frequency modulation

K. R. STURLEY

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The last word in radio—frequency modulation—is discussed in this concise book for the benefit of radio technicians.

The principles of frequency and phase modulated systems are outlined. The advantages and disadvantages of frequency and phase modulated transmission over amplitude modulation, the modulating methods and the various details of the frequency-modulation receiver are discussed in simple and clear terms.

A chapter is devoted to the conversion of the frequency into an amplitude change of carrier. There is an adequate treatment of the problems encountered in the design of frequency-modulation receivers. A detailed bibliography, many diagrams and a chart showing a typical frequency-modulation receiver circuit are additional features of the book.

FREQUENCY MODULATION

by

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CONTENTS

PREFACE TO THE FIRST AMERICAN EDITION

The purpose of this monograph is to present a brief review of the theoretical aspects of frequency-modulated radio waves and a concise and continuous discussion of the basic problems encountered in the design of frequency-modulation receivers. The presentation is made at the professional level. While it is assumed that the reader possesses considerable technical maturity, occasionally he may desire additional information, particularly in that part of the text which is devoted to the review. To meet such requirements, a complete bibliography is included and the text contains references to it. Much of the material listed in the bibliography is American; the rest is British literature readily available in the United States.

The text of this monograph was prepared in England and, therefore, the vacuum tubes mentioned are British. Generally, each mention of a tube includes also the operating requirements in the application discussed, so that intelligent substitution is readily possible, should the reader wish to apply the design procedure in any market.

A list of American tubes, approximately equivalent to the ones mentioned in the text, is included at the end of the last chapter. This list was furnished by Mr. C. R. Potter of the Radio Corporation of America. The same list is included in the chart of the typical frequency-modulation receiver circuit at the back of the book.

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Chapter 1

GENERAL NATURE OF THE SYSTEM

THE direct transmission and reception of speech or music over long distances, although not impossible, is impracticable, and propagation of audiofrequencies is usually accomplished by using them to modulate a radio-frequency wave acting as a carrier, *i.e.*, the audiofrequencies are used to control one of the three characteristics, amplitude, frequency, or phase, of the carrier. The most common method is by modulation of the carrier amplitude. The rate at which the amplitude is changed is directly proportional to the frequency of the original sound and the magnitude of the amplitude change is directly proportional to the intensity, a low intensity producing a small change of amplitude. This is illustrated in figures 1a and 1b, the unmodulated carrier amplitude in each case being constant at 1 volt. Figure 1a corresponds to a low-intensity sound, the carrier amplitude varying between 0.9 and 1.1 volts, whereas figure 1b corresponds to a high-intensity sound, the carrier being 100 per cent modulated, its amplitude varying from 0 to 2 volts.

In frequency modulation, the carrier amplitude remains constant and its frequency is varied at a rate corresponding to the modulation frequency (1,000 times per second if $f_{mod} = 1{,}000$ c/s). The frequency deviation — rise and fall from the central unmodulated carrier frequency — is controlled by the intensity of the audiofrequency. For example, if a 1,000 c/s note is being transmitted on a carrier frequency of unmodulated value, 1,000 kc/s, then the variation of the carrier frequency takes place at the rate of 1,000 changes per second and the frequency limits may be $\pm$ 100 c/s (the carrier frequency changing between 1,000.1 and 999.9 kc/s) for a low-intensity rate to ±100,000 c/s for a high-intensity note. These two conditions are illustrated in figures 2a and 2b for a half cycle of the modulation frequency. With phase modulation, the amplitude of the carrier remains constant and its phase angle with respect to its unmodulated condition is advanced and retarded at the frequency of the audiosignal. The magnitude of the phase change is determined by the intensity of the audiofrequency. Phase modulation has an effect on the carrier similar to that of frequency modulation, and while the phase of the carrier is varying, its frequency is also varying. There is a difference between the two, but this will be discussed later.

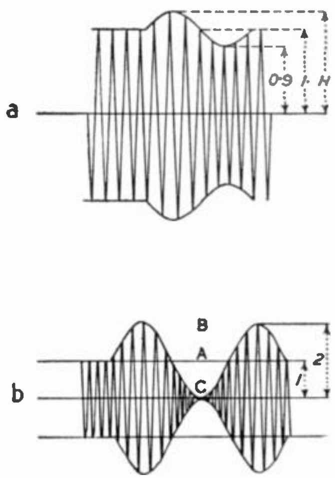

Fig. 1. *Amplitude-modulated wave*

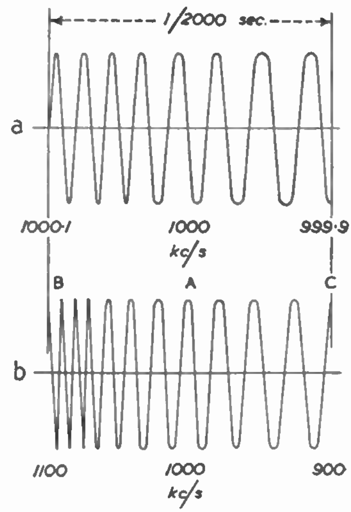

Fig. 2. *A half-modulation cycle of frequency-modulated wave*

Until recently, neither frequency nor phase modulation has been used to any great extent, largely because more complicated apparatus is required for reception and transmission than for amplitude modulation and they appeared to offer few advantages.

Armstrong[13] has, however, demonstrated that frequency modulation can, under certain conditions, give better fidelity, greatly improved signal-to-noise ratio* and larger broadcast service area than amplitude modulation and it seems probable that frequency modulation may be considerably developed in the future.

The Principles Involved in Frequency-Modulated and Phase-Modulated Transmission[5,39]

For understanding the operation of any type of modulation, a vector diagram is most useful, and amplitude modulation is quite simply illustrated if the modulation envelope is sinusoidal. The amplitude-modulated signal is represented mathematically by

$$\hat{E} \sin 2\pi f_c t \, [1 + M \sin (2\pi f_m t)] \qquad (1)$$

where $\hat{E}$ is the unmodulated carrier peak value, f_c is carrier frequency, f_m is modulating frequency, and M is the modulation ratio, the maximum value of which is 1.

The vector (expression 1) rotates at a speed of f_c c/s and varies in amplitude f_m times per second. An observer rotating with the carrier vector would see it as stationary, with amplitude varying between the limits $\hat{E}\,(1+M)$ and $\hat{E}\,(1-M)$ as shown in Figure 3a. Expression (1) can be expanded to

$$\hat{E} \sin 2\pi f_c t + \frac{\hat{E}M}{2} \cos 2\pi\,(f_c - f_m)\,t - \frac{\hat{E}M}{2} \cos 2\pi\,(f_c + f_m)\,t \qquad (2)$$

which consists of a carrier vector $\hat{E}$ rotating at f_c c/s and two equal sideband vectors $\frac{\hat{E}M}{2}$ rotating at (f_c+f_m) and (f_c-f_m) c/s about the same point as the carrier vector. Since these two sideband vectors only influence the carrier amplitude it is clear that their resultant must always be in line with the carrier. This is shown in figure 3b for successive instants of time corresponding to points A, B and C

* Credit must be given to H. J. Round, who first envisaged the possibilities of an improved signal-to-noise ratio from a frequency-modulated system in a letter to the *Radio Review* (Vol. 2, p. 220), in 1921.

in figure 1b. The upper-frequency sideband vector S_u is rotating around the carrier vector in a counter-clockwise direction — the accepted positive direction of frequency — at f_m c/s, while the lower frequency sideband vector S_l rotates in a clockwise direction around the carrier vector at the same frequency. The carrier vector in figure 3b is, for convenience, shown as being stationary, although actually, it moves round in a counter-clockwise direction at f_c c/s.

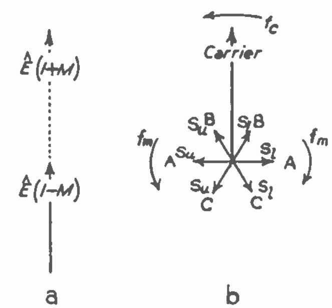

FIG. 3. *Vector diagram for amplitude-modulated wave*

The problem of representing frequency modulation by a single vector is complicated by the fact that the frequency of the carrier varies in accordance with the amplitude of the modulating frequency, and this means that the carrier vector rotates at varying speeds. Taking the example previously given, for a frequency deviation of ±100,000 c/s, the carrier varies from 1,100 kc/s to 900 kc/s. If we, acting as observers of the carrier vector, were to rotate at a frequency f_c c/s, the carrier unmodulated value, in the same direction as the carrier vector, the latter would appear to oscillate backward and forward like a metronome, at the freqency of the modulating wave, *i.e.*, 1,000 times per second. This condition is illustrated in figure 4a. Again, like the metronome, the vector is stationary at the extremes, positions x' and x'', of its stroke so that the carrier frequency is instantaneously at its unmodulated value f_c (point A in figure 2b), while it is moving at its fastest, backward or forward at the center x of its swing. Movement of the vector in a counter-clockwise direction means that the frequency is greater than f_c and in a clockwise direction, less than f_c. Therefore, when the vector moves counter clockwise, point x corresponds to B in figure 2b, but when the vector is travelling clockwise, x corresponds to C. In frequency modulation, the carrier deviation is fixed for a given amplitude of modulating input, so that the speed of the vector, as it passes through x, is constant and independent of the frequency of oscillation backward and forward, *i.e.*, that of the modulating input. Treating the problem as a mechanical one, the initial velocity at x' or x'' is zero, and the final velocity v at x is constant, so that the distance travelled from x' to x or to x'' is proportional to $v \times t$. However, time t is inversely proportional to the modulating frequency. Thus,

$$x'x = x''x = Kvt = K_1t = \frac{K_2}{f_m} \propto \frac{1}{f_m}$$

Therefore, the angle swept out by the carrier vector is large for low modulating frequencies and small for high modulating frequencies. Dotted positions x' and x'' correspond to a lower modulating frequency than the full-line positions. This again is analogous to the metronome, which increases its angle of sweep as the frequency decreases. The mathematical expression for a frequency-modulated wave is

$$\hat{E} \sin \left(2\pi f_c t - M \cos 2\pi f_m t\right) \tag{3}$$

where M is $\dfrac{f \text{ deviation}}{f_{\text{mod}}}$, sometimes called the modulation index, and the angle swept out by the vector is M radians or $57.3\ f_{dev}/f_m$ degrees. This is inversely proportional to f_m. Expression (3) can be treated in the same way as expression (1) for amplitude modulation was turned into a carrier and sidebands. Instead of only two sidebands per fundamental modulating frequency, it is found that there is a large number [1,2] spaced from the carrier by frequencies of $\pm f_m$, $\pm 2f_m$, $\pm 3f_m$, etc.

The theoretically infinite number of sidebands is, for all practical purposes, fortunately limited, as the amplitudes of the sidebands more distant from the carrier normally decrease very rapidly. Frequency modulation differs from amplitude in the number of sidebands as well as in their position with respect to the carrier. All odd-numbered sideband vectors, $\pm f_m$, $\pm 3f_m$, etc., from the carrier are so placed that their resultant falls on a line at right angles to the carrier [4,5] vector, and all even-numbered sidebands have a resultant in line with the carrier vector. The addition of a carrier and two sideband vectors spaced $\pm f_m$ from the carrier having a resultant 90° displaced from the carrier vector is shown in figure 4b. These two sideband vectors rotate around the carrier vector point O at f_m c/s. Their resultant adds to the carrier to cause the latter to oscillate backward and forward along the line AB. In fact, it is the same as figure 4a except for the amplitude variation of the vector. By adding suitable amplitudes of wider-spaced sidebands, we can neutralize the amplitude variation caus-

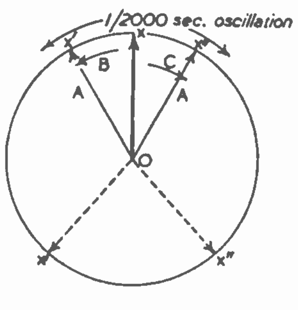

FIG. 4A. *Vector diagram for frequency-modulated wave*

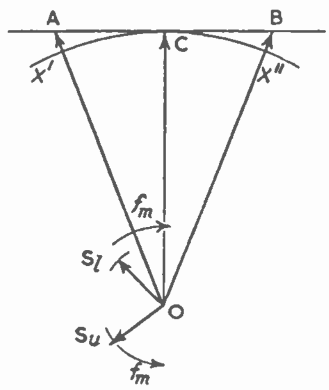

FIG. 4B. *Frequency modulation caused by 90° phase shift in sideband-vector resultant with respect to the carrier*

ing the point of the vector to describe an arc x' x'' giving only frequency modulation. Another important point to note with regard to frequency-modulation sidebands is that the individual amplitudes are not directly proportional to the amplitude of the original modulating frequency as in amplitude modulation, but actually vary widely, sometimes becoming zero, as the angle swept out by the carrier vector changes. The amplitude of the carrier component also varies widely, and for certain modulation-index values falls to zero. The amplitudes are actually Bessel-coefficient functions of the modulation index. The variation of the carrier and first three pairs of sidebands is plotted against modulation index in figure 5. When the modulation index is small, *i.e.*, the angle swept out by the vector in figure 4a is small (f_m high), all sideband pairs except those nearest the carrier are so small in amplitude that they can be neglected. However, for a large modulation index, *i.e.*, a large angle of vector oscillation (f_m low), there are many sidebands of appreciable amplitude. For example, if $f_m = 50$ c/s and its amplitude is such as to give a frequency change of carrier of $\pm 1{,}000$ c/s, the angle swept out by the vector moving from x' to x (figure 4a) is $1{,}000/50 = 20$ radians or 1,146°, whereas for $f_m = 5{,}000$ c/s and the same frequency change, the angle swept out by the vector is $1{,}000/5{,}000 = 0.2$ radians or 11.46°. It should be noted that the carrier vector may make a number of revolutions for low modulating frequencies.

Figure 5 shows that the carrier component is zero at modulation indexes of 2.405, 5.52, 8.654, etc. (the zero points continue at modulation-index separations of approximately π, *i. e.*, 3.14), and this fact is used for determining the frequency deviation[33] of a frequency-modulation signal.

The vector representation of phase modulation is similar to that of frequency modulation with one important difference; the angle swept out by the vector is constant for a given amplitude of all modulating frequencies and is dependent only on the amplitude of the latter. Essentially this means that the velocity of the vector at the center x (figure 4a) increases as the speed of oscillation increases,

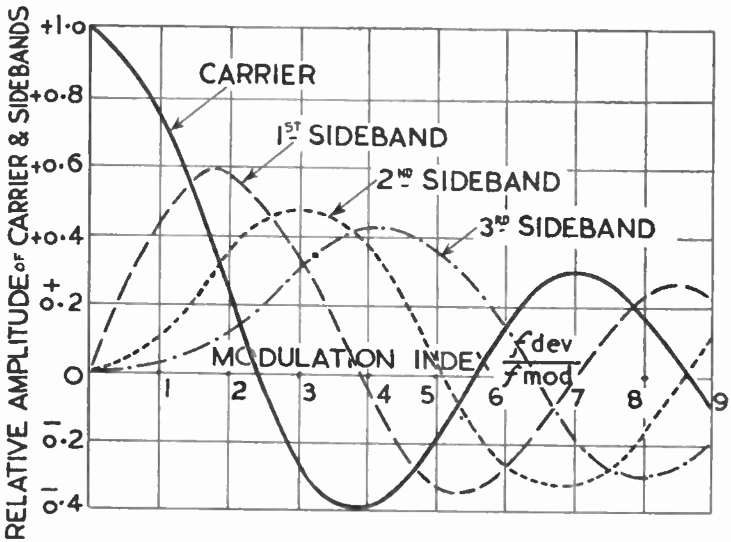

FIG. 5. *Relative amplitude of carrier and sideband components of frequency-modulation carrier in terms of the modulation index*

i.e., the rise and fall in carrier frequency (frequency deviation) is directly proportional to the modulating frequency. Thus, for a given amplitude of modulating voltage, a low frequency produces a small frequency deviation whereas a high frequency causes a frequency deviation n times as large, where n is the ratio of the high to the low modulating frequency. This means that phase modulation is equivalent to frequency modulation with a bass cut circuit in the modulation-frequency amplifier. Conversely, frequency modulation is equivalent to phase modulation with a treble-cut circuit in the modulation-frequency amplifier. It is, therefore, possible to turn phase modulation into frequency modulation by inserting in the modulation-frequency amplifier a network having an amplification characteristic inversely proportional to frequency. This gives a low frequency a large amplitude and a high frequency a small amplitude, *i.e.*, the phase angle is no longer constant, but increases as the modulation frequency falls. The mathematical expression for phase modulation is

$$E \sin \left(2\pi f_c t + \phi\, M \sin 2\pi f_m t\right) \tag{4}$$

and the phase angle swept out is equal to $\phi\, M$ where ϕ is a constant and M is proportional to the amplitude of f_m, but independent of its frequency (see Chapter III). Phase modulation has an infinite number of sidebands spaced $\pm f_m$, $\pm 2f_m$, etc., from the carrier, but since, unlike frequency modulation, the phase-angle change is constant, the wider-spaced sidebands for low as well as high modulating frequencies can usually be neglected. The resultants of all odd-numbered sidebands are 90° out of line with the carrier, and those of the even-numbered ones are in line with the carrier vector. The amplitude of each sideband vector is dependent only on the amplitude of the modualting frequency.

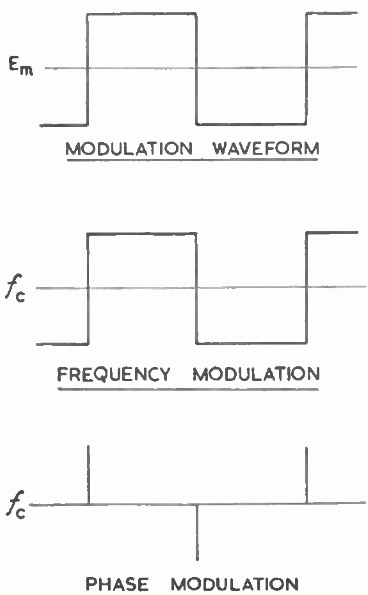

FIG. 6. *Illustration of the essential difference between frequency modulation and phase modulation*

The clearest distinction between frequency modulation and phase modulation has been indicated by Professor Howe.[29] He considers a square-shaped modulating wave as in figure 6 and finds that for frequency modulation, the carrier frequency varies above and below its unmodulated value in accordance with the amplitude of the modulation. Phase modulation, however, shows that apart from the sudden, instantaneous changes of phase when the carrier frequency becomes $+\infty$ or $-\infty$ the latter is constant at its unmodulated value.

The following table summarizes the control exercised by the modulation frequency upon the carrier by the three types of modulation discussed in this chapter.

TABLE I

Type of Modulation	*Frequency Modulation*	*Amplitude Modulation*
Amplitude	Rate of change of carrier amplitude	Extent of carrier amplitude change
Frequency	Rate of change of carrier frequency	Extent of carrier frequency change
Phase	Rate of change of carrier phase	Extent of carrier phase change

Chapter II

THE ADVANTAGES AND DISADVANTAGES OF FREQUENCY-MODULATED AND PHASE-MODULATED TRANSMISSION

WE can now consider the particular advantages and limitations of frequency modulation and phase modulation. Since frequency-modulated transmission is more likely to be widely adopted, we will deal with it first and in more detail.

Four important advantages[13] can be obtained with frequency modulation in comparison with amplitude modulation, *i.e.*, greater signal-to-noise ratio, lower transmitter power for a given audio-frequency output from the receiver, less amplitude compression of the audiomodulating voltage, and larger service area with little interference between stations having adjacent carrier frequencies. These advantages can, however, be obtained only under certain conditions of operation, the most important one of which is that reception must be confined to the direct ray from the transmitter. Indirect-ray communication, as in short-wave transmission over long distances, is subject to selective fading of carrier and side-bands caused by several rays arriving at the receiver by different routes. The time delay on the longer-path rays may result in certain frequency components arriving partly or completely out of phase with the same components in the shorter-path ray, thus causing a reduction in amplitude of these components. This selective fading causes serious distortion[3] of the audiofrequency signal at the receiver and it is worst when a large number of sidebands is being transmitted. One of the characteristics of frequency modulation is the large number of sidebands produced by low modulating audiofrequencies, and selective fading makes the output almost unintelligible. This effect and the need for a large pass-band to accommodate the frequency deviation render frequency modulation impracticable except on ultrashort waves.[13, 14] Phase modulation has fewer low-frequency sidebands so that selective fading is much less serious and it may be employed for short-wave transmission.[20] Amplitude modulation is least affected by multipath selective fading because there are only two sidebands per modulating frequency.

