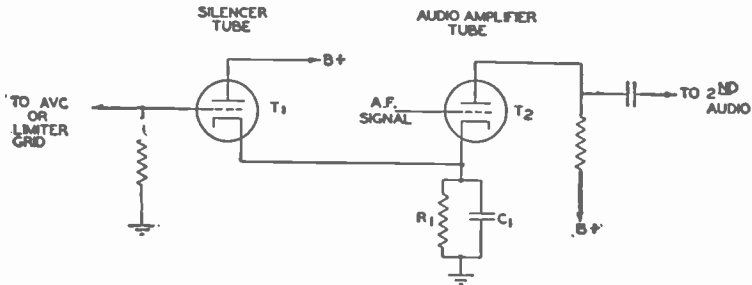


FIG. 6.23. A simple silencer arrangement.

condition is maintained until the signal strength of the input of the receiver again increases.

A simple circuit illustration of this principle is shown in Fig. 6.23.  $T_1$  is the silencer, or squelch tube.  $T_2$  is the first audio amplifier. Both tubes have a common cathode resistor and condenser,  $R_1$  and  $C_1$ . As long as  $T_1$  is maintained at cut-off by a negative voltage on its grid,  $T_2$  functions normally. Removal of the negative grid voltage at  $T_1$ , or at least a decrease to permit an appreciable current to flow, quickly brings  $T_2$  to cut-off because of the excessive voltage appearing across  $R_1$ . This holding action continues until  $T_1$  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. Triodes like the 6J5 ( $g_m = 2500$  micromhos) and the 6L5-G (1900 micromhos) 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 a Meissner receiver, shown in Fig. 6.24, the grid of the silencer tube is directly connected into the grid circuit of the 6SJ7 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

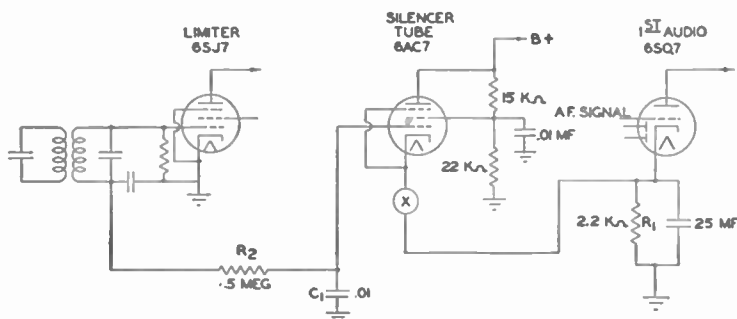


FIG. 6.24. Another possible silencer circuit.

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, for example, might be needed on weak signals when the presence of the silencer could prevent reception.

In the Espy model (2170), Fig. 6.25, 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 operate 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

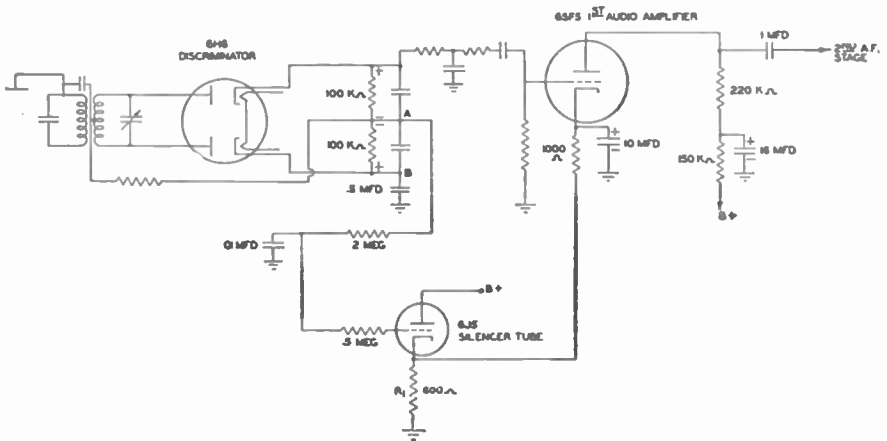


FIG. 6.25. A silencer functioning on the voltage obtained from the discriminator.

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.

Silencers and squelch circuits are necessary only in F-M sets using the limiter and discriminator type of arrangement. The ratio detector and the locked-in oscillator, which require no limiter, do not need any silencer system because the noise level is quite low between stations.

**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 overloading 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 only suffer 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.

Practically the same reasoning applies to receivers possessing the other types of detectors. In the Philco lock-in oscillator circuit, the minimum I.F. input signal required to operate the detector is approximately one-half volt rms for full deviation. The voltage required for other off-center frequencies varies linearly with deviation. Thus, if a minimum of one-half volt is required for full deviation, correspondingly less voltage is needed for smaller frequency swings. Inclusion of an A.V.C. voltage could only decrease, not increase, the sensitivity of the receiver and thus little is gained by its inclusion. One possible advantage, perhaps, of an A.V.C. voltage is to decrease the possibility of overloading in the R.F. stages, which can serve as a source of spurious response. This was mentioned previously. Receivers employing ratio detectors (see Chapter 7) have a ready source of A.V.C. voltage, and, therefore, advantage is taken of this to apply this control voltage to the R.F. stage and the first I.F. amplifier. In receivers utilizing other forms of detectors, where the A.V.C. voltage is not as readily obtained, most manufacturers omit the A.V.C.

## CHAPTER 7

### F-M DETECTORS

**Discriminator Action.** The heart of the F-M receiver, where the frequency variations are reconverted to intelligible sound, is at the discriminator. Since the intelligence is contained in the different

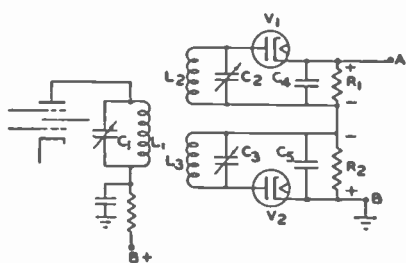


FIG. 7.1. An early form of discriminator. The secondary contains two separate windings.

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. 7.1.

$L_1$  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. 7.1, 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 above the center I.F.

value, while  $C_3L_3$  is peaked to a frequency the same number of kilocycles below 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. 7.2A, the frequency response curves of both tuned circuits are shown with respect to frequency. In Fig. 7.2B, 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. 7.1) it can be seen that the polarity of the voltage obtained at the output terminals  $A$  and  $B$  will depend upon which

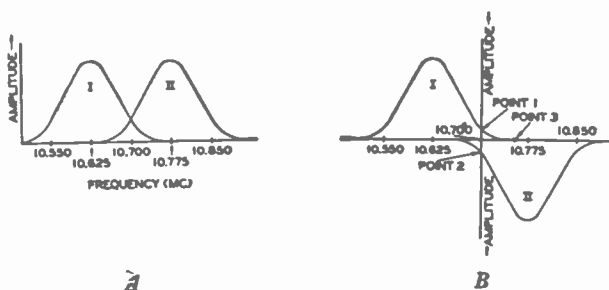


FIG. 7.2A. The secondary response curves. B. The same curves as in A in a position corresponding to the voltage developed across the load resistor.

resistor develops the larger voltage. If the voltage 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 to 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. 7.2B, 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, due to 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.

Since the output voltage represents the difference between the potentials present across  $R_1$  and  $R_2$ , we can combine both response

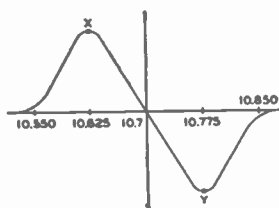


FIG. 7.3. The resultant curve derived by combining both separate curves of Fig. 7.2B

curves into one resultant curve. This has been done in Fig. 7.3. 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  $\pm 75$  kc. In most cases, the curve extends to  $\pm 100$  kc. The methods for aligning a discriminator are given in Chapter 11.

Condensers  $C_4$  and  $C_5$ , Fig. 7.1, 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 condensers, 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 wave form change, it would look as shown in Fig. 7.4. 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 recon-

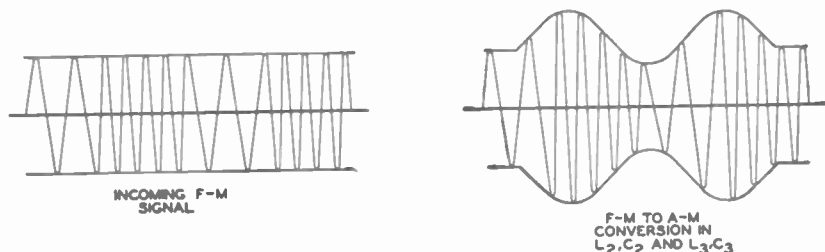


FIG. 7.4. The first step in the process of converting the F-M signal to the final audio voltage.

vert 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 circuit. It will be shown later that with a ratio detector it is possible to eliminate the effect of amplitude modulation without resort to a limiter.

**A Modified Discriminator.** The foregoing discriminator contains one primary and two secondary condensers. A more compact unit, one that requires only two adjustments, is the discriminator

shown in Fig. 7.5. Actually, what has been done here was to combine both secondary windings into one coil and use a single condenser to peak the secondary to the mid-frequency of the I.F. band. Now, the discriminator output voltage does not depend on the different response of two tuned circuits to an incoming frequency;

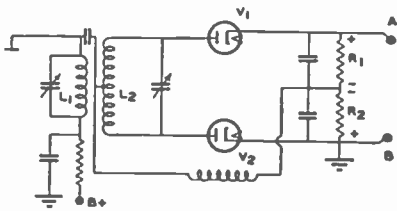


FIG. 7.5. A modified discriminator in widespread use today.

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 proceed with the explanation of this circuit, let us review some fundamental considerations of resonant circuits and the phase relationships of coupled circuits. In Fig. 7.5 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 due to 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.

When a voltage is induced into a coil, such as in the resonant circuit shown in Fig. 7.5, the current that flows will develop across

the coil a voltage that does not necessarily have to 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. 7.6. Due to 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. 7.7A, 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 into the secondary coil has a complete circuit consisting of the coil and the

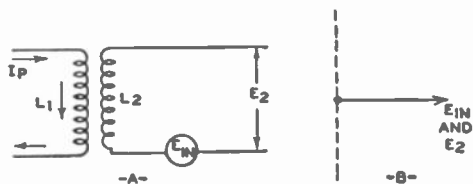


FIG. 7.6. The induced voltage in  $L_2$ , due to  $I_p$ , is in phase with the voltage appearing across  $L_2$ .

both in phase and amplitude. The voltage that is induced into the secondary coil has a complete circuit consisting of the coil and the

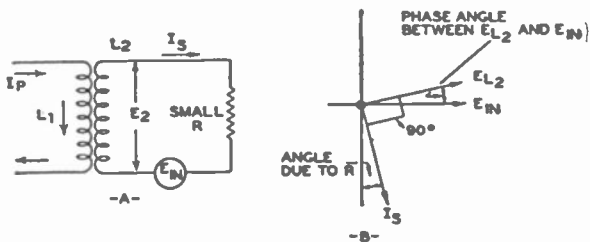


FIG. 7.7. The current and voltage phase relationship in the secondary circuit.

resistance. 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  $90^\circ$  (this from elementary electrical theory). This is shown graphically by the two vectors,  $E_{in}$  and  $I_s$ , in Fig. 7.7B.

Now this current, in flowing around the circuit and through  $L_2$ , will develop a voltage drop across  $L_2$  due to the coil's inductive reactance. As is true of every coil, the voltage will lead the current

by  $90^\circ$ . On the vector graph it can be placed in the position marked  $E_{L_2}$ ,  $90^\circ$  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  $E_{in}$ .

Following the same line of reasoning, we get the vector graph of Fig. 7.8 when the inserted resistance is high in comparison to the

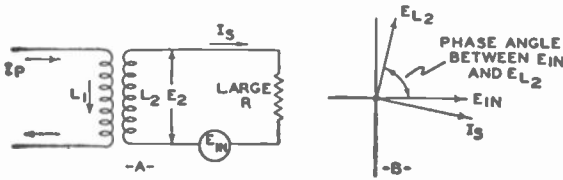


FIG. 7.8. The same circuit as shown in Fig. 7.7 except that  $R$  is now large. Note the difference in the phase angle between  $E_{in}$  and  $E_{L_2}$ .

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  $90^\circ$  out of phase with the current through it.

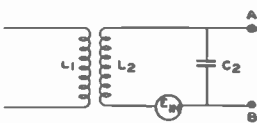


FIG. 7.9. A secondary resonant network.

*Secondary Resonant Circuit.* The next step is to replace the resistor,  $R$ , with a condenser,  $C_2$  (see Fig. 7.9). 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.

As far as the induced voltage is concerned, the coil and condenser 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, present 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. 7.10. At the resonant frequency,  $f_0$ , the opposition is least and the current largest. On either side of the resonant point the current decreases because the circuit impedance increases.

Fig. 7.10 shows only the variation in current amplitude with frequency. Now let us determine how the phase angle between the

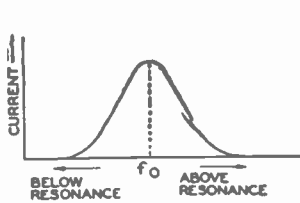


FIG. 7.10. The variation in current in a series resonant circuit.

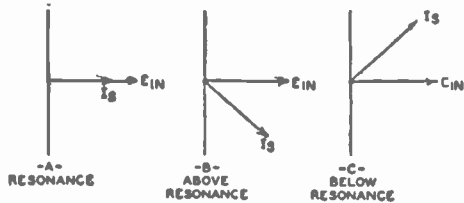


FIG. 7.11. The phase displacements between the induced voltage and the secondary current for different signal frequencies.

current ( $I_s$ ) and the induced voltage ( $E_{in}$ ) varies, as the frequency changes. At resonance, the sole 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. 7.11A. 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. 7.11B. Below resonance, the opposite occurs, and the current *leads*  $E_{in}$ . See Fig. 7.11C.

The foregoing phase shifting with frequency is the basis of operation of the discriminator shown in Fig. 7.5. 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

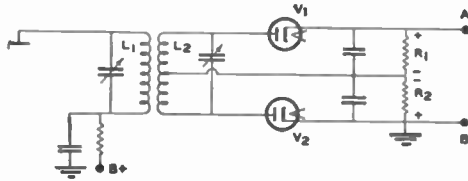


FIG. 7.12. Development of the modern discriminator. The inclusion of a center-tapped secondary and two separate rectifiers produce a balanced network.

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,  $R_2$  would predominate.

In this manner, swinging above and below the center frequency, we obtain positive and negative output audio voltages.

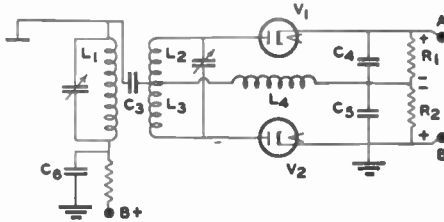


FIG. 7.13. The complete discriminator.

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. 7.12, we can obtain a balanced arrangement. This will 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. 7.13. If we compare this arrangement with the discriminator circuit of Fig. 7.5, we note they are identical.

**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

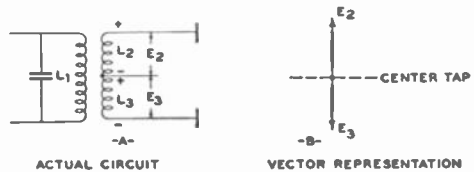


FIG. 7.14. Effect of center-tapping the secondary coil in the discriminator.

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. 7.14. 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.

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 volt-

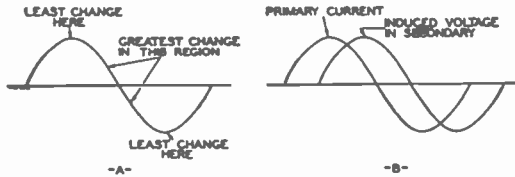


FIG. 7.15. (A) The regions of least and greatest change in an a-c wave. (B) The phase relationship between primary current and induced secondary voltage.

age,  $E_{in}$ . 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 occurs when the current is passing through zero, going from one polarity to another (see Fig. 7.15A). On the other hand, 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  $90^\circ$ , with the current leading. This is shown in Fig. 7.15B.

The voltage,  $E_1$ , developed across  $L_1$  will always be  $90^\circ$  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. 7.16A, then it must be placed  $90^\circ$  ahead of  $I_1$ , the current through  $L_1$ .  $E_{in}$ , on the same vector graph, must be placed  $90^\circ$  behind  $I_1$  as determined above. Hence, we see that the primary voltage and the induced secondary voltage differ by  $180^\circ$ . This is true at all frequencies. Remember that we are discussing  $E_1$  and the induced secondary voltage. We have not, as yet,

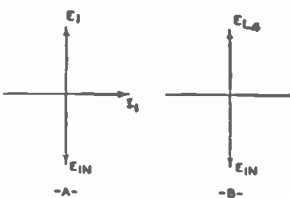


FIG. 7.16. (A) The phase differences between  $E_1$ ,  $E_{in}$  and  $I_1$ . (B)  $E_1$  replaced by its equivalent  $E_{L_4}$ .

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.

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. 7.16B. To the vectorial diagram, we must still add  $E_2$  and  $E_3$  and then combine them with  $E_{L_4}$ . Fig. 7.13 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 fully cover the operation of the discriminator, frequencies above, below, and at the mid-point value must be considered. The

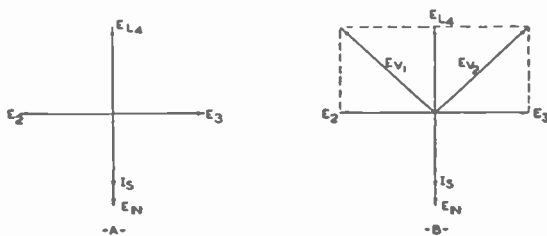


FIG. 7.17. (A) Vector diagram for an unmodulated signal. (B) The vector addition of  $E_2$  with  $E_{L_4}$  and  $E_3$  with  $E_{L_4}$ .

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. This will cover the entire range of an F-M signal. The output voltage from the discriminator will give the circuit response to input frequency variations.

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. 7.17A, 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  $90^\circ$ . Again, this 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  $90^\circ$  from  $I_s$ .  $E_2$  and  $E_3$  are on opposite sides of  $I_s$ , due to the center tap on the secondary coil (see Fig. 7.14).  $E_2$  and  $E_3$ , taken with reference to the center tap, are  $180^\circ$  out of phase with each other. To show this graphically, in Fig. 7.17A, 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  $E_{L_4}$  derive  $E_{V_2}$ , the voltage applied to  $V_2$ . It is evident from Fig. 7.17B 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

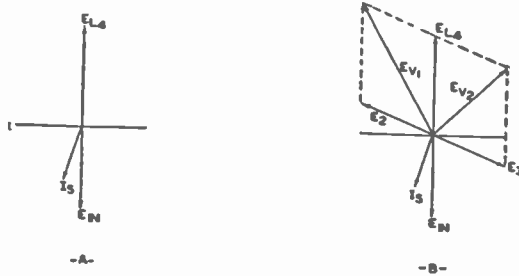


FIG. 7.18. Circuit phase conditions when the signal frequency is above the I.F. center value.

$R_2$ . 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  $180^\circ$  out of phase with the voltage across the primary. Hence, to start the vector diagram, we can insert these two voltages (see Fig. 7.18A). 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 condenser. Hence, in the present instance, the coil will cause the current to lag behind  $E_{in}$  (see Fig. 7.18A). Now we can draw in  $E_2$  and  $E_3$ , since they differ from the current ( $I_s$ ) by  $90^\circ$ .  $I_s$  is responsible for  $E_2$  and  $E_3$  appearing across the secondary coil and there is always a  $90^\circ$  phase difference between voltage and current in a coil.

Adding  $E_2$  and  $E_{L4}$ ,  $E_3$  and  $E_{L4}$ , vectorially, we see that the resultant vectors are no longer equal. In this case,  $E_{V1}$  is greater than  $E_{V2}$ . Consequently,  $R_1$  will develop a larger voltage than  $R_2$

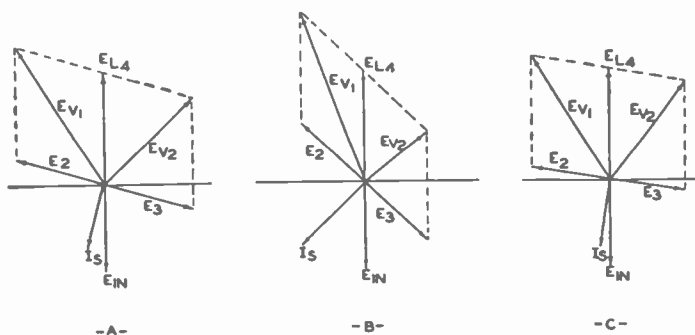


FIG. 7.19. Several additional off-frequency conditions. The input frequency is high.

and the output voltage 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. 7.19, illustrating the phase shifting that occurs.

*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  $E_{L4}$ .  $I_s$  follows next. Its position is ahead of  $E_{in}$  because as the frequency drops below resonance, the capacitive reactance of the condenser becomes greater than the impedance of the coil. In a capacitive circuit, current leads voltage. This places  $I_s$  ahead of

$E_{1n}$ .  $E_2$  and  $E_3$  are  $90^\circ$  out of phase from  $I_s$ , and these can be added to the vector diagram. Addition with  $E_{L4}$  shows the unbalance of voltages, this time favoring  $V_2$  (see Fig. 7.20). 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

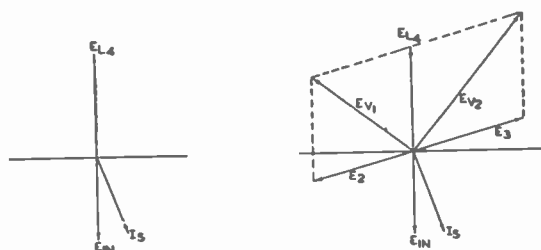


FIG. 7.20. The shift in vectors when the signal is below the I.F. midpoint.

be a faithful reproduction of the modulation as effected at the transmitter. The same discriminator "S" shaped curve derived for the previous discriminator (Fig. 7.3) 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. 7.21. The difference between this unit and the preceding circuit is to be found in the elimination of  $L_4$  and the use of only one condenser across the output instead of two. It may appear that  $E_{L4}$ , the reference voltage so necessary for the proper operation of the discriminator in Fig. 7.13, 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 from point  $A$ . However, being only on the order of 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. 7.21 than that of Fig. 7.13.

The circuit diagram of Fig. 7.21 may be drawn in many ways, one variation being shown in Fig. 7.22. At first glance there

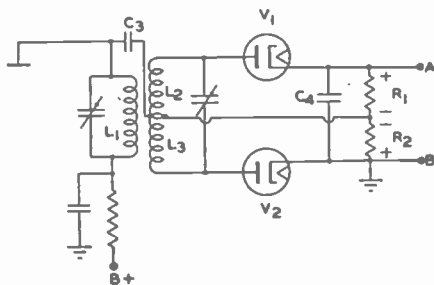


FIG. 7.21. A modified discriminator.

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

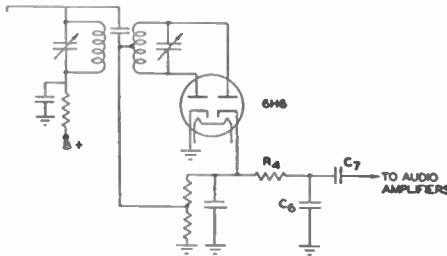


FIG. 7.22. The same discriminator as shown in Fig. 7.21, but slightly rearranged.

along its own peculiar lines and sometimes it requires a little patience to recognize the similarity of identical units.

$C_6$  and  $R_4$  of Fig. 7.22 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.

**A Modified Discriminator.** In standard A-M broadcasting receivers, it is customary to use a duo-diode triode or pentode for the dual function of detector and first audio stage. If we attempt to use the same tube for an F-M receiver, we run into difficulties because the duo-diode section of the tube contains a common cathode (see Fig. 7.23). Although there are two separate diode plates, there is only one common cathode. In the discriminator circuits thus far encountered, each cathode of a diode rectifier was connected to a different portion of the circuit. Hence, a conventional circuit could not be handled by a duo-diode. General Electric has, however, employed such tubes in their F-M sets with a modified discriminator circuit.

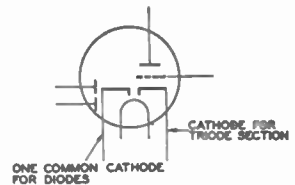


FIG. 7.23. A typical duo-diode tube.

The basic circuit is shown in Fig. 7.24.  $C_1$  and  $L_1$  form the primary tuner,  $L_2$ ,  $L_3$  and  $C_3$ , the secondary resonant circuit. The bottom of  $L_2$  and the top of  $L_3$  are connected by means of a small 200 mmf condenser. This condenser presents essentially no oppo-

sition to high I.F. and the bottom of  $L_2$  and the top of  $L_3$  may be considered as at the same potential. At the audio frequencies, the opposition of the 200 mmf condenser is high and one side is not at the same audio potential as its other side.  $R_1$  is the load resistor for  $V_1$ ;  $R_2$  is the load resistor for  $V_2$ . As before,  $V_1$  and  $V_2$  form two distinct and separate rectifier circuits.

A ground is placed at the left-hand side of  $R_2$ , this being equivalent to the ground at one end of  $R_2$  in the previous discriminator circuits. It represents one output terminal, and placement of a ground connection at the point shown has absolutely no effect on the operation of the circuit. The other output terminal, just as we had before, is taken from the opposite end of  $R_1$ . Here, too, the

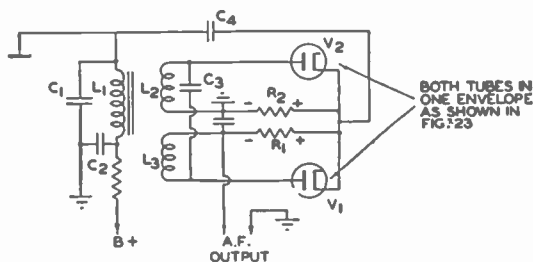


FIG. 7.24. A discriminator suitable for use with duo-diode tubes, containing a common diode cathode.

output represents the difference between the audio voltages developed across  $R_1$  and  $R_2$ .

The one outstanding departure from the other conventional design is the method of introducing the primary reference voltage into the secondary circuit. This is accomplished by  $C_4$ . For tube  $V_1$ ,  $E_1$  is brought to the secondary through  $C_4$  and  $R_2$ . Since  $C_4$  offers very little opposition to these high intermediate frequencies, the full primary voltage  $E_1$  appears across  $R_2$ . For  $V_1$ , the path is from the top of  $L_1$ , through  $C_4$  and  $R_1$  to ground. Again, in this path, only  $R_1$  offers any appreciable resistance, and the full primary voltage  $E_1$  appears across  $R_1$ . It is to be noted that  $R_1$  and  $R_2$  are both in parallel with  $L_1$ , as in the discriminator circuit of Fig. 7.21.

In operation, the circuit is exactly equivalent to either of the other two discriminators. At resonance, no output is obtained,

whereas above and below resonance, the output voltage will vary, the polarity being dependent upon the position of the incoming frequency. The rate at which the output voltage changes depends upon the rapidity with which the frequency changes position.

**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 is 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 affected. It would mean that perhaps one, at most two, I.F. stages would be needed. Recently, such a discriminator has been developed, one that does not respond to amplitude variations in the input signal over fairly wide limits. Due to its operation, it is known as a ratio detector.

To see why a ratio detector is immune to amplitude variations in the incoming signal, let us compare its operation with that of the ordinary discriminator.

In the discriminator circuit of Fig. 7.13, 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. Their 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.

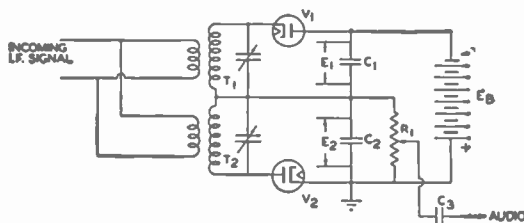


FIG. 7.25. Preliminary form of the ratio detector.

An elementary circuit of a ratio detector is shown in Fig. 7.25. In this form the detector is similar to the detector of Fig. 7.1 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. 7.1. 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  $C_2$ . Condenser  $C_3$  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 condenser — 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.

The practical form of the ratio detector is shown in Fig. 7.26. It uses the phase-shifting properties of the discriminator of Fig.

7.13.  $R$  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 condenser across  $R$  we maintain a fairly constant voltage. Thus, momentary changes in carrier amplitude are merely absorbed by the condenser. 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. 7.26. With this detector, it becomes possible to design an F-M receiver containing only two I.F. stages.

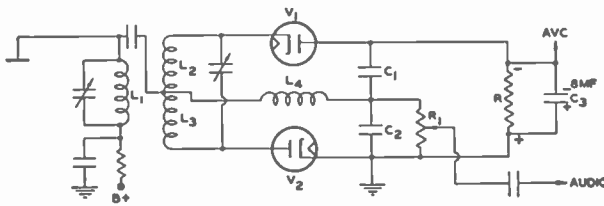


FIG. 7.26. Practical form of the ratio detector.

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. 7.26 represents but one form of this circuit. In Fig. 7.27 there is a modification 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. 7.27, 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

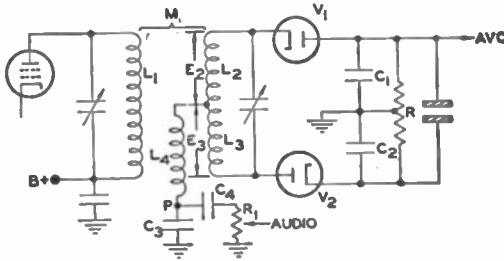


FIG. 7.27. A symmetrical ratio detector.

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 better understand the other modifications, the circuit of Fig. 7.27 has been rearranged (see Fig. 7.28). The path from the

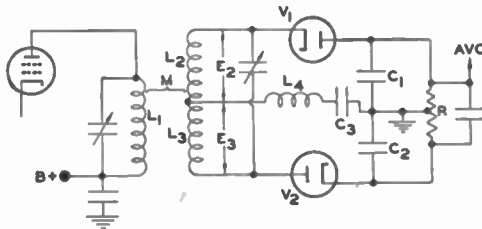


FIG. 7.28. The circuit of Fig. 7.27 rearranged to better indicate its mode of operation.

center tap on the secondary coil to the connection between  $C_1$  and  $C_2$  contains  $L_2$  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  $L_2$  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  $L_3$  to  $V_1$  and through  $C_1$  and  $C_2$  back to  $V_2$ . At mid-frequency, point  $P$ , Fig. 7.27, 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.

Since the voltage at point  $P$  is zero when the mid-frequency is received and either positive or negative at other frequencies, it can be used to control a tuning-eye indicator.

Another arrangement of the ratio detector is shown in Fig. 7.29A. This is similar to the circuit of Fig. 7.26 except that condenser  $C_2$  has been retained and  $C_1$  eliminated. The audio voltage is obtained from  $C_2$ , the same as in Fig. 7.26.  $L_4$  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. 7.29B 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. 7.18 and 7.20). 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.

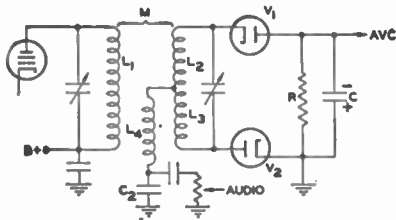


FIG. 7.29A. Another form that the ratio detector can assume.

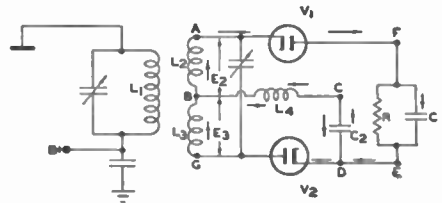


FIG. 7.29B. A rearrangement of Fig. 7.29A.

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 attach-

ing 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. 7.5. 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 input resonant circuit of the diodes is designed to present voltages to each diode which differ with frequency. In this manner, the output voltage will likewise be dependent upon frequency. The limitation of this discriminator rests in the fact that amplitude variations in the signal will appear in the detected audio voltages.

It was discovered that response to amplitude variations can be eliminated when the discriminator output is derived as a ratio instead of a difference voltage. Toward this end the ratio discriminator was developed. 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. Due to 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 that can be obtained. Thus, if we connect a small condenser in series with each tube, the voltage appearing across each will depend upon the input frequency (see Fig. 7.26). However, at all times, their sum is equal to the control voltage. Changes in signal frequency will merely alter the ratio of their voltages.

In a modified form of this detector, Fig. 7.28, the same control voltage is maintained, but the two condensers are replaced by one which is in series with both tubes. Its voltage is still affected by the current through each tube plus the control voltage, and hence it operates upon the same basic principle.

**The Beers Receiver.** The ratio detector was one attempt to simplify the construction of the F-M receiver. There is yet another solution available. This is the locked-in oscillator arrangement developed by G. L. Beers. A block diagram of the receiver is

shown in Fig. 7.30. Note that the conventional limiters have been eliminated and in their place a locked-in oscillator has been substituted. A discriminator closely resembling the discriminator of Fig. 7.13 is used, except that the linear range of response of this unit is considerably less than the 200 kc of the former unit. Besides these two units, the F-M receiver is similar to any previously discussed.

The locked-in oscillator is frequency-modulated by the incoming F-M signal. In other words, the incoming signal causes the locked-in oscillator to shift in frequency in exact step with the frequency variations of the incoming signal. However, the locked-in oscillator operates at one-fifth the frequency of the incoming I.F. signal. This reduces not only the center I.F., but also the deviation of the F-M signal about its central point. Thus, the  $\pm 75$ -kc swing

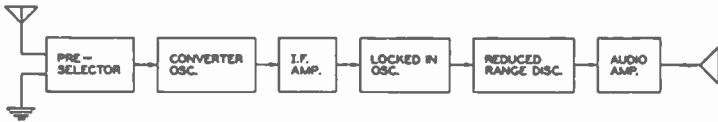


FIG. 7.30. A block diagram of the Beers receiver.

of the incoming signal is reduced to a  $\pm 15$ -kc swing at the oscillator output. Because of this, the following stage — the discriminator — can operate over a reduced range. In the process, the rate at which the frequency of the incoming signal shifts from one point to another is not altered in the output. This, of course, is the modulation frequency and has nothing to do with the compression of the frequency swing of the signal. A limiter is unnecessary in this circuit because the output from the locked-in oscillator to the following discriminator remains constant for a wide range of input voltages. However, if the input signal does not attain a certain minimum value, the oscillator is not “locked-in” by the incoming signal and no output is obtained.

A diagram of a lock-in oscillator is shown in Fig. 7.31. The output of the I.F. stages is applied to the oscillator tube, a 6SA7, at grid 1. The frequency here would be any of the current I.F. values, say 9.1 mc, or 9100 kc. In the output circuit, the plate tank ( $L_1$  and  $C_1$ ) is tuned to one-fifth of the I.F., or 1820 kc.  $L_1$  and  $C_1$

are coupled to  $L_2C_2$ , the latter being tuned to the second harmonic of 1820 or 3640 kc. Grid 3 receives this voltage and, with the plate, forms the oscillator. As we shall see presently, the coupling between  $L_1$  and  $L_2$  forces the oscillator to lock-in with the I.F. signal.  $C_3$  and  $R_1$  provide grid-leak bias for the oscillator.  $C_4$  and  $R_2$  do the same for grid 1. The reduced-range discriminator is identical with the discriminator in Fig. 7.13. For the discriminator we obtain the familiar "S" shaped curve, except that the linear portion extends for a range one-fifth the linear distance of an ordinary discriminator. Following the discriminator is a de-emphasis network leading to the audio amplifiers.

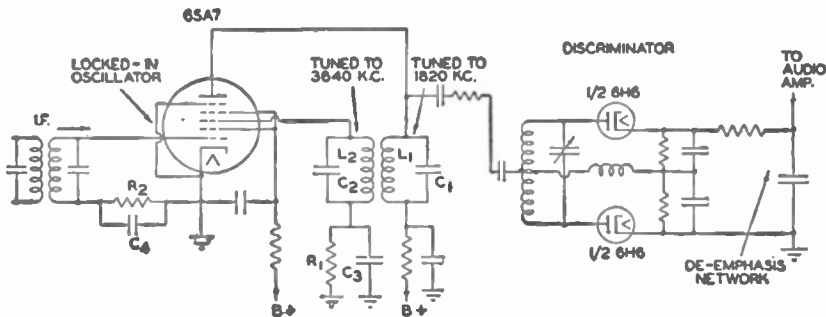


FIG. 7.31. The locked in oscillator and discriminator.

*Operation of Locked-In Oscillator.* Oscillations in any oscillator will, after a few moments, reach a steady state. With grid-leak bias, the oscillating voltage will, on each positive peak, drive the grid positive and cause grid current to flow. The grid-leak condenser becomes charged and establishes an average steady bias across the grid resistor during the continued operation of the oscillator. The grid current flows in pulses, and, since a pulse is actually a highly distorted sine wave, there will be many harmonic frequencies present in the oscillator output. This is true of almost all oscillators employing this form of bias. The higher the harmonic, the smaller its amplitude because the impedance offered to it by the tuned circuit decreases. When the impedance is small, less voltage is developed for feedback and subsequent amplification. However, with care in design, strong harmonics up to the 6th or 8th can be obtained.

In the absence of any incoming signal, the oscillator will produce the strongest output at 1820 kc. This is fed to the discriminator, but since the discriminator is designed to give no output at 1820 kc, no voltage is forwarded to the succeeding audio amplifiers.

Suppose that an I.F. signal at 9100 kc, the 5th harmonic of 1820 kc, is received. Present also in the oscillator is the 4th harmonic of 1820 kc, or 7280 kc. The fourth harmonic will beat with the incoming 9100 kc to produce a difference frequency of 1820 kc in the plate circuit. Due to the coupling between  $L_1$  and  $L_2$ , the difference frequency voltage will appear in the oscillator to lock-in with the incoming signal. When two signals lock-in, a fixed, unchanging relationship is maintained between them. Thus, for example, in a television receiver, the incoming synchronizing pulses trigger oscillators at the same point in each cycle. The frequency of the oscillators is kept fixed by the pulses.

When the incoming signal is unmodulated, its I.F. is exactly at 9100 kc. The output from the locked-in oscillator, under this condition, is 1820 kc. However, since 1820 kc represents the center frequency of the discriminator, no audio output is obtained. It is only when the oscillator shifts above and below 1820 kc that an audio voltage will appear at the discriminator output terminals.

Consider, now, an incoming signal at say, 9130 kc. When this voltage mixes with the 4th harmonic of 1820 kc, we obtain a difference frequency which is greater than 1820 kc. The effect of the latter voltage is to introduce an out-of-phase component into the oscillator-plate tuned circuit. As a result, the out-of-phase current will either aid or oppose the reactive current already flowing in the circuit. For example, if the incoming frequency has increased, the current introduced into  $L_1C_1$  due to the mixing of the incoming signal with the fourth harmonic will aid the inductive current already flowing in the circuit. The result is the same as if we had added some additional inductance in parallel with  $L_1$ . As is well known, the total value of inductances in parallel is less than either one. Decreasing the total inductance in any tuned circuit raises its operating frequency. Thus, when the incoming frequency rises, the oscillator frequency also rises. The oscillator is thus kept "locked-in" with the arrival signal.

It is a simple matter to determine the extent of the oscillator increase. If the incoming signal changes from 9100 kc to 9130 kc, then the oscillator increases from 1820 kc to 1826 kc. We know this because the incoming signal mixes with the 4th harmonic of the oscillator to produce the oscillator frequency. One-fifth of a 9130-kc signal is 1826 kc. The fourth harmonic of 1826 is 7304 kc and subtracting 7304 kc from 9130 kc gives us the required 1826 kc at which the oscillator is now functioning.

When the incoming signal drops below 9100 kc, the effect on  $L_1C_1$  is as though the capacitance had increased. The oscillator frequency decreases.  $L_2C_2$  follow increases and decreases of frequency in step with  $L_1C_1$ . In this way the output frequency fed to the discriminator is an exact duplicate of the incoming signal, but with less frequency deviation about the center point. From  $9.1 \text{ mc} \pm 75 \text{ kc}$  we get  $1.82 \text{ mc} \pm 15 \text{ kc}$ .

It may occur to many that actually the 6SA7 functions as a reactance tube. The shifting phase of the plate current causes the frequency of the oscillator to shift. Within limits, the oscillator will follow and lock-in with the incoming signal. However, if the incoming frequency moves too far in either direction, we find that the oscillator snaps out of control and returns rapidly to its natural resonant frequency of 1820 kc. In a sense this is an advantage because it eliminates interference from adjacent channel stations — something which the conventional discriminator is unable to do. In design, care must be taken to make certain that the oscillator will lock-in over the proper range. Such items as the type of tube, the amount of input signal and the loading effect of the discriminator all have a direct bearing on the range of the oscillator lock-in.

The tube must have a high mutual conductance between grid 1 and the plate and a fairly high plate current. It is the purpose of the tube to supply the reactive current which produces the oscillator frequency shift. If a small change in voltage at grid 1 causes a relatively large change in plate current (high  $g_m$ ), then the range over which the oscillator is locked-in will be increased. The important point in the lock-in operation is the production of a fairly large 4th harmonic component to mix with the incoming signal to produce the oscillator frequency. Large pulses of plate current will do this.

In addition, it is helpful if  $L_2C_2$  resonate at twice the frequency of  $L_1C_1$ . If  $L_2C_2$  resonated at the same frequency as  $L_1C_1$ , then the 4th harmonic would be quite small. However, by designing  $L_2C_2$  to resonate to the 2nd harmonic of  $L_1C_1$ , a greater amount of 4th harmonic voltage is produced. The reason for the increase is due to the fact that the 4th harmonic is only twice removed from the resonant frequency of  $L_2C_2$ . If  $L_2C_2$  had been designed for optimum operation at the fundamental, the 4th harmonic would be four times removed and, consequently, less voltage would have developed.

