FUNDAMENTALS OF FREQUENCY MODULATION

Finished Sept 10, 1960

46RC

NATIONAL RADIO INSTITUTE

ESTABLISHED 1914

WASHINGTON, D. C.

PIONEER HOME STUDY RADIO SCHOOL

NRI

STUDY SCHEDULE No. 46

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

☐ 7. Answer Lesson Questions.

☐ 8. Start Studying the Next Lesson.

FM1M156 1956 Edition Printed in U.S.A.

FUNDAMENTALS OF FREQUENCY MODULATION

Basic Principles of F.M.

WIDE-BAND frequency modulation, as you have learned, is a broadcasting system in which the frequency of the carrier of the broadcast station varies in accordance with the modulating signal. This method of modulation is used for high-fidelity broadcasting because it permits a higher signal-to-noise ratio and a larger dynamic (volume) range than does a.m. These advantages are gained because an f.m. system is insensitive to many amplitude variations that in a.m. cause background noise, static, and other interfering signals.

One disadvantage of wide-band f.m. broadcasting is that the r.f. channels must be 200 kc. wide, as compared to 10 kc. for standard a.m. This problem has been solved by assigning this service to the v.h.f. band where more channels are available. The present 88-108 mc. f.m. band, for example, accommodates 100 f.m. channels.

REVIEW OF A.M.

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. Let's review briefly what you have learned about a.m.

The signal peaks of an unmodulated r.f. carrier 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 rise and fall in amplitude in accordance with variations in the audio signal, as shown at the right in Fig. 1A.

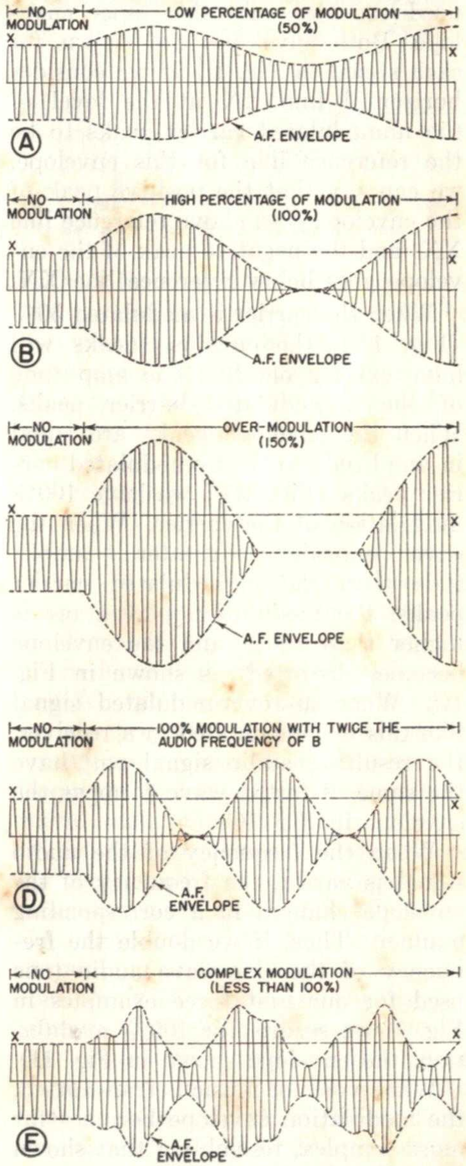


FIG. 1. Examples of amplitude-modulated signals.

A dashed line drawn through the positive peaks of the modulated r.f. signal in Fig. 1A gives 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 dashed 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.) If we consider horizontal line XX at the level of the unmodulated carrier peaks to be the reference line for this envelope, 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.

When the carrier is modulated 50% (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 carrier peaks (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 overmodulated signal like this is demodulated in a receiver, the resulting audio signal will have the same distorted wave form as the envelope in Fig. 1C.

When the frequency of the audio signal is varied, the frequency of the envelope changes in a corresponding 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 envelope 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 that 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), and 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. Overmodulation occurs whenever the sum of all the side frequency amplitudes is greater than the carrier amplitude.

CHARACTERISTICS OF AN F.M. SIGNAL

An r.f. signal can also be modulated by varying the frequency of the r.f. signal in accordance with an audio

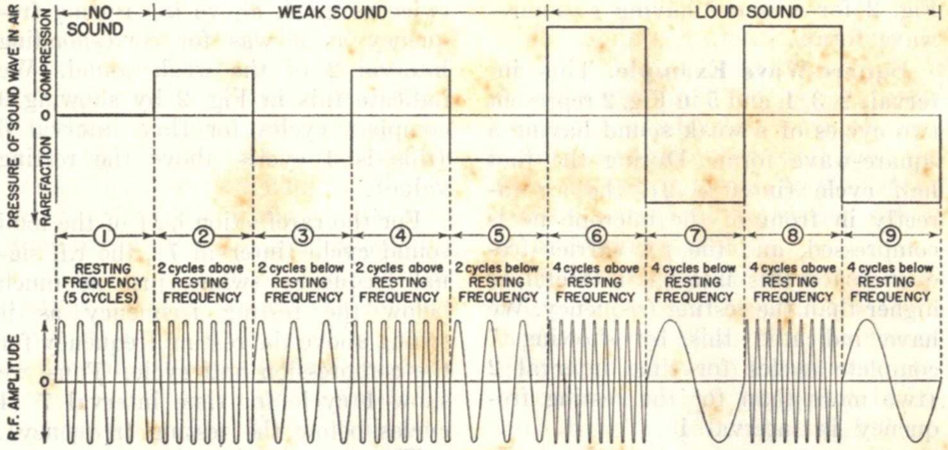


FIG. 2. Example showing how a sound wave having a square-wave form, makes the r.f. signal frequency of an f.m. transmitter vary above and below the assigned resting frequency.

signal 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 that is radiated by the transmitter when no sound is being transmitted is at an assigned r.f. value called the resting frequency or center frequency. 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 chose 5 cycles to represent the resting frequency in this diagram.)

Frequency Deviation. When an a.f. signal is fed into an f.m. transmitter, the frequency of the r.f. signal swings above or below the resting value in proportion to the increases or decreases in air pressure at the microphone that cause the audio signal. We will assume that an increase in air pressure at the microphone (compression) causes an increase in frequency. (Actually a compression wave could cause either an increase or decrease in frequency, depending on the number of a.f. stages and phase-reversing transformers between the microphone and the modulator output, but it is common practice to consider that compression causes a frequency increase.) 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).

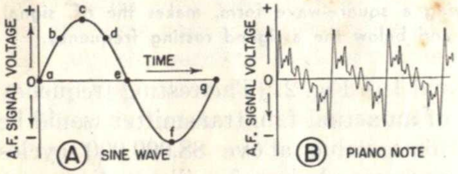


FIG. 3. A pure sine-wave signal like that at "A" or a complex audio signal like that at "B" both make the frequency of an f.m. transmitter swing above and below its resting value exactly in proportion to the positive and negative swings of the audio signal.

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, an increase in sound that doubles the audio voltage doubles the frequency deviation. Time intervals 6, 7, 8, and 9 in Fig. 2 illustrate the effect of a loud, square-wave sound.

During the first (compression) half-cycle of this louder sound wave (interval 6) 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 (4 cycles below 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 shows 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 represent either 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 in Fig. 3A.

As the audio signal increases in amplitude from a to b to c, the frequency of the r.f. signal will rise proportionately above the resting value 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 proportionately back to the resting value. Notice 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.

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 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 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 interstation interference at receivers make necessary a high value for the maximum frequency deviation. The greater the maximum frequency deviation used for desired signals, the less noticeable to the listener will be a frequency deviation due to an interfering signal.

The channels that have been made available to f.m. broadcast stations in the United States and its possessions by the Federal Communications Commission are .2 mc. apart in the v.h.f. band between 88 and 108 mc. 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 FCC 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.

A frequency deviation of 75 kc. corresponds to 100% modulation. However, most f.m. transmitters are capable of producing 100-kc. swings without distortion. This deviation, which corresponds to 133% modulation, is used for test purposes, in fact, the point indicating 133% modulation is marked on the VU meters used in f.m. transmitters. Modulations over 100% do not necessarily cause distortion in f.m. receivers, as you will learn when we study f.m. receiver operation in detail in a later Lesson.

The monitor engineer in an f.m. studio has meters before him that show when the loudness level at the microphone is exceeding the value that 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, he increases the a.f. gain when the loudness level at the microphone becomes very low for appreciable periods of time. Thus, the monitor engineer does essentially the same thing in both f.m. and a.m. systems: he keeps the audio level 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 65 kc. (.065 mc.) above and below the resting value. (Although a maximum deviation of 75 kc. is permitted by the FCC, many f.m. stations are operated with a maximum swing of only 65 kc. This provides a larger margin of safety so that on unanticipated loud peaks there is less likelihood of "splatter" into the adjacent channels.) Thus, the r.f. signal frequency will go up to 97.365 mc. for the positive peak of each audio cycle, and will drop down to 97.235 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 97.365 to 97.3 to 97.235 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 .065 mc. divided by 100, or .00065 mc. For this weak signal, then, the r.f. signal frequency will vary from 97.30065 mc. to 97.29935 mc., and there will be 1000 complete swings like this in each second.

If the f.m. station in our example next transmits a 100-cycle audio signal that is as loud as the original 1000-cycle signal, the deviation will still be 65 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.

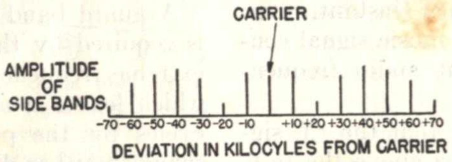


FIG. 4. This shows how frequency modulation of a carrier produces a large number of pairs of side bands even when the modulating signal is a pure sine wave. In this case the frequency of the modulating signal is 10 kc. and the modulating index is 5.

F.M. Side Bands. In our study of f.m., we may sometimes find it necessary to refer to the side bands generated by a frequency-modulated signal.

First let us consider the case when the modulating signal is a simple sine wave. It has been found both by mathematical analysis and by actual measurements, that such a frequency-modulated signal consists, as shown in Fig. 4, of the carrier plus two sets of side bands, one on each side of the carrier. These side bands are spaced the frequency of the modulating signal apart. Thus, if the modulating signal is 10 kc., the side bands are 10 kc., 20 kc., 30 kc., 40 kc., etc., above and below the carrier. Theoretically, these side bands extend to infinity; practically, however, the amplitudes of the side bands beyond the frequency deviation of the transmitter (± 75 kc.) are so small that they can be ignored. As you can see, the amplitudes of these side bands, even for sine-wave modulation, do not follow a very easily determined pattern.*

As the amplitude of the modulating signal is varied, the amplitudes of the side bands and of the carrier will vary; it is even possible for the carrier or any pair of side bands to drop to zero as the amplitude of the modulating signal varies. In general, the amplitudes of the side bands farthest from the carrier will increase as the modulating signal is increased. This is another reason for keeping the modulating signal amplitude to a value that prevents "splatter" into the next f.m. channel.

In standard f.m. transmitters, the maximum frequency deviation (F_D) is 75 kc. and the maximum frequency of the audio signal (F_M) to be transmitted is 15 kc. The ratio of F_D to F_M is called the *modulation index* or the *deviation ratio*. This ratio is given mathematically by:

$$\beta = \frac{F_D}{F_M}$$

The *modulation index* (deviation ratio) of standard f.m. stations is 5 (the frequency deviation is 75 kc. and the maximum frequency of the modulating signal is 15 kc.)

When the modulating signal is a complex wave, the number of side bands generated is greatly increased; however, the important side bands are still inside the 75-kc. deviation from the carrier in normal f.m. transmitter operation, hence ± 75 kc. is the practical and satisfactory band limit.

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 D, as shown in Fig. 5A. The amplitude (length) of this desired-signal vector D will be constant, and the rate at which the 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.

