

**FREQUENCY-MODULATED SIGNALS**

**THE F.M. RECEIVER**

**34FR-2**

**NATIONAL RADIO INSTITUTE**

ESTABLISHED 1914

**WASHINGTON, D. C.**



# Study Schedule No. 34

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

1. Brief A.M. Review; Characteristics of a Frequency-Modulated Signal - - - - - Pages 1-9

Spend at least twice as much time as usual on this first study step, because in it is the basis for the entire frequency modulation system of broadcasting. Once you get a clear picture in your mind of what an f.m. signal is like, you'll find that f.m. systems are no more difficult than the ordinary a.m. systems you have previously studied. Answer Lesson Questions 1 and 2.

2. An Elementary F.M. System; Additional Requirements in an F.M. System - - - - - Pages 9-13

Here is a brief survey of an entire f.m. system from the microphone through the f.m. transmitter and through the f.m. receiver. Simplified circuits and explanations are used so that you can concentrate upon getting a broad general picture first. Answer Lesson Question 3.

3. The F.M. Receiver; The Preselector; The Frequency Converter Pages 13-18

The block diagram of a combination f.m.-a.m. receiver in this section will prove a valuable reference guide for studying the remainder of this lesson, because it shows at a glance the relationships between the various stages of the receiver. Study each circuit slowly and carefully, referring to the circuit diagram each time a part is mentioned.

4. The I.F. Amplifier - - - - - Pages 18-27

Careful, slow study is required here also. You'll learn how the same i.f. tubes can be used both for f.m. and a.m. signals, and you'll learn how the limiter removes amplitude variations from f.m. signals. Answer Lesson Questions 4, 5, 6 and 7.

5. The F.M. Detector - - - - - Pages 27-32

Here's another section in which careful study will be well worth while if you intend to prepare yourself for any job whatsoever connected with f.m. systems. Answer Lesson Questions 8, 9 and 10.

6. Modern F.M. Transmitters; The Future of F.M. - - - - - Pages 32-37

This bird's-eye view of an f.m. transmitter completes your background knowledge of f.m. systems. With this knowledge, you should have no difficulty in mastering the details of any f.m. receiver maintenance job once you have completed the N.R.I. Course.

7. Mail your Answers for this Lesson to N.R.I. for Grading.

8. Start Studying the Next Lesson.

# Frequency-Modulated Signals

## Brief Review of Amplitude-Modulated Signals

**A** THOROUGH understanding of the characteristics of a conventional amplitude-modulated (a.m.) signal will help you to picture in your mind the characteristics of a frequency-modulated (f.m.) signal.

The signal peaks of an unmodulated r.f. carrier will all have the same amplitude, as shown at the left in Fig. 1A. When the r.f. carrier is amplitude-modulated with a single sine-wave audio frequency (such as that corresponding to a pure audio tone), the peaks of the r.f. signal will rise and fall in amplitude in accordance with variations in the audio signal, as shown at the right in Fig. 1A.

A dash-dash line drawn through the positive peaks of the modulated r.f. signal in Fig. 1A will give a curve known as the *envelope*, which has the same wave form as the original audio signal. (Another envelope having this same audio wave form is obtained by drawing a dash-dash line through the negative peaks, but at this time we need consider only the upper envelope. Both envelopes are shown for each signal in Fig. 1.)

Horizontal line *XX* at the level of the unmodulated carrier peaks will be the reference line for this envelope. Thus, we can say that the positive peak of the envelope goes above reference line *XX*, and the negative peak of the envelope goes below reference line *XX*. Furthermore, with 50% modulation as in Fig. 1A, the envelope peaks will have exactly one-half the amplitude of the unmodulated carrier peaks.

When the envelope peaks are equal in amplitude to the unmodulated car-

rier peaks, as in Fig. 1B, we have 100% modulation of the carrier.

When the envelope peaks are greater in amplitude than the unmodulated carrier peaks, the modulation percentage is higher than 100% and the envelope becomes distorted, as shown in Fig. 1C. When an over-



Courtesy General Electric Co.

A million-volt discharge of man-made lightning created a crashing roar which completely drowned out the musical program when this combination f.m.-a.m. receiver was tuned to a broadcast band a.m. station. When the set was tuned to an f.m. station carrying the same program, however, the music emerged from the loudspeaker clear and strong, with only a hardly noticeable static buzz in the background even though a million volts of electricity was dancing and sputtering a few feet away. Truly this is dramatic proof of the noise-reducing characteristics of an f.m. system.

modulated signal like this is demodulated in a receiver, the resulting audio signal will have the same distorted wave form as the dash-dash envelope in Fig. 1C.

When the frequency of the audio signal is varied, the frequency of the envelope changes in a corresponding

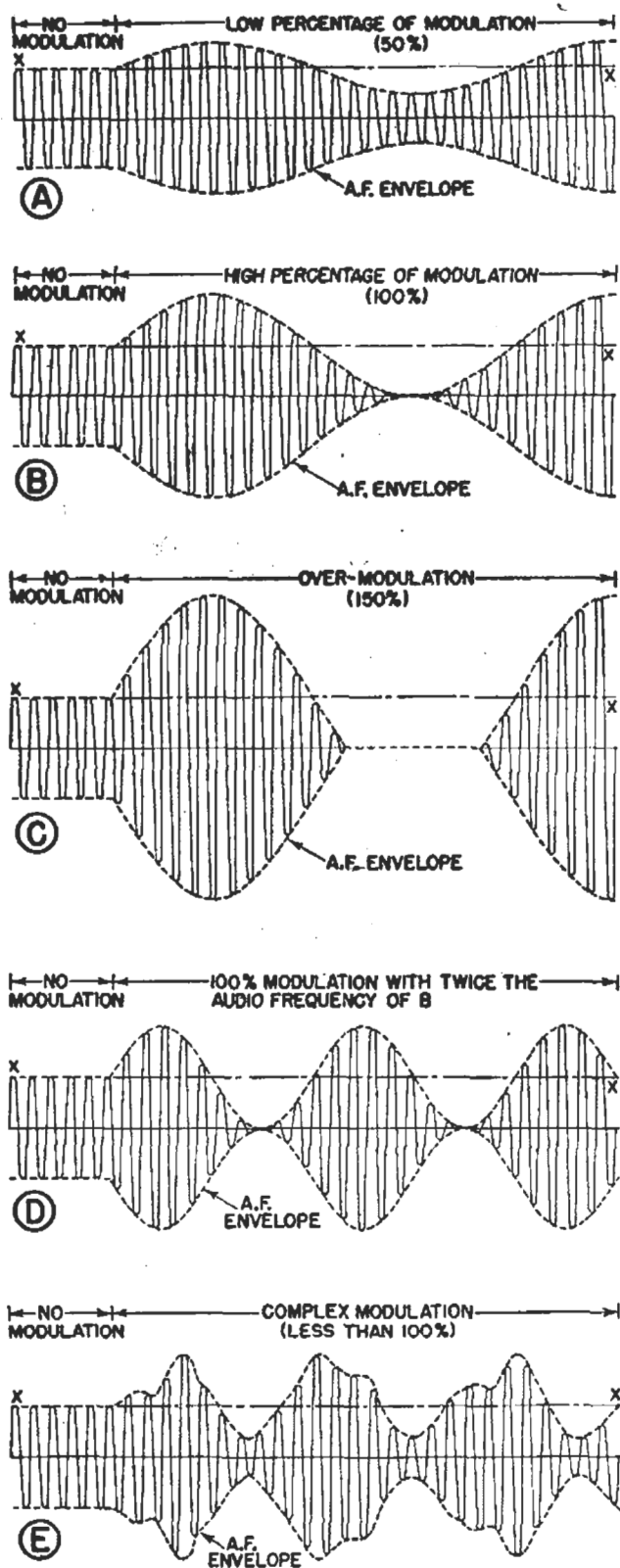


FIG. 1. Examples of amplitude-modulated signals.

manner. Thus, if we double the frequency of the sine-wave audio tone used for our first three examples in Fig. 1, we secure for 100% modulation the condition shown in Fig. 1D.

When voice or music is transmitted, the modulation envelope becomes far more complex, resembling that shown in Fig. 1E. When this envelope is

analyzed, it is found to have many different audio frequency components.

**Side Frequencies.** As you already know, an r.f. carrier signal which is modulated with a single fixed-frequency sine wave tone is equivalent to three different pure r.f. signals. One will have the assigned carrier frequency and constant amplitude. The other two, called *side frequencies*, will be respectively above and below the carrier frequency by the audio frequency value, and each one will vary in amplitude between zero and one-half the carrier amplitude as the percentage modulation varies between 0 and 100%.

Here is an example. If the highest frequency we wish to transmit is 5000 cycles (5 kc.), and the r.f. carrier frequency is 1000 kc., the three pure r.f. signals will be 1000 kc. (the carrier), 995 kc. and 1005 kc. (the side frequencies). If the lowest frequency to be transmitted is 100 cycles (.1 kc.), the two side frequencies going out with the 1000-kc. carrier will be 999.9 kc. and 1000.1 kc.

If a complex audio signal having many frequencies in the range from 100 cycles to 5000 cycles is being transmitted, there will be a 1000-kc. r.f. carrier signal traveling through space along with side frequencies ranging from 995 kc. to 999.9 kc. and from 1000.1 kc. to 1005 kc.

The percentage modulation for a complex signal varies from instant to instant; it is 100% when the sum of all the side frequency amplitudes at a particular instant of time is exactly equal to the carrier amplitude. Over-modulation occurs whenever the sum of all the side frequency amplitudes is greater than the carrier amplitude.

### Characteristics of a Frequency-Modulated Signal

Modulation of an r.f. signal can also be accomplished by varying the

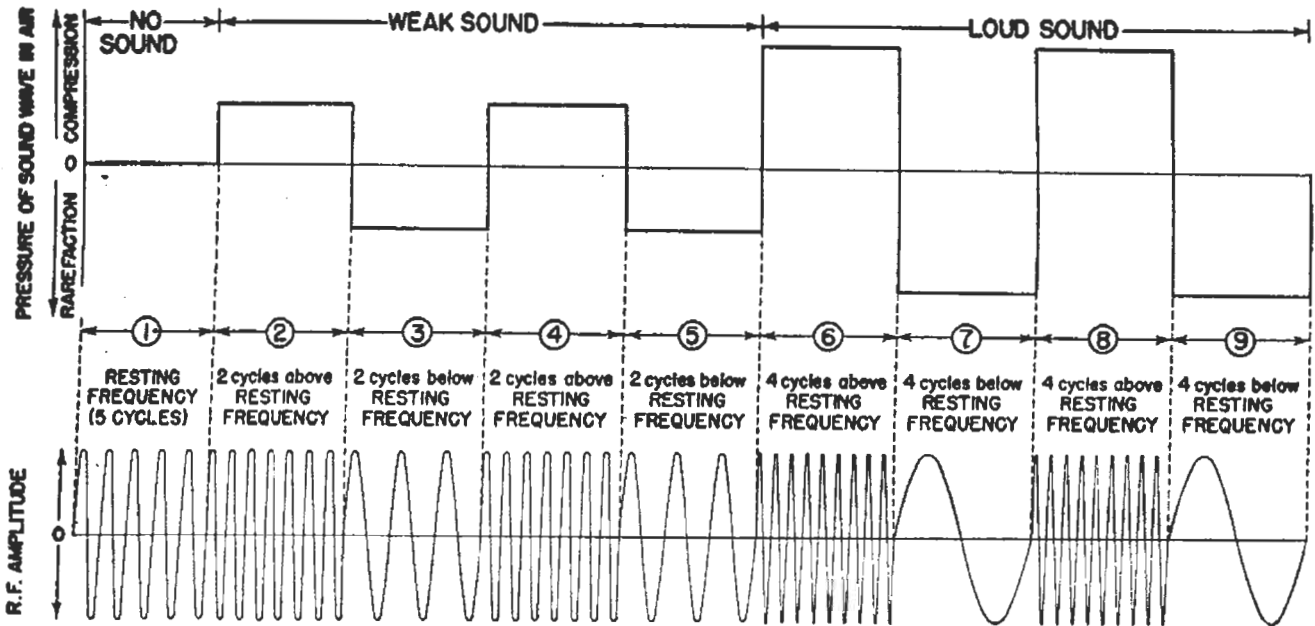


FIG. 2. Example showing how a sound wave having a square-wave form makes the r.f. signal frequency of an f.m. transmitter vary above and below the assigned resting frequency.

frequency of the r.f. signal in accordance with sound pressure, while keeping the r.f. amplitude essentially constant. This, basically, is *frequency modulation*.

Sound is an alternate compression and rarefaction of air particles. This means that air in the path of a sound wave is alternately being increased and decreased in pressure with respect to normal atmospheric pressure. When a sound wave in air is converted into its electrical equivalent by a microphone, we have a corresponding audio signal with positive and negative alternations, one alternation representing compression and the other rarefaction of air. The amplitude of an alternation depends upon the loudness of the original sound in air, and the frequency of the audio signal depends upon the pitch (frequency) of the original sound.

*The Resting Frequency.* With frequency modulation, the r.f. signal which is radiated by the transmitter is at an assigned r.f. value called the *resting frequency* or *center frequency* whenever no sound is being transmitted. The resting frequency thus corresponds to the instant when the air at the microphone diaphragm is at

*normal atmospheric pressure.* This is illustrated graphically by time interval 1 in Fig. 2. (The resting frequency of an actual f.m. transmitter would be some value above 88,000,000 cycles per second, but for illustrative purposes we arbitrarily choose 5 cycles to represent the resting frequency in this diagram.)

Although the compression half of a sound wave cycle in air could cause either an increase or decrease in the frequency of an actual f.m. transmitter (depending upon the number of a.f. stages and phase-reversing transformers which are between the microphone and the output of the modulator), it is a convenient and common practice among radio men to associate compression with an *increase* in f.m. transmitter frequency. We will follow this practice.

*Frequency Deviation.* When an a.f. signal is fed into an f.m. transmitter, the *frequency* of the r.f. signal thus swings *above* the resting value in proportion to increases in air pressure from the normal atmospheric value (compression). Likewise, the transmitter frequency swings *below* the resting value in proportion to decreases in air pressure from the nor-

mal atmospheric value (rarefaction). The amount of this swing above or below the resting frequency (the amount by which the instantaneous r.f. value differs from the resting value) is known as the *frequency deviation*. We illustrate these r.f. signal frequency changes in Fig. 2 for a sound having a square-wave form.

*Square-Wave Example.* Time intervals 2, 3, 4 and 5 in Fig. 2 represent two cycles of a weak sound having a square-wave form. During the first half cycle (interval 2), the air directly in front of the microphone is compressed, and the r.f. carrier frequency for this interval is therefore *higher* than the resting frequency. We have indicated this by showing 7 complete cycles for time interval 2 (two more than for the resting frequency in interval 1).

During the second half of the first sound cycle, we have rarefaction at the microphone, and the r.f. carrier frequency for this interval is *lower* than the resting frequency. We indicate this by showing 3 cycles for time interval 3 (two less than for interval 1).

In the second cycle of the weak sound, the process repeats itself, with the r.f. carrier frequency going above the resting value for time interval 4, and going below the resting value for interval 5.

In an f.m. system, doubling the sound pressure of the original sound doubles the deviation in transmitter frequency from the resting value. Time intervals 6, 7, 8 and 9 in Fig. 2 illustrate this for the square-wave sound under consideration.

During the first half cycle of this louder sound wave (interval 6), we have compression, and the r.f. signal frequency is twice as much above the resting frequency as it was for corresponding interval 2 of the weak sound. We indicate this in Fig. 2 by showing

9 complete cycles for time interval 6 (this is 4 cycles above the resting value).

For the rarefaction half of the loud sound cycle (interval 7), the r.f. signal frequency swings just as much below the resting frequency as it swung above the resting frequency for the compression half cycle. Thus, we show 1 cycle for time interval 7, this being 4 cycles below the 5-cycle value for the resting frequency.

