FREQUENCY MODULATION

Fundamentals • Apparatus Servicing

by NATHAN MARCHAND

Consultant to New York University Cardiological Department; formerly Senior Engineer, Federal Telecommunications Laboratory; formerly Lecturer in Electrical Engineering, Columbia University



Technical Division

MURRAY HILL BOOKS, INC. NEW YORK TORONTO MURRAY HILL BOOKS, INC., is a subsidiary of RINEHART & COMPANY, INC.



COPYRIGHT, 1948, BY MURRAY HILL BOOKS, INC. PRINTED IN THE UNITED STATES OF AMERICA ALL RIGHTS RESERVED

Preface

The many applications of frequency modulation have spurred on its advancement so rapidly that it has come of age overnight. Because of this rapid advancement and the many new ideas both in fundamental thinking and in circuits it has grown into a separate field of study. The applications are many and varied. Frequency-modulation broadcasting brings noiseless, high-fidelity reception of broadcast signals into the home, in many cases into homes where reception had never been clear of interference. In those latter cases, frequency modulation has proved its worth many times over. The incorporation of frequency modulation into television broadcasting as its accompanying sound channel has further emphasized the importance of understanding the basic principles of f-m circuits. Thus, anyone building, designing, or servicing a television transmitter or receiver in its entirety must understand the fundamentals of f-m theory. The superiority of frequency modulation in mobile communication is unquestioned. Nearly every new mobile installation is now frequency-modulated and many old a-m installations are being converted to the superior f-m systems. Mobile telephone communication to private automobiles employing a system of frequency modulation is now an accepted fact. Truly, frequency modulation has come of age.

This book has been the outgrowth of many years of working and teaching in the field of frequency modulation. The emphasis is on an understanding and grasp of the fundamental principles and circuits with a minimum use of mathematics. In a subject such as this, it is very easy to lose our way in a mathematical woods and accomplish nothing. The breakdown of fundamental ideas and circuits into simple, known concepts is the major aim of the book. Nevertheless, enough formulas are given to permit fundamental design, modification, and adaption of f-m equipment. Profuse il-

v

lustrations, pictures, and diagrams have been used, since a single picture is often worth ten pages of text.

The concept of phasors (often called "plane vectors" or merely "vectors") has been used because the explanation of many of the ideas and circuits becomes very confusing without their use. However a basic discussion, in very simple language, is included as an appendix and should be read carefully by the reader if he has not got a firm grasp of the concept. It may be advantageous for any reader who has not used phasors for some time to read the appendix as a review.

The book can be divided into five parts. The first part covers the fundamentals, including the different types of modulation, fidelity, band width, noise, and interference. In this section the different types of modulation are compared in each of these aspects. The second section deals with transmitters, covering both the direct and the phase-to-frequency modulation types.

The third section is a full discussion of f-m receivers. The fundamentally different circuits, such as detectors and limiters, are examined in minute detail, while such circuits as r-f amplifiers, converters, i-f amplifiers, and audio circuits are discussed mainly to show where differences from their ordinary a-m usage and design lie. Circuit diagrams and discussions of the basic types of complete receivers are given in detail. Discussions of the various types of antennas for both transmitters and receivers are included. The fourth section is concerned with mobile equipment 1-1 receivers, transceivers, and relay equipment. Installation problems, layouts, and cable diagrams are included.

The last section is devoted to test equipment and servicing. The similarities and dissimilarities of a-m and f-m servicing are shown, and alignment, using either a-m or f-m signal generators, is covered. Servicing by means of S curves is also included. Review questions have been placed at the end of each chapter.

For the serviceman and technician this book is intended to convey first an understanding of the fundamentals and equipment operation. As many circuits as possible have been included with the detailed operation of the components to allow reasoning from causes and effects in any problem he may encounter. Enough formulas have been given to enable the serviceman to replace parts where values are unknown as well as to convert and modify various equipments.

Servicing and installation procedures, installation diagrams, and many other special features for the serviceman's use have been included. Particularly valuable is the discussion of the different types of equipment and their applications, which will allow the serviceman to expand his field, show him new applications, and in many ways help him to expand his sales activities.

For the engineer, the basic fundamentals are presented so that

PREFACE

he may grasp them without expending too much time, especially if frequency modulation is not his main field. Basic equations are given with full references, in the event that more detailed study is desired. Pictures and diagrams of actual construction are shown to enable the reduction of designs to practice.

The amateur radio operator will find the values given on the circuit diagrams as well as enough constructional detail to allow the construction of f-m apparatus for his own use.

The contents of the book has been so chosen that it may be used as a textbook in a course in frequency modulation following a fundamental course in basic radio. Answers to questions that require numerical computation are given in the back of the book. These answers should be especially useful for home study.

New York City, October, 1948

Nathan Marchand.

Acknowledgments

I am deeply grateful for the assistance given me by my many colleagues both in the field and at Columbia University in compiling the material for the book. I am very much indebted to <u>Communications</u> magazine for their permission to reprint so much material from my articles on f-m transmitters which appeared in the magazine, and I am very much indebted to Mr. Lewis Winner, editor of <u>Communications</u> and <u>Service</u>, for his encouragement and help without which I doubt very much if the book would have been written. To my wife Ernesta, my daughter Mary Ann and my son Anthony, I owe my appreciation for their patience with me while I worked into the night so many times and, of course, for my wife's invaluable help in reading proofs.

I have received a great deal of help from the various magazines and companies in the field, and I have tried to give full credit in all cases. Many of the figures were taken from or based on illustrations given to me, and the following list is a compilation of the sources with the illustrations involved:

SOURCE

Communications

Western Electric Corporation Radio Corporation of America Federal Telephone and Radio Corporation

General Electric Company Radio Engineering Laboratories

Ł

FIGURES

5-1 to 5-4, 5-7, 5-10, 6-5 to 6-7, 6-10, 6-12 to 6-17, 6-19 to 6-21, 6-23 to 6-25, 6-27, 7-2, 7-4 to 7-7, 7-16, 7-18 to 7-20, 8-3, 8-5, 8-7, 8-8, 8-16, 8-19, 11-5 to 11-9, 14-4 to 14-12, 14-14 to 14-21, 14-23, 14-24. 5-11 to 5-19, 6-11, 6-28. 6-18, 13-19 to 13-22, 14-33. 6-22, 6-26, 8-23, 8-24, 14-13, 14-26, 14-32, 16-6. 7-8 to 7-14, 14-27. 8-12 to 8-15.

ix

Electronics

<u>RCA Review</u> Galvin Manufacturing Co.

Proceedings of the IRE

Magnavox Hallicrafters Co. <u>Western Electric Oscillator</u> Bendix Aviation Corporation Radio Retailing

.

8-20 to 8-22, 9-22, 13-12 to 13-16, 17-1 to 17-9. 9-19. 11-11, 16-3 to 16-5, 16-7, 16-8, 16-10, 16-15 to 16-17. 12-7, 12-8, 13-1 to 13-9, 17-10, 17-11, 17-13. 12-9, 13-10, 13-11. 13-17, 13-18. 14-25, 16-11 to 16-14. 18-7. 18-10.

Contents

Page	P	a	ge	
------	---	---	----	--

Pr	eface	v
1.	Fundamentals of Modulation	1
2.	Fidelity and Band-Width Requirements	14
3.	Noise and Interference	37
4.	Direct Frequency Modulation	51
5.	Frequency Control Circuits	68
6.	Direct Frequency-Modulation Transmitters	92
7.	Phase-to-Frequency Modulation	129
8.	Frequency-Modulation Transmitters Using Phase Modulation	151
9.	Frequency-Modulation Detectors	180
10.	Amplitude Limiters	2 08
11.	Radio-Frequency Amplifiers, Oscillators, and Con- verters	217
12.	Intermediate Frequency and Audio Circuits	230
13.	Frequency-Modulation Receivers	251

CONTENTS

14.	Frequency-Modulation Transmitting Antennas	287
15.	Frequency-Modulation Receiving Antennas	319
16.	Mobile Frequency-Modulation Equipment	330
17.	Frequency-Modulation Test Equipment	355
18.	Frequency-Modulation Servicing	373
Ans	wers to Questions	392
App	endix	394
Inde	x	405

FREQUENCY MODULATION

.

WRH

WRH

CHAPTER 1

Fundamentals of Modulation

1-1. The Purpose of Modulation. The radio wave as we know it today is almost an indispensable part of our everyday life. From the very moment that scientists discovered that radio waves can be transmitted through space, careful investigation and research into the different methods of using the radio wave to communicate ideas and intelligence for both commercial and entertainment purposes have been pushed forward without interruption.

The first really practical system resembled the old Indian smoke signals where a column of smoke was interrupted in a recognized code, or, more nearly, the coded light signals still in use today for night communication between ships within binocular distance. Similarly, the code transmitter used in radio communication has a special code signal assigned to each letter, number, or symbol. The operator uses his contact key to turn the transmitter on and off in accordance with this code and thus uses one of the simplest types of modulation.

A broad definition of modulation is: Modulation is a process whereby some characteristic of a wave is varied as a function of the instantaneous values of another wave. Hence, in radio communication, the radio wave is modulated in some manner to transmit the intelligence that the sender desires to convey to the recipient.

The radio wave that is being modulated is called the "carrier"; it is the vehicle on which the message is being carried. In present-day terms, to avoid confusion, the carrier designates the radio-frequency (r-f) wave present when there is no modulation of the transmitted wave. The intelligence incorporated into the wave by the process of modulation is called the "modulating wave" or, for brevity, the "signal"; in sound transmission, the "audio signal." The signal may be on-and-off switching, as in the code transmitter, or a variation in the amplitude, frequency, or phase, of the carrier in accordance with a sound, television, or facsimile signal. Amplitude modulation, so adequately serving the popular broadcast needs for such a very long time, impressed many people as being the only practical method of modulation for use in broadcast radio transmission (code transmission being confined to commercial and amateur applications). Widespread recognition of the advantages of frequency modulation and its practical application to broadcast and communication systems has led to its general acceptance, all arguments opposing it having been adequately answered. Actually, the methods of modulation are many and varied. Besides the popular amplitude- (a-m) and frequency-modulation (f-m) methods, there are many others, such as phase and pulse modulation. Phase modulation is used in many f-m transmitters, while pulses are employed in one commercial f-m transmitter system for frequency regulation. For this reason, brief discussions of both phase and pulse modulation are included in this chapter.

1-2. Amplitude Modulation. For clarity, let us confine our discussion to the transmission of sound signals. The sound waves are transformed by means of a microphone into low-frequency (1-f) electrical variations which, after sufficient amplification, are used to modulate the radio wave.

Transmissions in the standard broadcast band employ amplitude modulation whereby the amplitude of the carrier wave is varied in accordance with the audio signal. For example, let us suppose that it is desired to transmit the audio signal shown in Fig. 1-1 by am-



Fig. 1-1. One cycle of an audio signal to be used to amplitude-modulate a radio wave.

plitude modulation. The radio wave, since it is at a much higher frequency than the audio signal, will go through many cycles per second (cps) during the time interval required for 1 cps of audio signal. An oscilloscope trace of the radio wave amplitude-modulated by this audio signal is shown in Fig. 1-2. The r-f power being fed into the transmitting antenna from the radio transmitter is increased and decreased, thereby causing the voltage being received at the receiving antenna to follow the contour of the audio-signal wave. The outline of the modulated wave is shown by dotted line in Fig. 1-2. We note here how the amplitude follows the original audio signal.

Inasmuch as the amplitude of the r-f wave can never become

negative (have less than zero amplitude), some provision has to be made for the transmission of the negative half of the audio-signal cycle. This contingency is taken care of by using a definite amplitude of the r-f carrier wavefor zero audio signal. The zero-modulation amplitude is indicated in Fig. 1-2 as the carrier without



Fig. 1-2. An r-f wave amplitude-modulated with the audio signal of Fig. 1-1.

modulation. When the audio signal goes positive, the r-f wave increases in amplitude above this value; when the audio signal goes negative, the r-f wave decreases in amplitude <u>below</u> this nominal value. In other words, the variation in amplitude above and below the normal carrier level constitutes amplitude modulation.

For this reason, there is an upper limit to the amount of amplitude modulation that can be impressed upon an r-f carrier wave. The amount of modulation can be increased until the amplitude of



Fig. 1-3. An a-m wave modulated to its full extent. This amount of amplitude modulation is called "100 per cent."

the r-f wave falls to zero at the bottommost point in the audio cycle. In Fig. 1-3 is shown an r-f wave modulated to its maximum extent. At the point indicated, the r-f wave actually reduces to zero amplitude. This degree of modulation is called "100 per cent modulation."

If the wave should be modulated with any larger amplitude audio signal, some of the audio cycle would be lost. In Fig. 1-4 is a trace



Fig. 1-4. An overmodulated a-m wave where the r-f signal is reduced to zero over a large part of the audio cycle. Since the r-f signal cannot go below zero amplitude, the portion of the audio signal between a and b is lost.

of an r-f wave which is overmodulated. It was modulated, or really overmodulated, with an audio signal greater in amplitude than the audio signal used for 100 per cent modulation. To transmit the audio signal in its entirety, the r-f wave would have to attain less than zero amplitude between <u>a</u> and <u>b</u>. Because this is impossible, the audio signal is lost between those two points and the result is distortion in the radiated signal.

1-3. Frequency Modulation. Frequency modulation is actually one of the earliest types of modulation known to radio engineers. It was first employed in code transmitters where, instead of turning the transmitter on and off in accordance with the coded signal, the transmitted signal was shifted from one frequency to another. A receiver, tuned to one of the frequencies, would receive the signal when the transmitter was on that frequency but would receive nothing when the transmitter frequency was shifted. In other words, the tuned circuit of the receiver changed the frequency shifting of the transmitter to regular amplitude variations that an ordinary a-m receiver could detect.

This system, although it has many advantages, does not really utilize all the advantages of f-m systems in use in the f-m broadcast and f-m communication bands. Frequency modulation itself was in disfavor among many radio engineers until Major E. H. Armstrong's paper¹ on the subject, presented before the Institute of Radio Engineers, first indicated its many advantages.

A brief summary of the various advantages of frequency modulation to be discussed in detail in various parts of the text follows:

1. The reduction of noise and interference from both natural and man-made sources.

2. The ability of an f-m station to use the same frequency as another station, provided that it overrides the interfering signal by about double the signal strength.

3. Higher transmitter efficiency.

4. Lower background noise and greater fidelity by using wide band widths.

Major Armstrong built and operated an f-m broadcast station and demonstrated the reduction in noise and disturbances when a comparatively large frequency shift, commonly called "frequency deviation," is used. Up to that time, most engineers were arguing the pros and cons of frequency modulation in terms of small frequency deviations. On the basis of small deviations, those engineers had reached the conclusion, demonstrated as false by Major Armstrong, that there were no advantages in frequency modulation.

Frequency modulation is just what the name implies. The frequency of the carrier wave is varied in accordance with the audio signal, while the amplitude of the carrier is maintained constant. This is illustrated in Fig. 1-5.



Fig. 1-5. An f-m wave and its modulating audio signal. As the audio signal goes positive, the frequency increases; as the audio signal goes negative, the frequency decreases.

The r-f wave shown in Fig. 1-5 is frequency-modulated by the audio signal shown at the top of the figure. As a reference, the unmodulated (constant-frequency) r-f wave is illustrated to the left of the f-m wave. At instant a, the audio signal is zero; the frequency of the modulated wave is the same as the frequency of the unmodulated wave. At instant b, the audio signal has increased in the positive direction in the figure; the frequency of the modulated wave is also shown to have increased. At instant c, where the audio signal has reached the maximum positive value, the modulated-wave frequency has increased to its maximum value. At instant d, the audio signal has returned to its zero value; hence, the frequency of the modulated wave has returned to the carrier frequency, the frequency of the unmodulated wave. At instant <u>e</u>, the audio signal has decreased to a negative value; the modulated wave has decreased to a frequency below that of the unmodulated wave.

Thus, in frequency modulation the frequency at any instant is varied an amount proportional to the audio signal. As an example, using actual figures, let us assume that we want to frequency-modulate a 100-megacycle carrier with a 15-v peak-amplitude audio signal. To obtain the maximum allowable frequency deviation of ± 75 kc (set for the f-m broadcast band by the FCC), we will use a transmitter which will result in a 5-kc frequency shift for each volt of audio signal.

At the instant when the instantaneous value of the audio signal is zero, the transmitted frequency will be 100 megacycles, the unmodulated frequency. When the audio signal reaches ± 10 v, the transmitted frequency will shift by 10 x 5 = 50 kc; it will be 100.050 megacycles. When the audio voltage reaches its peak value of 15 v, the transmitted frequency will increase from its unmodulated value by 15 x 5 = 75 kc; it will reach its maximum frequency of 100.075 megacycles. Following this rule of proportional frequency shifting, when the instantaneous audio voltage descends to -5 v, the transmitted frequency will decrease by 5 x 5 = 25 kc; it will reduce to 99.975 megacycles. When the audio voltage reaches its peak negative value of -15 v, the transmitted frequency will decrease by 5 x 15 = 75 kc; the transmitted frequency will reach its minimum value of 99.925 megacycles.

To illustrate the action mathematically, let us assume that $\underline{f}_{\underline{O}}$ is the carrier frequency—the center frequency—of an f-m wave. Let us call \underline{d} the frequency shift for each volt of audio-signal input (for instance, \underline{d} in the example stated in the previous paragraph is equal to 5 kc per volt of audio signal). If we let \underline{V} represent the value of the modulating audio voltage (a plus voltage increasing the frequency) at any instant, then

Instantaneous transmitted frequency = $\underline{\mathbf{f}}_{\mathbf{O}} + \underline{\mathbf{V}} \times \underline{\mathbf{d}}$ (1-1)

Instantaneous transmitted frequency = 105 - 9x0.008 = 104.928 megacycles Ans.

Example. A frequency-modulated wave is transmitted at a carrier frequency of 105 megacycles. The frequency shift for each volt of audio signal is 8 kc. What is the transmitted frequency at the instant when the audio voltage being transmitted has a value of -9 v?

Equation 1-1 can be used directly. The carrier frequency is 105 megacycles, or f_0 is equal to 105 megacycles in Eq. 1-1. The value of <u>d</u> in the equation is 8 kc and <u>V</u> is equal to -9. To convert the value of <u>d</u> into megacycles, we divide by 1,000, making <u>d</u> 0.008 megacycles. Substituting into Eq. 1-1 we get

In the case of amplitude modulation, the maximum modulation that can be impressed on the radio wave is called "100 per cent modulation." In the case of frequency modulation, the design of the apparatus and government regulations determine the maximum amount of modulation, or, more specifically, the maximum frequency deviation that is allowable. For instance, in the f-m broadcast band, the maximum allowable frequency deviation is ± 75 kc. For this reason the apparatus, both the transmitting and receiving equipment, is designed to take only the allowable deviation of ± 75 kc. If the receiving equipment were designed to accept a wider frequency deviation, it would receive extraneous interfering signals from the adjacent channels. The maximum frequency deviation—the maximum allowable swing of the transmitted frequency—is called "100 per cent modulation" for frequency modulation; anything over that deviation is called "overmodulation."

1-4. Phase Modulation. Another of the pertinent types of modulation is called "phase modulation." Although this type of modulation, in its pureform, is very little used as a system of communication, it is important in some f-m transmitters as a means of producing an f-m wave. For this reason, phase modulation will be discussed and studied.

Phase modulation is a variation in phase of the transmitted wave in accordance with the signal being transmitted. The amplitude of the wave in this type of modulation is constant. Thus, as in the case of frequency modulation, the power in the transmitted wave does not vary.



Fig. 1-6. A phase-modulated wave and its modulating audio signal. Every time the audio signal goes positive, the phase of the modulated wave advances; when the audio signal goes negative, the phase of the modulated wave is retarded.

A phase-modulated wave and its modulating audio signal is shown in Fig. 1-6. The audio signal is the sine wave shown at the top of the figure. The unmodulated carrier is shown in the lower portion of the figure as a dotted wave; it is used as the reference wave. The phase-modulated wave, shown by the solid line, has its phase shifted in reference to the unmodulated carrier--the dotted wave.

Let us trace the variation in phase with the instantaneous audio signal. At instant <u>a</u>, the voice signal increases in the positive direction; the modulated wave advances in phase by the amount indicated as 1. At instant <u>b</u>, when the audio signal voltage is still larger, the phase difference has increased to the amount indicated on the figure as <u>m</u>. As the audio signal decreases in instantaneous value, the phase shift follows it, decreasing proportionately. Thus, when the instantaneous audio-signal voltage is zero, at instant <u>f</u>, the modulated wave and the reference carrier wave are in phase. Similarly, when the audio voltage assumes a negative value, for example, at the instant <u>g</u>, the phase of the modulated wave is negative; it lags by the amount indicated as <u>n</u>. To represent accurately the audio voltage, the phase shift of the modulated wave has to be proportional to the audio or other signal being transmitted.

An interesting and useful feature is observed by carefully examining Fig. 1-6. The time interval for each cycle of the modulated wave varies with the audio signal; hence, during phase modulation the frequency of the transmitted signal is varying. It is for this reason that many engineers have argued that phase modulation and frequency modulation are one and the same thing. However, there is one major difference that cannot be overlooked. The frequency varies in phase modulation as the phase is advanced or retarded; the cycles have to stretch or compress. During this advancement or compression the frequency is changed. Thus, the frequency changes in phase modulation only when the modulating audio voltage is varied.

In the case of frequency modulation, on the other hand, the frequency of the transmitted wave is directly proportional to the amplitude of the audio signal. The difference may be clarified by means of an example. Let us suppose that we are frequency-modulating a carrier with a square-wave modulating signal. At the bottom of the square wave the frequency would have one value, while at the top of the square wave the frequency would assume another value. Thus, the square wave acts as a keying signal, shifting the frequency from one value to another. Now, in the analogous case, let us assume that we are phase-modulating a carrier with a squarewave modulating signal. At the bottom of the square wave the phase will assume one value. Since the phase is not varying, the transmitted frequency would be the carrier frequency. When the square wave jumps to its upper value, the transmitted phase will quickly assume another phase relationship and stay there for the duration of the flat top of the square wave. It is only during the shift in phase that the transmitted frequency varies, while it returns to the carrier frequency as soon as the phase relationship remains constant. The transmitted frequency will be constant at the carrier frequency during the flat portions of the square wave, in the case of phase modulation, and will change to a different frequency only during the interval that the modulating voltage is varying from one value to another. Thus, a phase-modulated wave and a frequency-modulated wave may look alike to the eye, but the frequency of one is proportional to the change in audio voltage, while the frequency of the other is directly proportional to the audio voltage.

For the numerical case let us represent the phase shift per volt in phase modulation by the Greek letter θ (theta). Calling V the instantaneous value of the modulating audio voltage, we obtain the following equation for the instantaneous shift in phase of the modulated voltage:

Instantaneous phase shift =
$$V \times \theta$$
 (1-2)

<u>Example.</u> What is the instantaneous phase shift of a phase-modulated wave at the instant when the modulating audio voltage is +8 v? The phase shift per volt for this particular transmitter is 20° .

From Eq. 1-2, where θ is equal to 20 and <u>V</u> is equal to 8,

Instantaneous phase shift = $8 \times 20 = 160^{\circ}$ leading Ans.

Because of the system design, there is usually a maximum phase shift that is allowable; as long as the phase shift is less than this value, no distortion is noticeable and the equipment functions properly. Usually, if this maximum phase shift is exceeded, distortion of some type or interference with other signals will take place. When the transmitted wave is shifted the maximum amount in phase, the wave is said to be "100 per cent modulated." With a greater phase shift, the wave is said to be "overmodulated." We note that in this case also the apparatus and not the wave limits the amount of modulation that is employed.

1-5. Pulse Modulation. The use of pulses in radio communication is not new. However, because of the use of ultrahigh frequencies and the tremendous amount of work done on pulses during World War II (especially in the field of radar), modulation by means of pulses has become practical and useful. The methods of pulse modulation are also being used for other purposes, as in the case of one type of commercial f-m transmitter where pulses are used in the carrier-frequency regulation circuits. It is for this latter reason that a short discussion of pulse modulation is included.

The ordinary pulse consists of a short burst of r-f energy. This is illustrated in Fig. 1-7a. We see how very similar the pulse shape is to an ideal dot in code transmission, the pulse being created by turning the transmitted carrier on and then off after a very short interval of time. Pulse modulation is employed most often at ultrahigh frequencies, frequencies in the hundreds or thousands of megacycles. The pulse duration or width is usually extremely short, being a few millionths of a second or even shorter than a tenth of a millionth of a second in some cases. The height of the pulse, or "peak voltage," as it is sometimes called, is normally referred to



Fig. 1-7. A pulse consisting of a burst of r-f energy. It is created like a telegraph dot by turning the transmitted carrier on and then off after a very short interval of time.

as the "amplitude" of the pulse. The pulse itself may be represented by its envelope, or shape, as shown in Fig. 1-7b. This envelope represents the voltage necessary for turning the transmitted carrier on and off to create the pulse.

These pulses, whose creation (basically) is an amplitude modulation of the r-f wave, may be used to transmit messages by vary-



Fig. 1-8. Three methods of pulse modulation. The pulses may be varied in amplitude, in repetition rate, or in time position.

ing some property of the pulse, or its relationship to other pulses. In Fig. 1-8 are shown three possible methods of pulse modulation. The audio signal to be transmitted is shown in Fig. 1-8<u>a</u>. Figure 1-8<u>b</u> illustrates amplitude modulation of pulses. Here the pulses are varied in height in accordance with the signal being transmitted. This method is more difficult to use than the other methods of pulse modulation because, besides turning the transmitter on and off to create the pulses, the strength of the transmitted signal has to be varied at the same time.

In Fig. 1-8c is illustrated one method of frequency modulation of pulses. The repetition rate of the pulses (the number of pulses per second) is varied in accordance with the audio signal. Thus, when the signal goes positive the number of pulses per second increases; and when the signal goes negative, the number of pulses per second decreases. This method is usually simpler to apply than amplitude modulation of pulses; it involves only turning the transmitter on and off at a different rate as the modulating signal varies in voltage.

Figure 1-8d illustrates the principle of what is called "pulse time" or "pulse position modulation."² This method is quite similar to ordinary phase modulation. However, a timing or reference pulse is used. The timing pulse is either wider, or different in some manner, from the signal pulses which follow. In this illustration the timing pulse was made wider than the signal pulse. The signal pulse follows the timing pulse by an interval which is determined by the audio signal being transmitted. In other words, the position of the signal pulse in relation to the timing pulse transmits the intelligence.

As the audio signal increases in a positive direction, the time interval separating the two pulses also increases. This is shown as the distance, or time interval g. As the audio signal goes negative, the time interval decreases, as shown by the time interval h. By detecting the variation in position of the signal pulse with respect to the timing pulse, the audio signal is reproduced at the receiver.

The method of pulse-time modulation can be used also to transmit more than one message at a time over the same carrier. This is called "multiplex transmission." It is accomplished by using more than one signal pulse with the same timing pulse. Enough space has to be allowed between the timing pulses so that the various signal pulses can change position without interfering with one another.

One method of multiplex pulse-time modulation is illustrated in Fig. 1-9. A four-channel multiplex system is shown. After the timing pulse, signal pulse 1 is generated. This pulse can move back and forth only between the dotted lines <u>a</u> and <u>b</u>. Signal pulse 2, the following pulse, can move back and forth only between the limits <u>b</u> and <u>c</u>; signal pulse 3, between <u>c</u> and <u>d</u>; and signal pulse 4,

between \underline{d} and \underline{e} . At e the next timing pulse occurs and the train of pulses is repeated. At the receiver the signal pulses are separated;



Fig. 1-9. Multiplex pulse-modulation transmission by the use of more than one signal pulse for each timing pulse.

the position of each one is detected, and a separate audio signal obtained for each signal pulse.



Fig. 1.-10. Pulses of frequency modulation as used to transmit an audio signal. The frequency of the radio frequency constituting the pulse is varied with the modulating signal.

Another method of pulse modulation is shown in Fig. 1-10. In this case the signal is transmitted by actually changing the frequency of the radio wave that goes into creating the pulse; it is a regular f-m signal being sent out in short bursts instead of in one continuous steady wave. In a normal f-m receiver, pulses of voltage are obtained whose amplitudes vary with the audio-signal modulation. If the number of pulses per second is high enough, the pulses themselves will be inaudible and only the undistorted audio signal will be heard at the output of the receiver. This method is employed in a system of television transmission where the voice is transmitted on the same carrier as the picture signal. The voice is sent as short bursts of frequency modulation between the transmission of the picture signals.³

REFERENCES

1. E.H.Armstrong, "A Method of Reducing Disturbances in Radio Signaling by a System of Frequency Modulation," <u>Proc. IRE</u>, May, 1936, p. 689.

2. Donald Phillips, "Pulse Time Modulation," <u>Communications</u>, November, 1945, p. 46.

3. This method is used by the Columbia Broadcasting System in color-television transmission for the simultaneous transmission of picture and sound.

QUESTIONS

1. What is the purpose of modulation in radio?

2. What is amplitude modulation and how is it created? Illustrate your answer with a diagram.

3. What is meant by 100 per cent modulation in amplitude modulation?

4. Explain how a radio wave is frequency-modulated.

5. An f-m transmitter has a carrier frequency of 104 megacycles. If 1 v of audio signal will shift the output frequency 3 kc, what will be the output frequency at the instant when the voice signal has a positive voltage of 18 v?

6. What is meant by 100 per cent modulation in frequency modulation?

7. Describe the meaning of phase modulation.

8. If 1 v of audio-signal voltage will shift the phase of a phase-modulated wave by 24 deg, how many volts are necessary to shift it 180 deg?

9. Explain what determines 100 per cent modulation in phase modulation.

10. Describe the general characteristics of pulse modulation.

11. How may more than one message at a time be transmitted over the same carrier with pulse modulation?

12. Describe the method of pulse modulation where the frequency of the radio wave constituting the pulse is varied.

CHAPTER 2

Fidelity and Band-Width Requirements

2-1. Sound. Sound waves, we know, are produced by the mechanical vibrations of material objects in an elastic medium, air being the most important of the elastic mediums. Sound waves travel through air by means of compressions and rarefactions of the air. When these sound waves reach our ears, we experience the sensation of hearing.

In sound broadcasting, the sound vibrations are picked up by a microphone and transformed into electrical variations - an audio signal. An r-f wave, modulated by this audio signal, is broadcast. At the receiver the r-f wave is selected, amplified, and demodulated. The resultant electrical variations cause the cone of a loudspeaker to vibrate and recreate the sound waves that were picked up by the microphone.

The sound wave as it appears on the screen of an oscilloscope is normally a very intricate wave. Sound waves may be analyzed, however, by considering them to be the sum of a number of waves of different frequency and different amplitude. An example of this type of analysis is illustrated in Fig. 2-1. Figure 2-1a is a representation of a complex sound wave which has many variations and resembles very little the ordinary simple sine wave. Actually the wave is not complicated to analyze. It is composed of a fundamental (Fig. 2-1b), a second harmonic (Fig. 2-1c), and a third harmonic (Fig. 2-1d), each having a particular amplitude. When the three waves are added together with the proper phase relationships, the complex wave illustrated in Fig. 2-1a is obtained. The names given to the various components are as follows: The "fundamental" is a sine wave having the same base frequency as the original complex wave. The "second harmonic" is a sine wave at double the frequency. The "third harmonic" is a sine wave at three times the frequency; and so on, till the necessary number of components are obtained. The amplitudes of these waves are such that, when added together in the proper phase relationship, the result will duplicate the original wave. Each of

14

these sine waves, as noted above, is referred to as a component of the original wave.



Fig. 2-1. Wave <u>a</u> is a complex sound wave while waves <u>b</u>, <u>c</u>, and <u>d</u> are its components, <u>a</u> fundamental, a second harmonic, and <u>a</u> third harmonic.

All analyses concerning sound waves will be made with respect to pure sine waves because, no matter what shape, frequency, or amplitude a repetitive complex wave may have, it can be duplicated by the sum of a fundamental and a series of harmonics, each having the proper amplitude and phase relationship. The waves need not be sound waves to be duplicated by means of a fundamental and a series of harmonics. Moreover, analysis of a wave by means of harmonics is not the only method that is employed. Sometimes it is expedient to analyze waves as the sums of sine waves that are not in harmonic relationship. (This latter method will be used in the discussions on sidebands.)

To amplify a complex sound wave, for instance, in a publicaddress system, the equipment should be designed as though it were to amplify each of the components separately. If it will faithfully amplify each of the components of the wave, the equipment will amplify the complete complex sound wave without distortion. If the equipment will attenuate, or amplify, some of the component waves differently than others, then it will distort the complex sound wave. In the case of sound waves, accurate phase relationships do not have to be maintained because phase distortion is undetectable by the ear.

2-2. Fidelity. One of the definitions given for fidelity in the diction is "Exactness, as in a copy." This definition can be ap-

plied accurately to the use of the term in radio. For instance, in sound broadcasting low fidelity would mean a relatively poor reproduction of voice and musical sounds. As the fidelity becomes better, the quality of the reproduced sound also becomes better, until a practically perfect reproduction of the original sound is obtained. This excellent reception is referred to as high-fidelity reproduction. It signifies the reproduction of the sound impinging on the microphone with so little distortion that a person hears the sound at the receiver with the same quality and fullness as the audience at the studio. In the so-called medium-fidelity receiver, the voices and musical renditions are reasonably good and quite enjoyable, but the reproduction lacks the true realism present in such high-fidelity systems as are found in f-m broadcasting.

The requirements for high-fidelity reproduction include low distortion of transients, adequate dynamic range, and good loudspeaker equipment, in addition to good wave-form reproduction as obtained by adequate frequency response of the equipment. This latter factor usually limits the fidelity obtainable in any system. For instance, if the fundamental frequency of the complex wave depicted in Fig. 2-1<u>a</u> were 2,500 cps, to transmit the complete wave would require transmitting the fundamental at 2,500 cps, the second harmonic at 5,000 cps, and the third harmonic at 7,500 cps.

If the system transmitted only frequencies up to 3,000 cps, none but the fundamental would be transmitted and a very poor reproduction of the original wave obtained. Even though all other requirements for high-fidelity reproduction were met, the resultant signal would still be very poor-fidelity reproduction. If the system transmitted frequencies up to 5,000 cps, the fundamental and second harmonic would be transmitted. The reproduced wave would be a better reproduction of the original wave, and the resultant signal would possess reasonably good fidelity. In a high-fidelity system, where the transmitted band width includes frequencies up to 15,000 cps, all three components would be transmitted and the wave accurately reproduced at the receiver.

Figure 2-2 shows a chart that gives the fundamental frequencies of the notes on a piano, as well as the fundamental-frequency ranges of various musical instruments and the ranges of the human voice. We must not forget, however, that it is the harmonics of these fundamental frequencies which give the sounds quality and timbre and enable us to distinguish between the same notes on different instruments.

The accoustical range of console-type a-m broadcast receivers, about 60 to 5,000 cps (a limitation imposed by the frequency separation between adjacent stations), covers a good deal of the spectrum and, for most listening purposes, serves very well. In f-m highfidelity broadcasting, where adequate band width has been allowed, the receivers cover the range from about 50 to 15,000 cps, practically the total frequency range of the sound waves impinging on the microphone. The range includes practically all upper harmonics of musical



Fig. 2-2. A chart showing the fundamental frequencies of the various keys and notes on a piano keyboard as well as the fundamental frequency ranges of the various musical instruments and the human voice.

instruments, bringing out all the full quality and timbre of the notes. Also realized are the fine intonations of the human voice which are not obtained in lower-fidelity systems.

2-3. Radio-Frequency Sidebands. In designing r-f circuits, which must amplify modulated r-f waves, a problem corresponding to that of the complex sound wave is encountered. The modulated wave is no longer a pure sine wave but includes a variation in amplitude, phase, or frequency, depending upon the type of modulation being employed. The modulated wave can also be considered to be composed of a number of ordinary sine-wave components. The components, in this case, are not in harmonic relationship but are related by the modulating voltage. They consist of a carrier and a series of sidebands which are removed only slightly in frequency from the carrier.

One of the important considerations in radio transmission is the maximum band width necessary for the transmission of the modulated wave. In other words, it is the frequency band width within which are contained all the frequencies present in the transmitted wave. In a-m broadcasting, for instance, the assigned frequency band width is ± 5 kc; hence, all the frequencies that make up a modulated wave in the a-m broadcast band must be kept within ± 5 kc of the assigned carrier frequency. This is accomplished by suitable cutoff filters at the transmitter.

In f-m broadcasting the frequency band width assigned to each station is \pm 100 kc; hence, all the frequencies that constitute a modulated wave in the f-m broadcast band must be within \pm 100 kc of the assigned frequency. This is a band width 20 times that used in a-m broadcasting.

A study of the sideband characteristics of the various types of modulation is necessary to understand the reasons for the various band widths employed. In transmission, sideband frequencies outside the limits would interfere with the signals on adjacent channels. For reception, the sidebands created at various frequencies and with various types of transmissions should be known for the design and construction of the various amplifier circuits.

2-4. Band-Width Requirements for Amplitude-Modulation Transmission. Let us review first the band-width requirements for the a-m wave. An example of this type of modulation was given in Sec. 1-2 (Fig. 1-2) of Chap. 1. The frequency and phase of the r-f wave remains fixed while the amplitude varies in accordance with the audio signal.

An a-m wave and its components are illustrated in Fig. 2-3. In Fig. 2-3<u>a</u> is shown the modulating audio signal; its frequency will be designated by <u>F</u>. The unmodulated carrier is pictured in Fig. 2-3<u>b</u>; Its frequency will be designated by <u>f</u>.

When modulation of the r-f carrier takes place, the amplitude is varied in accordance with the audio signal. In other words, the amplitude will vary (in the example) with a frequency of \underline{F} . This amplitude variation can also be obtained by adding to the unmodulated carrier two other frequencies called the "high sideband" and the "low sideband." The high sideband is a constant-amplitude sine wave at a frequency of ($\underline{f} + \underline{F}$). The low sideband is a constant-amplitude sine wave at a frequency of ($\underline{f} - \underline{F}$), having the same amplitude as the high sideband.

One sideband is shown in Fig. 2-3c, and the other in Fig. 2-3d. Each of the sidebands alone would create amplitude modulation, but without the other sideband the phase and frequency of the resultant wave also would vary. What happens, though, is that as one of the sidebands tends to shift the frequency or phase forward, the other



Fig. 2-3. The resolution of an a-m wave into a carrier and two sidebands, one of the sidebands being higher in frequency and the other sideband being lower in frequency than the carrier by the value of the audio-signal frequency.

tends to retard it, and vice versa. One sideband compensates exactly for the phase and frequency distortion of the other.

This compensation is shown by the uniformity of the wave at the vertical dotted lines in Fig. 2-3. A vertical dotted line is drawn at every point where the carrier wave crosses the zero axis. We note that where the dotted lines cross the sideband waves, when the upper sideband has a positive value, the lower sideband has a negative value, and vice versa, the two values canceling one another. Hence, the two sidebands equalize one another at the points where the carrier wave crosses the zero axis and, as a result, the modulated wave crosses the zero axis at the same instants as the carrier, indicating

that the modulated wave has the same frequency and phase as the original carrier.

In this manner, the a-m wave can be duplicated by the simple addition of a carrier and sidebands. For example, at instant 1 all three waves in Fig. 2-3 are positive; the resultant amplitude $\underline{A_T}$ is obtained by adding the instantaneous amplitudes \underline{A} , $\underline{A_1}$, and $\underline{A_2}$. Thus, the maximum peak amplitude of the modulated wave is equal to the peak amplitude of the carrier plus the peak amplitudes of the two sidebands. Similarly, at instant 2, where the carrier is positive and the two sidebands are negative, the resultant amplitude of the modulated wave is equal to the peak amplitude of the carrier minus the peak amplitudes of both the sidebands. Thus, at instant 2 the minimum amplitude of the modulated wave is obtained.

We note how the average amplitude of the a-m wave is equal to the amplitude of the unmodulated carrier wave. The variation in the amplitude, on the other hand, is determined by the sum of the sideband amplitudes. Inasmuch as the sideband amplitudes must be equal in order to eliminate any phase or frequency shift, the variation from the average amplitude—with single-frequency modulation—is equal to twice the amplitude of either of the sidebands.

Calling the maximum amplitude of the unmodulated wave \underline{V}_{max} and the minimum amplitude of the wave \underline{V}_{min} , we can obtain the average amplitude—the amplitude of the carrier—by dividing the sum by 2, thus,

A-m carrier amplitude =
$$\frac{V_{max} + V_{min}}{2}$$
 (2-1)

 \underline{V}_{max} , from the previous discussion, is equal to the carrier amplitude plus twice the amplitude of the sidebands; and \underline{V}_{min} is equal to the carrier amplitude minus twice the amplitude of the sidebands. Thus, the over-all variation in amplitude from \underline{V}_{max} to \underline{V}_{min} is equal to four times the amplitudes of the sidebands. Solving for the sideband amplitudes,

A-m sideband amplitude =
$$\frac{\underline{V}_{max} - \underline{V}_{min}}{4}$$
 (2-2)

Example. An a-m wave, as seen on an oscilloscope, has a maximum peak amplitude of 12 v and a minimum peak amplitude of 2 v. The frequency of the wave is 540 kc, and it is being modulated by a 2-kc sine wave. What are the amplitudes and frequencies of the carrier and sidebands?

The frequency of the carrier is the same as the nominal frequency of the a-m wave. Thus, the frequency of the carrier is 540 kc. The amplitude of the carrier is obtained by using Eq. 2-1; thus,

A-m carrier amplitude =
$$\frac{12+2}{2} = 7 v$$

The frequencies of the sidebands are obtained by first adding the modulating frequency to the carrier frequency, and then subtracting the modulating frequency from the carrier frequency. Thus, the sideband frequencies are equal to 540 kc plus 2 kc, and 540 kc minus 2 kc. The amplitudes of the sidebands are obtained by substituting into Eq. 2-2; hence, A-m sideband amplitudes = $\frac{12 - 2}{4}$ = 2.5 v

Thus, the components of the modulated wave are

Wave	1	A carrier of 7 v at 540 kc
Wave	2	An upper sideband of 2.5 v at 542 kc
Wave	3	A lower sideband of 2.5 v at 538 kc Ans

When the modulating signal consists of a complex wave, the same method of analysis may be applied. As discussed previously, a complex audio wave is composed of a fundamental and a series of harmonics. For each component of the complex wave, a pair of sidebands is created. Thus, if the complex wave is made up of 10 components, 10 pairs of sidebands would exist in the modulated wave.

Example. A 500-kc carrier wave is modulated by a complex audio signal. The audio signal, with a fundamental frequency of 1 kc, consists of a fundamental, a second, a third, and a fourth harmonic. What are the resultant frequencies present in the modulated wave?

The frequencies present in the complex audio signal are the fundamental at 1 kc, the second harmonic at 2 kc, the third harmonic at 3 kc, and the fourth harmonic at 4 kc. Each will create a pair of sidebands equal to the carrier plus and minus the frequency of the harmonic. The resultant frequencies are

Carr	ier	•		•			500 kc 🖡
1-kc	sidebands	•	•	•		•	501 and 499 kc
2-kc	sidebands						502 and 498 kc
3-kc	sidebands	•			•		503 and 497 kc
4-kc	sidebands				•		504 and 496 kc Ans.

The band width necessary for the transmission of the a-m signal must include all the sidebands generated by the a-f range to be transmitted. Examining the above example, we can see that the higher the frequency of the modulating signal, the greater the band width necessary for transmission. If the band width will handle the highest frequency, it will also take care of all the lower frequencies in the modulating signal.

The sidebands for the highest audio frequency to be transmitted are separated from the carrier frequency by the number of cycles per second equal to the highest audio frequency. In other words, to transmit a 4-kc audio signal, the sidebands would extend to 4 kc above and 4 kc below the carrier frequency. The band width necessary would be 8 kc for the 4-kc audio signal. In determining the band width necessary for the transmission of an a-m wave, only the highest modulating frequency has to be considered. The band width necessary is twice this highest frequency.

Example. What band width is necessary for an a-m wave to be used in transmitting an audio signal covering the frequency range of 60 to 5,000 cps?

The band width necessary is twice the highest modulating frequency. The highest modulating frequency is 5,000 cps; thus,

Band width for amplitude modulation, 60- to 5,000-cps modulating signal = $2 \times 5,000 = 10$ kc wide; $\pm 5,000$ cps Ans.

It is interesting to construct the phasor diagram of an a-m wave.¹ The phasor may be pictured as an arrow on a wheel, the rotation of the arrow generating a sine wave. The same analogy may be extended to take in sine waves at different frequencies. Each phasor representing a different frequency is considered to be on a different wheel rotating at a different speed. The angles between the different phasors are changing continuously.



Fig. 2-4. Aphasor diagram of an a-m wave, the two sideband phasors $\underline{PA_1}$ and $\underline{PA_2}$ rotating in opposite directions.

In Fig. 2-4 is the phasor diagram for an a-m wave. The phasor <u>OP</u> represents the carrier wave, which is used as the reference phasor. In other words, this phasor remains stationary throughout all the diagrams. The other phasors move past the reference phasor at a rate equal to the difference in speed, the faster ones moving ahead counterclockwise and the slower ones receding clockwise. (The counterclockwise direction is taken as positive.)

For clarity, the sideband phasors are drawn at the end of the carrier arrow. $\underline{PA_1}$ represents the upper sideband. Since the upper sideband is at a higher frequency, the phasor representing it is rotating faster than the carrier phasor. This phasor will move ahead of the carrier phasor at a speed equal to the difference in frequencies. The counterclockwise rotation of the upper-sideband phasor is indicated by the rotational arrow in the figure.

The phasor <u>PA2</u> represents the lower sideband. Since this sideband is at a lower frequency, its phasor will rotate in the reverse direction (clockwise) at a speed equal to the difference in frequencies. This clockwise rotation is also indicated by a rotational arrow on the phasor. At any instant the phasors <u>PA1</u> and <u>PA2</u> can be added to give

22

a resultant phasor \underline{PM} . It is this result \underline{PM} which is added to the carrier to obtain the final modulated-wave phasor.

The modulated wave is thus represented by the phasor OM. Figure



Fig. 2-5. The a-m phasor diagram (Fig. 2-4) shown for succeeding instances of time. The resultant, phasor OM, only varies in size - not in position.

2-5 shows the phasor diagrams for several succeeding instants of time. In Fig. $2-5\underline{a}$ the two sideband phasors are almost vertical and, when added together, yield a large phasor in phase with the carrier. This results in a large magnitude for the resultant phasor <u>OM</u>. In Fig. 2-5b both phasors have rotated a quarter of a revolution and equalize one another. In Fig. $2-5\underline{c}$ the two sideband phasors have begun to point downward and, when added to the carrier, result in a phasor smaller than the carrier. In Fig. $2-5\underline{d}$ the two sideband phasors have swung past one another and are beginning to climb again, resulting in a rising amplitude for the final phasor OM.

We see, from Fig. 2-5, that at any instant the resultant phasor OM is in phase with the carrier phasor. The modulated-wave phasor OM thus remains stationary in the diagram as concerns rotation, but varies in amplitude; hence, it meets the requirements of an a-m wave. Inasmuch as the sideband phasors move past the carrier phasor at a rate equal to the difference in frequencies (the frequency of the modulating audio signal), they will make one rotation for every cycle of the modulating signal. The modulated-wave phasor will, therefore, vary in amplitude in accordance with the modulating audio signal.

2-5. Bessel Functions. Before discussing other methods of modulation, let us first discuss briefly the subject of Bessel functions. Bessel functions are special functions which determine the sideband amplitudes in both frequency and phase modulation. No attempt will be made to discuss the mathematical proof of their existence or derivation; we will confine the discussion to an explanation of their meaning and use.

One of the simplest of the tables of functions is the well-known multiplication table; it consists of a series of tables, usually memorized for convenience of use. Let us examine the construction and meaning of the multiplication table because of the similarity in the construction of all tables. A brief multiplication table is given in Table 2-1. The tables are known as the zero table, the one table, the two table, the three table, the four table, and so on up through all the numbers. In the illustration, for brevity, whole-number tables from one through nine have been shown.

Consider the four table as an example. Four times a certain number yields a result. For similarity to Bessel functions let us refer to the number as the argument. Thus, 4 times the argument yields a certain result. If the argument is 3, we glance across the

Table	Argument													
Number	1	2	3	4	5	6	7	8	9					
0	0	0	0	0	0	0	0	0	0					
1	1	2	3	4	5	6	7	8	9					
2	2	4	6	8	10	12	14	16	18					
3	3	6	9	12	15	18	21	24	27					
4	4	8	12	16	20	24	28	32	36					
5	5	10	15	20	25	30	35	40	45					
6	6	12	18	24	30	36	42	48	54					
7	7	14	21	28	35	42	49	56	63					
8	8	16	24	32	40	48	56	64	72					
9	9	18	27	36	45	54	63	72	81					

TABLE 2-1. A SIMPLE MULTIPLICATION TABLE SHOWING ITS CON-STRUCTION AND ITS SIMILARITY TO THE TABLE OF BESSEL FUNC-TIONS GIVEN IN TABLE 2-2

four table until we see, under 3, that 4 times 3 equals 12. There are two factors necessary to obtain an answer: the choice of the table and the value of the argument; both are used to determine the result.

Bessel functions, although arranged in a very similar table, are derived in a very complicated manner. (If further information is desired, the reader can refer to advanced mathematical books concerning the functions.) The table itself, however, is quite easy to use because of its similarity to the ordinary multiplication table.

The capital letter \underline{J} is used to designate a Bessel function. Two things have to be shown: the table to be used and the argument. Mathematicians customarily use shorthand methods of designating various things and, for Bessel functions, they do the following: A subscript on
the <u>J</u> is used to designate the table to be used. Following the <u>J</u> and enclosed in parentheses is the argument to be used. Thus $J_2(4)$ means use table <u>J</u>₂ and argument 4. The table is also referred to as the order. Thus, wherever the order of a Bessel function is given, it signifies the table to be used.

Table 2-2 is a short table of Bessel functions covering the orders from 0 to 4 and arguments from 1 to 6. Only whole number arguments are shown. The construction of the table is similar to the multiplication table shown in Table 2-1. Each order is on a horizontal line while the argument designates a vertical column. Thus, for any order the correct horizontal row of figures is chosen, and for any argument the correct intersecting vertical row is chosen. The number at the intersection common to both of these rows is the answer. For instance, let us locate the value of $J_2(4)$. Bessel functions of order

Argument										
Order	(1)	(2)	(3)	(4)	(5)	(6)				
<u>1</u> 0	0.76	0.22	-0.26	-0.40	-0.18	0.15				
\underline{J}_1	0.44	0.58	0.34	-0.07	-0.33	-0.28				
<u>J</u> 2	0.11	0.35	0.49	0.36	0.05	-0.24				
<u>J</u> 3	0.02	0.13	0.31	0.43	0.36	0.11				
<u>J</u> 4	0.002	0.03	0,13	0.28	0.39	0.36				

TABLE 2-2. A SHORT TABLE OF BESSEL FUNCTIONS COVER-
ING ORDERS FROM 0 TO 4 AND ARGUMENTS 1 TO 6

2 are shown along the horizontal row labeled \underline{J}_2 . Moving across this row until the figure under the argument 4 is reached, we obtain as the answer 0.36. Thus, the value of $\underline{J}_2(4)$ is 0.36.

In many tables, the multiplication table being an example, the numbers always increase as the value of what we call the argument increases. However, in Bessel functions, the results both increase and decrease as well as go negative.

In Fig. 2-6 are shown some curves, or graphs, of the Bessel functions. The Bessel function values are plotted on the vertical scale while the argument is used as the horizontal scale. The zero order Bessel function is the only order which starts at 1; all others start at zero. The zero order function first decreases with an increase in the value of the argument, until it reaches zero at an argument value of 2.40. It then goes negative, in a manner resembling slightly a sine-wave curve, and returns to zero again at a value of 5.52. It oscillates around the zero axis in smaller and smaller cycles until, if it were continued out far enough, it would just become a small wiggle around the axis. The other orders shown, the first and second, start at zero for a zero value of the argument. They also rise and fall, going through zero in a manner resembling the zero order function. We note here



Fig. 2-6. Curves of three Bessel functions, the zero-, first-, and second-order functions. We note that the zero-order function is the only one to have a value of 1 for a zero value of argument.

how the lengths of the cycles, the distances between the points at which the curves cross the zero axis, become larger and larger as the argument increases.

No attempt should be made to memorize the Bessel functions or their behaviour. They are always referred to in tables if a value or characteristic is desired.

2-6. Band-Width Requirements for Frequency Modulation. The f-m wave is a constant-amplitude wave varying in frequency in accordance with the modulating audio signal. (This was previously discussed in Sec. 1-3 in Chap. 1.) The f-m wave can also be represented and analyzed as a carrier plus a series of sidebands, as we shall now see.

The first major difference between the f-m and a-m analysis is encountered in the amplitude of the carrier component. In f-m transmission, the amount of power in the transmitted wave (its amplitude) is the same whether or not modulation is present; hence, any r-f energy which appears as sidebands has to be deducted from the energy in the original unmodulated carrier. The carrier component, therefore, does not remain constant but varies in amplitude as the modulation is varied. Ideally, in f-m transmission, the final transmitted wave never varies in amplitude because the sidebands always add enough energy to keep the transmitted wave at the same amplitude as the unmodulated carrier.

In analyzing the f-m wave, one of the two important factors is the frequency deviation—the maximum shift in frequency from the car-

rier frequency. It is represented by

In other words, as an f-m wave varies, increases and decreases in frequency from the carrier frequency, the maximum change in frequency is the deviation. The instantaneous deviation is determined by the amplitude of the modulating audio signal.

The other important factor is the audio frequency. The sideband amplitudes and frequencies are determined by the audio frequency and the frequency deviation.

The above two factors are combined to form the determining factor for the f-m sidebands. This factor is called the "f-m factor" and is designated by $\underline{m_f}$. It is equal to the frequency deviation divided by the audio frequency.

$$\underline{m}_{\underline{f}} = \frac{\text{frequency deviation}}{\text{audio frequency}}$$
(2-4)

The factor $\underline{m}_{\underline{f}}$ is used to obtain the amplitudes of the sidebands involved.

For example, if the frequency deviation is 75 kc and the modulating frequency is 15 kc, $\underline{m}_{\underline{f}}$ is equal to 75 divided by 15; $\underline{m}_{\underline{f}}$ is 5. It is interesting to note that if the frequency deviation is reduced to 25 kc and the modulating audio frequency is also reduced to 5 kc, then the value of the modulation factor will remain 5.

Unlike the case of amplitude modulation where each audio frequency produces only two sidebands, each audio frequency produces, in an f-m wave, a whole series of sidebands.

Let us assume that we have an f-m wave with a carrier frequency



Fig. 2-7: An unmodulated f-m carrier showing its constant amplitude and constant frequency.

of <u>f</u> and an audio frequency of <u>F</u>. Figure 2-7 shows an oscilloscope trace of the unmodulated wave, a constant-frequency constant-amplitude wave. The modulated wave, to be a pure f-m wave, must also have a constant amplitude. The frequency deviation in combination with the audio frequency F will determine the value of the f-m modula-

27

tion factor $\underline{m_{f}}$.

Figure 2-8 shows the frequencies and amplitudes of the sidebands when the wave shown in Fig. 2-7 is frequency-modulated. A value of $\underline{m_f}$ equal to 2 is used; the frequency deviation is twice the modulating audio frequency. For instance, if the audio frequency were 10 kc, the frequency deviation would be 20 kc.



Fig. 2-8. The carrier and sideband components of an f-m wave which has a modulating factor \underline{m}_f of 2.

We note first that the carrier component, shown in Fig. $2-8\underline{a}$, has been reduced to 0.22 of its unmodulated value. This reduction in amplitude is caused by the transfer of energy from the carrier to the sidebands.

The first pair of sidebands is shown in Fig. 2-8b. One is higher in frequency than the carrier by an amount equal to the audio frequency, and the other is lower than the carrier frequency by the same amount. For instance, if the audio frequency were 10 kc, one sideband would be 10 kc above the carrier frequency and the other would be 10 kc below the carrier frequency.

This pair, called the "first pair of sidebands," resembles the sidebands generated in amplitude modulation. However, there is one major difference. In amplitude modulation, it was pointed out, the phases of the sidebands are so oriented that the phase and frequency of the modulated wave remain the same as the phase and frequency of the unmodulated carrier. At every point where the carrier wave passes through zero, the voltage contributed by one sideband is canceled by the voltage contributed by the other. This does not occur in frequency modulation. The sidebands in frequency modulation are so oriented, with respect to the carrier component, that the frequency of the resultant wave will shift a maximum amount: the sidebands are shifted 90 deg in phase. In the Fig. 2-8b, at the point noted as A, where the carrier-wave component crosses the zero axis, both sidebands have negative values noted as A_1 and A_2 . Adding these values to the carrier component results in shifting the intersection with the zero axis a large amount -causing the resultant frequency to vary.

In Fig. 2-9<u>a</u> is shown the result of adding together only the carrier component and the first pair of sidebands. We note that the length of the cycle from <u>C</u> to <u>D</u> is very much larger than the length of the cycle from <u>E</u> to <u>F</u>, indicating that the wave is frequencymodulated. However, as a frequency-modulated wave, it is not very good, since there is also a great deal of amplitude variation present in the wave.

The result can be improved by adding the second pair of sidebands, as given in Fig. 2-8c. These sidebands are at the carrier frequency plus twice the modulating frequency, and at the carrier frequency minus twice the modulating frequency. For instance, with the 10-kc audio frequency, the second pair of sidebands would be 20 kc above the carrier and 20 kc below the carrier. Their resultant phase removes a great deal of the amplitude distortion shown in the first result.

Figure 2-9b depicts the result of adding the second pair of sidebands as well as the first pair of sidebands; much less amplitude distortion is present and the waves begin to look more like the final f-m waves should look.

The third pair of sidebands, shown in Fig. 2-8d, occur at frequencies differing from the carrier frequency by three times the audio frequency. For the 10-kc audio frequency, one of the sidebands would be 30 kc above the carrier frequency and the other would be 30 kc below the carrier frequency. The fourth pair of sidebands, shown in Fig. 2-8g, are at frequencies equal to the carrier frequency plus four times the audio frequency. For the 10-kc audio frequency, they would be 40 kc above the carrier frequency and 40 kc below the carrier frequency. The remaining sidebands—the fifth, sixth, and aboveare so small, for an $\underline{m_f}$ of 2, that they can be disregarded without any noticeable distortion.

Adding in these remaining sidebands results in a practically



Fig. 2-9. The development of an f-m wave by adding, to the carrier component, one pair of sidebands at a time.

perfect f-m wave. This final resultant wave is illustrated in Fig. 2-9c; the frequency varies in accordance with the modulating audio voltage, and the amplitude remains constant.

For every value of $\underline{m}_{\underline{f}}$, the amplitudes of a carrier component and a set of sidebands are determined. When these specified waves, determined by $\underline{m}_{\underline{f}}$, are added together, the correct f-m wave will result.

Let us now consider the general case of sideband formation in an f-m wave when employing sine-wave modulation. If a complex audio signal wave should be used, the same method of analysis may be employed by breaking down the audio signal into its sine wave components. Each component sine wave will generate its own series of sidebands, the final result being the sum of all the sidebands.

The sideband frequencies are determined by the carrier frequency \underline{f} and the audio frequency \underline{F} . The sideband frequencies obtained for each modulating frequency \underline{F} will be as follows:

Carrier					f		
1st pair of sidebands					$(\overline{f} + \overline{F})$	and (f.	- F)
2d pair of sidebands.					$(\underline{f} + \overline{2}\underline{F})$	and $(f \cdot$	$\overline{2F}$
3d pair of sidebands.					$(\underline{f} + 3\underline{F})$	and (f -	- 3 <u>F</u>)
4th pair of sidebands					(f + 4F)	and (f	- 4F)
And so on.							_

Any sideband pair is located at the carrier frequency plus and minus the pair number times the modulating frequency.

The number of sidebands necessary to reproduce accurately a single audio frequency in frequency modulation may be a thousand or more, these large numbers occurring when the value of $\underline{m_f}$ is very large. The amplitudes of the carrier and the sidebands are determined by the value of $\underline{m_f}$.

In solving for the amplitudes of the carrier and sidebands in frequency modulation, engineers found that their amplitudes were determined by the Bessel function of the proper order with the modulation factor m_f as the argument. The order of the Bessel function used to obtain the amplitude of a sideband pair is the same as the sideband-pair number. For instance, the first pair of sidebands has an amplitude determined by the value of $J_1(m_f)$. With m_f equal to 2, as in the graphical example, the amplitude of the first pair would be determined by the value of $J_1(2)$. From Table 2-2, the value of $J_1(2)$ is 0.58; hence, the amplitude of each of the first pair of sidebands is equal to 0.58 times the over-all amplitude of the final f-m wave. Similarly, the amplitude of the second pair is determined by $J_2(m_f)$, the amplitude of the third pair by $J_3(m_f)$, the amplitude of the fourth pair by $J_4(m_f)$, and so on. The carrier amplitude, since it really is the zeroth pair, is determined by $J_0(m_f)$. For convenience, Table 2-2 can be rewritten, substituting the sideband-pair number for the order of the Bessel function and m_f for the argument. In this manner, the sideband amplitudes can be read off the table directly.

A table of the amplitudes of the carrier and sidebands, for relatively small values of \underline{m}_{f} , is given in Table 2-3. In the table, the value of \underline{m}_{f} is noted at the top of the vertical column tabulating the amplitudes of the various sidebands for that specific value of \underline{m}_{f} . The amplitude of the sideband is equal to the decimal value given in the table times the amplitude of the over-all f-m wave, the amplitude of the final f-m wave being the same as the amplitude of the unmodulated carrier. A negative value in the table merely indicates TABLE 2-3. A TABLE GIVING THE FREQUENCIES AND AMPLITUDES (AS A FRACTION OF THE AMPLITUDE OF THE OVER-ALL F-M WAVE) OF THE SIDEBANDS AND CARRIER FOR A FEW VALUES OF THE F-M FACTOR $m_{\rm f}$

	(1)	(2)	(3)	(4)	(5)	(6)
Carrier (<u>f</u>)	0.76	0.22	-0.26	-0.40	-0.18	0.15
Sideband Pair:						
1st pair $(\underline{f} \pm \underline{F})$	0.44	0.58	0.34	-0.07	-0.33	-0.28
2d pair (<u>f</u> ± 2 <u>F</u>)	0.11	0.35	0.49	0.36	0.05	-0.24
$\frac{3d \text{ pair}}{(\underline{f} \pm 3\underline{F})}$		0.13	0.31	0.43	0.36	0.11
4th pair $(\underline{f} \pm 4\underline{F})$			0.13	0.28	0.39	0,36
5th pair $(\underline{f} \pm 5\underline{F})$				0.13	0.26	0.36
$\begin{array}{l} \mathbf{6th \ pair} \\ (\underline{\mathbf{f}} \pm \mathbf{6F}) \end{array}$					0.13	0.25
7th pair (<u>f</u> ± 7 <u>F</u>)	-				0.05	0.13
8th pair (<u>f ± 8F</u>)						0.06

 $\underline{\mathbf{m}}_{\underline{\mathbf{f}}} = \frac{\mathbf{frequency deviation}}{\text{audio frequency}}$

that those sidebands are reversed in phase; the sideband is shifted 180 deg, and therefore subtracts from, instead of adds to, the final result. When the amplitude of a sideband pair and all sideband pairs above it have values less than 0.05 (times the amplitude of the overall f-m wave), that particular sideband pair and all others above it can be omitted from the f-m wave without any noticeable distortion. For instance, in the table, for an $\underline{m_f}$ of 1, the third pair and higher-numbered sideband pairs are lower than 0.05; hence, they are not listed as necessary sidebands in the table and may be disregarded.

An interesting feature of the f-m wave is not shown in the table inasmuch as it contains only whole-number values for $\underline{m_f}$. When the value of $\underline{m_f}$ is 2.40 (the frequency deviation is 2.40 times the audio frequency), then the amplitude of the carrier component becomes zero —it disappears. This fact is very useful for checking the deviation of a transmitter. The audio frequency is varied up from zero until the carrier component becomes zero; the amplitude and frequency of the audio signal is noted at that point. The deviation is then 2.40 times the audio frequency. For instance, if with a modulating audio voltage of 1 v the modulating frequency is found to be 10 kc at the point where the carrier component disappears, then the deviation for 1 v of modulating signal is 24 kc (2.40 times 10).

The band-width requirement for any type of modulation is that the band must be wide enough to pass all the sidebands necessary to transmit the wave undistorted in amplitude and in phase, phase being an important factor in frequency modulation. The actual band width for frequency modulation can be found upon reference to Table 2-3. For an $\underline{m_f}$ of 4, for instance, the last significant pair of sidebands is the fifth pair. Since all lower sideband pairs in this wave are closer to the carrier in frequency, this last significant pair determines the band width; the band width has to be large enough to pass the fifth pair of sidebands—the carrier frequency plus 5 times the audio frequency. Thus, the band width should be 10 times the modulating audio frequency to transmit without distortion an f-m wave with an $\underline{m_f}$ of 4. As an example, if the modulating frequency were 1 kc, the band width for an $\underline{m_f}$ of 4 (a deviation of 4 kc) would be 10 kc.

⁻ For very large values of \underline{m}_{f} , the band width is only slightly larger than twice the frequency deviation. Thus, for values of \underline{m}_{f} larger than 20, the band width need never be more than about 10 per cent larger than twice the frequency deviation of the f-m wave. As an illustration, when the frequency deviation is 75 kc, the frequency in the f-m wave varies from 75 kc above the carrier frequency to 75 kc below the carrier frequency; the total variation in frequency, twice the frequency deviation, is 150 kc. For large values of \underline{m}_{f} (above 20) a band width 10 per cent greater than 150 kc, namely, 165 kc, would be adequate for undistorted transmission.

Example. Find the spacing between f-m stations if each station transmits a voice-signal range of 30 to 15,000 cps with a frequency deviation of ± 75 kc.

The value of mf will vary between

and

For 30 cps,
$$\frac{75,000}{30} = 2,500$$

For 15,000 cps, $\frac{75,000}{15,000} = 5$

For values of \underline{mf} above 20 the band width need be only 10 per cent greater than twice the frequency deviation. For ± 75 -kc deviation the band width should be 165 kc. For an \underline{mf} of 5, according to Table 2-3, the seventh pair of sidebands should be passed. Plus and minus 7 times 15,000 cps means a band width of 210 kc. The larger of the two, 210 kc, is the band width required. (Actually, in practice, the 15,000-cps signal is a harmonic and never approaches 100 per cent modulation, a frequency deviation of $\pm 75,000$ cps. Hence, the full 210-kc band width is not considered necessary for f-m broadcasting under ordinary conditions.)

The phasor diagram for the f-m wave is obtained in a manner similar to the method used for amplitude modulation. The only major differences are that more than one pair of sidebands are added for each modulating audio frequency, and that all sideband phases have to be taken properly into account. Each sideband, being at a frequency different from the reference carrier frequency, can be pictured as being on a wheel rotating at a speed equal to the frequency of the sideband. At any instant all the phasors can be added together to obtain the resultant phasor representing the final f-m wave. At some instants the effect of the sidebands will be to speed up the frequency of the resultant wave; at other instants to slow it down, thus imparting frequency modulation.



Fig. 2-10. A phasor diagram of an f-m wave at one instant of time. Phasors showing the result of adding in one set of sidebands at a time are also shown on the diagram.

Figure 2-10 depicts a phasor diagram of an f-m wave at one instant of time. A dotted circle shows the amplitude of the final over-all f-m wave. In other words, the phasor representing the f-m wave will be located somewhere around the circle with the tip of its arrow on the circle. We note that the carrier component, upon modulation, is smaller than this circle. Adding the first pair of sidebands shifts the resultant phasor closer to the circle; the second pair, still closer; and adding in the remainder of the sidebands puts the tip of the final resultant phasor on the circle. As the sideband sine waves vary with time, the final over-all phasor is shifted back and forth, creating frequency modulation.

2-7. Band-Width Requirements for Phase Modulation. Phase modulation resembles frequency modulation a great deal, especially in the calculation of the sidebands. In fact the same table (Table 2-3) can be used in determining the sidebands, as well as the band width. Instead of \underline{m}_{f} , the factor \underline{m}_{p} is used. In phase modulation, the determining factor for \underline{m}_{p} is the amount of phase shift being employed. For instance, if the phase-modulated wave is shifting in phase from 50 deg leading to 50 deg lagging, the determining value is 50 deg translated into radians by the factor 0.0175. Thus, for a phase deviation of ± 50 deg, \underline{m}_{p} would be equal to 50 times 0.0175 or 0.875.

To illustrate the use of Table 2-3, let us assume a phase deviation of 114 deg; \underline{m}_p would be equal to 0.0175 times 114 or just about 2. Looking under (2) in Table 2-3 (\underline{m}_f equal to 2), we find that the first, second, and third pairs of sidebands should be transmitted, the frequencies of the sidebands being the same as indicated on the table. Hence, the band width should be six times the modulating frequency.

Very often value of \underline{m}_p obtained is a decimal. A good approximation is to use the sideband and band widths determined in Table 2-3 under the nearest full-number value of \underline{m}_p (\underline{m}_f in the table).

2-8. Comparison Between Band Widths for Amplitude Modulation and Frequency Modulation. In both the f-m and the a-m cases, the frequencies of the sidebands are determined by the frequency of the modulating audio signal and the carrier frequency. However, in a-m transmission only one pair of sidebands is created for each modulating audio frequency. Thus, the band width is equal to twice the highest audio frequency being transmitted.

In the case of frequency modulation, on the other hand, a whole series of sidebands is created for each modulating audio frequency. The number of significant pairs of sidebands and their values is determined by the modulation factor \underline{m}_{f} , the modulation factor being obtained by dividing the frequency deviation by the audio frequency. Usually, in practice, the band width used for frequency modulation is 10 to 30 per cent larger than twice the maximum frequency deviation. This is used because an \underline{m}_{f} greater than 5 and reaching up to the thousands is usually employed.

In a-m broadcasting, audio frequencies from about 60 to 5,000 cps are utilized, the 5,000-cps high-frequency limitation resulting from an assigned 10 kc. However, in high-fidelity f-m broadcasting the audio frequencies used range from 50 to 15,000 cps, with the frequency deviation being ± 75 kc. For f-m broadcasting a band width of 200 kc is found necessary for proper transmission of the f-m wave—twenty times greater than the band width used in a-m broadcasting. However, the gain in fidelity and the reduction in background noise and interference (to be discussed in the next chapter) more than compensate for the added band width. The subject of f-m sidebands is quite complex and many discussions may be found in the literature.²

REFERENCES

1. See the appendix for a fundamental discussion of phasors. (These phasors are referred to as "vectors" in some texts. However, in this book they will be called "phasors," since they are not true vectors in the sense that a vector is a spatial term having magnitude and direction.)

2. For further discussion of the subjects of sidebands and band width, see

B. van der Pol, "Frequency Modulation," Proc. IRE, 1930, p. 1194.

S. W. Seely, "Frequency Modulation," RCA Review, April, 1941.

H. J. Scott, "Frequency Versus Phase Modulation," <u>Communications</u>, August, 1940, p. 10.

H. Roder, "Amplitude, Phase and Frequency Modulation," Proc. IRE, 1931, p. 2145.

QUESTIONS

1. Describe the composition of a complex sound wave.

2. How is a complex wave considered when analyzing radio equipment?

3. What is meant by fidelity in radio broadcasting?

4. Explain the relationship between frequency response and r-f band width.

5. What is meant by a sideband in r-f transmission?

6. What are the frequencies of the sidebands generated by amplitude modulation of a 1-megacycle carrier with a modulating signal of 3 kc?

7. Explain why the sideband amplitudes in amplitude modulation can never be greater than one half the carrier amplitude.

8. What are the amplitudes of the carrier and sidebands when the maximum peak amplitude of the a-m wave is 18 v and the minimum peak amplitude is 2 v?

9. What are the frequencies of the sidebands in an a-m wave when a 1.2-megacycle carrier is modulated by a 2.5-kc complex audio signal which contains a fundamental and first, second, and third harmonics?

10. Construct a phasor diagram representing the a-m wave as described in Question 8.

11. What is meant by the order of a Bessel function?

12. Obtain the value of $J_3(4)$ from Table 2-2.

13. What is meant by frequency deviation in frequency modulation?

14. Explain what is meant by \underline{m}_{f} .

15. Determine the value of $\underline{m_{f}}$ for an f-m wave with a carrier frequency of 30 megacycles, a frequency deviation of 30 kc, and a modulating frequency of 5 kc.

16. If the amplitude of the unmodulated carrier is 10 v, determine the frequencies and amplitudes of the carrier and sidebands present in the modulated wave of Question 15.

17. What determines the band width necessary for f-m transmission?

18. Approximately what band width is necessary for an f-m wave with a frequency deviation of 50 kc and a modulating frequency of 60 cps?

19. What modulating frequency will cause the carrier component of an f-m wave, using a frequency deviation of 50 kc, to go to zero?

20. What are the frequencies and amplitudes of the carrier and sidebands generated by phase modulation of a 2-megacycle carrier? The modulating-signal frequency is 2 kc, a phase shift of 57 deg is employed, and the amplitude of the final phase-modulated wave is 100 v.

CHAPTER 3

Noise and Interference

3-1. Sources of Radio Noise and Interference. The terms "noise" and "interference," as used in radio terminology, are closely related and are sometimes used interchangeably; but, for purposes of discussion and to avoid confusion, each will here be used to designate a definite type of unwanted radio signal obtained at the output (usually the loudspeaker). The spurious voltages that represent energy more or less uniformly distributed over the band of frequencies being used are termed "noise voltages"; whereas, the reception of a single average-frequency, unwanted, extraneous signal voltage that <u>interferes</u> with the reception of the desired signal is called "interference."

Interference may come from many sources. The interfering signal may be another f-m station on the same or on an adjacent channel; it may be a far removed f-m or a-m transmission, perhaps even on another band, being picked up directly in the receiver circuits because of insufficient shielding; or it may be a hum or circuit oscillations internally generated by the receiver. These, and nearly all other sources of interference, can often be eliminated by a careful choice of operating frequencies and by careful construction of the receiver.

Noise voltages may be subdivided into two groups: randomnoise voltages and impulse-noise voltages. Random-noise voltages are characterized by a large number of small sinusoidal voltages with a completely haphazard distribution in both frequency and phase. Impulse-noise voltages are also a large number of small sinusoidal voltages occurring, however, at regular, closely spaced intervals of frequency and possessing an in-phase relationship with one another.

Random-noise voltages are produced by a multitude of tiny generators; separately insignificant perhaps, but in combination overspreading the whole band. One of the primary sources of random noise in an amplifier is thermal agitation. The free electrons

37

in any conductor, or resistor, are never stationary, but are being continually moved about by thermal forces. As each electron moves in a conductor or resistor, it varies the charge concentration within the material, thereby generating a small potential difference across the terminals of the conductor or resistor. Another type of random noise is known as "tube noise." This results from haphazard voltages generated by the variations in the electron stream arriving at the plate of the tube. Sometimes, in practical applications, the tube noise is greater than the thermal agitation noise. It has been found that the only important random-noise voltages are those generated in the input circuit of the first r-f tube or in the tube itself. The noise and signal voltage obtained at the output of the first tube are usually amplified to a level which far exceeds the noise voltages produced by subsequent tubes. Thus, the contribution to the noise level made by the remainder of the tubes in the receiver is of negligible importance.

When a direct current is passed through a carbon resistor, however, the electrical noise generated greatly exceeds the thermal agitation noise. The added disturbance is caused by fluctuations in contact resistance between the myriad carbon granules of which the resistor is composed. For this reason, carbon resistors are seldom used in the construction of input circuits to amplifiers with low-level signal inputs and <u>never</u> in the first stages of a receiver.

An impulse or spark of high intensity, either natural or manmade, will create the regular pattern of impulse-noise voltages. When caused by natural sources, such as lightning flashes occurring in thunderstorms, the resulting sharp crack or, in excessive cases, the steady crackling is called "static." These disturbances can usually be distinguished from the man-made impulse noise occurring in relays, motors, elevators, or automobile spark plugs. When located, man-made noise can usually be suppressed to a large ex-



Fig. 3-1. A short-duration high-current discharge, a common source of impulse-noise interference.

tent, but in a metropolitan area it is practically hopeless to try to eliminate the many sources of noise. Consequently, in an urban area, impulse noise usually overshadows all other types of noise or interference.

To analyze impulse noise, let us consider the usual source—a high-current, short-duration discharge as shown in Fig. 3-1. When broken down into its components, this single, sharp, steeply rising pulse actually consists of a band of frequencies covering the whole spectrum of radio frequencies—from zero frequency, since it nearly always contains some direct current, to the highest frequency thus far measured. In other words, the noise contains an extended series of closely spaced harmonics which, since they are generated



Fig. 3-2. The uniform magnitudes and even distribution of impulse-noise voltage harmonics in the frequency band f_1 to f_2 .

by a single impulse, are all in phase with one another. In Fig. 3-2 are depicted, by means of a line representation, the approximate distribution and the amplitudes of the impulse-noise voltage harmonics intercepted between the frequency limits of f_1 and f_2 . For purposes of analysis, these harmonics can be considered to be equal in amplitude, similar in phase, closely spaced, and evenly distributed over the whole band of frequencies.

3-2. Signal-to-Noise Ratio. The noise energy present at the output of a receiver actually limits its maximum usable sensitivity. The signal voltage at the input of the receiver has to produce enough signal energy at the output of the receiver to override the noise disturbance. The voltage signal-to-noise ratio is the ratio of the signal voltage to the noise voltage at the output of the receiver. When the signal voltage is equal to the noise voltage, the voltage signal-to-noise ratio is 1. When the signal voltage is 10 times the noise voltage, the signal-to-noise ratio is 10. An important characteristic of a receiver is the amount of signal-voltage input necessary to produce a specified signal-to-noise ratio at the output. The values of the ratio usually used are 1 and 10. Hence, as the noise energy is reduced; weaker signals are capable of being received; the signal voltage necessary to produce the specified signal-to-noise ratio is decreased. Sometimes the ratio of signal power to noise power is used; it is known as the "power signal-to-noise ratio."

The following procedure may be used to measure the signal

voltage necessary to produce a specified signal-to-noise ratio in a receiver. If an outside antenna is used, the antenna should be disconnected and a dummy antenna-an impedance equal to the input impedance of the antenna is usually sufficient-connected across the antenna terminals. If a built-in loop antenna is employed, the receiver should be tuned to a point where no station is received. The volume control is then turned up until the noise energy is measurable across the loaded output terminals, or loudspeaker. if one is employed, with an output meter. With the volume control set at this point, and after the amount of noise voltage as indicated on the meter has been noted, an r-f signal voltage from a modulatedsignal generator, at the frequency to which the receiver is tuned, is introduced across the antenna terminals. This signal-voltage output of the generator is increased until the output meter indicates that the voltage output-the generator plus the noise voltage-is double what the voltage was with noise alone. The amount of signal voltage necessary to accomplish this doubling of output is the signal voltage needed to produce a signal-to-noise ratio of 1. The procedure used for obtaining the signal voltage necessary to produce a signal-to-noise ratio of 10 is the same, except that the output voltage has to be increased until it is eleven times the noise voltage - ten times the noise voltage plus the noise voltage.