To understand the reason for the greater signal-to-noise ratio obtained from a frequency-modulated system, it is necessary to state the characteristics of noise. Disturbances in, or external to, the receiver can produce noise in its output. Noise from sources outside the receiver is mainly of the impulse type and is caused by atmospheric disturbances or interference from electrical machinery, *e.g.*, the ignition system of a car, switching surges transmitted by the mains supply wiring, etc. It often has high peak voltages and may be periodic, continuous, or spasmodic. In a well-designed receiver, having no faulty contacts, internal noise is due to the random motion of electrons in the conductors and in the valves. The important sources are the first-tuned circuit and the radio-frequency valve. Thermal agitation (conductor) and shot (valve) noise usually cause a hissing sound and the frequency components cover a very wide range and are continually varying in amplitude and phase. In a receiver for amplitude modulation (in which no carrier is being received), these noise voltages beat among themselves, the beats being made audible by detection, and the wider the band, the worse is the noise. For example, if the receiver has a band width of 900 to 1,100 kc/s, a beat of 5 kc/s is produced by two noise voltages at 900 and 905 kc/s as well as by two at 1,095 and 1,100 kc/s and this can be repeated for all the noise components. If a carrier is applied and is large enough to insure linear detection, the noise voltages act as sidebands to the carrier and audible beats are now only produced between carrier and noise, *i.e.*, beats between the noise components themselves are suppressed.

Thus, only the noise components within audiorange of the carrier contribute to the noise output, and the noise output should fall. In practice, we more often find that the application of a carrier increases noise and this may be caused by noise on the carrier itself, *i.e.*, from the transmitter, and also by the fact that the noise voltages alone are not large enough to cause linear detection. However, it is still true that in the presence of a carrier, only the noise components within audiorange of the carrier contribute to the output.

A device called limiter is always incorporated in a frequency-modulation receiver to suppress any amplitude changes of carrier so that noise cannot have the same effect as in an amplitude-modulated receiver. For the sake of clarity, let us consider the action of a single noise-frequency component f_n kc/s spaced an audiofrequency $f_m \pm f_n - f_c$ kc/s from the carrier. By taking the carrier vector as stationary, the noise is a vector rotating round the carrier at f_m kc/s as illustrated in figure 7. We see that there is amplitude change and phase change of the carrier vector. The former,

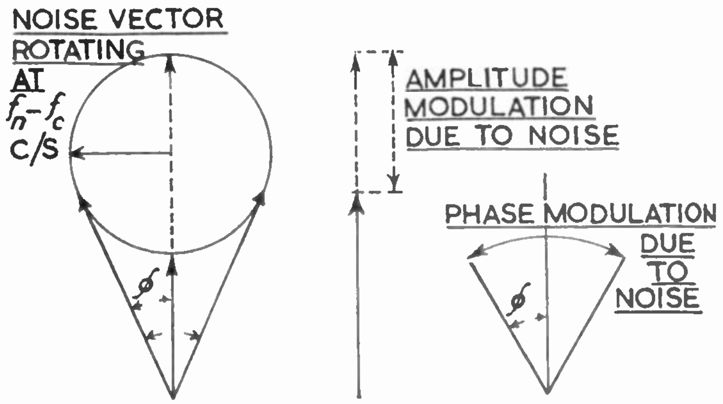

Fig. 7

which causes the noise output in the amplitude-modulation receiver, is suppressed by the limiter, and the latter, which produces phase modulation and thus, frequency change of the carrier, gives noise at the output of the frequency-modulation receiver. An important feature of phase modulation is that the carrier-frequency deviation is directly proportional to the frequency of a constant amplitude-, modulating signal so that noise sidebands near the carrier give much less frequency deviation and consequently much less audio-output from a frequency modulation receiver than those distant from the carrier. This "triangular" distribution of effective noise makes the signal-to-noise power ratio with a maximum frequency deviation equal to the audiorange $\pm$ 15 kc/s. This is 2.9 times higher than for an amplitude-modulation receiver working under corresponding conditions.[13] In the case of impulse noise from motor-car ignition systems, the signal-to-noise ratio is four times that of amplitude modulation. It is not necessary to confine the frequency deviation of the carrier to the audiorange and by increasing this, the signal output can be increased. Thus, the signal-to-noise ratio may be further raised by the ratio of frequency deviation to maximum audiomodulating frequency, *e.g.*, if the carrier deviation is $\pm$ 75 kc/s and the maximum audiofrequency is 15 kc/s, the signal-to-noise ratio is increased 25 times. Taking the conservative figure of 2.9 to 1 improvement caused by "triangular" noise distribution, we have a total improvement in signal-noise ratio of 72.5:1 (18.6 db). The increase in receiver-band width to accommodate the greater frequency deviation introduces extra noise sidebands, but if the carrier is large in comparison with the noise (at least twice the peak value of noise), there is no increase in noise output because the phase modulation of the carrier by the additional sidebands is outside the audible range. When the peak carrier-to-noise ratio is less than 1, interaction occurs between noise components, and noise is increased and signal-to-noise ratio decreased by increasing the frequency deviation. This causes a well-defined threshold area[30] to appear around a frequency-modulated transmitter. Outside this area, better reception is obtained with a lower frequency deviation and narrower receiver pass-band.[43] Inside this area, the reverse is true.

Signal-to-noise ratio can be still further improved by the use at the transmitter of "preemphasis,"[40] *i.e.*, increased amplitude of the higher modulation frequencies, followed at the receiver by "de-emphasis," *i.e.*, restoration of aural balance by reduced amplification of the high modulation frequencies. Preemphasis and de-emphasis can also be applied to amplitude-modulated transmission, but the improvement in signal-to-noise ratio is not so great as in the case of frequency modulation which, because of phase modulation of the carrier by the noise, produces greatest noise in the higher audio-ranges. An improvement in signal-to-noise power ratio of about 5.5/1 (7.4 db) is obtained with preemphasis giving a total increase of 400 to 1 (26 db) for frequency modulation in comparison with amplitude modulation. Field tests performed in England[55] suggest that these values are too high and improvements of 4.5 db with random noise and 1.5 db with ignition noise seem more practically attainable figures.

The second advantage of frequency modulation is that less power is taken from the mains supply for a given audiopower at the receiver output. In the power-amplifier stage of an amplitude-modulated transmitter, the direct current must be sufficient to allow 100 per cent, modulation without serious distortion, *i.e.*, it must be able to accommodate a carrier of twice the unmodulated amplitude. This means that the direct current must be twice the value which would be required if the carrier were maintained at its unmodulated amplitude. Since a frequency-modulated carrier has constant amplitude, it follows that the power required from the mains is half that of its amplitude-modulated counterpart. Alternatively, for a given mains power, frequency modulation can give an audiosignal power at the receiver of twice that for a corresponding amplitude-modulated system. This means a further increase in signal-to-noise ratio of 2 to 1 giving a total of 800 to 1 (29 db).

Successive stages of improvement in signal-to-noise power ratio are illustrated in figure 8.

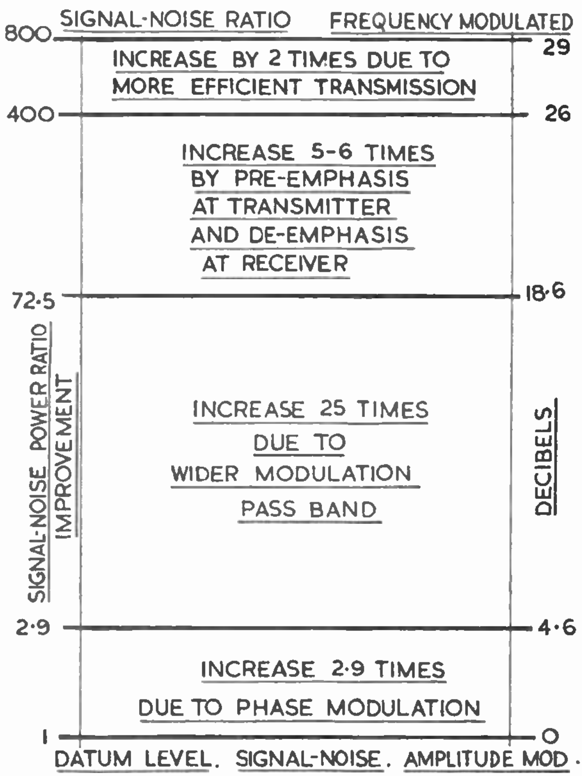

FIG. 8

Reduced compression of the audiosignal in a frequency-modulated transmitter really arises out of the increased signal-to-noise ratio. For an amplitude-modulated system, the maximum modulation percentage for reasonable distortion (less than 5 per cent) is 90 per cent and a suitable minimum value is 5 per cent if low level sounds are not to be marred by noise, so that the maximum possible change in audiooutput power is limited to 320 to 1 (25 db). Note that percentage modulation represents voltage, which must be squared to give power. Clearly, the maximum change of 10,000,000 to 1 (70 db) between fortissimo and pianissimo orchestral passages would sound unnatural in a normal room and some compression is essential. A power ratio higher than 320 to 1 is, however, desirable, and it can be raised to 32,000 to 1 (45 db) because of the higher signal-to-noise ratio (shown above at least 100 times better) in frequency modulation.

In addition to noise, a very important problem in wireless communication is the separation of desired from undesired programs. In an amplitude-modulated system, this limits the closeness of spacing between the carrier frequencies. If the separation between the two carriers is equal to an audiofrequency, an audible note is produced in the receiver output, causing serious interference with the desired program unless the desired signal at the receiver antenna is at least ten times that of the undesired signal. This limits the service area of either transmitter, and between the two there is a large area in which reception of one program is marred by the other. Increased separation of the carrier frequencies can remove the carrier-separation frequency outside the audiorange, but the desired signal service area is still limited by sideband overlap[45] from the undesired area. The sidebands of the latter react with the desired carrier, causing the characteristic frequency-inverted "monkey chatter." Powerful transmitters need to be separated by at least 50 kc/s if the interference area between them is not to be large.

A different state of affairs exists with two frequency-modulated systems, because the receiver suppresses amplitude change. Interference, as in the case of noise, occurs because of phase modulation of the desired carrier by the undesired carrier. This phase modulation results in an audiooutput of frequency equal to the carrier separation whose amplitude is directly proportional to the separation. Phase modulation is equivalent to frequency modulation with modulating amplitude being directly proportional to the modulating frequency. Thus, for small carrier separations, the interference is small. It is actually most noticeable for about 5 kc/s separation,[40] for although greater separations give greater equivalent modulation, the resultant output becomes less audible. We, therefore, find that two frequency-modulated systems can be operated with small carrier spacing with quite a small interference area (where the desired-carrier to undesired-carrier ratio is less than 2 to 1) between them. Interference is worst when both carriers are unmodulated. Although it is possible to operate with small carrier spacing, it is usually considered better to adopt a spacing slightly beyond the audiorange. This does not modify in any way the statements on the smaller interference area obtained with frequency as compared with amplitude modulation.

An effect to which reference should be made is that known as "capture." It occurs when two frequency-modulated transmitters are operating at the same carrier frequency. If the field strength of the desired signal is greater than about twice the undesired signal, interference from the latter is not serious. It is greatest when both field strengths are equal. Increase of the undesired field strength beyond this point causes the desired program to be reduced and the undesired program increased. At a field strength ratio of undesired to desired of about 2 to 1, the desired program is almost completely suppressed, *i.e.*, the undesired program has "captured" the receiver. This same action occurs between noise and the desired program. When the noise voltage approaches twice the desired voltage, the latter is suppressed and noise captures the receiver producing the threshold area already referred to. The net result is that a frequency-modulated transmission has a well-defined area in which signal-to-noise ratio is very good and outside which it is poor. With amplitude modulation, signal-to-noise ratio steadily deteriorates as the distance from transmitter is increased.

Phase modulation possesses much the same advantages as frequency modulation. Signal-to-noise ratio is greater than for amplitude modulation because of the increased frequency deviation coverage, although it is less than for frequency modulation since the triangular noise-spectrum effect is absent because noise itself phase-modulates the carrier. Lower transmitter mains power is required because of constant amplitude, and less audiocompression is needed. This means that figure 8 is applicable to phase modulation if the first step of 4.6 db is omitted. As stated before, it may be used for short-wave transmission when multipath selective fading is experienced, and it then has the advantage of lower transmitter power in comparison with amplitude modulation.

Chapter III

METHODS OF MODULATING THE FREQUENCY OR PHASE OF A CARRIER

THE close relationship between frequency modulation and phase modulation has already been shown. Obviously only a very slight modification is necessary to convert frequency modulation into phase modulation and *vice versa*.[7] The modification consists of inserting a frequency-discriminating network in the modulator stage. A resistance R and inductance L connected between the audiofrequency source and the frequency modulator (figure 9a) result in a phase-modulated output, because the RL circuit makes the audio-frequency voltage amplitude directly proportional to its frequency. This means that a constant-amplitude variable-frequency voltage at AB gives an amplitude proportional to its frequency (f_m) across CD and the frequency deviation of the carrier is directly proportional to f_m, the condition for phase modulation. Similarly, the RC circuit shown in figure 9b gives an amplitude across CD inversely proportional to f_m for a constant amplitude at AB, and this in conjunction with the phase modulator, produces a phase change of carrier inversely proportional to f_m, which is the characteristic of frequency modulation.

There are three main problems to be faced in frequency-modulation transmitting-apparatus design: (1) The frequency of the central carrier component must remain at a constant value and must not change when modulation is applied, (2) the frequency deviation of the carrier must be directly proportional to the amplitude of the modulating frequency and independent of its frequency, and (3) the amplitude of the varying carrier-frequency vector must be constant and independent of modulation. Frequency modulation of an oscillator can be accomplished by direct or indirect means. Direct modulation involves varying the frequency of the oscillator itself by varying the equivalent inductance or capacitance of its tuning elements. The most important problem in this case is the first, *i.e.*, maintaining constant central carrier-component frequency. With reasonable care in the design of the variable-reactance modulator, neither the second nor the third problems are serious. Indirect frequency modulation is achieved by altering the relationship of the carrier and sidebands of an amplitude-modulated carrier (as described in chapter I), or by varying the tuning elements of an amplifier stage following a crystal-controlled master oscillator, *i.e.*, by phase modulation in association with the RC correcting network described previously. The first problem now presents no difficulties, but the second and third assume importance.

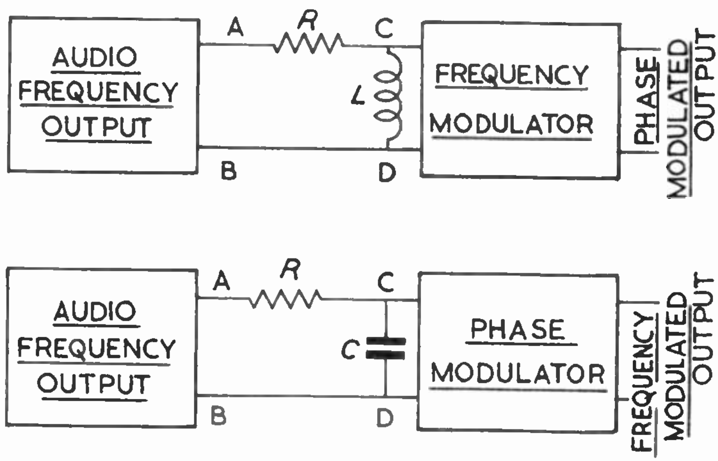

FIG. 9A. *Phase-modulated output from a frequency modulator*
B. *Frequency-modulated output from a phase modulator*

Direct frequency modulation is usually obtained by means of a variable-reactance valve, the anode and cathode of which are connected across the oscillator-tuned circuit, or a part of it. A voltage derived from the oscillator-tuned circuit which is at 90° to the anode-cathode voltage is applied to the grid of this valve. The variable-reactance valve anode current is in phase with the grid voltage and, therefore, leads upon, or lags behind the anode-cathode voltage by 90° to give the valve the characteristics of a capacitance or an inductance. An $\frac{1}{8}\lambda$ line,[10] terminated in a variable resistance controlled by the modulating voltage, is also a possibility, but for satisfactory operation, line losses must be low. A third method[47] employs a coil coupled to the oscillator-tuned circuit and shunted by a valve acting as a variable resistance and controlled by the modulating voltage. Control of saturation of iron-cored coils forming part of the oscillator-tuned circuit is a fourth method. It is also possible to frequency modulate a multivibrator[11] oscillator by using the modulating voltage as a bias on one of the valves.

Some form of carrier-component-frequency control must be adopted with all methods of direct frequency modulation. An automatic frequency-correcting circuit, similar to that used in receivers to maintain correct oscillator frequency, is very suitable for this purpose.

Indirect frequency modulation (corrected phase modulation) can be obtained by using any of the variable-reactance devices described for direct frequency modulation, but the variable reactance is placed across a tuned circuit in the anode of an amplifier valve following the crystal-controlled master oscillator. The other method is transposing sidebands of an amplitude-modulation carrier. Both suffer from the disadvantage of requiring a large degree of multiplication after the phase modulator in order to obtain the desired degree of frequency deviation at the output. The maximum angle of vector travel must not normally exceed ±25° (it can be raised by special compensating circuits to ±50°), and the final maximum angle may have to be of the order of 140,000° requiring a multiplication of over 5,000 times. Larger phase-angle deviations may be obtained by means of the phasitron,[56] for which a phase-angle change of $\pm 200°$ is claimed when modulation-content distortion is limited to 1.5 percent.

The most popular method of attaining direct frequency modulation is with the variable reactance valve and automatic-frequency correction of the carrier component. The L or C change of the oscillator-tuning elements must be such as to produce instantaneous changes of carrier frequency directly proportional to the instantaneous modulating-voltage amplitude. The required relationship between the frequency deviation of carrier and the L or C change is dependent on the magnitude of the ratio of the former to the carrier frequency. Probable values of frequency deviation and carrier are ± 75 kc/s and 40 Mc/s so that $f_{dev}/f_{carrier} = \pm .001875$. This low value makes simplifications in the analysis possible because it means that the ratio of the change of the inductance ΔL to the total inductance L of the tuned circuit is also small. For example, let

$$f_c = \frac{1}{2\pi\sqrt{LC}} = \text{initial carrier frequency} \tag{5}$$

If the frequency deviation of carrier is Δf, we have:

$$f_c + \Delta f = \frac{1}{2\pi\sqrt{(L-\Delta L)C}} \tag{6a}$$

and

$$f_c - \Delta f = \frac{1}{2\pi\sqrt{(L+\Delta L)C}} \tag{6b}$$

Combining (5) and (6a) the result is the same if we use (5) and (6b)

$$\frac{f_c+\Delta f}{f_c} = 1 + \frac{\Delta f}{f_c} = \frac{1}{\sqrt{1-\frac{\Delta L}{L}}} = \left(1-\frac{\Delta L}{L}\right)^{-\frac{1}{2}} \tag{7}$$

Expanding by the binomial theorem

$$1 + \frac{\Delta f}{f_c} = 1 + \frac{\Delta L}{2L} - \frac{3}{8}\left(\frac{\Delta L}{L}\right)^2 + \text{etc.} \tag{8}$$

but since $\Delta f/f_c$ is very small, it follows that $\Delta L/L$ is also small.

Thus $(\Delta L/L)^2$ is negligible, and (8) becomes

$$\frac{\Delta f}{f_o} = \frac{\Delta L}{2L} \tag{9a}$$

or

$$\Delta f \propto \Delta L \tag{9b}$$

Similarly, for a small change of capacitance ΔC

$$\frac{\Delta f}{f_c} = \frac{\Delta C}{2C} \tag{9c}$$

or

$$\Delta f \propto \Delta C \tag{9d}$$

Undistorted frequency modulation can, therefore, be obtained by making the change in L or C directly proportional to the amplitude of the modulating frequency. It is not always convenient to make a direct change of L and it may be necessary to obtain it by varying an inductance, or its equivalent, placed in parallel with L. The following equations then result:

$$f_c = \frac{1}{2\pi\sqrt{\frac{LL_1}{L + L_1}C}} \tag{10}$$

$$f_c + \Delta f = \frac{1}{2\pi\sqrt{\frac{LL_2}{L+L_2}C}} \tag{11}$$

where L_1 is greater than L_2.