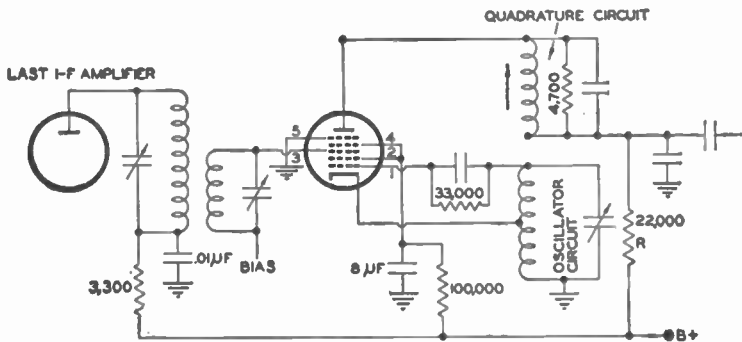


FIG. 7.32. The Philco single-stage F-M detector

**The Philco Single-Stage F-M Detector.** There has recently appeared an F-M receiver which utilizes a locked-in oscillator in a manner somewhat similar to the Beers receiver. However, due to the nature of the feedback between the mixer section of the unit, and the oscillator, no discriminator is required. Instead, the audio output is obtainable directly from the plate-current variations of the mixer tube.

The basic circuit of the F-M detector is shown in Fig. 7.32. The tube, which was specially designed for this purpose, is a heptode. The cathode and the first two grids serve as an electron-coupled oscillator with grid 2 as the plate. Grids 2 and 4 are connected together, but both grids are at I.F. ground potential. The input I.F. signal is fed in at grid 3, whereas the feedback or quadrature circuit is connected to the plate of the tube. Both the oscillator and the quadrature circuit are resonant to the I.F. In



order to insure the proper kind of feedback, the quadrature circuit is damped with a relatively low resistor (4700 ohms) so as to have a bandwidth which is roughly six times that required by the F-M signal.

The oscillator grid is closest to the cathode and it will exert the greatest influence over the flow of electrons. The oscillator functions class C and permits electrons to flow past grid 1 only in short pulses. Throughout the remainder of the cycle the grid is biased beyond cut-off. The second controlling element in the tube is grid 3. Its voltage will determine how much of the pulse passed by the first control grid reaches the plate. There are thus two factors controlling the amount of plate current that flows to the quadrature circuit. As we shall see presently, it is through this dual control that the circuit is able to function.

The basic operation of the circuit depends upon the fact that the pulses of current that the oscillator grid permits to flow is further modified in magnitude by the second (or signal) control grid 3. If, at some instant, the potential of grid 3 is more positive than its normal value, an increased plate current will flow through the quadrature coil. Conversely, if grid 3 is more negative, less than the average current will flow. Through the intersection of the oscillator and the quadrature coils, the change in current will alter the frequency of the oscillator. From the oscillator frequency variation, F-M demodulation occurs. Essentially, then, lock-in between the oscillator and the frequency variations of the incoming F-M signal must occur. In this respect, the circuit is similar to the Beers receiver. However, here the similarity ends.

The circuit consists of three main sections: the oscillator, the quadrature circuit, and the second control grid (No. 3) where the incoming signal is applied. The oscillator is a Hartley, designed to generate a frequency equal to the center frequency of the I.F. band. This is in the absence of any signal voltage. However, when a signal appears (assume for the moment that it is not modulated), the oscillator adjusts itself to lock-in with the signal. Throughout the lock-in range, the frequency of the oscillator and the signal will be identical.

The lock-in of the oscillator is brought about by the transfer

of energy (coupling) between the oscillator coil and the quadrature coil. The quadrature coil is designed so that its impedance does not appreciably change over the range of frequencies of the incoming signal. To insure this, the coil is made to possess a bandwidth approximately six times that which is actually required. This is the reason for the low-valued damping resistor (4700 ohms).

The quadrature coil is inductively coupled to the oscillator and any change in plate current in the quadrature circuit is reflected immediately into the oscillator. The reflected impedance consists of two parts, a resistive component and a reactive component. The resistive loading remains constant because of the low-valued resistor. The reactive component, however, will vary directly with the plate current. It is this reflected component which causes the oscillator frequency to vary. It is important that only the reactive feedback vary; otherwise, the amplitude of the oscillator signal will fluctuate producing an additional A-M component which is not part of the original signal.

The incoming signal controls the flow of plate current and this, in turn, because of the quadrature circuit, affects the oscillator frequency and causes it to lock-in with the incoming signal. The oscillator frequency is thus kept the same as the signal.

The control of the plate current by the signal grid (No. 3) is illustrated in Fig. 7.33. In A of Fig. 7.33, the incoming signal has the same frequency as the oscillator and the pulse of current passes through the tube when the signal grid is going through zero. Under this condition the two frequencies are locked together and the average plate current (of the pulses) is such as to maintain this condition. If, upon application of the signal, the exact phase relationship between the signal and the oscillator pulses is not as shown in Fig. 7.33A, then the average value of the plate current will change sufficiently to swing the oscillator into line. Thereafter, the two are locked in synchronization and the situation shown in Fig. 7.33A prevails.

When the signal frequency changes frequency — with modulation — the phase of the signal voltage (with respect to the pulses of plate current) changes, either increasing or decreasing the average value of the current. This change, reflected in the oscil-

lator circuit, appropriately increases or decreases its frequency to the new value. When the new frequency is attained, the phase between signal voltage and oscillator is again as shown in Fig. 7.33A.

Now the question is: How do these variations in oscillator frequency bring about demodulation of an F-M signal? Simply this — the average value of the current over any cycle is directly proportional to the peak value of the fundamental component of the pulse. (A pulse will contain a strong fundamental and many

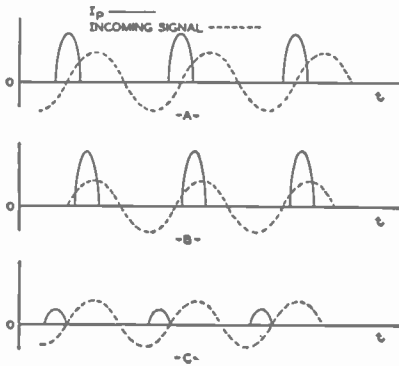


FIG. 7.33. Interaction between the signal voltage and the oscillator pulses to regulate the oscillator frequency. In (A) the two signals are identical; in (B) and (C) the signal frequency is either slightly higher or lower than the oscillator.

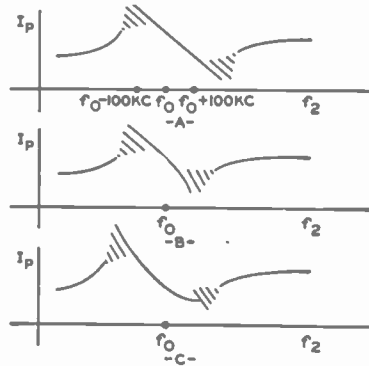


FIG. 7.34. (A) The proper response characteristics from the demodulator. (B) and (C) are the result of an improper adjustment.

weaker harmonics. The fundamental, of course, is the oscillator frequency.) Now, if the signal frequency rises, so will the oscillator frequency and the average value of the plate current will change. In this circuit, the current decreases. See Fig. 7.34. Conversely, if the signal frequency decreases, the oscillator frequency will decrease. This time Fig. 7.34 indicates that the plate current rises. Thus, plate current will vary with frequency and the circuit is designed so that the current varies in a linear manner (see Fig. 7.34). By placing a resistor in the plate lead,  $R$  of Fig. 7.32, we can obtain an audio voltage to feed the audio amplifiers that follow.

Note that the entire action will prevail only so long as the

oscillator is locked-in with the incoming signal. If the signal voltage is too small (below one-half volt for full 75-kc deviation) or the signal frequency shift too large, then the oscillator will slip out of control and return to its free-running frequency. This is shown in Fig. 7.34 by the series of broken dashes at the extremes of the discriminator response curve. An advantage of the requirement of a minimum voltage needed for control of the oscillator is the elimination of interstation noise. This is common with the conventional limiter discriminator arrangement. Its disadvantage is the loss of control that may occur with weak signals. However, weak signals tend to be noisy (even with F-M) and therefore the loss may not be too disadvantageous. With sufficient strength, amplitude variations in the signal do not cause change in the average plate current. What happens is that with a change in signal amplitude, the oscillator phase readjusts itself slightly to maintain the plate current required for that frequency.

Adjustment of the circuit consists in aligning the quadrature circuit for a linear response characteristic, Fig. 7.34A. If the circuit is detuned, the nonlinear curves of Figs. 7.34B and 7.34C are obtained.

## CHAPTER 8

### 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 this 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. 8.1. 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

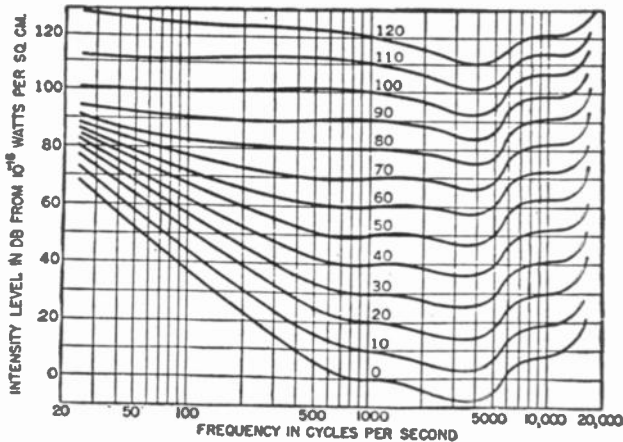


FIG. 8.1. Frequency and loudness characteristic curves of the ears.

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. 8.1 tend to be fairly flat.

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 levels. Beyond this, however, very little has been done and the listener is obliged to resort to his tone controls for further adjustments.

The dependence between volume and the frequency response is indicated more clearly in Fig. 8.2. 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 using people of

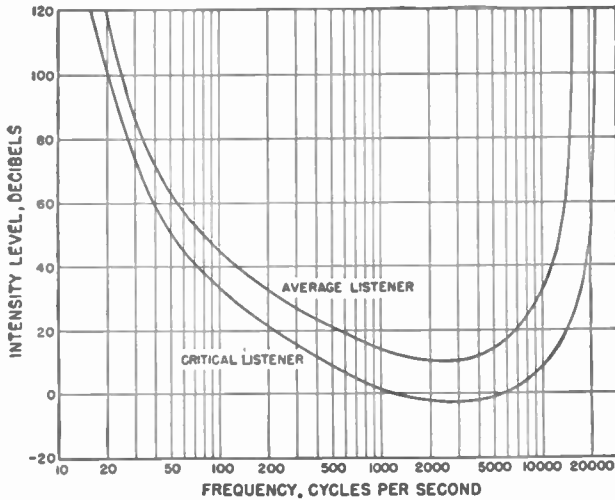


FIG. 8.2. A comparison between the hearing ability of average and critical listeners.

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.

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

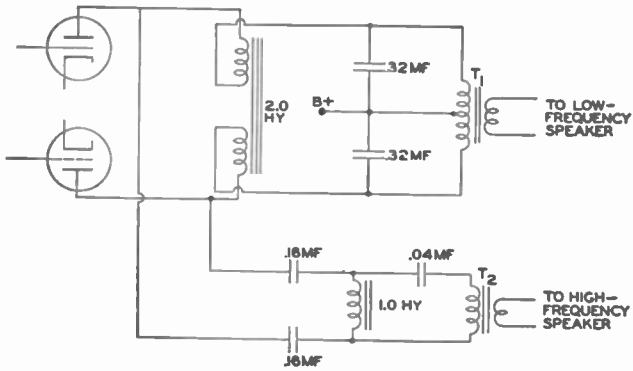


FIG. 8.3. A frequency-dividing network designed to feed two speakers.

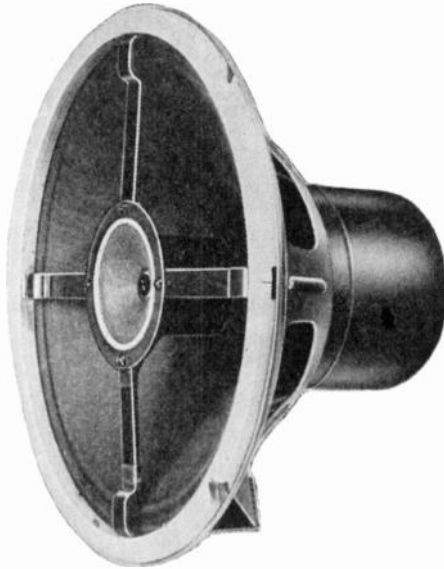


FIG. 8.4. A coaxial type of speaker.  
(Courtesy of Jensen.)



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 Loud-Speaker.* As the translator 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 4000–5000 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 one solution to the problem of efficient reproduction has been through the use of two speakers. A frequency-dividing network, shown in Fig. 8.3, 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 recent innovation has been the use of a coaxial speaker, where one small cone is positioned at the center of a much larger cone (see Fig. 8.4). 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 coverage of 30 cycles 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 frequen-

cies. 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

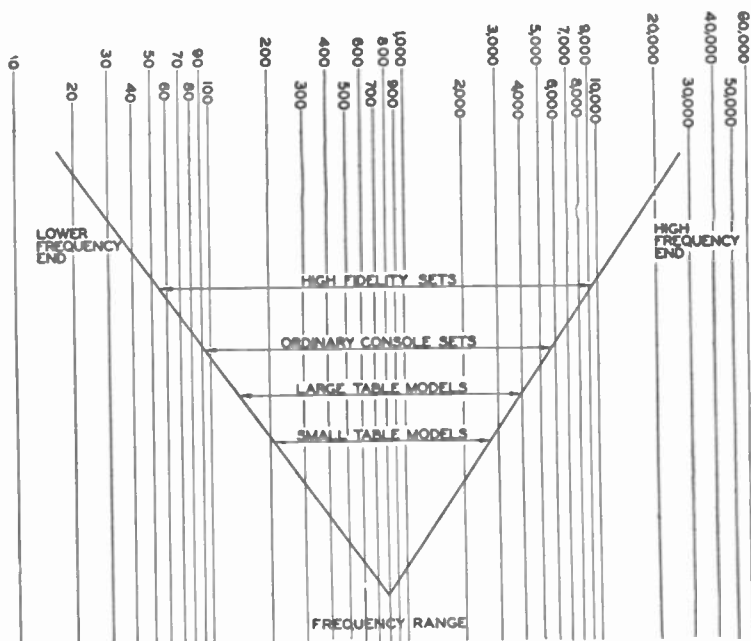


FIG. 8.5. Desirable frequency ranges for various types of receivers in order to obtain the proper tonal balance.

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. 8.5 shows how the limits vary from a small table 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 loud-speaker 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 center of the carrier (and 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 due to 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 band-pass 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 loud-speaker. 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 further accentuate hum. In a balanced push-pull amplifier, a-c ripple from the power supply is eliminated in the plate transformer.

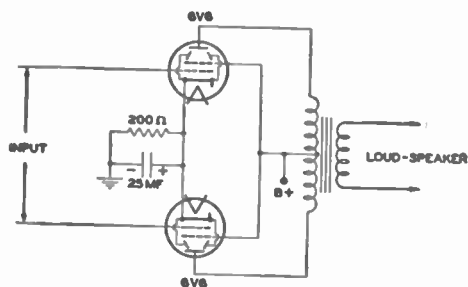


FIG. 8.6. A push-pull amplifier.

Ordinary transformers, however, have a poor frequency response, and compensated transformers are quite costly. The best solution

If we use a push-pull stage, such as the typical unit shown in Fig. 8.6, then the input voltages to each tube must be  $180^\circ$  out-of-phase with each other. The simplest, although not the cheapest, means of accomplishing this is by using input transformers.

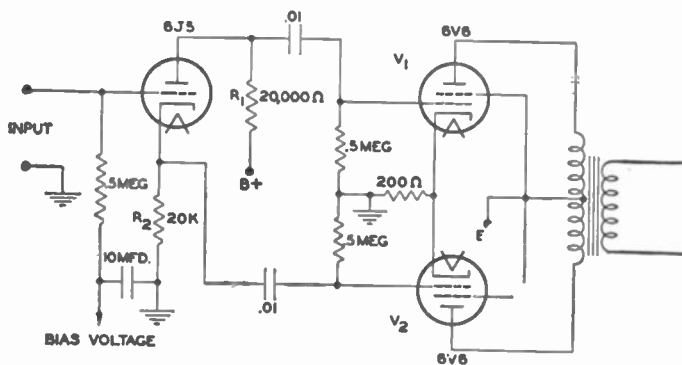


FIG. 8.7. Phase inversion supplied by a single triode.

is the use of a phase inverter whereby a portion of the signal is taken and inverted so that it becomes  $180^\circ$  out-of-phase with its input signal. This makes available two properly phased signals for the push-pull input.

Present phase inverters fall into two general classes. The simplest arrangement is shown in Fig. 8.7, where a single triode

(6J5, 76, 6C5) 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  $180^\circ$  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. 8.8

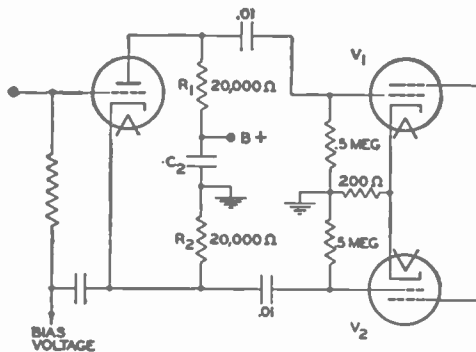


FIG. 8.8. Fig. 8.7 rearranged to emphasize the balance of voltage across  $R_1$  and  $R_2$ .

the 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. 8.9. In Fig. 8.9A 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  $R_2$  and,

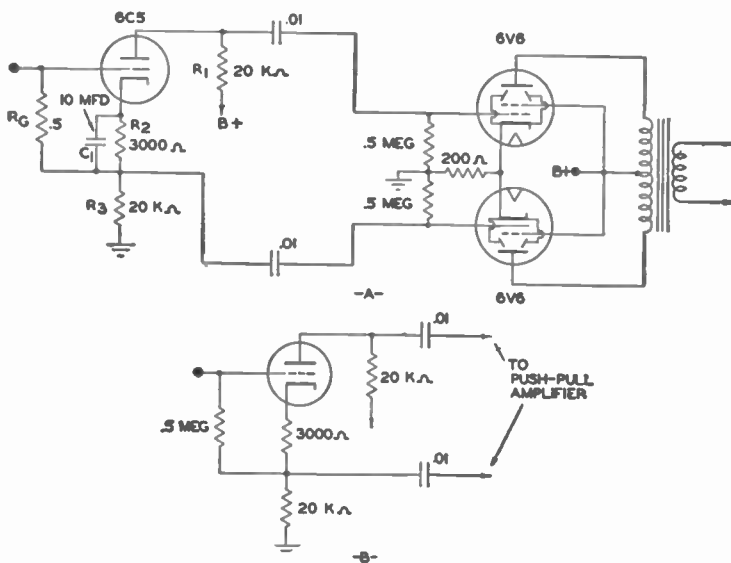


FIG. 8.9. Two alternate phase-inverter circuits.

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 due to the removal of the degeneration.

In Fig. 8.9B, 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.

The second basic phase inverter circuit is shown in Fig. 8.10.

Here, either  $V_1$  or  $V_2$  may be separate tubes or else each may be part of a duo-triode (6C8, 6F8, 6N7, 6SC7). 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  $180^\circ$ , we have the output voltage of  $V_2$  differing from the output voltage of  $V_1$  by  $180^\circ$ . 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

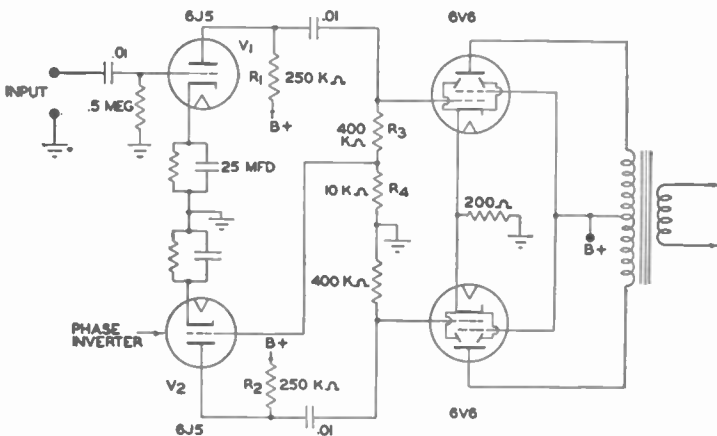


FIG. 8.10. Phase inversion using a separate tube.

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.

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 therefor), some unbalance will always be present.

Partial compensation for unbalance is obtained by the circuit of Fig. 8.11. A change has been introduced in the grid circuit of the

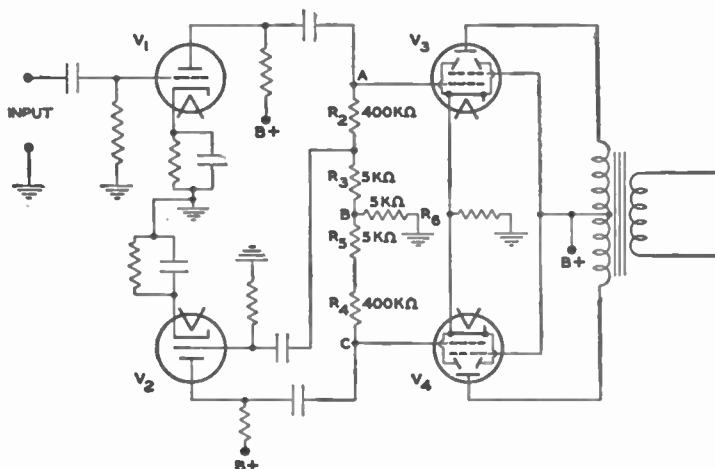


FIG. 8.11. A modified phase inverter to produce automatic balancing.

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  $V_3$  appears between points  $A$  and  $B$ ; for  $V_4$ , between points  $A$  and  $B$ ; for  $V_4$ , between points  $B$  and  $C$ . The voltage present at  $R_6$ , under normal conditions, is zero because the output 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 due to the greater 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 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. Within limits, the common resistor  $R_6$  tends to maintain a balanced circuit, although at a small sacrifice in overall gain.

**Negative Feedback.** The beneficial effects of feeding back some voltage which is out of phase with the input voltage were mentioned briefly above in connection with phase inverters. Actually in the case of Fig. 8.7, 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. 8.12 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

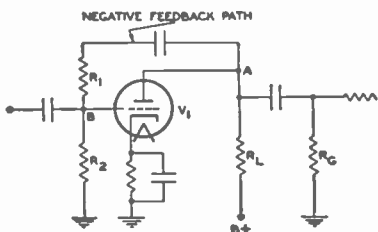


FIG. 8.12. A simple negative feedback circuit.  $R_1$  and  $R_2$  form a voltage-divider network across  $R_L$ .

fed back is in phase opposition to the incoming voltage appearing across  $R_2$ .

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 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 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 distortion. However, because the voltage fed back from point  $A$  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. 8.12 is determined

by the ratio of  $R_2$  to  $R_1 + R_2$ . (The two resistors 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.

One very simple form of negative feedback is obtained by removing the cathode by-pass condenser (see Fig. 8.13). The

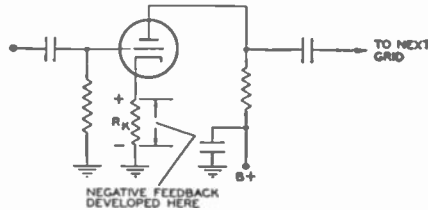


FIG. 8.13. Negative feedback obtained from an unby-passed cathode resistor.

a-c portion of the plate current must now flow through the cathode resistor  $R_k$  instead of being by-passed. This means that 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. 8.14. 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.

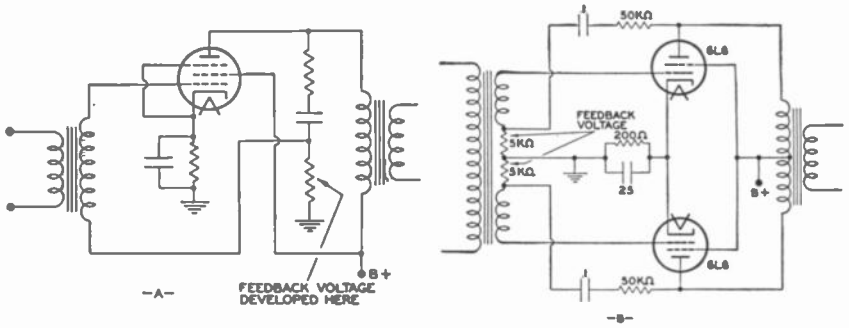


FIG. 8.14. Two additional methods for obtaining negative feedback.

**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, most manufacturers provide an adjustment which will accomplish this 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 actually involves a boosting of the frequencies that we desire to accentuate.

*Simple Bass-Boost Control.* A simple tone control that is found in many sets is shown in Fig. 8.15. It is placed in the plate circuit of one of the audio amplifiers and functions through its ability to

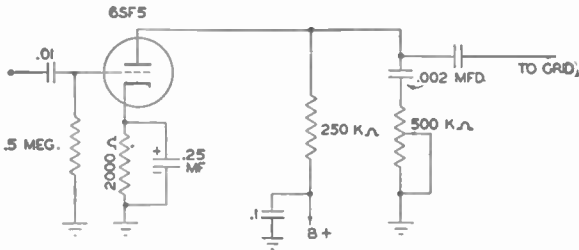


FIG. 8.15. A simple bass-boost tone control. The variation of control is accomplished with the 500,000-ohm potentiometer, but the entire tone control circuit includes also the 0.002-mf condenser.

decrease the relative strength of the high frequencies that are permitted to reach the speaker. When the center arm of the resistor is at the end nearest the condenser, the decrease of the high frequencies is maximum because only the condenser 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.

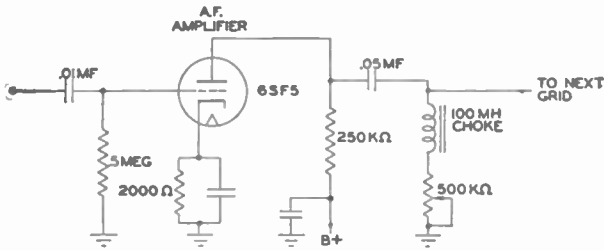


FIG. 8.16. A simple treble tone control.

*Treble Control.* The opposite effect can be achieved by replacing the series condenser of Fig. 8.15 with a series choke coil, as shown in Fig. 8.16. 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. This position 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

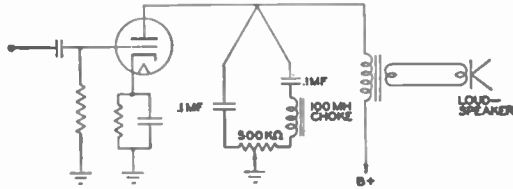


FIG. 8.17. A combination "treble-bass" tone control.

flexible control, we have the arrangement of Fig. 8.17. This unit is particularly useful at low 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. 8.1 that, as the volume is decreased, the high and low frequencies decrease more rapidly than the middle range frequencies. In the control of Fig.

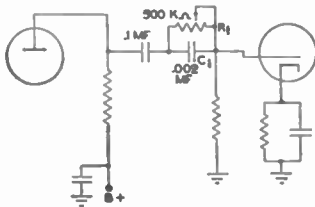


FIG. 8.18A. A series treble tone control.

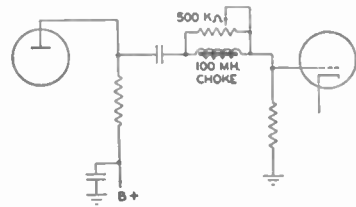


FIG. 8.18B. A series bass-boost tone control.

8.17, the middle frequencies are attenuated more than those at either end of the audio range. The overall result is a more uniform distribution.

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. 8.18A is possible. The low frequencies lose more voltage across the condenser  $C_1$  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 condenser can be replaced by a choke (see Fig. 8.18B).

*Tone Control by Inverse Feedback.* Tone variation can be also accomplished by inverse or negative feedback methods. Two popular methods are shown in Fig. 8.19. In Fig. 8.19A, the 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

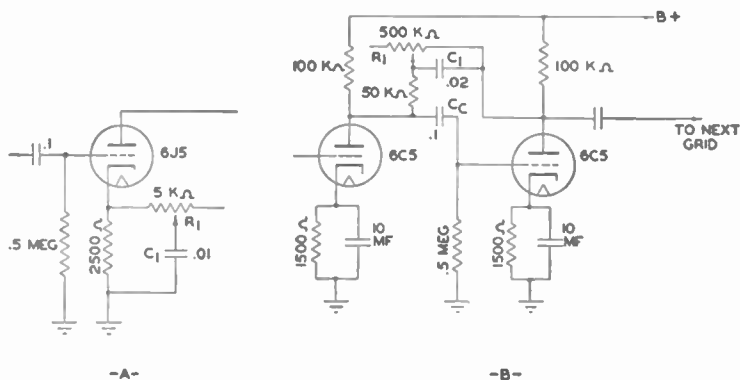


FIG. 8.19. Two common methods of applying negative feedback to obtain tone control.

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 frequencies 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 condenser. 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 loud-speaker. 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. 8.19B. Actually, what we do here is to feed back a voltage from the plate of the second 6C5 to the grid of the same tube. We use the plate-to-plate connection to keep the positive B+ voltage of the 6C5 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

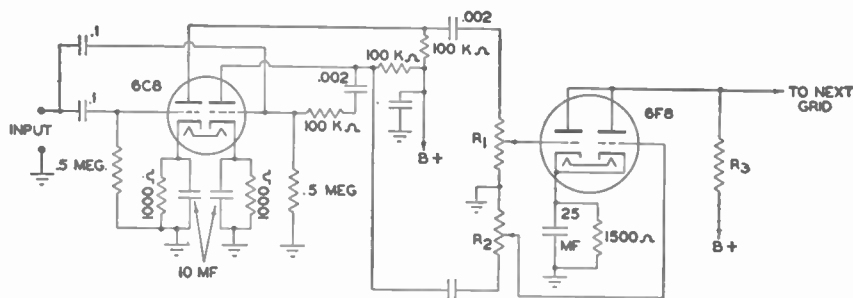


FIG. 8.20. A circuit in which the high or low frequencies may be varied independently of each other.

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 any other frequencies and, hence, a proportionately greater amount of voltage from this range passes through.

All of the preceding tone-control circuits have been so designed that in order to obtain either a high- or low-frequency output response, the opposite range of frequencies were attenuated. The circuits have the advantages of economy in construction and simplicity of design. Much more complex are the booster arrangements whereby a definite boost or increase is given to the section of



frequencies that we desire to accentuate. A typical circuit is shown in Fig. 8.20. The separate triodes of the 6C8 receive the input signal equally. However, in the second triode section, high-frequency negative feedback from the plate to the grid causes the signal that reaches  $R_2$  to be predominantly low in frequency. The 0.002-mf condenser in the plate circuit of the first triode of the 6C8 attenuates the low frequencies. Hence,  $R_1$  is the treble control booster, whereas  $R_2$  controls the bass response. By proper manipulation of each of these controls, we can derive almost any desired tone in the loud-speaker. The two tone controls,  $R_1$  and  $R_2$ , vary the amount of voltage that is applied to each triode section of the 6F8. This, in turn, will determine what percentage of the total output voltage contains either the high or the low frequencies. In the plate circuit of the 6F8, a common load resistor,  $R_3$ , recombines the two sections of the signal again.

*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. 8.21 a series resistor and condenser is 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. To offset partially 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.

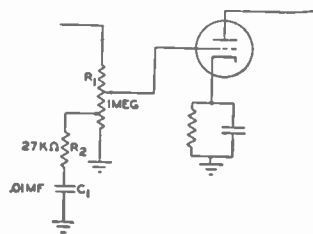


FIG. 8.21. A fixed-tone compensation circuit.

*Speaker Cross-Over Networks.* In an attempt to obtain as high a degree of faithful reproduction as possible, many F-M receivers employ two loud-speakers in place of the customary one. One

speaker is designed to respond especially to the high frequencies, the other to the low frequencies. Recently, there has become available a dual speaker constructed as one unit and known as a coaxial speaker (see Fig. 8.4). The name is derived from the placement of the smaller, high-frequency speaker at the center of the larger, low-frequency unit. Although mounted as one unit, both speakers are entirely distinct of each other.

To transfer the output from the audio amplifier to each of the speakers, a special cross-over network is used. This circuit separates the high and low frequencies and applies them to their respective speakers. The point of separation, or the cross-over point, as it is known, is placed generally near 400 cycles. From the common terminals in the network, Fig. 8.3, a low-pass filter admits all frequencies from 30 to 400 cycles, whereas frequencies over 400 cycles are quickly attenuated. On the other side, we have a sharp cut-off, high-pass filter, and frequencies from 400 cycles to 10,000 cycles are directed along this path. The high-pass filter is designed to cut-off quite sharply below 400 cycles in order to prevent the formation of undesirable harmonics. This form of distortion has been found to occur when frequencies below the acoustic cut-off are permitted to reach the speaker.<sup>1</sup>

<sup>1</sup> Paul W. Klipsch, *Electronics*, Nov., 1945, Vol. 18, No. 11, p. 144.

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*SECTION III*  
*F-M Transmitters*



CHAPTER 9  
 COMMERCIAL F-M TRANSMITTERS  
 (PART 1)

To develop frequency-modulated signals, three general methods are currently in use. One employs a reactance tube, another is based on phase modulation and the third method is new, involving a specially designed tube. Each represents a different approach to the problem and unless one is definitely and decisively shown to be superior to the others, all are likely to remain. We shall investigate each in turn, analyzing their basic circuit arrangement and then demonstrating a commercial application.

**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. 9.1. 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.

One crude and elementary method of obtaining a variation in capacitance is by the attachment of a condenser microphone in parallel with the tuning condenser  $C$ . Such an arrangement is

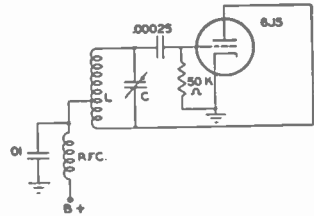


FIG. 9.1. A Hartley oscillator.

shown in Fig. 9.2. 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 capacity of the microphone plus the capacity of the tuning condenser.

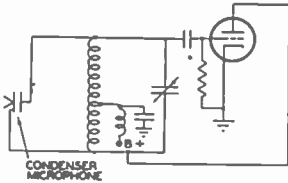


FIG. 9.2. An elementary method of frequency modulating an oscillator.

When the microphone is actuated by sound, the movable condenser plate is forced to vibrate back and forth about its rest or center position. At some time, under pressure of the moving air, the condenser 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 condenser 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 condenser 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 condenser 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 learned early in our study of radio that for many common frequencies the opposition that a resistor offers is constant. But when we insert a condenser or an inductance into the circuit, we note that the opposition depends upon the frequency of the applied voltage. For the condenser, 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, fact. 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 a condenser, inductance or resistance is actually present. In other words, if we can take a tube and so connect it that the voltage between its plate and cathode is  $90^\circ$  lagging the plate current, then it will appear as a condenser 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. 9.3. 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

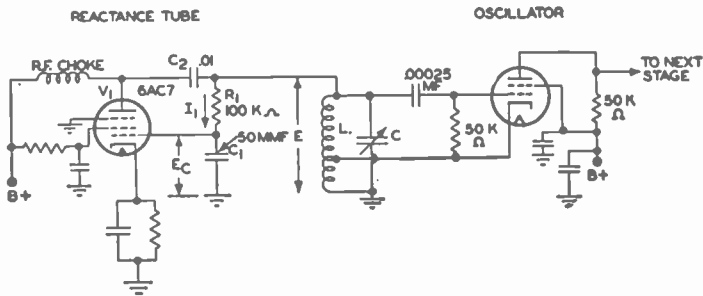


FIG. 9.3. A reactance-tube modulator.

the frequency of an oscillator to vary, we will incorporate an audio signal.

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. 9.3). 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. 9.4A.

Examine the voltage that is developed across the condenser  $C_1$ . We know that with every condenser, the current and the voltage (across that condenser) are out of phase (see Fig. 9.4B). In this instance, since the current through the series branch is practically in phase with the total applied voltage  $E$ , and, at the condenser, the current and  $E_c$  (the condenser voltage) are  $90^\circ$  out of phase, we



can add  $E_c$  to the vector diagram of Fig. 9.4A to obtain the result shown in Fig. 9.4C.  $E_c$  lags the current and is placed behind it.

Returning to Fig. 9.3, 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

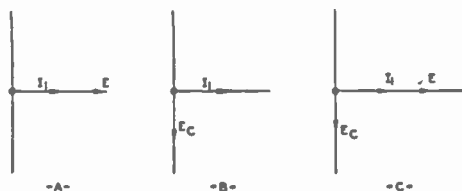


FIG. 9.4. (A) Phase relationship between oscillator tank voltage and current through  $R_1$  and  $C_1$ . (B) Current and voltage phase for a condenser. (C) The phase of  $E_c$  with respect to  $E$  and  $E_1$ .

exist for the grid voltage of the 6AC7. And, since the grid voltage directly controls the plate current, we can derive a relationship between  $E_c$ ,  $E_g$  (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. 9.5A. The plate current, being in phase with  $E_c$ , is placed  $90^\circ$  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

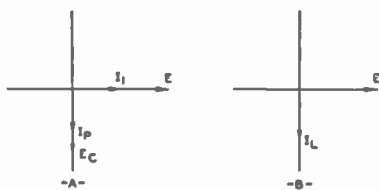


FIG. 9.5. (A) The phase angle between  $E$  and  $I_p$  of the 6AC7. Its resemblance to the phase relationship between  $E$  and  $I_L$  in the oscillator tank circuit (B) is evident.

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 farther. 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. 9.5B.  $E$  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.

Due to the placement of the 6AC7 across the oscillator tank circuit,  $I_p$  must flow through the oscillator coil and add to the

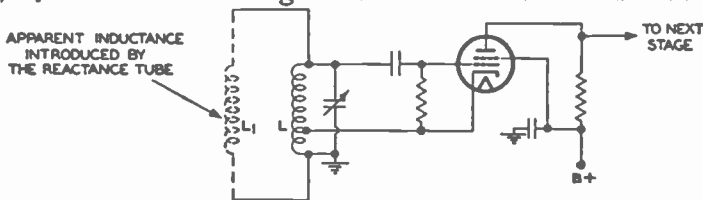


FIG. 9.6. Equivalent circuit of Fig. 9.3.

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. 9.6.

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. 9.7, which is essentially the same as the diagram of Fig. 9.3. 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  $E$ .

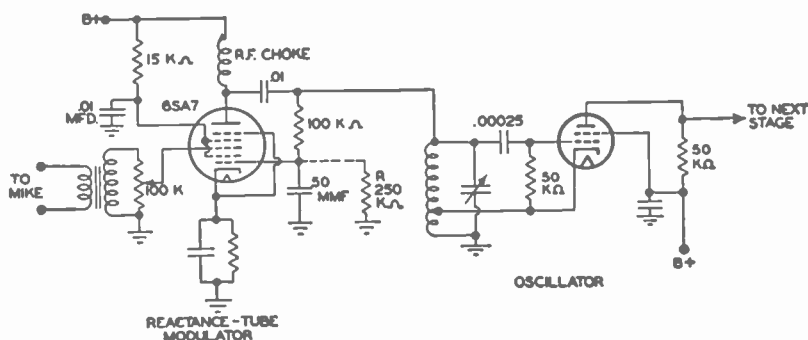


FIG. 9.7. A complete reactance-tube modulator.  $R$  may be connected across the 50-mmf condenser to provide a d-c path to ground for any stray electrons trapped on the grid.

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 6SA7 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 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 condenser 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. 9.8. 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 condenser (3 mmf) in series with a resistor, but their impedance values are reversed. The condenser, because of its very small capacity, pre-

sents a much greater opposition to the flow of current through this branch than the 1500-ohm resistor. Hence, the current will be

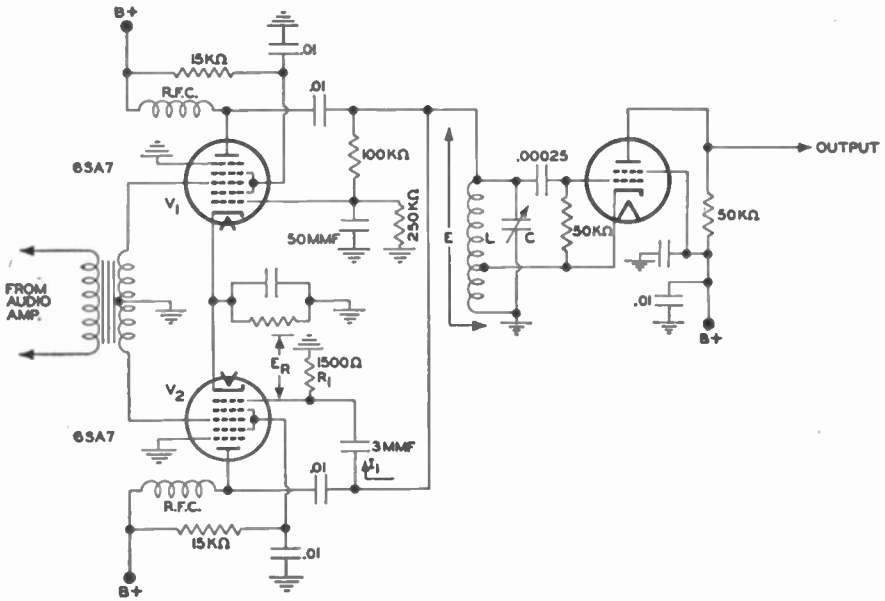


FIG. 9.8. A balanced modulator.

almost entirely capacitive and will lead the applied voltage  $E$  by practically  $90^\circ$ . This is shown vectorially in Fig. 9.9A. The voltage that is developed across  $R_1$  is in phase with the current

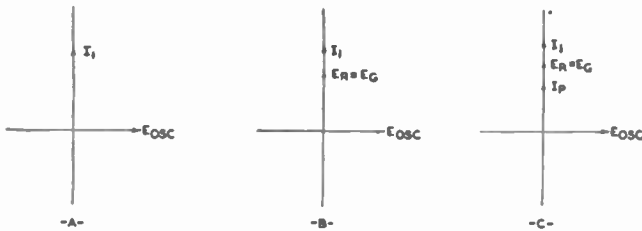


FIG. 9.9. The current and voltage phase relationship in the grid network of  $V_2$  (Fig. 9.8).

because, at all resistors, current and voltage are in phase. We may add  $E_R$  to the vector diagram, as in Fig. 9.9B. 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. 9.9B we can place  $E_o$  at the same point as  $E_R$ .

