# F.M. RECEIVERS

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# STUDY SCHEDULE No. 48

**For each study step, read the assigned pages first at your usual speed, then reread slowly one or more times. Finish with one quick reading to fix the important facts firmly in your mind. Study each other step in this same way.**






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# General Characteristics of F.M. Receivers

PERHAPS the best way to start our study of f.m. receivers is to compare them with the conventional a.m. receivers with which you are already familiar.

All f.m. receivers are superheterodynes, because for satisfactory operation the receiver generally must have more gain and selectivity than can be obtained in a t.r.f. receiver at f.m. frequencies.

The preselector, local oscillator, and mixer-first detector stage of an f.m. receiver are similar to those of an a.m. receiver. The high frequency of operation (88-108 mc.), however, means that special tubes and circuits must be used to obtain the necessary gain and band width needed at these frequencies.

The i.f. frequency of f.m. receivers is much higher than that of a standard a.m. receiver. The primary reason for this is to reduce image interference. As you know, a signal separated from the desired signal by twice the i.f. frequency can cause image interference. Since the f.m. band is 20 mc. wide, an i.f. more than half this value will prevent any f.m. station from causing image interference with any other f.m. station. For this reason, an i.f. of 10.7 mc. is generally used. Another reason for using such a high i.f. value is that the necessary band width (150 kc.) can be obtained more easily at that frequency. To secure this band width at 455 kc., for example, the i.f. amplifier would have to be so very heavily resistance-loaded and broadly tuned that very little signal gain could be obtained.

It is in the second detector that an f.m. receiver differs considerably from an a.m. receiver. The purpose of the f.m. second detector is to convert frequency variations in the input signal (which occur at an audio rate) into an electrical wave form whose amplitude varies in accordance with the modulating signal. If the set is to enjoy the f.m. advantages of freedom from noise and freedom from interference, any amplitude variation in the input signal must not cause an amplitude variation in the output.

There are two general types of f.m. detectors. The first, of which the slope, double-tuned, and Foster-Seeley detectors are examples, are sensitive to both frequency and amplitude variations in the input signal. All f.m. receivers using these detectors must therefore have a limiter section that removes amplitude variations from the f.m. signal before it reaches the detector.

No limiter stages are needed in f.m.

receivers using the other general type of f.m. detector, because such detectors are insensitive to amplitude variations. The ratio and Bradley detectors are examples of this type.

The audio section of an f.m. receiver is generally a high-fidelity, low-distortion amplifier capable of reproducing the high-quality, noise-free signal transmitted in wide-band f.m. broadcasting. Since you have already studied such audio systems, we will not discuss them in this Lesson.

# Noise Reduction in F.M. Systems

Before we start our study of noise reduction in f.m. systems, we need first to review the kinds of interference encountered in a.m. systems.

## SOURCES OF INTERFERENCE

The principal disturbances in a.m. reception can be divided into random and impulse noises. Random noises, which cause hissing sounds in the receiver output, can be either thermal agitation noise, arising from the small potentials set up by the random motion of electrons in the conductors of the first stage of the receiver, or tube noise, caused by random fluctuations in the rate at which electrons arrive at the plates of the vacuum tubes in the early stages of the receiver.

Impulse noises cause crackling or popping sounds in the receiver output. They may be static arising from electrical discharge in the atmosphere, or man-made interference that occurs when there is spurious radiation from such sources as electrical power equipment and automobile ignition systems.

There are two other types of interference encountered in a.m.: first, interference resulting from the reception of signals from stations other than the one whose program is desired; and, second, hum modulation of the signal caused by inadequate plate supply filtering or by poorly designed or defective circuits in which a.c. is used to heat cathodes. These noises can generally be considered to be random noises.

In an f.m. system, *all these types of interference can be practically overcome or at least greatly reduced, provided the voltage of the desired f.m. signal is somewhat greater than the voltage of the disturbance.*

## ANALYSIS OF NOISE VOLTAGES

Since many noises encountered in a.m. and f.m. systems are random in nature, let us study random noise first.

**Random Noise.** It has been found experimentally (and can be proved mathematically) that when two or more noise components of equal amplitude are combined, the resultant noise is equal to the amplitude of any of them times the square root of the



number of components present. For example, 9 noise components will combine to form a resultant signal only 3 times as large as the amplitude of any of the original signals. Since noise components are distributed equally throughout the frequency spectrum, the resultant noise is proportional to the square root of the band width. Thus an r.f. channel 160 kc. wide will have 4 times as much random noise as a 10-kc. channel.

**Impulse Noise.** Impulse noise, that is, noise produced by an electrical surge or repeated surges (such as auto ignition, motor sparking, or lightning) consists of a fundamental component and a number of other components harmonically related to the original. Since these componets all start in phase (at the time the surge occurs), the amplitude of the pulse received is proportional to the number of components. Again assuming that the components are equally distributed in the frequency spectrum, we find that the impulse noise output is directly proportional to the band width. Hence, any given impulse will cause 16 times as much noise in a 160-kc. band as in a 10-kc. band.

## NOISE IN AN F.M. SYSTEM

In a discussion of noise in an f.m. system, it is convenient to consider each component of the noise frequency (whether random or impulse) as a separate carrier, of which there are many present at all times. Since a combination of two station carriers that differ in frequency is equivalent to a station carrier and a single noise voltage, both cases may be considered at the same time.

The effect is most easily shown by the vector diagram in Fig. 1, in which the carrier vector is assumed to rotate counterclockwise at the rate, or frequency, of one complete 360° rotation for every cycle, and the noise vector is assumed to rotate around the end of the carrier vector at a rate equal to the difference between the frequencies of the carrier and the noise. The resultant vector, which represents the combined carrier and

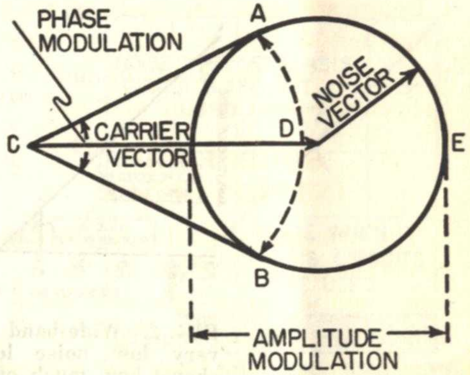


**FIG. 1. How an interfering signal or noise will combine with an f.m. carrier to produce amplitude and phase (frequency) modulation.**

noise voltages, varies continually both in length and in angle with respect to the carrier. The length variation corresponds to a variation in amplitude of the signal, and the variation in angle, as you learned in earlier Lessons, corresponds to a variation in frequency. Thus, the resultant voltage is both amplitude-modulated and frequency-modulated. Amplitude variations will be removed in the receiver, but the frequency variations will cause some noise in the receiver output.

The greater the difference between

the frequency of the carrier and the frequency of the noise, the greater the frequency deviation of the resultant voltage. Since the frequency deviation of the received signal determines the amplitude of the receiver output, it follows that the noise produced in the output of an f.m. receiver becomes louder the farther removed the noise frequency is from the carrier frequency.

This fact is shown in Fig. 2, in which line k-n-l-p-m represents the

In the reception of a.m. signals, there is no such reduction of noise at low frequencies. All noise components combine with the carrier equally, resulting in a rectangular noise spectrum. As a result, there is effectively an *18.75 db reduction in noise* in wide-band f.m. compared with a.m. for programs of similar quality—that is, for programs in which audio frequencies up to 15 kc. are used to modulate the carrier. Actually, of course, few a.m. stations use modu-

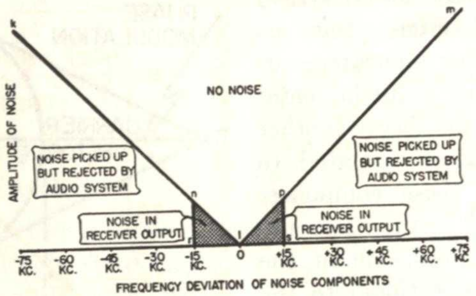


**FIG. 2. Wide-band f.m. signals have very low noise levels. This diagram shows how much of the noise picked up is inherently rejected by wide-band f.m.**

amplitude of the noise caused by noise components of various frequencies. As you can see, the noise produced becomes greater in amplitude as the frequency deviation (the difference between the noise frequency and the carrier frequency) increases. However, noise above a 15-kc. deviation is of no practical account, because the audio amplifier of the usual f.m. receiver will not pass frequencies above 15-kc. The actual noise in the receiver output is therefore represented by the shaded triangles n-r-l and l-p-s. As you can see, there is a considerable reduction in noise at the low audio frequencies.

lating frequencies higher than 5 kc. The reduced frequency range of the usual a.m. station reduces the noise output of a.m. receivers; therefore, if you compare the usual 15-kc.-modulated f.m. program with the usual 5-kc.-modulated a.m. program, the noise reduction in the former is not as great as 18.75 db.

**Comparative Interference in A.M. and F.M.** It has been found that if two a.m. stations are on the same carrier frequency, the signal of the desired station must be 100 times (40 db more than) the amplitude of the undesired signal for the interference to be unnoticeable. In narrow-band



(±15 kc.) f.m., used in mobile services, the similar ratio is 24 db or 10.5 times the amplitude. In wide-band f.m., the desired signal need be only twice the amplitude (6 db) of the undesired signal. For this reason, stations assigned the same frequency can be much closer together without causing interference if they are f.m. stations than they can be if they are a.m. stations.

## NOISE REDUCTION BY PRE-EMPHASIS AND DE-EMPHASIS

In the above discussion of the effects of interfering signals and noise components, you learned that noise and interference effects in f.m. systems are very much less than in a.m. systems. Furthermore, the residue of these disturbances that appears in the f.m. receiver output is, as shown by the height of curve A in Fig. 3, concentrated in the upper audio-frequency range. Since noise frequencies in the upper register are more irritating to the human ear, the noise concentrated at high frequencies is more objectionable than the same amount of noise energy uniformly distributed over the whole audio-frequency range. This unfavorable situation can be easily corrected in f.m. circuits by the use of pre-emphasis and de-emphasis.

Pre-emphasis refers to the use of a single network in the audio system of the transmitter to amplify the higher frequency components of the program much more than the lower frequency components. The standard 75-microsecond pre-emphasis curve calls for a gain - versus - frequency

characteristic that is flat to 500 cycles, then rises to 17 db at 15,000 cycles. Such a characteristic is shown in Fig. 4. Since a 17-db increase represents a voltage step-up of about 7, the frequency swing of the transmitter on a 15,000-cycle sound is 7 times as great as without pre-emphasis; therefore, the intelligence of the modulation overshadows noise in the receiver output far more effectively when pre-emphasis is used.

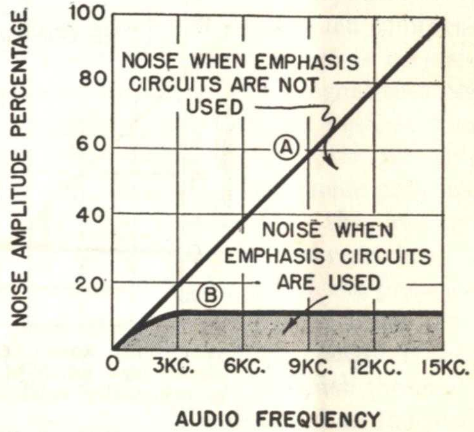


**FIG. 3. Pre-emphasis in an f.m. transmitter and corresponding de-emphasis in an f.m. receiver will greatly reduce the noise in the output of the f.m. receiver. The area under line A represents the noise when no emphasis circuits are used; the area under curve B represents the noise with pre-emphasis and de-emphasis.**

There is, however, some danger of over-modulating the transmitter seriously on the high frequencies with pre-emphasis, because the energy content of the high-frequency components of some program material is as much as that of the low-frequency components, particularly in the case of drums, cymbals, gourds, tambourines, muted trumpets, bells, and chimes. Spanish or South American music must be controlled at lower levels than other types of music for this reason.

In the receiver, a simple de-emphasis network is used following the detector to bring the highs down to a proper relationship with the lows. The gain-frequency characteristic of this network is the exact inverse of that of the pre-emphasis network; for example, the amplitude of a 15,000-cycle component is reduced to 1/7 of the value fed into the de-emphasis network. As a result, the highs and lows of the audio signal are brought into a proper relationship. At the same time, the high-frequency noise is reduced by the de-emphasis network; noise frequencies around 15 kc. are reduced to 1/7 their original amplitudes, others of lower frequencies are reduced somewhat less.