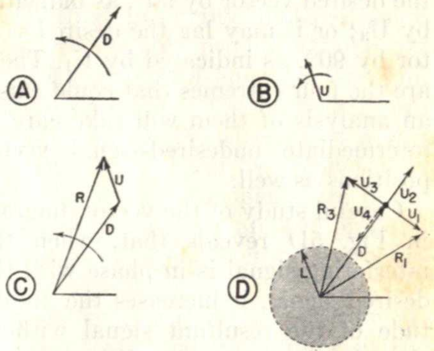


FIG. 5. Vector diagrams showing how an undesired signal "U" can affect the frequency and amplitude of a desired f.m. signal "D".

Now, let us suppose that we also have an undesired signal at the input of our f.m. receiver. It may be due either to noise interference or to an undesired r.f. carrier signal from some other station. Let us say that at the instant of time under consideration for Fig. 5A, the undesired signal has the amplitude and phase represented by vector U in Fig. 5B.

When we combine undesired-signal vector U with desired-signal vector D, as shown in Fig. 5C, we obtain resultant vector R. This represents the actual signal received at the antenna. The amplitude of vector R is greater than that of desired-signal D,

*The amplitude of the carrier and side bands can be determined only by using advanced mathematics.

showing that an interfering signal can affect the amplitude of a desired incoming signal. Furthermore, vector R is ahead of vector D; this indicates, as you will learn later in this Lesson, that the frequency of the received signal appears higher than it should be.

Interference vector U in Fig. 5B may have any phase relationship whatsoever throughout the entire range of 360°. It may be in phase with desired-signal vector D, as indicated by vector U_2 in Fig. 5D; it may be 180° out of phase, as indicated by vector U_4 in Fig. 5D; it may lead the desired vector by 90°, as indicated by U_3; or it may lag the desired vector by 90°, as indicated by U_1. These are the four extremes that could exist; an analysis of them will take care of intermediate undesired-signal vector positions as well.

Careful study of the vector diagram in Fig. 5D 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 that an f.m. receiver does not respond to amplitude variations in the input signal. This lack of response is produced by using a second detector whose output does not depend on amplitude variations or by using limiter stages that remove all amplitude variations ahead of the second detector. Amplitude variations of the received signal due to any interference will therefore not affect the output of the receiver.

Any frequency change due to noise will show up in the output of an f.m. receiver, for such frequency changes are converted into corresponding changes in amplitude by the "frequency discriminator" section of the receiver. However, the amount of frequency deviation produced by noise signals is quite small when compared to the high frequency deviation used in standard f.m. broadcasting. For this reason, the output noise in an f.m. receiver is generally negligible even though the original interference may be quite severe. For example lightning, which causes considerable noise in a.m. systems, produces practically no interference in a properly designed f.m. receiver.

Let us now consider what happens when the interference is another f.m. station operating on the same frequency. As we saw in Fig. 5, the received signal will vary in amplitude and in frequency. Again, any amplitude variation will be removed in the receiver. Whether or not the difference in frequency is noticeable depends on the relative amplitudes of the two signals. In standard wide-band f.m. it has been found that the interference is unnoticeable in the output of the f.m. receiver if the desired station is at least twice the amplitude of the undesired station. Remember, in comparison, that in a.m. the desired station must have an amplitude at least 100 times that of the undesired station to eliminate interference.

This important characteristic of a wide-band f.m. station means that there is a minimum of interference

between stations operating on the same frequency and that, therefore, channel assignments can be duplicated more often than in a.m. if necessary.

MULTIPLEX F.M. OPERATION

Frequency modulation has the added advantage of permitting multiplexing, that is, 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 facsimile 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 facsimile, and this modulated 20-kc. secondary carrier (called a "subcarrier") could then be 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 facsimile signal would be taken out ahead of the audio amplifier, and demodulated by conventional means in a separate detector stage to operate the facsimile receiver. Filters would be used to keep the 20-kc. signal out of the regular a.f. amplifier, to prevent overload of a.f. stages in the broadcast portion of the f.m. receiver.

AUDIO PRE-EMPHASIS AND DE-EMPHASIS

In standard f.m. broadcasting, there is deliberate pre-emphasis of high frequency components of the modulating signal at the transmitter and subsequent de-emphasis of them at the receiver. Let us see how this makes high fidelity reproduction possible with a minimum of background noise in the receiver output.

For high fidelity reproduction of speech or music, it is necessary that the harmonics or overtones in the upper audio frequency range be accurately reproduced. An f.m. system is capable of transmitting this full audio range because its band width depends on the *amplitude* but not the *frequency* of the modulating signal. Practically, however, the amplitude of these high-frequency components is very small, and they produce only a small amount of modulation. In an f.m. receiver with a flat frequency response to the upper end of the audio range, there are many high-frequency noise components produced by resistor noise and "shot-effect" in the f.m. detector circuit. These noises are present even in the absence of a signal. When a signal is received, its high-frequency components are likely to be lost in the noise already present; the result is unsatisfactory reproduction of signal.

However, if the high-frequency response of the audio amplifier in the f.m. receiver is reduced, as shown in Fig. 6A, the high-frequency noise components are greatly reduced and are no longer objectionable. The circuit used to de-emphasize the high audio frequencies is called a "de-emphasizer." It consists, in a standard f.m. receiver, of a low-pass RC filter with a time constant* of 75 microseconds. Fig 6B shows such a filter (R_1C_1) and how it is connected in the circuit of the f.m. receiver.

Because of the de-emphasis of high audio frequencies in the receiver, it is necessary to pre-emphasize them in

*The time constant of an RC circuit in seconds is equal to the resistance R in ohms times C in farads, this is, $t = RC$. Since a microsecond is a millionth of a second, the time constant in microseconds is the resistance R in ohms times C in microfarads. In the example given, $50{,}000 \times .0015 = 75$ microseconds. The same result can be obtained by multiplying R in megohms by C in micromicrofarads.

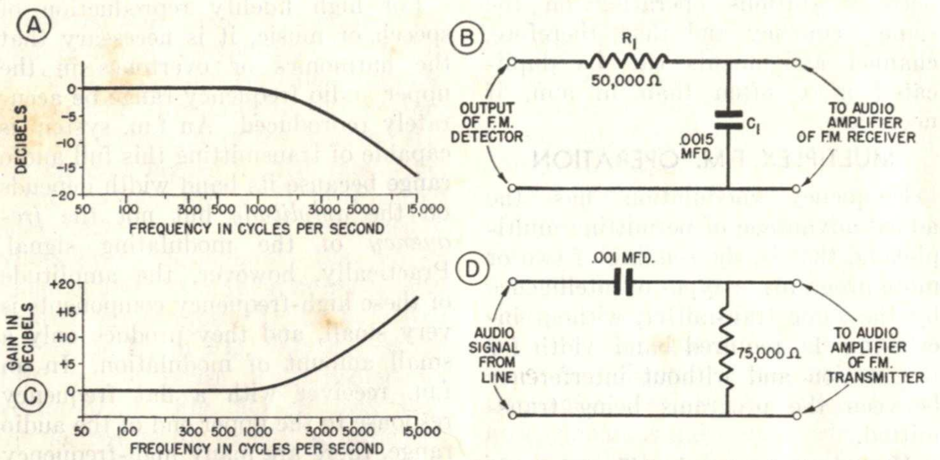


FIG. 6. In A the standard 75 microsecond de-emphasis curve for an f.m. receiver is shown, and in B a simple low-pass RC filter to obtain it is given. In C, the standard 75 microsecond pre-emphasis curve used in f.m. transmitters. D, a simple high-pass RC filter for obtaining this pre-emphasis.

the f.m. transmitter to obtain a flat over-all frequency response. The standard pre-emphasis curve shown in Fig. 6C is just the opposite of the receiver curve. The circuit producing this response is shown in Fig. 6D; it is a 75-microsecond high-pass RC filter that is connected in the transmitter as shown.

This pre-emphasis of high audio frequencies in the transmitter and corresponding de-emphasis in the receiver results in a flat over-all frequency response from microphone to loudspeaker with a minimum of noise in the output of the receiver. The resultant high-fidelity, noiseless sound reproduction is an important advantage of f.m.

DYNAMIC RANGE IN F.M. BROADCASTING

You recall that it is necessary to compress the volume range to about 25 db for efficient transmission in standard a.m. broadcasting. If the minimum signal is more than 25 db below 100% modulation, the amount of modulation is so small that it is lost in the inherent noises of an a.m. system.

In f.m., however, as we have seen, background noises are reduced to a minimum. This means that the permissible dynamic range can be much greater than 25 db. As a matter of fact, the full 70 db range of a symphony orchestra can be handled in a well-designed f.m. system. Practically, of course, the usual range is less than this. Most f.m. stations, however, are capable of transmitting a dynamic range of 45 db, which is the maximum that the acoustics of the usual location of an f.m. receiver will permit.

Direct F.M. Transmitters

Now that we have studied the basic characteristics of f.m. signals, let us see how frequency modulation is obtained.

Briefly, there are two general methods, the *direct* and the *indirect*. In the direct method, the frequency of oscillation of an LC oscillator is varied in accordance with the modulating signal. A block diagram of a direct f.m. transmitter is given in Fig. 7. The audio signal from the audio amplifier is applied through a pre-emphasis network to the frequency modulator, which varies the frequency of the LC oscillator. Frequency multiplier stages are used to increase the small amount of frequency deviation obtained in the LC oscillator to the large value needed for f.m. broadcasting. Class C high-efficiency power amplifiers capable of passing the full 150-kc. band are used to increase the power to the desired level.

Although this method is fundamentally the simplest one, there is a grave objection to it: the normal frequency drift of a free oscillator at these frequencies is more than can be tolerated in actual practice. The FCC has specified that f.m. stations should stay within 2000 c.p.s. of their assigned frequency. For a station on 100 mc., this means that the unmodulated carrier frequency must not vary more than .002% from the assigned value.

Special precautions must be taken to maintain such a frequency stability in an LC oscillator. The general method used is indicated in Fig. 7; the resting frequency of the LC oscillator is compared to a reference crystal oscillator, and any drift in frequency is automatically corrected through a frequency-correcting circuit. The various specific methods of doing this will be studied in detail in another Lesson.

In other transmitters, crystal oscillators are used as the master oscillators to obtain this high degree of stability. Since, however, it is not possible to vary the oscillating frequency of a crystal oscillator by the amount necessary for f.m. work, *indirect* methods of obtaining f.m. must be used. These methods generally consist of *phase-modulating* the crystal-controlled carrier and converting this phase modulation to frequency modulation. The relationship between p.m. and f.m. will be studied later in this Lesson and in another Lesson of your Course.

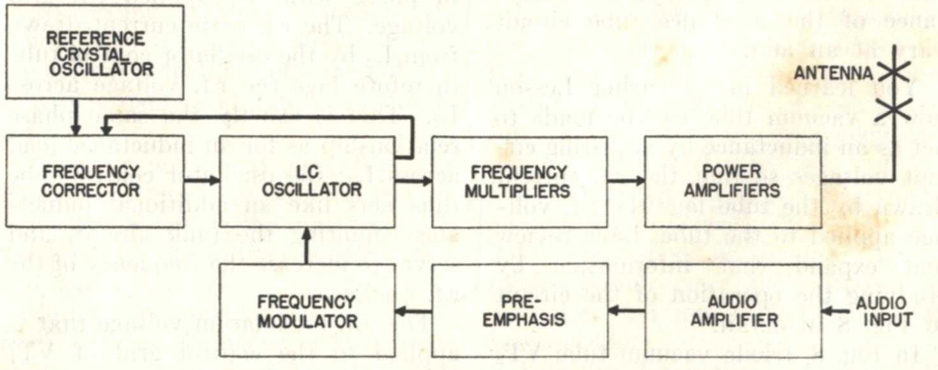


FIG. 7. The general block diagram of a direct f.m. transmitter.