The lower deviation of carrier $f_c - \Delta f$ yields a similar equation, but need not be considered.

$$1+\frac{\Delta f}{f_c} = \sqrt{\frac{1+\frac{L}{L_2}}{1+\frac{L}{L_1}}} = \left(1+\frac{L}{L_2}\right)^{\frac{1}{2}} \times \left(1+\frac{L}{L_1}\right)^{-\frac{1}{2}} = 1+\frac{L}{2L_2}-\frac{L}{2L_1}$$

if L_1 and $L_2 \gg L$

$$\text{thus } \frac{\Delta f}{f_c} = \frac{L}{2}\left(\frac{1}{L_2} - \frac{1}{L_1}\right)$$

$$\text{or } \Delta f \propto \left(\frac{1}{L_2} - \frac{1}{L_1}\right) \qquad (12)$$

For Δf to be proportional to the modulating-frequency amplitude, the latter must be proportional to $(1/L_2 - 1/L_1)$, and we shall see later that this result can be achieved.

An electrical method for varying the inductance or capacitance of the carrier oscillator-tuned circuit is obviously preferable to a mechanical one. The variable-reactance valve, as used in automatic frequency-correcting circuits, is particularly suitable for this purpose since it acts as a reactance to any voltage source connected

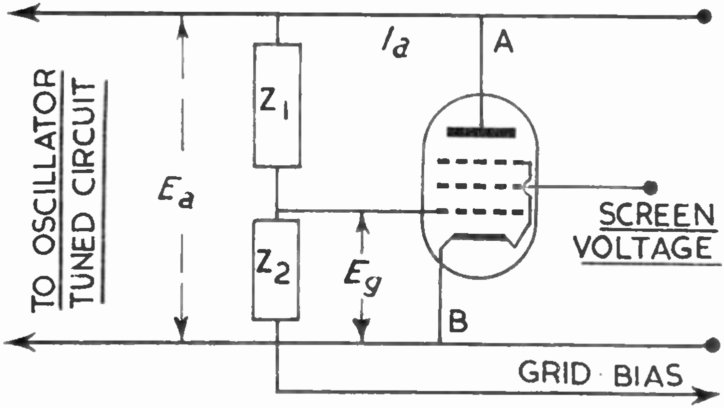

FIG. 10. *Variable-reactance valve circuit.*

between its anode and cathode, the value of the reactance depending on the grid bias of the valve. To utilize this property, it is necessary to supply the grid with a proportion of the anode voltage which has been given a phase shift of 90°. The basic circuit is that of figure 10. Impedances Z_1 and Z_2 act as a potentiometer to step-down and phase-shift the grid voltage. Thus if Z_1 is a resistance R, Z_2 a capacitance C, and the valve is a pentode or tetrode with a high internal resistance, the admittance across its anode and cathode (points AB) is given by

$$Y_{AB} = \frac{I_a}{E_a} = \frac{g_m E_g}{E_a}$$

where g_m is mutual conductance of the valve

$$Y_{AB} = \frac{g_m E_a Z_2}{(Z_1 + Z_2) E_a} = \frac{g_m Z_2}{Z_1 + Z_2} = \frac{g_m}{1 + \mathrm{j}\, R\omega C}$$

$$= \frac{g_m}{1 + (R\omega C)^2} - \mathrm{j}\frac{g_m R\omega C}{1 + (R\omega C)^2} \tag{13}$$

which is equivalent to a resistance $R_{AB} = \dfrac{1 + (R\omega C)^2}{g_m}$ in parallel with an inductance $L_{AB} = \dfrac{1 + (R\omega C)^2}{g_m R\omega^2 C}$

Three other combinations of R and L or C are possible and the resultant parallel resistance and reactance components of Y_{AB} are given in table II.

TABLE II

Z_1	R	C	R	L
Z_2	C	R	L	R
R_{AB}	$\frac{1 + (R\omega C)^2}{g_m}$	$\frac{1 + (R\omega C)^2}{g_m (R\omega C)^2}$	$\frac{R^2 + \omega^2 L^2}{g_m \omega^2 L^2}$	$\frac{R^2 + \omega^2 L^2}{g_m R^2}$
X_{AB}	$L_{AB} = \frac{1 + (R\omega C)^2}{g_m R\omega^2 C}$	$C_{AB} = \frac{g_m RC}{1 + (R\omega C)^2}$	$C_{AB} = \frac{g_m RL}{R^2 + \omega^2 L^2}$	$L_{AB} = \frac{R^2 + \omega^2 L^2}{g_m R\omega^2 L}$

From the first and fourth columns in table II we see that the equivalent inductance is inversely proportional to g_m so that expression (12) becomes

$$\Delta f \propto g_{m2} - g_{m1}$$

and if the $g_m E_g$ curve of the variable-reactance valve is a straight line $$\Delta f \propto E_{g2} - E_{g1}$$

Therefore, by inserting a modulating voltage in series with the d-c grid bias, the resultant variation of inductance produces a deviation of carrier frequency directly proportional to the modulating-voltage amplitude. Similarly, if g_m is proportional to E_g, the change in capacitance in columns 2 and 3 of table II means that Δf is proportional to the modulating-voltage amplitude (see expression 9d).

It is not usual to employ the reactance valve to modulate an ultrahigh-frequency oscillator. A lower carrier frequency is commonly used and frequency-multiplier stages inserted between it and the antenna to step-up the frequency to the required 40 Mc/s. This has the advantage of simplifying reactance-valve circuit design and of reducing interaction between the oscillator stage and succeeding amplifier stages. Furthermore, greater oscillator stability is possible at lower frequencies. The resistance component R_{AB} is undesirable because it causes amplitude modulation, but by a suitable choice of R and L or C and a reduction of the resonant impedance (L/CR) of the carrier-tuned circuit, this effect can be reduced to small proportions. A limiter may be inserted before the multiplier stages to reduce still further any amplitude modulation. For example, if the oscillator frequency is 1,000 kc/s, Z_1 a capacitance of 5 $\mu\mu$F, and Z_2 a resistance of 5,000Ω, mean g_m is 1 mA/volt, and oscillator-tuning capacitance 200 $\mu\mu$F.

$$C_{AB} = \frac{g_m CR}{1 + (R\omega C)^2} = 24.4\ \mu\mu\text{F}$$

The total tuning capacitance is 224.4 $\mu\mu$F. The required change of capacitance

$$\Delta C = \pm\ 224.4 \times .00375 = \pm\ 0.84\ \mu\mu\text{F}$$

Note from the first part of the chapter and expression (9c) that $\Delta f/f_o = 0.001875$ and $\Delta C/C = 2\,\Delta f/f_o = 0.00375$.)

This means a mutual conductance change of $\pm 0.84/24.4 = \pm 0.0344$ mA/volt. The value of R_{AB}, when g_m is 1 mA/volt, is 41,500Ω and it varies from 43,000Ω to 40,100Ω. If the oscillator coil has a Q of 50, giving a resonant impedance of 35,500Ω, the variation in amplitude due to the variation of R_{AB} corresponds to a modulation percentage of approximately 1.5 per cent. The variable reactance valve can be placed across only a part of the oscillator coil if desired, and this calls for a greater change of g_m to give a specified frequency modulation. Amplitude modulation by R_{AB} is slightly increased. The parallel resistance component R_{AB} can be increased to infinity if the phase-splitting device can be made to give an exact 90° phase difference between the anode and grid voltages of the variable-reactance valve. This can be achieved by replacing R in the phase splitter by a tuned circuit or by injecting an extra voltage into the grid circuit.[54] Push-pull reactance valves can reduce amplitude modulation to negligible proportions. Thus, two reactance valves may be connected across the Hartley oscillator of figure 11, one valve having an RC phase-splitting network, and the other, across the opposite half of the tuned circuit, having a CR network. Application of the modulating voltage in push-pull, *via* a center-tapped transformer, produces cancellatory variations in the resistance components, but the reactance variations are additive. The first valve has a resistance component of $R'_{AB} = \dfrac{1 + (R\omega C)^2}{g_{mo} + \Delta g_m}$ and the second has one of

$$R''_{AB} = \frac{1 + (R\omega C)^2}{(g_{mo} - \Delta g_m)\,(R\omega C)^2}$$ where g_{mo} is unmodulated g_m value,

$+\Delta g_m$ is change in g_m on one valve, and $-\Delta g_m$ is change in g_m on the other valve at the same time. The reactance components are

$$L_{AB} = \frac{1 + (R\omega C)^2}{(g_{mo} + \Delta g_m)R\omega^2 C} \text{ and } C_{AB} = \frac{(g_{mo} - \Delta g_m)CR}{1 + (R\omega C)^2}$$

and it will be noted that L_{AB} is decreased by increase of g_m and, as it is in parallel with the oscillator-tuning inductance, the oscillator frequency is increased. At the same time C_{AB} is decreased, also raising the frequency.

When a reactance valve is used as a modulator, it is essential to prevent change of g_m by means other than the modulating-voltage amplitude. Variation of high tension, hum voltages due to the low tension, aging of the valve can all contribute to changing g_m. The first two effects are reduced to small proportions by using two reactance valves in push-pull, but the oscillator frequency itself may vary, due to high-tension or low-tension supply-voltage variations or to changes of temperature, and the carrier-frequency stability can only be successfully maintained by using an automatic frequency-control[37, 41] circuit to control the d-c bias on the reactance valve. A basic circuit is shown in figure 11. Valve V_1 is the carrier oscillator (Hartley circuit) and V_2, a hexode valve, is the variable-reactance device. The audio-frequency voltage is applied at the first grid (the signal grid when V_2 is used as a frequency

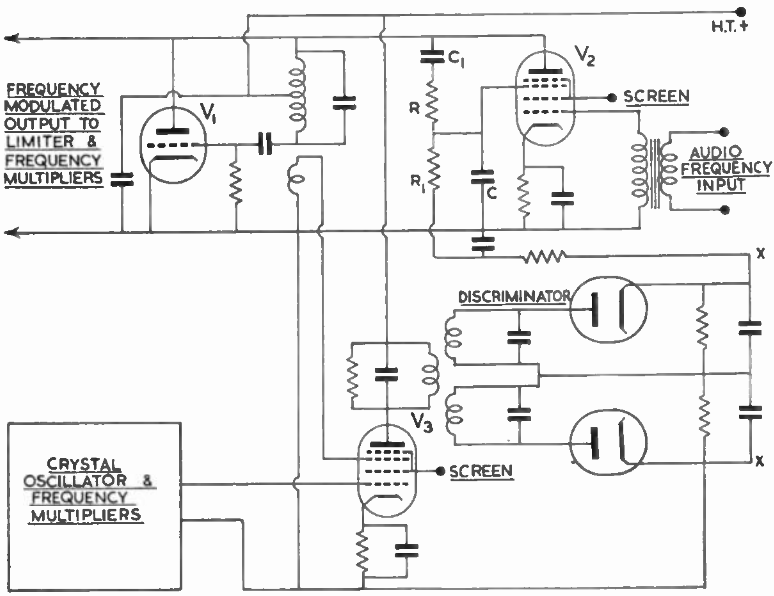

FIG. 11. *Frequency modulation with stabilized variable-reactance valve*

changer) to control the mutual conductance of the third grid, to which the RC phase-splitting network is connected. The third grid, normally the oscillator grid, is used because its $I_a E_g$ curve is more linear for large applied voltages than that of the first grid, and to get the required reactance change, the valve must be connected across the major portion of the oscillator coil. Condenser C_1 is the a-c coupling between anode and grid and has a large value (0.1 μF) while R_1 has a resistance of about 0.5 MΩ which is large compared with the reactance of condenser C. In the first chapter, we showed that a frequency-modulated wave consisted of a central carrier frequency, equal to the unmodulated value, and sidebands. This "central" component is applied to another hexode V_3, acting as a frequency changer, together with the output from a very stable crystal oscillator which is multiplied up if necessary. The difference frequency is passed to a discriminator, which is, in this case, two circuits tuned about 2 kc/s above and below the difference frequency, and two detectors giving opposing d-c voltages which provide bias for V_2 through the grid leak R_1. When the original carrier frequency is correct, the difference frequency is exactly centered between the two discriminator circuits and there is no d-c voltage across XX. If the carrier-frequency component wanders, a voltage, positive or negative, is produced and automatically adjusts the d-c bias on the reactance valve to correct for this. The carrier frequency component then has practically the same stability as that of the crystal oscillator plus discriminator, and by placing both these in a simple temperature-controlled oven, an adequate degree of frequency stability is obtained. It is generally sufficient to maintain an output carrier of 40 Mc/s within 1,000 c/s of its correct value for all normal operating conditions.

The push-pull variable-reactance valve circuit is capable of producing comparatively large frequency deviations at the initial carrier frequency ($\pm$20 kc/s at a carrier frequency of 5 Mc/s) and this means that the sideband amplitudes may be comparable with

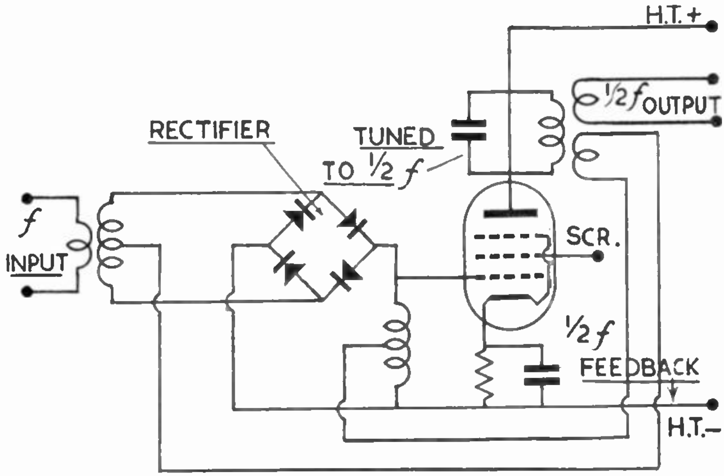

FIG. 12. *A circuit for frequency division*

or greater than the carrier component. It is, therefore, advisable to increase the latter relatively to the sidebands before application to the discriminator, and this is achieved by frequency division.[17] A frequency-dividing circuit is, in essence, a frequency changer having an input frequency of f and an output circuit tuned to $\frac{1}{2}f$. Part of the output voltage from this circuit is fed back to the input to react with the original input to provide a difference frequency of $\frac{1}{2}f$. Initially, shock excitation of the output $\frac{1}{2}f$ circuit caused by switching on the f input, noise, or any transient disturbance, produces the $\frac{1}{2}f$ frequency, and so long as f persists, $\frac{1}{2}f$ is generated in the output. An example of the frequency-divider circuit is shown in figure 12.

Automatic frequency correction of the oscillator can be achieved by motor control [41] of a trimmer capacitor in parallel with the main tuning capacitor. The motor is operated by the rotating field set up from the beat frequency between the initial carrier-component frequency, or a submultiple of it, and a crystal oscillator. The rotating field drives the motor backward or forward until the beat frequency is zero.

Indirect frequency modulation (corrected phase modulation) can be obtained by separating the two sidebands of an amplitude-modulated carrier from the carrier and passing them through a phase-adjusting network which places their resultant at 90° to the carrier vector, as shown in figure 13. This method does not completely suppress amplitude modulation (a limiter can be incorporated to remove amplitude modulation) and the phase change is not exactly proportional to the modulating-voltage amplitude, but if its maximum value is limited to about $+25°$, distortion of the modulation does not exceed 5 per cent.[18] Now if 30 c/s modulation is to produce a carrier frequency deviation of ± 75 kc/s at 40 Mc/s, a phase change of 143,200° ($75{,}000 \times 360/30 \times 2\pi$) is required, so that the phase change of 25° must be multiplied 5,720 times, *i.e.*, the original carrier frequency must be $40/5{,}720 = 7$ kc/s. Generally the minimum carrier frequency is about 200 kc/s, and it is multiplied first up to about 12 Mc/s. The latter is then applied to a frequency changer with a crystal-controlled local oscillator, or the initial unmodulated 200 kc/s carrier voltage suitably multiplied, to give a difference frequency of about 900 kc/s, which has the full frequency modulation of the 12 Mc/s input and can be multiplied up again. A circuit[13] showing this type of phase modulator is given in figure 14. The carrier frequency obtained from the crystal-controlled master oscillator is fed into two channels. One is an amplifier and the other a balanced push-pull modulator. This channel suppresses the carrier and changes the phase of the resulting sideband by 90° with respect to the original carrier. Suppression of the carrier is obtained by connecting the carrier input in the common grid lead, thus producing cancelling voltage changes in the anode circuit.

The sideband frequencies which are obtained by addition of the carrier voltage and the audio-frequency voltage in T_1 are added at the output circuit (primary of T_2). The condensers C_1 and C_2 in the primary circuit of T_2 cancel its inductive reactance so that it behaves as a pure resistance. The sideband current I_1 in the primary is thus in phase with the carrier input to the grid.

The required phase shift of 90° or 270° for the sidebands is, however, obtained at the secondary because the voltage across the points AB is out of phase with the current I_1 by 90° or 270°, depending on the direction of winding of the secondary.

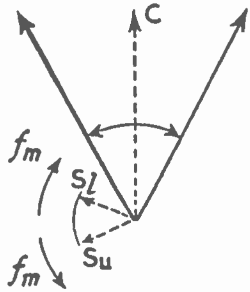

FIG. 13

The sideband voltage is amplified in V_3 and mixed with the original carrier frequency in the anode of V_4, after which the resulting phase-modulated wave is amplified and passed through frequency multipliers to bring the carrier up to its required ultrahigh-frequency value.

As shown in chapter I, the limit of ±25° placed upon the maximum possible angle of initial phase modulation is caused by the fact that the locus of the carrier vector is the tangent AB rather than the arc X' CX'' (figure 4b). If the carrier-component vector OC in figure 4b could be reduced in value as the sidebands S_L and S_U approach their "inphase" condition to the right or left, then the equivalent modulation distortion would be reduced, or alternatively, the angle could be increased for the same distortion. It has been found possible to increase the maximum initial phase angle of modulation to about 50° by using the modulating voltage to reduce the carrier vector at the extremes of the vector swing.[48] This is performed by taking the output across a resistance load in a full-wave rectifier, supplied by the modulating voltage, to bias the carrier-amplifier valve V_4 in figure 14. A resistance load (not shunted by a reservoir capacitance) must be employed because the bias must follow the instantaneous changes of modulating voltage, and full-wave rectification is necessary in order that the carrier vector may be reduced by both positive and negative halves of the modulating voltage. A delay bias is introduced in series with the rectifier so that there is no reduction of carrier amplification when the carrier vector is in the center (point C in figure 13) of its swing. Phase modulation may also be obtained by combining three carrier voltages,[54] two of which, E_1 and E_2, at $\pm$ 90° to the third E_3, are differentially amplitude-modulated. Rectified bias voltage, obtained from the modulation voltage, may be applied to the amplifier for E_3, as described previously, so as to reduce the amplitude modulation at the extremes of the vector swing.

A very important measurement on a frequency-modulation transmitter is that of frequency deviation and this can be made by

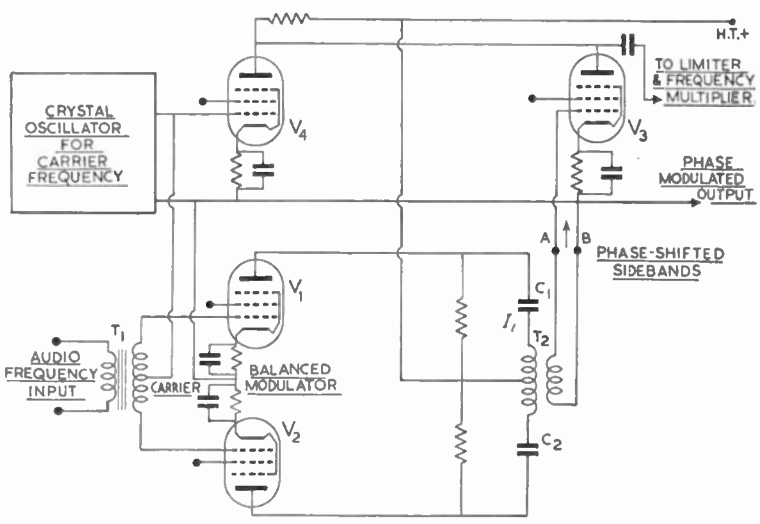

FIG. 14. *Circuit diagram showing production of phase modulation*

applying a tone-modulating source and listening on a receiver, designed for amplitude-modulation reception, and tuned to the carrier frequency. The latter is heterodyned and the resultant beat note is passed through a filter. As the modulating voltage (and frequency deviation of the carrier) is increased, there is a succession of zero amplitudes of the beat frequency as shown in figure 5.[33] Thus, if the modulating frequency is 1,000 c/s, zero amplitudes occur for frequency deviations corresponding to 2405, 5520, 8654 c/s, etc. For monitoring purposes, a peak voltmeter may be used across the output of a frequency-modulation receiver and calibrated in terms of frequency deviation as measured by the first method. Gas-filled relay valves may be connected across the receiver output to trigger an alarm when the positive or negative peak of the audio-frequency output exceeds an amplitude corresponding to the maximum deviation of 75 kc/s.