The 4-cycle swing above and below the resting frequency is repeated for intervals 8 and 9 in Fig. 2, to give the second cycle of the loud sound.

*Sine-Wave Example.* The square-wave audio signal in Fig. 2 showed how sudden changes in the amplitude of an audio signal affect the output signal frequency of an f.m. transmitter. Now let us see how gradual changes in amplitude, such as those occurring in the sine-wave audio signal shown in Fig. 3, will affect an f.m. transmitter. This diagram can either represent variations in air pressure from a normal atmospheric value (as in Fig. 2) or positive and negative a.f. signal voltage swings.

First of all, we can say that the r.f. signal will be at its resting value whenever the audio signal passes through zero, such as at points *a*, *e* and *g*.

As the audio signal increases in amplitude from *a* to *b* to *c*, the frequency of the r.f. signal will rise above the resting value in a similar manner, to a maximum r.f. value corresponding to peak amplitude *c*. As the audio signal decreases gradually in amplitude to zero again at point *e*, the r.f. signal frequency will drop gradually back to the resting value in a similar manner. Note particularly that for the entire interval of time from *a* to *e* when the a.f. signal is positive, the transmitter frequency is *above* the resting value.

When the a.f. signal goes through its negative half cycle from *e* to *f* to *g*, the r.f. signal frequency will decrease in a correspondingly gradual manner from its resting value to a minimum value corresponding to point *f*, then rise gradually again to the resting value to complete the sine-wave cycle.

**Complex Voice or Music Signals.** When a complex audio signal like that shown in Fig. 3B is transmitted by an f.m. system, the transmitter frequency will vary above and below the resting value in accordance with the amplitude and polarity of the audio signal at each instant, even though this voice or music signal contains many different audio frequencies.

f.m. as they are used today.

Theoretically, the full audio frequency spectrum with a full range of volume can be handled satisfactorily by an f.m. system regardless of how small or how large the maximum deviation value may be. In actual practice, however, the added requirements of a high signal-to-noise ratio and minimum inter-station interference at receivers make necessary a *high* value for the *maximum frequency deviation*. The greater the maximum frequency deviation employed for desired signals, the less noticeable to the listener will be a given frequency deviation due to an interfering signal.

The channels which have been made available to f.m. broadcast stations in

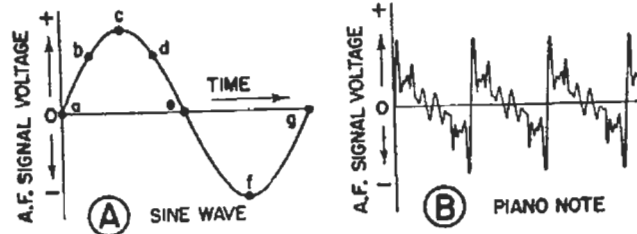


FIG. 3. A pure sine wave signal like that at A or a complex audio signal like that at B both make the frequency of an f.m. transmitter swing above and below its resting value exactly in proportion to the positive and negative swings of the audio signal.

Thus, we can say that the r.f. signal frequency will be *above* the resting value whenever the audio signal is *positive*, with the deviation from the resting value being proportional to the positive amplitude at each instant. Likewise, the r.f. signal frequency will be *below* the resting value whenever the audio signal is negative, with the deviation being proportional to the negative amplitude at each instant.

**Amount of Frequency Deviation.** Keeping in mind the fundamental f.m. fact that the instantaneous deviation in the frequency of the r.f. signal is proportional to the instantaneous amplitude of the audio signal, let us now consider actual frequency values for

the United States and its possessions by the Federal Communications Commission are .2 mc. apart in the very high frequency band between 88 and 108 mc., with the lower portion of the band from 88 to 92 mc. being reserved for non-commercial educational f.m. stations. Each channel assignment represents *the assigned resting frequency* of the station (the unmodulated signal frequency of the station).

A guard band at least 25 kc. wide is required by the F.C.C. beyond each extreme frequency swing of a station, which leaves a maximum of 75 kilocycles for the permissible frequency swing in either direction from the assigned resting frequency.

Percentage modulation, as we know it in connection with the a.m. system of broadcasting, does not exist in an f.m. system. Since the greatest permissible frequency deviation for a loud sound is 75 kc., this deviation corresponds to 100% modulation of an f.m. transmitter. Over-modulation (a deviation greater than 75 kc.) will not cause distortion in receivers, however, unless it makes the signal sweep beyond the linear portion of the S curve for discriminator action (to be considered later in this lesson).

The monitor engineer in an f.m. studio has meters before him which tell when the loudness level at the microphone is exceeding the value which gives the full 75-kc. deviation. Whenever necessary, he reduces the gain of the studio a.f. amplifier enough to prevent the transmitter from swinging more than 75 kc. off from the resting frequency. Likewise, the monitor increases the a.f. gain when the loudness level at the microphone drops way down for appreciable periods of time. Thus, the goal of the monitor engineer is essentially the same, for both f.m. and a.m. systems: To keep the audio level at receivers as high as is practical without making the program sound unnatural and without causing over-modulation at the transmitter.

*Actual Example.* Suppose that an f.m. station is assigned the 97.3-mc. channel, and its microphone is picking up a loud 1000-cycle sound. The resting frequency of this station will be 97.3 megacycles, and the maximum permissible deviation will be 75 kc. (.075 mc.) above and below this resting value. Thus, the r.f. signal frequency will go up to 97.375 mc. for the positive peak of each audio cycle, and will drop down to 97.225 mc. for the negative peak of each audio cycle. Since the audio signal passes through 1000 complete cycles in each second,

the f.m. transmitter will go through 1000 complete swings (from 97.3 to 97.375 to 97.3 to 97.225 to 97.3) in each second in order to follow the positive and negative amplitude variations of the audio signal.

If the lowest loudness level to be transmitted in this example is 1/100 of the loudest level, the deviation for this weakest audio signal will be .075 mc. divided by 100, or .00075 mc. For this weak signal, then, the r.f. signal frequency will vary from 97.30075 mc. to 97.29925 mc. and there will be 1000 complete swings like this in each second.

If the f.m. station in our example is transmitting a loud 100-cycle audio signal, the deviation will still be 75 kc. on each side of the resting frequency, but now there will be only 100 complete swings back and forth between the two extreme frequencies in each second.

*F.M. Side Frequencies.* Mathematical computations as well as actual measurements tell us that the continually varying r.f. signal frequency for an f.m. transmitter is equivalent to a carrier frequency and an infinitely large number of side frequencies. Fortunately, however, the essential side frequencies needed for high-fidelity reproduction are within the 150-kc. deviation range over which the r.f. signal sweeps. Receivers are designed for reception of this range, and consequently they receive the essential side frequencies.

Side frequencies more than 75 kc. off from the resting frequency might create interference with adjacent-channel stations if these stations were in the same locality, but the Federal Communications Commission prevents interference of this nature by keeping channel assignments at least 400 kc. (.4 mc.) apart for f.m. stations in the same service area.



### Technical Data on Side Frequencies.

When the carrier of an f.m. transmitter is modulated with an audio signal having a constant frequency of  $f$  cycles per second, the r.f. signal swinging above and below the resting frequency is equal to the following fixed-frequency signals: A signal having the resting frequency of the transmitter; two side frequencies, one above and one below the resting frequency by the audio value  $f$ ; two side frequencies, above and below the resting frequency by twice the audio value  $f$ ; two side frequencies, above and below the resting frequency by three times the audio value  $f$ ; etc.

Here is an example. If the resting frequency of an f.m. transmitter is 103.5 mc. and the audio modulation frequency is 10,000 cycles (.01 mc.), the equivalent individual frequencies will be: 103.5 mc., 103.49 mc. and 103.51 mc., 103.48 mc. and 103.52 mc., 103.47 mc. and 103.53 mc., etc., down to zero and up to infinity.

In your work with f.m. apparatus, however, you will find it more convenient to think of an f.m. signal as a single r.f. signal which is continually varying in frequency above and below its resting frequency. (We do this in amplitude modulation when we consider a modulated r.f. signal as a single r.f. signal which is varying in amplitude from instant to instant as in Fig. 1A.)

**Multiplex F.M. Operation.** Frequency modulation has the added advantage of permitting multiplexing, the sending of two or more programs or types of intelligence by the same transmitter, without increasing the required band width for the station and without interference between the programs being transmitted.

Here is an example: To transmit code messages along with a broadcast program, a 20-kc. oscillator signal (above the audio range) could be made to vary in amplitude in accordance with the code modulation, and this modulated 20-kc. secondary carrier then fed into the transmitter along with the regular voice or music program. At the receiver, the voice or music program would pass through the stages in the conventional manner, and the 20-kc. modulated code signal would be taken out ahead of

the audio amplifier, and demodulated by conventional means in a separate detector stage. Filters would be used to keep the 20-kc. signal out of the regular a.f. amplifier, to prevent overloading of a.f. stages.

**How F.M. Systems Reduce Interference.** At any instant of time, we can consider that an f.m. signal has a definite frequency, above or below the resting frequency of the transmitter. We can therefore represent a desired f.m. signal at the input of an f.m. receiver by a vector, as shown in Fig. 4A. The amplitude (length) of this desired-signal vector  $D$  will be constant, and the rate at which the

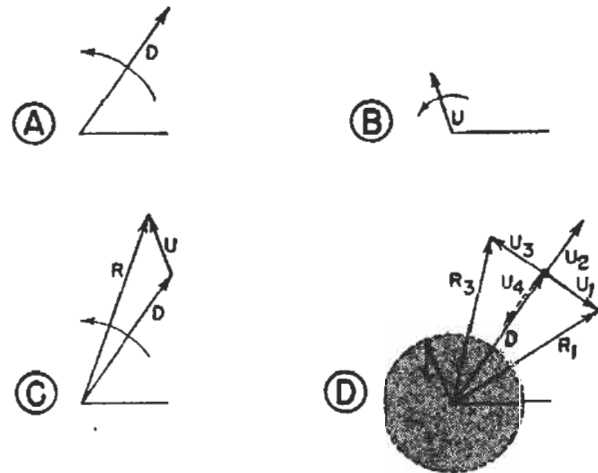


FIG. 4. Vector diagrams showing how an undesired signal  $U$  can affect the frequency and amplitude of a desired f.m. signal  $D$ .

vector rotates counter-clockwise will correspond to the frequency of the r.f. signal at the instant of time under consideration. The position of the vector with respect to the horizontal reference line in the diagram is unimportant.

Now, let us suppose that we also have at the input of our f.m. receiver an undesired signal. It may be due either to noise interference or to an undesired r.f. carrier signal from some other station, but nevertheless we know that this undesired signal has both frequency and amplitude. Let us say that at the instant of time under consideration for Fig. 4A, the

undesired signal has the amplitude and phase represented by vector  $U$  in Fig. 4B.

When we combine undesired-signal vector  $U$  with desired-signal vector  $D$ , as shown in Fig. 4C, we obtain resultant vector  $R$ . The amplitude of vector  $R$  is greater than that of desired-signal  $D$ , showing that an interfering signal can affect the amplitude of a desired incoming signal. Furthermore, vector  $R$  is ahead of vector  $D$ , indicating that the sudden arrival of this interfering signal makes desired-signal vector  $D$  increase its rotational speed suddenly; this means that during the time interval in which vector  $D$  is moving to its new position corresponding to vector  $R$ , the frequency of the desired signal is higher than it should be.

Interference vector  $U$  in Fig. 4B may have any phase relationship whatsoever throughout the entire range of  $360^\circ$ . It may be in phase with desired-signal vector  $D$ , as indicated by interference vector  $U_2$  in Fig. 4D; it may be  $180^\circ$  out of phase, as indicated by interference vector  $U_4$  in Fig. 4D; it may lead the desired vector by  $90^\circ$ , as indicated by  $U_3$ , or it may lag the desired vector by  $90^\circ$ , as indicated by  $U_1$ . These are the four extremes which could exist, and an analysis of them will take care of intermediate undesired-signal vector positions as well.

Careful study of the vector diagram in Fig. 4D reveals that when the interfering signal is *in phase* with the desired signal, it *increases* the amplitude of the resultant signal without changing its frequency. When the interfering signal is *180° out of phase* with the desired signal, it *decreases* the amplitude of the resultant signal without changing its frequency. When the interfering signal leads the desired signal, the resultant will be speeded up in frequency, and the amplitude

will generally also be altered. Finally, when the interfering signal lags the desired signal, the resultant will be slowed down in frequency and will likewise be altered in amplitude. We thus come to the conclusion that an interfering signal in an f.m. system can change both the *frequency* and the *amplitude* of the desired f.m. signal.

You will learn later in this lesson that the limiter stage of an f.m. receiver reduces all incoming signals to a constant low amplitude corresponding to radius  $L$  in Fig. 4D. Since changes in amplitude due to noise vectors are all outside of this acceptance area, the limiter stage of a properly

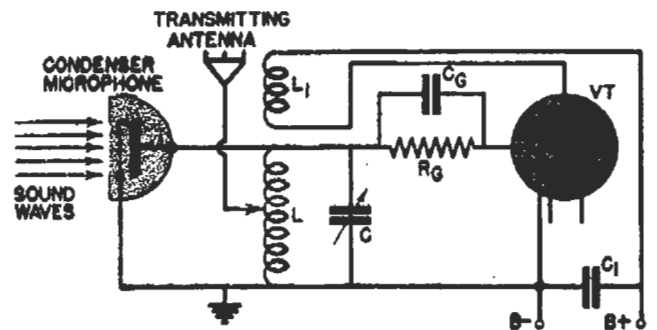


FIG. 5. Simple f.m. transmitter circuit, employing a condenser microphone shunted across the tank circuit of a conventional tuned-grid r.f. oscillator.

designed and adjusted f.m. receiver wipes out amplitude changes due to noise.

Any change in frequency due to noise will show up in the output of an f.m. receiver, for changes in frequency are not affected by the limiter section of the receiver, and are converted into corresponding changes in amplitude by the following frequency discriminator section. By using a high maximum deviation value, however, changes in the frequency of the desired signal due to an interfering signal in the receiver can be made negligibly small with respect to frequency changes due to the desired audio program.

Field tests have proved that the ratio of desired to undesired signals increases as the maximum frequency

deviation of the system is increased. Tests have also proved that with a maximum deviation of about 75 kc. (a total swing of 150 kc.), interfering signals become quite imperceptible at the loudspeaker when the desired signal is at least twice as strong as the undesired signal.

### An Elementary F.M. System

*The F.M. Transmitter.* In beginning our study of actual f.m. systems, let us first consider one of the simplest possible transmitter circuits capable of producing frequency-modulated signals.

A conventional vacuum tube oscillator circuit employing a coil-condenser resonant circuit can be tuned to frequencies in either direction from a reference frequency without appreciably changing the amplitude of the output signal. By taking a simple oscillator circuit like this and shunting its resonant circuit with a condenser microphone in the manner shown in Fig. 5, a simple but effective f.m. transmitter is obtained. Sound waves apply to the movable diaphragm of the microphone a varying pressure which makes this diaphragm move alternately towards and away from the fixed plate. This changes the capacity of the condenser microphone in accordance with variations in sound pressure, thereby alternately raising and lowering the frequency of the r.f. oscillator to give the desired frequency-modulated signal.

There are a number of drawbacks to this simplified f.m. transmitter circuit. The condenser microphone would have to be close to the oscillator, to prevent pick-up of stray signals by the microphone leads. Each condenser microphone would have to be designed to give a definite frequency deviation with its particular oscillator, and even with this precaution the amount of deviation would vary with

the particular performer using the microphone. Undoubtedly these problems could be solved by engineers if necessary, but fortunately there are more practical means for securing frequency modulation of a carrier signal. One such scheme will now be considered.