REACTANCE TUBE MODULATORS

One of the simplest methods of obtaining f.m. directly is shown in Fig. 8. It depends on the fact that the frequency of operation of an LC oscillator can be varied by connecting a variable inductance across the LC tank circuit. Instead of a coil, however, a vacuum tube, known as a "reactance" tube, is used. This tube is connected in a circuit that acts as a variable inductance across the resonant circuit of an r.f. oscillator;

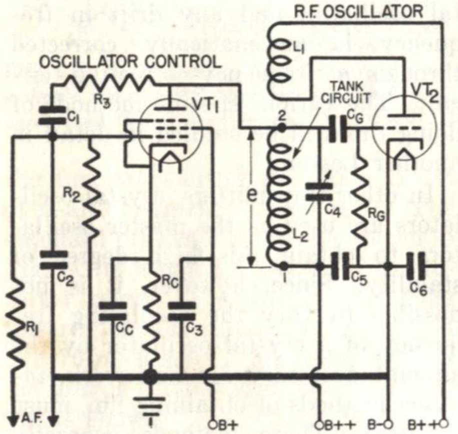


FIG. 8. This practical f.m. transmitter circuit employs an oscillator control circuit consisting of a vacuum tube that acts like an inductance shunted across the r.f. tank circuit. The effective inductance value varies from instant to instant in accordance with variations in the amplitude of the a.f. signal, thereby giving frequency modulation.

f.m. is secured by making the inductance of the reactance tube circuit vary at an audio rate.

You learned in an earlier Lesson how a vacuum tube can be made to act as an inductance by adjusting circuit voltages so that the r.f. current drawn by the tube lags the r.f. voltage applied to the tube. Let's review and expand that information by studying the operation of the circuit in Fig. 8 in detail.

In Fig. 8, triode vacuum tube VT_2 is connected into a conventional tuned-grid r.f. oscillator circuit, with L_2 and C_4 forming its tank circuit. Pentode tube VT_1 serves as the oscillator control tube that acts like an inductance; its plate is connected directly to terminal 2 of tank inductance L_2, and its cathode is connected to terminal 1 of this coil through r.f. by-pass condensers C_c and C_5 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_2-C_4. First of all, r.f. tank voltage e in Fig. 8 must be considered to be 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_3-C_1-C_2-C_5.

At radio frequencies, path R_3-C_1-C_2-C_5 is essentially resistive (the reactances of all three condensers are low with respect to the resistance of R_3), so 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 that 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 that is in phase with the applied r.f. grid voltage. The r.f. plate current drawn from L_2 by the oscillator control tube therefore lags the r.f. voltage across L_2. This is exactly the same phase relationship as for an inductance load across L_2; 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.

The a.f. modulation voltage that is applied to the control grid of VT_1 through R_1 varies the transconductance of the tube, and hence makes the a.c. plate current vary at an audio rate. Consequently, the effective 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 fequency, which is determined by the size of inductance L_2, the normal inductance of the oscillator control circuit, and the tank circuit capacity.

When an a.f. signal is applied, the frequency deviation maximum can be adjusted by setting the maximum a.f. voltage level fed to the oscillator control tube.

Resistor R_1 prevents the a.f. source from shorting the input of the reactance tube; it is really an isolating resistor.

For these reasons, the circuit in Fig. 8 is suitable as a signal source for an f.m. system that is to transmit intelligence. The f.m. signal would be taken from the tank circuit inductance L_2, either by a link coupled to this coil near terminal 1, or by capacitive coupling from terminal 2.

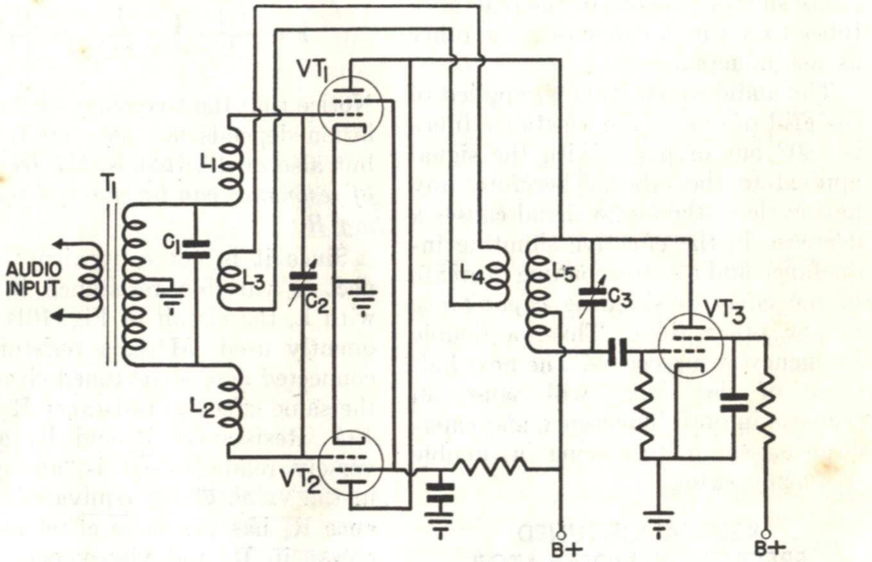


FIG. 9. A push-pull balanced reactance tube modulator circuit. Note that the 90° phase shift is obtained by link coupling L_5 to L_1 and L_2 through L_3 and L_4. The two reactance tubes VT_1 and VT_2 are 180° out of phase so that one acts as an inductance and the other as a capacitance.

BALANCED REACTANCE TUBE MODULATORS

In many commercial f.m. transmitters, two tubes are used, as shown in Fig. 9, in a balanced reactance tube circuit. The advantages of such a circuit are, first, that the frequency deviation can effectively be doubled, and, second, the two reactance tubes maintain the balance of the audio input with respect to ground. These factors permit more linear frequency deviation and low distortion of the audio signal.

This circuit contains a conventional

Hartley oscillator, using VT_3. Part of its output is inductively coupled by links L_4 and L_3 to the tuned circuit L_1, L_2, C_1, and C_2. This tuned circuit causes the r.f. voltage applied to the grids of reactance tubes VT_1 and VT_2 to be 180° out of phase, with one tube current effectively leading the r.f. voltage in L_3-C_3 by 90° and the other lagging by the same amount. These phase shifts cause one of the reactance tubes to act as a condenser, the other as an inductance.

The audio signal from T_1 applied to the grid of one of the reactance tubes is 180° out of phase with the signal applied to the other. Therefore, any half cycle of the audio signal causes a decrease in the effective shunting inductance and a corresponding decrease in the effective shunting *capacitance* of the other tube. Thus, a double frequency swing occurs. The next half cycle of the signal will cause an *increase* in both inductance and capacitance, again producing a double frequency swing.

RESISTANCE-TUNED FREQUENCY MODULATOR

Another scheme for direct frequency modulation uses resistance tuning of an LC circuit. Before we study this

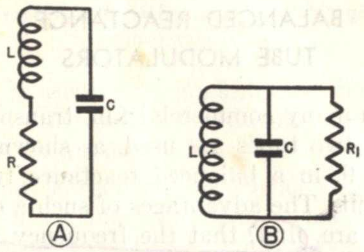


FIG. 10. Resistance tuning. The resonant frequency of a tuned circuit can be varied slightly by changing the value of the series resistance R as shown in A. Likewise, as shown in B, a variation of the shunt resistance R_1 across the tuned circuit will cause a variation in frequency of resonance.

system, let's see how a variable resistance will affect the resonant frequency of a tuned circuit.

In Fig. 10A is shown an LC circuit with a resistance R connected in series with the inductance. It has been found by experiments and by mathematical analysis that the resonance frequency of such an LCR circuit is

$$F = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{L^2}}$$

Notice that the frequency of the oscillation depends not only on L and C but also on R; that is, *the frequency of resonance can be changed by varying R.*

Since it is not convenient to connect a variable resistance in series with L, the circuit of Fig. 10B is frequently used. Here, a resistance R_1 connected across the tuned circuit has the same effect as resistance R in Fig. 10A. Resistances R and R_1 are inversely related—that is, an increase in the value of the equivalent resistance R_1 has the same effect as a decrease in R, and vice versa.

When R is negligibly small (as it generally is in most high Q tuned circuits), the frequency of resonance is simply

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

which is the formula generally used to calculate the resonance frequency of an LC circuit.

A basic circuit for producing f.m. by varying a resistance is shown in Fig. 11. VT_3 is used in a tuned-grid, tuned-plate LC oscillator circuit. The plate tank L_2-C_3 is tuned to be inductive at the frequency of operation. The grid tank circuit L_1-C_2 determines the frequency of operation of the oscillator; diode VT_2 is used as a variable resistance across this grid

circuit to produce frequency modulation of the oscillator.

The d.c. plate current for tube VT_1 (which amplifies the audio signal) flows through VT_2 and R_1. An audio signal applied to VT_1 causes its plate current and hence the plate current of VT_2, to vary. This causes the a.c. plate resistance of diode VT_2 to change. Since C_1 is an r.f. by-pass condenser this a.c. resistance is effectively connected across L_1-C_2, as the diagram shows. Since this resistance varies when an audio signal is applied to VT_1, the frequency of the oscillation of VT_3 will likewise vary.

The advantage of this circuit is that the diode VT_2 is non-microphonic and therefore less likely to cause any undesired r.f. carrier noise than is a reactance-tube modulator.

MILLER-EFFECT FREQUENCY MODULATOR

As you learned in your study of video and audio amplifiers the apparent input capacitance of a voltage amplifier depends on the voltage gain of the stage; this is known as the "Miller Effect." Therefore, the input capacitance of such an amplifier can

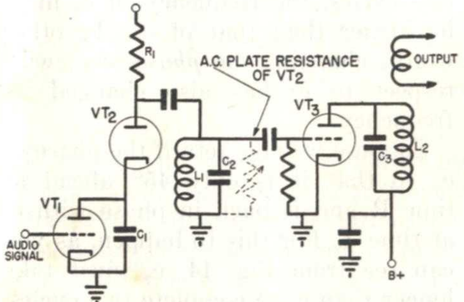


FIG. 11. A basic resistance tuning f.m. circuit. VT_3 is used in a tuned grid-tuned plate oscillator where L_1C_2 determines the frequency of operation. An audio signal applied to VT_1 causes a variation in the a.c. plate resistance of VT_2. This a.c. resistance, because C_1 is an r.f. bypass, is effectively in parallel with L_1C_2 and causes the frequency of oscillation to vary at the audio rate.

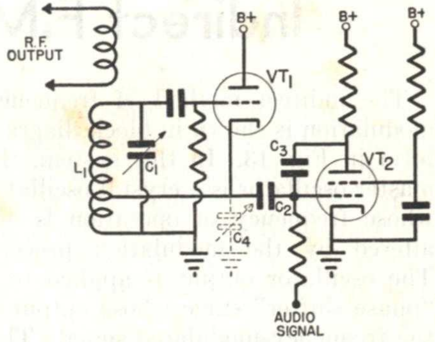


FIG. 12. The basic Miller effect frequency modulator. The oscillator is a conventional Hartley with L_1C_1 determining the frequency of operation. VT_2 is an r.f. amplifier whose gain can be varied by changing the d.c. voltage on the grid. This varies the apparent size of C_4 effectively connected across a portion of L_1 and thus varies its frequency.

be made to vary by varying the stage gain. This variation in capacitance can be used to produce frequency modulation.

The basic circuit using the "Miller Effect" to vary the frequency of an LC oscillator is shown in Fig. 12. The oscillator in this example is a conventional Hartley using VT_1. (Any other LC oscillator circuit could be used.) The input capacitance of VT_2 (which is equal to the plate-to-grid capacitance C_3 times one plus the gain of the stage) is effectively connected from the tap of coil L_1 to ground. (It is shown as C_4 in Fig. 12.) It will, therefore, with C_1, control the frequency of operation.

When VT_2 is a pentode the gain is approximately the transconductance of the tube times the plate load resistance. Since its transconductance varies with grid voltage the gain of the stage can be varied by applying an audio signal to the grid of VT_2. These variations in stage gain cause similar variations in the input capacitance of VT_2 and thereby produce frequency modulation of the oscillator.