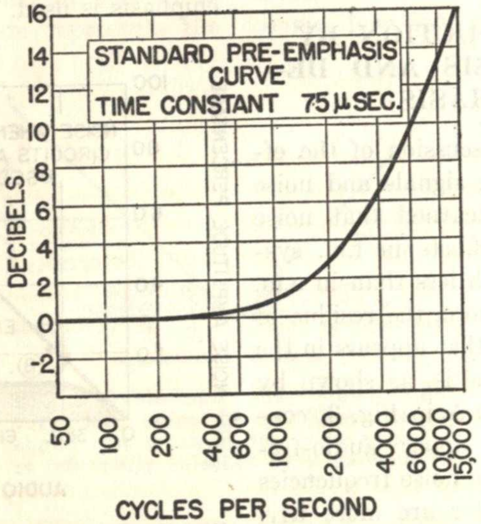


**FIG. 4. The standard pre-emphasis curve used in f.m. wide-band broadcast services.**

The result is shown in Fig. 3, which shows the noise found at various frequencies when emphasis circuits are not used (curve A) and when pre-emphasis and de-emphasis networks are used (curve B). As you can see, the greatest amount of noise reduction occurs at the highest frequencies; in other words, the pre-emphasis and de-emphasis are most effective in reducing the amplitudes of the noise components at the frequencies where the noise is most likely to be annoying.

## HIGH FIDELITY POSSIBLE IN F.M.

In the discussion of noise reduction above, we used a typical value of 15,000 cycles for the highest frequency handled by the audio amplifier of the f.m. receiver. This is in contrast to the 5,000 cycles or less found in the typical a.m. receiver.

There are two chief reasons why f.m. sets are designed to reproduce higher audio frequencies than a.m. sets are. One is that a.m. receiver circuits are not capable of increasing the signal-to-noise ratio at the receiver output over that at the input. The only way in which noise can be reduced in an a.m. receiver is by narrowing the audio channel — in other words, by cutting off the higher audio frequencies. The f.m. system, on the other hand, inherently reduces noise, so it is possible to widen the audio channel of an f.m. set to 15,000 cycles without raising the noise to an objectionable level.

The other reason is that interference becomes appreciable in the output of an a.m. receiver if the desired signal is not at least 100 times as strong as the interfering signal. An a.m. receiver is therefore usually built to accept side bands no wider than 5 kc. from the desired signal; if it were to accept a wider band, it would frequently pick up interfering signals from adjacent stations that would be strong enough to cause noticeable "monkey chatter" in the receiver output. Interference is not noticeable in an f.m. receiver, however, as long as the desired signal is more than twice as strong as the undesired one, a fact that makes it possible for an f.m. set to reproduce a 15-kc. audio range without trouble from adjacent-channel interference.

## DYNAMIC RANGE

The term dynamic range refers to the difference in sound levels between the loudest and softest portions of a program. For symphonic music, the range may be on the order of 70 db, corresponding to a voltage ratio of about 3000 to 1. If it were possible to transmit the full volume range to a receiver that could use it, the set would reproduce the loud and soft passages just as they occur at the studio.

In a.m. broadcasting, the volume range is reduced by the monitor operator to 35 db or less. This is done by advancing the amplifier gain control on weak passages and reducing the gain on the strong passages. A.M. transmitters also use compressor amplifiers to limit peaks. Such compression of the volume range is necessitated by the inherent signal-to-noise ratio in a.m., which, at best, seldom exceeds 40 db. Even if the full dynamic range of a symphonic program could be transmitted, the average a.m. receiver could not reproduce it satisfactorily because of the presence of hum and noise at the lower passages. If the volume control is adjusted to a setting at which the loudspeaker is not overloaded on the loud passages, reception can be satisfactory at high and medium levels of the music in spite of the presence of noise, because the program energy greatly exceeds the noise energy and thereby renders the noise unnoticeable to the human ear. On the soft passages of music, however, hum and noise are very apparent. In fact, the music amplitude may be well below the amplitude of the disturbances, so that the music is not even heard during the soft passages and the noise level appears to have risen. Reasonably noise-free reception can be achieved in an a.m. receiver only by narrowing the audio channel of the receiver to reduce the noise and by operating with a reduced dynamic range at the a.m. transmitter.

Such restrictions on the possible dynamic range of transmission and reception are not present in f.m. Most f.m. transmitters are capable of transmitting the full dynamic range of a symphony, and the better f.m. sets, which have a very low noise level, can reproduce practically the full range.

However, this does not mean that f.m. receivers are actually operated with such a range. In a home, where maximum volume must generally be kept at some reasonable value, it is usually impossible to make use of a dynamic range of more than 35 or 40 db; therefore, some of the potential abilities of an f.m. set are seldom brought into actual use. However, the range used is almost always greater than that of an a.m. set. This is particularly evident on soft notes; a listener to an f.m. set can hear soft passages that would be lost in the inherent noise of an a.m. set.

# F.M. Receivers Using Amplitude Limiters

In the Armstrong system of f.m. reception, separate limiter stages in the receiver are used to remove amplitude variations in the signal. Let us study this type of f.m. receiver in detail.

A block diagram of this type of receiver is shown in Fig. 5. As you can see, it resembles the diagram of a conventional a.m. set from the antenna to the input of the limiters. However, even in these stages, there are certain important differences between f.m. and a.m. circuits. For example, an automatic volume control acting upon all the r.f. and i.f. amplifier stages, which is common in a.m. receivers, is neither necessary nor desirable in f.m. circuits. It is unnecessary because the f.m. limiter maintains the amplitude of the signals applied to the discriminator at a fixed level. It is undesirable because, in general, any system that reduces the r.f. gain preceding the limiter tends to lower the signal-to-noise ratio at the limiter. However, a few receivers have been designed in which a limited amount of automatic or manually adjustable volume control is incorporated to prevent the signal amplifier grids from being driven positive on very strong signals.

Another difference between a.m. and all f.m. broadcast superheterodyne receivers lies in the comparative band width of the i.f. amplifier system. In a.m., at any one modulating frequency, there is but one pair of side band components having frequencies respectively higher and lower than the carrier by the amount of the modulating frequency. If the highest modulating frequency is 5 kc., an i.f. band width of 10 kc. centered at the carrier frequency is adequate.



In f.m., on the other hand, a large number of pairs of side band components of appreciable amplitude may be present along with the center frequency component. As we said before, the maximum band width is required when the f.m. signal is fully modulated at the highest modulating frequency. For example, if the frequency is varying over the range of plus or minus 50 kc. per second (corresponding to 67% modulation) at a rate of 10 kc. per second, eight important pairs of side bands are present, requiring a theoretical band width of 8 x 2 x 10 or 160 kc. centered at the intermediate frequency.

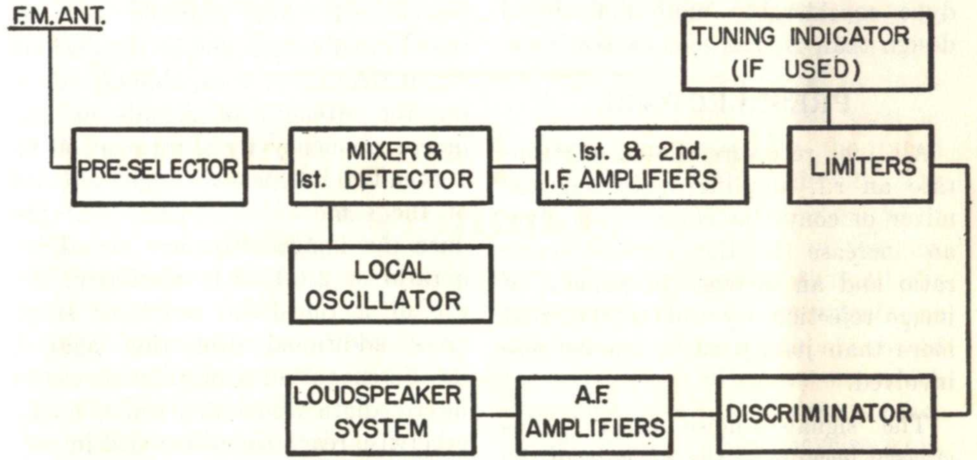


**FIG. 5. Block diagram of the Armstrong type of f.m. receiver.**

Actually the band width of an f.m. receiver need not be made as wide as theoretically required, because: 1, side band components at frequencies near the limits of the theoretical band are of rather small amplitude and appear only when the transmitter is strongly modulated at the higher audio frequencies; 2, a worthwhile increase in r.f. gain per amplifier stage is obtained as the band width is decreased; and 3, with a somewhat narrower band, adjacent channel interference is reduced. The selectivity curves of i.f. amplifiers in f.m. broadcast receivers are down about 6 db at 75 kc. above and below the intermediate frequency. This represents a compromise between the opposing aims of obtaining maximum gain per stage and excellent suppression of adjacent channel interference on the one hand and of having uniform amplification of all side band components on the other. This comparatively narrow i.f. band width can cause appreciable distortion when the incoming f.m. signal is strongly modulated at high audio frequencies.

In general, f.m. receivers should have a higher over-all r.f. gain than a.m. receivers. This is necessary so that the weakest signal from which satisfactory f.m. reception is desired will be amplified sufficiently to permit amplitude-limiting in the receiver. For example, if the receiver is to operate satisfactorily on a signal voltage of, say, 10 microvolts (.00001 volt) at the antenna terminals, and if 4 volts are necessary at the limiter to obtain amplitude-limiting action, the over-all r.f. gain in the r.f. amplifier, mixer, and i.f. amplifier should amount to at least 4/.00001 or 400,000. Such a gain at the high r.f. and i.f. frequencies involved, together with greater band width in the i.f. stages, makes the signal amplification section of the f.m. superheterodyne considerably more difficult to design than is that of a.m. receivers.

## PRESELECTORS

All f.m. receivers should incorporate an r.f. amplifier preceding the mixer or converter stage. This gives an increase in the signal-to-noise ratio and an increase in amount of image rejection, two advantages that more than justify the extra expense involved.

The signal-to-noise ratio is increased because of the r.f. gain of the preselector. A well-designed coil has a voltage set-up at resonance of about 3 to 5, and careful circuit design and choice of the r.f. tube can produce a tube gain of about 6 to 10, so the r.f. amplifier can have an over-all gain of about 20 to 50. Although a large gain is generally not obtainable in the r.f. amplifier at frequencies in the 88-108 mc. band, the gain of the r.f. amplifier does serve to reduce materially the amount of gain required in the mixer and i.f. amplifier stages. The introduction of gain *before* the mixer stage is especially desirable because it serves to improve the signal-to-noise ratio of the receiver. Most of the noise due to tube hiss is introduced by the mixer. Thus, if the amplitude of the signal can be raised before it reaches the mixer grid, the usable sensitivity of the receiver can be improved.

The amount of image rejection is increased by an r.f. amplifier stage because the image frequencies are passed through two independent tuned circuits and the rejection of the image frequency is therefore more pronounced. For example, if one tuned circuit, resonant at the desired signal frequency, is capable of reducing the strength of signals at the image frequency by a ratio of 50 to 1, then two independent tuned circuits of the same characteristics will reduce the image frequency signal by a ratio of 2500 to 1. Similarly, the use of a tuned r.f. amplifier stage gives additional protection against interference from strong signals at the intermediate frequency, which might otherwise reach the mixer grid in sufficient strength to cause serious interference at the mixer output.

## THE OSCILLATOR AND MIXER

As in all superheterodynes used for v.h.f. operation, a separate tube, not a part of the mixer, is used for the local oscillator stage. The constant-amplitude oscillator voltage and the signal voltage from the r.f. amplifier are applied to separate control elements of the mixer tube. The component of plate current at the difference frequency is used to excite the high gain i.f. amplifier that follows the mixer stage.



Whatever tube and circuit arrangement is used, it is important that the mixer tube furnish as much signal gain as possible, and that there be a minimum of interaction between the oscillator and signal circuits. For example, adjustment of the trimmer condenser in the tuned signal input circuit should not affect the oscillator frequency. The output voltage of the oscillator should be as large as possible without overloading the mixer and should be fairly constant over the frequency range.