Proceeding with the analysis, we note that  $I_p$  of  $V_2$ , being governed directly by  $E_o$ , is also in phase with it. Hence, we arrive at a situation where, for  $V_2$ , the plate current is  $90^\circ$  ahead of the applied oscillator voltage  $E$  (see Fig. 9.9C). Thus, by rearranging the condenser-and-resistor combination and altering their relative values we obtain a plate current that leads instead of lags the voltage  $E$ . Then,  $V_2$  appears capacitive.

Within the tank circuit, this capacitive current combines in phase with the capacitive current flowing through the tuning condenser  $C$ . 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 condenser 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 6SA7'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. 9.7.

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.

**The Complete Transmitter.** Our progress, to this point, in the forming of an F-M signal has consisted solely in examining the action is 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 arises. 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. 9.10. 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 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

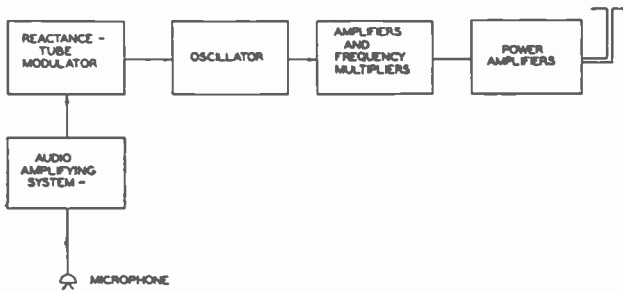


FIG. 9.10. The block diagram of an F-M transmitter using a reactance-tube modulator.

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 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. 9.11. This network is arranged to produce an output voltage from the dis-



crinator 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

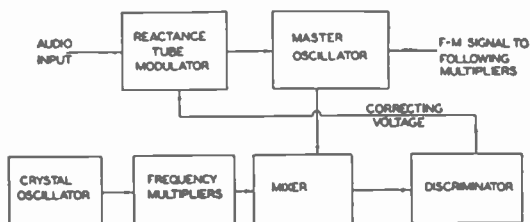


FIG. 9.11. A block diagram of a discriminator control system which keeps the master oscillator at its assigned center frequency.

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 4.7–6 mc. The frequency variation, due to the modulation, is generally near 4.2 kc. This is the initial maximum swing, destined to become  $\pm 75$  kc at the output of the transmitter. Under present frequency allocations, the output carrier is confined to the range 88–108 mc. Hence, we must increase the relatively low oscillator frequency to some value between 88–108 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,  $\pm 4.167$  kc, increased 18 times, gives the maximum  $\pm 75$  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. 9.12, 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

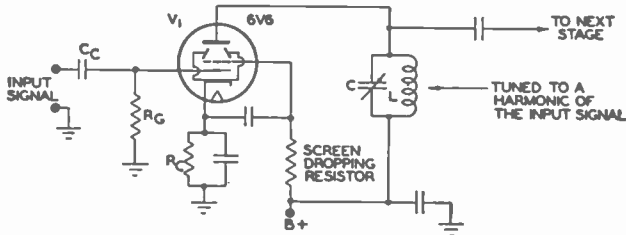


FIG. 9.12. A frequency multiplier. The main grid bias is developed by  $R_g$ ;  $R_o$  is merely for protective bias.

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 fully appreciate 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 actually 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 condenser becomes charged. During the negative portions of the signal the condenser 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. 9.13. With this as the operating point, the incoming voltage will vary above and below this value. At the positive peaks, the

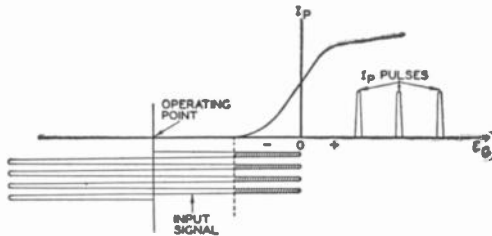


FIG. 9.13. The operating characteristics of a frequency multiplier. Only the shaded portion of the input signal is effective in producing plate current.

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 total grid voltage is more positive than the plate-current cut-off value. This region is shown shaded in Fig. 9.13. 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. 9.12, tuning  $L$  and  $C$  to the third harmonic of the input will produce this component across the circuit.

It may be wondered why 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  $\frac{1}{3}$  or less than the fundamental if  $L$  and  $C$  were tuned to the fundamental frequency. Attempting to reach 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 condenser 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  $\pm 4.167$ -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  $\pm 75$ -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. 9.14, and the corresponding schematic appears in Fig. 9.15.

The push-pull modulators differ from the previous push-pull 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-capacitance 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  $90^\circ$ . If the plate current of one modulator tube was  $90^\circ$  leading, the current of the other tube was  $90^\circ$  lagging.

In the RCA transmitter, the R.F. voltage from the oscillator is transferred to  $L_1$  inductively. The voltage developed across  $L_1$

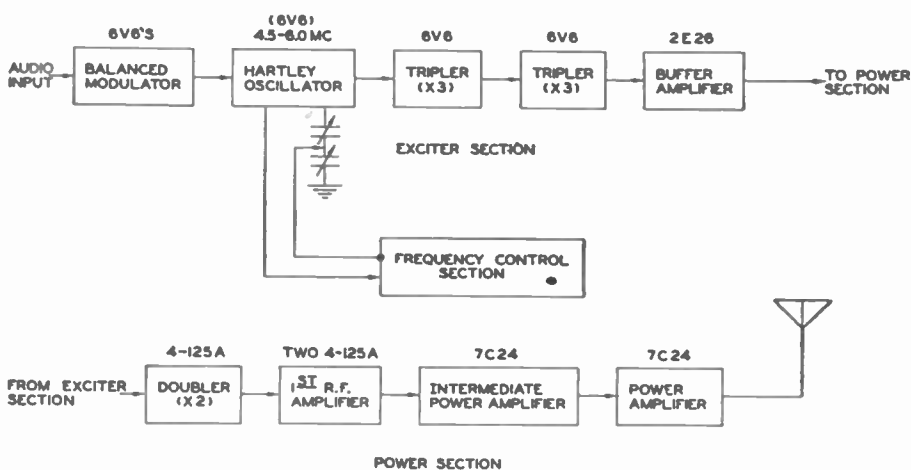


FIG. 9.14. The outline of a modern RCA F-M transmitter. The notation above each block designates the tubes used in that stage.

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 circuit of the two modulator tubes, we note that they form essentially a push-pull circuit. The two inner ends of  $L_3$  and  $L_4$  are placed at the same R.F. potential by means of  $C_5$ .  $C_5$  offers only slight opposition to R.F. currents and thus we may consider  $L_3$  and  $L_4$  as being directly connected.  $C_3$  and  $C_4$  resonate  $L_3$  and  $L_4$  to the same frequency as the oscillator tank circuit. A simplified illustration demonstrating the direct resemblance of this arrangement to a push-pull circuit is shown in Fig. 9.16. Since the grids are so connected, each receives R.F. voltages

that are  $180^\circ$  out of phase with respect 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 —  $L$ ,  $L_3$ , and  $L_4$  — so that the R.F. voltage transferred from the oscillator is shifted  $90^\circ$ . Thus, the voltage is first shifted  $90^\circ$  and then applied to  $L_3$  and  $L_4$  to make each grid  $180^\circ$  out of phase with each other. As a result of this arrangement, each tube presents to the oscillator tank a different form of reactance. However, with no audio input signal at  $V_1$  and  $V_2$ , the push-bell connection of each modulator acts to cancel the reactance produced by each tube at the oscillator tank.

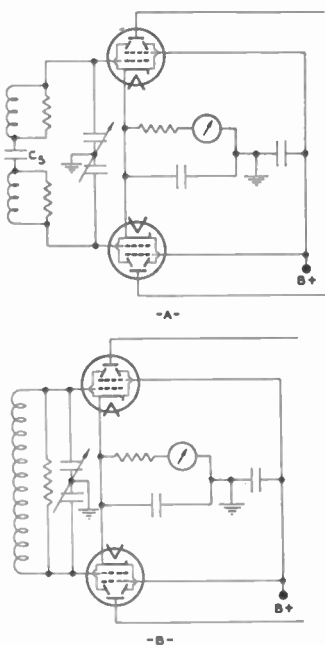


FIG. 9.16. (A) The input circuit of the RCA modulator.

(B) An equivalent diagram.

(inductive or capacitive), thereby producing a frequency shift. The two tubes, in push-pull, 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  $C_6$ . The output coil 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, whereas  $V_6$  is merely an isolating or buffer ampli-

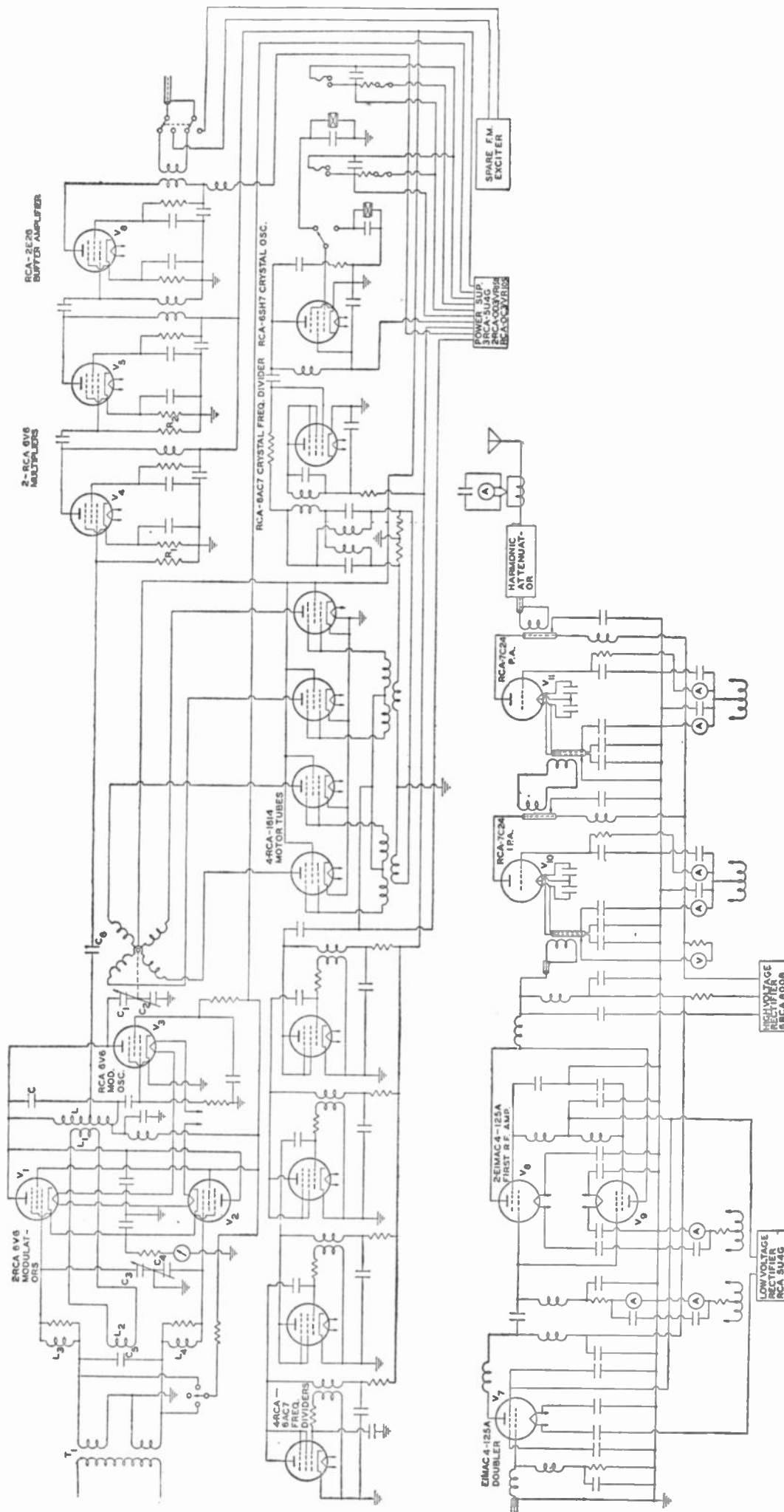


FIG. 9.15. The schematic diagram of one of RCA's newest F-M transmitters.





fer. The layout, to this point, represents the exciter unit and is completely contained within one cabinet (see Fig. 9.17). Included

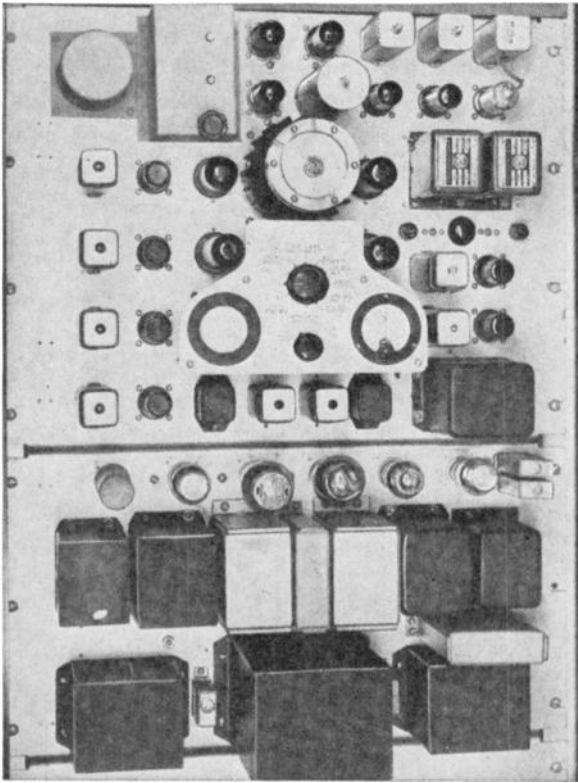


FIG. 9.17. A view of the exciter unit used in RCA F-M transmitter.

also is a spare exciter assembly for emergency use, a power supply and a frequency control unit.

As a preventative measure,  $V_4$  and  $V_5$  have cathode resistors. These are used to protect the tubes in case something should cause the oscillator to cease operating. The two tubes, in ordinary operation, obtain practically their full bias from the grid-leak input resistors,  $R_1$  and  $R_2$ . However, if the oscillator should fail, the voltages across  $R_1$  and  $R_2$  drop to zero. If no other protection existed for the tubes, the zero-biased grids would permit sufficient current to flow and actually cause the plates to become red hot. By

inserting cathode resistors, a bias is automatically developed and serves to prevent excessive heating at the plate.

The cathode resistor of  $V_6$  is actually used for operative bias, since this amplifier, being a buffer amplifier, is operated class A.  $V_4$  and  $V_5$  are biased for class C operation.

The center frequency, at the output of  $V_6$ , is somewhere between 44 and 54 mc. This is now doubled to the final operating frequency by the first stage of the second unit of the transmitter ( $V_7$  is the doubler, using an Eimac 4-125A tube). Beyond this there are two tubes in parallel,  $V_8$  and  $V_9$ , functioning as the first or inter-

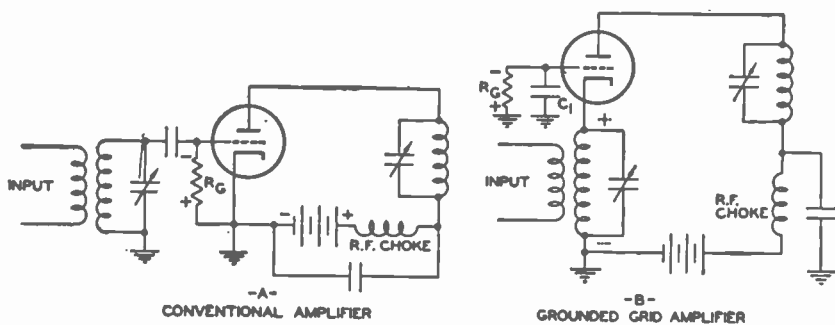


FIG. 9.18. A comparison between conventional and grounded grid amplifiers.

mediate power amplifiers.  $V_{10}$  (an RCA-7C24 tube) is the first or intermediate power amplifier and  $V_{11}$ , another RCA-7C24 tube, is the final power amplifier. From the latter stage, the signal is fed to the antenna system. The complete assembly of tubes will produce a 3-kw signal. By removing the last RCA-7C24 tube and attaching the antenna directly to the output of the intermediate power amplifier, we reduce the output to 1 kw. This arrangement makes for greater flexibility since it permits a station using the lower-powered circuit to increase its output merely by the addition of one stage.

**Grounded-Grid Amplifiers** A close inspection of  $V_{10}$  and  $V_{11}$  will reveal that they are not connected as conventional amplifiers (Fig. 9.18A) but as grounded-grid amplifiers (Fig. 9.18B). In a conventional amplifier, the input signal is fed to the grid of the tube, with the cathode kept at a fixed potential. In a grounded-

grid amplifier, the grid is connected directly to ground (through condenser  $C_1$ ), while the cathode receives the incoming signal. As far as the current flow through the tube is concerned, it makes no difference whether the grid is kept at zero R.F. potential and the cathode voltage is varied, or the cathode is placed at R.F. ground potential and the grid is varied. Tube current is, at all times, subject to the voltage difference between the grid and cathode.

The distinct advantage of a grounded-grid amplifier arises from its excellent operation at high frequencies. By placing the grid at R.F. ground potential, we effectively shield the output and input circuits from each other and thereby minimize undesired feedback and oscillations. In the conventional amplifier, this can only be accomplished by special neutralizing precautions. It is only when the power of a grounded-grid amplifier is above 10 kw that any special neutralizing steps are required.

A disadvantage of a grounded-grid amplifier is the requirement of more driving power than we need for the usual amplifier. The reason for the increase is due to the degeneration that is present in the circuit. When the input signal drives the cathode increasingly negative, its effect is the same as though we had kept the cathode fixed and made the grid more positive. The plate current flow increases, producing a larger current flow through the cathode circuit. A larger voltage drop is developed across the circuit in the cathode leg with a polarity as shown by the dotted signs in Fig. 9.18B. The result, therefore, is to produce a voltage drop which opposes the driving voltage of the signal. Naturally, this opposition reduces the effect of the input voltage. Hence, to obtain the same output as we have with the conventional amplifier, more input is needed in the grounded-grid amplifier to overcome the effect of the degeneration.

Although degeneration necessitates a higher input voltage, it does increase the stability of the amplifier. This is especially desirable at frequencies near 100 mc.

The RCA-7C24 tube, shown in Fig. 9.19, is designed especially for use in high-frequency, grounded-grid amplifiers. It possesses a grid structure that presents a maximum of shielding between plate and filament electrodes. This results in a low plate-to-

filament capacity. The grid terminal is a disk-seal brought out through the glass completely around the tube. When used in connection with an external shield, the input and output circuits of the stage are well isolated from each other.

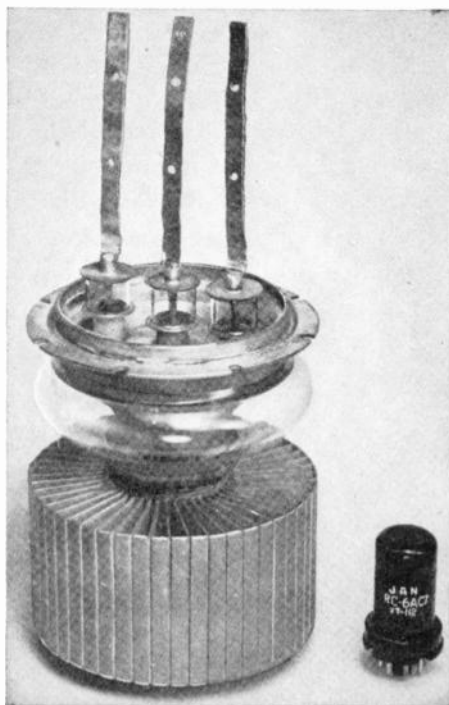


FIG. 9.19. A recently developed tube (RCA-7C24) designed especially for grounded-grid operation.

**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, due to 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.

In recent reactance-tube transmitters, two general types of frequency-control systems have been used. One utilizes motor control, in which a 2-phase motor is attached to one of the oscillator tuning condensers (see Fig. 9.20). Any center-frequency drifting causes the motor to rotate the tuning condenser in a direction to counteract the drift. In the second control system, a discriminator produces a voltage that is fed to the reactance tubes. By varying the grid-bias voltage of the reactance tube, we can also vary its effect on the oscillator and thereby produce a shift in frequency sufficient to recenter the wandering oscillator. Each system is explained in the following paragraphs.

*Motor Control.* In the RCA transmitter shown in Fig. 9.15, condensers  $C_1$  and  $C_2$  are motor-controlled. The condensers are connected across the oscillator tuning coil  $L$  and any rotation of  $C_1$  and  $C_2$  will result in a change in frequency. A block diagram of the frequency correction network and its place in the transmitter is shown in Fig. 9.21. 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 4.5–6.0 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, 88–108 mc, it is necessary to provide the values of 4.5–6.0 mc for the oscillator.) The divider output is then fed to a balanced modulator.

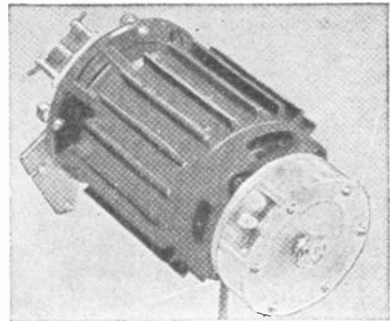
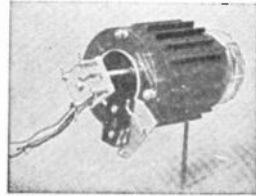


FIG. 9.20. Two views of the frequency-determining capacitor mounted on the shaft of the motor. An oil dampening unit mounted on the front end of the motor shaft, prevents overshooting. (Courtesy of RCA.)

The other input to the modulators is obtained from a highly accurate crystal oscillator. The crystal oscillator frequency is lowered by  $\frac{1}{2}$ , at which value it will be exactly  $1/240$ th of the correct main oscillator frequency. This frequency is then compared

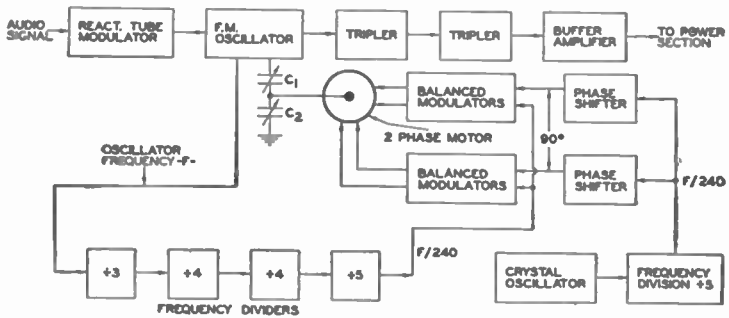


FIG. 9.21. An outline illustrating the several circuits of the frequency control section.

in the balanced modulator with the frequency obtained from the main oscillator. If any difference exists, the motor is actuated and the tuning condensers rotated until the oscillator is again centered at its assigned spot.

This is briefly the operation of the system. Before we consider the complete schematic, it would be advisable to review the operation of several of the units in order to gain a better appreciation of their function in this particular network.

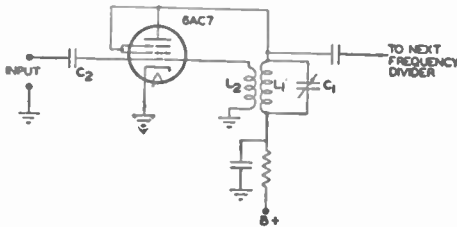


FIG. 9.22. A frequency divider similar to those used in the RCA F-M transmitter.

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 condenser  $C_2$ . Let us say that its value is 6.0 mc. An output one-fourth of this is desired, so the oscillator functions at 1.5 mc. At the tube the 6.0 mc 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. Due to 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 enable it 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. Due to 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. 9.21, 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 mc or 0.25 mc or 25 kc. The reason we require this much frequency division is due to the 2-phase motor, as we shall presently see.

The overall purpose of the frequency divider is to bring to the balanced modulators 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 include the crystal oscillator. After the output of the crystal oscillator is decreased by  $\frac{1}{3}$ th, 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. The comparison of the two frequencies takes place in the balanced modulator. Disregard, for the moment, the phase-shifting network between the crystal oscillator and the balanced modulator.

*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

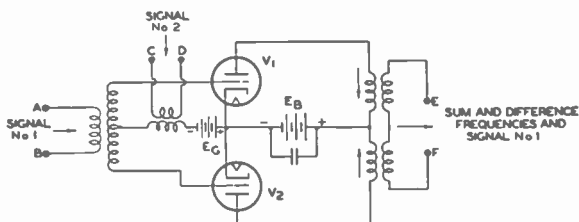


FIG. 9.23. A balanced modulator.

signals at the output. The other input frequency does not appear at the output. In the circuit of Fig. 9.23, 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 and either set of terminals may be used.

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 effects 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 and the result is a complete cancellation of the 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 this 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  $90^\circ$ . 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  $90^\circ$ . 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 condensers. 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.

The schematic diagram of the control system is shown in Fig. 9.15. Included with the crystal oscillator is a heating unit to maintain a constant temperature. The power supply is presented in block form.

*Discriminator Control.* We have studied the operation of a motor-control network. Now, let us see how control of the main oscillator center-frequency may be obtained with a discriminator. In Fig. 9.11, 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 discriminator 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.

A frequency stabilization circuit is shown in Fig. 9.24. 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 6SA7 tube, connected as a mixer.

The 5.0-mc oscillator voltage is picked off by means of a small condenser and brought to grid 1 of the 6SA7. 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

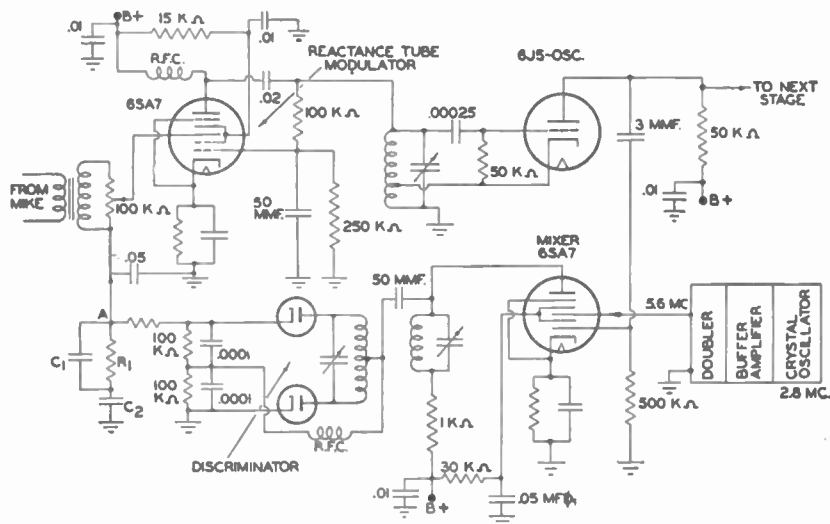


FIG. 9.24. A discriminator frequency-stabilization circuit.

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 7.

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 vary with 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. 9.24, 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 cycles per second.

## CHAPTER 10

### COMMERCIAL F-M TRANSMITTERS

#### (PART 2)

**F-M Through Phase Modulation.** It was previously seen that, indirectly, frequency modulation is produced when a carrier is phase-modulated. Major Armstrong employed this fact when he first produced frequency modulation. Since then, several large manufacturers have designed broadcast transmitters based fundamentally upon the same system. The advantages of the Armstrong system are derived from the control of the transmitter frequency by a crystal oscillator. This provides a high degree of accuracy and eliminates the need for any frequency-controlling network, such as required in transmitters producing F-M by the reactance-tube method. 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 the Armstrong system, as we shall presently see, the F-M is produced beyond the oscillator. Hence, crystal control is feasible.

Before we investigate the operation of the Armstrong 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. 10.1, is made to shift back and forth as it spins around at its fixed frequency. Due to 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. 10.1, and

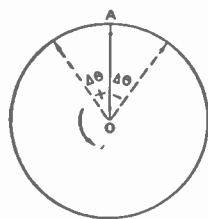


FIG. 10.1. Phase modulator represented by vectors.  $\Delta\theta$  is the maximum phase shift of the carrier.

(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.

Through the use of vectors, we can readily demonstrate the formation of phase modulation. From this, it requires only a simple transition to bridge the gap to the electrical apparatus. However, let us begin with something which is more familiar, namely, amplitude modulation.

**A-M Versus P-M.** Fig. 10.2 illustrates the conventional and the vectorial representation of amplitude modulation. In conventional notation, the low-frequency audio signal at  $A$  varies slowly from  $0^\circ$  to  $360^\circ$ . 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, due to 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.

In direct 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 an other, that is illustrated by the vector in Fig. 10.1. 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  $180^\circ$  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 can be done by adding the audio voltage vectorially at right angles to the carrier voltage or, electrically, by combining them  $90^\circ$  out of phase with each other.

Let us examine this critically. In Fig. 10.3A, the unmodulated carrier and the audio-modulating voltage are shown separately.

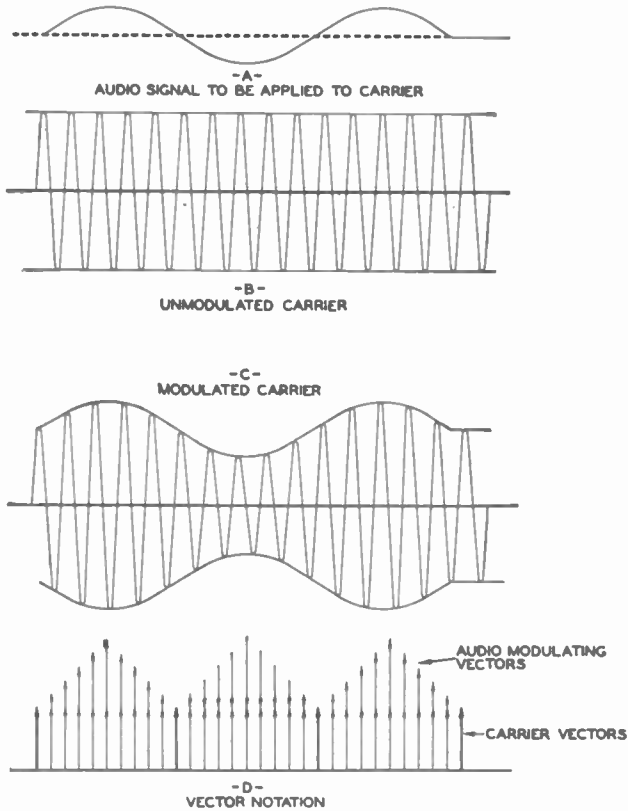


FIG. 10.2. A comparison of the conventional and vector notation for amplitude modulation.

If we combine them  $90^\circ$  out of phase as in Fig. 10.3B, 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 ampli-

tude 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-modulating voltage cycle is shown in Fig. 10.4. 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.

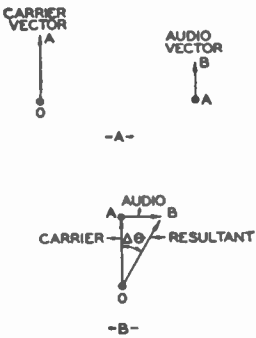


FIG. 10.3. (A) The separate audio and carrier vector ( $OA$  and  $OB$ ) and their vector addition to produce P-M. (B)  $\Delta\theta$  is the phase difference between the unmodulated carrier and the modulated resultant.

resultant vector (or the modulated vector) 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

The formation of frequency modulation, or the variation in frequency in the phase-modulated wave, is clearly indicated in Fig. 10.4. 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 vector) 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

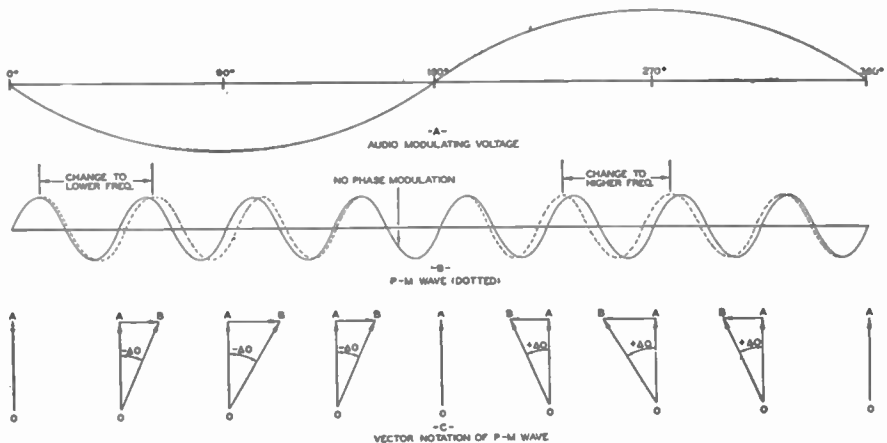


FIG. 10.4. Illustrations of phase modulation.



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 and no equivalent frequency formation.

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 ( $360^\circ$ ), the modulation disappears and the frequency of the carrier is once more at its central value.

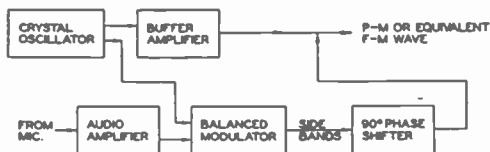


FIG. 10.5. The basic Armstrong system for generating F-M waves.

**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. 10.4. However, we obtain the same results. We start with a fixed source of high frequency voltage, a crystal oscillator (see Fig. 10.5). The output from the 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 due to the operation of the balanced modulator. The sidebands are then shifted in phase by  $90^\circ$  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  $90^\circ$  and combined with the carrier from the buffer amplifier.

Let us pause for a moment and compare the Armstrong method with the previous system for obtaining phase modulation. In one instance we apply the audio-modulating voltage directly to the carrier (maintaining  $90^\circ$  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's 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  $30^\circ$  or less. Expressed in radians (one radian is equal to  $57.3^\circ$ )  $30^\circ$  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  $30^\circ$  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. 10.6. Further, each sideband will be  $90^\circ$  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  $90^\circ$  and combined with the remaining half of the carrier (which was fed into the buffer amplifier, Fig. 10.5) to produce the phase or frequency modulation.

To recapitulate, the Armstrong method can be used because

1. A carrier, phase-modulated at angles of  $30^\circ$  or less, will produce only two sidebands.

2. Each sideband is  $90^\circ$  displaced from the carrier.

In the Armstrong system we take (a) one carrier and (b) two sidebands and combine them  $90^\circ$  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. 10.5. The equivalent schematic is shown in Fig. 10.7. 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 positive. At the same time, the screen grid of  $V_4$  is negative since it is attached to the opposite side of the transformer secondary winding. Due to the 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

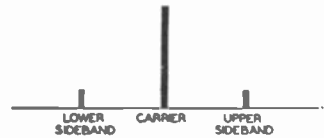


FIG. 10.6. A low value of phase shift produces only two sidebands.

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  $L_3$ .

On the next audio half cycle, when the screen grid of  $V_3$  is going negative and the screen grid of  $V_4$  is driven positive, the opposite

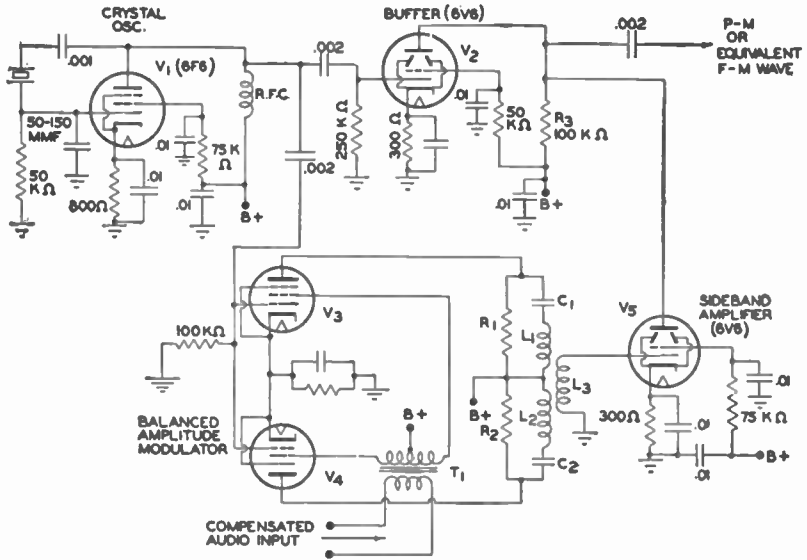


FIG. 10.7. The schematic diagram of the basic Armstrong modulation.

set of conditions prevail. 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 produce a  $90^\circ$  phase shift. This is accomplished through the insertion of condensers  $C_1$  and  $C_2$  in series with  $L_1$  and  $L_2$ . These condensers 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 voltage 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 occurs when the field is going from positive to negative (or negative to positive) values. On the other hand, the induced voltage is minimum or zero when the flux is hardly changing at all. This occurs at the positive and negative peaks (see Fig. 10.8). If we plot the relationships of induced voltage to magnetic flux variations, we immediately see that a phase difference of  $90^\circ$  is intro-

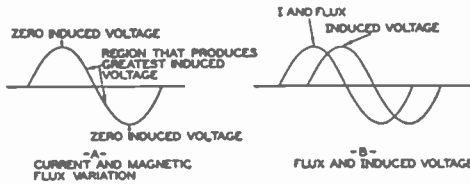


FIG. 10.8. The relationship between a magnetic field and the induced voltage.

duced. In this manner we produce the necessary sidebands and shift them  $90^\circ$  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 take place.

Frequency Swing Compensation. Before we turn our attention to the succeeding frequency multipliers, it is first necessary to ascertain how great a frequency shift we obtain from the combination of sidebands and carrier. We know that the equivalent F-M derived from phase modulation is based on the equation

$$\Delta = f \cdot \Delta\theta$$

where  $f$  = frequency of audio modulating voltage  
 $\Delta\theta$  = maximum phase shift.

Actually, in the above 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  $57.3^\circ$ ;  $f$  is expressed in cycles.

The maximum phase swing is restricted to  $30^\circ$  in commercial equipment. At the highest audio frequency, utilizing the full phase swing of  $30^\circ$  (0.524 radians), 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.

To achieve this, a predistorter circuit is inserted in the audio amplifier stages (see Fig. 10.9). The resistor  $R$  is made very large in comparison to the reactance of condenser  $C$  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 condenser 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.

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 (which is  $\pm 26.2$  cycles) will determine how much multiplication is to be applied to obtain the final  $\pm 75$  kc.

Let us return to the circuit of Fig. 10.7. Utilizing the maximum phase shift of  $\pm 30^\circ$ , we discovered that, at 50 cycles, the carrier has a frequency variation of  $\pm 26.2$  cycles. The carrier frequency, at this point in the circuit, is the crystal oscillator frequency, or 200 kc. Hence, our F-M signal is 200 kc with a frequency swing of  $\pm 26.2$  cycles. What we desire, at the antenna, is the assigned station frequency, say, 90 mc  $\pm 75$  kc. The problem is to increase the 200 kc  $\pm 26.2$  cycles to 90 mc  $\pm 75$  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.

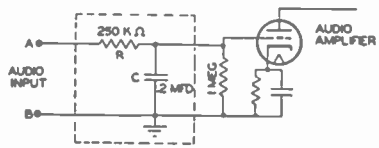


FIG. 10.9. A predistorter network.

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  $\pm 75$  kc for the sidebands.

A method to accomplish this is shown in Fig. 10.10. The entire transmitter is arranged in block diagram, with each unit numbered

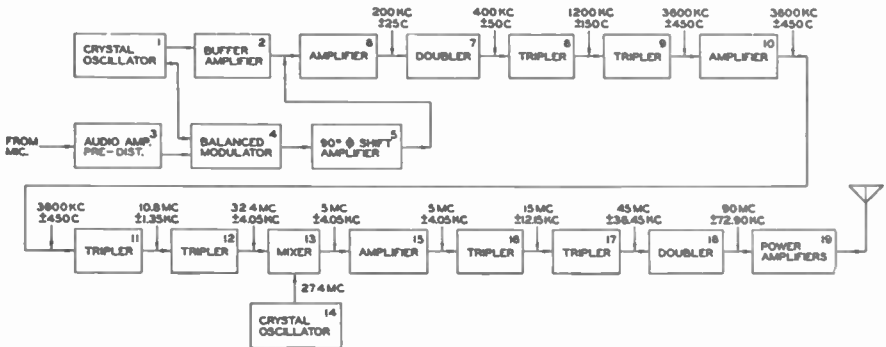


FIG. 10.10. A complete block diagram of the basic Armstrong F-M transmitter.

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  $\pm 4.05$  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.

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  $\pm 4.05$  kc. The range, then, of the input signal is from 32.39595 mc ( $32.4$  mc  $- 4.05$  kc) to 32.40405 mc ( $32.4$  mc  $+ 4.05$  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:

32.39595	32.40405
<u>-27.40000</u>	<u>-27.40000</u>
4.99595	5.00405

or, what is the same thing,  $5$  mc  $\pm 4.05$  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  $\pm 72.9$  kc. By assuming a starting frequency variation of 25 cycles, we obtain only a  $\pm 72.9$ -kc frequency shift at the output. However, with the 26.2 cycles proposed originally, we would come much closer to  $\pm 75$  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 predistorter, 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 de-emphasis circuit returns the higher frequencies to their proper level. A commercial audio-amplifier network, containing both pre-emphasis and audio correction, is shown in Fig. 10.14.

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.