A cathode-ray tube monitor[9] can be made by connecting the modulating voltage to the horizontal deflecting plates, and the transmitted carrier to one of the vertical plates, thus producing a band across the screen. If two radiofrequency voltages corresponding to the carrier instantaneous frequency at the limits of frequency deviation ($f_c \pm 75$ kc/s), are inserted in the other vertical plate, a peak occurs in the band when the carrier deviation brings the instantaneous carrier frequency to, or beyond, these limits.

Chapter IV

THE FREQUENCY-MODULATION RECEIVER[53]

IN the second chapter, it was stated that frequency modulation could not be satisfactorily employed for short-wave operation because selective fading, caused by multipath transmission, resulted in an almost unintelligible signal at the receiver. Its sphere of usefulness is limited to the ultra-short-wave band where, communication is mainly by the direct ray. Radio-frequency amplification is very limited at ultra-high frequencies and the superheterodyne method of reception is essential in order to achieve sufficient overall amplification.

A schematic diagram of the probable form of receiver is shown in figure 15, and we see that it differs from that of an amplitude-modulation receiver only in the inclusion of the limiter and frequency-amplitude converter stages. The dipole aerial is connected to a radio-frequency stage followed by a frequency changer with a local oscillator. A series of intermediate-frequency stages, the output of which supplies the limiter,[15] is after the frequency changer. A converter changing the frequency modulation into amplitude modulation follows the limiter, and the amplitude-modulated wave after detection forms the input to a high-fidelity audiofrequency amplifier covering a range from 30 to 15,000 c/s. The radio-frequency, frequency-changer, and local-oscillator stages will be practically identical with those for amplitude modulation at the same carrier frequency, but because of the comparatively wide pass-band of ±100 kc/s which is needed to accept the carrier deviation, a higher intermediate frequency is required. The limiter stage is necessary to suppress amplitude changes of the carrier which are caused by transmission variations and noise because the

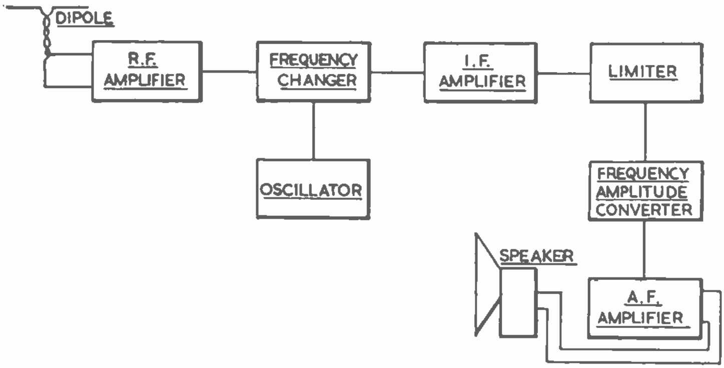

FIG. 15. *Diagrammatic layout of frequency-modulation receiver*

detector following the frequency-amplitude converter is sensitive to amplitude variations. There is no need to comment on the frequency-amplitude converter. It is obviously required because the character of the original audiofrequency voltage modulating the carrier is that of amplitude variation. The audiofrequency stages will be identical in design to those used for high-fidelity amplitude-modulated transmission.

We will now turn to a more detailed examination of the various stages.

THE AERIAL

The aerial consists of a dipole so dimensioned as to act as a half-wave resonant aerial at the center of the band of frequency-modulated transmissions. Its over-all length is about 5 per cent. This is less than the wave length of the resonant frequency, since end effects cause its electrical length to be always greater than its actual length. If a reflector is used, it is usually spaced about one-quarter of the wave length away from the aerial and it may be a half wave length long or greater. A length greater than a half wave length helps to give a more constant response[46] over the wave band. The center of the dipole is taken to a center-tapped coil at the receiver by a twin wire feeder. Motor-car ignition interference, a serious problem on ultra-short waves, is mainly vertically polarized so that best signal-to-noise ratio is usually obtained by horizontal placement of the dipole aerial. Ultrahigh-frequency aerial design is a specialized subject which cannot be treated in detail in this book, but the bibliography ([27, 32, 35, 46, 51]) contains a selected set of references to the subject.

THE RADIO-FREQUENCY AMPLIFIER STAGE

The advantages of including a radio-frequency stage before the frequency changer are increased sensitivity, increased signal-to-noise ratio, and increased selectivity against image signal and spurious intermediate frequency responses due to interaction between undesired signals and harmonics of the oscillator frequency. The first two factors, which are inter-related, are the more important ones. The use of a limiter stage requires a high degree of over-all amplification and as most of this must be obtained in the intermediate-frequency amplifier, instability is a real danger. Additional amplification at the signal frequency, however small, (it will probably not greatly exceed three times in general-purpose receiving valves or twelve for acorn[24] valves), is desirable because it allows the intermediate-frequency gain to be reduced for the same over-all sensitivity.

Feedback of voltage developed in the inductance of the cathode-earth lead [31] through the grid-cathode capacitance of the valve and, to a less extent, electron-transit time, contribute to a low grid-input resistance component, the formula for which is

$$R_g = \frac{1}{g_m \omega^2 C_{gk} L_k} \qquad (14)$$

where g_m is mutual conductance of the valve, C_{gk} is grid cathode capacitance, and L_k is inductance of the cathode-earth lead. Taking $g_m = 3\text{mA/volt}$, $C_{gk} = 3.5\ \mu\mu\text{F}$ and $L_k = 0.2\ \mu\text{H}$, $f = 45\text{Mc/s}$, the input resistance caused by cathode inductance is $6{,}000\Omega$. An average for the general-purpose type of radio-frequency valve (including electron-transit-time damping) is $4{,}000\Omega$. Acorn valves having smaller transit times and shorter cathode-earth leads usually have an input resistance of the order of 20,000 ohms.

The heavy damping from the valve prevents the realization of a high degree of selectivity in the signal stages. This fact, together with the restricted range of ultrahigh-frequency transmission, makes it possible to consider preset tuning of the radio-frequency stage to the center of the desired range. Discrimination against adjacent transmission is achieved in the intermediate-frequency amplifier, and variable tuning is obtained by oscillator frequency adjustment. Under these conditions, optimum coupling may be used between the aerial and the first-tuned circuit. The aerial coupling then gives maximum voltage transfer, but at the same time reflects a resistance component into the first-tuned circuit equal to that already existing before the aerial is coupled to it. If we take 4,000Ω as an average valve grid-input resistance, the total parallel resistance of the first-tuned circuit cannot exceed 2,000Ω when aerial coupling is at its optimum. A value of first-circuit tuning capacitance of 20 $\mu\mu$F (valve and stray capacitances make it impossible to consider much less) gives an inductance value of 0.626 μH and a magnification Q* of 2,000/176.5 = 11.3. The off-tune frequency at which the response of the circuit is 0.707 of its maximum value is given by

$$\Delta f = \pm \frac{f_r}{2Q} = \pm \frac{45}{22.6} \backsimeq \pm 2 \text{ Mc/s} \tag{15}$$

where f_r is the resonant frequency.

Thus, such a circuit could easily accommodate transmissions covering a range from 43 to 47 Mc/s when the circuit is tuned to 45 Mc/s. It is clear that signal tuning would confer little advantage over this range, in which twenty frequency-modulated transmissions having frequency deviations not exceeding ±100 kc/s could be located. When signal-circuit tuning is employed, selectivity can

* $Q = \omega L/R_s = R_p/\omega L$, where R_s and R_p are respectively the equivalent series or parallel resistance components of the tuned circuit.

be improved by using acorn valves or by tapping the grid of the radio-frequency valve down the coil. The reflected parallel resistance across the coil is thus reduced and its Q increased. The increase in Q may more than offset the decrease in grid voltage resulting from tapping down and a net increase in sensitivity may be registered. It may be noted that for a pass-band of 200 kc/s (normal maximum frequency deviation at 45 Mc/s = ±100kc/s), formula (15) calls for a Q of $f_r/2\Delta f = 45/0.2 = 90$, and signal tuning therefore requires a considerable reduction in damping if full advantage is to be derived from it. The input resistance of the valve may be increased by including a resistance[28] in the cathode-earth lead to reduce the phase angle between the cathode voltage and the current. The insertion of a series resistance in the cathode lead to reduce grid-circuit damping from the valve requires a small capacitance of about 5$\mu\mu$F from cathode to earth in order to be effective. The inclusion of a resistance reduces the over-all gain of the stage because of negative feedback in the cathode circuit, but this is usually more than offset by the reduction in damping on the tuned circuit and the net result is a slight increase in over-all amplification.

Adjustment of tuning may be by variable air condensers, but variation of inductance is preferable since fixed condensers are less susceptible to aging and temperature effects than variable condensers, which are themselves much less stable than variable inductances. The latter consist of a few turns of copper wire with a metal plunger capable of being screwed into the coil former. The plunger, which acts as a short-circuited turn to reduce inductance, must be of high-conductivity material (copper, brass, aluminum) if it is not to alter appreciably the Q of the coil. The usual precautions appropriate to ultrahigh-frequency operation must be taken, *i.e.*, leads must be as short as possible, all earth connections taken to the same point on the chassis, and adequate decoupling by small mica condensers of electrodes normally carrying only d-c or a-c supply voltages (screens and heaters). A probable form of variable-tuned radio-frequency amplifier is shown in figure 16, which is

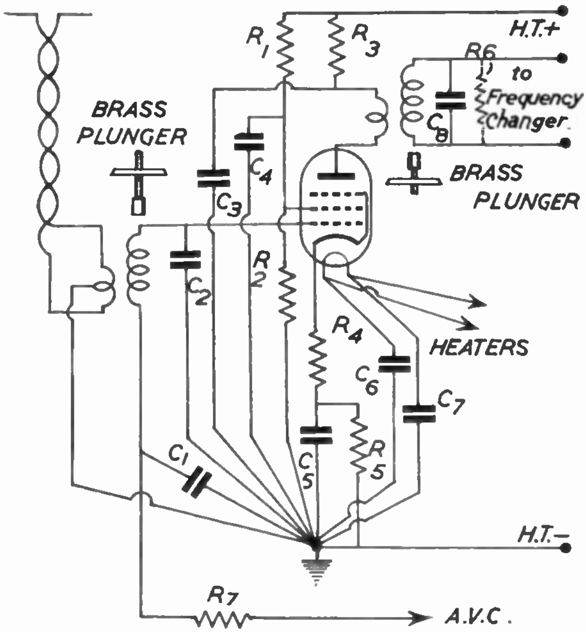

FIG. 16. *Circuit of variable-tuned radio-frequency amplifier using tuning plungers in the aerial and anode coils*

drawn to emphasize the points enumerated above. The second tuned circuit may require a damping resistance R_6 (shown dotted) if the frequency changer is a heptode with the signal grid further from the cathode than the oscillator grid. The input-resistance component of this valve is often negative[22] at high frequencies. The resistance will not be required for a hexode frequency changer which has a low positive input-resistance component comparable with that of the radio-frequency valve. Condensers C_1, C_3, C_4, C_5, C_6 and C_7 (mica, 0.01 μF) are for by-passing radio frequencies to earth. The automatic-gain-control decoupling condenser C_1 need not be as large as for an amplitude-modulated system, since feedback of any amplitude change (which should in any case be small) along the automatic-gain-control line is in such a direction as to help suppress it. C_2 and C_8 are fixed tuning condensers; R_1 and R_2 have values appropriate to the screen voltage required, generally 30,000 and 20,000Ω respectively. R_3 is a 1,000Ω decoupling resistance and R_5 a self-bias resistance of about 300Ω. R_4 (20 to 40Ω) would not actually be included in a preset-tuned amplifier operating over a range of frequencies and is the antidamping resistance which increases the input-resistance component of the valve. R_7 is an automatic-gain-control filter resistance of about 1MΩ.

THE FREQUENCY CHANGER

The frequency changer follows normal practice, a triode-hexode, or a separate triode with hexode or heptode comprising the local oscillator and frequency changer and small mica decoupling condensers and a common earth point are necessary as for the radio-frequency amplifier. The intermediate-frequency transformer in the anode circuit is damped to secure the necessary pass-band width. A typical frequency-changer oscillator circuit is shown in figure 17. Condensers C_2, C_3, C_4, C_6 and C_7 (0.01 μF mica) are for by-passing radio frequencies to earth. C_1 and C_{10} are fixed tuning condensers for signal and oscillator respectively, and C_5 and C_8 tune the primary and secondary of the intermediate-frequency transformer. Resistances R_1 and R_2 form the screen potentiometer, R_3 is damping resistance to give the required band width at the intermediate frequency, and R_4, (1,000Ω) and R_5 (300Ω) are decoupling and self-bias resistances. Details of the oscillator components and intermediate-frequency transformer are given in their appropriate sections. If a pentode acorn valve (there is no hexode or heptode type available in this series) is employed as a frequency changer, the oscillator voltage is usually applied to the suppressor grid. Cathode[38] application has been used, but is normally less satisfactory because of the higher inter-electrode-capacitance coupling between oscillator and signal circuits.

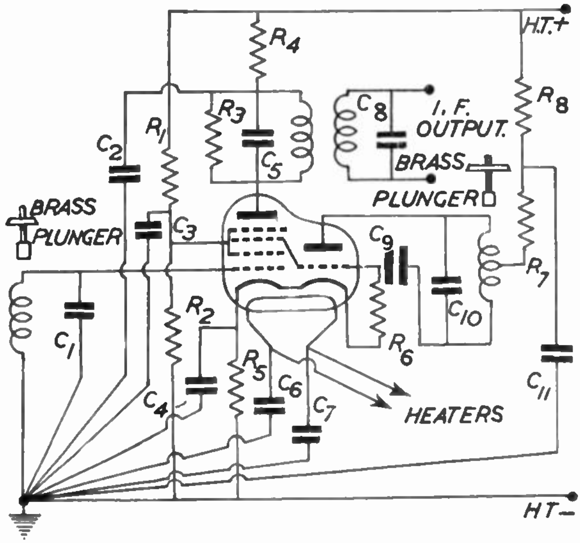

FIG. 17. *Typical frequency-changer oscillator circuit using a triode hexode*

THE OSCILLATOR

The great difficulty in ultrahigh-frequency oscillators is to obtain sufficient oscillation amplitude without squegging or dead spots in the tuning range. A modified Colpitts circuit[16, 21] (the interelectrode anode-cathode and grid-cathode capacitances act as the splitting capacitance to cathode) is often favored as it uses no separate reaction coil. The grid-coupling condenser C_9 in figure 17 is used to control oscillation amplitude. A value of about 20 $\mu\mu$F is suitable. R_6, the self-biasing grid leak, has a value of about 50,000Ω. R_7 prevents the center-tap of the coil from being earthed as far as radio frequencies are concerned. If it is short circuited, we have a Hartley oscillator. R_8 and C_{11} are decoupling components to reduce feedback from the audiofrequency stages and to smooth out variations of high-tension voltage. An alternative circuit is the electron-coupled oscillator of figure 18,[36] but this requires a separate triode-valve oscillator because there is an oscillating voltage between the cathode and the earth. It is very suitable for capacitance-tuned circuits when one side of the capacitor must be earthed. The oscillator grid of the hexode frequency changer is generally coupled by an RC circuit to the grid of the oscillator valve in figure 18. A suitable value for the grid-leak resistance in the RC circuit is 50,000Ω, and the capacitance should generally not exceed 10μμF because the oscillator-grid input of the hexode has a low input resistance, which may be sufficient to prevent oscillation altogether if placed right across the oscillator-tuned circuit. A small coupling capacitance of 10 μμF acts to increase the parallel resistance reflected from the hexode—a capacitance of 50μμF will usually cause oscillation to cease. C_3 and C_4 are radio-frequency by-passing condensers. C_2, the grid coupling condenser, has a value of 50μμF, R_1, the self-bias grid leak is 50,000Ω, and C_1 is the tuning condenser.

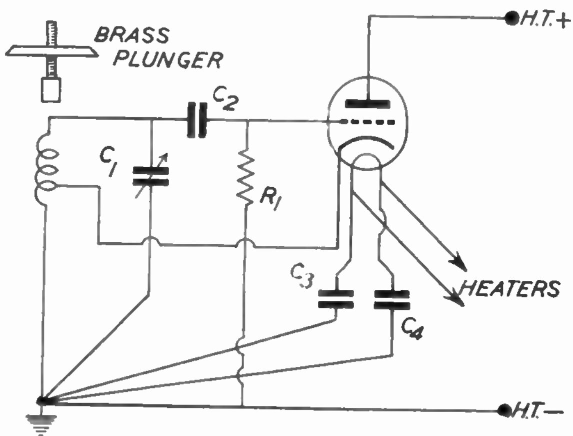

FIG. 18. *Alternative form of oscillator circuit suitable for use with one side of the condenser earthed*

The oscillator is a key point in the ultrahigh-frequency receiver and satisfactory over-all performance demands a high degree of frequency stability. Frequency error of the oscillator has a different effect on frequency modulation from that on amplitude modulation. In the latter case, unless the error is large, detuning results mainly in attenuation distortion of the audiofrequency output with accentuated high-frequency components producing high-pitched shrill reproduction. Frequency modulation of the oscillator by hum or interfering voltages has no effect since the detector is responsive only to amplitude variation. In frequency modulation, oscillator error limits the permissible frequency deviation of the carrier since it off-centers the latter with respect to the frequency-amplitude converter, and harmonic distortion of the audiooutput—flattening of the top or bottom half of the wave shape—results at high-level modulation. The action of the limiter largely prevents the attenuation-distortion effect present with amplitude modulation. Frequency modulation of the oscillator by hum, noise, etc., is a serious matter, since the converter changes this to amplitude variation and produces an audiosignal. We see, therefore, that oscillator long-period or short-period frequency stability is of much greater importance in a frequency-modulated system.

Causes of Frequency Instability

The causes of frequency instability, separating the long-from the short-period effects, will now be considered. Slow drift of oscillator frequency is caused by heat and humidity, the former generally predominating, while rapid changes are chiefly caused by high-tension supply-voltage fluctuation due to hum, mains interference, or feedback from the audiofrequency stages. The increase in tuning inductance and capacitance with increase in temperature is mainly responsible for long-period changes, although heating of the valve also contributes its quota, and the chief problem is to reduce these variations to a minimum. It is generally more difficult to produce a variable condenser with low temperature coefficient than a variable inductance, so that inductance tuning with fixed condensers of the silvered mica type is preferable. Inductance variations with temperature result from an increase of radius and length, the former increasing and the latter decreasing L. Reduced variation is therefore possible by suitably proportioning the coefficient of radial and axial expansion (radial-expansion coefficient should be about half that of axial). This can be achieved by winding the coil turns loosely on a former and fixing the ends firmly to the former so that radial expansion is determined by the conductor and axial by the former. An alternative is to reduce both axial and radial expansion by shrinking the coil on a former having low coefficients of expansion, *e.g.*, ceramic material which has a coefficient of about 7×10^{-6} as compared with copper at 17×10^{-6} and the dimensional change of a coil shrunk on a ceramic former will be largely determined by the coefficients of the ceramic material.

Capacity Compensation

Capacitance temperature changes are caused by expansion and dielectric-insulation variation. Careful mechanical design, such as accurate centering of the rotor plates of a variable condenser or the use of silvered-mica plates to reduce capacitance variation caused by pressure change in fixed condensers, can assist in controlling expansion effects, while the use of ceramic material reduces dielectric changes. Connecting leads should be short, securely fixed, and not in tension. Preliminary cyclical heating is often an aid to frequency stability. Certain types of condensers can be constructed to give a negative temperature coefficient, *i.e.*, capacitance falls as temperature rises, and they may be used to compensate for the positive temperature coefficient of tuning inductance or capacitance. Compensation is only complete, however, at one particular frequency and the temperature of the corrector condenser must follow that of the component it is intended to compensate. Thus, it is still essential to aim at the highest possible stability before applying correction.

Oscillator Stability

Valve temperature effects due to interelectrode dimensional changes (the capacitance variation is of the order of 0.02 to 0.04 $\mu\mu$F) can be reduced by loose coupling between active electrodes and the tuned circuit. The use of a harmonic of the oscillator for combining with the signal to produce the intermediate frequency helps to reduce capacitance variations in inverse ratio to the harmonic employed, *i.e.*, using the second harmonic of the oscillator as the active frequency reduces the frequency drift for the same tuning inductance to one-half. There are disadvantages of oscillator-harmonic operation since signals spaced the intermediate frequency away from the fundamental and other harmonics will produce spurious responses. Greater frequency stability may be obtained by operating the oscillator at a frequency lower than the signal by an amount equal to the intermediate frequency, and this confers no disadvantages when the signal circuits have preset tuning.