Indirect F.M. Transmitters

The indirect method of frequency modulation is shown in block diagram form in Fig. 13. In this system, the master oscillator is a crystal oscillator whose frequency of operation is *not* altered by the modulation process. The oscillator output is applied to a "phase shifter" stage whose output is the frequency-modulated signal. This signal is then applied to frequency multipliers to obtain the desired carrier frequency and frequency deviation, next, to power amplifiers to obtain

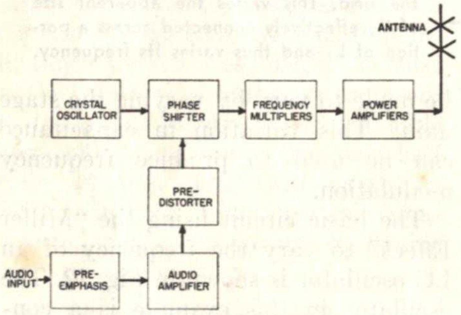


FIG. 13. This is a general block diagram of an indirect f.m. transmitter.

the desired power output, and finally, to the radiating antenna.

The audio signal is applied through a standard pre-emphasis network to the audio amplifier, as it is in a direct f.m. system. The audio output is then applied to a "pre-distorter" network that we will study shortly. Finally, the output of this network is applied to the phase-shifter circuit where its action changes the phase, and therefore the frequency, of the carrier in accordance with the amplitude of the audio signal. The output of the phase-shifter circuit as we have said, is the basic f.m. signal of the transmitter.

To understand the operation of this type of transmitter and the need for a pre-distorter, we must first study the relationship between frequency modulation (f.m.) and phase modulation (p.m.), and learn how p.m. can be used to obtain f.m.

PHASE MODULATION

Let us assume that we have a phase-shifting device (Fig. 14A) that can shift the phase of a sine-wave voltage so that the output (e_o) can vary from 90° ahead (+90°) to 90° behind (—90°) the input voltage e_i. Further, let us assume that e_i is a sine-wave voltage, as shown in Fig. 14B, and that at time M there is no phase shift in the phase shifter of Fig. 14A. Both e_i and e_o (shown in Fig. 14C) are at zero at that time. Now let us see what happens when we operate the phase shifter.

Suppose we first advance the phase of output voltage e_o in such a manner that it is 45° ahead of e_i at the end of one cycle (time N) and 90° ahead at the end of two cycles (time P). Since e_o has completed two cycles in less time than e_i required to complete two cycles, the frequency of e_o must be higher than that of e_i. In other words, changing the *phase* of e_i with respect to e_o has also changed its frequency.

Suppose we now retard the phase of e_o so that it is only 45° ahead at time R and is back in phase with e_i at time T. For this to happen, as you can see from Fig. 14, e_o must take longer than e_i to complete two cycles, and therefore must be lower in frequency. Again, a phase shift has produced a frequency change.

This simple example illustrates a fact you must remember to understand p.m. **Changing the phase of a sine wave also changes its frequency during the time that the phase is varying.** Thus, if we continually vary the phase of a sine wave, its frequency will vary continually too.

This is the basic principle used in all indirect f.m. transmitters. The modulating audio signal changes the phase of the carrier produced by the crystal oscillator and so creates a frequency-modulated carrier.

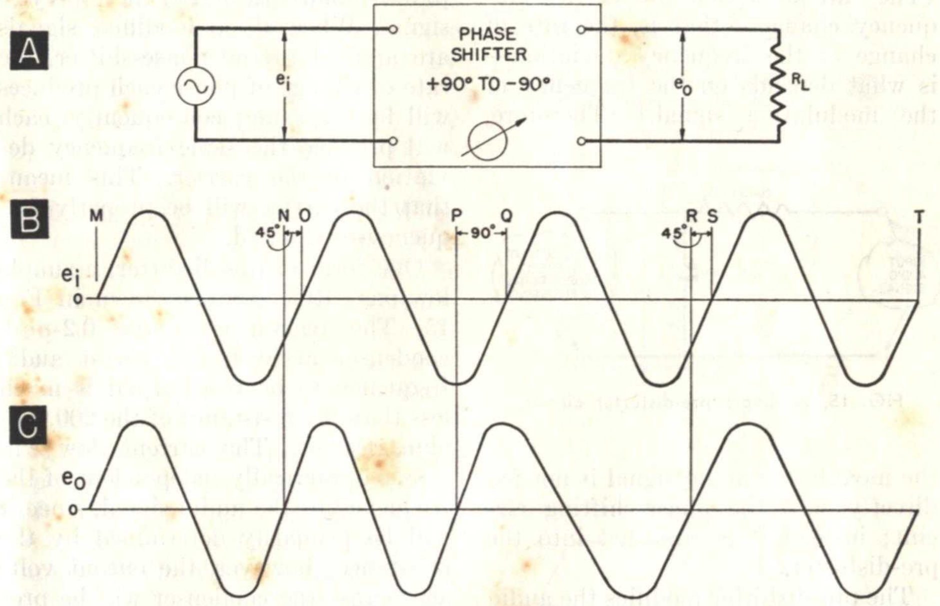


FIG. 14. A varying phase shift of a stable r.f. voltage e_i as shown in B, will produce an output e_o which is effectively frequency modulated. This is the basic principle used in all indirect f.m. transmitters to obtain f.m.

When a sine wave is modulated by this method, the amount of frequency change (the frequency deviation) depends upon both the amount of phase shift and the speed with which the shift occurs. In other words, the frequency deviation depends upon the *rate of change* of phase shift. If the rate of change is high, the frequency deviation is large; if the rate is low, the frequency deviation is small.

Unfortunately, any given rate of change of phase shift can be created in many ways. Say, for the sake of an example, that a phase shift of 10 degrees occurs in 1 second. (These figures bear no relationship to any practical case; we are using them merely to illustrate a point.) The rate of change is then 10/1, or 10. We can get the same rate of change if we have a phase shift of 20 degrees that occurs in 2 seconds, or a shift of 5 degrees that occurs in ½ second: in fact, there are an infinite number of combinations of shifts and speeds that will produce the same rate of change.

This means that different modulating audio signals can produce the same frequency deviation in the carrier signal. As you will learn in a few moments, the phase-shifting circuit is so arranged that the amplitude of the modulating signal determines the *amount* of phase shift in the carrier, and the frequency of the modulating signal determines the *speed* with which the phase shift occurs. Therefore, a modulating signal of small amplitude and high frequency will cause the same frequency deviation in the carrier as will a signal of lower frequency and greater amplitude.

Such a state of affairs cannot be permitted if we are to have frequency modulation, because the deviation of an f.m. signal must depend only on the amplitude of the modulation signal and not at all on its frequency. (The rate at which the carrier frequency changes—that is, the rate of change of the frequency deviation—is what depends on the frequency of the modulating signal.) Therefore,

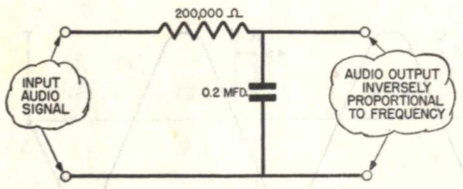


FIG. 15. A simple pre-distorter circuit.

the modulating audio signal is not fed directly into the phase-shifting circuit; instead it is first fed into the pre-distorter.

The pre-distorter modifies the audio signal, making its amplitude inversely proportional to its frequency. To see why this is done, let's suppose we have three audio signals of equal amplitude, one of 100 cycles, one of 1000 cycles, and one of 10,000 cycles. To produce an f.m. signal, each of these signals should cause the same frequency deviation in the carrier, since they are all equal in amplitude.

The signals will not produce the same deviation in the carrier if they are applied directly to the phase-shifter. All of them will produce the same phase shift in the carrier, but the rate of change of phase shift each will produce will depend upon the frequency, as you just learned. Therefore, the 10,000-cycle signal will produce the greatest deviation, the 100-cycle signal will produce the least.

If the signals are fed into the pre-distorter first, however, their amplitudes will be changed in inverse proportion to their frequencies: the amplitude of the 1000-cycle signal will be reduced to one-tenth that of the 100-cycle signal, and the amplitude of the 10,000-cycle will be reduced to one-hundredth that of the 100-cycle signal. When these modified signals are applied to the phase-shifter, the rate of change of phase each produces will be the same; consequently, each will produce the same frequency deviation in the carrier. This means that the carrier will be properly frequency-modulated.

One form of pre-distorter, a simple low-pass RC filter, is shown in Fig. 15. The reactance of the 0.2-mfd. condenser at even the lowest audio frequency to be transmitted is much less than the resistance of the 200,000-ohm resistor. The current flow will thus be practically independent of the frequency of the audio signal, since it will be primarily determined by the resistance; however, the output voltage across the condenser will be proportional to its reactance. Since the reactance is inversely proportional to frequency, the output voltage will also be inversely proportional to the frequency; in other words, it will be pre-distorted in the desired manner.

BASIC P.M. METHODS

Now that you've learned the general theory of how indirect f.m. transmitters work, let's see how an audio voltage can be used to change the phase of a carrier.

One method is shown in simplified form in Fig. 16A. The output of a stable r.f. source (generally a crystal oscillator) is applied to the LCR network shown. The internal resistance of the source (R_p) is much higher than the reactance of the LCR network. At the frequency of the r.f. source, the capacitive reactance of C_1 is one-half the inductive reactance of L_1.

This LCR network is the phase-shifting circuit. Its output differs in phase from its input by an angle that depends on the setting of the variable resistor R_1. The phase of the output can be shifted from 90° ahead of the input to 90° behind the input by varying R_1.

When R_1 is open, for example, only L_1 is effectively in the circuit; therefore, the output voltage (across L_1) *leads* the source voltage by 90°. When R_1 is a short circuit, the network is essentially capacitive (because the reactance of C_1 is less than the reactance of L_1) and the output *lags* the input by 90°. Intermediate values of phase shift are produced at intermediate settings of R_1.

A practical basic circuit using this type of phase shifter is shown in Fig. 16B. The r.f. signal is applied to the grid of VT_1, a conventional pentode amplifier. The variable resistance R_1 of Fig. 16A is replaced by the output circuit of a cathode follower stage using VT_2. An audio signal applied to the grid of VT_2 will change the value of the output impedance of the cathode follower* and thus the phase of the output r.f. voltage. When the input audio voltage to VT_2 is inversely proportional to the audio frequency, the output r.f. voltage will be frequency-modulated.

To prevent non-linear frequency deviation, the amount of phase deviation per phase-shifting stage is generally quite small. In the example above, the phase shift is generally limited to 25° in either direction, even

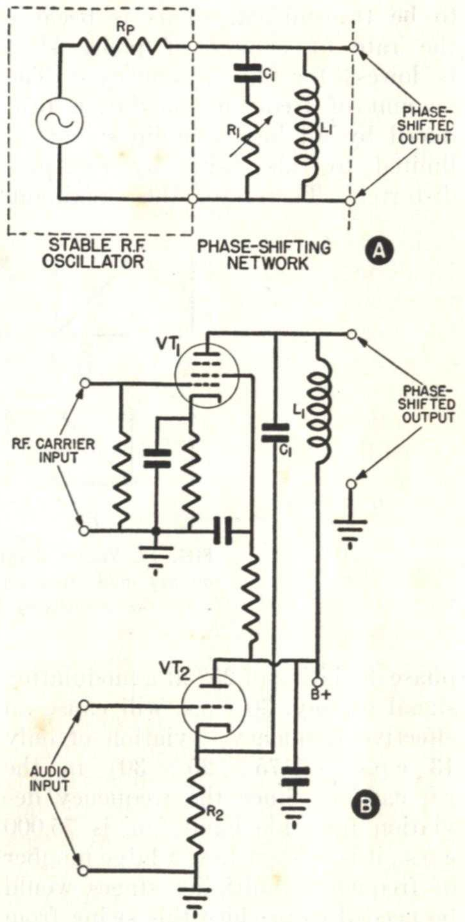


FIG. 16. A basic phase modulation method. In A is the fundamental circuit, in B a practical version where R_1 is replaced by the output impedance of a cathode follower VT_1.

though the circuit is capable of producing a 90° swing.