**The Dual Channel Armstrong System.** An improvement over the preceding system with respect to frequency stability and distortion has been achieved through the design of 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  $90^\circ$  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. 10.11. 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. 10.11B). 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  $90^\circ$  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 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 condensers 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 practically  $90^\circ$  ahead in phase, with respect to the voltage across  $L_1$ . Since the voltages

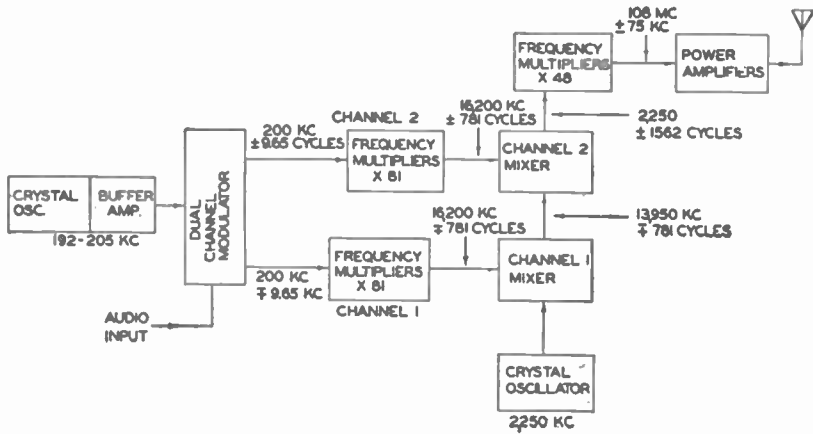


FIG. 10.11A. A block diagram of the dual channel modulator circuit.

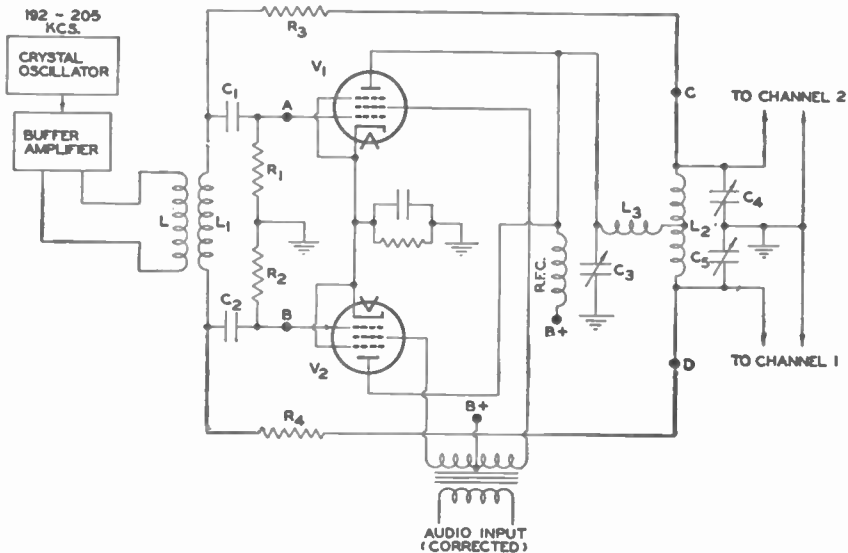


FIG. 10.11B. The modulating circuit of the dual channel F-M transmitter. (Courtesy Communications Magazine.)

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  $90^\circ$ . The required  $90^\circ$  phase shift has thus been acquired. Now to the production of the sidebands.

The grids of  $V_1$  and  $V_2$  are excited  $180^\circ$  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  $C_5$  form 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  $90^\circ$ -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. 10.11B have been rearranged as shown in Fig. 10.12. 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  $L_2$ .

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. 10.11B. 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. 10.12, 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. On the other hand, 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 fre-

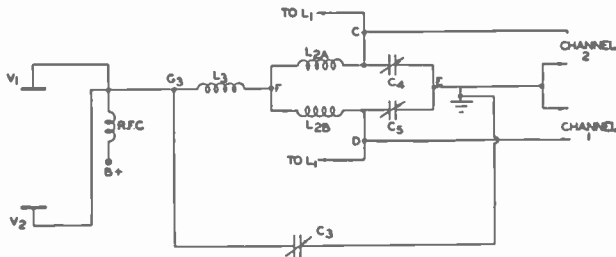


FIG. 10.12. The electrical network wherein the sideband and carrier voltage combine.

quency. 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. 10.12 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  $90^\circ$ . 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 appear 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 condenser, the voltage lags the current by  $90^\circ$ . Hence, the voltage that appears across  $C_4$  and  $C_5$  is  $90^\circ$  (due to its flow through  $L_3$ ) plus an additional  $90^\circ$  (due to the voltage and current relationships across a condenser) out of phase with respect to the sideband voltages applied by the modulator to terminal  $G$  and ground. And, due to the modulator input network,  $C_1$ ,  $C_2$ ,  $R_1$ , and  $R_2$ , another  $90^\circ$  shift was introduced. The overall result is this: the sideband voltages appearing across  $C_4$  and  $C_5$  are  $180^\circ + 90^\circ$ , or  $270^\circ$ , out of phase with the carrier voltages at  $C_4$  and  $C_5$ . But  $270^\circ$  in one direction (say, clockwise) is equal to  $90^\circ$  in the opposite (or counterclockwise) direction. Thus, the sideband frequencies and the carrier are combined  $90^\circ$  out of phase with each other. This is the desired result.

To recapitulate, we have the sideband frequencies formed at  $V_1$  and  $V_2$  and shifted  $90^\circ$  (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  $180^\circ$ . Since  $180^\circ + 90^\circ$  is equal to  $270^\circ$ , and  $270^\circ$  clockwise is the same as  $90^\circ$  counterclockwise, we find that the sidebands and the carrier are combined  $90^\circ$  out of phase with each other. This produces phase modulation and, with it, equivalent frequency modulation.

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. 10.13A 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. 10.11B). They differ, in other words, by  $180^\circ$ . This is shown in Fig. 10.13A. Adding the vectors for the sidebands and the carrier voltages, we obtain the result shown in Fig. 10.13B. 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. 10.11A.

The output of each channel is fed into a separate mixer. Feeding into the first mixer is the voltage from a 2250-kc crystal oscillator and the 16,200 kc  $\mp$  781 cycles derived from channel No. 1.

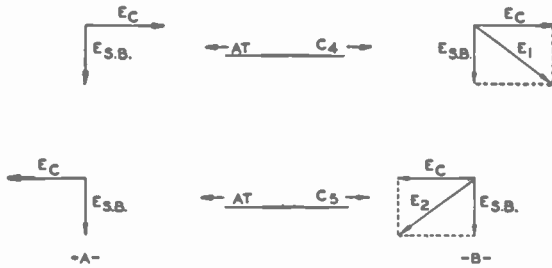


FIG. 10.13. The vectorial addition of the sideband and carrier voltage.

As a result of the mixing process, a difference frequency of 13,950 kc  $\mp$  781 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 kc, or 2250 kc. For the sidebands we have  $\pm 781 - \mp 781$  cycles, or  $\pm 781 + \pm 781$  cycles, or  $\pm 1562$  cycles. Note what has been done here. With the negative or minus sign, we have converted  $\mp 781$  cycles to  $\pm 781$  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  $2 \times 2 \times 2 \times 2 \times 3 = 48$ ) increase 2250 kc  $\pm$  1562 cycles to 108 mc  $\pm$  75 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 desired 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 88–108 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 kc. If the control oscillator is set at 2250 kc, then the difference frequency output from the first mixer is 13,869 kc. At the second mixer, 13,869 kc and 16,119 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 only tend to reduce the amplitude of the signal from each channel due to 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.

**A Commercial Unit.** The arrangement of a commercial dual channel modulator is shown in Fig. 10.14. The audio channel contains both pre-emphasis and audio correction. The crystal oscil-



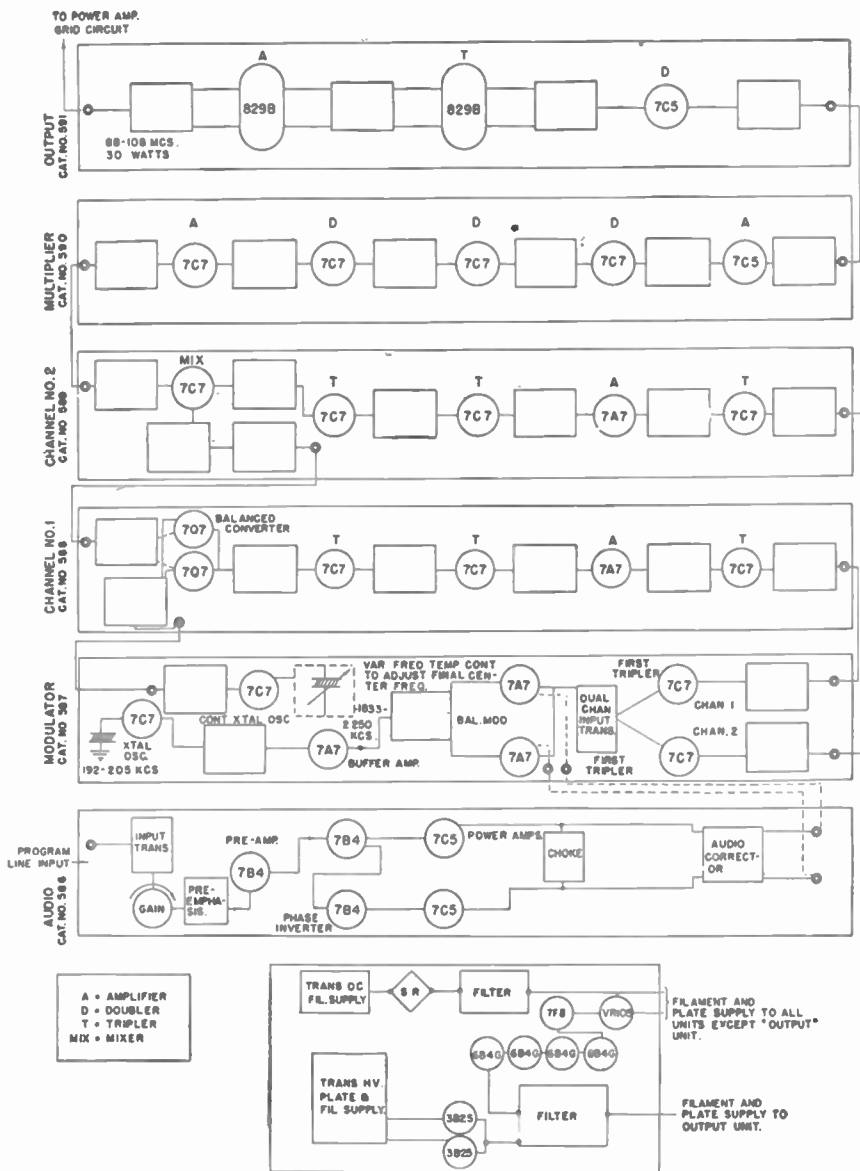


FIG. 10.14. The block diagram of a commercial dual channel modulator.

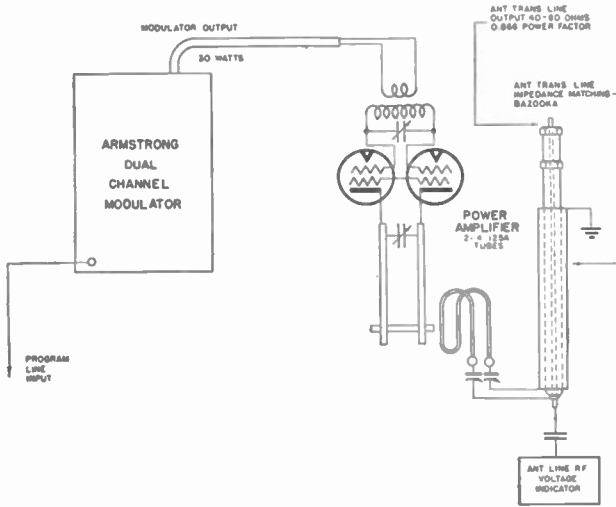


FIG. 10.15. The arrangement of a 250-watt F-M transmitter using the dual channel modulator. (Courtesy of REL.)

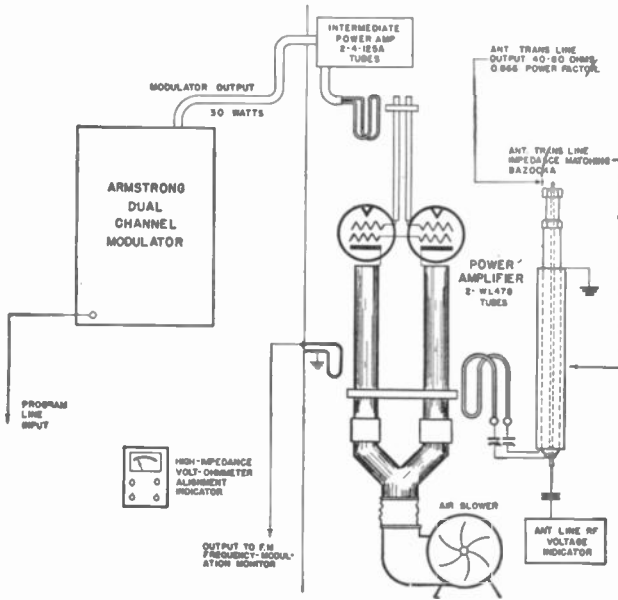


FIG. 10.16. The same modulator driving a higher-power transmitter. (Courtesy of REL.)

lator feeding the balanced modulator 7A7 tubes is somewhere in the frequency range of 192–205 kc. A 7A7 serves as a buffer amplifier. Beyond the balanced modulator we find the circuit that combines the carrier and the sidebands, after which the separation into the two channels is made. Each channel consists of four triplers and a straight amplifier.

Channel No. 1 output is applied to a balanced converter where mixing with the control crystal-oscillator voltage occurs. The crystal frequency is 2250 kc. From the balanced converter, the difference frequency is fed to the second mixer where the output from channel No. 2 is applied. It is here, in this mixer, that the original 2250 kc of the control crystal oscillator is formed again, with a frequency modulation of  $\pm 1562$  cycles.

The final function of the modulator unit is to produce the output carrier frequency (between 88–108 mc) and the  $\pm 75$ -kc sideband variation. This is accomplished with four doublers, a tripler, and three straight amplifiers. The circuits of the power amplifiers that follow the modulator unit will depend upon the power output desired. For a 250-watt carrier, the modulator could feed into the arrangement shown in Fig. 10.15. The modulator has an output of 30 watts, which is sufficient to drive the pair of 4-125A power amplifiers to produce 250 watts of radiated energy.

If 1000 watts of output power is required, then the arrangement of Fig. 10.16 is satisfactory. Because of the additional heat dis-

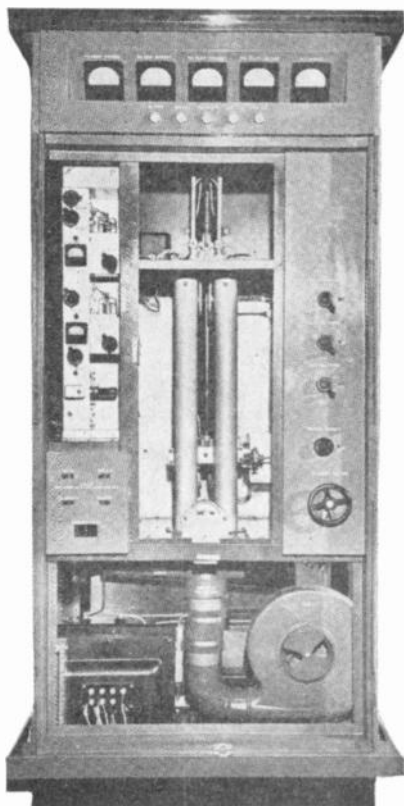


FIG. 10.17. The power amplifier cabinet showing the blower needed to cool the power tubes. (Courtesy of REL.)

sipation at the plates of the power tubes, an air blower is necessary to carry away the heat and prevent the collapse of the tube (see Fig. 10.17). Notice that quarter-wave tuning stubs have replaced the conventional coil-and-condenser resonant circuit arrangement. The air blower is plainly visible at the base of the transmitter cabinet, and the plate quarter-wave tuning stubs are located at the center of the unit. The additional output power would require the use of intermediate power amplifiers between the modulator unit and the power amplifier.

**A New Tube.** The Armstrong phase-shift method and the Crosby reactance-tube system of producing F-M have both been known for some time. Recently, however, an entirely new system

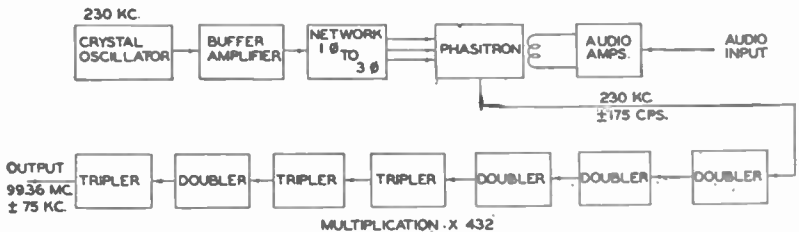


FIG. 10.18. A block diagram of the exciter unit of an F-M transmitter utilizing the Phasitron.

for producing F-M has been devised by Dr. Robert Adler and utilized commercially by the General Electric Company. This newer system employs a specially designed tube, which produces the preliminary F-M. The tube is known as the "Phasitron," and is the modulator unit for the transmitter.

A block diagram of the exciter unit, which includes all the circuits except the power amplifiers, is shown in Fig. 10.18.

The stability of the carrier is dependent upon the 230-kc crystal oscillator. In this respect, the circuit resembles the Armstrong unit. Indeed, the resemblance between the two extends even farther because this system also utilizes phase shifting to obtain equivalent F-M. However, the difference lies in the fact that all of the phase shifting due to modulation occurs within one tube, the Phasitron, whereas in the Armstrong system several tubes are required to accomplish the same purpose.

The entire transmitter is designed around the Phasitron. An internal view of the Phasitron is shown in Fig. 10.19A, with a labeled drawing in Fig. 10.19B. Although specific reference is made in what follows to the labeled schematic drawing, the reader will be able to gain a clearer idea of the entire functioning of the tube if he makes a continual cross-reference between both diagrams of Fig. 10.19. For every part labeled in Fig. 10.19B, stop and locate that part on the tube shown in Fig. 10.19A.

The tube is designed so that electrons emitted from the cathode reach

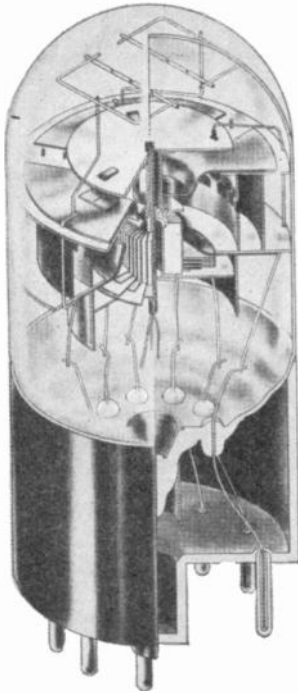


FIG 10.19A. Internal view of Phasitron.

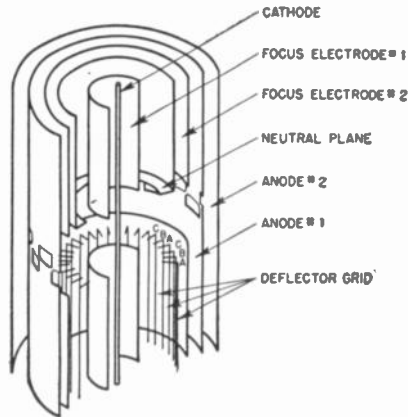


FIG. 10.19B. A labeled schematic drawing of the internal segment of the Phasitron.

the positive plates or anodes 1 and 2 in one definite plane. At all other points the electrons are prevented from reaching these anodes by shields. From Fig. 10.19A, we see that the only exposed section of the cathode is at the center. Above and below the center, two shields prevent any emitted electrons from reaching either anode 1 or 2.

Since the cathode at the center section emits electrons in all directions, or  $360^\circ$ , the emitted electrons will assume the appearance of a circular disk. The formation of this electron disk is shown in

Fig. 10.20. In order to keep the electron disk very thin, especially at the edges, two focusing electrodes are inserted. The focus electrodes 1 and 2 are shown in Fig. 10.19B.

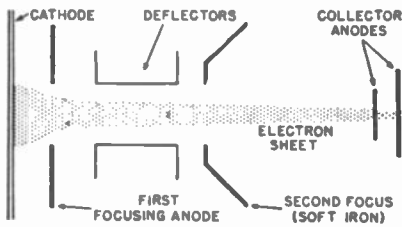


FIG. 10.20. The electron-optical arrangement in the tube for producing an electron disk.

The electron disk extends out from the cathode until it reaches the positive anode 1. When we examine this plate, we see that it has holes punched in it, each hole separated from its neighbor as shown in Fig. 10.21. The dividing line between the upper and lower holes represents the line where the edge of the electron

disk impinges on anode 1. If the edge of the electron disk impinges directly on this line, no electrons will be in the plane of the punched holes and be able to reach anode 2, which is located directly behind anode 1. Except for these holes in anode 1, no other direct path is available for the electrons to reach anode 2. This, of course, is purposely done, and fits in with the action of the tube.

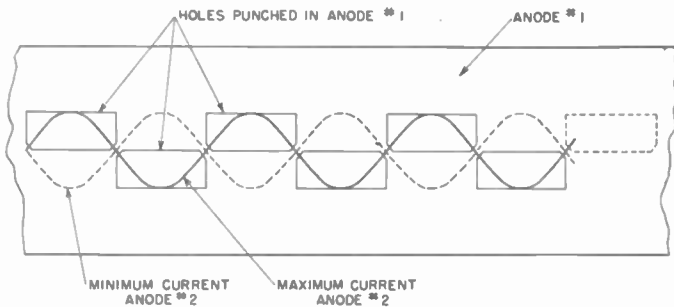


FIG. 10.21. The placement of the holes punched in anode 1.

A specially constructed set of grid wires are situated below the level of the electron disk. The grid wires are near the outer edge of the circular electron disk, just before the electrons reach anode 1. However, because the grid wires are below the electron disk, any voltage on the grid wires will only act to do one of two things: either pull the edge of the electron disk down with a positive charge or else repel the edge of the electron disk up with a negative charge.

Thus, if we connected all the grid wires together and placed an alternating voltage on them, the edge of the electron disk would rise and fall vertically (flip up and down) as the voltage changed. But whether they rose or fell, they would reach either anode 1, or, if they passed through its holes, anode 2.

The foregoing would result if all the grid wires were tied together. Actually they are not. There are 36 distinct grid wires. For the sake of explanation, every three wires have been labeled

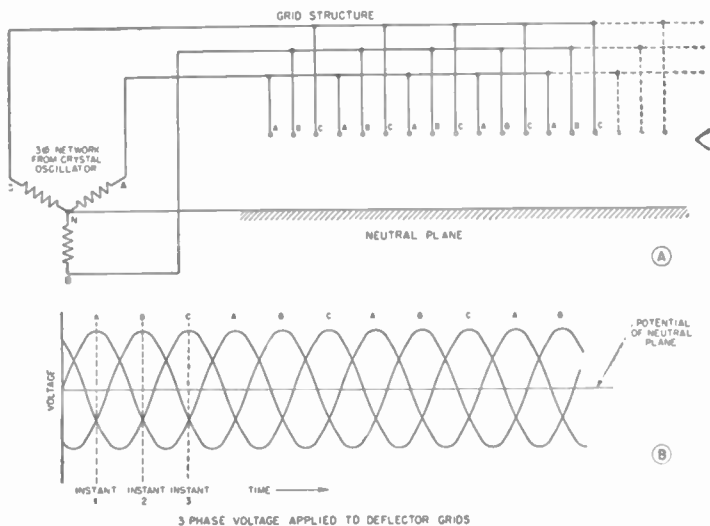


FIG. 10.22. (A) The application of the three-phase voltage to the grid structure. (B) The voltage variation of each set of grids.

*A*, *B*, and *C*, successively, until all of the 36 wires have been lettered. In construction, all the wires marked *A* are connected together electrically, as are all the *B* wires and all the *C* wires. We obtain, in this manner, a repetition in potential for every third wire, and every third wire will possess the same polarity. In the overall view, there are three sets of grid wires, each set containing 12 grid leads.

The grid wires are arranged in this manner in order that a three-phase voltage may be fed to them, one phase to group *A*, one to group *B*, and the third phase to group *C*. Fig. 10.18 indicates that the three-phase voltage is obtained from the crystal oscillator branch. The crystal oscillator generates a single-phase voltage,

which when amplified is converted to three-phase by an appropriate network. The three-phase voltage is then applied to the three sets of grids (Fig. 10.22A). Each phase is shown applied to a separate group of grids, with the common terminal of the three-phase system attached to the electrode marked "neutral plane." This electrode is directly above the group of grid wires. Any voltage between the neutral plane and the set of grid wires will act directly on the electron disk passing between these two elements.

To represent the alternating three-phase voltage, Fig. 10.22B is helpful. Each phase is shown separately, with its variations from instant to instant. The response of the edge of the electron disk to these changing grid voltages will depend upon the voltage polarity on the different sets of grid wires at any instant with respect to the neutral plane electrode.

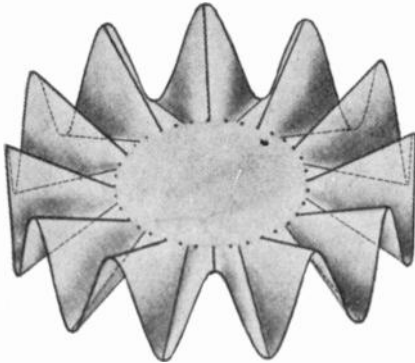


FIG. 10.23. The ruffled appearance of the edge of the electron disk.

At instant 1, Fig. 10.22B, the voltage at all the *A* wires is at its maximum positive value, while the voltages at the *B* and *C* set of grid wires are equally negative. Hence, above each *A* wire, the electron disk is attracted downwards while above each *B* and *C* wire, the edge of the electron disk is repelled upwards. The resultant shape of the disk edge possesses a ruffled appearance as shown in Fig. 10.23.

Since the grid voltage is always changing, at instant 2 we find that all the *B* wires are at their maximum positive value, while all the *A* and *C* wires are negative. Now those sections of the edge of the disk that were attracted downward previously to all the *A* wires will have gradually shifted around until they are attracted to the *B* wires. Extending the same line of reasoning, it is readily seen that at instant 3, the edges above the *B* wires will shift to the *C* wires. After this, the sequence occurs over again, with *A*, *B*, and *C* following in order as outlined.



If we were standing inside the tube, perched above one grid wire, it would appear to us as though the electron disk was rotating rapidly, with the ruffled edges of the disk rising and falling in step with the rise and fall of the controlling crystal-oscillator frequency. Since there are 12 sets of 3 wires each, we would find 12 maxima and 12 minima in the ruffled edges of the electron disk. This is evident from Fig. 10.23.

The effect of the variation in the position of the edges of the disk is to vary the amount of current that reaches anode 2, situated directly behind the hole openings in anode 1. Anode 1 has 24 holes punched in it. Twelve are above the plane of the electron disk and 12 below. Suppose that at the instant when all the  $A$  grid wires are positive, the ruffled edge assumes such a position that all the electrons pass through the anode 1 openings. This condition is indicated by the heavy, dark line in Fig. 10.21. At this instant of time, anode 2 is receiving its maximum current. But, as the ruffled edges of the disk rotate, anode 2 gets less and less current because less and less electrons reach the openings of anode 1. The second anode receives its least amount of current, which is practically zero, when the ruffled edges of the disk assume the position indicated by the dotted lines in Fig. 10.21. At this moment the electrons, at all points, are striking the solid sections of anode 1. Between the maximum and minimum amounts of current flow to anode 2 is one-half cycle. Thus, the current flowing to anode 2 varies sinusoidally at a frequency determined by the crystal oscillator.

It is now possible to see one reason for employing the three-phase voltage. Without it, we would not have any ruffled edges on the electron disk, and we could not produce the sinusoidal current variation arriving at anode 2.

*Modulating the Phasitron.* The above represents the basic operation of the circuit with no modulation. The output from GL-2H21, the Phasitron tube, would merely consist of the 230-kc frequency due to the crystal oscillator.

Before we consider the modulation, let us note the effect of a magnetic field on the rotating ruffled edge of the electron disk. The electron disk is horizontal in position, starting at the central cathode and extending to the anodes. Suppose we wind a coil about the

tube, as shown in Fig. 10.24. Then its magnetic field will extend vertically, or up and down. Hence, the electron disk and the applied magnetic field are at right angles to each other.

It can be proven, both mathematically and experimentally, that when an electron travels at right angles to a magnetic field (as it is obviously doing here) the electron will be deflected at right angles to both the direction of the magnetic field and its direction of travel. Within the Phasitron the electrons originally travel directly from cathode to anode 1 or 2. This is shown by the arrow in Fig. 10.25. Upon the application of a vertical magnetic field, the beam begins to shift at right angles to its former direction. The shift is in the

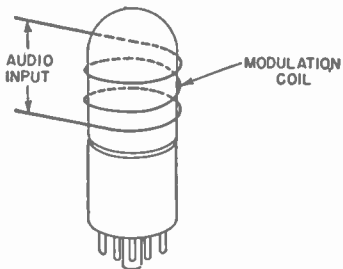


FIG. 10.24. Modulating the Phasitron tube.

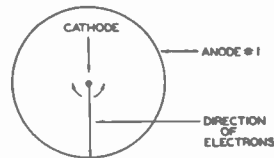


FIG. 10.25. Illustrating the effect of the modulation coil on the electron disk within the Phasitron tube.

direction of one of the two smaller arrows in Fig. 10.25. (Two arrows are shown, depending upon whether the magnetic flux is up or down.) These small arrows, it can be seen, are at right angles to the path of travel of the electrons and to the direction of the magnetic lines of force. How far the electrons shift either to the right or left of their original position depends upon the strength of the applied magnetic field. For a strong field, a large shift occurs; for a small field, a small shift.

Now let us modulate the current flowing through the coil by the audio voltages. The audio current will alternate in direction from instant to instant. In response to this periodic change in direction, the shifting of the electron disk will also be periodically to the right or left. Thus, superimposed on the regular rotation of the electron

disk due to the three-phase voltage will be this alternate shifting back and forth. Here is exactly the same type of action that occurred in the vector diagram used originally to explain phase modulation. The effect, then, of the modulation coil wound about the tube is to produce, through the fluctuating magnetic field set up by the audio currents, a back-and-forth shifting of the rotating electron disk.

Due to the shifting of the electron disk, the current reaching anode 2 will likewise vary. Since the shifting of the disk represents a phase modulation wave, the current received by anode 2 will also be phase-modulated. And from phase modulation we can obtain equivalent frequency modulation. The modulation coil is connected to a push-pull voltage amplifier tube. This, in turn, obtains its input from a 6SL7 audio amplifier. The frequency modulation produced as a result of the foregoing action is, at its maximum, equal to  $\pm 175$  cycles. By means of 4 doublers and 3 triplers, it becomes possible to raise the  $\pm 175$  cycles to  $\pm 75$  kc. This represents a multiplication of 432 ( $2 \times 2 \times 2 \times 2 \times 3 \times 3 \times 3 = 432$ ). At the same time, of course, the crystal oscillator frequency is also increased 432 times, from 230 kc to 99.36 mc. For any other frequency between 88–108 mc, the crystal oscillator frequency would change accordingly. To find the crystal frequency needed, merely take the assigned frequency of the carrier and divide this by 432.



*SECTION IV*

*Receiver Alignment*



## CHAPTER 11

### F-M RECEIVER ALIGNMENT PROCEDURE

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 proper alignment of the tuned circuits within the receiver. When a station is tuned in manually on an F-M receiver, the speaker output gradually increases in volume, accompanied by static and more receiver noise than is customary 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. At this point the signal is slightly weaker than it was on the approach. Likewise, the signal strength will increase on the other side of the correct tuning position, but with the background noise evident. Continued turning of the dial will cause the signal noise to disappear entirely. Interstation noise, unless removed by a special noise-limiter circuit, is much greater with F-M than with A-M. The reasons for this have been discussed previously.

Two methods for aligning F-M receivers are currently in use. The better and quicker method is by the use of an F-M signal generator and an oscilloscope. The alternate method requires less extensive equipment — merely a single-signal generator and either a sensitive current meter or a high-ohmage voltmeter. Since both methods are likely to find wide usage, both will be considered in detail. The serviceman can then choose the one best suited to his purpose.

At the present writing, commercial F-M receivers can be divided into three categories. In one group we have the familiar limiter and discriminator combination. To date this circuit has been used most widely, possibly because it is the oldest and most tried. Then there is the ratio detector, a recent circuit developed by RCA. The adjustments for this network are more critical than for the previous

discriminator and, until more experience has been acquired, many manufacturers have abandoned it in favor of the more familiar detector. Finally, there are those receivers that derive F-M demodulation from some form of a synchronized oscillator. At present only one large manufacturer is using this last method and it is difficult to gauge how widespread its use will become.

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 alignment can be determined by either of the methods outlined above, we will commence with the method requiring the least amount of equipment and then progress to more extensive visual procedure.

#### LIMITER-DISCRIMINATOR RECEIVERS

**Point-to-Point Method.** For the purposes of alignment, the receiver can be divided into three main sections: the R.F. stages, the I.F. and limiter stages, and the discriminator. Some manufacturers recommend alignment of the discriminator first, others the I.F. and limiter network. Actually, experience has indicated that it makes little difference which is adjusted first, provided a final test is made on both sections to check their correspondence over the total bandpass. It is very important that zero output from the discriminator occur at the midpoint of the I.F. system. Negligence in this respect will greatly impair the noise-reducing properties of the F-M signal.

To align the discriminator, the signal generator is connected to the grid of the last limiter stage (if more than one is employed) preceding the discriminator. The indicating device, either a voltmeter or an ammeter with a series resistor, is placed across *AB* or *BC*, as shown in Fig. 11.1. The meter must be connected with observance to polarity. Other than this, however, either load resistor may be used. The signal amplitude at this point in the alignment need not be kept at any particular level.

The signal generator frequency is set at the I.F. carrier value, say 10.7 mc. Then the primary tuning adjustment of the dis-



criminator transformer is varied until the voltmeter reading is maximum. In approaching maximum deflection, it is best to under-shoot and overshoot the maximum point, gradually narrowing the

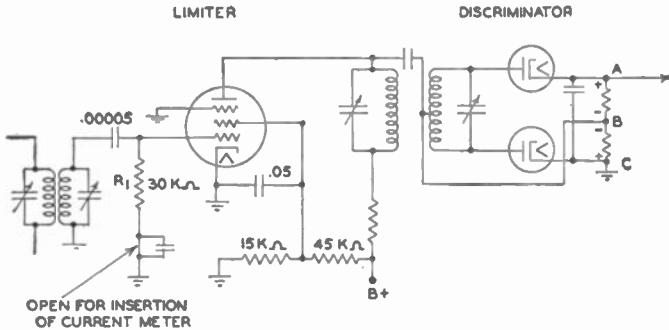


FIG. 11.1. A typical limiter-and-discriminator circuit illustrating the placement of indicators.

difference until the peak value is reached. In this way the possibility of adjusting the transformer for optimum results is enhanced.

With the signal generator untouched, place the indicator across both load resistors, or between *A* and *C*, Fig. 11.1. Adjust the secondary trimmer screw until the indicator reads zero. The voltage across both load resistors must be equal to obtain this indication and this is the proper output at the center I.F.

As an alternate method of adjustment of the secondary, an amplitude-modulated signal can be used. An audible output will be obtained from the speaker when the secondary trimmer is improperly positioned. At the proper setting, the output of the discriminator will decrease to zero and the sound will disappear.

The foregoing adjustments should have centered the discrimi-

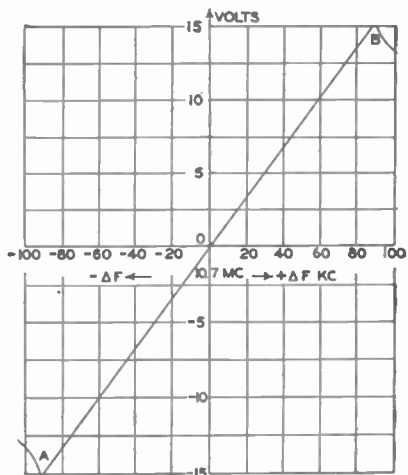


FIG. 11.2. The proper output characteristic of a discriminator.

nator 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  $\pm 75$  kc. Most manufacturers design the discriminator transformer to be linear up to  $\pm 100$  kc, but  $\pm 75$  kc will suffice. Vary the frequency of the signal generator to obtain several points such as  $\pm 20$  kc,  $\pm 40$  kc,  $\pm 60$  kc, and  $\pm 75$  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 the primary wind-

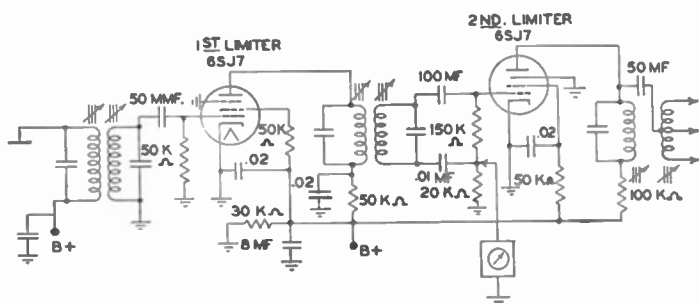


FIG. 11.3. When two tuned limiters are used, the indicator is placed in the grid circuit of the second tube.

ing trimmer must be reset. The linearity of the discriminator response is a function of the primary trimmer, whereas the cross-over point (or symmetry) of the curve is determined by the secondary trimmer.

**I.F. System Alignment.** In the adjustment of the I.F. stages, the limiter comes first. If two tuned limiters are present, the indicating meter should be put in the grid circuit of the second tube, as shown in Fig. 11.3. If the coupling between the limiters is of the untuned type, the indicator is placed in the grid circuit of the first limiter. An ammeter is inserted in series with the grid resistor. If the resistor is high in value, a microammeter is necessary; if the resistor is small, a milliammeter may be used. In the absence of an ammeter, it is possible to connect a high-resistance voltmeter from

the grid to ground. A low-resistance voltmeter will change the characteristics of the grid circuit considerably unless the signal level is kept very low. Most limiters function by having a certain amount of grid current flow on the peaks of the input signal. However, this flow of current lowers the impedance presented by the tube to the input circuit. Thus, there is a shift in the resonant peak of the tuned circuit. The smaller the resistance of the voltmeter shunted from grid to ground, the greater the circuit loading and the greater the possibility of frequency shift with even low testing signals. Many manufacturers provide a low-valued resistor in series with the grid-leak resistor. The low-resistance voltmeter

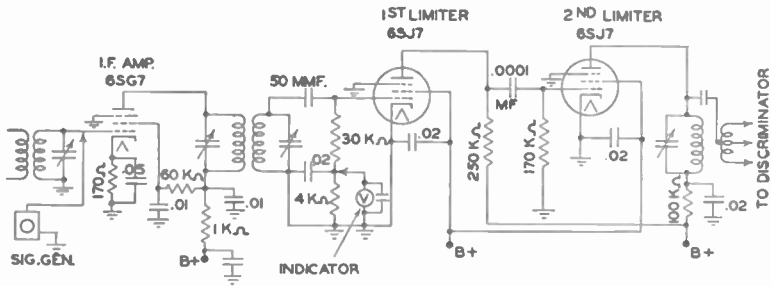


FIG. 11.4. Placement of the signal generator and indicator with a two stage R-C network.

may then be placed across the smaller resistor without seriously affecting the input circuit (see Fig. 11.4). At all times it is best to use as low an input signal as possible, consistent with good results. The stages will thus be aligned for maximum response at low signals. Upon the application of strong voltages, a certain shifting in the peak of the limiters will result, producing some distortion. However, strong signals can deal more effectively with interference and distortion and the effect of this peak shifting is then far less noticeable than it would be if the shifting occurred on weak signals.

If two limiter stages are used and they are coupled by means of tuned circuits, the signal generator is placed at the grid of the first limiter and the meter is kept in the grid circuit of the second limiter. The signal generator is still at the I.F. carrier value — the

previously mentioned 10.7 mc. The tuned transformer between the limiters is then adjusted for maximum meter indication.

When the coupling between the two limiters is not adjustable — perhaps because of resistance-capacitance coupling — then the signal generator is inserted at the grid of the last I.F. stage preceding the limiters (see Fig. 11.4). The indicator is in the grid circuit of the first limiter.<sup>1</sup> It must be cautioned again that at no time in the alignment of the I.F. system should the input signal become so strong that either limiter (if two are used) is driven into saturation. In this event, attempting to peak the transformers for sharp maximum response on the meter becomes impractical.

The alignment is continued, working back stage by stage until all of the I.F. stages have been peaked. To align the first I.F. stage, attach the signal generator to the grid of the mixer tube. Since the mixer grid tuned circuit is resonant to the much higher input signal, its impedance to the I.F. is very low. Consequently, a very high setting of the signal generator amplitude control will be required to push a small amount of signal into the I.F. system. The best method is to unsolder the lead of the grid of the mixer tube and substitute a 100,000-ohm resistor from the mixer grid to ground. The signal generator voltage is then applied across this resistor. The actual operation requires less time than it takes to describe it, and the results are excellent.

The question is often asked as to 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 oscillator is used or short the plates of its tuning condenser, providing we do not ground out the  $B+$ . In crystal oscillators, removal of the crystal itself will prevent operation of the oscillator.

<sup>1</sup>Throughout the chapter the foregoing suggestions for the placement of the indicator in the limiter grid circuit should be kept in mind. In order to avoid repetition, instructions will state only that the meter be placed in the limiter grid circuit. It will then depend upon the circuit of the receiver itself to determine whether the first or second limiter grid circuit is to receive the indicator. Furthermore, unless otherwise stated, the generator output is unmodulated.

**Overall Check on Alignment.** The entire I.F. system and the discriminator have now been adjusted. As a final test we should check the spread about the center point to determine whether the discriminator, in conjunction with the I.F. circuits, is symmetrical about the center frequency. It will be sufficient if three to five frequencies on either side of the center position are used for checking. Leave the signal generator at the grid of the mixer tube, remove the meter from the grid circuit of the limiter and place a voltmeter across the load resistors of the discriminator, points *A* and *C*, of Fig. 11.1. Next, set the signal generator to a frequency 20 kc above the center I.F. value. Note the reading on the voltmeter. Lower the signal generator frequency to a point 20 kc below the center frequency of the I.F. system. Again note the reading (with the meter leads reversed). If the circuits are in proper alignment, the two readings should be identical. Follow the same procedure for various other frequencies about the center position, noting each time whether equal indications are obtained for equal frequencies above and below the center frequency. If this is so, the system is in complete alignment. If not, it becomes necessary to readjust first the discriminator and then the I.F. circuits.