*Inductance-Tube Type of F.M. Transmitter.* A vacuum tube circuit can be made to act like an inductance, simply by adjusting circuit voltages so that *the r.f. current drawn by the tube will lag the r.f. voltage applied to the tube.* (This scheme is widely employed in the automatic frequency

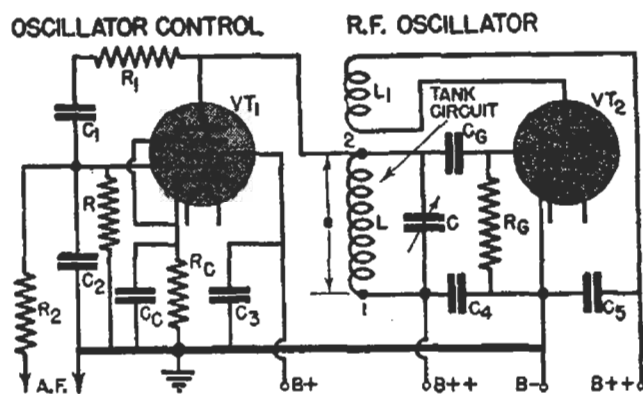


FIG. 6. This practical f.m. transmitter circuit employs an oscillator control circuit consisting of a vacuum tube which acts like an inductance shunted across the r.f. tank circuit. The effective inductance value varies from instant to instant in accordance with variations in the amplitude of the a.f. signal, thereby giving frequency modulation.

control circuits of some radio receivers.) By connecting across the coil-condenser resonant circuit of an r.f. oscillator the vacuum tube circuit which is acting like an inductance, and by making the inductance of this vacuum tube circuit vary at a desired audio rate, we secure frequency modulation.

An inductance-tube type of f.m. transmitter circuit is given in Fig. 6. Triode vacuum tube  $VT_2$  is connected into a conventional tuned-grid r.f. oscillator circuit, with  $L$  and  $C$  forming its tank circuit. Pentode tube  $VT_1$  serves as the oscillator control tube which acts like an inductance; its plate is connected directly to termi-

nal  $\mathcal{Z}$  of tank inductance  $L$ , and its cathode is connected to terminal 1 of this coil through r.f. by-pass condensers  $C_0$  and  $C_4$  and the grounded chassis.

Now let us see how oscillator control tube  $VT_1$  can act as an inductance in shunt with tank circuit  $L-C$ . First of all, r.f. tank voltage  $e$  in Fig. 6 must be considered as the r.f. voltage source acting upon the oscillator control circuit. The two r.f. signal paths connected in parallel across r.f. voltage source  $e$  are the plate-cathode path of oscillator control tube  $VT_1$  and path  $R_1-C_1-C_2-C_4$ .

At radio frequencies, path  $R_1-C_1-C_2-C_4$  is essentially resistive (the reactances of all three condensers are low with respect to the resistance of  $R_1$ ), and hence the r.f. current flowing over this path is *in phase with its r.f. source voltage  $e$* . This r.f. current develops across condenser  $C_2$  an r.f. voltage which *lags* the r.f. current, and hence lags r.f. voltage  $e$ . (The a.c. voltage across a condenser always lags the condenser current.)

The r.f. voltage across  $C_2$  acts on the control grid of  $VT_1$ , making the tube pass an r.f. plate current which is *in phase with the applied r.f. grid voltage*. The r.f. plate current drawn from  $L$  by the oscillator control tube *therefore lags the r.f. voltage across  $L$* . This is exactly the same phase relationship as for an inductance load across  $L$ ; the oscillator control tube thus acts like an additional inductance shunting the tank circuit, and serves to *increase* the frequency of the r.f. oscillator.\*

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\* When two inductances are in parallel, the combined inductance is less than that of the smaller inductance. Lowering the total tank circuit inductance raises the oscillator frequency. If one of the parallel inductances is reduced in value, the oscillator frequency will increase; if one of the parallel inductances is increased in value, the oscillator frequency will decrease.

The a.f. modulation voltage which is applied to the control grid of  $VT_1$  through  $R_2$  varies the transconductance of the tube, and hence makes the a.c. plate current vary at an audio rate. Consequently, *the inductance of this tube also varies at an audio rate*. This in turn makes the frequency of the r.f. oscillator swing above and below its resting value at an a.f. rate, giving frequency modulation of the r.f. carrier without appreciable variation in the r.f. amplitude. When the a.f. signal is removed or drops to zero, the r.f. oscillator returns to its resting frequency, which is determined by the size of inductance  $L$ , the normal inductance of the oscillator control circuit, and the tank circuit capacity.

Resistor  $R_2$  prevents the a.f. source from shorting the input of the inductance tube; it is really an isolating resistor.

By adjusting the initial C bias on the oscillator control tube (varying the ohmic value of  $R_C$ ) and by monitoring properly the a.f. voltage fed into the oscillator control tube, the maximum deviation in frequency can be made any desired amount. The circuit of Fig. 6 is therefore a suitable signal source for an f.m. system which is to be employed in transmitting intelligence. The f.m. signal would be taken either from terminal 2 or from a link which is coupled to tank circuit inductance  $L$ .

*The F.M. Receiver.* Up to the point at which frequency-modulated signals are converted into amplitude-modulated signals, an f.m. receiver uses essentially the same circuits as a corresponding a.m. receiver. The a.f. amplifier and loudspeaker are likewise essentially the same for both systems. The important new action in an f.m. receiver is the conversion of the f.m. signal into the desired audio signal, so let us consider now the basic

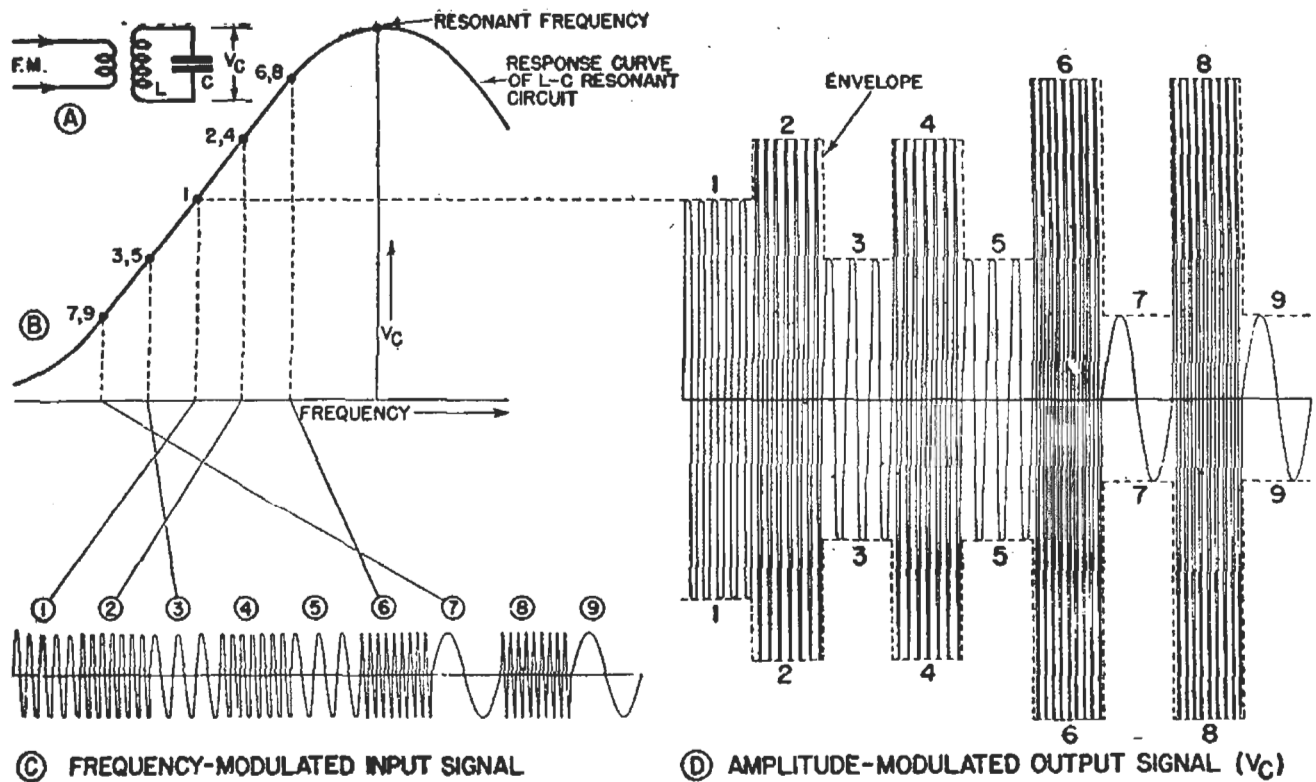


FIG. 7. These diagrams illustrate how a simple resonant circuit can convert an f.m. signal into an a.m. signal.

principles involved in this conversion.

If a frequency-modulated r.f. signal is introduced into an  $L-C$  resonant circuit which is tuned slightly above the highest deviation frequency, the r.f. voltage developed across the resonant circuit will vary with the frequency of the induced f.m. signal. An ordinary resonant circuit like the  $L-C$  circuit in Fig. 7A is thus a simple means for converting an f.m. signal into an a.m. signal.

A portion of the resonant response curve for this  $L-C$  circuit is shown in Fig. 7B. By applying to this curve in the proper manner the f.m. signal shown at the bottom of Fig. 2 (having a square-wave audio modulation), we can get a graphical picture of how the amplitude of r.f. output voltage  $V_C$  varies with the frequency of the incoming f.m. signal.

Let us assume that when the f.m. transmitter is at its resting frequency (time interval 1 in Figs. 2 and 7C), the operating point is at 1 in Fig. 7B. The vertical distance from 1 down to the horizontal reference line then de-

termines the amplitude of r.f. voltage  $V_C$  across the resonant circuit, so we show the r.f. output voltage for time interval 1 as an r.f. signal having this same amplitude and the same resting frequency value, as at 1 in Fig. 7D.

When the frequency of the f.m. signal increases to the value for time interval 2 in Fig. 7C, we move up to point 2 on the response curve, and thereby secure the output signal at 2 in Fig. 7D. Likewise, the input signal frequencies for time intervals 3, 4, 5, 6, 7, 8 and 9 in Fig. 7C give the output signal amplitudes shown at 3, 4, 5, 6, 7, 8 and 9 respectively in Fig. 7D.

Note that the signal frequency is unchanged by the resonant circuit, but the amplitude of the output signal varies in proportion to the frequency deviations of the input f.m. signal. A dash-dash line drawn through the positive peaks of the r.f. output voltage in Fig. 7D gives the original square-wave audio modulation of Fig. 2, showing that we now have an amplitude-modulated r.f. signal which can be demodulated with a conventional am-

plitude detector circuit. The fact that our output signal also varies in frequency does not matter, for the amplitude detector circuit removes all r.f. components.

A complete circuit capable of converting an f.m. signal into an audio signal is shown in Fig. 8; this circuit can appropriately be called an *f.m. detector*. The final i.f. amplifier stage is also shown, and employs a 6SK7 pentode, with a double-tuned i.f. transformer ( $C_3-L_3-L_2-C_2$ ) in its input circuit. Resistor  $R_L$  across the primary of this i.f. transformer provides the loading required to flatten the response over the 150-kc. range through which

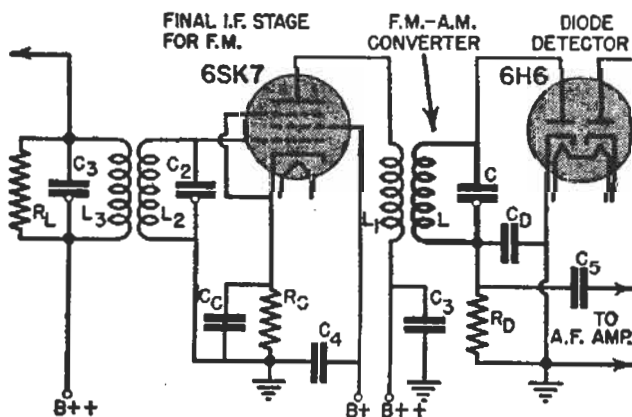


FIG. 8. This diagram shows how a resonant circuit can be used to convert the f.m. output signal of the i.f. amplifier into an a.m. signal which can be demodulated by a conventional diode detector.

the f.m. signal is varying during loud audio loudness levels.

Plate current of the final i.f. tube flows through coil  $L_1$ , inducing an f.m. signal voltage in coil  $L$  of the  $L-C$  resonant circuit in which frequency-to-amplitude conversion takes place. The r.f. voltage across this  $L-C$  circuit therefore varies in amplitude in accordance with variations in the original audio signal amplitude. This amplitude-modulated r.f. voltage is demodulated by one section of the 6H6 tube, and the resulting a.f. output voltage appears across diode load resistor  $R_D$ . Condenser  $C_D$  removes r.f. variations across  $R_D$ , so that only the desired a.f. voltage is fed into the a.f. amplifier.

### Additional Requirements in an F.M. System

The f.m. transmitter and receiver arrangements just described represent the absolute minimum required in an f.m. system. Transmitters must have some type of crystal control, so as to maintain the assigned resting frequency value within the limits prescribed by law, and receivers must also have additional features which insure high fidelity and operating stability, reduce noise and simplify tuning.

The remainder of this lesson will be devoted chiefly to practical f.m. receiver circuits, but basic principles of highly stable modern f.m. transmitters will be covered briefly at the end of this lesson text to make your f.m. training more complete.

Resonant circuits have proved unsatisfactory for f.m.-a.m. separating purposes, chiefly because the slopes of their response curves are not sufficiently linear for high-fidelity purposes. Instead, we find in modern f.m. receivers a variation of the discriminator circuit used in a.f.c. circuits. This discriminator circuit is simple to adjust and gives linear operation. It uses a push-pull circuit to convert frequency variation to amplitude variations at the i.f. frequency and then detects this signal.

In a practical f.m. receiver, an additional section called the limiter is required just ahead of the discriminator to restore constant amplitude, so that all amplitude rises due to noise or station interference will be removed.

The limiter is a vacuum tube amplifier which is so designed that all desired signals will cause plate current saturation. This action of the limiter stage also levels out the response of the r.f. system.

Since the limiter in an f.m. receiver delivers the same amplitude for both

weak and strong incoming signals, the loudspeaker volume level does not change when tuning from one f.m. station to another. This means that it is unnecessary to use an a.v.c. system in an f.m. receiver to prevent blasting or counteract fading. Most f.m. receivers do have a.v.c., but this is primarily to prevent extremely strong input signals from driving the grids of r.f. or i.f. amplifier tubes positive and thereby causing blocking or interference. If grid current were allowed to flow during positive grid voltage values, the

resulting distortion of the incoming f.m. signal would produce harmonics which could beat with oscillator harmonics and other station signals to produce interfering signals at the i.f. value for f.m.

Another requirement in a modern f.m. receiver is a tuning indicator (either a meter or a cathode ray tuning indicator tube). F.M. resonant circuits are so broad due to loading that the average person would have difficulty in tuning in an f.m. station by ear alone.

## The F.M. Receiver

*Types of F.M. Receivers.* Radio receivers which are capable of receiving f.m. signals can be divided into three groups, as follows:

1. Sets designed exclusively for f.m. reception. These can be table models or consoles, for listeners who already have good a.m. receivers.

2. Combination f.m.-a.m. receivers, providing all-wave a.m. reception along with f.m. reception. These will usually be large console sets and may also have automatic record changers, television, facsimile or sound-recording features.

3. F.M. converters, which are simply f.m. receivers without audio amplifiers or loudspeakers. They provide f.m. reception when connected

to the a.f. input terminals of an ordinary a.m. receiver.

All f.m. receivers should have high sensitivity in order to operate the limiter at saturation for the widest frequency swings of the weakest signal to be received, and hence superheterodyne circuits are invariably employed.

Typical circuits used in combination f.m. and a.m. receivers will now be analyzed.