The frequency of the oscillator may be made either higher or lower than the signal frequency. Both arrangements have been used in f.m. receivers. With the oscillator frequency higher than the signal frequency, it is easier to obtain good tracking over the tuning range. With the oscillator frequency lower than the signal frequency, it is easier to obtain good frequency stability from the oscillator.

**Frequency Stability.** In f.m. broadcast reception, where the maximum transmitter deviation is ±75 kc., the maximum permissible receiver oscillator drift after the oscillator has completed its warm-up is 10 kc. Drifts in excess of this amount cause excessive audio distortion of the output signal and cause the signal-to-noise ratio to decrease because the linear portion of the S curve of the discriminator is exceeded. (Frequency discriminators will be studied in detail shortly.)

## THE I.F. AMPLIFIER

The intermediate frequency amplifier in an f.m. receiver, as it does in an a.m. receiver, contributes the major part of the r.f. gain and provides the selectivity necessary to avoid adjacent channel interference. A low i.f. is desirable from the standpoint of obtaining good selectivity and high gain per stage, and also because it will give a more sensitive discriminator action (since any given frequency deviation, such as ±75 kc., will become a comparatively large percentage of the i.f. frequency at the discriminator). However, to eliminate interference from signals at the image frequency, a fairly high i.f. must be used. The image frequency that differs from the desired signal by twice the amount of the i.f. will then be effectively suppressed in the tuned r.f. amplifier and mixer circuits. The i.f. should preferably be equal to at least one-half the width of the f.m. receiver tuning band so that all the image frequencies will lie outside the f.m. band. However, the i.f. should not itself be a frequency on which strong signals are encountered. Since the f.m. band is 20 megacycles wide, an i.f. value of 10.7 mc. is generally used.

## LIMITERS

The purpose of the limiter section in an f.m. receiver is *to remove amplitude variations from the f.m. signal* at the output of the i.f. amplifier, so that the signal fed into the discriminator will have constant amplitude. In serving this purpose, the limiter automatically corrects deficiencies in the frequency response of the preceding r.f. and i.f. stages. The limiter also provides *an a.v.c. voltage* for use during f.m. reception.

A typical single-tube limiter stage is shown in Fig. 6. To understand how the 6SJ7 pentode in this circuit can function as a limiter, we must first consider the $E_g$-$I_p$ characteristic curves for this tube under various operating conditions.

When a 6SJ7 pentode is operated as an amplifier at normal voltages, such as with a 250-volt d.c. plate voltage and a 100-volt d.c. screen grid voltage, the static $E_g$-$I_p$ characteristic of the tube will be like curve 1 in Fig. 7. This curve is essentially linear up to the highest plate current that can safely be passed by the tube, and consequently the plate current increases in proportion to positive grid voltage values. This is of no value for limiter action; we desire a characteristic that will make the plate current essentially constant regardless of how much the grid is driven positive, and hence the tube must be operated in such a way that plate current saturation occurs at a low positive grid voltage value.

When the d.c. plate and screen grid voltages of the 6SJ7 pentode are about 60 volts, we secure the desired condition (saturation at a fairly low positive grid voltage), as indicated by curve 2 in Fig. 7. This curve is for the tube alone, and would be obtained from the circuit of Fig. 6 only if resistor R and condenser C were shorted out so as to give static conditions.

If a strong sine-wave signal is applied to the grid of the 6SJ7 tube when it is operated with characteristic curve 2 in Fig. 7, the operating point will be at o, negative peaks of the input signal will swing beyond cut-off (beyond c), and positive peaks will swing into the saturation region of the curve (beyond point x at which saturation begins). As a result, both the positive and negative peaks of the plate current will be cut off, and we will secure a certain amount of the desired limiting action on strong signals.

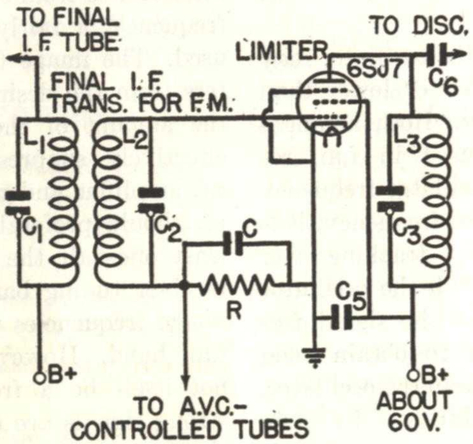


**FIG. 6. Simplified one-tube limiter circuit for an f.m. receiver.**



Now let us return to the basic limiter circuit in Fig. 6, and consider its operation first with resistor R in the grid circuit but with condenser C removed. First of all, we can see that during half-cycles that swing the grid negative, there is no grid current flowing through resistor R, and consequently the circuit acts essentially the same as if resistor R were shorted. In other words, our characteristic curve with R in the circuit is the same for negative grid voltages as it is without R. For half-cycles that make the grid swing positive; however, the flow of grid current through R develops across it a voltage drop that opposes the applied positive grid voltage, and that consequently serves to provide a negative bias that is at each instant proportional to the positive applied voltage. The characteristic curve for this condition is like curve 3 in Fig. 7, which is flatter than curve 2. The saturation effect is considerably more pronounced now, and plate current during positive half-cycles is limited by the negative bias across R as well as by the saturation characteristic of the tube at the low d.c. operating voltages used.

With R alone in the limiter circuit, the bias developed across it would follow individual r.f. positive peaks, and would consequently be varying continually. In a practical limiter circuit, resistor R in Fig. 6 is shunted by condenser C, with the time constant of R and C being equal to the time of several r.f. cycles. Under this condition, the R-C circuit assumes a definite negative bias voltage that is maintained relatively constant over several cycles of the r.f. input voltage, and therefore exists for negative half-cycles as well as for positive half-cycles.

An automatic bias of this nature gives a characteristic curve somewhere between curves 2 and 3 in Fig. 7; with R and C large in electrical size so as to give a long time constant, the dynamic characteristic curve approaches static curve 2 because we are approaching a fixed-bias condition, but with low electrical values for R and C, the dynamic characteristic approaches curve 3 because the bias now varies almost instantaneously with input signal strength.

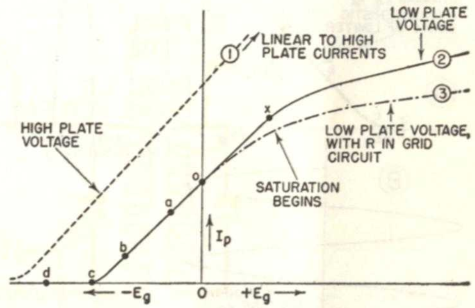


**FIG. 7. Static characteristic curves of a vacuum tube for high (1) and low (2) plate voltages, and dynamic characteristic curve (3) for a limiter circuit using a low plate voltage and a resistor R (but no condenser C) in the grid circuit. The plate load resistance for the static curves is assumed to be equal to the resistance of the plate load resonant circuit at resonance during dynamic operation.**

A typical dynamic characteristic curve for a single-tube limiter stage is shown in Fig. 8. In a stage having this characteristic, the bias voltage developed across R and C is proportional to the average signal strength over several r.f. cycles, and consequently the operating point may be at a, b, c, or even beyond cut-off at d, depending upon the r.f. signal strength. Thus, with a sine-wave input voltage $e_g$ that places the operating point at point b, the wave form of the resulting plate current $i_p$ is as shown at C in Fig. 8. This plate current is far from being sinusoidal like the input voltage, but we will still secure a sinusoidal output voltage across plate resonant circuit $L_3$-$C_3$ in Fig. 6 because this circuit is tuned to the 10.7-mc. i.f. value for f.m. It has a high Q factor, and therefore responds only to the desired fundamental frequency of the plate current pulses, rejecting the harmonics.

When a limiter stage has the dynamic characteristic curve of Fig. 8, two things happen when the strength of the input signal increases: 1, the increased negative bias makes the negative half-cycle swing beyond cut-off for a longer period of time during each cycle, thereby reducing the operating angle during which plate current does flow; 2, the amplitude of the plate current pulses increases slightly, because an automatically produced bias can never completely counteract an increase in signal strength. The increased positive voltage on the grid produces a slight increase in the amplitude of the plate current pulses because the characteristic curve rises slightly in the saturation region, rather than being perfectly flat.

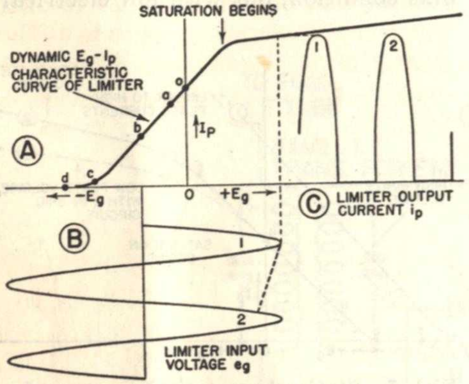


**FIG. 8. These diagrams will help you to understand the action of the basic one-tube limiter circuit in Fig. 6.**

It is possible to design a limiter circuit so that, for any reasonable increase in the r.f. input voltage to the limiter, the operating angle for plate current will decrease just enough to counteract the increase in the amplitude of the plate current pulses. The energy fed into plate resonant circuit $L_3$-$C_3$ in Fig. 6 at the fundamental intermediate frequency for f.m. will then be constant, and the desired i.f. output voltage across this limiter resonant circuit will likewise be constant.

When this goal in limiter circuit design is reached, the over-all dynamic characteristic curve for the limiter is like curve 1 in Fig. 9A, in which the r.f. output voltage of the limiter is plotted against the r.f. input voltage to the limiter. With this characteristic, the i.f. output voltage remains essentially constant regardless of input voltage for all input voltage values that reach saturation (swing beyond point s in Fig. 9A).

If the design of the limiter circuit is such that the operating angle decreases faster than the plate current pulse amplitude increases, we have over-compensation and secure the over-all characteristic represented by curve 3 in Fig. 9A. Likewise, if the amplitude of the plate current pulses



increases faster than the operating angle decreases, we have under-compensation and secure the over-all characteristic represented by curve 2 in Fig. 9A. The values employed for R and C in the limiter circuit of Fig. 6 determine which over-all characteristic will be obtained, and hence these two parts in a limiter circuit are highly important.

Since the d.c. voltage produced across limiter grid resistor R is proportional to the strength of the f.m. signal at the limiter input, this d.c. voltage can be used for a.v.c. purposes during f.m. reception.

A limiter must be *fast-acting* (must have a *short time constant*) if it is to block out sudden noise surges. This assumes that there is sufficient r.f. amplification in the f.m. receiver so that the weakest desired signal will swing the limiter grid beyond the point at which saturation begins.

To show how the limiter can flatten the response characteristic of the r.f. system, let us assume that the limiter stage under consideration has a flat over-all grid voltage-plate voltage characteristic curve like 1 in Fig. 9A. Assume further that, under this condition, all limiter input signals having peaks higher than 5 volts cause plate current saturation. (Positive grid swings will then go beyond point s, at which saturation begins.)

Now suppose that the r.f. system ahead of the limiter has the sharply peaked response characteristic shown in Fig. 9B. The vertical scale gives peak amplitude values; thus, if a strong input signal gave a 20-volt output peak amplitude at the resting frequency, this peak amplitude at the limiter input would drop down to 5 volts during the maximum deviation of 65 kc. (from b to a to c). With the limiter characteristic of Fig. 9A, all signal peaks above 5 volts would be cut down to 5 volts by the limiter, and the solid-line curve in Fig. 9C would then represent the response characteristic of the r.f. system and limiter combined, for a strong f.m. input signal. Since this is flat over the desired band width of 130 kc., it is obvious that the limiter has flattened the highly peaked response of the r.f. system.