In a practical installation, the signal-to-noise ratio may be determined by the pick-up noise rather than by the internally generated noise in the receiver. This is often true in urban areas and in mobile installations. In those cases, the noise energy should be measured with the operating antenna in its final position, connected across the input terminals of the receiver, and with conditions as close to operating conditions as possible. The receiver is then detuned from any received signal, tuned to a frequency where no station is received, and the foregoing procedure repeated. This final signal-to-noise ratio voltage is very important; it determines how the receiver will operate in the field. In any practical case, it will be higher than the signal-to-noise ratio voltage determined in the laboratory.

3-3. Interference by an Adjacent Signal. When an adjacent carrier, with or without modulation, is not rejected by the tuned circuits of a receiver but is allowed, in addition to the desired signal, to pass through the receiver, interference with the desired signal will result. This happens every time that an extraneous frequency, an unwanted signal, is present within the admitted frequency band of the received signal. It will be assumed, for purposes of discussion and calculation, that the interfering signal is small in comparison with the desired signal. This is nearly always the case in practice.

The method of analysis is straightforward. An unmodulated interfering signal and an unmodulated desired signal are assumed. The two are mixed together and the resultant modulation determined. This mixed signal is fed into the detector and, depending upon the type of detector employed, a resultant voltage is produced at the output. The magnitude of this output voltage is the magnitude of the interference produced by the interfering signal.



Fig. 3-3. At a is shown the desired carrier which is being received; at b, the interfering signal; and at \underline{c} , the resultant wave present in the receiver. The wave shown in c is obtained by adding the waves shown in a and b.

Figure 3-3 illustrates, graphically, interference by an adjacentcarrier signal. The desired carrier, at a frequency \underline{f}_0 , is shown in Fig. 3-3a. The interfering signal, at a slightly displaced frequency of \underline{f}_1 , is depicted in Fig. 3-3b. Assuming that both signals are passed through the tuned circuits of the receiver, we know that the two signals will combine. The resultant is drawn in Fig. 3-3c. It will contain both amplitude and frequency modulation, as shown.

In the case of amplitude modulation, we can consider the result as a beat between two frequencies. Inasmuch as the interfering signal is at a slightly different frequency from the desired carrier, their phase relationship will be changing continuously; at one moment the interfering voltage will add to and a little while later subtract from the desired signal voltage. It will form the modulation envelope shown by dotted line in Fig. 3-3c, the frequency of the envelope being equal to the difference in the frequencies of the carriers, $(f_0 + f_1)$ or $(f_1 - f_0)$ --the positive result being chosen. A measure of the interference is obtained by calculating the a-m factor $\underline{m}_{\underline{a}}$ of the interference envelope. The percentage modulation is equal to 100 times $\underline{m}_{\underline{a}}$. The factor $\underline{m}_{\underline{a}}$ is obtained by dividing the amplitude of the detected wave—either of the two dotted waves that comprise the envelope shown in Fig. 3-3<u>c</u>—by the amplitude of the desired carrier. Practically, it determines the amplitude of the interfering signal obtained at the output of the receiver.

Let us call the carrier-wave amplitude \underline{E}_0 and the interferingwave amplitude \underline{E}_1 . Inasmuch as the envelope wave is produced by the addition and subtraction of the interfering voltage, the amplitude of the modulation envelope will be \underline{E}_1 ; the factor \underline{m}_a will be obtained by

$$\underline{\mathbf{m}}_{\underline{\mathbf{a}}} = \frac{\underline{\mathbf{E}}_{1}}{\underline{\mathbf{E}}_{0}} \tag{3-1}$$

and the percentage modulation by

Percentage modulation =
$$100 \frac{\underline{E}_1}{\underline{E}_0}$$
 per cent (3-2)

Example. Determine the a-m factor and the percentage modulation of the interference caused by an adjacent unmodulated carrier signal of $10 \mu v$ (microvolts) when the amplitude of the desired signal is $80 \mu v$.

The modulation factor is obtained by substituting into Eq. 3-1

$$\underline{m}_{\underline{a}} = \frac{10}{80} = 0.125$$

and the percentage modulation, from Eq. 3-2, is 12.5 per cent . Ans.

3-4. Adjacent-Signal Interference in Frequency Modulation. In the case of frequency modulation, where the receiver is insensitive to amplitude modulation, the amount of interference (the magnitude of the detected interfering signal) is determined only by the variation in frequency of the resultant wave. The frequency modulation present in the wave, as shown in Fig. 3-3c, is a little more apparent in Fig. 3-4 where all of the amplitude modulation has been removed and only the frequency modulation depicted.

The amount of frequency modulation present in the wave is determined by two factors. One is the ratio of the amplitude of the interfering wave to the amplitude of the desired signal, and the other is the magnitude of the frequency difference between the desired signal and the interfering signal. This latter factor and its relation to the frequency deviation used in the system are among the most important considerations for noise reduction in frequency modulation. Although there is a great deal of amplitude modulation present in the wave shown in Fig. 3-3c, the amount of frequency modulation present in the wave, as seen in Fig. 3-4, is quite small.



Fig. 3-4. The frequency modulation present when both the desired carrier and an adjacent signal are received. It is equivalent to the wave shown in Fig. 3-3c with the amplitude modulation removed.

The percentage modulation can be expressed by

Percentage modulation = 100
$$\frac{\underline{E}_1 (\underline{f}_1 - \underline{f}_0)}{\underline{E}_0 \underline{f}_{\underline{d}}}$$
 per cent (3-3)

where $\underline{f}_{\underline{d}}$ is the frequency deviation for 100 per cent modulation and the other symbols have the same significance as in the previous discussions. If the interfering signal is lower in frequency than the desired carrier frequency, a minus-sign answer will be obtained. This reversal in sign should be disregarded inasmuch as only the magnitude of the frequency difference is used in the equation. The frequency of the resultant interference signal in the f-m receiver will also be (f_1-f_0) , the difference frequency.

Let us compare the equations, Eq. 3-2 for the percentage modulation of interference in a-m reception, and Eq. 3-3 for the percentage modulation of interference in f-m reception, and determine their significance. We see that a certain magnitude interfering signal will cause the same amount of interference in a-m receivers, irrespective of its frequency, provided that it passes the tuned circuits of the receiver. In the case of f-m reception, however, the interference is reduced by a factor equal to the frequency difference between the two signals divided by the frequency deviation necessary to produce 100 per cent modulation at the receiver—a substantial reduction. Here we have the reason for broad-band f-m systems: the broader the band (the greater the frequency deviation), the greater the reduction in interference.

Example. An interfering signal of $4\mu v$ amplitude removed 2 kc from the desired carrier is causing interference. Determine the percentage modulation of this interference in both a-m and f-m reception. The desired carrier in both cases has an amplitude of 50 μv , and the f-m set employs a frequency deviation of 75 kc for 100 per cent modulation.

For the case of a-m interference, Eq. 3-2 is used.

A-m percentage modulation = 100 x
$$\frac{4}{50}$$
 = 8 per cent

And for the case of f-m interference, Eq. 3-3 is used, where $(\underline{f_1} - \underline{f_0})$ is equal to 2 kc and $\underline{f_d}$ is equal to 75 kc.

F-m percentage modulation = $100 \times \frac{4}{50} (2,000) = 0.21$ per cent Ans.

Quite a reduction!

3-5. Noise Reduction in Frequency Modulation. Noise, as has been discussed previously, consists of many small voltages distributed uniformly over the frequency spectrum. Each of these voltages that passes the tuned circuits of the receiver can be considered to be an adjacent-signal disturbance; the sum of all of the small disturbances will be the noise obtained at the output of the receiver. In impulse noise, where all of the voltages in the noise spectrum are in phase, the sum is found by simple addition; but in the case of random noise, where the phases of the voltages are haphazardly related, the sum is found by taking the square root of the sum of the squares. Hence, in the case of random noise, the condition that the voltages are not all in phase is taken into account.



Fig. 3-5. A graphical representation of the impulse noise in frequency modulation and in amplitude modulation where the audio band width in both cases is 30 to 15,000 cps; the f-m receiver employs, in this case, a frequency deviation of 75 kc.

For the process of addition, we shall employ the graphical method. Let us determine the ratio of the noise voltages produced in a theoretical a-m and a theoretical f-m receiver, both of which have an a-f band width of 30 to 15,000 cps and where the f-m receiver uses a frequency deviation of 75 kc. Impulse noise, being the simpler of the two to work out, is considered first.

Figure 3-5 shows a graphical representation of the impulse

noise as a function of the audio frequency for both frequency modulation and amplitude modulation. These are graphs of the percentagemodulation equations (Eqs. 3-2 and 3-3). The only voltages which are passed through to the output are those which lie within the audio band width (for comparison purposes both amplitude modulation and frequency modulation are taken as passing the maximum band width of 30 to 15,000 cps); all others will be cut off in the audio amplifiers. The a-m impulse noise, since it is equal to the sum of all of the individual voltages, will be determined by the area of the rectangle abcd-the noise within the audio band. The f-m impulse noise, since it is modified by a factor of frequency divided by the frequency deviation, will be determined by the trapezoidal area shown shaded in Fig. 3-5. The ratio of these two areas will show the impulse-noise reduction in a straightforward f-m receiver when it employs the same audio band width as an equivalent a-m receiver, namely, 30 to 15,000 cps, and employing a frequency deviation of +75 kc. The ratio is approximately 10. In other words, the f-m receiver will have one tenth the impulse noise obtained in ordinary high-fidelity a-m reception. Neglecting the lower-frequency cutoff in the audio system, because it is so close to zero, and calling the high-frequency audio cutoff \underline{F} , we find that the equation for impulsenoise voltage reduction in an f-m receiver from that in an a-m receiver with the same audio band width is

Impulse-noise voltage reduction factor =
$$2\frac{\frac{1}{d}}{F}$$
 (3-4)

where $\underline{f}_{\underline{d}}$ is the frequency deviation for 100 per cent modulation in the f-m receiver.

The calculations for random noise are similar to those for impulse noise, except that the squares of the individual voltages are used in the graphical representation, since, as discussed previously, the random-noise voltages are haphazardly arranged in phase. In Fig. 3-6 are drawn the squared percentage-modulation curves for both a-m and f-m reception. The noise reduction is now determined by the square root of the two areas within the audio band: the area of the rectangle abcd, the random noise squared for amplitude modulation; and the shaded area, the random noise squared for frequency modulation. (This is equivalent to taking the ratio of the square root of the sum of the squares.) The ratio of the two areas is approximately 75; therefore, the noise reduction, the square root of the ratio, is about 8.7. In other words, the random noise in an f-m receiver is about 0.115 of what it would be in an a-m receiver when they both have an audio band width of 30 to 15,000 cps and the f-m receiver employs a frequency deviation of 75 kc for 100 per cent modulation. Again, neglecting the audio 1-f cutoff, we find that the equation for the random-noise reduction factor, with equal audio band widths in both receivers, is

Random-noise reduction factor =
$$1.73 \frac{\underline{fd}}{F}$$
 (3-5)

where the symbols have the same meaning as previously defined. This factor is just a little smaller than the impulse-noise reduction factor, but is still a large reduction in the noise value.



Fig. 3-6. A graphical representation of a-m random noise squared and of f-m random noise squared for an audio band width of 30 to 15,000 cps, using a frequency deviation, in the f-m receiver, of 75 kc.

Example. Determine the noise-reduction factors for f-m reception in relation to a-m reception when both receivers employ an audio band extending up to 10,000 cps and the f-m receiver uses a frequency deviation of 30 kc for 100 per cent modulation.

Equations 3-4 and 3-5 are used where $\underline{f}_{\underline{d}}$ is put equal to 30,000 cps and \underline{F} to 10,000 cps.

Impulse-noise reduction factor =
$$2\frac{30,000}{10,000} = 6$$

Random-noise reduction factor = 1.73 $\frac{30,000}{10,000} = 5.2$ Ans.

3-6. De-emphasis. A process called "de-emphasis" is employed to decrease still further the noise level in f-m receivers. Referring to Figs. 3-5 and 3-6, we find that the major contributions to both impulse and random noise occur in the higher-frequency bracket of the audio band. The noise could be decreased a great deal by cutting down the h-f response of the receiver, but the fidelity would be lowered if some means of maintaining it were not employed. However, it was found that de-emphasis of the high frequencies can

be employed at the receiver and the fidelity still maintained by emphasizing the high frequencies at the transmitter. This is possible because the noise picked up by a receiver depends only on the receiver circuits and the receiver-response characteristics—not on the transmitter.



Fig. 3-7. The compensation of de-emphasis as demonstrated by the frequencyresponse curves of both transmitter and receiver which equalize one another.

Figure 3-7 illustrates the basic principle of de-emphasis and its compensation. The receiver-response curve indicates that the high-frequency response in the receiver has been cut down. This would normally decrease the fidelity, but, with a transmitter audioresponse curve emphasizing the highs as shown, this decrease does not occur. The transmitter characteristic compensates for the receiver characteristic as far as fidelity is concerned, while the attenuation of the highs in the receiver has diminished the amount of noise in the receiver output.

3-7. Conclusions. Let us first compare the characteristics, as far as noise is concerned, of the narrow-band a-m and wideband f-m broadcast systems. On listening to both types of receivers, we find that the f-m receiver is remarkably quiet-there is no background hiss. The hiss, caused by random noise, has been decreased by a very large factor in the f-m receiver. Pickup noises, such as light-switch clicks and refrigerator motors, have practically disappeared in the f-m reception. These are impulse noises which have been greatly reduced. We find that in the a-m receiver, on weak stations, it is sometimes possible to hear another station in the background. We notice that in the f-m receiver this does not occur. There are two reasons: first, the f-m stations, being in the 100-megacycle band, are not normally received beyond the line of sight (which decreases interference); and second, whatever interference is present is removed by the inherent reduction of adjacent-signal interference.

In commercial applications, especially mobile services such as police, fire, and ambulance intercommunication, the reduction in noise and interference is very advantageous. Reducing the background and pickup noise means that the call can get through in noisy low signal-strength areas where communication was heretofore impossible: under bridges, in streets between skyscraper buildings, and in heavy-static commercial areas. Because of the reduction in adjacent-signal interference, a station on the same or adjacent channel is completely cut out when the desired signal is only about twice the magnitude of the interfering signal, instead of about 50 to 100 times the magnitude as required in an a-m system. This, in combination with the limited propagation present in the higher frequencies, permits the use of the same channel by many groups throughout the country.

QUESTIONS

1. Describe the different types of radio noise and interference.

2. What type of radio noise would be generated by a d-c buzzer? Explain.

3. What type of radio noise would be generated by a carbon microphone? Explain.

4. What characteristics differentiate impulse noise from random noise?

5. What is signal-to-noise ratio?

6. How is the signal voltage necessary to obtain a signal-to-noise ratio of 10 measured?

7. If $30\mu v$ of signal produce a signal-to-noise ratio of 10, what signal voltage is necessary to produce a signal-to-noise ratio of 1?

8. If $25\mu v$ of signal are necessary to produce a signal-to-noise ratio of 8, what signal voltage is necessary to produce a signal-to-noise ratio of 3?

9. Why is the voltage necessary to produce a specified signal-to-noise ratio higher when measured on an actual installation than when measured using a dummy antenna in the laboratory?

10. Determine the percentage modulation of the interference on a desired carrier of 60 μ v amplitude by an interfering signal of 10 μ v amplitude and 3,000 cps removed from the carrier on (a) an a-m receiver and (b) an f-m receiver employing a frequency deviation of 30 kc for 100 per cent modulation.

11. Determine the impulse-noise reduction factor of an f-m receiver over that of an a-m receiver when they both have an audio band width extending up to 15,000 cps. and the f-m receiver employs a frequency deviation of 35 kc for 100 per cent modulation.

Determine the random-noise reduction factor for the receiver of Question 11.
 What is meant by de-emphasis and what does it accomplish?
 How is de-emphasis prevented from decreasing the fidelity of the f-m broad-

casting system?

15. What are the advantages of wide-band frequency modulation, as far as noise is concerned, for broadcast applications?

16. Why do the reductions in noise and interference in frequency modulation make the system so desirable for mobile applications?

CHAPTER 4

Direct Frequency Modulation

4-1. Oscillator Frequency Variation. A direct method of obtaining frequency modulation is to vary the frequency of the oscil-



Fig. 4-1. A simple oscillator circuit where the frequency depends on the values of \underline{L} and \underline{C} . When the capacitor \underline{C}^{I} , shown dotted, is added to the circuit, it will change the frequency of oscillation.

lator being used as the initial source of r-f energy. In Fig. 4-1 is shown a simple Hartley oscillator. The frequency of this oscillator depends on the values of the inductance \underline{L} and the capacitor \underline{C} that comprise the tank circuit. In fundamental radio theory it was shown that

$$\underline{f} = \frac{1}{2\pi} \underline{LC} \tag{4-1}$$

where \underline{f} is the frequency generated by the oscillator.

Let us add, now, in parallel, the capacitor \underline{C} ', shown by dotted line in Fig. 4-1. This capacitor will add to the capacitance of the capacitor \underline{C} so that the resultant frequency of the oscillator will now be determined by the value of ($\underline{C} + \underline{C}$ '); consequently, the frequency of oscillation, from Eq. 4-1, will decrease. The effect of

adding a parallel inductance instead of a capacitance is similar, but, since two inductances in parallel are equivalent to an inductance which is smaller than either of the original inductances, the frequency of the oscillator will increase.

The variation of the frequency of the oscillator may be accomplished by changing any of the tank-circuit components \underline{L} , \underline{C} , or \underline{C} '. In practical usage, however, \underline{L} and \underline{C} are usually employed for manual adjustment of the oscillator frequency while \underline{C} ' is employed to produce the frequency modulation. In other words, the value of \underline{C} ', usually the effective capacitance of an electronic device, is varied with the modulating voltage to produce the variation in frequency which constitutes frequency modulation.

This type of oscillator cannot employ, as do a-m transmitters, a quartz crystal as its tank circuit, since crystal control would prevent the necessary frequency variation. Nevertheless, stringent stability requirements still have to be met; therefore, some other method of center-frequency stabilization must be used. The usual method is to compare the center frequency of the f-m wave with the stable frequency generated by a crystal oscillator, and use the result for controlling the center frequency of the f-m oscillator. Either a motor may be used to adjust the value of the capacitor C, or a voltage may be used to adjust the center value of the electronic capacitance C'; both of these methods or their equivalent have been found sufficient to meet the FCC requirements. (These methods will be discussed in detail in the chapter on frequency-control circuits.)



Fig. 4-2. A basic reactance-tube circuit showing the audio-input terminals and the simulated-reactance terminals.

4-2. The Reactance Tube. A reactance tube is a tube wired into a particular type of circuit so that its a-c impedance from plate to ground is similar to the impedance of a reactance. The impedance of a pure reactance is characterized by a quadrature relationship: it causes the current flowing through the reactance to be 90 deg out of phase with the voltage impressed across it. By reproducing this relationship, a tube can simulate a reactance.

A basic reactance-tube circuit is shown in Fig. 4-2. The simulated-reactance terminals are from plate to ground; as shown, the circuit passes through a coupling capacitor whose impedance at the frequencies to be used is negligible. An r-f choke is used as a means of impressing B+ voltage on the plate. The screen-grid circuit is standard. The grid circuit, consisting of \underline{R}_1 and \underline{C}_1 , is the key to the tube's behavior.

Any r-f current which flows from cathode to plate in the tube will also flow through any circuit connected across the simulatedreactance terminals; it is assumed that the paths through the r-f choke and through C_1 have a very high impedance and draw negligible r-f current. The object is to make the r-f current flow 90 deg out of phase with whatever voltage is impressed across the simulated-reactance terminals.

The voltage <u>e</u> across the simulated-reactance terminals is also impressed (by a direct connection, as shown) across the capacitor $\underline{C_1}$ and the resistor $\underline{R_1}$ in series. The capacitor $\underline{C_1}$ is a small capacitor whose reactance is very much larger than the value of the resistor $\underline{R_1}$; hence, the current through the combination will be determined by the capacitor only, and will lead the impressed voltage <u>e</u> by 90 deg. Since the voltage across a resistor is in phase with the current through it, the voltage across the resistor will also lead the impressed voltage <u>e</u> by 90 deg. But the voltage across the resistor is the r-f voltage which is impressed from grid to ground. Consequently, the grid voltage, and therefore the plate current, will be 90 deg out of phase with the voltage <u>e</u> across the simulated-reactance terminals. Thus, the r-f current flowing across the simulated-reactance terminals will be 90 deg out of phase with the voltage impressed across the terminals.

The amount of alternating plate current which the tube draws will depend on the gain of the tube. The variation in gain, caused by varying the grid bias, is used to control the amount of plate current drawn by the tube and, hence, controls the value of the simulated reactance. The grid bias is varied by the audio voltage impressed through an isolating r-f choke. As the audio voltage increases, the grid bias increases; the amount of r-f plate current increases; the value of the simulated reactance in this case decreases; and, if the tube is used as a portion of an oscillator tank circuit, the frequency of the oscillator will either increase or decreases; the amount of r-f plate current decreases; the value of the simulated reactance increases; and the oscillator frequency will thus change in the opposite sense. The result is frequency modulation.

4-3. The Capacitive-Reactance Tube Modulator. Inasmuch as

the circuit illustrated in Fig. 4-2 is used to produce modulation, it is called a "modulator"; and, because it resembles a capacitive reactance, its full name is the "capacitive-reactance tube modulator." In a capacitive reactance, the current flowing through the reactance leads the voltage impressed across the reactance by 90 deg. In the modulator circuit of Fig. 4-2, the voltage across the grid, from the foregoing discussion, leads the impressed voltage by 90 deg; consequently, the plate current, the current drawn by the simulated reactance, leads the impressed voltage \underline{e} by 90 deg and the simulated reactance resembles a capacitance. Calling the value of the simulated reactance C, we find that its value is given by

$$C' = C_1 R_1 g_m \tag{4-2}$$

where $g_{\underline{m}}$ is the transconductance of the tube in <u>mhos</u>, \underline{R}_1 is the value of the resistance in ohms, and \underline{C}_1 is the value of the capacitor in farads. This equation assumes that the current drawn by the $\underline{C}_1\underline{R}_1$ circuit is negligible and that the voltage across the resistor \underline{R}_1 is shifted a full 90 deg. To meet the latter requirement for all practical purposes, we must make the reactance of \underline{C}_1 very much larger than the value of \underline{R}_1 . The value of \underline{C}_1 is not critical, but it should be chosen carefully. The larger the value of \underline{C}_1 , the larger the value of C (apparent from Eq. 4-2). But this is not obtained in practice, inasmuch as \underline{R}_1 would have to be proportionately smaller. Let us make \underline{R}_1 one tenth the reactance of \underline{C}_1 , one tenth being a small enough value for practical application. This value of \underline{R}_1 can now be substituted into the equation for \underline{C} betained:

$$\underline{C}' = \frac{0.1\underline{g}_{\underline{m}}}{2\pi f}$$
(4-3)

This shows that the value of the simulated capacitance under these conditions depends only on the transconductance of the tube $g_{\rm m}$ and on the frequency <u>f</u> being used. The only added requirement that the value of <u>C</u>₁ must meet is that the value of its reactance should be large with respect to the input impedance of the r-f voltage source impressed across the simulated-reactance terminals.

EXAMPLE. Determine the value of capacitive reactance obtainable at 5 megacycles with a capacitive reactance tube modulator when using a tube which has a transconductance of $8,000\mu$ mhos.

Using the circuit depicted in Fig. 4-2 with the resistance in the grid circuit one tenth of the value of the reactance of C_1 , we can substitute directly into Eq. 4-3.

$$\underline{C'} = \frac{0.1 \times 0.008}{2 \times 3.14 \times 5,000,000}$$

 $\underline{C}^{\dagger} = 25.5 \ \mu\mu f$ Ans.



Fig. 4-3. A capacitive-reactance tube modulator being used as a variable capacitance across the tank circuit of an oscillator.

Figure 4-3 shows the reactance tube being used as the variable reactance across the tank circuit of an oscillator. The two simulated-reactance terminals are connected directly across the terminals of the tank circuit of the oscillator. As the reactance of the reactance tube varies, it will vary the frequency of the oscillator.

EXAMPLE. Determine the variation in frequency of an f-m oscillator employing a capacitive reactance tube circuit at 5 megacycles wherein the transconductance of the tube varies from zero to 9,000 μ mhos. An inductance of 30 μ h is used in the tank circuit of the oscillator.

From the equation for the resonant frequency of a tank circuit (Eq. 4-1), it is found that a capacitance of 33.8 $\mu\mu$ f is necessary in the tank circuit to resonate it to 5 megacycles. At zero transconductance the reactance presented by the reactance tube will be infinite. It will not draw any reactive current, and its effect on the resonant frequency of the oscillator will be nil. The oscillator frequency will remain at 5 megacycles. The effective capacitance of the reactance tube when the transconductance is 9,000 μ mhos can be obtained by substituting in Eq. 4-3. Since the frequency of operation is not known, as an approximation let f equal 5 megacycles.

$$\underline{C}' = \frac{0.1 \times 0.009}{2 \times 3.14 \times 5,000,000}$$

$$\underline{C}' = 28.6 \ \mu\mu f$$

Hence, there is a total capacitance of 33.8 plus 28.6 $\mu\mu f$, or 62.4 $\mu\mu f$, in the tank circuit to resonate with the 30 μ h. The final tank-circuit resonant frequency may be obtained by substituting in Eq. 4-1.

$$\frac{f'}{2 \times 3.14 \times \sqrt{30 \times 10^{-6} \times 62.4 \times 10^{-12}}}$$

f! = 3.68 megacycles

Actually, the frequency shift is even greater because a value of 5 megacycles was used in calculating the value of C'instead of final frequency value. For a more accurate result, the calculations should be repeated using 3.68 megacycles for <u>f</u> in the calculation of <u>C</u>!. Only a small portion of this frequency variation is actually used in practice because of nonlinearity of the tube characteristics near the extremes and the desirability of allowing a linear portion on each side of the operating portion of the characteristic for control purposes.



Fig. 4-4. A capacitive-reactance tube circuit employing an inductance and resistor in the grid circuit to obtain the 90 deg phase shift.

Another reactance-tube circuit which will behave like a capacitive reactance is shown in Fig. 4-4. In this case the value of \underline{R}_2 is made very much larger than the reactance of \underline{L}_2 , causing the current through \underline{L}_2 to be determined, for all practical purposes, by the value of the resistor \underline{R}_2 . Consequently, the current through \underline{L}_2 will be in phase with the voltage impressed across the simulated-reactance terminals. But the voltage across an inductance leads the current through it by 90 deg. Hence, the voltage on the grid, and therefore the current through the tube, will lead the voltage impressed across the simulated-reactance terminals by 90 deg. The result will be a simulated capacitive reactance whose value \underline{C} is given by

$$C' = \frac{\underline{L}_2}{\underline{R}_2 g_m}$$
(4-4)

If we make the value of \underline{R}_2 ten times the value of the reactance of \underline{L}_2 , then the equation for the value of \underline{C} becomes the same as the equation for the previous case (Eq. 4-3).

equation for the previous case (Eq. 1-5). 4-4. The Inductive Reactance-Tube Modulator. The simulated reactance of a reactance-tube modulator can also be made inductive. The only difference in its design is that the grid voltage is made to lag the impressed voltage by 90 deg, instead of leading 90 deg as in the previous cases. This will cause the plate current to lag 90 deg behind the voltage impressed across the simulated-reactance terminals and its effect will duplicate that of an inductive reactance.



Fig. 4-5. An inductive-reactance tube modulator employing the phase-shift characteristics of a capacitor and resistor in the grid circuit.

Figure 4-5 shows the circuit diagram of an inductive reactance tube modulator which employs the phase-shift characteristics of a capacitor and resistor in the grid circuit. In this case, the value of R₃ has to be very much greater than the magnitude of the reactance of C₃. Hence, the current through the capacitor C₃ will be determined by the resistor R₃. The current will be in phase with the voltage \underline{e} impressed across the simulated-reactance terminals. But the voltage across a capacitor always lags the current through it by 90 deg. Consequently, the voltage across the capacitor, the voltage impressed on the grid of the tube, will lag the voltage \underline{e} by 90 deg. Thus, the plate current drawn by the tube lags the voltage impressed across the simulated-reactance terminals by 90 deg and simulates an inductive reactance.

If \mathbb{R}_3 is made large enough so that the amount of plate current passing through it is negligible, then the value of the simulated inductive reactance \underline{L} will be given by

$$\underline{L}' = \frac{\underline{R}_3 \underline{C}_3}{\underline{g}_m}$$
(4-5)

where \underline{g}_{m} is the transconductance of the tube being used. Now, let us make \underline{R}_{3} ten times the value of the reactance of \underline{C}_{3} . Using this condition, we find that the equation for the effective inductance can be simplified to

$$\underline{L}' = \frac{10}{2\pi \underline{f}g_{m}}$$
(4-6)

showing that the simulated inductance varies inversely with the frequency and transconductance. When used across the tank circuit of an oscillator, this inductance is in parallel with the inductance in the tank circuit and reduces the effective inductance of the circuit.

EXAMPLE. Determine the value of simulated inductance that can be obtained with a reactance tube that has a transconductance of 5,000 μ mhos and is being used at a frequency of 5 megacycles.

Using the circuit shown in Fig. 4-5 and making the value of \mathbb{R}_3 ten times the reactance of \mathbb{C}_3 , we can substitute directly into Eq. 4-6.

 $\frac{L'}{2 \times 3.14 \times 5,000,000 \times .005}$ L' = 63.7 \mu h <u>Ans</u>.



Fig. 4-6. An inductive-reactance tube modulator employing the phase-shift characteristics of an inductance and resistance in the grid circuit.

Another circuit for an inductive reactance tube modulator is shown in Fig. 4-6. In this case, an inductance and a resistance are used in the grid circuit to provide the necessary phase shift. The inductive reactance of \underline{L}_4 is made very much larger than the value of the resistor \underline{R}_4 . Consequently, the current flowing through the resistor \underline{R}_4 will be determined by the value of the inductance \underline{L}_4 . This current will lag the voltage impressed across the simulatedreactance terminals by 90 deg. Hence, the voltage impressed on the grid and, therefore, the current drawn by the plate of the tube will lag the impressed voltage by 90 deg, the requisite for an inductive-reactance effect. If the value of the reactance of the inductance \underline{L}_4 is made 10 times the value of the resistor \underline{R}_4 , then the inductance is given by the same equation (Eq. 4-6).

4-5. A Balanced Reactance-Tube Modulator. In Fig. 4-7 is shown a circuit with two reactance tubes. The audio inputs to the tubes are in push-pull while the reactance terminals are in parallel. One of the tubes simulates a capacitive reactance, while the other simulates an inductive reactance. Paralleling the reactance terminals is accomplished by connecting the two plates in parallel



Fig. 4-7. A balanced-reactance tube modulator employing two reactance tubes with the reactance terminals in parallel, one of the tubes being a capacitive-reactance tube and the other an inductive-reactance tube.

across the tank circuit of the oscillator, as shown. The plate voltage is supplied through the inductance of the oscillator tank circuit. This eliminates the need for the r-f choke and coupling capacitor that would otherwise be used in the plate circuits. The audio input is applied through an audio transformer; a bias voltage is used to obtain the operating point of the reactance tubes. The circuit shown is merely a combination of the circuits given in Figs. 4-1, 4-2, and 4-5.

The operation of the circuit is based on the variation of the reactances with a variation of the transconductances, as given in Eqs. 4-3 and 4-6. In the capacitive-reactance tube the capacitance \underline{C} varies directly with $\underline{g}_{\underline{M}}$, increasing with an increase in $\underline{g}_{\underline{M}}$, while in the inductive-reactance tube the inductance \underline{L} ' varies inversely with $\underline{g}_{\underline{M}}$, decreasing with an increase in $\underline{g}_{\underline{M}}$. To obtain the maximum shift in frequency, both the inductance and the capacitance \underline{L} ' and \underline{C} ' should vary together. Hence, the transconductance of one tube should increase when the transconductance of the other tube decreases, and vice versa. This is accomplished by having a push-pull input; when the grid voltage of one tube increases, the grid voltage of the other tube decreases, yielding an additive effect on the frequency of the oscillator.

The balanced reactance tube modulator offers two advantages: First, since one tube is operating on a portion of the transconductance curve above the operating point when the other tube is operating on a portion of the transconductance curve below the operating point, the use of the push-pull input circuit tends to decrease any distortion due to curvature of the transconductance characteristics. Second, since the circuit consists of two tubes in parallel, it reduces the amount of swing on each tube necessary to produce the desired amount of frequency shift.

4-6. The Input Capacitance Modulator. Another means of accomplishing frequency modulation is by utilizing the input capacitance variation of a tube. The fundamental circuit for the input



Fig. 4-8. The basic circuit of an input-capacitance tube modulator.

capacitance modulator is shown in Fig. 4-8. The input capacitance \underline{C}_{in} of a tube depends on the grid-to-plate capacitance, the capacitance \underline{C}_{gp} shown by dotted line in the figure; thus,

$$\underline{C}_{in} = \underline{C}_{\underline{gp}} \left(1 + \underline{g}_{\underline{m}} \underline{R}_{\underline{L}}\right) + C_{\underline{gk}}$$
(4-7)

where $g_{\underline{m}}$ is the transconductance, C_{gk} the grid-to-cathode capacitance (usually negligible), and $\underline{R}_{\underline{L}}$ is the load resistance. Hence, the input capacitance will vary with the transconductance, and therefore with the grid voltage. In Fig. 4-8 the reactance terminals are from grid to ground. A grid leak is used to keep the grid from accumulating charge and to provide a path for the bias and modulating audio voltages to reach the grid.

EXAMPLE. Determine the input capacitance of a tube whose plate-to-grid capacitance is 4 $\mu\mu$ f and whose transconductance varies from 1,000 to 2,000 μ mhos. The plate-load resistance used is 100,000 ohms. (An inductor is used to tune out any shunt capacitance.)

We can substitute directly into Eq. 4-7.

$$\underline{C}_{in} = 4 \times 10^{-12} (1 + 0.001 \times 100,000)$$

$$\underline{C}_{in} = 404 \ \mu\mu f$$
 Ans.

And similarly for a transconductance of 2,000 µmhos

$$\underline{C}_{in} = 804 \ \mu f$$
 Ans.



Fig. 4-9. A pentode input-capacitance modulator used with a Hartley oscillator. The capacitor $\underline{C}_{\underline{a}}$ augments the grid-to-plate capacitance, and $\underline{R}_{\underline{L}}$ contains a tuned circuit to obtain a high value of resistance.

In Fig. 4-9 appears the circuit of a pentode input-capacitance modulator used to vary the frequency of a Hartley oscillator. The oscillator could be a 6J5 tube and the modulator, a 6AB7 tube. To obtain a high plate resistance, which is necessary for a high input capacitance, the load resistance consists of a variable inductance and capacitance in parallel with the load resistor. By tuning this circuit to resonance at the frequency being used, any stray capacitance, as well as the output capacitance of the tube, is tuned out, yielding a pure resistance for $\underline{R}_{\underline{L}}$. The plate-to-grid capacitance is augmented with the capacitance. Basically, however, it is still the same circuit as illustrated in Fig. 4-8.

The modulator varies the frequency of an ordinary Hartley oscillator. To reduce the amount of r-f voltage on the grid of the tube, the input-capacitance modulator is connected across only a portion of the inductance in the tuned circuit. A variable-coupling capacitor, C_c is used to couple the modulator to the oscillator tank circuit; thus, a fixed-frequency swing for a given change in the input capacitance may be obtained as the oscillator capacitance is varied for use at different points in the frequency band. The audio voltage is fed into the grid of the input-capacitance tube and, as the voltage varies, the transconductance of the tube is varied. Thus, the input capacitance is changed and, in turn, frequency modulation is produced in the oscillator.

4-7. The Transmission-Line Modulator. The variation of the input impedance of a transmission line as the load impedance varies can also be used to produce frequency modulation.¹ When a transmission line approaches in length an appreciable fraction of a wavelength, the impedance looking into the input terminals of the line is quite different from the actual load impedance connected across the



Fig. 4-10. A transmission line of length \underline{l} terminated in a load resistance \underline{R}_m .

load terminals. In Fig. 4-10 is shown a transmission line of length 1 terminated in a load resistance $\underline{R}_{\underline{m}}$. The terminals <u>a</u> and <u>b</u> are the input terminals across which a voltage <u>E</u> is impressed. The length <u>l</u> is chosen to be an eighth of a wavelength. For a wavelength of 10 ft (a frequency of about 100 megacycles), the length <u>l</u> would be 15 in. This is a usable length which can be incorporated into a transmitter as one of its component parts.

Transmission-line theory is, in itself, a complex field of study and we could not expect to cover it completely in a few paragraphs. However, some of the theory which deals with our specific problem may be discussed briefly.

When an a-c voltage, like the voltage \underline{E} in Fig. 4-10, is impressed across the input terminals of a transmission line, the effect of that voltage is not felt instantaneously along the whole length of the line; the effect of the voltage actually travels down the line at a definite speed—the speed of light, 186,000 mps. When the voltage reaches the end of the line (the load impedance), the voltage may be completely absorbed, or part of it may be reflected, depending upon the load impedance. The reflected voltage then travels back to the input terminals affecting the input impedance. At low frequencies, transmission lines are usually very short with respect to a wavelength, and these effects are not very noticeable. However, at the higher frequencies, where the transmission lines are no longer short with respect to a wavelength, these effects are very important.

For an ordinary transmission line there is one value of load resistance for which no reflection takes place, all of the energy being absorbed in the resistor. This value of resistance is called the "characteristic impedance" of the transmission line and is usually noted as Z_0 . The value of \underline{Z}_0 can be anywhere from 40 ohms or
less for a coaxial transmission to over 500 ohms for a high-power balanced-transmission line. When the load resistance has any other value, either above or below the characteristic impedance, a reflection of voltage will take place. For different values, different amounts of reflection will occur.

Let us consider now the transmission line pictured in Fig. 4-10. The characteristic impedance of this line is 100 ohms. Suppose that we terminate it in an impedance of nominal value of about 250 ohms. A reflection will definitely take place. Since the effect of the voltage has to travel down the line, be reflected, and then travel back again, it will be out of time phase with the impressed voltage at the input terminals. This effect will cause the input impedance to have a reactive component. For a given transmission line, the reactive component will be determined by the reflected wave which, in turn, is determined by the load-impedance value. When the load resistance $\underline{R_{m}}$ is varied from 200 to 300 ohms, the reactive component of the input impedance varies from about 75 to 130 ohms capacitive reactance. There is also present, in the input impedance, a resistive component of about 145 ohms which remains constant as the load resistance is varied between the aforesaid values.

This variation in the input impedance of a transmission line may be utilized in an f-m modulator. The circuit for an f-m oscillator



Fig. 4-11. An f-m modulator employing a transmission line with a cathode-follower variable load resistance.

is shown in Fig. 4-11. Instead of an ordinary load impedance, a cathode follower is used. The impedance from cathode to ground of a cathode follower varies with the grid bias. The audio input is impressed on the grid of the cathode follower; as the audio voltage varies the grid bias, the impedance from cathode to ground of the cathode follower will vary proportionately. The input terminals to the transmission line, <u>a</u> and <u>b</u>, are connected across a break in the oscillator tank circuit. With a change in audio voltage (hence, a

change in the impedance of the cathode follower), the input reactance of the transmission line will vary; this, in turn, will vary the frequency of the oscillator and thus produce frequency modulation. If the loading on the tank circuit is too great because of the resistive component of the input impedance of the transmission line, a shunting resistor should be placed across the terminals \underline{a} and \underline{b} . The variation in reactance will still be sufficient for the necessary frequency modulation.

4-8. The Resistance-Capacitance F-M Generator. Nearly all resistance-capacitance oscillators may be used to produce frequency modulation by replacing one of the frequency determining resistors with a variable-resistance tube whose resistance is regu-



Fig. 4-12. A resistance-capacitance oscillator where $\underline{R}_1, \underline{R}_2, \underline{C}_1$, and \underline{C}_2 are the frequency-determining constants.

lated by the audio voltage.² In Fig. 4-12 is shown the circuit diagram of a resistance-capacitance oscillator. It consists of a pair of resistance-coupled amplifiers with the output of the second amplifier fed back into the input of the first amplifier through an <u>RC</u> network. This network consists of two capacitors C_1 and C_2 and two resistors R_1 and R_2 . By connecting them as shown, with R_1 and C_1 in parallel and with R_2 and C_2 in series, the fed-back voltage is in correct phase for the circuit to oscillate at one frequency, the frequency \underline{f} , where

$$\underline{f} = \frac{1}{2\pi \sqrt{C_1 R_1 C_2 R_2}}$$
(4-8)

Hence, the frequency of oscillation depends on the values of these constants; if any one of them is varied, the frequency will vary.

EXAMPLE. Determine the frequency of oscillation of an <u>RC</u> oscillator of the type shown in Fig. 4-12 if the two frequency-determining capacitors are 100 *H*⁴f each, and the two frequency-determining resistors are 20,000 ohms each.

The values given can be substituted directly into Eq. 4-8, 100×10^{-12} for $\underline{C_1}$ and $\underline{C_2}$ and 2×10^4 for $\underline{R_1}$ and $\underline{R_2}$, giving

$$\frac{f}{2 \times 3.14 \times \sqrt{100 \times 10^{-12} \times 100 \times 10^{-12} \times 2 \times 10^4 \times 2 \times 10^4}}$$

f = 79 kc Ans.

To produce frequency modulation, one or both of the frequencydetermining resistors may be replaced by an electronicly variable resistance; such a resistance may be a cathode follower whose grid voltage is controlled by the audio-input voltage. This is illustrated



Fig. 4-13. An <u>RC</u> oscillator used as an f-m generator by replacing one of the frequency-determining resistors by a cathode follower.

in the circuit shown in Fig. 4-13. The resistor \underline{R}_1 of the oscillator shown in Fig. 4-12 has been replaced by a cathode follower. A grid leak \underline{R}_g of high resistive value has been added to the circuit to prevent a static charge from being built up on the grid. The audio input is fed into the grid of the cathode follower. The cathode resistor \underline{R}_c has a value which, when divided into 1, yields a result that is very much smaller than the transconductance of the tube used in the cathode follower. This value gives high sensitivity of control.

The cathode follower simulates an apparent resistance \underline{R}_t across a and b, the terminals of \underline{R}_1 . Its value is determined by the transconductance of the cathode-follower tube. This resistance \underline{R}_t is equal to 1 over the transconductance of the tube; thus,

$$\underline{\mathbf{R}}_{\underline{\mathbf{t}}} = \frac{1}{\underline{g}_{\underline{\mathbf{m}}}} \tag{4-9}$$

when the value of $\underline{R}_{\underline{C}}$ is comparatively high. Hence, the cathode follower produces a variable resistance across the terminals of the frequency-producing resistor which, in turn, causes the frequency of the oscillator to change with the audio voltage. In other

words, the simulated resistance $\underline{R}_{\underline{t}}^{*}$ is substituted for the resistor $\underline{R}_{\underline{1}}$ in the frequency-determining equation (Eq. 4-9). Making this substitution we obtain

$$\underline{\mathbf{f}} = \frac{\sqrt{\underline{g}_{\underline{\mathbf{m}}}}}{2^{\underline{\gamma}}\sqrt{\underline{C}_{1}\underline{C}_{2}\underline{\mathbf{R}}_{2}}} \tag{4-10}$$

The frequency of the <u>RC</u> oscillator will vary with the square root of $\underline{g_m}$. If both resistors $\underline{R_1}$ and $\underline{R_2}$ of the <u>RC</u> oscillator were replaced by cathode followers, the frequency would then vary directly with the transconductance. The equation would be similar to the one given in Eq. 4-10, except that $\underline{R_2}$ would be removed from the denominator and the square-root sign removed from the $\underline{g_m}$.

$$\underline{\mathbf{f}} = \frac{g_{\underline{\mathbf{m}}}}{2\pi\sqrt{\underline{\mathbf{C}}_{1}\underline{\mathbf{C}}_{2}}} \tag{4-11}$$

EXAMPLE. Determine the frequency variation of an <u>RC</u> oscillator whose frequency-determining capacitors are each 100 $\mu\mu$ f. They are used in a circuit like that shown in Fig. 4-12 but with both frequency-determining resistors replaced by cathode followers. The transconductance of the cathode followers is varied by the audio voltage of 1,500 to 2,000 μ mhos.

This problem meets the requirements for Eq. 4-11. Substituting into Eq. 4-11 we obtain

$$\underline{f}_{1} = \frac{1,500 \times 10^{-0}}{2 \times 3.14 \times \sqrt{100 \times 10^{-12} \times 100 \times 10^{-12}}}$$

and

$$\underline{f}_2 = \frac{2,000 \times 10^{-6}}{2 \times 3.14 \times \sqrt{100 \times 10^{-12} \times 100 \times 10^{-12}}}$$

where the frequency varies between \underline{f}_1 and \underline{f}_2 . Solving the two equations we find that the frequency varies from 2.4 to 3.2 megacycles.

There are many other variations of the <u>RC</u> oscillator, besides the one shown in Fig. 4-12, which may be used as an f-m generator. Any of the <u>RC</u> oscillators in which a frequency-determining element, a resistor or a capacitor, can be replaced by an electronicly simulated impedance may be so used. In the case of a capacitor, a reactance tube can be used as the variable impedance. Any of these cases can be analyzed in a manner similar to the foregoing case. It will be found that with the use of television pentode amplifier tubes, nearly all the <u>RC</u> oscillators will work well up to about 4 megacycles. If it is desired to use them at still higher frequencies, compensating circuits similar to those used in video amplifiers should be employed in the amplifier circuits of the oscillator. Very often amplitude modulation of the output radio frequency may be reduced to a negligible amount by the proper choice of operating constants.

4-9. The Inductively Coupled F-M Generator. A variable reactance may be introduced into the tank circuit of an oscillator by inductive coupling and without the use of phase-shifting circuits as



Fig. 4-14. A modulator circuit wherein a variable reactance is introduced into the tank circuit by means of inductive coupling.

required in many of the other methods.³ In Fig. 4-14 is shown the basic modulator circuit. The tank circuit, used for the oscillator, consists of the inductance \underline{L} and the capacitor \underline{C} . Coupled into the inductance \underline{L} with a mutual inductance of \underline{M} is the modulating circuit made up of an inductance $\underline{L}_{\underline{S}}$, a capacitor $\underline{C}_{\underline{S}}$, and a variable resistor $\underline{R}_{\underline{S}}$. This modulating, or secondary, circuit couples into the tank circuit a reactance which is dependent on the value of the resistor $\underline{R}_{\underline{S}}$. As this resistor varies in value, it causes the reactance coupled into the tank circuit to vary in value and thus causes



Fig. 4-15. An f-m generator employing a variable reactance coupled into the tank circuit by inductive coupling. The plate resistance of a 6F6 tube is used as a variable resistance to control the frequency of oscillation.

the frequency of the oscillator to change. A practical f-m modulator is obtained by replacing the variable resistance with an electronic resistance which can be varied by the audio voltage.

Figure 4-15 shows a practical circuit employing the inductively

coupled reactance. An 802 tube is used as the oscillator at a mean frequency of 2.5 megacycles. An inductance of 16.78 μ h is employed. Inductively coupled to this inductance, with a mutual inductance of 1 μ h, is the modulator circuit. The modulator circuit consists of a 100- μ f capacitor (maximum value), a 37.4- μ h inductance, and the plate circuit of a 6F6, all in parallel. The plate impedance of the 6F6 tube is used as the variable resistance. The audio modulating voltage is fed into the grid circuit of the 6F6, and as the grid voltage is varied the a-c plate resistance varies, causing a variation in the value of the reactance coupled into the tank circuit of the oscillator which, in turn, causes the output frequency to vary. About a 20-kc variation can be obtained at 2.5 megacycles.

REFERENCES

1. A.V.Eastman and E.D.Scott, "Transmission Lines as Frequency Modulators," Proc. IRE, July, 1934, p. 878.

2. M.Artzt, "Frequency Modulation of Resistance Capacitance Oscillators," Proc. IRE, July, 1944, p. 409.

3. Bruce E. Montgomery, "An Inductively Coupled Frequency Modulator," Proc. IRE, October, 1941, p. 559.

QUESTIONS

1. Determine the frequency of oscillation of an oscillator whose tank circuit consists of an inductance of 40 μ h shunted by 80 $\mu\mu$ f of capacitance.

Why cannot a crystal-controlled oscillator be used in a direct f-m generator?
Explain how the reactance-tube circuit simulates a reactance.

Explain how the reactance tube is used to produce frequency modulation.

Explain how a reactance tube is used to produce frequency incadation.
Determine the variation in simulated capacitive reactance in a reactance-tube

5. Determine the variation in simulated capacitive reactance in a reactance tast circuit employing a tube whose transconductance varies from 1,000 to 1,200 μ mhos at a frequency of about 1 megacycle.

6. What is the variation in simulated inductive reactance for a circuit employing the tube of Question 5 and used at the same frequency?

7. Determine the variation in the frequency of the oscillator of Question 1 when the capacitive reactance tube of Question 5 is used to modulate it. (Note that the frequency at which it is being used is no longer 1 megacycle.)

8. Discuss the operation and advantages of the balanced reactance tube circuit.

9. What variation in input capacitance is possible with a tube whose transconductance can vary from 1,200 to 1,400 μ mhos and which employs a plate resistance of 25,000 ohms? The over-all grid-to-plate capacitance is 3 μ f.

10. Explain how a variation in input capacitance can be employed in an f-m generator.

11. Describe the operation of the transmission-line modulator.

12. What is meant by the characteristic impedance of a transmission line?

13. How can a resistance-capacitance oscillator be converted into an f-m generator?

14. In an <u>RC</u> oscillator of the type shown in Fig. 4-12, what is the frequency of oscillation if $\frac{1}{1000}$

$$\underline{R}_1 = \underline{R}_2 = 1,000 \text{ ohms}$$

 $C_1 = C_2 = 40 \ \mu\mu f$

and

15. What frequency variation can be expected from the oscillator of Question 14. when the resistor \underline{R}_1 is replaced by a cathode follower whose transconductance can vary from 1,000 to 1,500 /mhos?

16. Determine the frequency variation when both \underline{R}_1 and \underline{R}_2 of the oscillator of Question 15 are replaced by cathode followers of the type described in Question 15. 17. From the literature on RC oscillators obtain the circuit of another type of

RC oscillator and show how it may be used as an f-m generator.

18. Describe the operation of the inductive type of f-m generator.

CHAPTER 5

Frequency-Control Circuits

5-1. The Need for Frequency Control. F-m transmitters employing direct f-m generation utilize a frequency-control circuit for carrier-frequency stability in order to meet the stringent requirements of the FCC. In practically all direct f-m generators, the oscillator frequency is modulated by varying one of the frequencydetermining elements in the resonant circuit of the oscillator. To allow for large deviation during modulation, the frequency-determining circuit should have a low Q; this requires a low-capacitance high-inductance circuit. However, accurate frequency control requires a very high value of Q. These two requirements are conflicting and are only met by the use of a frequency-control circuit.

The FCC regulations specify that the carrier frequency must be kept within $\pm 2,000$ cps of the assigned frequency in the f-m broadcast band; at 100 megacycles this means a stability of 0.002 per cent. Using ordinary capacitors and inductors in the resonant circuit of an oscillator with the highest value of Q (which will be the best condition for stability), and with the best temperature-compensation technique known, the oscillator still could not meet the stability requirement. The required order of stability is only obtainable with a crystal oscillator acting through a frequency-control circuit.

The carrier frequency control circuit should be insensitive to frequency variations caused by modulation so that the wave itself will not be demodulated by the frequency-control system; only the center, or carrier, frequency should be controlled. Thus, in addition to the problem of frequency regulation, there is the added problem of differentiating between the desired frequency variation (the frequency modulation) and the carrier frequency drift. This is not very difficult to do since the frequency drift of the carrier takes place at a very low frequency, much lower than the lowest audio frequency used for modulation. Thus, a low-pass filter in the control circuit can eliminate the modulation frequencies.

5-2. Carrier Amplitude in F-M Signals. Basically, the general

68

control method consists of comparing the f-m carrier frequency with the output frequency of a crystal-controlled oscillator which is operating well within the permitted $\pm 2,000$ cps to allow for any inexactness in the control circuit. The output of the crystal-controlled oscillator and the carrier of the f-m wave are compared in what may be called a "comparator circuit" whose output, in turn, is used to regulate the center frequency of the direct f-m generator.

One suggested method of obtaining the carrier is to employ a very narrow band filter that would pass only the carrier. However, with some degrees of modulation the carrier is actually zero (see discussion in Chap. 3). No carrier would be available at such times for operation of the comparator.

At other degrees of modulation the amplitude of the carrier is so much smaller than its unmodulated value that its use is limited. This reduction in amplitude of the carrier is illustrated in Fig. 5-1, where the ratio of frequency deviation to modulating frequency is only 15. As we have seen previously, the modulated wave contains the same amount of energy as the unmodulated carrier; the energy found in the sidebands has been extracted from the original carrier energy. In



Fig. 5-1. Relative amplitudes of an f-m wave carrier and sidebands when an audio frequency \underline{F} is used and the ratio of frequency deviation to modulating frequency is only 15.

Fig. 5-1 the carrier amplitude is only 0.014 of its unmodulated value, while the amplitudes of many of the sidebands are very much larger. As the frequency and amplitude of the audio signal are varied (a practical consideration), the amplitude of the carrier will vary enormously. It could not be reliably employed in a frequency-regulating system.

Fortunately, there is a simple solution. If the frequency of the

transmitted wave should be halved, usually by means of a frequencydivider circuit, the frequency deviation would also be halved. The audio frequency, the frequency of modulation, would not be affected. Hence, the ratio of frequency deviation to modulating frequency would also be halved. In other words, if the frequency of the transmitted wave should be divided by any factor, the ratio of the frequency deviation to the modulating frequency would also be divided by the same factor. When this ratio is decreased, as was discussed in Chap. 3, the amplitude of the carrier is increased. This is illustrated in



Fig. 5-2. Relative amplitudes of an f-m wave carrier and sidebands when an audio frequency \underline{F} is used and the ratio of frequency deviation to modulating frequency has been reduced to 0.2. The amplitude of the carrier is now adequate for use in a frequency-control circuit.

Fig. 5-2, where the ratio of frequency deviation to modulating frequency is 0.2. The carrier, in this latter case, is almost the same as its unmodulated value. It can be obtained from the f-m wave whose sideband distribution is shown in Fig. 5-1 by dividing the frequency of the wave with a factor of 75. In practical transmitters a frequency division of as high as 1,000 may be employed, being accomplished by passing the f-m wave through successive stages of frequency dividers. Because the carrier is so much larger in amplitude whan the sidebands, the output of the dividers can be used directly in the frequency-comparator circuit without the use of any narrow-band filters, although a low-pass filter to prevent demodulation is sometimes used. This low-pass filter may be employed at the output of the comparator to filter out the frequency variations caused by the modulating voltage, as discussed previously.

5-3. Discriminator Control. One type of circuit that may be used for frequency control is shown in Fig. 5-3. This is one of the discriminator circuits used in f-m receivers. It will be discussed in detail later (Chap. 9) but, briefly, it operates in this manner: A d-c voltage is obtained at the output which is dependent on the frequency of the input r-f signal. The circuit consisting of \underline{L}_1 and \underline{C}_1 and the circuit consisting of \underline{L}_2 and \underline{C}_2 are both tuned to the midfrequency of the f-m wave. \underline{C}_3 is a coupling capacitor which adds the voltage across the $\underline{L}_1\underline{C}_1$ circuit to half the voltage induced in \underline{L}_2 , opposite halves being used for each diode. As the frequency of the input r-f signal varies, the amplitudes of the signals introduced across the



Fig. 5-3. A frequency-discriminator circuit which is tuned to the mid-radio frequency. The polarity and amplitude of the d-c output is dependent on the deviation of the input frequency from this mid-frequency.

two diodes vary, giving rise to a voltage at the output which is proportional to the frequency of the input signal.

Without a crystal-controlled frequency, the stability rests solely in the stability of the circuits involved. The necessary circuit stability is very difficult to obtain. For this reason, when the ordinary discriminator is used for frequency control in a broadcast f-m transmitter, the transmitted frequency is converted to a low frequency by means of an oscillator (crystal-controlled) converter circuit so that the percentage variation is increased in the control circuit.

The other method of discriminator control is called a "phase discriminator." In actuality, the phase discriminator is a phase comparator, inasmuch as it compares the phase of the crystal-generated constant-frequency wave with that of the mid-frequency wave of the f-m signal. By maintaining a constant phase relationship, it automatically maintains a constant frequency for the carrier of the f-m wave.

Let us picture the two frequencies as two rotating wheels. A phase discriminator would lock them in constant relative position, in other words, in constant phase. One wheel may precede the other wheel, for instance, by a constant, let us say 10 deg, or one thirty-sixth of a rotation. To maintain this lead, never increasing or decreasing it, both wheels would have to turn at the same speed. In the same manner, if the phase between two r-f waves is kept constant, their frequencies will be equal. The phase discriminator does not have to keep the phase constant but near enough to constant to prevent a frequency variation of greater than $\pm 2,000$ cps, the allowable deviation.

Figure 5-4 shows the circuit diagram of a phase discriminator. Two diodes, the two sections of a 6H6, are employed with two inputs. both untuned. One of the two signals to be compared is applied at input <u>A</u>, and the other at input <u>B</u>. For instance, one might be the crystal-generated constant-frequency wave; while the other would be the f-m wave divided sufficiently so that the ratio of frequency



Fig. 5-4. Circuit of the phase discriminator where the output voltage is dependent on the phase between the two input voltages at <u>A</u> and <u>B</u>.

deviation to modulating frequency is very small. We see in Fig. 5-4 that the voltage from input <u>A</u> is applied to the two diodes in pushpull, while the voltage from input <u>B</u> is applied to the two diodes in parallel. For reference, the diodes are labeled diode I and diode II. The push-pull transformer has a 1:2 voltage ratio while the parallel transformer has a 1:1 voltage ratio. Hence, the circuit is equivalent to applying the sum of the two r-f input voltages across diode I while applying the difference between the two voltages across diode II.

Since these inputs are a-c voltages, the sum and difference take into account the phase relationships between the two voltages. First, briefly, the circuit operates as follows: Diode I rectifies the sum voltage and produces a d-c voltage across \underline{R}_1 equal to the amplitude of the sum voltage. Diode II rectifies the difference voltage and produces a d-c voltage of opposite polarity across \underline{R}_2 equal to the amplitude of the difference voltage. The output taken across both \underline{R}_1 and \underline{R}_2 is direct voltage proportional to the amplitude of the sum voltage minus the amplitude of the difference voltage.

To analyze the operation of this type of circuit, it is helpful to use the phasor diagrams. In Fig. 5-5 is shown a phasor diagram where the voltage at input <u>B</u> leads the voltage at input <u>A</u> by 90 deg. The voltage V_{O2} across the points oa is added to the voltage V_{d0} across the points <u>do</u>. It is equivalent to adding the voltage at input <u>A</u> to the voltage at input <u>B</u>. It yields the voltage V_{I} across diode I. Similarly, the voltage \underline{V}_{Ob} across the points <u>ob</u> is added to the voltage \underline{V}_{d0} across the points <u>do</u>. Inasmuch as the voltage \underline{V}_{Ob} is the negative of the voltage across input <u>A</u> (lags it by 180 deg), this addition is equivalent to obtaining the difference between the voltage \underline{V}_{II} across diode II. In this case the magnitudes of the two voltages impressed across the diodes are equal; the difference between the two d-c voltages obtained at their outputs, the output voltage of the comparator, is zero. The detected voltages cancel one another.

Let us suppose, now, that the voltage at input <u>B</u> begins to lead the voltage at input A by an angle θ deg which is less than 90 deg. This





Fig. 5-5. Phasor diagram of the voltages in the phase discriminator when the d-c voltage output is zero.

Fig. 5-6. A phasor diagram of the voltage relationships within the phase discriminator when the voltage at input <u>B</u> lags the voltage at input <u>A</u> by an angle $\overline{\Theta}$ deg, less than 90 deg.

is shown in the phasor diagram of Fig. 5-6. The magnitude of the voltage V_I across diode I is now greater than the magnitude of the voltage \underline{V}_{II} across diode II. When these voltages are rectified by the diodes and their difference obtained across the resistors \underline{R}_1 and \underline{R}_2 , a positive output voltage is obtained. This output voltage is then fed back to the f-m generator to halt the change in phase. Similarly, if the angle θ should be greater than 90 deg, a negative output voltage will be obtained which will cause the f-m generator frequency to shift the other way and halt the forward phase shift. In this manner, by locking the phase, the two frequencies are synchronized.

The output circuit of the phase discriminator is so constructed that, with the proper constants, it acts as a low-pass 10-cps filter. Hence, the output voltage contains only the slow mean-frequency drifts – all of the modulation variations being removed.

In Fig. 5-7 is shown a curve of the output voltage \underline{V}_R as a function of the phase angle between the two voltages being compared in the phase discriminator. The vertical axis is scaled in terms of the input r-f voltage amplitude \underline{V} . Thus, as the phase between the two voltages is varied from 0 to 180 deg, the output voltage — the difference in amplitude between \underline{V}_I and \underline{V}_{II} —varies from +2<u>V</u> to -2<u>V</u>. Below 0 deg and above 180 deg, the curve bends back toward the axis, indicating that control is lost at those points. The phase relationship between the two voltages has to be kept within those 180 deg. Within that variation, the curve is linear, showing that very good control is obtainable.

5-4. Motor Control. Controlling the frequency of an f-m generator of the reactance-tube, or similar, type by varying the bias voltage on the modulating tube necessarily limits the amount of frequency excursion allowable for the audio modulating voltage. Enough of the linear characteristic of the modulating source has to be allowed, above and below the audio-voltage excursion, to permit the control



Fig. 5-7. A curve showing the difference in amplitude between V_{I} and V_{II} . It will be noted that control is lost at $\Theta = 0$ deg and 180 deg.

source to shift the bias and thereby correct the carrier frequency of the f-m wave. This is illustrated in Fig. 5-8. Not all the linear



Fig. 5-8. A modulating characteristic for bias-frequency control. The whole linear portion of the characteristic cannot be used for modulation because allowance has to be made for the control-voltage shift of the operating point.

characteristic is used for the f-m frequency excursion, because allowance has to be made for the shift in bias (the operating point) by the control voltage. At all times there is a portion of the linear characteristic which is not being used.

By the use of motor control of center frequency, the necessity for bias variation is removed, and the complete modulating characteristic can be employed.¹ Instead of varying the bias on the modulating tube, a motor changes the setting of a variable capacitor in the tank circuit of the oscillator. The position of the motor control is determined by the difference in frequency between the carrier of the f-m wave and the output of a crystal oscillator.



Fig. 5-9. A block diagram of a method of comparing two frequencies, one at input <u>A</u> and the other at input <u>B</u>, to obtain a rotary field for a two-phase control motor.

In Fig. 5-9 appears the block diagram of a method of comparing two frequencies, one at input A and one at input B, to obtain a rotary field to actuate a control motor, the control motor being a small two-phase a-c motor. In a normal transmitter circuit, the voltage for input A would be obtained from a crystal oscillator and the voltage for input \overline{B} would be obtained from the f-m generator, proper dividing circuits being used in both cases. The voltage from input A is separated into two parts, one being shifted 90 deg in phase with respect to the other. Each of these is fed into a balanced modulator. Also fed into each balanced modulator is the voltage from input B. The balanced modulator is a mixer which takes two inputs and mixes them together to yield the beat, or difference, frequency. In other words, the output of each balanced modulator is a voltage at a frequency equal to the difference in frequency between the voltage at input A and the voltage at input B. The output of the balanced modulator is so arranged that only the beat frequency is obtained at the output. Inasmuch as one of the balanced modulators has the input from A shifted 90 deg, the beat frequency in its output will also be shifted 90 deg. Thus, the two outputs of the balanced modulators supply two voltages, 90 deg out of time phase, necessary for running a twophase motor. Within the motor are two coils in 90-deg space phase (at right angles to one another) which, when supplied with the proper voltages - one 90 deg out of phase with the other - generate a rotary field that turns the armature.

The motor is actually the controlling device. When the two frequencies are not the same, a beat note is generated in the balancedmodulator comparator. This beat note actuates the motor turning the armature until the beat note disappears. At that point no voltage is available to run the motor and the motor stops, leaving the two frequencies in synchronism. The motor is reversible; when the frequency shifts the other way and reverses the phase relationships, the motor again corrects the frequency of the f-m generator by turning in the opposite direction.



Fig. 5-10. Stabilizing circuit for use with motor control of a variable capacitor in the oscillator circuit.

In Fig. 5-10 is shown the circuit diagram for a stabilizing circuit using as a motor control a variable capacitor in the tank circuit of the f-m generator. The signal from input A first goes through the phase-shifting network. Each leg of the phase-shifting network consists of a resistor and capacitor whose impedances are equal in magnitude so that the current flowing through each leg leads the voltage impressed across it by 45 deg. (The crystal-oscillator voltage, the voltage impressed across input A, is constant in frequency in order that this requirement may be met exactly.) The voltage across R1 is in phase with the current passing through it; hence, the voltage impressed on balanced modulator I, the voltage across R_1 , leads the voltage impressed across input A by 45 deg. The voltage across C2, on the other hand, lags the current flowing through it by 90 deg. But the current through the capacitor leads the voltage impressed across input A by 45 deg; hence, the voltage across C_2 , the voltage impressed on modulator II, lags the voltage impressed across input A by 45 deg. Since the magnitude of the capacitor reactance and the value of the resistor are equal, the two output voltages of the phase shifter are also equal. Thus, we have equal voltages, with a phase relationship of 90 deg to one another, impressed on each of the balanced modulators.

The balanced modulators consist of two nonlinear tubes in pushpull. The voltage from input B is impressed on the tubes in pushpull by using a push-pull transformer. The phase-shifted voltage from input <u>A</u> is impressed on the tubes in parallel, from the center tap of the push-pull transformer to ground. The nonlinear characteristics of the tubes produce, at their outputs, the beat frequency between the two input voltages. The two output beat-frequency voltages are fed into the field coils of a small two-phase motor which turns the rotor of a variable capacitor in the tank circuit of the f-m generator.

This method of control possesses the feature of the motor armature turning only when a beat frequency exists to produce a rotation of the field. Changes in the intensity of either the crystal-oscillator voltage or the f-m signal voltage will not actuate the motor armature; hence, the adjustment of the frequency and, therefore, the frequency of the f-m generator are unaffected by any variations in circuit gain. If a progressive change in the relative phase of the two frequencies takes place, no matter how slight it may be, the currents through the motor windings will slowly rise and fall, causing the armature to rotate until the difference between the two frequencies is zero.

Because an actual mechanical adjustment of frequency is employed, the balance and intensity of the electric currents in the modulator are undisturbed from their optimum adjustment even though the control should be called upon to correct a frequency drift of many megacycles in the transmitted wave. A failure of the frequencycontrol system will not interrupt the program but will leave the frequency at the last corrected point.



Fig. 5-11. A graphical representation of a system of frequency control where the total time the f-m frequency is above the assigned frequency is made equal to the total time it is below.

5-5. Integrated-Pulse Control.² The foregoing systems of regulation depend on maintaining a correct center frequency in an f-m oscillator. Other systems of frequency control have been suggested which would correct to a different point. One of these would correct to where the total time that the f-m generator frequency is above the assigned frequency would be equal to the total time it is below the assigned frequency. This is illustrated graphically in Fig. 5-11. If the modulation were a pure sine wave, as shown in the figure, where the time interval \underline{T}_1 is equal to the time interval \underline{T}_2 , such a system would correct to the assigned frequency. With unsymmetrical modulation, like the nonsinusoidal wave shown in the figure, some of the sidebands would spill over into the adjacent channel; the carrier frequency would have shifted.

Another system has been proposed which would adjust the frequency so that the maximum excursion above the assigned frequency would be equal to the maximum excursion below the assigned fre-



Fig. 5-12. A frequency regulation system where the frequency is adjusted so that the frequency excursion of the f-m wave above the assigned frequency is equal to the frequency excursion below the assigned frequency.

quency. This is illustrated in graphical form in Fig. 5-12. \underline{E}_1 and \underline{E}_2 are the peak amplitudes of the modulating voltage, and the mid-frequency of the f-m generator is adjusted so that the frequencies generated at these voltages are equidistant from the mid-frequency. Such a system would keep the sidebands within the assigned frequency spectrum under steady-state conditions, even though the wave form of the modulating voltage were unsymmetrical. Some difficulties may be encountered under transient conditions, such as are found in audio-program material where a nonsymmetrical wave form may by followed immediately by another nonsymmetrical wave form of opposite polarity. The speed of correction has to be exact because if it is too slow there may be considerable spilling over of the sidebands into adjacent channels, and if it is too fast unwanted frequencies may be generated. Very often one or the other may take place. Figure 5-13 depicts a still different system of frequency regulation; this system is based on the area enclosed on each side of the assigned frequency, the area being based on the actual frequency shift and the time interval of the shift. It is equivalent to equalizing the areas under the modulating voltages, areas $\underline{A_1}$ and $\underline{A_2}$ as shown



Fig. 5-13. A frequency regulation system based on the area enclosed on each side of the assigned frequency.

in Fig. 5-13, and will maintain the correct carrier frequency.

If a crystal oscillator is built to operate at the assigned center frequency and mixed with the output of the f-m oscillator in a nonlinear circuit, a beat note will be produced which has an instantaneous frequency equal to the instantaneous deviation of the modulated oscillator. The total number of cycles of beat note produced while the oscillator is on one side of the assigned frequency will be exactly proportional to the area enclosed by that part of the curve. It follows, therefore, that if the total number of cycles produced while the oscillator is on the high side of the assigned frequency is equal to the total number of cycles produced while the oscillator is low in frequency, then the transmitter is operating at the correct point.

The integrated-pulse system of control has been developed to take advantage of this latter method. Each cycle of beat frequency between the signal frequency and the reference frequency is used to generate a pulse. The pulses are separated into two circuits, one circuit receiving the pulses when the signal frequency is higher than the reference frequency, and the other receiving the pulses when the signal frequency is lower. A pulse-counting circuit is so arranged that when a pulse appears on one of these circuits, a definite charge is transferred from a point of fixed potential to a storage capacitor. When a pulse appears on the other circuit, a charge of the same magnitude is transferred from the storage capacitor to a point of fixed potential. Since no bleeder is employed across the storage capacitor, the charge on the capacitor tends to remain constant during any period when there are no pulses. The voltage appearing across the capacitor, which is proportional to the charge stored in it, is used as a control on the reactance tube which controls the frequency of the master f-m oscillator.



Fig. 5-14. A block diagram of the phase shifter and mixer portion of the integrated-pulse control.

Figure 5-14 shows a block diagram of the phase shifter and mixer circuits of the control system. The signal from the f-m oscillator is designated by \underline{F}_{s} , and the reference signal from the crystal oscillator by $\underline{F_r}$. The crystal frequency signal is fed through two 45-deg phase-shift networks, each consisting of a resistor and capacitor of the same number of ohms (similar to the phase-shift network shown in Fig. 5-10). One part of the signal is shifted forward in phase by 45 deg and the other retarded by 45 deg. Mixer A and mixer B are used to mix these quadrature voltages with the f-m signal-frequency voltage. Figure 5-15a shows the relative output of the two mixers when the frequency of the modulated oscillator is higher than that of the crystal oscillator. It will be noticed that the output of mixer B leads the output of mixer A by 90 deg. In Figure 5-15b, the signal frequency is lower than the reference frequency, and consequently the output of mixer B lags the output of mixer A by 90 deg, the phase of the output being dependent on the relative frequencies.

The output of mixer <u>A</u> is used to trigger a direct-coupled multivibrator. This multivibrator serves as an electronic switch to make square waves that correspond with the sine-wave input. This is illustrated in Fig. 5-16. Since the input to the multivibrator is much greater than the amount required to trigger it, the time at which the multivibrator turns over (begins the square wave) will be approximately the time at which the voltage of the output of mixer <u>A</u> passes through zero. At this time the output of mixer <u>B</u> is at either a positive or negative peak. This may be seen by following one of the dotted lines down Fig. 5-16.

The voltage on each of the two multivibrator plates is differentiated by a series capacitor and shunt resistor. The resultant two voltages appear as a series of narrow pulses of opposite polarity at the output points \underline{D} and \underline{E} . We note how the relative polarities of the pulse voltages are the same for both frequency relationships.



Fig. 5-15. The outputs of the two mixers shown in Fig. 5-14 when the signal frequency is higher than the reference frequency and vice versa.

In Fig. 5-17 is shown the pulse-selection circuit. D and E refer to the same points as in the circuit of Fig. 5-16. The voltages at point \underline{F} and \underline{G} show the result when the pulses are superimposed on the output of mixer B. It will be noticed that when the pulses appearing at point D are superimposed on the output of mixer B, the pulses subtract from the sine wave when the signal frequency is higher than the reference frequency, and add to the sine wave when the signal frequency is lower than the reference frequency. In the case of the pulses appearing at point E, the pulses add to the sine wave if the signal frequency is higher, and subtract if it is lower. These two signals, appearing at F and G in Fig. 5-17, are passed through biased diodes which are used as pulse discriminators. The bias on these diodes is set just above the peak value of the output of mixer B. The result is that when the pulses add to the sine wave, the bias is overcome and the pulse is passed through the diode. When the pulse subtracts from the sine wave, the bias prevents the diode from conducting and the pulse is not passed. This arrangement serves to separate the pulses into two circuits, one circuit energized by one pulse for each cycle of beat frequency when the signal frequency is high, and the other circuit energized by one pulse for each cycle of beat frequency when the signal frequency is low.

The output from <u>H</u> and <u>J</u> of the pulse discriminator is fed into the two pulse counters, as shown in Fig. 5-18. The two pulse counters,



Fig. 5-16. The generation of pulses from the output of mixer \underline{A} for both frequency conditions.

really integrators, are arranged in a balanced circuit to control the charge on the storage capacitor \underline{C}_{S} . Two 3-v sources are used, one positive and one negative. When a pulse actuates point \underline{K} from the source \underline{H} , it causes a current to flow through the diode \underline{M} , which is replenished by a current from the -3v source. This causes the capacitor \underline{C}_{S} to lose electrons, and creates a positive voltage at the point \underline{P} . When a pulse arrives at the point \underline{L} from the source \underline{J} , it causes a current to flow through the diode N, which it replenishes from the +3v source. Diode N is reversed, inasmuch as the plate is connected to the storage capacitor \underline{C}_{S} ; hence, a charge will be removed from the capacitor by electrons flowing onto the capacitor plates. The voltage on the capacitor at the point \underline{P} will decrease. The charge conveyed by each pulse, whether at \underline{K} or \underline{L} , is constant; therefore, the net remaining charge on the capacitor \underline{C}_{S} , the voltage

at the point <u>P</u>, is a measure of the difference in the number of pulses from each input, <u>H</u> and <u>J</u>.



Fig. 5-17. Pulse separation circuit. Point <u>H</u> receives pulses when the signal frequency is lower than the reference frequency, and point <u>J</u> when it is higher.

The voltage across the storage capacitor is fed into a cathode follower which, in turn, controls the bias on the modulator tube. Since the modulator tube controls the frequency of the modulated oscillator, the frequency is a direct function of the charge on the storage capacitor C_S . It will be noticed that there is no bleeder resistor across the storage capacitor, hence, the system has no natural frequency which the frequency control must overcome.

Figure 5-19 shows a block diagram of the complete system. Oscillograms of the voltages at the interconnection points are indicated on the diagram for both frequency conditions, the upper waves for the case where the signal frequency is higher than the reference crystal-oscillator frequency, and the lower waves for the case where the signal frequency is lower. If the average frequency of the modulated oscillator is different from the reference frequency, the charge on the storage capacitor is







NOTE: Upper wave trains when the signal frequency is higher than the reference frequency. Lower trains when the signal frequency is lower than the reference frequency.

Fig. 5-19. A block diagram of the complete integrated-pulse frequency-control system showing the interconnection of the various circuits.

continually changing in the proper direction to overcome this difference. When the difference has been overcome, the system becomes balanced, the only tendency to pull off being because of stray leakage, which is negligible.

In the practical application of this control method, the modulated oscillator is operated on one ninth of the assigned frequency of the transmitter. It has been found that simple tuned circuits in the multiplier stages provide adequate selectivity without cutting the sidebands when the modulation is applied. When very much lower frequencies are used, special band-pass filters must be employed to provide proper band width and selectivity at the same time.

5-6. Frequency-Divider Circuits. A frequency divider is necessary in an f-m transmitter circuit (as has been discussed previously) when it is desired to lower the ratio of frequency deviation to modulating frequency. The necessary frequency division may be anywhere from about 100 to over 1,000. Fortunately, the amplitude of the wave remains constant in f-m so that only the frequency need be considered in the division. Hence, devices like multivibrators, relaxation oscillators, and pulse counters can be used.



Fig. 5-20. A multivibrator frequency-divider circuit with the output taken off across a small value of cathode resistor.

In Fig. 5-20 is shown a multivibrator circuit for use in frequency division. The multivibrator, consisting of two resistance-capacitance coupled amplifiers with the output of the second tube coupled back into the input of the first tube, has its frequency determined by the values of the grid resistors and the coupling capacitors \underline{R}_1 , \underline{C}_1 , \underline{R}_2 , \underline{C}_2 . The free frequency of oscillation of the multivibrator (no input-synchronizing voltage) \underline{f}_m is given approximately by

$$\underline{\underline{f}}_{\underline{\mathbf{m}}} = \frac{1}{\underline{\mathbf{R}}_{1}\underline{\mathbf{C}}_{1} + \underline{\mathbf{R}}_{2}\underline{\mathbf{C}}_{2}}$$
(5-1)

The circuit is designed so that this frequency $\underline{f_m}$ is 1/n the input frequency \underline{f} ; hence, in operation a frequency division of \underline{n} is obtained. A frequency division of anywhere from 2 to about 7 may be obtained

under stable conditions.

The operation of the multivibrator frequency divider can best be understood by examining the voltage and current relationships as



Fig. 5-21. The voltages and currents in the multivibrator shown in Fig. 5-20.

shown in Fig. 5-21. Without any synchronizing voltage the grid voltages would be similar; they would have the shape shown for the grid voltage of tube II (Fig. 5-21c). The plate current of one tube, say tube I, flows while the coupling, or grid, capacitor of the other tube, tube II, is discharging through its grid resistor. When the capacitor is discharged, the action switches and the plate current flows in tube II while tube I is having its grid capacitor discharged. The cathode resistor of tube I has flowing through it the plate current of tube I; therefore, its voltage has the same shape as the plate current.