### Combination F.M.-A.M. Receiver

A block diagram of a typical combination a.m. and f.m. superheterodyne receiver is given in Fig. 9. The circuits now to be considered will be

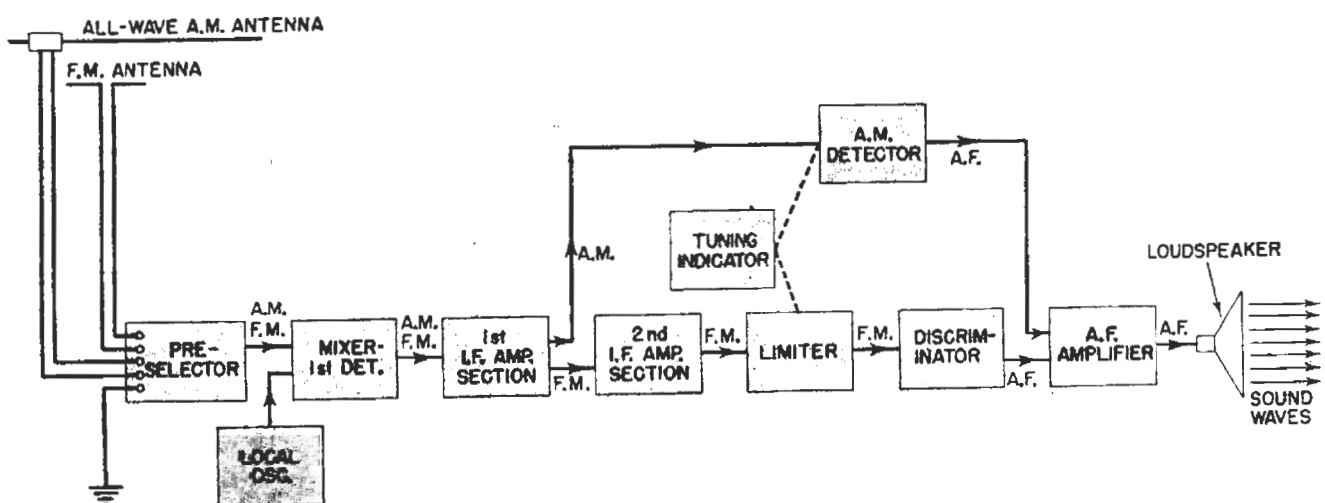


FIG. 9. Block diagram of a typical combination f.m. and a.m. receiver.

those employed in this particular receiver arrangement.

Note first of all that there are two antennas, one for regular all-wave a.m. signal pick-up, and the other for ultra-high-frequency f.m. signal pick-up. The band-selecting switch automatically connects the correct antenna to the receiver. (Fairly satisfactory results can be obtained with a single combination antenna on some sets, as you will later learn.)

For amplitude-modulated signals, the local oscillator will operate about 455 kc. above the incoming a.m. signal frequency, while for frequency-modulated signals, it will operate at 4.3 mc. below the incoming f.m. signal frequency.\* The i.f. signal produced by the mixer-first detector (about 455 kc. for a.m., or 4.3 mc. for f.m.) passes through the first i.f. amplifier section to a switching circuit.

For f.m. reception, the signals are fed into a second i.f. amplifier section, which raises the signal level the required amount for the limiter stage. The constant-amplitude output of the limiter is fed into the discriminator, and the resulting a.f. output signal is fed directly into the input of the a.f. amplifier.

For a.m. reception, the output of the first i.f. amplifier section is fed directly into a conventional a.m. detector, and its a.f. output is likewise fed into the input of the a.f. amplifier. The tuning indicator, if used, would

\* Although 10.7 mc. is the recommended and widely used i.f. value for f.m. receivers at the present time, you may also encounter other values. Some f.m. receivers are capable of operating not only on the 88-108-mc. f.m. band, but on the older 42-50-mc. band as well. This low-frequency band is no longer in use. These sets particularly may have a different value of i.f., generally lower in frequency. The higher value now in use minimizes chances for interference between the various signals which are picked up by the receiving antenna or produced in the receiver by superheterodyne action.

be connected through the band-changing switch to the a.v.c. source for either the f.m. or a.m. channel. A common power pack serves all sections.

### F.M. Receiving Antennas

The antenna for f.m. reception will ordinarily be a doublet of the correct length to give half-wave operation at an average frequency in the f.m. band. Since commercial f.m. stations are assigned to frequencies between 88 and 108 mc., a doublet which resonates at about 98 mc. will ordinarily be used. With proper design, its response will normally be broad enough to give satisfactory coverage of the entire f.m. band.

A frequency of 98 megacycles corresponds to a wavelength of 3.06 meters (wavelength in meters is equal to 300 divided by frequency in megacycles). A half-wave f.m. doublet antenna will therefore have a total length of about 1.5 meters; multiplying this value by 3.28 gives about 5 feet as the total length for an average f.m. doublet receiving antenna.

The simplest possible f.m. doublet is that shown in Fig. 10A, in which the transmission line is connected to the center of the 5-foot long doublet. Since a half-wave doublet has a resistance of about 72 ohms at its center, a transmission cable having a surge impedance of approximately this same value would ordinarily be used. Slight mismatches can be tolerated, however; in fact, a cable with a surge impedance of about 100 ohms will make the antenna response characteristic more uniform over the entire f.m. band. The transmission cable is connected to the receiver through a matching transformer.

At very-high frequencies, line-of-sight paths give the best transmission of radio waves, but, fortunately, reliable f.m. reception is possible con-



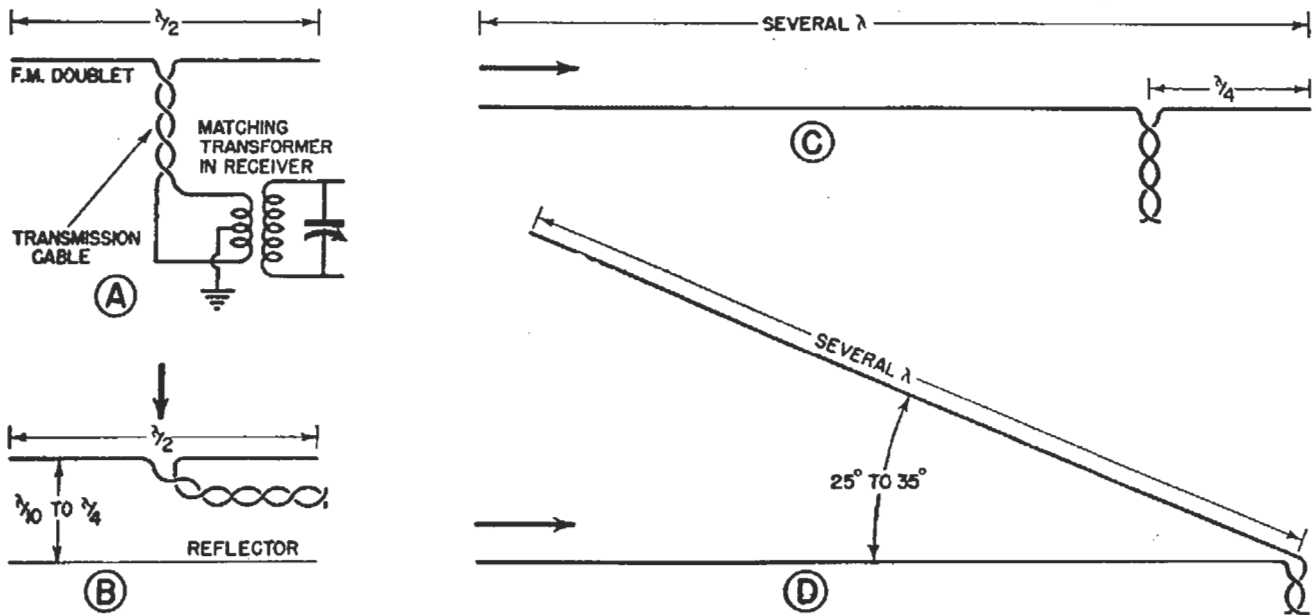


FIG. 10. Types of f.m. receiving antennas. The heavy arrow in each case indicates the direction (along the surface of the earth) from which signals are best received. The symbol  $\lambda$  represents "wavelength."

siderably beyond the line-of-sight range from a transmitter. (With a line-of-sight path, it should be possible to see the transmitting antenna with a telescope on a clear day, from the position of the receiving antenna.)

A compact very-high frequency antenna built into the console cabinet of an f.m. receiver will often give adequate signal pick-up of stations up to about ten miles away. Even a quarter-wave vertical antenna can sometimes be used when signals are strong.

In locations considerably beyond the line-of-sight limit, the position of the receiving antenna becomes important. The half-wave horizontal f.m. doublet should then be as high as possible, and broadside (at right angles) to the path over which f.m. signals are arriving.

In remote locations where all of the desired f.m. signals are coming from essentially the same direction, such as from a single large city, a reflector like that shown in Fig. 10B can be used with a conventional doublet to increase the signal pick-up from the desired direction. The doublet is always placed between the reflector and the transmitter.

The antenna arrangements shown

in Figs. 10C and 10D give even better signal pick-up at remote locations. In Fig. 10C, one section of the doublet has the usual length of  $\lambda/4$  (one-quarter wavelength), while the other section is several wavelengths long. The long section is aimed directly at or slightly above the transmitter.

In the V doublet arrangement of Fig. 10D, both horizontal sections are equal, several wavelengths long, and aimed in the general direction of the transmitter.

Before mounting an f.m. doublet permanently, it is always a good idea to move the antenna toward or away from the transmitter a distance of about 5 feet, and tilt it at various angles while an assistant is checking signal strength at the receiver by watching the tuning indicator or using a meter, to find the position which gives maximum signal strength. Cancellation due to reflected waves arriving over different paths will then be a minimum.

### The Preselector

A preselector having at least one r.f. amplifier stage is just as desirable in an f.m. superheterodyne receiver as in an a.m. set, for r.f. amplification

ahead of the mixer-first detector in an f.m. receiver provides much needed extra sensitivity and minimizes converter noise. Just as with a.m. receivers, the higher-priced f.m. receivers will usually have an r.f. stage in the preselector. Less expensive f.m. sets will have a simple antenna transformer with tuned secondary, feeding directly into the mixer-first detector.

The circuit of a typical r.f. amplifier stage for a combination f.m.-a.m. receiver is shown in Fig. 11A. To simplify the diagram, only one all-wave a.m. range is shown; additional contacts on the three sections of the band-changing switch and additional input and output r.f. transformers would be provided for other a.m. ranges.

When the three switches ( $S_1$ ,  $S_2$  and  $S_3$ ) are set to position 1, as shown in Fig. 11A, the entire preselector is connected for f.m. reception. The half-wave f.m. doublet feeds its f.m. signal into the grid circuit of the 6SK7 r.f. amplifier tube through f.m. antenna matching transformer  $L_1-L_2$ , with secondary winding  $L_2$  being tuned to the incoming f.m. signal frequency by gang condenser section  $C_1$ . The amplified r.f. plate current of the 6SK7 tube flows through winding  $L_3$  of the f.m. input transformer for the mixer, inducing in secondary winding  $L_4$  a corresponding r.f. voltage. Gang condenser section  $C_2$  tunes  $L_4$  to resonance at the desired station frequency, and the resulting f.m. signal voltage across this resonant circuit is applied to the grid and cathode of the mixer-first detector tube. Automatic C bias for the 6SK7 tube is provided by  $R_2$  and  $C_9$ .

Setting the three switches in the preselector circuit of Fig. 11A to position 2 automatically connects the circuit for a.m. reception on one band. Switch  $S_1$  disconnects the secondary

of the f.m. matching transformer from the r.f. amplifier tube grid, and connects the grid instead to the secondary of the a.m. antenna matching transformer. The all-wave a.m. doublet feeds its a.m. signal to the r.f. amplifier tube through a.m. matching transformer  $L_5-L_6$ , with secondary winding  $L_6$  being tuned to resonance by gang condenser sections  $C_3$  and  $C_1$  in parallel. (The conventional condenser section for a.m. is split into two sections,  $C_3$  and  $C_1$ , so that one low-capacity section will be available for f.m. to give a favorable L-to-C ratio.) The r.f. output of the 6SK7 tube is fed to the mixer through a.m. input transformer  $L_7-L_8$ , with secondary winding  $L_8$  being tuned by  $C_4$  and  $C_2$  in parallel.

When the switches are at position 2 for a.m. reception, a.v.c. voltage is applied to the r.f. and mixer tubes through  $L_6$  and  $L_8$  respectively. No a.v.c. is used on these two tubes during f.m. reception in this circuit, but engineers are now recommending that a.v.c. be provided for the r.f. amplifier tube of an f.m. receiver to prevent overloading of any tubes by extremely strong f.m. signals. The a.v.c. voltage for f.m. would be supplied by the *limiter* section.

A variation of the preselector circuit of Fig. 11A which makes possible the use of one doublet antenna for both f.m. and a.m. reception is shown in Fig. 11B. The remainder of the preselector circuit is not shown, since it is identical with that in Fig. 11A. The connections are such that doublet antenna action is secured for f.m. reception, but the antenna is automatically converted into a plain vertical antenna for a.m. reception. Resonant circuit L-C is tuned to the mid-frequency of the f.m. band (about 98 mc.); since it is a series resonant circuit, it has a low impedance at or near resonance, and has the

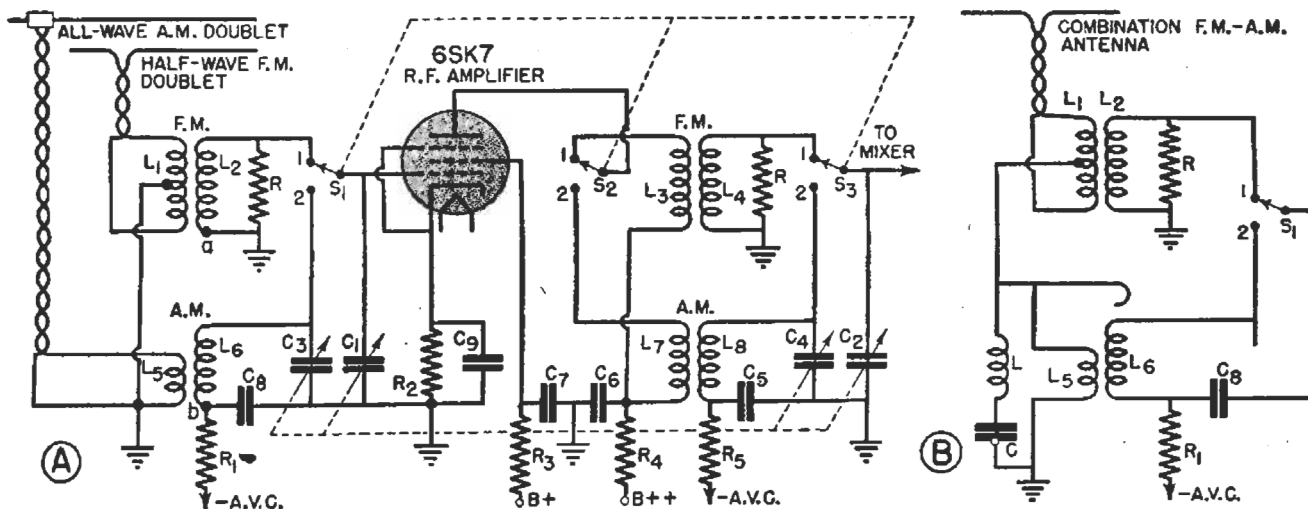


FIG. 11. Typical preselector circuit arrangements for a combination f.m.-a.m. receiver. If the r.f. amplifier tube is to have a.v.c., terminal *a* of f.m. coil  $L_2$  is connected to terminal *b* of a.m. coil  $L_6$  instead of to ground. A switch at the two a.v.c. sources (shown in Fig. 20) insures that the correct a.v.c. voltage (for f.m. or for a.m.) will be fed to this tube.

effect of grounding the center tap of winding  $L_1$ . Under this condition, the antenna acts as a very-high-frequency doublet for the f.m. band.