As you learned earlier, the amount of frequency deviation (usually abbreviated dF) produced by phase modulation of an r.f. carrier depends on the amount of phase shift in degrees (θ) and the frequency of the

*The output impedance of a cathode follower is essentially the cathode resistance (R_2) shunted by a second resistance whose value is equal to the reciprocal of the transconductance of the tube. The output impedance of the circuit is therefore varied when the audio signal varies the transconductance of the tube.

modulating signal (F_M). The exact relationship is given by the equation

$$dF = .0175\ \theta\ F_M$$

This equation shows that the smallest amount of frequency deviation is obtained at the lowest audio frequency to be transmitted. (This is because the rate of change of phase shift is lowest for this frequency.) The amount of frequency deviation produced by all higher audio signals is limited to this value by the predistorter. Therefore, the maximum phase deviation of 25° at a modulating signal of, say, 30 c.p.s. will cause an effective frequency deviation of only 13 c.p.s. (.0175 × 25 × 30) in the r.f. carrier. Since the frequency deviation for wide band f.m. is 75,000 c.p.s., it is evident that a large number of frequency multiplier stages would be needed to produce this swing from the original frequency deviation.

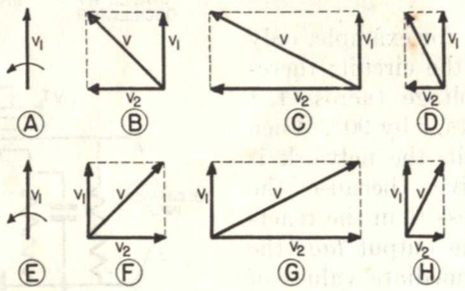


FIG. 17. Vector diagrams showing how a frequency modulated signal can be produced by the Armstrong phase-shift method.

Therefore, in a commercial transmitter using the above type of phase shifter, six stages are used in cascade so that the total phase deviation is 150°. This produces a 78-cycle frequency deviation for a 30-cycle signal. A series of frequency multipliers are used to multiply this deviation by 972, producing a total deviation that is over 75 kc.

Of course, the r.f. carrier is also multiplied by 972, so the crystal oscillator frequency must be the desired carrier frequency divided by 972—that is, about 100 kc.

There are other methods of producing frequency modulation by phase modulation. In one method, the variations in the speed of rotation of a plane of electrons in a special tube (a Phasitron) produce the desired phase modulation and subsequent frequency modulation. We will study this method in detail in a later Lesson.

Still another method of producing f.m. by phase modulation is used in the Armstrong system. Let's see how this system works.

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 of the same frequency but 90° 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 with the vector diagrams in Fig. 17.

Vector V_1 in Fig. 17A represents the original signal, having a definite frequency and a constant amplitude. (It is customary, in discussing such a vector, to assume that it is rotating counter-clockwise at a rate corresponding to its frequency. In our discussion, however, we shall consider that we are examining the vector at an instant of time so brief that the vector does not move.)

Let us assume that we suddenly introduce into the circuit a voltage V_2 that has the same frequency as V_1, but leads it by exactly 90°. At the instant this voltage is introduced, the vector diagram for the circuit will be as shown in Fig. 17B. Notice that the resultant vector V, which represents the combined voltage in the circuit, is to the left of vector V_1. In other words, the vector representing the circuit voltage has shifted simultaneously from the direction of vector V_2 to the direction of vector V. Such an instantaneous shift in the direction of a vector means that the quantity represented by the vector has momentarily changed in frequency; a shift to the left indicates a momentary increase, a shift to the right a momentary decrease. Fig. 17B, therefore, represents a momentary increase in the frequency of the circuit voltage.

Thus, introducing a voltage that leads the original voltage by 90° produces an increase in the frequency.

The amount of the frequency increase produced by introducing an out-of-phase voltage in this manner depends upon the relative amplitudes of the two signals. In Fig. 17B, V_2 is the same length as V_1 (in other words, the two voltages have equal amplitudes), so the angle between V_1 and resultant vector V is 45°. Fig. 17C shows what happens if V_2 is considerably larger than V_1. The angle between V_1 and V is now greater than 45°; this represents a greater momentary frequency increase than does Fig. 17B. Conversely, Fig. 17D shows that the angle between V_1 and V is less than 45° if V_1 is larger than V_2; under these conditions, the momentary frequency increase is less than in the other cases.

You can see what these facts lead to—if the amplitude of the out-of-phase signal is continually varied, the amount of frequency increase caused by injecting the out-of-phase signal will vary also.

So far, we have considered what happens when the injected signal leads the original signal by 90°. Now let's take a look at Figs. 17E through 17H, which show what happens when V_2 lags V_1 by 90°. As you probably expect, the results are the opposite of those that occur when V_2 leads V_1: injecting a lagging voltage causes a momentary frequency decrease in the circuit voltage (Fig. 17F) that is larger when V_2 is greater than V_1 (Fig. 17G) than it is when V_2 is smaller than V_1 (Fig. 17H). We can again draw the conclusion that a continual variation in the amplitude of the out-of-phase voltage V_2 will cause a continual variation in the amount the frequency of the circuit voltage changes (the frequency change this time being a decrease).

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

To make the deviation in frequency dependent only upon the *amplitude* and *polarity* of V_2, a pre-distorter circuit like the one you have already studied must be introduced into the audio amplifier of the transmitter to make the audio output voltage inversely proportional to the audio frequency. 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 have relatively low frequencies, and frequency multipliers are used to bring the f.m. signal up to the correct frequency for the transmitting antenna.

The basic Armstrong circuit for securing frequency modulation with this phase-shift method is shown in 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°; in other words, the plate loads for VT_2 and VT_3 are essentially pure inductances.

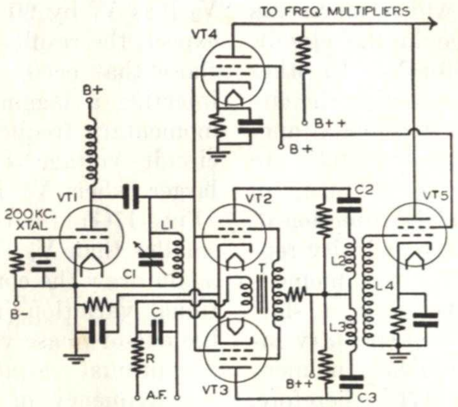


FIG. 18. Basic circuit used in the Armstrong phase-shift type of f.m. transmitter.

Fig. 18. Vacuum tube VT_1 in a crystal oscillator circuit produces the initial r.f. carrier (200 kc. is a typical value) with negligible frequency drift. The output of this crystal oscillator is fed into an r.f. amplifier tube VT_4 and also to the grids of the balanced r.f. stage containing tubes VT_2 and VT_3. Notice that the grids of this balanced stage are fed in *phase,* and that 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

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 varies correspondingly. We can change the a.c. plate resistance of a tube by varying the screen grid voltage.

In the circuit of Fig. 18, 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 makes the screen grid voltage on one tube increase and makes 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 r.f. voltage induced in secondary winding L_4 will either lead or lag the a.c. grid voltage by 90°, 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. 17. This voltage is amplified by amplifier stage VT_5.

The original carrier voltage, corresponding to vector V_1 in Fig. 17, is, as we said earlier, amplified by VT_4. The output of VT_4 (corresponding to V_1) and the output of VT_5 (corresponding to V_2) are fed in parallel into the frequency multiplier system. There they combine to give the desired frequency-modulated signal.

Propagation of F.M. Signals

We learned in our studies of standard a.m. broadcast transmission that the r.f. signals are propagated mostly by the ground wave because the antennas are generally only a fraction of a wavelength above the ground. A ground wave is attenuated as it travels over the earth, the amount of attenuation depending on the ground conductivity and the frequency of operation. Since attenuation increases as the frequency increases, a station on 600 kc. has a much greater field strength than a station on 1500 kc. for a given amount of power. As a matter of fact, impractically large amounts of power would be necessary for reliable ground wave communication at frequencies much above the broadcast band. At these frequencies, sky wave radiation reflected back to earth from the Heaviside layers must be used for communication.

DIRECT RADIATION

In the 88-108 mc. f.m. band, not only is propagation by ground wave impractical, but also there are no reflections of the sky wave. Propagation in this band must, as shown in Fig. 19, depend on direct radiation oı ground-reflected waves or a combination of both. Notice that the useful radiation is essentially in the horizontal direction.

To minimize ground losses, the transmitting and receiving antennas must be placed as high above ground

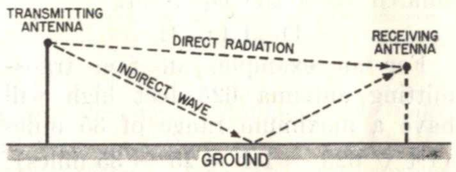


FIG. 19. In the v.h.f. band where f.m. stations are allocated, the useful radiation is free space radiation which travels either by a direct path or is ground-reflected in an indirect path. Note that to have this type of radiation, both the transmitting and receiving antennas should be as high above ground as is possible.

as possible. For this reason, all f.m. transmitting antennas are placed on high towers, buildings, hills, or even on mountain tops.

Reception of f.m. signals is possible somewhat beyond a direct line-of-sight path because the signals bend slightly in the same direction as the earth's curvature (see Fig. 20). This occurs because the density and moisture content of the air become less with altitude; this difference (although very small for antennas a few hundred feet high) causes refraction (bending) of v.h.f. signals. The result of this refraction is that the maximum range is, on the average, the same as if we calculated the distance to the optical horizon but assumed that the earth's radius were increased about 33 per cent.

Taking into account the curvature of the earth and the refraction just mentioned, we find that the maximum range in miles of an f.m. antenna

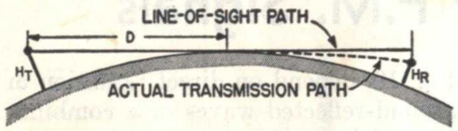


FIG. 20. V.H.F. signals are refracted slightly by the earth's atmosphere, so a receiving antenna slightly below the true line-of-sight path from a transmitting antenna may still receive direct radiation.

(distance D in Fig. 20) can be calculated from the equation:

$$D = 1.4\sqrt{H}$$

For an example, an f.m. transmitting antenna 625 feet high will have a maximum range of 35 miles ($1.4\sqrt{625} = 1.4 \times 25 = 35$ miles). This maximum distance assumes that the receiving antenna is at ground level; practically, of course, it will generally be above ground level, meaning that the maximum distance will be greater than indicated by the equation. For example, a receiving antenna 100 feet above ground would permit reception 14 miles beyond the 35-mile limit for a ground-level antenna, making a total of 49 miles. You can see from this how important it is to mount both transmitting and receiving antennas as high as possible.

The transmission paths at very-high frequencies are very stable as long as a straight line is possible between transmitter and receiver. There is then no fading and the polarization of the wave at the transmitter is maintained very accurately. However, when the transmission distance is so great that it is necessary to depend upon the refraction in the earth's atmosphere to obtain direct ray transmission to the receiver, fading frequently occurs as a result of variations in the earth's atmosphere that change the amount the wave is refracted. This becomes very severe when the transmission distance approaches or exceeds the maximum of direct ray transmission that we just calculated.

Polarization. It has been found that when both transmitting and receiving antennas are several wavelengths above the ground level, horizontally polarized transmission gives a higher signal-to-noise ratio than does vertical polarization. The reason is that man-made noise near the receiver is generally vertically polarized. Horizontal polarization is therefore used in f.m. broadcasting.

Horizontal Pattern. Because of the limited range and the necessity for mounting the antenna on a high point, most f.m. antennas are mounted on a tower or building in the center of the service area. The horizontal radiation pattern must be a circle to cover this area, so f.m. antennas are designed for such a pattern.

Frequency Modulation Antennas

You have already learned that the useful radiation from an f.m. transmitting antenna is in a horizontal direction. The energy that is generally radiated in the sky wave can, if directed downward, be used to increase the radiation in a horizontal direction. This is done by using two or more antenna radiating elements (called "bays") arranged vertically and fed power in a manner that causes a cancellation of the skyward radiation. The manner in which extra coverage is obtained by suppressing the sky wave is illustrated in Fig. 21. Fig. 21A shows the radiation pattern, in the vertical plane, from a single-element f.m. Pylon antenna. (The Pylon antenna will be studied shortly.) As the arrows indicate, considerable power is radiated toward the sky, where it is lost.