When a synchronizing voltage is to be put on, the grid resistor of tube I is increased slightly so that the grid voltage is not completely discharged in the time allotted for that cycle. The synchronizing voltage, as seen from Fig. 5-20, is then superimposed on the grid-voltage characteristic. The synchronizing, or input, voltage will lock the multivibrator in at the point where the positive cycle of the synchronizing voltage will force the grid of tube I to positive at the beginning of the multivibrator cycle, the positive peak making up for the decrease caused by the larger grid resistor. In Fig. 5-20 this point is noted as the "sync" point. The figure shows the multivibrator being used as a divider with a factor of 3. The voltage across the cathode resistor, because it is similar in shape to the plate current, does not have any of the distortion present in the grid curves. A sine-wave output may be obtained by using a tuned circuit at this point. Example. What values of coupling capacitors and grid resistors should be used in a multivibrator that is to be used to divide a frequency of 90 kc by 5?

The free-oscillation frequency of the multivibrator $\underline{f_m}$ would be 90 divided by 5, or 18 kc. Using Eq. 5-1 and letting $\underline{R_1C_1}$ equal $\underline{R_2C_2}$, which is usually done, we obtain

$$\underline{\mathbf{f}}_{\underline{\mathbf{m}}} = \frac{1}{2\underline{\mathbf{R}}_{1}\underline{\mathbf{C}}_{1}}$$

And assuming a value of 20,000 ohms each for the grid resistors, we can solve for the value of the capacitors. These come out to be

$$C_1 = 27.5 \mu \mu f$$

To obtain a slightly lower frequency of free oscillation, as required by the action of synchronization, we choose the nearest larger round value, $30\mu\mu d$. Hence, one answer would be grid resistors of 20,000 ohms and grid capacitors of 30 $\mu\mu d$. These are only two of the many values which would yield the correct result.



Fig. 5-22. The circuit diagram of a blocking oscillator where $\underline{R}_{\underline{g}}$ and \underline{C} are the frequency-determining components.

Another type of frequency divider is the blocking oscillator shown in Fig. 5-22. The frequency of free oscillation in this type of oscillator is determined again by the discharge of a capacitor, in this case the capacitor C. This capacitor is discharged through the grid resistor \underline{R}_g . The transformer T is usually an iron-cored transformer which, with its distributed capacitance, resonates at a high frequency — high with respect to the frequency of oscillation of the blocking oscillator.

The operation of the blocking oscillator may be understood from the grid-voltage curve shown in Fig. 5-23. When the grid voltage is zero, the tube oscillates at the resonant frequency of the transformer \underline{T} - but only for a half cycle or less. During that part of the cycle the grid draws so much current that it plunges to negative, cutting off the oscillations. A large time constant is employed for the grid-capacitor — grid-resistor circuit. The grid capacitor then discharges through the grid resistor \underline{Rg} , as shown by the dotted curve in Fig. 5-23, provided that there is no synchronizing-voltage input. When the discharge curve reaches zero, the cycle begins over again. With no synchronizing voltage this would occur at the point <u>D</u>. The synchronizing voltage, however, is superimposed on the discharge-voltage curve and forces the grid to positive at one of its positive peaks. This is illustrated at the point called the "sync point" in



Fig. 5-23. The variation in grid voltage on a blocking oscillator. The dotted curve shows the capacitor discharge as it would occur without any synchronizing voltage.

Fig. 5-23. Since the natural period of discharge is longer than the desired output frequency, the synchronizing voltage will always trip off the cycle and produce an exact frequency division.

If the clearance $\underline{V}_{\underline{C}}$ between the previous positive peak and the zero (or trip) voltage is not great enough, the oscillator may trip on the wrong cycle and produce the wrong frequency. This often occurs when the frequency-division factor is large, greater than about 5. To prevent this, a tuned circuit called a "stabilizing circuit" is sometimes inserted in the grid circuit between the input and the transformer. This circuit is tuned to about 1 1/2 times the desired output frequency. With each pulse of grid current it produces a damped oscillation which, since it is added to the grid characteristic, pulls down the curve near the synch point and improves the clearance. A blocking oscillator with a stabilizing circuit is illustrated in Fig. 5-24.

Many other types of relaxation oscillators which can be used as frequency dividers have been designed. Any ordinary oscillator can be expected to produce relaxation oscillations when high-impedance tank circuits and close coupling between grid and plate circuits are used. By inserting the synchronizing voltage at the correct point, many of these may be used as dividers.

Still another type of frequency divider is the one employing the ring modulator (Fig. 5-25). In this type of divider, the ring modulator is used in a straightforward fashion to mix two frequencies, the input at a frequency of \underline{f} , and the modulating signal at a frequency of $\underline{f}/2$. These are mixed by applying the input signal across two corners

of the ring of rectifiers and applying the modulating signal across center-tapped resistors connected diagonally across the ring between



Fig. 5-24. A blocking oscillator with a stabilizing circuit in the grid input tuned to about 1.5 fm.



Fig. 5-25. A ring modulator with feedback used as a frequency divider with a factor of 2.

opposite corners. Calling the four terminals of the ring of rectifiers <u>A</u>, <u>B</u>, <u>C</u>, and <u>D</u>, we connect the input across <u>C</u> and <u>D</u>, and the modulation voltage across <u>E</u> and <u>F</u>, the center taps of resistors across <u>A</u> and <u>B</u> and <u>C</u> and <u>D</u>. The output is taken off across <u>A</u> and <u>B</u> and amplified. This output signal will be at the difference frequency, the difference between f and f/2 being also f/2. In the plate circuit of the amplifier is a tuned circuit, tuned to f/2, which is coupled to the output coil. The modulating voltage, which is at the same frequency, is also obtained from this tuned circuit. It sounds very much like lifting oneself up by one's boot straps, but it is not so and it works. Any small disturbance is sufficient to start the tuned circuit oscillating, and the feed-back circuit builds it up. Simply closing the switch for the circuit is sufficient to start it up.

Figure 5-26 depicts a pulse-counter type of frequency divider. This type of circuit is very similar to the blocking oscillator where the transformer T has a high resonant frequency and causes the grid to draw a pulse of current. This pulse of current causes the grid



Fig. 5-26. A pulse-counter type of frequencydivider circuit. This resembles the circuit of the blocking oscillator except that the capacitor C is discharged through a counter circuit instead of a resistor.

capacitor to become charged and makes the grid voltage very negative. Instead of allowing the charge to leak off on a resistor, as in the case of the blocking oscillator, a counter circuit is used. For every positive peak or pulse of voltage from the input, a small amount of charge is removed from the capacitor through the double diode. Hence, the capacitor is discharged in steps, and when a sufficient number of pulses has been applied to the input, the capacitor is discharged and another cycle started. If n pulses are needed to discharge the capacitor, the output frequency will be 1/n the input frequency.

REFERENCES

1. J. F. Morrison, "A New Broadcast Transmitter Circuit Design for Frequency Modulation," Proc. IRE, October, 1940, p. 444. 2. Westinghouse Electric Corp.

QUESTIONS

1. Why is a frequency-control system necessary with a direct f-m generator?

2. How is a reliable carrier amplitude obtained for frequency control? Explain. 3. Why is the use of a standard f-m discriminator without the use of a crystal- .

oscillator converter unreliable for frequency control?

4. How does the phase discriminator operate?

5. Why is a crystal-controlled oscillator necessary with a phase discriminator?

6. Explain why keeping the phase relationship between two waves constant keeps their frequencies in synchronism.

7. What are the advantages of motor control over bias control of frequency in an 1-m generator?

8. Describe how the two-phase voltage relationship is obtained for motor control.9. Explain the operation of the balanced modulator.

10. Why do not changes in the intensities of the two signals being compared by the motor control affect the control position?

11. What is the basic principle of operation of the integrated-pulse control method?

12. Draw the complete circuit diagram of the integrated-pulse system of frequency control by interconnecting the various circuits described in this chapter.

13. Describe the operation of the pulse-separator circuit in the integrated-pulse control method.

14. How does the pulse counter in the integrated-pulse control method operate?

15. Explain the operation of a multivibrator as a frequency divider.

16. Analyze the function of the blocking oscillator as a frequency divider.

17. How is a ring modulator used as a frequency divider?

18. Describe the circuit employing a pulse counter as a frequency divider.

CHAPTER 6 Direct Frequency-Modulation Transmitters

6-1. Basic Units. A direct f-m transmitter generates the f-m signal by using, as a fundamental source, a variable-frequency oscillator. It may be any one of the various types discussed in Chap. 4, "Direct Frequency Modulation." Because this type of f-m generator is not inherently stable enough to meet the $\pm 2,000$ -cps stability requirement of the FCC, a method of frequency control, as discussed in Chap. 5, has to be employed. Many combinations of different f-m generators and different control methods have been successfully used in the construction of commercial transmitters.



Fig. 6-1. A block diagram of a direct f-m transmitter showing the basic-unit divisions.

In Fig. 6-1 is drawn a block diagram of a direct f-m transmitter showing the basic unit divisions. The audio input is first amplified in the audio-amplifier stages. Since modulation takes place at lowpower level (at the oscillator), merely one or, at the most, two stages of low-level amplification are usually employed. It is assumed that the signal has already been monitored and is ready for transmission. In the case of mobile and commercial transmitters, where fidelity is not too important, no monitoring is employed and the amplifiers are usually quite narrow-band. In the case of high-fidelity

WRH

transmission, the audio amplifiers have to amplify frequencies up to 15 kc.

Following the audio amplifier, and controlled by it, is the direct f-m generator. It usually consists of a modulator and variable-frequency oscillator whose frequency deviation is controlled by the modulator. The maximum frequency deviation may be anywhere from ± 15 kc or less for mobile transmitters to ± 75 kc for high-fidelity f-m broadcast transmitters. In the case of broadcast transmitters, a pre-emphasis unit is incorporated into the input of the f-m generator to modify properly the characteristics of the audio signal. The carrier frequency of the f-m generator is usually in the neighborhood of 5 megacycles; the f-m signal has to be passed through a series of frequency multipliers before it can be used to excite the power amplifiers. These frequency multipliers, or "harmonic generators," as they are sometimes called, are ordinary class C type multipliers used to obtain the proper carrier-frequency signal for the power amplifiers.

Part of the output of the f-m generator is also fed into the frequency-control unit. This is the portion of the signal that is compared with the output of a stable crystal oscillator. In some cases, like the integrated-pulse control method, the f-m signal for the control unit may be at a high frequency. For other methods, the frequency comparison takes place at a low frequency, and the control unit actually has incorporated into it a number of divider stages for both the f-m signal and the crystal-generated signal. Upon comparison of the two signals, a control voltage is obtained which is used to regulate the center frequency of the f-m generator.

These five units, enclosed in the dotted box in Fig. 6-1, are usually built into one manufactured unit know as the "modulator" or the "exciter unit." It produces a low-power f-m signal at the proper frequency and with the desired regulation and deviation. One of the outstanding advantages of an f-m transmitter is that class C amplifier stages can be used in all the power-amplifier stages; the amplitude of the signal is always constant. This constancy of amplitude also allows the amplifier to be used at maximum efficiency at all times. Inasmuch as the class C amplifier is one of the most efficient types, the power-amplifier stages of an f-m transmitter are thus always very efficient. Another point to consider is that the same exciter unit can be used for a whole series of different transmitters, each with a different power output obtained by using different power amplifiers.

In Fig. 6-2 is shown, in block-diagram form, the composition of various transmitters with different power outputs but still using the same exciter unit. Let us assume that the exciter itself has a power output of 50 w. The exciter, without any additional power amplifiers, could be used as a low-power 50-w transmitter. For a 250-w transmitter, a power-amplifier stage is added, as illustrated. This is a straightforward class C operated stage. The 50-w exciter unit is

used as a driver unit. For a 1-kw transmitter, if 50 w of driving power is sufficient, the 250-w power-amplifier stage can be replaced



Fig. 6-2. A series of transmitters with different power outputs created by simply adding class C amplifiers.

by a 1-kw power amplifier. For a 10-kw power output, a 10-kw power amplifier is used, following the 1-kw amplifier which acts as the driver. This assumes that 1 kw of driving power is necessary for the 10-kw amplifier. The driver stage need supply only the amount of power, usually with just a little extra for good measure, that is necessary to produce full power output from the final power-amplifier stage. Hence, any desired power output can be produced from the same exciter by using a sufficient number of class C amplifiers, until the final power-amplifier stage, which will deliver the output power, has sufficient driving power delivered to it.

The output power of an f-m transmitter, operating at a permanent installation, may be increased very simply. The installed transmitter may be used as a driver unit to drive a new higher-power class C amplifier. For instance, a 1-kw f-m transmitter may be converted to 10-kw power output merely by adding a 10-kw power amplifier and using the 1-kw transmitter to drive the power amplifier. This additive method of increasing power permits low-cost expansion of the service area over which the transmitter is to be received.

6-2. Review - Class C Amplifiers. A class C amplifier 1 may be defined as an amplifier operating under such conditions that the plate

current flows in pulses of less than a half-cycle duration. It employs tuned circuits for its plate-load impedance and has a relatively high plate efficiency. In this type of amplifier the plate current flows in pulses. The output power is constant in amplitude (the complete cycle being accomplished by the flywheel effect of the tank circuit) and is at the same frequency as the grid-input voltage. Although the class C amplifier is quite efficient, it cannot be used for the amplification of a-m waves but may be used, and is used, for the amplification of f-m waves.



Fig. 6-3. A typical class C amplifier using a triode tube. A Rice neutralization circuit is shown by dotted line.

Figure 6-3 shows a typical class C triode amplifier circuit. This type of circuit, employing a grounded cathode, would oscillate of its own accord when the grid and plate circuits are tuned to the same frequency. This may be eliminated by some method of neutralization, by which a voltage fed back to the grid circuit cancels the regenerative voltage fed through the tube itself from the plate to grid. In the diagram is shown by dotted line the Rice system of neutralization to prevent self-oscillation.

The current and voltage relationships existing in a class C amplifier are shown in Fig. 6-4. The plate voltage consists of the a-c tank voltage superimposed on the B+ voltage. The total voltage never goes to zero but comes down to quite a small minimum \underline{E}_{min} . At the same time, the input voltage is causing the grid voltage to vary in a sinusoidal manner around the grid bias. The grid voltage is usually sufficient to drive the grid to positive for a small amount of the cycle. The grid voltage reaches its maximum at the same instant that the plate voltage reaches its minimum. The plate current that flows is a combination of the action of both the plate and grid voltages. The grid is biased well beyond cutoff and only when the grid voltage tends to swing the grid towards zero does the plate current begin to flow. Hence, the plate current only flows when most of the B+ voltage is used as a drop across the tank circuit; only a small portion is wasted as the voltage drop at the plate of the tube, thus decreasing the plate

dissipation and increasing the efficiency of the amplifier. These pulses of plate current act on the tank circuit like a push on a swing



Fig. 6-4. The current and voltage relationships in a class C amplifier showing the platecurrent pulse at minimum plate voltage.

and maintain the sinusoidal output voltage.

Power is absorbed from the input because the grid actually draws current during the interval that it swings positive. Part of the absorbed power is dissipated in the grid electrode of the tube and the rest absorbed by the grid-bias source. Care must be taken when adjusting the input voltage that the grid dissipation is not exceeded; it might permanently injure the tube.

6-3. Multipliers (Harmonic Generators). The ordinary multiplier circuit is a class C amplifier in which the plate-tank circuit is tuned to a harmonic of the input frequency. Because the plate and grid circuits are no longer tuned to the same frequency, neutralization is not necessary; otherwise the circuits are very similiar. The voltage across the tank circuit in the plate lead now generates 2 or more cps for each pulse of plate current which is drawn. It is like pushing the swing every second cycle (for a doubler) instead of at the end of each cycle. The flywheel effect of the tank circuit carries it through the correct number of cycles with a negligible loss in amplitude.

The plate-current pulse, in this case, has to be shorter in duration than in the case of the straight amplifier. This is necessary because the frequency of the voltage across the plate tank circuit is a multiple of the frequency of the grid-exciting voltage; hence, the plate voltage is in the neighborhood of the minimum for only a small portion of the grid-voltage cycle. Its actual duration is a compromise between
various factors. If it is very short, the driving power and driving voltage have to be large but the plate efficiency is very high. If the pulse is too short or too long, however, the output power drops off from the maximum possible under optimum conditions. A short pulse is obtained by using a large driving voltage and a high value of grid bias. Thus, only a small portion of the positive peak of the grid voltage is used to draw plate current. It is for this reason that the amount of exciting power necessary for a harmonic generator is greater than for a corresponding straight class C amplifier. The power necessary also increases rapidly with the order of the harmonic.

The amount of power output available from a harmonic generator is less than would be obtained from the same tube used as an ordinary class C amplifier. For a multiplication factor of 2, the output decreases to about 65 per cent of that of the class C amplifier; for 3, to about 40 per cent; for 4, to about 30 per cent; and so on, decreasing with each harmonic.

6-4. The Grounded-Grid Amplifier. The grounded-grid amplifier² is a modified form of the ordinary class C amplifier. In a grounded-grid amplifier the grid is at a-c ground potential and the driving voltage is applied between cathode and ground. The d-c and a-c potentials and currents used are the same as in the conventional circuits so far as the tube itself is concerned; hence, the ratings that apply to one will apply equally to the other. It will be found, however, that the input driving impedance is quite low in the groundedgrid circuit.

The advantages claimed for this type of circuit are

1. The circuits are simpler and require fewer components than the conventional circuits.

2. Neutralizing, when necessary, is very simple and is not required at all for low powers.

3. The circuits are stable and do not need critical adjustments, even at 100 megacycles.

4. The output is greater than the output from any other amplifier circuit using a tube of the same size and rating.

5. It is possible to use the same tubes in drivers and power amplifiers, reducing the number of different types required in a transmitter.

In Fig. 6-5 is shown the simplified circuit diagram of a groundedgrid amplifier. No neutralizing circuit is shown here; neutralization will be discussed later. The grid is actually bypassed to ground; it is at ground r-f potential. <u>Rg</u> is a grid resistor through which d-c grid current flows to produce an added grid-leak bias. <u>E_c</u> is the gridbias voltage and <u>E_B</u>, the B+ or plate voltage. The input voltage <u>E_g</u> is applied between cathode and ground (grid potential) so that it appears across grid to ground as in the conventional circuit. Thus, the input voltage is actually impressed on the grid of the tube and can be dealt with like the grid voltage in the case of the conventional circuit. The grid acts as a shield between the input and output circuits and, in



Fig. 6-5. A circuit of a grounded-grid amplifier showing \underline{E}_{L} , the a-c voltage across the load; \underline{E}_{p} , the a-c voltage across plate to cathode; and \underline{E}_{g} , the a-c input voltage.

tubes made for use in the grounded-grid circuit, the shielding is very complete. The output, however, is taken across plate to grid (ground) instead of plate to cathode, as in the conventional amplifier. The a-c output voltage $\underline{E}_{\underline{L}}$ is in series with the input voltage $\underline{E}_{\underline{g}}$; their sum is the voltage from cathode to plate in the tube.



Fig. 6-6. The load and grid voltage swings for the grounded-grid amplifier. <u>P</u> is the plate d-c voltage; <u>K</u>, the cathode d-c voltage; and <u>G</u>, the grid d-c voltage.

Figure 6-6 shows oscillograms of the load and grid voltages during one cycle of frequency. In accordance with the diagram given in Fig. 6-5, the cathode d-c potential, the line <u>K</u>, is zero while the grid potential, the line <u>G</u>, is at $-\underline{E}_{C}$ and the plate potential, the line <u>P</u>, is at <u>EB</u>. The grid remains constant at zero voltage while the cathode potential varies. When the cathode swings negative, it produces the same effect as the grid voltage swinging positive in a conventional circuit. Hence, the cathode voltage \underline{E}_g is in phase with the plate voltage \underline{E}_p . The limit of the downward swing in the a-c plate voltage (determined by the condition that the plate voltage should never be less than the grid voltage at any instant) is now $-\underline{E}_c$, instead of the peak positive-grid swing as in the case of the conventional amplifier. It is not affected by the input voltage \underline{E}_g , because the input voltage causes both the grid and plate potentials to vary.

The output voltage \underline{E}_{L} is larger than the output voltage obtained in the ordinary class $C \overline{a}$ mplifier. EL, the output voltage between ground and plate, is actually the sum of two voltages: the voltage E_g between ground and cathode and the voltage \underline{E}_p between cathode and plate. This is the normal output voltage of the conventional class C amplifier. Inasmuch as they are both in phase, the output voltage EL is the direct sum of \underline{E}_g , the input voltage, and \underline{E}_p , the tube output voltage. Assuming that we draw the same a-c plate current Ip from the tube in both cases, we find that when using a tube with the same ratings first in a conventional circuit and then in the grounded-grid circuit, the power output can be increased by a factor of one half $\underline{E}_{gI_{p}}$ where both are peak values. This increase in power does not come from the tube itself but comes from the driver. The driving power must be increased by that amount, sometimes a difficult problem. However, it will be noted that even though the power required from the driver increases, the amount of power fed into the grid of the tube remains the same and the extra power appears as useful power in the output circuit.



Fig. 6-7. The equivalent circuit of a groundedgrid amplifier. We see how both $\underline{E}\underline{g}$ and $\underline{E}\underline{p}$ contribute to the output voltage.

In Fig. 6-7 appears the equivalent circuit of a grounded-grid amplifier. In this equivalent circuit, \underline{E}_g appears between cathode and ground; $\underline{E}_{\underline{L}}$ is clearly seen to be made up of the sum of \underline{E}_g and \underline{E}_p . This agrees with the foregoing discussion. Because of the higher voltage available across the load, the load impedance $\underline{R}_{\underline{L}}$ is larger than would be encountered in a conventional circuit.

Neutralization, when required, may be accomplished by inserting

an inductance in series with the grid. In ordinary class C amplifiers, self-oscillations are caused by feedback from the output to the input through the plate-to-grid capacitance. This feedback is balanced out by providing a similar path for a signal of opposite polarity from the plate-to-grid circuit. In the grounded-grid circuit, two paths for a feedback signal already exist: the feedback from plate to cathode through the plate-to-cathode capacitance, and the feedback through the impedance of the grid lead. The impedance of the grid lead, being common to both the cathode and plate circuits, offers a feedback path by providing coupling between the output and input circuits. When the signals fed back through these two paths cancel one another in the input circuit, the tube is neutralized. For values of plate-to-cathode capacitance normally encountered in tubes adaptable to grounded-grid use, the inductance in the grid-to-ground circuit is sufficient; in some cases even series capacitance may be required to reduce the tube's value to the neutralization point. At frequencies where the physical size of the tube becomes appreciable with respect to a quarter wavelength, it may become necessary to use a tunable-bias blocking capacity to reduce the tube-element inductance to a value low enough for neutralization.

6-5. Grounded-Plate Amplifier. Very often a low-capacity tank circuit is desirable in the output circuit to obtain adequate gain with a sufficiently broad tank circuit. The large stray capacitance across the output circuit may be reduced by taking advantage of the low inherent filament-to-ground capacity of a tube. This is accomplished by



Fig. 6-8. A simplified schematic diagram of a groundedplate amplifier.

using a grounded-plate circuit as shown in Fig. 6-8 in simplified form. The plate is grounded through a bypass capacitor and the output circuit is placed between ground and cathode. It is equivalent to moving the r-f ground in an ordinary class C amplifier from its normal position at the cathode to the plate of the tube. The driving voltage is still impressed from cathode to grid, and the output is taken off across the output tank circuit. Now the stray plate-to-ground capacitance is in parallel with the grounding, or bypass, capacitor and has no effect on the operation of the circuit. We note, however, that the plate-to-cathode capacitance is still in parallel with the output tank circuit as in the straightforward amplifier; this grounded-plate circuit substitutes the stray-cathode-to-external-ground capacitance for the stray-plate-to-external-ground capacitance across the output tank circuit, taking advantage of the lower value obtained from cathode to ground.

In this circuit the cathode of the tube is at a high r-f potential, and the driving voltage to the grid has to be supplied above this high potential. In a filament-type tube the filament voltage has to be supplied at this high r-f potential without shunting the tank circuit with a large capacitance.



Fig. 6-9. A grounded-plate circuit utilizing a transmission-line tank circuit and a coaxial line-driver input.

Let us consider first the modified circuit shown in Fig. 6-9. The circuit is inherently the same as the circuit shown in Fig. 6-8 except that a coaxial-line tuned circuit is substituted for the tank circuit. A coaxial-line segment shorted at one end and fed at the other resonates just like an ordinary coil- and capacitor-tuned circuit. The length of the coaxial-line segment is approximately a quarter wavelength at the resonant frequency. Thus, the shorted coaxial line is substituted for the ordinary tank circuit shown in Fig. 6-8, the cathode being connected to the inner conductor and the outer conductor being grounded. The grid-driving voltage is provided by a coaxial feed line coming up through the center of the inner conductor of the cathode tank circuit. Since a coaxial line provides a voltage between the inner and outer conductors at its termination point, the inner conductor of the feed line is connected to the grid and the outer conductor is automatically connected to the cathode. Thus, the drive voltage appears between grid and cathode as depicted in Fig. 6-8. The output power is still taken off across the tank circuit between cathode and ground.

The feedback capacitance in this type of amplifier which will cause self-oscillations is the grid-to-plate (or grid-to-ground, since the plate is bypassed to ground) capacitance. An inductive reactance, usually a shorted transmission line, which is less than a quarter wavelength long at the frequency being used, is employed and is connected between the grid and ground to tune out the feedback capacitance. It is made adjustable by means of a movable short, so that the neutralization adjustment may be made after installation.



Fig. 6-10. A grounded-plate circuit as used in practice showing the neutralization inductance, the load-matching inductance, and the filament-input leads.

Figure 6-10 depicts the circuit of a grounded-plate amplifier as used in practice. The tube used is a filament-type triode. The output tank circuit is again a shorted coaxial transmission line with an adjustable short for tuning purposes. The grid-driving voltage is brought in through the center of the tank-circuit inner conductor, as discussed in reference to the previous circuit. There is now added, between grid and ground, a short length of shorted transmission line which acts as the neutralizing inductance — neutralizing the gridto-ground capacitance. This inductance is also adjustable by means of an adjustable short. The filament leads are brought up to the filament through additional holes in the center conductor of the tank circuit, while the r-f currents are bypassed to the conductor proper through a pair of bypass capacitors. Thus, the filament is tied, for r-f purposes, to the inner conductor of the tank-circuit line. The output is still taken off between filament (cathode in the previous case) and ground. Also shown in Fig. 6-10 is a load-matching inductance which matches the output load to the circuit. The outer-conductor current of the r-f output line, instead of being able to flow directly to ground, has to flow through the coaxial-line segment, called the ''load-matching inductance,'' and then to ground. Hence, the load-matching inductance is in series with the r-f output line and can be used to tune it to a purely resistive load.

The problem of a ground at about 100 megacycles is difficult to grasp because a ground implies a wire or conducting sheet which throughout its length is at the same potential—equal amplitude and equal phase. But at 100 megacycles the time necessary for the effect, or wave, to travel a few inches will introduce an appreciable phase shift. The nearest approach to the ground concept in the u-h-f region is an all-enclosing shield which keeps extraneous currents out and keeps the circuit currents in. Thus, the square of lines cornered by the four ground symbols can be an enclosing shield around the tube with the three coaxial lines of the circuit penetrating through it. It should be as small and complete as possible.

Figure 6-11 shows the type 5541 WE tube which may be used in a grounded-plate amplifier to produce 10 kw of useful power. It has a single thoriated-tungsten filament which has a nominal rating of 55 amp at 7 to 5v. The tube has an amplification factor of 26, a maximum plate voltage of 8,000 v, and a plate dissipation of 10 kw. It is capable of operating at full ratings up to 110 megacycles. The necessary driving power to obtain 10 kw output is somewhat less that 500 w. It is air-cooled, employing fins for contact with the air stream. The anode itself is short and stubby, a characteristic of high-frequency power tubes. If desired, the tube can also be used in the conventional grounded-cathode or in the grounded-grid types of circuit.

6-6. Reactance-Tube Motor-Control Transmitter.³ One possible way of constructing an f-m transmitter employing direct f-m frequency generation is to combine the reactance-tube modulator with motor frequency control. In Fig. 6-12 appears a block diagram of an exciter unit employing those devices. The oscillator operates at a center frequency between 4.5 and 6 megacycles. This frequency is noted as f in the diagram. In the output circuit, two triplers and a buffer amplifier are used, giving an output frequency of 9f. Hence, the output carrier frequency will be between the frequency limits of 40.5 and 54 megacycles, one half the desired output frequency for the f-m broadcast band. It is necessary, for this reason, to incorporate a doubler in the final power-amplifier stages.

The output from the modulated oscillator, following the theory of motor control, is divided by a factor of 240 in a series of divider circuits and fed into the balanced modulators of the control system. Thus, the carrier frequency of the f-m signal at the input to the



Fig. 6-11. The type 5541 WE tube which may be used in a final power amplifier of the grounded-plate type to deliver 10 kw of f-m power.

balanced modulators will be between 18.75 and 25 kc and will have a ratio of frequency deviation to modulating frequency of the proper diminutive value. The crystal oscillator, for simple stable operation, is designed for a frequency range between 94 and 125 kc. Dividing the crystal frequency by 5 should yield the same frequency as <u>f</u> divided by 240. Thus, for a carrier frequency of 6 megacycles at the f-m generator, an oscillator frequency of 125 kc would be used, inasmuch as they both would result in the same frequency at the balanced modulator. Any difference between the two frequencies being compared by the control unit will cause the control motor to turn and vary the center frequency of the modulated oscillator, by means of a variable capacitor in the tank circuit, until the difference is zero. A circuit diagram of the complete exciter unit is shown in Fig. 6-13. Two 6V6 tubes in a push-pull arrangement are used as modu-



Fig. 6-12. A block diagram of an exciter unit employing a reactance-tube modulator and a two-phase motor control for frequency stabilization.

lator tubes. The audio input is taken in through a tapped audio transformer T. The plates of the tubes are connected in parallel across the tank circuit L and C. The 90-deg phase-shifted voltage is fed back from the tank circuit to the grids of the two modulator tubes through the link L¹L¹ which uses transformer coupling to introduce the phase shift. By coupling each of the two modulator tubes with different polarities, one operates as a capacitive-reactance tube and the other as an inductive-reactance tube, as should be in a pushpull arrangement. The oscillator itself is another 6V6 tube which has paralleled across its tank circuit, from plate to ground, the variable capacitor C_{v} , which is the frequency-controlling capacitor operated by the two-phase motor. The output of the modulated oscillator is then fed into a series circuit of two 6V6 buffer multipliers and a 2E26 buffer amplifier. If it is so desired, the 2E26 tube may be used as a multiplier also. After the 2E26, the signal is taken out and fed into the power-amplifier units.

Through the connecting line <u>A</u> a portion of the modulated-oscillator signal is fed into a series of four frequency dividers. Each of these frequency dividers employs a 6AC7 tube connected as a triode with a tuned circuit in the plate lead. A simplified circuit of the divider is shown in Fig. 6-14. It is a locked-in type of frequency divider where the frequency <u>f</u> to be divided is fed into the grid circuit and the output frequency <u>f</u>/<u>n</u> is taken off across a tuned circuit tuned to that frequency. The output is also fed back to the input through a coupling coil. The divider is actually an oscillator which is locked into a subharmonic of the input voltage. This divider circuit permits division ratios of as much as 5 to 1 with stable operation. The wave shape of the output, while not perfect, is sufficiently good for its use in the control circuit. The dividers cover



Fig. 6-13. A circuit diagram of a reactance-tube modulated motor-controlled exciter unit.

the required frequency range by means of an adjustable iron slug in the tuned circuit. The lock-in range claimed for this type of divider is as high as ± 5 per cent. A similar type of divider is used to divide the oscillator frequency by five.



Fig. 6-14. A locked-in oscillator frequency divider using a 6AC7 as a triode oscillator.

Four 1614 tubes, called the "motor tubes," are used in a pair of balanced modulators to mix the divided modulated-oscillator signal and the divided crystal-oscillator signal. Any difference in frequency causes the motor to change position and adjust the center frequency of the modulated oscillator. The second crystal circuit shown in the diagram is a spare; only one crystal is used at a time. The crystal circuits are so designed that they may operate at any frequency between 94 and 125 kc without any tuning adjustments by merely changing the crystal.

To allow for rapid checking of the circuits, a test equipment unit is built into the exciter. A 2AP1 cathode-ray oscilloscope and selector switch is provided to permit checking of the operation of each divider and multiplier by means of lissajous figures. A threeposition selector switch enables the operator to apply a d-c potential to either reactance-tube grid, causing the frequency of the modulated oscillator to shift either high or low over a considerable range. The behavior of the frequency-control motor and the relative rotation of the shaft required to correct the artificial frequency shift introduced by the switch may be observed on a dial on the motor shaft. This operation gives a rapid check of the performance of the reactance tubes and the frequency-control mechanism. A meter is also provided to read the plate current of the reactance tubes and the modulated oscillator. A buzzer, operated by a cam on the frequency-control capacitor shaft, provides warning if, for any reason, the frequency control is about to fail because of passing through a minimum or maximum of the oscillator frequency-adjustment capacitor. For instance, if for some reason the frequency of the modulated oscillator keeps climbing so that the centrol motor keeps increasing the value of the frequency-adjustment capacitor, when the capacitor reaches its maximum value it can no longer increase and the control system will lose control over the

FREQUENCY MODULATION

frequency; at this point the buzzer will sound, warning the operator that something is wrong.



Fig. 6-15. Front and rear views of the two-phase motor with the frequency-determining capacitor mounted directly on the shaft.

DIRECT FREQUENCY-MODULATION TRANSMITTERS 109

Front and rear views of the two-phase motor used for the frequency-control system are shown in Fig. 6-15. It was found advisable to mount the tuning capacitor on an insulator at one end of the motor shaft. The fixed plates of the capacitor are split in half, one half grounded and the other half connected to the master oscillator circuit. By using this type of construction, all backlash



Fig. 6-16. A front view of the reactance-tube motor-controlled exciter unit showing the oscilloscope and meter test panel.

and lost motion, as well as friction other than that of the motor bearings themselves, are eliminated. The full range of control requires the motor armature to turn only ± 45 deg. The rate of frequency correction is limited by one requirement: It must not cause fre-



Fig. 6-17. The circuit of a 250- to 1,000-w f-m transmitter using a 7C24 grounded-grid power amplifier.

quency demodulation at the lowest modulation frequency. The motor itself is an induction type with viscous damping to prevent overshooting. The damping unit also can be seen in Fig. 6-15; it is enclosed in the lucite case on the front end of the motor.

A front view of the exciter unit is shown in Fig. 6-16. A vertical type of construction is used, with the tubes and main components mounted on the front of the panel. The wiring for the components is in the rear and is available through a rear door. One advantage of this vertical type of mounting is that the air enters through filters on the bottom of the cabinet and travels upward past all of the components in an unobstructed manner. In this it is in contrast with the shelf type of mounting where care must be taken that the shelves do not obstruct the free flow of air. On the left center of the chassis, in Fig. 6-16, can be seen the oscilloscope for visible checking of the wave shapes; in the center is the selector switch, while on the right is a meter for magnitude checks. The chassis in the lower half of the picture is the power supply for the exciter unit.

It is now possible to combine the exciter unit with a series of amplifiers, including a doubler to attain the correct frequency, and thus obtain a complete transmitter. A simplified circuit diagram of such a transmitter is shown in Fig. 6-17. The output of the exciter unit is fed into a 4-125A doubler to obtain frequencies in the 100-megacycle f-m broadcast band. This is a medium-power thoriated-tungsten-filament v-h-f triode which can deliver about 375 w at 120 megacycles when driven to full output. The next stage consists of two 4-125A tubes in a conventional grounded-cathode parallel circuit. If the transmitter output were taken from these tubes, a range in power from 50 to 250 w would be obtained. For tuning purposes and for monitoring, meters are provided in the filament returns of the 4-125A tubes, so that the total cathode current for each stage may be read. A meter is also provided in the grid-return lead of the push-pull stage to indicate the total grid current drawn by the two 4-125A tubes. The last stage is a grounded-grid power amplifier employing a 7C24. Transmission-line tuning is employed in both the cathode and plate circuits. Although it cannot be seen in the diagram. the filament power leads are brought in through the cathode line to keep the radio frequency from being bypassed to ground. Meters are provided for both grid and plate currents.

The 7C24 tube, shown in Fig. 6-18 with a 6AC7 metal tube for comparison, was designed for use as a stable amplifier tube at 100 megacycles in a grounded-grid circuit. It resembles in size and appearance the 827-R but differs inasmuch as it is a triode, whereas the former is a tetrode. Moreover, the construction is quite different. The 7C24 is provided with a grid structure specifically designed to offer a maximum of shielding between the plate and filament electrodes, resulting in a very low plate-to-filament capacity. The grid connection is a disk seal brought out through the glass all the way around the tube. When this is utilized in connection with an external shield, the input and output circuits of the amplifier can be well isolated.



Fig. 6-18. The 7C24 tube especially designed for 100-megacycle transmitters shown with a 6AC7 metal tube for size comparison.

In the output circuit of the transmitter is a device known as a "harmonic attenuator." It is a band-pass filter which reduces the harmonic content of the transmitter output by blocking the harmonics, thereby keeping them out of the antenna. This improves the wave shape of the radiated signal and prevents interference with other stations having frequencies in harmonic relationship with the transmitter frequency. Finally, there is delivered to the antenna, through a coaxial transmission line, 1 kw of f-m power.

Thus, we have a complete f-m transmitter. It is possible to mount all of the components in two standard transmitter cabinets: in one cabinet would be mounted the high-voltage rectifier and exciter unit while in the other would be mounted the r-f amplifiers. In the commercial transmitter, there is still additional space in the r-f amplifier cabinet to permit the addition of more amplifiers in case a power greater than 1 kw is desired at any future time. This transmitter will operate on any specified frequency between 88 and 108 megacycles. It has an output impedance of 35 to 75 ohms and the stability is better than the $\pm 2,000$ cps required by the FCC. It is capable of being modulated with a frequency deviation of up to ± 100 kc.

To obtain a higher power output, it is necessary only to continue adding power-amplifier stages. For instance, to increase the power output of the transmitter to as high as 3 kw, only one power amplifier need be added. The power amplifier actually used is another 7C24 tube in a grounded-grid circuit. The 1,000-w transmitter



Fig. 6-19. The block diagram of an f-m exciter unit employing an input-capacitance modulator and a phase-discriminator center-frequency control.

acts as a driver stage for this higher-power transmitter but also contributes, by virtue of the grounded-grid circuit, to the output power. To increase the power still further, it is necessary only to add more stages of amplification. All of the stages are class C stages and, even though they are amplifying a modulated wave, they are always operating at maximum efficiency. This, as pointed out before, is one of the advantages of using frequency modulation.

6-7. Input-Capacitance Phase-Discriminator Transmitter. Another combination of units used in the construction of an f-m transmitter is the input-capacitance (also called the "Miller-effect") modulator combined with the phase-discriminator method of frequency control. A block diagram of the exciter unit is shown in Fig. 6-19. An audio input with an input impedance varying between 150 and 600 ohms is employed with the pre-emphasis circuit following the input transformer. A modulated oscillator, operating between the frequencies of 3.66 and 4.5 megacycles, is included with a modulator tube and a buffer amplifier in the dotted rectangle called the "modulator unit." The output of the buffer, called the "r-f output," is fed into the necessary multipliers to obtain frequencies in the f-m broadcast band. Some of the output of the buffer is also shunted off through a series of frequency dividers and buffers to be divided by a factor of 256. The carrier frequency of the f-m signal, after division, is between 14.3 and 17.6 kc. Inasmuch as the frequency deviation in the frequency range of 3.66 to 4.5 megacycles has to be about +3.1 kc to obtain a 75-kc deviation in the f-m broadcast band, the deviation at 14.3 to 17.6 kc would be about 12 cps. With that small a deviation, the phase discriminator, in combination with a low-pass filter to remove harmonics, will operate very satisfactorily to control the frequency.

The crystal operates between 114.4 and 140.8 kc in a simple stable oscillator circuit. The output frequency is divided by eight to obtain a frequency that is the same as the center frequency of the divided modulated-oscillator signal. A 12H6 double diode is used as the phase discriminator. Its output, after passing through a low-pass filter to prevent any demodulation of the audio signal, is fed into the Miller-effect tube to adjust the center frequency of the modulated oscillator.

A circuit diagram of the modulator unit is shown in Fig. 6-20. This is the unit enclosed in dotted lines in Fig. 6-19. The 12J5 tube is used as an ordinary Hartley oscillator, the tank circuit being tuned by means of a variable powdered-iron core instead of the conventional variable capacitor. The resistor in the grid circuit is used for stability. The 6AB7 pentode tube is used as the input-capacitance (or Miller-effect) modulator. Modulation is effected by reflection of the effective capacitive variation in the grid circuit of this tube, which is connected across the oscillator-tuned circuit. The capacitive change is proportional to the total actuating voltage on the grid of the tube. This voltage is the sum of the program a-f voltage input plus the regulation voltage obtained from the phase discriminator.



Fig. 6-20. The circuit of an input-capacitance modulator employing a 12J5 and two 6AB7's.



116



Fig. 6-21. The circuit of a center-frequency control unit using a phase discriminator.

Thus, the modulator not only varies the frequency of the transmitted signal to produce frequency modulation, but also adjusts the center operating frequency. A capacitor has been shunted from plate to grid of the modulator tube. This tends to increase the input capacitance of the tube (see Sec. 4-6). To obtain a fixed-frequency swing for a given change in input capacitance as the oscillator capacitance is changed to cover the frequency band, an adjustable capacitance the r-f signal on the modulator grid, the capacitance of the modulator is coupled across only a portion of the tank circuit. The buffer amplifier, employing a 6AB7 tube, is a straightforward grounded-cathode amplifier, a portion of whose output is shunted into the divider chain for use in the phase discriminator.

Figure 6-21 is a schematic of the center frequency control unit showing the dividers, buffers, crystal oscillator, and discriminator. This circuit follows the block diagram shown in Fig. 6-19, excluding only the audio and modulator circuits. The 12SN7-GT tubes are used as multivibrator dividers, with 6AB7 and 12SJ7 tubes as buffers in between divider stages. Two crystal circuits are used, one being in use at all times and the other being held in readiness as a spare. The signal enters at the top left 12SN7-GT tube and proceeds through the various dividers and buffers until it reaches the lower left 12SN7-GT tube, the final buffer. The divided f-m signal is fed into one section of this tube, while the divided crystal-oscillator signal, which originated in the crystal oscillator at the lower right of the diagram and passed through two 12SN7-GT dividers separated by a 12SJ7 buffer, is fed into the other section of the tube. The output of this final buffer is the two divided signals which are now passed through individual low-pass filters and fed into a 12H6 phase discriminator. The output of the phase discriminator is the control voltage which is first passed through a low-pass filter and then added to the audiosignal voltage; the sum of the two then is fed into the modulator grid, accomplishing both frequency deviation and frequency control. The low-pass filter is designed to cut off at 10 cps, passing no frequencies above that and thereby avoiding demodulation.

Figure 6-22 shows an input-capacitance modulated oscillator mounted in a phase-discriminator control chassis. The modulator unit is built on a separate small chassis and is removable from the main control chassis. The picture illustrates how the modulator unit is fastened on the chassis.

Thus, we have available at the output of the modulated oscillator an f-m wave at a carrier frequency of between 3.66 and 4.5 megacycles. To transmit in the f-m broadcast band, the 100-megacycle band, the signal has to be multiplied by a factor of 24. The unit which accomplishes this is a seven-tube unit including a buffer and amplifier which is capable of delivering a power output of 250 w in the frequency range of 88 to 108 megacycles. The unit, called the "intermediate power amplifier and multiplier," can be used with no further power amplification when a transmitter of only 250 w power output is desired.



Fig. 6-22. An input-capacitance modulated oscillator mounted in a phase-discriminator center-frequency control chassis. The modulated oscillator is mounted on a separate chassis and is removable.

A simplified schematic diagram of the intermediate power amplifier and multiplier unit is shown in Fig. 6-23. It employs three 1614 single-ended doublers in the lower-power stages, followed by two 815's, the first of which is used as a push-pull tripler and the second as a push-pull buffer amplifier. Ordinary transformer coupling is used to convert from single-ended operation to push-pull at the first 815 tube. The final stage of the intermediate power-amplifier and multiplier unit consists of two 4-250A tubes in push-pull. An ordinary grounded-cathode circuit is used with single-turn hairpin coils tuned by movable shorts. The output of these two tubes is used as driver power for the following power-amplifier stages.

For both the 1-kw and the 3-kw transmitters, two 7C26 tubes in



Fig. 6-23. A simplified circuit of the intermediate power amplifier and multiplier unit capable of delivering 250 w of useful power.

a grounded-cathode push-pull circuit are employed as the power amplifier. For the 1-kw output the tubes operate with a 2,000-v plate voltage, while for the 3-kw output the plate voltage is raised to 3,000 v. However, in line with the RMA (Radio Manufacturers Association) recommendations, the primary power of the 1-kw transmitter is obtained from a single-phase 220-v source, while the primary power for the 3-kw transmitter is obtained from a three-phase 220-v source.

A block diagram for the 20- or 50-kw transmitters using the



Fig. 6-24. A block diagram of a 20-kw transmitter which employs a 3- and a 10-kw amplifier in the driver stages.

same exciter unit is shown in Fig. 6-24. The only items added are the power-amplifier stages necessary to produce the desired power output. The crystal oscillator is noted as a separate unit, but actually it is incorporated in the control unit, as discussed previously. Figure 6-25 depicts a simplified-circuit diagram of the power-amplifier stages for a 20-kw transmitter. The first pair of tubes, two 7C26's, are used in the conventional grounded-cathode push-pull circuit with

DIRECT FREQUENCY-MODULATION TRANSMITTERS 121

transmission-line tuning of the plate and grid circuits. These produce 1 kw of useful power which drives two 7C27 tubes in a push-



Fig. 6-25. A simplified circuit diagram of the power amplifier stages of a 20-kw transmitter.

pull, grounded-grid circuit. Thus, some of the output power from the 7C26's is passed on to the output of the 7C27's. At the output of the 7C27 tubes there is available 10 kw of f-m power. For a 10-kw transmitter, the power amplifier terminates in these 7C27 tubes. For a 20-kw output, the 7C27's drive four 3X2500A3 tubes in a pair of push-pull grounded-grid circuits. Again, by virtue of the grounded-grid circuit, some of the output power is obtained directly from the output of the 7C27 tubes. All eight tubes contribute to the power output, and thereby decrease the amount of power that has to be generated in the final amplifier. The antenna-coupling unit is a balanced converter used to couple the push-pull output stage to a coaxial transmission line.⁴

For a power output of 50 kw, two D-16 tubes are used in the final stage instead of the four 3X2500A3 tubes. Whereas in all the lower-power transmitters air cooling is used, water cooling is used in the 50-kw transmitter.

The transmitters are all built in vertical-rack panels in practice. A rear view of the 1-kw transmitter is shown in Fig. 6-26. It is com-



Fig. 6-26. The rear view of a 1-kw input-capacitance-modulated phase-discriminator-controlled transmitter.

plete in two cabinets. The right-hand cabinet contains the regulated power supplies in the lower chassis, the modulated oscillator and control unit in the center chassis, and the intermediate power amplifier and multiplier unit in the upper chassis. The left-hand cabinet contains the power-amplifier stages and output-conversion transformer, converting the push-pull output to single-ended power.

The f-m noise present in the transmitted wave from transmitters using the input-capacitance phase-discriminator exciter unit, measured under normal broadcasting conditions, is -65 db below full carrier conditions. The 10-kw transmitter requires 24.5 kva for full power output; the 20-kw transmitter, 45 kva; and the 50-kw transmitter, 100 kva. The 10-kw transmitter weighs approximately 5,000 lb; the 20-kw transmitter, approximately 10,000 lb; and the 50-kw transmitter, approximately 15,000 lb. They all operate well within the $\pm 2,000$ -cps carrier-frequency requirement and are capable of frequency deviations beyond ± 75 kc.

6-8. Integrated-Pulse-Control Transmitter. A transmitter using an integrated-pulse method of frequency control is shown schematically in Fig. 6-27. Enclosed in dotted lines are the three basic units: the frequency-stabilizer unit, the frequency-modulatedoscillator unit including the multiplier stages, and the power amplifier for a 3-kw output. To change the amount of power output, it is necessary, as with the other transmitters, only to change or add to the existing power amplifier.

The f-m oscillator and multiplier unit includes the input audio transformer and pre-emphasis network. The pre-emphasis network, consisting of a resistor and capacitor in parallel, may be switched out of the circuit, and a resistance pad with the same loss but with no frequency discrimination may be switched into the circuit. Thus, the transmitter may be operated with or without pre-emphasis. A 6SJ7 tube is used as an audio amplifier feeding into the 1614 modulator-control tube.

The modulator-control tube obtains its plate voltage through the modulator tube, two sections of a 6H6 tied in parallel. The plates of the 6H6 are coupled across the tank circuit of the 1614 oscillator by means of a small capacitor. Since the cathodes of the 6H6 (also the plate of the 1614 modulator-control tube) are bypassed to ground. the amount of current flowing through the capacitor is determined by the r-f impedance of the 6H6 tube, which varies as the tube conducts current over a smaller or larger portion of the r-f cycle. This capacitor constitutes a reactance across the tank circuit; as the amount of current flowing through the capacitor varies, it will vary the generated frequency of the oscillator. The modulator-control tube has fed into its grid both the audio and the control voltages. These voltages vary the impedance of the tube which, in turn, varies the voltage drop across the tube. Since the d-c voltage is fed to the modulatorcontrol tube through the 6H6, as the voltage across the modulatorcontrol tube changes, it will leave more or less of the d-c voltage available for the 6H6. Thus, the operating point and therefore the a-c impedance of the 6H6 is altered, causing it to modify the amount of current flowing through the small capacitor, thereby varying the frequency of the oscillator.

The oscillator employs a 1614 tube in an oscillator-tripler circuit. The oscillator circuit is of the electron-coupled type, with one side of the tank circuit grounded, the other side connected to the grid, and the center tap connected to the cathode. Instead of using the plate circuit for straight amplification, however, it is tuned to the third harmonic, thereby acting to triple the frequency of the



Fig. 6-27. The circuit diagram of a 3-kw transmitter using pulse control of frequency.

tank circuit. The f-m oscillator operates at one ninth the output frequency of the transmitter, making only one additional tripler necessary. The second tripler is an 829B tube with its two sections connected in push-pull. Its output is fed by direct coupling into another 829B in a push-pull amplifier circuit called the "intermediate amplifier." From here the signal is transmitted, by means of a coaxial line, to the power-amplifier section of the transmitter.

An interesting addition to the modulated-oscillator circuit is the 6H6 feedback discriminator. This circuit picks up a portion of the f-m signal from the oscillator tank circuit and detects it, thereby obtaining the audio-modulation signal. This detected signal is fed back in inverse relationship to the incoming signal at the input of the audio amplifier. Thus, there is an inverse feedback loop around the oscillator, audio amplifier, and modulator which improves the fidelity and reduces the noise.

The frequency control, or stabilizer, unit follows the circuit discussed in Sec. 5-5. An extra crystal oscillator is built into the unit as a spare. The output of the crystal oscillator and the output of a buffer amplifier, whose input is obtained from the f-m oscillator, are mixed in a pair of 6SA7's, the phase of the oscillator voltage being properly shifted by means of resistance-capacitance networks. The beat-frequency outputs are then passed through a pair of amplifiers, sections of a 6SN7, while another 6SN7 is used to generate the pulses. A 6H6 tube is then used as a pulse selector, selecting pulses for frequencies above and below the carrier frequency. To assure equal amplitudes for all the pulses, the pulses are passed through amplifiers and then through limiters which even out the amplitudes and ensure the proper operation of the counter circuit. Two 6H6's are used in the counter circuit and are followed by a 6SL7 cathode follower for stable amplification down to direct current. A threeposition high-low switch is used in the output of the cathode follower to vary the amount of direct current superimposed on the control voltage. In this manner the static deviation, the deviation before the control voltage is impressed, is varied. The control voltage and the static-deviation voltage are then impressed on the grid of the audio-amplifier tube in conjunction with the audio-input signal.

The power amplifier consists of two stages. The first, called the driver, is a grounded-cathode push-pull amplifier using two 4-125 tubes. The output is coupled to two WL473 tubes in a push-pull grounded-grid circuit utilizing transmission-line tuning. The 3=kw output power is delivered to a coaxial transmission line of between 40 and 80 ohms. It is monitored by an r-f meter using a 9006 diode detector. An additional line is provided to tap off power for modulation and frequency monitoring. A switch is provided in the platevoltage lead to delete the plate voltage when tuning the transmitter

voltage lead to delete the plate voltage when tuning the transmitter. For other power ranges the output stages are changed. In the 1-kw transmitter, only one 4-125 tube in a single-ended circuit is used as a driver, while a single WL473 in a single-ended groundedgrid circuit is used as the power amplifier. For 10 kw of output power, the 3-kw transmitter is used as a driver to drive two WL479 tubes in the final stage. The power input for the 1-kw transmitter is 3.6 kw; for the 3-kw transmitter, 8.5 kw; and for the 10-kw transmitter, 31 kw: all at a 90 per cent power factor. The transmitters have a frequency stability of $\pm 1,000$ cps, twice as good as required by the FCC.



Fig. 6-28. A rear view of the pulse-control transmitter with doors open and dust covers removed.

Figure 6-28 shows a rear view of the 3-kw transmitter with the doors open and the dust covers removed. The transmitter is completely contained in a cubicle with over-all dimensions of 66-in. width, 34-in. depth, and 74-in. height. It weighs about 1,900 lb. complete. External connections are brought out for the 208 to 240-v, three-phase input-power line; the 115-v a-c power input for the crystal

heaters and the convenience outlet; the program-input line; the r-f output line for connecting the modulation monitor and frequency meter; and the coaxial line to the radiating system. The modulated oscillator and the frequency-control units are of the plug-in type so that they can be easily removed from the cubicle if necessary. The other transmitter circuits are wired into the cubicle in such a manner that all portions are readily accessible.

REFERENCES

1. See Terman, F. E., <u>Radio Engineers' Handbook</u>, New York, McGraw-Hill Book Company, Inc., 1943, for an excellent discussion of class C amplifiers and harmonic generators.

2. C. J. Starner, "The Grounded Grid Amplifier," <u>RCA Broadcast News</u>, January, 1946.

3. N. J. Oman, "A New Exciter Unit," RCA Broadcast News, January, 1946.

4. N. Marchand, "Transmission Line Conversion Transformers," <u>Electronics</u>, December, 1944, p. 142.

QUESTIONS

1. Describe the basic units that comprise a direct f-m transmitter.

2. Why can the output power of an f-m transmitter be increased with a minimum of rebuilding?

3. What is a class C amplifier?

4. Why is a class C amplifier so efficient?

5. Describe the current and voltage relationships in a class C amplifier.

6. How does a multiplier circuit differ from a straightforward class C amplifier? 7. Explain why the amount of power obtained from a multiplier is usually less

than in a straight amplifier.

8. Describe the operation of a grounded-grid amplifier.

9. What are the advantages of the grounded-grid amplifier?

10. Explain how the grounded-grid amplifier utilizes some of the driving power in its own output.

11. How is neutralization accomplished in a grounded-grid amplifier? Explain its operation.

12. Describe the grounded-plate amplifier.

13. What are the advantages of the grounded-plate amplifier?

14. How is neutralization accomplished in the grounded-plate amplifier?

15. Describe, by means of a block diagram, a transmitter employing a resistancecapacitance oscillator and motor control of carrier frequency. The output desired is 1 kw.

16. Draw a circuit diagram of the exciter unit for the transmitter described in Question 15.

17. Why is it advantageous to have an oscillocope included as a part of the test equipment built into an f-m transmitter?

18. Design, by means of a block diagram, a reactance-tube phase-discriminatorcontrolled transmitter, with output of 3 kw.

19. Draw a schematic diagram of the transmitter described in Question 18.

20. Design, by means of a block diagram, an input-capacitance-modulated and integrated-pulse-controlled transmitter; output desired, 10 kw.

21. Draw a schematic of the transmitter described in Question 20.

CHAPTER 7

Phase-to-Frequency Modulation

7-1. Phase and Frequency Modulation Compared. So far we have discussed only the direct method of producing frequency modulation, but, historically, it was not the first to be practically applied to wideband high-fidelity f-m communication. Major E. H. Armstrong with his ingenious amplitude-to-phase-to-frequency modulation¹ was first to provide a system producing the f-m signal as used today. The system differs from the direct f-m generators (variable oscillators with some type of frequency-control circuit); in this case an f-m signal is generated directly from a primary crystal-controlled source. The generation of frequency modulation by employing phase modulation has also been termed an indirect method, since one type of modulation is employed to produce another.

Let us examine carefully the resemblance (mentioned previously in Chaps. 1 and 2) between phase modulation and frequency modulation. In Fig. 7-1 are shown two phase-modulated waves, one modulated by a voltage at the frequency \underline{F} , and the other modulated by a voltage at double the frequency 2F. By examining the modulated waves and not knowing whether they were phase- or frequency-modulated, we could not tell which was employed; the resultant waves for both types of modulation look alike. Let us investigate the reasons for this resemblance. The unmodulated wave is shown dotted in Fig. 7-1 and is used as the reference for the phase-modulated wave. As the modulating voltage increases, the phase-modulated r-f wave starts to lead in phase, moving ahead of the dotted reference wave. In so doing, the wave has to stretch - like a pair of lazy tongs. By this process of stretching, the wavelength is increased and the frequency proportionately decreased. Similarly, when the phase is decreased to a lagging value by the modulating voltage going negative, the wave contracts like the contraction of a pair of lazy tongs. The wavelength decreases with the contraction, and the frequency proportionately increases. Hence, in the phase-modulated wave, as the phase

129

is shifted back and forth the frequency is increased and decreased during each cycle of modulation. This seems to fit our visual con-



Fig. 7-1. Two phase-modulated waves: one phase modulated by a voltage at the frequency \underline{F} and the other by a voltage at the frequency $2\underline{F}$, the resultant frequency modulation obtained being doubled.

ception of frequency modulation.

By carefully studying the modulated waves and their modulating voltages, however, we will find that phase modulation differs from frequency modulation. Figure 7-1a is an oscillogram of the modulating voltage at a frequency of <u>F</u> that is phase-modulating the r-f voltage shown in Fig. 7-1b. For clarity a low-frequency r-f wave is used. The half wavelength of the unmodulated wave is noted as $\lambda_0/2$. At the positive and negative peaks of the modulating wave, the points <u>P</u> and Q, the voltage has reached its maximum values and, at those instants, is unchanging in value. Hence, referring now to Fig. 7-1b (the modulated wave), we find that the phase is also unchanging at those points; at those points the modulated wave has the same wavelength and the same frequency as the unmodulated wave, being displaced only in time. However, at the point <u>H</u> in the modulating-voltage the voltage is changing at its most rapid rate. To change the phase of the wave rapidly, the modulated wave must contract

a maximum amount, thereby causing the half wavelength to decrease to its lowest value $\lambda_{\underline{h}}/2$. Hence, the frequency of the phase-modulated wave reaches its maximum value, not when the modulating voltage is at its peak as in the case of frequency modulation, but rather when the modulating voltage is changing most rapidly. Similarly, at K where the modulating voltage is now changing in the opposite direction, the modulated wave reaches its minimum frequency. Thus, we have the major difference between the two: in a frequency-modulated wave the frequency is determined by the instantaneous value of the modulating voltage, whereas in a phase-modulated wave the frequency is determined by the <u>rate of change</u> of the instantaneous value of the modulating voltage.

An interesting point of variation between frequency modulation and phase modulation resulting from this dependence on rate of change is illustrated by doubling the frequency of modulation as demonstrated in Figs. 7-1c and 7-1d. The amplitude of the modulating voltage is unchanged. Because the frequency is twice as high, the phase now has to change twice as rapidly. Therefore, the minimum half wavelength obtained, $\lambda_h'/2$, is half of $\lambda_h/2$ (the minimum half wavelength obtained with a modulating frequency of <u>F</u>). Thus, in a frequencymodulated wave the frequency deviation is independent of the modulating frequency, whereas in a phase-modulated wave it is proportional to the frequency.

This latter point was also demonstrated in Chap. 2 where the sidebands for both types of modulation were studied. The phase-modulation factor \underline{m}_p was dependent only on the phase deviation, while the frequency modulation factor \underline{m}_f not only was dependant on the frequency deviation but was also inversely proportional to the modulating frequency.

7-2. Frequency Modulation from Phase Modulation. It is possible to produce a frequency-modulated wave, therefore, with a phase modulator which compensates for characteristic differences between the two types by correcting the wave shape of the modulating voltage, or, to state it more precisely, by properly altering the magnitudes and phases of the various frequency components of the modulating voltage. The audio input signal, in effect, is altered in two ways: First, considering only the amplitude, the audio voltages are reduced in amplitude in proportion to their frequency. (If components are involved, each component amplitude is reduced in proportion to its frequency.) Thus, in Fig. 7-1 the modulating voltage at a frequency of 2F would be reduced to one half its original amplitude and therefore would produce the same frequency deviation as the modulating voltage at a frequency of F. Thus, even though the phase shift is twice as fast, shifting the phase half the amount produces the same frequency deviation. Second, considering the phase and frequency, the signal is transformed into a modulating voltage whose rate of change is proportional, at every instant, to the amplitude of the audio signal at that same instant. Mathematically, the first requirement follows upon the second, but usually it is best to consider each separately. Incorporating these corrections into the modulating voltage will produce, at the output of the phase modulator, a signal which is frequency-modulated by the audio signal input.

At first glance it appears that a very complicated circuit will be needed to accomplish these corrections. Fortunately, that is not true,



Fig. 7-2. A simple audio-correction network consisting of a series resistor \underline{R} and a parallel capacitor \underline{C} .

and the simple resistance-capacitor network shown in Fig. 7-2 will accomplish both corrections. It consists, simply, of a resistor \underline{R} in series with the input and used in conjunction with a capacitor \underline{C} in parallel with the output. The resistance \underline{R} is made at least five times larger than the value of the reactance of \underline{C} . Hence, the current through the circuit, including the capacitor, is determined only by the value of resistor \underline{R} ; the current has the same wave shape as the audio input voltage. However, the voltage drop across a capacitor is inversely proportional to the frequency of the current through it, being determined by its reactance $\underline{X}C$, where

$$\underline{\mathbf{X}}_{\underline{\mathbf{C}}} = \frac{1}{2\pi \underline{\mathbf{F}}_{\underline{\mathbf{C}}}}$$
(7-1)

Thus, since the impedance of the capacitor varies inversely with the frequency \underline{F} , the magnitude of the voltage across it also varies inversely with the frequency.

In addition to this property, the capacitor possess another characteristic feature: The voltage across the capacitor is a measure of the charge upon its plates. The rate of change of the charge on the plates of a capacitor determines the current in the circuit. Inasmuch as the current through the circuit has the wave shape of the audio signal, the rate of change of the voltage across the capacitor, the output voltage, is proportional at every instant to the audio input voltage. This property provides for the second requirement.

Consequently, to produce an f-m wave at the output of a phase modulator, it is necessary only to insert a resistance-capacitor correction network, as shown in Fig. 7-2, into the modulating-voltage input lead. When examining, with an oscilloscope, the voltage input and the voltage output of the correction network, we find that a complex wave form will be distorted in the following manner: All the
components of the wave will be shifted 90 deg in their phases. This is the rate-of-change effect. The ratio of the second-harmonic amplitude to the fundamental amplitude will be reduced to one half; the ratio of the third harmonic amplitude to the fundamental amplitude will be reduced to one third; and so forth. This is the reductionin-amplitude effect.

7-3. The Armstrong Phase Modulator. From Sec. 7-2 we see that the problem is thus reduced to one of obtaining a crystal-generated phase-modulated wave. Major Armstrong solved it by converting a crystal-controlled a-m wave into a phase-modulated wave. Thus, he used amplitude modulation to obtain phase modulation which, with its proper correction, created the desired f-m wave.

To understand the theory behind the conversion from amplitude to phase modulation, let us refer to Fig. 2-3 in Chap. 2, where the composition of an a-m wave as the sum of a carrier and two sidebands is demonstrated. It is there shown that at every point of zero voltage of the carrier (the dotted lines) the amplitude of the high sideband just cancels the value of the low sideband; the resultant sum does not shift the phase of the a-m wave at any point. Notice how the points $\underline{A_1}$ and $\underline{A_2}$ occur directly under \underline{A} and how the points $\underline{B_1}$ and $\underline{B_2}$ occur directly under \underline{B} . Let us now shift the two sidebands,



Fig. 7-3. The creation of a phase-modulated wave from an amplitude-modulated wave (that of Fig. 2-6) by shifting both sidebands 90 deg with respect to the carrier wave.

as shown in Fig. 7-3, so that points $\underline{A_1}$ and $\underline{A_2}$ occur at a later time under the carrier wave—under the point \underline{C} which leads the point \underline{A}

by 90 deg. Similarly, points B1 and B2 have also been shifted forward in phase by 90 deg. To accomplish this shift in phase, we have to separate the sidebands from the carrier and shift either the carrier or sidebands 90 deg in phase while keeping the other constant. Let us now recombine the carrier with the out-of-phase sidebands. Since the sidebands are now 90 deg out of phase with the carrier, they no longer combine, so that the amplitude of the high sideband cancels the amplitude of the low sideband at the zero points of the carrier: the amplitudes actually add at those points of zero carrier amplitude, tending to shift the phase of the resultant wave a maximum amount, as shown in Fig. 7-3d. In addition, because the peak amplitudes of the two sidebands no longer occur in phase with the peak amplitude of the carrier, but at a point 90 deg advanced, the amplitude variation of the result is very much smaller. When this amplitude variation is removed, we have a practically perfect phase-modulated wave. The carrier wave is shown dotted in Fig. 7-3d as the reference for the phase shift which varies in magnitude from a minimum of θ_{lag} to a maximum of θ_{lead} .

This method of obtaining a phase-modulated wave from an amplitude-modulated wave can be described quite graphically with the



Fig. 7-4. A phasor diagram of an amplitude-modulated wave where \underline{OC} is the peak amplitude of the carrier; \underline{CD} and \underline{CE} , the peak amplitudes of the sidebands; and \underline{OW} , the resultant wave. In this case the phase of \underline{OW} remains constant and only its amplitude varies.

use of the phasor diagram. In Fig. 7-4 is shown a phasor diagram of an ordinary a-m wave. <u>OC</u> represents the carrier, which remains stationary in the diagram; <u>CD</u> represents the low sideband, which rotates clockwise about the point <u>C</u>; and <u>CE</u> represents the high sideband, which rotates counterclockwise about the point <u>C</u>. The two sidebands, being at the same frequency difference above and below the carrier frequency, rotate at the same speed in their respective directions with respect to the carrier phasor; hence, they always make equal angles with the carrier <u>OC</u>. Adding the two produces a length CW which adds to or subtracts from, depending on its direction, the length OC. Therefore, the result <u>OW</u> remains in phase with the carrier (superimposed on the carrier \underline{OC}) while varying only in amplitude to produce the a-m wave.

Let us now shift the sidebands CE and CD 90 deg with respect



Fig. 7-5. A phasor diagram of an amplitude-modulated wave wherein the sidebands have been shifted 90 deg. The resultant $\underline{OW'}$ varies both in amplitude and phase relative to the carrier OC.

to the carrier phase. The result is shown in Fig. 7-5 where the two sidebands are now rotating symmetrically about a line at 90 deg to the carrier \underline{OC} . Adding the two sidebands results in a line \underline{CW}^{\dagger} which is at 90 deg to the carrier; the final modulated wave \underline{OW}^{\dagger} , \underline{OC} plus \underline{CW}^{\dagger} , varies much less in amplitude than its counterpart \underline{OW} in Fig. 7-4 and now varies in phase, making an angle θ deg with the carrier. During one cycle of rotation the point \underline{W}^{\dagger} will move from B to A and back again, creating the desired phase modulation.

In this method of phase modulation, the actual variation in phase and therefore the variation in frequency—is quite small. To obtain wide frequency variation, many stages of frequency multiplication are necessary. Modulation usually takes place at about 200 kc and, to obtain the final high frequency, various methods are employed to obtain high multiplication values. These methods will be discussed in the next chapter, Chap. 8.



Fig. 7-6. A block diagram of the Armstrong modulator with the input and output frequencies of the several stages shown. \underline{F} is the audio frequency and \underline{f}_0 the fundamental carrier frequency.

In Fig. 7-6 is shown a block diagram of the complete Armstrong modulator. The audio input, noted as the frequency \underline{F} , is taken in

through an audio amplifier. The audio signal then passes through the correction network so that frequency modulation may be generated by means of phase modulation. The corrected audio signal modulates a carrier generated at a constant frequency \underline{f}_0 in a crystal-controlled constant-frequency oscillator. The modulator employed is of the balanced type, which suppresses the carrier and produces only the sidebands in its output. These sidebands are at the frequencies $(\underline{f}_0 + \underline{F})$ and $(\underline{f}_0 - \underline{F})$ as shown in the diagram. They are shifted 90 deg in phase and are combined with the pure carrier obtained from a separate center-frequency amplifier, also fed by the constant-frequency oscillator. Upon combination, the result is the desired crystal-controlled f-m wave which is passed on to the multiplier stages. The multipliers and amplifiers, being of the overdriven class C type, remove all amplitude variation and yield a pure f-m wave at the output.



Fig. 7-7. The circuits of a balanced modulator where tubes 1 and 2 are excited by the r-f oscillator in parallel, but are modulated by the corrected audio input in push-pull.

The circuit of the balanced modulator is shown in Fig. 7-7. It consists of two tubes, tube 1 and tube 2, in a parallel-series combination circuit. The r-f signal from the crystal-controlled oscillator is fed to both grids in parallel. The corrected audio signal, however, amplitude-modulates the r-f signal in push-pull, being fed into the screen grids through a push-pull transformer. The output, in the plate circuits of the two tubes, is derived from a push-pull tank circuit. When the carrier voltage is +E on the grid of tube 1, it will also be +E on the grid of tube 2, since they are in parallel, and the resultant plate current caused by the carrier in the push-pull tank will cancel. This is true for any value of carrier impressed on the grid. But the modulating voltage is impressed in push-pull and will be of opposite sign and will add in the push-pull tank circuit. Consequently, there will be available, at the output, only the side-

bands of the a-m wave.

To accomplish the 90-deg phase shift of the sidebands, a loosely coupled transformer secondary is employed. The voltage available across the secondary of a loosely coupled r-f transformer is always shifted 90 deg in phase with respect to the current in the primary. There is one small discrepancy, inasmuch as the theory calls for a shift in phase of the sidebands equal to 90 deg of the carrier wave, while in this circuit we are shifting each of the side bands 90 deg of their own cycle. (We note in Fig. 7-4 that the sidebands were advanced a quarter cycle of the carrier wave, not a quarter cycle of their own waves.) However, the difference is so small, because of the comparatively small difference in frequency, that the amount of distortion introduced is completely negligible.

Reviewing the complete process, we find that (1) the input audio signal is distorted in a predetermined manner; (2) the <u>corrected</u> audio signal is employed in a balanced modulator to modulate a crystal-controlled carrier where only sidebands are obtained in the output; (3) the phase of either the sidebands or the carrier is shifted 90 deg to obtain the desired phase relationship; and (4) the two resultant signals are combined to obtain an f-m signal which may be multiplied up to any deviation.

7-4. The Phasitron.² The phasitron system affords a means of obtaining wide-angle phase modulation. It employs a specially designed tube called the "phasitron." If the input signal is properly corrected by means of a correction network, the output of the phasitron will be an f-m signal.

Thus, the phasitron tube provides another method of phase-modulating a crystal-controlled carrier where wide phase deviations, and therefore wide frequency variations, are possible without a great deal of multiplication. A schematic diagram of the phasitron tube is shown in Fig. 7-8. Plates 1 and 2 are at a positive potential and draw electrons from the vertical cathode in the center of the tube. By means of two focus electrodes which surround the top and bottom of the cathode, the emitted electrons are formed into a tapered thinedge disk. This disk, with the cathode for its axis, lies between the neutral plane and the deflector-grid structure and extends out to plate 1—the inner of the two positive electrodes. The electrons in this disk are traveling outward from the cathode to the plate as in any normal tube. They first reach plate 1, but, if they are lined up with any of the holes in plate 1, they will travel on to reach plate 2.

The deflector grids consists of 36 symmetrically spaced wires, the active portions of which run radially out from the cathode between the inner and outer focusing electrodes. These wires are labeled <u>A</u>, <u>B</u>, and <u>C</u>. A three-phase constant-frequency voltage is then applied to these wires, one phase to every third wire, so that all the <u>A</u> wires in the figure will have one phase; all the <u>B</u> wires, another phase;

and all the \underline{C} wires, the third phase. The return lead for the three phases is the neutral plane, a conducting surface occupying the same



Fig. 7-8. The schematic diagram of the phasitron showing the internal construction and the staggered holes in plate 1.



Fig. 7-9. The developed or unrolled view of the grid structure and neutral plane of the phasitron tube.

area as the grid deflector but on the opposite side of the electron disk. Thus, a voltage existing between the deflector grids and the neutral plane will cause the electrons in the disk to be deflected up or down, depending on the instantaneous voltage at the grid past which the electron is traveling.

A developed view of the grid structure on the neutral plane is shown in Fig. 7-9, where the whole structure has been unrolled into a flat sheet. The deflecting action of these three-phase voltages can now be seen: at instant 1 in Fig. 7-9, the grid wires <u>A</u> are positive with respect to the neutral plane, while the grid wires <u>B</u> and <u>C</u> are negative. Consequently, the electrons passing grid wires <u>A</u> will be deflected downward toward the grid, while the electrons passing grid wires <u>B</u> and <u>C</u> will be deflected upward toward the neutral plane. This is shown as the sine-wave edge of the disk at instant 1 in Fig. 7-9a. Actually, the edges of the disk, as shown in the per-



Fig. 7-10. The undulated electron disk obtained when the three-phase voltage is connected to the deflector grids of the phasitron.

spective view of Fig. 7-10, have been distorted into a sine-wave ruffle. At instant 2 in Fig. 7-9, one third of a cycle later, deflector wire B is positive and wires A and C are negative. The resulting effect will be that shown for the edge of the disk at instant 2 in Fig. 7-9a. The undulate disk on Fig. 7-10 would then appear to have moved the space of one grid wire during the time interval between instant 1 and instant 2. From this explanation, it can be seen that with the three-phase voltages applied to the grids, the undulate electron disk appears to rotate at a rate determined by the applied frequency and the number of deflector grids. The frequency of rotation will be the input frequency divided by one third the number of grid wires, since the disk moves the distance between one grid wire for each third of a cycle. For 36 wires the undulations will rotate at one twelfth the frequency of the three-phase voltage. Actually, the electrons in the disk do not rotate but the undulations move around the circumference like waves on the surface of a body of water.

A developed view of a portion of plate 1 is shown in Fig. 7-11. This anode has 12 holes punched above the undeflected electron disk and 12

holes punched below. The rotating undulate edge of the electron disk therefore impinges on this series of holes. At an instant when the



Fig. 7-11. A developed view of the holes in the first plate showing how the edge of the electron disk either goes through the holes in the plate or terminates on the plate.

disk edge is lined up, as shown by the solid line, most of the electrons pass on through the holes to plate 2, the second anode. As the undulations in the electron disk rotate, fewer and fewer electrons will go through the holes until, a half cycle later (of the three-phase rotating voltage), the disk will assume the position shown dotted in Fig. 7-11. At this latter position few, if any, electrons get through to plate 2. Thus, the current flowing to plate 2 varies sinusoidally at the threephase voltage frequency. Also, we see that any variation in the angular velocity of rotation of the undulations will result in phase and frequency modulation in this output current.

Suppose now that we introduce a magnetic field perpendicular



Fig. 7-12. The method of impressing the magnetic field on the phasitron to introduce phase modulation.

to the electron disk. It may be done by placing a magnetic coil (or solenoid) around the tube, as shown in Fig. 7-12. Now, from the

basic theory of magnetic deflection, the electrons traveling radially outward from the cathode will have a force exerted on them in a direction perpendicular to their path and perpendicular to the direction of the magnetic field. The electrons will then travel in a spiral about the cathode, causing the undulations in the edge of the disk to twist, thus introducing a phase shift in the output current—the current reaching plate 2. Audio-frequency current flowing in this coil causes a-f angular displacements of the undulations in the disk, at the a-f rate, to be superimposed on the rotation of the undulations. Thus, phase shifts which vary at the a-f rate are obtained in the output current. The result—a phase-modulated wave. The average, or center frequency, is the frequency of the three-phase voltage. In the actual transmitter, this voltage is crystal-controlled. Also, since the amount of electrons reaching the plates does not change, there should be no amplitude modulation present in the output.

Our two requirements for obtaining an f-m wave at the output of a phase modulator were: The audio voltages (and their components) have to be reduced in amplitude proportional to their frequencies; and the signal has to be transformed into a modulating voltage whose rate of change is equal to the amplitude of the frequency-corrected signal (Sec.7-2). Both of these corrections can be accomplished, within the modulating coil, provided that the impedance of the coil can be made a purely inductive reactance. When a constant-voltage source—the voltage being independent of the frequency—is used to supply current to a purely inductive reactance, the current flowing through the inductance will be inversely proportional to the frequency, inasmuch as the impedance of an inductance is proportional to the frequency. Therefore,

$$\underline{X}_{\underline{L}} = 2\pi \underline{FL}$$
(7-2)

where $\underline{X}_{\underline{L}}$ is the value of the reactance; \underline{F} , the frequency being used; and \underline{L} , the value of the inductance in henrys. In addition, since the voltage across an inductance is equal to the rate of change of the current through it, the rate of change of the current will be equal to the amplitude of the impressed audio voltage—the second requirement. The last link in the chain is that the magnetic field within the tube be directly dependent on the current in the coil, as it actually is; consequently, the modulating force—the magnetic field—will be actuated by a corrected audio-wave shape—the current in the coil, and the output of the phasitron will be frequency-modulated.

The physical size of the tube (GL-2H21) is shown in Fig. 7-13, where the tube is being held in front of a diagram showing its internal construction. All leads to the tube elements are brought out through the base of the tube so that an ordinary type tube socket may be used. The phase deviation obtained from a phasitron (GL-2H21) tube as a function of the magnetic field strength in gausses is shown in Fig. 7-14. With this wide phase-angle modulation, the amount of frequency modulation produced in the output is quite large.



Fig. 7-13. The phasitron tube being held in front of a diagram showing its internal construction.

A frequency deviation of ± 175 cps at 230 kc per sec is obtainable. Hence, for a frequency deviation of ± 75 kc at the desired frequency, a multiplication factor of only 432 is needed to obtain a crystalcontrolled f-m signal in the f-m broadcast band.