For a.m. bands, usually below 20 megacycles, the series  $L$ - $C$  circuit is considerably off resonance, and hence has a high capacitive reactance. Winding  $L_1$  is therefore ungrounded, and signal currents picked up by the two sections of the doublet flow through the two sections of  $L_1$  to the center tap, then through winding  $L_5$  to ground ( $L_5$  has much lower reactance than the  $L$ - $C$  circuit). The a.m. signal current through  $L_5$  induces a corresponding r.f. voltage in  $L_6$ , and this is fed through the r.f. amplifier tube to the mixer in exactly the same manner as for the circuit of Fig. 11A.

Observe that a resistor ( $R$ ) is shunted across each f.m. resonant circuit in Figs. 11A and 11B. These resistors serve to load the resonant circuit, providing broad resonance which gives essentially uniform response at all frequencies within the deviation range of an incoming f.m. signal. Ordinarily, resistor  $R$  has an ohmic value somewhere between 10,000 and 20,000 ohms, with the higher value being preferred in order that the preselector will have good selectivity. When no r.f. amplifier stage is

used, a band-passed preselector circuit ahead of the frequency converter is highly desirable, to broaden the response and at the same time improve selectivity.

### The Frequency Converter

In any superheterodyne receiver, the local oscillator and the mixer-first detector constitute a section commonly called the *frequency converter*. The local oscillator feeds into the mixer-first detector an unmodulated r.f. signal which beats with the incoming modulated signal to produce the correct i.f. signal.

A number of different circuit arrangements can be employed in the frequency converter of a combination f.m.-a.m. receiver. Some designers will use a pentagrid converter tube, some will use a combination triode-pentode tube, and some will use separate tubes for the oscillator and mixer-first detector.

It is customary to operate the local oscillator *above* the incoming signal frequency for f.m. reception, to secure greater stability and prevent local oscillator signals from creating interference in the television band below 88 mc. On a.m. bands also the oscillator is usually operated *above* the incoming signal frequency.

Switching sections must be provided on the band-changing switch of a combination f.m.-a.m. receiver to insert the correct inductance and variable capacity in the oscillator tuning circuit of the frequency converter. The oscillator would thus be tuned 10.7 mc. (or whatever other i.f. value is used for f.m.) *above* the incoming signal frequency for f.m. reception, and about 455 kc. *above* the incoming signal frequency for a.m. reception.

An example of a conventional frequency converter circuit employing a type 6K8 combination triode-pentode tube is given in Fig. 12. This circuit could be connected directly to the

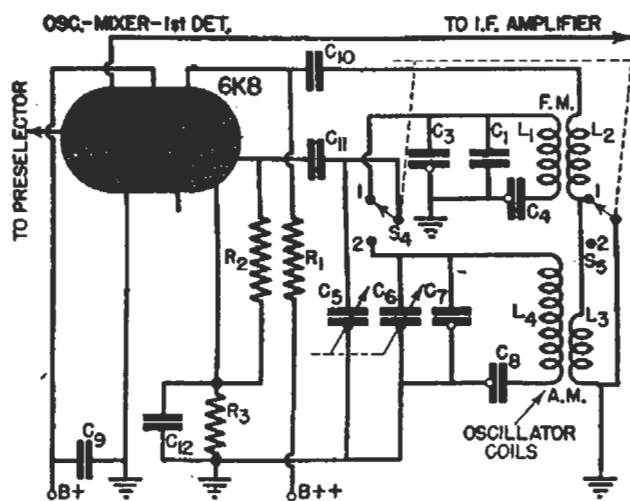


FIG. 12. Typical frequency converter circuit for a combination f.m.-a.m. receiver.

preselector circuit of Fig. 11, with all tuning condenser sections being controlled by one tuning knob and all switches being controlled by the band-changing knob.

When switches  $S_4$  and  $S_5$  in Fig. 12 are at position 1 for f.m. reception, oscillator tank circuit inductance  $L_1$  is tuned 10.7 mc. above the incoming f.m. signal frequency by tuning condenser section  $C_5$ . The high-frequency trimmer is  $C_3$ , and the low-frequency padder is  $C_4$ , while  $C_1$  serves as a temperature-compensating condenser. (Condenser  $C_1$  changes its capacity in a definite manner with temperature, thereby minimizing oscillator frequency drift by com-

pensating for other capacity changes which occur as the set comes up to normal operating temperature after being turned on.)

Feed-back from the oscillator plate circuit is provided by inductance  $L_2$ . Automatic C bias for the oscillator is provided by the flow of grid current through grid resistor  $R_2$  and grid condenser  $C_{11}$ . D.C. blocking condenser  $C_{10}$  prevents the plate supply voltage of the oscillator triode from sending direct current through oscillator feed-back coil  $L_2$  to ground.

All leads in the f.m. oscillator section of the circuit should be as short as possible, so that resonance effects will be localized to the desired  $L_1$ - $C_5$  tuning circuit. All grounded elements in a section should be connected to a common point on the chassis. At very-high frequencies, long leads have appreciable inductance and capacity to the chassis, and hence might form additional resonant circuits which create trouble.

When switches  $S_4$  and  $S_5$  in Fig. 12 are at position 2 for reception on one of the a.m. bands, the oscillator tank circuit inductance is  $L_4$ . The tank circuit is tuned 455 kc. above the incoming a.m. signal frequency by tuning condenser sections  $C_5$  and  $C_6$  in parallel.  $C_7$  serves as the high-frequency trimmer, and  $C_8$  serves as the low-frequency padder. Feed-back energy from the plate circuit is now supplied by  $L_3$ , for removal of the short across  $L_3$  by  $S_5$  allows r.f. plate current to flow through both  $L_2$  and  $L_3$  to ground.

### The I.F. Amplifier

We cannot use a standard 455-kc. i.f. value for both f.m. and a.m. reception. The gain of a 455-kc. i.f. amplifier drops entirely too much when it is band-passed sufficiently to pass a 150-kc. wide f.m. signal. Fur-

thermore, the selectivity of this band-passed 455-kc. i.f. amplifier would be way too poor for standard a.m. reception. A 10.7-mc. i.f. amplifier is satisfactory for f.m., but its selectivity is likewise too poor for a.m. reception.

To meet these conflicting requirements in a combination f.m.-a.m. receiver, separate i.f. transformers can be used for f.m. and a.m., so that a.m. signals will pass through standard 455-kc. i.f. transformers, and f.m. signals will be served by 10.7-mc. i.f. transformers. With the exception of the first i.f. transformer primary, the connections can be such that signals will take the correct transformers in traveling through the i.f. amplifier. It is even possible to mount both the f.m. and a.m. transformers for a given stage in the same shield can.

When corresponding windings of the i.f. transformers for f.m. and a.m. are connected in series as shown in Fig. 13, signals will automatically take the correct path. Thus, when a 455-kc. a.m. signal is coming through, the transfer of signals will be accomplished by the 455-kc. a.m. transformer  $L_3-C_3-L_4-C_4$ , because  $L_1$  will be essentially a short-circuit path. When 10.7-mc. f.m. signals are coming through, signal transfer will be through f.m. transformer  $L_1-C_1-L_2-C_2$ , because  $C_3$  will be essentially a short-circuit path. In each case, the i.f. transformer not in use will be so far off resonance that its effect will be negligible.

*I.F. Switching.* During broadcast band a.m. reception, the local oscillator produces a signal somewhere between about 1000 and 2000 kc. (1 and 2 mc.) when the local i.f. value for a.m. is 455 kc. The sixth harmonic of the oscillator will then cover 6 to 12 mc., and there will be one tuning dial position at which this oscillator signal coincides with the 10.7-mc. i.f. value for f.m.

With the series arrangement of 455-kc. and 10.7-mc. i.f. windings, this oscillator harmonic (if strong enough) could get through the first i.f. transformer for f.m., and drive the self-bias so far beyond cut-off (due to increased plate current through the cathode resistor) that the desired 455-kc. i.f. signal would be blocked. The result would be a "dead spot" in the broadcast band at the setting which made this oscillator harmonic equal to 10.7 mc.

Other harmonics of the local oscillator can likewise create dead spots at different broadcast band dial settings,

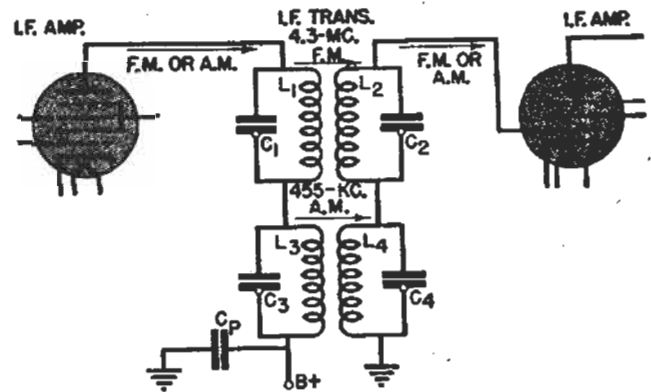


FIG. 13. I.F. transformer arrangement suitable for a combination f.m.-a.m. receiver. Separate i.f. transformers with corresponding windings connected in series are used for f.m. and a.m. No switches are employed, for desired signals automatically take the correct path due to circuit impedance values.

and hence it is necessary either to disconnect or short out the 10.7-mc. first i.f. transformer during broadcast band reception. When a switch is provided at the i.f. amplifier input for this purpose, only the desired i.f. signal passes through the first i.f. tube to the other i.f. transformers, and hence no additional i.f. switches are needed.

An example of a complete i.f. amplifier section which has provisions for disconnecting the unused first i.f. transformer primary is shown in Fig. 14. Section  $S_1$  of the band-changing switch connects to the converter the correct i.f. transformer primary winding ( $L_1$  or  $L_7$ ) for the particular signals being received. All other pairs

of i.f. windings are still connected in series just as in Fig. 13, however.

The i.f. input switch can just as well be between the grid of the first i.f. amplifier tube and the secondary windings of the first i.f. transformers, or can be used to short out the primary or secondary winding of the unused first i.f. transformer.

Note that the i.f. transformers for a.m. have fixed condensers, with tuning being accomplished with adjustable iron-core inductances. Pulverized iron cores are not effective at the 10.7-mc. i.f. value for f.m., but at a high frequency like this a low-resistance non-ferrous core (such as copper, aluminum or brass) can be used instead to vary the inductance. Non-ferrous cores of this type are inserted from the grounded end of the i.f. coil; they vary the inductance due to eddy currents induced in the core, and also add capacity to the tuning circuit. Non-ferrous cores can also be used for r.f. coils in f.m. circuits.

Sometimes the final i.f. stage is used only during f.m. reception. A simple method for eliminating this stage during a.m. reception involves grounding point *a* in Fig. 14 permanently, and connecting secondary  $L_{10}-C_{10}$  of the

second 455-kc. i.f. transformer directly to the a.m. second detector. Transformer  $L_{11}-C_{11}-L_{12}-C_{12}$  would be omitted.

Both i.f. amplifier tubes in Fig. 14 have resistors in their cathode leads to provide automatic C bias for both f.m. and a.m. reception. During a.m. reception, however, the diode a.m. detector supplies an a.v.c. voltage to the first i.f. amplifier tube, as well as to preceding tubes. (During f.m. reception, the a.v.c. voltage is provided by the limiter stage.)

*The Limiter.* 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 15. 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

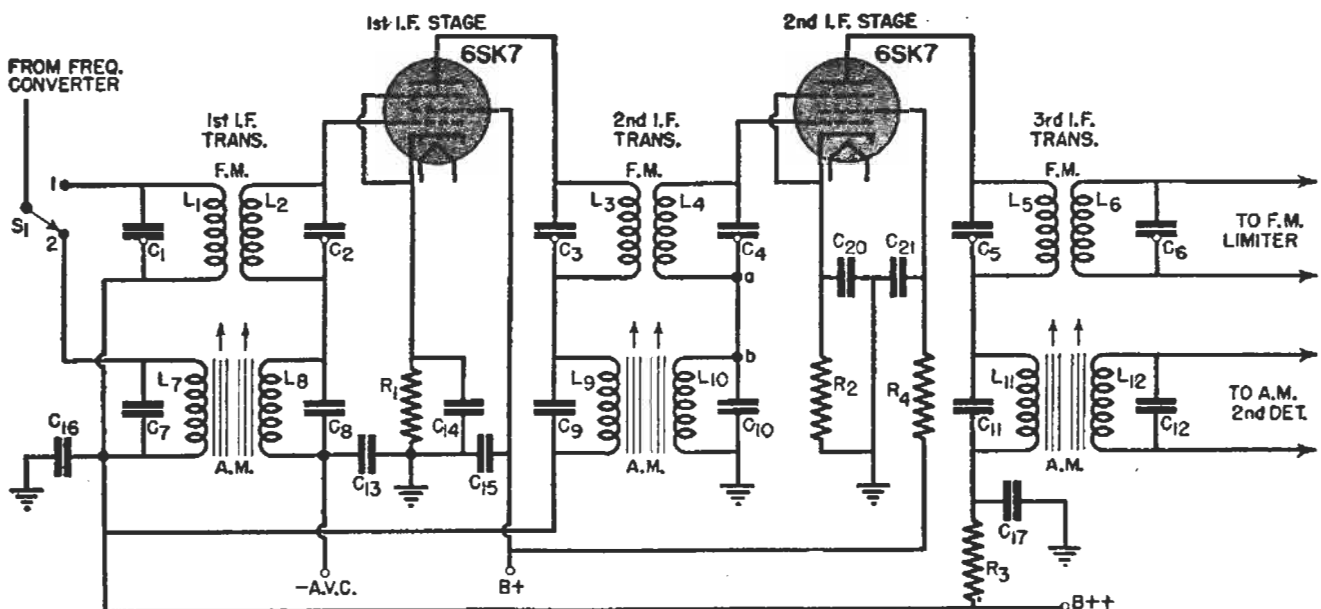


FIG. 14. Circuit diagram of a complete i.f. amplifier for a combination f.m.-a.m. receiver. Note the use of a switch for the primary windings of the first i.f. transformers. The groups of three parallel lines with arrows on a.m. transformers represent adjustable pulverized iron cores; these are adjusted in the same way as ordinary i.f. trimmers during i.f. alignment.

operating conditions.

When a 6SJ7 pentode is operated at normal voltages for amplifying purposes, 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. 16. This curve is essentially linear up to the highest plate current which can safely be passed by the

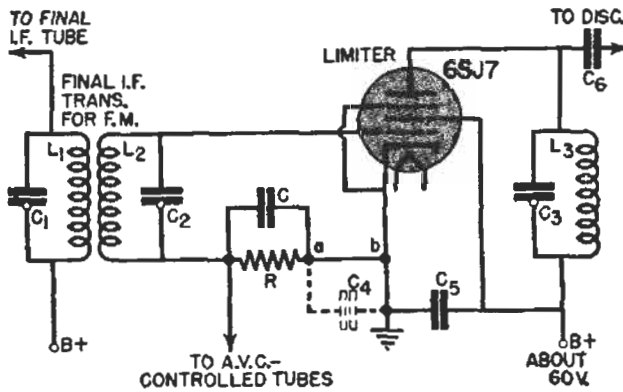


FIG. 15. Simplified one-tube limiter circuit for an f.m. receiver. Condenser  $C_4$  is provided for alignment purposes, as will be explained later.

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 which 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 whereby saturation begins at a fairly low positive grid voltage value, as indicated by curve 2 in Fig. 16. This curve is for the tube alone, and would be obtained from the circuit of Fig. 15 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. 16, 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.