**Double-Peak Transformers.** The foregoing system of alignment is applicable to those receivers which are designed with the I.F. coils peaked to one point. There are, on the other hand, many I.F. transformers which possess a double peak. For these, a slightly modified approach is necessary. The simplest case arises when the manufacturers' data are on hand, providing information of the two peak frequencies. Suppose the peaks occurred at plus and minus 60 kc, as shown in Fig. 11.5. Then after the discriminator was aligned as described above (without regard to these peak frequencies), we would align the limiter (or limiters) and each of the I.F. amplifiers at each peak. A meter in the grid circuit of the limiter would be used, as before, for the indicating device. Each peak should develop approximately the same reading at the meter. If this condition is not obtained and the difference between the readings is relatively large, it is necessary to readjust the entire I.F. network to bring both peaks to approximately the same ampli-

tude. Sometimes the various trimmer adjustments must be juggled a little back and forth, but in no event should one trimmer be made to compensate for a gross difference in another trimmer. This indicates a trouble other than mere misalignment, say possibly a shorted turn. The I.F. transformer responsible for the improper

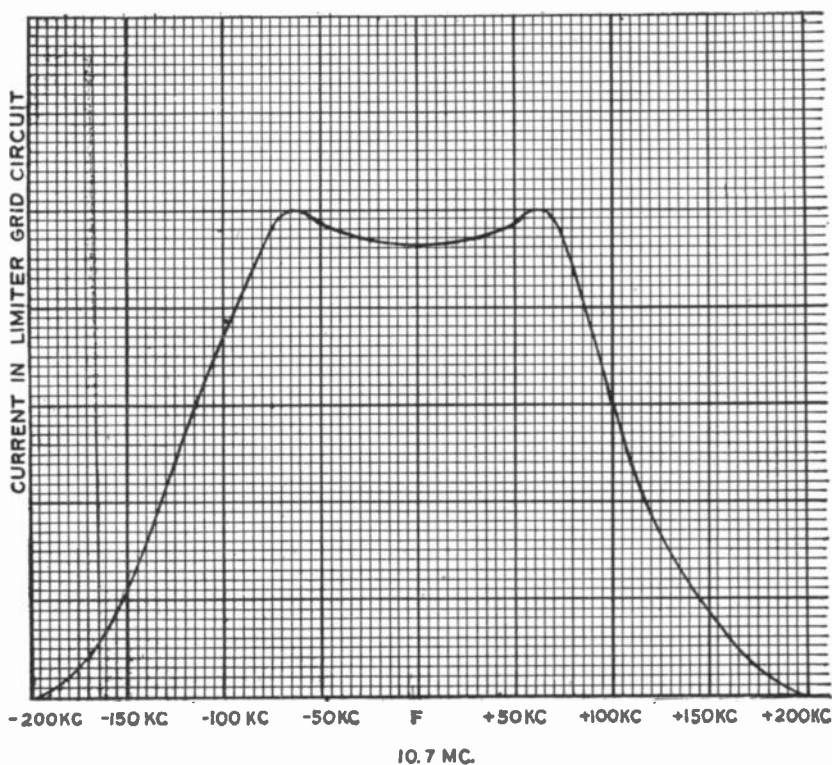


FIG. 11.5. A double-peak response curve for an F-M, I.F. system.

peaking should be examined. It may be necessary to replace it entirely.

Once the I.F. stages have been satisfactorily adjusted, several frequencies above and below the I.F. carrier value should be chosen from the signal generator and applied to the grid of the mixer. A voltmeter across the discriminator load resistors will then reveal whether frequencies equally distant above and below the I.F. carrier produce equal (and opposite) voltages at the discriminator output.

**Peaking Coils of Unknown Characteristics.** The foregoing procedure is straightforward and will cause little trouble. However, the occasion may well arise when the serviceman has a set to repair and align with no information available as to the form of the I.F. response curve. In such instances, the discriminator should be aligned first as previously described. This method of adjusting discriminators is essentially the sole method and applies to all conventional units. Following this, the meter (or other indicator) is transferred to the grid circuit of the limiter and the signal generator placed at the grid of the preceding tube. For two limiters that are transformer coupled, the signal generator is placed at the grid of the last I.F. amplified. We wish to send a signal through the first tuned circuit nearest the discriminator in order that its peaking indications may be obtained.

With the generator producing a signal at the I.F. value, peak the coils according to the indications on the meter. Use as low a signal as possible in order not to saturate the limiter. If a peak is obtained, do not immediately assume that the coils are of single peak design. To check this, move the meter to the discriminator load resistors (points *A*, *C*, Fig. 11.1). Vary the signal generator frequency equal distances above and below the center frequency. If the indications on the meter demonstrate that the response is linear, the coils are properly aligned. If the readings are considerably off, the coils may be assumed to be double peaked. The question now is — where are these peaks located?

The answer is only found by the cut-and-try method. Assume that the peaks are located 60 kc on either side of the carrier. Peak the I.F. coil we are now adjusting at these points (with the meter back in the grid circuit of the last limiter). Then test the alignment by sending in voltages at other frequencies (both plus and minus about the carrier) to determine whether equal frequency variations produce equal deflections on the meter. For this latter test, the meter is returned to the discriminator output. If the results are unsatisfactory, repeat the test with a higher peaking frequency; if this fails, use a lower one. By shifting back and forth, it should be possible to obtain the proper peaking points.

What we are doing here is to test our alignment as soon as one

stage is adjusted. When the correct peaking frequencies are given, we can afford to wait until all the I.F. stages have been adjusted before we run the overall test. But when this information is not available, we must first properly peak one transformer before we attempt the others. In situations of this type, keep the number of adjustable components to a minimum.

To summarize: When the form of the I.F. response characteristic is known, peak all the stages first before testing the overall response. When this information is not available, determine the I.F. carrier and then use the last tuned network preceding the discriminator to find the proper characteristics. Follow through with the adjustment of all the other I.F. stages. Whenever an I.F. stage (which includes the limiters) is being aligned, the indicating meter is placed in the grid circuit of either the first or last limiter, depending on the coupling between the limiters. Whenever the discriminator or the overall system is under test, we use the output voltage of the discriminator for the indications.

**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 the present commercial sets. In addition, nearly all R.F. tuned circuits are single-peaked, requiring that we adjust the various trimmers merely for maximum response.

F-M receivers are designed with one of two types of tuning; either continuous tuning, where the listener brings in a station by listening to the speaker and at the same time watching the indications on a "magic-eye" tube or else some form of push-button tuning where the frequencies of the stations that can be received are preset by the manufacturer. The alignment procedures for both are essentially the same except that in one case we can arbitrarily choose our points of alignment, whereas in the other case we are restricted to the chosen frequencies.

To start the alignment, connect the signal generator to the input terminals of the receiver. (Unless the circuits are completely out of alignment it is perfectly feasible to adjust all the R.F. circuits and the oscillator at the same time.) The indicating meter is placed in the grid circuit of the limiter. Set the signal generator to some frequency at the high end of the band, say 104 or 105 mc.



Adjust the front dial of the receiver accurately to read the same frequency. With the signal generator on, peak the oscillator until the maximum indication is obtained. Next, adjust the R.F. trimmers, while at the same time rocking the dial knob. Rocking consists of adjusting the R.F. trimmers while rotating the dial a small amount back and forth through peak output. The purpose is to find the maximum peak. If the point of maximum response occurs at a reading of the dial other than that of the incoming frequency, the oscillator trimmer will have to be adjusted while the dial is gradually brought back to correct position. Once at the correct position, the R.F. trimmers are aligned for maximum response. As previously mentioned, it is important to remember that the indicating meter, in its present position, will give usable readings only if the signal does not saturate the limiter. Keep the signal generator output at its lowest point consisting with good indications on the meter.

Next, decrease the frequency of the signal generator and again turn the dial of the F-M receiver until it reads the same frequency. Adjust the trimmers of the oscillator and R.F. circuits for maximum indication at the meter. If only one trimmer is available for each unit, a compromise position will have to be determined — one that gives the most uniform reading over the dial.

Continue decreasing the frequency of the signal generator and repeat the foregoing procedure until the low end of the band (88 mc) has been reached.

The foregoing procedure is applicable to those receivers that possess continuous manual control over the entire F-M range from 88–108 mc. Many receivers, however, employ fixed push-button tuning. For these sets the procedure is simplified in the respect that alignment is required only at the fixed frequencies. The restriction, of course, applies only to the R.F. and oscillator stages. The I.F. is untouched by the fact that push-button tuning is employed. Sets that contain crystal oscillators, operated at a harmonic, remove the necessity of a special alignment for the oscillator. In these instances adjustment of the oscillator requires only peaking the resonant circuit tuned to the harmonic. At the grid of the mixer tube feed the proper frequency signal as specified

for that particular position of the push-button. Peak the tuned circuit in the oscillator circuit and the R.F. trimmers for maximum reading at each of the push-button frequencies.

**Visual Alignment.** Without doubt, adjustment of any receiver can be accomplished more quickly and efficiently with an F-M signal generator and an oscilloscope than by means of the substitute point-by-point procedure outlined above. In one case we are viewing the complete response whereas in the other we obtain several points and the remainder of the graph is assumed. Results of the adjustments of the various trimmers may be viewed immediately, whereas in the point-by-point method, the effect of an adjustment can only be ascertained by its effect at the particular frequency and its effect on the overall response must remain unknown.

**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 through this indirect method 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 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. Remem-

ber 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

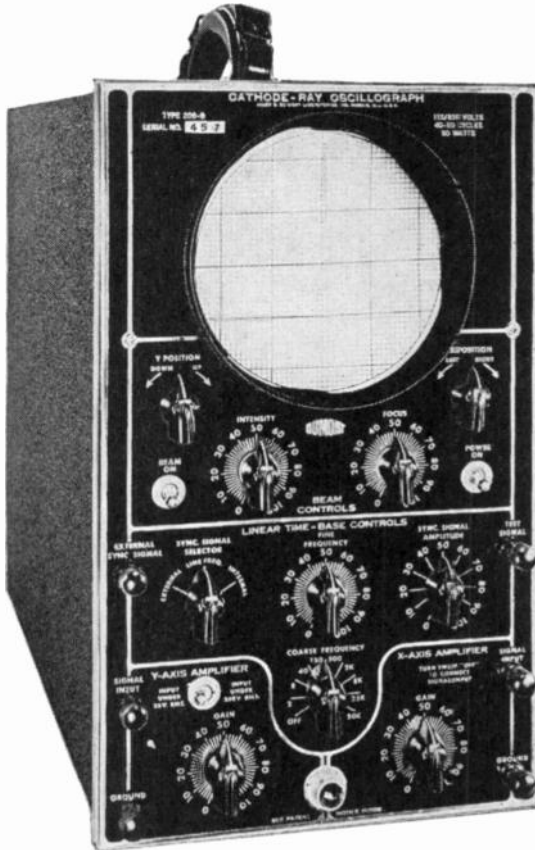


FIG. 11.6. A modern service oscilloscope. (Courtesy of DuMont.)

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

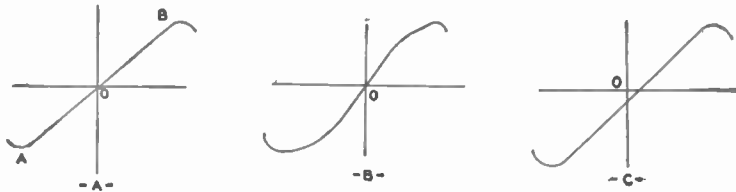


FIG. 11.7. Several response curves obtained from a discriminator (A) correct alignment; (B) improper primary adjustment; (C) incorrect secondary trimmer setting.

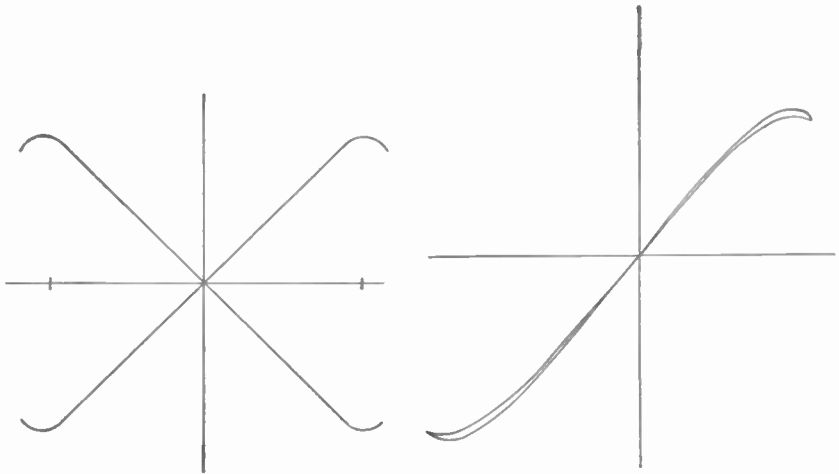


FIG. 11.8. A double discriminator response as seen on the oscilloscope screen.

FIG. 11.9. A discriminator characteristic obtained with a sine-wave sweep.

scope. If the entire I.F. system is properly aligned, the signal generator may be placed at the grid of any of the tubes from the first I.F. stage to the last limiter. If the discriminator is aligned before the other stages, the signal generator must be placed at the grid of the limiter preceding the discriminator. The vertical input terminals of the scope, for discriminator alignment, connect between point *A* and ground (Fig. 11.1).

With the signal generator in operation, the curve that should appear on the scope screen is shown in Fig. 11.7A. 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.



FIG. 11.10. A signal generator covering the F-M and television bands.

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. The curves may appear either as shown in Fig. 11.8 or Fig. 11.9, depending upon the system used by the F-M signal generator. In the signal generator shown in Fig. 11.10, the 60-cycle sine-wave input from the power line is used to modulate the reactance-tube modulator. 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 saw-tooth sweep of the indicating scope with this 60-cycle sine-wave voltage.

A saw-tooth voltage, shown in Fig. 11.11, when applied to the horizontal 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

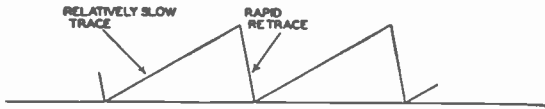


FIG. 11.11. The application of a saw-tooth wave to the deflecting plates of a scope causes a relatively slow left-to-right motion and rapid retrace.

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. 11.12 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 deflecting plates would naturally not effect the beam. Hence, when the sine wave of Fig. 11.12 is applied to the deflecting plates, and the voltages are at point *A*, the electron beam will pass through the center of the deflecting system unaffected and strike the fluorescent screen at its center.

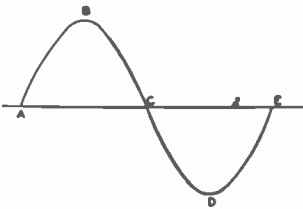


FIG. 11.12. A sine wave.

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 hence 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 reactance tube is modulated 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 we apply the F-M signal to the discriminator circuit, 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 saw-tooth sweep of the oscilloscope itself is in the off position and without effect. If the entire system (which includes the F-M signal generator and the receiver) were without distortion, the left to right motion of the electron beam should be exactly the same as the right to left motion. Since even a slight amount of discrepancy will cause one trace to be slightly displaced from the other trace, two "S" shaped curves will be obtained as shown in Fig. 11.9.

The foregoing double trace represents one of two possible types. There is yet another form that the double trace may take and this is shown in Fig. 11.8. In one instance, a sine sweep is employed, causing the electron beam to trace forward and then backward. The second trace form is obtained when, in place of a sine sweep, we use the saw-tooth sweep generated in the scope itself. To understand how the double sweep of Fig. 12.8 is obtained, let us analyze the action of the overall system.

**Saw-Tooth Deflection.** With signal generators which utilize the oscilloscope saw-tooth deflecting voltage, a linearly rising and falling modulating voltage is driving the reactance-tube modulator and producing the frequency modulation. (Whether or not a reactance-tube modulator is used makes little difference. The frequency variation to be described is obtained from 95 per cent of such F-M signal generators marketed commercially.) In one cycle, the frequency output of the signal generator will swing twice across the frequency range to be covered. As an illustration, suppose we desire a frequency carrier from the generator of 90 mc, with a frequency variation of plus and minus 150 kc. Then in one cycle

of the modulating voltage, we would obtain from the signal generator a sweep from 89.850–90.150 mc and then back to 89.850 mc again. In sweeping from 89.850–90.150 mc, the discriminator response will produce one “S” shaped curve, while in sweeping back from 90.150–89.850 mc, another “S” shaped curve will be obtained.

Two “S” shaped response curves appear on the oscilloscope (Fig. 11.8) because the saw-tooth deflecting voltage retraces rapidly when it reaches the right-hand side of the screen. Thus, in the interval when the frequency is swinging from 90.150–89.850 mc, after one “S” curve has been traced on the screen, the electron beam does not move slowly from right to left; rather it snaps back

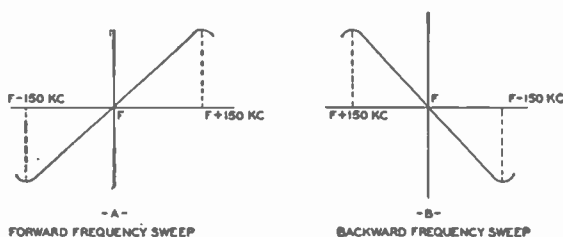


FIG. 11.13. The forward and backward frequency sweeps when saw-tooth deflection is employed.

to the left-hand side of the screen and traces out the discriminator characteristic for the frequency range, 90.150–89.850 mc. The first and second “S” shaped curves are shown in Fig. 11.13. When both these curves are placed on one screen, we obtain the result shown in Fig. 11.8.

Note that although both curves are placed along the same horizontal axis, both do not have the same horizontal axis values. While the beam traces forward on the first frequency sweep, the horizontal axis represents an increasing frequency (see Fig. 11.13A). On the second trace, the frequency is decreasing (see Fig. 11.13B). In the actual alignment this shift in axis values is unimportant as long as the “S” shaped curve is sufficiently linear and the proper frequency bandwidth is obtained.

To summarize: F-M signal generators may have one of two types of output. In one instance, the signal frequency variation occurs sinusoidally. In the other instance, the variation rises



and falls linearly. With a sinusoidal variation, the oscilloscope deflecting voltage is obtained from the generator. With a linear frequency change, the saw-tooth deflecting voltage of the oscilloscope itself is used. A sine-wave deflecting potential produces two curves, one slightly displaced from the other but otherwise similar at all points. For saw-tooth wave deflection, the two curves are mirror images of each other.

**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. Attach a wire from the "Vertical" post of the oscilloscope to terminal *A* 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 practically all F-M signal generators (see Fig. 11.10) to permit a portion of the modulating voltage to be tapped off and applied as synchronizing voltage to the terminal labeled "External Sync" on the oscilloscope front panel (see Fig. 11.6). In this way it is possible to synchronize the sweep of the oscilloscope with the modulating voltage of the signal generator and obtain a stationary pattern on the oscilloscope screen.

It is understood that synchronization is required only when the saw-tooth deflection voltage of the oscilloscope itself is utilized. For sine-wave deflection, the F-M signal generator furnishes the deflecting voltage and no additional synchronizing is needed.

A phase control is commonly incorporated in signal generators to enable the serviceman to regulate the phase of the modulating voltage. With this control it is possible to adjust the point of frequency shift and thus produce a conventional response curve. As an illustration, suppose the carrier frequency is set at 90 mc with  $\pm 150$ -kc deviation. If the phase control is then placed in such a position that the frequency shift is zero when the electron beam begins its left-to-right trace, the pattern shown in Fig. 11.14B will be obtained. The top pattern (Fig. 11.14B) is obtained on the first forward sweep of the beam on the assumption that the frequency of the signal generator starts at the carrier position (here assumed to be 90 mc), drops to 89.850 mc, and then returns to the

carrier frequency again. At this point the beam has reached the extreme right-hand side of the screen. Under the influence of the saw-tooth wave, it is brought back swiftly to the left-hand side of the screen again. Now the bottom pattern (Fig. 11.14B) is traced out when the signal generator's frequency drops to 89.850 mc and then back to the carrier position.

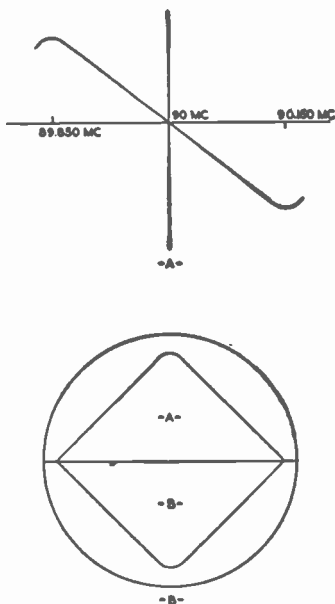


FIG. 11.14. (A) The correct pattern. (B) A possible pattern when the phase control is improperly set.

By adjusting the phase of the modulating voltage, we can alter the point of frequency shift at the start of the oscilloscope's electron beam's motion and obtain the proper "positioning" of the "S" shaped curve. Anyone who has used an oscilloscope to observe waveforms is familiar with the fact that, by adjusting the phase of the input wave, the point at which it starts at the left-hand side of the screen can be altered. Several possible positions for an ordinary sine wave are shown in Fig. 11.15. These may readily be observed with any oscilloscope.

Returning to the alignment of the discriminator, after all the controls have been properly set, adjust the primary and secondary trimmers of the discriminator coil until the proper response characteristic is obtained. The desired figure, of course, is the curve shown in Fig. 11.14 A. This completes the alignment of the discriminator.

**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 last limiter stage. The signal generator is placed between grid and ground of each successive tube working from the tube just preceding the last limiter back to the grid of the mixer stage. At each position, the interstage transformer nearest the signal generator is adjusted until the desired bandpass response curve is obtained on

the screen. If double peaking is used, care must be exercised to adjust each stage for the symmetrical curve of Fig. 11.5. The peaks must be positioned at equal distances from the center, or carrier-frequency, point. With single peaking, the midpoint of the peak should occur directly at the central carrier point. The level of the input signal is kept as low as possible in order not to drive the limiter beyond saturation.

Double curves will be seen on the oscilloscope screen, just as in the alignment of the discriminator. The remarks made at that time apply equally here.

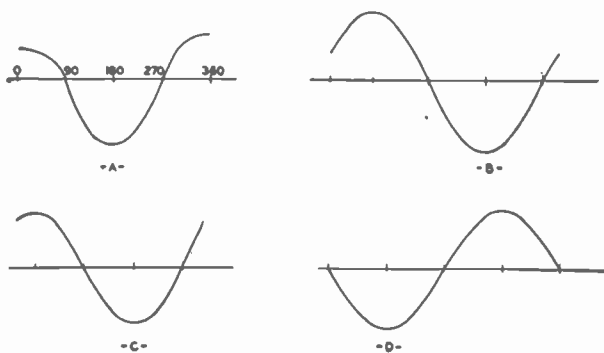


FIG. 11.15. Several sine waves, each starting at a different point in its cycle.

**R.F. System Alignment.** The alignment procedure for the R.F. stages consists of the following: Attach the signal generator to the antenna terminals of the receiver, leaving the oscilloscope across the grid-leak resistor of the limiter stage. Starting at 108 mc, adjust the oscillator and input trimmer condensers until the response curve seen at the oscilloscope is a maximum. The same procedure is repeated at several points on the dial in order to ascertain whether the response is uniform over the 88–108 mc range. Practically all R.F. coils are of single-peak design and the foregoing method of tuning for one maximum response at each test point is adequate.

**Marker Frequencies.** In using the oscilloscope, it would aid considerably if the frequency at specific points on the curve obtained at the screen were identified in some manner. Such points of indication are known as marker frequencies or marker points. Some

signal generators, like the one shown in Fig. 11.10, produce internal marker points at either 1-mc or 10-mc points. The indication, on the screen, of these points is either a slight wiggle in the curve being viewed or a slight break in the curve at that point. Both indications are illustrated in Fig. 11.16.

With the signal generator shown in Fig. 11.10, two crystal oscillators generate fundamental frequencies of 1 and 10 mc.

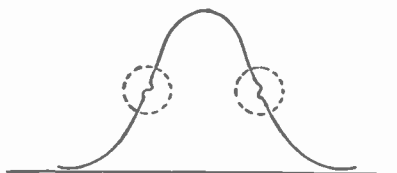


FIG. 11.16*A*. Marker indications as an aid to locating specific frequency points on the oscilloscope pattern.

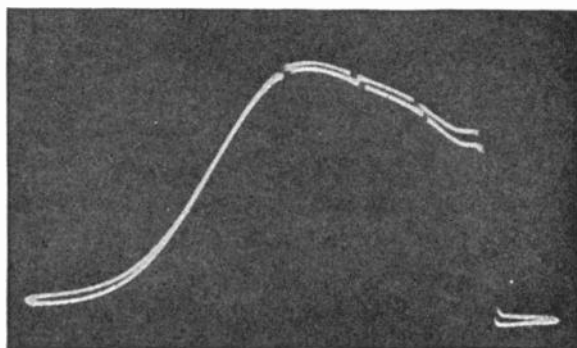


FIG. 11.16*B*.

These oscillators are designed, however, so as to produce outputs that are rich in harmonics. Hence, 1- and 10-mc marker points appear over the entire range of the unit.

It might seem that the identification of the various marker points would be extremely difficult, but this is not so when the dial reading is taken into account. Suppose, for example, that a signal of 90 mc is applied to the F-M receiver, along with 10-mc marker points. On the screen of the scope, a marker indication should appear at the center of the screen, this being the ninth harmonic of the original fundamental 10-mc oscillator, or the pip at 90 mc. If we also inject the 1-mc. marker pips, we can locate all the points on either

side of the 90-mc marker, such as 87 mc, 88 mc, 89 mc, 91 mc, 92 mc, etc. Several moments spent in becoming familiar with the dial markings of the signal generator and observing the change in position of the 10-mc marker pips will readily familiarize one with the identification of the various frequency locations on any curve produced on the screen of the oscilloscope.

The disadvantage of the foregoing system is that marker pips cannot be placed at points which are intermediate to the 1- and 10-mc values mentioned. The F-M intermediate frequency value is 10.7 mc and a marker pip could not be obtained at this point.

There is, however, still another system whereby we may obtain the marker indication at any point, but it requires the use of an additional single-signal generator which is placed somewhere along the path of the signal voltage from the other F-M generator. It 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 wiggle at the proper point in the characteristic curve.

As an illustration, suppose the I.F. system of the receiver is being aligned. For a double-peaked response, it is necessary that each peak be equidistant from the center frequency. Connect the F-M signal generator across the grid circuit of the first I.F. stage (or the mixer stage) and, at the same time, connect the generator that will supply the marker frequency in parallel with the F-M generator. Let the F-M generator sweep over the band in the usual manner. Set the other generator — which is a single-signal unit — at the frequency of one of the peaks. The oscilloscope, connected into the grid circuit of the limiter stage, will indicate a wiggle at the point where one of the peaks should be located. By switching the single-signal generator to the other peak frequency, we can then determine whether or not the system has been properly adjusted.

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.



*SECTION V*  
*Commercial Receivers*





## CHAPTER 12

### R.F. TUNERS FOR F-M RECEIVERS

**Introduction.** In conventional, sound-broadcast receivers, it is current practice to employ variable tuning condensers, ganged together, to tune across the allotted band of frequencies. At these frequencies, 600 to 1700 kc and even at the higher short-wave frequencies, such an arrangement proves to be completely satisfactory. However, when we attempt to employ the same units, properly modified for the higher F-M frequencies, we find that the results leave much to be desired. Of prime consideration is the problem of inductance. 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 becomes 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 broadcast coil's actual inductance. Hence, it may be (and is) disregarded. At the higher frequencies, however, the inductance of the wiring forms a substantial 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 condenser structure itself. Effects such as microphonic howls and inductive coupling in the condenser structure are two of the undesirable results that occur.

Undoubtedly, the most difficult job of design in an F-M receiver concerns the "front-end" or R.F. section. In Chapter 4 we touched briefly upon some of the problems that would be encountered in the mixer and oscillator stages and how these differed from the circuits in conventional broadcast receivers. It was pointed out that the increase in frequency necessitated consideration of factors

which generally did not warrant special attention in the lower frequency sets. And, to a great extent, common solutions were given.

Of particular interest to us in this chapter will be some of the tuning units which are currently being employed in commercial receivers. These, in common with the other components used in these circuits, reflect the altered requirements which the higher frequencies impose. The serviceman, in working on these receivers, must possess more than a superficial knowledge if he is to achieve any success. Toward that end, we will investigate several of the more common methods in commercial receivers.

**Inductuner.** Although this book is concerned primarily with F-M, cognizance must be taken of the fact that the F-M and

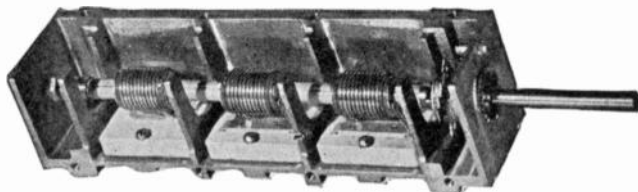


FIG 12.1. The Mallory Inductuner.

television frequencies are closely adjacent. It would be highly desirable if some tuner could be used to tune over both bands. Not only would this make for compactness, but it would also increase the usefulness of the set since the listener would be able to receive all F-M and television stations within his community. A tuner possessing these characteristics is shown in Fig. 12.1. It was designed by Paul Ware, manufactured by the P. R. Mallory Company and is commonly known as the Inductuner.

As can be seen from Fig. 12.1, the Inductuner consists of three separate variable inductance units on a common shaft. The coils are wound on ceramic forms, with movable trolley sliders for making contact at each point on the coil. At the coil end, an internal stop mechanism limits rotation to ten turns. Each coil is tunable continuously for ten turns which permits an inductance variation from approximately 0.02 to 1.0 microhenries. Note that this represents an inductance ratio of 1:50 and accounts for the

extremely wide band coverage of 44 to 216 mc. The range covers all the present television and F-M channels.

The movable trolley contact divides each coil into a used and unused section. As the trolley moves along, the low frequency end of the coil is shorted to the contact. This raises the natural frequency of the unused part and a maximum occurs when the contact is closest to the high frequency end of its travel. As a consequence, the natural frequency of the unused portion must be higher than the upper limit of the tuning range. In the unit shown, the natural frequency is approximately 300 mc.

The advantages of using a variable coil instead of a variable air condenser lies in the wider frequency coverage that is possible. A length of wire and its inductance can be reduced to a smaller value than the minimum capacitance of a variable condenser. Furthermore, as the inductance decreases, the losses reduce proportionately. In fact, it is even possible for the losses to decrease faster than the inductance, giving a rising  $Q$  with frequency. This, in turn, produces a desirable rise in gain at the high frequency end of the band.

The electrical digram of each winding of the Inductuner is shown in Fig. 12.2.  $C_A$  and  $C_L$  are stray capacitances from the high and low frequency ends of the coil to the case. In shunt with the inductance is a third capacitance,  $C$ , which consists of two parts:

1. A fixed portion representing interelement stray capacitance.
2. A variable portion which is the distributed capacitance of the winding itself. As turns are progressively shorted out, the distributed capacitance of the coil changes.

The variation in  $Q$  with frequency is shown in Fig. 12.3, illustrating the relative constancy of  $Q$ .

Each Inductuner contains three windings, which permits three circuits to be controlled at one time. These would be the R.F. amplifier, the oscillator and the mixer input circuits.

The Inductuner does not necessarily have to be used over its entire range. The internal stop mechanism can be set so as to

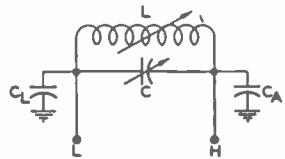


FIG. 12.2. The equivalent electrical circuit of the Inductuner.

restrict the movement of the trolley to only the desired portion of the unit. Other possible combinations are shown in Fig. 12.4 and

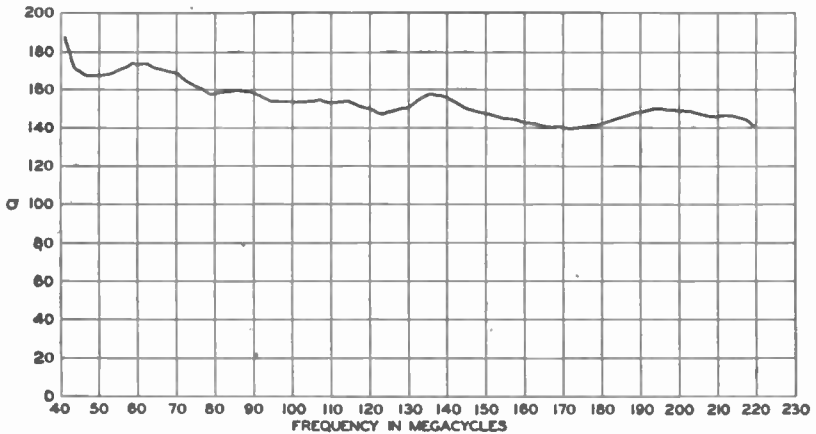


FIG. 12.3. The variation in  $Q$  with frequency for the Inductuner.

Fig. 12.5. In Fig. 12.4, the addition of a fixed end coil,  $L_1$ , permits us to obtain a better  $Q$  characteristic than if the variable coil were run out to the limit in order to reach the highest frequency. With

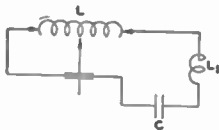


FIG. 12.4. The addition of a fixed end coil,  $L_1$ , is often employed to obtain a better high frequency response.

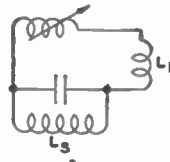


FIG. 12.5. The connection of a shunt coil,  $L_1$ , across the Inductuner provides a simple method of bandspreading a desired section of the dial.

the addition of the end coil, the trolley on the variable coil is turned to the end and the separate end inductance,  $L_1$ , provides the highest frequency. Properly designed, it is possible to produce a rising  $Q$  with frequency, which is highly desirable in view of the poorer

overall performance of receivers at the high frequency end of the band. For the circuit in Fig. 12.4, the tuning ratio is given by

$$\frac{f_2}{f_1} = \sqrt{\frac{L(\text{max}) + L_1}{L_1}}$$

where  $f_2$  = the highest frequency it is desired to reach  
 $f_1$  = the lowest frequency.

When one section of the Inductuner is used for the oscillator and one section for the mixer, a shunt coil is placed across the Inductuner oscillator unit in order that its frequency be higher than that of the other unit. In television receivers, for example, the oscillator is above the signal frequency by an amount equal to the intermediate frequency. The same is true in some F-M receivers. Hence, the

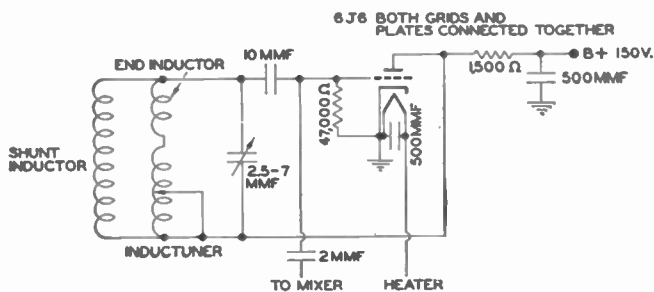


FIG. 12.6. A high frequency oscillator circuit using the Inductuner.

circuit for the oscillator would be as shown in Fig. 12.6, whereas the Inductuner section serving the mixer would be simply as shown in Fig. 12.4. To adjust the circuit of Fig. 12.6, we adjust the capacitor trimmer at the low frequency end and adjust the end inductor until the highest frequency of the band is obtained.

**Permeability Tuning.**<sup>1</sup> 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

<sup>1</sup>Z. Bevin, "A Permeability-Tuned 100 Mc Amplifier." A paper presented at the National Electronics Conference, Chicago, 1946.

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.

In the design of a H.F. permeability tuner, cognizance must be taken of the stray or circuit capacity 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 thought 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 5, 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 capacity. At resonance, the impedance of the tuned circuit is given by

$$\frac{Q}{\omega C}$$

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  $\omega$  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 capacity 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 capacity. Solving for  $Q/\omega C$  now, we obtain a value of 5000 ohms. This, in parallel with the tube (3500 ohms), produces a total input impedance 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

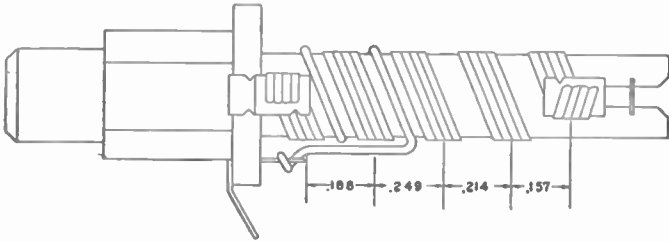


FIG. 12.7. A four-wire permeability-tuned coil. The variable pitch produces a linear characteristic.

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.

In the particular permeability-tuned coils shown in Fig. 12.7, 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 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. 12.7.

*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 will be found that the winding is not uniform in pitch. Fig. 12.8 illustrates the shape of the tuning curve for a permeability unit if the winding is made uniform in pitch. The

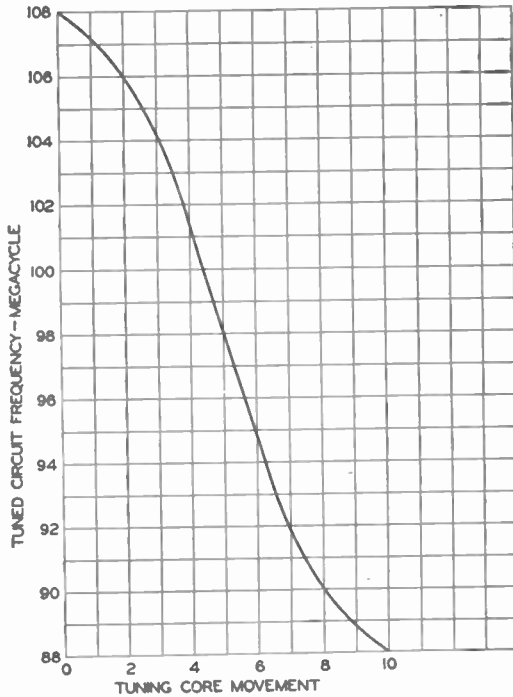


FIG. 12.8. The nonlinear shape of the tuning curve when the winding of the permeability-tuned coil is uniform in pitch.

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 further insure 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. 12.9 is the result.

In the mass production of permeability-tuned coils, it must be expected that a certain variation between units will occur. Since

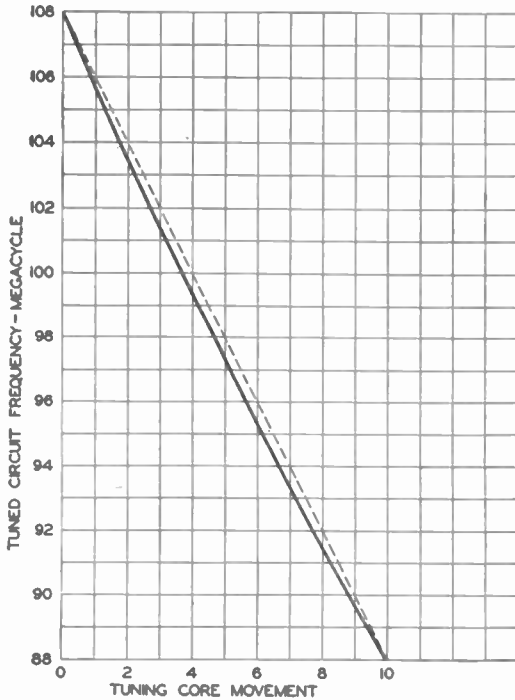


FIG. 12.9. The improved turning curve (solid line) when the turn spacing is made logarithmic. The dotted line, which is exactly linear, is used for comparison.

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 condenser is possible, but a variable condenser is highly susceptible to change with temperature and would be highly detrimental to receiver stability. It was discovered, however, 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.

Photographs of the entire tuning assembly are shown in Fig. 12.10. The antenna, mixer and oscillator coils are mounted on a bracket fastened to the side of a conventional variable condenser. The latter is used, in this receiver, to tune the A.M.-broadcast and

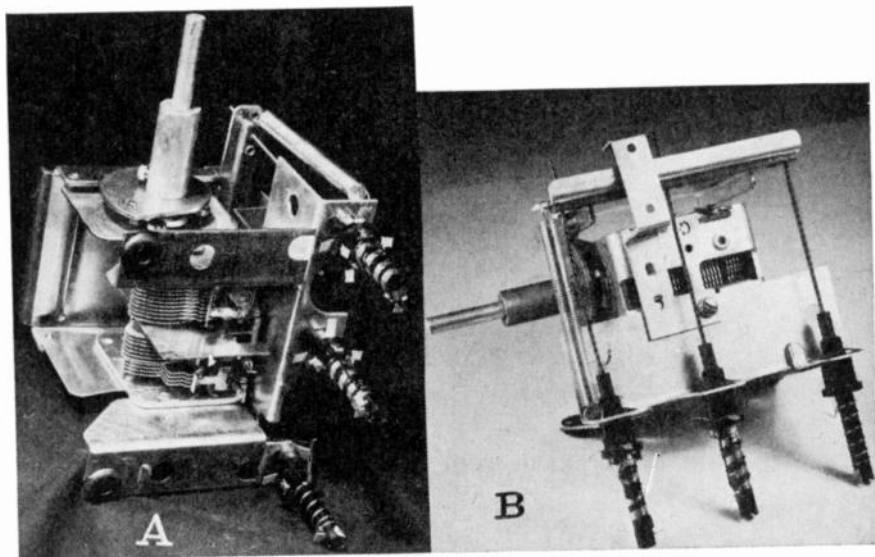


FIG. 12.10. (A) Receiver turning assembly showing the mounting of the permeability coils. (B) Underside view.

short-wave bands. A cam is mounted on the shaft of the variable condenser 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 Guillotine Tuner.** In the preceding section, tuning over the F-M band (88–108 mc) was accomplished with permeability tuning. However, a definite knack is required in winding the coils to obtain the proper pitch in production so that satisfactory calibration and tracking may be maintained. The coils have few turns, and a small displacement from the desired position is sufficient to throw off the calibration noticeably. Another solution to this

method of tuning has been developed by the General Electric Company. It uses a modified form of variable inductance tuner which, it is claimed, is more easily adaptable to production methods.