7-5. Phase Modulation by Resistance Variation.³ Another way of converting a stabilized carrier signal into a phase-modulated signal is by the use of a series circuit. Let us consider the circuit shown in Fig. 7-15. A voltage \underline{V}_{in} is impressed across a series circuit consisting of a variable resistance <u>R</u> and a fixed capacitor <u>C</u>. The phase of the current <u>I</u> through the circuit will depend on the values of <u>C</u> and <u>R</u>. If <u>R</u> is small with respect to the reactance of <u>C</u>, the current will lead by practically 90 deg. As the value of R is increased, the circuit will become more and more resistive, causing the leading phase angle of the current to become less and less until, when R is



Fig. 7-14. The phase deviation in degrees of the phasitron (GL-2H21) versus the field strength in gausses.

large with respect to the reactance of <u>C</u>, the phase angle is practically zero. The voltage drop across a resistance is in phase with the current through it; thus, the output voltage \underline{V}_{out} is taken off across the resistor, as shown. As the resistance <u>R</u> is varied, therefore, the current <u>I</u> will vary in phase with respect to the input voltage; the output voltage \underline{V}_{out} will be phase-modulated. If the resistor is replaced by a linear element, such as a vacuum tube, acting as a vari-



Fig. 7-15. A simple series circuit, a capacitor \underline{C} and a resistor \underline{R} , used to phase-modulate a stabilized carrier.

able resistor, phase modulation may be accomplished at the signal frequency.

A circuit employing two such phase shifters is drawn in Fig. 7-16. The phase-shift network consists of the capacitance reactance \underline{X} in series with the plate impedance of a 6J5 as the variable resistance. The B+ voltage is supplied through a high enough resistance or, if necessary, an r-f choke, so that the amount of r-f current flowing through that path to ground is negligible. As the grid voltage is varied by the corrected audio voltage, the phase of the voltage drop across the 6J5 varies, and phase modulation of the crystal-controlled carrier is accomplished. The effect of using two stages is additive, thus doubling the phase shift. This circuit is the simplest type of phase-shift network, but it suffers from the disadvantage that the attenuation per section (the loss of voltage across the capacitor) is considerable and results in a rapid decrease in voltage along the string of phase shifters.

Another interesting circuit, which maintains a constant-amplitude output while shifting the phase, employs the so-called "phase-shifting



Fig. 7-16. A phase shifter using a series circuit of a resistor and a capacitor, where the plate impedance of a 6J5 is used as the resistor. Two cascaded stages of phase shift are shown.

bridge." The bridge is illustrated in Fig. 7-17. Two similar fixed resistors $\underline{R}_{\underline{A}}$ and $\underline{R}_{\underline{B}}$ are used to obtain a center tap across the input voltage \underline{V}_{in} . When the voltage across $\underline{R}_{\underline{A}}$ is positive at any instant (from the center tap O to G), it will be negative, or 180 deg out of phase, across $\underline{R}_{\underline{B}}$ (from the center tap O to H). Also across the input, and completing the bridge, is a series circuit consisting of a capacitor C and a variable resistance R. The output voltage is taken off from the center tap O to the tap between the capacitor C and the variable resistor R.

Let us examine the extreme variation points. When <u>R</u> is zero, the output voltage will be the voltage across <u>RB</u> from <u>O</u> to <u>H</u>; and when <u>R</u> is very large, the output voltage will be practically equal to the voltage across <u>RA</u> from <u>O</u> to <u>G</u>. Since the two resistors are equal in value, the output voltage will have shifted 180 deg, but will not have changed amplitude. By varying <u>R</u> gradually between these two points (zero resistance and a very large magnitude with respect to the magnitude of the reactance of <u>C</u>), it is possible to shift the phase uniformly from zero to 180 deg and maintain a constant amplitude for the output voltage.

A circuit utilizing a bridge-type phase shifter is depicted in Fig.

7-18. The input voltage is impressed across the bridge by a cathode follower (one half of a 6SL7 tube) which is used to maintain a constant



Fig. 7-17. A phase-shifting bridge where the phase of the output voltage is shifted between 0 and 180 deg while maintaining a constant-amplitude output.

voltage across the bridge. The input to the cathode follower is obtained from a crystal-controlled oscillator. The value of the ef-



Fig. 7-18. A bridge-type phase shifter as used in a circuit with the resistance variation being accomplished by a cathode follower.

fective resistance across the terminals of <u>R</u> is varied by shunting it with a cathode-follower 6AC7. As the corrected audio signal varies, the grid voltage of the shifter—the effective resistance from ground to cathode—varies, thus accomplishing the phase shift. The output voltage from the phase-shifter bridge will be balanced to ground; it has to be converted into an unbalanced voltage before it can be fed into the next stage. This conversion is accomplished by the repeater (the other half of the 6SL7). From there the signal may be passed through additional stages of phase shift. This bridge circuit has the advantage of almost complete freedom from amplitude modulation, but the fact that each stage is balanced to ground and must be converted to an unbalanced voltage to supply the next stage is a disadvantage. This conversion can be accomplished by means of a balancing transformer, but the transformer must have a band width at least twice the highest signal frequency. The balance requirement adds to the complexity of the adjustments and to the number of components required. Four or five stages of this type of shifter give acceptable results.

Another interesting circuit is the constant-impedance type of



Fig. 7-19. A constant-impedance phase-shift circuit with its characteristic phasor diagram. A constant-amplitude varying-phase output is obtained when the circuit is fed with a constant current.

phase shifter shown in Fig. 7-19. It consists of an inductance in parallel with a resistor-capacitor series circuit. R_2 , the resistor, is made variable. The reactance of the capacitor is half the reactance of the inductance.

Now let us examine the extreme points of variation. When R is very large, an inductive current will flow (the current through R being negligible), and the impedance of the circuit will be inductive. When R is zero, a capacitive current of twice the magnitude of the inductive current will flow through the capacitor. Half of this current will neutralize the inductive current and the other half will flow through the input source. The input source will see an impedance of the same magnitude as the foregoing inductive impedance, but it will be a capacitive impedance. As R_2 is varied between these two values, zero resistance and maximum resistance, the impedance of the total circuit across the input Ein will vary from a pure capacitance to a pure inductance while maintaining at all times a constant magnitude. If a constant-current source is used, for instance, a pentode, the voltage across the circuit will vary from 90 deg lagging, for the capacitive value, to 90 deg leading, for the inductive value, as shown in the diagram.

Figure 7-20 shows a circuit diagram of a constant-impedance type of phase modulator. The constant impedance consists of the in-



Fig. 7-20. A constant-impedance type of phase modulator which makes use of the cathode-to-ground resistance of a vacuum tube (cathode follower) to vary the phase angle of the constant-impedance network.

ductive reactance $\underline{X}_{\underline{L}}$, the capacitive reactance $\underline{X}_{\underline{C}}$, and the resistance \underline{R} shunted by a cathode follower called the "shifter." The crystalwith a constant current. The effective resistance across the terminals of the resistor \underline{R} varies as the corrected audio voltage varies the grid bias of the cathode follower. Consequently, the voltage across the constant-impedance circuit is frequency-modulated. With the proper operating point on the shifter tube, it is possible to obtain an f-m signal output containing very little distortion.

7-6. Cathode-Ray-Tube Phase Modulator. Another type of phase modulator makes use of the electron beam in a cathode-ray tube. The beam is rotated in a circle by impressing on the deflecting plates of the cathode-ray tube a two-phase voltage (similar to the operation of a split-phase motor). Instead of allowing the cathode-ray beam to impinge on a fluorescent screen as in an oscilloscope, it is focused on a special target of the shape shown in Fig. 7-21a. The target consists of two similar conducting plates A and B, which are separated from one another by a pair of spiral cuts. Each plate is connected to a ground return through its own resistor, plate A through RA and plate B through RB. When the electron beam strikes plate A, a current will flow from ground to plate A as shown by the arrow on R_A . This current consists of the electrons in the beam returning to ground. When the electron beam strikes plate B, a current will flow from ground to plate <u>B</u> as shown by the arrow on \underline{R}_{B} . Hence, the output voltage will reverse, depending on which plate is being struck by the electron beam.

To operate the tube as a phase modulator, the electron beam is rotated in a circle by impressing a two-phase voltage on the deflection circuits, one phase on the vertical deflection plates and the other phase on the horizontal deflection plates. The diameter of the circle obtained may be varied by either varying the deflection sensitivity or by varying the magnitudes of the deflecting voltages. Let us assume that at instant 1 the circle is small, as shown in Fig.



Fig. 7-21. A cathode-ray tube target as used to generate a phase-modulated wave. It consists of two conducting plates separated by a pair of symmetrically placed spiral insulating air gaps.

 $7-21\underline{a}$. As the beam rotates it will generate a square-wave voltage in the output, as shown in Fig. 7-21<u>b</u>. If the diameter of the circle on the target plates is changed to, let us say, the diameter shown as instant 2, the phase of the generated square-wave voltage will change because the spiral has put the two plates of the target in a new position with respect to the circle. This phase shift for the circle of instant 2 is shown in Fig. 7-21<u>c</u>. Hence, by changing the diameter of the scanning circle it is possible to produce phase modulation. A tank circuit in the output will convert the square wave to a sinusoidal signal with low harmonic content.

A block diagram of a possible method of using the cathode-raytube phase modulator as an f-m generator is shown in Fig. 7-22. The corrected audio signal is used to amplitude-modulate an r-f signal generated by a crystal-controlled oscillator. The a-m signal is then divided into two parts, one of which is shifted 90 deg in phase. Thus, we have an a-m two-phase voltage amplitude-modulated by the corrected audio signal. This two-phase voltage is impressed on the deflection plates of the cathode-ray-tube phase modulator and varies the diameter of the circle described by the electron beam on the target plates. From the tank circuit connected across the target plates, a crystal-controlled f-m signal will be obtained.



Fig. 7-22. A block diagram of a possible method of using the cathode-ray tube phase modulator as an f-m generator.

7-7. The Serrasoid Modulator.⁴ Another very interesting type of phase-to-frequency modulator is the serrasoid modulator which employs pulse-phase modulation. A crystal oscillator in the neighborhood of 100 kc is used to generate a series of very linear sawtoothed waves. The saw-toothed wave is then directly coupled to the grid of a tube that is cathode-biased so that conduction begins about halfway up the saw tooth. A pulse is generated at the point of conduction. The bias is then varied with the corrected audio voltage so that the point of conduction is varied up and down the saw tooth. Thus, the pulse generated at the point of conduction is phase-modulated. The pulse can then be used to excite a tank circuit, which, since the pulses are phase-modulated with the corrected audio voltage, yields an f-m wave that can be multiplied up to the desired carrier frequency with the desired deviation. The circuit is capable of phase modulation of peak deviation of 150 deg. Noise 80 db down with only 0.25 per cent distortion for 100 per cent modulation is claimed for the modulator which, in its final form, is very compact.

REFERENCES

1. Armstrong, E. H., "A Method of Reducing Disturbances in Radio Signalling by a System of Frequency Modulation," <u>Proc. IRE</u>, May, 1936, p. 689

2. General Electric Company.

3. W. E. Phillips and M. Marks, "Phase Modulators," <u>Sixth Annual Broadcast</u> Engineer's Conference, 1946.

4. Day, J. R., "Serrasoid F-M Modulator," Electronics, October, 1948, p. 72.

QUESTIONS

1. Is it possible to tell whether an r-f wave is phase- or frequency-modulated by examining the wave itself? Explain.

2. How does phase modulation differ from frequency modulation?

3. Explain how an f-m wave may be obtained with the use of a phase modulator.

4. Describe a correction network for the audio voltage in a phase-to-frequency modulation system; explain how it operates.

5. Describe the principle of the Armstrong phase modulator.

6. Explain the Armstrong phase modulator with the use of phasor diagrams.

7. Describe, with the use of a block diagram, the operation of the complete Armstrong amplitude-to-phase-to-frequency modulation.

 ${\bf 8.}\,$ Explain the operation of the balanced modulator, and show how it suppresses the carrier.

9. Draw a schematic diagram of the phasitron tube and explain briefly the functions of the various electrodes.

10. Discuss the function and operation of the multiple-grid deflection wires in the phasitron.

11. What rotates at the crystal-controlled frequency inside the phasitron tube?

12. At what speed does the rotation discussed in Question 11 take place, and upon what does it depend?

13. How is the output voltage generated in the phasitron tube?

14. How is modulation accomplished in the phasitron tube?

15. Why is it not necessary to use a separate correction network with the phasitron tube?

16. Describe the series-type phase-modulation circuit using resistance variation.

17. Draw and explain the circuit of a phase modulator using the phase-shifting bridge circuit.

18. Explain the operation of the constant-impedance phase-modulation circuit.

19. Draw a circuit diagram of a phase modulator using the constant-impedance circuit.

20. Describe the operation of the cathode-ray-tube phase modulator.

21. Explain how the cathode-ray-tube phase modulator may be used to generate frequency modulation.

CHAPTER 8

Frequency-Modulation Transmitters Using Phase Modulation

8-1. Original Armstrong Modulators. In Sec. 7-3 we discussed the theoretical aspects of the Armstrong phase modulator and one



Fig. 8-1. A circuit diagram of the modulator unit for a balanced-modulator type of Armstrong f-m generator.

of its components - the balanced modulator. In Fig. 8-1 is shown a circuit diagram of the complete modulator unit in a balanced-

151

modulator type of Armstrong f-m generator. The audio input is taken in through an RC correction network which performs the proper audio correction for an f-m signal output. The output of the audio amplifier furnishes the push-pull input for the balanced modulator. The signal from a crystal-controlled oscillator is divided into three channels, one channel feeding the carrier amplifier, and the other two channels feeding the two balanced-modulator tubes in parallel. The plate circuit of the balanced modulator is tuned to resonance (made purely resistive impedance) by means of the two series capacitors C_1 and C_2 . These balance the two circuits and keep the plate current in phase with the carrier input voltage. The 90-deg phase shift, as discussed previously, takes place across the transformer feeding into the sideband amplifier. The sidebands and the carrier are combined in the resistor R1 located in the plate circuit of the carrier amplifier. An ordinary resistance load is used because the frequency employed is only about 200 kc. The combined wave is the desired f-m wave. It is amplified in the r-f amplifier and multiplied the necessary number of times to obtain an f-m signal at the assigned frequency and with the proper frequency deviation.



Fig. 8-2. A balanced-modulator circuit of sideband generator using push-pull inputs and paralleled outputs.

Another way of generating sidebands is shown in Fig. 8-2. This is really another type of balanced modulator wherein the carrier and audio inputs are both fed into the two tubes in a balanced pushpull fashion while the output is connected across both plates in parallel. The oscillator output may be converted to push-pull (balanced) voltages by means of a transformer, resistance-inductance networks, or resistance-capacitor networks. In this case a resistancecapacitor network is employed. For tube 1 a small (high-impedance) capacitor C_1 is used in series with a low value resistor R_1 ; hence, the current in the circuit is determined by the capacitor - it leads the voltage by 90 deg. By taking the voltage drop across the resistor R_1 , which is in phase with the current, and applying it to the grid of tube 1, we obtain for the tube a voltage input that is leading the voltage output of the oscillator by 90 deg.

For tube 2 a high-value resistor and low-impedance capacitor is used; the current is in phase with the voltage output of the oscillator. To maintain equal voltage magnitudes, the value of the reactance of $\underline{C_1}$ divided by the resistance of $\underline{R_1}$ is made equal to the value of resistance of $\underline{R_2}$ divided by the reactance of $\underline{C_2}$. Thus, the voltage dividing action of both networks is the same. In the case of tube 2, the grid voltage is obtained across the low-impedance capacitor; this voltage lags the current (and hence the oscillator voltage) by 90 deg. Since tube 1 is fed with a grid voltage that leads 90 deg, the two voltages are 180 deg out of phase - the correct relationship for push-pull operation. These networks are similar to those employed in obtaining the 90-deg phase shifts in the reactance-tube circuits and, as in those circuits, resistance-inductance networks can also be used.

The two plates of the modulator tubes are connected in parallel through the load resistor \underline{R} . With no modulation, when tube 1 causes the current in the resistor to increase, tube 2, 180 deg out of phase, causes the current to decrease, thereby canceling the current from tube 1. Consequently, the carrier balances out. With modulation, the magnitude of the r-f output of tube 1 will increase when the magnitude of the r-f output of tube 2 decreases. The difference between the two will be the sideband voltages, and they will be available across the load resistor \underline{R} . Also combining in the resistor \underline{R} is the carrier obtained from the carrier amplifier. With the properly corrected audio voltage, this combination will result in the desired f-m output signal ready to be multiplied in the following stages.

8-2. Frequency-Deviation Multiplication. As mentioned previously, the source of r-f in the Armstrong method of f-m generation is a crystal oscillator operating in the neighborhood of 200 kc. The phase deviation employed is less than 0.2 radians to insure negligible distortion in the output. This results in a very small frequency deviation and a multiplication of anywhere from 6,000 to 13,000 is needed to secure the 75-kc deviation normally used in the broadcast band. If the signal were multiplied 8,000 times directly, it would yield an output frequency for the transmitter of over 1,000 megacycles. For an output frequency in the 100-megacycle band, a method of double conversion may be employed. The first method



Fig. 8-3. The original method of large-scale multiplication employed with the Armstrong modulator. Any deviation of either crystal affects the stability of the output.

to be used is illustrated in the block diagram of Fig. 8-3. Also shown are some frequencies and multiplication figures which may be used to obtain an output frequency of 105 megacycles. The original oscillator, at a frequency of 200 kc, and the modulator, with a phase deviation of 0.2 radians, generate a signal with a frequency deviation at the lowest frequency of, let us say, about ± 10 cps. To obtain a frequency deviation of 75 kc it is necessary to use frequency multipliers with a resultant multiplication of 7,500. As shown in Fig. 8-3, frequency multipliers with a multiplication factor of 100 are first used to obtain a signal at 20 megacycles with a frequency deviation of 1 kc. A converter used at this point steps the frequency down to one seventy-fifth of the final carrier frequency. For instance, if a carrier frequency of 105 megacycles is desired, the frequency is stepped down to 1.4 megacycles by beating it with a crystal-generated frequency of 18.6 megacycles. Since this is equivalent to subtracting the beating frequency, the deviation remains the same at the output of the converter as at the input. Subtracting 18.6 megacycles from 20 megacycles plus 1 kc (20,001 kc) yields 1,401 kc, 1.4 megacycles plus 1 kc. Thus, a 1.4-megacycle signal is obtained with a 1-kc deviation. Another series of multipliers with a multiplication factor of 75 is then utilized to produce the final desired signal at 105 megacycles with a frequency deviation of 75 kc. This signal is then fed into the power amplifiers for amplification before transmission.

This method of multiplication suffers from one inherent defect: the final frequency is dependent on the stability of both crystal oscillators. If either one drifts in frequency, the output frequency will also drift; the stability of the output frequency is thus only about one half the stability of a single crystal-oscillator system.

8-3. Two-Channel Multiplication. There is another method of multiplication which has been used successfully, producing a final carrier which is dependent on the stability of only one crystal. This is illustrated in block-diagram form in Fig. 8-4. A 200-kc crystal-

controlled oscillator is utilized, as before, but two separate modulators are employed. The crystal-oscillator output is fed into both



Fig. 8-4. The two-channel system shown in block diagram form. Only the stability of the second crystal oscillator $\underline{f_2}$ determines the stability of the final carrier output.

of the modulators in parallel. The corrected audio signal, on the other hand, is reversed in phase before it is fed into modulator 2. This phase reversal is accomplished by merely reversing the connections of the audio input. Because of the phase reversal, when modulator 1 shifts the phase forward, modulator 2 retards the phase an equal amount. In other words, when the output from modulator 1 is increasing in frequency, the output from modulator 2 is decreasing in frequency. This is indicated by using the \pm sign in front of the frequency deviation obtained from modulator 1 and the reversed \mp sign in front of the frequency deviation obtained from modulator 2.

To transmit very low audio frequencies, the frequency deviation at the output of the 200-kc oscillators is only 6.43 cps. Separate frequency multipliers are used for each channel. These multiply each of the outputs by a factor of 81, as indicated on the diagram. Channel 1 produces 16.2 megacycles ± 521 cps, and channel 2 produces 16.2 megacycles ∓ 521 cps. At this point a stable crystal-controlled oscillator is employed. This oscillator is adjusted to produce one seventy-second of the final carrier frequency desired. This frequency is noted as f_2 in Fig. 8-4. It is combined with the output of channel 1 in the channel 1 mixer. Similar to the heterodyning in the previous discussion, the frequency produced is the difference between the two outputs, $(16.2 - f_2)$ megacycles ± 521 cps. This output from the channel 1 mixer is combined with the output of channel 2 in the

ł



Fig. 8-5. The circuit diagram of a modulator and multiplication unit for the twochannel system of phase-to-frequency modulation.



channel 2 mixer. Again the difference frequency is obtained. Subtracting, we obtain a carrier frequency of \underline{f}_2 with twice the deviation of each of the original frequencies: namely, ± 1.042 kc. When one frequency increases with modulation the other frequency decreases, causing the difference deviation to be the sum of the two, or, since they are equal, twice the deviation of each. This output frequency is then multiplied by 72 to obtain the desired carrier frequency with a frequency deviation of ± 75 kc. A class C power amplifier is then used to produce the required amount of power. The effect of varying the 200-kc frequency at the first crystal oscillator does not influence the final carrier frequency \underline{f}_0 , since it is effectively canceled out at the second modulators. The stability of the carrier is dependent only on the stability of the \underline{f}_2 crystal oscillator.

8-4. Two-Channel Transmitter. A circuit diagram of a modulator using two-channel multiplication and a modified Armstrong modulator is shown in Fig. 8-5.1 This circuit follows the block diagram of Fig. 8-4 where a multiplication of 11,664 is employed to produce a 75-kc deviation in the f-m broadcast band. These figures are based on an audio response of 30 to 15,000 cps and a phase deviation in the modulators of 0.214 radian.

In the crystal oscillator (200 kc) a 6J5 is used in a conventional circuit. Since the final carrier frequency does not depend on this unit, the crystal is not critical as to accuracy; however, a low-drift CT cut, which does not vary its resonant frequency with small variations in temperature, is used. A 6SJ7 buffer amplifier, lightly coupled to the oscillator, contains in its output circuit a resistance-capacity network to produce two voltages in phase quadrature (phase difference = 90 deg) required to drive this type of phase modulator.

A three-stage push-pull amplifier using 6J5's forms the audio section. The input transformer matches a 600-ohm balanced line to the grids of the first pair of 6J5's. Located on the secondary side of the transformer is a resistance-capacitor 75- μ sec pre-emphasis network which is standard (the product of <u>R</u>, in ohms, times <u>C</u>, in farads, is equal to 75x10⁻⁶). These are the paralleled resistor and capacitor circuits which are inserted in the grid leads. Each side of the amplifier drives two modulator tubes. To obtain the necessary inverse relationship between modulating voltage and frequency, a corrector network is placed at the end of the amplifier. A 0.25- μ f capacitor connected from each modulator screen to ground makes up the capacitor section and also serves as an r-f screen bypass. Placing the correction network at this point in the circuit enables all stages of the audio system to be operated at reasonably high levels, resulting in a lower noise level.

Each of the modulator and multiplier sections uses two 6SK7's as modulators. In operation, one of the 200-kc voltages is applied to the control grids of one tube in each modulator pair and the other quadrature 200-kc voltage is applied to the other tube of the pair.

A very simplified sketch of the modulation system is shown in Fig. 8-6. It is quite similar in circuit appearance to a balanced modulator



Fig. 8-6. Simplified sketches of the modulator circuits as used in the circuit of Fig. 8-5. The phasor diagrams for each of the modulators are also shown, illustrating the resultant oppositesign deviations.

except that the circuit is so arranged that the r-f signal is fed into the the two tubes of each pair in quadrature instead of in push-pull. The plates of the tubes of each modulator are parallel-connected, the plate loads being double-tuned band-pass transformers. Referring to the phasor diagrams shown next to the modulators in Fig. 8-6, we find that <u>A</u> is the voltage developed across the load from the tube <u>A</u>, and <u>B</u> the voltage from the tube <u>B</u>. These combine to produce the resultant <u>R</u> in both cases, no modulation being present. R is the result of adding two equal voltages 90 deg out of phase; it will be located midway between the two voltages at 45 deg.

When a modulating voltage is applied to the modulator screens, connected as shown in Fig. 8-6, the output of each of the modulator tubes will vary. If at one instant the screen of tube <u>A</u> of modulator 1 is being driven more positive, the output of this tube will increase. Simultaneously, the output of tube <u>B</u> will decrease. Consequently, as the dotted line shows us, the phase of the resultant <u>R</u>! will be advanced. Because the phase of the audio is reversed at modulator 2, the resultant voltage in the load of modulator 2 is retarded in phase an equal amount. By using the corrected audio input, this produces an f-m signal at the output of modulator 1 with a \pm frequency deviation and a signal at the output of modulator 2 with a \mp frequency deviation. We see that by feeding the r-f signal into the modulators in quadrature, we actually produce a carrier in the output which is shifted 90 deg from the sidebands, eliminating the necessity for separate carrier amplification and recombination with the sidebands.

Each modulator is followed by its own multiplication channel: two tripler stages, a buffer-amplifier stage, then two more tripler stages. All stages use 6SJ7's.

The tank circuits are designed to pass a band of ± 20 kc from the center frequency. This band pass is more than sufficient, since the modulation factor is still small enough in those stages that only the first sideband is significant. It allows a margin of safety so that frequency alignment is unnecessary. These are the two X 81 multipliers shown in channels 1 and 2 of Fig. 8-4. Their outputs have frequency deviations of 521 cps at a carrier of 16.2 megacycles for 100 per cent modulation.

The master crystal oscillator, controlling the carrier frequency, operates at one seventy-second of the carrier frequency, as previously stated. The crystal can be of the standard-broadcast type, using a thermostatically-controlled oven-type holder, with an adjustable air gap. An amplifier stage, loosely coupled to the oscillator tank, provides push-pull output at the crystal frequency.

The output of the master oscillator is applied to the signal grids of the two 6SA7's connected as a balanced mixer. The control grids, connected in parallel, are excited by the output of modulator and multiplier 1. The 6SA7 plates are connected in push-pull and balance out any of the 16.2-megacycle frequencies that may get through. An isolating amplifier using a 6SJ7 follows the mixer; its output is applied to the signal grid of a second 6SA7—the second mixer. The control grid of this second mixer receives its drive from the modulator and multiplier 2.

To follow the action of the mixer section, let us assume a crystal frequency of 1.5 megacycles. This is mixed in the first mixer with the 16.2-megacycle ± 521 -cps signal, the output being tuned to the difference signal. Therefore, we are applying to the second mixer through the isolating amplifier a frequency of 14.7 megacycles ± 521 cps. This is then mixed with the 16.2-megacycle ± 521 cps signal from section 2 and the 'output circuit is again tuned to the difference frequency, a frequency of 1.5 megacycles ± 1.042 kc—the master oscillator frequency and the sum of the two frequency deviations. It will be noted that the frequencies produced from the 200-kc oscillator are balanced out; the 200-kc oscillator exercises no control over the output carrier frequency.

In the final multiplier chain, four 6SJ7 stages, an amplifier, a tripler, a doubler, and an amplifier, are used. This last amplifier is followed by two 6AG7 doublers, the output of the second 6AG7 driving an 807 amplifier.

All tank circuits have to be designed to pass a band sufficiently wide to transmit all sideband pairs which are significant at the particular frequency concerned, thus avoiding sideband cutting and consequent distortion. A multiplication of 24 is achieved in this multiplier section; the output of the multiplier contains a frequency deviation equal to 3,888 times the original deviation produced in each of the modulators. This amount of multiplication, when followed by an 829 tripler stage in the power amplifier (not shown in Fig. 8-5), provides the total multiplication of 11,664 as required by the system.

The power amplifier can consist of any of the tubes or circuits discussed in Chap. 6, provided that a tripler is employed as the driver. For a 250-w output, an 829 tripler driving four tetrodes, 4-125A's, can be employed. A front view of such a transmitter is



Fig. 8-7. A front view of the two-channel modulator and multiplier unit mounted on a vertical hinged chassis. Note the careful individual shields to prevent interaction.

shown in Fig. 8-7. Enclosed within the pipe support is the modulator unit. Vertical construction and individual shields are employed to prevent any interaction which might distort the output.

8-5. Dual-Channel Modulator. The dual-channel modulator method of f-m generation is a combination of the two modulators of

the two-channel system into a single modulator circuit. It has the same advantages as the two-channel system, plus simplification. A block diagram of the dual-channel modulator as used in an f-m



Fig. 8-8. A block diagram of a transmitter employing a dual-channel modulator. The stability of the output carrier is dependent only on the stability of crystal oscilator \underline{B} .

transmitter circuit is shown in Fig. 8-8. We see how similar this diagram is to the two-channel system as shown in Fig. 8-4, except that the two modulators have been replaced with the dual-channel modulator. This particular circuit, which is used in practice, yields an effective multiplication factor of 7,776; hence, a frequency deviation of about ± 9.65 cps is necessary for a deviation of ± 75 kc at the output. The operation of the dual-channel modulator is similar to the operation of two separate modulators using the same oscillator and the same audio input. It produces two outputs, but, although they both have the same center frequency, when one deviates in the plus direction, the other deviates in the minus direction. As in the two-channel method, it is equivalent to reversing the phase of the audio for one of the outputs, again indicated in Fig. 8-8 by using \pm for one output and \mp for the other.

The rest of the transmitter is quite similar to the two-channel transmitter discussed previously. The outputs are fed into two separate channels, 1 and 2. Each channel consists of a number of multiplier stages with an over-all multiplication of 81. Hence, at the output of each channel the signals have a frequency deviation of 781 cps, each deviation still being the reverse of the other. For an output carrier frequency of 108 megacycles, a crystal oscillator <u>B</u> at 2,250 kc is used to convert the frequency of channel 1 to 13,950 kc with a frequency deviation of \mp 781 cps. The signal is then mixed with the out-

put signal of channel 2. The output of this channel 2 mixer is, as discussed in the previous transmitter, at the frequency of the crystal oscillator <u>B</u> and is unaffected by any variation or drift of the frequency of crystal oscillator <u>A</u>.

Also, since the deviations obtained at the output of the channel 1 mixer and at the input from channel 2 are opposite in sign, the two deviations will add, resulting in a signal at 2,250 kc with a frequency deviation of $\pm 1,562$ cps. The final frequency-multiplier stages have a resultant frequency multiplication of 48. Accordingly, the input to the power-amplifier stages is at 108 megacycles with a frequency deviation of ± 75 kc.

A circuit diagram of the dual-channel modulator is shown in



Fig. 8-9. A circuit diagram of the Armstrong dual-channel modulator, showing the production of two outputs at the same carrier frequency but with opposite sign frequency deviations.

Fig. 8-9. The output of the crystal-controlled oscillator, operating at a frequency of about 200 kc, is first amplified with a buffer amplifier whose output is inductively coupled to the input coil L of the modulator unit. Two tubes \underline{T}_1 and \underline{T}_2 are employed in the circuit. Both the r-f voltages and the audio voltages are applied in push-pull, the r-f voltages to the control grids and the audio voltages to the screen grids. A balanced modulator is created by combining the two outputs in parallel in the plate circuits; the carrier is eliminated and only the sidebands are obtained. The sidebands are shifted 90 deg in phase by having the carrier input to the modulator tubes shifted 90 deg through the two similar resistor-capacitor networks C1R1 and $C_{2}R_{2}$. The reactances of the capacitors C_{1} and C_{2} are very much greater than the value of the resistors \underline{R}_1 and \underline{R}_2 ; hence, the voltages fed into the two tubes are 90 deg out of phase with the voltage across the inductance L. The common cathode connections of the two tubes are connected to the junction of the two resistors through a bias resistor and bypass capacitor. The grids, being connected to

the extremities of the resistors, are thus excited in push-pull by voltages which have been shifted 90 deg with respect to the voltage across the inductance \underline{L} ; consequently, the generated sidebands at the output of the modulator are shifted 90 deg.

The carrier appearing across the inductance \underline{L} is now combined with the sidebands at the output in such a manner as to create two separate outputs with opposite deviations. The carrier-frequency voltage is led through resistors \underline{R}_3 and \underline{R}_4 from the input inductance \underline{L} to the output inductance \underline{L}_4 across the terminals A and B (the terminals of the combined input circuit of the two channels). By tuning this circuit to resonance with the parallel capacitors \underline{C}_4 and $\underline{C}_5\underline{C}_6$, the impedance from A to B is made purely resistive at the carrier frequency; hence, the carrier-frequency voltage appearing across the points AB is in phase with the voltage across the modulator input coil \underline{L} . The common junction of \underline{C}_5 and \underline{C}_6 (equal value capacitors) is grounded, resulting in equal carrier voltages of opposite polarity being applied across each of the separate channel outputs.

The sidebands are mixed with the carrier in the coil \underline{L}_4 . This coil, with its center tap, can be regarded as a three-terminal net-



Fig. 8-10. The equivalent circuit of the inductance \underline{L}_4 , a threeterminal network between terminals <u>A</u>, <u>B</u>, and <u>C</u>—the center tap of the inductance.

work as shown in Fig. 8-10a. Each half of the coil can be represented as an inductance of its own. It is the inductance offered by each half when the other half is open-circuited, \underline{L}_1 across channel 1, and \underline{L}_2 across channel 2. When a voltage is impressed across <u>AB</u>, the two sections in series, the mutual inductance (indicated by <u>M</u> with its double arrow in the diagram) causes the inductance of the entire coil to increase by 2<u>M</u>, <u>M</u> being the effect of \underline{L}_1 on \underline{L}_2 and also the effect of \underline{L}_2 on \underline{L}_1 . The total inductance between <u>AB</u> is thus $\underline{L}_1 + \underline{L}_2 + 2\underline{M}$. The equivalent circuit is shown in Fig. 8-10b. Since the inductance between the center tap and each of the extremities does not contain the factor \underline{M} , an inductance of $-\underline{M}$ is shown in the centertap lead of the equivalent circuit. (It is perfectly legitimate to use a minus value inductance in calculations in equivalent circuits but, of course, a true minus inductance cannot exist as an independent device.)

With this equivalent circuit for the inductance \underline{L}_4 , we can now draw a simplified circuit of the mixer network where the sidebands



Fig. 8-11. The equivalent circuit of the mixer network where the sidebands are mixed with the carrier. The capacitor \underline{C}_4 is shown by dotted line because no sideband current flows through it.

are combined with the carrier (Fig. 8-11). The inductances and capacitances of the original circuit are shown horizontally to better picture the operation of the circuit. Thus, we have for the sideband tuned circuit the capacitor C3 in parallel with a series circuit made up of L3, -M, and an element consisting of the two halves of the center-tapped inductance with their respective tuning capacitors. When we compare this equivalent circuit with the circuit shown in Fig. 8-9, we can now see how this inductance L4 fits into the circuit. Since the two halves are similar, the points A and B are at the same potential in relation to the sidebands. As a point of interest, since the capacitor C4 is connected across A and B, no sideband current flows through this capacitor. C3 is used to resonate the circuit.

Thus, the two points <u>A</u> and <u>B</u> are at the same potential with respect to the sidebands or, expressing it another way, the sidebands are in phase at <u>A</u> with respect to <u>B</u>. But, referring to Fig. 8-9, we find that the carrier is 180 deg out of phase at <u>A</u> with respect to <u>B</u>. Consequently, because of the reversal of phase of the carrier but not of the sidebands, the deviation produced for one channel is of the opposite sign to that produced for the other channel, since, when one is adding, the other is subtracting. This dual-channel modulator thus does the work of two Armstrong phase modulators placed back to back and producing deviations of opposite signs.

8-6. Dual-Channel Transmitter.² The dual-channel transmitter is similar to the two-channel transmitter except that the dual-channel

modulator is employed, with its resultant efficiency of operation. In Fig. 8-12 is shown a block diagram giving the detailed construction of a dual-channel modulator and multiplier unit. This modulator



Fig. 8-12. A detailed block diagram of the dual-channel modulator and multiplier unit showing the various tubes used for each of the stages.

follows the pattern indicated in Fig. 8-8. Thirty watts of power at anywhere from 88 to 108 megacycles is available at the output of the unit. The center-frequency stability limits the drift to one fifth the permissible amount of $\pm 2,000$ cps from the assigned frequency. Frequency-modulation noise level on the carrier within the band of 50 to 15,000 cps is more than 70 db below 100 per cent modulation. Amplitude modulation on one of the transmitters of this design was measured to be 60 db below the 100 per cent modulation.

Like all other f-m broadcast transmitters, the dual-channel transmitter also takes advantage of the class C amplifier which is



Fig. 8-13. The power output stage for a 1-kw output employing two 4X-500A's in push-pull.

used in the power-amplifier stages. In Fig. 8-13 is shown the poweramplifier stage of one of the transmitters using a dual-channel modulator. Inasmuch as the frequencies to be amplified are in the neighborhood of 100 megacycles, transmission-line tuning is used throughout. The 30 w of power from the modulator is coupled from the dualchannel modulator and multiplier unit into the grid transmission lines, the input circuit of two 4X-500A tubes in push-pull. This circuit is tuned by means of a movable shorting bar. The output circuit is also a tuned transmission line, but this time the lines are made large to carry the power and also to conduct the cooling air to the plates of the tubes. A small amount of power (2w) is taken off for use at the monitor, and the remainder is coupled into a transmission-line conversion and matching transformer. The transformer matches the output into a 7/8-in diameter single-concentric antenna transmission line. This power amplifier has a plate efficiency of 70 per cent under normal operating conditions and produces 1,000 w of usable power.

A front view of the 1-kw dual-channel transmitter is shown in Fig.



Fig. 8-14. A front view, with the doors open, of a 1-kw dual-channel broadcast transmitter showing the carefully isolated power-amplifier chamber.

8-14. The front doors are open to show the interior of the poweramplifier chamber and the modulator final multiplier section. Meters are in an inclined position for convenient reading by the operator. The power-amplifier chamber is fully isolated from the other circuits. Between the power-amplifier lines the antenna conversion transformer, also called the "bazooka," may be seen. In the lower right-
hand part of the power-amplifier chamber is the high-frequency vacuum tube voltmeter for antenna-line r-f output indication. The vertical panel at the right side contains all the tuning controls, including a crystal vernier for adjusting the transmitter to center frequency. The large wheel in this panel is a variable-transformer control for all filaments.

A right-side view of the same transmitter with doors and panels



Fig. 8-15. A right-side view, with doors open and panels removed, of a 1-kw dualchannel broadcast transmitter showing the constructional details.

removed is shown in Fig. 8-15. The power supply and the power controls are located in the right side on vertical chassis, which are hinged, providing complete accessibility of components and chassis

wiring. This view also shows the main transmitter wiring, which runs entirely through vertical and horizontal ducts, allowing ease in repair or replacement of connections in the equipment.

For other power outputs, only the power-amplifier stage has to be changed. To obtain an output of 250 w the output of the dualchannel modulator is fed into a pair of 4-125A's. This transmitter can be converted into a 3-kw transmitter by feeding the output from the pair of 4-125A's into a pair of WL478's. As in all f-m transmitters, it is possible to pyramid the transmitter into a higherpowered one by adding class C stages.



Fig. 8-16. A block diagram of an exciter unit employing the phasitron-type phase-tofrequency modulator.

8-7. Phasitron Transmitters. In Fig. 8-16 is shown the block diagram of a phasitron transmitter employing the phasitron GL-2H21 as the phase-to-frequency modulator. In this circuit a frequency deviation of ± 175 cps at 230 kc is obtained at the phasitron. A straight multiplication of 432 produces a frequency deviation of ± 75 kc at a carrier frequency of 99.36 megacycles. The crystal-oscillator frequency is adjustable between 203 and 250 kc so that carriers between 87.7 and 108 megacycles are available. The circuit is quite normal except for the modulator, and any class C amplifiers can be used at the output for power amplification.

For use in the phasitron, the single-phase crystal-controlled r-f input voltage has to be converted into a three-phase equal-amplitude actuating voltage. This may be accomplished by the simple circuit shown in Fig. 8-17. The single-phase voltage from the crystal oscillator is fed into the primary of a tapped-secondary transformer. The voltage from O to \underline{B} is, therefore, 180 deg out of phase with the voltage from O to \underline{A} . This is indicated by the + and - signs on the transformer. The object of the circuit is to transform these two voltages into three voltages equally spaced in phase around the 360 deg of a complete cycle. The action of the circuit is best demonstrated by the phasor diagram shown in Fig. 8-18. Here, the voltages OA and OB are drawn back to back, 180 deg out of phase. Referring now to both the circuit



Fig. 8-17. A simple circuit for transforming a single-phase input into a three-phase output voltage for use with the phasitron.

shown in Fig. 8-17 and the phasor diagram, we see that \underline{V}_1 will be in phase with the voltage \underline{OA} , inasmuch as the voltage across a portion of a resistor is in phase with the voltage across the whole



Fig. 8-18. A phasor diagram showing the action of the one-to three-phase conversion circuit depicted in Fig. 8-17.

resistor. Hence, \underline{V}_1 is shown superimposed on the voltage \underline{OA} in the phasor diagram. The purpose of \underline{R}_1 is merely to adjust the magnitude of \underline{V}_1 . To obtain \underline{V}_2 , which lags \underline{V}_1 by 120 deg, the series combination of \underline{R}_2 and \underline{C}_2 across the voltage OB is employed. By making the reactance of \underline{C}_2 equal to 1.73 times the magnitude of the resistor \underline{R}_2 , the current through the series combination will lead the voltage across it by 60 deg. The voltage \underline{V}_2 , being the voltage drop across the resistor \underline{R}_2 , will be in phase with the current. It will lead the voltage \underline{OB} by 60 deg. This is depicted in the phasor diagram by drawing \underline{V}_2 60 deg counterclockwise from the voltage \underline{OB} .

The series combination of \underline{R}_3 and \underline{C}_3 is used to obtain the voltage \underline{V}_3 , which leads the voltage \underline{V}_1 by 120 deg. The resistance of \underline{R}_3 is made 1.73 times the magnitude of the reactance of \underline{C}_3 . Hence, the

current through the series combination will lead the voltage across it by 30 deg. However, the voltage across the condenser always lags the current through it by 90 deg. Consequently, the voltage across C_3 will lag 90 - 30 or 60 deg behind the voltage OB. This is illustrated in the phasor diagram by drawing V_3 60 deg clockwise from the voltage OB. The net result is three voltages equally spaced 120 deg apart. By choosing the proper values for the components, it is possible to obtain three equal-magnitude voltages.

In the diagram of Fig. 8-17, a ground \underline{G} is shown at \underline{O} . In a balanced three-phase system no current would flow through to this ground return. It is advantageous to omit it in the actual model, since omitting it tends to force balance in the circuit.

There are other acceptable methods of converting a single-phase voltage into a three-phase voltage. One method, which has often been employed, is to convert the single-phase input into a two-phase voltage by means of a resistor-capacitor network and then use the well-known (to power engineers) Scott-connected transformer system.³ Any system which yields a fairly well-balanced three-phase output may be employed for the conversion unit.

A circuit diagram for the GL-2H21 phasitron modulator is shown in Fig. 8-19. The single-phase 203- to 250-kc crystal-controlled



Fig. 8-19. A circuit diagram for the GL-2H21 phasitron modulator. The audio voltage is applied to a coil \underline{L} surrounding the tube.

frequency, 1/432 of the assigned carrier frequency, is converted to three phases by means of the resistance-capacitor network. These voltages are impressed on the proper terminals of the phasitron and the audio voltage is fed into the coil surrounding the tube proper. Between plate 2 and B+ is placed the output tuned circuit. Since the phasitron is modulated by means of a magnetic field, any stray or extraneous field will cause noise modulation of the carrier. To prevent this noise modulation, the tube and its modulating coil are carefully shielded by a magnetic shield of a high-permeability material and, as a further precaution, a small selenium rectifier and filter is provided to filter the d-c filament supply.

Two push-pull amplifier stages consisting of twin triodes, a 6SL7 followed by a 6SN7, are used in the audio system. A pre-emphasis network of 75 μ sec is used at the input, and the audio output is fed into the phasitron modulating coil through an output transformer. This modulating, or field, coil is designed to present, as closely as possible, a purely inductive reactance to the voltage impressed across it. For high-fidelity purposes, it should appear so over the entire a-f band from 50 to 15,000 cps. As discussed previously, if this is done, the magnetic field in the phasitron will be inversely proportional to the modulating frequency, thus converting the output of the phasitron from phase to frequency modulation. To compensate for any resistance of the field coil, a feedback circuit having a frequency characteristic which slightly reduces the feedback, thereby increasing the gain, at the lower frequencies may be used in the audio amplifier. This is done by using in the feedback circuit an RC network having the same time constant as the field coils being used - the time constant being the product of the resistance and the capacitance used in the circuit.

The r-f circuits following the modulator consists of two doublers, two triplers, a doubler, and finally a tripler, as shown in Fig. 8-16. The tubes are all 6SJ7 tubes except for the last doubler, which is a 6V6, and the final tripler, which is an 815. The interstage couplings are designed to have the proper band width at their respective frequencies with high attenuation outside the band.

A 250-w output may be obtained by adding two class C amplifiers - an 829B push-pull driver, followed by two Eimac 4-250-A tetrodes in a push-pull circuit as a power amplifier. This 250-w transmitter is used as the basic unit for obtaining higher power outputs. These higher ratings are obtained, as in the other transmitters discussed, by using the 250-w transmitter to drive other class C stages of power amplification.

8-8. Cascade Phase-Shift Transmitter. The ordinary phaseshift circuits employing standard components, as discussed previously, may be employed in a phase-to-frequency modulation transmitter.⁴ Figure 8-20 shows a block diagram of an f-m transmitter employing a phase-shift modulator. A crystal oscillator at 100 kc is shifted \pm 77.16 cps by six cascaded phase shifters. The modulating voltage is obtained from an a-f input that has had incorporated into it the standard 75 μ sec pre-emphasis, as well as the necessary frequency correction for the generation of frequency modulation from phase modulation. This latter correction is accomplished in the section marked "inverse-frequency network." Any of the power amplifiers discussed previously may be used at the output to obtain the desired power output.

The amount of phase shift obtainable from a single phase-shift stage is not very great; however, it is possible to increase the useful phase shift of a modulator by adding together the individual phase

173

shifts of two or more stages connected in cascade. The block diagram of Fig. 8-21 illustrates the manner of connecting the phase-shift



Fig. 8-20. A block diagram of an f-m broadcast transmitter employing a cascade phase-shift modulating circuit.

stages in cascade for the transfer of r-f energy, but in parallel as far as the audio signal is concerned. The progressive increase



Fig. 8-21. A block diagram showing the individual stages of a cascade phase-shift modulator, each stage shifting the phase sufficiently to obtain a frequency deviation of ± 12.5 cps.

of frequency deviation from stage to stage, an increase of 12.5 cps, is also shown.

The design of a cascade phase-shift modulator is shown in the schematic diagram of Fig. 8-22. The phase shifters employ 6SJ7 tubes with a single section of a 6SN7GT tube as the resistance tube. Frequency multipliers with an over-all multiplication factor of 12 are employed in the modulator proper. The final multiplier, not shown, provides a multiplication of 81, placing over-all multiplication at 972. A built-in tuning meter enables the operator to make adjustments without the aid of external measuring equipment.

8-9. Transmitter-Installation Layout. The actual installation of an f-m transmitter is not very much different from its counterpart, the a-m transmitter. This is true whether a direct or phase-tofrequency f-m transmitter is used. Figure 8-23 shows a proposed layout for a 10-kw f-m transmitter using the ground floor of a specially built building. A four-frame transmitter is shown being mounted in a wall that separates the operator from the rear of the transmitter. The operator is seated at the console, near which is shown the monitoring speaker and two turntables for recordings. Also



뇌

1

Ż

Fig. 8-: ploying six 22 > stages schematic diagram tages of cascaded <u>e</u> phase-shift ىو modulator modulation. for an -÷. 3 broadcast transmitter em 1

flush used. culation through the transmitter 110-v power with service is Cool air circuits ىو checkered plate must is not available, mounted be provided ť . The ىم the transformer, cabinets. in rear panel the rear of the board For this compartment for also shown, separating showing purpose, an exthe has wall. various 5 cirbe Ħ





All of the available

to the operator is

ىم

test-equipment bench for routine

trench below the floor

and

covered checks

S

enclosed

İn

ø

haust fan capable of removing at least 1,000 cu ft of air per minute is shown mounted in the top rear of the building.



Fig. 8-24. A proposed installation-unit layout for a 50-kw f-m broadcast transmitter.

The layout a 50-kw transmitter is shown in Fig. 8-24. Logically, the layout and building are more elaborate. It is supposed that

FREQUENCY MODULATION

a water cooling system is used, so a room is provided for the water cooler and water pumps. Spares should be provided to permit continuous operation in the case of any failure of the cooler system. The high-voltage rectifier and the rear of the transmitter are accessible only through access doors having automatic interlocks for personnel safety. A separate room with its own interlocked access door is provided for the power transformers and filter reactors. The same type of trench construction can be used here as was employed for the 10-kw unit; the layout, of course, is much more complex. Exhaust fans are provided for air circulation in all of the rooms.

REFERENCES

1. J. H. Martin, "Experimental 88 to 108-mc 250 Watt F-M Broadcast Transmitter," <u>Communications</u>, September, 1946, p.22.

2. Frank A. Gunther, "REL F-M Broadcast Transmitters," <u>FM and Television</u>, March, 1946.

3. For a discussion of the Scott, or T, connection See C. E. Magnusson, <u>Alternating Currents</u>, New York, McGraw-Hill Book Company, Inc., 1931, pp. 215-217.

4. M. Marks, "Cascade Phase Shift Modulator," <u>Electronics</u>, December, 1946, p. 104.

QUESTIONS

1. Explain the operation of the balanced modulator.

2. Discuss three ways of obtaining a 90-deg phase shift for use in the Armstrong phase-to-frequency modulator.

3. Why is frequency-deviation multiplication necessary in the Armstrong method of phase-to-frequency modulation? Explain.

4. Why does the heterodyning method of reducing the frequency of an f-m signal (a converter) not reduce the frequency deviation?

5. Explair the method of straightforward single-channel frequency-deviation multiplication employing a converter tube.

6. What is the disadvantage of single-channel multiplication when used with the Armstrong phase-to-frequency modulator?

7. Explain the two-channel method of frequency-deviation multiplication.

8. What basic advantage is there in the two-channel method?

9. Draw a block diagram of the two-channel transmitter where each of the phase-to-frequency modulators produces a frequency deviation of ± 15 cps for a ± 75 -kc deviation at 90 megacycles.

10. Choose a resistor and capacitor combination which may be used for a $75\,\mu\,\mathrm{sec}$ pre-emphasis network.

11. Describe the operation of a phase-to-frequency modulator where the two tubes of a balanced-modulator circuit are fed with the carrier signal in quadrature.

12. What are the advantages of the dual-channel modulator?

178

13. Describe the operation of the dual-channel modulator.

14. Explain in detail the operation of the combining network, in the dual-channel modulator, where the carrier is combined with the sidebands.

15. Drawa block diagram of a dual-channel transmitter where the modulator produces a frequency deviation of 12 cps for a \pm 75-kc deviation at 108 megacycles.

16. Describe a circuit for obtaining a three-phase voltage from a single-phase source.

Explain the operation of the phasitron modulator circuit.
Why is magnetic shielding so important in the phasitron modulator?
Describe the operation of the cascade phase-shift oscillator.

20. Describe a possible transmitter layout for the 1-kw transmitter shown in Fig. 8-15.

CHAPTER 9

Frequency-Modulation Detectors

9-1. Introduction to the Frequency-Modulation Receiver. The function of the f-m receiver, like that of the familiar a-m receiver, is to select the desired signal from a crowded ether, amplify the signal, detect it, and reproduce the intelligence carried upon it. Basic-



Fig. 9-1. A block diagram of an f-m receiver, the f-m detector being responsive to frequency modulation only and incorporating a limiter if necessary.

ally, however, it differs in many ways. Figure 9-1 shows a block diagram of an f-m receiver. Except for the detector, it resembles an ordinary superheterodyne receiver. It contains an r-f amplifier section, a converter, an intermediate-frequency (i-f) amplifier section, a detector, and an audio-amplifier section.

Let us start at the antenna and mention briefly some of the design considerations for the various components. The frequencies used in f-m transmission are usually very high, very often in the ultrahigh-frequency (u-h-f) region. As mentioned previously, the f-m broadcast band is in the 100-megacycle band. At these frequencies an antenna consisting of a wire stretched between a couple of glass insulators and connected to the receiver by an ordinary lead-in wire usually cannot be used where field strengths are nominal. Special attention has to be given to the design of the antenna, the polarization being used, and the transmission line used between antenna and receiver. The frequencies being used also make the choice of tubes

WRH

for the r-f amplifiers, the converter, and the oscillator an important consideration. The circuit design must again follow the principles used in the u-h-f region, avoiding frequency distortion of the signal rather than amplitude distortion, as in the case of a-m receivers.

The i-f amplifiers are usually centered at a comparatively high frequency for that type of circuit; 10.7 megacycles is a common i-f frequency. With heterodyne conversion, the frequency deviation in f-m broadcast reception remains \pm 75 kc; the band width of the i-f stages has to be sufficient to pass all of the important sidebands to avoid distortion. It means a band width of about 165 kc (as discussed in Chap. 2). Where the broadcast stations are spaced 200 kc apart, the i-f tuned stages have to reject all signals beyond \pm 100 kc away from the carrier. Again care has to be taken to avoid frequency distortion rather than amplitude distortion.

The f-m detector is the radically different circuit in the receiver. The first requisite is that it be as insensitive as possible to any amplitude modulation. This may be accomplished by the use of a limiter stage preceding the f-m detector. This limiter stage strives to remove all of the amplitude modulation which may have been picked up by the signal, including any a-m noise or interference signal. The insensitivity to amplitude modulation may be accomplished by special design of the detector or previous (limiter) stages. A number of such circuits are available. The detector proper, of course, must linearly translate the f-m signal into the original intelligence, or voice signal, for amplification by the audio stages and reproduction by the speaker. For high-fidelity broadcast reception, the audio stages have to be wide band, from about 50 to 15,000 cps. Similarly, the speaker or speakers have to be carefully designed so that the fidelity is not lost upon reproduction.

All in all, the f-m receiver has to be carefully conceived, carefully designed, and carefully built to meet the stringent requirements of high-frequency wide-band high-fidelity broadcasting, in addition to using new and novel circuits for wide-band frequency modulation. Each of the basic components will be discussed separately, with specific application to the f-m receiver. Because many of the design considerations of the preceding components in the f-m receiver depend upon the characteristics of the detector employed, it is advantageous to discuss detector-circuit design first.

9-2. Single Tuned-Circuit Detector. One of the basic principles used in f-m detection is to convert the f-m signal into an a-m signal and then employ an ordinary a-m detector. In its simplest form, the conversion circuit consists of a single tuned circuit; a diagram of the circuit is shown in Fig. 9-2. The f-m signal is fed into a resonant circuit consisting of an inductance \underline{L} in parallel with a variable capacitance \underline{C} . The voltage across the resonant circuit is fed into a diode detector circuit and the output voltage appears across the load re-

sistor $\underline{R}_{\underline{L}}$. The circuit is similar to a straightforward a-m detector, except for the tuning of the resonant circuit.



Fig. 9-2. A single tuned-circuit f-m detector using a detuned resonant circuit and a straightforward diode detector.

Instead of being tuned to the center frequency of the f-m signal, the resonant circuit is detuned slightly. In this way we are actually operating on the slope of the resonant circuit characteristic, as il-



Fig. 9-3. The characteristic curve of a resonant circuit being used in an f-m detector circuit. This is a generalized curve given in terms of the Q and the resonant frequency of the circuit.

lustrated in Fig. 9-3. This shows the characteristic curve in terms of the resonant frequency <u>f</u> and the <u>Q</u> of the circuit, the <u>Q</u> being equal to the inductive reactance $2\pi \underline{fL}$ divided by the effective series resistance of the tuned circuit <u>R</u>.¹ The voltage-amplitude output, the voltage impressed across the diode, is plotted against the frequency of

the input signal in cycles above or below resonance expressed in terms of $\underline{f/Q}$. For instance, if the resonant frequency of the circuit is 10 megacycles and the \underline{Q} of the circuit is equal to 100, $\underline{f/Q}$ will represent 100 kc, $\underline{2f/Q}$ 200 kc, and so on. The curve varies very little between a \underline{Q} of 25 and a \underline{Q} of infinity as demonstrated by the slight variation of the curve of \underline{Q} equal to 25 from the curve of \underline{Q} equal to infinity.

The object of the resonant circuit is to convert the f-m signal into an a-m signal containing the same intelligence. This is accomplished by using that portion of the characteristic which is linear. By superimposing a straight line on the resonant characteristic, it is found to be virtually linear from 0.15f/Q to 0.55f/Q, the center of the linear portion being located at 0.35f/Q. When the carrier of the f-m signal is located at 0.35f/Q cps below the resonant frequency, as the frequency of the f-m signal increases, the amplitude of the output voltage will increase linearly, and as the frequency decreases, the output voltage will decrease linearly, thus converting the f-m signal into an a-m signal which may be detected by the a-m diode detector. If the f-m signal should shift off the linear portion of the characteristic either by an increase in the frequency deviation beyond the bounds of the linear portion or by a shift in the frequency relationship between the resonant frequency and the f-m carrier frequency (the f-m carrier being no longer in the center of the linear portion), distortion of the signal will be evident in the output.

The action of the circuit may be better understood by examining



.

Fig. 9-4. Three oscillograms showing the conversion of an f-m signal to an a-m signal by means of a detuned resonant circuit and the final output voltage.

the three oscillograms shown in Fig. 9-4. The pure f-m signal, with amplitude modulation previously removed by a limiter, is shown in Fig. 9-4a. This is the input signal to the resonant circuit. The

resonant circuit is now tuned so that the range of frequencies in the f-m signal falls on the straight line portion of the resonant curve. The portion on either side of resonance may be used. Suppose we choose the portion below resonance, as shown in Fig. 9-3. The output voltage, the voltage across the resonant circuit, will increase as the frequency of the f-m signal increases, and it will decrease as the frequency of the f-m signal decreases. The output voltage from the detuned resonant circuit detector is shown in Fig. 9-4b. We see now that the r-f signal has been amplitude-modulated in accordance with the variation in frequency. This is the signal which is impressed across the diode for amplitude detection. Upon detection by the diode, we obtain the audio signal shown in Fig. 9-4c.

EXAMPLE. Determine the resonant frequency and the \underline{Q} of a resonant circuit to be used for the detection of a 10-megacycle f-m signal with a frequency deviation of ± 75 kc.

From Fig. 9-3 we see that to obtain from an f-m signal the maximum amount of amplitude variation without distortion, the frequency should vary from about $0.15\underline{f/Q}$ to $0.55\underline{f/Q}$. Hence, the difference, $0.40\underline{f/Q}$, is equated to twice the frequency deviation - the total frequency excursion.

$$0.40 \frac{f}{Q} = 150,000$$

or

 $\underline{Q} = \frac{\underline{f}}{375,000}$

Since the resonant frequency will be quite close to 10 megacycles, we can use 10 megacycles for f in the calculation of Q. (The true value of the resonant frequency is greater than 10 megacycles, as seen from the diagram.) Substituting 10 megacycles for \underline{f} and solving for Q, we obtain

The true resonant frequency of the tuned circuit will be equal to 10 megacycles plus $0.35\underline{f}/\underline{Q}$ cps. Again using 10 megacycles for \underline{f} we obtain

Resonant frequency =
$$10,000,000 + 0.35 = \frac{10,000,000}{28.7}$$

Resonant frequency = 10,131,000 cps

Recalculating the results will show that using 10 megacycles instead of 10.131 megacycles for <u>f</u> does not introduce any appreciable error. Thus, for a single tuned-circuit f-m detector, we can use a resonant circuit tuned to 10.131 megacycles and having a Q of 26.7.

Returning now to Fig. 9-4, we note that the amount of amplitude modulation superimposed on the f-m signal is quite small; it is actually only about 18 per cent amplitude-modulated, close to the maximum amount of amplitude modulation that can be obtained with this system. This is both inefficient and detrimental to the operation of the f-m system. The effect of the f-m signal has been cut down to about one fifth, while any extraneous interfering a-m signal will come through with full force. We can readily understand why this system of single tuned-circuit f-m detection is so very seldom employed.

9-3. Detector Employing Two Tuned Circuits. A much improved method of f-m detection employs two tuned circuits, one displaced in frequency above the f-m carrier, and the other below the f-m carrier.



Fig. 9-5. A double tuned-circuit type of f-m detector where one tuned circuit is tuned to a frequency slightly above the carrier frequency and the other tuned circuit is tuned to a frequency slightly below the carrier frequency.

A circuit diagram of the detector is shown in Fig. 9-5. The r-f input, from the preceding limiter stages, is fed into a tuned circuit consisting of \underline{L}_p and \underline{C}_p , the primary resonant circuit. This circuit is tuned to the carrier frequency of the f-m signal. Coupled to this primary circuit, but not coupled to one another, are two tuned circuits. One circuit, consisting of \underline{L}_1 and \underline{C}_1 , is tuned to a frequency slightly above the f-m carrier frequency, while the other, consisting of \underline{L}_2 and \underline{C}_2 , is tuned to a frequency slightly below the f-m carrier frequency. The outputs of the two tuned circuits are rectified, or detected, individually by separate diodes. Since the load resistors \underline{R}_1 and \underline{R}_2 are arranged back to back, the audio signal output is the difference between the two rectified tuned-circuit voltages.

In Fig. 9-6 is shown the method of determining the over-all detection characteristic of the double tuned-circuit detector. The voltage across \underline{R}_1 in the circuit diagram is determined by the tuned circuit $\underline{L}_1\underline{C}_1$; it will be a resonance curve whose peak is displaced in frequency above the f-m carrier frequency as shown. It is shown positive because, referring to Fig. 9-5, we see that the d-c current will always flow down resistor \underline{R}_1 to ground — being determined by the polarity of rectification of the diode. The voltage across \underline{R}_2 , a resonance curve whose peak is displaced in frequency below that of the f-m carrier, is shown negative because, again referring to Fig. 9-5, the d-c current in this resistor always flows up — away from ground.

With no frequency modulation, the voltage across \underline{R}_1 will can-

cel the voltage across \underline{R}_2 , inasmuch as the resonant curves are symmetrically displaced from the f-m carrier. In this manner any residual amplitude modulation that has remained on the signal will tend to cancel. As soon as any frequency variation (frequency modulation) takes place, the voltage across the resistors will vary. If the frequency increases, the positive voltage across \underline{R}_1 will increase while the negative voltage across \underline{R}_2 will decrease — resulting in a positive-voltage output. When the frequency decreases, the opposite takes place - resulting in a negative-voltage output. The over-all f-m detection characteristic is thus the sum of the two characteristics, one positive and the other negative. This is shown



Fig. 9-6. A diagram illustrating a method of determining the f-m detection characteristic of a double-tuned f-m detector.

as the over-all detection characteristic in Fig. 9-6. When this characteristic is compared with the small amount of linear characteristic in the single tuned-circuit characteristic (Fig. 9-3), the much greater sensitivity of this circuit can be better appreciated.

The advantages of this type of f-m detector are greater sensitivity, reduction in the response to amplitude modulation, and, because of the greater expanse of linearity, greater ease in tuning the receiver. The main difficulty lies in adjusting the individual tuned circuits to different frequencies and adjusting properly the coupling of the circuits.

The proper operation of the two-tuned-circuit detector is dependent on two factors: the Q's of the three tuned circuits and the displacement of the resonant frequencies of $\underline{L1C1}$ and $\underline{L2C2}$. The Q's of the secondary circuits,² the two displaced in frequency, are given by Q_s where

$$\underline{Q}_{\underline{S}} = \frac{2}{3} \times \frac{f - m \text{ carrier frequency}}{\text{frequency deviation}}$$
(9-1)

The \underline{Q} of the input primary circuit \underline{Q}_p is given by

$$\underline{Q}_{\underline{p}} = \frac{1}{2} \underline{Q}_{\underline{s}}$$
(9-2)

A condition for practical linearity of the detection characteristic is that the separation between peaks of the secondary circuits be about 2 1/2 to 3 times the frequency deviation. For best results, however, the process shown in Fig. 9-6 may be followed. Two similar resonant curves for the Q_S given by Eq. 9-1 should be plotted, one above the axis and the other below, as shown. The over-all detection characteristic, obtained by taking the sum of the two resonant curves, is determined for different displacements of the peaks. The displacement giving the maximum linearity in the over-all detection characteristic is the displacement to use.

EXAMPLE. Determine the three \underline{Q} 's of the tuned circuits in an f-m detector employing a primary tuned-input circuit and two frequency-displaced secondary resonant circuits. The f-m signal to be detected has a frequency deviation of 75 kc about a carrier frequency of 5 megacycles.

To determine $Q_{\underline{S}}$, the Q of the secondary coils, we substitute into Eq. 9-1.

$$\underline{Q_{s}} = \frac{2}{3} \times \frac{5,000,000}{75,000}$$
$$\underline{Q_{s}} = 44 \quad \underline{Ans}.$$

And from Eq. 9-2

$$\underline{Q}_{\underline{p}} = \frac{1}{2} \times \underline{Q}_{\underline{S}} = 22 \quad \underline{Ans}.$$

9-4. The Frequency-Modulation Discriminator. The f-m discriminator operates on another principle: the phase shift between



Fig. 9-7. The phase relationships between the voltage across the primary and the voltage across the secondary of two resonant loosely coupled circuits, both resonant at the same frequency.

coupled resonant circuits. This is illustrated in Fig. 9-7. Figure 9-7<u>a</u> shows two resonant circuits, the primary, consisting of <u>Lp</u> and <u>Cp</u>, coupled with a mutual inductance of <u>M</u> to the secondary, con-

sisting of $\underline{L}_{\underline{S}}$ and $\underline{C}_{\underline{S}}$. Both are resonant at the same frequency $\underline{f}_{\underline{r}}$. $\underline{E}_{\underline{1}}$ is the input voltage across the primary and has a frequency which we will call \underline{f} . $\underline{E}_{\underline{2}}$ is the output voltage across the secondary.

Figure 9-7b shows the phase relationship between the two voltages \underline{E}_1 and \underline{E}_2 when the input frequency \underline{f} is equal to the resonant frequency \underline{f}_r . \underline{E}_2 leads \underline{E}_1 by 90 deg. When the input frequency \underline{f} falls below the resonant frequency, the secondary voltage begins to lead by more than 90 deg, as shown in Fig. 9-7c. Within the operating limits of the discriminator, the lower the input frequency, the greater will be the angle by which \underline{E}_2 leads \underline{E}_1 . When the input frequency increases above the resonant frequency, the phase angle between \underline{E}_2 and \underline{E}_1 begins to fall below 90 deg. This is depicted in Fig. 9-7d. Again, within the operating limits of the discriminator, the higher the input frequency, the smaller will be the angle between \underline{E}_2 and \underline{E}_1 . Since the phase variation takes place in the neighborhood of resonance (around the peak of the resonant curve), we have a relatively constant voltage which varies in phase with the input frequency.

The f-m discriminator changes this phase variation into an amplitude variation in a manner similar to that employed in the phase-



Fig. 9-8. A circuit diagram of the f-m discriminator where both primary and secondary tuned circuits are resonant at the same frequency.

discriminator control circuit discussed in Sec. 5-3. Figure 9-8 shows a circuit diagram of the f-m discriminator. The primary circuit, consisting of \underline{L}_p and \underline{C}_p , and the secondary circuit, consisting of a center-tapped inductance \underline{L}_s and \underline{C}_s , are both tuned to resonance at the f-m carrier frequency. The secondary voltage is split into two equal parts, \underline{E}_1 and \underline{E}_2 . Connected to the center tap is a coupling capacitor \underline{C}_c which also connects to the primary circuit, as shown. The r-f choke acts as a d-c return and prevents shorting the center tap to ground. Two diodes are utilized for detection, one in each leg of the secondary circuit.

Let us now determine what this circuit accomplishes. Impressed across diode I with its attendant load resistor \underline{R}_1 is the sum of the

primary voltage plus one half the secondary voltage. The polarity of the diode is so arranged that the output is a varying positive voltage, the current always flowing toward ground. Impressed across diode II with its load resistor \underline{R}_2 is the sum of the primary voltage plus the second half of the secondary voltage \underline{E}_2 . We see here that \underline{E}_2 is the negative of \underline{E}_1 ; when the top of the secondary coil is positive, the bottom of the coil will be negative. The rectified voltage across \underline{R}_2 being a varying negative voltage with the current in the resistor always flowing away from ground. The output voltage, the sum of the voltages across \underline{R}_1 and \underline{R}_2 , is thus the difference between the two rectified voltages.

The actual operation of the circuit can best be understood from



Fig. 9-9. The phase diagrams for the f-m discriminator showing how the output voltage is generated as the input frequency varies.

a consideration of the phasor diagrams shown in Fig. 9-9. Figure 9-9<u>a</u> shows the phasor diagram of the voltages when the input frequency \underline{f} is equal to the resonant frequency \underline{f}_r of the two tuned cir-

cuits. \underline{E}_1 leads the primary voltage \underline{E}_p by 90 deg while \underline{E}_2 , the second half of the secondary voltage which is reversed in phase, lags E_p by 90 deg. Referring now to Fig. 9-8 (the circuit diagram), we see that the r-f voltage impressed across diode I is equal to the sum of \underline{E}_p and \underline{E}_1 . This is indicated in the phasor diagram as \underline{E}_I , the sum of \underline{E}_p and \underline{E}_1 . Similarly, the r-f voltage impressed across diode II is equal to the sum of \underline{E}_p and \underline{E}_2 . This is shown in the phasor diagram as \underline{E}_{II} . The rectified voltage across resistor \underline{R}_1 is a positive voltage proportional in magnitude to E1, while the voltage across resistor R_2 is a negative voltage proportional in magnitude to E_{II} . Thus, with no modulation on the carrier, the output voltage will be zero, the voltage across \underline{R}_2 equalizing the voltage across \underline{R}_1 . Again, we see in this circuit that any residual amplitude modulation will tend to cancel out when there is not frequency modulation of the carrier. This result does not eliminate the effect of amplitude modulation completely but tends to reduce it, as in the case of the two tuned-circuit detector. The effect of amplitude modulation will be demonstrated with phasor diagrams.

Figure 9-9b shows the phasor diagram for the case where the input frequency \underline{f} is less than the resonant frequency $\underline{f_r}$ of the tuned circuits. In this case $\underline{E_1}$ will lead by more than 90 deg. $\underline{E_2}$, on the other hand, being the negative of $\underline{E_1}$, will lag by less than 90 deg. ($\underline{E_2}$ is $\underline{E_1}$ reversed.) The voltage across diode I, $\underline{E_I}$, being the sum of $\underline{E_p}$ and $\underline{E_1}$, has decreased in magnitude, while the voltage across diode II, $\underline{E_{II}}$, being the sum of $\underline{E_p}$ and $\underline{E_2}$, has increased in magnitude. Upon detection we find that the positive voltage across $\underline{R_1}$ has decreased, while the negative voltage across $\underline{R_2}$ has increased. The output voltage is thus a negative voltage equal to the difference between the two rectified voltages.

In Fig. 9-9 is shown the phasor diagram for the case where the input frequency f is greater than the resonant frequency of the circuits. Now E1 leads by less than 90 deg, while E2, the negative of E1, lags by more than 90 deg. Conditions are reversed from those shown in Fig. 9-9b. The positive voltage across R1 is greater than the negative voltage across R2 and the resultant output, the magnitude difference between the two, is therefore a positive voltage. Thus, we now have a circuit which, within its operating limits, results in a voltage output proportional to the frequency.

The effect of amplitude variation, or amplitude modulation, of the input can be seen by examining Figs. 9-9b and c. If Ep, the primary input voltage, should increase in magnitude, then both E1, and E2 will also increase proportionally. This effect is illustrated in Fig. 9-10. In Fig. 9-10a the amplitude (for the same frequency relationship as shown in Fig. 9-9a) has been decreased to one half its former value. Ep, E1, and E2 are all decreased to one half their former values so that the resultants EI and EII are also reduced to one half their previous values. Hence, the voltages across the two resistors R_1 and R_2 fall to one half the values shown in Fig. 9-9b resulting in the output voltage, the difference between the two, drop-



Fig. 9-10. Two phasor diagrams showing the effect of amplitude variation on the output of the f-m discriminator.

ping to one half its former magnitude.

In Fig. 9-10<u>b</u> the amplitude of the input signal has been greatly increased, while keeping the frequency relationships the same. Again it is shown that the amplitude of the output signal varies proportionally to the magnitude of the incoming wave.

Thus, we see that even though the amplitude modulation tends to cancel out under zero-frequency modulation, the magnitude of the detected f-m output is proportional to the magnitude of the incoming wave; in this manner it is possible for amplitude modulation to distort the output of the detector. The same effect is present in the previous f-m detector using two tuned circuits and, in both of these detectors, careful elimination of any amplitude modulation from the incoming signal is necessary. Both of these detectors employ amplitudelimiter stages for the elimination of amplitude modulation. In Fig. 9-11 is shown a characteristic curve of an f-m discriminator. It is a symmetrical curve about the carrier frequency, re-



Fig. 9-11. The characteristic curve of an f-m discriminator showing its symmetry about the carrier frequency and its peak separation.

sembling very much the two tuned-circuit characteristic. In the circuit diagram (Fig. 9-8), the resonant circuits, consisting of $\underline{L}_p\underline{C}_p$ and $\underline{L}_S\underline{C}_S$, are both tuned to the carrier frequency. The important design considerations in this case are the Q's of the circuits and the coupling, or mutual inductance, between the primary and secondary circuits. Using the symbols of Fig. 9-8, we know that by symmetry \underline{R}_1 should be equal to \underline{R}_2 . These resistors, since they load the circuit through the diode, affect the Q's of the resonant circuits.

In designing the f-m discriminator circuit, the secondary inductance should be twice the primary inductance. Assuming a perfect rectifier, we find that the effective loading of the load resistors is one half their resistive value. Hence, in determining the Q's of the primary and secondary, we have to shunt the resonant circuits with the proper effective resistances. The loading of the diode and load resistor is equivalent to shunting the whole secondary tuned circuit with a resistance equal to \underline{R}_1 and to shunting the primary circuit with a resistance equal to \underline{R}_1 and to shunting the primary circuit with a resistance equal to $\underline{R}_1/4$. The Q's of the two circuits, calculated on the basis of these shunting resistors, are quite close to the true Q's of the circuits.³ For a coupling between the primary and secondary not exceeding three quarters of the critical coupling, the separation of the two peaks, as shown in the characteristic curve of Fig. 9-11, may be approximately determined by

Separation between peaks =
$$\frac{\text{carrier frequency}}{\underline{Q}_{S}}$$
 (9-3)

where \underline{Q}_S is the \underline{Q} of the secondary circuit (the circuit consisting of $\underline{L}_S \underline{C}_S$ shunted by the effective load resistance \underline{R}_1). When the coupling is greater than three quarters of the value of the critical coupling, the peak separation is actually greater than the value given by Eq. 9-3. The discriminator characteristic is quite linear up to about three quarters of the peak-to-peak separation, but sometimes, where a high degree of linearity is desired, as little as one quarter of the separation between peaks is used for actual demodulation. In other words, the frequency deviation of the signal may vary from one eighth to three eighths of the peak-to-peak separation. A value of about one quarter is a safe value to use in the ordinary case.

Example. Determine the design data for an f-m discriminator to be used with an intermediate frequency of 5 megacycles and a frequency deviation of 75 kc.

In Fig. 9-12 is shown a circuit diagram of the f-m discriminator with nominal values of 40,000 ohms for the load resistors and $100 \mu\mu$ for the bypass capacitors. A $20 - \mu\mu$ coupling capacitor is also assumed. These are noncritical values which can be assumed from practice.

To make the frequency deviation one quarter of the peak-to-peak separation, we need a peak-to-peak separation of four times 75 kc (300 kc). Using a coupling of three quarters the critical coupling, we know that Eq. 9-3 will apply. Substituting 5 mega-cycles for the carrier frequency and 300 kc for the peak-to-peak separation, we obtain for Q_s , the Q of the secondary circuit,

$$\underline{Q}_{\underline{S}} = \frac{5,000,000}{300,000}$$

so that

$$Q_{\rm S} = 16.7$$

However, in a parallel circuit, the \underline{Q} of the circuit is equal to the parallel resistance divided by the inductive reactance, therefore,

$$\underline{Q}_{\underline{S}} = \frac{\underline{R}_1}{2\pi \underline{I} \underline{L}_{\underline{S}}}$$

where \underline{f} is the carrier frequency. This assumes that the losses within the tuned circuit are negligible in comparison with the shunting resistor.

Solving for \underline{L}_{S} , we obtain

$$\underline{\mathbf{L}}_{\underline{\mathbf{S}}} = \frac{-\underline{\mathbf{R}}_{1}}{2m\underline{\mathbf{I}}\underline{\mathbf{Q}}_{\underline{\mathbf{S}}}}$$

We can now substitute into this equation and obtain the value of \underline{L}_s .

$$\underline{L}_{\underline{S}} = \frac{40,000}{2 \times 3.14 \times 5,000,000 \times 16.7}$$

Hence,

$$\underline{L}_{s} = 76 \,\mu h$$
 Ans.

The primary reactance is half the secondary reactance or

$$\underline{L}_{D} = 38 \ \mu h \ \underline{Ans}.$$

If for any reason these values should be too high, for instance, when the stray capacitance across the coils is so large that they cannot tune to the carrier frequency, the magnitude of the inductances may be reduced by reducing the values of the load resistors. The inductances will reduce in proportion to the reduction in the value of the resistors. The capacitances to be used in the tuned circuits are those necessary to tune the circuits to resonance at 5 megacycles, in this example.

9-5. The Ratio Detector. We have shown (Fig. 9-10) how the f-m

discriminator responds to amplitude modulation as well as to frequency modulation; hence, a limiter must be employed. It would be a decided advantage to get an f-m detector which would be sensitive only to the f-m signal and would reject all amplitude modulation. This would eliminate the necessity for a limiter stage and would result in an increase in gain for the tubes and circuits used.

This final result is accomplished in the ratio detector.⁴ The ratio detector uses an r-f circuit resembling an ordinary discriminator r-f circuit, but the output, instead of being proportional to the difference between the absolute values of the voltages impressed on the diodes, is proportional only to the ratio between the magnitudes of the two voltages. We can see how the use of the ratio of the two voltages instead of the difference between them accomplishes the rejection of amplitude modulation by examining Fig. 9-10. In Fig. 9-10b the amplitude of the incoming signal has increased threefold and the output of the discriminator also has increased threefold.



Fig. 9-12. A circuit diagram of the f-m discriminator showing load resistances for calculating the values of L_s and J_p .

However, we also note that, logically, the ratio between \underline{E}_{I} and \underline{E}_{II} has not changed at all. If the output were proportional only to this ratio, it would not have varied with the amplitude of the incoming signal.

We have said that the output is proportional to the ratio between the magnitudes of \underline{E}_{I} and \underline{E}_{II} . Obviously, though, if the difference between the two voltages is proportional to the frequency, as determined by the r-f circuit, then the ratio of the two voltages will not be directly proportional to the frequency. But it was found that the magnitude of the ratio divided by <u>1 plus the ratio</u> is proportional to frequency in the same manner as is the difference between the two voltages. It is this latter relationship that is used, therefore, to determine the output of the ratio detector. Again, since only the ratio is involved and the ratio does not vary with the incoming amplitude, the output is not affected by amplitude modulation.