Now let us return to the basic limiter circuit in Fig. 15, 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 which 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 will be the same for negative grid voltages as it was without  $R$ . For half cycles which make the grid swing positive, however, the flow of grid current through  $R$  will develop across it a voltage drop which opposes the applied positive grid voltage, and which consequently serves to provide a negative bias which

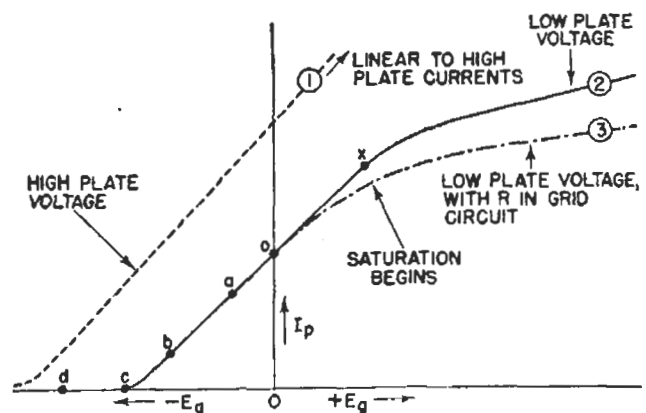


FIG. 16. Static characteristic curves of a vacuum tube for high (1) and low (2) plate voltages, and dynamic characteristic curve (3) for a limiter circuit using a low plate voltage and a resistor  $R$  (but no condenser  $C$ ) in the grid circuit. The plate load resistance for the static curves is assumed to be equal to the resistance of the plate load resonant circuit at resonance during dynamic operation.

is at each instant proportional to the positive applied voltage. The characteristic curve for this condition will

be like curve 3 in Fig. 16, 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. 15 would be shunted by condenser  $C$ , with the time constant of  $R$  and  $C$  being equal to the time of several r.f. cycles. Un-

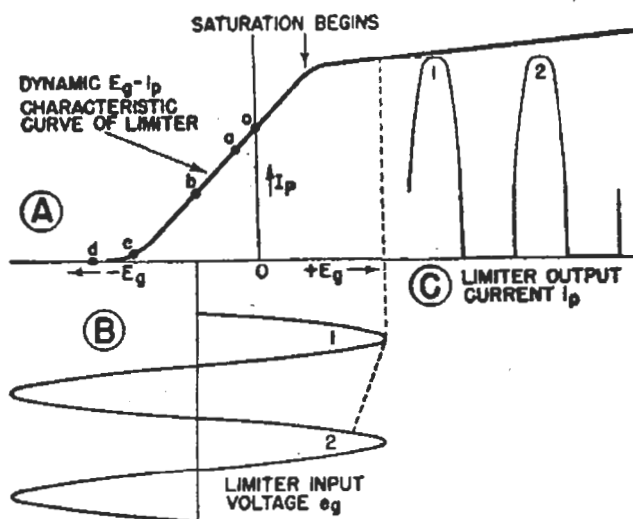


FIG. 17. These diagrams will help you to understand the action of the basic one-tube limiter circuit in Fig. 16.

der this condition, the  $R$ - $C$  circuit assumes a definite negative bias voltage which is maintained relatively constant over several cycles of the r.f. input voltage, and which hence exists for negative half cycles as well as for positive half cycles.

An automatic bias of this nature gives a characteristic curve which will be somewhere between curves 2 and 3 in Fig. 16; with  $R$  and  $C$  large in electrical size so as to give a long time constant, the dynamic characteristic curve will approach static curve 2 because we are approaching a fixed-bias condition, but with low electrical values for  $R$  and  $C$ , the dynamic char-

acteristic will approach curve 3 because the bias will now vary almost instantaneously with input signal strength.

A typical dynamic characteristic curve for a single-tube limiter stage is shown in Fig. 17. With this characteristic, the bias voltage developed across  $R$  and  $C$  will be proportional to the average signal strength over several r.f. cycles, and consequently the operating point might 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  $e_g$  which places the operating point at point  $b$ , the wave form of the resulting plate current  $i_p$  will be as shown at  $C$  in Fig. 17. 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. 16 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.

With the dynamic characteristic curve of Fig. 17 for a limiter stage, two things happen when the strength of the input signal increases: 1. The increased negative bias makes negative half cycles 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.

It is possible to design a limiter cir-



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. 15 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 realized, the over-all dynamic characteristic curve for the limiter will be like curve 1 in Fig. 18A, in which the r.f. output voltage of the limiter is plotted against the r.f. input

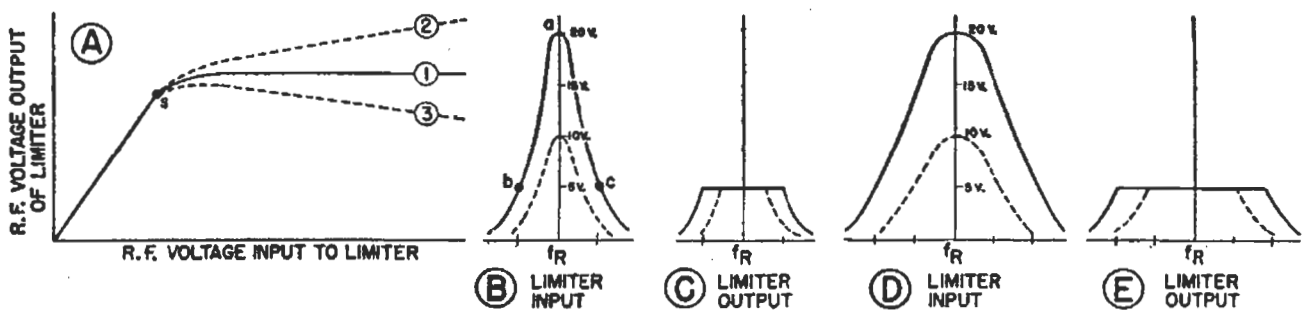


FIG. 18. These curves show how the limiter can correct the over-all response characteristic of the entire r.f. system (including the i.f. amplifier and the receiving antenna system) in an f.m. receiver.

voltage to the limiter. With this characteristic, the i.f. output voltage remains essentially constant regardless of input voltage, for all input voltage values which reach saturation (swing beyond point  $s$  in Fig. 18A).

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. 18A. 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. 18A. The values employed for  $R$  and  $C$  in the limiter circuit of Fig. 15

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. 18A. 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. 18B. 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 75 kc. from *b* to *a* to *c*). With the limiter characteristic of Fig. 18A, all signal peaks above 5 volts would be cut down to 5 volts by the limiter, and the solid-line curve in Fig. 18C 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 150 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

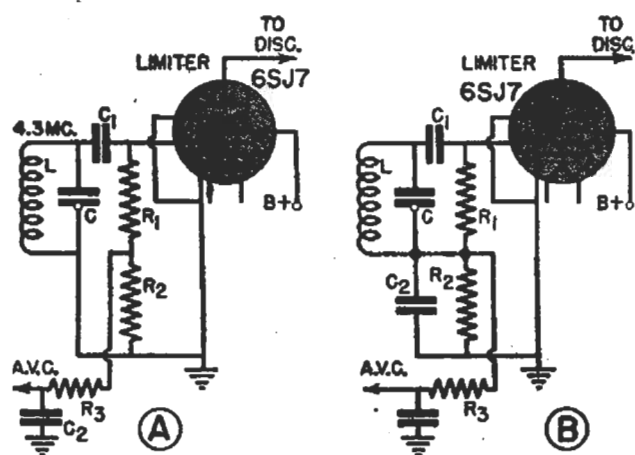


FIG. 19. Typical one-tube limiter circuits.

resting-frequency peak of only 10 volts at the output of the r.f. system. The dash-dash response curve in Fig. 18B would then give peak values for various deviations. When this weaker signal is fed into the limiter, the output response would be as shown by the dash-dash line in Fig. 18C. 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. 18D, a signal having a 20-volt peak would give a combined r.f.-limiter response corresponding to the solid line in Fig. 18E 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 dash-dash lines in Fig. 18D and 18E; even at this lower signal level, however, the combined r.f.-limiter response is still flat over a wide enough range to allow the entire frequency swing of 150 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 which is varying in frequency from 75 kc. below to 75 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. 18B, 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 would be far too great to give the desired constant output amplitude when

the limiter input drops to 5 volts. In order to keep the limiter output amplitude constant over the entire deviation range of 150 kc., the C bias should automatically reduce itself as the limiter input signal amplitude drops.

Noise surges which 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. 19A 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 which keeps r.f. components out of the a.v.c.-controlled tubes.

**Dual-Action Limiter Circuit.** In the limiter circuit arrangement shown in Fig. 19B, 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. 19A.  $R_2$  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, while  $R_1$  and  $C_1$  take care of changes in signal amplitude due to a peaked r.f. re-

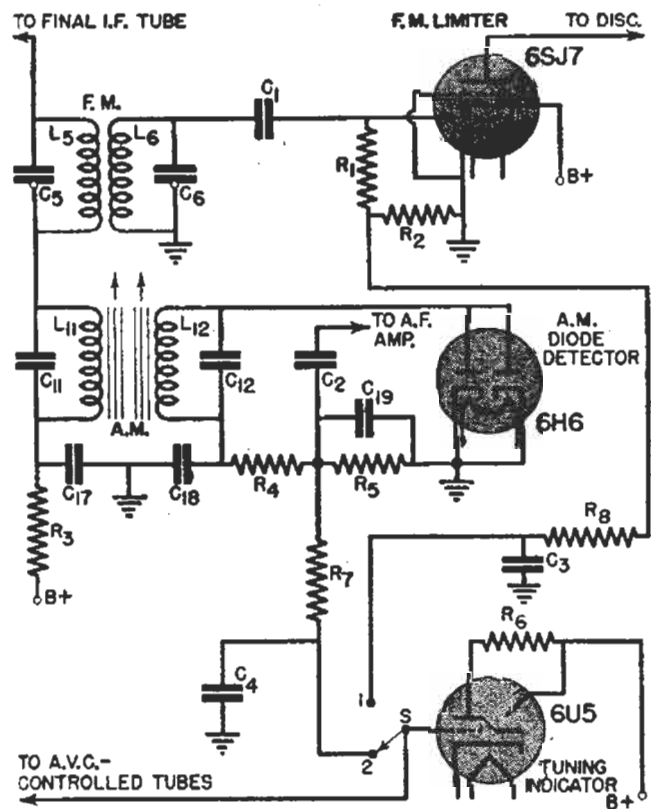


FIG. 20. One-tube limiter circuit with tuning indicator. I.F. amplifier connections to the conventional diode detector for a.m. reception are also shown.

sponse characteristic or to noise interference. This circuit arrangement thus provides independent control over two of the important factors which affect the design of a limiter circuit.

**Limiter with Tuning Indicator.** A more complete single-tube limiter circuit is shown in Fig. 20. This arrangement includes an electric eye for tuning purposes during both a.m. and f.m. reception. The output circuit of the i.f. amplifier is also shown, and is identical with that in Fig. 14. The

limiter input circuit is the same as that shown in Fig. 19A.

The a.v.c. voltage developed across  $R_2$  in Fig. 20 is fed to the tubes which are to be a.v.c.-controlled during f.m. reception through a.v.c. filter  $R_8-C_3$  and switch  $S$  (in position 1 for f.m. reception), and is also applied to the control grid of the type 6U5 cathode ray tuning indicator tube.

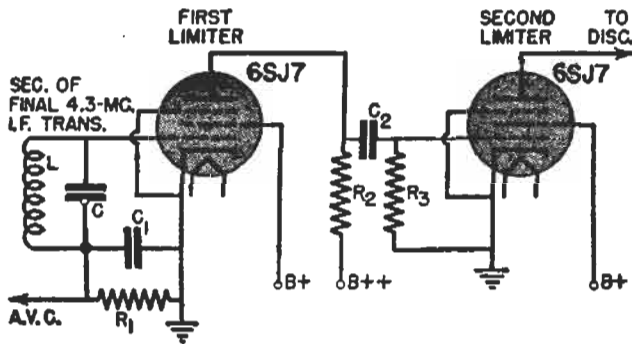


FIG. 21. Cascade limiter circuit employing two limiter tubes. The a.v.c. voltage is obtained from the first tube. The General Electric Model JFM-90 f.m. converter is one example of f.m. units employing this cascade limiter arrangement.

For a.m. reception, the output of the i.f. transformer for a.m. is fed into the diode detector (a 6H6 tube with both plates tied together), and the a.f. voltage component across  $R_5$  is fed to the a.f. amplifier through d.c. blocking condenser  $C_2$ . The d.c. voltage across  $R_5$  is fed to the a.v.c.-controlled tubes through a.v.c. filter  $R_7-C_4$  and switch  $S$  (in position 2 for a.m. reception), and is also applied to the tuning indicator tube.

*Cascade Limiter.* Improved limiter operation both for weak and strong signals, along with considerably higher gain, can be secured with two limiter tubes connected as shown in Fig. 21, 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. 19A, 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 tube, so that the i.m. signal which 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.

## The F.M. Detector

The f.m. detector or discriminator used in an f.m. receiver to convert f.m. signals into audio signals is much like the frequency discriminator circuit employed in the a.f.c. system of a conventional a.m. superheterodyne receiver. Technically speaking, the discriminator converts deviations in frequency from an assigned resting value into positive and negative d.c. voltages corresponding to the original audio signal.

Although the action of a.f.c. discriminator circuits has been covered elsewhere in the course in connection with a.m. receivers, the action will be reviewed briefly here in connection with the typical f.m. discriminator circuit given in Fig. 22.

Discriminator transformer  $T_1$  is of special design, with its primary winding  $L_2$  tuned to the 10.7-mc. i.f. value by  $C_2$ , and with a center-tapped secondary winding  $L_3$  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 resonant step-up produces across the two sections of  $L_3$  the voltages  $e_1$  and

$e_2$ , which are always equal in magnitude and are  $180^\circ$  out of phase with each other when considered with respect to center-tap  $p$ .

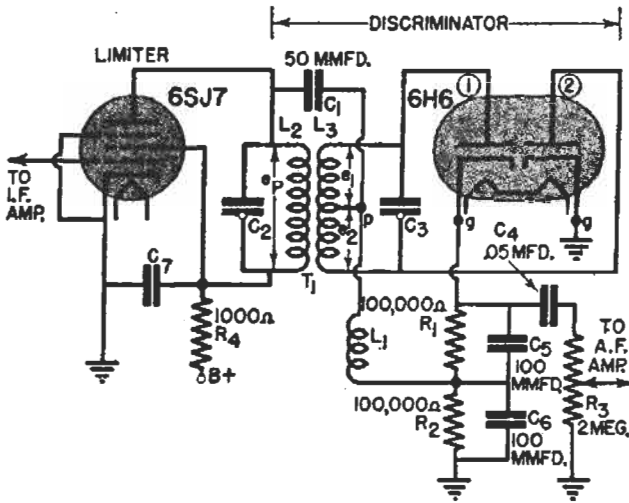


FIG. 22. Typical discriminator circuit for an f.m. receiver. The discriminator load resistors are the 100,000-ohm resistors  $R_1$  and  $R_2$ .

Limiter output voltage  $e_p$  in series with  $e_1$  is applied to diode section 1 of the 6H6 double-diode tube. (D.C. blocking condenser  $C_1$  provides an r.f. path from the upper end of  $L_2$  to  $p$ , while 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 diode section 1. Thus,  $e_p$  exists across choke coil  $L_1$ , and acts in series with  $e_1$  or  $e_2$  on either diode.) Likewise, 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 section 1 is  $R_1$ , and the load resistor for diode section 2 is  $R_2$ . The chassis provides the connecting path between the lower end of  $R_2$  and the cathode of diode section 2.

Electrons flow in opposite directions through  $R_1$  and  $R_2$ , as you can readily

verify 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 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 which 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 conditions: 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.

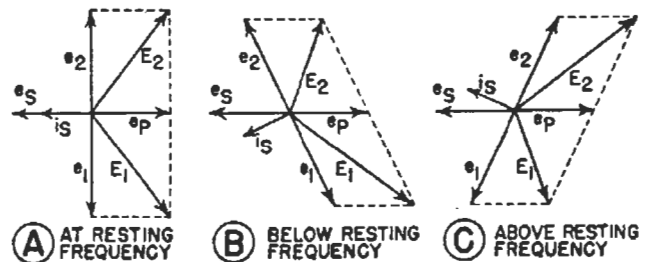


FIG. 23. These vector diagrams will help you to understand the action of the discriminator in an f.m. receiver.