Now suppose the receiver is tuned to a weaker f.m. signal, which gives a resting-frequency peak of only 10 volts at the output of the r.f. system. The broken-line response curve in

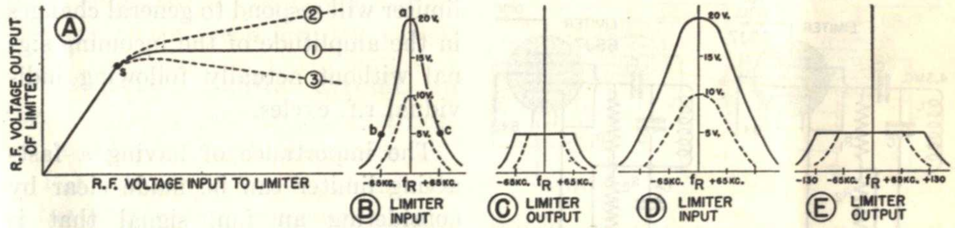


**FIG. 9. These curves show how the limiter can correct the over-all response characteristic of the entire r.f. system (including the i.f. amplifier and the receiving antenna system) in an f.m. receiver.**

Fig. 9B will then give peak values for various deviations. When this weaker signal is fed into the limiter, the output response will be as shown by the broken line in Fig. 9C. The band width over which we have uniform response is now obviously insufficient for standard f.m. signals, and distortion will occur during loud portions of the program.

On the other hand, if the r.f. system has the broadly peaked response characteristic shown in Fig. 9D, a signal having a 20-volt peak will give

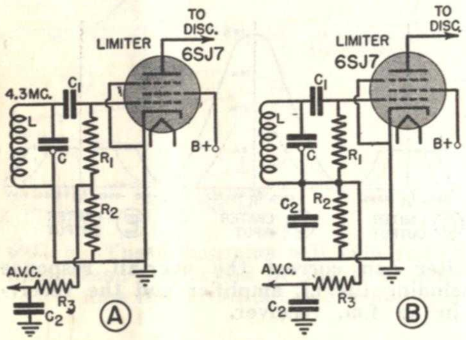


FIG. 10. Typical one-tube limiter circuits.

a combined r.f.-limiter response corresponding to the solid line in Fig. 9E, after passage through the limiter. This response is ideal, being flat for a frequency swing of about 250 kc. If the input signal should now drop to a maximum peak value of 10 volts, the input and output response characteristics would be as shown by the broken lines in Fig. 9D and 9E; even at this lower signal level, however, the combined r.f.-limiter response would still be flat over a wide enough range to allow the entire frequency swing of 130 kc. to exist at the limiter output without changes in amplitude.

We thus arrive at the important conclusion that a broad over-all response for the r.f. and i.f. sections enables the limiter to handle weaker signals satisfactorily. Almost any sort of peak response is permissible if the r.f. and i.f. sections have sufficient gain, however, for *the limiter will then flatten the over-all response of the preceding sections over the entire range of deviation frequencies.*

**Time Constant of Limiter.** The time constant of the R-C circuit in the limiter is ordinarily made equal to the time of a few r.f. cycles, so the limiter will respond to general changes in the amplitude of the incoming signal without actually following individual r.f. cycles.

The importance of having a fast-acting limiter can be made clear by considering an f.m. signal that is varying in frequency from 65 kc. below to 65 kc. above its resting frequency (maximum deviation, corresponding to maximum loudness). With a receiver having a sharply resonant r.f. response like that in Fig. 9B, and with a peak limiter input of 20 volts at the resting frequency, the amplitude of the signal fed into the limiter will vary from 5 volts (at points b and c) to 20 volts (at a).

With the 20-volt input, the limiter will probably be operating near plate current cut-off, but this cut-off bias will be far too great to give the desired constant output amplitude when the limiter input drops to 5 volts. To keep the limiter output amplitude constant over the entire deviation range of 130 kc., the C bias should automatically reduce itself as the



limiter input signal amplitude drops.

Noise surges that enter an f.m. receiver are oftentimes strong but of extremely short duration. If these surges are to be blocked out, the limiter must be able to increase its own negative bias automatically for the duration of each surge. This action can occur only if a fast-acting limiter (with a short time constant for R and C) is used.

The C bias produced by the R-C grid network of the limiter stage can change fast enough to flatten the r.f. response and squelch noise only if its time constant is kept very short (within the time of a few r.f. cycles).

**Fast-Acting Limiter Circuit.** In Fig. 10A is a practical limiter circuit employing a grid resistance ($R_1 + R_2$) between the grid and cathode of the limiter, with condenser $C_1$ and the grid-cathode capacity of the tube acting with the grid resistance to provide the required time constant for fast limiter action. This is attained by using a low capacity value for $C_1$ and a low ohmic value for the grid resistance. A portion of the d.c. voltage developed across the grid resistance (the d.c. voltage across $R_2$) is used for a.v.c. purposes during f.m. reception. $R_3$ and $C_2$ form the conventional a.v.c. filter that keeps r.f. components out of the a.v.c.-controlled tubes.

**Dual-Action Limiter Circuit.** In the limiter circuit arrangement shown in Fig. 10B, both $C_1$ and $C_2$ have low reactances at very high frequencies. $R_1$ and $R_2$ in series form the grid return path. $R_1$ and $C_1$ together have a time constant equal to a few r.f. cycles, and hence provide a rapidly changing bias equivalent to the fast limiter action of the circuit of Fig. 10A. $R_2$ and $C_2$ have a much longer time constant, and act to change the C bias voltage in accordance with the average strength of an incoming f.m. signal. In other words, $R_2$ and $C_2$ take care of the major changes in signal amplitude, such as those occurring when tuning from a weak distant f.m. station to a strong local f.m. station, and $R_1$ and $C_1$ take care of changes in signal amplitude due to a

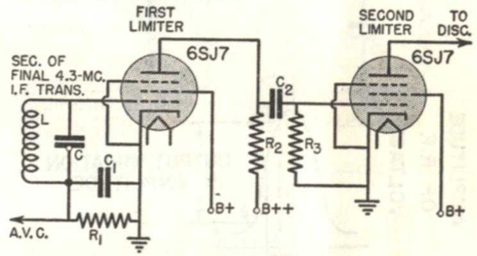


FIG. 11. Cascade limiter circuit employing two limiter tubes. The a.v.c. voltage is obtained from the first tube. The General Electric Model JFM-90 f.m. converter is one example of f.m. units employing this cascade limiter arrangement.

peaked r.f. response characteristic or to noise interference. This circuit arrangement thus provides independent control over two of the important factors that affect the design of a limiter circuit.

**Cascade Limiter.** Improved limiter operation for both weak and strong signals, along with considerably higher gain, can be secured with two limiter tubes connected as shown in Fig. 11 in what is known as a *cascade limiter* circuit. The action of the first limiter tube is essentially like that of the limiter circuit in Fig. 10A, with the grid resistor and condenser being chosen to give an even shorter time constant so this first limiter will effectively reduce impulse noise. Whatever amplitude variations exist in the output of the first limiter tube are removed by the second limiter

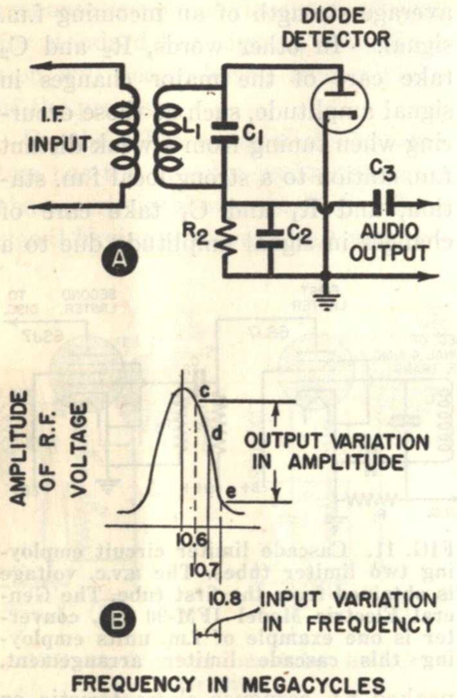


**FIG. 12. A, a simple f.m. detector in which the slope of tuned circuit $L_1C_1$ shown in curve B is used to produce a variation in output amplitude for a variation in input frequency.**

tube, so that the f.m. signal that the limiter section feeds into the discriminator has essentially constant amplitude. Listening tests have shown that this cascade limiter reduces impulse noises far more than any single-tube limiter circuit.

One engineer recommends that the r.m.s. value of limiter output voltage be at least 20 volts for effective noise reduction. The minimum voltage gain for the entire r.f. system including the limiter should be at least 20,000,000 times to permit reception of weak desired f.m. signals without noise interference. To provide this amount of amplification, an r.f. amplifier stage ahead of the frequency converter is highly desirable.

## A SIMPLE DETECTOR

A circuit capable of converting an f.m. signal into an audio signal is shown in Fig. 12A. (For simplicity, this diagram shows the detector connected to the output of the last i.f. stage—actually, as you just learned, one or two limiters are inserted in most commercial receivers between the last i.f. stage and the detector circuit to suppress any amplitude noise pulses or amplitude modulation that may have been sent out by the transmitter.) The f.m. signal flows from the plate circuit of the last i.f. stage to $L_1C_1$, which can be tuned to either a higher or lower frequency than the center i.f. value of 10.7 mc. In the example given, $L_1C_1$ is tuned to a lower frequency.

The response of the tuned circuit, shown in Fig. 12B, decreases as the frequency of the input signal decreases. Between c, d, and e, the slope of the response curve is rather straight and is the portion used in "slope" or "off-resonance" f.m. detectors. When the input i.f. is frequency modulated, the r.f. voltage across $L_1C_1$ *varies in amplitude* in accordance with the original audio signal amplitude impressed on the transmitter. This amplitude-modulated r.f. voltage is demodulated by the diode, and the resulting audio voltage appears across the diode resistor $R_2$. Condenser $C_2$ removes the r.f. variations across the resistor $R_2$ so that only the desired a.f. is fed to the audio circuits.

This circuit is a simple, basic one; refinements of it have been used in some f.m. receivers, but it is not satisfactory if high fidelity is desired — the response curve is not sufficiently linear to produce accurate reproduction of the original modulation.

## DOUBLE-TUNED F.M. DISCRIMINATOR

A somewhat better type of f.m. discriminator is shown in Fig. 13A. As you can see, it is a push-pull version of the slope detector just studied. It has the advantage of being capable of producing a more nearly linear response characteristic. Let's see how it works.

The circuit $L_1$-$C_1$ is tuned to the i.f. "carrier" or resting signal, but responds well to frequencies on both sides of this value. The input signal induces voltages in both $L_2$ and $L_3$, thus causing electrons to flow through diodes $VT_1$ and $VT_2$.

Since electrons move only from the cathode to the plate, electrons flowing through $VT_1$ come upward through resistor $R_1$, producing across this resistor a voltage drop with the polarity shown. Similarly, electrons flowing through $VT_2$ move downward through resistor $R_2$; the voltage drop across this resistor is therefore of opposite polarity from that across resistor $R_1$.

The resonant circuit M ($L_2$-$C_2$) is tuned to a frequency somewhat lower than $L_1$-$C_1$, whereas N ($L_3$-$C_3$) is tuned to a frequency somewhat higher. If, for example, the frequency of 4 mc. in Fig. 13B represents the $L_1$-$C_1$ resonance frequency, then 3.9 mc. may represent the resonance frequency of M, and 4.1 mc. that of N. Hence, neither M nor N is tuned to

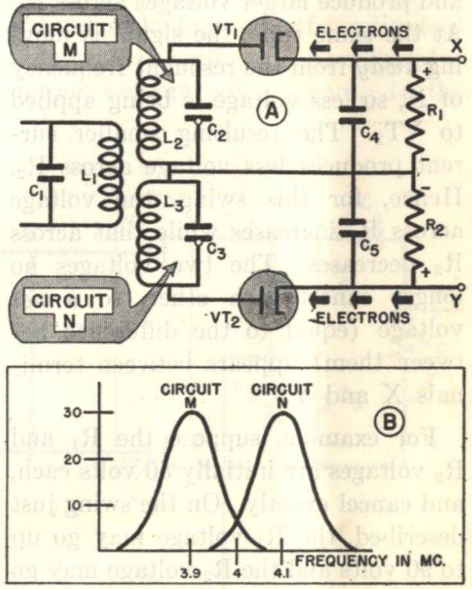


**FIG. 13. The double-tuned type of f.m. discriminator. (Sometimes called the Conrad or Travis discriminator.)**

*favor* the center frequency, although both pass it along to a certain extent.