Humidity effects necessitate the use of a nonhygroscopic insulation material, such as ceramic.

Supply-voltage changes may cause slow or rapid changes of frequency. Low-tension heater change is generally comparatively slow in action affecting valve temperature, cathode emission and cathode-heater resistance and capacitance (this is important in the electron-coupled oscillator of figure 18). Loose coupling to the tuned circuit may also cloak the effect. Variation of high-tension supply controls frequency because of its effects on the mutual conductance and internal resistance of the valve, and it is largely responsible for frequency-modulation troubles. Adequate decoupling and smoothing is essential to stability, and feedback from the audiofrequency amplifier is lessened by using a push-pull output stage to the loudspeaker or even a separate high-tension supply.

If point-to-point communication or broadcasting at a fixed frequency is employed, the oscillator can be controlled by a crystal, and frequency stability adequate for normal purposes will be obtained. Alternatively, frequency drift can be corrected by automatic frequency correction from a variable-reactance valve connected across the oscillator-tuned circuit and provided with a d-c correcting bias from the frequency discriminator (chapter VI) of the frequency-to-amplitude converter.

Distortion in Frequency-Modulated Reception

Attentuation distortion caused by too narrow a receiver pass-band can occur in frequency-modulated reception if the receiver pass-band is narrower than the maximum modulating frequency. Thus, for a frequency deviation of $\pm$ 5 kc/s and a maximum modulating frequency of 15 kc/s, a receiver pass-band of $\pm$ 5 kc/s causes attenuation distortion of the modulating frequencies exceeding 5 kc/s. This type of distortion is unlikely to be a problem with wide-band frequency-modulated transmission.

Harmonic distortion may occur in a frequency-modulation receiver, but the cause is not nonlinear operation of the valves as in amplitude modulation. The cause is nonlinear phase change with frequency change over the pass range of the amplifier circuits. Amplitude change over this range is removed by the limiter, but the former is generally accompanied by the nonlinear phase change. A nonlinear frequency-change or phase-change characteristic in the frequency-to-amplitude converter can produce harmonic distortion. Only the detector and subsequent audiofrequency amplifier valves can cause harmonic distortion. The importance of centering the carrier in the pass range of the receiver and of preserving stability of the L and C components of the tuned circuit will, therefore, be apparent.

Chapter V

THE FREQUENCY-MODULATION RECEIVER

(*Continued*)

INTERMEDIATE-FREQUENCY AMPLIFIER[52]

THE actual value of the intermediate frequency must naturally be settled first. The comparatively wide pass-band required (200 kc/s) limits the minimum intermediate frequency to 2 Mc/s, but the question is whether or not a higher value would be preferable. The lowest possible value of intermediate frequency has the advantages of greater amplification and selectivity with stability, but the possibility of spurious responses is more serious. Spurious responses, generated by the frequency changer, are given in the order of their importance.

(1) The image interference is caused by interaction between the local oscillator and an undesired signal at a frequency as much above or below the oscillator frequency as the desired signal is below or above. If image response is only likely to be serious over a given band of frequencies covering the desired signal, it can be avoided by making the intermediate frequency at least half of this band, *e.g.*, if we assume frequency-modulated transmissions to cover a band from 40 to 50 Mc/s, an intermediate frequency above 5 Mc/s will prevent image interference from transmissions in this band.

(2) Oscillator-harmonic response results from combinations of oscillator harmonics and undesired signals.

(3) Signal-harmonic response results from the interaction of undesired signal-frequency harmonics with the oscillator fundamental.

(4) Signal-harmonic and oscillator-harmonic combinations may also occur. Interaction between equal harmonics of both, *e.g.*, second harmonic of signal and oscillator, is likely to be more serious than unequal harmonics since the former are nearer to the desired signal.

(5) Intermediate frequency harmonic response caused by the desired signal being close in frequency to an intermediate-frequency harmonic may also interfere. It is usually caused by feedback along the automatic-gain-control or high-tension line or by stray coupling between the limiter or detector and the aerial.

(6) Direct intermediate-frequency response, caused by a signal at the fundamental or submultiple of the intermediate frequency, the latter being converted to the intermediate frequency by the frequency-changer stage.

(7) Interaction between undesired signals separated by the intermediate frequency.

(8) Cross modulation may occur.

A high value of intermediate frequency assists in reducing interference from 1, 2, 3, 4, and 7 because it removes the interfering signal farther from the desired and allows radio-frequency selectivity to be more effective. Interference from 5 and 6 is increased by raising the intermediate frequency, but the effect of 5 can be mitigated by adequate intermediate-frequency decoupling of the limiter anode circuit, the detector-first audiofrequency amplifier connection and automatic-gain-control line. It is not likely to be serious since the probable maximum value of intermediate frequency (10 Mc/s) requires fourth-harmonic feedback to cause interference in the 40 to 50 Mc/s band. Cross modulation is rarely a serious problem in the amplitude-modulation receiver and Wheeler[42] states that it has little interference capability in frequency-modulated reception.

Since the limiter requires a certain minimum input voltage (2 to 5 volts) to remove amplitude variation, the gain of the intermediate-frequency amplifier must be sufficient to bring the weakest probable signal up to the limiter-input minimum. In general, much greater intermediate-frequency amplification is required for frequency modulation than for amplitude modulation so that the maximum intermediate frequency must be limited to that value which gives the required over-all gain without approaching instability. A value between 4 and 5 Mc/s is a reasonable compromise and in subsequent calculations, we shall assume an intermediate frequency of 4.5 Mc/s. There are other methods of reducing spurious responses besides that of a high intermediate frequency: the reduction of input voltage to the frequency changer by automatic-gain-control on the radio-frequency stage and increased radio-frequency selectivity decrease effects from 3 and 4, while the reduction of oscillator voltage to the lowest level consistent with satisfactory frequency changing decreases responses from 2 and 4.

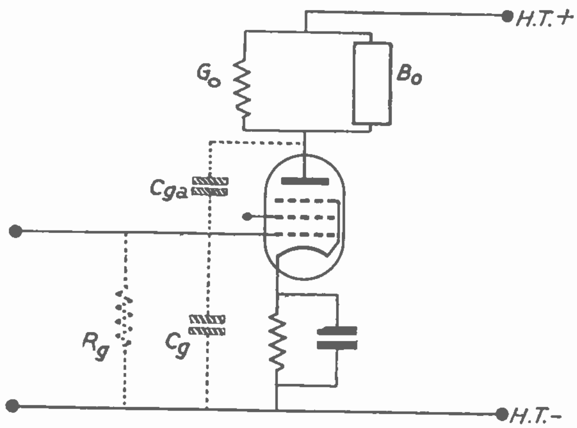

FIG. 19. *Equivalent grid-input admittance of a valve with grid-anode capacitance feedback*

The problems to be solved in the design of the intermediate-frequency amplifier are, therefore, to obtain highest over-all amplification with freedom from self-oscillation, and level pass-band with rapid attenuation outside this band. Sources of instability are input-output coupling, common impedance to the intermediate frequency in valve-electrode leads normally carrying only direct or mains alternating currents (anode high-tension supply, screen, grid bias, cathode and heaters) and grid-anode inter-electrode capacitance. The first two can be reduced to negligible proportions by suitable shielding and decoupling circuits. Common impedance coupling can largely be eliminated by connecting decoupling condensers for each stage to a common earth point as for the radio-frequency amplifier. Thus, we come to the basic fact that grid-anode capacitance feedback sets a limit to maximum over-all amplification. This feedback effect is most conveniently specified in terms of grid-input admittance, and analysis shows that the input admittance is equivalent to a resistance and capacitance in parallel (figure 19), the approximate formulae for which are

$$R_g = \frac{G_o^2 + B_o^2}{g_m B_o \omega C_{ga}} \tag{16}$$

and

$$C_g = C_{ga}\left(1 + \frac{g_m G_o}{G_o^2 + B_o^2}\right), \text{ where } G_o \text{ is the conductance} \tag{17}$$

of the anode circuit, B_o is the susceptance of the anode circuit, and C_{ga} is grid-anode capacitance.

It should be noted that the anode susceptance B_o is positive and equal to ωC_o when the anode-circuit parallel capacitance is C_o, and negative and equal to $-1/\omega L$ when the parallel inductance element is L_o, *i.e.*, regeneration occurs when the anode circuit is inductive and instability is possible, but degeneration results when the anode is capacitive. Now a parallel-tuned circuit is inductive at frequencies below resonance and capacitive at frequencies above resonance, so that an amplifier having a tuned anode circuit tends to increase the amplitude of frequencies in its grid circuit below its resonant frequency and decrease those above it. If, therefore, a similar tuned circuit is supplying the input signal, the otherwise symmetrical over-all frequency response is given an asymmetrical character with low frequencies "boosted" and high frequencies depressed as shown in figure 20. This distortion of the frequency response occurs before the amplifier reaches an unstable condition and maximum usable amplification is, therefore, limited to a value very much less than that causing self-oscillation. The actual value must be such that the minimum negative resistance component, R_g (*min*), of the grid-input admittance is at least ten times the parallel-resistance component of the grid-tuned circuit if frequency distortion is to be negligible. The minimum resistance R_g is obtained by differentiating expression (16) with respect to B_o and equating to 0, from which

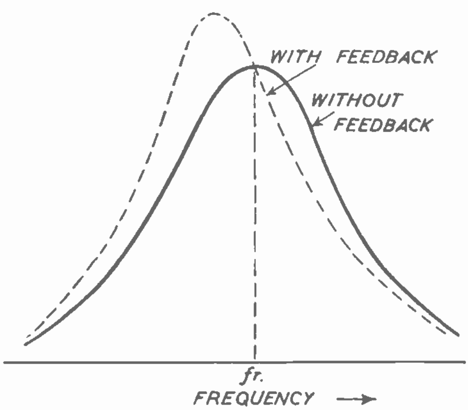

FIG. 20. *Effect of grid-anode coupling on the overall frequency response of an amplifier with single-tuned circuits in grid and anode*

$$B_o = \pm G_o \tag{18}$$

B_o is treated as the variable since it changes rapidly in the region of resonance, from a high negative value below, through zero at resonance, to a high positive value above it. G_o, over the same frequency range, remains practically constant and equal to the reciprocal of the dynamic impedance ($R_{DO} = \omega_r L_o Q_o = Q_o / \omega_r C_o$, where ω_r is the resonant angular velocity) of the anode tuned circuit. Thus, the minimum value of resistance is

$$R_g\ (min) = \frac{2G_o}{g_m\,\omega C_{ga}} = \frac{2}{g_m \omega C_{ga} R_{DO}} \tag{19}$$

and this must be at least $10 \times R_{DI}$, where R_{DI} is the dynamic impedance of the input-tuned circuit. If instead of single-tuned circuits, we have double-tuned transformers, calculation may be based on the assumption that coupling is never less than critical, and under these conditions one circuit reflects into the other a resistance equal to its initial resistance, *i.e.*, the actual dynamic impedance is half that of one tuned circuit alone and expression (19) becomes

$$R_g\ (min) = \frac{4}{g_m \omega C_{ga} R_{DO}} \tag{20}$$

To obtain a flat frequency response over the pass-band, it is necessary to combine single-tuned circuits with overcoupled circuits with double-peaked response,[12] the peak of the single cir-

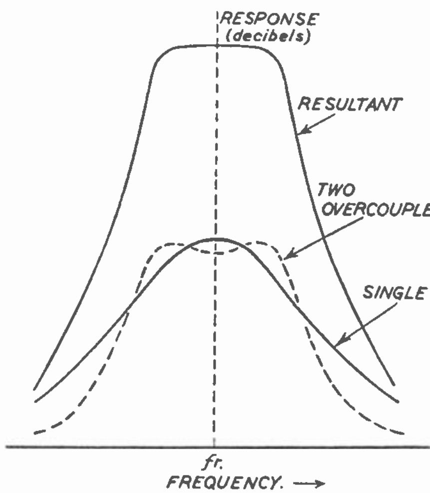

FIG. 21. *Addition of the frequency response of a single-coupled and two over-coupled circuits to give a flat pass-band response*

uit filling in the trough of the two overcoupled circuits as shown in igure 21. The combination of a pair of overcoupled circuits with a air of under-coupled or critically-coupled circuits with single-eaked response has much the same effect and we shall use these rinciples in the design of the intermediate-frequency amplifier, he diagram of which is given in figure 22. V_1, the frequency-hanger valve, has a pair of critically-coupled tuned circuits in its node circuit, the first and third intermediate-frequency amplifier alves (V_2 and V_4) a pair of overcoupled circuits and the second V_3) a single-tuned circuit. Three stages of intermediate-frequency mplification are employed as this is about the minimum number or adequate gain. By using the generalized curves developed by Beatty[8] for the frequency response of single-tuned and double-uned circuits and assuming that the primary and secondary ircuits of T_2 and T_4 are identical, we find that an almost flat pass-and response can be obtained by combining two pairs of over-oupled circuits (T_2 and T_4) of constant $Q_2k_2=2$, with one pair of ritically-coupled circuits (T_1) of $Q_1k_1=1$, and a single-tuned ircuit (T_3), where k is the coupling coefficient (M_1/L_1 for T_1, and M_2/L_2 and M_4/L_4 for T_2 and T_4) and Q is the magnification of one f the circuits in the absence of coupling from the other. The vercoupled circuits T_2 and T_4 have maximum response at $Q_2 2\Delta f/f_r = \pm 1.8$ where Δf is the frequency off-tune from f_r, the esonant or trough frequency, and the trough-to-peak ratio is —2db. By selecting Q_2 to satisfy the above expression when $\Delta f = \pm 100$ kc/s (the maximum frequency deviation of the fre-uency modulation), we have

$$Q_2 = \frac{1.8 f_r}{2\Delta f} = \frac{1.8 \times 4.5}{0.2} = 40.5 \tag{21}$$

nd the two transformers T_2 and T_4 with this Q value will give a eak at 100 kc/s on either side of the central frequency or trough osition, at which there is a 4db loss. The frequency response of he two circuits is the dotted curve 2 of figure 23. Similarly, if Q_1

and Q_3 are chosen to satisfy the condition

$$Q_1 = Q_3 = \frac{f_r}{2\,\Delta f} = 22.5 \tag{22}$$

at $\Delta f = 100$ kc/s, we have the broken-line and full-line curves 1 and 3 with a loss of 3 and 1 db respectively at 100 kc/s off-tune, thus exactly counterbalancing the gain of 4 db due to T_2 and T_4. There

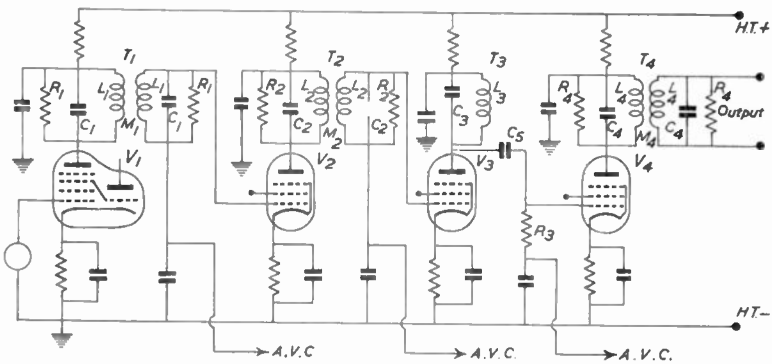

FIG. 22. *Circuit diagram for a typical intermediate-frequency amplifier*

is no exact compensation at all frequencies in the pass-band, but the variation in the over-all response curve (4 in figure 23) does not exceed 0.7 db. Having determined the Q values of the various circuits, we now have to select the L and C values to give a grid-input resistance component R_g not less than ten times the dynamic resistance of the grid circuit. In designing the amplifier we will assume that valves V_2, V_3, and V_4 are identical with $g_m = 2$mA/volt, $C_{ga} = 0.02$ $\mu\mu$F and an internal resistance $R_a \gg R_D$. Therefore, maximum gain is given by $g_m R_D$ and $g_m R_D/2$ for single-tuned and double-tuned circuits respectively. Since $R_D = Q\omega L = Q/\omega C$, and Q is fixed by frequency-response considerations, it is clear that maximum gain per stage requires minimum tuning capacitance (C), which in practice is about 50 $\mu\mu$F. Using this value and starting at the last intermediate-frequency amplifier, we find that $C_4 = 50\ \mu\mu F$; $L_4 = 25\mu\ H$, $f_r = 4.5$ Mc/s and from expression (21), $Q_4 = 40.5$, so that $Q_4 k_4 = 2$ gives $k_4 = 0.0495$ and $M_4 = k_4 L_4 = 1.235\ \mu H$.

$$R_{D4} = \frac{Q_4}{\omega_r C_4} = \frac{40.5 \times 10^{12}}{6.28 \times 4.5 \times 10^6 \times 50} = 28{,}700\ \Omega$$

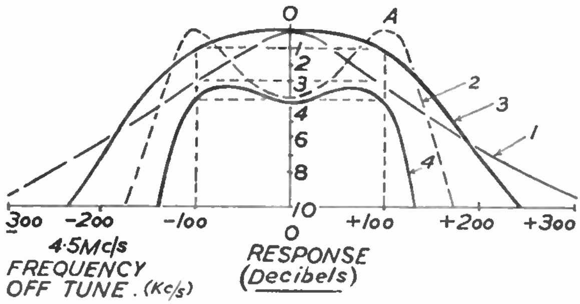

FIG. 23. *Frequency response of (1) a single-tuned circuit, Q=22.5; (2) two over-coupled circuits, Q=40.5, Qk=2; (3) two critically-coupled circuits, Q=22.5; (4) overall for (1)+(3)+2 of (2)*

Maximum gain at the frequency-response peak (A in figure 23) is

$$G_4\ (max) = g_m \frac{R_{D4}}{2} = 28.7$$

Expression (20) gives the minimum input resistance at the grid of V_4 as

$$R_{g4}\ (min) = \frac{4}{g_m \omega C_{ga} R_{D4}} =$$

$$\frac{4 \times 10^{12}}{2 \times 10^{-3} \times 6.28 \times 4.5 \times 10^6 \times 0.02 \times 28{,}700} = 123{,}200\Omega$$

Therefore, dynamic impedance of T_2 must not exceed $12{,}320\Omega$ if the frequency response is not to be seriously affected by feedback.

$\therefore\ R_{D3} = 12{,}320\Omega$ and from (22) $Q_3 = 22.5$ and $C_3 = \dfrac{Q_3}{\omega R_{D3}}$

$$= \frac{22.5 \times 10^{12}}{6.28 \times 4.5 \times 10^6 \times 12{,}320}\ \mu\mu F = 64.5\ \mu\mu F$$

$$L_3 = 19.4\ \mu H$$
$$G_3\ (max) = g_m R_D = 24.64$$

Expression (19) gives

$$R_{g3}\ (min) = \frac{2}{g_m\, \omega\, C_{ga}\, R_D} = 143{,}000\ \Omega$$

The dynamic resistance of the transformer T_2 secondary is $R_{D2}/2 = R_{D4}/2 = 14{,}350\Omega$ and this fulfils the condition that $\frac{1}{2}R_{D2}$ shall not be greater than $R_{g3}\ (min.)/10$. All circuit constants are identical with those of T^4. Therefore,

$$R_{g2}\ (min) = R_{g4}\ (min.) = 123{,}000\Omega$$

Transformer T_1 must have a dynamic impedance not exceeding $12{,}300\Omega$, *i.e.*, $R_{D1} \not> 24{,}600\Omega$. The maximum dynamic impedance is, however, fixed for us because Q_1 is to be 22.5 and C_1 not less than $50\mu\mu$F.

$$\therefore R_{D1} = \frac{Q_1}{\omega C_1} = \frac{22.5 \times 10^{12}}{6.28 \times 4.5 \times 10^6 \times 50} = 15{,}900\ \Omega$$

This value of R_{D1} cannot be exceeded without reducing C_1, but as it is less than the maximum R_{D1} required by feedback considerations, it simply means that feedback has even less effect. The constants for T_1 are, therefore,

$$C_1 = 50\mu\mu F,\quad L_1 = 25\mu H,\quad Q_1 = 22.5$$
$$M = \frac{L}{Q} = 1.11\mu H \quad (Q_1 k_1 = 1),\ R_{D1} = 15{,}900\Omega$$

$$G_1\ (max) = g_c \frac{R_{D1}}{2}$$

where g_c is the conversion conductance of the frequency changer valve V_1; a suitable value is 0.3 mA/volt.

$$\therefore G_1\ (max) = 0.3 \times 7.95 = 2.385$$

The over-all gain of the intermediate-frequency amplifier from the grid of V_2 is 28.7 (gain of V_2) $\times$ 24.6 (gain of V_3) $\times$ 28.7 (gain of V_4) divided by 1.587 (this is the ratio corresponding to the 4db loss from peak to trough in V_2 and V_4).