Fig. 21B is an approximation of the vertical radiation patterns from an antenna consisting of two Pylon elements stacked vertically. In this case, skyward radiation is greatly reduced and the radiation at low angles and along the horizon is greatly increased. Thus, the field strength and effective power are much greater, even though the total power radiated is the same.

In Fig. 21C is shown the vertical pattern for a four-element Pylon antenna. The pattern is still further squashed down and the radiation in the horizontal direction increased. Still further gains can be obtained by using more elements. However, in most types of f.m. antennas, the amount of gain per added element decreases so that the number of bays used is usually limited to four.

This increase in effective power does not affect the horizontal radiation pattern of the antenna; it remains circular.

FIELD GAIN AND POWER GAIN

In comparing the advantages of various types of multi-element antennas, engineers use the terms field gain and power gain.

The *field* gain is the ratio of the field intensity (in volts) produced by

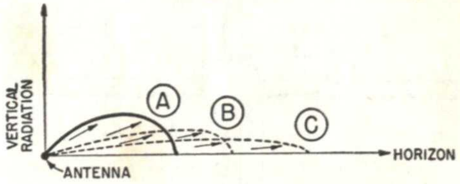


FIG. 21. By adding more elements to an f.m. antenna, the energy radiated in a vertical direction and thence lost can be concentrated toward the horizon where it is more useful. This results in a gain of field strength at the receiver as if effectively the power output of the transmitter were increased. A shows the vertical radiation from a single element; B, a two element, and C, a four element pylon array. Note that four elements double the field strength which is the same as a power gain of four. (The skyward lobes of curves B and C, because they are small, are not shown here.)

a multi-element antenna to the field intensity produced by a vertical half-wave antenna fed the same power. For example, the field intensity produced by a six-bay turnstile antenna (which we will study shortly) is 2.07 times as great as that produced by a half-wave vertical antenna; the field gain of the former is therefore 2.07.

The *power* gain of an f.m. antenna is defined as the ratio of the power in a half-wave antenna to the power in the multi-element antenna that will produce radiated signals of the same field intensity.

Since power is proportional to the square of voltage, the power gain of an f.m. antenna is equal to the *square* of the field gain. Thus the antenna with a field gain of 2.07 will have a power gain of 4.28. A power gain of 6 in an antenna produces a field gain of 2.45.

In studying propagation characteristics, engineers find field gain the easier to use in calculations.

Notice that the power gain of all

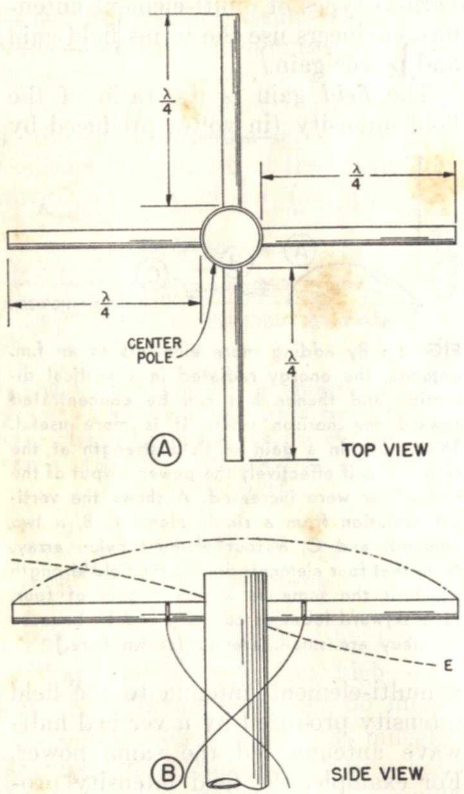


FIG. 22. The basic turnstile array consists of two shunt-fed half-wave dipoles at right angles to each other and fed 90° out of phase. In A is shown a top view, and in B a side view of this type of antenna. One dipole has been left out at B to simplify the drawing.

multi-element antennas is compared to that of a half-wave vertical antenna and not to that of a single-bay antenna of the type specified. Single-bay antennas of the types used for f.m. broadcasting generally have a power gain of less than one, that is, they produce a smaller field intensity than does a half-wave antenna. For instance, a single-bay turnstile has a gain of .5. This means that a six-bay turnstile with a power gain of 4.3 over a vertical half-wave has a power gain of 8.6 over a single-bay turnstile.

TRANSMITTING ANTENNAS

You have just learned that an f.m. antenna should, first, have a very nearly circular horizontal radiation pattern, second, concentrate as much of the radiated energy as possible in a horizontal direction (have a good power gain), and, third, be horizontally polarized. Let us now study the types of antennas meeting these requirements and in general use in f.m. broadcast stations.

The types of f.m. transmitting antennas in general use at present are: (1), the turnstile antenna and versions of it; (2), various forms of circular or ring antennas; (3), forms of square loop antennas; and (4), the Pylon antenna.

THE TURNSTILE ANTENNAS

The basic turnstile antenna consists of four horizontal quarter-wave elements arranged as shown in Fig. 22 to form two half-wave antennas at right angles to each other. Fig. 22B shows the current wave and the voltage wave on each of the dipoles. The center of the dipole is at zero voltage with respect to ground, so the radiating rods can be fastened directly to the grounded pole at this point. The r.f. power is shunt-fed to each of the dipoles by either open-wire lines or coaxial lines connected as shown in Fig. 22B. The feed lines are spaced at the proper distance from the pole to provide an impedance match from line to antenna.

The horizontal field of a single one of these dipole antennas is shown in Fig. 23A. Obviously, such a pattern is not satisfactory when uniform transmission in all directions is desired. This is the reason for the second set of rods in each bay. When this second set is fed an equal amount of power that is 90° out of phase with that fed to the first set, the patterns of the two radiating dipoles are those shown by the oval lines of Fig. 23B, and the combined field is the heavy line. The latter, as you can see, is very nearly a circle.

To achieve high horizontal gain,

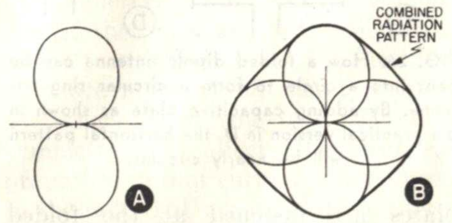


FIG. 23. A, the horizontal radiation pattern of a single dipole antenna. B, how the radiation pattern of two dipole antennas at right angles to each other and fed 90° out of phase form practically a circle.

several bays of dipoles must be used, and the dipoles in each bay must be fed in phase with the power applied to each of the other bays. If the bays are spaced a half-wave apart, this can be conveniently done with a transmission line that is crossed over between each bay, thereby counterbalancing the phase shift that occurs along this line between bays. Two such lines, one for each set of dipoles, run up the tower, twisting around it as they go. At the base of the tower the two lines are fed currents 90° out of phase by any of several methods.

The Super-Turnstile. To broaden the frequency response of a turnstile antenna, the four straight radiating elements of Fig. 22 are each replaced in the super-turnstile antenna with the radiator shown in Fig. 24. This type of radiating element is simply a framework made of tubing (a construction that is used to lower wind resistance); *electrically,* however, it acts as if it were a flat, solid sheet of metal. These radiators are called "current sheets." Notice that each element is fed at the center and grounded to the center supporting pole at each end.

This antenna is fed in the same manner as other turnstile antennas, and when more than one bay is used, as shown in Fig. 25, there is a half-wave spacing between centers of the bays.

CIRCULAR OR RING ANTENNAS

The circular or ring antenna is essentially a folded dipole antenna bent around into a circle. In Fig. 26A, the folded dipole is shown in its simplest form. Essentially it consists of two half-wave radiators, one of which

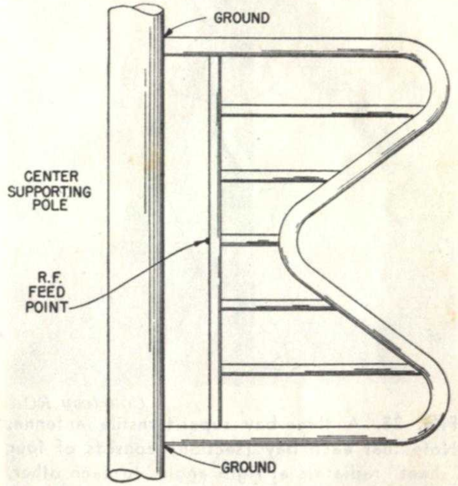
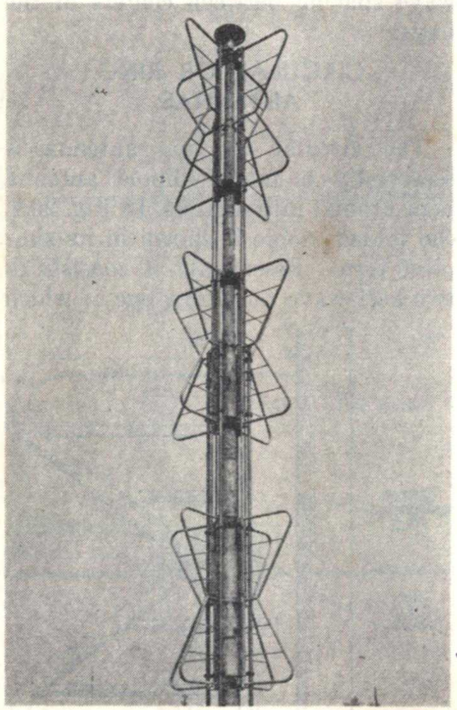


FIG. 24. A side view of a single "current sheet" radiator of a super-turnstile antenna. Although made of tubing to reduce wind resistance, it acts as if it were a solid sheet of metal and thus has a broad frequency response characteristic. Four of these elements arranged and operated as a turnstile form one bay.

is broken at the center; the system is fed at this point. Since the two radiators are mounted very close together, the currents in them flow in the same direction, and the current distribution on both is a sine wave. Since the voltage to ground at the center is zero, the unbroken radiator can be attached directly to the supporting pole at this point. The ends of the lower radiator can be fed power by an open balanced line or by a pair of concentric lines oppositely phased.

The radiation characteristics of this straight dipole are the same as those



Courtesy RCA

FIG. 25. A three-bay super-turnstile antenna. Note that each bay (section) consists of four "sheet" radiators at right angles to each other, and that the bays are separated vertically by one half wavelength.

of one pair of radiators on the turnstile, that is, the horizontal pattern is a figure 8, which is undesirable for broadcast purposes. To approach uniformity of transmission in all directions, the dipole is bent into a circle as shown in Fig. 26B. This, however, will not of itself give a circular pattern, because the current distribution is not uniform around the radiator. To improve it, a pair of large metal

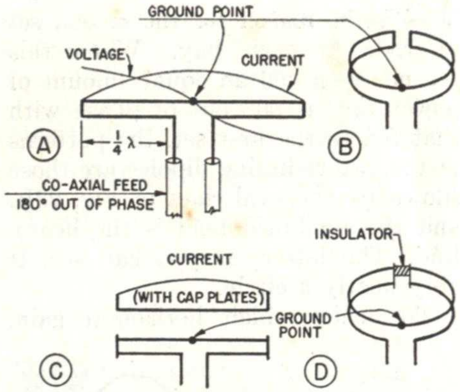


FIG. 26. How a folded dipole antenna can be bent into a circle to form a circular ring antenna. By adding capacitive plate as shown in the practical version in D, the horizontal pattern will be nearly circular.

plates are fastened at the folded points. These plates have the effect of adding end capacity to the radiator, changing the current distribution to that shown in Fig. 26C. The current is now approximately uniform around the loop, and the signal radiated approaches a circular pattern to the same degree.

The circular antenna, Fig. 26D, presents a neat appearance and has a higher gain per bay than does the turnstile. However, to keep down the mutual impedance, the bays must be placed a full wavelength apart. Thus

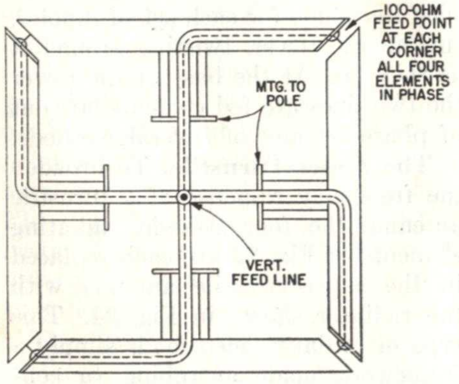


FIG. 27. Top view of a square loop f.m. antenna which is frequently used where mounting on an a.m. tower is desired.

the *gain per height* of a circular loop antenna is less than that of a turnstile when more than one bay is used. For instance, a three-bay circular antenna that is two wavelengths high, 20 feet at 90 mc., has a power gain of 2.6, whereas a five-bay turnstile having the same height has a power gain of 3.5.