The circuits are adjusted so that, at the resting frequency, the voltages applied to $VT_1$ and $VT_2$ are equal, and equal currents flow through $R_1$ and $R_2$. Since these are equal resistances, their voltage drops are equal and since their voltage drops are *opposite in polarity* (see Fig. 13A), there is NO VOLTAGE between terminals X and Y when only the resting frequency is fed into this stage.

Now let us suppose the signal frequency is modulated so that it swings from 4 to 3.9, back through 4 to 4.1, then back toward 3.9, etc.

As the frequency approaches 3.9, more voltage will be developed by M. This causes larger voltages to be applied to $VT_1$; larger currents result and produce larger voltages across $R_1$. At the same time, the signal is moving *away* from the resonant frequency of N, so less voltage is being applied to $VT_2$. The resulting smaller current produces less voltage across $R_2$. Hence, for this swing, the voltage across $R_1$ increases while that across $R_2$ decreases. The two voltages no longer cancel each other, so a *net* voltage (equal to the difference between them) appears between terminals X and Y.

For example, suppose the $R_1$ and $R_2$ voltages are initially 50 volts each, and cancel exactly. On the swing just described, the $R_1$ voltage may go up to 90 volts and the $R_2$ voltage may go down to 10 volts. The difference is 90 minus 10 or 80 volts, which will appear between X and Y. Terminal Y will be positive, since the polarity will be that of the voltage across resistor $R_1$.

When the swing reverses toward 4.1 mc., exactly the reverse action occurs. Now, N develops the larger voltage, $VT_2$ passes the larger current, and the drop across $R_2$ is larger than that across $R_1$. This reverses the polarity of the difference voltage existing across X and Y, since the polarity now is that of the voltage across $R_2$.

**Summary.** A signal, shifting in frequency, is introduced into this stage. The frequency variations produce a varying voltage across terminals X and Y. Small frequency changes produce small voltages, since frequencies near 4 mc. do not "climb" as high on $L_2$-$C_2$ and $L_3$-$C_3$ resonance curves. Larger changes, out to the limits set by the resonance points (3.9 and 4.1 in Fig. 13B), produce larger voltages. Thus, we have a circuit that produces an amplitude variation from a varying frequency signal.

This same circuit also gives us demodulation; in this, it is like the diode detector we just discussed. The diodes $VT_1$ and $VT_2$ rectify the r.f. signal, and condensers $C_4$ and $C_5$ pass the r.f. variations around resistors $R_1$ and $R_2$. Thus, the original intelligence signal appears across terminals X and Y.

## THE FOSTER-SEELEY DISCRIMINATOR

The Foster-Seeley discriminator is often used to obtain the very nearly linear amplitude variation corresponding to frequency deviation that is needed in high-fidelity f.m. receivers. Although we have already studied this circuit, it is of such importance that we will review its operation at this time.

Discriminator transformer $T_1$ in Fig. 14 is of special design. Its primary winding $L_2$ is tuned to the 10.7-mc. i.f. value by $C_2$, and its center-tapped secondary winding $L_3$ is tuned to the same i.f. value by $C_3$. The limiter output current flowing through $L_2$ induces a corresponding f.m. voltage in $L_3$, and the resonant circuit produces across the two sections of $L_3$ voltages $E_1$ and $E_2$ which are always



equal in magnitude and 180° out of phase with each other when considered with respect to the center tap.

The limiter output voltage $E_p$ in series with $E_1$ is applied to diode section 1 of the 6AL5 duplex diode tube. (D.C. blocking condenser $C_1$ provides an r.f. path from the upper end of $L_2$ to p, and the chassis and r.f. by-pass condensers $C_5$, $C_6$, and $C_7$ complete the path for r.f. signals from the lower end of $L_2$ to the cathode of

section 1 is $R_1$ and the load resistor for diode section 2 is $R_2$. Electrons flow in opposite directions through $R_1$ and $R_2$, as you can easily see by tracing the diode circuits. This means that the combined voltage across both $R_1$ and $R_2$, which is the output voltage of the discriminator, will at each instant be the difference between the individual voltages. If the individual voltages are equal, the discriminator output voltage will be zero. If the

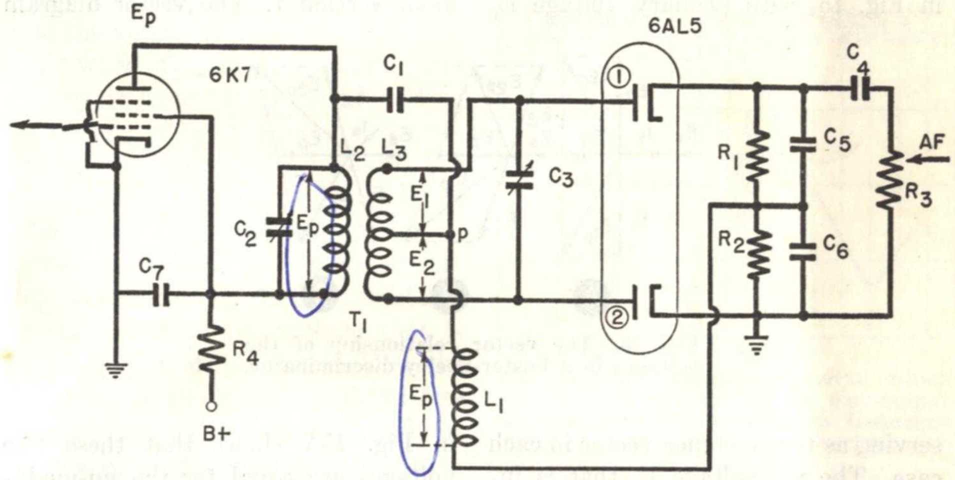


**FIG. 14. The basic Foster-Seeley discriminator circuit.**

diode section 1. Thus the voltage $E_p$ exists across choke coil $L_1$ and acts in series with $E_1$ or $E_2$ on either diode.) Likewise, the limiter output voltage $E_p$ in series with $E_2$ is applied to diode section 2.

The net voltage applied to each diode section is therefore the *vector sum* of the two individual voltages acting on that section. Each diode section rectifies its net applied r.f. voltage and produces a proportional d.c. output voltage across its load resistor. The load resistor for diode

voltages across $R_1$ and $R_2$ are different, the combined voltage will have the polarity of the larger of the two individual voltages and will be equal in magnitude to their numerical *difference.*

Let us consider now the factors that make the output voltage of one diode higher than that of the other. First of all, we must choose some voltage or current for reference purposes. Since $E_p$ is common to all circuits under study, we can use it as our reference voltage.

Phase relationship in this discriminator circuit must be considered for three different combinations: 1, when the limiter output signal frequency is equal to the i.f. resting frequency to which the discriminator resonant circuits are tuned; 2, when the limiter output frequency is less than the i.f. resting frequency; 3, when the limiter output frequency is higher than the i.f. resting frequency. The vector diagrams for these three conditions are shown at A, B, and C respectively in Fig. 15, with primary voltage $E_p$ serving as the reference vector in each case. The r.f. voltage $E_s$ that is induced in secondary winding $L_3$ is 180° out of phase with the primary r.f. voltage $E_p$, so it is shown 180° out of phase with reference vector $E_p$ in each of the vector diagrams.

When the limiter output signal is exactly at the i.f. resting value to which the discriminator circuits are tuned (in other words, when no sound is being transmitted), secondary tuned circuit $L_3$-$C_3$ is at resonance and secondary current $I_s$ flowing through $L_3$ is in phase with $E_s$, as indicated in Fig. 15A. The voltage produced across the entire secondary winding by this secondary current (the sum of voltages $E_1$ and $E_2$) therefore leads both $I_s$ and $E_s$ by 90°. $E_1$ and $E_2$ are used only in connection with the center tap of $L_3$, however, so if we show $E_1$ as leading $I_s$ by 90°, we must show $E_2$ as lagging $E_1$ by 180°; they are shown this way in Fig. 15A.

Adding $E_p$ and $E_2$ vectorially gives $E_{2p}$ as the resultant voltage acting upon diode section 2. Likewise, adding $E_p$ and $E_1$ vectorially gives $E_{1p}$ as the resultant voltage acting upon diode section 1. The vector diagram in Fig. 15A shows that these two voltages are equal for the no-modulation condition; therefore, the d.c. voltages developed across $R_1$ and $R_2$ by the two diode sections are equal in magnitude and their sum is zero. This is just as it should be, since no a.f. signal should be obtained when there is no a.f. modulation at the transmitter.

When the limiter output signal frequency is *lower* than the i.f. resting value to which resonant circuit $L_3$-$C_3$ is tuned, this circuit becomes *capacitive*, and $I_s$ leads $E_s$ as shown in Fig. 15B. Since voltages $E_1$ and $E_2$ must be 90° out of phase with $I_s$, the resultant voltages $E_{2p}$ and $E_{1p}$ are unequal, with $E_{1p}$ (the voltage applied to diode section 1) the larger. Referring to Fig. 14, you can see that with diode section 1 getting the higher r.f. voltage, we secure a higher d.c. voltage across $R_1$ than across $R_2$, and the combined voltage across $R_1$ and $R_2$ is therefore *positive* with respect to the ground. The more the limiter output frequency swings below the i.f. resting frequency, the greater will be this positive voltage applied to the a.f. amplifier input.

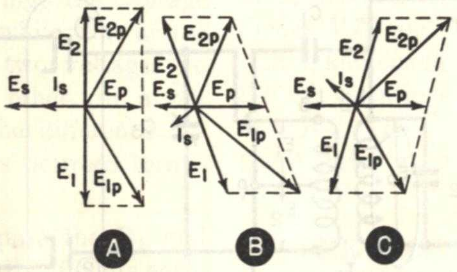


**FIG. 15. The vector relationship of the voltages in a Foster-Seeley discriminator.**



By a similar analysis, we can obtain the vector diagram shown in Fig. 15C when the limiter output frequency is *higher* than the i.f. resting frequency. When this occurs, the net voltage applied to the input of the a.f. amplifier by $R_1$ and $R_2$ combined is *negative* with respect to ground.

The frequency discriminator circuit shown in Fig. 14 thus produces at its output a d.c. voltage that is at each instant proportional to the deviation in the incoming signal frequency from its resting value and has a polarity determined by the direction in which this frequency deviation occurs. The discriminator thus converts an f.m. signal directly into the original audio signal voltage used to modulate the f.m. transmitter.

R.F. by-pass condensers $C_5$ and $C_6$ in Fig. 14 must have a low reactance at the i.f. resting frequency and yet must have a high reactance at audio frequencies so there will be no serious shunting effect upon the a.f. voltage developed across $R_1$ and $R_2$. This a.f. voltage is fed to volume control $R_3$ through d.c. blocking condenser $C_4$, which prevents the C bias voltage of the first a.f. amplifier tube from entering the discriminator circuit and prevents the d.c. discriminator output voltage from acting on the grid of the first a.f. tube.

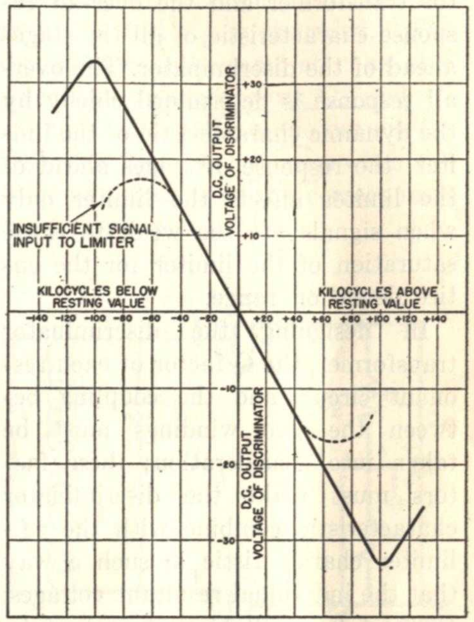


**FIG. 16. The so-called S curve which represents the variation in d.c. output corresponding to variation in frequency input to a Foster-Seeley discriminator.**

## DICRIMINATOR'S CURVE

The solid-line curve in Fig. 16 shows the relationship between the incoming r.f. signal frequency of an f.m. receiver (with respect to the resting value) and the d.c. output voltage of the discriminator. Because of the similarity of this curve to the letter S, it is commonly known as an S curve.