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. 23, with primary voltage  $e_p$  serving as the reference vector in each case.

The r.f. voltage  $e_s$  which is induced in secondary winding  $L_3$  is  $180^\circ$  out

of phase with the primary r.f. voltage  $e_P$ , hence  $e_S$  is shown  $180^\circ$  out of phase with reference vector  $e_P$  in each of the vector diagrams in Fig. 23.

When the limiter output signal is *exactly* at the i.f. resting value of 10.7 mc. to which the discriminator circuits are tuned (when no sound is being transmitted), secondary tuned circuit  $L_3-C_3$  is at resonance, and secondary current  $i_S$  flowing through  $L_3$  will be in phase with  $e_S$ , as indicated in Fig. 23A. The voltage produced across the entire secondary winding by this secondary current will therefore *lead* both  $i_S$  and  $e_S$  by  $90^\circ$ . The resulting secondary voltage ( $e_1+e_2$ ) across  $L_3$  is utilized only in connection with the center tap of  $L_3$ , however, so if we show  $e_1$  leading  $i_S$  by  $90^\circ$ , we must show  $e_2$  as lagging  $e_1$  by  $180^\circ$ , just as in Fig. 23A.

Adding  $e_P$  and  $e_2$  vectorially gives  $E_2$  as the resultant voltage acting upon diode section 2. Likewise, adding  $e_P$  and  $e_1$  vectorially gives  $E_1$  as the resultant voltage acting upon diode section 1. The vector diagram in Fig. 23A shows that these two voltages are equal for the no-modulation condition, and hence the d.c. voltages developed across  $R_1$  and  $R_2$  by the two diode sections are equal in magnitude. The resultant voltage across both  $R_1$  and  $R_2$  is therefore zero, 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. 23B. Since voltages  $e_1$  and  $e_2$  must be  $90^\circ$  out of phase with  $i_S$ , we have the unequal resultant voltages  $E_2$  and  $E_1$ , as in Fig. 23B. With diode section 1 getting the higher r.f. voltage  $E_1$ , 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 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.

By a similar analysis, we obtain the vector diagram shown in Fig. 23C for the condition wherein the limiter output frequency is *higher* than the i.f. resting frequency. The net voltage applied to the input of the a.f. amplifier by  $R_1$  and  $R_2$  combined is now *negative* with respect to ground.

The frequency discriminator circuit shown in Fig. 22 thus produces at its output a d.c. voltage which is at each instant proportional to the deviation in the incoming signal frequency from its resting value, and having 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 for modulation purposes at the f.m. transmitter.

R.F. by-pass condensers  $C_5$  and  $C_6$  in Fig. 22 must have a low reactance at 10.7 mc., 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.

*S Curve for Discriminator Action.* The solid-line curve in Fig. 24 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, for otherwise amplitude distortion would be present in the a.f. output of the discriminator.

The linearity of the S curve in Fig. 24 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 are particularly important; these factors must make the discriminator characteristic combine with the r.f.-limiter characteristic in such a way that the individual resultant voltages,  $E_1$  and  $E_2$ , will at each instant be proportional to the frequency deviation. When this condition is attained, the combined discriminator output voltage across  $R_1$  and  $R_2$  will be proportional to the frequency deviation, thereby giving across these resistors the desired a.f. voltage without 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 dash-dash S curve in Fig. 24 illustrates this condition.

When the S curve for discriminator action has too short a linear region for weak signals, reproduction will still be satisfactory at medium and low program loudness levels, but amplitude distortion will occur during

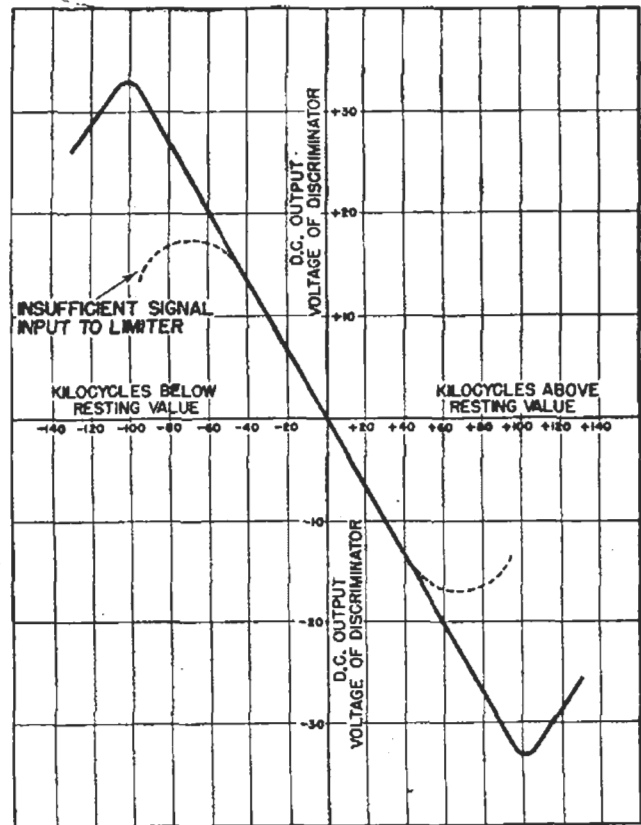


FIG. 24. S curves for an f.m. receiver, showing how the discriminator output voltage varies with frequency deviation. Curves like this are obtainable only if the over-all grid voltage-plate voltage characteristic of the limiter is flat beyond the point at which saturation begins (like curve 1 in Fig. 18A). When this over-all limiter characteristic is not flat (is like 2 or 3 in Fig. 18A), the angles which the curves make with the horizontal axis will be different (giving either lower or higher discriminator output voltages), and the curves will no longer be linear (resulting in distortion of the audio output signal). You will often find an S curve like this drawn from the lower left to the upper right. Reversal of connections to the discriminator diode plates gives this result; either connection is satisfactory, for the polarity of an a.f. signal cannot be detected by human ears.

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 dash-dash curve in Fig. 24. This 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.

**Alignment of Limiter and Discriminator Stages.** During servicing of an f.m. receiver, the limiter and discriminator can only be aligned for best possible operation. Little can be done to alter the linearity of the discrim-

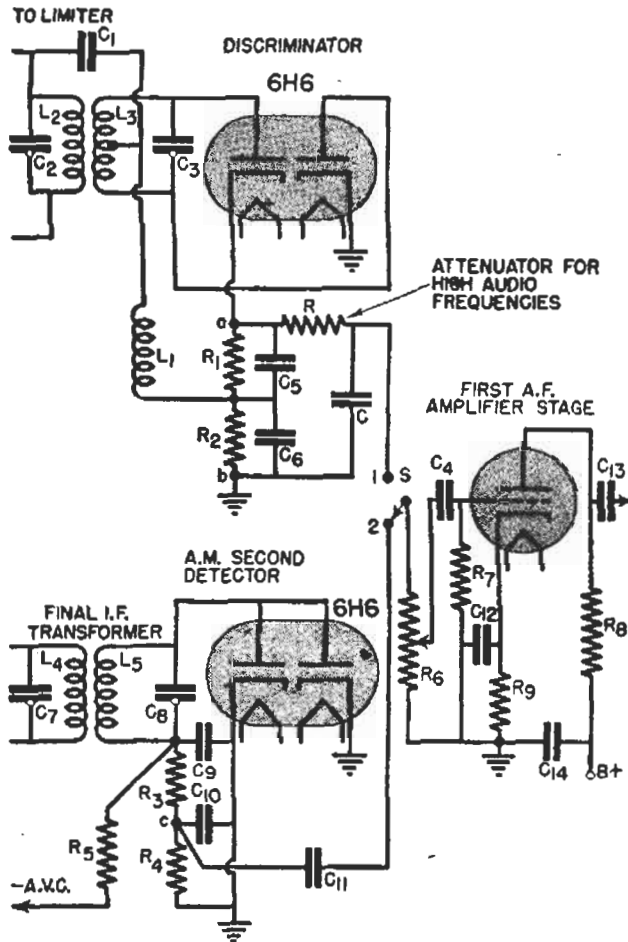


FIG. 25. This diagram shows how section S of the band-changing switch in an f.m.-a.m. receiver can be used to connect either the discriminator or the a.m. second detector to the input of the audio amplifier. Note that volume control  $R_0$  is effective at both switch settings.

inator, for this linearity is controlled chiefly by the limiter characteristics and the design and construction of the discriminator transformer.

**Alignment Procedure.** Connect a signal generator to the input of the final i.f. amplifier stage for f.m., set this s.g. to the i.f. resting frequency (usually 10.7 mc.), then open the grid return lead of the limiter circuit, insert a .01-mfd. condenser at this

point, and connect across the condenser a 0-200-microampere d.c. meter.\* (In circuit diagrams of f.m. receivers, you will often find in the grid return lead of the limiter stage a condenser, about .01 mfd., shorted out by a lead. This condenser is inserted for measuring the grid current during alignment of the limiter; the service technician simply unsolders the shorting lead, then connects his microammeter across the condenser. This is illustrated in Fig. 16, where  $C_4$  would be the .01-mfd. condenser provided for measuring purposes. The microammeter would be connected to points a and b, after first opening the lead between these two points.)

With these preliminary connections made, turn on all apparatus, adjust the limiter input trimmers ( $C_1$  and  $C_2$  in Fig. 16) for maximum grid current as indicated by the meter, adjust the signal generator output voltage so that the meter reading is at least 50 microamperes, then readjust  $C_1$  and  $C_2$  a final time for a maximum meter reading. This 50-microampere or larger current is necessary so the limiter will provide normal loading on the preceding resonant circuit.

Without changing the s.g. settings, remove the microammeter (or vacuum tube voltmeter) from the limiter input circuit and restore the connection between points a and b. Now connect a high-resistance voltmeter across one of the diode load resistors (across either  $R_1$  or  $R_2$  in Fig. 22), and adjust the discriminator transformer primary trimmer ( $C_2$  in Fig. 22) for maximum output voltage.

Next, connect the high-resistance voltmeter across both  $R_1$  and  $R_2$  (a

\* With limiter circuits which also provide an a.v.c. voltage, a vacuum tube voltmeter connected to the a.v.c. source can be used in place of the microammeter. The adjustment would be made for maximum a.v.c. voltage, as this would then correspond to maximum grid current.



connection between the cathodes of the diode sections in Fig. 22 will do this), and adjust the discriminator transformer secondary trimmer ( $C_3$  in Fig. 22) for zero d.c. output voltage. This last adjustment is usually quite critical.

This completes the alignment of the limiter and discriminator stages. The remainder of the tuned circuits in an f.m. receiver are then aligned, usually for peak response, by adjusting each alignment trimmer in turn either for maximum a.v.c. voltage or for maximum grid current in the limiter.

*The Audio System.* The over-all fidelity of an f.m. receiver is determined to a great extent by its audio system. If the receiver is intended only for f.m. reception with a minimum of noise, and high fidelity is not required, an ordinary a.f. amplifier and loudspeaker will probably be employed. When high-fidelity f.m. reception is desired, however, the audio amplifier will be designed to handle a frequency range from about 30 cycles to about 15,000 cycles, and a specially designed reproducing system will be used to handle this extremely wide range of audio frequencies.

The sound-reproducing system in a high-fidelity f.m. receiver will usually employ at least two loudspeakers, one to handle low and intermediate frequencies, and the other to handle the higher frequencies. These loudspeakers will be coupled to the audio amplifier output stage through special networks, in order to secure proper division of output energy to the two loads. The entire sound-reproducing system will be housed in an acoustically corrected cabinet. The normal rating of the audio system will ordinarily be at least 15 watts, to secure sufficient power for proper reproduction of loud bass notes.

In the transmission of f.m. signals, it is customary to accentuate or pre-

emphasize the higher-frequency notes at the transmitter, so that these will over-ride noise. To cut these high frequencies back down to normal, high-fidelity f.m. receivers have a high-frequency attenuator in the audio amplifier. This high-frequency attenuator (usually a simple resistor-condenser circuit) compensates for the pre-emphasis at the transmitter and thus restores the normal balance between highs and lows.

In a combination f.m.-a.m. receiver, a single audio system will serve for both types of reception. A switch like that shown in Fig. 25 will ordinarily be provided to connect the audio amplifier input circuit to the detector stage being employed for a particular program. This switch will be ganged to the main band-changing switch in the r.f. system, so as to make detector switching automatic.

In the circuit of Fig. 25, 50,000-ohm resistor  $R$  and .001-mfd. condenser  $C$  in the f.m. channel form the attenuator which counteracts f.m. transmitter pre-emphasis of high audio frequencies. All commercial f.m. transmitters will have the same amount of pre-emphasis, and this will be such that an attenuator having a time constant of 75 micro-seconds will restore normal receiver response for high audio notes. Normal attenuation of the a.f. amplifier at high audio frequencies, due to stray shunt capacities between circuit leads and ground, must be taken into account when designing the attenuator. Thus, the values used for  $R$  and  $C$  in Fig. 25 give a time constant of only 50 micro-seconds, but the a.f. amplifier has stray shunt capacities which make the total attenuation equivalent to that of a 75-micro-second attenuator.

If no switch were employed at the a.f. amplifier input (if both points 1 and 2 in Fig. 25 were connected to the a.f. amplifier input permanently), the

diode detector not in use would place an additional load on the other diode during positive peaks, causing amplitude distortion of the audio signal. For example, if an f.m. signal were being received, point *a* in Fig. 25 would be made alternately positive and negative with respect to point *b* and the chassis by the audio signal. This in turn would make point *c* in the unused a.m. detector alternately positive and negative with respect to the chassis. Since point *c* is connected to the plates of the unused diode through resistor  $R_3$ , the plates would likewise be positive for half of each cycle, and would be conductive. The path through  $R_3$  and the diode would

limiter. Comparing the circuit in Fig. 26 to the discriminator, you will find that one diode is reversed, so that we have a half-wave rectifier rather than a push-pull type. Now the voltage across  $C_2$  and  $C_3$  is the sum of the two rather than their difference. Hence, the audio must be taken off in a bridge arrangement from terminals *m* and *n*.

When the signal input is such that *k* is positive with respect to *l*, the diodes both conduct, causing  $C_4$  to be charged (in parallel with  $C_2$  and  $C_3$  in series) to a value corresponding to the average carrier level of the signal. With no modulation, the voltages  $e_1$  and  $e_2$ , added vectorially to  $e_3$ , are equal, so the  $C_2$  and  $C_3$  drops are

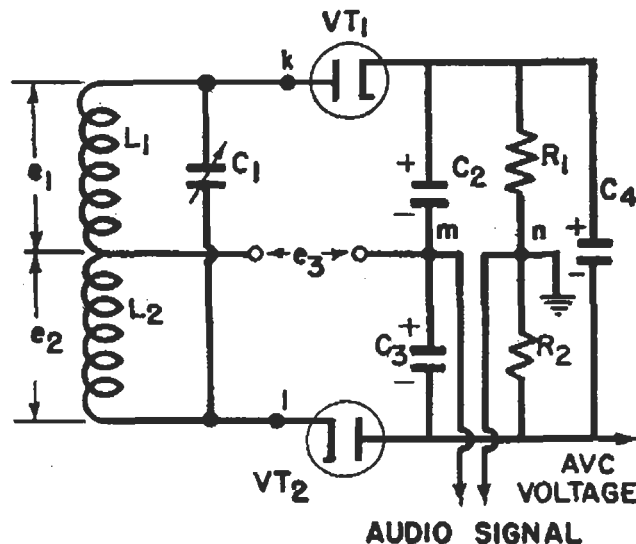


FIG. 26. A ratio detector.

then be a load on the f.m. detector during each positive a.f. peak, thus reducing the positive peaks. To prevent this condition, the detector switch is used.