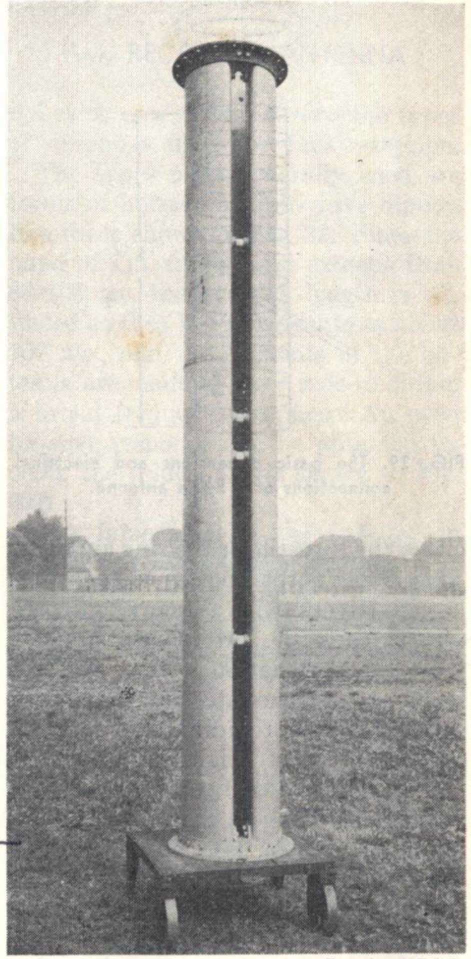
The off-center mounting of the rings is also a disadvantage in that it makes for mechanical and electrical dissymmetry. Thus, while the loops are of the same approximate weight as the turnstile elements, the fact that they are mounted off-center requires a stronger supporting pole. The dissymmetry also affects the electrical properties in that currents are induced in the pole that are opposite in phase to those in the radiators. Because of these mechanical and electrical difficulties, it is believed impractical to go beyond three or four bays in this type of antenna.

SQUARE-LOOP ANTENNAS

The antennas previously described are all mounted on supporting poles of the flagpole variety. Where such a pole can be mounted on an existing structure or where the ground height is in itself sufficient, one of these standard types of antenna should definitely be used. Some difficulties sometimes arise when it is desired to mount an f.m. antenna on an existing a.m. tower. Most such towers were not built for and will not support the heavy pole used with multi-element turnstiles or ring antennas. In such cases, several variations of what may be called, for want of a better term, a square-loop antenna, have been used.

The square-loop antenna consists of four dipole radiators arranged in the form of a square that may or may not be closed at the corners. They may take the form of folded dipoles or of simple dipoles fed at the center, according to the method used to obtain impedance matching. Several antennas of this type have been designed and are now in operation.

One form of square-loop antenna is



Courtesy RCA

FIG. 28. The basic radiating element of a Pylon antenna.

shown in Fig. 27. It is made up of four hollow half-wavelength radiators of rectangular cross-section. Each of these is supported at its center by two pieces of tubing that secure it to the mounting pole or tower. Each element, as shown in Fig. 27, is end-fed by a coaxial line that enters the center of the next element and runs through it to the end. All four elements are fed in phase.

The *gain per bay* of the square-loop antenna is greater than that of

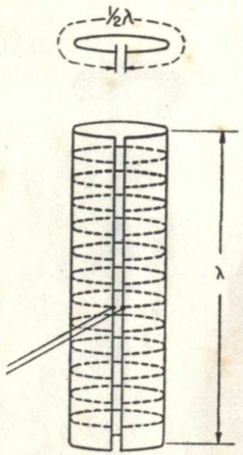


FIG. 29. The basic dimensions and electrical connections of a Pylon antenna.

either the turnstile or the ring antenna. This is not surprising, since each bay has effectively twice as many radiators as the turnstile. Moreover, because the vertical radiation is very low, the bays can be mounted at half-wave intervals, thus providing high gain for a given antenna height.

The three disadvantages of the square-loop antenna are, first, such an antenna must be laid out and especially built for each installation, because each one will be slightly different in arrangement and mounting details; second, the tuning is quite critical and must be done with the radiators in place; and third, it is very difficult to design such an antenna to withstand a heavy ice load.

THE PYLON ANTENNA

An f.m. antenna unusual in mechanical construction is the RCA Pylon antenna shown in Fig. 28. A Pylon section consists of a cylinder about 13 feet high and 19 inches in diameter that has a vertical slot in one side. The main advantage of this type of antenna is that the mechanical mounting is quite simple and the antenna is self-supporting. Sections can be added by bolting them to the top of the basic section.

As shown in Fig. 29, a Pylon section is approximately one wavelength long and about a half-wavelength in circumference. If it is fed at the center of the slot, the cylinder acts as if it were a large number of parallel circular half-wave radiators. The slot serves as a transmission line to feed all parts of the section.

The horizontal directivity pattern

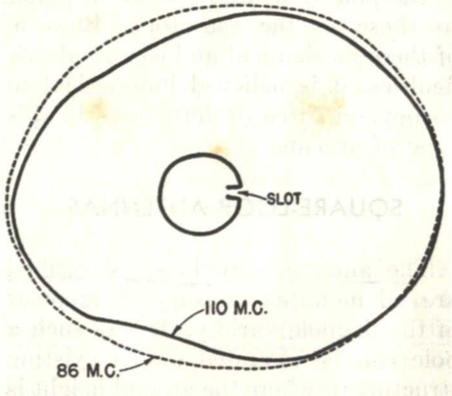


FIG. 30. The horizontal directivity pattern of a Pylon antenna. Two patterns are given, one for 86 mc. and another 110 mc., both beyond the limits of the f.m. band.

of a Pylon antenna, shown in Fig. 30, is not exactly a circle and depends on the frequency of operation. The two patterns show the horizontal directivity at 86 mc. and 110 mc., that is, beyond the limits of the f.m. band. At all f.m. frequencies, however, the pattern does not differ more than 2.5 db from a perfect circle.

Multiple Pylon arrays generally consist of two, four, or eight sections stacked vertically. The method of

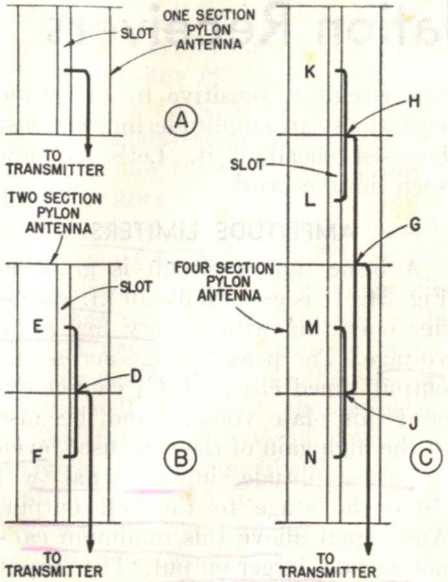


FIG. 31. How r.f. energy is fed to a Pylon antenna. In A we see that the transmission line terminates at the center of the section. In B and C the sections are each fed in the center by transmission lines that are an equal length from the transmitter. This provides equal power distribution to each of the sections.

connecting the transmission lines for one-, two-, and four-section arrays is shown in Figs. 31A, 31B, and 31C, respectively. In the two-section arrangement the transmitter power is fed first to point D then to E and F, the midpoint of the two sections. In the four-section array, the power goes to G, then to H and J, and finally to K, L, M, and N, the centers of the four sections. This method of connection makes the length of transmission lines to each radiator the same so that equal power is fed to each. Transmission lines themselves are mounted inside the antenna section near the slot.

The power gain of a Pylon antenna as compared to a half-wave dipole is 1.5 per section; the gain of a four-section array, for example, is 6. This power gain of 1.5 per wavelength is a greater gain than is obtained in the other types of f.m. antennas we have studied.

F.M. RECEIVING ANTENNA

Let us now briefly review the types of antennas used for f.m. reception. The types most generally used are forms of horizontal half-wave dipoles like those shown in Fig. 32. Since the band of f.m. frequencies extends from 88-108 mc. the antenna length is adjusted so that it will resonate at about 100 mc., and the elements of the antenna are made of large rods to obtain a broad frequency response. An even broader response can be obtained by using the folded dipole shown in Fig. 32B.

The impedance of the antenna in Fig. 32A is low—about 72 ohms. The impedance of the folded dipole is about 300 ohms, so it can be connected directly to a standard 300-ohm transmission line.

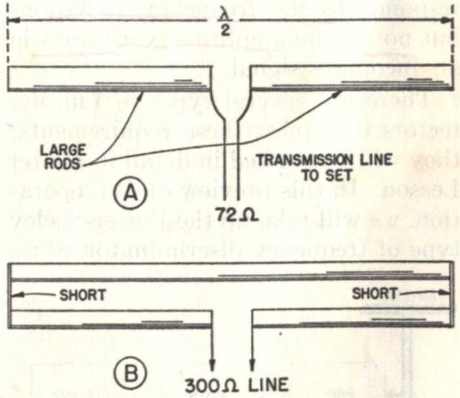


FIG. 32. The basic f.m. receiving antenna is a half wave dipole. A conventional dipole is shown in A, and a folded dipole in B.

Frequency Modulation Receivers

Let us now briefly preview the essential characteristics of f.m. receivers. We will study them in detail in a later Lesson.

The superheterodyne circuit shown in the block diagram in Fig. 33 is generally used in f.m. receivers. The preselector, local oscillator, and mixer-first detector differ from the corresponding circuits in a.m. receivers only in that they must be built to operate at the very-high frequencies used in f.m. The i.f. amplifiers generally operate on 10.7 mc. in order to pass the wide band (150 kc.) of an f.m. signal. The audio amplifier in a good f.m. receiver is a high-fidelity amplifier capable of reproducing the full audio range with low distortion.

It is in the second detector and associated circuits that an f.m. receiver differs radically from an a.m. receiver. Remember that the advantages of static-less reception and a minimum amount of interstation interference are obtained *only* when the second detector in an f.m. receiver responds to the frequency variations but not to the amplitude variations in an incoming signal.

There are several types of f.m. detectors that meet these requirements; they will be studied in detail in a later Lesson. In this preview of f.m. operation, we will take up the Foster-Seeley type of frequency discriminator. Since this circuit is sensitive to amplitude variations, an amplitude limiter must be used ahead of it. Let's see how such limiters work.

AMPLITUDE LIMITERS

A basic limiter circuit is given in Fig. 34. It is essentially an i.f. amplifier operating with a very low plate voltage. The peak voltage across the output tuned circuit L_3C_3 cannot exceed this plate voltage, and, because of the high gain of the tube used, even a low-amplitude input signal will drive the stage to the full output. Any signal above this minimum cannot cause a larger output. The output is thus effectively limited, since variations in amplitude of the input do not produce any variation in the amplitude of the output. The grid network RC is included in the circuit primarily so that when the grid goes positive on large amplitude signals, the grid current flow produces a negative bias that further aids the limiter action. The network also provides an a.v.c. voltage for the receiver.

Sometimes two limiter stages are connected in cascade so that any small amplitude variations that pass the first stage will be removed by the second limiter stage.

F.M. DETECTORS

The d.c. output of the detector in an f.m. receiver is proportional at all times to frequency variations of the i.f. signal applied to it. Thus, when an f.m. signal is received, the output of the detector in the receiver is an a.c. voltage that corresponds to the original modulating signal.

Let us now study different types of f.m. detectors.