This curve, representing the characteristics *of the f.m. receiver up to the input of the audio amplifier*, should be linear over the entire deviation range of the incoming f.m. signal; otherwise, amplitude distortion will occur in the a.f. output of the discriminator. The linearity of the S curve in Fig. 16 depends upon two things—the design of the discriminator transformer and the over-all response characteristic of all the stages ahead of the discriminator. The over-all response is determined chiefly by the dynamic characteristic of the limiter; the response of stages ahead of the limiter affects the limiter only when signals are too weak to cause saturation of the limiter for the entire deviation range.

In designing the discriminator transformer, the Q factor of each resonant circuit and the coupling between the two windings must be taken into consideration; these factors must make the discriminator characteristic combine with the r.f.-limiter characteristic in such a way that the individual resultant voltages, $E_{1p}$ and $E_{2p}$, will at each instant be proportional to the frequency deviation. When this result is achieved, the combined discriminator output voltage across $R_1$ and $R_2$ will be proportional to the frequency deviation and will be free from amplitude distortion.

It is desirable to have the S curve linear over a somewhat wider range than the maximum deviation to compensate for inaccuracies in tuning, frequency drift in the local oscillator, or misalignment of the discriminator resonant circuits.

A falling off in the amplitude of the limiter output near the deviation limits due to too weak a signal at the limiter input will reduce the frequency range over which the S curve is linear. The broken S curve in Fig. 16 illustrates this condition. When the S curve for discriminator action has too short a linear region, reproduction will be satisfactory at medium and low program loudness levels but amplitude distortion will occur during loud sounds. Thus, all sounds loud enough to cause a deviation of more than 40 kc. will swing the signal frequency beyond the linear region of the broken curve in Fig. 16. That is why it is so important that the limiter in an f.m. receiver deliver a constant amplitude output signal over the entire deviation range. The goal of the f.m. receiver designer is to provide sufficient voltage gain ahead of the limiter so that the limiter can maintain constant output amplitude for the weakest desired incoming signal.





# F.M. Receivers Without Limiters

In the Armstrong type of f.m. receiver, the amplitude limiter stages are separate from the f.m. discriminator stage. To simplify f.m. receivers, some manufacturers use detector circuits that are insensitive to amplitude variations, thus eliminating the limiter section. The amplitude limiting and frequency discriminating actions are thus combined in one section.

## THE RATIO DETECTOR

One of these amplitude-insensitive detectors is the "ratio" detector shown in Fig. 17. As you can see, its circuit is similar to that of the Foster-Seeley discriminator. $L_1$-$L_2$ and $C_1$ are tuned to the i.f. frequency, and voltages $e_1$ and $e_2$ are 180° out of phase. Voltage $e_3$, which is 90° out of phase with the voltages across $L_1$ and $L_2$ when the input r.f. is equal to 10.7 mc., is connected so that it is in series with $e_1$ across $C_2$ and $VT_1$ and in series with $e_2$ across $C_3$ and $VT_2$. (Voltage $e_3$ can be obtained by coupling through a d.c. blocking condenser to the plate of the i.f. amplifier.) I inductive coupling

When the input i.f. signal remains at the frequency that corresponds to the center resting frequency of the f.m. signal, the voltage $e_1 + e_3$ is equal to $e_2 + e_3$, so the rectified d.c. voltage across $C_2$ is equal to that across $C_3$. (Notice that one of the diodes, $VT_2$, is reversed from the connection used in a Foster-Seeley discriminator. This makes the d.c. voltage across $C_2$ and $C_3$ the sum of the two voltages. In the Foster-Seeley, it is the difference.)

The audio output of this stage is developed between points m and n. (Point n, the common connection between $R_1$ and $R_2$, is grounded.) The d.c. voltage is the same at m as at n when the voltages cross $C_2$ and $C_3$ are equal. Thus the audio output (as shown in Fig. 18A) is zero when the received signal is at the correct resting frequency.

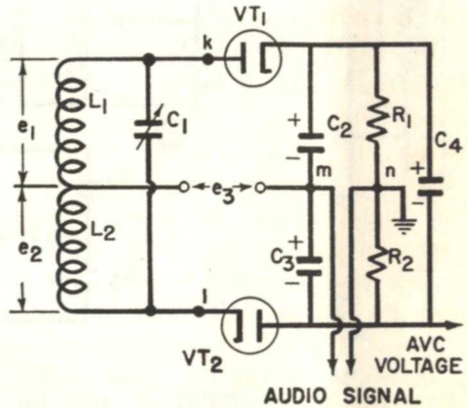


**FIG. 17. The basic ratio detector circuit.**

For the sake of illustration, we have assumed that there are 5 volts across each condenser $C_2$ and $C_3$ in Fig. 18A. The total voltage across condenser $C_4$ is therefore 10 volts.

Now suppose that the frequency of the input i.f. signal changes, with the result that the voltage across $C_2$ increases to 8 volts and the voltage across $C_3$ decreases to 2 volts. This means that (as shown in Fig. 18B) the voltage at m will be 3 volts negative with respect to n. Similarly, if the input frequency should deviate the same amount in the other direction, the voltage across $C_2$ decreases to 2 volts and across $C_3$ increases to 8 volts. As shown in Fig. 18C, m will then be 3 volts positive with respect to n.

The action so far is the same as that of the usual Foster-Seeley circuit. The polarity of the direction of frequency deviation and its amplitude depends on the amount of deviation.

to variations in the amplitude of the input signal. *If the input signal increases or decreases at an audio rate, the condenser voltage is not able to follow these variations.* The d.c. voltage across $C_4$ can, however, follow slow (sub-audio) variations in the amplitude of the input signal.

Let us assume, for example, that a certain value of frequency deviation

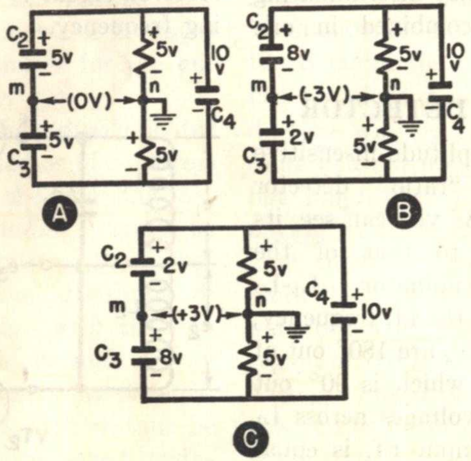


**FIG. 18. How the d.c. output varies in a ratio detector.**

**Discrimination Against A.M.** Let us now observe the action of condenser $C_4$ in Fig. 17 and see how it is used to minimize the effect of amplitude variations in the input. When terminal k is positive, and l is negative, diodes $VT_1$ and $VT_2$ conduct, charging $C_4$ with the polarity shown. The average voltage across $C_4$ is proportional to the f.m. signal strength, and, hence, can be used as a source of a.v.c. voltage. The time constant of $C_4$ (generally 5 to 10 mfd.) and $R_1$ and $R_2$ (about 20,000 ohms total) is about 0.1 to 0.2 seconds. This means that $C_4$ is slow in responding

produces an 8-volt drop across $C_2$ and 2 volts across $C_3$ as shown in Fig. 18B. The output voltage will be —3 volts. Now assume that for some reason (a burst of static or an interfering station) the amplitudes of the voltages $e_1$, $e_2$, and $e_3$ all increase. The voltages across $C_2$ and $C_3$ will both try to increase. However, since the voltages across $C_2$ and $C_3$ must always be equal to that across $C_4$, which cannot change quickly, the voltages across $C_2$ and $C_3$ must remain constant despite the sudden increase in input voltage. (Actually, the extra voltage will be dropped across $VT_1$



and $VT_2$, because more current flows through them as they attempt to charge $C_4$ to the increased voltage.)

This discrimination against noise bursts occurs at *any* signal level input. Thus, there is *no* "threshold" effect in a ratio detector—that is, there is no minimum signal value

means signal does not have to reach a certain point for noise reduction

input signal is coupled capacitively through the .001-mfd. condenser to $L_2$-$C_2$, which is coupled inductively to $L_1$-$C_1$. These two tuned circuits are adjusted for the 90° phase difference necessary for proper operation. The stabilizing condenser is 5 mfd. and its shunt resistor is 22,000 ohms, giving

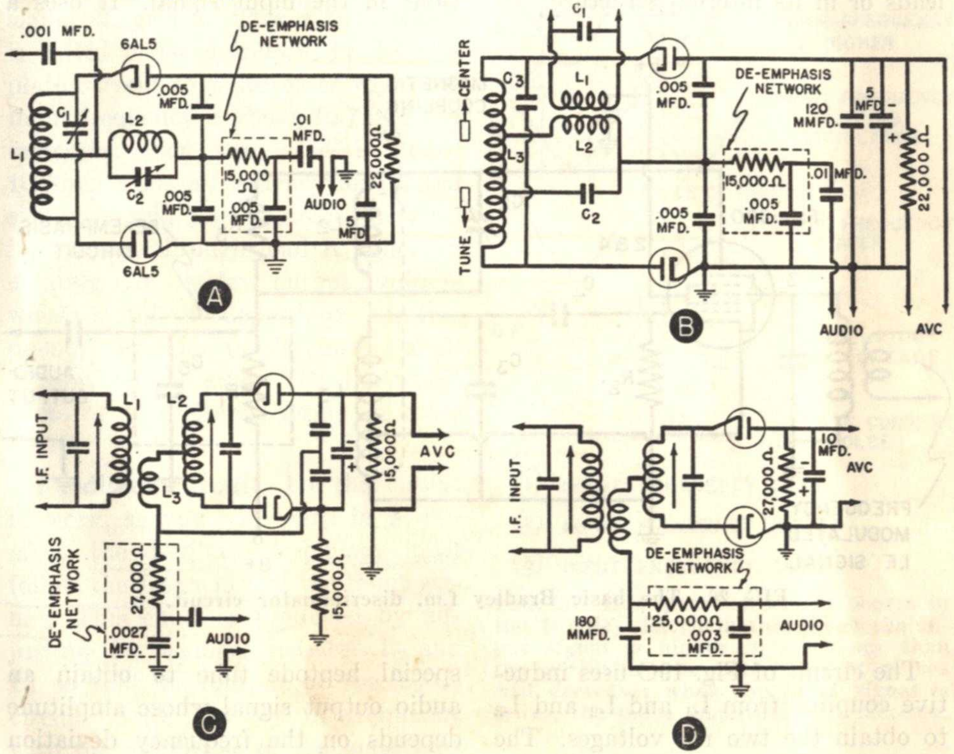


**FIG. 19. Four practical variations of the basic ratio detector are shown here.**

necessary before the noise reduction becomes effective.

**Practical Circuits.** Since the audio output voltage does not depend on the d.c. voltage across $R_1$ and $R_2$, the basic circuit of Fig. 17 can (when there is no direct d.c. path around $C_2$ or $C_3$) be simplified to the forms in Fig. 19A, B, C, and D. In Fig. 19A, the

a time constant of .11 seconds. The de-emphasis network, like those used in the other circuits in this figure, has the standard 75-microsecond time constant.

In Fig. 19B, the i.f. input is coupled inductively from $L_1$-$C_1$ to $L_2$-$C_2$ and thus to $L_3$-$C_3$. Two slug tuners are used to vary the inductance $L_3$, one to tune $L_3$-$C_3$ to the proper i.f. frequency and the other to center the discriminator curve, that is, adjust for the 90° phase difference. An .11-second time constant is used in this circuit also. The 120-mmfd. condenser by-passes r.f. that may not be by-passed through the 5-mfd. electrolytic condenser because of inductance in its leads or in its internal structure.

the two condensers that shunt the stabilizing condenser (the 180-mmfd. condenser is the other).