The tuner, shown in Fig. 12.11, consists of two identical brass frames which form a two-turn inductance when connected at their

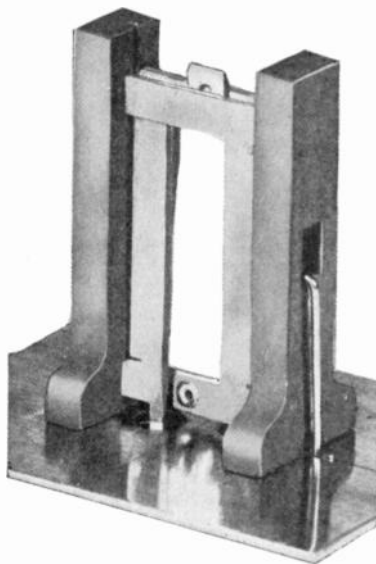


FIG. 12.11. The guillotine tuner assembly. (Courtesy of G. E.)

open ends. To vary the inductance of these two turns, a brass blade is inserted between the frames. The effect of the brass blade is to reduce the inductance of each turn or frame as the blade is moved down in the slot between the frames. To obtain the desired tuning curve, slots are cut in the blade. Once the form of the slot has been determined, the entire unit can be mechanically assembled, reducing the possibility of error. After the unit has been combined, the two terminals of the tuner project through the receiver chassis. This, in addition to providing a rugged tie point for soldered connections, makes possible short leads to the unit.

The various sections of the tuner are shown separately in Fig. 12.12. Because of the physical appearance of the unit it has been nicknamed the "guillotine tuner." In G.E. receivers, it is used to tune the F-M bands and two or more short-wave bands. The

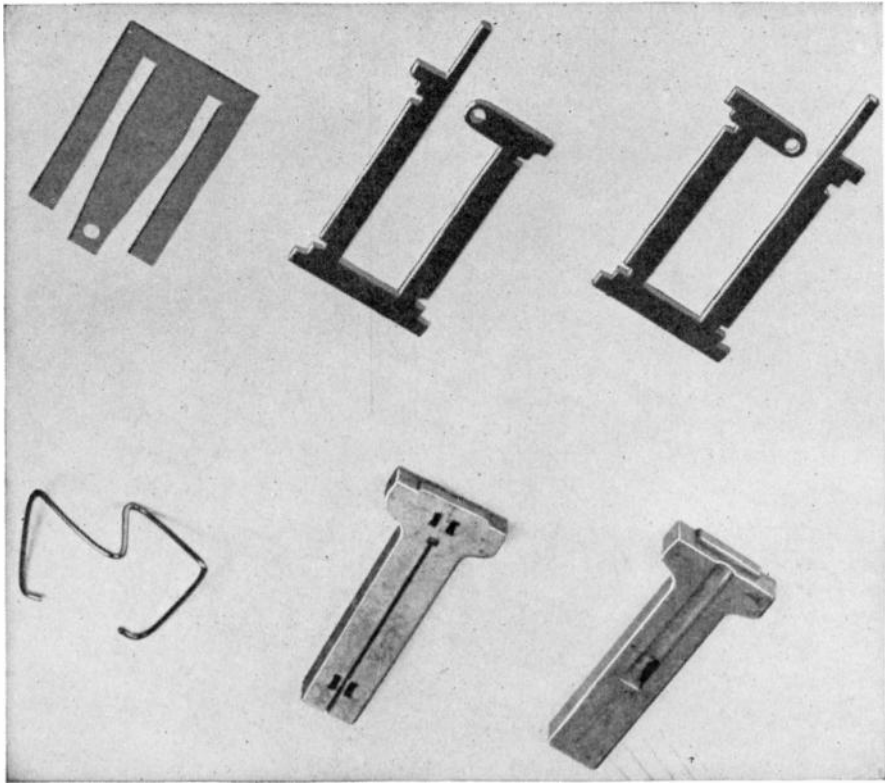


FIG. 12.12. The several components of the guillotine tuner. Note the relatively simple construction permitting easy fabrication.

low frequency ranges are tuned by adding shunt capacity and by inserting series inductances in the guillotine circuit. Fig. 12.13 shows inductance and  $Q$  characteristics of the unit over the tuning range of 88–108 mc.

A photograph of an assembly using the guillotine tuner is shown in Fig. 12.14. The blades of the tuner are raised and lowered by a plastic elevator which, in turn, is attached to a windlass. At the right-hand side of the assembly there can be seen the broadcast-band coils. These are permeability-tuned, with the iron core raised and lowered by the same assembly.

The entire tuner in final form is enclosed in a metal box which shields the unit and keeps out dust. To date, no trouble has been

encountered from microphonic howl. Further, since the blade of the tuner is ungrounded, there are no sliding contacts to produce noise.

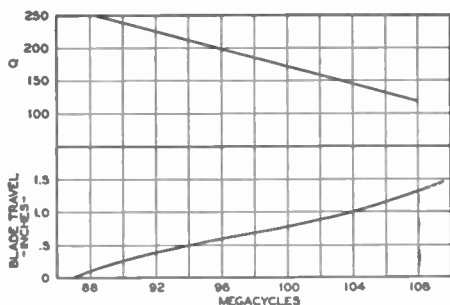


FIG. 12.13. Inductance and  $Q$  characteristics of the guillotine tuner.

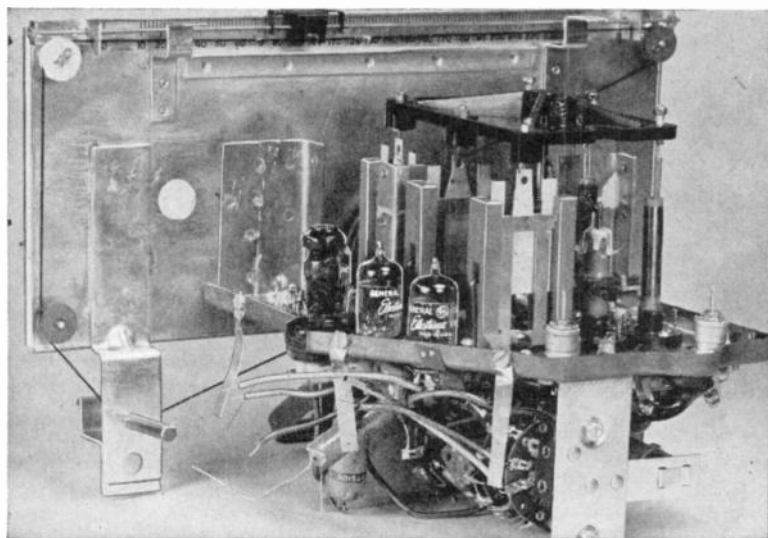


FIG. 12.14. The guillotine tuner used in a G. E. receiver.

**Coaxial Tuners.**<sup>2</sup> One of the prime reasons for using a high intermediate frequency is that a receiver becomes less sensitive to image signals. Image signals, it will be remembered, are separated

<sup>2</sup> From a paper presented by G. Wallin and C. W. Dymond at the National Electronics Conference, Chicago, 1946.

from the desired signal by twice the I.F. value. In equation form,

$$\text{Image frequency} = \text{Desired signal} \pm 2(\text{I.F.}).$$

The plus or minus depends upon whether the mixing oscillator frequency is above or below the signal frequency. As the I.F. value decreases, the frequency separation between the image and desired signals becomes less. This requires, consequently, that the R.F. tuned circuits of the receiver possess a higher  $Q$  (or selectivity) in order to reject image signals. It has been proved in practice that the input tuning circuits must impose a 60-db attenuation on the undesired signals in order to eliminate any possibility of their producing a noticeable effect in the receiver. It can be shown, when the I.F. value is 10.7 mc, that each tuned circuit must possess a  $Q$  of 75, whereas when the I.F. is 4.3 mc the effective  $Q$  per circuit rises to 200.

The advantage of a lower I.F. lies in the greater stage gain. An I.F. amplifier designed for 4.3 mc is capable of 1.57 times more gain than an I.F. amplifier functioning at 10.7 mc. More gain means less stages and a resulting greater economy.

The problem of tube loading at 100 mc was previously treated in the section on Permeability Tuning. Covered, too, was the effect of increasing the capacitance across the circuit. Since neither of these is capable of producing the desired high  $Q$  of 200 (if an I.F. of 4.3 mc is to be employed), some other method of approach is necessary.

When the input signal from the antenna is fed directly into an R.F. amplifier, the input or grid impedance is nominally in the neighborhood of 3000 to 3500 ohms. However, if the signal is fed directly into the grid of a converter tube, the input impedance may be on the order of 7000 to 12,000 ohms. To obtain the desired effective  $Q$ , it becomes necessary to feed the signal directly into the converter instead of an R.F. amplifier, to use a high value of tuning capacity, and to use high  $Q$  elements. For the last component, a high  $Q$  tuner, transmission lines are employed. Before we actually analyze the application of these lines to tuners, let us briefly review 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 components 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

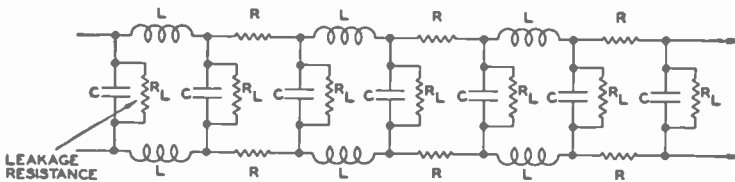


FIG. 12.15. The electrical components of a transmission line.

is formed whenever two conductors are separated by a dielectric. In this instance the dielectric is air, although it may be any non-conductor. 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 total number of magnetic lines of force that encircle 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.

To represent the foregoing inductance, capacitance and resistance in schematic form for analytical purposes, a diagram such as shown in Fig. 12.15 is often used in engineering books on transmis-

sion 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 each. 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

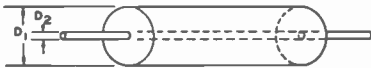


FIG. 12.16. A coaxial cable,  $D_1$  is the inside diameter of the outer conductor;  $D_2$  is the outside diameter of the inside conductor.

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 recently developed polyethylene transmission line used as lead-in cables for F-M and television antennas. Finally, there is the coaxial cable, shown in Fig. 12.16, in which one conductor is placed inside and at the center of an outer conductor. It is this final type which is used as a tuner in the F-M receivers manufactured by the Galvin Manufacturing Co. (Motorola).

**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 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 condenser placed across the end of the line. However, since the magnitude of the inductance is low, the use of a gang condenser — one section

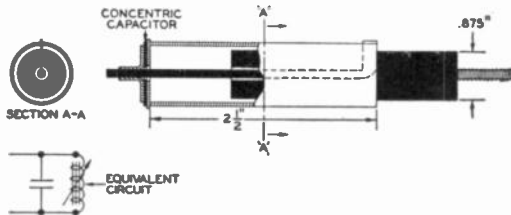


FIG. 12.17. The construction of the coaxial tuner. The equivalent coil and condenser is also shown.

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 condenser, 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 any of the previous units.

The coaxial line, with an iron core, is shown in Fig. 12.17. 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. 12.17. 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 only  $\pm 0.5$  per cent. The coaxial line and capacitor type of construction forms

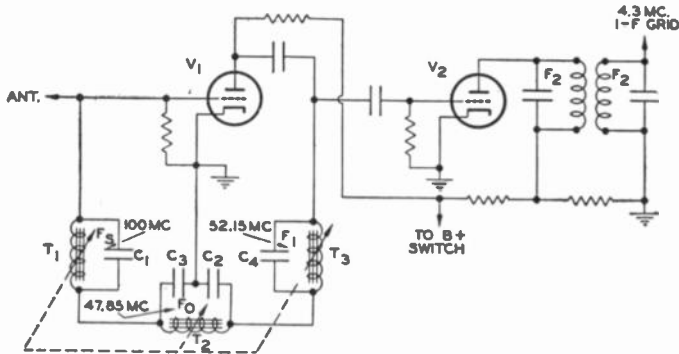


FIG. 12.18. Coaxial tuners employed in a superheterodyne circuit. The three ganged coaxial units are:  $T_1$  and  $C_1$ ;  $T_2$ ,  $C_3$  and  $C_5$ ;  $C_4$  and  $T_3$ .

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.

**Transmission Tuner Application.** Application of the tuner to a superheterodyne F-M receiver is shown in Fig. 12.18. The circuit is novel in that it employs one oscillator in a double superheterodyne network.  $C_1$  and  $T_1$  (the coaxial tuner) tune over the range of 88–108 mc, and the input signal is applied to the grid of  $V_1$ . At the same time,  $V_1$ , using  $C_3$ ,  $C_2$ , and  $T_2$ , is functioning as a Colpitts oscillator and generating a voltage which mixes with the incoming signal. For a station signal of 100 mc, the frequency of the Colpitts oscillator is 47.85 mc. The difference between these

two frequencies is 52.15 mc and this is the frequency at which  $C_4$  and  $T_3$  are resonant. This frequency now mixes with the oscillator voltage again, 47.85 mc, this time in  $V_2$ , and the resulting difference frequency of 4.3 mc is obtained and fed to the intermediate frequency amplifiers. Here, then, one Colpitts oscillator feeds two circuits. The incoming signal is lowered twice to reach the final

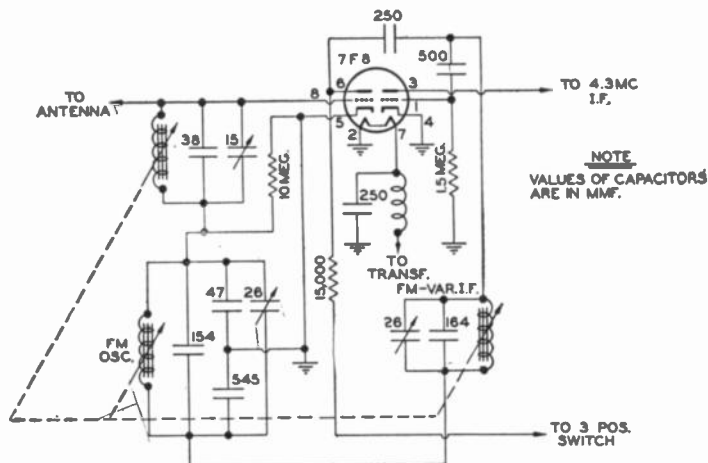


FIG. 12.19. The commercial application of Fig. 12.18.

4.3-mc I.F. The moveable iron cores of each coaxial cable are ganged together in a mechanical arrangement and moved up and down in unison.

$F_s$ , the signal frequency,  $F_o$ , the oscillator frequency and  $F_1$ , the frequency of  $C_4$  and  $T_3$  are related according to the following expressions:

$$F_s - F_o = F_1 \quad (1)$$

(The difference frequency,  $F_1$ , obtained by mixing the signal and the oscillator voltage.)

$$F_1 - F_o = \text{I.F.} \quad (4.3 \text{ mc})$$

Then  $F_1$  is mixed with the oscillator to step down the signal to 4.3 mc.

The final circuit used in the F-M receiver is shown in Fig. 12.19. A duo-triode, 7F8, replaces the two previously separate tubes.

The coaxial line used in the oscillator requires two condensers across the inductance. For this, a dual concentric condenser was constructed as shown in Fig. 12.20. Fig. 12.21 is a curve of the warm-up drift characteristic of the oscillator. The temperature

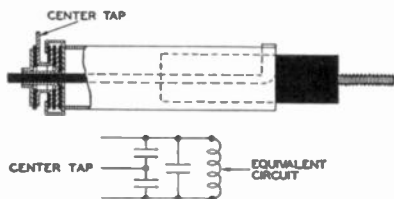


FIG. 12.20. The Colpitts oscillator (Fig. 12.19) requires a double condenser, mounted as shown above. The equivalent conventional circuit is also shown.

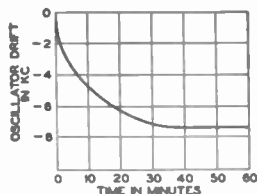


FIG. 12.21. The warm-up drift characteristic of the coaxial-tuned oscillator.

coefficients of the coaxial line and the iron core were found to be +32 and +55 parts per million respectively. A ceramic capacitor with a negative temperature coefficient is used for compensation.

In the receiver using this system for tuning, only one I.F. stage is used. The output of the amplifier then feeds into a ratio detector.

## CHAPTER 13

### COMMERCIAL F-M RECEIVERS

We have studied the various sections of an F-M receiver in considerable detail. In this chapter 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 — the type of R.F. tuner employed and the F-M detector or discriminator. The remaining sections of the circuit, which include the I.F. stages, the audio amplifiers and the power supply, are designed along closely similar lines. Although each of these stages are not exact duplicates of each other, complete understanding of one circuit will readily lead to the same comprehension of other models.

In regard to discriminators, all of the currently available receivers employ either a conventional discriminator, a ratio detector, or a synchronized oscillator, the last exemplified by the Philco circuit. For complete coverage, one commercial model of each type is included in the chapter and each is analyzed in detail, both as to the circuit content and alignment procedure.

**The Conventional Discriminator.** The schematic diagram of a Stromberg-Carlson A-M-F-M receiver is shown in Fig. 13.1. It is an eleven-tube receiver covering the broadcast range (540–1620 kc), the short-wave bands (8.8–10.2 mc), and the old and the new F-M frequencies (42–50 mc and 88–108 mc). The inclusion of the 42–50 mc band enables listeners to receive those F-M stations which have not fully converted to the newly opened band and which are still broadcasting on the old frequencies. This is actually an interim measure and these F-M frequencies will, without doubt, be dispensed with as soon as all F-M activity on the lower bands has ceased.

**Band Switches:** In examining a schematic diagram, the serviceman will generally find that, because more than one band is covered

by the receiver, selector switches are employed. Furthermore, since many of the input circuits are fully connected only through these switches, there is a maze of wires going to or leading from each section of the switch. For the sake of compactness, the various switches are mounted on one shaft to form one composite unit. As a distinguishing characteristic, however, each separate unit of the switch is generally labeled as a section, for example, section 2, section 3, etc. In the circuit of Fig. 13.1, there are six separate sections, all mounted on the same shaft, and, in addition, each section has a front (*F*) and rear (*R*) plate.

Each switch contains a moveable, circular metallic strip which, when rotated, contacts different switch terminals. In Fig. 13.1, all the switches are shown in the short-wave position. If we trace from the short-wave input coil to terminal 1 at the rear of section 1, we find that the moveable strip of this section is also contacting terminal 3. A wire from terminal 3 goes to the grid of the R.F. amplifier tube, a 6SG7. Thus, through the switch, connection is made between the short-wave antenna coil and the 6SG7 tube. Note that the arrow (denoting contact arm) from terminal 3 to the moveable strip is longer than the arrows from the other terminals of the same section. This indicates that as long as the moveable contact is not rotated completely past terminal 3, there will be electrical contact between the terminal and the moveable strip. Other terminals—those which are drawn with short arrows (denoting shorter contact arms)—do not always touch the moveable metallic strip. If an indentation or groove in the moveable strip is facing one of these other terminals, no contact is made. Thus, for example, terminal 2 of section 1-R does not contact the center strip when the switch is in the short-wave position.

Curved arrows at each section indicate that the switch is to be rotated in a clockwise direction. In Fig. 13.1, when the switch is rotated once in a clockwise direction, the broadcast (B.C.) band is switched in. Beyond this we have, in order, F-M-1 (88–108 mc), F-M-2 (42–50 mc) and a phonograph. To aid in following the schematic readily, the active or connecting terminals at each switch position have been tabulated in Table 1. In each block of the table, the numbers which are separated by a comma are electrically con-

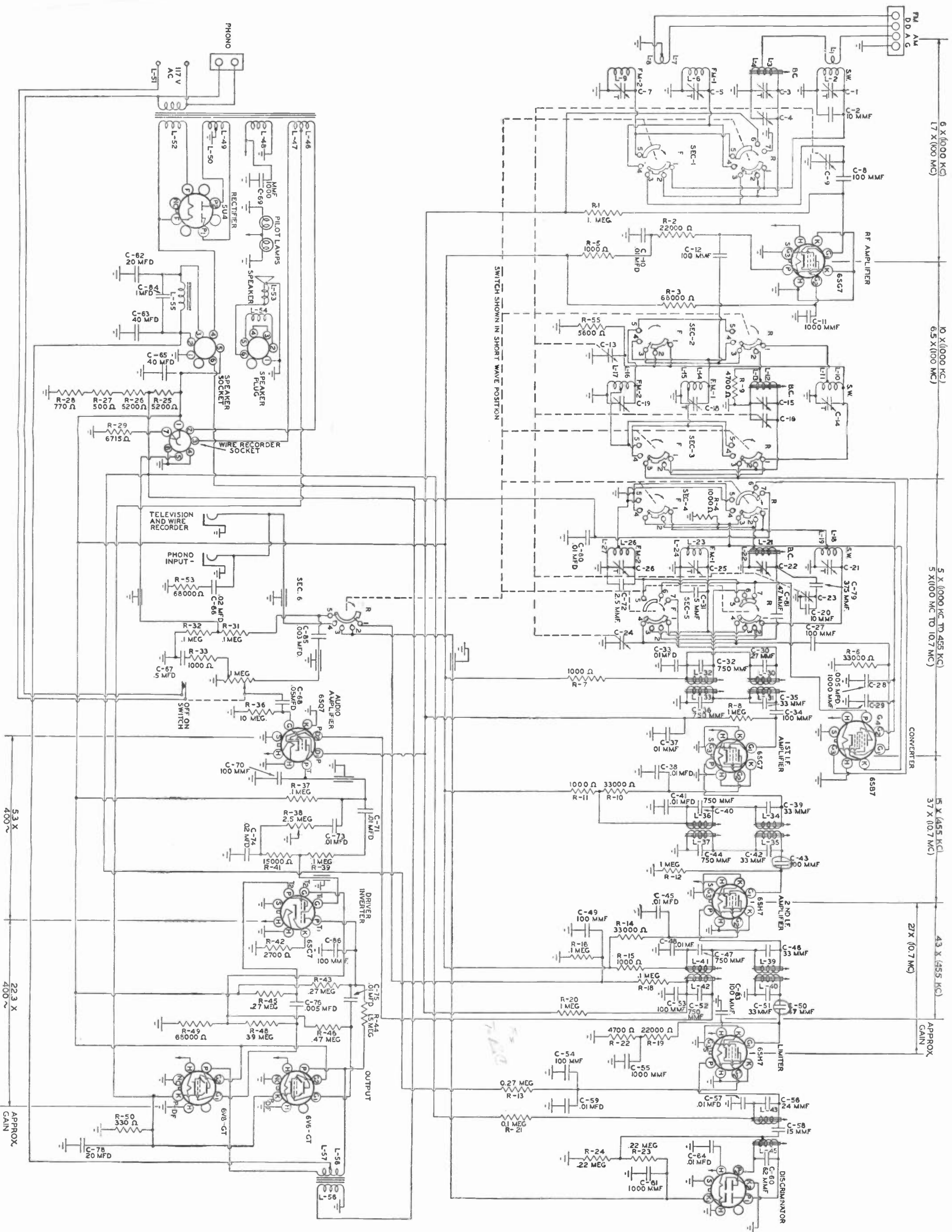


FIG. 13-1. An eleven-tube Stromberg-Carlson A-M-F-M combination receiver. (Courtesy of Stromberg-Carlson.)

5.3 X 400 ~  
22.3 X 400 ~  
APPROX. GAIN

6 X (1000 KC)  
17 X (100 MC)  
10 X (1000 KC)  
6.5 X (100 MC)  
5 X (1000 KC TO 455 KC)  
5 X (100 MC TO 10.7 MC)  
15 X (455 KC)  
37 X (10.7 MC)  
43 X (455 KC)  
21 X (10.7 MC)  
APPROX. GAIN





nected when the switch is at the position indicated at the left-hand side of the table. If two sets of numbers are given, such as 3, 4 and 6, 7, then 3 and 4 connect together and 6 and 7 connect together, but there is no connection at that switch between each set. However, it is entirely possible that some connection is made at some other portion of the receiver.

TABLE 1

Band	Switch Section										
	1R	1F	2R	2F	3R	3F	4R	4F	5R	5F	6R
Short wave	1, 3	1, 3	1, 3	1, 2	1, 3	1, 3	1, 3 6, 7	1, 3	1, 4 6, 7	1, 4	1, 4, 5
Broad-cast	2, 3	2, 3	2, 3	....	2, 3	2, 3	2, 3 6, 7	2, 3	2, 4 6, 7	2, 4	1, 4, 5
F-M-1	3, 4 6, 7	3, 4	3, 4	2, 3, 4	3, 4	3, 4	3, 4 6, 7	3, 4	3, 4	3, 4 6, 7	2, 3, 4, 5
F-M-2	5, 3 6, 7	3, 5	3, 5	2, 3, 5	3, 5	3, 5	3, 5 6, 7	3, 5	4, 5	3, 5 6, 7	3, 4, 5

A careful inspection of the connections at each switch section is important. Unless this is carefully and systematically done, it will be extremely difficult to trace the various circuits.

*Receiver Analysis:* The schematic diagram of the receiver in Fig. 13.1 indicates that the stage arrangement is as follows:

- 1 — 6SG7 ..... R.F. amplifier
- 1 — 6SB7 ..... Converter
- 1 — 6SG7 ..... First I.F. amplifier
- 1 — 6SH7 ..... Second I.F. amplifier
- 1 — 6SH7 ..... Limiter
- 1 — 6H6 ..... Discriminator
- 1 — 6SQ7 ..... A.V.C. demodulator and first audio
- 1 — 6SC7 ..... Driver and inverter
- 2 — 6V6GT ..... Output
- 1 — 5U4G ..... Rectifier

The input circuits, four in number corresponding to the four bands of the receiver, are all condenser-tuned. C-9, C-4 tune the

antenna input coil, C-13 and C-16 are for the converter input, and C-24 with C-23 resonate the various oscillator coils. All six condensers are ganged in accordance with conventional design. A simple, single-wire antenna is provided on the S.W. and B.C. bands. For the F-M stations, a dipole antenna using a balanced transmission line is required.

The R.F. amplifier, a 6SG7, operates with A.V.C. on the B.C. and S.W. bands, and with essentially zero bias on the two F-M channels. No. A.V.C. is active when the receiver is set for F-M, the A.V.C. line being grounded by the contacts on the section I-R

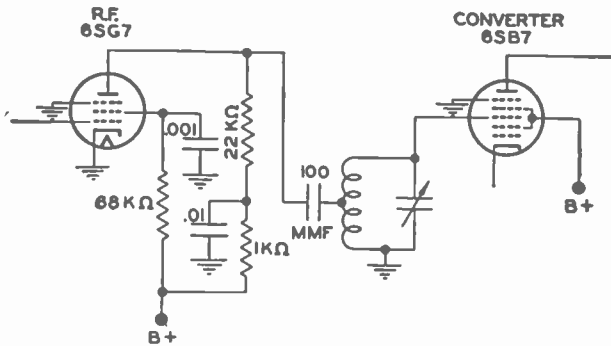


FIG. 13.2. A simplified diagram of the coupling network between the R.F. amplifier and the converter tube.

switch. Coupling between the R.F. amplifier plate circuit and the converter grid is accomplished by means of a modified impedance-coupled network (see Fig. 13.2). The R.F. voltage developed across the 22,000-ohm plate resistor is capacity-coupled to the proper tuned circuit which, in turn, connects to the grid of the 6SB7 converter. In each instance, the coil resonates with all the capacity (fixed or stray) across this portion of the circuit and maintains good response to the incoming signals. Due to this arrangement, any significant variation in interelectrode capacitances of different 6SG7 tubes might require realignment when the tubes are changed. This is especially true on the F-M-1 band, 88–108 mc. The proper converter coil is selected with section 3 switch.

The 6SB7 converter tube is one which has been especially designed for converter operation at the higher frequencies. Ord-

narily, it is not recommended that a single tube both mix and generate the oscillator voltage at the higher frequencies. The reasons for this were previously noted. However, due to special construction, it is entirely feasible to use the 6SB7 in this manner. Each band possesses a separate oscillator coil, each connected to a Hartley oscillator. Switch sections 4 and 5 control the connections within the oscillator coils. Two terminals which, when connected, supply the screen-grid voltage to the 6SB7 are included in the section 4-R switch. When the switch is at the extreme right position, bringing in the phonograph, the screen-grid B+ voltage is disconnected and the front section of the receiver becomes inoperative. This eliminates all possibility of interference from outside sources while the phonograph is in use.

In the plate circuit of the converter, we have the 10.7-mc I.F. for the F-M bands and the 455-kc I.F. for the S.W. and B.C. channels. If the circuit is examined at this point, it will be seen that the windings are switched according to the band. In the stages following the converter, switching is not employed and either I.F. signal passes through both windings. Switching is employed in the first I.F. transformer to eliminate undesired beat-frequencies that have been found to appear if both windings are kept in the plate of the converter at the same time. However, beyond this section of the receiver, the large difference between intermediate frequencies make further switching and separation unnecessary. The I.F. stages are conventional in design, with the A.V.C. voltage applied only to the first I.F. amplifier. This, however, is only for the S.W. and B.C. bands. All the coils are permeability-tuned. Adjustment consists in properly positioning the iron cores with a small screwdriver.

At the output of the second I.F. amplifier, the S.W. and B.C. signals are fed to a diode detector where they are detected and applied to the triode section of a 6SQ7. A.V.C. is also developed in this circuit and used to control the gain of the first I.F. amplifier and the R.F. amplifier on the S.W. and B.C. bands only. Both circuits follow conventional design and require no special attention. For F-M signals, the second I.F. amplifier is followed by a 6SH7 limiter and a 6H6 discriminator. The limiter operates with grid-leak bias and lowered electrode potentials, producing complete

saturation for a 2-volt signal. In the grid circuit of the limiter, a portion of the grid resistor, 4700 ohms, is by-passed with a 1000-mmf condenser. During F-M alignment, it is the d-c voltage appearing across this resistor which serves as an indication. The by-pass condenser prevents the I.F. voltage from passing through the indicating meter and possibly injuring the movement. The discriminator has already been fully described in a previous chapter. The audio output of the unit is connected through section 6-R switch to the 1-megohm volume control. A 1000-ohm resistor and series 0.5-mf condenser are placed across the lower portion of the volume control and provide automatic tone compensation at low audio

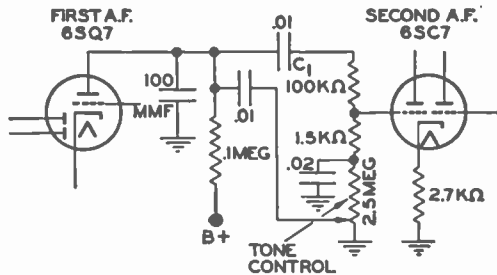


FIG. 13.3. The arrangement of the tone control network when the low frequencies are desired.

volume. The compensation is fixed, although an adjustable tone control is available in the plate circuit of the 6SQ7 tube.

The adjustable tone control is a refinement of a simple bass-boost control circuit previously analyzed. In most receivers employing this form of control, the 0.01-mf condenser and series 2.5-megohm resistor are shunted from the plate of an audio amplifier to ground. Here, however, we find that the grid return lead of the following tube is also made a part of the circuit. Due to this, the accentuation of the highs and lows becomes more pronounced. Two equivalent circuits of this network are shown in Figs. 13.3 and 13.4. Fig. 13.3 is the equivalent circuit when the lows are at their maximum; Fig. 13.4, when the highs predominate. In the bass position, the high audio frequencies are completely by-passed to ground by the 0.01-mf condenser. In addition, the full 2.5-megohms of the tone control are in the grid network of the

6SC7. Due to the added resistance, the 0.01-mf coupling condenser ( $C_1$ ) becomes less effective in reducing the low frequencies that are applied to the circuit ( $X_c = 1/2\pi fC$ ), and more lows reach the grid of the 6SC7. At the other extreme, when the full 2.5 megohms are in the plate by-passing network, only 15,000 ohms is present at the grid of the 6SC7. Now the 0.01-mf coupling condenser substantially reduces the lows reaching the 6SC7 and indirectly aids the high or treble frequency response.

The remainder of the audio amplifier network is conventional, providing push-pull output. Stability is aided by the negative feedback network from the plate of one 6V6GT amplifier, through a

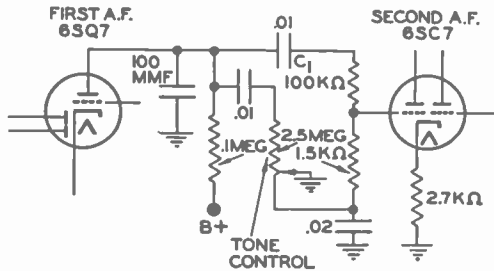


FIG. 13.4. The tone control network of Fig. 13.3 accentuating the treble frequencies.

0.01-mf condenser and 1.5-megohm resistor, to the grid of the same tube. B+ power is provided for the set and a wire recorder by a 5U4G full-wave rectifier.

The placement of the radio components on the receiver chassis can be studied from the photographs of Fig. 13.5.

**Alignment Procedure.** A modern commercial receiver is a delicately adjusted mechanism and should never be aligned unless it is absolutely necessary. The procedure to be given here is the procedure which has been recommended by the manufacturer. However, it can be employed for almost all receivers having the same type of discriminator. In other sets, the condenser and transformer identification numbers will naturally be different, but the adjustment that is made at each stage will be identical in each receiver.

Allow the set to warm up for 15 to 20 minutes before beginning the alignment. Always align a receiver using the smallest signal

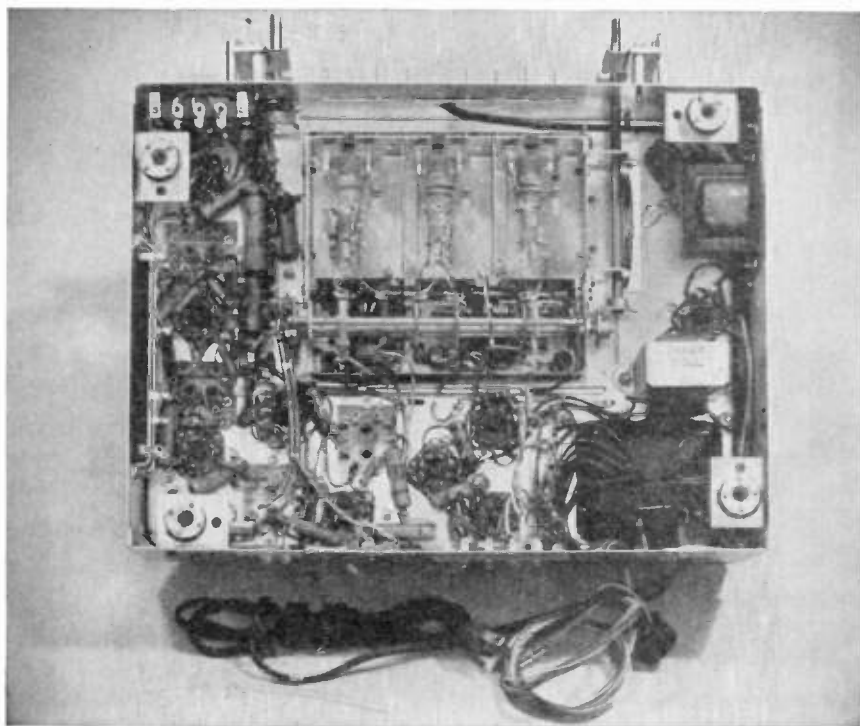
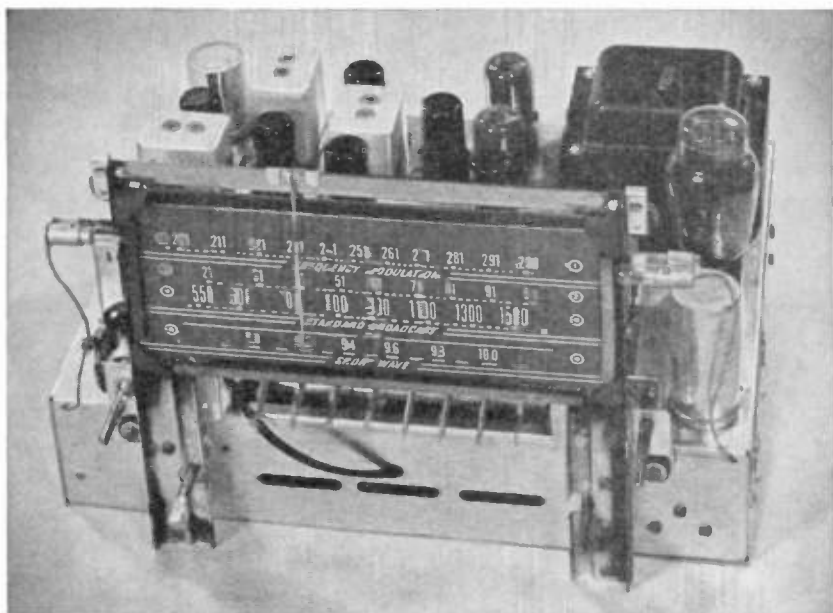


FIG. 13.5A & B. Views of the parts placement and circuit-wiring of the Stromberg-Carlson A-M-F-M receiver.

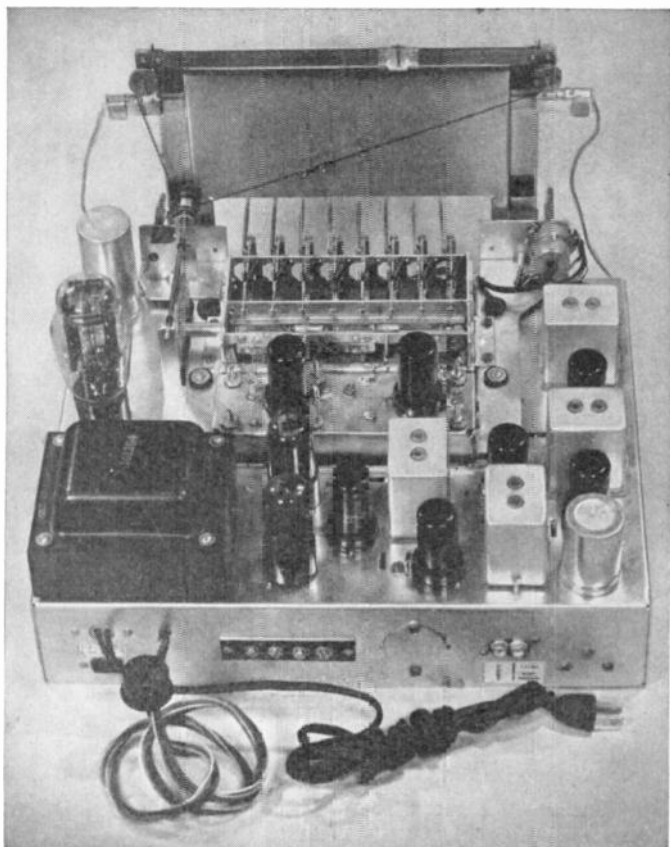


FIG. 13.5C. Another view of the parts placement of the Stromberg-Carlson A-M-F-M receiver. (Courtesy of Stromberg-Carlson.)

amplitude from the signal generator which will give a readable indication. The volume control should be set to maximum position. The first section of the receiver to align is the A-M circuits.

*A-M-I.F. Adjustment:* Unless there is reason to believe that the entire I.F. system is out of alignment, it is perfectly permissible to insert the test signal at the converter tube and adjust the entire system at one time. The first step is to connect the signal generator to the converter signal grid. This is terminal 8 of the 6SB7 tube. The ground of the signal generator is connected directly to the receiver ground. The output lead of the generator is connected to pin 8 (and other points to be specified later) through a 0.05-mf con-

denser. Next, place an output meter across the voice coil of the speaker. Most signal generators use a 400-cycle note for modulation and this should be turned to the on-position. If there is control of the degree of modulation, set it at 30 per cent. The tone control is set at the maximum high position. If a separate bass knob is available, it is set to the lowest (counterclockwise) position. Adjust the signal generator to the proper I.F. value (for this receiver, 455 kc) and adjust each of the I.F. cores for maximum output using as low a test signal as possible. During this operation, the tuning dial is set to approximately 600 kc.

I.F. transformers which are slug-tuned must be adjusted slowly and without undue force, otherwise the threads within the coil form may be stripped. The adjustment is made for maximum indication on the output meter. In the receiver shown in Fig. 13.1, the I.F. cores adjusted are:

<i>Stage</i>	<i>Coils Adjusted</i>
Converter output . . . . .	L-32, L-33
First I.F. . . . .	L-36, L-37
Second I.F. . . . .	L-41, L-42

*A-M-R.F. Adjustment:* We will consider the B.C. band first, then the S.W. band. If the receiver contains a loop antenna, it should *not* be disconnected during the alignment. The signal from the generator, in this instance, is coupled to the loop by means of another loop. Construct a 6-turn, 6-inch diameter loop using ordinary hook-up wire. Connect the loop to the terminals of the signal generator and place it near the radio loop. If the receiver does not contain a built-in loop, connect the generator to the antenna post, using a 200-mmf condenser. The modulation is again 30 per cent. The output meter remains connected across the voice coil of the speaker. The procedure, then, is as follows:

1. Set the signal generator at 1600 kc.
2. Set the station selector at 1600 kc.
3. Range switch should be at A-M.
4. Adjust the oscillator, R.F. and antenna trimmers for maximum indication on the output meter. The trimmers in Fig. 13.1 are C-3, C-15, and C-22, each marked with the letter "T."



5. Reduce the input signal and readjust the trimmers until the maximum output is obtained for the least input voltage.

6. Rotate the station selector dial to 600 kc.

7. Set the signal generator to 600 kc.

8. Adjust the iron cores in oscillator, R.F. and antenna coils for maximum output. Note that the low frequency response of an iron-core coil is governed by the iron cores, while a trimmer adjusts the high frequency end. If the coils are air-core units, only the high frequency adjustment is possible.

9. Repeat the 1600-kc and the 600-kc alignments until no further change is necessary.

For the short-wave band the following steps apply. Do not remove any built-in loop antennas.

1. Connect the signal generator to the antenna and ground terminals of the receiver using a 400-ohm series resistor.

2. Place range switch to short-wave.

3. Adjust the signal generator to 9.5 mc.

4. Set the dial pointer to 9.5 mc.

5. Adjust oscillator, R.F. and antenna trimmer for maximum output. The trimmer condensers are: C-21 (osc.), C-14 (R.F.), and C-1 (antenna).