A simplified circuit of the ratio detector is illustrated in Fig. 9-13. First, let us compare this circuit with the circuit for the f-m discriminator as shown in Fig. 9-8. The r-f section consisting of two coupled circuits, both resonant at the carrier frequency and



Fig. 9-13. A simplified schematic diagram of the ratio detector showing its resemblance to the f-m discriminator in the r-f section and its radically different detection circuit.

coupled together with the coupling condenser $\underline{C}_{\underline{C}}$, is analogous to that used in the f-m discriminator. Hence, the voltages impressed across the diodes for this circuit resemble those obtained for the f-m discriminator. The resemblance, however, stops there, and the detection circuit is quite different. We note that both diodes are now connected with the same polarity, diode I being reversed from its connection in the f-m discriminator; instead of a pair of load resistors, a battery is connected across the two diodes; and, finally, the output is taken off across one of the capacitors instead of across the load resistors.

The effect of the battery is to keep the voltage across $\underline{C}_1, \underline{V}_1$, plus the voltage across $\underline{C}_2, \underline{V}_2$, always equal to \underline{V}_B , the voltage across the battery. Assuming no internal resistance for the battery, we find that this equality means that the sum of \underline{V}_1 and \underline{V}_2 is a constant. The battery would keep this sum constant during all the time of detection.

Let us now see how this constancy eliminates the effect of amplitude modulation and results in a linear-detected output across \underline{C}_2 . Each of the two diodes may be replaced by an effective resistance depending on the magnitude of the signal voltage impressed across it. In effect, then, the equivalent circuit of the detection section of the ratio detector is shown in Fig. 9-14. Here each of the diodes is represented by an effective variable resistance. The magnitude of each of these resistors is determined by the magnitude of the r-f voltage impressed on each of the diodes. Let us designate the effective resistance of diode I as \underline{R}_{I} and the effective resistance of diode II as \underline{R}_{II} .

The output voltage across the capacitor \underline{C}_2 will be determined by the voltage-dividing action of the effective resistances \underline{R}_I and \underline{R}_{II} . The capacitance \underline{C}_2 , being shunted across \underline{R}_{II} , will have the same voltage across it as \underline{R}_{II} . Calling the voltage output \underline{V}_O , we obtain

$$\underline{\mathbf{V}}\underline{\mathbf{O}} = \frac{\underline{\mathbf{R}}_{\mathbf{I}}}{\underline{\mathbf{R}}_{\mathbf{I}} + \underline{\mathbf{R}}_{\mathbf{I}}} \quad \underline{\mathbf{V}}\underline{\mathbf{B}}$$
(9-4)

Dividing the numerator and denominator by R_{I} , we get

$$\underline{\mathbf{V}}_{\underline{\mathbf{O}}} = \frac{\underline{\mathbf{R}}_{\mathbf{I}\mathbf{I}}/\underline{\mathbf{R}}_{\mathbf{I}}}{1 + (\underline{\mathbf{R}}_{\mathbf{I}\mathbf{I}}/\underline{\mathbf{R}}_{\mathbf{I}})} \underline{\mathbf{V}}_{\underline{\mathbf{B}}}$$
(9-5)



Fig. 9-14. An effective equivalent circuit of the ratio detector where each diode is represented by a variable resistance.

Thus, we see that the output is determined by the ratio of \underline{R}_{II} to \underline{R}_{II} . However, as we have stated before, these effective resistors are determined by the voltages \underline{E}_{II} and \underline{E}_{III} impressed across the two diodes. Consequently, the output voltage is determined by the ratio of the voltages divided by 1 plus the ratio, this relationship being correct for the detection of f-m signals.

The actual operation of the ratio detector may be demonstrated with the use of phasor diagrams. Figure 9-15 shows three such phasor diagrams, with the same conditions as shown in Fig. 9-9 for the f-m discriminator. The determination of the magnitudes of the voltages EI and EII follows the same principles as in the f-m discriminator. In Fig. 9-15a is shown the condition of the input frequency f being equal to the resonant frequency of the tuned circuits f_r . E₁ and E₂, for the condition of resonant frequency, are at right angles to \underline{E}_{p} , resulting in two equal voltages, \underline{E}_{I} and \underline{E}_{II} , being impressed across the diodes. The effective resistance of both diodes will be the same $(\underline{\mathbf{R}}_{I}=\underline{\mathbf{R}}_{II})$ so that the battery voltage $\underline{\mathbf{V}}_{\mathbf{B}}$ will divide into two equal parts, \overline{V}_1 and \underline{V}_2 . This maintains the ratio of $\underline{V}_2/\underline{V}_1$ equal to the ratio of E_{I}/E_{II} ; the output voltage V_O is V_2 . Inasmuch as this voltage is equal to one half the battery voltage, $1/2 V_B$, plus the audio output, we obtain the instantaneous audio output by subtracting one half \underline{V}_B from \underline{V}_2 . In this first case of <u>f</u> equal to \underline{f}_r , \underline{V}_2 is equal to one half \underline{V}_{B} ; the instantaneous audio output is zero $(1/2 \underline{V}_{B})$ being the d-c component).

In Fig. 9-15b is shown the phasor diagram for the condition of \underline{f}

being less than $\underline{f}_{\underline{r}}$, that is, the input signal frequency being less than the carrier frequency of the f-m input. In this case \underline{E}_1 leads by more



Fig. 9-15. Phasor diagrams of the ratio detector showing the generation of the output voltage for various frequency inputs.

than 90 deg and \underline{E}_2 lags by less than 90 deg, resulting in a voltage \underline{E}_{II} across one diode which is larger than the voltage \underline{E}_{I} across the other diode. The effective resistance of diode II will now be more than the effective resistance of diode I; hence, the voltage across diode II will be more than that across diode I. The sum of the two voltages \underline{V}_1 and \underline{V}_2 is still maintained equal to the battery voltage $\underline{V}_{\underline{B}}$, but the ratio between them has now changed to equal the ratio of \underline{E}_{I} to \underline{E}_{II} . The voltage across \underline{C}_2 , \underline{V}_2 , being more than one half $\underline{V}_{\underline{B}}$, will make the instantaneous audio output a positive voltage.

Repeating the same set of phasor diagrams for the case of the input frequency which is less than the resonant frequency, we find that the output voltage V_2 is less than one half V_B (Fig. 9-15c). This results in an instantaneous audio voltage which is negative. We also

see that as the frequency deviation increases, the ratio of the diode voltages \underline{E}_{II} and \underline{E}_{III} also increases, which, in turn, increases the voltage output. The characteristic detection curve of the ratio detector is quite similar to the curve shown for the f-m discriminator (Fig. 9-11).

Let us now consider the effect of amplitude modulation on an f-m ratio detector. As shown in Fig. 9-10, the effect of amplitude modulation is to change the magnitudes of \underline{E}_{I} and \underline{E}_{II} , but <u>not</u> to change the ratio between the two. By using a ruler on the two diagrams of Fig. 9-10 we can check this point experimentally. As shown in Fig. 9-15, however, the output voltage is determined only by the ratio and nothing else. Thus, even though amplitude modulation should change the amplitudes of the input signals, the output will be determined solely by the variation in frequency, the f-m intelligence. No amplitude limiters or other means of removing amplitude modulation from the f-m wave are necessary when the ratio detector is employed.



Fig. 9-16. A ratio detector schematic where a resistor-capacitor combination RC_3 takes the place of the battery. The time constant RC_3 should be very large - about 0.25.

Figure 9-16 shows an operating circuit where the battery is replaced by a resistor-capacitor combination <u>RC3</u>. <u>C4</u> is a very large capacitor, about $8\mu f$, and tends to keep the voltage across the resistor <u>R3</u> a constant. By keeping the time constant very large, <u>RC3</u> being equal to about 0.25, the voltage across <u>R</u> will remain constant with audio variation in signal, establishing a constant voltage across the two capacitors, <u>C1</u> and <u>C2</u>. Thus, it is the audio equivalent of the battery shown in Fig. 9-13. In one application, a capacitor of $8\mu f$ shunting a resistor of 30,000 ohms may be employed. This combination yields a time constant of 0.24. Capacitors of 0.004 μf are used for each of the two capacitors <u>C1</u> and <u>C2</u>. The output is taken off across a volume control <u>VC</u> whose total resistance is about 1 megohm. The reason for such a large value is that the loading across the capacitor <u>C2</u> must be negligible to avoid distortion. The voltage across <u>R</u> is proportional to the average value of the carrier so that is can be used as the automatic volume control (a-v-c) voltage. The voltage is referred to as constant, but actually it is only comparatively constant with respect to the rapid variations of the audio signal; however, it does follow the slow drift of changing carrier amplitude.



(b) Ratio detector audio signal

Fig. 9-17. The audio-voltage outputs of the f-m discriminator and the ratio detector when the input signal is momentarily cut off.

Another interesting difference in the operation of the f-m discriminator and the ratio detector is shown in Fig. 9-17. This figure illustrates the audio-voltage outputs of both types when the r-f signal is momentarily cut off. In the case of the f-m discriminator (Fig. 9-17a), the signal will fall to zero at those points — points 1, 2, and 3 — while in the case of the ratio detector, the signal has only a small variation from its former value. In the f-m discriminator, as soon as the signal is cut off and no current flows through the diodes, the voltage drop across the load resistors will fall to zero. But in the case of the ratio detector, the capacitor C_2 will tend to maintain its charge until the signal comes on again. Although on an oscillogram the difference in operation with momentary signal cutoff is very evident, to a person listening to the receiver, the disturbance generated by momentary cutoff is about the same in both detectors.

In many cases, the r-f choke is left out and the primary coil is connected directly between the center tap of the secondary and the junction between \underline{C}_1 and \underline{C}_2 . This is effectively the same circuit, but it means that the primary has to be isolated from the previous circuit d-c voltage by means of a coupling capacitor. The analysis and operation of the circuit is similar to the circuit discussed.

In actual operation of the circuit, as shown in Fig. 9-16, some amplitude modulation got through to the output. The output voltage varied as the amplitude of the input signal was changed from a very low value to a very high value. It was determined that harmonic distortion, introduced by i-f amplifiers, will cause a great deal of trouble. This, of course, can be removed by degenerating the second harmonic in the last i-f stage. The rest of the a-m signal can be removed by inserting a small resistance in series with the large capacitance $\underline{C3}$. This resistor allows compensating currents to flow through the diodes, causing a stepup in amplification when the signal amplitude is high, thereby straightening out the amplitude variation characteristic to zero level at the very high amplitude inputs. Other methods accomplishing the same result may also be used.



Fig. 9-18. A circuit diagram of a full-wave ratio detector employing two double diodes and a compensating resistor R_c .

Figure 9-18 shows another arrangement of a ratio detector. In this case two double diodes are used to accomplish full-wave detection. This eliminates any effect of harmonic distortion, but a compensating resistance $\underline{R}_{\underline{C}}$ is still employed in the constant voltage section in series with \underline{C}_3 .



Fig. 9-19. A ratio-detector circuit using an impedance-matching winding L_t and feeding a 6AL5 high-perveance double diode.

The circuit for a ratio detector operating at a center frequency of 10.7 megacycles is shown in Fig. 9-19.5 To improve the sensitivity,

a form of impedance matching is used between the high impedance of the driving pentode and the relatively low impedance which the diodes reflect across the winding that injects the primary voltage into the diode circuit. This is the winding \underline{L}_t , which is normally wound over the cold end of the primary winding \underline{L}_p and tightly coupled to the primary. Slug tuning is used for both primary and secondary windings. A bifilar construction is used for the secondary winding, the slug penetrating each half of the winding to essentially the same degree; hence, variation in the position of the slug does not unbalance the two halves of the secondary winding.

The major part of the rectified output voltage is stabilized by means of the 8- μ f capacitor C₁. This capacitor, in conjunction with the two load resistors R₁ and R₂, gives a discharge-time constant of approximately 0.1 sec. The rectified voltage drop across R₃ and R₄ is not stabilized against changes in diode current. This permits minimizing the residual balanced component of amplitude modulation for this particular circuit design. The fact that the two resistors are not equal in value makes it possible to produce a compensating unbalanced component to overcome that which might otherwise appear in the output. This unbalance is principally due to the variation in the dynamic input reactance of the diodes, and is not necessarily an indication that the circuit is unbalanced in any way. The 47-ohm resistor R₅ modifies the peak diode currents and further reduces the unbalanced a-m component, particularly at high signal levels.

The capacitor \underline{C}_2 by passes the secondary system to ground at the intermediate frequency. The conventional de-emphasis circuit, which is placed conveniently at this point, provides further filtering against intermediate frequency getting into other circuits and causing feedback and its accompanying difficulties.

Good ratio-detector performance can be obtained with either high-perveance diodes, such as the 6AL5, or medium-perveance diodes, such as the 6H6. Diodes having less perveance than the 6H6 give less satisfactory performance in the ratio detector, because the residual balanced component of amplitude modulation becomes appreciably large. It is desirable to make the secondary inductanceto-capacitance ratio as high as possible, consistent with keeping the secondary tuning capacitance high enough that stray capacitances and variations in tube capacitances do not have an excessive effect on the tuning. The secondary Q should, in general, be as high as possible while consistent with the peak separation. At frequencies in the order of 10 megacycles, this will mean a Q of about 75 to 150. The primary inductance-to-capacitance ratio should also be as large as possible to obtain the greatest sensitivity. The limiting factor here is the maximum stable gain between the grid and plate of the ratio-detector driving tube.

9-6. Frequency-Modulation Detector Employing an Oscillator.

One type of f-m detector utilizing an oscillator⁶ is the type suggested by G. L. Beers. The limiter is replaced by a locked-in oscillator, the oscillator being at about one fifth the input intermediate frequency. The oscillator is actually a divider circuit dividing the input f-m wave instantaneous frequency by 5, or by whatever other division factor is employed. The divider circuit is then followed by an f-m discriminator at one fifth the intermediate frequency and with one fifth the deviation.



Fig. 9-20. A block diagram of an f-m detector employing a locked-in oscillator divider instead of the conventional limiter.

A block diagram of the final stages of an f-m receiver employing this system is shown in Fig. 9-20. The signal from the final i-f stage is led into the locked-in oscillator. Since the signal is merely a synchronizing voltage, the output of the locked-in oscillator, besides remaining constant in amplitude, is always a constant times the instantaneous frequency, one fifth if the division factor is 5. Thus, the output of the locked-in oscillator does not contain any of the amplitude modulation that may have been on the incoming signal, but does contain all of the frequency modulation. The oscillator is then followed by a conventional discriminator at one fifth the intermediate frequency which, in turn, is fed into the audio amplifier.

For instance, let us suppose that the i-f signal had a carrier frequency of 5 megacycles with a frequency deviation of 75 kc. After passing through the locked-in oscillator, all of the amplitude modulation that may have been present on the wave will have been removed, and the signal will have a carrier frequency of 1 megacycle (if a division factor of 5 is employed) with a frequency deviation of 25 kc. The frequency discriminator is designed for these latter frequencies.

A single-stage f-m detector employing an oscillator is shown in Fig. 9-20.7 This circuit is also responsive only to frequency modulation and not to amplitude modulation; no limiter is necessary. The circuit employs a special heptode tube having a high transconductance and a sharp cutoff characteristic. The second and fourth grids of the tube are used to shield the third grid—the input grid from the rest of the tube.

Figure 9-21 shows the basic circuit diagram of the single-stage f-m detector employing an oscillator. It is desirable to drive the detector from a source of low impedance because of the feedback of energy from the electron stream to the input grid; hence, the input circuit consists of a three-to-one stepdown transformer using circuits tuned to the i-f carrier frequency. The input is fed into the third grid, which is shielded, by means of two screen grids, from the rest of the tube. The cathode and first grid are connected as an oscillator, using the inductance \underline{L}_g and the capacitance \underline{C}_g ; the



Fig. 9-21. A basic circuit of the single-stage f-m detector employing a locked-in oscillator.

combination is resonant at approximately the i-f carrier frequency. The voltage generated by the oscillator appears on the cathode and first grid. The current through the tube, being regulated by the oscillator, is a series of short-duration pulses—short with respect to the space between pulses.

However, the amount of space current reaching the plate is dependent on the potential of the input grid, the third grid, at the instant that the pulse arrives there. The potential on the input grid is the sum of the bias voltage and the input voltage which, depending on its polarity, either adds to or subtracts from the bias voltage.

In the plate lead is the quadrature circuit, consisting of an inductance \underline{L}_p , a capacitance \underline{C}_p , and loading resistor \underline{R}_p . This quadrature circuit is also tuned to the input carrier frequency; with resistor damping, it has a band width about 6 times the frequency excursion—12 times the frequency deviation. The circuit is so adjusted that the fundamental component of the plate current is fed back to the oscillator tank circuit in such a manner that its effect is reactive at all times. In this manner the plate current variations affect only the frequency of the oscillator and not its amplitude. The screen grids, grids 2 and 4, completely shield the oscillator is determined solely by the tank-circuit constants and the quadrature fundamental component fed back from the plate circuit. Thus, the result is pure frequency modulation of the oscillator by the input signal that controls the magnitude of the pulse reaching the plate circuit.

For each incoming frequency, the phase relation of the oscillator signal and the incoming signal tends toward equilibrium. Hence, the oscillator locks in at the same frequency as the input signal. The feedback is so damped that it is noncritical with frequency.

Now, as long as the amplitude of the oscillator voltage is unaffected with frequency, the relation of plate current to oscillator frequency is linear. In other words, as the incoming frequency varies, it results in a different magnitude of pulse flowing in the plate circuit. The relationship between the magnitude of the pulse and the frequency of the input signal is linear over the desired range of plus and minus the frequency deviation. The amplitude of the incoming signal, so long as it is large enough to maintain synchronism, acts merely to shift the phase of the lock-in point of the oscillator which does not affect the magnitude of the pulse in the plate circuit—the magnitude of the pulse being determined by the frequency. Thus, the magnitude of the pulse is unaffected by variations in amplitude of the incoming signal. The voltage across the load resistor is equal to the average value of the pulses, the r-f components being bypassed to ground through a capacitor. Therefore, the voltage across R_L is the detected voltage of the f-m wave. No limiter is necessary because the detector is insensitive to amplitude modulation.

The minimum signal needed to operate this detector is about 1/2 v rms for full 75-kc deviation. Since this input voltage acts merely as a synchronizing signal, the detector can possess a gain. The output for 75-kc deviation is about 15 v rms.

With careful attention to shielding, it is possible to reduce the response to amplitude modulation to about 50 to 60 db below the response to frequency modulation. There may be a small response to amplitude modulation because of stray coupling between the input signal and the oscillator circuit. This stray coupling, consisting of stray and interelectrode capacitance, may be reduced to a very small value by careful design of the tube and careful wiring layout. If necessary, a neutralizing circuit consisting of a blocking capacitor in series with a nearly self-resonant choke can be connected between the plate and the input grid to remove all residual amplitude response.

9-7. The Fremodyne Detector. Described as a superregenerative superheterodyne, the basic Fremodyne receiver and detector circuit, shown in Fig. 9-22, consists of a double-triode tube, one section of which is used as a local oscillator of the Colpitts type displaced 21.75 megacycles above the incoming signal frequency.⁸

The local oscillator, the lower section in the diagram, is fed onto the grid of the other triode section (upper section) along with the antenna input. This upper grid circuit is tuned to the signal frequency. The plate tank circuit of the upper section forms a Colpitts oscillator using \underline{L}_2 and the two 30- $\mu\mu$ f capacitors tuned to 21.75 megacycles. The same triode section operates as a superheterodyne converter and i-f amplifier. It is also a self-quenching superregenerative detector with a quenching frequency in the region of 17 to 30 megacycles by virtue of the 150,000-ohm resistor \underline{R}_1 returned to B_+ . By these means, the oscillator frequency that gives the strongest audio signal is recovered across a 22,000 ohm resistor in the lead from cathode to B_- .
9-8. Miscellaneous Frequency-Modulation Detectors. Fundamentally, any device which yields a d-c output proportional to fre-



Fig. 9-22. A circuit diagram of the basic Fremodyne receiver and detector circuit.

quency can be used as an f-m detector. For instance, a pulse counter, a device for indicating the pulse rate at any instant, can be used as an f-m detector by first converting each cycle of the f-m wave into a pulse. An ordinary multivibrator can be used for the conversion. Once constant-amplitude constant-width pulses are obtained, all the counter need do is filter out the r-f components and indicate the average power contained in the pulses (the higher the pulse rate, the greater the power).

Another f-m detector contains an artificial line through which a portion of the input signal is passed.⁹ This delayed wave is then combined with the original wave. Since the delay in an artificial line is a constant time, the delay will vary with frequency, being greater in phase for the higher frequencies — the phase delay being proportional to frequency. Upon recombining with the original wave, the resultant instantaneous amplitude is proportional to the instantaneous phase. As the frequency increases above the center frequency, the amplitude decreases; and as the frequency decreases below the center frequency, the amplitude increases. Passing this resultant signal through an ordinary a-m detector will yield the f-m signal. Since this system is responsive to amplitude modulation a limiter has to be used.

Still another method of detection can be obtained by reversing the Armstrong method of generating an f-m signal. The frequency de-

viation can be reduced by means of dividers to the proper low ratio of frequency deviation over modulating frequency. The carrier is then separated from the signal by a sharp filter, shifted in phase by 90 deg, and then recombined with the sidebands to yield amplitude modulation. The signal is then detected in an ordinary a-m detector, passed through an inverse correction network, and is ready for audio amplification.

Any ordinary reactance can be used to convert an f-m wave to its equivalent a-m wave since, basically, all reactances are frequency sensitive. Let us take the case of an ordinary inductance. If a varying-frequency constant-voltage signal is impressed across an inductance, the current through the inductance will be proportional to frequency. Passing this current through a small resistance, we can obtain an a-m wave which may be impressed on an ordinary a-m detector. In fact, there are many devices, such as wave guides, transmission lines, phase-shift networks, and so on, all frequency sensitive, which can be used as f-m detectors.

REFERENCES

1. See F. E. Terman, <u>Radio Engineering</u>, 2d ed., New York, McGraw-Hill Book Company, Inc., 1937, p. 54.

2. M. G. Crosby, "Reactance Tube Frequency Modulators," <u>RCA Rev.</u>, July, 1940. p.

3. Hans Roder, "Theory of the Discriminator Circuit for Automatic Frequency Control," Proc. IRE, May, 1938, p. 590.

4. S. W. Seeley, "Ratio Detectors for F-M Receivers," paper presented before the New York Section of the Institute of Radio Engineers, Oct. 3, 1945.

5. Stuart Wm. Seeley and Jack Avins, ""The Ratio Detector," <u>RCA Rev.</u>, June, 1947. p. 201.

6. G. L. Beers, "A Frequency-Dividing Locked-in Oscillator Frequency Modulation Receiver," Proc., IRE, December, 1944. p. 730.

7. William E. Bradley, "Single-Stage F-M Detector," <u>Electronics</u>, October, 1946, p. 88.

8. "Fremodyne F-M Receivers," Electronics, January, 1948, p. 83.

9. F. E. Terman, <u>Radio Engineers' Handbook</u>, New York, McGraw-Hill Book Company, Inc., 1943, p. 588.

QUESTIONS

- 1. What are the functions of an f-m receiver?
- 2. What are some of the design considerations of an f-m receiver?
- 3. Describe the operation of the single tuned-circuit f-m detector.

4. Find the resonant frequency and the \underline{Q} of a resonant circuit to be used in the detection of an f-m signal at 5 megacycles with a frequency deviation of ± 30 kc.

5. What are the disadvantages of the single tuned-circuit f-m detector?

6. Explain the operation of the f-m detector employing two tuned circuits, each detuned slightly to either side of resonance.

7. Determine the Q's of the tuned circuits to be used in an f-m detector employing two tuned circuits, each detuned slightly to either side of resonance, and used to detect the signal described in Question 4.

8. Describe the relationship between the primary and secondary voltages in a loosely inductively coupled resonant circuit.

9. Explain the operation of the f-m discriminator.

10. Discuss the variation of the phasor diagrams with frequency of the f-m discriminator.

11. Describe the effect of an amplitude variation of the input signal on the output of an f-m discriminator.

12. Why is a limiter used with the f-m discriminator?

13. Determine the Q of the secondary and the primary and secondary inductance of an f-m discriminator to be used to detect the f-m signal of Question 4.

14. Explain why the ratio detector is insensitive to amplitude modulation.

15. Describe the operation of the ratio detector.

16. Explain the variation in the phasor diagrams of the ratio detector.

17. How does a parallel resistor and capacitor with a large time constant duplicate, for audio variations, the effect of a battery?

18. Compare the action of a ratio detector with the f-m discriminator when the signal is momentarily cut off.

19. Describe the f-m detector employing a locked-in oscillator as a divider.

20. Describe the single-stage f-m detector employing a locked-in oscillator in a heptode circuit.

21. Describe the operation of the Fremodyne detector circuit.

CHAPTER 10

Amplitude Limiters

10-1. The Purpose of the Limiter. As shown in the previous chapter on f-m detectors, some detectors are insensitive to amplitude modulation, whereas others are not. The amplitude limiter is employed to remove the amplitude modulation from a signal before it is impressed on an f-m detector which does not inherently suppress amplitude modulation. Thus, the single tuned-circuit detector, the double tuned-circuit detector, the f-m discriminator, and all other f-m detectors which respond to amplitude modulation have to be preceded by some sort of limiter. The basic advantages of f-m systems—the gain in signal-to-noise ratio, the low background-noise level, the suppression of interference, and so on—are all based on the premise that amplitude modulation is rejected in the receiver.

The purpose of the limiter stage in an f-m receiver is to remove, as completely as possible, any amplitude modulation which may be present on the incoming signal without affecting or distorting the frequency modulation. Thus, we have a nonlinear circuit with respect to amplitude variation, while providing for the linear transmission of the f-m signal. The limiting may be accomplished in the grid circuit, called a "grid limiter," or it may be accomplished by saturation in the plate circuit of the output current. The cutoff limiter, based on the cutoff point of the grid-voltage plate-current characteristic, may be classed as a saturation limiter, inasmuch as the plate current is being cut off by too negative a grid voltage.

The amplitude limiter may be compared to the overflow in a reservoir. As long as there is sufficient water in the reservoir to carry off the excess through the overflow, the level of the water will be practically unaffected by the amount flowing into the reservoir. The more water that flows in, the more will run off through the overflow, maintaining a constant level in the reservoir. But, as soon as the water falls below the line where the excess runs off through the overflow, the water level in the reservoir will rise and fall with the influx of water; its level will no longer be constant.

In much the same manner, an amplitude limiter needs a certain threshold value before which the limiter will not function. Before this threshold is reached, the amplitude modulation in the incoming signal will get through to the detector, but after it is reached—after sufficient signal strength is obtained—the limiter prevents any amplitude modulation from reaching the detector. After the threshold has been reached, all higher signal strength is rejected, or discarded, by the limiter. To make sure that the input signal to the limiter always exceeds the threshold value, sufficient amplification usually is provided prior to the limiter stage. However, like the overflow water in the reservoir, a portion of the amplification in the receiver is lost by the limiting of the received signal, mainly because of the desire to operate well above the threshold value.

10-2. The Grid-Circuit Limiter. The grid-circuit limiter employs a grid leak and grid capacitor, operating very much like an ordinary class C amplifier. A circuit diagram of a grid-circuit



Fig. 10-1. A circuit diagram of a grid-circuit limiter employing a grid leak $\underline{R}_{\underline{g}}$ and a grid capacitor $\underline{C}_{\underline{g}}$.

limiter is shown in Fig. 10-1. The input is taken in through a tuned cirucit consisting of the inductance L_p and the capacitor C_p . This circuit can be the secondary of the last i-f coupling transformer. Rather than a fixed bias, only a grid leak R_g and grid capacitor C_g , usual in a class C amplifier, are employed. Each positive cycle of r-f voltage will cause the grid to draw current which, in turn, will create a bias voltage across the resistor $\underline{R_g}$. The polarity of this bias voltage, as shown, will cause the operating point of the tube to shift negative, reducing the plate current. Under proper operating conditions only the tip of the input cycle causes a current to flow in the plate circuit (very much like the case of the overdriven class C amplifier). If the amplitude of the input signal should increase, as might happen with the presence of amplitude modulation, it will cause more grid current to flow, charge the grid capacitor $\underline{C_g}$, increase the voltage drop across $\underline{R_g}$, shift the operating point more negative, and thus maintain practically the same magnitude of pulse

amplitude in the plate circuit. The operation of the grid limiter may be better understood by



Fig. 10-2. The grid-voltage plate-current operational characteristics of a gridcircuit limiter showing its limiting action on amplitude modulation.

examining Fig. 10-2. This figure indicates diagrammatically, on the grid-voltage plate-current characteristic of the limiter tube, the voltages and currents existing in the circuit. A sharp cutoff tube like the 6J7, operating at reduced plate and screen potentials, is employed. The grid input voltage is shown plotted vertically with time progressing downward, while the plate-current pulses are shown plotted horizontally with time progressing in the plus gridvoltage direction. The first few cycles of grid input voltage contain no amplitude modulation. The very tip of each wave draws grid current, which charges the capacitor \underline{C}_{g} and creates a negative bias voltage as the current is discharged through the grid resistor \underline{R}_{g} . The amount of current drawn by the tip of each wave is just sufficient to replenish the charge lost by the capacitor through the gridleak resistor, thus maintaining the equilibrium value shown as the dotted straight line in the figure. This circuit operates in the same manner as the grid-leak and grid-capacitor biasing of a class C amplifier. With the bias below cutoff, as is necessary for proper operation, the plate current consists of a series of pulses, the heights of which are determined by the voltage between the cutoff point and the peak of the grid voltage.

As soon as amplitude modulation is introduced, as shown in the diagram, the positive peak of the grid voltage is increased. This, obviously, will increase the amount of grid current which is being drawn. The increase in grid current will cause the charge on the

grid capacitor to increase, sending the bias voltage more negative. Only a very small increase in the positive peak is necessary to send the grid bias very negative. The new positive peak at the equilibrium point is but slightly more positive than the original positive peak with no modulation. Thus, the plate-current peak is increased only very slightly, eliminating practically all the effect of the amplitude modulation.

A similar effect takes place when the magnitude of the incoming signal decreases. The peak of the grid voltage is less positive than the previous cycles and it therefore draws less grid current. This smaller current does not replenish the charge on the grid capacitor, and the bias voltage decreases until a new equilibrium point is reached. The new equilibrium point brings the peak of the grid voltage to almost the same level, maintaining practically the same magnitude of plate-current pulse.

In summing up the operation of the grid-circuit limiter, we can say that the effect of the grid leak and grid capacitor is to bring all of the positive peaks into line; the cutoff point of the tube, brought to a low negative value by using low plate and screen potentials, skims off the peaks of the waves like a scythe cutting off the tops of grass. We note how the bias voltage follows the amplitude modulation as in a grid-leak detector circuit. The bias voltage must follow this variation in order to compensate for it. The mean value of the voltage will follow the amplitude of the received signal and can be used as an a-v-c voltage or tuning-indicator voltage.

The output circuit is an ordinary tuned circuit where the platecurrent pulses are changed into full cycles by the flywheel effect of the resonant circuit. Inasmuch as the pulses are all of practically the same amplitude, the amplitude of the wave produced by the tuned circuit will also be constant.

The time constant of the grid leak and grid capacitor must allow the bias to follow the a-m envelope exactly; otherwise some amplitude modulation will get through. As in driving an automobile, to keep within the lane of traffic, we have to follow the contour of the road. A time constant of about 10 μ sec, provided, for example, by a grid resistor of 100,000 ohms with a grid capacitor of 100 $\mu\mu$ f, can be used. If the time constant is too high, the grid-bias variation will not follow the a-m envelope of the incoming signal and, if it is too low, the change in bias is not great enough to reduce the amplitude modulation sufficiently.

In Fig. 10-3 is shown the input-output curve of a typical gridcircuit limiter. The shape of the curve between zero input voltage and the threshold voltage is dependent on the characteristic of the tube employed in the circuit. The output voltage raises to a maximum and, in spite of the appearance of increased plate-current pulse amplitude (as shown in Fig. 10-2), the output voltage drops with an increase in input voltage. This is caused by the decrease in the amount of fundamental energy contained in the plate-current



Fig. 10-3. The input-output voltage curve for a typical grid-circuit limiter.

pulses. The higher the input voltage, the smaller the portion of it which is clipped; hence, the plate-current pulse becomes narrower as the input voltage is increased, and even though the pulse is slightly greater in magnitude, it is also narrower, and the amount of fundamental energy contained in the pulse is less. Thus, the wave produced by the plate tank circuit decreases with an increase in the amplitude of the input wave. We note, however, that the decrease is quite small compared to the large variation in the input amplitude.

10-3. Gate-Type Limiter. One of the disadvantages of the gridcircuit limiter is that a short-duration high-voltage noise impulse will cause the grid capacitor to charge up to a high negative value and cut off the reception of the f-m signal until the capacitor has discharged back to its normal equilibrium value. If the limiter were a gate affair passing only a small amount of voltage above and below the zero axis, instead of a clipping circuit, this defect would be eliminated.

One method of accomplishing gate clipping is shown in Fig.



Fig. 10-4. A gatetype of limiter circuit employing two direct-coupled tubes with low plate and screen voltages.

10-4.¹ This circuit employs two tubes directly coupled together. Both tubes are operating at reduced plate and screen voltages so that cutoff occurs at a few volts negative. The cathode of the second tube is operating near the plate voltage of the first tube but, since the voltages involved are small, sufficiently large plate and screen voltages are supplied to the second tube to provide the proper operating conditions. They are usually so arranged that the cutoff voltages for the two tubes are equal. Both tubes are biased by means of cathode resistors to about 1 v above cutoff.

Basically, the circuit operates to remove all signal voltage below 1 v in the first tube and then, because of the reversal in phase taking place at the output of the first tube, removing all voltage above 1 v on the other side of the input cycle. This is illustrated



Fig. 10-5. The grid-voltage platecurrent operational characteristics of the two tubes used in the gate type of limiter shown in Fig. 10-4.

in Fig. 10-5, which shows the operational characteristics of the two tubes. An input wave containing amplitude modulation is shown as the grid voltage of the first tube (Fig. 10-5<u>a</u>). The bias voltage is quite close to cutoff so that the negative peaks of the wave are cut off in a straight line, as shown by the shape of the plate-current wave. However, each of the cycles varies in height on one side so that the wave still contains amplitude modulation. Direct coupling employing a resistor load <u>R</u>₁ is employed between stages so as not to shift the operating point with respect to the second tube. If capacitor coupling were employed, the bottom half of the wave, shown as the plate-current wave in Fig. 10-5<u>a</u>, would no longer line up. The reversal of polarity is normal in a straightforward resistancecoupled vacuum-tube amplifier.

The second tube has impressed on its grid the voltage drop across the resistance \underline{R}_1 . This wave will have the same shape as the plate-current output wave of the first tube. We note, however, that the cutoff portion is transmitted without distortion, while the remainder of the amplitude variation, this time on the other side of the cycle, is removed by the second tube. Hence, the plate-current wave in the second tube is a squarish wave, owing to the amplitude variation of the input wave being sliced off above and below the axis of the wave.

Thus, the fundamental operation of this circuit is quite different from the operation of the grid-circuit limiter. The grid circuit lines up all the peaks and clips off an equal amount of each peak, while this gate-type limiter cuts off the wave above a given voltage, removing the amplitude variation and resulting in a squarish output wave. When this squarish wave is fed into the output resonant circuit, it results in a sine-wave output to the discriminator.

The operating characteristic of this type of limiter is quite similar to the characteristic shown in Fig. 10-3 for the grid-circuit limiter, except that beyond the threshold point the output of the gate type of limiter rises slightly instead of falling, as shown in the figure. As the input voltage to a gate type of limiter is increased, the output wave being fed into the output resonant circuit tends to approach a square wave, the sides of the wave becoming steeper and steeper with the increasing input. The steeper the sides--the squarer the wave--the more fundamentalit contains, increasing the output of the resonant circuit proportionately. Checking back on the gridcircuit limiter, we find that in that case the pulses became narrower and narrower with increased input, quite in contrast to the foregoing analysis for the gate-circuit limiter.

In some applications the cathode circuit of the second tube is grounded and the operating voltages, including the bias voltage, are increased in order to obtain equivalent operating conditions. Sometimes, instead of direct coupling being used to maintain the constant level of the cutoff portion of the output of the first tube as it is being fed into the second tube, a grid leak and grid capacitor can be used in the grid circuit of the second tube to line up each cycle with the zero grid-voltage axis. After it is lined up at the input, the cutoff characteristic slices off the amplitude modulation contained on the opposite side.

10-4. Dual Limiters. In practical application, many designers prefer to use two stages of grid-circuit limiting rather than two tubes in a single-gate limiter, although, for many home-receiver designs, a single stage of grid-circuit limiting has been found quite sufficient. By using two stages of cascaded grid-circuit limiting, the small amplitude variation, indicated in Fig. 10-3, can be removed and the threshold lowered to a smaller value. An input-output voltage characteristic of a double or dual limiter is shown in



Fig. 10-6. An input-output voltage characteristic for a dual limiter employing grid-circuit limiting.

Fig. 10-6. We note how very straight the output characteristic is, yielding excellent limiting. Of course, it can never be perfect but the effect of amplitude modulation in a signal above the threshold value is reduced to negligibility.



Fig. 10-7. The circuit diagram of a dual-limiter operating at 10.7 megacycles and employing $10-\mu$ sec time constants.

In Fig. 10-7 is shown the circuit diagram of a dual limiter as used with an i-f amplifier of 10.7 megacycles. The i-f transformers are resistance-loaded with 100,000 ohms to obtain the necessary 200-kc band width. Limiting is accomplished in the grid circuit of the 6SH7 tubes by means of the 50- $\mu\mu$ f 200,000-ohm circuit. Thus, this limiter employs a 10- μ sec time constant in both tubes. The d-c voltage drop across the first 200,000-ohm resistor is taken off through a 1-megohm resistor and is used as the a-v-c voltage. This removes some of the burden of regulating amplitude from the limiters, since the a-v-c voltage varies the gain of the previous stages, tending to keep the level of the input voltage constant.

FREQUENCY MODULATION

In those cases where a great deal of impulse noise is expected, such as ignition noise in mobile apparatus, it is often advisable to reduce the time constant in the limiter to a very small value, as low as 1 or 2 μ sec. This allows the limiting voltage to follow the impulse-noise variation and eliminates the otherwise slow recovery time. In ordinary broadcast receivers, a ten- μ sec time constant is usually adequate.

REFERENCE

1. René T. Hemmes, "Introduction to FM Receivers and Discussion of FM Signal Amplifiers and Limiters," <u>FM and Television</u>, July, 1945, p. 40; a limiter which appears to use the same principle is shown in this paper.

QUESTIONS

- 1. When does an amplitude limiter have to be used?
- 2. Describe, in general terms, the operation of a limiter.
- 3. Explain the operation of a grid-circuit limiter.
- 4. What part of the cycle does the amplitude limiter pass?
- 5. Why is a sharp cutoff tube used in the grid-circuit limiter?
- 6. Why does the output voltage from a grid-circuit limiter decrease slightly with an increase in input voltage?

7. Explain what is meant by the threshold voltage and why a grid-circuit limiter possesses a threshold.

- 8. What determines the value of the time constant used in a grid-circuit limiter?
- 9. How does a gate-type limiter differ from a grid-circuit limiter?
- 10. Describe the operation of a gate-type limiter.
- 11. Why are dual limiters used?

216

CHAPTER 11

Radio-Frequency Amplifiers, Oscillators, and Converters

11-1. The Radio-Frequency Amplifier. The reasons for incorporating an r-f amplifier in an f-m superheterodyne receiver are similar to those justifying its incorporation in an a-m superheterodyne receiver--and are even stronger because of the much higher frequencies involved. The use of the r-f amplifier increases the sensitivity of the receiver, improves the signal-to-noise ratio, and rejects spurious signals.

The increase in gain, even though quite small as in the 100megacycle band, is useful in an f-m receiver, especially those receivers employing a limiter. When a limiter stage is used, a large over-all amplification of the incoming signal is necessary. Most of this amplification is obtained in the i-f amplifier and, if it is considerable, the danger of instability—parasitic oscillations—is increased. Any decrease in the amount of amplification necessary in the i-f amplifier (and gain in an r-f amplifier decreases it) simplifies the design of the circuits.

The amount of noise generated by a tube functioning as a converter is always greater than when it is operated as an r-f amplifier. For instance, a 6AC7 tube generates about twice the noise voltage when it is operated as a converter as when it is operated as an r-f amplifier.¹ Hence, if sufficient gain is obtained in the r-f amplifier to override the noise in the converter, the signal-to-noise ratio would be quite improved. However, any amplification at all in the r-f amplifier will usually improve the signal-to-noise ratio of the receiver. In f-m systems, where one of the basic reasons for employing frequency modulation is the improved signal-to-noise ratio, the incorporation of an r-f amplifier takes on added importance.

Finally, even in those cases where the gain of the r-f stage is negligible, the rejection of spurious signals is a very important contribution. The intermediate frequency is generated by the oscillator beating with the desired signal. If the desired signal is higher in frequen-

217

cy than the oscillator (by an amount equal to the intermediate frequency), another signal, lower infrequency than the oscillator by the same amount, will produce a spurious response. Similarly, if the desired signal is lower infrequency than the oscillator frequency, a spurious response will result from an extraneous signal higher in frequency. Spurious signals may also be generated by the interaction of two undesired signals, or between harmonics of the oscillator output and undesired signals. Sometimes signals within the i-f band itself are passed directly through the converter and cause serious interference. All these spurious signals may be rejected by the selectivity characteristics of r-f amplifiers. In ordinary cases, a single r-f amplifier stage is sufficient to reject interfering signals normally encountered in practice. For clear reception the r-f amplifier can be discarded only where the desired signal is very strong.

The selectivity of the r-f amplifier is usually determined by the tuned circuit at its input and the tuned circuit at its output. The band width, the frequency band included between points at which the frequency response of the circuit drops to 0.707 of its maximum value in voltage, is a measure of the selectivity; for each tuned circuit it is dependent on the resonant frequency of the tuned circuit— the carrier frequency of the desired signal—and on the effective Q of the circuit. Calling the band width Δf and the resonant frequency f_0 , we obtain for a single tuned circuit

$$\Delta \underline{\mathbf{f}} = \frac{\underline{\mathbf{f}}_{\underline{\mathbf{O}}}}{\mathbf{Q}} \tag{11-1}$$

This \underline{Q} not only incorporates the tuned-circuit inductance and the tuned-circuit resistance, but also the shunting impedance of any tubes or circuits (such as the impedance introduced by the antenna). At a carrier frequency of 100 megacycles, Eq. 11-1 indicates that a \underline{Q} of 500 would be necessary to obtain a band width of 200 kc. This \underline{Q} is extremely difficult to obtain at these frequencies, and the designer usually accepts the highest \underline{Q} obtainable with normal circuit components. For those applications where a high \underline{Q} is necessary, there are available special resonant cavities, butterfly circuits, and other similar devices, or, in special cases, more complex bandpass networks may be used. The important consideration, though, is the tube, the input resistance of which is usually the determining factor for the input circuit \underline{Q} .

Conventional r-f amplifiers nearly always use pentode tubes. The pentode tube possesses a very low grid-plate capacitance and thus removes the possibility of self-oscillation. If triodes were employed, they would have to be neutralized to prevent self-oscillation. The gain obtainable with a pentode is usually much higher than that obtainable with a triode, especially in the f-m frequency bands. In the 100-megacycle band, the input resistance of nearly all available tubes is still comparatively low and the actual selectivity is not too great. However, this property affects only the adjacent-channel reception normally rejected by the i-f amplifiers. For the spurious-signal rejection, the signals being quite far removed in frequency from the desired signal, sharp selectivity is not necessary and the r-f amplifier still performs its important functions.

Sometimes, where it is found necessary to decrease the damping effect of the r-f amplifier tube and thus increase the Q of the input circuit, a small resistance, from about 10 to 50 ohms, is introduced in series with the cathode lead. The negative feedback, introduced by this resistor, usually decreases the gain of the tube; but, by increasing the Q of the input circuit, it may actually provide an increase in over-all gain as well as provide better selectivity.²

11-2. Radio-Frequency Amplifier Circuits. A circuit diagram of an r-f amplifier operating in the 100-megacycle f-m band is



Fig. 11-1. The circuit diagram of an r-f amplifier employing a 6AG5 pentode tube and operating in the 100-megacycle f-m band.

shown in Fig. 11-1. In the broadcast band a horizontal dipole is usually employed as the antenna, with a balanced transmission-line lead-in to the receiver. This line is transformer-coupled to the input resonant circuit in the grid lead of the 6AG5 pentode amplifier. A tuning capacitor, as well as a trimming capacitor for alignment purposes, is shunted across the input coil. A $35-\mu\mu$ f capacitor isolates the grid for direct current and allows the a-v-c voltage to be impressed on it through a 200,000-ohm resistor. Cathode bias is provided by a bypassed 150-ohm cathode resistor. The plate circuit consists of a resonant circuit with its accompanying tuning and trimmer capacitors. The two tuning capacitors are ganged together with the oscillator tuning capacitor to provide single-dial control. Coupling to the converter is direct through a $35-\mu\mu$ f capacitor with a 50,000-ohm grid resistor in the converter circuit. When more than one band of frequencies is desired, and the f-m broadcast band may be split into two bands, it is found expedient to switch the tuned circuits either by push-button or rotary-switch control. Both the grid and plate resonant circuits have to be switched and, if we include the oscillator resonant circuit, it means a minimum of three switches ganged together. Very often it is desirable to change the bias by switching cathode resistors in the r-f amplifier. If this is included, it will result in a four-gang switch. For best operation, there are many items that could be switched, but it is best to employ as few switches as possible, particularly in the r-f leads.

The r-f amplifiers used in the 30- to 40-megacycle communication band are quite similar. For mobile equipment, the input is usually obtained from a whip antenna; the antenna coupling coil is singleended, one end being grounded and the other end connected to the lead from the whip antenna. The gain of the r-f amplifier at these lower frequencies is, of course, higher, and the types of tubes available for use in the amplifier are greater in number.

The grounded-grid amplifier has also been used as an r-f voltage amplifier.³ The use of a triode results in a better signal-tonoise ratio for the receiver as compared with a pentode tube. The triode cannot be used in a conventional circuit because, as mentioned before, it would have to be neutralized to prevent self-oscillation. The grounded-grid amplifier removes the necessity for neutralization by interposing the grid as a shield between the plate and input circuits. The input circuit in this case is the cathode circuit.

A circuit diagram of a grounded-grid voltage amplifier is shown



Fig. 11-2. The circuit diagram of a grounded-grid r-f voltage amplifier employing a triode tube.

in Fig. 11-2. The input, or cathode resonant circuit, which may or may not be tuned, is made up of the inductance \underline{L}_1 and the capacitor \underline{C}_1 , while the output, or plate resonant circuit, consists of \underline{L}_2 and \underline{C}_2 . It is evident that the current flowing through the cathode resonant circuit is the same as the current flowing through the plate resonant circuit—the current flowing down through the cathode circuit, across the bypass capacitor, and up through the plate circuit. Thus, the tube acts as a medium for transposing the input current from the input, which may be a low-impedance source, to the output, which may be a high-impedance load. Hence, the gain of this type of amplifier is the ratio of the resonance resistance of the output tuned circuit divided by the resonance resistance of the input resonance circuit.

11-3. Oscillator Requirements. The oscillator is one of the most important sections of an f-m receiver. It must be in about the 90-megacycle region when used in the broadcast band—a difficult oscillator to design and construct. Stable operation is a requisite for satisfactory operation of the receiver. One of the difficulties with oscillators in this frequency range is the presence of dead spots, points of zero or very low amplitude output. These difficulties must all be eliminated for acceptable operation.

Frequency error in the oscillator caused by tracking troubles in an f-m receiver not only will cause distortion by shifting the signal band outside the pass band of the i-f amplifiers, but will also cause distortion by shifting the signal off the linear portion of the f-m detector. The wider the i-f pass band and the greater the linear portion of the f-m detector, the less critical will be the oscillatorfrequency adjustment. With a strong signal, any amplitude distortion caused in the i-f amplifiers is usually removed by a limiter or by insensitivity of the f-m detector to amplitude variation. Thus, the important consideration in this case is the linear portion of the f-m detector.

Frequency modulation of the oscillator has no effect in a-m receivers, while it is of prime importance in an f-m receiver; in the latter, obviously, it is the frequency modulation which is detected. Frequency modulation of the oscillator may be caused by hum or noise voltages or plate-voltage variation. Thus, slow oscillator frequency variations, such as drift or tracking capacitor shift, as well as rapid oscillator frequency variations in the form of frequency modulation of the oscillator output, will cause serious interference in an f-m receiver.

Frequency modulation of the oscillator may be caused by platevoltage variations in the oscillator circuit. These variations may be caused by hum from the 60-cps line or by feedback from the audio stages. These may be eliminated by careful filtering or voltage regulation and by careful shielding of the circuits involved.

Frequency drifting of the oscillator is usually caused by temperature or humidity effects. It can be reduced a great deal by using temperature-compensated circuits. Inductance variations can be reduced by suitably proportioning the radius and length of the coil, each of these dimensions having a compensating effect upon one another. Temperature-compensated capacitors are available, as well as capacitors with negative coefficients to compensate complete circuits.

Heating of the oscillator tube will often affect the frequency of the oscillator. This is quite detrimental, since it means that stable operation will not be reached until some time after the set has been turned on. This effect may be minimized by loosely coupling the tube to the oscillator tank circuit. It may also be minimized by operating the oscillator on a subharmonic of the desired oscillator frequency in inverse ratio to the harmonic used.⁴ This, in effect, reduces the frequency shift caused by capacitance variations.

In the case of communications receivers where only one, or at the most, a few channels are used, a crystal oscillator may be employed. This assures constancy of frequency and removes many of the difficulties encountered with tunable oscillators. In cases where one channel is used most of the time and tunable operation is also desired, a crystal oscillator may be switched out and a tunable oscillator switched in; hence, for the main channel of communication very good oscillator performance is assured.

11-4. Oscillator Circuits. Any of the standard oscillator circuits can be employed in f-m receivers, provided they operate with satisfactory stability. For many ordinary applications, the conventional pentagrid converter, the triode-hexode converter consisting of a triode oscillator and a hexode mixer in the same envelope, or separate oscillator and mixer, may be used. The important requirements are uniformly good output and good stability.

One method of increasing the stability of an oscillator is shown



Fig. 11-3. One method of stabilizing an oscillator and making it independent of the plate-voltage variations.

in Fig. 11-3.⁵ Here a center-tapped inductance is used, being tuned to resonance by two similar series capacitors, both labeled \underline{C} . The center tap, instead of being returned directly to the cathode as in an ordinary Hartley oscillator, is connected to the junction of the two capacitors by a tapped resistance \underline{R} . The tap \underline{t} is returned to the cathode, as shown. This tap point is varied back and forth until the most stable operation of the oscillator is obtained. If desired, a tapped coil may be used instead of the resistance R.

Figure 11-4 shows the oscillator section of a 6SB7Y pentagrid converter as may be used in a receiver in the 100-megacycle f-m broadcast band. This is a tickler-coil oscillator with the grid tapped down on the tank circuit to improve the stability. The second grid, instead of being at ground r-f potential, is used to obtain the



Fig. 11-4. The oscillator section of a pentagrid converter as used in the 100-megacycle f-m band.

tickler feedback voltage. A tuning capacitor, ganged to the r-f and converter tuning capacitors, is employed in the tank circuit, as well as a trimmer for tracking purposes. Many of the oscillators used in f-m broadcast receivers, as well as those used in the 30- to 40megacycle communication band, are conventional in design and give quite satisfactory results.

11-5. Converter Systems.⁶ In this discussion the frequencies used will be the 88- to 108-megacycle f-m broadcast band, although these systems are applicable to other bands including the 30- to 40-megacycle communication band. When applying the circuits to another band, the frequencies have to be changed proportionately.



Fig. 11-5. A conventional type of superheterodyne arrangement for the f-m broadcast band using a 10-megacycle i-f system.

Figure 11-5 shows a block diagram of a conventional type of f-m superheterodyne conversion arrangement employing an r-f stage. Both the r-f stage and the converter-stage inputs are tuned over the frequency range of 88 to 108 megacycles. To obtain an intermediate frequency in the order of 10 megacycles, the oscillator

covers the frequency range of 78 to 98 megacycles. A lower frequency than the carrier is used because, even though the tracking problem is a little more difficult, the inherent stability of the oscillator is better at the lower frequencies.



Fig. 11-6. A superheterodyne system employing an 11-megacycle intermediate frequency where the injected local oscillator signal is at one half the usual frequency.

A variation of this system is shown in Fig. 11-6 where the second harmonic of the oscillator is mixed with the incoming signal in the converter. Better stability of operation with cheaper construction usually is obtained when using the harmonic of the oscillator frequency as the beat frequency. In this case, the oscillator frequency varies from 38.5 to 48.5 megacycles to produce an 11-megacycle intermediate frequency.

Another system which may be used is the double-superheterodyne arrangement. A simple type employing a single-oscillator source to develop both frequencies is shown in the block diagram



Fig. 11-7. A double-superheterodyne arrangement utilizing a single local oscillator to develop both beat frequencies.

of Fig. 11-7. Here a tunable local oscillator, varying between the frequencies of 45 to 49 megacycles, is combined with the incoming

signal at the first converter. The result is a signal between the frequencies of 39 to 49 megacycles. This means that the local oscillator frequency does not remain fixed in difference from the incoming signal, but rather varies more slowly in frequency to create the rising resultant output frequency. The image frequency is located in the range of 137 to 167 megacycles, well outside of the r-f selectivity curve. When the same oscillator frequency is combined with the output of the first converter in another converter, called the "second converter," the resultant output frequency is a constant, in this case, 10 megacycles. A minimum of four variable tuned circuits are necessary for this system: one for the r-f amplifier, one for the variable-frequency oscillator, and one for each of the converters. If any amplifier stages are used between the first and second converters, they too will have to be tunable over the band of 39 to 49 megacycles.

When two oscillators are used in a double-superheterodyne system, the design can take the form of the block diagram shown in



Fig. 11-8. A double-superheterodyne arrangement with the first oscillator a fixed-frequency crystal-controlled circuit.

Fig. 11-8. A crystal oscillator is used to convert the incoming 88to 108-megacycle signal to a 20- to 40-megacycle intermediate frequency. The second oscillator is a variable frequency oscillator between the frequencies of 24.3 to 44.3 megacycles. Thus, the output of the second converter is a constant 4.3-megacycle intermediate frequency. Comparing this system with the single-oscillator double-superheterodyne method shown in Fig. 11-7, we see that the same number of variable tuned stages is needed, as well as variable tuned circuits for any amplifier stages between the two converters. However, if the variable-frequency local oscillator is located at the first converter, as shown in Fig. 11-9, the number of variable tuned circuits necessary is reduced. In this case, only three variable tuned circuits are used, as in the conventional singlesuperheterodyne receiver: one in the r-f amplifier, one in the first



Fig. 11-9. A double-superheterodyne system for f-m reception employing a 4.3megacycle intermediate frequency with the second oscillator crystal-controlled.

converter, and one in the first local oscillator. The first intermediate frequency is a fixed 26 megacycles, the result of mixing the 88to 108-megacycle incoming signal with a 62- to 82-megacycle local oscillator signal. The second converter uses a crystal oscillator at 21.7 megacycles to produce a second intermediate frequency of 4.3 megacycles. In this system, 26-megacycle fixed tuned amplifiers can be used between the first and second converters, as in an ordinary i-f amplifier.

For fixed-frequency reception, such as in police or fire communication, the double-superheterodyne receiver can be used with two crystal oscillators. For a communication receiver in the 30to 40-megacycle range, the first intermediate frequency can be 4.3 megacycles and the second, 455 kc. The second intermediate frequency can be low, because frequency deviations as low as ± 15 kc are used for 100 per cent modulation.

11-6. Converters. The converter design can follow any of the conventional designs: a pentagrid converter with its oscillator section, a hexode converter or mixer with a separate triode or pentode oscillator—in some cases the oscillator-tube elements are contained in the same envelope as the converter, or an ordinary pentode can be used for the mixer tube. In any case, the signal input grid to the converter tube has to be tuned to the incoming signal, so the input capacitance of the tube should be low.

Figure 11-10 shows a simplified circuit diagram of a pentagrid converter for an f-m broadcast receiver. The output from the plate of the r-f amplifier is fed into a tuned circuit consisting of the inductance \underline{L}_1 and the capacitance \underline{C}_1 (the Hartley oscillator circuit). This circuit is the plate load of the r-f amplifier tube. It is coupled directly through a coupling capacitor and a grid resistor



Fig. 11-10. A simplified circuit diagram of a pentagrid converter as used in an f-m broadcast receiver with a 10-megacycle intermediate frequency.

 R_1 into the third, or input, grid of a 6SB7-Y pentagrid converter. The cathode and first grid of the tube are connected in a conventional Hartley oscillator, the inductance \underline{L}_0 and the capacitance \underline{C}_0 being the resonant circuit. The third grid of the tube acts as the plate of the oscillator, at the same time acting in conjunction with the fourth grid to form a shield for the input grid. The oscillator operates over the frequency range of 78 to 98 megacycles, yielding a 10-megacycle intermediate frequency. The plate output is led directly into the first i-f transformer of the i-f amplifier section.

Figure 11-11 shows the circuit diagram of the r-f amplifier and two converters in a double-superheterodyne receiver. This communication receiver is designed for single-channel reception: therefore, two crystal oscillators are used and the frequency deviation being received is only +15 kc maximum. A single-ended vertical antenna is used to receive the signal at a frequency somewhere in the range of 30 to 40 megacycles. A 6SD7GT tube is used as anr-f amplifier before the first converter. The tuned circuit is in the plate return of this tube and is coupled, by means of a coupling capacitor and grid resistor, to the input grid of a 6SD7GT tube. Also coupled into this grid, through the capacitor \underline{C}_0 , is the output of a crystal oscillator. This crystal oscillator also employs a 6SD7GT tube in a conventional crystal-oscillator circuit. Its frequency is quite low, a fourth harmonic being used for the reception of frequencies from 30 to 37 megacycles, and a fifth harmonic for the reception of frequencies from 37 to 40 megacycles. For instance, for the reception of a signal on 35 megacycles, a fourth harmonic

would be used. The input frequency, 35 megacycles, plus the intermediate frequency, 4.3 megacycles, divided by 4, yields the os-



cillator frequency, 9.825 megacycles.

The first intermediate frequency is a fixed frequency, 4.3 megacycles as mentioned previously, so that a fixed tuned i-f stage can be used for amplification. As shown in the diagram, this stage also employs a 6SD7GT tube. (By using as many similar-type tubes as possible, the number of spares that have to be carried with the equipment is smaller.) Following the first i-f amplifier, we note that a hexode-triode combination tube, a 6K8GT, is used as a converter. The triode section is used as a crystal oscillator at a frequency of 4.755 megacycles, which is mixed, in the hexode section, with the 4.3-megacycle first intermediate frequency. This yields a second intermediate frequency of 455 kc, which is passed on to the second i-f amplifiers. To change the tuning of the receiver, the crystals for the oscillators have to be changed and the tuned circuits tuned to the new frequencies.

REFERENCES

1. F.E.Terman, Radio Engineers' Handbook, New York, McGraw-Hill Book Company, Inc., 1943, p. 573.

 K.R.Sturley, Frequency Modulation, London, Hulton Press, Ltd., 1942, p. 26.
M.C.Jones, "Grounded-Grid Radio-Frequency Voltage Amplifiers," <u>Proc. IRE</u>, July, 1944, p. 423.

4. Sturley, op. cit., p. 30.

5. Y.Kusunose and S.Ishikawa, "Frequency Stabilization of Radio Transmitters," Proc. IRE, February, 1932, p. 310.

6. Sidney X. Shore, "Crystal Oscillators in F-M and Television," <u>Communica-tions</u>, September, 1945.

QUESTIONS

1. What are the advantages of using an r-f amplifier in a superheterodyne receiver?

2. What spurious signals can be rejected by an r-f amplifier?

3. Where is the selectivity of an r-f amplifier determined?

4. Why are pentode tubes used so often in r-f amplifiers?

5. Draw a circuit diagram of an r-f amplifier.

6. Describe the grounded-grid voltage amplifier.

7. What are some of the advantages of a grounded-grid amplifier?

8. What are the oscillator requirements for satisfactory receiver operation?

9. How is the maximum frequency drift of an oscillator in an f-m receiver determined?

10. What causes frequency-drifting of the oscillator?

11. Name some of the sources of frequency modulation of the oscillator.

12. When can a crystal oscillator be used in a receiver?

13. Describe one method of improving the frequency stability of an oscillator.

14. What is a double-superheterodyne receiver?

15. How can a single oscillator be used for both converters in a double-superheterodyne receiver?

16. Discuss the advantages of employing a crystal oscillator for either the first or second converter in a double-superheterodyne receiver.

17. Draw a circuit of a pentagrid converter for the f-m broadcast band.

18. When can two crystal oscillators be employed in a double-superheterodyne receiver?

CHAPTER 12

Intermediate-Frequency and Audio Circuits

12-1. Intermediate-Frequency Amplifier Requirements for Frequency Modulation. The basic requirement of the i-f stages in f-m receiver is the same as that for any other superheterodyne receiver: it is to pass and amplify the desired signal and to reject any adjacent undesired signals. Thus, it has to pass a band of frequencies determined by the desired signal and its modulation sidebands, allowing for drift of the oscillator, calibration inaccuracies, and tracking errors. The sides of the pass-band characteristic have to be sharp enough to reject the undesired adjacent signal. Experience has shown that good practice allows for a variation of 6 db in the pass band and calls for an attenuation of 60 db or more for the interfering adjacent signal.



In Fig. 12-1 is shown a generalized transmission characteristic of an i-f amplifier. It is plotted in terms of frequency versus attenuation in decibels. The pass band extends up to $\pm f_1$, wherein all signals are attenuated less than 6 db, while all signals beyond $\pm f_2$ are rejected, being attenuated more than 60 db. Thus, all sidebands of the desired signal should normally fall within f_1 cps of the desired signal, while sidebands from the adjacent signals should not

WRH

be within \underline{f}_2 cps of the desired signal. For f-m broadcast reception, where the spacing between stations is 200 kc and the sidebands from each station extend out to about 80 kc on both sides of the carrier, \underline{f}_1 should be at least 80 kc to include all of the necessary sidebands, and \underline{f}_2 should be less than 120 kc to cut out the nearest sidebands of the adjacent signals. In any communication system, the larger that \underline{f}_1 can be made and the smaller that \underline{f}_2 can be made, the more leeway will be allowed for oscillator drift and the other mistuning effects, \underline{f}_1 approaching closer and closer to the value of \underline{f}_2 . In the case of the f-m receiver, however, the pass band is effective only as long as the detector is linear over the necessary range. If the frequency should drift out of the linear portion of the f-m detector, distortion will result even though the signal is still being passed by the i-f amplifier.

Upon examination we will find that there are many f-m receivers with i-f pass bands that are much too narrow for unattenuated passage of all the necessary sidebands. As the signal swings back and forth in frequency it will pick up amplitude modulation because of attenuation introduced by the i-f characteristic. These receivers depend upon the insensitivity of the detector to amplitude modulation, or upon the removal of amplitude modulation by a limiter stage, for undistorted f-m reception. However, it is a general characteristic of i-f amplifiers that there will be phase shift introduced by the amplifier when it introduces attenuation. Thus, the signal will be shifted in phase as it passes through the narrow i-f amplifier and, as we have seen in the consideration of transmitters, phase shift is the equivalent of frequency shift. This frequency distortion caused by the narrow-band intermediate frequency cannot be removed by any insensitivity of the detector to amplitude modulation, or by a limiter stage, since it is actual distortion of the frequency modulation. The distortion is not very noticeable at the lower modulating frequencies but at the higher modulating frequencies this extraneous phase shift causes appreciable distortion. Hence, the real importance of a flat wide-band i-f characteristic in an f-m receiver lies in its phase characteristic and not in its amplitude characteristic. However, since these characteristics usually go together, the i-f stages in frequency modulation are very often discussed in terms of their amplitude characteristic.

Considerations of gain and interference determine the intermediate frequency to be used. There is no doubt that a low intermediate frequency will result in a high gain per stage as well as a sensitive detector stage, but it calls for careful r-f selection to avoid interference from image and other extraneous frequencies.

Let us consider first the effect of the image frequency. When employing a 5-megacycle intermediate frequency, we have to adjust the oscillator frequency to differ from the received signal by 5 megacycles. For the reception of a signal at 100 megacycles, the oscillator would be adjusted to a frequency of 95 megacycles, if we keep the oscillator frequency on the low side. The 95-megacycle oscillator beating with the 100-megacycle desired signal would produce a 5-megacycle intermediate frequency. But another signal, an undesired one, at 90 megacycles would also create a 5-megacycle beat frequency when it is mixed with the 95-megacycle oscillator This is the image frequency. The image frequency is aloutput. ways caused by an extraneous signal which is twice the intermediate frequency removed from the desired signal and on the same side of the signal as the oscillator in terms of frequency. For instance, for an intermediate frequency of 8.5 megacycles, where the oscillator being used is lower in frequency than the desired signal, the image frequency would be located at twice 8.5 (17) megacycles lower than the desired signal frequency. If the oscillator were higher in frequency than the desired signal, the image frequency would be 17 megacycles above the desired signal frequency.

To reduce the chances of interference, it is desirable to locate the image frequency outside of the band of frequencies being received. This should be true for any tuning point of the receiver. The f-m broadcast band extends, in its complete form, from 88 to 108 megacycles, a band of 20 megacycles. If the intermediate frequency is greater than 10 megacycles, let us say 10.7 megacycles, the image frequency would be located twice 10.7, or 21.4, megacycles away from the received signal. Even at the extreme tuning points, at 88 and at 108 megacycles, the image frequency would always be outside of the f-m band. Thus the image signal would not be a wide-band f-m broadcast signal and the interference would probably be reduced.

Another point to consider is the interference caused by the mixing of two received signals which are separated in frequency by the i-f value. Thus, for a 10.7-megacycle intermediate frequency, two stations separated by 10.7 megacycles would mix together and create an i-f signal. One means of eliminating this occurrence is to increase the i-f frequency to a value greater than the band width of the whole assigned spectrum which, for the f-m broadcast band, is 20 megacycles. This, however, does not eliminate beats between f-m stations and carriers outside the band. Another means of eliminating this type of extraneous signal is to choose an intermediate frequency which is different from the spacings between any of the signals in the band. Since f-m stations in an area of reception in the broadcast band are spaced at their closest about 200 and, if possible, as far as 400 kc apart, an intermediate frequency consisting of an odd number of 100 kc would be preferable. It is for this reason that a frequency of 10.7 megacycles is usually preferred to a frequency of 10.8 megacycles for the i-f stages.

All of these precautions in the choice of an intermediate frequency do not eliminate the need for an r-f amplifier preceding the converter stage. The r-f amplifier is very necessary to reduce any image pickup from stations outside the band if a high intermediate frequency is being used, and to avoid extraneous pickup from beats between stations within the band and other signals outside the band. It is still desirable to avoid direct pickup from stations located within the i-f band itself which would cause annoying interference that cannot be tuned out by operating the normal controls of the receiver.

Once the frequency for the i-f stages has been decided upon, the problems that remain are to obtain the necessary amplification with the least number of stages, a flat band-pass characteristic over the necessary band width, and rapid attenuation outside the band to avoid adjacent-channel interference. These should be obtained without any instability such as self-oscillations or motorboating (intermittent blocking).

The gain of the i-f amplifiers is determined by the characteristics of the detector or limiter, if one is used, the desired sensitivity of the receiver, and the gain of the r-f amplifier and converter stages.

Let us assume that a sensitivity of $10 \ \mu$ v at the antenna terminals is desired, with an input to the detector stage of 1 v. This sensitivity means an over-all gain from the antenna terminals to the detector of 100,000. Upon design of the r-f and converter stages, it will be found that a gain of 5 is normally obtainable, leaving an over-all gain of 20,000 for the i-f stages. Thus we see that an increase in either the sensitivity or input to the detector stage will increase the amount of gain necessary in the i-f stages, whereas an increase in the gain through the r-f and converter stages will decrease the necessary i-f gain.



Fig. 12-2. A band-pass circuit consisting of two resonant circuits coupled together with a mutual inductance of M.

12-2. Band-Pass Circuits. The i-f circuits in an f-m receiver have to pass a band of frequencies which is comparatively wide as compared to the i-f circuits in an a-m receiver. The band width and the cutoff, or shape, of the i-f stages is determined fundamentally by the circuits used to couple one i-f tube to the next. One of the simplest types of band-pass circuit is the ordinary coupled circuit, the common i-f transformer found in a-m receivers. Thus, a response characteristic having a relatively flat top and steep sides can easily be obtained with two resonant circuits coupled by transformer action from the primary to the secondary. This circuit is illustrated in Fig. 12-2 where the primary is made up of C_1 and L_1 and the secondary consists of C_2 and L_2 , the two inductances being coupled together with a mutual inductance of <u>M</u>. The coefficient of coupling k is given by

$$\underline{\mathbf{k}} = \frac{\underline{\mathbf{M}}}{\sqrt{\underline{\mathbf{L}}_{1}\underline{\mathbf{L}}_{2}}} \tag{12-1}$$

By choosing the correct values of Q for the primary and secondary, and the correct value of k, it is possible to obtain a large variety of band width and flatness.¹ Although the analysis of the circuit is quite complex and is dependent on all factors, the pass band is very closely related to the coefficient of coupling k, while the response over the pass band, or flatness, is determined primarily by the Q's of the primary and secondary. A large value of k is necessary in an f-m i-f stage to yield a wide band between the cutoff sides; but if the Q's of the circuits are large, the characteristic will have double peaks, one on each side of the characteristic, with a pronounced valley in the center. Decreasing the two Q's to a low enough value to eliminate the valley will also round off the top. As was mentioned previously, the actual falling off of the signal because of curvature in the top of the i-f characteristic is not too important, since the resultant amplitude distortion is eliminated at the detector; however, the resultant phase-shift variation should be kept at a minimum to avoid f-m distortion.

An interesting effect is that, with similar Q's in the primary and secondary, a slight amount of detuning will produce the same increase in band width and other effects as an increase in <u>k</u>. With detuning, however, the magnitude of the output is less than with an increase in <u>k</u>. This method of increasing the band width is very seldom employed because of the difficulties in adjustment and alignment.

Another method of obtaining wide band width in the i-f amplifier is to use a series of single resonant circuits, each one coupling two i-f amplifier tubes together, as shown in Figure 12-3. Thus, each coupling stage will have a single resonant-response characteristic. To obtain a wide-band characteristic, each of the resonant circuits has to be tuned to a slightly different frequency. Let us assume that we are using a three-tube i-f amplifier. The converter would be coupled to tube 1 with a circuit resonant at f_1 ; tube 1 would be coupled to tube 2 with a circuit resonant at f_2 ; and tube 2 would be coupled to tube 3 with a circuit resonant at \underline{f}_3 . The three resonant characteristics are shown in Fig. 12-4a. Notice how \underline{f}_2 is located



Fig. 12-3. An i-f single-resonant coupling circuit; a series of such circuits each tuned to a different frequency will result in an over-all wide-band characteristic.

centrally between $\underline{f_1}$ and $\underline{f_3}$. The resultant amplifier characteristic will be the product of the three curves; the gain at any frequency is the product of the three gains at that frequency. The result is



Fig. 12-4. Three single-tuned resonant circuits used in successive stages of i-f amplification; each is tuned to a slightly different frequency so that the result is an over-all wide-band characteristic.

plotted in Fig. $12-4\underline{b}$; it is a wide-band characteristic with a flat response over the pass band and steep sides to cut out adjacent-channel interference.

The advantage of using single-resonant circuits in the i-f stages lies in the ease of alignment and tuning. Each stage is adjusted to a maximum at the picked frequency, $\underline{f_1}$ being the carrier frequency of the i-f signal, and $\underline{f_2}$ and $\underline{f_3}$ being frequencies on either side near the edges of the band. If the band width to be used is not very great, a combination of two instead of three circuits can be used. Another method of obtaining a broad-band response for an i-f amplifier is to employ a double-tuned circuit to couple one stage, followed by a single-tuned circuit in the next stage; this is illus-



Fig. 12-5. A wide-band i-f amplifier consisting of a double-tuned stage followed by a single-tuned stage which compensates for the valley in the overcoupled double-tuned stage.

trated in the circuit diagram of Fig. 12-5. A double-tuned circuit is used to couple tube I to tube II, while a single-tuned circuit is used to couple tube II to tube III. The Q's of the primary and secondary of the double-tuned circuit are kept large and the coefficient of coupling is increased until the proper band width is obtained. This will result in a response characteristic for the double-tuned



Fig. 12-6. The response characteristics of the combination double-tuned and single-tuned circuits used in the i-f amplifier shown in Fig. 12-5.

stage having a deep valley in its center (Fig. 12-6a). The singletuned circuit characteristic will have a single peak which will fall in the center of the valley of the double-tuned circuit. Thus, the three tuned circuits, two in the double-tuned and one single-tuned, will be resonant at the mid frequency of the pass band. The resultant characteristic (Fig. 12-6b) will be a wide-band flat curve, as shown.

Triple-tuned circuits, three mutually-coupled tuned circuits used between stages, have been employed to obtain very wide-band amplifiers for special application, but they are difficult to design, build, and adjust. They are seldom used in f-m receivers except when extra care and performance are necessary for a very special purpose.

12-3. Intermediate-Frequency Amplifier Design. Let us now follow through the design of an i-f amplifier for an f-m receiver.² The receiver, in this case, used an f-m discriminator as the detector stage. Two stages of i-f amplification were desired to amplify a voltage of $5 \mu v$ to the voltage necessary for effective operation of the limiter which, in this case, was 2 v. Thus, an over-all amplification of 400,000 was necessary. Assuming an over-all gain of 10 through the converter, we find that the necessary gain through the i-f amplifiers was 40,000.

As discussed previously, the amplifier must allow the passage of modulation frequencies to ± 100 kc from the carrier and must attenuate as rapidly as possible thereafter. However, experience has indicated that an attenuation of 6 db, 75 kc removed from the carrier, will result in no observable output distortion. Under this condition, an attenuation of 30 db, 200 kc removed from the carrier, can be achieved in design and will result, from a usable standpoint, in a practical degree of selectivity. To meet these requirements, an operating frequency of 8.25 megacycles was chosen in this instance for the i-f amplifier.

Several factors govern the gain and the selectivity of the i-f amplifier when the value of the intermediate frequency is estab-



Fig. 12-7. A schematic diagram of an intermediate amplifier operating at 8.35 megacycles.

lished. These are principally:

1. the transconductance of the amplifier tube and its interelectrode capacitance,

2. the number and Q of the tuned circuits, and

3. the ratio of inductance to capacitance in the tuned circuits.

The figure of merit of an amplifier tube may be considered as directly proportional to the transconductance \underline{g}_{m} of the tube, and inversely proportional to the plate-to-grid capacitance \underline{C}_{gp} . The 6SG7 tube with a \underline{g}_{m} of 4,000 and a \underline{C}_{gp} of 0.003 $\mu\mu$ f was chosen. The schematic diagram of the i-f amplifiers, including the converter stage, limiter, and detector, is shown in Fig. 12-7. It is comprised of three amplifying tubes, including the converter, using three stages of double-tuned circuit design. By making the inductance of the primary equal to the inductance of the secondary in the double-tuned transformer, and by making the Q's of both circuits equal in value, the stage gain at the frequency <u>f</u>, in terms of the <u>gm</u> of the amplifier tube, is given by

Stage gain =
$$\pi g_m f L Q$$

(12-2)

where \underline{L} is the inductance of the primary, the secondary having the same value.

But the value of \underline{Q} also determines the selectivity of the amplifier stage. It thus becomes necessary to balance the influence of \underline{Q} in stage gain against its effect on selectivity. It is found that a \underline{Q} of 45 will meet the necessary selectivity characteristic, as stated previously.

The ratio of inductance to capacitance that is chosen is also a compromise between amplifier-stage gain and stability with respect to the frequency to which the stage tunes. The tuning inductance should be high to provide adequate stage gain, but, on the other hand, the tuning capacitance should be high in order to minimize the effects of changes in capacitance due to changes in temperature in the tube, the socket, and the associated wiring. Experience has indicated that it is advisable to include at least 35 μ / μ f in the total shunt capacitance to allow stable operation and optimum gain. At 8.25 megacycles, the circuit inductance is thus 10.5 μ h.

Using these constants and substituting into Eq. 12-2, we obtain

Stage gain = $3-14 \times (4,000 \times 10^{-6}) \times (8.25 \times 10^{6}) \times (10.5 \times 10^{-6}) \times 45$

Stage gain = 49

This can be experimentally verified from the actual interstage i-f amplifier. In the stage preceding the limiter a gain of about 30 was obtained. By suitable choice of coupling in the converter stage, a conversion gain of 27 was realized, making the net over-all gain of 39,800 very close to the required 40,000.

The coil was constructed as a solenoid single-layer coil on a 5/16-in. coil-form diameter. The coils were wound of No. 36 B&S gauge enamel wire, 32 turns for the primary and the same number for the secondary. The distance between inside turns of the windings was 23/64 in. and the shield container was 1-3/8 in. on a side. The <u>Q</u> of the windings in air was 75. It was found necessary to use

electrostatic shielding between the transformer windings. In the absence of a shield, the capacitive coupling was approximately the same order of magnitude as the inductive coupling. Capacitive coupling will tend to distort the selectivity curve and, also, the effective coupling between the windings will change rapidly with spacing; the desired coupling is thus difficult to obtain and to maintain in production. It was found that the simple expedient of a brass screw inserted axially in the coil and connected to a frame proved a satisfactory means of minimizing the disturbing electrostatic



Fig. 12-8. An i-ftransformer for 8.25 megacycles; the 2-56 brass screw shown by dotted lines is used as an electrostatic shield.

BRASS MACHINE SCREW GROUNDED TO FRAME

coupling. A typical construction is shown in Fig. 12-8. The design of an i-f transformer for use at 10.7 megacycles is



Fig. 12-9. An i-f transformer for 10.7 megacycles using variable-reactance (slug) tuning, a fixed capacitance of $40 \,\mu\mu$ f, and a loading resistor of 120,000 ohms.

shown schematically in Fig. 12-9. In this transformer a fixed cap citance of 40 $\mu\mu$ f is used as the tuned-circuit capacitance and is shunted by a variable inductance, a slug-tuned iron-core inductance, which is varied by varying the position of a powdered-iron core within the inductance. The Q's of the coils are reduced to the desired value by means of a shunting resistance which, in this case, is 120,000 ohms. Using a 6SG7 tube as the i-f amplifier, we can obtain a stage gain of about 34 with this transformer.