$\therefore$ Total gain $= 12{,}800$

or including the frequency changer, the gain from the grid of V_1 to the output of V_4 is 30,500.

No attempt has been made to specify the values of the resistances R_1, R_2, etc., shunting T_1, T_2, etc., because they will depend on the initial Q of the coils. For example, if the Q of the coils of T_1 is 150, the equivalent dynamic resistance of the tuned circuit is

$$R_D = \frac{Q_o}{\omega C} = \frac{150 \times 10^{12}}{6.28 \times 4.5 \times 10^6 \times 50} = 106{,}000\Omega$$

R_1 must be such that when paralleled with 106,000Ω, the total is 15,900Ω,

$$i.e., R_1 = \frac{106{,}000 \times 15{,}900}{106{,}000 - 15{,}900} = 18{,}700\ \Omega$$

A similar procedure can be used to find R_2, R_3, and R_4. It should be noted that R_3 is the grid leak for V_4 and that the coupling condenser C_5 (0.001 μF) is sufficiently large for R_3 to be effectively in parallel with L_3 and C_3.

The Amplitude Limiter Stage

In order to take full advantage of frequency-modulation transmissions, some circuit must be included to reduce to negligible proportions any amplitude modulation of the carrier due to noise, interference or variation in the over-all frequency response of the receiver pass-band. This is essential because the audiofrequency content of the frequency-modulation signal is extracted by means of an amplitude detector, such as a diode, after its frequency variation has been converted to an amplitude variation. An initial audiofrequency-amplitude variation of the carrier is detected at the same time and produces an undesired audio-output.

There are five possible types of amplitude limiter:

(1) A saturated amplifier, having an amplification factor inversely proportional to the amplitude of the input signal.

(2) A controlled local oscillator, locked by the frequency of the frequency-modulation input, but having an output-voltage amplitude independent of the amplitude of the controlling signal.

(3) An integrating device, having an output voltage dependent upon the frequency, but independent of the amplitude of the input signal.

(4) A negative-feedback system, which detects the amplitude modulation and uses it to supply automatic-gain-control bias to the intermediate-frequency valves to reduce envelope as well as carrier variations.

(5) A neutralizing device which detects the amplitude modulation and supplies it in reversed phase to the frequency-modulation audiooutput so as to cancel the initial amplitude variation.

The saturated amplifier is the most common type of limiter and only this form will be considered. Details of the other kinds are to be found in chapter 15, part II, of the author's book, *Radio Receiver Design.* A typical circuit for the saturated amplifier limiter is shown in figure 24. The carrier input is detected by the I_gE_g characteristic of the valve and automatic bias is produced across

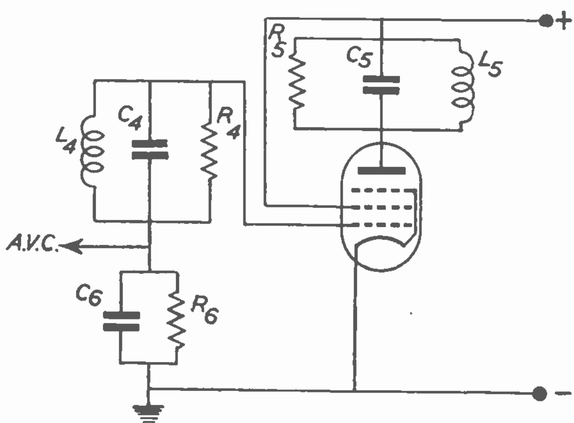

FIG. 24. *An amplitude-limiter circuit*

R_6. Any change in carrier amplitude causes a corresponding change in bias, *e.g.*, increase of carrier increases the negative bias across R_6. When the gain of the valve to carrier fundamental frequency is inversely proportional to the grid bias, and the bias voltage is a faithful reproduction of the amplitude-modulation envelope, amplitude modulation is absent from the output. This condition can be approached by operating the valve with a low cut-off ($I_a = o$) bias causing it to act as a Class C amplifier. Thus, the valve in figure 24 has low voltages on screen and anode (about 40 volts), and for satisfactory operation, a nonvariable-mu, short-grid base valve is essential. The resemblance of the limiter stage to a "leaky-grid" detector may be noted. It is, in fact, this same type of detector working under saturated conditions, with an anode circuit tuned to the carrier fundamental instead of an aperiodic circuit accepting audiofrequencies. In the "leaky-grid" detector, the time constant of the self-bias resistance R_6 and condenser C_6 must allow the bias change to follow exactly the modulation envelope and this also applies to the limiter. A suitable time constant is 10 to 20 microseconds with $R_6 = 100{,}000\Omega$ to $200{,}000\Omega$ and $C_6 = 100$ $\mu\mu$F. If the time constant is too high, the bias variation is not proportional to the amplitude modulation, and if it is too low, bias change is reduced and gain-control compensation is inadequate. Typical limiter input-output curves are shown in figure 25. The output variation is about 2.5 per cent (curve 1), 2.4 to 2.34 volts for a 2.0 to 20 input voltage change with $E_a = E_s = 36$ volts. Increasing E_a (curve 2) raises the output level, but does not change the general shape of the curve, while increasing E_s (curve 3) moves the point of level output to a higher input voltage (roughly in proportion to the increase in E_s) and tends to greater variation in output voltage. In all cases, the output voltage rises to a maximum and then falls slightly as input is increased. This is caused by the fact that the valve is distorting the input wave and producing more harmonic and less fundamental. The rate of fall of output is very largely controlled by R_6, high values of which increase the rate of fall.

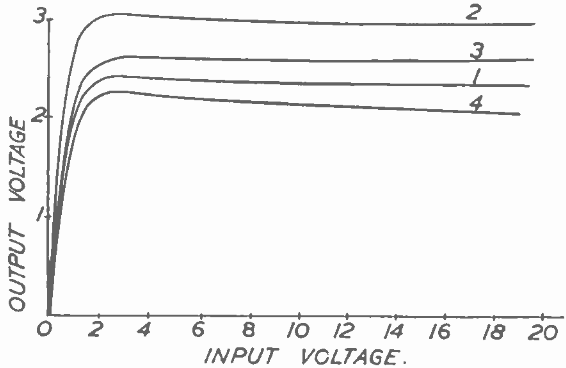

FIG. 25. *Typical input-output voltage curves for an amplitude limiter*

(Compare curves 1 and 4.) The tuned circuit in the anode of the limiter must by-pass satisfactorily harmonics of the input voltage without appreciably affecting fundamental amplitude over the pass range of 4.4 to 4.6 Mc/s (4.5 ±0.1 Mc/s). Reduction of the amplitude of frequencies at the edges of the pass-band results in harmonic distortion of the audiofrequency output from the frequency-amplitude converter detector. If the reduction is the same at each end of the pass range, the distortion consists mainly of odd harmonics (3rd, 5th, etc.). It may be noted that variations of pass-band response before the limiter are compensated by its action, but subsequent variations result in harmonic distortion of the audiofrequency output. A suitable value of Q for the limiter anode-tuned circuit is 4.5 which gives a loss of 0.1 db (representing 1 per cent change of amplitude) at 4.6 Mc/s and a loss of 19 db (representing a reduction to 1/10 amplitude) at 9 Mc/s the second harmonic frequency. The frequency-amplitude converter circuit which follows, may constitute the anode load of the limiter, or it may have a separate valve amplifier. If a very high degree of noise suppression is required, two limiters may be employed in the intermediate-frequency amplifier.

Since across R_6 in figure 24, there is a negative voltage proportional to carrier amplitude, it may be used as a source of automatic gain-control for the radio-frequency and intermediate-frequency stages of the receiver. Overloading of the frequency change can thus be prevented. Automatic-gain-control of the frequency-changer stage is not usually employed because of electron coupling between signal and oscillator circuits, any change of which causes oscillator frequency drift.

Another very satisfactory form[54] of saturated amplifier limiter uses two valves with a common unbypassed resistor in their cathode circuits. The anode of the first valve is connected directly to high-tension positive, and the output voltage is derived across a resistor or tuned circuit in the anode of the second. Limiting is caused by anode current cut-off. The first valve acts on the negative half of the input voltage and the second upon the positive half, the common cathode resistance providing phase reversal for the grid of the second valve.

Chapter VI

FREQUENCY TO AMPLITUDE CONVERSION

THE chief difference between frequency-modulated and amplitude-modulated reception lies in the method of making intelligible the audiofrequency signal conveyed in the modulation. The principle underlying the methods of detecting frequency modulation is the conversion of the frequency into an amplitude change of carrier, which is then applied to an amplitude detector such as a diode. The conversion must be accomplished in a linear manner, *i.e.*, the amplitude change is directly proportional to the frequency change, and it must also be efficient so that the resultant amplitude modulation is high. Many of the advantages of frequency modulation disappear if the frequency-amplitude conversion efficiency is low. One of the earliest methods[6] was to apply the frequency-modulated wave to a circuit off-tuned from the carrier unmodulated value. For example, a parallel-tuned circuit, connected in the anode of a tetrode or pentode valve, produces an output voltage-frequency curve as shown in figure 26, when a constant-amplitude, variable-frequency voltage is applied between grid and cathode. If this circuit is detuned (above or below) from the carrier unmodulated frequency, frequency modulation results in an output voltage whose amplitude is proportional to the frequency deviation of carrier. It will, however, only be linearly proportional if the carrier-frequency deviation is confined to the linear part AB of the curve. By applying the output voltage to an amplitude detector, such as a diode, an audio-frequency signal, corresponding to the original signal modulating the transmission, is obtained. Although variation of frequency as well as amplitude of the output voltage is occurring, it is only the amplitude change which is detected by the diode, for the latter cannot differentiate against change of frequency. It is clear that this method of detection is very inefficient, partly because it is dependent on the slope of the output voltage-frequency curve, which for practical circuits is not very high, but mainly because the tuned circuit is operated in the detuned condition where over-all amplification is low. An amplitude-modulated signal is amplified at the flat-top portion of the output curve where amplification is maximum. Furthermore full advantage cannot be taken of increased deviation of carrier for the circuit must be damped (see the dotted curve in figure 26) to increase the linear part of the curve, and conversion efficiency is thus reduced.

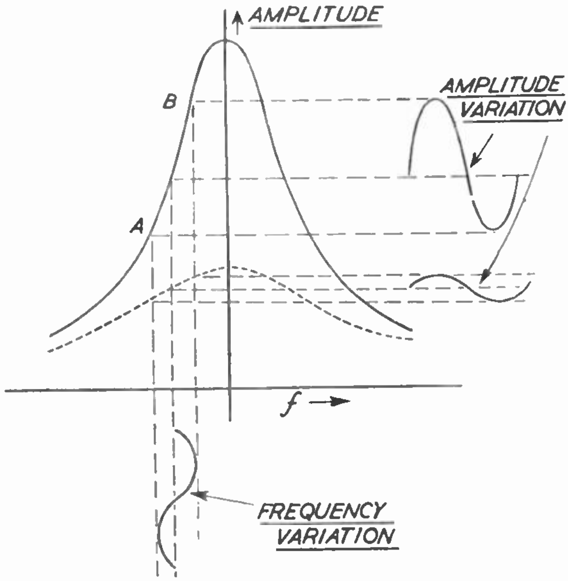

FIG. 26. *Frequency-amplitude conversion of a detuned circuit*

A second method of detection suppresses one set of sidebands. The principles involved are best understood by taking the carrier and sideband analysis set out in chapter I. There we showed that a frequency-modulated wave was represented by a carrier frequency (equal to the central unmodulated value) and pairs of sidebands spaced $\pm f_m$, $\pm 2f_m$, etc., from the carrier, the odd-numbered sideband pairs combining to give a resultant at 90° to the carrier and the even ones having a resultant in line. The addition of the first pair of sidebands to the carrier gives the frequency-(and partially amplitude-) modulated carrier of figure 27 and taking this as a

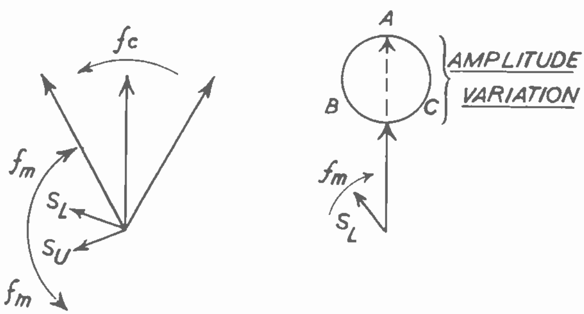

FIG. 27. *Frequency-amplitude conversion by suppression of one sideband*

basis, we see that suppression of one of the sidebands results in the mainly amplitude-modulated carrier, whose locus of operation is the circle ABC. The amplitude modulation is not directly proportional to the original frequency modulation even when all sidebands (instead of one pair) are considered and detection of the amplitude variation by a diode produces a distorted audiofrequency output containing mainly second harmonic. The suppression of one half of the sidebands is clearly inefficient since the transmitted energy in these is not used. Both disadvantages may be overcome by applying the frequency-modulated wave to two channels,[20] one containing a filter suppressing the upper set of sidebands and the other a filter suppressing the lower set. Figure 28 illustrates this method schematically. The diode detector outputs are connected in opposition so that an unmodulated carrier produces zero volts across AC. The amplitude variations of the carrier at the outputs of the two filters are in phase opposition (the upper sideband in figure 27 is subtracting from the carrier when the lower is adding) so that modulation causing the volts to rise across AB reduces the volts across BC and there is a double increase in the audiofrequency output-voltage change across AC. This phase opposition also leads to cancellation of the second-harmonic distortion in the amplitude variation, and the resultant voltage across AC is, therefore, a reproduction of the audiofrequency signal modulating the transmitter.

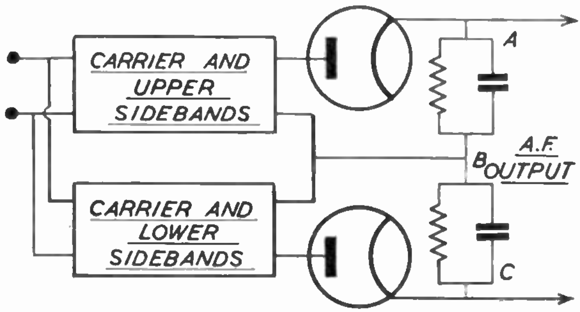

FIG. 28. *Frequency-amplitude conversion and detection by means of sideband suppression using both sidebands*

A third (the most popular) method employs a frequency discriminator similar to the one used for automatic frequency correction of the oscillator in amplitude-modulated superheterodyne receivers. In an automatic-frequency-control system, this discriminator translates intermediate-frequency carrier-frequency error into a d-c bias voltage, which is, for example, increasingly negative for frequencies below the correct carrier setting, zero at the correct setting and positive for frequencies above it. If, therefore, we apply a sinusoidal frequency-modulated carrier to such a circuit, the actual d-c bias voltage will remain constant, provided the unmodulated carrier frequency is located correctly, but there will be a sinusoidal a-c component in the output, which will correspond in amplitude to the original frequency modulation.

There are two types of discriminator, one known as an amplitude discriminator and the other as a phase discriminator, and both can be equally effective as frequency-amplitude converters.

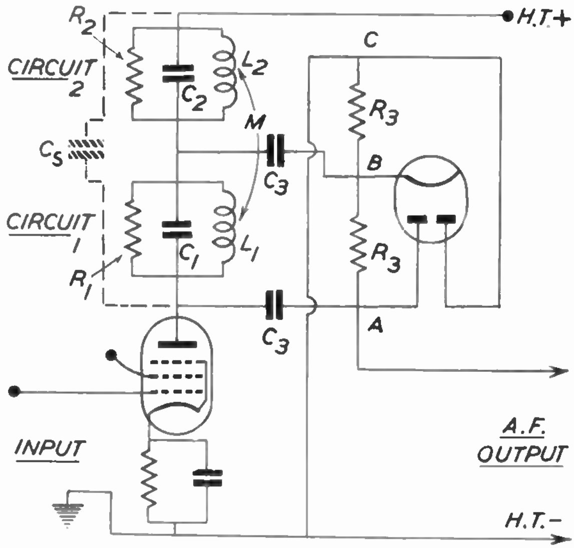

FIG. 29A. *Amplitude discriminator as a frequency-amplitude converter*

The first type consists essentially of a valve (to the grid of which is applied the frequency-modulated signal) having two series-connected anode circuits, one tuned above and the other below the unmodulated carrier or central frequency by equal amounts. The outputs from the two circuits are connected to diode detectors (figure 29a) the d-c load resistances of which are joined to give opposing voltages. Thus the total d-c voltage across AC is zero when the input voltage has a frequency equal to the central frequency midway between the resonant points of circuits 1 and 2. When the input frequency is changed to bring it closer to the resonant point of 1 the d-c voltage across AB rises and that from 2 across BC falls, *i.e.*, there is a double increase in volts across AC similar to that obtained by the second method. The action of the discriminator is illustrated in figure 29b, which shows the frequency-response curves for circuits 1 and 2. Since the detected voltages are in opposition, the active voltage is represented by the difference between the two curves, and this is indicated by the

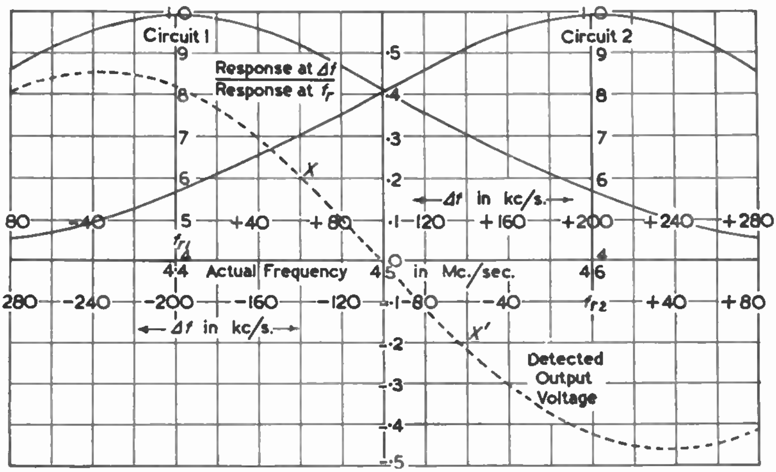

FIG. 29B. *Frequency-response curves and detected output voltage for the amplitude discriminator*

Note—The polarity of the dashed curved must be reversed to line up with figure 29A

dotted curve. By applying a frequency-modulated wave (unmodulated carrier equal to the central frequency 4.5 Mc/s) an amplitude-modulated output is obtained. The principle of operation is similar to that for the first method except that the carrier deviation is accommodated on two tuned circuits, with consequent increase in efficiency and frequency range over which the frequency-amplitude conversion is linear. The frequency response curves in figure 29a are obtained from the generalized curve of a single-tuned circuit due to Beatty,[8] who shows that the ratio of maximum response at the resonant frequency f_r to that at any off-tune frequency Δf is given by

$$R\Delta f = \frac{\text{Response at } f_r + \Delta f}{\text{Response at } f_r} = \frac{1}{\sqrt{1 + \left(\frac{Q2\Delta f}{f_r}\right)^2}}$$

where Q is the magnification of the circuit. The horizontal frequency scale is marked in Δf, the off-tune frequency from the resonant frequencies of 1 and 2 (4.4 and 4.6 Mc/s), and in actual frequencies, but the curves are calculated on the assumption that the resonant frequency for each tuned circuit is 4.5 Mc/s, *i.e.*, the curve for circuit 1 is identical with that for circuit 2 except that it is displaced horizontally by 0.2 Mc/s. For maximum conversion efficiency, the slope XOX' of the dotted curve, and thus the slope of the response curves of 1 and 2, must be as steep as possible. However, maximum slope XOX' is not the sole criterion of performance, because the resultant amplitude variation must be linearly proportional to the frequency deviation of the carrier from its unmodulated value. Greatest range of linearity is achieved by satisfying the condition

$$Q = \frac{f_r}{2\Delta f} \tag{24}$$

where f_r is resonant frequency of either tuned circuit, $\Delta f = f_c - f_{r1}$ or $f_{r2} - f_c = 100$ kc/s, and f_c is the central (*i.e.*, carrier) frequency of the discriminator at which the d-c output voltage is zero, viz., the intermediate frequency of 4.5 Mc/s.
Thus, for circuit 1

$$Q_1 = \frac{4.4}{0.2} = 22$$

and for circuit 2

$$Q_2 = \frac{4.6}{0.2} = 23$$

In practice, the result is satisfactory if $Q_1 = Q_2 = 22.5$. The composite dotted curve in figure 29b shows a turn-over at top and bottom, and if the frequency deviation exceeds about ± 80 kc/s the positive and negative peaks of the audiofrequency output signal are flattened with consequent production of odd harmonics (3rd, 5th, etc.) If larger frequency deviations are to be accommodated, the separation of the resonant frequencies of circuits 1 and 2 must be increased and their Q value decreased.