## BRADLEY F.M. DISCRIMINATOR

The Bradley f.m. discriminator is another type of f.m. discriminator that is insensitive to amplitude variations in the input signal. It uses a

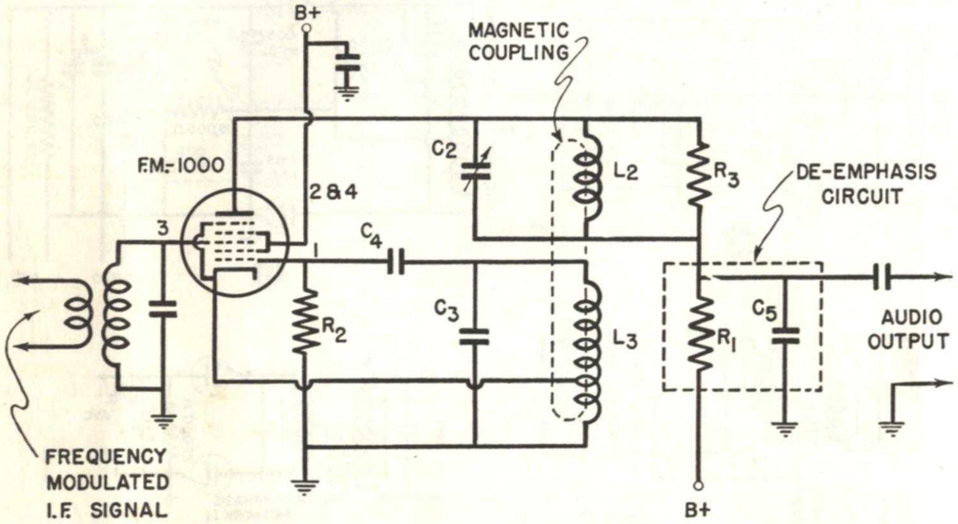


**FIG. 20. The basic Bradley f.m. discriminator circuit.**

The circuit of Fig. 19C uses inductive coupling from $L_1$ and $L_2$ and $L_3$ to obtain the two r.f. voltages. The shunt resistor consists of two 15,000-ohm resistors and therefore is similar to the basic circuit of Fig. 17. The de-emphasis network of 27,000 ohms and .0027 mfd. plus the stray capacity of the wiring gives the desired 75-microsecond time constant.

The last example of a ratio detector is shown in Fig. 19D. Inductive coupling is again used, and the circuit is further simplified by using distributed wiring capacitance for one of

special heptode tube to obtain an audio output signal whose amplitude depends on the frequency deviation of the incoming signal but is independent of its amplitude. This tube is similar to the pentagrid converter tubes used as mixers in superheterodyne receivers. It is, however, specially designed for use in this circuit.

Before we go into the details of circuit operation, let us first briefly preview the action of this discriminator. The basic Bradley circuit as it is used in Philco f.m. receivers is shown in Fig. 20.



**Circuit Operation.** The cathode, first and second grids of the heptode, and $L_3$, $C_3$, $C_4$ and $R_2$ form a conventional class C oscillator whose frequency of operation is 10.7 mc.—the center of the i.f. band. This oscillator produces plate current pulses whose amplitude depends on the phase difference between the oscillator voltage on grid 1 and the input signal on grid 3. This action, which is called *phase detection,* produces a plate current that decreases when the input frequency is above 10.7 mc. and increases when the input is below 10.7 mc. Thus, a frequency-modulated signal on grid 3 will produce a varying current through load $R_1$, thereby creating an audio output voltage whose amplitude depends on the frequency deviation of the input signal.

Now, let us study the important parts of this circuit action in more detail.

**Feedback Circuit.** For this circuit to work, as you will learn in a moment, the oscillator frequency must follow the frequency of the input signal. This effect is produced by adjusting the feedback between $L_2$ and $L_3$ so that the output of the $L_3$-$C_3$ circuit varies linearly in frequency with the amplitude of the plate current. In other words, the feedback is such that an increase in plate current causes a proportionate decrease in the oscillator frequency. Normally, when no signal is being received, the oscillator is tuned to the i.f. of the receiver (10.7 mc.).

**Phase Detection.** The plate current of the tube consists of a series of narrow pulses that occur at the oscillator frequency. When there is no input signal on grid 3, the height of these pulses—in other words, the amplitude of the plate current—is constant. When a signal is applied to grid 3, however, the height of the pulses is affected by the grid voltage, provided the pulses do not occur at

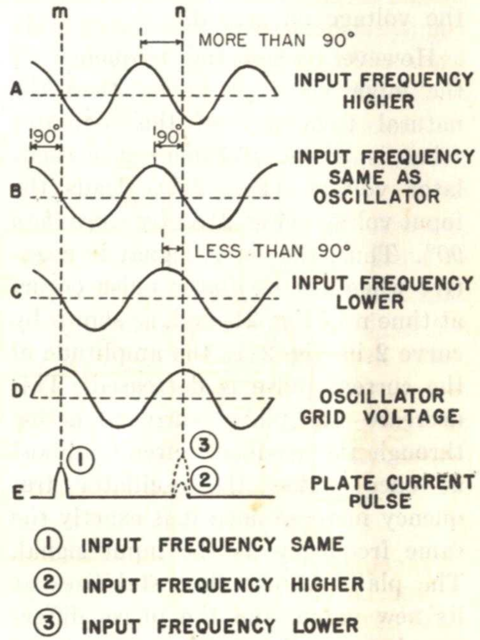


**FIG. 21. How phase detection occurs in the Bradley discriminator. When the input signal is higher in frequency than the oscillator, the plate current pulses will decrease; when the input signal is lower, the pulse amplitude will increase.**

the instant the grid voltage is going through zero. In normal operation, the signal applied to grid 3 is never large enough to cause plate current flow by itself; the grid 3 voltage acts solely as a modifier of the plate current pulses produced by the oscillator.

Fig. 21 shows in detail the voltages in the heptode tube. In B, the input voltage has the same frequency as the oscillator (10.7 mc.). The feedback circuit is adjusted so that the oscillator grid voltage (shown in D) is 90° behind the input signal. The plate current pulse of the oscillator (shown in E) thus occurs at the time when the input signal (B) is passing through zero. This means that the plate current pulse is unaffected by the voltage on grid 3.

However, when the frequency of the input signal is *higher* than the natural frequency of the oscillator (that is, above 10.7 mc.), the oscillator voltage (Fig. 21D) leads the input voltage (Fig. 21A) by *more than 90°*. Thus, the input signal is negative when the oscillator pulse occurs at time n in Fig. 21, and, as shown by curve 2 in Fig. 21E, the amplitude of the current pulse is decreased. This decrease in plate current, acting through the feedback circuit, almost at once makes the oscillator frequency increase until it is exactly the same frequency as the input signal. The plate current then stabilizes at its new value, and the phase difference between the oscillator signal and the input signal remains at the value that caused the increase in plate current. (You can easily see why this phase difference remains fixed. If it did not, the plate current would change again; since the plate current remains constant at its new value, however, it follows that the phase difference must remain constant also.)

Similarly, a decrease in the input frequency causes, as shown in Fig. 21C and curve 3 of Fig. 21E, an increase in the amplitude of the plate current pulse. This causes the oscillator frequency to decrease to the frequency of the input signal, and the plate current of the tube stabilizes at its new and higher value.

The result of these actions, as you can see, is that the amplitude of the plate current depends upon the frequency of the input signal. If the input increases in frequency, the amplitude of the plate current decreases; conversely, if the input decreases in frequency, the plate current amplitude increases. The plate current therefore consists of a series of pulses that vary in amplitude according to the frequency of the input signal. This current, flowing through the de-emphasis circuit $R_1$-$C_5$, develops across $R_1$ a voltage that follows the envelope of the pulses — in other words, an audio voltage that is a reproduction of the original modulating voltage.

**Discrimination Against Amplitude Modulation.** When the input signal frequency is above or below 10.7 mc., any increase in its amplitude tends to cause an increase in the amplitude of the plate current pulse. However, any current increase causes the oscillator frequency to decrease momentarily so that the phase difference between the two signals decreases; this causes the current to decrease, thereby correcting the original increase. The result is that changes in the amplitude of the input signal cause no change in the plate current. *The discriminator output thus is independent of the amplitude of the input signal.* This is a very important characteristic of the circuit.

Of course, when the input frequency deviates from 10.7 mc., it must have at least a certain minimum amplitude to cause the oscillator to lock-in with the input signal. For a deviation of 75 kc., the i.f. input must be at least ½ volt r.m.s.

Since this circuit is insensitive to amplitude variations only when the feedback from input signal to oscillator is through $L_2$, grid 3 of the heptode tube is carefully shielded from the other elements of the tube. This prevents any undesired capacitive coupling that would make the circuit respond to amplitude variations in the input.

**De-Emphasis.** The plate load resistance $R_1$ and condenser $C_5$ form the standard 75-microsecond de-emphasis network used in f.m. broadcast receivers.

**Review.** Let us briefly review the operation of this detector. The local oscillator section of this tube normally operates on 10.7 mc., the oscillator voltage is 90° behind the input signal, and the input signal therefore has no effect on the plate current of the tube. When the input frequency increases, the phase difference between the oscillator voltage and the input signal becomes more than 90°; the plate current then decreases, bringing the oscillator up to the frequency of the input signal. Its voltage remains more than 90° ahead of the input signal. When the input signal drops below the i.f. in frequency, the phase difference between the oscillator voltage and the input signal becomes less than 90°; the plate current then increases, and the oscillator frequency drops to that of the input signal. Its voltage remains less than 90° ahead of the input signal. As a result of these actions, changes in the input frequency cause the plate current to change linearly, but amplitude changes in the input signal do not affect the plate current.

# Narrow-Band F.M. Receivers

The f.m. receivers we have been studying are designed for wide-band (±75 kc.) f.m. broadcasting. There are, however, a number of point-to-point fixed and mobile services (police, taxicab, radio telephone, public utilities, etc.) where narrow band (±15 kc.) f.m. is used.

The general characteristics of an f.m. receiver for these services is illustrated by Fig. 22, which is a block diagram of the 152-162 mc. FM-40X manufactured by Kaar Engineering Company.

## A PREVIEW OF CIRCUIT OPERATION

Notice first that this is a double i.f. (triple-detection) superheterodyne receiver. The advantages of this are that image rejection is secured in the high i.f. section, and selectivity and high gain are obtained in the low i.f. section. The first i.f. is between 5.3 and 5.7 mc., depending on the frequency of the input signal. Since the mobile band is only 10 mc. wide (152-162 mc.) these i.f. values mean that no signal in this band can cause an image signal in this receiver. In fact, this high i.f. frequency and the two r.f. stages reduce image interference from any source to a minimum.

The second i.f. frequency is 455 kc., a value commonly used in a.m. receivers. At this frequency, high amplification of the signal is possible. Since the f.m. band width is only 30 kc., resistance loading the i.f. transformer will give the necessary pass band without much sacrifice in gain. The use of so low a second i.f. is practical only in narrow-band f.m.; there would be little point in attempting to use a similar system for wide-band f.m., because the 455 kc. i.f. section would give practically no gain if it were designed to have the 150 kc. pass band required for wide-band f.m.

This receiver is designed to operate on only one frequency, so a crystal-controlled local oscillator can be used. The stability of the crystal oscillator assures simplicity of control and reliable operation.

The crystal oscillator provides both the local oscillator signals, one of them directly, one through a frequency multiplier. For example, if the received signal is 152 mc., the crystal frequency of 4.88855 mc. is multiplied by 30 in the frequency multiplier stage. This produces a 146.6565-mc. local oscillator signal, which, when combined in the first mixer stage with the received signal, will produce a difference frequency of 5.3435 mc. (the first i.f. frequency). In the second mixer, the output of the first i.f. section is combined with the fundamental of the crystal oscillator; this gives the second difference frequency (455 kc.).

After amplification in the second i.f. stages, the signal is applied to two amplitude limiter stages, to the third detector (which is a double-tuned frequency discriminator), and then to the audio amplifier.

The gain of any f.m. receiver is so high that random noises may be amplified to a noticeable and objectionable degree when no carrier is being



received. (Remember that when there is no carrier, there is no suppression of the noise. Of course, when a station is being received, the noise, as you learned earlier, is effectively completely suppressed.)

To eliminate this noise, which is particularly noticeable in narrow-band f.m. receivers, a "squelch" circuit is used. When the noise output of the receiver is excessive, the squelch circuit operates to block out the audio signal. This is done by amplifying and rectifying the noise output of the discriminator and using the d.c. voltage produced to cut off the audio amplifier. Thus, when no signal is being received, the noise in the output of the discriminator causes the audio amplifier to be inoperative. When a signal is being received, the noise is greatly reduced and the received signal is amplified. The result is interference-free reception at all times.