12-4. Audio-Circuit Requirements for Frequency Modulation. The design of an audio system is always determined by the properties of the audio signal to be amplified. The use of high fidelity in f-m broadcasting has meant that the audio circuits in f-m broadcast receivers must be designed to handle this wider frequency range, preferably from a low of about 30 to 15,000 cps, although a lower limit of about 50 cps has been found acceptable in most practical cases. Since owners of f-m broadcast receivers are looking for fidelity of reproduction, the designer of the audio circuits has to be very careful to keep the distortion, cross modulation, and extraneous signal content of the audio-amplifier section as low as possible. In communication receivers, however, the audio band width is usually limited, since all that is desired is intelligibility.

One of the most bothersome extraneous signals present in audio amplifiers is hum. Because of the very low background-noise level, even a small amount of 60-cps power-supply hum is very noticeable. To meet the high-fidelity requirements of broadcast receivers, the audio circuits and the speaker system are designed to amplify and reproduce, without attenuation, any 60-cps signal which enters the system. Hence, very careful filtering and shielding are necessary to prevent the audio circuits from picking up any of the power-supply ripple or hum. When a full-wave rectifier system is employed, as in nearly all receivers being built today, it will be found that power-supply ripple will produce a 120-cps hum in the output. Stray transformer fields and filament current pickup introduce a 60-cps hum.

The speaker system in an f-m broadcast receiver must meet the same reproduction requirements as the audio system; otherwise, the fidelity of reproduction would be lost at this crucial point. The speaker system has to reproduce the whole high-fidelity band of frequencies without distortion or excessive beaming of the higher frequencies. By "beaming" is meant the concentration of the high frequencies into a narrow beam directly in front of the speaker, making them inaudible to any listener sitting to one side of the receiver.

12-5. Resistance-Capacitance Coupled Amplifier Considerations. The resistance-capacitance coupled amplifier is now used almost universally for a-f voltage amplification circuits in broadcast receivers. The main consideration, of course, is usually cost, but the <u>RC</u> coupled amplifier is also stable, easy to design, and easy to build. The important factor for the amplifiers used in f-m broadcast receivers is that the high-fidelity requirements can be met.

Figure 12-10 shows a circuit diagram of an <u>RC</u> amplifier stage employing pentode tubes. Tube I is coupled to tube II through the
network consisting of the load resistor $\underline{R}_{\underline{1}}$ of tube I, the coupling capacitor $\underline{C}_{\underline{C}}$, and the grid reisitor $\underline{R}_{\underline{g}}$ of tube II. Also shown in dotted lines are output, or plate, capacitance $\underline{C}_{\underline{p}}$ of tube I, and the



Fig. 12-10. A resistance-capacitance coupled amplifier showing, in dotted lines, the output capacitance \underline{C}_p of tube I and the input capacitance \underline{C}_g of tube II.

grid, or input, capacitance \underline{C}_{g} of tube II. These are the capacitances looking into the circuits under operating conditions, \underline{C}_{g} being equal to \underline{C}_{in} , as given in Eq. 4-7 of Chap. 4, where $\underline{R}_{\underline{L}}$ would be the load resistance of tube II. \underline{C}_{D} , for practical purposes, can be taken as the plate-to-cathode capacitance of tube I. The bypass capacitors are assumed to be large enough that their effect on the low-frequency cutoff point is negligible; to meet this requirement their reactances, at the lowest audio frequency to be amplified, should be small with respect to the impedances they are bypassing.

For design purposes the circuit of Fig. 12-10 may be simplified into three equivalent circuits³: one for the frequencies in the middle of the audio range, one for the low frequencies, and the third for the high frequencies. The nominal equivalent circuit of the <u>RC</u> amplifier for frequencies in the middle of the audio range is shown in



Fig. 12-11. The nominal equivalent circuit of the <u>RC</u> ^eg coupled audio amplifier shown in Fig. 12-10 for frequencies in the middle of the audio pass band.

Fig. 12-11. In this circuit we have assumed a transconductance \underline{gm} for tube I, assuming also that the plate resistance of the tube is so large with respect to \underline{R}_1 that it may be disregarded. If the plate resistance is not large enough to be disregarded, it should be shown connected in parallel with \underline{R}_1 in all of the equivalent circuits. In this middle range of frequencies, the effects of the coupling capacitor and the shunting capacitances can be disregarded. Thus, the voltage input to tube II \underline{eg} is obtained by taking the voltage drop

across the resistors which is caused by the tube current \underline{i} where \underline{i} is given by $g_{\underline{m}}\underline{e_{in}}$. Thus,

$$\underline{\mathbf{e}}_{\mathbf{g}} = \underline{\mathbf{e}}_{\mathrm{in}} \frac{\underline{\mathbf{g}}_{\underline{\mathbf{n}}} \underline{\mathbf{R}}_{\underline{\mathbf{l}}} \underline{\mathbf{R}}_{\underline{\mathbf{g}}}}{\underline{\mathbf{R}}_{\underline{\mathbf{l}}} + \underline{\mathbf{R}}_{\underline{\mathbf{g}}}}$$
(12-3)

Dividing both sides by the input voltage $\underline{e_{in}}$, we obtain the ratio of output to input voltage, the gain of the amplifier stage.

$$Gain = \frac{\underline{e}_{g}}{\underline{e}_{in}} = \frac{\underline{g}_{\underline{m}}\underline{R}_{\underline{l}}\underline{R}_{\underline{g}}}{\underline{R}_{\underline{l}} + \underline{R}_{g}}$$
(12-4)

This is the nominal gain of the amplifier stage, which drops off as the frequency becomes very high or very low.

The low-frequency equivalent circuit for the RC amplifier (Fig.



Fig. 12-12. The low-frequency equivalent circuit c^f the <u>RC</u>-coupled audio amplifier shown in Fig. 12-10.

12-10) is given in Fig. 12-12. At the low frequencies the reactances of the shunting capacitances \underline{C}_p and \underline{C}_g are so large with respect to the resistances \underline{R}_1 and \underline{R}_g that their effect may be disregarded. The coupling capacitor at these low frequencies, on the other hand, tends to block the current, and it is the value of this capacitor in relation to the values of the resistors that determines the low-frequency cutoff of the amplifier. We have, in effect, a load resistance \underline{R}_1 , across which is connected the series circuit consisting of the capacitance \underline{C}_c and the grid resistor \underline{R}_g , the output voltage being taken off across the resistor \underline{R}_g . As the frequency goes lower and lower, the reactance of the capacitor \underline{C}_c increases more and more, allowing less and less current to reach \underline{R}_g , thereby causing the response to drop off. The frequency at which the response, or gain, drops to 0.707 of the value given in Eq. 12-4 is usually called the "low-frequency cutoff point." Calling this frequency \underline{f}_{low} , we obtain, from the circuit of Fig. 12-12,

$$\underline{\mathbf{f}}_{1\text{ow}} = \frac{1}{2\pi \underline{\mathbf{C}}_{\underline{\mathbf{C}}}(\underline{\mathbf{R}}_{1} + \underline{\mathbf{R}}_{g})}$$
(12-5)

In Fig. 12-13 is shown the equivalent high-frequency circuit of the RC amplifier (shown in Fig. 12-10). At these high frequencies

242

the series reactance of the coupling capacitor $\underline{C}_{\underline{C}}$ is negligible, but the decreasing reactances of the shunting capacitances $-\underline{C}_{\underline{P}}$, $\underline{C}_{\underline{g}}$, and any stray capacitance across the circuit $\underline{C}_{\underline{S}}$ (the capacitance to

Fig. 12-13. The high-frequency equivalent circuit of the <u>RC</u>-coupled audio amplifier shown in Fig. 12-10. The total capacitance $\underline{C_T}$ includes all the capacitances shunting the circuit plus the stray, or wiring, capacitance $\underline{C_S}$.



ground of the wires and elements in the circuit)—tend to bypass the current across the resistors $\underline{R}_{\underline{l}}$ and $\underline{R}_{\underline{g}}$. This effect, of course, will cause the high-irequency response to drop off. Thus, the high-frequency equivalent circuit consists of the two resistors $\underline{R}_{\underline{l}}$ and $\underline{R}_{\underline{g}}$ in parallel, shunted by the total capacitance $\underline{C}_{\underline{T}}$ where

$$\underline{C}_{\underline{T}} = \underline{C}_{\underline{p}} + \underline{C}_{\underline{g}} + \underline{C}_{\underline{S}}$$
(12-6)

The value of \underline{C}_{S} , the stray capacitance, often has to be guessed at, since the design is usually made before the circuit has been put together in even a rough form. A value of between 2 and 5 $\mu\mu$ f can be taken, depending on the complexity of the circuit and the proximity to ground of the wiring and elements.

From the circuit of Fig. 12-13, we can calculate the upper-frequency cutoff, the upper frequency at which the gain drops to 0.707 of the value given in Eq. 12-4. This frequency <u>fhigh</u> is given by the following equation:

$$\underline{\mathbf{f}}_{\text{high}} = \frac{\underline{\mathbf{R}}_{\underline{\mathbf{l}}} + \underline{\mathbf{R}}_{\underline{\mathbf{g}}}}{2\pi \underline{\mathbf{C}}_{\underline{\mathbf{T}}} \underline{\mathbf{R}}_{\underline{\mathbf{l}}} \underline{\mathbf{R}}_{\underline{\mathbf{g}}}}$$
(12-7)

Thus, from Eqs. 12-4, 12-5, and 12-7 we can determine the nominal band width and amplification of an <u>RC</u>-coupled audio amplifier. By examining the equivalent circuits for the high- and low-frequency ends of the band, Figs. 12-12 and 12-13, we can see how the band width may be determined for design purposes. Since the value of the coupling capacitor $\underline{C}_{\underline{C}}$ determines the low-frequency cutoff, increasing the value of this capacitor will extend the amplifier to lower frequencies. From Eq. 12-5 we see that the same effect can be accomplished by decreasing $\underline{R}_{\underline{I}}$, or $\underline{R}_{\underline{G}}$, or both. The high-frequency response is determined by the relationship between the total shunting capacitance $\underline{C}_{\underline{T}}$ and the values of $\underline{R}_{\underline{I}}$ and $\underline{R}_{\underline{G}}$. Since the value of $\underline{C}_{\underline{T}}$ is usually an inherent property of the circuit and tubes (its value cannot be easily reduced), the high-frequency response is extended to higher frequencies by decreasing the parallel

impedance of $\underline{\mathbf{R}}_{1}$ and $\underline{\mathbf{R}}_{\underline{\mathbf{f}}}$. (See Fig. 12-13.) However, $\underline{\mathbf{R}}_{\underline{\mathbf{f}}}$, the grid resistance, is usually much larger than the load resistor $\underline{\mathbf{R}}_{\underline{\mathbf{l}}}$. Since its effect on the parallel impedance is relatively small, the high-frequency response can best be extended by decreasing the value of the load resistor $\underline{\mathbf{R}}_{1}$.

An example of the variation in gain as a function of frequency



Fig. 12-14. The variation in gain of an <u>RC</u>-coupled audio amplifier plotted as a function of frequency; a logrithmic scale is used to better show the high- and low-frequency characteristics.

of a wide-band audio amplifier is shown in Fig. 12-14. To show the variation at both ends of the band, a logarithmic scale is used for the

EXAMPLE. Determine the nominal gain, f_{low} and f_{high} of an audio amplifier, using the circuit shown in Fig. 12-10 with the following constants:

$$\begin{array}{rcl} \underline{\mathbf{R}}_{\underline{\mathbf{I}}} &= 400,000 \text{ ohms} \\ \underline{\mathbf{R}}_{\underline{\mathbf{g}}} &= 1 \text{ megohm} \\ \underline{\mathbf{C}}_{\underline{\mathbf{C}}} &= 0.0025 \ \mu \text{f} \\ \underline{\mathbf{C}}_{\underline{\mathbf{T}}} &= 30 \ \mu \mu \text{f} \\ \underline{\mathbf{g}}_{\underline{\mathbf{m}}} &= 1,200 \ \mu \text{mhos} \end{array}$$

The nominal gain is the gain in the middle-frequency range and is given by Eq. 12-4. Substituting into this equation, we obtain

Gain =
$$\frac{0.0012 \times 400,000 \times 1,000,000}{400,000 + 1,000,000}$$

Gain = 343 Ans. a

The value of <u>flow</u> is obtained by substituting into Eq. 12-5; thus,

$$\underline{f}_{10w} = \frac{1}{2 \times 3.14 \times 0.0025 \times 10^{-6} \times (400,000 + 1,000,000)}$$

$$\underline{f}_{10w} = 45 \text{ cps} \quad \underline{Ans} \cdot \underline{b}$$

The value f_{high} can be obtained by substituting into Eq. 12-7; hence,

$$\underline{f}_{high} = \frac{400,000 + 1,000,000}{2 \times 3.14 \times 30 \times 10^{-12} \times 400,000 \times 1,000,000}$$

$$\underline{f}_{high} = 18,600 \text{ cps} \quad \underline{Ans. c}$$

frequency. It results, in the normal case of an audio amplifier, in a symmetrical curve, as shown.

a

We have discussed, in this section, the desire to make the audio amplifiers in an f-m broadcast receiver wide-band, but we did not mention the need to keep the band width from being too wide. If the audio frequencies to be received lie between the frequency limits of 50 to 15,000 cps, as in the ordinary f-m receiver, the audio amplifiers should be designed to transmit that band width and no more; the wider the band width in an f-m receiver, the higher the noise level, as was pointed out in the discussion on noise and interference in Chap. 3. The gain decreases as the upper-frequency response of the amplifier is increased in the ordinary <u>RC</u> coupled amplifier; the efficiency of operation of the amplifier decreases with an increase infrequency response. When the frequency response is pushed very low, there is also more possibility of encountering blocking or motorboating. For these reasons, it is best to design the amplifier to cover the necessary band width and no more.

Another method of improving the frequency response of an audio amplifier, if there is amplification to spare, is to employ negative feedback. A small voltage from the output is fed back to the first stage of the audio-amplifier system in such a manner as to oppose the incoming signal. Depending on the amount of feedback, the gain is decreased, but the distortion is also decreased and the frequencyresponse band widened. The actual application of negative feedback in an f-m receiver audio system is no different from its application to any other audio system.

Shown in Fig. 12-15 is a diagrammatic representation of a feedback amplifier. Without any feedback, the amplifier has a nominal voltage gain of <u>A</u> times. In other words, if the feedback circuit were omitted and the gain of the amplifier were measured, the resultant gain of the amplifier would be A. The feedback circuit takes a fraction of the output voltage and feeds it back to the input. The value of the fraction is called β . For instance, if the output voltage were 5 v and the feedback circuit fed back 1/2 v of that output voltage to the input, the value of β would be 1/10.

The important value to consider is the value of the constant $(1 - A\beta)$. We might call this constant the "improvement constant" because the distortion with feedback has been reduced by this factor from the distortion which would nominally be present in the circuit without feedback, provided, of course, that negative feedback has been employed. By negative feedback is meant the fact that the fedback voltage opposes the input voltage and tends to counteract it. When negative feedback is employed, the value of β is taken as negative, and, since two negatives make a positive, the improvement constant is equal to 1 plus the quantity A times the magnitude of β . However, since in negative feedback the fed-back voltage tends to

oppose the input voltage, the resultant gain of the amplifier with feedback will be less than the nominal amplification without feed-



Fig. 12-15. A feedback amplifier using an amplifier with a nominal gain without feedback of <u>A</u> and a feedback circuit with a feedback factor of β .

back. The resultant amplification is actually equal to the amplification A without feedback divided by the same factor $(1 - \underline{A}\beta)$. Thus, the amplification of the amplifier is reduced by the same factor as the improvement in distortion.



A simple type of feedback circuit is shown in Fig. 12-16. The feedback voltage is taken off the secondary of the output transformer and the polarity is chosen so that it opposes the input voltage. The value of β is determined by the values of <u>R1</u> and <u>R2</u>, which act as a voltage divider—the value of β being equal to <u>R2/R1</u>.

Another simple type of feedback circuit that is often employed in a resistor-capacitor coupled amplifier is the unbypassed cathode resistor. Since the plate current, in flowing back to the amplifier tube, has to flow through the cathode resistor, if it is not bypassed it will feed back some of the output voltage to the grid of the tube. The value of β in this case is equal to the value of the cathode resistor divided by the value of the load resistor. (If the value of the cathode resistor is not very small with respect to the value of the load resistor, the value of β is given by the value of the cathode resistor divided by the sum of the load and cathode resistors.)

EXAMPLE. An amplifier has a nominal gain of 60 without any feedback. What is the reduction in distortion and the reduction in gain when a negative feedback circuit with a β of -1/12 is employed? (The minus sign is always used for β in a negative feedback circuit.)

The reduction in distortion and the reduction in gain is determined by the value of $(1 - \underline{A}\beta)$, which in this case is equal to $1 - 60 \times (-1/12)$ or 6. Thus, the final am-

plifier has one sixth the distortion of the amplifier without feedback and has a resultant gain of one sixth of 60, or 10.

It must be mentioned here that the feedback must be so arranged that at no frequency does the value of $\underline{A}\beta$ become equal to + 1, because, if it does, the amplifier will oscillate at that frequency.

12-6. Loudspeaker-Divider Networks. As mentioned previously, it is sometimes very difficult to obtain the necessary wide-band frequency response in a single speaker without beaming of the high frequencies. In many cases it is found more advantageous to employ two speakers, one for the high frequencies, called a "tweeter," and another for the low frequencies, called a "woofer." The two speakers may be installed separately, or they may be combined in a coaxial arrangement, the smaller high-frequency speaker being set coaxially within the larger low-frequency cone.

When two speakers are employed in a reproduction system, some type of dividing network has to be used in order to feed only the high frequencies to the tweeter and the low frequencies to the woofer⁴ in order to prevent cross modulation and distortion. The dividing point in the frequency spectrum is called the "crossover frequency." The low-frequency speaker is quite large in comparison with the highfrequency speaker, making it desirable to send most of the power in the audio signal to the woofer. Fortunately, the amount of power present in the higher frequencies is small in comparison with the power in the low-frequency range. A crossover frequency of about 400 to 2,000 cps is usually employed. Although the lower crossover frequencies are desirable, the higher ones result in more reasonable sizes and designs for the woofer and tweeter speakers.

There are many types of dividing networks. Two practical types



Fig. 12-17. Two simple practical types of divider networks used to feed a pair of speakers for high-fidelity reproduction.

with not too much complexity of circuit are shown in Fig. 12-17. These networks are designed to be placed between the output transformer and the voice coils of the speaker. Assuming that both voice

coils have the same resistance \underline{R} , and that a crossover frequency of \underline{fc} is used, we obtain for the values of inductance and capacitance

$$\underline{\mathbf{L}}_{1} = \frac{\underline{\mathbf{R}}}{\sqrt{2\pi}\underline{\mathbf{f}}_{\mathbf{C}}}$$
(12-8)

$$\underline{\mathbf{C}}_{1} = \frac{1}{2\sqrt{2}\pi} \underbrace{\mathbf{f}}_{\mathbf{C}} \underline{\mathbf{R}}$$
(12-9)

$$\underline{L}_{2} = \frac{\underline{R}}{2\sqrt{2}\pi \underline{f}_{c}}$$
(12-10)

$$\underline{C}_2 = \frac{1}{\sqrt{2\pi}\underline{f_c}\underline{R}}$$
(12-11)

Thus, to design either network shown in Fig. 12-17, the crossover frequency \underline{f}_{C} has to be decided upon and the voice-coil resistance of the speaker R has to be obtained. Using the above values and substituting into the proper equations, we can obtain the values of inductance and capacitance to be used. We must remember, however, that these networks carry the full output power, requiring that low-loss elements be used in the circuit. In commercial applications the loss is usually held to about 0.5 db for the network.

EXAMPLE. Design a dividing network with a crossover frequency of 1,000 cps for a pair of speakers, each having a voice-coil resistance of 10 ohms.

Let us use the network shown in Fig. 12-17a. We can now substitute into Eqs. 12-8 and 12-9, where <u>R</u> is taken as 10 and \underline{f}_{C} as 1,000.

$$\underline{L}_{1} = \frac{10}{1.4 \times 3.14 \times 1,000}$$

$$\underline{L}_{1} = 2.25 \text{ mh} \quad \underline{Ans. a}$$

$$\underline{C}_{1} = \frac{1}{2 \times 1.4 \times 3.14 \times 1,000 \times 10}$$

$$\underline{C}_{1} = 11.25 \text{ /f} \quad \underline{Ans. b}$$

and

12-7. Loudspeaker-Cabinet Requirements. The final result in any high-fidelity receiver is the sound which the listener hears, and even with the best receiver and the best loudspeaker, if the speaker baffle and enclosure are not correct, the resultant sound may be reproduced with poor fidelity. The cabinet should be designed so that the speaker enclosure and baffle are correct for the speaker to be employed.

One manufacturer⁵ recommends the following sized cabinets for its line of high-fidelity speakers: For an 8-in. dynamic speaker with a frequency response of 70 to 13,000 cps and a coverage angle of 70 deg, a sloping cabinet 16 in. wide, 21 in. high, with a top depth of 9 1/4 in. and a bottom depth of 12 in. is recommended. The actual enclosure required for this speaker is 2 cu ft. For the 10-in. speaker with a frequency response of 65 to 10,000 cps and a coverage angle of 60 deg, a sloping cabinet 19 in. wide, 22 in. high, with a top depth of 8 7/8 in. and a bottom depth of 11 13/16 in. is recommended. The enclosure required for this speaker is $2 \frac{1}{2}$ cu ft. For the 12 in. speaker with a coverage angle of 50 deg and a frequency response of 60 to 10,000 cps, a sloping cabinet 21 in. wide, 23 3/4 in. high, with a top depth of 9 1/4 in. and a bottom depth of 12 3/8 in. is recommended. This speaker requires an enclosure of 3 cu ft.

For a two-unit system with a frequency response of 60 to 15,000 cps and a coverage angle of 90 deg, a sloping cabinet 30 1/2 in. wide, 20 in. high, with a top depth of 11 1/4 in. and a bottom depth of 13 3/4 in. is recommended. This system is composed of a 12-in. dynamic speaker and a sectorial high-frequency horn, with the proper dividing network for frequency division between the speakers.

REFERENCES

1. F.E.Terman, <u>Radio Engineers' Handbook</u>, New York, McGraw-Hill Book Company, Inc., 1943, p. 170.

2. W.H.Parker, "The Design of an Intermediate-Frequency System for Frequency-Modulation Receivers," <u>Proc. IRE</u>, December, 1944, p. 751.

3. F.E.Terman, <u>Radio Engineers' Handbook</u>, New York, McGraw-Hill Book Company, Inc., 1943, p. 354.

4. J.K.Hilliard, "Loud Speaker Dividing Networks," <u>Electronics</u>, January, 1941, p. 26.

5. Western Electric Co.

QUESTIONS

What are the requirements for an i-f amplifier for use in an f-m receiver?
 What happens when the i-f band width in an f-m receiver is too narrow to handle the frequency deviation of the received signal?

3. Describe how the consideration of image frequency helps determine the intermediate frequency in an f-m receiver.

4. Determine the intermediate frequency that would be necessary to locate the

image frequency outside the 30- to 40-megacycle frequency range.

5. What determines the gain to be obtained in an i-f amplifier?

6. Find the necessary i-f gain for a sensitivity of 30 μ v at the input terminals to obtain a voltage of 2 v at the limiter. The gain from the antenna terminals to the converter output is 6.

7. Describe the response characteristic that can be obtained by two resonant circuits coupled together.

8. Describe the effect on the response characteristic of two coupled resonant circuits when the circuit Q's are varied; when the coefficient of coupling is varied.

9. What is the effect of detuning in two coupled resonant circuits?

10. Explain a method of obtaining a wide-band i-f characteristic by means of a series of single-tuned resonant circuits.

11. Describe how a combination of overcoupled double-tuned circuits and singletuned circuits can be used to obtain a wide-band i-f circuit.

12. What are the factors that govern the gain and selectivity of an i-f amplifier once the intermediate frequency has been chosen?

13. Find the stage gain of a double-tuned i-f amplifier circuit at 10.7 megacycles, where the transconductance of the tube employed is $3,600 \,\mu$ mhos, the circuit Q's of both resonant circuits are 50, and the coil inductance is 8 μ h.

14. Describe a method of electrostatic shielding between i-f transformer windings.

15. What are the audio-circuit requirements for f-m broadcast receivers?

16. Determine the nominal mid-frequency gain of an audio amplifier using the circuit shown in Fig. 12-10 with the following constants:

 $\frac{\underline{R}_{\underline{1}}}{\underline{R}_{\underline{g}}} = 500,000 \text{ ohms}$ $\frac{\underline{R}_{\underline{g}}}{\underline{C}_{\underline{c}}} = 1.5 \text{ megohms}$ $\frac{\underline{C}_{\underline{c}}}{\underline{C}_{\underline{T}}} = 20 \ \mu\mu\text{f}$ $\underline{g}_{\underline{m}} = 1,600 \ \mu\text{ mhos}$

17. Determine the low-frequency cutoff \underline{f}_{low} for the amplifier of Question 16.

18. Determine the high-frequency cutoff \underline{f}_{high} for the amplifier described in Question 16.

19. What is a loudspeaker-divider network and why is it used?

20. Design a loudspeaker-divider network for a crossover frequency of 1,500 cps and loudspeaker voice-coil resistances of 8 ohms.

CHAPTER 13

Frequency-Modulation Receivers

13-1. Receiver Requirements for Frequency Modulation. We have already discussed the specific requirements for the various units in an f-m receiver, but now let us consider the over-all receiver requirements. These over-all requirements are interrelated with the requirements of the individual units, since the units are designed only as a part of the complete receiver.

The sensitivity figure for an f-m receiver is the minimum input that will provide the freedom from noise and interference that is inherent in an f-m system. This may be expressed as the minimum input which, with full deviation (\pm 75 kc in broadcast receivers), delivers an audio output free from distortion and with the desired signal-to-noise ratio to insure freedom from noise caused by input resistors and r-f tubes. A signal-to-noise ratio of 10 or higher is desired for the minimum acceptable signal. The FCC considers the useful f-m broadcast area to extend out to the contour line where the field intensity drops to 50 μ v per meter; hence, f-m broadcast receivers should have sufficient gain to give satisfactory operation at field intensities below this value. Observations have shown that, with the best practical tubes and design, the maximum sensitivity obtainable in the f-m broadcast band is about 3 μ v and is limited by tube and antenna noise.¹

The r-f selectivity in the f-m broadcast receiver is also of concern, since the 400-kc separation between f-m stations in the same area in the 100-megacycle f-m band represents a separation of 0.4 per cent of the average signal frequency (in the standard a-m broadcast band the local station separation of 50 kc represents 5 per cent of the average signal frequency). Thus, the f-m stations are about 12 times closer together than the local a-m stations.

The r-f selectivity may be evaluated by the image ratio (the ratio of the receiver outputs for equal amplitude signals at the desired frequency and at the image frequency), which is given approx-

251

imately for a single r-f stage by

Image ratio =
$$4\underline{Q}\frac{\underline{f_i}}{\underline{f_s}}$$
 (13-1)

where $\underline{f_i}$ is the intermediate frequency, Q the property of the tuned circuit, and $\underline{f_S}$ the signal frequency.¹ For a 10-megacycle intermediate frequency and a 100-megacycle signal frequency, a circuit Q of 100 will yield an image ratio of 40 (one fortieth the sensitivity to the image frequency signal) for an r-f stage. For an antenna stage the ratio would be one half of this because of antenna loading. The image ratio is important, not so much as a measure of the direct image interference, but rather as a measure of front end selectivity against the spurious harmonic responses of a superheterodyne receiver.

A definite value of image ratio for satisfactory reception of f-m signals has not yet been definitely established. In a-m reception it has been felt that the ratio of desired signal to undesired signal should be at least 100 to 1 for satisfactory reception. Because of adjacent-signal discrimination, a ratio of f-m signals at the limiter (or detector, if no limiter is employed) of 3 to 1 would result in approximately the same amount of interference. Thus, an image ratio of 10 to 1 in frequency modulation would be equivalent to a ratio of over 300 to 1 in an a-m receiver.

13-2. Frequency-Modulation Tuning Indicators.² Since reception exists for an extended range around the correct tuning point in an f-m receiver, a good visual tuning indicator is a necessary adjunct. The case of mistuning in an f-m receiver is quite different from that in an a-m receiver. In an f-m receiver there is little change in audio volume level as the receiver is tuned in and out of the correct tuning point.

At the correct tuning point, the noise and distortion are both at a minimum. Slight mistuning may cause only a small increase in the ordinary type of background noise, but the high audio frequencies and all frequencies that are fully modulated would be distorted. This is because an f-m detector is a balanced circuit, and depends upon exact tuning to deliver undistorted signals to the audio amplifier. The balance also helps reject unwanted signals such as noise and interference. Just how effectively the f-m receiver eliminates interference is affected materially by this balance at the detector. With even slight detuning, the ignition of a passing automobile will produce annoying staccato that might not be present with the receiver tuned to the correct frequency. Thus, for the listener to take advantage of all the benefits that an f-m receiver affords, there should be provided some visual means of recognizing the correct tuning point.

252

To better illustrate the various tuning circuits, let us apply them to the conventional limiter and discriminator circuit shown



Fig. 13-1. A conventional limiter and discriminator circuit as used in an f-m receiver. \underline{V}_1 , \underline{V}_2 , and \underline{V}_3 represent the various d-c voltages on the discriminator, while \underline{V}_4 represents the d-c voltage on the limiter grid.

in Fig. 13-1. \underline{V}_1 , \underline{V}_2 , and \underline{V}_3 are the three d-c voltages, obtainable at the output of the discriminator, upon which the audio voltages are superimposed. \underline{V}_4 is the d-c voltage present on the grid of the limiter tube. The simplest type of tuning indicator is similar to that used in an a-m receiver. The grid of a tuning-eye tube can receive the d-c voltage \underline{V}_4 from the limiter grid and indicate when the signal is tuned to a peak in the over-all selectivity curve. This is an indirect method of approach, because the correct tuning point is the balance of the detector and this method assumes that the receiver is and remains perfectly aligned throughout; i.e., the selectivity curve is perfectly symmetrical and the discriminator curve is centered upon it. In the case of those detectors which do not employ a limiter, this method is obviously impossible to employ.

The direct method utilizes the voltages from the discriminator, and of necessity involves greater complication. Exact resonance occurs when the incoming wave is centered on the detector curve so that, according to Fig. 13-2—a graph of the variation of the three voltages \underline{V}_1 , \underline{V}_2 , and \underline{V}_3 — \underline{V}_3 , the sum of \underline{V}_1 and \underline{V}_2 , is zero. We are considering here only the direct voltages and assuming that any audio voltages present at these points are filtered out before the voltages are used for tuning-indicator operation.

By actually placing a voltmeter, preferably of the center-zero type, directly across the discriminator load, the receiver may be tuned accurately. However, the meter has been considered by most manufacturers unsuitable, and the next step is to learn how a tuning eye may be utilized. If the grid of the tuning-eye tube is connected to the high side of the discriminator load, the voltage V_3 will cause the aperture to increase and decrease, but there will be no way of determining where the zero-voltage point, the tuning point, occurs. It is therefore necessary to establish some zero-voltage reference point. This may be done in several ways.





The tuning-eye indicator may be operated with an initial fixed bias sufficient to close the aperture when the voltage \underline{V}_3 from the discriminator is zero. When \underline{V}_3 is positive, then, the eye will open, and when \underline{V}_3 is negative, the eye will overlap. Thus, the point of zero voltage, the correct tuning point, is established as the point at which the eye will just close. Some form of potentiometer must be provided with this method, since changes in line voltage, tube characteristics, and so on will require a readjustment of this reference point.

An alternate means of accomplishing this same function, but avoiding the necessity of regularly checking the zero points, consists in providing enough bias voltage to close the eye part way. A switch is then placed across the discriminator load so that when closed, the voltage on the eye grid is zero, and the amount the eye is closed is determined only by the bias. By tuning the receiver until the tuning indicator aperture is the same for both positions of the switch, the correct point for receiver tuning is indicated.

Since a manually-operated switch is undesirable, it may be replaced by a specially designed contacter, as shown in Fig. 13-3. It is a vibrating contact operating on alternating current to produce rapid intermittent short-circuiting of the tuning-indicator grid to ground. The result is two superimposed eye apertures, as shown in Fig. 13-3, one remaining fixed and representing the interval when the grid is grounded, and the other opening and closing according to the variation in the load voltage V_3 . Correct tuning will occur at the point where these two apertures coincide. A desirable variation is shown here by the dotted line and the dotted cross, where the negative voltage for the initial bias is sup-



Fig. 13-3. A circuit diagram of a contactor-type tuning-eye indicator. Two superimposed apertures are shown on the eye; correct tuning is indicated when they coincide.

plied by the limiter-grid circuit. Because the bias from the limiter is zero when no station is being received, the tuning indicator is wide open, but narrows as a station is approached on the dial. This circuit change makes operation less confusing, since there is a difference then between the condition where no station is being received and the case of reception with the receiver tuned exactly to resonance, both occurring when V_3 is zero.

Another method of using the tuning eye across the discriminator load is to put the positive and negative discriminator voltages through special circuits so that they will each produce a voltage in the same direction. This is shown in Fig. 13-4, where a double diode and a double triode are used in addition to the indicator tube.

In this circuit the detector voltage is impressed upon the cathode of one diode and the plate of the other, the two other corresponding electrodes being grounded through resistors across which voltages are developed and impressed upon the triode grids. It is apparent that through the unidirective action of each diode, one triode will receive only negative voltage and the other only positive. The latter, therefore, will have its plate current increased with a consequent lowering of the plate voltage through an increased drop in the plate resistor. This will decrease the voltage on the focusing electrode of the tuning indicator, since it is attached to this triode plate, and the tuning-eye aperture will increase.

In a like manner, a negative voltage at the discriminator will tend to decrease the plate current in the second triode and thereby reduce the voltage drop in the cathode resistor, effectively decreasing the bias on the first triode to produce the same effect as a positive voltage at the discriminator. Therefore, on either side of resonance the eye will open; good reception can be assured in the regular manner by tuning for the greatest closing of the eye. By proper choice of resistors, the eye can be just closed without overlapping for the condition of zero voltage across the detector. This is an extremely sensitive tuning device, more sensitive than a meter, and requires no preliminary setting up.



Fig. 13-4. A tuning-eye indicator circuit using a dual diode and a dual triode to produce a voltage in the same direction from both the positive and negative discriminator voltages.

The second portion of the eye in this system is actuated by the biasing of an auxiliary tube with the negative voltage derived from the limiter grid and may be omitted; in fact it must be omitted in detector circuits that employ no limiter tube.

Since the first section of the tuning indicator is closed when no signal is received as well as when exact resonance to a received signal is achieved, the operation may be confusing. However, this can be overcome by the addition of a resistor. shown by dotted line in Fig. 13-4. This reduces the voltage of the electrode controlled

by the detector when no station is being received. As a signal is approached, the voltage on the electrode connected to the single triode rises because of the increase in bias voltage supplied from rectification in the limiter grid circuit. The voltage on the detectorcontrolled electrode will also rise because of the connecting resistor; thus, when the station has been correctly tuned, the combination of effects will result in both apertures of the tuning indicator being closed.



Fig. 13-5. A relatively simple circuit whereby both the positive and negative detector voltages are converted to positive voltages and then impressed on the tuningeye tube grid.

Illustrated in Fig. 13-5 is a relatively simple circuit whereby the positive and negative detector voltages are both converted to positive voltages and then impressed on the tuning-eye-tube grid. It will be noticed that the d-c ground has been removed from the cathode of the detector diode and has effectively been placed at the electrical center of the load, so that the voltages, with respect to





ground, are now of the form shown in Fig. 13-6. These are the two voltages \underline{V}_5 and \underline{V}_6 shown in Fig. 13-5 across each half of the load. A bypass capacitor is used to ground the low side of the detector-

circuit load for audio frequencies, as it still is necessary to deliver to the audio amplifiers the full amount of detected audio frequencies.

To obtain sufficient sensitivity in this circuit, a sharp cutoff tuning-eye tube is needed. However, in combination a-m and f-m receivers this is undesirable, since a-m receivers require an in-



Fig. 13-7. A variation of the circuit shown in Fig. 10-5 where a dual sharp-cutoff triode receives the voltages from the circuit of Fig. 13-5.

dicator tube of the remote cutoff type. In Fig. 13-7 is shown a circuit where a dual sharp-cutoff triode receives the voltages of Fig. 13-6. It is seen from the curves of Fig. 13-6 that when one grid is positive, the other is negative by an equal amount, and at first thought it would seem that any effect produced on the tuning indicator by one triode would immediately be canceled by the other. However, since these triodes are operated in the region of plate-current cutoff, the negative grid voltage from the detector produces a negligible plate current as it is beyond cutoff, while a positive voltage on the other grid causes an abrupt plate-current rise, since it is in the conducting region of the tube characteristic. The plate currents flowing in the triodes are illustrated in Fig. 13-8. The sum of the two plate currents, the determining current which is flowing through the load resistor, is also shown. This results in an aperture variation as shown in Fig. 13-9.

This type of indicator is symmetrical and very sensitive. Here again, by using a portion of the limiter grid voltage, the tuning eye can be open when no station is being received and closed only when a signal is correctly tuned. This variation is shown by dotted line in Fig. 13-7.

Although the discriminator circuit was used for illustrative purposes in this discussion, the circuits illustrated can be adapted,





Fig. 13-8. The plate currents flowing in the two sharp-cutoff triodes shown in the circuit diagram of Fig. 13-7.

Fig. 13-9. The resultant tuningeye aperture variation for the circuit of Fig. 13-7 as the receiver is tuned through resonance.

with slight variations, to many other types of f-m detectors. For instance, in the ratio detector illustrated in Fig. 9-16 (Chap. 9), the variation of the voltages across the two capacitors $-C_2$, the output capacitor, and C1, the symmetrically placed capacitor across the other half of the circuit-resembles the two voltages shown in Fig. 13-6. They would be similar to those voltages if half of the constant voltage across both were subtracted from each. The constant voltage, as indicated in the ratio-detector circuit, is determined by the intensity of the signal and is used as an a-v-c voltage. This voltage is similar in magnitude and variation to the voltage on the limiter and can be used in place of it in the indicator circuits. Thus, by using the a-v-c voltage of the ratio detector instead of the limiter voltage, and by using the d-c voltages across the two capacitors instead of voltages \underline{V}_5 and \underline{V}_6 , we can obtain an indicator circuit similar to those shown. In a like manner, the principles of the indicator circuits can be adapted to other types of detectors.

Another method which does not necessitate finding the exact tuning, or balance, point is to employ a system of automatic frequency control (a-f-c).³ The system usually consists of two parts: an f-m detector for converting frequency departures from the intermediate frequency into a control voltage, and a control circuit for converting this control voltage into a variation in oscillator frequency. Thus, automatic frequency control is applied to the intermediate frequency produced at the converter and brings it into line



TUPHIS CONDENSER (10)

Fig. 13-10. An f-m tuner used to cover the frequency range of 87.1 to 108.9 megacycles. The numbers enclosed in circles are used to indicate the components in the layout shown in Fig. 13-11.



Fig. 13-11. Top and bottom views of the f-m tuner chassis showing the placement of the various components as labeled in the circuit diagram of Fig. 13-10.

261

with the chosen, or true, intermediate frequency.

In an f-m receiver the system is simplified because an f-m detector is an inherent part of the receiver. Thus, the d-c voltage output of the detector is used to regulate the frequency of the oscillator in the receiver by application of one of the methods discussed in Chap. 4 on direct f-m modulators. In this manner any mistuning of the receiver is corrected by operation of the a-f-c circuit within the receiver itself.

13-3. A Frequency-Modulation Tuner. An f-m tuner⁴ covering the frequency range of 87.1 to 108.9 megacycles, the f-m broadcast band, is shown in the circuit diagram of Fig. 13-10. This circuit shows all the tuner units: the r-f amplifier, oscillator, and converter, i-f amplifiers, limiter, and detector—everything (except the audio amplifiers and speakers) necessary to receive an f-m signal. It can be used in conjunction with any audio amplifier and reproducers having the proper fidelity of reproduction.

A 6AG5 tube is used in a straightforward r-f amplifier circuit yielding a gain, at 98 megacycles, of 17.5 from the r-f grid to the converter grid. The oscillator and mixer circuits are combined in a single pentagrid converter circuit using a 6SB7Y tube, and producing a gain of 8.3 at 98 megacycles, and 9.2 at 10.7 megacycles (the intermediate frequency) from the converter grid to the first i-f grid. The i-f circuits are inductively tuned with powdered-iron slugs and are loaded with loading resistors of 120,000 ohms. The first i-f stage has a gain of 34 to the limiter grid while the second has a gain of 33.

A 6SH7 tube is used as a grid-circuit limiter using a $47-\mu\mu$ f capacitor shunted with a 220,000-ohm resistor in the grid return. The d-c voltage on the limiter grid, across the 220,000-ohm resistor, is used as the a-v-c voltage on the r-f and first i-f tubes. It is taken off through a 1-megohm resistor to keep from loading down the grid circuit. It is also used as the source of the tuning-eye indicator voltage, being impressed on the grid of the 6U5 tuning indicator.

A 6H6 double diode is used in an f-m discriminator circuit for the detector. At the output of the detector a 100,000-ohm resistor and $0.00047-\mu f$ capacitor are used for de-emphasis. This capacitance, in parallel with the input capacitance of the first audio stage, is used to obtain the capacitance necessary for the required amount of de-emphasis.

Top and bottom views of the f-m tuner chassis are shown in the layout drawings of Fig. 13-11. The circled numbers refer to the components as labeled in the circuit diagram of Fig. 13-10. We notice, from these diagrams, the care required in the layout so that short leads may be used. This is necessary to avoid oscillation and feedback troubles in the completed circuit. 13-4. Combined Amplitude-Modulation and Frequency-Modulation Home Receivers.⁵ To satisfy the widest demand, home receivers sometimes have standard-broadcast and short-wave a-m bands, one or two f-m bands, and automatic tuning. The major item of interest to us is the 88- to 108-megacycle f-m band, which should provide maximum performance at the minimum price. Since all other bands are lower in frequency, the circuits and tubes are chosen for this band and the conventional a-m sections added.

We will discuss one particular design, and indicate other circuit possibilities. The receiver-circuit diagram is shown in Fig. 13-12. The tubes and circuits for the 100-megacycle f-m band were so chosen that they could be satisfactorily switched to conventional a-m service. High-gain tubes and the extra i-f stage necessary in an f-m receiver make the solution of gain and selectivity problems on the a-m band simple.

The f-m design in the following circuit is based on a limiter tube with an f-m discriminator circuit as the detector. The demodulators require three diodes, two for f-m detection, and one for a-m detection. To avoid circuit switching at the high i-f voltage point, separate diodes are used. At least one of the diodes must have a separate cathode to operate at the high-potential end of the discriminator. A type 6S8GT triple-diode high-mu triode tube is used for economy reasons. The circuit switching is accomplished at audio frequency, and the volume control is switched from the a-m audio load resistor to the f-m audio load resistor. This system is advantageous for any type of f-m detector that might be used.

A high audio or deviation sensitivity of the detector is necessary so that a-m to f-m switching shall occur with a minimum change of audio level. In amplitude modulation a mediocrea-v-c characteristic is tolerable and not easily avoidable with the use of high-gain semiremote-cutoff tubes and the removal of automatic volume control from the converter tube. The f-m output, due to the more stringent limiter requirements, results in a better equivalent automatic volume control and, while always higher than amplitude modulation at low input, it must also be equal to the highest a-m level.

To meet the above audio-level requirements, an i-f output voltage of approximately 10 v is necessary and should appear at the discriminator at all inputs, including the minimum usable signal. This contrasts with about 0.2 v required at the a-m demodulator diode. Thus, the gain in frequency modulation must be 50 times greater than the gain required in amplitude modulation.

In the limiter stage, a single-tube limiter is used, again for reasons of economy. The grid-bias limiter using a sharp cutoff tube with low screen voltage operates satisfactorily. The inherent contact potential of approximately 0.5 v on all thermionic tubes imposes a low limit on the voltage necessary for operation. A minimum of 1 v to the tube grid will permit a 50 per cent change in amplitude and still be on the operating range of the tube character-



Fig. 13-12. A schematic diagram of a complete receiver incorporating the standard broadcast and short-wave a-m bands, two f-m bands, and automatic tuning.

istic. This is about the minimum voltage for which the limiter can suppress amplitude modulation, owing to the i-f selectivity characteristic, and similarly, for the suppression of noise. This input



Capacitor values are given in micromicrofarads and resistor values in ohms, unless otherwise noted.

must also produce the required 10 v for the discriminator input.

Where further improvement is desired, overloading of the limiter grid may be reduced by designing the last i-f stage for some measure of additional limiting, such as the addition of a diode to the last transformer to limit the voltage level, or by the use of automatic volume control from the limiter to control the r-f tubes (as shown in the circuit of Fig. 13-10). Care must be taken in using automatic volume control on the i-f tubes, since it sometimes results in a change of effective grid-circuit capacitance and mistuning of the i-f circuit, with consequent dissymmetry. Such dissymmetry makes for poor interference rejection, f-m phase distortion, and undesirable manual-tuning characteristics. Since the peak noise voltage at the a-v-c filter network may be high, causing the grid to go positive and causing grid current to flow, a low timeconstant <u>RC</u> network will change a short-duration pulse to one of longer duration and result in audible noise.

Since the over-all sensitivity must be as high as possible to enable operation with the minimum antenna, as represented by the one installed within the radio cabinet, it is desirable that as much gain as possible be achieved at the intermediate frequency. The limiting factor is the equivalent grid-circuit noise voltage of the converter tube. For an advantageous signal-to-noise ratio at this point, it is found that the maximum desirable sensitivity is about $50 \ \mu v$ for 1 v at the limiter, or a gain of 20,000 times in the i-f amplifier.

The i-f tubes should be chosen for maximum voltage gain stability, as discussed in the previous chapter. This is given by the ratio of g_m/Cgp . Some typical tubes and their characteristics are given in Table I. The tubes are listed in their order of preference as determined by the ratio.

The highest stable gain obtained in the practical circuit was about 50 times. The stage feeding the limiter has a gain of about 40 times because it is loaded by the limiter grid, which draws current. The converter gain at the intermediate frequency of 8.3 megacycles was about 10 with the tubes used. The resultant is the product of $40 \times 50 \times 10$, or 20,000.

The selectivity characteristic of a broad-band i-f channel is required to be 180 to 200 kc at two times down and about 700 kc at 1,000 times down from nominal gain. The first figure is used to avoid phase distortion and to minimize undesirable amplitude modulation which makes tuning more critical and requires a greater degree of limiting (Chap. 12). The latter selectivity figure is due to the assignment by the FCC of local channels spaced 400 kc apart and a desire to achieve a high degree of adjacent-channel selectivity. Overcoupled transformers proved satisfactory for this characteristic.

The mechanical design of the combination transformer, includ-

ing the 455-kc and 8.3-megacycle transformers, is shown in Fig. 13-13. Because the transformers represent a large part of the per-



Fig. 13-13. A composite i-f transformer for 455 kc and 8.3 megacycles. The silvered-mica disks show at the bottom of the shield can.

formance and cost of the receiver, considerable work is devoted to making them simple and compact. In these transformers, permeability-tuned inductors and silvered-mica capacitors are used for good temperature-frequency stability.

In the mechanical construction of the transformer in Fig. 13-13, a single molded piece provides the four posts for the individual primaries and secondaries of the four tuned circuits, which work at 455 kc and 8.3 megacycles. A well in the molded piece below the posts contains a stack of silvered-mica disks which comprise the tuned-circuit capacitors. Contact to the silvered surfaces is made by double-ended formed-metal pieces, one end providing the coil terminal and the other end providing a terminal for chassis wiring. In each hollow coil post is a plastic liner into which is screwed a

TABLE 13-1

Тире Туре	<u>g</u> ^m	<u>C</u> gp	<u>g</u> m/Cgp
7W7	5,800	0.0025	2,300
6SH7	4,900	0.003	1,600
6SG7	4,000	0.003	1,300
6BA6	4,400	0.0035	1,250

A LIST OF TUBE TYPES AND THEIR CHARACTERISTICS

short threaded iron core to provide adjustment. The elasticity of the liner economically substitutes, in this case, for inside coilform threads.

Both primary capacitors are printed in silver upon a single mica disk, with another disk providing the secondary capacitors. Bypass capacitors associated with these circuits are also incorporated into the unit by means of additional mica disks. The drift for the complete unit is approximately 10 to 20 kc for a 30° C change in temperature.

The main requirements for a converter tube, which should be a single tube for economy, are a high-conversion conductance for gain, and a high oscillator g_m to support the oscillations with lowimpedance circuits at low-line voltages. The 6SB7 pentagrid converter, or the double triodes 6J6 and 7F8, will answer these requirements. The negative-signal grid input conductance of pentagrid converters improves r-f selectivity, which is much needed at 100 megacycles, and the internal oscillator modulation feature simplifies band switching; but the high cathode impedance, necessary for oscillator operation of this type tube, makes a suitable circuit difficult to design and prone to spurious oscillation. Double triodes result in higher conversion conductances but suffer from low signal-grid impedance as a result of plate-circuit degeneration through the triode mixer grid-to-plate capacitance.

Before a choice of tube is made, the oscillator frequency-temperature stability must be examined. This is of great importance in an f-m set, since a drift of 20 to 30 kc may be noticeable at the discriminator, affecting audio distortion and noise rejection. Such a drift is but a small fraction of 100 megacycles and necessitates careful design. Most of the frequency drift is caused by capacitance changes within the oscillator tube during the warm-up period, and by stray circuit changes as in the socket, band switch, or wiring. Since these total to an irreducible minimum capacity, regardless of frequency, it is desirable to make the change a small part of the total tuned-circuit capacitance. For this reason, second harmonic operation of the oscillator is sometimes used with a high lump-circuit capacitance as shown in Fig. 13-12.

Harmonic operation is usually not feasible with dual-triode tubes where mixer voltage injection is across an impedance. Since the oscillator-circuit to signal-circuit impedance is a considerable mismatch, more power is required for impedance injection. A high $\underline{g_m}$ pentode in the same tube envelope with a high $\underline{g_m}$ triode and interconnected by an injection grid would make an excellent converter.

Concerning the tuning system, a choice must be made between capacitor tuning, permeability tuning, and reactance tuning by means of vanes, metal slugs, or variations of these. For loop tuning in amplitude modulation, the variable-capacitor system produces the highest signal pickup sensitivity, especially where the loop size is limited by the cabinet.

For f-m tuning, a variable capacitor suffers from inductance in its plates and rotor wipers, and from its size, which makes it difficult to obtain short wiring. Circuit coupling through the common shaft may cause regeneration, and the structure and assembly of the gang capacitor cause instability with temperature change if great care is not taken.

For frequency stability, iron-core permeability tuning is quite good in the 100-megacycle band. There are available extremely fine particle-size iron powders which maintain circuit Q and permeability at these frequencies. To further enhance stability, in the circuit shown in Fig. 13-12, the normal trimmer capacitor was eliminated, although various silver-ceramic, glass, and air dielectric trimmers have been designed to improve stability of this adjustment.

If the inductance of the coil varies logarithmically with linear movement of the tuning element, then a fixed increment, plus or minus, of external circuit capacitance may be compensated for by a fixed displacement of the tuning element. After this adjustment at one point of travel, the inductance-versus-travel curve is the same as before. To satisfy such a tuning characteristic, the r-f and oscillator coils may be wound with a variable pitch, usually determined initially by experiment. Coils of this type are pictured in Fig. 13-14. Slugs for these coils are operated by the cam arrangement shown in Fig. 13-15.

The choice of an r-f tube involves gain, selectivity, and signalto-noise ratio to such an extent that as good a tube as can be afforded should be used. The gain and selectivity may be evaluated by examining the product obtained by multiplying the mutual conductance and the input resistance for each tube. The tube noise

FREQUENCY MODULATION

factor depends on mutual conductance divided by the square root of plate current. The ratio of plate current to total cathode current



Fig. 13-14. The winding of the r-f coil at the left is tinsel ribbon. The oscillator coil at the right is tuned to a frequency that is one half the signal frequency plus 8.3 megacycles.

should be high. The 6AG5 tube which was used in the circuit illustrated in Fig. 13-12 was chosen as a good compromise between cost and performance. The sensitivity of the receiver with this tube was found to be better than 10 μ v in production.

The audio channel is usually common to both a-m and f-m reception, which results in an amplifier that is capable of higher fidelity than the a-m system will allow. The limitation in the a-m system is usually loss of high audio frequencies caused by sideband cutting of selective circuits. Additional attenuation of higher audio frequencies is usually necessary to lower reception noise and hiss. Such attenuation of high frequencies does not occur appreciably in f-m selective circuits, and the noise rejection is a function of the system, making unnecessary such high-frequency attenuation. As a result, the necessary difference in attenuation is taken care of in the selective circuits, and the audio amplifier can be identical for both systems, the receiver selectivity accounting for the difference in high-frequency response. At the low-frequency end of a receiver audio spectrum, some trouble may be encountered in bass-boost design. For a-m reception, the audio-amplifier input network may need a rising gain characteristic for decreasing frequencies. This would be necessary to



Fig. 13-15. The complete tuning assembly of the receiver; the shaft of the capacitor gang for a-m tuning operates a cam that drives the tuning slugs of the f-m circuits.

make up for speaker and cabinet deficiencies and does not have to become excessive at extremely low frequencies, since transmitter modulation in amplitude modulation usually falls off rapidly at extremely low audio frequencies.

For f-m reception, such an audio-amplifier characteristic may be unsatisfactory. There is usually better low a-f modulation in f-m transmitters. The result could be an excessive bass boost in frequency modulation, and calls for a balance between the necessary low a-f accentuation in amplitude modulation and the added low frequencies present in frequency modulation. Where separate detector circuits are used, this can be adjusted by the low-frequency design of the f-m detector audio-output characteristic.

The complete chassis of the receiver whose circuit diagram is



Fig. 13-16. The complete chassis of the receiver whose circuit diagram is given in Fig. 13-12.

shown in Fig. 13-12 is pictured in Fig. 13-16.

13-5. All-Purpose Receiver.⁶ In Fig. 13-17 is shown the circuit diagram of a very-high-frequency superheterodyne receiver capable of accepting either a-m, f-m, or continuous wave code signals in the 27.8- to 143-megacycle range. Automatic-volume-control and automatic-noise-limiter circuits are incorporated, and the power-supply circuit allows the receiver to be operated from a 115- or 230-v 50/60 cps single-phase source or from an external supply that will provide direct current at 270-v and 6.3-v alternating

or direct current.

The circuit is that of the conventional superheterodyne receiver up to the second i-f amplifier stage. The output of the second i-f amplifier is fed to two channels, the f-m signal channel and the a-m signal channel. The f-m channel consists of the f-m limiter and discriminator detector, while the a-m channel consists of an additional i-f amplifier stage and second detector stage. The demodulated signal from both channels then feeds the same audio amplifier, being selected by the a-m-f-m switch.

The r-f amplifier stage employs a type 956 acorn-type pentode tube in a conventional class A amplifier circuit. The secondary of the input transformer is tuned by a section of a ganged capacitor shunted by a trimmer capacitor. The trimmer capacitor is controlled from the front panel by a control marked "Antenna" to provide accurate alignment of the r-f stage with varying antenna-load impedances. A resistor in series with the input grid is used to prevent parasitic oscillations. The output is transformer-coupled to the next stage. Shunting the transformer from the plate to the grid of the next tube is a 10- $\mu\mu$ f capacitor which provides a small amount of coupling to improve the response at the high-frequency ends of the bands, thus equalizing the r-f signal amplitudes over the tunable frequency ranges.

The mixer stage employs a type 954 acorn-type pentode tube in a cathode-coupled mixer circuit. A signal from a local oscillator 5.25 megacycles higher in frequency than the received signal on the low-frequency band, band 1, and 5.25 megacycles lower in frequency on the two high-frequency bands, bands 2 and 3, is fed to the mixer tube through the cathode and provides the difference frequency of 5.25 megacycles for the i-f amplifier stages. The oscillator circuit consists of a type 955 acorn-type triode in a tuned-plate untuned-grid type of oscillator circuit.

The first and second i-f amplifier stages employ type 6AC7 and 6AB7 pentodes respectively. The gain of the first and second i-f amplifiers is varied by the r-f gain control, connected in series with the cathodes of both tubes, to provide sensitivity control for the receiver, instead of the usual practice of varying the gain of the r-f amplifier stages. This method of control permits the r-f amplifier stages to operate at maximum gain. The a-v-c voltage is applied to the grids of the tubes in this section through a pair of decoupling networks. The a-v-c voltage is supplied by the second detector tube during a-m reception, and a small amount of voltage is also supplied for a similar purpose from the limiter tube during f-m reception.

The band width of the i-f channel is varied by means of a selectivity switch. It is a three-position switch which also turns on the a-c power. Using this switch, the operator can provide a relatively sharp frequency response for a-m reception (selectivity switch set at "Sharp"), or a relatively broad frequency response for f-m reception (selectivity switch set at "Broad"). The selectivity of the amplifier is controlled by switching in a third winding which varies the coupling between primary and secondary windings. In the Sharp position, the coupling winding is disconnected and only the coupling between primary and secondary windings determines



Fig. 13-17. The circuit diagram of a 27.8- to 143-megacycle receiver capable of receiving a-m, f-m, or continuous-wave code signals.

the band width of the i-f amplifier. In the Broad position, the coupling winding is introduced to increase the coefficient of coupling between the windings. This increase in coupling broadens the i-f amplifier frequency response to accept f-m signals. The signal voltage supplied by the second i-f amplifier is fed to the limiter and discriminator for f-m reception and to the third i-f amplifier stage



²⁷⁵

and second detector for a-m reception.

The last i-f amplifier stage, used for a-m reception, employs a type 6SK7 pentode connected in a conventional class A amplifier circuit. The gain of this stage is not varied as was the case for the first and second i-f amplifiers. The amplified signal voltage developed across the secondary of the output transformer is then fed to the second-detector stage.

Both the second-detector and automatic-noise-limiter stages employ a single-type 6H6 duo-diode. One section of the 6H6 serves as a detector for the a-m signals by rectifying the modulated carrier. The a-v-c voltage and the a-f voltage are obtained from the load and voltage divider resistors at the output. The remaining diode section of the 6H6 tube serves as automatic noise limiter as follows: The capacitor from plate to ground becomes charged by the rectified carrier voltage, and the time constant of this capacitor and the filter network associated with it is such that the a-f voltage variations do not alter this charge. During a severe noise pulse, however, the cathode of the diode plate connected to this capacitor becomes more negatively charged than the capacitor; hence, current flows, shorting the audio voltage to ground through the capacitor, until the cathode voltage of the automatic-noise-limiting diode again reaches a less negative potential than its plate and the capacitor acquires its normal charge again. By shorting the audio voltage to ground during a noise pulse, the automatic noise limiter prevents the objectionable noise pulses from reaching the audio-amplifier stages.

The beat-frequency oscillator, for continuous-wave code reception, employs a type 6J5 triode tube in a modified Hartley oscillator circuit. The oscillator frequency is adjusted by a movable powdered-iron core within the field of the tuned-grid circuit. This core adjustment sets the oscillator's frequency at 5.25 megacycles and is adjusted by a screw driver during alignment. The fine adjustment of the oscillator frequency required to provide control of the beat-note frequency is controlled by the variable-capacitor control labeled "Pitch Control" which tunes a small portion of the total oscillator coil.

The f-m detector consists of a limiter stage and a discriminator stage. The 6AC7 limiter tube operates as a saturated amplifier in which the output voltage remains constant over a large range of input-voltage levels. When operating as an f-m receiver, a-v-c action is obtained by using some of the voltage obtained in the gridcircuit return of the limiter. The discriminator circuit employs a 6H6 tube in a conventional hookup. The de-emphasis network, consisting of a 220,000-ohm resistor and a $560-\mu/\mu$ capacitor in the output of the discriminator, provides the necessary amount of deemphasis. From the de-emphasis network the audio signal is fed
to the a-f gain control in the same way as the audio signal from the a-m detector tube.

The audio-amplifier stage consists of a class A phase-inverter amplifier employing a type 6SL7GT twin triode driving a pair of 6V6GT/G pentodes in push-pull class A. The output of the audiopower amplifier is coupled to the load through a transformer, the secondary of which provides output impedances of 500 and 5,000 ohms to ground and 600 ohms balanced to ground. A tone switch is used to select the required network combination for the desired tone, or audio response.

The tuning meter serves two circuits in the receiver, depending upon the type of signals being received. It is switched from one circuit to the other by the a-m-f-m switch. When metering the reception of a-m signals, the tuning meter measures the plate current of the second i-f amplifier tube which varies with the strength of the signal carrier. When metering reception of f-m signals, the tuning meter measures the unbalanced current in the discriminator load resistors. As discussed in Sec. 13-2, the receiver is tuned to the exact center of the f-m carrier when the meter rests at zero.

In the power supply, a type 5U4G tube is employed in a conventional full-wave rectifier circuit. The "Send/Receive" switch is connected in series with the high-voltage lead from the rectifier filament to break the high-voltage circuit to the receiver's filter sections, thereby disabling the receiver but at the same time keeping the tube heaters hot, ready for instant use. To provide constant supply voltage for the oscillator, mixer, and screen grid of the second i-f amplifier stages, a voltage-regulator tube, type OD3/VR-150, is used. The voltage is supplied to the screen of the second i-f tube to provide accurate current control for the tuning meter connected in the plate circuit of this tube.

A front view of the complete receiver mounted in a cabinet 9 5/16 in. high, 19 1/8 in. wide, and 15 3/4 in. deep is shown in Fig. 13-18. It has a sensitivity of 2 μ v at 30 megacycles and 10 μ v at 135 megacycles for a 50 mw output when the input is derived from a signal generator modulated 30 per cent at 400 cps. The image ratio exceeds 1,000 to 1 at 30 megacycles, 300 to 1 at 58 megacycles, 100 to 1 at 80 megacycles, and 60 to 1 at 100 megacycles. The net over-all weight of the receiver is about 78 lb.

13-6. Ratio-Detector Frequency-Modulation Receiver. Receivers employing ratio detectors have a number of rather interesting characteristics. Since the rectified voltage across the long-time-constant load circuit of the ratio detector automatically adjusts itself to the input signal level, there is no fixed threshold, and the audio output and stabilizing voltage (which may be used for automatic volume control) are proportional to the input signal. Less

277

i-f gain is required, since the ratio detector in typical designs will provide appreciable amplitude rejection with as little as 10 to 50 μ v



Fig. 13-18. A front view of the combination a-m, f-m, and code receiver shown in the circuit diagram of Fig. 13-17.

of input signal to the grid of the ratio-detector driver tube. This will also tend to make the receiver quiet between stations. When automatic volume control is employed, the selectivity of the receiver is maintained for strong as well as weak station signals.

The tuning characteristic of a receiver employing the ratio detector is characterized by comparatively low side response because of its efficient rejection of amplitude modulation. (The side signal acquires amplitude modulation when it moves up and down the steep sides of the i-f selectivity characteristic.)

The circuit diagram, with the range switch in the f-m position, of a combination a-m-f-m receiver employing a ratio detector is shown in Fig. 13-19.⁷ This receiver covers the 540- to 1,600-kc a-m broadcast band, the 9.2- to 16-megacycle a-m short-wave C band, and the 88- to 108-megacycle f-m broadcast band. The intermediate frequency for a-m reception is 455 kc and for f-m reception is 10.7 megacycles. The receiver has an undistorted power output of 10 w into a 12-in. electrodynamic speaker whose voice-coil impedance is 2.2 ohms at 400 cps. Twelve tubes are used, including the power supply.

Few of the circuits, as shown in the diagram, deviate from the

conventional forms. The 6AT6 tube performs the function of second detector in the a-m broadcast band and the short-wave band only. Diode 5 of this tube functions as a device to prevent the a-v-c buss from becoming positive. The 6AU6 driver and the 6AL5 ratio detector are used only for f-m reception, the tubes being inoperative at all other times. The power-supply and output stages are shown



Fig. 13-20. A schematic diagram of the driver and output stage of the receiver shown in Fig. 13-19. Also included is the power-supply circuit.

in Fig. 13-20. A 6J5 tube is used to drive two 6F6G tubes in pushpull as the output stage.

The circuit of the ratio detector, as used in this receiver, is equivalent to the circuit previously described under f-m detectors. It differs only in the method of applying the i-f energy to the primary. A circuit diagram of the detector alone is shown in Fig. 13-22. We see here how the driver couples into the primary through an impedance-matching link and the primary is directly tied to the secondary of the detector input circuit. The a-v-c voltage is taken off across the long-time-constant circuit consisting of a 5- μ f capacitor and a 22,000-ohm resistor. The 75- μ sec time constant for de-emphasis is obtained in the resistor-capacitor combination made up of the 15,000-ohm resistor and the 0.005- μ f capacitor.



Fig. 13-19. A complete schematic for the radio chassis of a combination a-m, range switch is shown in the f-m position.



f-m, and short-wave receiver, employing a ratio detector for f-m reception. The



Fig. 13-21. Top and bottom views of the radio chassis



incorporating the ratio-detector circuit shown in Fig. 13-19.



Fig. 13-22. A schematic diagram of the ratio detector used in the receiver illustrated in Fig. 13-19. The tuning slugs for the various circuits are indicated beside the coils they tune.

REFERENCES

1. Z. Benin, "Modern Home Receiver Design," Electronics, August, 1946, p. 94.

2. John A. Rodgers, "Tuning Indicators and Circuits for Frequency-Modulation Receivers," <u>Proc. IRE</u>, March, 1943, pp. 89-93.

3. Charles Travis, "Automatic Frequency Control," Proc. IRE, October, 1935, p. 1125.

4. The Magnavox Co., Fort Wayne, Indiana.

5. Z. Benin, "Modern Home Receiver Design," Electronics, August, 1946, p. 94.

6. The Hallicrafters Co.

7. RCA Victor, Radio Corporation of America.

QUESTIONS

1. Discuss the receiver requirements for an f-m broadcast receiver.

2. Determine the approximate image ratio of an r-f stage using a tuned circuit with a Q of 90 and operating at 85 megacycles. The intermediate frequency in the receiver is 8.25 megacycles.

3. Why is it possible to receive satisfactory signals when employing a lower image ratio in an f-m receiver than in an a-m receiver?

4. Explain why a good tuning indicator is desirable in an f-m receiver which does not employ automatic frequency control.

5. How can the d-c voltage present on the grid of a limiter be used for the operation of a tuning indicator?

6. How can the voltages across the discriminator load resistors be used for the operation of a tuning eye?

7. Describe the circuit wherein both the positive and negative discriminator voltages produce a voltage in the same direction for the operation of a tuning eye.

8. Draw an a-f-c circuit for use in an f-m receiver.

9. In a combined a-m and f-m receiver, why does the design of the f-m broadcast circuits take precedence? 10. Why is it desirable to use separate a-m and f-m detectors in a combined a-m-f-m receiver?

11. Why should care be taken when using automatic voltage control on the i-f stages in an f-m receiver?

12. What determines the choice of an i-f tube in an f-m receiver?

13. Discuss the requirements of a converter tube for a combined a-m-f-m receiver.

14. Describe how trimmer adjustments may be made in a permeability-tuned oscillator without the use of a trimmer capacitor.

15. Do separate audio channels have to be used in a combined a-m-f-m receiver? Explain.

16. Describe how the band width of an i-f amplifier can be varied by the introduction of a third winding.

17. Explain the operation of the automatic noise limiter shown in the circuit of Fig. 13-17.

CHAPTER 14

Frequency-Modulation Transmitting Antennas

14-1. Electromagnetic-Wave Propagation.¹ Let us first review briefly some of the properties of electromagnetic waves in general. In free space (a fictitious concept where there are no ground planes or obstacles present) the electromagnetic wave spreads out from its source like a water wave after a disturbance in the water. However, the electromagnetic wave spreads out spherically in all directions; the wave front is a sphere instead of a circle as in the water wave. When the distance from the source, usually a transmitting antenna, is very large, the sphere is so large in diameter that, for all intents and purposes, the small amount of the wave front which is intercepted by the receiving antenna can be considered a flat surface, and a wave of this type is called a "plane wave." Since the distances between transmitting and receiving antennas are always comparatively large, the discussions and analyses of electromagnetic-wave propagation for broadcast and communication uses are normally confined to the discussions of plane-wave propagation.

The electromagnetic wave consists of an electric field and a mag-



Fig. 14-1. A diagram showing that the electric and magnetic fields in a propagated electromagnetic wave are at right angles to one another, and are both at right angles to the direction of propagation.

netic field at right angles to one another, as shown in Fig. 14-1. The two fields are represented by double arrows because they are a-c fields varying in magnitude and sign. The direction of propagation, also indicated in the figure by an arrow, is at right angles to both the

286

fields. When the direction of the electric field is vertical, the wave is said to be "vertically polarized"; when the direction of the electric field is horizontal, the wave is said to be "horizontally polarized." The receiving antenna wire should be parallel to the direction of polarization for strongest reception. In a purely polarized wave, one whose direction of polarization does not vary, a wire at right angles to the direction of polarization will have no voltage induced in it; there will be no reception.

Electromagnetic waves, like light waves, are reflected with an angle of reflection equal to the angle of incidence when they strike a surface like the ground, or the surface of a body of water, or the side of a mountain, or even the side of a building. When the reflecting surface is a perfect conductor, the reflected wave has the same magnitude as the incident wave and is either in phase, or 180 deg out of phase with the incident wave, depending on the polarization of the wave.¹

When the direction of polarization is parallel to the reflecting surface (for instance, a horizontally polarized wave striking a horizontal perfectly conducting reflector), the wave is reflected with a 180-deg phase shift, a reversal in sign. An interesting point is that, for the direction of polarization parallel to the reflector, the reflected wave cancels the incident wave at the surface, so that the intensity of the field at the surface is zero. As we move further and further away from the surface, the intensity increases and then decreases in a cyclic manner like the standing wave on a transmission line.

When the direction of polarization is perpendicular to the reflecting perfect-conductor surface (for instance, a vertically polarized wave striking a horizontal perfectly conducting reflector) the wave is reflected without any phase shift, in phase with the incident wave. For this type of reflection, the reflected wave reinforces the incident wave at the surface and the intensity of the wave at the surface is a maximum. As we move farther and farther away from the surface, the intensity of the wave decreases and then increases, again in a cyclic manner—a standing-wave pattern.

Figure 14-2 shows the case of reflection and refraction of an electromagnetic wave when it strikes the surface of a medium with different electrical properties than those of the medium in which it is traveling. If the wave is traveling through air it may hit a surface, like the side of a building, where the dielectric constant, the permeability, or both, are different from that of air. In this case some of the energy in the wave is reflected with an angle of reflection equal to the angle of incidence, while the rest of the wave is transmitted through the second medium with an angle of refraction that depends on the index of refraction, very much as in the case of light. The index of refraction is determined by the electrical properties of the two mediums. We note here the similarity between the behavior of light waves and radio waves.



Fig. 14-2. The refraction and reflection of an electromagnetic wave when it strikes the surface of a medium with different electrical properties than those of the medium through which it traveling.

The electromagnetic wave can travel from the transmitting antenna to the receiving antenna by several different routes. It may reach the receiving antenna as the ground wave, the electromagneticwave route which is affected by the presence of the ground; it may travel indirectly, being reflected by the ionosphere, the ionized region that exists in the upper atmosphere; or it may be reflected by the troposphere, the part of the atmosphere comparatively close to the earth where the formation of clouds takes place. Thus, there may be three different components at the receiving antenna: the ground wave, the ionospheric wave, which is sometimes called the "sky wave," and the tropospheric wave.

The ground wave may be divided into two parts: a surface wave and a space wave.² The surface wave travels along the surface of the earth and accounts for nearly all of the energy reaching the receiving antenna when both the transmitting and the receiving antenna are located at ground level. It is the main wave present in the daylight coverage for standarda-m broadcast band stations. The space wave may be divided into two components, the direct wave and the



Fig. 14-3. The two components of the space wave: the direct wave; and the wave reflected from the ground surface, called the "reflected wave."

reflected wave, reflected from the ground surface (Fig. 14-3). For horizontal polarization, the reflected wave and the direct wave tend to cancel one another at ground level since the wave is reflected with a 180-deg phase shift. As the receiving antenna is raised, however, the difference in the path lengths begins to compensate for the 180-deg phase shift and the magnitude of the space wave increases quite rapidly varying in a cyclic manner with height. After a few wavelengths of receiver antenna height, the space wave becomes the major part of the ground wave, the surface wave becoming negligible, and the magnitude of the received wave is proportional to antenna height. The magnitudes of the surface and space waves are affected by the ground constants, antenna heights, the condition of the atmosphere through which they are propagated, as well as the distance between receiving and transmitting antennas.

14-2. Electromagnetic-Wave Propagation in the Frequency-Modulation Bands. For f-m communication we are interested in electromagnetic propagation in the 30- to 40-megacycle and 88to 108-megacycle bands. The waves in the 88- to 108-megacycle band are not reflected by the ionosphere, while those in the 30- to 40-megacycle band are reflected only very seldom by the ionosphere and then only under exceptional conditions. Hence, in both of these bands, propagation takes place by means of the ground wave and, in some cases, by means of tropospheric reflections for reception beyond line of sight. Thus, the important reception area is within line of sight of the transmitting antenna, this area being determined mainly by the curvature of the earth. For reception in areas well within the line of sight, the field strength at the receiving antenna is practically inversely proportional to the square of the distance separating the receiving and transmitting antennas, as well as being proportional to the heights of both antennas. (When the distance separating the two antennas is not large with respect to the antenna heights, a condition not usually encountered in practice, this proportionality no longer exists and the received field strength assumes an oscillatory magnitude variation with distance.²)

For reception of waves in the f-m bands beyond line of sight, the propagation takes place through refraction of the waves in the earth's atmosphere and through tropospheric reflections. For this type of reception, the field strength decreases more rapidly than the inverse of the square of the distance between antennas, and increases more than normally with antenna height.

The height of the transmitting antenna is very important. By doubling it, the equivalent field strength of practically double the transmitter power is obtainable. For excessive height, of course, the construction and maintenance of the antenna tower as well as losses in the antenna feed lines remove any advantages that may be gained. It is thus an economic consideration to obtain the greatest height possible with the lowest cost while saving on the amount of power necessary for the transmitter.

The effect of polarization on field strength is another important consideration.³ The effect is only appreciable at very low antenna heights and when the antenna height is comparable with the distance

separating the receiver and transmitter, conditions which occur in mobile commercial communication. For low antenna heights the field strength is greater with vertical polarization and, since the amplitude of the ground-reflected wave is less with vertical polarization, the oscillatory variations in the magnitude of the field strength are less with vertical polarization. On the other hand, man-made interference such as that generated by automobile ignitions ordinarily produces stronger interference with vertically polarized fields than with horizontally polarized fields.

The f-m broadcast transmissions in the 88- to 108-megacycle band are standardized on horizontal polarization, it having been found that under normal conditions the signal-to-noise ratios will be greater than those obtainable with vertical polarization, particularly where the receiving-antenna height is not very great. For mobile transmissions, however, where the antennas are located very close to the ground, the added field strength obtainable with vertical polarization usually outweighs other considerations, even though signal-to-noise ratio may be no better than that obtainable with horizontal polarization. Since vertical mobile antennas are usually cheaper to install and maintain, nearly all ordinary mobile installations use vertical polarization.

The field strength obtainable from an f-m installation will be determined by the antenna height and the transmitter power. It is often necessary to predict the range of a particular installation or to determine the characteristics of the installation for a specified range. The predictions of these ranges can be simplified by the use of the chart shown in Fig. 14-4. With this chart it is possible to determine. the signals that can be expected with various effective radiated powers, antenna heights, and distances.⁴

The heights indicated in the chart are those of the transmitting antenna over the average terrain to the point in question, with a receiving-antenna height of 30 ft. This average height is ordinarily obtained by plotting a ground contour profile between the transmitter and the indicated point. Since such effects as shadows, reflection, and diffraction will be important, the values obtained can be taken as median values of intensity to be expected.

The chart really consists of two parts. One part is formed by the three scales: the transmitting-antenna height in feet, the distance in miles from the transmitter to the point in question, and the microvolts-per-meter field intensity created at that point for 1 kw of effective radiated power. Thus, a straight line connecting an antenna height with a distance from the transmitter will give the field intensity to be expected at that distance for 1 kw of power. The second part of the chart consists again of three scales: the microvolts per meter for 1 kw of power (a scale common with the first part), the effective radiated power in kilowatts, and the final field intensity in millivolts per meter. A straight line connecting the microvolts per meter for 1 kw with the effective radiated power in kilowatts will

Examples 1. The first example shown in Fig. 14-4b shows the range of an f-m broadcast station, provided by 20-kw effective radiated power at an antenna height of 500 ft, to the point at which the field intensity drops to 1 mv. In Ex. 1 we have a dotted line drawn between 20 kw and the 1-mv-per-meter field intensity. We note that this line intersects the microvolt plot for 1 kw at $224 \mu v$ per meter. Drawing a solid line between this $224 - \mu v$ point and the 500-ft transmitting-antenna height plot, we obtain a distance, or range, of 32 miles. Thus, we find that a station with an antenna height of 500 ft over the average ground and radiating 20 kw will give a 1-mv permeter signal at 32 miles, at the receiver-antenna height of 30 ft.

2. In Ex. 2 of Fig. 14-4c, we have the case of a transmitter site, 1,000 ft high, and the need for a 50- μ v (0.05 mv)-per-meter signal 70 miles away. The dotted line connecting 70 miles on the distance scale with 1,000 ft on the antenna-height scale indicates that a station with 1 kw of effective radiated power, operating at this height, would radiate 21.5- μ v per meter at a distance of 70 miles. Drawing a solid line between this 21.5- μ v point, the 0.05-mv-per-meter field-intensity point, and the power scale, we find that an effective radiated power of 5.5 kw would be required to produce the desired 50 μ v per meter at the point 70 miles distant with a transmitter-antenna height of 1,000 ft.



Fig. 14-4a. A range prediction chart for f-m broadcast stations.

intersect the field-intensity scale at the final field intensity at the receiving point.

14-3. Frequency-Modulation Broadcast Antennas. The f-m broadcast radiating system usually consists of an array of individual antennas, each antenna meeting a number of specific requirements. The radiation from each antenna has to be horizontally polarized



Fig. 14-4b. The graphical use of Fig. 14-4a for example 1.

to meet the requirements of the FCC; any vertically polarized energy transmitted will be wasted, inasmuch as the horizontally polarized receiving antennas will receive only horizontally polarized waves. Since it is preferable to install the radiating system on the highest site near the center of the area to be covered, the horizontal radiation pattern of each antenna should be omnidirectional. When installation in the center of the area to be covered is not possible, the horizontal radiation pattern should be such as to produce uniform radiation over the desired region. Each antenna should direct its radiated energy in the horizontal direction; the higher this directivity, the greater the effective radiated power and the fewer the antennas that need to be stacked in the final array to obtain the desired horizontal gain (the stacked array will be discussed later). Each antenna should be noncritical and easy to feed. It should be simple to adjust, preferably capable of adjustment at the factory before shipment so that no adjustments need be made upon installation.



Fig. 14-4c. The graphical use of Fig. 14-4b for example 2.