Off-Resonance F.M. Detectors. If a frequency-modulated r.f. signal is introduced into an L-C resonant circuit that 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. 35A 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. 35B. Let's see what happens if we apply to this curve the f.m. signal shown at the bottom of Fig. 2 (having a square-wave audio modulation).

First, assume that when the f.m. transmitter is at its resting frequency (time interval 1 in Figs. 35B and 35C), the operating point is at 1 in Fig. 35B. The vertical distance from 1 down to the horizontal reference line then determines the amplitude of r.f. voltage V_c across the resonant circuit, so we show the r.f. output voltage of the L-C circuit for time interval 1 as an r.f. signal having this same amplitude and the same resting frequency value. This is shown at 1 in Fig. 35D.

When the frequency of the f.m. signal increases to the value for time interval 2 in Fig. 35C we move up to point 2 on the response curve, and

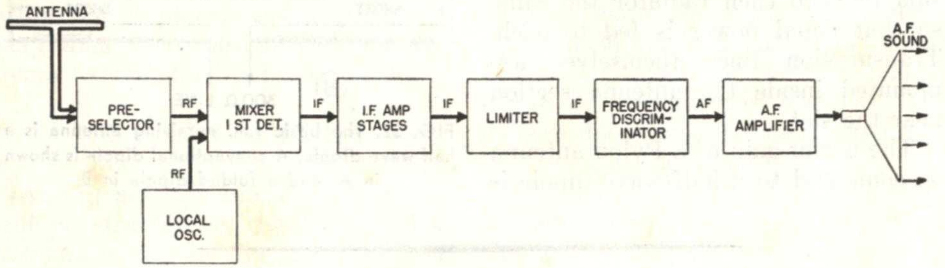


FIG. 33. The block diagram of an f.m. receiver.

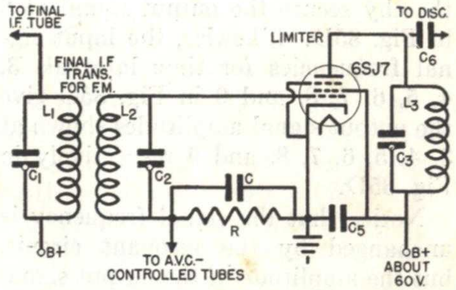


FIG. 34. Simplified one-tube limiter circuit of an f.m. receiver.

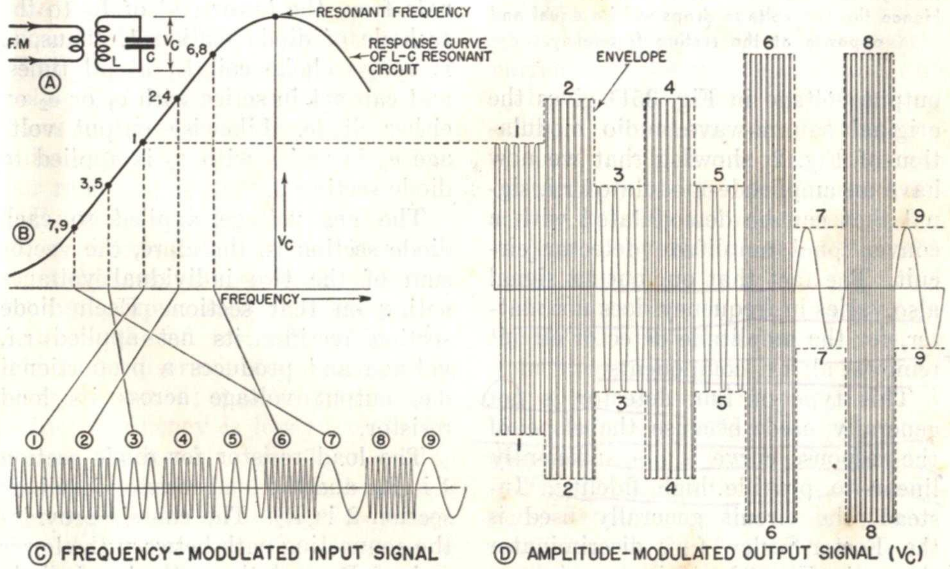


FIG. 35. These diagrams illustrate how a simple resonant circuit can convert an f.m. signal into an a.m. signal.

thereby secure the output signal at 2 in Fig. 35D. Likewise, the input signal frequencies for time intervals 3, 4, 5, 6, 7, 8, and 9 in Fig. 35C give the output signal amplitudes shown at 3, 4, 5, 6, 7, 8, and 9 respectively in Fig. 35D.

Notice that the signal frequency is unchanged by the resonant circuit, but the amplitude of the output signal of the circuit varies in proportion to the frequency deviations of the input f.m. signal. A dashed line drawn through the positive peaks of the r.f.

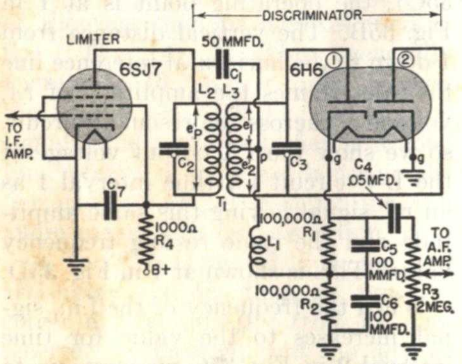


FIG. 36. Typical discriminator circuit for an f.m. receiver. Electrons flow from L_1 upward through R_1, but flow downward through R_2. Hence the two voltage drops will be equal and opposite at the resting frequency.

output voltage in Fig. 35D gives the original square-wave audio modulation of Fig. 2, showing that we now have an amplitude-modulated r.f. signal that can be demodulated with a conventional amplitude 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.

This type of f.m. detector is not generally used, because the slope of the response curve is not sufficiently linear to provide high fidelity. Instead, the circuit generally used is the Foster-Seeley f.m. discriminator shown in Fig. 36. It is very important to understand the basic operation of this circuit, because variations of it are used not only for detection in f.m. receivers but also in the center-frequency control circuits of f.m. transmitters. We will discuss it briefly here and in more detail in later Lessons.

F.M. Discriminator. The primary winding L_2 of the discriminator transformer T_1 in Fig. 36 is tuned to the 10.7-mc. i.f. value by C_2, and the center-tapped secondary winding L_3 is tuned to the same i.f. value by C_3. The limiter output current flowing through L_2 induces a corresponding f.m. voltage in L_3, and 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 180° out of phase with each other.

The limiter output voltage e_p in series with e_1 is applied to diode section 1 of 6H6 double-diode tube (in some receivers, a 6AL5 tube is used instead). D.C. blocking condenser C_1 provides an r.f. path from the upper end of L_2 to p, and the chassis and r.f. by-pass condensers C_7, C_6, and C_5 complete the path for r.f. signals from the lower end of L_2 to the cathode of diode section 1. Thus, e_p is across choke coil L_1 at all times, and can act in series with e_1 or e_2 on either diode. Likewise 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 see by tracing the diode circuits. This means that the combined voltage across both R_1 and R_2, which is the output voltage of the discriminator, will at each instant be the *difference* between the individual voltages. If the individual voltages are equal, the discriminator output voltage will be zero; if the voltage across R_1 and R_2 are different, the combined voltages will have the polarity of the larger of the two individual voltages, and will be equal in magnitude to their numerical difference.

voltage e_p serving as the reference vector in each case.

The r.f. voltage e_s induced in secondary winding L_3 is 180° out of phase with the primary r.f. voltage e_p, so e_s is shown 180° out of phase with reference vector e_p in each of the vector diagrams in Fig. 37.

When the limiter output signal is exactly at the i.f. resting value of 10.7 mc. to which the discriminator circuits are tuned (that is, when no modulation is being transmitted), secondary tuned circuit L_3-C_3 is at resonance, and the secondary current i_s flowing through L_3 will be in phase

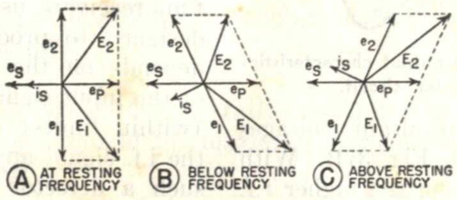


FIG. 37. These vector diagrams will help you to understand the action of the discriminator in an f.m. receiver.

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

Phase relationships 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. resting frequency; and 3, when the limiter output signal 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. 37, with primary

with e_s, as indicated in Fig. 37A. The voltage produced across the entire secondary winding by this secondary current will therefore lead both i_s and e_s by 90°. However, this secondary voltage is made up of two voltages, e_1 and e_2, which are referred to the center tap p of L_3; if we show e_1 leading i_s by 90°, we must show e_2 as lagging e_1 by 180°, just as Fig. 37A does.

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. 37A shows that these two voltages are equal for the no-modulation condition, so 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. 37B. Since voltages e_1 and e_2 must be 90° out of phase with i_s, we

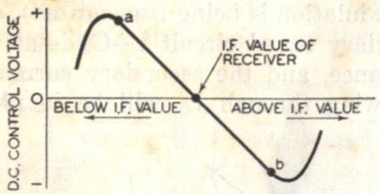


FIG. 38. S-curve showing output characteristics of a discriminator circuit.

have the unequal resultant voltages E_2 and E_1 shown in Fig. 37B. 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. 37C for the condition in which 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. 36 thus produces at its output a d.c. voltage whose amplitude, as shown in Fig. 38, is at each instant proportional to the deviation of the incoming signal frequency from its resting value. The polarity of this voltage is determined by the direction in which the frequency deviation occurs. The output can be made very linear over most of the curve (from a to b in Fig. 38). The discriminator thus converts an f.m. signal directly into the original audio signal voltage used to modulate the f.m. transmitter.

Other F.M. Detectors. As you will notice in another Lesson when we study f.m. receivers in detail, some f.m. receivers use detectors that are designed to produce an output that depends on the frequency deviation of the input signal but is independent (within limits) of the amplitude of the i.f. signal applied to them. When such a detector is used, the limiter stages preceding the f.m. detector are no longer needed. One such circuit, which is similar to the discriminator of Fig. 36, is called a "ratio" detector because the a.c. output is proportional to the *ratio* of the vector voltages applied to the two diode plates and hence independent of any amplitude variation of them.

In still other f.m. detectors designed to eliminate the need for amplitude limiter stages an oscillator that is a part of the detector circuit is frequency-modulated in accordance with the received signal; this action is used to produce an output proportional to frequency deviation but independent of the amplitude of the f.m. signal.

Lesson Questions

Be sure to number your answer sheet 46RC.

Place your Student Number on *every* Answer Sheet.

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

1. What two factors in high fidelity programming are more readily attained with f.m. than a.m.?

2. To secure a high signal-to-noise ratio with minimum interstation interference at receivers, should the maximum frequency deviation in an f.m. system be high or low?

3. If the frequency of the audio signal fed to an f.m. transmitter decreases, but its amplitude remains the same, does the frequency deviation *increase*, *decrease*, or *remain the same?*

4. In wideband f.m. broadcasting, what must be the ratio of the amplitude of the desired signal to the undesired signal in order for the interference between the two signals to be negligible?

5. Why is a pre-distorter used in indirect f.m. transmitters?

6. How many microseconds is the time constant of a standard pre-emphasis network?

7. If the maximum phase deviation in an indirect f.m. transmitter is 60° and the audio modulating frequency is 30 cycles per second, what is the frequency deviation produced?

8. Why are f.m. broadcast transmissions horizontally polarized?

9. Assuming that the receiving antenna is at ground level, what is the theoretical maximum range of an f.m. transmitter whose antenna is 400 feet above ground level?

10. What is the purpose of a limiter stage in an f.m. receiver using a Foster-Seeley discriminator?

ALL MEN WANT TO SUCCEED

Here's a quotation I ran across the other day that made me think of several fellows I know:

"All men want to succeed. A few men want success so badly that they are willing to work for it."

Isn't it true that almost every fellow you know *wants* success, *wants* more money, *wants* security?

But how many of these men are willing to buckle down and study — work — think — to get the good things they want?

It is very true that only *comparatively few men* are willing to work hard for success.

You are one of those few men. You have proved this fact by enrolling for the NRI course—by working to complete many of your Lessons. *You* are taking the first and most important step toward success.

J. E. Smith