This completes the alignment of the A-M section of the receiver. Next, we turn to the F-M section.

*F-M-I.F. Adjustment:* The F-M-I.F. value is 10.7 mc in this receiver. It is possible to align the entire F-M system at one time, just as was done on A-M, but because this system is more critical, it is best to align each stage separately. The indicator, preferably an electronic voltmeter, is connected across the 4700-ohm resistor in the limiter grid circuit. The voltmeter, in order to be most sensitive, is used on its lowest scale and the signal is always adjusted so that it does not cause the needle to swing to the extreme right-hand side of the scale. The sequence of steps is as follows:

1. Connect the signal generator to the grid of the 6SH7, second I.F. amplifier tube. This is pin 4.

2. Set the signal generator to 10.7 mc. No modulation is required since the signal itself produces a d-c voltage across the 4700-ohm resistor.

3. Adjust the cores (L-39 and L-40) of the third F-M-I.F. transformer for maximum output of the voltmeter. At all times keep the signal amplitude at the lowest possible value needed to give an indication on the meter.

4. Transfer the signal generator to the grid of the first I.F. amplifier tube, 6SG7. (Pin 4.)

5. Adjust the cores of L-34, L-35 for maximum indication of the meter.

6. Finally, transfer the generator to pin 8 (and ground) of the 6SB7 converter tube.

7. Turn the range switch to the low F-M band.

8. Adjust the cores of L-30 and L-31 for maximum indication of the meter.

Many manufacturers purposely overcouple one (or two) of the F-M-I.F. transformers in order to obtain a wide-band characteristic. When an overcoupled stage is to be aligned with a single 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 condenser 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.

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. After adjustment of all the windings in this manner, the load is removed.

*Discriminator Alignment (F-M):* 1. Connect output lead of the generator either to the grid of the 6SH7 (second I.F. amplifier) or to the grid of the limiter. The other generator lead is connected to the receiver chassis.

2. Connect the voltmeter between the center of the diode load resistors (R-23, R-24) and ground.

3. Adjust the primary of the discriminator transformer, L-43, for maximum meter indication.

4. Place the indicator across both diode load resistors.

5. Adjust the secondary of the discriminator, L-45, for zero output.

6. Swing the generator frequency 75 kc above and below 10.7 mc. Note the linearity of the discriminator. If not sufficiently linear, then the primary and secondary will have to be readjusted.

The only remaining adjustments are the alignment of the R.F. stages at the two F-M bands.

*R.F. Alignment—F-M:* In the alignment of the R.F. end of the receiver, the two terminals to which a dipole is connected are used. However, in the alignment all antennas connected to these terminals are disconnected. Note that we have a balanced input here and the

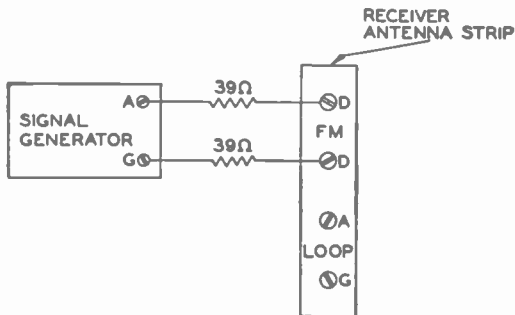


FIG. 13.6. The manufacturer's recommended method of connecting an A-M signal generator to a balanced input coil. (Courtesy of Stromberg-Carlson.)

signal generator must be connected accordingly. The manufacturer recommends two 39-ohm resistors as shown in Fig. 13.6. Other manufacturers may specify different values. For the lower F-M band (42–50 mc), set the signal generator to 46.1 mc or approximately at mid-band. On the upper F-M band (88–108 mc) set the signal generator to 100.1 mc. In each instance the range switch and the dial setting should be positioned at the proper point. Place a voltmeter across the 4700-ohm resistor in the limiter grid circuit. Adjust the oscillator, R.F. and antenna trimmers for maximum meter indication.

*Alignment with Oscilloscope:* It is preferable, when possible, to align F-M circuits with an F-M sweep generator and an oscil-

loscope. The oscilloscope is placed across the 4700-ohm resistor of the limiter grid circuit when the I.F. stages are being adjusted. The F-M sweep generator is moved from stage to stage toward the converter or mixer while the proper set of I.F. transformer cores are adjusted. For the discriminator adjustment, the oscilloscope is placed across the full output of this stage. Note that loading networks are not required with this method since the entire circuit response characteristic appears on the oscilloscope screen. The circuits are then adjusted until the proper waveform and bandpass are achieved.

**F-M Receiver with Philco Detector.** The synchronized oscillator type of F-M detector was recently developed by Philco and its principle of operation was described in Chapter 7. A complete A-M-F-M receiver incorporating this detector is shown in Fig. 13.7. Many sections of the receiver contain essentially the same type of circuits analyzed in the previous receiver and for these very little additional information is necessary. Those circuits which do differ will be examined in detail.

The switching arrangement of the receiver is conducted chiefly by push buttons. In the schematic of Fig. 13.7, all the push buttons are shown in the "out" position except the one governing the F-M frequencies. In order for a push button to be "in," it must be moved up. This is indicated by the arrowed notation at the extreme left-hand side of the push buttons. In addition to the push buttons controlling the short-wave, F-M, and broadcast bands, there are additional push buttons for five local B.C. stations, a push button for on-off and a button for the phonograph. When the phonograph is playing, the cathode circuit of the third I.F. amplifier is opened and the A-M detector connection to the volume control is disconnected. This insures that no outside stations reach the audio amplifier system while the phonograph is in operation.

There are also three rotary switches and these are actuated by the F-M push button. The rotary switches alter the circuit when F-M is to be received. More on this later.

*Receiver Circuits:* The F-M input is unbalanced, with one terminal of the dipole lead-in grounded. The R.F. amplifier is used only for the F-M, the input signal on all other bands being

NOTE: ALL PUSH BUTTONS SHOWN IN OUT POSITION EXCEPT FM  
ALL ROTARY SWITCHES ARE OPERATED BY FM PUSH BUTTON

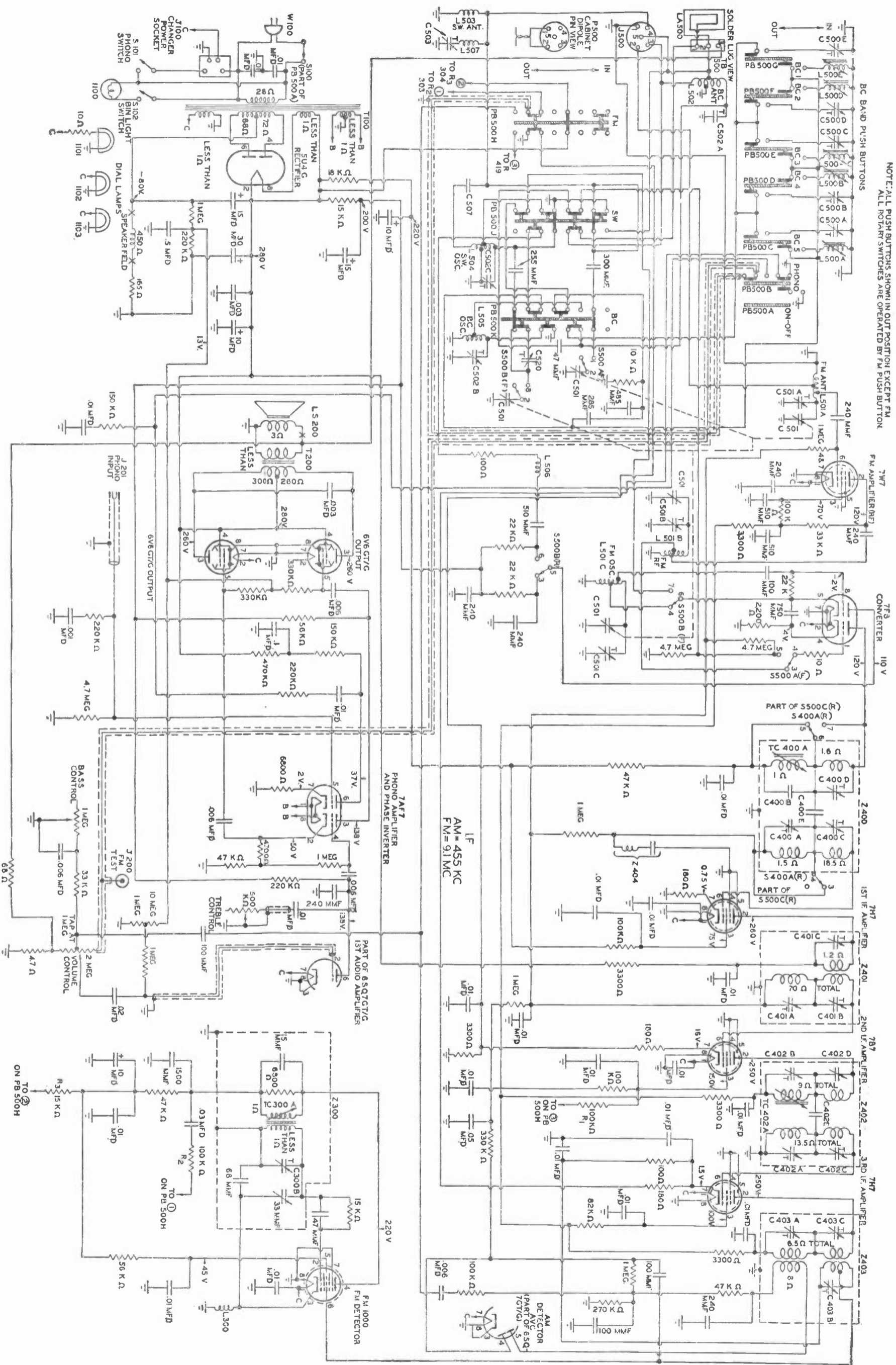


FIG. 13-7. The complete schematic Philco A-M-F-M receiver, model 46-1213. (Courtesy of Philco Corp.)  
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fed directly to the grid of the mixer triode of the 7F8 converter. This is evident by tracing from the antenna coils through the respective push-buttons and thence to the grid of the 7F8. The left-hand triode of the 7F8 converter functions as an oscillator on all bands; the mixing or conversion occurs in the other triode. The mixer output is applied to the first I.F. transformer. Note that in this receiver, in common with the previous set, a switch short-circuits the primary winding of the I.F. transformer not in use. The contacts of the switch are controlled by the F-M push button. Hence, on all but the F-M band, the contacts are in one position while, when the F-M push-button is operated, the contacts shift to the other position. At the first I.F. transformer, terminals 6 and 7 make contact on all bands except F-M. On F-M, connection is between terminals 5 and 6.

In the I.F. system, the first, third, and fourth I.F. transformers have two sets of windings each. The second I.F. transformer has a single primary winding, a secondary winding tuned to 9.1 mc, and another secondary tuned to 455 kc. The reason for employing an untuned primary is to prevent instability on the A-M bands. In the second I.F. amplifier, the screen-grid voltage is increased when the set is switched to F-M. This is accomplished by removing the ground connection on the 100,000-ohm resistor ( $R_1$ ). On all others bands, one end of the resistor is grounded, producing a voltage-divider arrangement and lowering the voltage applied to the screen grid. The increased voltage at the screen grid results in greater gain.

The A-M detector, using one diode plate of the 6SQ7 tube, is entirely conventional (see Fig. 13.8). The set also contains A.V.C., but the A.V.C. voltage is obtained using the F-M secondary winding of Z403. Both primary windings of Z403 are in series, and the currents of all I.F. signals must pass through both. Thus, whether the set is on A-M or F-M, a voltage is induced in *both* secondary windings. The second diode plate of the 6SQ7 receives all voltages and, from the amplitude of these voltages, develops a corresponding d-c voltage. This is the regulating A.V.C. voltage and is applied to the R.F. amplifier, the triode mixer on all bands except F-M, and the first and second I.F. amplifiers. Note that





0.03-mf condenser and 100,000-ohm resistor to the F-M push button and from here to the 2-megohm volume control. B+ voltage is applied to the various electrodes through the F-M push button and is active only when the switch is on.

In the audio amplifier, inverse feedback voltage is taken from the secondary of the output transformer and fed to the volume control. A bass control is connected to a tap on the volume control that is ordinarily used for the tone compensation network. A separate treble control is inserted in the plate circuit of the first audio amplifier. The 7AF7 following the 6SQ7 performs three functions. One triode section of the 7AF7 receives the output of the 6SQ7 and functions as an amplifier and phase inverter. The amplifier gain, however, is little more than unity. The other triode section of the same tube acts as a phonograph amplifier. A push-pull 6V6 amplifier drives the speaker and is capable of a maximum power output of 10 watts.

**Receiver Alignment.** The alignment of the broadcast and short-wave sections of the receiver follows identically the directions given for the preceding receiver. Consequently, we need examine only the F-M alignment procedure. This differs from what was previously given because of the Philco F-M detector circuit.

To set up the receiver for the alignment of the F-M bands, we use an A-M signal generator, connecting its ground lead to the radio chassis and the output lead through a small condenser (0.05 mf) to the points specified. Most manufacturers specify an A-M generator because it is the one most commonly found in repair shops. However, an F-M generator may be used if it is available. The use of the F-M sweep generator will be discussed after the alignment procedure, using the A-M signal generator, has been explained.

An output meter, serving as the indicator, is connected across the voice coil winding of the speaker. It remains in this position throughout the entire receiver alignment. The panel controls, treble tone and volume control are turned fully clockwise, bass tone control completely counterclockwise. Finally, the F-M push button is depressed bringing in the proper circuits.

The Philco F-M detector is a synchronized oscillator which is forced to follow the frequency variations of the input signal. To

align the I.F. stages, it is necessary to alter temporarily the F-M detector so that the output meter will indicate when each of the I.F. stages is properly peaked. By grounding pin 2 of the F-M 1000 tube or connecting it to the radio chassis, the Philco F-M detector is converted to an A-M detector. In this form, the voltage from an A-M generator can be used to align each of the F-M-I.F. amplifiers. The generator is adjusted for 400-cycle audio modulation so that the detected signal will pass through the audio amplifier stages and energize the output meter.

Before we start on the alignment of the I.F. stages, it is necessary to construct a loading network, since each I.F. transformer is overcoupled. The manufacturer specifies a loading network consisting of a 4700-ohm resistor and a 0.1-mf condenser in series. The network loads down one winding of a transformer while the other winding is being peaked. Set the generator to the I.F. value of 9.1 mc. Then, starting with the third I.F. amplifier, every transformer winding is adjusted for maximum indication on the output meter. There is no special discriminator transformer in this receiver, the final I.F. transformer (Z403, Fig. 13.7) being aligned as a normal I.F. transformer.

To align the F-M detector, remove the grounding lead at pin 2 of the F-M 1000 tube. This permits the oscillator to function again. However, in order to align the oscillator exactly to the center I.F., 9.1 mc, we short out the quadrature feedback network (see Chapter 7) by placing a jumper from pin 4 of the F-M 1000 tube to the top of the 47,000-ohm resistor in the plate circuit of the same tube. When this has been done, the signal generator, with the audio modulation off, is connected to the grid of the third I.F. amplifier tube. The frequency of the generator signal is 9.1 mc. The frequency of the oscillator is adjusted, using the trimmer across the oscillator coil (C300B in Fig. 13.7), until zero beat is obtained, as indicated by a zero reading of the output meter.

The indication on the output meter is the result of the formation of difference frequencies caused by the beating of the 9.1-mc signal voltage with the detector oscillator voltage inside the F-M 1000 tube. When the oscillator frequency is the same as the signal frequency, the difference frequency is zero and, consequently, the output

meter indicates nothing. On either side of this zero frequency point, the two mixing frequencies produce a difference frequency which can extend up to 15,000 cycles. Note, however, that if the difference is greater than 15,000 cycles, the meter will again record no deflection because the intervening audio amplifiers will not pass higher frequency signals.

The jumper is now removed and the iron core of the quadrature coil TC300A is adjusted for zero beat. With this final adjustment, the F-M detector is completely aligned. When making the last adjustment, it is necessary that the output from the generator be kept below the level where the oscillator locks-in. If this precaution is not observed, a false zero-beat indication is obtained. The proper adjustment is a single, very sharp zero-beat point.

After the alignment of the F-M detector, there remains only the R.F. coils to adjust. Using the A-M signal generator, we must again convert the F-M 1000 tube to an A-M detector; hence, ground pin 2 of the F-M 1000 tube. The signal generator is then connected, through a 0.05-mf condenser, to the ungrounded dipole input terminal. The ground of the generator is attached to the radio chassis. Set the signal generator (with the 400-cycle modulation on) to some frequency at the upper end of the F-M band, say 106 mc. Set the receiver dial to the same frequency. Adjust the F-M oscillator trimmer condenser (C501C in Fig. 13.7) for maximum deflection of the output meter.

Next, lower the signal frequency and set the receiver dial to a point at the low end of the band, say 90 mc. Adjust the oscillator coil itself for maximum response by physically adjusting the spacing of the turns. To determine whether the turns should be moved closer together or spread apart, take a tuning wand and insert the brass end in or near the oscillator coil (L501C in Fig. 13.7). If the output meter reading increases when this is done, spread the turns apart slightly. On the other hand, if the powdered-iron end of the wand, when inserted into the coil, produces an increase in meter reading, compress the turns slightly. The coil adjustments must be handled carefully since only a slight change is necessary at these high frequencies. The coil is properly set if either metal causes a decrease in output-meter reading. The adjustments at the high and

low ends of the F-M band are repeated until no further improvement is noted.

Adjust each of the other coils and trimmer condensers in the R.F. section (C501B, C501A, L501B, and L501A) similarly until the best possible readings are obtained.

**Oscilloscope Alignment.** While the manufacturer outlined the foregoing receiver alignment with an A-M signal generator, the job could have been accomplished with an oscilloscope and an F-M signal generator.

To start the alignment of the I.F. stages, pin 2 of the F-M 1000 tube is grounded. The vertical plates of the oscilloscope are then connected across the 47,000-ohm resistor in the plate circuit of this tube. Connect the F-M signal generator to the input grid of each I.F. stage and adjust the proper trimmer condenser until the best response is obtained. With this method, the full bandpass characteristics of the I.F. system is visible on the oscilloscope screen. Note that *no* loading networks are required in this method because the entire bandpass characteristic appears on the oscilloscope screen.

To adjust the F-M detector, the oscilloscope remains in the same position. However, the grounding lead to pin 2 is removed. With the F-M sweep generator at some intermediate stage, the detector-oscillator adjustments are made, while observing the F-M detector response to the full range of frequencies. The desired characteristic is a straight line, as shown in Chapter 7.

If the F-M sweep generator covers the 88–108 mc range, then the R.F. section of the receiver can also be aligned. Again ground pin 2 of F-M 1000 tube. Connect the signal generator to the dipole input terminals. Set both the generator and the receiver dial to the same frequency (say, first at 106 mc and then at 90 mc). Adjust all R.F. trimmers and coils for maximum response and desired bandwidth at each point.

**The Ratio Detector.** The ratio detector is a product of the research of the Radio Corporation of America and used in their model 612V3 A-M-F-M receiver. A schematic diagram of this receiver is shown in Fig. 13.10 and 13.11. The schematic is divided into two sections, this being the manner in which the set is constructed. The main unit contains everything up to the first audio





amplifier; the second, smaller chassis has the remaining audio amplifiers and the power supply. The entire receiver, including the cabinet, is shown in Fig. 13.12.

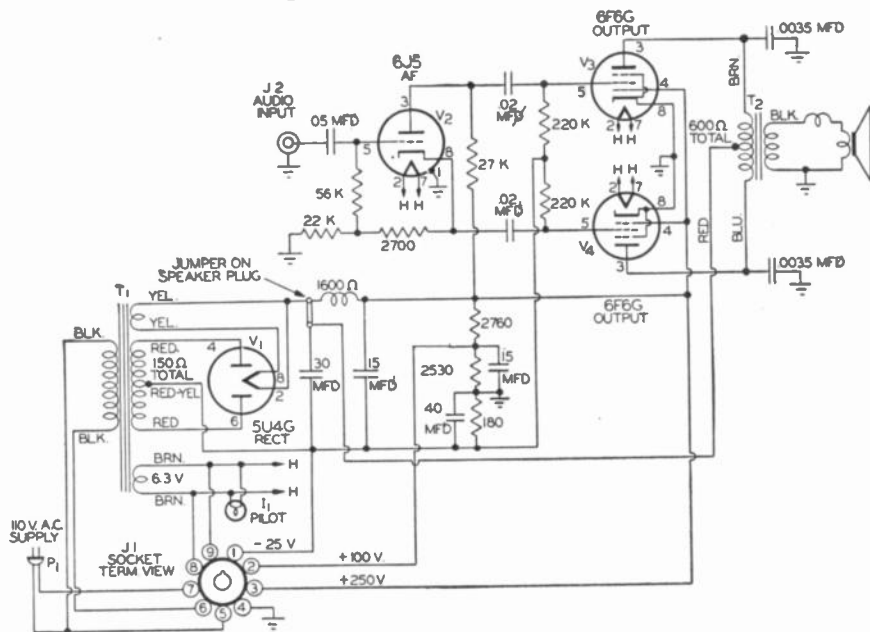


FIG. 13.11. The audio amplifier section and power supply of the RCA model 612V3 receiver. (Courtesy of RCA.)

The twelve tubes of the receiver are divided, functionally, as follows:

6BA6	.....	R.F. amplifier
6BA6	.....	Mixer
6BE6	.....	H.F. oscillator
6BA6	.....	First I.F. stage
6AU6	.....	Second I.F. stage and phonograph amplifier
6AU6	.....	Driver for ratio detector
6AL5	.....	Ratio detector
6AT6	.....	A-M detector, A.V.C.
6J5	.....	Second audio amplifier
6F6G (2)	..	Output power amplifiers
5U4G	.....	Power supply rectifier

This receiver contains many circuits which are similar to those already analyzed. As before, these will be omitted and the description confined to the circuits which differ from what has already been given. The switches are shown in their maximum clockwise position, in the F-M position. According to the system by which these



FIG. 13.12. The 612V3 A-M-F-M receiver described in the text. (Courtesy of RCA.)

switches are wired, the first position, in a counterclockwise rotation, is for the short-wave band. The next one, farther counterclockwise, is the broadcast band. The final position, beyond this, is for the phonograph.

The 6BE6 tube functions as an electron-coupled Hartley oscillator on the F-M band and as an electron-coupled Colpitts oscillator for the two lower bands. Simplified diagrams of the oscillator



connections under both conditions are shown in Fig. 13.13. Note that the filaments of the tube, at the F-M band, are connected to the source of power by means of a short section of transmission line. The purpose of the line is to bring the heater wires (both of them) within the tube to R.F. ground potential. It has been found that if this is not done, the oscillator will not develop the fullest possible output and will be fairly unstable in operation. At the lower

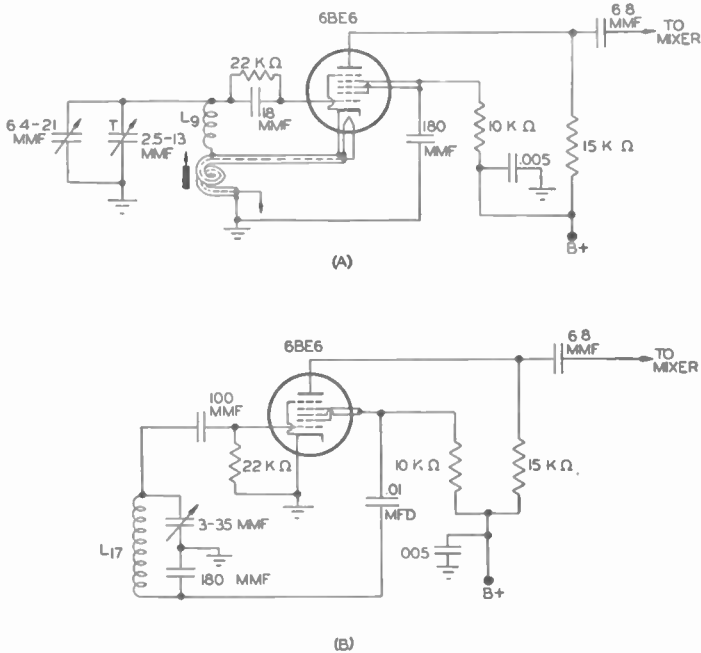


FIG. 13.13. Simplified diagrams of the oscillator circuit (A) for the F-M band (B) for the A-M bands.  $L_{18}$  replaces  $L_{17}$  on short-waves.

frequencies, the short length of this line is negligible and can be disregarded, placing the cathode and one side of the filament at ground potential.

The A-M section of the I.F. system uses only the 6BA6 tube for amplification. The F-M signal, however, is amplified by two stages of I.F. and one driver stage (also at the I.F. value) before reaching the 6AL5 ratio detector. The reason, of course, is the lower amplification that each stage is capable of providing for the 10.7-mc F-M signal.

In order to avoid the use of another tube, the 6AU6 second I.F.

amplifier also serves as the first amplifier for the phonograph attachment. The output of the phonograph is fed to the grid of the 6AU6 and the amplified signal is taken from the plate circuit, with the 22,000-ohm resistor serving as the plate load for the tube when it is being used as an audio amplifier. The signal is tapped off by means of a 0.1-mf condenser and brought to the volume control, after which it follows the same path taken by all the other audio signals. As a means of preventing interference from incoming signals, when the phonograph is in operation, the switches are connected so that they ground the grid of the R.F. amplifier and open the grid circuit of the mixer tube and the first I.F. amplifier. Still another precaution is the removal of the B+ voltage from the plate of the 6AU6 driver tube when the set is not in the F-M position.

Automatic volume control is developed and applied to the R.F. amplifier and first I.F. stage on all three broadcast bands. When the set is switched to the F-M position, the A.V.C. is obtained from the ratio detector. On the two remaining bands, it is obtained in the conventional manner, using a diode. No A.V.C. is available in the phonograph position.

In the path of the F-M signal, between the second I.F. amplifier and the ratio detector, we find a tube labeled "driver." This is the 6AU6. Upon closer examination, it is found to be operating as a partial limiter, employing a modified form of grid-leak bias provided by a 15,000-ohm grid resistor in parallel with a 5-mf condenser. The inclusion of a driver stage may appear to contradict the basic advantage of a ratio detector — namely, the ability to function without the use of limiters. It is possible to omit the driver stage and obtain satisfactory operation, and indeed many commercial sets utilizing this type of detector do so. However, through the use of a driver, or partial limiter, the fidelity of the unit can be improved.

The characteristic response curves of the ratio detector are shown in Fig. 13.14. These reveal that the output voltage of the detector decreases after the amplitude of the incoming signal has increased beyond a certain point. This behavior is due to the fact that the optimum coupling for the mean signal strength is such that the primary impedance increases as the signal-rise produces more secondary damping because of the presence of the diode tubes.

When a diode conducts, its internal resistance depends upon the amplitude of the applied voltage. The greater the voltage, the lower the tube's resistance. It is this decrease in resistance that produces the damping of the secondary winding. The lowered secondary impedance and increased primary impedance alters to some extent the phase relations between the primary and secondary circuits and accounts for the behavior of the characteristic curves.

The convergence of the response curves at high signal amplitudes means proportionately less audio output. The inclusion of a stage preceding the detector to limit the maximum strength of strong signals would minimize this form of distortion. Hence, a driver stage is used which functions as a partial limiter and is effective actually only with strong signals.

Instead of coupling the driver output directly into the ratio detector, link-coupling is employed. The use of this type of coupling at this point is closely related to the distortion which is produced in the driver stage and the effect of this distortion on the detector characteristic. It was discovered, upon investigation, that this reason the zero deviation curve (marked 0 in Fig. 13.14) is not a straight line at different signal amplitudes is due chiefly to the formation of I.F. harmonics in the stages preceding the detector. By removing the harmonics, we can straighten out, to a great extent, all of the curves.

To minimize the transfer of harmonics to the ratio detector, link-coupling is employed. Between any two inductively coupled circuits, there also exists capacitive-coupling. The capacitive-coupling is due to the incidental capacitance that exists between any two conductors. The transfer of harmonics occurs chiefly through this incidental capacitance. The usefulness of link-coupling is that it is accomplished with a turn or two closely coupled to the driver

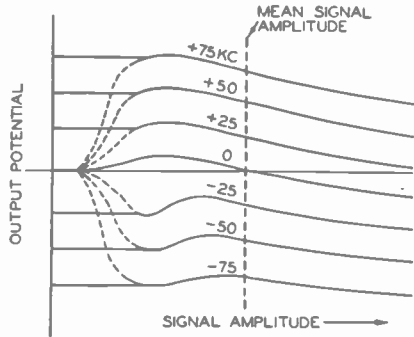


FIG. 13.14. The characteristics of the ratio detector as the signal amplitude varies. A perfect unit would give a constant output or, graphically, a series of straight lines.

plate circuit at a point where the R.F. potential is near zero. Link-coupling also reduces the effect of the secondary load variations on the driver plate circuit and thereby maintains a higher  $Q$  in the latter circuit. This further aids in the suppression of harmonics.

In the present circuit, the link-coupling coil serves yet another purpose. It functions as the "sense" or reference voltage for the phase detection of F-M signals. This was previously discussed. The ratio detector circuit is rearranged in Fig. 13.15 in order that the second purpose of the link coil,  $L_1$ , can be more readily seen.

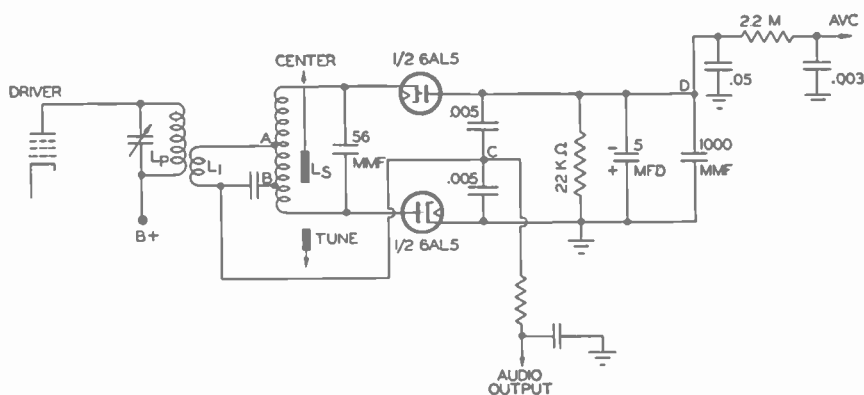


FIG. 13.15. The ratio detector circuit of Fig. 13.10 rearranged to indicate more clearly the use of its several components.

Point  $A$  of the link coil connects to the center of the secondary coil. From this point, we can follow the path through  $L_1$  to point  $C$ , between the two 0.005-mf condensers. The demodulated audio variations which appear across the lower 0.005-mf condenser is passed through a de-emphasis network and then applied to the volume control at the entrance of the audio amplifier system. The link coil connects to point  $B$  on the secondary coil in order that the signal in  $L_p$  appears in  $L_s$  also.

The regulating voltage for the detector operation is developed across the 22,000-ohm resistor and the 5-mf condenser. This latter condenser is shunted by a 1000-mmf condenser in order to neutralize the inductance existing in the relatively large 5-mf condenser at 10.7 mc. The negative voltage at point  $D$  is useful for A.V.C. purposes and is used as such. A filter, composed of a 2.2-megohm resistor,

## F-M RATIO DETECTOR ALIGNMENT

Set range switch to F-M position

Steps	Connect High Side of Osc. To	Tune Osc. To	Turn Vol. Cont. To	Adjust
1	Connect a 680-ohm resistor between pins 5 and 7 of the ratio detector tube 6AL5, (V7). Connect d-c probe of a voltohmyst to the negative lead of the 5-mf electrolytic capacitor $C_1$ , and the common lead of the meter to chassis.			
2	Driver grid pin 1 of 6AU6 (V6) in series with a 0.01-mf capacitor.	10.7 mc 30% modulated at 400 cycles A-M.	Maximum volume.	Driver transformer $T_6$ for maximum d-c voltage across $C_1$ (approx. 14.5 volts).
3	Remove meter leads and disconnect the 680-ohm resistor from the tube 6AL5, (V7). Connect two 68,000-ohm resistors (within 1% of each other) in series across the 22,000-ohm, ratio-detector load resistor $R_1$ . Connect the common lead of the voltohmyst to the center point of the 68,000-ohm resistors and the d-c probe to terminal $A$ of the ratio-detector transformer $T_7$ . Use the 30-volt scale.			
4	Same as Step 2.	Same as Step 2. (Approx. .25 volts output.)	Average volume level.	$T_7$ bottom core for zero d-c balance* $T_7$ top core for min. audio output (output meter across voice coil).**
5	Reconnect voltohmyst as in Step 1, omitting the 680-ohm resistor.			
6	Repeat Step 2 omitting 680 ohms.			
7	Remove all connections.			

\* Near the correct core position the zero point is approached rapidly and continued adjustment causes the indicated polarity to reverse. A slow approach to the zero point is an indication of severe detuning, and the bottom core should be turned in the opposite direction.

\*\*The zero d-c balance and the minimum A.F. output should occur at the same point: If such is not the case, the two cores should be adjusted until both occur with no further adjustment of either core. It may be advantageous to adjust both cores simultaneously, watching the voltohmyst and an output meter connected across the voice coil for the point at which both zero d-c and minimum output occurs.

NOTE: Two or more points may be found which will satisfy the condition required in Step 4.  $T_7$  top core should be correctly adjusted when approximately  $\frac{1}{8}$  inch of threads extend above the can. Therefore, it is desirable to start adjustment with the top core in its furthest "in" position and turn out, while adjusting the bottom core, until the first point of minimum A.F. and minimum d-c is reached.

## F-M ALIGNMENT

*Range switch in F-M position*

Steps	Connect the High Side of the Test Osc. To	Connect Ground Side of the Test Osc.	Tune the Osc. To	Radio Dial Tuned To	Adjust
1	Connect the d-c probe of a voltohmmyst to the negative lead of the 5-mf electrolytic capacitor $C_1$ , and the common lead of the meter to chassis ground.				
2	Mixer grid pin 1 of 6BA6 (V2) in series with a 0.01-mf capacitor.	To chassis ground.	*10.7 mc 30% modulated at 400 cycles a-m.	Maximum capacitance (fully meshed).	** $T_5, T_3, T_1$ top and bottom cores alternately loading primary and secondary of each transformer with 680 ohms while the opposite side of the same transformer is being adjusted. Adjust all transformers for maximum voltage across $C_1$ .
***3	F-M antenna terminals 1 in series with a 120-ohm resistor.	To F-M antenna terminal 2 in series with a 120-ohm resistor.	106 mc	106 mc	Osc., $C_2$ for max. voltage across $C_1$ .
4			88 mc	88 mc	Osc., $L_9$ for max. voltage across $C_1$ .
5	Repeat Steps 3 and 4 for exact calibration.				
6	Same as Step 3	Same as Step 3	106 mc	106 mc	R.F., $C_3$ for max. voltage across $C_1$ .
7	Same as Step 3	Same as Step 3	90 mc	90 mc	R.F., $L_{11}$ for max. voltage across $C_1$ .
8	Repeat Steps 6 and 7 for maximum output.				
9	Same as Step 3	Same as Step 3	106 mc	106 mc	Ant., $C_4$ for max. voltage across $C_1$ .
10	Same as Step 3	Same as Step 3	90 mc	90 mc	Ant., $L_8$ for max. voltage across $C_1$ .
11	Repeat Steps 9 and 10 for maximum output.				

\* When the windings are loaded, it is necessary to increase the 10.7-mc input since the gain will decrease and the voltage across  $C_1$  will be less.

\*\* This method is known as alternate loading which involves the use of a 680-ohm resistor to load the plate winding while the grid winding of the same transformer is being peaked. Then the grid winding is loaded with the 680-ohm resistor while the plate winding is being peaked.

\*\*\* Before proceeding with Step 3, set the pointer on MECHANICAL MAXIMUM CALIBRATION point at the low end of the band when the gang capacitor is fully meshed.

## A-M ALIGNMENT\*

*Signal-Generator.* For all alignment operations, connect the low side of the signal generator to the receiver chassis, and keep the generator output as low as possible to avoid A.V.C. action.

*Output Meter Alignment.* If this method is used, connect the meter across the voice coil, and turn the receiver volume control to maximum.

Steps	Connect the high side of the Test Osc. To	Tune Test Osc. to	Range Switch	Turn Radio Dial To	Adjust the following for maximum output
1	Mixer grid pin 1 of 6BA6 (V2)	455 kc	"A" band	High frequency end of dial.	**Top and bottom cores of $T_2$ and $T_4$ .
2	High side of loop primary.	1400 kc		1400 kc	Osc. — $C_5$ R.F. — $C_6$ Ant. — $C_7$
3	Through a dummy antenna comprising a 200-mmF capacitor.	600 kc		600 kc	Osc. — $L_{18}$ R.F. — $L_{13}$ Ant. — $L_6$
4	Repeat Steps 2 and 3 for maximum output.				
5	Disconnect dummy antenna and adjust antenna trimmer $C_7$ on loop when set is installed in cabinet.				
6	"C" band antenna terminal 3.	15.2 mc	"C" band	15.2 mc	Osc. — $C_8$ *** R.F. — $C_9$ *** Ant. — $C_{10}$
7	Through a dummy antenna comprising a 150-ohm resistor in series with a 25-to 30-mmF capacitor.	9.5 mc		9.5 mc	Osc. — $L_{17}$ R.F. — $L_{12}$ Ant. — $L_5$
8	Repeat Steps 6 and 7 for accurate alignment.				

\* It is absolutely necessary to adjust the 10.7-mc transformers before adjusting the 455-kc transformers.

\*\* It is necessary to alternately load the primary and secondary of each 455-kc I.F. transformer with 10,000 ohms while the opposite side of the same transformer is being adjusted.

\*\*\* To guard against the possibility of alignment of  $L_{17}$  and  $C_8$  to image frequencies, tune the test oscillator to 15.2 mc and turn the radio-dial to 15.2 mc. Then adjust the test oscillator to 16.11 mc (image frequency). By increasing the test oscillator output, a signal should be heard.

Tune the test oscillator to 9.5 mc and turn the radio dial to 9.5 mc; then adjust the test oscillator to 10.41 mc (image frequency). By increasing the test oscillator output, a signal should be heard. (If these image frequencies cannot be heard, the set is incorrectly aligned, therefore, repeat steps 6 and 7.)

0.05-mf and 0.003-mf condensers, prevents any of the I.F. voltages from reaching the A.V.C. line and the rest of the circuit.

**Receiver Alignment.** The step-by-step procedure for aligning an F-M receiver has already been noted. For the present RCA set, the reader should encounter no difficulty if the alignment procedure is tabulated in chart form. The first chart outlines the adjustment of the ratio detector, the second chart is for the remainder of the F-M circuits, and the third chart is for the A-M bands. The notes included with each chart tend to clarify the procedure and to indicate special precautions.<sup>1</sup>

<sup>1</sup> The diagrams of the receiver and the alignment procedure are used through the courtesy of the Radio Corporation of America.



## CHAPTER 14

### SERVICING F-M RECEIVERS

**Extension of A-M Servicing Methods.** The servicing of F-M receivers conforms essentially to the basic procedures currently in use for A-M sets, but has been modified sufficiently to meet the altered construction of the F-M receiver. In making a comparison of the function of corresponding stages in both types of receivers, it will be found that, except for the second detectors and limiters (if any), there is no functional difference between corresponding stages of the two receivers. 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, 88–108 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 each 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

$\rho$  = 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 increase. 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 common knowledge 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 governed 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. 14.1). Each small section

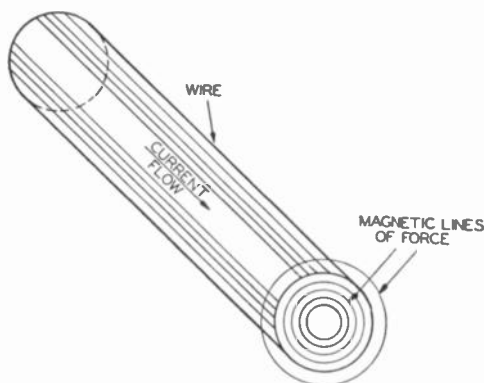


FIG. 14.1. Current flow through a wire and the magnetic lines of force encircling the wire.

of current flowing through the wire has magnetic lines of flux encircling it, but the sections of current at the outer edges of the wire obviously 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 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 item of interest to the serviceman is the fact that at the high frequencies, the equivalent circuit of so-called "pure" resistances becomes as shown in Fig. 14.2. The resistive and inductive affects of a resistor are due, of course, to the previous explanation. The capacity arises from the capacity between the terminals and the capacity between parts of the resistor itself.

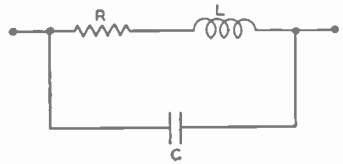


FIG. 14.2. The equivalent circuit of an ordinary resistor at higher frequencies.

*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 condenser, would not add any significant amount of inductance. This is evident from an examination of the expression governing the inductance of a straight length of copper wire:

$$L = 0.0051s \left( 2.303 \log_{10} \frac{4s}{d} - 0.75 \right) \text{ microhenries}$$

where  $s$  = length of wire in inches  
 $d$  = diameter of wire in inches

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. As the signal frequencies reach 100 megacycles, 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 condensers introduce inductance into the circuit and we find that trouble arises because of the interaction of the magnetic fields set up between ganged condensers connecting to different circuits. It is common practice to shunt paper and electrolytic condensers with small mica condensers in order to neutralize the 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. Condensers 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 frequency 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.