The electrical constants should be unaffected by changes in the weather; however, if they are affected by the elements, such as ice, then precautionary devices, such as de-icers, should be provided. The antenna should be light and sturdy to allow its being mounted on a tall mast; the taller the mast, the less power needed for the desired coverage.

Many antennas have been designed which meet the foregoing requirements. They may be divided into two categories, one consisting of variations of the horizontal-loop antenna, and the other of variations of the turnstile antenna.⁵ Antennas of both categories have been successfully used in practical installations.

14-4. The Loop Antenna. The horizontal-loop antenna, or "magnetic dipole" as it is sometimes called, has the proper omnidirectional horizontal radiation pattern, as well as some directivity to decrease the radiation of energy in the vertical direction. It can also be stacked easily in an array. However, the ordinary loop antenna, when used in the 100-megacycle band, has too low a radiation resistance to be useful. The ordinary loop antenna, as used in the lower-frequency bands, consists of a number of turns of wire around a loop whose dimensions are very small with respect to the wavelength at which it is being used. Actually, the total length of wire-

FREQUENCY MODULATION

from one terminal around the several turns and back to the other terminal — is usually negligible with respect to the wavelength. To meet this requirement in the 100-megacycle band even if only one turn were used, the loop dimensions would have to be so small that the radiation resistance would be practically too low in value to be used; it would be extremely difficult to feed and adjust.



Fig. 14-5. The current distribution, indicated by the height of the shading, around a loop antenna whose diameter is very small with respect to the wavelength at which it is being used.

Fig. 14-6. The current distribution around a loop antenna whose circumference is equal to a half wavelength at the frequency being used.

The problem of loop design is mainly one of current distribution in the radiating elements. In Fig. 14-5 is shown a loop antenna whose diameter is very small with respect to the wavelength at which it is being used. The magnitude of the current around the loop is indicated by the height of the shading around the circumference of the loop. A close approximation of the current distribution can be obtained by assuming that the distribution is the same as for a lossless transmission line. Inasmuch as this distribution follows a cosine curve, for very small values of loop diameter the current will be practically constant; this is illustrated in the diagram. The radiation pattern of this type of loop is omnidirectional in the plane of the loop. If the loop is situated horizontally, it will radiate with equal strength in all horizontal directions.

In the ultrahigh-frequency range, when the diameter of the loop in wavelengths or fraction of wavelength is increased to obtain a higher radiation resistance, the current no longer remains constant around the circumference of the loop and the radiation pattern begins to distort. For a diameter of about 20 in. at 100 megacycles, the current actually falls almost to zero at the input, as shown in Fig. 14-6. The pattern in the plane of the loop will no longer be omnidirectional.⁶

14-5. The Frequency-Modulation Circular Transmitting Antenna.⁷

In Fig. 14-7 is shown the evolution of the f-m circular antenna from an ordinary folded dipole. Figure 14-7a shows an ordinary folded



Fig. 14-7. The evolution of the f-m circular antenna from an ordinary folded dipole.

dipole. The current distribution is maximum in the center of the dipole and falls to zero at either end, as shown. The dipole can be shortened and the current made more uniform along the length by capacity loading on the ends of the dipole (Fig. 14-7b). The effect is similar to the effect of a hat (capacity loading with a top structure) on a vertical a-m broadcast band antenna. The capacity takes the place of a portion of the antenna, replacing the low current end segments. The elements of the shortened folded dipole are then bent into a circle, as shown in Fig. 14-7c. The two capacity elements now approach one another, forming an air capacitor where provision is made for adjustment of the capacity value. The current is not exactly uniform but is near enough to give a substantially omnidirectional pattern.

Figure 14-8 shows the constructional details of the final antenna. Normally, this type of antenna would be fed with a balanced line or, if a coaxial line were used, some type of a conversion transformer would have to be employed. Measurements made of the circular antenna, comparing the results obtained by feeding it directly with a coaxial line, grounding the other terminal, and the results obtained by feeding it with a conversion transformer, indicated that there is no practical need for including the transformer in the circuit.⁸ Hence,





one terminal of the input leads is grounded or shorted to ground (as shown by dotted line in Fig. 14-7) when a coaxial feed line is employed. With one terminal grounded, horizontal pattern measurements showed no evidence of redistribution of the currents in the important radiating elements.

In studying the final antenna, we find that the double current path serves two purposes. First, it permits a transformation of the radiation resistance to a terminal-resistance value in the general order of the nominal characteristic impedance of the coaxial transmission lines used to feed the antenna. Second, it permits direct mounting of the radiating system at a point of ground potential, thus eliminating disturbing capacity effects, which normally accompany insulated mountings, and securing protection against lightning. The adjustable capacitor permits tuning of the electrical circuit to resonance at the factory, upon receipt of the desired frequency of operation. An experimental horizontal pattern is illustrated in Fig. 14-9.

14-6. The Square-Loop Antenna. The square-loop f-m antenna is based on the ultrahigh-frequency loop antenna which is used for air navigation applications.⁹ The basic operation principles are illustrated in Fig. 14-10. Two small loops, small enough so that the current is quite uniform, are fed in the phase indicated in the figure by the plus and minus signs. The two plus terminals are connected together and the two minus terminals are connected together. The loops are semicircular and are placed back to back, as shown. The current flowing in the leg <u>BC</u> will cancel the effect of the current flowing in the leg <u>FE</u>. Points <u>B</u> and <u>F</u> and points <u>C</u> and <u>E</u> are connected together. The result is practically uniform current all around the loop.

This type of loop antenna can be fed by a coaxial line in the manner of a shielded loop wherein the shield acts as the antenna and the coaxial feed line is run up through the shield to a break in the continuity of the shield. This is illustrated in Fig. 14-11. It is called



Fig. 14-9. The experimental horizontal radiation pattern of the circular f-m antenna illustrated in Fig. 14-8.

a "two-element loop" and its circumference cannot exceed one wavelength. To increase the size of the loop, more elements are added. Figure 14-12 shows schematic diagrams of the three- and four-element loops. The circumference can be increased a half wavelength for each added element.

In the final design, a four-element loop was chosen, and, since the electrical characteristics of a loop do not vary as the shape is varied within reasonable limits, a square construction was employed. This type is illustrated in Fig. 14-13. The lines coupling the main feeder to the radiating elements are used as the matching device to match the antenna to the characteristic impedance of the feed line.



Fig. 14-10. The basic operation principle of the u-h-f loop antenna showing how two small loops are effectively placed back to back to create a larger loop.

It was found that the same sized mechanical structure can be used to cover all the frequencies in the f-m broadcast band. The



Fig. 14-11. The two-element u-h-f loop modified to permit coaxial line feed. The coaxial line is fed through the mast that supports the loop proper.

measured radiation pattern of the square-loop antenna for the f-m broadcast band is shown in Fig. 14-14.

14-7. The Cloverleaf Antenna.¹⁰ Another f-m antenna which creates an effective horizontal-loop antenna by the judicious placement together of four smaller loop antennas is known as the "cloverleaf antenna." The basic circuit for this type of antenna is shown in Fig. 14-15. It consists of four similar loops placed symmetrically about the point A like four pennies laid on a table with their centers forming a square. They just clear one another, so that no short circuits occur. The feed terminals of the loop are adjacent to A, and all four loops are fed in parallel. For convenience, a single line feed with a ground, or tower, return is utilized. Let us assume that the feed line is run up the center along A. All of the counterclockwise displaced terminals of the loops would be connected to the feed line at A and the remaining terminals grounded; actually, for symmetry, they are connected together at the tower support and connected to the tower. Thus, at any instant, the currents flowing through all four loops would be in phase, as indicated by the arrows

in the figure.

The currents flowing in the adjacent portions of the loops, as indicated by the two arrows in the right center of the figure, flow



Fig. 14-12. Three- and four-element u-h-f coaxial-fed loops. Each added element allows the circumference of the loop to be increased a half wavelength.

in opposite directions, thus canceling, and leaving only the effect of the current flowing in the outer portions. Hence, we have a symmetrical shape resembling a four-leaf clover which produces, in effect, a ring of uniform, in-phase current, the requisite of an omnidirectional loop radiation pattern.

Figure 14-16 shows the actual construction and installation of the cloverleaf antenna. The four elements, or individual.loops, are fed from a center 3-in. feed conductor, being fastened to it by means of clamps. This would be point <u>A</u> in Fig. 14-15. The opposite ends of the elements are connected to the tower, forming a tower return path for the radiating currents. The supporting tower itself is about 1 ft square. It is interesting to note that the heating elements for de-icing are run through the antenna proper, as shown.

Because of the displacement of the loops, the added length of the feed line for each loop, and the spurious currents which may flow in the vertical members of the tower, there will be a small amount of vertically polarized radiation. This does not hinder reception or interfere with the radiated horizontally polarized wave, but it does lead to a small waste of power -a field being radiated



Fig. 14-13. Constructional details of the square f-m loop antenna, a modification of the loop antenna shown in Fig. 14-12b.

which is not used for reception. This vertically polarized radiation is nullified, for that reason, by four small-diameter vertical cables

MEASURED HORIZONTAL PATTERN



Fig. 14-14. The measured horizontal radiation pattern of the loop antenna shown in Fig. 14-13.

connected between the cloverleaf elements, as shown in Fig. 14-16. 14-8. The Slot Antenna. The slot antenna¹¹, also called the "'pylon antenna," 12 is an ingenious method of creating horizontally polarized waves and approximately omnidirectional radiation. The



Fig. 14-15. Basic circuit of the cloverleaf antenna using a single feed line and a ground, or tower, return.

slot antenna, illustrated in Fig. 14-17, is a cylinder approximately 13 ft high and 19 in. in diameter with a narrow lengthwise slot cut from top to bottom. The cylindrical structure itself is the radiator. A single transmission line, running up the inside of the cylinder along the slot to the midpoint of its length, is used as the feed line.

The cylinder can be rolled from a single sheet of metal. It is capped on each end with a cast base that acts as an electrical short. These cast bases give the antenna great mechanical strength and provide a means of connecting the antenna to the supporting tower or to additional stacked elements.

The operation of the antenna can best be understood by considering first a thin horizontal section of the vertical cylinder (Fig. 14-18<u>a</u>). This thin section can be considered a horizontal loop antenna which is fed by the voltage existing across the opening of the slot. It is analyzed like an ordinary loop antenna and should not be made too big in diameter since, like the loop, if it is too large in diameter, an omnidirectional horizontal radiation pattern would not be obtained. The edges of the slot itself can be considered an openwire transmission line, shorted at the ends by the caps, and inductively loaded by the loops that comprise the cylinder.

The voltage distribution along the slot for a cylinder one wavelength long is shown in Fig. 14-18b. This is the voltage which is impressed on the radiating elements (the loops) all along the cylinder. Inasmuch as the voltage along a shorted transmission line is every-



Fig. 14-16. The construction and connections for the cloverleaf f-m transmitting antenna as installed on a tower for operation in an array.

where in phase, the loops are all radiating in phase — in an additive manner. This loop analysis is shown schematically in Fig. 14-19. The circumference of a horizontal section of the cylinder is chosen to be about one half wavelength in the f-m band. The individual loop elements are shown as dotted lines. When the currents all flow, they form, in fact, a veritable sheet of current flowing horizontally around the cylinder.

The horizontal radiation pattern of a pylon antenna is shown in Fig. 14-20. We see here how the radiation pattern varies as the frequency is increased. This is in accord with the theory of a loop antenna, the antenna becoming more directional as the frequency increases. Theoretically, for a perfectly omnidirectional horizontal pattern, the diameter of the cylinder should be a very small value in comparison to the wavelength being used; but, of course, this small diameter would produce a diminutive radiation resistance, while the actual variation in field strength around the antenna obtained by using a reasonable diameter is quite acceptable. Figure 14-21 shows the standing-wave ratio obtained on a 51.5-ohm line using a pylon antenna, curve <u>a</u> for a direct connection to the line, and curve <u>b</u> with a stub matching section. Hence, using only one size antenna, the whole f-m broadcast band can be covered. Mechanically, the antenna is a single-element self-supporting structure which is easy to erect. It is simply bolted to the building,



b

Fig. 14-17. The slot, or pylon, antenna, a cylinder with a narrow slot cut from top to bottom and capped at each end with a conducting cast base.

tower, or other structure by means of bottom flanges. Additional sections may be stacked on top by means of these same bottom flanges.

There are a number of advantages claimed for this type of antenna. A single antenna is quite light, each unit weighing about 350 lb. Ice problems are negligible because the transmission lines are run inside the cylinder where ice formation is unlikely. Moreover, there are no arms or other external elements that must be braced.

14-9. The Turnstile Antenna. The turnstile antenna is another type which has been used very successfully for the production of horizontally polarized radiation in an omnidirectional horizontal pattern.⁵ It consists fundamentally of two half-wave dipole antennas placed at right angles in the horizontal plane and fed with currents 90 deg out of phase with each other (Fig. 14-22). If the radiation patterns of the two dipoles are assumed sinusoidal in shape, the nominal pattern for a very short dipole in the horizontal plane, the radiation pattern of the turnstile is found to be perfectly omnidirectional for the horizontal plane. In the actual case, the radiation patterns of the half-wave dipoles are not perfectly sinusoidal in shape and the radiated field drops off about 10 per cent midway between antennas; it results in a slight "squaring" of the pattern, as shown in Fig. 14-23.



An improved type of turnstile antenna is known as the "super turnstile."¹³ This antenna (Fig. 14-24), uses current-sheet radiators as the radiating elements. They are not actually solid sheets but are constructed of tubing in a framework which closely approximates the effect of a solid sheet. This design aids in cutting down the wind resistance which would otherwise be very large. It is known that the use of large radiating elements increases the obtainable band width and, in the case of the super turnstile, the band width more than adequately covers the entire f-m band; no field adjustments have to be made.

The current sheets are fastened directly to the tower at the top and bottom, thereby creating a slot between the sheet and the tower. The transmission line is then connected so that its outer conductor is attached to one current sheet and the inner conductor attached to the tower. The connections are made to adjacent points on opposite sides of the slot. Thus, a voltage is created along the slot similar to the voltage along the slot of the pylon antenna. Hence, an in-phase current sheet is created along the radiating element. The sheet radiator is so shaped to obtain uniform radiation.



Fig. 14-19. A graphical representation of the slot antenna considered as being made up of a large number of circular elements.

The normal power gain of one element is 1.25. The tolerance on the input impedance is such that the voltage standing-wave ratio is 1.5 or less, when fed by a 51.5-ohm transmission line between the frequencies of 88 to 108 megacycles. The horizontal radiation pattern is essentially circular, deviating from round by less than 1/2 db.

14-10. Stacked Frequency-Modulation Antenna Arrays. The field strength set up by any f-m transmitter increases with the ef-

fective radiated power in the horizontal direction, as illustrated by calculations made on the chart of Fig. 14-4. The effective radiated



Fig. 14-20. The horizontal radiation pattern of a slot antenna for frequencies at either end of the f-m broadcast band.

power, however, is equal to the transmitter power times the effective power gain of the antenna system. In other words, it is best to concentrate as much of the transmitted power as possible into the horizontal direction, since power is wasted by radiating at vertical angles not utilized for reception.

The array used is a series of antennas stacked one above the other and fed with in-phase currents. Most of the radiated energy is then concentrated in the horizontal plane. This is illustrated by the three-dimensional radiation pattern shown in Fig. 14-25. Here, a stacked array of five cloverleaf antennas with a half wavelength between antennas is utilized. The effect of the ground plane is the same for all types of antennas, so that calculations are made for antennas remote from ground, as shown in the figure. We note here that nearly all the energy is concentrated in the horizontal plane. The power gain of this type of a stacked array is 5; the effective radiated power in the horizontal direction is five times the transmitter output power if we neglect losses in the feed lines and antenna loss resistance. A series of curves showing the power gain of a linear array of vertically stacked loops is shown in Fig. 14-26. The power gain



Fig. 14-21. The variation of the standing-wave ratio on a single-section slot antenna between the frequencies of 80 and 102 megacycles, a being the curve obtained by connecting the antenna directly to the transmission line and <u>b</u> the curve obtained when a stub-matching section is employed.

in the horizontal direction, for different spacings between the stacked antennas (the spacing being given in electrical degrees) and for different numbers of loops is given. Interpolation can be used for odd numbers of loops.

Example 3. Determine the effective radiated power of a 10-kw transmitter feeding a stacked array of six loops with a separation of 300 deg between antennas.

Referring to the chart in Fig. 14-26, we see that the power gain of a stacked array using six loops separated by 300 deg is about 6.5. Neglecting losses, we obtain 6.5 times 10, or 65 kw of effective radiated power in the horizontal direction. To take into account losses, we would have to subtract any feed-line and other losses and only multiply the actual radiated power by 6.5.



Fig. 14-22. A turnstile antenna consisting of two half-wave dipoles set at right angles to one another in the horizontal plane and fed with currents 90 deg out of phase with one another.

A picture of a two-antenna stacked array using circular antennas is shown in Fig. 14-27. This is the array used by WCBS-FM in New York City. 14-11. Frequency-Modulation Vertical Transmitting Antennas. Where vertically polarized radiation is desired, a vertical radiator



Fig. 14-23. The radiation pattern of a half-wave dipole turnstile antenna in the plane of the antenna.

has to be used as the transmitting antenna. Thus, in communication with mobile stations, a fixed-station transmitter usually uses a ver-



Fig. 14-24. A super-turnstile antenna using current sheet radiators.

tical antenna. Similar problems to those encountered in f-m broadcasting have to be met: the antenna has to have, in very many cases, an omnidirectional horizontal radiation pattern; it should be light and sturdy and capable of being mounted high on a tower or building; it should not be affected by variations in the weather or, if it is, precautions against this must be taken; and it should be efficient, and easy to adjust.

The simplest type of antenna for vertically polarized radiation in

FREQUENCY-MODULATION TRANSMITTING ANTENNAS 309

an omnidirectional horizontal pattern is the vertical monopole used so much in a-m broadcast transmitter installations. The antenna normally consists, in the case of a-m broadcast transmission, of a vertical



Fig. 14-25. The three-dimensional radiation pattern of a stacked array of five cloverleaf antennas. For clarity, only half of the complete three-dimensional pattern is shown.

radiator, about a quarter wavelength long, set directly on the ground. It is fed by a coaxial cable, the inner conductor being connected to



Fig. 14-26. A family of curves showing the power gain of a stacked array of loop antennas for different numbers of antennas and different spacings between loops.

the base of the vertical radiator, which is insulated from ground, while the outer conductor is grounded at the feed point. This type of antenna cannot be directly employed for f-m transmissions because, even at 30 megacycles, antenna height is very important for proper coverage without the use of excessive power and, since the wavelength is small, the antenna height would be small. A variation of the vertical monopole which may be mounted on a tower is shown in Fig. 14-28. One type uses a flat conductor which



Fig. 14-27. A view of the tower and loops in a stacked array consisting of two circular f-m loop antennas as used for f-m broadcasting by WCBS-FM in New York. simulates the ground plane. It consists of a circular piece of metal, with a radius of a quarter wavelength, attached to the shield of the

coaxial line at its termination. The radiator is then attached to the inner conductor at that point, as shown. This flat shild, really a col-



Fig. 14-28. A vertical monopole using a simulated ground plane which may be a circular conductor a quarter wavelength in radius or a series of wires a quarter wavelength long, placed around the base of the antenna and grounded to the supporting structure.

lar, prevents any currents from flowing down the antenna support, or outside of the feed line, and confines the currents to the radiating



Fig. 14-29. A vertical monopole using a quarter wavelength sleeve connected to the outer conductor at the feed point and open at the other end.

element. The flat shield, in turn, may be simulated by a series of wires equally spaced in the plane of the shield, thereby reducing weight and wind resistance. Thus, the final antenna may consist of a coaxial feed line, which also acts as a support, with the inner conductor extending as the vertical radiator, and a series of horizontal wires, at the termination of the outer conductor, acting as a simulated ground plane. Higher input impedances and increased power gain may be obtained by increasing the radiator length anywhere up to a half wavelength, but a matching network would have to be employed at the feed point for proper matching of the transmission line.

Another vertical antenna, which has the characteristics of a vertical dipole, is shown in Fig. 14-29. A sleeve is used, which surrounds the outer conductor of the coaxial line and extends for a quarter wavelength back down around the coaxial feed line. The outer conductor of the feed line is connected to the sleeve at the antenna feed point. The sleeve is then insulated from the outer conductor for the rest of its length and left open at the far end. The vertical radiator is then connected to the inner conductor at the feed point. The inner conductor current flows up the vertical radiator while the outer conductor current flows up the sleeve surface. Thus, the sleeve and the vertical radiator act as a dipole antenna, having all the characteristics of an ordinary dipole. At the bottom of the sleeve the open space between the sleeve and the outside of the coaxial line acts as a choke and keeps any r-f current from flowing down the supporting members. Variations of this type are very often used in the 30- to 40-megacycle f-m communication band.

If a directive antenna is desired for concentration of the transmitted energy to one side of the antenna, parasitic directors or reflectors may be employed. An antenna employing both a director



and a reflector is shown in Fig. 14-30. The reflector is a conductor slightly longer than a half wavelength placed, parallel to the antenna,
FREQUENCY-MODULATION TRANSMITTING ANTENNAS 313

a little more than one eighth wavelength in back of the antenna. It is adjusted by varying its length until the maximum, or desired, amount of directivity is obtained. The director is another conductor slightly less than a half wavelength placed, also parallel to the fed antenna, a little less than one eighth wavelength in front of the antenna, in the direction of desired directivity. The director can also be adjusted by adjusting its length. Thus, for a small amount of directivity, either a director or reflector may be used; the radiation



Fig. 14-31. Horizontal-radiation pattern of a vertical antenna alone, with a reflector, with a director, and with both.

patterns for these cases are shown in Fig. $14-31\underline{a}$ and \underline{b} . The amount of directivity obtained with either of these parasitic antennas is about the same. Figure $14-31\underline{c}$ is an approximate pattern of an antenna array consisting of both a director and a reflector used with a fed vertical antenna. The directivity is very much greater; therefore, this type of array is used only when a comparatively limited area has to be covered to one side of the antenna.

14-12. Transmission Lines and Feeding Problems. A discussion of transmission lines and feeding problems in detail would normally require a book in itself.¹ However, we might mention here the specialized problems encountered in f-m installations.

Any losses in the transmission line are deducted from the amount of power available to be radiated, thus reducing the field strength. Hence, the transmission line should be chosen not only for its powercarrying capacity but, even more difficult to obtain, also from the point of view of low losses. Impedance matching of the antennas to the lines also affects the efficiency of the station while, if an array is used, the relative gain is a maximum only if currents of the proper phase and amplitudes are fed to the individual loops.

The characteristics of the transmission-line system should not be allowed to vary from time to time, as in this case the radiation characteristics will continue to vary and seriously affect the gain. Beaded transmission lines, very much favored because of their lowloss characteristics, are greatly affected when moisture seeps into the air dielectric within the line; a short circuit may even occur if the moisture should happen to collect on the surface of one of the separator beads. To overcome this defect, special equipment which pumps dry air or inert gas nitrogen through the transmission lines is sometimes used. To be effective, however, the lines must be leakproof. This requires carefully designed end seals and connectors to maintain airtight joints with proper line matching and at the same time allowing for expansion and contraction caused by changes in ambient temperature and r-f heating.



Fig. 14-32. A schematic diagram of the transmission lines used to feed a fourbay f-m loop antenna. The various surge impedances \underline{Z}_0 for proper matching are indicated.

Figure 14-32 shows how an array of four horizontal loops are fed to maintain equal in-phase currents in all of the loops and still match the input line. Each pair of loops is fed with 100-ohm transmission lines which are matched to the antennas by means of stubs. Midway between the two antennas the 100-ohm transmission lines are parallel-connected, creating a 50-ohm impedance. This junction is matched by using a 50-ohm feed line. The two 50-ohm transmission lines from two pairs of antennas are then connected together, creating a 25-ohm input impedance. Since 25 ohms is too low a value to use in the feeder, the 25 ohms is transformed by using a quarter-wave transformer. This is a quarter-wavelength piece of transmission line with a 35-ohm impedance (35 being the square root of 25 x 50). Thus the main coaxial feeder of 50-ohms characteristic impedance is matched, and, since the connections are perfectly symmetrical, all the antennas receive equal in-phase amplitudes.

Another way of obtaining the same result is to use separate transmission lines to a common junction and there use a transformer to match the main feed line. The individual lines, however, must either all be the same length, or differ by an integral number of half wavelengths. Care must be taken in this case to avoid reversals in phase in the radiated wave by reversing the connections to the antenna should the length differ by an odd number of half wavelengths.

Sometimes it is desirable to operate both the a-m and the f-m antennas on a common structure, placing the f-m broadcast antenna on top of the a-m antenna.¹⁴ For shunt-fed a-m tower antennas, which are occasionally encountered in practice, there is no problem other than the structural requirements imposed on the radiator by the addition of a top in the form of the f-m antenna. Since the base of such an antenna is grounded, the f-m feed line can be connected to ground on the horizontal run from the station and then fastened directly to the tower on the vertical run. The only important effect on the a-m station operation will be the modified input impedance caused by the increased height; this can generally be corrected by minor tuning adjustments.

The most commonly encountered a-m antenna, however, is a series-fed antenna, a bottom-fed tower insulated from ground at the base. An f-m feed line cannot be connected directly to the ground on the horizontal run and to the tower on the vertical run without causing a short circuit in the base insulator. Therefore, some means of isolating the two systems at the base insulator is required. Any scheme of isolation must eliminate the short circuit across the base insulator by raising the shunt impedance — the impedance to currents flowing across the base insulator by the path of the f-m feed line — to a high value compared with the base impedance of the a-m radiator itself.

Two methods of accomplishing this result are shown in Fig. 14-33. In Fig. $14-33\underline{a}$, the feed line for the f-m antenna is run along the ground until it reaches the a-m tower. It is then run up the tower on insulators for a distance of one quarter of the wavelength at the a-m frequency and then connected to the tower proper. For instance, if the a-m frequency is 1,000 kc, the distance would be one quarter of 300 m, or 75 m (about 245 ft). Thus, there exists, between the

tower and ground, a quarter-wavelength shorted transmission line – a very high r-f impedance — which prevents a short circuit across



Fig. 14-33. Two methods of installing an f-m antenna on top of a series-fed a-m tower without shorting the tower to ground for the a-m signal.

the base of the tower. The outside of the f-m cable acts as one side of the line while the tower acts as the other.

Figure $14-33\underline{b}$ shows another way of installing the transmission line to create the effect of a high impedance across the base of the tower. The f-m transmission line is run down the tower and its outside shield is tied to it electrically all the way down. At the base, the shield is insulated from ground for a distance of one quarter of the a-m wavelength and then tied to ground. Thus, again the outside of the shield and ground form a quarter-wavelength shorted transmission line across the base, thus preventing an r-f short of the a-m signal.

REFERENCES

1. For a detailed discussion of this subject as well as a fundamental discussion of ultrahigh-frequency antennas, see Nathan Marchand, <u>Ultrahigh Frequency Transmission and Radiation</u>, New York, John Wiley & Sons, Inc., 1947.

^{2.} See F. E. Terman, <u>Radio Engineers' Handbook</u>, New York, McGraw-Hill Book Company, Inc., 1943, pp. 674-769.

^{3.} George H. Brown, "Vertical vs. Horizontal Polarization," <u>Electronics</u>, Vol. 13, October, 1940, p. 20.

FREQUENCY-MODULATION TRANSMITTING ANTENNAS 317

4. Frederick C. Everett, "Range Prediction Chart for F-M Stations," Communications, November, 1945, p. 57. This chart was prepared from data supplied by the FCC in their "Standards of Good Engineering Practice," with 98 megacycles and horizontal polarization used as a basis of computation. It uses the effective radiated power which is the power of the transmitter times the antenna-power gain due to directivity.

5. George H. Brown, "The Turnstile Antenna," Electronics, April, 1936, p.14. 6. J. B. Sherman, "Circular Loop Antennas at Ultra-High Frequencies," Proc. IRE, September, 1944, p. 534.

7. M. W. Scheldorf, "F-M Circular Antenna," Gen. Elec. Rev., March, 1943.

8. A. G. Scheldorf, "Circular Antennas," Sixth Annual Conference of Broadcast Engineers, Columbus, Ohio, 1946.

9. A. Alford and A. G. Kandoian, "UHF Loop Antennas," Trans. AIEE Supplement, Vol. 59, 1940, p. 843.

10. L. C. Tyack, "The Clover-Leaf," Western Electric Oscillator, April, 1946.

11. Charles R. Jones, "Slotted Tubular Antenna," Communications, July, 1946, p. 36.

12. Robert F. Holtz, "The Pylon Antenna," RCA Broadcast News, October, 1946.

13. R. F. Holtz, "Super Turnstile Antenna," Sixth Annual Conference of Broadcast Engineers, 1946.

14. Robert F. Holtz, "Isolation Methods for FM Antennas Mounted on AM Towers," RCA Broadcast News, October, 1946.

QUESTIONS

1. Why are the discussions of propagation for broadcast and communication purposes usually confined to plane-wave propagation?

2. Discuss the relationships between the electric vector, the magnetic vector, and the direction of propagation that exist in an electromagnetic wave.

3. What takes place in an electromagnetic wave when it is reflected from a perfectly conducting surface? Discuss the effect of polarization.

4. Explain refraction of an electromagnetic wave. When does it take place?

5. Describe the different ways that an electromagnetic wave can travel from the transmitting antenna to the receiving antenna.

6. Discuss electromagnetic-wave propagation in the f-m bands.

7. Why is horizontal polarization employed in the f-m broadcast band?

8. Why do mobile f-m installations often employ vertical polarization?

9. Using the chart shown in Fig. 14-14, determine the distance at which the field intensity of an f-m transmitter, radiating an effective power of 50 kw at an antenna height of 300 ft, drops to $50 \mu v$.

10. What are the requirements of an f-m broadcast antenna?

11. Why cannot an ordinary loop antenna be used for an f-m broadcast antenna?

12. Describe the operation of the f-m circular transmitting antenna.

13. Explain the principles of operation of the square-loop antenna.

14. How does the cloverleaf antenna function?

15. Show how the slot antenna creates a horizontally polarized field.

16. What determines the voltage distribution along the slot of a slot antenna? 17. Explain the creation of an omnidirectional horizontal pattern for horizontally polarized waves by a turnstile antenna.

18. What is the super turnstile?

19. What is accomplished by stacking f-m antennas?

20. What is the effective radiated power of a 4-kw transmitter feeding a stacked array of four loops with a separation of 240 deg between loops?

21. Explain the operation of the vertical monopole mounted on a tower and using a simulated ground plane.

22. Explain the operation of a sleeve on a vertical radiator which prevents r-f currents from flowing down the supporting mast.

23. What should be the characteristics of a transmission-line system used to feed an f-m antenna array?

24. How should an f-m antenna, installed on top of an a-m series-fed tower, be fed to prevent short-circuiting the a-m signal?

CHAPTER 15

Frequency-Modulation Receiving Antennas

15-1. Fundamental Characteristics of a Receiving Antenna. The purpose of a receiving antenna is to intercept the desired electromagnetic wave, abstract energy from it, and deliver it (usually via a transmission line) to the receiver input. At ordinary a-m broadcast frequencies the antenna installation is usually quite simple because of the comparatively low frequencies involved. A short length of wire, or a built-in loop antenna, or even metal objects insulated from ground - copper screens, bed springs, and similar devices - have sufficient pickup for satisfactory operation of ordinary highly sensitive a-m broadcast receivers. In the u-h-f range, in fact, anywhere above 30 megacycles, the problem is not so simple; carefully designed antennas and installations should be employed to obtain the best results (or sometimes even just acceptable results) from the f-m receiver. To understand the reasons for this statement. we will first review some of the fundamental characteristics of a receiving antenna.

The actual operation of a receiving antenna can best be understood by studying the equivalent circuit of the receiving antenna.



Fig. 15-1. A dipole receiving antenna feeding into a load impedance $\underline{Z}_{\underline{L}}$, the effective impedance present across the terminals of the antenna.

Let us consider the ordinary dipole antenna shown in Fig. 15-1. This antenna may be feeding directly into the receiver, or —the more common case —it may be feeding into a transmission line which leads to the receiver. In either case there will be an effective load, which we have called \underline{ZL} in the figure, across the terminals of the antenna. The antenna, intercepting the transmitted field, will act as

a generator with a voltage of, let us say, \underline{V}_0 . This voltage will cause a current to flow through the load $\underline{Z}_{\underline{L}}$. But since the current has to flow from the antenna, it also has to flow through the internal impedance of the antenna. Calling this internal impedance \underline{Z}_0 , we



obtain as the equivalent circuit diagram the circuit shown in Fig. 15-2.

Thus, we see that the voltage picked up by the antenna is fed into two impedances in series, the load impedance and the internal impedance of the antenna. The load impedance $\underline{Z}_{\underline{L}}$ can be measured directly and, since it is usually a matched transmission line, it is often the characteristic impedance of the transmission line leading to the receiver. For instance, if a 72-ohm transmission line is used to connect the antenna to the receiver and is matched at the receiver end, the load impedance would be 72 ohms. The internal impedance, since it is the impedance to current flowing through the antenna, is the impedance that the antenna presents when used as a transmitting antenna. Thus, a half wave dipole, which has a radiation resistance of just about 72 ohms, presents (if we neglect losses) an impedance of 72 ohms when used as a transmitting antenna. This same antenna, when used as a receiving antenna, will have an internal impedance of 72 ohms.

The only remaining factor to be determined is the voltage \underline{V}_0 which is picked up by the antenna. This voltage may be written as a product of two factors, the field intensity \underline{E} of the electromagnetic wave at the receiving antenna, and a factor which is called the "effective height" of the antenna \underline{h}_e , thus,

$$\underline{V}0 = \underline{Eh}\underline{e} \tag{15-1}$$

The effective height, sometimes called the "effective length," of the antenna is really the main determining factor in the use of the antenna as a receiving antenna. It is a factor determined by the antenna size and configuration, although sometimes its derivation becomes very complex even for what would normally be called "simple" antennas.¹ Let us, however, consider the common case of a half-wave dipole — an antenna, fed in the center, which has an over-all length of one half wavelength at the frequency being received. For this type of antenna, when the wave is received parallel to the antenna, the effective height is given by

$$\underline{h}_{\underline{e}} = \frac{\lambda}{3.14}$$
(15-2)

where λ is the wavelength of the electromagnetic wave being received.

Thus, for a half-wave dipole, the effective height at 10 megacycles where the wavelength is 30 m would be 9.6, while at 100 megacycles where the wavelength is 3 m the effective height drops to 0.96. This may be explained by the fact that the half-wave antenna at 100 megacycles is only one tenth the length of a half-wave antenna at 10 megacycles and intercepts only one tenth the amount of energy. As we go up in frequency, the tuned lengths that are used as antennas decrease in size, and the voltage picked up drops proportionately. For this reason greater care must be used in the construction and installation of antennas for the u-h-f bands.

Example. Determine the voltage set up across the terminals of a 100-ohm characteristic-impedance transmission line by a half-wave dipole at 100 megacycles. The electromagnetic-field intensity parallel to the antenna is $50 \,\mu v$ per m and the 100-ohm transmission line is matched at the receiver input.

For this problem we can employ the equivalent circuit shown in Fig. 15-2. \underline{Z}_0 , the internal impedance of the dipole, can be taken as 72 ohms. $\underline{Z}_{\underline{L}}$, the input to a matched transmission line, is the characteristic impedance of the transmission line, 100 ohms. The effective height, as given by Eq. 15-2 for a wavelength of 3 m, is 0.96. Substituting into Eq. 15-1, we obtain

$$\underline{V}_0 = 50 \times 0.96$$

or

 $\underline{V}_0 = 48$

$$\underline{\mathbf{V}}_{\underline{\mathbf{L}}} = \underline{\mathbf{V}}_{0} \frac{\underline{\mathbf{Z}}_{\underline{\mathbf{L}}}}{\underline{\mathbf{Z}}_{\mathbf{L}} + \underline{\mathbf{Z}}_{0}}$$

where $\underline{V}\,\underline{L}$ is the voltageacross the input to the transmission line. Substituting into this equation, we obtain

$$\underline{V}\underline{L} = 48 \quad \frac{100}{100 + 72}$$

so that

 $\underline{\mathbf{V}}_{\underline{\mathbf{L}}} = \mathbf{28} \mu \mathbf{v} \quad \underline{\mathbf{Ans}}.$

The receiving antenna may also have directional properties. The radiation pattern of the antenna, or antenna array, is applicable when the same antenna, or antenna array, is used for the reception of electromagnetic waves. In the case of the receiving antenna, however, the radiation pattern shows the amount of voltage set up in the antenna for electromagnetic waves coming from different



Fig. 15-3. The directive pattern of a half-wave dipole receiving antenna. The voltage received from a wave coming from the direction 1 is <u>AO</u> and from the direction 2, if the field intensity remains the same, is <u>BO</u>.

directions. For instance, Fig. 15-3 shows the radiation pattern of a half-wave dipole when used as a transmitting antenna. When using this antenna as a receiving antenna, it indicates the voltage that would be induced into the antenna by electromagnetic waves coming from different directions. A wave coming from direction 1 would induce a voltage of magnitude <u>AO</u>. If the orientation of the antenna should be so changed that the electromagnetic wave were coming from direction 2, even though the field intensity at the antenna should remain the same, the voltage induced into the antenna would drop to the magnitude <u>BO</u>, about half of the voltage <u>AO</u> in this case. Thus, the radiation pattern of the receiving antenna shows its directivity properties much in the same manner as a transmitting antenna radiation pattern shows the directivity properties of a transmitting antenna.

It may be mentioned here that this reciprocity between transmitting and receiving antennas extends to other factors. If the antenna is matched to a transmission line as a transmitting antenna, it will be matched to the same transmission line as a receiving antenna for the same frequency. Its efficiency of operation will be the same for both reception and transmission. Hence, if the characteristics of an antenna, or antenna array, are known for its use in the transmission of electromagnetic waves, we can deduce quite accurately its behavior as a receiving antenna.

15-2. Receiver-Antenna Requirements for Frequency Modulation. The f-m broadcast receiving antenna is a u-h-f antenna used to cover the 88- to 108-megacycle band. This is a broad band to be

covered by one antenna, but there are quite a few different types of antenna used successfully in the field. The antenna should be sensitive and maintain as good a match to the transmission line as possible over the whole band. There are special cases, of course, where the signal strengths are so large that even an odd piece of wire attached to the terminals of the receiver will yield good reception, but these cases are the exception rather than the rule. Normally, the antenna has to be reasonably broad-band to obtain noise-free reception from stations available in the area.

By "broad-band" is meant an antenna that not only has a reasonably good effective height in the f-m broadcast band but also delivers most of its received energy to the receiver over the whole band. By referring back to the equivalent circuit in Fig. 2, we see that this means that the internal impedance of the antenna \underline{Z}_0 should not become excessive anywhere in the band; otherwise, it will absorb all the received energy and none will be available across the load. The maximum power is transferred to the load when the internal impedance is equal to the load impedance, a condition under which the antenna is said to be matched. Thus, broad-band reception often becomes a problem of maintaining a good match between antenna and load over the whole band.

The other requirements, from a practical standpoint, are simplicity, ease of installation, ruggedness, and weatherproofness. To meet the demands of most f-m broadcast receiver owners, the antenna has to be reasonably inexpensive. Only in very rare cases would an expensive installation be tolerated. To keep down the manufacturing costs, and thus the sale price, the antenna construction has to be simple. This will also simplify the installation. By facilitating the installation, both in equipment and in adjustment, it is possible not only to reduce the cost of the installation, but also to avoid special test equipment and costly adjustments after installation. A receiving antenna should operate for years under ordinary conditions without any maintenance or adjustments. Preferably, it should not be affected by the weather; under no conditions of weather should it be rendered inoperative.

A great deal depends upon the care with which the antenna is installed, its mounting, and its location. The problems here are no different than with the installation of an antenna for any type of modulation, which requires the same precautions. Common sense is by far our best ally for these factors.

15-3. Receiving Antennas for Frequency Modulation. As mentioned previously, where the field strength of the received signal is very high, inefficiency in the receiver antenna may be tolerated without impairing the received signal. In these cases the field intensity is so high that, even though the efficiency is poor and transfer of energy to the receiver is inefficient, enough power reaches the receiver to operate it satisfactorily. Cases like this are encountered when the receiver is close to the sites of the transmitting antennas. An ordinary thin-wire half-wave dipole with a directional pattern, as shown in Fig. 15-3, can often be utilized successfully as the receiving antenna under these conditions. To receive stations over the whole f-m broadcast band of frequencies, it should be tuned to about 98 megacycles. A thin dipole, however, is very narrow-band, which means that the reception efficiency at other than the frequency to which it is tuned will be low.

The normal installation, however, requires a broad-band antenna. The simplest way to obtain a broad-band match is to increase the size of the conductors in a half-wave dipole. This will decrease the resonant impedance of the antenna but will keep it constant over a



Fig. 15-4. A broad-band f-m dipole utilizing large-diameter thin-walled pipes as its elements.

greater band width. A broad-band dipole is illustrated in Fig. 15-4. A thin-walled copper pipe, the thin wall being used only for weight reduction (since the thickness of the wall of the pipe after the first few thousandths of an inch has no effect on its behavior at 100 mega-cycles), is used as the elements of the antenna. Diameters as small as 1 in. have been found to operate quite satisfactorily for ordinary reception, although larger diameters are often used for better matching purposes.

Instead of the pipe, flat plates or cones may be used to obtain excellent results.² To decrease the wind resistance, the solid sur-



Fig. 15-5. A broad-band f-m cone antenna where the cone is simulated by six wires forming the cone surface.

faces may be replaced by a series of wires. For instance, Fig. 15-5 shows a broad-band antenna simulated by six wires tracing the surface of a cone. The length of the wires is about 38 in, each for

reception in the f-m broadcast band, and the diameter of the base of the cone need only be about 3 in. For greater band widths, the diameter of the base of the cone should be increased. Sometimes the wires are merely fanned out in the horizontal plane, as indicated



Fig. 15-6. A broad-band f-m receiving antenna consisting of a number of wires fanning out from the two terminals and simulating a conducting surface.

in Fig. 15-6, and simulate a flat conducting surface. Four wires are shown in the figure which fan out about 4 in., about 1 in. between wires, and create a broad-band effect quite sufficient for reception over the whole f-m band. More wires, fanning out to larger dimensions, may be used to increase the broad-band effect of the antenna.

In all the antennas employing large diameter elements, or simulated large diameters, care should be taken to ascertain, either from the manufacturer or by measurements, the transmission line impedance to be used with the antenna for maximum efficiency of operation. Mismatching an antenna means a needless loss in sensitivity.

The folded dipole is another type of broad-band antenna which can be used in the f-m broadcast band. The folded dipole, as shown



Fig. 15-7. A folded dipole consisting of a halfwave dipole shunted by another half-wave conductor slightly displaced from it. Point <u>A</u> is at ground potential and may be used to mount the antenna.

in Fig. 15-7, consists of a half wave dipole shunted by another half wave conductor slightly displaced from it. Thus, it really is two antennas, one being fed from the conventional center and the other being fed from the ends of the first. The input impedance of this type of antenna is about four times the impedance of an ordinary dipole, which makes it about 290 ohms. Other impedances may be obtained by choosing different diameters for the antenna and the shunting conductor. The broad-band feature of the folded dipole makes it usable over the whole f-m broadcast band when it is tuned to resonance at about 98 megacycles An ordinary piece of open 300-ohm solid dielectric transmission line shorted at both ends with the output taken across the center of one side makes a good folded dipole for the f-m band. (This is essentially the arrangement shown in Fig. 15-7.)

The directive effects of a dipole may be eliminated by bending the two elements together to form a V or U shape with the output at the base. Care again must be taken to ascertain the new characteristic impedance of the antenna since changing its configuration also changes its input impedance. V's with an angle of about 75 deg between legs and U's with a base of about one half the length of the U have satisfactory omnidirectional directivity patterns. The folded dipole may also be bent into a V or U shape to make it omnidirectional.

Single-frequency installations are used in point-to-point or mobile installations. Since the broad-band feature is not necessary, ordinary thin-wire antenna installations are used.

In those cases where directivity is desired, the reflector and director, as discussed in a previous chapter, may be employed to advantage. The directivity pattern is the same for reception as for transmission so that the transmission directivity patterns given for the reflector and director may be directly applied to the antenna array as a receiving antenna.

15-4. Receiving-Antenna Installations. The ordinary antenna installation uses a twisted-pair transmission line leading from the antenna on the roof to the receiver at its point of installation. This



Fig. 15-8. An f-m receiver antenna installation using a twisted-pair lead-in wire as the transmission line from the antenna to the receiver.

To ground connection

is illustrated in Fig. 15-8. The twisted pair is used to avoid, as much as possible, extraneous signal pickup on the way down to the receiver. In cases where the noise pickup is very bad, a shield is used to surround the two lead-in wires. Special low-loss dualconductor shielded transmission lines may be purchased for this application. The cost, of course, is higher than that of an ordinary twisted pair. When a shield is employed, it should be connected to the ground terminal at the receiver and, if possible, grounded near the antenna. The latter ground is merely a precaution to keep r-f currents off the shield and may be dispensed with in some cases.

Sometimes it is desirable to use a coaxial transmission line from the antenna to the receiver. For a single-ended antenna, like the monopole antenna, the inner conductor is connected to the antenna while the coaxial shield is grounded. At the receiver, if the input is also single ended (between a single terminal and ground) the connections will also be very simple. The inner conductor is connected to the antenna input terminal while the outer shield is connected to the ground terminal. It is also advisable to connect the ground terminal to a good ground connection.

Let us now consider the case of a coaxial line installation when using a balanced antenna, such as a dipole and a receiver with a balanced input. This type of installation is inferior to a shielded balanced-line installation as discussed previously, but its cost is



Fig. 15-9. An f-m receiver-antenna installation using a coaxial transmission line with a dipole antenna and a balanced-receiver input.

usually a great deal less. Figure 15-9 shows a schematic diagram of the installation. We avoid the use of complex elements and structures to keep down cost and also keep the installation wide-band. At the balanced dipole, one side is connected to the inner conductor while the other side is connected to the shield. At the receiver, where the input is balanced, the inner conductor is connected to one of the antenna input terminals while the shield is connected to the other input terminal. This latter terminal is also grounded, converting the input to a single-ended circuit. The ground terminal should also be connected to a good ground connection. Although for a balanced antenna to balanced receiver connection this installation is not as efficient as the shielded twisted-pair or balanced-line installation, it does decrease the noise pickup.

Sometimes, if there is a long run of transmission line (like from the lower floors to the roof of some of the larger houses) the losses in the transmission line may make the ordinary type of installation impractical. One solution is to install the f-m receiver antenna outside a window near the receiver. It is best to use a window that is not hemmed in on all sides by other buildings. An installation using a





Fig. 15-10. An f-m receiver-antenna installation outside a window, thus eliminating the necessity for a long transmission line.

Fig. 15-11. An under-the-rug foldeddipole f-m receiver-antenna installation.

folded dipole is shown in Fig. 15-10. The supporting member should keep the antenna approximately a quarter wave away from the side of the building, about 2 1/2 ft away at 100 megacycles. This type of installation is cheaper and very often more efficient than an installation using a long length of transmission line. The noise pickup, because of its proximity to the noise sources, is usually higher. With a high signal field strength, however, this latter consideration is of little consequence.

Another type of installation which is cheap, simple and usually works well in unshielded installations is the under-the-rug installation illustrated in Fig. 15-11. By an unshielded installation is meant an installation in a building where not much metal has been used in its construction; thus, the f-m waves can penetrate through to the point of installation. Pickup of any extraneous signals within the building could not be avoided, so that another consideration would be a low interference level within the building. Its sensitivity would not be as high as an antenna installed on the roof but, in many cases, will prove more than adequate. The antenna need not be placed under the rug, but may be installed behind the receiver or other furniture or in any other convenient place.

REFERENCES

1. Ronald King and Charles W. Harrison, Jr., "The Receiving Antenna," Proc. IRE, January, 1944, p. 18.

2. S. A. Schelkunoff, Electromagnetic Waves, New York, D. Van Nostrand Company, 1943, Chap. XI.

QUESTIONS

1. Describe the equivalent circuit of a receiving antenna.

- 2. What determines the internal impedance of a receiving antenna?
- 3. Explain what is meant by the effective height of a receiving antenna.
- 4. How does the effective height of an antenna vary with frequency?
- 5. Why do u-h-f antennas have to be carefully designed and installed?

6. An antenna with an effective height of 0.80 is placed in an electromagnetic field of $30 \mu v$ per meter. Determine the voltage across a load impedance of 50 ohms placed across the terminals of the antenna if the internal impedance of the antenna is 70 ohms.

7. What is the relationship between the directional properties of a receiving antenna and its directional properties when used as a transmitting antenna?

- - 8. What is meant by matching a receiving antenna?
- 9. Discuss the receiving antenna requirements for an f-m antenna.
- 10. When can an inefficient receiving antenna be used?
- 11. Why are broad-band antennas employed for f-m broadcast reception?
- 12. How can a dipole be made broad band?
- 13. What is a folded dipole?

14. Describe an f-m dipole installation using a shielded balanced-transmission line.

15. Why should long transmission lines between the transmitter and antenna be avoided if possible?

16. Describe an antenna installation used to avoid a long transmission line between receiver and antenna when the distance to the roof of the building is very great.

CHAPTER 16 Mobile Frequency-Modulation Equipment

16-1. Basic Mobile-Communication Systems. The various types of mobile-communication systems may be separated into three basic groups: one-way, two-way, and three-way systems. The one-way system is, as the name implies, a one-way signaling service, or, more accurately, a broadcast service to mobile units. A dispatcher at the central transmitter broadcasts instructions which are picked up by receivers installed in the mobile units. This



Fig. 16-1. One-way mobile-communication system; the mobile units are equipped only with receivers.

is illustrated in Fig. 16-1. The mobile units are equipped only with receivers and are expected to follow the instructions given by the dispatcher. Since no immediate means of replying is provided for the mobile unit operator, he is not expected to reply to, or communicate with the dispatcher in most cases. Where a reply is necessary, the mobile unit operator has to leave his station and use the nearest telephone or other available means of reply communication. Although the latter factor is disadvantageous at some times, the one-way system provides a basic dispatching service with the minimum of equipment: a central station transmitter for the dispatcher and only receivers for the mobile units.

The two-way mobile communication system is used where the operator of the mobile unit has to be able to communicate immediately with the operator at the fixed station. The mobile units are thus equipped not only with receivers but also with transmitters that provide the mobile-unit operator with two-way communication between himself and the central station. This system is illustrated





in Fig. 16-2. This system requires a transmitter-receiver installation in each of the mobile units as well as at the central station. The general two-way telephone service between any vehicle and any wire telephone or other mobile unit may be included in this group. The central station may thus provide a means for tying into the normal telephone lines. The two-way mobile-communication system is necessary in all cases where the mobile-unit operator has to communicate with the central station without leaving his unit or without halting its progress.

The three-way system provides, in addition to all the conveniences of the two-way system, direct communication between the mobile units, without the necessity of going through the central station. The equipment is the same as for the two-way system except that two-frequency operation is necessary where communication between mobile units must not interfere with central station communications.

In any of these communication systems used in police, fire, or other emergency dispatching service, the mobile receiver is usually on as long as the vehicle is in service while the central station transmitter is on only for the duration of the broadcast, perhaps only for a few seconds at a time at widely spaced intervals. Since the receivers are usually very sensitive, the in-between-signal noise, the blast of noise heard when there is no signal from the transmitter, has to be muted. The signal from the transmitter has a quieting effect because it increases the a-v-c voltage, cutting down the sensitivity and thereby decreasing the noise level. Hence, nearly every receiver used under these conditions employs some type of "squelching" circuit to mute the output when there is no transmission from the central station, or other source, to be received.

Another feature which is sometimes incorporated into a mobile-

communication network is the selective call system. When the dispatcher so desires, all receivers are muted except the one he desires to communicate with. Thus, with the selection of a proper code



Fig. 16-3. A circuit diagram of a double-superheterodyne mobile f-m receiver including the vibrator power supply and the squelch circuit.

signal by means of a push button or dial, the central station operator can speak to a single mobile unit, a group of mobile units, or all the mobile units in his area. Hence, he is heard only by those mobile units which are concerned and he does not bother the others with messages that do not concern them. Another advantage of a

MOBILE FREQUENCY-MODULATION EQUIPMENT

selective call system is the removal of false signals or blasts of noise. Sometimes a station on an adjacent channel will spill over just enough to open the squelch circuit but with not enough signal



to quiet the noise. The result is a very disturbing blast of noise. With a selective call system, however, this false removal of the squelching action would not occur, since the receiver would not open up unless the proper code signal were received from the central station.

333

16-2. Mobile Transmitters and Receivers. Figure 16-3 shows a circuit diagram of a double-superheterodyne receiver, including the vibrator power supply and the squelch circuit.¹ As stated in a



Fig. 16-4. A circuit diagram of a mobile f-m transmitter using crystal-confollowing page.

previous chapter, the principal advantage of the double superheterodyne lies in the large gain obtainable with minimum tendency towards interstage oscillation. This receiver operates in the 30to 40-megacycle mobile-communication band and is a straightforward f-m receiver except for the inclusion of the squelch circuit. The signal picked up by the antenna is amplified by a 7C7 r-f amplifier and fed into a 7C7 first mixer. In the first mixer, the signal



trolled phase-to-frequency modulation. See the schematic diagram parts list on the

is combined with the fourth harmonic of the high-frequency oscillator, a crystal-controlled 7C7 tube. Crystal control can be used because the mobile receiver is set to receive only one frequency, that of the central station transmitter. The resultant beat signal, 4.3 megacycles, is amplified by the 4.3-megacycle first i-f am-

FIG. 16-4

SCHEMATIC DIAGRAM PARTS LIST

Diag. No.	Description	Diag. No.	Description
C1	Molded Mice Cond (5000MMED 200V) 100		
C2	Molded Mice Cond. (5000MMTD-300V.) 10%	RI	Carbon Resistor (470,000-1/2-10) Ins.
C3	Molded Mice Cond. (100MMFD-400V.) 10% *	RZ	Carbon Resistor (1000-1/2-5) Ins.
C4	Molded Mice Cond. (100MMFD-400V.) 10%	RJ	Carbon Resistor (10,000-1/2-10) Ins.
C5	Molded Mica Cond. (2000MMFD-300V.) 10%	R4	Carbon Resistor (47,000-1/2-10) Ins.
CG	Molded Mice Cond. (100MMFD-400V.) 10%	K5	Carbon Resistor (47,000-1/2-10) Ins.
C7	Molded Mice Cond. (TOUMMFD-400V.) 10%	R6	Carbon Resistor (100-1/2-5) Ins.
CR	Float Cond (20MED 25V)	R/	Carbon Resistor (4,700-1/2-10) Ins.
<u> </u>	Tub Cond & Bracket (05MED 600V)	R8	Carbon Resistor (330-1-10) Ins.
C10	Tub. Cond. & Bracket (OSMFD-600V.)	R9	Carbon Resistor (100,000-1/2-10) Ins.
CII	Molded Mice Cond (100MMED 400V) 100	RIU	Carbon Resistor (100,000-1/2-10) Ins.
C12	Molded Mica Cond. (100MMFD-200V.) 10%	R11	Carbon Resistor (33,000-1/2-10) Ins.
C13	Molded Mica Cond. (2000MMFD-500V.) 10%	RIZ	Carbon Resistor (33,000-1/2-10) Ins.
C14	Molded Mica Cond. (2000MMFD-500V.) 10%	RIJ	Carbon Resistor (220,000-1/2-10) Ins.
C15	Molded Mica Cond. (2000MMFD-500V.) 10%	R14	Carbon Resistor (100-1/2-5) Ins.
C16	Molded Mica Cond. (100MMFD-400V.) 10%	RIS	Carbon Resistor (1000-1/2-10) Ins.
C17	Molded Mice Cond. (2000MMFD-500V.) 10%	RIO	Carbon Resistor (150,000-1/2-10) Ins.
C18	Molded Mica Cond. (2000MMFD-500V.) 10%	R17	Carbon Resistor (220,000-1/2-10) Ins.
C19	Molded Mica Cond. (2000MMFD-500V.) 10%	R18	Carbon Resistor (100-1/2-5) Ins.
C20	Molded Mica Cond. (2000MMED 500V.) 10%	R19	Carbon Resistor (1000-1/2-10) Ins.
C21	Molded Mica Cond. (2000MMED 500V.) 10%	R20	Carbon Resistor (150,000-1/2-10) Ins.
C22	Molded Mica Cond. (2000MMFD 300V.) 10%	R21	Carbon Resistor (220,000-1/2-10) Ins.
C23	Molded Mica Cond. (2000MMFD-500V.) 10%	D92	Carbon Resistor (100-1/2-5) Ins.
C24	Molded Mica Cond. (50MMFD-400V) 10%	R24	Carbon Resistor (2 200 1 10) Ins.
C25	Tub. Cond. & Bracket (.05MFD-600V.)	R25	Carbon Resistor $(1/2, 1/2, 5)$ Inc.
C26	Molded Mica Cond. (5000MMFD-300V.) 10%	R26	W W Resistor (20.000-5-5) Inc
C27	Molded Mica Cond. (2000MMFD-500V.) 10%	R27	Carbon Resistor (22,000-1/2-10) Ins
C28	Variable Cond. (35MMFD)	R28	Carbon Resistor (220 $000-1/2-10$) Ins.
C29	Variable Cond. (140MMFD)	R29	Carbon Resistor (220,000- $1/2-10$) Ins
C30	Molded Mica Cond. (.01MFD-2500V.)	R30	Carbon Resistor (22,000-1/2-10) Ins.
C31	Molded Mica Cond. (2000MMFD-500V.) 10%	R31	Carbon Resistor $(470-1/2-10)$ Ins
C32	Molded Mica Cond. (2000MMFD-500V.) 10%	R32	W. W. Resistor (5000-10-5) Ins.
C33	Elect. Cond. (40MFD-25V.)	R33	Carbon Resistor (18-1/2-5) Ins.
C34	Molded Mica Cond. (2000MMFD-500V.) 10%	R34	Carbon Resistor (22,000-1-10) Ins
C35	Filter Condenser (6MFD-1000V.)	R35	Carbon Resistor (27,000-1/2-10) Ins.
C36	Filter Condenser (2MFD-1000V.)	RFC1	R.F. Choke
C37	Trimmer Condenser	RFC2	R.F. Choke
C38	Trimmer Condenser	RY1	"A" Power Relay (D.P.S.T.)
C39	Trimmer Condenser	RY2	"B" Power Relay (D.P.D.T.)
C40	Trimmer Condenser	RY3	Antenna Relay (D.P.D.T.)
C41 '	Trimmer Condenser	SW1	Meter Switch (S.P 8T)
F1	Fuse (50AMP) For 6V. Dynamotor Operation	T1	Osc. Plate Coil & Shield Assembly
	Fuse (5 AMP) For 117V, 60 Cycle Operation	T2	R.F. Coil & Shield Assembly
	Fuse (AMP) For 6V. Vibrator Pack Operation	Т3	1st Quad, Coil & Shield Assembly
J1 (Open Circuit Jack	T4	2nd Quad, Coil & Shield Assembly
L1 '	Tank Coil (30-35 Mc.)	T5	Diode Coil & Shield Assembly
,	Tank Coil (35-40 Mc.)	T6	Microphone Transformer
		XTAL	Crystal

* Used only in units which are tuned to L.F. end of band.

plifier section, another 7C7, and passed to the second mixer tube, a 7A8 tube—an octode converter. In the second mixer tube, the 4.3-megacycle signal is combined with the 6-megacycle signal from the oscillator section of this tube. The frequency difference signal, 1.7 megacycles, is amplified by the 1.7-megacycle second i-f section and fed to the first limiter stage.

Two stages of limiters are used, two 7C7 tubes, to reduce the amplitude variations of the input signal to as low a degree as possible. The two tubes are operating as saturated amplifiers with the limiting action taking place across the $100-\mu\mu$ f 100,000-ohm capacitor-resistor combinations in the grid circuits. The grid voltages that result from this limiting action may be measured at terminals 2 and 3 of the meter switch which is used during alignment.

The discriminator is of the normal type and operates at a center frequency of 1.7 megacycles; it results in a positive voltage when the applied frequency is higher than 1.7 megacycles, and a negative voltage when it is lower than 1.7 megacycles. This output voltage may be measured between terminal 5 of the meter switch and ground (chassis).

The squelch circuit is based on a characteristic of f-m reception to reduce the noise in the output when a signal, however weak, is received. It operates as follows: Noise from the discriminator, in absence of signal, is fed to the noise-amplifier tube, a 7C7. This noise is amplified and fed to the noise-rectifier tube, a 7A6 double diode. The rectified noise voltage develops a d-c voltage across the 1-megohm resistor R-1. This d-c voltage tends to bias the grid of the squelch-control section of the 7F7, first audio and squelch tube, in a positive direction. However, at the same time, voltage is fed to this same grid from the second limiter through another 1-megohm resistor R-2. This latter voltage tends to bias the grid negative. The difference voltage thus developed, approximately 1/2 v at the squelch-control grid, causes the plate to draw current and a voltage appears across the 1-megohm plate-resistor R-3.

The voltage across R-3, in the absence of signal, biases the grid of the first audio section of the 7F7 sufficiently to cut off (or squelch) the audio system. When signals are present and noise is reduced, the positive voltage from the noise rectifier is reduced and the negative voltage from the limiter increases. The differential voltage, approximately 1.5 v negative or greater, is then sufficient to bias the grid of the squelch-control section of the 7F7 to cutoff, and no plate current flows through R-3. Under this condition, the first audio grid is at normal bias and the audio system is in operation—the squelch is off.

Figure 16-4 shows the circuit diagram of a mobile transmitter for 30-w output in the 30- to 40-megacycle band. A crystal-controlled oscillator, a 7C7, is used to supply a regulated frequency to

337

a phase-to-frequency modulation circuit, two 7A8 modulators. Without temperature regulation of the crystal, the transmitted frequency holds within 0.02 per cent of the assigned frequency. For 100 per cent modulation, the transmitter has a frequency deviation of ± 15 kc.

As we see from examining the circuit diagram, it is a straightforward f-m transmitter adapted for mobile use. Either a dynamotor or vibrator power supply may be used to supply the necessary B+ voltages. If so desired, a-c operation may also be utilized. A schematic diagram of the vibrator power supply which may be



Fig. 16-5. A circuit diagram of a vibrator power supply used in conjunction with the mobile transmitter shown in Fig. 16-4.

used with this transmitter is shown in Fig. 16-5. Because of the large drain, two vibrator units operating with the inputs in parallel are employed. When using the vibrator power supply, jumper lead JU-1 (shown in Fig. 16-4) should be removed, breaking the parallel connection between the two filter capacitors.

For selective calling, the system illustrated in Fig. 16-6 may be employed.² For simplicity, a block diagram is employed where the complete receiver up to the first audio stage is represented as a single block. A tap from the grid of the first limiter tube is used to obtain the primary operating voltage. A switch is provided so that the receiver can operate either on the ordinary carrier squelch circuit or on the selective-calling system. For selective calling, the switch is left open. A tuned reed is used in the vibrating reed relay; this is a frequency-sensitive relay which responds to only one frequency. This actuating audio frequency is obtained from the output of the first audio amplifier tube. The contacts of this frequency-sensitive relay are in series with the contacts of another relay which is actuated by the rise in voltage on the grid of the first limiter tube.

When a signal is received, the voltage on the grid of the first limiter tube rises and closes one relay, the limiter relay. If an audio tone of the proper frequency is transmitted, it, in turn, closes the vibrating reed relay. When the two relays are closed, they actuate a third relay in the filament circuit of the final output tube. The output tube employs a quick-heating filament and, in a few seconds, the receiver is ready to receive the transmission. The relay is then held in place as long as the carrier is on, so that the tone can be removed and the signal transmitted. If the proper tone is not received when the carrier is on, one of the primary relays, the vibrating-reed relay, remains open and the receiver remains muted.

When operation without selective calling, reception of all trans-



Fig. 16-6. The block diagram of a selective-calling and squelch-circuit system used in a mobile receiver.

mitted signals, is desired, the switch shown in Fig. 16-6 can be closed and the selective-calling part of the system removed. With the switch closed, all that is necessary to operate the filament relay is a rise in voltage of the first limiter grid. The filament relay will then close and actuate the receiver. Operation with the switch closed is equivalent to ordinary squelching.

Thus, with selective calling, the receiver will not receive any signal except that one which is employing the proper tone signal for the particular receiver. Hence, a single transmitter may answer many purposes; one tone signal may turn on the receivers of the police patrol; another, the receivers of the fire patrol; still another, the receivers of the ambulance group; and so on.

16-3. Mobile F-M Installations. Before proceeding with any installation, it is desirable to have a definite plan wherein the exact

339

location of the various components is selected. The first step in such a plan is to obtain or make a cable and equipment layout of the apparatus to be installed. Such a layout for an f-m mobile re-



Fig. 16-7. A cable diagram for the installation of an f-m mobile receiver.

ceiver is shown in Fig. 16-7. The main components—the receiver, antenna (an ordinary vertical whip antenna is usually employed), control head, and speaker—are roughly sketched in first and then the proper connecting cables, with the connectors specified, are drawn between the various units. The sketch need not be as elaborate as the one illustrated, since a rough drawing showing all the necessary details will serve the purpose of planning.

The next step is to examine the vehicle in which the equipment is to be installed. A car layout should then be made showing the location of the equipment as well as the cable routes. Such a car layout is shown in Fig. 16-8 for the equipment shown in the cable diagram of Fig. 16-7. The receiver, in this case, is located in the rear trunk, with the antenna mounted close to it on the side of the trunk. The control head and speaker are mounted near the driver; the cables run either along the side or across the center of the vehicle.

The preliminary procedure for an f-m receiver-transmitter combination is the same except that the diagrams are more complex. A cable layout for a receiver-transmitter is shown in Fig. 16-9. In this installation, the power for the receiver is obtained from the transmitter chassis. The antenna is also fed into the trans-



Fig. 16-8. A car layout for the installation of an f-m mobile receiver whose cable diagram is shown in Fig. 16-7.

mitter chassis and then from the transmitter chassis to the receiver or chassis. A relay connects the antenna to the receiver or trans-



Fig. 16-9. The cable diagram for the installation of an f-m mobile transmitterreceiver combination.

mitter circuit, depending upon which is in operation at the moment. Either a military-type microphone or a handset may be employed with the unit while the installation of a speaker depends upon the choice that is made. A microphone and speaker combination leaves the operator free to go about his work while receiving a call and is usually preferred where a dispatching service does not require the mobile-unit operator to answer each call immediately. The handset, on the other hand, gives better reception in noisy locations as well as allowing privacy of communication. Push-to-talk buttons are used in both cases. The car installation, as shown in Fig. 16-10, is quite similar to the receiver installation shown in Fig. 16-8, except that more space is required for the equipment and a microphone mounting has to be provided.

A few other arrangements of the units in various types of vehicles is shown in Fig. 16-11. Figure 16-11a shows an installation in a van type truck. The equipment is installed in a box under the body, and the antenna rises above the body of the truck. Figure 16-11b shows an installation which is quite similar to those previously discussed except that the antenna is installed on top of the vehicle. This allows maximum sensitivity of reception and max-



Fig. 16-10. A car layout for the installation of the mobile f-m transmitter-receiver combination whose cable diagram is shown in Fig. 16-9.

imum power gain for transmission. The last installation shown (Fig. 16-11c) is for a panel-type delivery truck. The equipment is installed inside the body while the antenna is fastened to the top of the body.

When proceeding with the installation, care should be taken when drilling mounting holes to prevent puncture of the gas tank or gas line. Hence, the location of the tank and line should be carefully noted before drilling the mounting holes. In cases where uneven floor space prevents direct mounting of the equipment to the floor,



Fig. 16-11. Three car layouts for the installation of mobile f-m equipment in various types of vehicles.

it may be necessary to use blocks under the base plate of the equipment. In cases such as this, where the equipment chassis is not directly grounded by the mounting screws, a heavy flexible ground strap should be soldered between the equipment and car frame to supply the ground path.

In mounting the control head either on or below the instrument panel, the grounding should again be carefully considered. To reduce interference, it may be necessary to scrape the finish off the under panel and from the mounting lips of the control head to get

WRH

better electrical connection.

In installing the cable assemblies, care should be taken to keep them out of harm's way. They should be dressed down to the floor and routed under the floor mat or other protective covering. The cables should also be protected from damage by rubber grommets or friction tape at points where it may be necessary to run them through holes in the car frame or over steel seat guides. Care should be taken to avoid sharp bends in the r-f cables to the antenna or between chassis, since most r-f cables are susceptible to damage and shorting when sharply bent. A ground connection is very essential at each end of the coaxial line to the antenna. The inner conductor should be securely fastened to the antenna proper, and the connection should be periodically checked.

If there is any chance that water may get into the trunk or other part of the car where the equipment is installed, the equipment should be protected by mounting it a few inches above the floor on a solid platform. Again, the equipment should be carefully grounded by means of a heavy shield braid to the car frame.

For installations in a wooden station wagon, it may be found advantageous to install the equipment on a mesh screen or sheet metal surface. The screen should be as large as possible and grounded to the car frame. Care should again be taken that all the equipment is properly grounded, or bonded, to the car frame.

Although one of the inherent characteristics of f-m reception is the ability to reduce noise in the presence of the received carrier, it is advisable to keep the noise level as low as possible. These interfering noises may be of a local nature such as produced by sparking motors, fluorescent lights, diathermy apparatus, electric fences, and other similar sources. When located, these disturbances can be cut down and perhaps eliminated by shielding and filtering. Passing automobiles may also radiate interference, but this is of a temporary nature and is usually not serious.

In the case of a mobile installation, work will be necessary, usually, on the car electrical system to limit the production of electrical interference. The level of interference created by the car in which the mobile equipment is installed will be the limiting level for satisfactory reception; the noise produced by the car should be reduced to the lowest possible level.

Generator noise can be detected by coasting the car in gear, with the ignition shut off. The installation of a generator filter will usually stop the generator noise. Installation of spark-plug suppressors will usually reduce the ignition interference to an acceptable degree. Bypass capacitors to ground, which are carefully shielded, can be tried at various points on the battery system such as at the distributor, ignition switch, gauges, and so on. (Do not connect the capacitor across the generator field.) The installation of hood-grounding wipers, bonding strips from the motor block to the frame, or bonding the exhaust pipe or muffler to the frame, will sometimes help. There has been no set procedure formulated for the elimination of noise in a car and what may be a successful method in one car may not work on the next. The only solution is to keep trying various things until the offending noise is eliminated or reduced to negligibility.

16-4. General Telephone Service Using Frequency Modulation. Frequency modulation is also being used to supply general telephone service to mobile vehicles.³ Three classes of service are offered: a general two-way telephone service between any vehicle and any wire telephone or other mobile unit; a two-way dispatching service between an operator's office and his own mobile units; and a one-way signaling service to mobile units to notify the driver that he should comply with prearranged instructions.

The equipment is of two general types, one operating in the 152to 162-megacycle band and its counterpart operating in the 30- to 40-megacycle band. The following description refers to the 152to 162-megacycle equipment, but is applicable, for the most part, to the lower-frequency equipment also.

The equipment consists of a radio receiver and transmitter, antenna, selective signaling device, and telephone instrument. The controls and the instrument, shown in Fig. 16-12, may be mounted under the instrument panel of the vehicle in a single unit. The controls are few in number: one on-off power switch and a push-totalk switch in the handle of the handset. No tuning or adjustments are made by the operator.

The transmitter has a power output of 20 w or thereabouts, depending on the area being served. The high voltages are supplied by a dynamotor which receives its power from the battery in the vehicle. Quartz crystal control of the transmitter is provided, with a thermostatic-controlled constant-temperature oven enclosing the quartz plate, thus assuring constancy of transmitted carrier frequency. A phase-to-frequency modulation circuit is employed. The minimum deviation plus frequency tolerance of the carrier under speech modulation is normally a swing of ± 15 kc. For test purposes by the installation or test man, all test and measuring points are brought out to a convenient jack on the front of the panel; a meter test set is used with a plug to fit the test jack, and the value of the currents in the various sections of the transmitter circuit is measured by simply rotating the test-set switch to various positions.

The receiver uses a double-superheterodyne circuit, fixed-tuned with crystal control for both conversion oscillators, thus giving effective image suppression with no loss in selectivity. A non-synchronous vibrator is used so that no concern need be had as to whether the positive or negative side of the vehicle battery is grounded. A switch is provided for either 6- or 12-v battery operation. With less than a $1-\mu v$ input to the receiver, the signal-to-



Fig. 16-12. The control unit for the mobile station in a mobile-telephone installation. It is mounted under the instrument panel of the vehicle.

noise ratio is about 25 db. Negative feedback is provided in the audio circuits for fidelity of reception. A view of the receiver with the cover removed is shown in Fig. 16-13.

The selector shown in Fig. 16-14, with its associated filters and relays, provides full selective calling. Each mobile station is assigned a five-digit number, prefixed by a channel designation. The number 1 is used as the selector "return to normal" pulse. An example of a mobile station number would be WJ3-4727, WJ being the channel designation and 34727 the mobile radiotelephone station number, this designation being in accordance with standard telephone practice for telephone numbers.

The mobile-telephone traffic operator dials the digits for the called station. In doing so, she causes the control terminal equip-

ment to modulate the land transmitter with a series of pulses of 600 and 1,500 cps. These pulses, received and amplified at the



Fig. 16-13. A top view of the receiver used in mobile-telephone service with its top cover removed.

mobile stations, cause all selector sets assigned to the same channel to be actuated. Stop pins in the code wheel of the selector in the receiver will cause only that selector to be positioned to ring a bell and light a call lamp in the vehicle. All other selectors will, of course, fail to reach the final or signaling position. The bell stops ringing after a short interval, but the lamp remains lighted until the subscriber answers the call by removing the handset from its cradle.

16-5. Relay Equipment. In some cases it is impossible to install the central-station receiver in a situation where its antenna can receive the broadcasts from the mobile units over the whole area that has to be covered. In those cases, where coverage is necessary, relay transmitters have to be used. Relay stations are usually needed only to relay transmissions from the mobile units because of the limited amount of power and low antenna height associated with mobile equipment. Transmissions from the central station can usually reach the mobile units when adequate transmitter power and antenna height is employed.
Figure 16-15 shows a block diagram of a relay transmitter installation using the 116- to 119-megacycle band for the relay fre-



Fig. 16-14. The selector being used so that each vehicle is signaled only for its own incoming calls.



Fig. 16-15. A block diagram of a relay-system installation for incoming calls from mobile units.

quency. In this case, the relay transmitter is shown situated on top of a mountain where, because of its favorable location, it can readily pick up and rebroadcast messages not receivable by the unfavorably situated central station receiver. Since the rebroadcast frequency is in the 116- to 119-megacycle band, the relay station antenna should be within line of sight of the central station.



Fig. 16-16. The circuit diagram of the ultrahigh-frequency transmitter used to

The normal equipment needed for a relay-station installation includes the following units: a receiver to pick up the mobile unit's message, a relay transmitter to retransmit the message, a receiving antenna, and a transmitting antenna. A receiver at the relaytransmitter frequency is needed at the central station to receive the relayed message. Directional antennas are usually used for both the relay transmitter at the relay station and the relay receiver at the central station.

The relay station can employ a standard receiver to pick up the



relay calls at 116 to 119 megacycles. See schematic diagram parts list on next page.

messages from the mobile units, provision being made, of course, to have it operate on the type of power available at the relay station.

The receiver at the relay station is operative at all times, but the ultrahigh-frequency transmitter is off the air unless a signal is being received from the mobile units. A control relay mounted in

FIG, 16-16

SCHEMATIC DIAGRAM PARTS LIST

Diag	a	Diag.	
No.	Description	No.	Description
R1	Carbon Resistor (470,000-1/2WATT-10%)	C16	Molded Mica Cond. (2000MMF-20%-500W.V.)
R2	Carbon Resistor (1000-1/2WATT-5%)	C17	Molded Mica Cond. (2000MMF-20%-400W.V.)
R3	Carbon Resistor (10,000-1/2WATT-10%)	C18	Molded Mica Cond. (2000MMF-20%-500W.V.)
R4	Carbon Resistor (47,000-1/2WATT-10%)	C19	Molded Mica Cond. (2000MMF-20%-500W.V.)
R5	Carbon Resistor (47,000-1/2WATT-10%)	C20	Molded Mica Cond. (100MMF-20%-500W.V.)
R6	Carbon Resistor (100-1/2WATT-5%)	C21	Molded Mica Cond. (2000MMF-20%-500W.V.)
R7	Carbon Resistor (4,700-1/2WATT-10%)	C22	Molded Mica Cond. (2000MMF-20%-400W.V.)
R8	W. W. Resistor (500 OHMS, 10%, 10 WATT)	C23	Molded Mica Cond. (1000MMF-20%-500W.V.)
R9	Carbon Resistor (33,000-1WATT-10%)	C24	Molded Mica Cond. (50MMF-20%-500W.V.)
R10	Carbon Resistor (39.000-1WATT-10%)	C25	Oil Filled Condenser(.05MFD-600W.V.)
R11	Carbon Resistor (68.000-1WATT-10%)	C26	Molded Mica Cond. (2000MMF-20%-400W.V.)
R12	Carbon Resistor (68,000-1WATT-10%)	C27	Paper Canacitor: .0005 MDF. 400V.
R13	Carbon Resistor (220,000-1/2WATT-10%)	C28	Neutralizing Condenser
R14	Carbon Resistor (100-1/2WATT-5%)	C29	Neutralizing Condenser
R15	Carbon Resistor (12.000-1/2WATT-10%)	C30	Variable Condenser (10MMF)
R16	Carbon Resistor (18,000-1/2WATT-10%)	C31	Molded Mica Cond. (2000MMF-10%-400W.V.)
R17	Carbon Resistor (220,000-1/2WATT-10%)	C32	Molded Mica Cond. (2000MMF-10%-400W.V.)
R18	Carbon Resistor (100-1/2WATT-5%)	A	
R19	Carbon Resistor (390-1/2WATT-10%)	C34	Molded Mica Cond. (2000MMF-20%-400W.V.)
R20	Carbon Resistor (39,000-1/2WATT-10%)	C35	Filter Condenser (6MFD-1000W.V.)
R21	Carbon Resistor (1000-1WATT-5%)	C36	Filter Condenser (2MFD-1000V.)
A		C37	Trimmer Condenser T1H
R23	Carbon Resistor (1000-1/2WATT-10%)	C38	Trimmer Condenser T2
R24	Carbon Resistor (47.000-1/2WATT-10%)	C39	Trimmer Condenser T3
R25	Carbon Resistor (10,000-1/2WATT-10%)	C40	Variable Condenser
R26	W. W. Resistor (20,000-10WATT-10%)	C41	Variable Condenser
R27	Carbon Resistor (22,000-1/2WATT-10%)	C42	Variable Condenser
R28	Carbon Resistor (220,000-1/2WATT-10%)	C43	Molded Mica Cond. (1000MMF-10%-300W.V.)
R29	Carbon Resistor (220,000-1/2WATT-10%)	C44	Paper Capacitor: .002 MFD. 500V.
R30	Carbon Resistor (220,000-1/2WATT-10%)	C45	Paper Capacitor: .002 MFD. 500V.
R31	Carbon Resistor (470-1/2WATT-10%)	C46	Paper Capacitor: .002 MFD, 500V.
R32	W. W. Resistor (2,500-25WATT-ADJ.)	C47	Paper Capacitor: .01 MDG, 2500V.
R33	Carbon Resistor (18-1/2WATT-5%)	TIH	Osc. Plate Coil & Shield Assembly
R35	Carbon Resistor (1/2-1/2WATT-5%)	T2H	R.F. Plate Coil & Shield Assembly
(A)	· · · · · · · · · · · · · · · · · · ·	тзн	1st Quad. Plate Coil & Shield Assembly
R36	Carbon Resistor (50-1/2WATT-5%)	Т6	Microphone Transformer
R38	W. W. Resistor (3.000-25WATT-10%)	RY2	"B" Power Relay (D.P.D.T.)
C1	Molded Mica Cond. (5000MMF-20%-500W.V.)	RY4	Control Relay (D.P.D.T.)
C2	Molded Mica Cond. (50MMF-20%-500W.V.)	J1	Open Circuit Jack
C3	Molded Mica Cond. (50MMF-20%-500W.V.)	RFC1	F.F. Choke
C4	Molded Mica Cond. (2000MMF-20%-500W.V.)	RFC2	R.F. Choke
C5	Ceramic Condenser(8MMF) 5% NPOK	RFC3	R.F. Choke
C6	Molded Mica Cond. (2000MMF-20%-400W.V.)	L1	Quadrupler Tank Coil
C7	Molded Mica Cond. (2000MMF-20%-400W.V.)	L2	Doubler Driver Tank Coil
C8	Elect, Cond. (20-40MFD-25W,V.)	L3	Link Coil Assembly
C9	Oil Filled Condenser(.05MFD-600W.V.)	L4	Final Grid Coil
C10	Oil Filled Condenser (.05MFD-600W,V.)	L5	Final Tank Coil, Strap & Grid Cans Assembly
C11	Molded Mica Cond, (100MMF-20%-500W.V.)	L6	Antenna Link Coil
C12	Molded Mica Cond. (2000MMF-20%-500W.F.)	SW1	Meter Switch (S.P 8T.)
C13	Molded Mica Cond. (2000MMF-20%-400W.V.)	F1	Fuse (5AMP) For 117V, 60 Cycle Operation
C14	Molded Mica Cond. (2000MMF-20%-500W.V.)	STAL	Crystal (Frequency Specific to Order)
C15	Molded Mica Cond. (100MMF-20%-500W.V.)		

the transmitter is closed by current furnished by the control receiver whenever a signal is being received; this relay then turns on the B power of the transmitter. The transmitter itself, a schematic diagram of which is shown in Fig. 16-16, is similar to the 30- to 40-megacycle transmitters except that it contains the control relay and is designed to operate at the proper relay frequency. A 6L6 tube is used to drive an 815 dual-beam power tube in push-pull. The output is about 10 w at the relay frequency. It is mounted in a cabinet separate from the receiver at the relay station to prevent interaction between the two.