To determine frequency-amplitude conversion efficiency we must first find the slope of the composite characteristic XOX' in figure 29b, by differentiating expression 23 with respect to Δf. Thus the slope of XOX' is

$$\text{Slope} = \frac{dR\Delta f}{d\,(\Delta f)} = \frac{\frac{\frac{1}{2}4Q^2}{f_r^2}2\Delta f}{\left[1+\left(\frac{Q^2\Delta f}{f_r}\right)^2\right]^{3/2}} = \frac{4Q2\Delta f}{f_r^2\left[1+\left(\frac{Q2\Delta f}{f_r}\right)^2\right]^{3/2}} \quad (25)$$

Conversion efficiency is best expressed in terms of peak-output voltage per 1 kc/s frequency change from the central value (4.5 Mc/s) per 1 volt peak-input at the grid of the discriminator valve. It is obtained by multiplying expression (25) by $2g_m R_D n_d$; the factor 2 is necessary since expression (25) gives the slope of the response of one circuit alone and the composite dotted curve, which is the difference between the slopes of 1 and 2, is twice that of either. R_D is the resonant impedance of either circuit in figure 29a, including damping from its diode and diode load resistance R_3, *i.e.*, each circuit is damped by the resistance R_3 [C_3 is a coupling condenser (0.0001 μF) of low reactance] in parallel with $R_3/2$, that caused by diode conduction current, giving a total damping resistance of $R_3/3$. The voltage-detection efficiency n_d of the diode detectors is the ratio of the audiofrequency voltage-output change to the amplitude-modulated carrier input-amplitude change. An average value for this is 0.85.

The value of R_D is limited by the anode-earth capacitance, which forms a coupling reactance between the two circuits 1 and 2, and modifies their frequency response. For example, an anode-earth capacitance of 10 $\mu\mu$F has a reactance of 3540 ohms at 4.5 Mc/s, so that we cannot allow R_D to exceed about 2000 ohms. Assuming $g_m = 2m\text{A/volt}$, $R_D = 2000\Omega$ and $\eta_d = 0.85$, conversion efficiency

$$= 2 \times 2 \times 10^{-3} \times 2000 \times 0.85 \times \frac{4 \times 22.5^2 \times 100}{(4.5 \times 10^3)^2 \left[1 + \left(\frac{22.5 \times 200}{4.5 \times 10^3}\right)^2\right]^{3/2}}$$

$$= \frac{2.72 \times 10^2 \times 22.5^2}{(4.5 \times 10^3)^2 \times 2^{3/2}} = 0.024 \text{ peak volts per kc/s off tune per 1}$$

volt peak input.

The stray-capacitance coupling between the circuits can be neutralized by providing mutual inductance coupling between the coils (see M in figure 29a), so that

$$\frac{M}{\sqrt{L_1 L_2}} = \frac{C_s}{\sqrt{C_1 C_2}}$$

The direction of M is important, and it should be positive according to the convention[52] adopted for M as a shunt arm in the equivalent T network and not according to the convention[15] adopted with regard to the series arm. In a negative direction M adds to the coupling due to C_s and makes the frequency response even worse.

In calculating component values for circuits 1 and 2, we may assume that they are identical except for C_1 and C_2, which are adjusted for resonance at 4.4 and 4.6 Mc/s respectively, and also that the error involved in taking the resonant frequency of both circuits as 4.5 Mc/s is negligible. Thus

$$L_1 = L_2 = \frac{R_D}{\omega_m Q} = \frac{2000}{6.28 \times 4.5 \times 10^6 \times 22.5} = 3.14\mu\text{H}$$

$C_1 = 416\mu\mu\text{F}$ ($f_{r1} = 4.4$ Mc/s)

$C_2 = 380\mu\mu\text{F}$ ($f_{r2} = 4.6$ Mc/s)

Assuming R_3 to be 0·1 MΩ, the resistance contributed by the diode circuits is $\frac{1}{3}R_3$, *i.e.*, 33,333 ohms, so that additional damping resistances of approximately 2130 ohms are required across each circuit to bring R_D down to 2000 ohms.

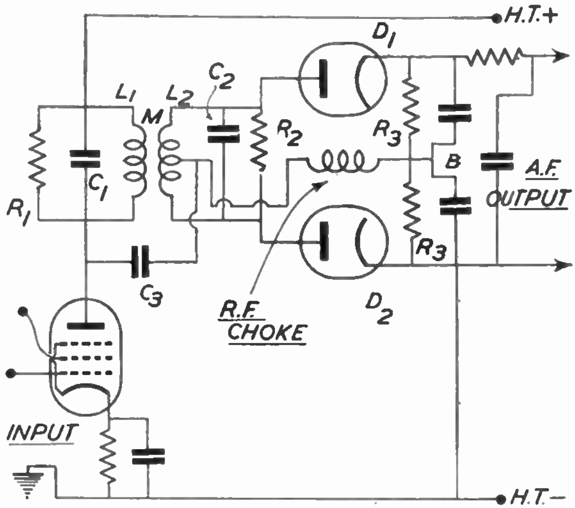

Fig. 30. *The phase discriminator as a frequency-amplitude converter*

The phase discriminator,[19] a later development, uses the fact that the voltage across the secondary circuit of a double-tuned transformer is 90° or 270° out of phase with that across the primary at the resonant frequency to which both are tuned. In one of its simplest forms (see figure 30), the secondary is center-tapped and half its voltage in series with the primary voltage is applied to one diode D_1 and the other half also in series with the primary voltage is applied to a second diode D_2. The primary voltage is developed across the radio-frequency choke between the center tap of the secondary and the center point of the diode-load resistances R_3 and R_3 by means of the coupling condenser C_3. The d-c outputs of the two diodes are connected in opposition, and an over-all frequency-response curve, similar to the dotted composite curve of figure 29b, is obtained when both primary and secondary are tuned to the central frequency, f_c. An understanding of the operation of the phase discriminator giving two voltage peaks off-tuned from the resonant frequency of the primary and secondary circuits is best gained by reference to the vector diagram of figure 31. The primary voltage vector is E_1 and the half-secondary-voltage vectors $\pm E_2/2$ are shown in phase opposition because of the center-tap. At resonance, the primary and secondary vectors are at right angles, but for frequencies above and below f_c, the two secondary vectors are tilted either to positions $\pm E_2'/2$ or to positions $\pm E^{2''}/2$. The amplitudes of the secondary voltages decrease as the off-tune frequency increases as shown in figure 31. The primary-voltage vector amplitude also changes when the off-tune frequency increases, and for couplings approaching critical value or greater, it increases at first as the frequency departs from the resonant value of primary and secondary.

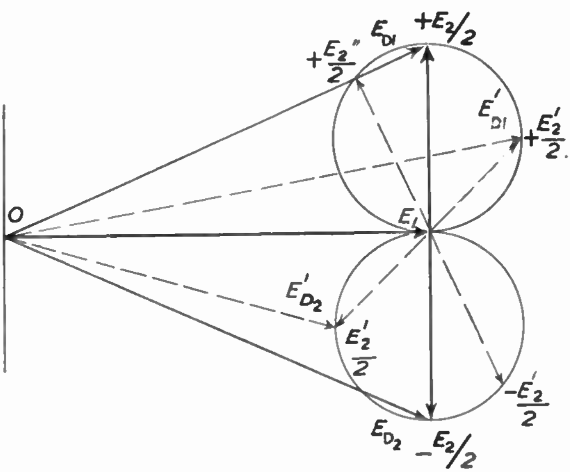

FIG. 31. *Vector diagram of phase discriminator*

This means that the frequency response of the primary is double-humped with peaks on either side of resonance, and the frequencies at which they occur move further from resonance as the coupling is increased between primary and secondary. If the amplitude of the primary voltage vector E_1 in figure 31, is assumed to remain constant as the off-tune frequency is increased, it is found that the output voltages from the diode detectors D_1 and D_2 have maximum values in the region of the 45° positions of the half-secondary-voltage vectors. This causes the frequency response of the input voltage E_{D1} to be similar to that of circuit 1 of the amplitude discriminator with maximum response below that of the resonant frequency (4.5 Mc/s). Conversely, the frequency response of E_{D2} is similar to that of circuit 2 with a maximum above 4.5 Mc/s. The actual off-tune frequencies at which the maxima occur are determined by the Q of the primary and secondary circuits, and also the coupling between them, decrease of Q and increase of coupling increasing the off-tune frequencies of the maxima. If E_1 remains constant, the numerical difference between the voltage vectors E_{D1} and E_{D2} is never linearly proportional to the off-tune or frequency deviation of the carrier. The difference is always less than required and the characteristic is curved as shown by curve 1 in figure 33. If, however, E_1 is allowed to increase as the off-tune frequency is increased, it increases the voltage difference between D_1 and D_2 cathodes, and results in a more linear discriminator characteristic.

A detailed analysis of the phase discriminator can be found elsewhere,[49] but the formulae involved are as follows

The diode voltages are

$$E_{D1} = \frac{g_m\, E_g\, R_{D1}\left[1 + j\, Q_2\left(\frac{2\Delta f}{f_c} + \frac{k}{2}\sqrt{\frac{L_2}{L_1}}\right)\right]}{\left(1 + j\, Q_1\frac{2\Delta f}{f_c}\right)\left(1 + j\, Q_2\frac{2\Delta f}{f_c}\right) + Q_1Q_2k^2} \qquad (26a)$$

where g_m is mutual conductance of the valve preceding the disminator, R_{D1} is resonant impedance of the primary when not coupled to the secondary, Q_1, Q_2 is over-all magnification of primary and secondary circuits, Δf is off-tune frequency from f_c, k is $\frac{M}{\sqrt{L_1L_2}}$ (coupling coefficient), and L_1, L_2 is inductance of and secondary coils.

$$E_{D2} = \frac{g_m E_g R_{D1}\left[1 + j\,Q_2\left(\frac{2\Delta f}{f_c} - \frac{k}{2}\sqrt{\frac{L_2}{L_1}}\right)\right]}{\left(1 + j\,\frac{Q_1\,2\Delta f}{f_c}\right)\left(1 + j\,\frac{Q_2\,2\Delta f}{f_c}\right) + Q_1Q_2k^2} \tag{26b}$$

The reversal of sign before the k term in the numerator is the only difference between E_{D2} and E_{D1}. The slope of the composite discriminator-characteristic curve in figure 29b at the central frequency f_c is obtained by differentiating $2E_{D1}$* with respect to Δf and then putting $\Delta f = 0$.

$$S_f = f_c = \frac{2\,g_m E_g R_{D1}\,Q_2^2 k\sqrt{\frac{L_2}{L_1}}}{f_c\left(1 + Q_1Q_2k^2\right)\left(1 + \frac{Q_2^2k^2L_2}{4L_1}\right)^{\frac{1}{2}}} \tag{27a}$$

or assuming $Q_1 = Q_2$

$$S_f = f_c = \frac{2\,g_m E_g R_{D1} Q^2 k\sqrt{\frac{L_2}{L_1}}}{f_c\left(1 + Q^2k^2\right)\left(1 + \frac{Q^2k^2L_2}{4L_1}\right)^{\frac{1}{2}}} \tag{27b}$$

Maximum range of linearity of characteristic does not give maximum slope, and optimum design, taking both factors into consideration, is given by

$$\frac{L_2}{L_1} = 1.77, Qk = 1.5, \frac{E_2}{E_1}\text{ (at resonance)} = 2$$

Replacing the above values in expression 27b we get

$$\text{conversion efficiency} = \frac{2g_m E_g R_{D1}\,\eta_d Q\,1.5\sqrt{1.77}}{f_c\left(1 + 1.5^2\right)\left(1 + \frac{1.5^2 \times 1.77}{4}\right)^{\frac{1}{2}}}$$

$$= 0.868\,\frac{g_m E g R_{D1}\eta_d Q}{f_c}$$

and the range of off-tune frequency over which the characteristic is linear is from 0 to $\pm 0.4\, f_c/Q$. Distortion is not excessive if the range is extended to $\pm 0.5\, f_c/Q$. To illustrate design features, let us

* The slope of the composite curve is twice the slope of the E_{D1} or E_{D2} frequency curve.

assume that $f_o = 4.5$ Mc/s, $g_m = 2$mA/volt, $E_g = 1$ volt peak, $n_d = 0.85$ $R_3 = 0.1$ MΩ, and that the frequency deviation of ±100 kc/s is accommodated over the range 0 to $\pm 0.5 f_o/Q$. Then

$$Q = \frac{0.5 \times 4.5 \times 10^3}{0.1 \times 10^3} = 22.5$$

$$k = \frac{1.5}{Q} = 0.066$$

For maximum conversion efficiency R_{D1} is required to be as high as possible, which means that the highest value of L_1 is needed. However, L_1 is limited by L_2, the maximum value of which is determined by the minimum value of C_2. Assuming the latter to be 50μμF

$L_2 = 25$ μH, $L_1 = 14.1$ μH, $C_1 = 88.5$ μμF,

$M = k\sqrt{L_1 L_2} = 1.252$ μH, and $R_{D1} = \omega_c$, $L_1 Q = 9000\Omega$

therefore,

$$\text{conversion efficiency} = \frac{0.868 \times 2 \times 10^{-3} \times 9000 \times 0.85 \times 22.5}{4.5 \times 10^3}$$

$= 0.0665$ volts per 1 kc/s off-tune per 1 volt peak input

The discriminator output-voltage-frequency characteristic for the above component values is shown in figure 32.

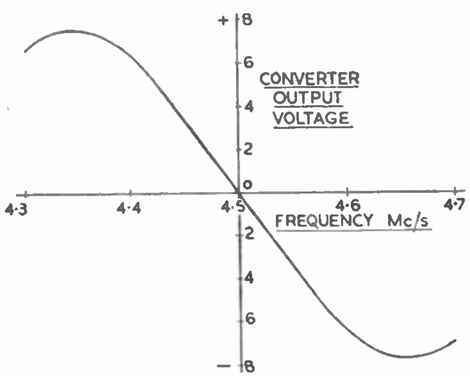

FIG. 32. *Voltage-frequency characteristic for phase discriminator of optimum design*

In calculating the damping resistances required to give the final Q of 22.5 to primary and secondary circuits, we must note that damping of the primary is produced by the conduction current of both diodes, *i.e.*, the cause of this damping is $R_3/2$, in parallel with $R_3/2$, *i.e.*, $R_3/4$. Damping from diode conduction current across the secondary is equivalent to R_3; that from each half-secondary $R_3/2$ is stepped up due to the center-tap to $2R_3$ across the complete secondary, thus giving $2R_3$ in parallel with $2R_3$ or a total of R_3 for both diodes.

Correct alignment of the discriminator is not difficult when the effect of the various factors on the characteristic is understood. Primary tuning mainly affects the symmetry of the positive and negative halves of the characteristic. A primary resonant frequency less than f_c makes the lower-frequency peak of greater amplitude than the higher. Secondary tuning controls the frequency of zero voltage, a secondary resonant frequency greater than f_c raising "zero" frequency above f_c. Mutual inductance varies the upper and lower peak frequencies moving them farther from f_c as M is increased. Adjustment is best carried out in the following manner. The coupling capacitor C_3 in figure 30 is disconnected and, with mutual inductance coupling less than critical, the primary and secondary are tuned for maximum voltage across either R_3 resistances when the frequency is 4.5 Mc/s. C_3 is now connected and the primary retuned for equal positive and negative maxima across the two resistances R_3 at approximately equal off-tune frequencies on either side of 4.5 Mc/s. The secondary is next adjusted to give zero volts across the same points at 4.5 Mc/s. Finally, the mutual inductance is increased until equal positive and negative maxima are obtained at 4.35 and 4.65 Mc/s (these correspond to the off-tune frequency peaks in the curve of figure 32). The required linear characteristic should then be obtained.

Variation of mutual-inductance coupling (Qk) causes the characteristic to pass through the phases illustrated by curves 1, 2 and 3

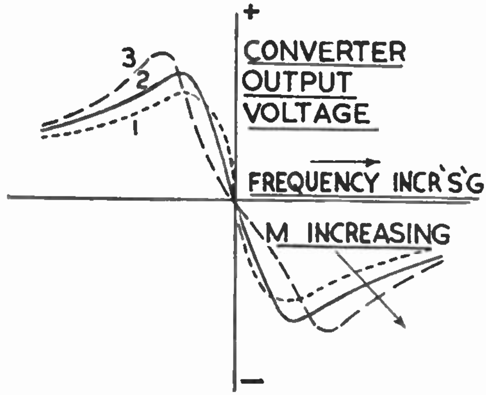

FIG. 33. *Voltage-frequency characteristic of a phase discriminator with increasing mutual-inductance coupling*

in figure 33. Curve 1 is obtained when Qk is less than unity (*i.e.*, less than critical), the linear range of the characteristic is restricted and the peaks are close to off-tune frequencies of $\Delta f = \pm 0.5\, f_c/Q$. Curve 2 illustrates the correct value of Qk (1.5) while curve 3 shows how linearity is lost when optimum value of Qk is exceeded, a double S-shaped characteristic being obtained. As Qk is increased, the positive and negative peaks continue to increase in amplitude.

The output from the phase discriminator is connected *via* a radio-frequency filter and, if necessary, a deemphasizing circuit attenuating the higher audiofrequencies to the audiofrequency amplifier. The latter should follow standard high-fidelity practice with flat frequency response from 30 to 15,000 c/s and low distortion. The design of such stages is well known and selected references are given in the bibliography.[23, 24, 25, 26, 34]

Summarizing the result of our examination of the frequency modulation, we see that there are no inherent difficulties. In the case of the receiver, an amplitude limiter and frequency-amplitude converter are the only additions required to the ultrahigh-frequency amplitude-modulation counterpart, and these added complications are more than justified by the noise-free, high-fidelity reception that is possible.

Chapter VII

THE COMPLETE RECEIVER

The circuit diagram is in the nature of a summary of the design procedure outlined in the previous chapters, and to complete the analysis, the method of calculating the component values is briefly described below. The receiver is assumed to have preset-signal circuits, tuned to a central frequency of 45 Mc/s, and having a pass-band of 43 to 47 Mc/s. An intermediate frequency of 4.5 Mc/s is employed.

The Aerial Input Circuit

The details of the aerial input circuit are as follows:

Radiation resistance of dipole aerial	$=80\Omega$
Characteristic impedance of the feeder to match with the aerial	$=80\Omega$
Total inductance of the center-tapped coil L_1 (any suitable value, preferably less than L_2 may be chosen)	$=0.28\ \mu$H
Total capacitance across the secondary coil. This is the sum of C_1, the valve, wiring, and coil self capacitances. (20 $\mu\mu$F is suggested in chapter IV, but this is too low for the high g_m radio-frequency valve Z62, which has a high input capacitance)	$=30\ \mu\mu$F
Total inductance of the secondary coil L_2	$=0.416\ \mu$H
Undamped Q of the secondary circuit Q_o	$=150$
Grid-input resistance (R_{g1}) of the radio-frequency valve Z62, at 45 Mc/s	$=3{,}000\Omega$
Total equivalent series resistance R'_{2s} of the secondary circuit, including the valve grid-input resist-	

ance $=0.785+4.52$

$= \dfrac{\omega L_2}{Q} + \dfrac{\omega^2 L^2{}_2}{R_{g1}}$ $=5.405\Omega$

The Q of the secondary circuit damped by the valve resistance R_{g1}

$= \dfrac{\omega L_2}{5.405}$ $=21.75$

Optimum coupling between the feeder and secondary circuit calls for the following value of mutual inductance M_1[44] between L_1 and L_2.

$$M_1 = \frac{Z_{a1}}{\omega}\sqrt{\frac{R'_{2s}}{R_{a1}}}$$

where Z_{a1} is total impedance of the primary circuit and is equal to $\sqrt{R^2{}_{a1} + X^2{}_{a1}}$ and R_{a1} is characteristic impedance of the feeder.

$$X_{a1} \doteqdot \omega L_1 = 6.28 \times 45 \times 0.28 = 79\Omega$$

$$\therefore Z_{a1} = \sqrt{80^2 + 79^2} = 112.3\Omega$$

$$\therefore M_1 = \frac{112.3}{6.28 \times 45}\sqrt{\frac{5.405}{80}} = 0.104\ \mu\text{H}$$

If the coils L_1 and L_2 are wound on a half-inch-diameter former with 16 British Standard wire gage at 10 turns per inch, the total number of turns is approximately 4 and 6 respectively.