*Tube Operation at the Higher Frequencies:* Not only are resistors, coils and condensers subject to modification with frequency, but also the conventional vacuum tube. At the low frequencies, the time it takes an electron to travel the interelectrode distance is negligible, because this time represents a very small portion of the interval necessary to complete one cycle of the grid or input a-c wave. As the frequency of the input wave is increased, however,

the time for one complete cycle of the a-c wave becomes less and less, and soon the electron transit time becomes comparable to the period of the alternating wave. An example will make this clearer. Suppose that it takes an electron 0.000,000,001 sec. to travel from the filament to the plate. This may be ignored when the frequency of the input wave to the grid is, say, 1000 cycles per sec., for then one complete cycle requires 0.001 sec. This is a very long interval compared to the electron transit time given just above. But suppose that the frequency of the a-c wave is raised to 100 megacycles. Now, one cycle is completed in 0.000,000,01 sec. which is not far from the time that it takes the electron to travel from filament to plate. When this happens, something occurs in the grid circuit that causes the grid-to-filament (or cathode, as the case may be) impedance to drop from its low-frequency value (which is so high it can be considered as infinite). This affects the whole grid tuning circuit because the grid-to-cathode impedance is really in parallel with this tuner. Thus, if this high resistance is lowered in value, the same result is obtained as though a low resistance were placed in parallel with the resonant circuit. The  $Q$  is lowered, and with it goes the efficiency of the circuit.

*Electron Flow — Revised:* In order to help visualize the entire process, some ideas on electron flow will be revised or, rather, extended. Ordinarily, it is unusual to think of current flow in either the plate or grid circuits until the electrons from the cathode hit these elements. Actually, however, this is the point that completes a cycle begun when the electron first left the cathode. The charge of the electron is negative, and at the moment it leaves the cathode the conditions shown in Fig. 14.3A prevail. Here it can be seen that on leaving the cathode the electron left an equal and opposite (positive) charge on the cathode. At the same time a very small positive charge is induced on the plate. This is because the cathode electron is slightly closer to the plate and an electron on the plate has been repelled a small distance from the anode surface. As the cathode electron gets closer and closer to the plate, the positive charge of this anode will correspondingly increase, and the electron that was originally at the plate will be farther and farther repelled from the anode toward the cathode through the wire (see Fig.

14.3B). When the cathode electron finally reaches the plate, it neutralizes the positive charge there, and likewise the anode electron by this time has reached the cathode. The positive charge on the cathode is now also neutralized. The circuit is then in equilibrium and the current flow has ceased (see Fig. 14.3C). Multiply this electron by the billions that actually flow and there are the comparatively large plate currents observed in tubes. Some theorists may demand that it be explained

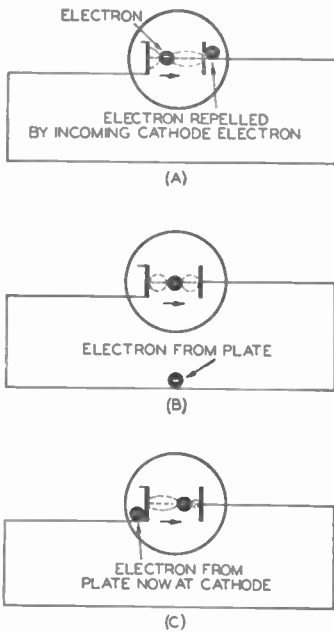


FIG. 14.3. A simplified illustration of the modified conception of current flow in a tube at the higher frequencies.

that the electron that left the plate be shown as only going a small distance and pushing other electrons in the plate lead until one finally goes to the cathode and neutralizes the positive charge left there by the cathode electron that left for the plate. The plate electron has been described as going the full distance from plate to cathode. Be that as it may, the fact still remains that, when the cathode electron reaches the plate, the current stops. This is the important thought to keep in mind.

The above concepts are not new but are seldom mentioned because, at the low frequencies, results obtained agree with the ordinary ideas of electron flow in tubes. With increase of frequency, however, the transit time effects require that this more general idea be used. Now the effect on the grid will be determined.

When an electron approaches the grid, there is induced in the grid a small charge, with the result that a small current starts to flow. This is due to the displacement of the electrons in the grid wires caused by the oncoming electron. If the time it takes any oncoming signal voltage to change is slow compared to the speed of the electron, the current induced in the grid by the approaching electron will be equal and opposite to the induced current when the

electron passes the grid and goes on to the plate. The two currents — that due to the approaching electron and that due to the same electron when it is on the other side of the grid — will thus cancel. As the frequency of the a-c grid voltage is increased, however, the signal voltage will change appreciably between the time the electron is approaching the grid and the time when it is receding on the other side of this element, and the induced currents due to this moving electron will modify the phase and amplitude of the signal. This, it has been found, causes losses in the grid circuit. This is known as the transit-time effect in tubes and begins to assume importance when the value of the signal voltage at the grid changes appreciably in the time it takes the electrons to move through the tube. A specific example might make this much clearer.

A wave, having a frequency of 100 mc/sec, has one tenth of a cycle occur in 0.000,000,001 sec. In ordinary tubes, the time of flight of an electron from the cathode to the plate is approximately 0.000,000,001 sec. Thus an appreciable change in grid voltage will occur while the electron moves from the cathode to the plate region. Any current induced in the grid by this electron as it approaches this element will occur at a different time in the a-c grid voltage cycle than when the electron recedes from the grid on the other side. It is this effect that gives rise to losses in the grid circuit. The end result is a lowering of the usually high resistance between the grid and cathode. If the grid-to-cathode resistance drops, any tuned circuit placed between these elements will be affected and the selectivity will drop.

Electron transit-time action in a tube is a complicated concept to grasp and, yet, it is very important. Perhaps another impression, stated a little differently, might give rise to a mental picture of what takes place. Consider Fig. 14.4A, where the cathode and grid are considered as the plates of a small condenser equal in value to the interelectrode capacitance. The a-c generator inserted across this condenser is the input signal. Now, it is well known that, if the current responds instantly to voltage changes across a condenser, the current will lead the voltage by  $90^\circ$ . The graph of Fig. 14.4B shows this. But suppose, for one reason or another, that the current cannot follow the voltage variations instantly. Then the current will not be  $90^\circ$  out of phase with the voltage but will have

some other angle, something less than  $90^\circ$ . This is equivalent to a condenser and resistor combination, such as shown in Fig. 14.4C.

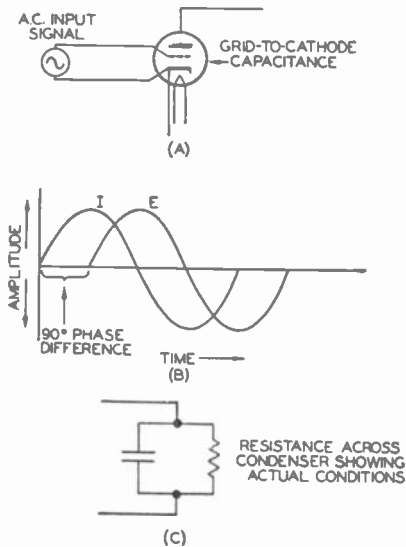


FIG. 14.4. Another method of illustrating the effect of transit time on the input impedance of a tube.

lower its  $Q$  and selectivity and increase its losses. The tuning circuit may even stop functioning as such if the resistance becomes small enough. It is because of this last-mentioned effect that the transit time in a tube is important.

Several formulas have been derived to give the exact relationship between conductance of the grid and the frequency. One such relationship is given here and is due to W. R. Ferris:

$$G_g = K g_m f^2 T^2$$

where  $G_g$  = the input conductance between the grid and cathode (conductance is the reciprocal of resistance)

$K$  = a constant of the tube

$g_m$  = the mutual conductance of the tube

$f$  = the frequency of the input signal

$T$  = the time it takes an electron to move from the cathode to the plate.

For the first case (A) we can see that, if the current and voltage are  $90^\circ$  out of phase, the circuit is purely capacitive. But if the angle is something less than  $90^\circ$ , the equivalent circuit consists of a condenser with a resistor shunted across it. We have the same effect when the electron transit time modifies the signal phase and amplitude.

To carry the illustration one step farther, replace the a-c grid generator by a resonant circuit which can produce the same type of voltage. The resistance shown in Fig. 14.4C and mentioned just above will be in parallel with this tuned circuit, thus tending to



Inspection of the equation reveals that the conductance increases rapidly as the frequency rises due to the fact that the frequency termed is squared.

*Minimizing Transit-Time Effects:* Since the resistance between the grid and the cathode decreases with frequency, and since this resistance shunts all tuned circuits placed between these two elements, it is desirable to attempt to minimize these adverse effects. One method which suggests itself is to increase the electron velocity until the time required to reach the plate from the cathode is made negligible in comparison with any alternating voltage applied to the grid. A means of accomplishing this consists in placing higher voltages on the plate to attract the electron more strongly, causing it to travel with greater speed. The electron, however, now that it has this additional speed, will hit the plate harder and cause it to become hotter. Means must be devised to cope with this increased plate dissipation.

As another possible method, we can move the electrodes closer together. This would again counteract transit-time effects of electrons because the electron has less distance to travel and any voltage placed on the plate will be more effective. There is, however, a limit to the spacing of the electrodes without having each affect the other adversely. Again, with elements closer together, less heat can be tolerated at the plate, and the amplification factor of the tube becomes less. Tubes which have been specifically designed for operation at 100 mc represent compromises of the foregoing conflicting factors.

**The Effect of Cathode Inductance.** Transit time of the electron is not the complete answer to the lowering of the input impedance of a tube at the higher frequencies. Of practically equal importance is the inductance existing in the cathode leads of the tube. The current of a tube must flow through the cathode lead wires and in the process develops a voltage across

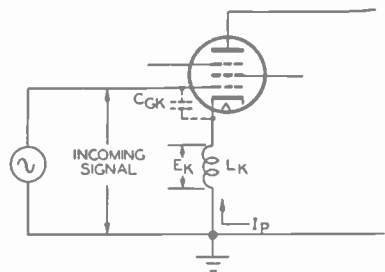


FIG. 14.5. Decrease of the input impedance of a tube is due, in part, to the inductance in the cathode lead of a tube.

this inductance (see Fig. 14.5). The alternating component of this current, like the d-c component, is impressed between the grid and the cathode. As a result, the effective signal voltage is decreased because of the opposition of the cathode lead voltage to the input signal voltage. The situation is analogous to inverse feedback. However, this voltage is present even though the cathode of the tube is grounded at the tube socket. The lead inductance occurs in the wires within the tube itself. The input resistance resulting from the voltage developed across the cathode lead inductance is

$$\text{Input resistance due to cathode inductance} = \frac{1}{\omega^2 g_m L_k C_{gk}}$$

where  $\omega = 2\pi \times$  frequency of the input signal

$g_m =$  mutual conductance of tube

$L_k =$  cathode lead inductance

$C_{gk} =$  interelectrode capacitance between grid and cathode

Note that the voltage which is developed across this cathode-lead inductance is due to the plate current. As far as the plate circuit is concerned, this voltage,  $E_k$ , is of little significance. It is at the grid, where the signal voltage is impressed that the effect becomes important.

To eliminate the effect of the lead inductance voltage on the input grid circuit, tube manufacturers have designed tubes with two wires leading directly from the cathode structure inside the tube to the tube base. In this manner, one terminal is available for the grid circuit and one for the plate circuit and its current and the two circuits are divorced from each other. In the circuit of Fig. 14.6, the 6AG5 R.F. amplifier tube possesses two cathode terminals. Even though both cathode terminals are grounded, pin 2 would be connected to the grid tuning condenser and coil. Pin 7 is the cathode connection for the plate circuit. To it would be connected the screen-grid and plate by-passing condensers. The d-c component of the plate current divides between both paths, but this is of no consequence since it does not contribute to the degenerative effect.

Servicemen should be cautious, in this respect, in accepting the connections of the two cathode terminals as shown on the manufac-

turer's schematic diagram. The diagram is not always an exact reproduction of the circuit, as laid out in the chassis, especially with regard to the separate connections of the same cathode. In many schematics, the cathode is grounded and the diagram indicates this fact without stating that separation exists as explained in Fig. 14.6.

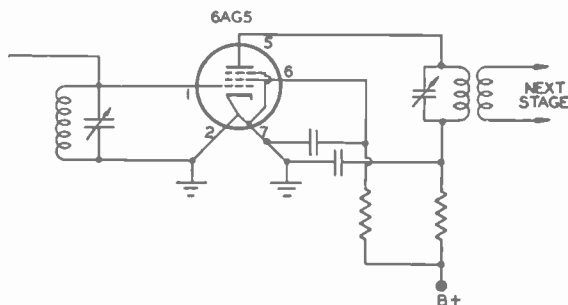


FIG. 14.6. The use of two cathode wires to eliminate the adverse effect of cathode inductance.

If someone, not familiar with the reason for the separate cathode terminals, connected them together, the result would be a decrease in receiver sensitivity due to a lower input resistance.

**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 successfully surmount 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 the less critical A-M receivers, it will seldom be 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. The standard octal socket can accommodate most tubes.

As a routine measure, the serviceman should replace all tubes in the set with test tubes known to be in good operating condition. 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 completely 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 rub his finger back and forth across the phonograph needle and note whether any scraping noise is heard from the loudspeaker.

These are but a few of the little tricks and preliminary tests that the serviceman has at his disposal to simplify the process of repairing receivers.

**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 oftentimes noisy. 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 condenser. 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:* There are three types of commercial F-M detectors currently in use and each was analyzed in Chapter 7. 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 the conventional discriminator, Fig. 14.7, connection of the meter across either load resistor will result in a deflection if the A-M signal voltage 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  $R_1$  or  $R_2$  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. 14.7. The ground terminal of the signal generator is connected to the receiver chassis. The output lead is connected through a 0.05-mf condenser 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.

The linearity of the detector to F-M signals can be determined by connecting the voltmeter across both load resistors and observing

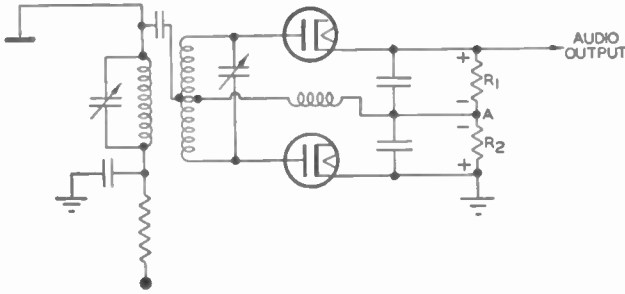


FIG. 14.7. A conventional discriminator.

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  $\pm 75$  kc from this I.F. point.

In the ratio detector, reproduced in Fig. 14.8, the voltmeter is connected between the point marked A.V.C. and ground. Since

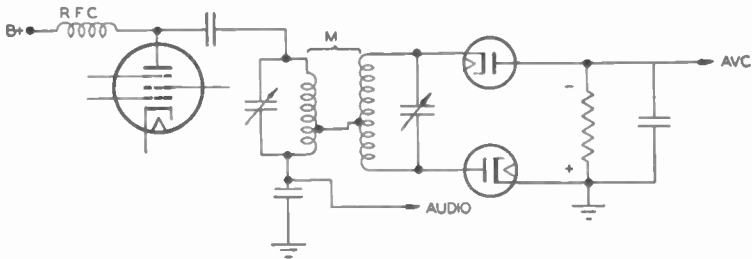


FIG. 14.8. The ratio detector.

the A.V.C. point 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 signal generator is connected to point *D*, with a 0.05-mf condenser in series with the output lead of the generator. The ground terminal of the generator connects directly to the receiver chassis. For this test the output of the generator should be, at least, 0.5 volt. If it is not possible to obtain this much signal voltage, the signal generator will have to be connected to the grid of the last I.F. amplifier, or even the one previous to that. It is understood, of course, that these stages are functioning properly.

Connect the d-c voltmeter across  $R_1$ , observing the marked polarity. Use an unmodulated signal. Now swing the generator frequency approximately 80 kc above to 80 kc below the 9.1 mc I.F. The d-c voltage across  $R_1$  is normally 15 volts and this exists either in the absence of a signal or when the signal is at 9.1 mc (representing no modulation). At 80 kc above 9.1 mc, the reading becomes 8 volts; it is 23 volts at 80 kc below 9.1 mc. If these indications are not obtained, it is a sign that trouble exists within this portion of the receiver.

The next step, if the foregoing test indicates that further testing is necessary, is to connect an audio signal generator to point *D*, Fig. 14.9. A separate generator is not necessary if the R.F. signal generator can provide any audio note. It is customary to incorporate a 400-cycle audio note in most good R.F. signal generators. With the audio signal applied to point *D*, a loud, clear signal should be heard from the speaker. This indicates that any audio voltage appearing in the plate circuit of the F-M detector is able to reach the audio amplifiers. If an audio output is absent, check the plate voltage of the F-M detector tube. After the d-c potentials have been tested — and found to be correct — then the coupling condensers and resistors leading to the audio channel are next. Resistances are tested by means of continuity checks; condensers are best tested by direct substitution.

It is understood that each step in the procedure is given with the thought that the trouble exists at some point which has not yet been reached. In other words, if the second step just given reveals the source of trouble in the F-M detector there is obviously no need to continue. However, if the indications fail to reveal the defect, the next step in the testing of the detector must be taken.

For the third test, short-circuit pin 2 of the F-M 1000 tube to the chassis. The tube is now functioning as an A-M detector. Connect the R.F. generator to point *D*, Fig. 14.9, with the audio modulation in the "on" position. The incoming amplitude-modulated signal will be detected, and a loud, clear audio note heard from the speaker. The signal generator is set to 9.1 mc and the signal voltage should be, at least, 0.5 volt. Absence of the audio output means that there is something defective in this circuit. Check tube, plate and screen voltages, condensers and resistors.

A final check, providing the receiver is still not working, is at the F-M oscillator. Possibly the simplest test to determine whether any oscillator is functioning, is to measure its grid-bias potential. Oscillators used in superheterodyne receivers are self-biasing, using a grid-leak bias form of circuit. A voltage will be present at this grid — with respect to ground — only when the oscillator is operating. The grid (pin 2) should be approximately 2.5 volts negative. Note, however, that presence of the negative voltage does not indicate that the oscillator is on frequency. It is merely an indication that the oscillator is functioning.

The proper indication, in all of these tests, means that the F-M detector is operating normally.

*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, no such limiter exists. In this instance, 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.

In a Philco receiver, using their special detector circuit, it becomes necessary to convert the F-M detector to an A-M detector in order to be able to use an A-M signal for testing. When pin 2 of the F-M 1000 tube is connected to the chassis, the detector-oscillator becomes inoperative, and the circuit responds to A-M signals. As an indicator, we can place an output meter across the voice coil of the speaker or else listen for an audio note from the loud-speaker, providing, of course, that a modulated A-M signal is used.

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-condenser 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 (see Philco diagram). A tube has recently become available, a 6SB7, which will function satisfactorily as a converter at frequencies of 100 mc. In the past, it has usually been necessary to have separate tubes for the mixing and oscillator functions, in order to avoid undesirable interaction between the two. However, with the appearance of special tubes, such as the 6SB7, it is likely that all manufacturers will eventually use a single tube. At the present writing, they appear to be evenly divided between one tube and two tubes.

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 for this 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 or not 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 condensers. Their purpose is to compensate for changes in the electrical characteristics of other condensers and/or the coils. If one of these special condensers 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.

Another defect, which is relatively new, is the howl that is heard on the F-M band. This howl is caused by the speaker vibrating the oscillator tuning slug. A fiber spacer placed between the oscillator and the mixer slug shafts will eliminate the vibration. This same effect may only become noticeable as the volume control is advanced, but the remedy remains the same. Howling may also be due to speaker vibrations reaching the oscillator sections of the gang condenser. Mounting the gang condenser on a rubber cushion or a felt strip will help to eliminate this howl. Also, floating the speaker on rubber grommets is helpful.

## F-M CONVERSION CHART

(88 to 108 mc)

Fre- quency	Channel Number	Fre- quency	Channel Number	Fre- quency	Channel Number	Fre- quency	Channel Number
88.1	201	93.1	226	98.1	251	103.1	276
88.3	202	93.3	227	98.3	252	103.3	277
88.5	203	93.5	228	98.5	253	103.5	278
88.7	204	93.7	229	98.7	254	103.7	279
88.9	205	93.9	230	98.9	255	103.9	280
89.1	206	94.1	231	99.1	256	104.1	281
89.3	207	94.3	232	99.3	257	104.3	282
89.5	208	94.5	233	99.5	258	104.5	283
89.7	209	94.7	234	99.7	259	104.7	284
89.9	210	94.9	235	99.9	260	104.9	285
90.1	211	95.1	236	100.1	261	105.1	286
90.3	212	95.3	237	100.3	262	105.3	287
90.5	213	95.5	238	100.5	263	105.5	288
90.7	214	95.7	239	100.7	264	105.7	289
90.9	215	95.9	240	100.9	265	105.9	290
91.1	216	96.1	241	101.1	266	106.1	291
91.3	217	96.3	242	101.3	267	106.3	292
91.5	218	96.5	243	101.5	268	106.5	293
91.7	219	96.7	244	101.7	269	106.7	294
91.9	220	96.9	245	101.9	270	106.9	295
92.1	221	97.1	246	102.1	271	107.1	296
92.3	222	97.3	247	102.3	272	107.3	297
92.5	223	97.5	248	102.5	273	107.5	298
92.7	224	97.7	249	102.7	274	107.7	299
92.9	225	97.9	250	102.9	275	107.9	300

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.

# QUESTIONS AND PROBLEMS

## CHAPTER 1

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  $\pm 75$  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. In the foregoing example, what would be the relative amplitude of each sideband?

17. Assuming that the unmodulated carrier has a value of 20 amperes, compute the exact value of each sideband and the carrier during the foregoing modulation.

18. It was determined that a carrier and its sidebands possessed the following current values.

Carrier .....	13.00 amps.
1st set .....	16.95 amps.
2nd set .....	24.31 amps.
3rd set .....	15.45 amps.
4th set .....	6.60 amps.
5th set .....	2.15 amps.
6th set .....	0.57 amps.
7th set .....	0.03 amps.

What was the value of the unmodulated carrier?

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

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

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

21. 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 15 kc, 9.375 kc, and 7.5 kc.

22. In the previous example, compute the actual values of the amplitudes for each of the three frequencies given, assuming that the unmodulated carrier value was 60 amperes. Draw these sidebands, with their carrier, to scale.

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

24. 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

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  $50^\circ$ . 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  $180^\circ$  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

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  $60^\circ$ ?
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 components be employed for A-M transmission? Explain.

#### CHAPTER 4

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 capacitively 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. 4.3, 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? Illustrate your answer using the circuits in Chapter 13.
12. Which three oscillator circuits are commonly employed in F-M receivers? Draw one and couple its output capacitively 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 5

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 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 is found in the I.F. amplifiers of an F-M receiver?
25. How are most I.F. stages tuned? Why?

## CHAPTER 6

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. Are special tuning devices necessary in F-M receivers for proper reception of signals? Explain.
19. What types of tuning devices are used? How do they function?
20. Draw the circuit of a simple tuning device connected to the limiter.
21. Explain how the circuit drawn in Question 20 operates.
22. What limitations does this method of connecting a tuning device possess?
23. Draw the schematic diagram of a tuning device connected to the discriminator output.
24. Explain the operation of the circuit drawn for Question 23.
25. Why are interstation noise suppression systems used? What is the basic idea behind most of these systems?
26. Draw the schematic diagram of a simple silencer circuit and explain its operation.
27. What type of tube would be most desirable in silencer networks? Why?
28. Does the silencer network function when a loud burst of interference is being received? Explain your answer.
29. Draw the schematic circuit of the I.F. and limiter section of an F-M receiver. Use two I.F. stages, two resistance-coupled limiter stages, and a tuning device.

## CHAPTER 7

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. Describe the exact function of each component in the discriminator.
9. Are there any modifications that can be made in the circuit? Explain.
10. Draw the schematic diagram for an F-M discriminator using a duo-diode tube having a common cathode.
11. What limitations of the Foster-Seeley type of discriminator resulted in the development of a ratio detector?
12. Draw the circuit of a ratio detector.
13. Explain briefly the operation of a ratio detector.
14. Describe the operation of the Beers locked-in discriminator circuit.
15. Why does the discriminator used with this arrangement possess a limited frequency range?
16. Draw the schematic circuit of the Beers F-M discriminator system.
17. What is the purpose of the 6SA7 tube used in the Beers arrangement?
18. What happens in the Beers discriminator when the incoming signal deviates too far from its center frequency? How does this affect the audio output? Contrast this with the effect obtained on the Foster-Seeley discriminator using the same off-frequency signal.
19. Which of the foregoing arrangements does the Philco F-M detector resemble? Why?
20. Draw the circuit diagram for the Philco F-M detector.
21. What is the purpose of the quadrature circuit in the Philco circuit?
22. Explain the operation of the Philco F-M detector.
23. Can the Philco circuit and the Beers arrangement be used interchangeably? Explain your answer.
24. Compare the response curves of all the F-M detectors described in Chapter 7. From this, what conclusion can we draw concerning the response curves of F-M detectors now in development?

## CHAPTER 8

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 receivers? What other type is there? Which is the most desirable?
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. What is automatic tone compensation? Why is it used? How is it achieved?

## CHAPTER 9

1. What three 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  $90^\circ$  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. What advantages do grounded-grid amplifiers possess?
16. Draw the circuit of a grounded-grid amplifier. Could such a unit be employed in an F-M receiver? Explain.
17. Explain briefly the operation of the frequency-control circuit in the RCA F-M transmitter. Illustrate by means of a block diagram.
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. Why are grounded-grid amplifiers used only in high-frequency circuits? Could they be used in standard broadcast A-M receivers? Explain.
22. 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 10

1. What is the Armstrong system of generating frequency modulation? Why is it useful?
2. What factors govern the amount of F-M produced through phase modulation? How are these factors combined in a formula?
3. Draw a block diagram of the basic Armstrong system. Explain the function of each stage.
4. What limitations must be observed when employing phase modulation to produce F-M? How does the Armstrong system observe these limitations?
5. What is a predistorter network? Draw a diagram of such a network and connect it to an amplifier.
6. Why is a predistorter network required in the Armstrong modulator?
7. When a full phase swing of  $30^\circ$  is utilized, how much equivalent F-M is produced when the audio signal frequency is 1000 cycles?
8. Differentiate between predistorter networks and accentuator circuits. Where is each used?
9. Why is predistorter circuit design based upon the lowest audio frequency used in the modulator?
10. 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?
11. How does the Armstrong Dual Channel Modulator differ from the basic circuit?
12. What stage determines the accuracy of the final carrier frequency in the Dual Channel Modulator Transmitter? Explain.

13. Explain by means of a block diagram the operation of the Dual Channel Transmitter.
14. What type of modulation actually occurs in the Phasitron tube? Why?
15. Describe briefly the operation of the Phasitron tube.
16. How is F-M produced with the Phasitron tube?
17. Would a predistorer network be required in the Phasitron circuit? Explain.
18. What advantages does the Armstrong Dual Channel Transmitter possess over the basic Armstrong circuit?
19. 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.
20. What does an F-M receiver contain to counteract the effect of the predistorer and attenuator networks in the transmitter? Explain your answer in detail.

### CHAPTER 11

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 double-peak 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 saw-tooth 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 saw-tooth 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?

### CHAPTER 12

1. Why do conventional-type tuning circuits prove unsuitable for high frequency receivers?

2. What general characteristics should high frequency tuners possess?
3. How does the Mallory Inductuner operate?
4. Draw the circuit schematic of a high frequency oscillator circuit using the Inductuner. Explain how this circuit is aligned.
5. Explain why inductance tuning is better suited for high frequency circuits than condenser tuning.
6. What ordinarily happens to the  $Q$  of a tuning circuit as its frequency is raised? Why?
7. How does permeability tuning differ from the method employed in the Inductuner? What is slug tuning?
8. Why is a high  $L$  to  $C$  ratio desirable in a tuning circuit? When is it purposely lowered?
9. Why do we find four parallel wires used on permeability-tuned coils? How could a single wire be made to produce the same effect?
10. What is the reason for the uneven turn spacing found on some coils?
11. How does the "guillotine tuner" differ from the Inductuner and the permeability tuner?
12. Describe the structure and operation of the G.E. "guillotine tuner."
13. What properties of a transmission line enables it to be used for tuning?
14. What advantages are offered by the use of a low intermediate frequency? What disadvantages?
15. Why does the choice of a coaxial transmission line tuner permit us to use a low intermediate frequency?
16. What is the double superheterodyne principle?
17. Draw the schematic diagram of a coaxial tuned circuit employed in a double superheterodyne receiver.

## CHAPTER 13

1. Why are band switches employed in modern receivers? What are their disadvantages, especially in high frequency receivers?
2. What are the average gains per stage that can be expected in F-M receivers? Start at the antenna and work toward the F-M detector.
3. Why are pentagrid converters used sparingly in high frequency receivers?
4. How can one I.F. system be made to serve both F-M and A-M signals? Draw the schematic diagram of a 2-stage I.F. amplifier suitable for serving 10.7-mc F-M signals and 455-kc A-M signals.
5. In A-M-F-M combination receivers, the first I.F. coil not in use is shorted out. Explain. Why is the same procedure unnecessary in subsequent stages?
6. How is a de-emphasis filter added to an F-M discriminator?
7. How many actual audio amplifiers are there in Fig. 13.1, p. 292?
8. List the recommended order of alignment in the Stromberg-Carlson A-M-F-M receiver.
9. Describe in detail the alignment of the F-M-I.F. system of the Stromberg-Carlson receiver shown in Fig. 13.5.
10. How are overcoupled I.F. transformers aligned? What effect does the type of indicator used have on the method of alignment? Explain.
11. How can the output of a signal generator be loosely coupled to the input terminals of a receiver?
12. Draw the F-M detector network used in the Philco A-M-F-M receiver of Fig. 13.7.
13. Draw the A-M detector network of the same receiver.



14. How is the A.V.C. voltage of this receiver obtained? On what bands does it function?
15. What similarities exist between the Philco and Stromberg-Carlson sets?
16. Draw the R.F. section of the receiver when the receiver is set for A-M. For F-M.
17. When an A-M signal generator is used to align the F-M circuits of the Philco receiver, what changes must be made in the circuit?
18. Describe the alignment of the Philco F-M detector using an A-M signal generator and an indicating meter.
19. In what respects does the I.F. system of the Philco receiver differ from the I.F. system of the Stromberg-Carlson set?
20. What is the exact procedure for aligning the R.F. stages of the Philco receiver?
21. Draw the schematic diagram of the ratio detector used in the RCA model 612V3 receiver.
22. Illustrate how one of the I.F. amplifiers also serves as a phono amplifier.
23. Through which stages must all F-M signals pass? All A-M signals?
24. What is the purpose of the driver stage?
25. Explain, in detail, the alignment of the ratio detector.
26. Where is the indicator placed when the F-M-I.F. stages are being adjusted? Why is this position chosen?
27. Describe the alignment of the A-M sections of any one of the three receivers discussed in this chapter.
28. How would we align the A-M stages of the RCA receiver, given a sweep signal generator and an oscilloscope?
29. Trace out and draw the circuits used to receive the short-wave band. Only the R.F. amplifier, mixer and oscillator circuits need be drawn.
30. Compare all three receivers as to the number of similar stages and dissimilar stages.

## CHAPTER 14

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. Why is the conventional concept of current flow in a vacuum tube unsuitable at the high frequencies?
5. What effect does a tube have on its associated network as we raise the frequency of the signal?
6. In what ways are the newer tubes better suited for high frequency operation than the older, conventional tube?
7. Define electron transit time. What is its effect on the operation of a tube?
8. Why are some tubes designed with two cathode terminals?
9. Illustrate how tubes possessing two cathode terminals should be connected into a circuit.
10. What preliminary tests should be made before any actual work of servicing is begun?
11. How is the audio system of a receiver tested?
12. Outline the method you would employ to check the Foster-Seeley discriminator, the ratio detector and Philco F-M detector.
13. Would a-c hum in the power supply affect an F-M receiver? Explain.

14. How does testing of the I.F. system depend upon the type of F-M detector used?
15. Must the indicating meter always be placed in the output of the F-M detector when the I.F. system is being tested? Explain.
16. List some of the defects which could possibly affect an I.F. system.
17. How is the R.F. section of an F-M receiver tested?
18. What indication do we obtain from a normally functioning oscillator? •
19. 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?
20. Microphonic tubes produce what effect in a set?

## INDEX

- Adjacent channel interference, 46
- Alignment, discriminator, 246, 263, 302
  - double-peak transformer, 251
  - I.F. system, 248, 299
  - point-to-point method, 246
  - R.F. section, 254, 265
  - receiver, 245, 297, 308, 317
  - visual method, 256, 303
- Amplifiers, adjustment, 93
  - grounded grid, 200
  - I.F., 78
- Amplitude, modulation, 4
  - sidebands, 7
  - vectors, 30
- Armstrong system, 215
  - dual-channel, 224
  - frequency swing compensation, 219
- Audio, amplifiers, 157, 307
  - audio balance, 161
  - bandwidth limits, 162
  - high fidelity, 157
  - negative feed-back, 169
  - phase inverters, 163
  - push-pull, 163
  - servicing, 334
  - tone control, 172
- Audio balance, 161
- Automatic tone compensation, 177
- Automatic volume control, 120, 306
  
- Balanced modulators, 189, 206
- Band switching, 291
- Bass-boost, 173
- Beers receiver, 147
  - locked-in oscillator, 149
- Bessel function charts, 18, 21, 22
  
- Capacitance at high frequencies, 324
- Cascaded limiters, 106
- Circuit selectivity, 276
- Coaxial tuner, 283
- Commercial receivers, 291
- Converter, frequency, 61
  - gain, 68
- Coupled circuits, 126
- Coupling, oscillator, 68
- Cross-over networks, 177
- Crystals, 74
  - temperature coefficient, 75
- Crystal oscillator, 71
  - grid-plate circuit, 72
  
- De-emphasis, 55
- Detectors, 10, 122, 291
  - Philco, 152, 304
  - ratio, 140, 311
- Deviation ratio, 18
  - interference suppression, 38
  - noise, 53
- Discriminator, 10, 122, 291
  - alignment, 246, 258, 263, 302
  - frequency centering system, 192
  - G.E. model, 138
  - modified, 125, 136
  - operation, 122, 131
  - servicing, 334
- Dual channel Armstrong transmitter, 224
  
- Electron-coupled Hartley oscillator, 70
- Electron transit time, 327
  - effect of cathode inductance, 329

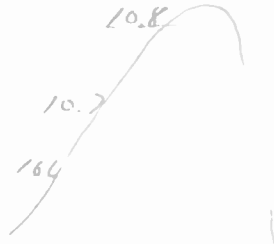
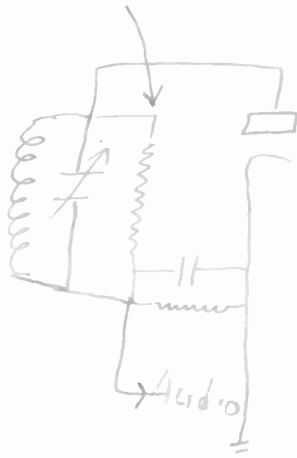
- Figure of merit, 91
- F-M detector, 122, 131  
 Beers, 122, 131  
 Philco, 152, 304  
 ratio, 140  
 servicing, 334
- F-M and interference, 38  
 frequency separation of interfering signals, 42  
 random noise, 51  
 suppression of interference, 43
- F-M oscillators, 181, 184
- F-M receivers, guillotine tuner, 280  
 Inductuners, 272  
 noise response, 52  
 permeability tuning, 275  
 Philco, 304  
 RCA, 311  
 R.F. tuners, 271  
 servicing, 321  
 Stromberg-Carlson, 292  
 tuning, 109
- F-M signals, deviation ratio, 18  
 frequency variation, 9  
 from phase modulation, 32  
 guardbands, 11, 13  
 modulation index, 17  
 sidebands, 11  
 sideband power, 13  
 sideband variation, 15
- F-M transmitters, 181, 191  
 Armstrong system, 215  
 balanced modulators, 189  
 dual channel, 224  
 frequency centering system, 192  
 frequency control, 202  
 frequency division, 204  
 frequency multipliers, 191, 193, 198, 221  
 grounded-grid amplifiers, 200  
 modulators, 184  
 Phasitron, 234  
 RCA transmitter, 196  
 reactance-tube, 181
- Frequency centering system, 192, 209
- Frequency control, 202, 206  
 motor control, 203
- Frequency converter, 61  
 gain, 68
- Frequency division, 204
- Frequency drift, 63, 230  
 countermeasures, 6+
- Frequency mixer, 61
- Frequency multipliers, 191, 193, 198, 221
- Gain, I.F. amplifier, 87
- Grid-leak bias limiter, 102
- Grounded-grid amplifier, 200
- Guardbands, 11, 13, 48
- Guillotine tuner, 280
- Harmonic spurious responses, 85
- Hartley oscillator, 70
- Hearing response characteristics, 158
- High-fidelity, 157
- High-frequency oscillators, 61
- Hum, 56
- I.F. amplifiers, alignment, 248  
 servicing, 338  
 visual, 26+
- Image response, 80
- Impedance-coupled limiters, 108
- Impulse noise, 56
- Indirect F.M., 34, 211
- Inductance, 323
- Inductuners, 272
- Interference, F-M, 38  
 frequency separation of interfering signals, 42
- Intermediate frequency amplifiers, 78, 9  
 adjustment, 93  
 design factors of I.F. amplifiers, 78  
 gain, 79, 87  
 image response, 80

- Intermediate frequency amplifiers,  
selectivity, 79, 87  
spurious responses, 80
- Intermediate frequencies, choice of,  
79
- Interstation noise suppression, 117  
silencer tubes, 118
- Inverse feedback tone control, 175
- Limiters, 97**  
impedance-coupled, 108  
in cascade, 106  
operation, 97  
resistance-coupled, 108  
time constants, 103  
types of, 102
- Loading networks, 302, 308
- Locked-in oscillator, 149  
Philco single-stage detector, 152
- Lowered voltage limiter, 105
- Magic-eye tuning indicators, 109**
- Marker frequencies, 265
- Mixer, frequency, 61  
gain, 68
- Modulator, Armstrong, 215  
balanced, 189, 206  
dual channel, 224  
reactance-tube, 181, 184
- Modulation index, 17
- Mutual conductance, 91  
figure of merit, 91
- Negative-feed-back, 169**
- Noise, deviation ratio, 53  
impulse, 56  
limiters, 98
- Oscillator, 61**  
coupling, 68  
crystal, 71  
electron-coupled, 70  
frequency drift, 63
- Oscillator, frequency placement, 75  
Hartley, 70  
separate, 65  
servicing, 340  
stability, 62, 210  
tuned-grid tickler, 70
- Oscilloscope, 256  
alignment, 303, 310  
discriminator alignment, 258  
saw-tooth deflection, 261  
sine-wave deflection, 260
- Permeability tuning, 275**  
coaxial tuner, 283  
coil spacing, 278  
Guillotine tuner, 280  
R.F. selectivity, 277
- Phase inverters, 163
- Phase modulation, 25, 210  
Armstrong system, 215  
dual channel, 224  
frequency swing compensation, 219  
Phasitron, 234
- Phase shift, controlling factors, 27,  
210
- Phasitron, 234  
modulation, 239
- Philco F-M detector, 152, 304  
servicing, 335
- Philco receiver, 304
- Point-to-point alignment, 246  
I.F. system, 248
- Power supply servicing, 334
- Pre-emphasis, 55
- Push-pull amplifiers, 163
- Q of a coil, 78**  
**Q variation with frequency, 78**  
**Q of a tuned circuit, 87**  
**Q inductuner, 273**
- Random noise, 51**
- Ratio detector, 140, 311  
modifications, 143

- Ratio detector, servicing, 335
- RCA transmitter, 196
- Reactance tube, 181
  - RCA transmitter, 191, 196
- Receiver, noise response, 52
  - permeability tuning, 275
  - Philco, 304
  - RCA, 311
  - R.F. tuners, 271
  - Stromberg-Carlson, 291
  - tuning of, 109
- Receiver alignment, 245
  - double-peak transformers, 251
  - I.F. system, 248
  - point-to-point method, 246
  - R.F. section, 254
  - servicing, 321
- Receiver analysis, 291
  - Philco, 303
  - RCA, 311
  - Stromberg-Carlson, 293
- REL transmitter, 229
- R.F. tuners, 271
  - coaxial tuner, 283
  - Guillotine tuner, 280
  - inductuner, 272
  - permeability tuning, 275
- R.F. amplifiers, alignment, 254, 303
  - servicing, 339
  - visual, alignment, 265
- Resistance at high frequencies, 321
- Resistance-coupled limiters, 108
- Selectivity, 276
- Servicing F-M receivers, 321
  - audio amplifiers, 334
  - F-M detector, 334
  - general procedure, 331, 333
  - I.F. amplifiers, 338
  - power supply, 334
  - preliminary tests, 332
  - R.F. system, 339
- Sidebands, A-M, 7
  - F-M, 11
- Sidebands, significant, 19
  - variation, 15
- Signal generator, 259
- Silencer tubes, 118
- Speakers, 161
  - coaxial type, 161
  - frequency-dividing networks, 160
- Speaker cross-over networks, 177
- Spurious response, 80
  - direct I.F. response, 84
  - harmonic, 85
  - image response, 81
  - stations separated by I.F., 84
- Static, 49
- Thermal agitation, 49
  - random noise, 51
  - static, 49
  - tube hiss, 49
- Tone compensation, 177
- Tone control, 172
  - bass-boost, 173, 296
  - by inverse feedback, 175
  - treble, 173, 297
- Transmission lines, 285
  - application for tuning, 288
- Transmitters, 181, 191
  - Armstrong system, 215
  - balanced modulator, 189
  - dual channel, 224
  - frequency centering system, 192
  - frequency control, 202
  - frequency multipliers, 191, 193, 198, 221
  - grounded-grid amplifiers, 200
  - Phasitron, 234
  - RCA transmitter, 196
  - reactance-tube, 181
  - reactance-tube modulator, 184
- Treble control, 173
- Trouble shooting procedures, 331
- Tube hiss, 49
- Tube input impedance, 325, 330
  - cathode inductance, 329

Tube input impedance, electron transit time, 326  
 Tube operation at high frequencies, 324  
 Tuned-grid tickler oscillator, 70  
 Tuning indicators, 109  
 Vectors, 27

Vectors, applied to modulation, 31  
 Vector addition, 30  
 Visual alignment, 256  
   discriminator, 263  
   I.F. amplifiers, 264  
   marker frequencies, 265  
   oscilloscope, 256  
   R.F. system, 265



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