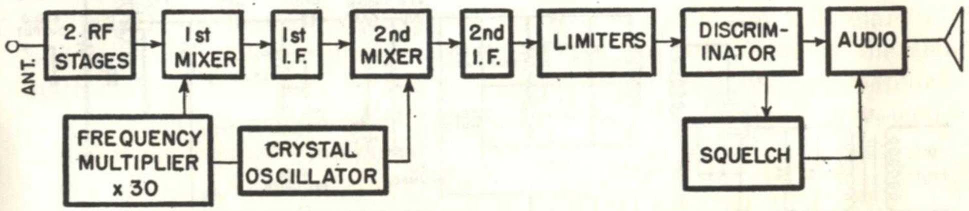


**FIG. 22. The block diagram of the Kaar FM-40X 152-162 mc. narrow-band mobile receiver.**

## CIRCUIT OPERATION

The output limiter, discriminator, noise amplifier, and squelch circuits of this receiver are shown in detail in Fig. 23.

The i.f. signal is applied through $T_1$ to $VT_1$, the last amplitude limiter stage. The output is then applied through $T_2$ to the f.m. discriminator. Two germanium crystal diodes (1N34's) are used in the double-tuned f.m. discriminator circuit. Since $C_1$ is 125 mmfd. and $C_2$ 175 mmfd., $L_1$-$C_1$ will tune to a frequency higher than 455 kc., $L_2$-$C_2$ to one lower than 455 kc. Thus, when the input signal varies from 440 to 470 kc., the d.c. output will vary in accordance with the frequency modulation of the input r.f. signal.

The audio output signal of this detector stage is applied to both the audio amplifier and the squelch circuit. Since the transmitters in so-called narrow-band f.m. are actually phase-modulated without pre-distortion (meaning that no inverse frequency network is used), the high audio frequencies in the received signal will be louder than they should be for natural reproduction. This frequency distortion is not very objectionable because of the limited range of modulating signals (250 - 3000 c.p.s.) and because of the other distortion in the system. To overcome this frequency distortion and make the frequency response more natural, $R_3$, $R_4$, and $C_3$ are used as a low-pass RC filter that reduces the amplitude of the high audio frequencies. $R_3$, $R_5$, and rheostat $R_6$ form the volume control circuit. A rheostat is used instead of a potentiometer because the volume control in many mobile receivers is in a remote unit some distance from the receiver, and only one wire and a ground return are needed for a rheostat, whereas a potentiometer requires two wires and a ground return. $VT_2$ is a conventional audio amplifier.

pends on the amplitude of the noise present. When the amount of noise is objectionable, this d.c. voltage biases the audio amplifier $VT_2$ beyond cut-off, thus "squelching" the noise signal. A rheostat varies the gain of $VT_3$ and can therefore be used to adjust the signal level that will produce this squelching action—in other words, it is the squelch sensitivity

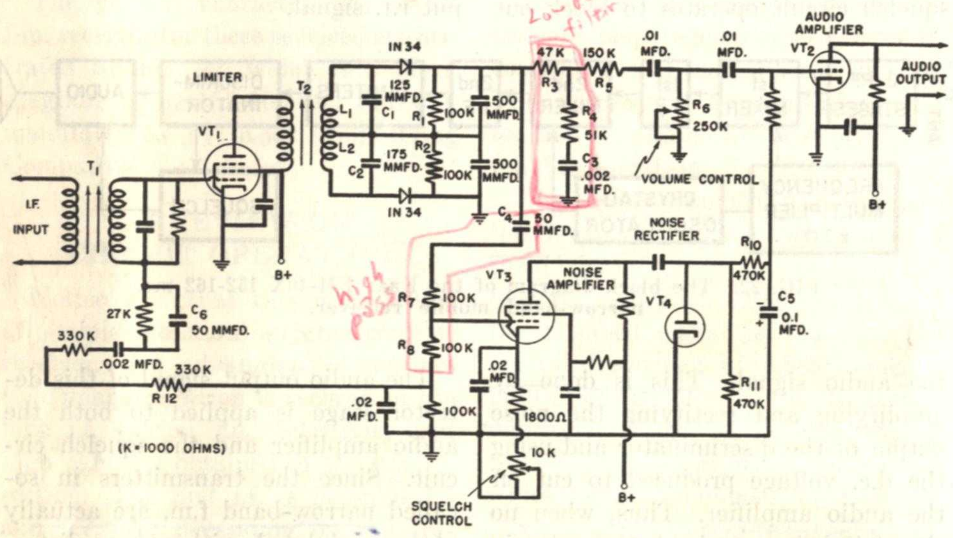


**FIG. 23. The final limiter, discriminator, audio amplifier and squelch circuit of the Kaar FM-40X narrow-band f.m. receiver. This circuit has many features that are typical of this type of receiver.**

**Squelch Circuit.** Since the range of voice frequencies transmitted in this system extends only from 200 to 3000 cycles, the higher frequency noise components of the audio output of the discriminator can be fed to $VT_3$ through a simple high-pass RC filter $C_4$, $R_7$, $R_8$ to $VT_3$. $VT_3$, therefore, amplifies only the noise present in the signal and is not affected by the voice frequencies. This amplified noise is rectified by diode $VT_4$ and produces a negative voltage across $C_5$ that de-

control. When a signal is being received, however, the noise is suppressed by the presence of the signal; $C_5$ then discharges through $R_{10}$ and $R_{11}$, allowing $VT_2$ to amplify the signal.

Notice that the limiter grid bias voltage across condenser $C_6$ is applied through $R_{12}$ and $R_8$ to the grid of noise amplifier $VT_3$. This is done to prevent the squelch circuit from operating on very strong audio signals. When the audio signal across $R_1$-$R_2$





is quite high, enough audio signal may be coupled through $C_4$ to cause the squelch to operate. However, a strong signal across $R_1$-$R_2$ means that limiter $VT_1$ is being overdriven and that there is therefore a negative d.c. voltage across $C_6$. This d.c. voltage, applied through $R_{12}$ and $R_8$, reduces the gain of the noise amplifier and prevents the squelch from operating.

# Antennas For F.M. Reception

A relatively strong, noise-free signal from the antenna is needed for best operation of any radio receiving system. For this reason, the signal field strength and signal-to-noise ratio should be as high as possible.

The three factors to be considered in choosing an antenna for f.m. reception are signal strength pickup, directivity, and inherent ability to reject interference. From the standpoint of signal pickup, a structure that is resonant or nearly resonant to the frequency to be received is always preferable. Since the f.m. band is relatively narrow—about 8 to 10 percent above and below the mean frequency—a resonant structure such as a half-wave dipole is very well suited for f.m. reception.

The requirements for directivity vary with the locality. In general, however, the antenna should be non-directive except when it is desirable to receive a particular station that delivers a relatively weak signal at the receiver location or when two f.m. stations operate on the same frequency and the only way of separating them is to have a directional antenna. Aside from these particular cases, which are the exception rather than the rule, it is desirable to have a non-directional f.m. antenna.

From the standpoint of the third requirement (rejection of interference from sources close to the receiver) the antenna should be horizontally polarized, because most noise, particularly man-made noise originating at ground levels, is vertically polarized. A horizontal dipole coupled to the receiver by a transmission line meets this requirement admirably.

From these considerations, it appears that a dipole is the most suitable antenna for f.m. reception.

## HALF-WAVE DIPOLES

Half-wave dipoles have a center impedance of about 75 ohms. Most f.m. receivers, however, are designed for a 300-ohm input. There are two ways of using dipoles to match this higher impedance. One way, shown in Fig. 24A, is to use a quarter-wave matching stub made of 150-ohm transmission line to match the 75 ohms to 300 ohms. A 300-ohm transmission line is used to connect the matching section to the receiver.

A second and more generally used scheme is shown in Fig. 24B. Here the antenna is a folded dipole whose

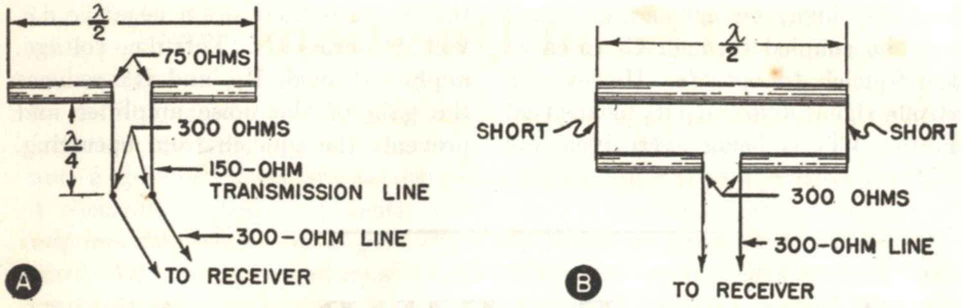


FIG. 24. A, half-wave dipole with 150-ohm quarter-wave matching stub to match it to a 300-ohm transmission line. B, a folded dipole can be connected directly to a standard 300-ohm line.

center impedance is 300 ohms (4 × 75 ohms); a 300-ohm line can be connected directly to this point.

**A.M. and F.M. Combination.** The standard a.m. broadcast band can be received on f.m. antennas, but the operation is not very efficient or satisfactory. However, a simple modification of the antenna can be made that will not affect its operation at very high frequencies, but will permit a longer conductor to be used at the a.m. broadcast frequencies only. This modification is shown in Fig. 25. It consists of adding a much longer wire to one arm of the dipole, and interposing a wave trap or a circuit that is anti-resonant to the f.m. band between the dipole and the added wire.

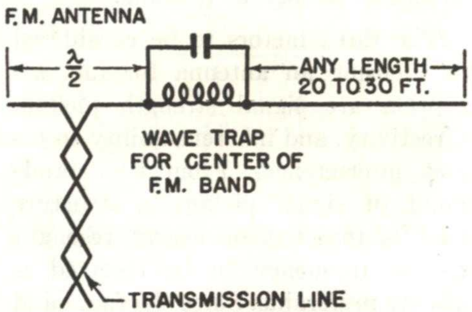


FIG. 25. Method of extending ½ wave dipole for broadcast reception on the standard broadcast band.





# Lesson Questions

**Be sure to number your Answer Sheet 48RC.**

**Place your Student Number on *every* Answer Sheet.**

**Most students want to know their grade as soon as possible, so they mail their set of answers immediately. Others, knowing they will finish the next Lesson within a few days, send in two sets of answers at a time. Either practice is acceptable to us. However, don't hold your answers too long; you may lose them. Don't hold answers to send in more than two sets at a time or you may run out of Lessons before new ones can reach you.**

1. What section of an Armstrong type of f.m. receiver removes amplitude variations from the f.m. signal?

2. If the deviation ratio of an f.m. system is increased, will the signal-to-noise ratio *increase, decrease,* or *remain the same?*

3. What are the two advantages of using a preselector stage in an f.m. receiver?

4. What is the main advantage of a high i.f. frequency in an f.m. receiver?

5. If the i.f. in an f.m. receiver is 10.7 mc., the local oscillator operates on a frequency higher than the input signal, and the receiver is tuned to 93.9 mc., what is the frequency of a signal that will cause an image?

6. What characteristics should a limiter have in order to block out sudden noise surges?

7. Why is it permissible to have a peaked over-all response for the r.f. and i.f. sections in a properly designed Armstrong type of f.m. receiver?

8. Is there a "threshold" effect in a ratio detector?

9. What are the advantages of using a double superheterodyne in narrow-band f.m. receivers?

10. When does the squelch circuit of a narrow-band f.m. receiver operate to block out the audio signal?

# TO BE INDEPENDENT, PRACTICE ECONOMY

To become truly independent, the practice of *simple economy* is necessary. And *economy* requires neither superior courage nor great virtue. It requires only ordinary energy and consistent attention. Essentially, *economy* is the spirit of orderliness applied to the administration of your own *personal affairs*. It means management, regularity, prudence and the avoidance of waste.

*Economy* also requires the power to resist present gratification of your wants, in order to secure future benefits. And even wild animals practice this *economy* when they store food for the winter!

Yes—the practice of *economy* is necessary unless and until you figure out some fool-proof way to make money faster than you can spend it! I'll admit that a few men have been able to do this—but until you can discover this golden secret, your best road to independence will be the day in and day out practice of *reasonable economy*.

J. E. Smith