*Ratio Detectors.* Recently, there have been introduced several hybrid circuits that perform both the functions of limiting and discrimination in a single tube. One example of this is the "ratio" detector, also known as the Seeley detector. Basically, it is like the discriminator you have been studying, except that it is insensitive to amplitude changes and hence acts as its own

equal. Hence *m* is at the same potential as *n*, and there is no audio. When the  $e_1$ — $e_3$  vector is longer than the  $e_2$ — $e_1$  one, the drop across  $C_2$  is greater than that across  $C_3$ , and vice versa.

If we assume that the initial  $C_4$  voltage is 10 volts, then the  $C_2$  and  $C_3$  sum must equal this, or be 5 volts each at the resting frequency. Then, should the signal make the  $C_2$  voltage 8 volts, and the  $C_3$  voltage 2 volts, point *m* will be 3 volts more negative than formerly. (Point *n* remains fixed by the  $C_4$  drop.) When the signal swings the other way, then point *m* moves positive. Hence, we get our audio out-

put from the frequency swings, as in the discriminator circuit.

The  $C_4-R_1-R_2$  time constant is made purposely long, so that the voltage across it follows the average amplitude of the incoming signal. This fixes the potential of terminal  $n$ . Also, should there be any sudden amplitude change, such as may be produced by noise, there can be no quick change in the  $C_4$  voltage; hence the  $C_2-C_3$  sum remains relatively fixed. The noise pulse is dropped across the diodes because they pass more current trying to charge  $C_4$  to a new value. As long as the  $C_2-C_3$  sum is constant, only frequency variations can cause an audio output by shifting the  $C_2$  and  $C_3$  volt-

tion at the transmitter, and that the frequency deviation from this resting frequency be at all times proportional to the sound level at that instant.

Two different methods for securing frequency modulation with the required stability and linearity are widely used. One involves the use of an inductance tube and an automatic crystal monitoring control, while the other, suggested by Major Armstrong, starts with a crystal oscillator and utilizes a phase-shifting circuit. Let us analyze each method in turn.

*Inductance Tube Transmitter.* A block diagram of a commercial f.m. transmitter of the inductance tube

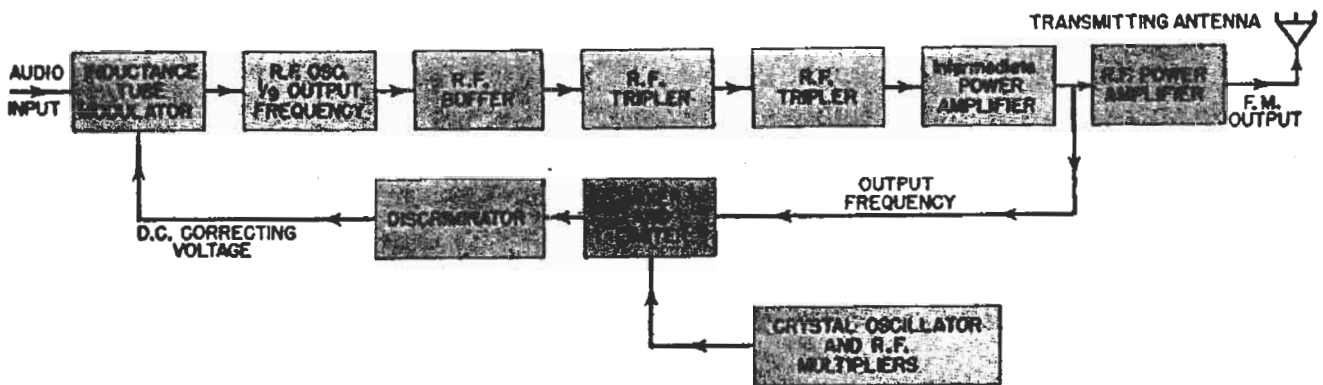


FIG. 27. Block diagram of an f.m. transmitter employing an inductance tube in the modulator stage. The main oscillator stage is of the ordinary L-C type, but its frequency is corrected automatically by the auxiliary crystal oscillator acting through an automatic frequency control system.

ages individually. Hence, this is a combination limiter-discriminator that is popular in the less expensive sets particularly.

### Modern F.M. Transmitters

Although the method shown in Fig. 6 for producing frequency modulation of a carrier signal is simple and effective, it lacks frequency stability; an r.f. oscillator shunted by an inductance tube in this manner will drift in frequency with variations in temperature and with variations in the d.c. supply voltage.

In commercial f.m. broadcast stations, it is essential that the transmitter return to its assigned resting frequency whenever there is no modula-

tion is given in Fig. 27. In this system, the oscillator is of the conventional L-C type, with temperature-compensated parts in its tank circuit (no crystal). This r.f. oscillator is tuned to some fraction of the assigned resting frequency of the f.m. station, and frequency-multiplying stages are employed ahead of the final amplifier to bring the frequency up to the assigned resting value. Thus, in the f.m. transmitter chosen as an example for Fig. 27, the r.f. oscillator is operated at one-ninth the assigned resting frequency, and its output is fed into two frequency tripler stages which bring the signal up to the correct frequency.

The inductance tube in the modu-

lator stage is connected across the tank circuit of the L-C oscillator, and hence any change in the grid voltage of the inductance tube will change the frequency of the oscillator. Two voltages act upon the grid of this inductance tube; one is the audio signal voltage applied by the speech input amplifier, and the other is a d.c. voltage produced by the discriminator in the following manner.

Let us assume that the L-C oscillator is at exactly one-ninth of the assigned resting value, and that there is no audio modulation. By means of weak capacitive coupling to some point in the r.f. amplifier, the signal at the assigned resting value is fed into a mixer stage, where it beats with a lower fixed r.f. signal produced by a temperature-controlled crystal oscillator acting through frequency-multiplying stages. The result is a signal at the difference frequency, corresponding to the i.f. signal in a superheterodyne receiver.

The mixer output signal is fed into a discriminator circuit which is tuned exactly to this difference frequency, and which therefore produces a d.c. voltage which is proportional to deviations from this difference frequency. With the L-C oscillator exactly correct, there is negligible deviation, and the d.c. output voltage which the discriminator applies to the grid of the inductance tube is therefore zero.

When the L-C oscillator drifts in frequency, the discriminator produces a d.c. voltage which is proportional to the magnitude of the frequency drift and has a polarity determined by the direction of the drift. This d.c. voltage changes the inductance of the inductance tube exactly enough to correct the frequency drift.

The action of this automatic frequency-correcting circuit is the same when audio modulation is present, for a condenser filter in the discriminator

prevents audio signals from traveling from the discriminator to the inductance tube. Only the d.c. voltage actually due to drift from the assigned resting frequency can act upon the inductance tube.

Sometimes a portion of the audio output of the discriminator is utilized to provide inverse feed-back which compensates for distortion occurring in the transmitter. In this case, a special coupling circuit would be used to apply both the d.c. and a.f. components of the discriminator output to the inductance tube grid.

The frequency-tripling stage is simply an r.f. stage operated as a class C amplifier, with its plate resonant

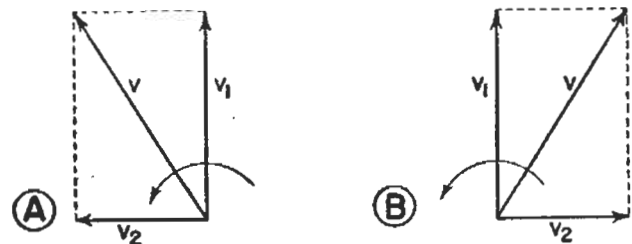


FIG. 28. Vector diagrams showing how a frequency-modulated signal can be produced by the Armstrong phase-shift method.

circuit tuned to the third harmonic of the input frequency.

Any tendency for the f.m. transmitter to drift from its assigned resting frequency thus produces a d.c. voltage which acts upon the grid of the inductance tube in such a manner as to correct the tendency toward drift. A zero-center type of voltmeter connected across the discriminator output indicates when the drift is getting dangerously great, thus warning the operator to retune the L-C oscillator manually before the discriminator loses control.

*Armstrong Phase-Shift Transmitter.* The basic principle underlying the Armstrong method is: When a signal having a definite frequency is flowing through a circuit, and another signal having this same frequency but  $90^\circ$  out of phase is suddenly sent into the circuit, there is an instantaneous phase

shift in the original signal, with the result that the frequency of the combination either increases or decreases. We can illustrate this principle by means of the vector diagrams in Fig. 28.

Vector  $V_1$  in Fig. 28A represents the original signal, having a definite frequency and a constant amplitude. As this vector rotates in the conventional counter-clockwise direction at a rate corresponding to its frequency, let us suddenly introduce into the circuit a voltage  $V_2$  which has the same frequency as voltage  $V_1$  but leads  $V_1$  by exactly  $90^\circ$ . The sudden introduction of this new voltage  $V_2$  causes the net voltage  $V$  of the combination to shift in a counter-clockwise direction, corresponding to a momentary increase in frequency. If voltage  $V_2$  is suddenly increased in amplitude after it is introduced into the circuit, resultant vector  $V$  speeds up again, producing another momentary increase in frequency.

On the other hand, if the suddenly-introduced  $V_2$  lags  $V_1$  by  $90^\circ$ , the resultant vector  $V$  will move clockwise, corresponding to a momentary decrease in frequency, as indicated by the vector diagram in Fig. 27B. If vector  $V_2$  is varying continually in both amplitude and polarity in accordance with a sine-wave audio signal, resultant voltage vector  $V$  will swing above and below its original frequency in a corresponding manner.

All this means that we can produce frequency modulation of a carrier signal by combining it with another signal which has the same carrier frequency but is alternately  $90^\circ$  leading and  $90^\circ$  lagging the original signal in accordance with positive and negative swings of an audio signal, and varies in amplitude in accordance with variations in the amplitude of the audio signal.

In order to make the deviation in frequency dependent only upon the amplitude and polarity of  $V_2$ , an attenuator is introduced into the audio amplifier of the transmitter so as to make the audio output voltage decrease gradually as the audio frequency increases. This insures that a 1-volt audio signal at 10,000 cycles will produce the same frequency deviation as a 1-volt audio signal at 1000 or 100 cycles.

As a rule, both  $V_1$  and  $V_2$  in an f.m. transmitter of this type will have relatively low frequency values, and fre-

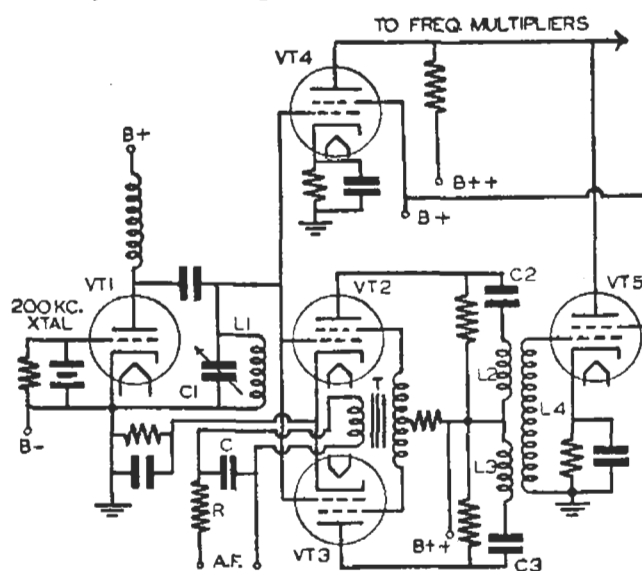


FIG. 29. Basic circuit used in the Armstrong phase-shift type of f.m. transmitter.

quency multipliers will be used to bring the f.m. signal up to the correct frequency for the transmitting antenna.

The basic Armstrong circuit for securing frequency modulation by means of this phase shift method is given in Fig. 29. Vacuum tube  $VT_1$  in a crystal oscillator circuit produces the initial low r.f. carrier value (200 kc. would be a typical value) with negligible frequency drift. The output of this crystal oscillator is fed into r.f. amplifier tube  $VT_4$  and also into the balanced r.f. stage including tubes  $VT_2$  and  $VT_3$ .

Note that the grids of this balanced stage are fed in phase, while the output of this stage goes to an untuned

r.f. transformer having a center-tapped primary  $L_2$ - $L_3$ . With in-phase plate currents flowing in opposite directions through the primary to the center tap, and with the tubes balanced so that the currents in the two sections of the primary are normally equal in magnitude, the resultant flux linked with secondary winding  $L_4$  is zero, and no voltage is induced in this winding for transfer to the frequency multipliers through amplifier stage  $VT_5$ .

The values of the parts in the plate circuits of the two balanced amplifier tubes are such that the r.f. current in each half of the primary lags the a.c. grid voltage by approximately  $90^\circ$ ; in other words, the plate loads for  $VT_2$  and  $VT_3$  are essentially pure inductances.

The operation of this circuit depends upon the basic fact that when the a.c. resistance of a tube is varied, its a.c. plate current will vary correspondingly. We can change the a.c. plate resistance of a tube by varying the screen grid voltage.

In the circuit of Fig. 29, the screen grid voltages for both  $VT_2$  and  $VT_3$  in the balanced amplifier stage are applied through the secondary winding of audio transformer  $T$ . With this arrangement, an a.f. voltage applied

to the primary winding will make the screen grid voltage on one tube increase, and make the screen grid voltage on the other tube decrease. This causes an unbalance in the r.f. plate currents flowing through  $L_2$  and  $L_3$ , and consequently we secure a resultant flux which links with  $L_4$ .

The resultant current flowing through both  $L_2$  and  $L_3$  will either lag or lead the a.c. grid voltage by  $90^\circ$ , depending upon which coil current is greatest. This means that the r.f. voltage induced in secondary winding  $L_4$  will either lead or lag the a.c. grid voltage by  $90^\circ$ , depending upon whether the a.f. input voltage is swinging positive or negative, and the amplitude of this r.f. voltage in  $L_4$  will vary in accordance with the amplitude of the a.f. input signal. The voltage across  $L_4$  thus corresponds to vector  $V_2$  in the diagrams of Fig. 28.

Amplifier stage  $VT_5$  merely provides a means for securing the proper relationship between the amplitudes of r.f. voltages  $V_1$  and  $V_2$ . These two voltages, from tubes  $VT_4$  and  $VT_5$ , are fed in parallel into the frequency multiplier system. They combine to give the desired frequency-modulated signal, as was explained in connection with the vector diagrams in Fig. 28.

## TEST QUESTIONS

Be sure to number your Answer Sheet 34FR-2.

Place your Student Number on every Answer Sheet.

Send in your answers for this lesson immediately after you finish them. Doing this insures that the graded answers will reach you while the subject matter is still fresh in your mind, and you will get the greatest possible benefit from our speedy personal grading service. Never hold up a set of lesson answers.

1. To secure a high signal-to-noise ratio with minimum inter-station interference at receivers, should the maximum frequency deviation in an f.m. system be high or low?
2. In what two ways can an interfering signal affect a desired f.m. signal?  
*It can change both the frequency and the amplitude of the f.m. signal.*
3. What is the purpose of the discriminator in an f.m. receiver?
4. What section in an f.m. receiver produces the a.v.c. voltage?  
*P12  
The limiter section.*
5. What section in an f.m. receiver removes amplitude variations from the f.m. signal?  
*The limiter section.*
6. What characteristic should a limiter have in order to block out sudden noise surges?  
*It must be fast-acting.*
7. Why is it permissible to have a peaked over-all response for the r.f. and i.f. sections in a properly designed f.m. receiver?  
*Then the limiter will handle weak as well as strong signals without distortion.*
8. Describe the type of distortion which occurs when the S curve for discriminator action has too short a linear region for weak signals.
9. Why is it essential that grid current be drawn by the limiter tube during alignment of the resonant circuit ahead of the limiter?  
*P29*
10. Why is an attenuator for high audio frequencies used in the audio system of an f.m. receiver?  
*P30  
It reduces the ratio between high and low frequencies that were disturbed at that the transmitter.*