Because of optimum coupling, the over-all Q of the tuned secondary circuit is halved, *i.e.*, is 10.87, so that the band width (over which the loss does not exceed 2db) is

$$\Delta f = \pm f_r/2Q = \pm\frac{45}{21.75} = \pm 2.07 \text{ Mc/s}$$

and this satisfies the requirement for preset-signal tuning.

Optimum coupling gives a voltage step up from the feeder to grid of V_1 and the expression [44] is

$$\text{Amplification} = \frac{M_1}{2C_{1T}Z_{a1}R'_{2s}} = \frac{0.104 \times 10^{-6}}{2 \times 30 \times 10^{-12} \times 112.3 \times 5.405} = 2.85$$

Automatic gain control is not applied to V_1, since change of signal grid bias affects the grid input resistance and capacitance, thus varying the band width and tuning of the secondary circuit. There is a decrease of about 4 $\mu\mu$F in the grid input capacitance from normal to maximum negative bias, and the operating input capacitance is of the order of 15 $\mu\mu$F. Wiring and coil (L_2) self-capacitance account for about 5 $\mu\mu$F so that capacitance C_1 must be about 10 $\mu\mu$F.

The Anode Circuit of the Radio-frequency Valve V_1

The preset-signal circuit leading to the frequency changer may be inserted in the anode of the radio-frequency valve, and coupled by capacitance and grid leak to the grid of V_2. This has the advantage of simplicity and generally highest stage gain from the grid of V_1 to the grid of V_2, but it has the serious disadvantage of high "stray" capacitance, since there are two valve capacitances (from V_1 and V_2) across the tuned circuit. This can be mitigated by using a smaller inductance for the signal coil, or by using transformer coupling. The first is the simpler solution, for it is difficult to obtain a very high mutual inductance between the primary and secondary of the transformer. Thus, the value of L_3 is selected to be 0.28 μH, which gives a tuning capacitance of 44.5 $\mu\mu$F. Of this total, approximately 15 $\mu\mu$F is contained in the grid input capacitance of the frequency changer, 10 $\mu\mu$F in the anode output capacitance of the radio-frequency and 5 $\mu\mu$F in stray and coil self-capacitance, leaving a capacitance for C_4 of approximately 15 $\mu\mu$F. The resistance R_4 is the grid leak of the frequency-changer valve, and it also acts as the damping resistance for providing the wide pass-band. Its value is calculated as follows We will assume that the undamped circuit Q_0 is 150, as for the aerial circuit, that the frequency-changer grid input resistance is the same as that of the radio-frequency valve, viz., 3,000Ω (this is very nearly so because the valve is a hexode), and that the internal slope resistance of the radio-frequency valve can be neglected. The over-all Q_d of the damped circuit is therefore given by

$$Q_d = \frac{\omega L_3}{\omega L_3/Q_o + (\omega L_3)^2/3{,}000} = \frac{1}{1/Q_o + \omega L_3/3{,}000} = \frac{1}{0.0066+0.0263}$$
$$= 30.4$$

To reduce Q to $Q_T = 11.25$, the value required for a pass-band of ± 2 Mc/s, the resistance R_4 is

$$R_4 = \frac{\omega L_3.Q_dQ_T}{Q_d - Q_T} = \frac{79 \times 30.4 \times 11.25}{19.15} = 1410\Omega$$

The mutual conductance of the radio-frequency valve, Z62 is 7.5 mA/volt, and the resultant dynamic impedance of the anode-tuned circuit is $\omega L_3 Q_T = 890\Omega$.

The amplification from the grid of V_1 to the grid of $V_2 =$
$7.5 \times 10^{-3} \times 890 = 6.67$

THE OSCILLATOR CIRCUIT

A separate triode valve is employed as it is generally more stable and easier to maintain in oscillation than the triode section of the triode-hexode. The anode of the triode section of V_2 is returned to cathode. The electron-coupled form of oscillator is employed because it is easy to oscillate. Also, negative feedback, caused by the portion of the tuning coil between cathode and earth, assists amplitude and frequency stability, and one side of the tuning capacitance C_{15} can be earthed. The inductance of the coil L_6 is 0.416 μH, and the cathode tapping on this six-turn coil occurs at approximately two turns up from the earthed end. The tuning capacitance is made up of the grid input capacitance of V_2 and V_3 (about 17 $\mu\mu$F), wiring and coil self-capacitance (about 5 $\mu\mu$F), the fixed capacitor C_{13} (7 $\mu\mu$F), and the series combination of C_{14} and C_{15}. C_{14} is a fixed capacitor of 30 $\mu\mu$F restricting the range of C_{15}, a variable air-dielectric capacitor with ceramic insulating supports and a maximum and minimum value of 20 and 5 $\mu\mu$F respectively. in order to obtain greatest stability, the oscillator frequency is selected to be lower than the signal frequency, and variation of C_{15} then covers the desired frequency range from 38.5 and 42.5 Mc/s.

Constant high-tension supply is an essential requirement for frequency stability and two decoupling capacitors are used from the anode of V_3 to earth. C_{17} by-passes radio frequencies and C_{18} any audiovoltages or hum voltages in the high-tension supply. Better high-tension regulation can be obtained with a gas-filled device such as a neon tube (shown dotted in the diagram), and then C_{18} becomes unnecessary. R_{12} must be reduced to about 10,000Ω.

Intermediate-Frequency Amplification

Details of the design of the intermediate-frequency amplifier were given in chapter V. We shall only state the values of the inductances and capacitances, and show the method of calculating the damping resistances required to provide the band width of ± 100 kc/s at the intermediate frequency of 4.5 Mc/s.

The First Intermediate-Frequency Transformer in the Anode of the Frequency-Changer Valve V2

Total inductance of primary or secondary (L_4 and L_5) = 25 μH

Total capacitance of primary or secondary = 50 μμF.

Required Q for ± 100 kc/s band width = 22.5

Mutual inductance between L_4 and L_5, $M_2 = 1.11$ μH

If the initial Q of the undamped primary or secondary is $Q_o = 150$, the equivalent parallel resistance of either circuit is $\omega L_4 Q_o = 106{,}000\Omega$ (assuming that the grid input resistance of the next valve V_4 can be neglected).

The extra damping resistance (R_6 and R_{10}) required

$$= \frac{106{,}000\, Q_T}{Q_o - Q_T} = \frac{106{,}000 \times 22.5}{150 - 22.5} = 18{,}700\Omega$$

The amplification of the frequency changer V_2, when $g_c = 0.3$ mA/volt.

$$= \frac{g_c \omega L_4\, 22.5}{2} = 2.385$$

Automatic gain control is not applied to the frequency-changer valve V_2 because variation of signal-grid bias varies the input resistance and capacitance of the signal and oscillator grids, causing variation of tuning and damping and, most serious of all, change of oscillator frequency.

The Second Intermediate-Frequency Transformer

This is an over-coupled transformer ($Qk=2$) with nearly twice the Q of the transformer in the frequency-changer anode circuit. The inductance and capacitance values are the same as for the first transformer.

The required $Q=40.5$.

Mutual inductance between L_7 and L_8, $M_3=1.235\ \mu$H.

Extra damping resistance is

$$(R_{15} \text{ and } R_{19}) = \frac{\omega L_7 Q_o Q_{1T}}{Q_o - Q_{1T}} = \frac{106{,}000 \times 40.5}{150 - 40.5} = 39{,}200\Omega$$

The double-humped frequency response of this transformer has a trough at 4.5 Mc/s, 2 db below the peak, thus the amplification of valve V_4 at 4.5 Mc/s is

$$\frac{g_m \omega L_7 Q_T}{2 \times 1.26}$$

[KTW63, $g_m = 1.5$ mA/volt, -2 db $= 1/1.26$]

$$= \frac{1.5 \times 10^{-3} \times 28{,}700}{2.52} = 17.1$$

Automatic gain control is applied to the valve V_4, since the input resistance and capacitance change resulting from signal-grid-bias variation is designed to be small. The automatic-gain-control decoupling capacitance C_{19} is smaller than its counterpart in the amplitude-modulation receiver since undesired amplitude change of the frequency-modulated carrier, when fed back along the automatic-gain-control line, tends to cancel the amplitude variation of the carrier. It is, in fact, a form of negative feedback.

The Isolator Stage

An isolator stage is necessary in the intermediate-frequency amplifier to reduce the regenerative and degenerative effect of the anode-grid capacitance of the intermediate-frequency valves. Its single-tuned circuit component values are:

Total inductance $L_9 = 19.4\ \mu H$

Total capacitance $= 64.5\ \mu\mu F$

Required dynamic resistance of circuit $= 12{,}320\Omega$

Extra damping resistance $R_{25} =$

$$\frac{\omega L_9 Q_o \times 12{,}320}{\omega L_9 Q_o - 12{,}320} = \frac{82{,}000 \times 12{,}320}{69{,}680} = 14{,}500\Omega$$

Amplification of valve V_5 (KTW63)

$$= g_m R_D = 1.5 \times 10^{-3} \times 12{,}320 = 18.5$$

The Last Intermediate-Frequency Amplifier Stage

This is identical with the first amplifier stage (valve V_4 and the second intermediate-frequency transformer), except that the limiter stage which follows reflects a load on to the secondary of the transformer, necessitating an increase in the secondary damping resistance R_{34}. Damping caused by grid current in the limiter valve V_7 can be taken as half the d-c load resistance, viz., $\frac{1}{2}R_{33} = 50{,}000\Omega$.

The required total secondary-damping resistance is $39{,}200\Omega$

Extra damping resistance R_{34} is

$$\frac{39{,}200 \times 50{,}000}{50{,}000 - 39{,}200} = 182{,}000\Omega$$

The Limiter Stage

The resistance R_{33} and capacitance C_{37} provide self-bias for the limiter valve V_7, which takes grid current. The combination of grid-current bias and low anode and screen voltage provides the limiting action to suppress amplitude modulation. The resistances R_{35} and R_{36} act as a potentiometer to reduce screen and anode voltages to approximately 40 volts. C_{39} is a large decoupling capacitor (8 μF) to by-pass the low-frequency components of any amplitude modulation present at the grid of the limiter valve.

The limiter anode circuit contains the phase discriminator for changing the frequency modulation into amplitude modulation.

Details of the component values for the phase discriminator have been given in the last chapter, and were as follows:

Inductance of the primary $L_{12} = 14.1\ \mu H$
Inductance of the secondary $L_{13} = 25\ \mu H$
Capacitance of the primary $C_{40} = 88.5\ \mu\mu F$
Capacitance of the secondary $C_{42} = 50\ \mu\mu F$
Q of primary and secondary $= 22.5$
Mutual inductance between L_{12} and L_{13}, $M_5 = 1.252\ \mu H$

Heavy damping is required across the primary and secondary in order to obtain a final Q of 22.5, and in the circuit diagram radio-frequency choke coupling (as shown in figure 30) is dispensed with, the center point of the secondary being connected to the junction of the diode load resistances R_{39} and R_{40}, and a single capacitance C_{43} is used from one diode cathode to the other. The primary circuit is, therefore, damped by the two load resistances in parallel as well as the diode-conduction current, *i.e.*, there is a total damping resistance of $\frac{1}{6}\ R_{39}$ (16,666Ω). If $Q_o = 150$, to reduce the over-all Q to 22.5 requires a total damping resistance of

$$\frac{\omega_c L_{12} \times 150 \times 22.5}{150 - 22.5} = 10{,}600\Omega$$

so that the additional damping resistance required is

$$R_{37} = \frac{16{,}666 \times 10{,}600}{6{,}066} = 29{,}200\Omega$$

The damping resistance across the secondary is that caused by conduction current only, viz., $0.1 M\Omega$. The total damping resistance to reduce Q from 150 to 22.5 is

$$\frac{\omega_c L_{13} \times 150 \times 22.5}{127.5} = 18{,}800\Omega$$

so that

$$R_{38} = \frac{100{,}000 \times 18{,}800}{81{,}200} = 23{,}100\Omega$$

The Detector Stage

Little comment is needed on the detector stage. The resistance R_{41} and capacitance C_{44} act as a radio-frequency filter between the detector output and first audio-frequency valve. If preemphasis (increased amplification of the high audiofrequencies contained in the modulation) is used at the transmitter, C_{44} is increased to .0005 μF in order to obtain deemphasis at the receiver. This corresponds to a preemphasis of 100 microseconds. Recent British Broadcasting Co. field tests[55] suggest 50 microseconds as a more suitable value and either R_{41} or C_{44} would have to be halved.

The Audiofrequency Amplifier

The push-pull output stage necessitates either a transformer, with center-tapped secondary in the output of the audio-frequency valve, or an extra phase-reversing valve. The latter method is adopted in this receiver on the score that RC coupling gives better frequency response than transformer coupling.

The second audio-frequency amplifier valve, V_{11}, acts simply as a phase reverser, and its input voltage is derived from a potentiometer across the output of the first valve V_{10}. The degree of tapping-down is equal to the amplification of V_{10}, and the resistances R_{49} and R_{50} are calculated as follows:

Slope resistance of V_{10} (H63) $= 66{,}000\Omega$

Amplification factor $= 100$

$$\text{Stage gain of } V_{10} = \frac{100 \times 100{,}000}{100{,}000 + 66{,}000} = 60$$

$$\text{Ratio of } \frac{R_{50}}{R_{49} + R_{50}} = 1/60$$

In the circuit diagram R_{50} is 10,000Ω and variable, so that a maximum ratio of 1/51 is obtainable. The correct position on R_{50} can be found by inserting a transformer in the high-tension lead from the center tap of the output transformer primary. Phones across the secondary of the first transformer in the high-tension lead give minimum volume when the tapping point on R_{50} is correctly adjusted.

The Output Stage

This uses two tetrodes (KT66), triode-connected to act as push-pull triode valves. The resistances R_{52} and R_{53}, in the screen circuits, guard against high-frequency parasitic oscillation, to which push-pull circuits are often very prone. A common-bias resistance equal to $\frac{1}{2}R_{54}$ may be used for the two valves if desired, and this has the advantage of reducing hum and audiofrequency negative feedback in the cathode circuit. The disadvantage is that independent bias adjustment for matching the valves is not possible. If separate bias resistances are used, matching is performed by adjusting one bias resistance to give minimum volume in phones connected to a transformer in the high-tension lead to the center-tap of the output transformer.

The anode-to-anode load for the two KT66 valves in push-pull is 2,500Ω, so that a step-down transformer ratio of approximately 13 to 1 is required if the loudspeaker speech coil impedance is 15 ohms.

APPROXIMATE AMERICAN EQUIVALENTS OF VACUUM TUBES USED

Marconi	*RCA*
Z62	6AB7
X65	6K8-G
H63	6SF5
KTW63	6KT-G
KTZ63	6J7-G
D63	6H6-GT
KT66	6L6-G

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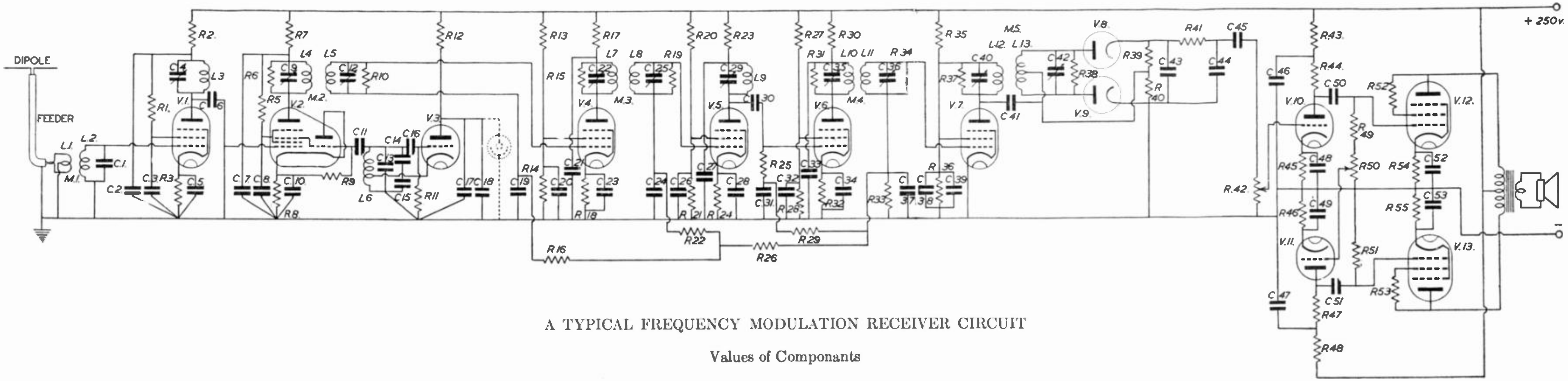

A TYPICAL FREQUENCY MODULATION RECEIVER CIRCUIT

Values of Components

(Resistances in ohms)

R 1	50,000	R20	30,000	R38	23,100
R 2	1,000	R21	20,000	R39	0.1 meg
R 3	200	R22	0.1 meg	R40	0.1 meg
R 4	1,410	R23	1,000	R41	0.2 meg
R 5	25,000	R24	300	R42	0.5 meg (variable)
R 6	18,700	R25	14,500	R43	10,000
R 7	1,000	R26	0.25 meg	R44	0.1 meg
R 8	250	R27	30,000	R45	2,000
R 9	50,000	R28	20,000	R46	2,000
R10	18,700	R29	0.25 meg	R47	0.1 meg
R11	50,000	R30	1,000	R48	10,000
R12	30,000	R31	39,200	R49	0.5 meg
R13	30,000	R32	300	R50	10,000 (variable)
R14	20,000	R33	0.1 meg	R51	0.5 meg
R15	39,200	R34	182,000	R52	100
R16	0.1 meg	R35	40,000	R53	100
R17	1,000	R36	10,000	R54	180
R18	300	R37	29,200	R55	180
R19	39,200				

Capacitances

C 1	10μμF (approx.)	C19	0.001 μF	C36	20 μμF (fixed) and 30 μμF (max., variable, air)
C 2	0.01 μF	C20	0.1 μF	C37	100 μμF
C 3	0.01 μF	C21	0.1 μF	C38	0.1 μF
C 4	15 μμF (approx.)	C22	20 μμF (fixed) and 30 μμF (max., variable, air)	C39	8 μF (electrolytic)
C 5	0.01 μF	C23	0.1 μF	C40	70 μμF (fixed) and 30 μμF (max., variable, air)
C 6	100 μμF	C24	0.001 μF	C41	50 μμF
C 7	0.01 μF	C25	20 μμF (fixed) and 30 μμF (max., variable, air)	C42	30 μμF (fixed) and 30 μμF (max., variable, air)
C 8	0.01 μF	C26	0.1 μF	C43	50 μμF
C 9	20 μμF (fixed) and 30 μμF (max., variable, air)	C27	0.1 μF	C44	100 μμF
C10	0.01 μF	C28	0.1 μF	C45	0.05 μ
C11	10 μμF	C29	30 μμF (fixed) and 30 μμF (max., variable, air)	C46	2 μF
C12	20μμF (fixed) and 30 μμF (max., variable, air)	C30	500 μμF	C47	2 μF
C13	7 μμF	C31	0.001 μF	C48	25 μF (electrolytic)
C14	30μμF	C32	0.1 μF	C49	25 μF (electrolytic)
C15	20μμF (max., variable, air)	C33	0.1 μF	C50	0.05 μF
C16	50μμF	C34	0.1 μF	C51	0.05 μF
C17	0.01 μF	C35	20 μμF (fixed) and 30 μμF (max., variable, air)	C52	25 μF (electrolytic)
C18	16 μF (electrolytic)			C53	25 μF (electrolytic)

Inductances (microhenries)

L 1	0.28
L 2	0.416
L 3	0.28J
L 4	25
L 5	25
L 6	0.416
L 7	25
L 8	25
L 9	19.4
L10	25
L11	25
L12	14.1
L13	25

Valves

V 1	6AC7, 6AJ4
V 2	6K8
V 3	6F5
V 4	6K7, 6U4
V 5	6K7, 6U4
V 6	6K7, 6U
V 7	6J7
V 8	6H6
V 9	6H6
V10	6F5
V11	6F5
V12	6L6
V13	6L6

Mutual inductances (microhenries)

M 1	0.104
M 2	1.11
M 3	1.235
M 4	1.235
M 5	1.252

E

F