The ultrahigh-frequency receiver at the central station consists of a standard 30- to 40-megacycle receiver with an ultrahigh-frequency converter to convert the relayed signal of 116 to 119 megacycles to a signal in the 30- to 40-megacycle band for reception by



Fig. 16-17. The schematic diagram of the converter employed to convert the 116- to 119-megacycle received signal to a signal in the 30- to 40-megacycle range.

the 30- to 40-megacycle receiver. The converter (Fig. 16-17) consists of a 6AC7 type mixer which heterodynes the 116- to 119-megacycle signal with an 80-megacycle signal produced by a crystaloscillator and multiplier circuit. The oscillator consists of a 6SC7 tube operating with one section as a triode oscillator at 8 megacycles, and the other section as a five times multiplier yielding an output of 40 megacycles. This 40-megacycle signal is doubled in a 6AC7 tube and the resultant 80-megacycle signal fed into the mixer.

In those cases where direct wire connection between the relay station and the central station is available, the relay transmitter and station receiver may be dispensed with and a direct wire substituted. All that is needed at the relay station is a receiver to pick up the message being transmitted by the mobile unit. The received signal may then be fed into a direct wire connection and made available to the central station operator at the other end of the wire.

REFERENCES

Galvin Manufacturing Co.
Federal Telephone and Radio Corp.

3. Robert Bright, Jr., and Stanton Vanderbilt, "Telephone on Wheels," Western Electric Oscillator, July, 1946.

QUESTIONS

1. Describe the different basic types of mobile-communication systems.

2. When should each of the different basic types of mobile-communication systems be employed?

3. What is meant by "squelching" in a mobile receiver, and why is it employed?

4. Describe what is meant by selective calling in mobile communication.

5. What are the advantages of employing selective calling?

6. Describe the circuits and operation of a squelch circuit based on noise amplification.

7. Describe, by means of a block diagram, the operation of the tuned-reed method of selective calling,

8. Why should a definite plan be formulated before proceeding with a mobile installation?

9. Explain what is meant by a cable or equipment layout.

10. Observe the plan of your own or a friend's car and make a car layout for the installation of a mobile receiver-transmitter combination.

11. What precautions should be taken concerning the car equipage when installing mobile equipment?

12. Describe the special precautions that should be taken when installing the cables—particularly the antenna cable. 13. What precautions can be taken to reduce the noise pickup from the auto-

mobile's electric system?

14. When is relay equipment necessary in mobile-communication-system installations?

15. Describe the apparatus necessary for a mobile-communication-system relay station.

16. What advantage is there to a direct wire connection to the relay station?

CHAPTER 17

Frequency-Modulation Test Equipment

17-1. Signal-Generator Requirements for Frequency Modulation. The basic requirements of any signal generator are stability and accuracy of calibration. In f-m signal generators, especially in the f-m broadcast band, these two requirements introduce many complications which have to be considered in the design, construction, and use of the generator.

In the f-m signal generator, again, as in the f-m transmitter, we have the problem of stability in a variable-frequency sourcevariable carrier frequency, in this case, as well as variable frequency because of the method of modulation. The stability is interlinked with the accuracy of calibration since the carrier frequency and the frequency deviation should remain constant within the practical requirements once the generator output is set; they should also return to the same value if the dials are turned and reset to the same calibration. The frequencies used in f-m signal generators, especially for the f-m broadcast band, are usually much higher than those employed in a-m signal generators. Again, the problems encountered in building an f-m signal generator for the f-m broadcast band are similar to those encountered in the construction of any 100-megacycle band equipment. Careful consideration of grounds, regeneration, and circuit components are necessary to avoid troubles in the finished equipment.

One of the most difficult problems in the design of an f-m signal generator is the problem of modulation calibration. The usual types of modulation circuits produce frequency deviations which vary with the oscillator frequency. Since it is necessary to calibrate permanently the deviation obtained from the signal generator, some means must be employed to maintain a constant modulator sensitivity over the whole frequency range of the instrument.

17-2. Frequency-Modulation Signal Generators.¹ There are two basic systems employed in the design of f-m signal generators:

the heterodyne system and the constant-deviation variable-oscillator system. The basic operating principle of the heterodyne sys-



Fig. 17-1. A block diagram showing the essential stages of an f-m signal generator employing the heterodyne system.

tem is indicated in the block diagram of Fig. $17-1.^2$ In this system. a fixed oscillator is frequency-modulated by an audio frequency. In the diagram, the fixed oscillator is controlled by a reactance-tube modulator unit to which the modulating voltage is being applied by means of a deviation-control potentiometer. Actually, the frequency modulation may be accomplished by means of any of the f-m systems previously discussed, or by the regular variation of the capacitance in the tank circuit of the fixed oscillator by a motor driving a variable capacitor. The fixed oscillator maintains a constant center frequency of 40 megacycles so that the output of the fixed oscillator is an f-m wave at a constant carrier frequency of 40 megacycles. Since the oscillator center frequency does not vary in this system, it is possible to calibrate the deviation directly in terms of the audio-modulation voltage. The output of the f-m fixedcenter-frequency oscillator is combined with a variable oscillator in a mixer tube. The sum frequency is then amplified, passed through an attenuator, and used as the output frequency. When the desired deviation has been chosen, it is necessarily held constant. regardless of the output frequency; it is the fixed oscillator that is modulated and its frequency deviations are transferred unaltered to the sum frequency voltage (or difference frequency voltage, if that is employed) generated in the mixer stage. It is also possible to modulate the variable oscillator and produce a constant deviation by ganging the deviation-control potentiometer to the variable tuning element of the variable oscillator. The necessary output frequency is then produced by heterodyning the signal with an unmodulated fixed-frequency oscillator. However, modulation of the fixed oscillator offers less complications and is usually preferred.

If a constant-deviation modulation system is utilized, the signal does not require the step of heterodyning. Either an oscillator operating directly at the output frequency or a master-oscillator multiplier arrangement may be used. To make the system work, however, the inherent variation of the deviation with frequency has to be corrected or removed. Various means may be applied to the reactance-tube circuit for correcting the inherent variation of the deviation with frequency. One method is to gang a modulation-output potentiometer with the



Fig. 17-2. A reactance-tube circuit where constant deviation with centerfrequency variation is obtained by ganging a modulation-input potentiometer with the tuning capacitor.

tuning dial, as shown in Fig. 17-2. Potentiometer <u>P</u> is designed to have a resistance taper such that, when ganged to the shaft of tuning capacitor <u>C</u>, the modulation sensitivity is maintained constant as the frequency is varied. For the circuit shown, the potentiometer reduces the modulation input as the frequency is increased.





Figure 17-3 shows how the modulation sensitivity may be maintained constant by ganging an element of the reactance-tube phase shifter to the shaft of the tuning dial. For the inductive-reactance tube and variable-capacitance tuned circuit shown, phase-shifter capacitor \underline{C}_1 is ganged to tuning capacitor \underline{C} in such a manner that \underline{C}_1 is increased as \underline{C} is decreased. This varies the phase-shifter constants as the frequency is increased, so that the degree of modulation produced by the reactance tube falls off with frequency sufficiently to produce constant deviation. For linear operation, a particular shaping of the plates of \underline{C}_1 with relation to those of capacitor \underline{C} is required. In the circuit of Fig. 17-3, the constants of the phase-shifter circuit are mechanically varied to produce constant-deviation operation. It is also possible to arrange constant-deviation sensitivity over a given frequency range by means of a particular phase-shifting network which inherently produces the required phase shift and has the proper characteristics.



Fig. 17-4. A constant-deviation reactance-tube circuit obtained by having the phase shifter <u>R</u>, <u>C</u>₁, and <u>L</u>₁ produce an attenuation inversely proportional to the square of the frequency over the tuning range.



Fig. 17-5. A constant-deviation network where the required inverse-square attenuation is obtained by the bridged-T phase-shifting network $\underline{R}_1, \underline{R}_2, \underline{C}_1, \underline{C}_2$, and \underline{C}_3 .

Figures 17-4 and 17-5 show two capacitance-tuned circuits and inductive-reactance-tube arrangements which produce constant frequency deviation over a range of frequencies. In the standard reactance-tube circuit, the grid voltage is obtained across a small capacitor between grid and ground. The voltage across the capacitor is inversely proportional to frequency; hence, the grid voltage in a standard inductive-reactance-tube circuit is inversely proportional to frequency. Therefore, we see that the normal output produced by a two-element phase shifter is an output amplitude which is inversely proportional to the frequency of operation. However, to obtain constant-frequency deviation over a range of frequencies, the phase-shifter characteristic (the output amplitude) has to vary inversely with the square of the frequency to which the tuned circuit is resonant. In Fig. 17-4, this inverse-square-law characteristic is obtained by using a series-tuned circuit consisting of \underline{L}_1 and \underline{C}_1 in the grid circuit. This circuit L_1C_1 is series-tuned to a frequency above the range in which the circuit is to be used. Hence, the tuned circuit $\underline{L}_1\underline{C}_1$ operates on the capacitive portion of its resonant characteristic and therefore appears as a capacitance to produce an inductive-reactance circuit. In this manner the circuit produces an inductive reactance inversely proportional to the square of the operating frequency over a frequency range of about 1.5 to 1.

Similarly, the circuit shown in Fig. 17-5 uses a bridged-T network $\underline{R_1 R_2}$ and $\underline{C_1 C_2 C_3}$ to obtain the proper attenuation characteristic. The network is tuned to locate its rejection frequency above the range for which the circuit is to be used. Using this network, it is possible to obtain constant-deviation operation over a frequency range of about 2 to 1.

The frequency coverage can be doubled without switching oscillator and modulator circuits by following the oscillator with a class C stage, whose tank coil is switched to operate either as an amplifier or as a doubler. This stage permits the oscillator and modulator to operate at half the frequency of the high-output range. At lower frequencies, the effects of stray and residual reactances are not as troublesome as they would be at the higher frequencies. It also permits amplitude modulation of the output stage if desired.

Sometimes it is found that a single class C stage is not entirely satisfactory, because some undesired amplitude modulation may be produced during frequency modulation of the oscillator; and, if the oscillator voltage is not sufficient, the class C stage will not remove this amplitude modulation by its normal process of limiting. A second objection is that amplitude modulation of the class C stage may produce a considerable amount of spurious frequency modulation of the oscillator. Both of these objections may be removed by the addition of a second class C stage.

A simplified circuit diagram of an f-m signal generator incorporating two class C stages is shown in Fig. 17-6. The reactancemodulated oscillator operates from 27 to 54 megacycles. This stage is then followed by a doubler stage and an output stage. The output stage operates either as an amplifier or as a second doubler (quadrupler when referred to the oscillator frequency), thus providing frequency coverage from 54 to 216 megacycles. The oscillator and modulator operate at a relatively low frequency and the only r-f switching required is to switch a ground contact on the output coil.

The reactance tube uses a bridged-T type network similar to that of Fig. 17-5, in which \underline{C}_2 and \underline{C}_3 are replaced by the input capacitance and grid-to-plate capacitance of the reactance tube. Capacitor \underline{C}_1 is made variable and ganged with the tuning capacitor to provide the precision of deviation calibration required in a signal generator.

The mechanical details of the r-f assembly are shown in Fig. 17-7. The slotted stator of capacitor C_1 is visible in the foreground



of the picture. By bending the sections of the stator, the deviation can be made as constant as desired over the entire frequency range.

Fig. 17-6. A simplified diagram of an f-m signal generator using a constantdeviation reactance-tube modulator and incorporating two class C amplifier stages.

Resistor \underline{R}_4 (as shown in Fig. 17-6) is bypassed for r-f signals only, causing degeneration at modulation frequencies, and is used as a modulation-sensitivity control. Resistor \underline{R}_4 controls the cathode bias by bleeding current through the cathode resistor, and is used to adjust the modulator tube to its most linear operating point. With this circuit, it is possible to obtain a total root-mean-square distortion of less than 1 per cent at ± 75 -kc deviation when using a modulating signal containing less than one half per cent distortion.

Amplitude modulation is produced by modulating the screen circuit of the output stage. Since the power required is small, compared with that required for frequency modulation, a simple resistance-tuned modulating oscillator serves for either amplitude or frequency modulation. Frequency-range switching is accomplished by providing the output tank coil with two contact points, one or the other of which is grounded by spring-contact fingers actuated by the frequency-range switch.

In the output tank circuit, a high \underline{Q} is desired to reduce the harmonically related spurious output signals, while a low \underline{Q} is desired to reduce the amplitude modulation introduced on f-m signals by the selectivity of the circuit. A compromise is necessary. One value that may be used is to adjust the damping to reduce spurious signals to about 35 db below the desired signal. The resulting amplitude modulation at \pm 75-kc deviation for this adjustment is about 2 per cent.



Fig. 17-7. Mechanical details of the r-f assembly of the f-m signal generator shown in Fig. 16-6. The arrow indicates the slotted stator of C_1 .

The design of the output stage is shown schematically in Fig. 17-8. A mutual-inductance (piston) attenuator with a coupling coil indicated in the diagram as \underline{L}_2 is coupled to a D-shaped output coil shown schematically as \underline{L}_1 . Two 53-ohm resistors are used, one in series with \underline{L}_2 , and the other across the output terminals. The equivalent circuit of the attenuator is shown in Fig. 17-8b. Neglecting the inductance \underline{L}_2 , which is usually quite small, the generator can be represented as a source \underline{e} in series with a resistance of 26.5 ohms (Fig. 17-8c). The piston attenuator, a cylinder wherein the separation between input and output coils is varied, is a simple and accurate attenuator for the v-h-f range, since the rate of attenuation is dependent only on the inner diameter of the attenuator tube after sufficient attenuation has been introduced to suppress undesired modes of propagation of the electromagnetic field within the tube. With this type of attenuator, a range from 0.1 μ v to 0.2 v can be obtained.

Front and rear views of the signal generator shown schematically in the circuit of Fig. 17-6 are shown in Fig. 17-9. The frequency deviation and percentage of amplitude modulation are controlled by two potentiometers below the modulation meter switch. On the tuning dial, the vernier scale changes the frequency by about 26 kc on the low-frequency range, and 52 kc on the high-frequency range. The choice of the modulating frequency is provided by the dial at the lower left in the front view of the generator, and the out-



Fig. 17-8. An output circuit of an f-m signal generator employing a piston attenuator. Circuit <u>a</u> is the output system in which a D-shaped coil \underline{L}_1 is used. The equivalent circuit of the attenuator is shown at <u>b</u> and the equivalent circuit of the signal generator at <u>c</u>.

put is controlled by the attenuator dial at the lower right.

17-3. Requirements for Frequency-Modulation Station Monitors.³ The FCC requires each f-m broadcast station to have in operation, at the transmitter, an approved-frequency monitor independent of the frequency control of the transmitter. The frequency monitor shall be approved by the commission and shall have a stability and accuracy of at least one half (\pm 1,000 cps) of the permitted center frequency deviation of the f-m broadcast station. Visual indication of the operating frequency must be provided.

In general, a frequency monitor for f-m broadcast stations requires a stable source of r-f energy whose frequency is accurately known, and a means of comparing the transmitter center frequency with this stable source. The visual indicator is calibrated to indicate the deviation of the transmitter center frequency from the correct assigned frequency.

Some of the general specifications that frequency monitors have to meet before they are approved by the commission are as follows:

1. The unit shall have an accuracy of at least $\pm 1,000$ cps under ordinary conditions (temperature, humidity, power-supply variations, and other conditions which may affect its accuracy) encoun-

tered in f-m broadcast stations throughout the United States, for any channel within the f-m broadcast band.

2. The range of the indicating device shall be at least from 2,000 cps below to 2,000 cps above the assigned center frequency.

3. The scale of the indicating device shall be so calibrated as to be accurately read within at least 100 cps.

4. Means shall be provided for the adjustment of the monitor indication to agree with an external standard.

5. The monitor shall be capable of continuous operation and its circuit shall be such as to permit continuous monitoring of the transmitter center frequency.

6. Operation of the monitor shall have no deleterious effect on the operation of the transmitter or the signal emitted therefrom.

Each f-m broadcast station is also required to have an approved-modulation monitor in operation at the transmitter. This monitor may or may not be a part of the f-m broadcast frequency monitor. Some of the specifications that the modulation monitor is required to meet are as follows:

1. It should have means of ensuring that the transmitter input to the modulation monitor is proper.

2. It should have a modulation peak-indicating device that can be set at any predetermined value from 50 to 120 per cent modulation, \pm 75-kc swing being 100 per cent modulation, and for either positive or negative swings—either above or below the carrier frequency.

3. The accuracy of the percentage-of-modulation reading should be within ± 5 per cent modulation at any percentage of modulation up to 100 per cent modulation.

4. The frequency-characteristic curve should not depart from a straight line more than $\pm 1/2$ db from 50 to 15,000 cps. Distortion shall be kept to a minimum.

5. The monitor should not absorb appreciable power from the transmitter.

6. Operation of the monitor should have no deleterious effect on the operation of the transmitter.

17-4. Direct-Indicating Frequency-Modulation Station Monitors. Practically all the methods of frequency regulation as discussed in Chap. 5 can be used as the basis for a frequency monitor. Consider the one based on the principle of frequency division. As we have discussed previously, when a frequency-modulated wave is passed through a series of frequency dividers, the carrier-component amplitude progressively increases until, with enough division, practically all the energy in the divided wave rests in the carrier subharmonic. The frequency of this component, which may now be easily measured, is equal to the f-m carrier component divided by the division factor. Another possible method is to use



Fig. 17-9. Front and rear views of an f-m signal generator incorporating the circuit shown in Fig. 17-6. The frequency deviation and the percentage of amplitude modulation are controlled by two potentiometers below the modulation-meter switch.



Choice of the modulating frequency is provided by the dial at the lower left and the output is controlled by the attenuator dial at lower right.

the integrated-pulse system and compare the center frequency to an independent standard. In fact, any of the frequency-control systems may be so employed.

Still another method is to measure the average of the instantaneous carrier frequency, using a circuit which in principle is a modified precision f-m receiver.⁴ A block diagram of this system



Fig. 17-10. A block diagram of an f-m station monitor including facilities for monitoring frequency, percentage modulation, and audio quality.

is shown in Fig. 17-10. The transmitted wave is noted as $f_{t\pm}75$ $kc_{\pm} \Delta f_t$ where f_t is the assigned center frequency while Δf_t is the amount that the carrier frequency of the transmitted wave differs from the assigned carrier frequency, the plus or minus sign indicating whether it is above or below the correct frequency. This f-m wave is fed into the monitor from the transmitter and is immediately limited and converted into an intermediate frequency of approximately 5 megacycles in the mixer tube. The beat-oscillator frequency is derived from a crystal oscillator and multiplier which has an output frequency of $f_t - f_{IF}$ where f_{IF} is the intermediate frequency. Subtracting this frequency from the transmitted f-m wave, we obtain, for amplification in the power amplifier, the f-m wave f_{IF} + 75 kc + Δf_t . This f-m i-f signal is then fed into a frequency discriminator. The discriminator circuit is designed to deliver a current which is proportional to the instantaneous deviation of the carrier from its assigned value, being zero at the assigned value. The zero-output-current point of the discriminator may be calibrated, when necessary, by a separate precision crystal oscillator that has a frequency of f_{IF} . The discriminator output current is averaged over the audio cycle and the resultant direct current is

proportional to the shift of the mean frequency from its assigned value ($\Delta \underline{f}_{\underline{t}}$ being the magnitude of the shift). A zero-center d-c instrument with a linear scale marked from -2,000 to +2,000 cps indicates the difference between the mean carrier frequency and its assigned value.

The a-c component of the output of the discriminator circuit has the form of the modulating wave because of the linear relation between current and frequency in the discriminator. Since the peak value of the signal is proportional to the peak frequency deviation. the percentage of modulation is indicated by a peak voltmeter. This is operated by the discriminator audio output after amplification, and it is so calibrated that 100 per cent modulation is indicated when the instantaneous peak frequency deviation is 75 kc. A gas tetrode 2051 tube is arranged with the peak rectifier output applied to its grid so that when the modulation peaks exceed some preset value, which is under the control of the operator, the tube is "fired." Since the plate-supply voltage for the gas tube is 60 cps alternating current in this case, the plate is negative every 1/60 sec with respect to the cathode. This allows the grid to regain control of the plate current every 1/60 sec. The "firing" of the tube causes a red flasher lamp to light, and also closes a relay circuit which may be connected to an external alarm device or to a counter to record the number of peaks of modulation which exceed the value for which the control is set.

Audio-quality monitoring is also provided in the monitor. Since it is standard practice to pre-emphasize the high frequencies of the audio modulating signal, a de-emphasis circuit is necessary for recovering the original audio signal. The output of the discriminator is passed through this circuit and thence into a low-distortion output tube designed to feed an external 600-ohm system.

One of the problems in a frequency monitor is to obtain sufficient current from the discriminator for mean-frequency indication. The conventional discriminator, as used in receiver applications, is designed primarily to have a high-voltage sensitivity; that is, to deliver an audio voltage as large as possible for a given swing in frequency. This voltage is then applied to the loudspeaker through a suitable amplifier. In a frequency monitor, however, it is also required that the direct output current, which is proportional to the variation in the mean frequency, be available at a useful level. The range of mean-frequency variation is 4,000 cps (±2,000 cps), whereas the range of the instantaneous frequency (the frequency deviation, ± 75 kc) is 150,000 cps, making the amplitude of the mean-frequency indicating signal only 2.7 per cent of the amplitude of the audio signal from the discriminator. Furthermore, the mean-frequency indication is a slowly varying d-c signal, and a stable d-c amplifier is not a simple matter. These considerations led to the

modified discriminator circuit shown in Fig. 17-11. This discriminator circuit is especially adapted to giving a high-current sensi-



Fig. 17-11. The schematic diagram of a current-sensitive discriminator showing the mean-frequency meter and level-indicating meter for use in a monitor.

tivity rather than a high-voltage sensitivity. Moreover, this modified discriminator is balanced to ground and allows an audio signal of either polarity to be obtained from it. Thus, indication of either positive or negative percentage shift is possible. A d-c instrument which is bypassed for audio frequencies can then be used to indicate the average output current directly from the discriminator, and d-c amplifiers are not required.

In the conventional discriminator \underline{E}_1 and \underline{E}_2 are equal and opposite voltages when the center frequency is applied to the discriminator. Thus, the sum of these voltages, in the normal discriminator, is zero. As the frequency deviates to either side of the center frequency, the output voltage, the sum of the two voltages obtained from each diode output resistor, is either positive or negative. A d-c instrument to measure mean frequency, in that case, would have to be shunted across both diode output resistors and should not impose too much load upon them. This circuit results in a low-current sensitivity.

In the current-sensitive discriminator circuit (Fig. 17-11), on the other hand, the diode voltages \underline{E}_1 and \underline{E}_2 are in the same direction and proportional to \underline{I}_1 and \underline{I}_2 , respectively, the currents flowing through the two load resistors \underline{R}_1 and \underline{R}_2 . The difference current $\underline{I}_{\underline{L}}$ between \underline{I}_1 and \underline{I}_2 flows through the center-frequency instrument through a small resistor to ground, and back up to the secondary coil of the discriminator through another resistor and an r-f choke. At the center frequency, \underline{I}_1 equals \underline{I}_2 and $\underline{I}_{\underline{L}}$ is zero. At frequencies above and below center frequency, $\underline{I}_{\underline{I}}$ has positive and negative values, respectively, and follows a linear characteristic with frequency as does the voltage $\underline{E}_{\underline{L}}$ in a conventional discriminator circuit. The resistors $\underline{R}_{\underline{a}}$ and $\underline{R}_{\underline{b}}$ are inserted to give peak voltages proportional to positive and negative frequency swing during modulation. Voltages from these resistors supply the modulation-monitor circuits.

Since the center-frequency indication is a function of both the frequency and the amplitude of the signal at the discriminator, it is required that a particular signal level exist there for correct indication. This may be indicated conveniently by a high-resistance d-c instrument connected across \underline{R}_1 and \underline{R}_2 . For any frequency in the pass band, \underline{E}_1 is as much greater than its value at center frequency as \underline{E}_2 is less than its value at center frequency. Thus, the sum of \underline{E}_1 and \underline{E}_2 is constant over the pass band, and $\underline{E}_{\underline{L}}$ is a direct voltage even during modulation. For reasons of economy, the percentage-modulation instrument can be switched into the level-measuring circuit when calibration is desired.

The de-emphasis circuit is a standard resistance and capacitance circuit; it is shown between $\underline{R}_{\underline{b}}$ and point \underline{C} in the figure. The amplifier which raises the level across $\underline{R}_{\underline{a}}$ or $\underline{R}_{\underline{b}}$ to a value sufficient for percentage-modulation measurements is highly stabilized by negative feedback; the gain of this amplifier is only one thirtieth of what it would be without the feedback. This results in a calibration of the percentage-modulation instrument which is virtually independent of aging or changing of the percentage-modulation-circuit tubes.

17-5. Frequency-Modulation Spectrum Analyzer. The f-m spectrum analyzer makes an instantaneous plot of the component frequencies and their amplitudes and presents them on the screen of a cathode-ray oscilloscope. The screen of the tube may be calibrated in frequency deviation so that the frequencies of the individual sidebands can be determined.

Figure 17-12 shows a block diagram of the equipment.⁵ It is simply an a-m superheterodyne receiver capable of operating at the f-m signal frequency. The center frequency of the h-f oscillator is 2 megacycles below the input signal carrier and, using a linear sweep voltage, is swept across a band of ± 100 kc, by means of a reactance-tube modulator, at the rate of 30 cps. This "scans" the spectrum 100 kc above and below the carrier frequency and "reports" to the i-f amplifier any signal it encounters during its excursion. In the mixer circuit, the sideband components of the f-m input signal beat with the sweeping frequency of the h-f oscillator. Thus, the components are all translated to the intermediate frequency and pass through a very narrow band i-f amplifier. A narrow-band amplifier is necessary to separate the individual components from the complete signal. The individual components are then detected by an a-m second detector which, since it is sensitive to amplitude modulation, will



Fig. 17-12. A block diagram of the frequency-spectrum analyzer showing its resemblance to an ordinary a-m superheterodyne.



Fig. 17-13. Sketches of the frequency spectrum of an f-m wave as obtained on a cathode-ray oscilloscope screen using the system shown in Fig. 17-12.

deliver an output equal to the amplitude of the component which happens to be passing through i-f amplifier. Thus, a sharp unidirectional pulse is obtained for each component in the frequency spectrum of the incoming signal. These pulses are then applied to the vertical deflection plates of an oscilloscope.

The pulses representing the components are properly spaced on the oscilloscope screen by applying to the horizontal plates of the oscilloscope the same 30-cps linear sweep voltage applied to the reactance-tube modulator. The resulting pattern on the screen will show the amplitude of the carrier and sidebands with their relative spacings and frequencies.

Sketches of the traces obtained on the oscilloscope for different f-m waves are shown in Fig. 17-13. The components are indicated by sharp peaks which are determined by the band width of the i-f amplifier, the peak being an actual trace of the amplifier response curve. Figure 17-13a shows the trace obtained with only a carrier input; (b) indicates the trace obtained with a 7.5-kc deviation and a 15-kc modulating frequency; (c) depicts the trace obtained with a 36-kc deviation and a 15-kc modulating voltage, under which conditions the carrier component reduces to zero; and, finally, (d) is the trace obtained with a 75-kc deviation and a modulating audio frequency of 2.5 kc.

REFERENCES

1. Donald M. Hill and Murray G. Crosby, "Design of F-M Signal Generator," Electronics, November, 1946, p. 96.

2. A. W. Barber, C. J. Franks, and A. G. Richardson, "A Frequency-Modulated Signal Generator," <u>Electronics</u>, April, 1941, p. 36.

3. These requirements are based on the <u>F-M</u> Standards of Good Engineering Practice issued by the FCC as of November, 1946.

4. H. R. Summerhayes, Jr., "A Frequency-Modulation Station Monitor," <u>Proc.</u> IRE, September, 1942, p. 399.

5. Roger J. Pieracci, "A Frequency-Modulation Monitoring System," Proc. IRE, August, 1940, p. 374.

QUESTIONS

1. What are the basic requirements for a signal generator?

Why is modulation calibration so difficult in an f-m signal-generator design?
Describe the heterodyne system as used in an f-m signal generator.

4. Explain the operation of the constant-deviation variable-oscillator system as used in an f-m signal generator.

5. Describe how a reactance-tube modulator for a constant-deviation variableoscillator f-m generator may be corrected for constant deviation by means of an output potentiometer. 6. Describe how a phase-shift network in a reactance-tube modulator can be used to produce constant deviation in an f-m generator.

7. How can a class C amplifier be used to double the frequency coverage of an f-m signal generator?

8. What advantage is there to using two class C amplifiers following the f-m oscillator in a signal generator?

9. What are the requirements for an f-m station-frequency monitor?

10. Enumerate the requirements for an f-m station-modulation monitor.

11. How can a frequency regulation method used in f-m transmitter construction be used as a station-frequency monitor?

12. Explain the use of a precision-type receiver as a frequency monitor in an f-m station.

13. Describe, by means of a block diagram, the various components of an f-m station monitor operating on the principle of a precision-type receiver.

14. Explain, using a circuit diagram, the operation of the current-sensitive discriminator.

15. Explain, by means of a block diagram, the operation of an f-m frequency-spectrum analyzer.

CHAPTER 18

Frequency-Modulation Servicing

Except for the detector, the f-m receiver is basically the same as the ordinary a-m superheterodyne receiver. In addition, the detector employed in the f-m receiver has to employ a limiter to remove any effect of amplitude modulation, or it has to be inherently insensitive to amplitude modulation. Those circuits preceding the detector in an f-m receiver differ only in band width and the fact that they operate at higher frequencies than those of the ordinary broadcast a-m superheterodyne receiver.

Thus, many of the servicing procedures for the a-m receiver are directly applicable to the f-m receiver, and the servicing problems are very much the same. A thorough knowledge of a-m receiver servicing must be acquired before attempting to learn the specialized problems of f-m receiver servicing, such as the problems of noise and distortion.

18-1. Alignment Using an Amplitude-Modulation Signal Generator.¹ In many cases an f-m receiver has to be aligned when no f-m signal generator is available. The manufacturer's instructions nearly always specify an a-m signal generator method, and wherever possible the manufacturer's instructions should be followed as closely as possible. However, very satisfactory results can be obtained with the general instructions which follow when more specific instructions are not available.

Usually in a combination a-m - f-m receiver the general rule is to align as many tuned circuits as possible in the proper sequence with one signal generator setting. The order of alignment may be as follows: the i-f stages of the a-m section, the i-f stages of the f-m section, the detector of the f-m section, the r-f and oscillator stages of the a-m section, the r-f and oscillator stages of the shortwave band section, and the r-f and oscillator stages of the f-m section.

The essential equipment required for the alignment of an f-m

receiver when only a conventional a-m signal generator is available is as follows:

1. An a-m signal generator covering the frequency range of the receiver to be aligned. For a broadcast f-m receiver, the signal generator has to cover a frequency range to 108 megacycles.

2. A vacuum-tube voltmeter. If a vacuum-tube voltmeter is not available, a d-c microammeter in series with a 100,000-ohm resistor and a double-pole double-throw switch to reverse the meter connections may be used; or, if available, a cathode-ray oscilloscope may also be used.

18-2. Alignment of the Intermediate-Frequency Stages of the Frequency-Modulation Receiver. The following steps may be employed to align the i-f stages of an f-m receiver:

1. First, determine the intermediate frequency of the f-m receiver. The frequency is usually marked on the chassis or transformer can. The RMA standard is 10.7 megacycles, but 4.3 megacycles, 9.1 megacycles, or other frequencies may be encountered in some sets. The intermediate frequency of the a-m section is usually 455 kc. In many cases, the i-f transformers for the f-m and a-m sections are housed in the same shield can and it is necessary to determine which of the trimmers are for the f-m transformer and which are for the a-m transformer.

2. The vacuum-tube voltmeter is then connected between ground and the grid side of the first limiter-tube grid resistor. This is



Fig. 18-1. The connections for a standard a-m signal generator and vacuumtube voltmeter when used for the alignment of the i-f stages.

illustrated in Fig. 18-1. The negative d-c voltage developed when the signal is applied will serve as the output indicator during the alignment. Thus, the limiter is employed as an a-m detector and the i-f alignment can be made as in an a-m receiver.

3. The input from the signal generator is then connected to the receiver. The preferred connection is to connect the signal generator to the receiver through a capacitor of about 0.05μ f to the grid of the last i-f tube, and align the transformer feeding the limiter stage (Fig. 18-1). Then shift the signal generator to the preceding stage

and align the next transformer, working backward stage by stage to the grid of the converter. It is sometimes possible, or even necessary, because of inadequate output from the signal generator, to connect the signal generator to the converter grid at the start and align the set, starting from the back. But the stage-by-stage connection is the safest, particularly when overcoupled stages are encountered. The signal generator is used unmodulated unless an oscilloscope is employed as the output indicator.

4. The trimmers of the i-f transformer are adjusted for maximum output, as indicated on the vacuum-tube voltmeter. Interaction between the trimmers is indicated when the setting of one trimmer has an effect upon the setting of the other. If there is little or no interaction between the trimmers, the stage is not overcoupled and the alignment is as straightforward as the usual a-m alignment. The higher intermediate frequencies usually make overcoupled transformers less necessary and simplify the alignment. If the i-f transformers are of the flat-topped or overcoupled type, two maximum responses will be observed as the frequency of the signal generator is tuned through the frequencies from about 100 kc above and below the intermediate frequency. For overcoupled transformers, the trimmers are aligned for equal response of the two peaks equally spaced on each side of the center frequency, the response in the center to be not less than 90 per cent of the peak response. Each stage is aligned in succession, back to the converter stage.

18-3. Alignment of the Discriminator. To align a discriminator circuit, the following steps may be employed:

1. The vacuum-tube voltmeter is connected between the high side of the discriminator load resistors and ground, measuring the d-c voltage across the resistors. This is illustrated diagrammatically



Fig. 18-2. The connections for the vacuum-tube voltmeter in the alignment of a discriminator.

in Fig. 18-2. The signal generator is connected to the grid of the last i-f transformer through a $0.05-\mu f$ capacitor. No modulation of the signal generator is required.

2. The frequency of the signal generator is then varied about 100 kc above and below the intermediate frequency. A plus and

minus voltage should be developed on the voltmeter by the frequencysensitive discriminator circuit. With the signal generator set at the intermediate frequency, the secondary of the discriminator transformer is set at <u>nearly</u> zero. (If the voltmeter reads exactly zero when the signal generator is tuned to the intermediate frequency, the secondary is detuned slightly so that a reading is obtained.) Now the primary trimmer is adjusted for a maximum voltage reading on the vacuum-tube voltmeter. The secondary trimmer is readjusted so that the output voltage as indicated the vacuumtube voltmeter is exactly zero.

3. The frequency of the signal generator is then swung above and below the intermediate frequency, producing a variation in the output voltage. The maximum positive voltage produced at the output should nearly equal in magnitude the maximum negative voltage produced at the output. These maxima should also occur at equal frequency intervals on either side of the intermediate frequency. For accurate alignment, a curve of voltage output versus frequency, as



Fig. 18-3. An f-m detector characteristic with the tuning points for a 10.7-megacycle i-f transformer as used in an f-m broadcast receiver.

shown in Fig. 18-3, can be plotted. The intermediate frequency for this figure is 10.7 megacycles, and the peaks occur at 10.8 and 10.6 megacycles, both equally displaced 100 kc from the intermediate frequency. With correct alignment the magnitude of the positive peak will be equal to the magnitude of the negative peak, as indicated in the figure. If the resultant characteristic of the discriminator circuit is not correct, the alignment procedure should be repeated.

18-4. Alignment of Receivers Employing the Ratio Detector. Receivers that employ the ratio detector do not employ a limiter stage, and the alignment procedure given in the previous sections must be modified as follows:

1. In place of connecting the vacuum-tube voltmeter across the limiter grid resistor, it is connected from the negative end of the load resistor to ground, as shown in Fig. 18-4. Otherwise the alignment of the i-f stages, as discussed in Sec. 18-2, is the same. 2. The ratio-detector primary is tuned for maximum output on the vacuum-tube voltmeter, using the same connections as in step 1



Fig. 18-4a. Voltmeter and signal-generator connections for a ratio detector.



Fig. 18-4b. The connections for a vacuum-tube voltmeter employing two resistors for the alignment of a ratio detector with one side of the load resistor grounded.

and with the signal generator set at the intermediate frequency.

3. The secondary of the ratio-detector transformer is aligned with the voltmeter disconnected from the load resistor but connected to the audio output before the coupling capacitor, as shown. The secondary is now adjusted for zero d-c voltage.

If the center tap of the load resistor is not grounded, but one end is grounded as shown in Fig. 18-4a, it is expedient to connect two equal resistors across the load resistor and connect the ground terminal of the vacuum-tube voltmeter to the center tap of these resistors, when aligning the secondary of the ratio-detector transformer for zero output at the intermediate frequency. The added resistors should be about 10 times the sum of the load resistors used in the circuit.

18-5. Alignment of Receivers Employing the Single-Stage Oscillator Detector. The single-stage f-m detector employing an oscillator is characterized by the absence of the conventional limiter, diode detectors, and discriminator circuits, whose functions are combined in a single tube, a special heptode type FM1000 (see Sec. 9-6). The first and second grids of this tube are used as grid and anode of an oscillator which nominally operates at the intermediate frequency. The output of the i-f amplifier is fed into the injection grid of the heptode. The plate is reactively coupled to the oscillator circuit and causes the oscillator to lock in and follow the variations in the frequency of the i-f signal. This, in turn, causes variations in the plate current which are linear with respect to frequency deviation and the audio signal is obtained from the plate circuit.

To align receivers using this circuit, an a-c range on the vacuumtube voltmeter is employed. If an oscilloscope is available, it could be used instead of the vacuum-tube voltmeter. The vacuum-tube voltmeter or oscilloscope is connected across the voice coil of the receiver or some other convenient point in the audio amplifier whereby audio output from the detector can be measured. A jumper is then connected from the first grid (pin 2) of the heptode to ground. This jumper shorts the oscillator section and the tube functions as an a-m detector. With modulation on the signal generator, the i-f stages are aligned, using the audio output as the indicator.

To align the f-m detector, an unmodulated signal at the intermediate frequency is applied to the grid of the last i-f stage. The jumper from grid 1 to ground is disconnected and another jumper across the tuned circuit in the plate circuit of the heptode is inserted. The trimmer across the oscillator circuit is then adjusted for zero beat between the oscillator frequency and the applied intermediate frequency from the signal generator. This is the point at which the audio output drops sharply to zero. The jumper across the plate-tuned circuit is then removed. The output of the signal generator is kept low, below the level which causes the oscillator to lock in, and the trimmer in the plate circuit is adjusted for zero beat. A single very sharp zero-beat point should be obtained.

18-6. Alignment of the Radio-Frequency Section. The alignment of the r-f section is straightforward, the same type of indication being used for these circuits as was used for the alignment of the i-f stages. For a receiver employing a discriminator circuit, the vacuum-tube voltmeter is used to measure the d-c voltage across the limiter-grid resistor as the output indicator. The signal-generator output is connected through a 300-ohm resistor to the antenna terminal with its low end going to the ground on the chassis. If the antenna winding is balanced to ground, it may be desirable to connect the antenna terminals to the signal generator through two 150ohm resistors.

With the signal generator unmodulated and tuned near the high end of the band, the dial of the receiver is set to the signal-generator frequency by adjusting the oscillator trimmer. The antenna and r-f trimmers are adjusted for maximum output. The signal generator is then tuned to a frequency near the low end of the band and the oscillator padder is adjusted, if one is provided, or the oscillator inductance is adjusted by spacing the turns on the oscillator coil nearer or farther apart until the dial reading coincides with the signal-generator frequency. Adjust the padders or inductance of the antenna and r-f coils for maximum output. The adjustment at the high end of the band should be repeated and the low end of the band checked again. These two tests should be repeated until satisfactory alignment is obtained.

The alignment of the f-m receiver should now be complete.

18-7. Visual Alignment Using the Oscilloscope.² For the visualalignment procedure, a wide-band signal generator covering the i-f band of the receiver with a sweep circuit of ± 200 kc is required. In addition, a cathode-ray oscilloscope of the wide-range type, using preferably a 5-in. tube, is needed. The alignment of the r-f sections is not usually performed visually and requires the use of the standard signal generator and vacuum-tube voltmeter as described in Sec. 18-6.

Alignment should be started by connecting the vertical amplifier of the oscilloscope across the grid resistor of the limiter circuit, if a discriminator and limiter are employed or, if one of the other types of f-m detector is used, across the equivalent a-m detection point as described in the previous sections. The low side of the vertical amplifier of the oscilloscope should connect directly to ground, and the high side should connect to the grid load resistor through a 1/2-megohm resistor to isolate the connecting lead. A shielded lead should be used on the oscilloscope connections to the receiver to prevent oscillation of the i-f stages caused by feedback through the added leads. In some receivers the oscilloscope and signal generator have to be grounded to the same point on the receiver chassis to prevent oscillation.

The wide-band sweep-signal generator is adjusted so that the middle or center of the sweep is exactly at the intermediate frequency of the receiver to be aligned. The sweep width is then adjusted to ± 200 kc. The high side of the signal generator is then connected to the grid of the preceding i-f tube through a $0.05-\mu f$ capacitor. The low side of the signal generator should be connected to the same ground point as the oscilloscope ground. These connections are shown in the block diagram of Fig. 18-5. The horizontal sweep of the oscilloscope should be adjusted to synchronize with the synchronizing pulses developed in the signal generator in order to keep the pattern developed on the oscilloscope stationary. This is accomplished by feeding the synchronizing pulses from the signal generator into the synchronizing or trigger input of the oscilloscope.

Two curves will be obtained on the oscilloscope. One curve will be the trace as the signal generator sweeps from the low end of the band to the high end, and the other curve, the trace as the frequency sweeps from the high end of the band to the low end. The last i-f transformer (or "limiter-input transformer," as it is sometimes called) is now aligned by adjusting the primary and secondary



Fig. 18-5. A block diagram showing the connections for the visual alignment of the i-f stages in an f-m receiver.

trimmers so that the curves on the oscilloscope screen coincide and are as flat-topped as can be obtained. By making the curves coincide, the i-f curve is made symmetrical about the intermediate frequency. To check the width of the flat-topped portion, the sweeping range of the signal generator is reduced until the trace on the oscilloscope is a magnified picture of only the flat portion of the i-f curve. Depending on the design of the i-f transformer, the flat portion should be in the neighborhood of ± 150 kc, which will be indicated by the setting of the sweep control when only the flat portion of the i-f curve is traced on the oscilloscope.

After the last i-f transformer is adjusted, the input of the signal generator should be connected to the grid of the next preceding i-f tube (if there are two i-f tubes, which is the usual case). The trimmers on the i-f transformer connected to the plate of this tube are adjusted for coincidence of the forward and reverse sweeps, and the curve is again made as nearly flat-topped as possible without the loss of amplitude. The width of the flat-topped portion should again be checked by decreasing the sweep of the signal generator. Sometimes, if desired, the curve may be made flatter over the range of the i-f band by sacrificing amplitude and, in the case of overcoupled transformers, by taking a greater drop in amplitude in the center of the band.

The next step is to connect the signal generator to the grid of the converter tube through an isolating capacitor, and the trimmers on the first i-f transformer are adjusted as before. This completes the alignment of the i-f transformers and the next step is to align the detector transformer. The i-f transformers have been aligned stage by stage, and no over-all adjustments should be made after completing the stage-by-stage alignment.

For the alignment of the detector stage the signal generator is left connected to the grid of the converter tube, and its frequency setting remains at the intermediate frequency. The leads from the vertical deflection circuits of the oscilloscope are now placed across the audio output of the detector. The output can be taken from across the load resistor of the detector or from any convenient point in the audio amplifier. The trimmers on the detector are now adjusted so that two crossing f-m detector curves are obtained (one for each direction of sweep). The crossover point <u>a</u> should be at the center of the characteristic, and the straight-line portions of the curves, to cover the required ± 75 kc, should extend over the required proportion of the curve. This may be checked by reducing the sweep to ± 75 kc and noting whether the oscilloscope traces become two crossing straight lines.

In the case of the discriminator and ratio detector, the primary trimmer is adjusted by reducing the sweep range to a low value and adjusting the primary trimmer for maximum value. The secondary trimmer is adjusted by using the wide-frequency deviation sweep and adjusting for the symmetrical crossover point \underline{a} .

The alignment of the r-f stages is performed with a standard signal generator and a vacuum-tube voltmeter, as described in the previous sections.



Fig. 18-6. Discriminator-alignment curve. At a we have the crossover point which should be midway between peaks.

18-8. Preliminary Steps in Servicing the Frequency-Modulation Receiver. As stated previously, the servicing of an f-m receiver follows basically the same procedure as the servicing of an a-m receiver. Many of the steps in the servicing procedure are the same, but others are radically different. Let us assume that a receiver with an f-m band has been brought to us for repair, with the request that we check the complete f-m section to make sure that it is operating at its best. The radio itself is completely dead.

The preliminary steps are the same as for any electronic device. First, check to see if the filaments of the tubes are lighting. If not, the line input and input fuses should be checked as in the ordinary a-m receiver case. If the filaments are lit when the set is turned on, it should be turned off immediately and all tubes tested in a tube tester. Very often the complete trouble lies in a defective or weak tube. Although the tests in a tube tester are never absolutely positive (tubes which test perfectly good on a tube tester



Fig. 18-7. A circuit diagram of a combination a-m





-f-m receiver employing a ratio detector for f-m.

are sometimes inoperative in the receiver), the preliminary step of tube testing should not be omitted in any case.

After eliminating the line input and the tubes as a source of trouble, the next steps involve actual checking of voltages at points in the circuit of the receiver, and for these tests the circuit diagram of the receiver should be obtained. Let us assume that the circuit diagram shown in Fig. 18-7 is a circuit diagram of the receiver under test. This receiver is a combination a-m-f-m receiver with a built-in loop antenna for the a-m section and a built-in dipole antenna for the f-m section. A ratio detector is employed in the f-m section for demodulation of the f-m signal.

As an aid to servicing the receiver, the manufacturer has included the voltages to ground as well as the resistance to ground from the pins of each of the tube sockets in the receiver. Thus, with an ordinary multitester having the specified 20,000 ohms per volt for the d-c scales, and 1,000 ohms per volt for the a-c scales, all of the operating voltages in the receiver can be rapidly checked. Nearly all radio receiver manufacturers supply the operating voltages in one form or another for servicing purposes. If there is no B+ voltage available at any of the tubes, the trouble is probably in the power-supply circuit. The power-supply circuit can be tested in the standard manner by removing the rectifier tube and checking for a-c voltage and then checking the filter circuit for open load resistors, or voltage dividers, or shorted filter capacitors. An overload of the power supply will usually be indicated by a lowering of the output voltage and can be located by resistance measurements since overload takes place when the load falls in resistive value.

After these preliminary checks, if the trouble has not yet shown up by a voltage or resistance check, some method of signal tracing has to be employed.

18-9. Servicing the Audio Amplifier. To test the audio amplifier by signal insertion, it is necessary to have a good wide-range audio oscillator with sufficient output to operate the speaker directly through the output transformer. Thus, the audio signal is inserted between the plate of the output 6V6 and ground to test the output audio circuit of the receiver. A shorted bypass capacitor, an open transformer winding, or an open speaker coil will result in no audible output. By applying sufficient audio signal, it is also possible to check the speaker for rattles and distortion at high output levels. If it is found that the low-impedance audio-amplifier transformer tends to short-circuit the audio amplifier, a high resistance in the order of 25,000 to 50,000 ohms should be inserted in series with the audio-oscillator output leads.

To test the operation of the last audio stage — the 6V6 audiopower amplifier — the audio signal is fed into the grid of the 6V6 tube. The audio-oscillator output is now reduced in proportion to
the gain of the 6V6 stage. No output from the speaker, or a distorted output, indicates that the trouble is in that stage and all components in the circuit of the tube should be checked.

If all the tests on the 6V6 power amplifier show that it is operating properly, the next step is to test the operation of the first audio amplifier which, in this case, is the triode section of the 6SQ7 tube. However, since this tube is in a rather complex circuit with the tone and volume controls, let us first examine the operation of



Fig. 18-8. The circuit of the 6SQ7 triode section in the audio amplifier of the receiver of Fig. 18-7.

the circuit. Figure 18-8 is a simplified diagram of the circuit. A 2-megohm volume control with a 1-megohm tap is used for a volume control at the input. If the 4-megohm control were not connected to the junction of the 100,000-ohm resistor and the $0.002-\mu f$ capacitor, it would be a straightforward tone-compensated volume-control circuit accenting the bass at low levels. The tone control consists of the 4-megohm control in series with a $0.01-\mu f$ capacitor across the plate circuit of the 6SQ7 tube. As the series resistance is reduced, the high frequencies are bypassed to ground through the 0.01-µf capacitor, thereby cutting down the high frequency response of the amplifier. This occurs when the tap on the 4-megohm resistor is moved to the far right in Fig. 18-8. At the same time, since 4 megohms is very large with respect to the reactance of the 0.002-µf capacitor, the tone-compensation circuit is operating at full capacity. As the tone-control tap is moved to the left in the figure, the high-frequency audio signals are allowed through the audio amplifier circuit, and the tone compensation is kept in operating condition until the tap on the tone control approaches the far left. At that point, the 0.002-µf capacitor is shunted to ground and any tone compensation is removed. Thus, in effect, the highs are accented still more. De-emphasis is accomplished by shunting the plate resistor with a $220-\mu\mu f$ capacitor.

The tone-control circuit may be tested by inserting the audio

signal on the plate of the 6SQ7 tube and varying the tone control. A 5,000-cps signal from the signal generator should be progressively cut down as the tone control is turned toward the bass accentuation.

The signal-generator output leads should now be shifted to the grid of the 6SQ7 tube and the tests repeated. With both the volume and tone controls set at their center points, bass-tone compensation should be evident. If the tone-compensation capacitor were shorted, the volume would be seriously reduced but not cut off. However, no bass compensation would be present. This can be tested by moving the tone-control tap to the far left in the figure (maximum h-f response), and then moving it only a small amount away from its maximum point. If the volume control is set at its center and the audio signal being fed into the grid circuit is around 60 cps, an increase in volume should be checked.

18-10. Servicing the Detector and Intermediate-Frequency Stages. Let us now confine ourselves to the f-m section of the receiver, since the a-m section can be tested in the ordinary manner.

The next stage to be checked, proceeding back toward the input, is the detector stage which, in this case, is a ratio detector. The ratio-detector transformer uses a fixed-slug capacitively tuned secondary. Capacitor tuning is used for the secondary to prevent unbalancing of the two sections, which may take place with slug tuning. However, the slug is used not only to increase the Q of the transformer, but also to aid in balancing the two sections in manufacture.³

The large time constant for the operation of the ratio detector is established by the 25 μ f and 10 μ f capacitors and their associated resistors. The audio output is taken off the tertiary winding through a filter network and across a 100,000-ohm resistor to ground. The a-v-c voltage is taken off the 1,000-ohm resistor in the large timeconstant circuit.

The detector circuit can best be tested with a combination f-m-a-m signal generator. If an oscilloscope is available, it can also be used to good advantage. The signal generator should be adjusted to deliver a center frequency equal to the intermediate frequency. For the receiver shown in Fig. 18-7, the frequency would be 10.7 megacycles. The signal is then inserted before the f-m detector transformer. This point in the receiver of Fig. 18-7 would be between the plate of the 7AG7 tube and ground. An isolating capacitor of about 0.05µf should be used as in alignment procedure.

The input signal should now be frequency-modulated, starting at a very low level and increasing the modulation until full deviation is used. For broadcast receivers this would mean a deviation of ± 75 kc. The signal should come through the speaker without distortion. A better check than just listening to the signal can be obtained by watching the audio signal on an oscilloscope, connecting the oscilloscope across any convenient point in the audio circuit.

The frequency modulation should then be removed and amplitude modulation inserted on the input signal. Very little signal should now get through to the output. The a-v-c circuit is checked by putting a vacuum-tube voltmeter from the grid of any of the tubes with automatic volume control on them (for instance, the 7AH7 tube in the diagram) to ground and noting the d-c voltage as the intensity of the signal from the signal generator is varied. For this latter test no modulation is necessary. Any troubles or misalignment of the detector stage (or limiter stage, if one is employed) should show up under these tests. When a limiter stage is employed, the a-v-c voltage is usually obtained from the limiter and, of course, the amplitude modulation is removed by the limiter.

To test the i-f stages, the signal generator is moved back stage by stage, and the f-m and a-m modulation tests are repeated. If one of the i-f stages should be badly detuned, the discrimination between f-m and a-m signals may be affected. If stagger-tuned i-f stages are used (single-tuned i-f transformers tuned at various frequencies to get the necessary band width), it is best to test the complete i-f amplifier at once by inserting the signal at the grid of the converter stage. The attenuation of the input signal as the input is moved back stage by stage is an indication of the gain of each stage. If the signal has to be increased instead of attenuated as the input from the signal generator is moved back a stage, it is an indication that the stage is not functioning properly and should be carefully checked.

Distortion at high levels of deviation is an indication of improper alignment of the i-f stage. However, the alignment of any of the stages should never be touched until all other factors in the circuit have been carefully checked and it is definitely certain that the trouble is in misalignment.

It is always a wise move to make the local oscillator inoperative when testing the i-f stages in order to keep out extraneous signals. This can usually be accomplished by putting a shorting lead across the oscillator tuning capacitor.

18-11. Servicing the First-Detector, Oscillator, and Radio-Frequency Sections. Before discussing the servicing of the firstdetector, oscillator, and r-f stages, let us first examine the operation of the circuits shown in Fig. 18-7 for these stages. The f-m antenna coil is tapped at 300 ohms for matching with a transmission line of the same impedance. The tuning section of the f-m portion is a three-gang capacitor and a 6AG5 is used as an r-f amplifier. The control grid of this r-f amplifier is coupled through a capacitor to the antenna coil, and series-fed from the a-v-c system through a 1-megohm resistor. The 6AG5 plate is fed into an r-f transformer, and a tap on the secondary of the transformer feeds the control grid of the converter section of the 7F8 tube. The control grid is tapped down on the coil for two reasons. First, the input capacitance of a triode is much higher than that of a pentode. And second, the input impedance of any tube is much lower at high frequencies than at broadcast frequencies. Therefore, since both the tube-input capacitance and impedance will vary, they will affect the frequency stability of the circuit. In addition, the Q of the coil, and, consequently, the selectivity of the circuit, is improved by reducing the tube-loading effect. The oscillator uses the other section of the 7F8 tube as a Hartley oscillator with the cathode tapped off on the oscillator coil. By connecting the cathode return of the converter stage to the cathode of the oscillator, the oscillator signal is cathodecoupled into the converter stage.

The first step in testing these sections is to tune the receiver to the high end of the band and feed an r-f signal corresponding to the calibration of the receiver dial into the converter grid through an isolating capacitor, as in the alignment procedure. The r-f signal is then frequency-modulated to about full deviation. The first thing to find out is whether the oscillator is oscillating and, second, whether it is tracking. If the oscillator is operating correctly at the correct frequency, the demodulated signal should appear at the output.

If no signal appears, either the oscillator is not oscillating at all, or it is not oscillating at the correct frequency. To determine whether the oscillator is oscillating off frequency, the r-f signal input should be shifted slowly above and below the calibrated frequency on the receiver dial. A signal output will appear when the signal generator is shifted an amount in frequency which is determined by the amount that the oscillator is off frequency. This would indicate the need for realignment of the oscillator tracking capacitors. If the oscillator is not functioning, no signal will appear at the output, no matter what frequency in the r-f band is employed at the signal generator.

If it is thought that the oscillator is not functioning, then the proper check to make is to substitute the signal generator for the oscillator and note whether the receiver functions. If the oscillator is a fundamental-frequency oscillator, which is tuned to the input frequency minus the intermediate frequency, as in the ordinary receiver, then the signal generator is tuned to that frequency and, with all modulation removed, the signal-generator output is fed into the converter. For the receiver shown in Fig. 18-7, the signal would be fed into the cathode of the triode converter section of the 7F8 tube. If the receiver is tuned to a broadcast station, the station signal will be heard at the output if the only trouble with the receiver is an inoperative oscillator section. The trouble in the oscillator should then be located with ordinary meter checks of the component parts. The signal generator is then connected to the grid of the r-f tube and the f-m signal again put into the receiver. The decrease in necessary signal is again a measure of the gain of the stage. Image rejection can be tested by decreasing the radio frequency of the signal generator until it is twice the intermediate frequency below the calibrated frequency on the receiver dial. The output of the receiver should be very low and is a measure of the image rejection. If all tests are successful, the signal generator is then placed at the antenna terminals with an appropriate dummy-antenna network, as suggested by the receiver manufacturer. (An example of a dummy-antenna load for the receiver of Fig. 18-7 is





shown in Fig. 18-9.) If the antenna coils are in good condition, we can check the operation of the r-f amplifier and check again on the operation of the oscillator-converter and i-f sections of the receiver.

It is always wise to check the antenna lead-in wire and antenna itself before giving the final approval. It can be checked for shorts between leads and shorts to ground roughly with a multitester, but actual visual inspection of the antenna installation should not be omitted. A small test receiver is very useful in testing the antenna installation and the expected field intensity at the receiver.

18-12. Servicing with the S Curves.⁴ The S-curve patterns, illustrated in Fig. 18-10, are f-m detector output characteristics which are obtained by applying the r-f output of an f-m signal generator (or sweep generator) to the receiver input, while feeding a portion of the generator's modulating voltage to the oscilloscope.

If the generator's modulating voltage is a saw-toothed wave, and provision is made for connecting a portion to some external apparatus, the modulation may be applied directly to the horizontal amplifier terminals of the oscilloscope. If the sweep voltage is a sine wave, however, better results may be obtained by connecting to the external synchronizing input, and using the internal sawtooth of the oscilloscope.

The vertical plates are connected to the output of the detector. As a result of the repetitive sweeping of an r-f signal through the f-m receiver, the detector output voltage varies in value and polarity, producing what is called an S curve; the normal f-m detector S curve being illustrated in Fig. 18-10a.

In Fig. 18-10b the pattern reveals an S curve that is acceptable, except for its narrowness, indicating a narrow over-all band width.



Fig. 18-10. S curves showing characteristic troubles in an f-m receiver.

which would result in loss of high-pitched tones. The insufficient band width will also result in audio clipping at the higher modulation levels and a distortion very noticeable to the listener. Repeating the alignment procedure will usually locate troubles causing this type of S curve.

The severe unbalance of the pattern shown in Fig. 18-10c shows the result of a discriminator that is not tuned to the center frequency of the r-f-i-f circuits, while the next view, Fig. 18-10d, shows the case of a local oscillator that is not tuned to the center frequency of the i-f and detector circuits. It is presumed, of course, that a well-calibrated f-m generator is being used.

The fifth diagram, Fig. 18-10e, shows the pattern observed when the f-m generator sweeps through an i-f amplifier with a bad tendency to regeneration. The last S curve, shown in Fig. 18-10f, is the result of a detuned, or nontracking, r-f amplifier.

Thus, by means of an S curve, a very rapid determination of the over-all characteristics of the f-m receiver up to the audio amplifier can be obtained very quickly, and very often a great deal of time can be saved in the servicing operation.

REFERENCES

1. Fred Dalasta and W. P. Mueller, "Servicing with a Modern Oscilloscope," Sylvania News (Tech. Sec.), Vol. 14, No. 8, September, 1947.

2. "F-M Alignment," Service, May, 1947, p. 16.

3. Alvin A. Baer, "Portable F-M/A-M Receiver," Service, October, 1947, p. 22. 4. "Servicing FM Receivers with the Oscilloscope," Radio and Television Retailing, May, 1947.

QUESTIONS

1. What equipment is necessary to align an f-m receiver when using only a conventional a-m signal generator?

2. Describe the alignment of the i-f stages of an f-m receiver using only a conventional a-m signal generator.

3. Describe the alignment of a discriminator detector when only a conventional a-m signal generator is used.

4. How is a ratio detector aligned, using a conventional a-m signal generator?

5. Explain how the single-stage oscillator detector is aligned.

6. Describe the alignment of the r-f sections of an f-m receiver.

7. Describe the method of visual alignment employing an oscilloscope.

8. Why is it important to ground the oscilloscope and signal-generator grounds to the same point on the receiver chassis?

9. What are the preliminary steps to be taken in the servicing of an f-m receiver?

10. How should the audio amplifier of an f-m receiver be checked?11. Why should an f-m detector be checked for both a-m and f-m reception?

12. Describe the test procedure to be used in testing the i-f stages of an f-m receiver.

13. How can it be determined whether the oscillator section is operating?

14. How can the converter section of an f-m receiver be tested?

15. Describe the final tests on the r-f sections of the receiver.

Answers

Numerical Answers, Where Required, to Questions That Appear at the End of Each Chapter

Chapter 1

5. 104.054 megacycles

•

8. 7.5 v

Chapter 2

6.	1.003 and 0.997 megacycles						
8.	Carrier amplitude - 10 v - Amplitude of each sideband - 4 v						
9.	9. First pair of sidebands - 1.2025 and 1.1975 kc						
	Second pair of sidebands - 1.205 and 1.195 kc						
	Third pair of sidebands - 1.2075 and 1.1925 kc						
	Fourth pair of sidebands - 1.210 and 1.190 kc						
12	0.43						
15	6						
16	Carrier components = 30 megacycles = 1.5 v						
10.	Eisst pair of sidebands $= 30$ megacycles $= 5$ kc $= 2.8$ v						
	First pair of sidebands $= 30 \text{ megacycles} \pm 10 \text{ kc} = 2.0 \text{ v}$						
	Second pair of sidebands = 30 megacycles ±10 kc = 2.4 v						
	Third pair of sidebands $= 30$ megacycles ± 10 kc $= 1.1$ v						
	Fourth pair of sidebands -30 megacycles ± 20 kc -3.0 v						
	Fuch pair of sidebands -30 megacycles ± 23 kc -3.6 v						
	Sixth pair of sidebands -30 megacycles ± 30 kc -2.5 v						
	Seventh pair of sidebands - 30 megacycles ± 35 kc - 1.3 v						
	Eighth pair of sidebands -30 megacycles ± 40 kc -0.6 v						
18.	110 kc						
19.	20.8 kc						
20. Carrier component - 2 megacycles - 76 v							
	First pair of sidebands -2 megacycles ± 2 kc -44 v						
	Second pair of sidebands - 2 megacycles ± 2 kc - 11 v						
	Chapter 3						
	Chapter 5						
2	Impulse noise 3. Bandom noise						
7	3 / m 8, 9,375 / m						
10	(a) 16 5 nor cont (b) 1.65 nor cont 11.47						
10.	(a) 10.5 per cent (b) 1.05 per cent 11. π .						
12.	<u>4</u>						

393

		Chapter	4				
1. 6. 9. 15.	2.65 megacycles 1.1 to 1.6 μh 93 to 108 μμf 3.98 to 4.87 megacycles	Chapter	1 1 9	5. 7. 14. 16.	16 to 24 μμf 2.53 to 2.57 megacycles 3.98 megacycles 3.98 to 5.96 megacycles		
4. 7. 13.	$\begin{array}{l} \underline{Q} = 33 \\ \overline{Resonant \ frequency} = 5.07 \\ \underline{Q}_{S} = 167 \\ \underline{Q}_{p} = 83 \\ \underline{Q}_{S} = 42 \\ \overline{R}_{1} = 40,000 \ ohms \\ \underline{L}_{S} = 32 \ \mu h \\ \underline{L}_{p} = 16 \ \mu h \end{array}$	53 megacy	ycles	5			
Chapter 12							
4. 13. 17. 20.	Over 5 megacycles 48 53 cps For the circuit shown in	Fig. 12-1	5 <u>a</u> : <u>I</u>	6. 16. 18. =1 =	11,000 600 21 kc 1.2 mh; $\underline{C}_1 = 9.36 \mu f$		
		Chapter	13				
2.	35.5						
		Chapter	14				
9.	63 mi			20.	14 kw		
		Chapter	15				

APPENDIX

Phasor Calculations

A-1. The Sine Wave. Before studying the methods of phasor calculations, let us review some of the properties of the ordinary sine wave to which phasor calculations are confined. An oscillograph trace of a sine-wave current flow, more commonly referred to as an alternating current, is shown in Fig. A-1. The wire through which the current is flowing is also shown in the figure and, to represent the sine wave adequately as a graph, one direction of current flow is noted as positive and the other as negative.

The current flow starting at any point of time, for instance at the point <u>A</u>, and going through a complete recurring period — in the figure this would be, if we start at <u>A</u>, from <u>A</u> to the positive peak <u>B</u>, then down to the negative peak <u>D</u>, and back to the zero value <u>E</u> would occupy an interval known as "one cycle." In other words, a cycle represents a complete wave. The number of cycles recurring per second is called the "frequency in cycles per second."

The maximum value that the voltage, or current, reaches during its flow in either direction is called the "peak amplitude" of the wave, or, sometimes, just the "amplitude." These are the values reached at the points B and D in Fig. A-1.

The mathematical expression for a sine wave is

$$\underline{\mathbf{i}} = \underline{\mathbf{I}} \sin 2\pi \, \mathrm{ft} \tag{A-1}$$

where <u>i</u> is the instantaneous current; <u>I</u>, the peak amplitude of the current; <u>f</u>, the frequency of the wave in cycles per second; <u>t</u>, the time in seconds; and $\tau\tau$, a constant equal to 3.14.

A-2. The Phasor. In Fig. A-2 is shown a wheel rotating about its axis. Inserted into the wheel is a pin located some distance from the center; this pin is more clearly visible in the sectional view. The pin rides in a horizontal slot of a yoke which is confined to vertical motion by means of a set of guides. The horizontal slot

WRH

APPENDIX

in the yoke allows the pin to move freely in the horizontal direction without moving the yoke, but any vertical motion of the pin will move the yoke along the vertical axis. Thus, like the well-known



Fig. A-1. An oscillogram showing the instantaneous variation in current flow in a wire conducting an alternating current. The signs + and - are used to indicate the direction of current flow.

crankshaft, the yoke extracts one of the linear components, in this case the vertical component, of the rotary motion of the pin.



Fig. A-2. A mechanical device for obtaining a sine-wave shape from a rotating wheel. A pin in the wheel drives a yoke with a pencil attached. The pencil traces a sine wave on a moving piece of paper.

The end of the yoke carries a pencil arranged to write on a strip of paper moving steadily in the horizontal direction (at right angles to the direction of movement of the yoke). As the wheel rotates and the recording paper moves horizontially, the pencil will trace a sine wave whose shape will be the same as the wave seen on an oscilloscope tracing the wave form of a pure sine wave. When the pin is lined up horizontally with the center of the wheel we have what we shall call the "zero axis" to correspond with the zero value of a normal sine wave. Above this axis in the direction of <u>B</u> is called "positive," <u>B</u> being the positive peak; below the axis in the direction of <u>D</u> is called "negative," <u>D</u> being the negative peak. For every rotation of the wheel, the pencil will draw one cycle of the sine wave. Thus, if the speed were 500 rotations per second (if that were possible in a mechanical device), the pencil would draw 500 cycles of the sine wave per second. In other words, the speed of rotation of the wheel rotates, the higher the frequency of the wave that is drawn.



Fig. A-3. The sine wave generated by a wheel with the pin located close to the center. The amplitude of the wave is small.

In Fig. A-3 is shown a wheel on which the pin is very close to the center; consequently, the wave which is drawn is very small. Its amplitude is a maximum at the point marked B, this maximum amplitude being equal to the length of an arrow drawn from the center of the wheel <u>A</u> to the pin <u>P</u>. This equality exists because the greatest distance the pencil can move is from its position when the pin is at the bottom of the rotation to its position when the pin is at the top of the rotation. These, of course, are the peak positive-amplitude and the peak negative-amplitude points. The arrow drawn from the center to the pin is called a "phasor."

In Fig. A-4 are shown two arrows <u>AP</u> and <u>AG</u>. Consider the waves that would be generated if pins were placed at the ends of these arrows and the arrows rotated. (For mechanical reasons, to avoid interference, one pin can be placed in front of the wheel and the other pin in back of the wheel, or in some other similar arrangement.) The two waves that would be drawn would be displaced from each other by a certain amount. In the arrow, or phasor, diagram the two arrows are drawn an eighth of a rotation apart. In asmuch as a circle is divided into 360 parts, called "degrees," one eighth of the circle is equivalent to one eighth of 360 degrees, or 45 deg. Similarly, each sine-wave cycle is divided into 360 parts also called "degrees," or sometimes, more exactly, "electrical degrees." The wave generated by <u>AP</u> in Fig. A-4 will reach its peak one eighth of a cycle, or 45 deg, before the wave generated by AG reaches

its peak value. Consequently, the wave \underline{AP} is said to "lead" the wave AG by 45 deg. This angular difference is called the "phase dif-



ference" between the two waves. Because of this characteristic of phase relationships, the representation of the sine waves by arrows is called a "phasor diagram," and the arrows themselves are called "phasors."

This phasor concept is a very interesting device, inasmuch as now a sine wave or, in particular, a sine-wave alternating current or voltage can be represented by an arrow. In our minds we have to picture the arrow on a wheel rotating at a speed equal to the frequency of the wave and with a pin located at the tip of the arrow. The pin can then be pictured as driving the mechanism shown in Fig. A-2 to generate the sine wave. The distance from the pin to the center of the wheel, equal in magnitude to the length of the arrow, will determine the amplitude of the wave. The phase difference in degrees between any two waves may be indicated by drawing two arrows displaced by an equivalent number of geometric degrees. The difference in amplitude between the two waves may be indicated by drawing the two arrows in proportional sizes.

Example. Shown in Fig. A-5 is a circuit consisting of a resistor and an inductor with an alternating current flowing through it. The amplitude of the voltage across the resistor is 25 v and across the inductor is 15 v. The voltage across the resistor lags the voltage across the inductor by 90 deg. Also shown in the figure is a double-trace oscilloscope which is capable of showing on its screen two waves simultaneously, so that their amplitudes and phase may be compared. Let us draw the two waves as they would be depicted on the screen of the oscilloscope, as well as the phasor diagram for the two waves. We will choose 1/4 in. as equivalent to 5 v in the drawing for the phasor diagram, thus keeping the sizes of the phasors proportional to their electrical values.

The wave across the resistor will reach a peak value of 1 1/4 in. whereas the wave across the inductor will reach a peak of 3/4 in. Inasmuch as 360 deg is a complete cycle, 90 deg is a quarter of a cycle. A phase difference of a quarter cycle, or 90 deg, means that one wave will have reached its peak value while the other (the lagging wave) is going through zero value. Assuming that the waves progress to the left

on the illustration, we note that the lagging wave will be displaced to the right, reaching its peak value 90 deg, or one quarter of a cycle, later than the leading wave. Calling the voltage across the resistor $\underline{V}_{\underline{R}}$ and the voltage across the inductor $\underline{V}_{\underline{L}}$, we can draw the two waves as shown in Fig. A-6a. The voltage $\underline{V}_{\underline{L}}$ is smaller than $\underline{V}_{\underline{R}}$, but it reaches its maximum a quarter cycle, or 90 deg, sooner than $\underline{V}_{\underline{R}}$.

The phasor diagram is shown in Fig. A-6b. Two arrows are drawn from the same center point, one 1 1/4 in. long to represent \underline{V}_{R} , and the other 3/4 in. long to represent \underline{V}_{L} . These are always pictured as rotating in a counterclockwise direction; hence, the lagging voltage \underline{V}_{R} is drawn first. The leading voltage \underline{V}_{L} must be drawn 90 deg, or 1/4 of a rotation further counterclockwise, as shown on the diagram.



Fig. A-5. The circuit used in Ex. A-1. It consists of a resistor and an inductor in series conducting an alternating current. The double-trace oscilloscope shows both waves at once, indicating their relative amplitudes and their phase relationship.

Let us review now what a phasor diagram is meant to represent and what can be learned upon examining a phasor diagram.



Fig. A-6. The wave and phasor diagrams for the two voltages $\underline{V}_{\underline{R}}$ and $\underline{V}_{\underline{L}}$ as shown in Fig. A-5. (Answer to Ex. A-1.)

The phasor diagram is a simplified picture of one or more sine waves (a-c waves) showing all of their properties and the relationships between the different waves. Phasors are all drawn as ar-

APPENDIX

rows to indicate the waves. The arrows all start at a center point and point outward like the spokes of a wheel. The length from the center to the tip of the arrow (drawn to some definite scale) is proportional to the peak amplitude of the wave. The angles between the arrows, in other words their position around the circle, show the phase relationships between the waves. If they are drawn so that they coincide, it means that they are "in phase," since they will both reach their peak values simultaneously and their zero values simultaneously. If one arrow is advanced in a counterclockwise direction by a certain angle, it means that the wave it represents is leading in phase by that angle.

A-3. The Addition and Subtraction of Phasors. One of the important contributions of the phasor method of representing sine. waves is that it leads to a simple way of adding and subtracting sine waves. In d-c circuits, if two steady-state currents are caused to flow in a wire, the total current is equal to the sum of the two cur-





Fig. A-7. Two direct currents, 2 amp flowing from battery 1 and 4 amp flowing from battery 2, flowing in the same direction through a resistor. Since they are both flowing in the same direction, they are added directly. Fig. A-8. Two a-c sources causing two currents \underline{I}_1 and \underline{I}_2 to flow in the resistor \underline{R} . The resulting current is noted as \underline{I}_R .

rents. For instance, in Fig. A-7 is shown a resistor conducting current from two battery sources. One battery is producing a current flow of 2 amp and the other a current flow of 4 amp. Since the two currents are both flowing in the same direction through the resistor, the total current flowing in the resistor is the sum of the two currents, or 6 amp.

However, where alternating currents are concerned, the process of addition is more difficult. In Fig. A-8 is shown a circuit similar to the one used in the above example, except that now alternating currents are involved. Source 1 causes a current \underline{I}_1 to flow in the resistor, and source 2 causes a current \underline{I}_2 to flow in the resistor. Both of these currents are alternating currents. The total current flowing in the resistor is noted as \underline{I}_R .

All alternating currents flow first in one direction and then in the other direction in a wire. To compare two currents, it must be stated clearly which direction of flow will be called positive in each case. In Fig. A-8 arrows are shown underneath the symbols for the currents. These arrows do NOT represent the direction of flow of the currents. They do indicate which direction is to be taken as positive when connecting an oscilloscope or when drawing a graph



Fig. A-9. The currents \underline{I}_1 and \underline{I}_2 of Fig. A-8 as seen on an oscillograph trace. Reversing the connections that pick up I_2 would cause the wave to reverse, as shown by the dotted trace.

of the various currents in the circuit. For instance, in Fig. A-9 are shown two a-c waves as they might appear on an oscilloscope. For clarity, a double-trace oscilloscope is used so that both waves can be seen simultaneously. Wave 1 is shown heavy and wave 2 is shown as a light-line trace. In this case we note that I₂ reaches its peak after I1 reaches its corresponding peak, and so it is said to "lag" I1. If, however, the probes from the oscilloscope used to pick up I₂ are reversed, then I₂ will appear as shown by the dotted line in the figure. It is equivalent to switching the wave around, since the halves of the cycles that were positive become negative, and the negative parts become positive in the figure. Now the dotted wave, 12 reversed, would lead I1, since it reaches its peak value before I1. Thus, depending on how the technician operating the oscilloscope connects the pickup probes, it is possible to get two different (but related) results for the phase relationships of the waves. For this reason, the positive direction is sometimes indicated in a-c circuits by an arrow. In those cases where only the final wave is considered and its phase relationships to other waves in the circuit are not discussed, the arrows are usually omitted.

Returning now to the problem of obtaining I_R in Fig. A-8, we can obtain the resulting current by superimposing the two waves I_1 and I_2 . We can consider them to be like two water waves. When they are both rising, the result is the sum of the two and is also rising. When they are both falling, the result is again the sum of the two and is now in the negative direction. If one wave is rising while the other is falling about the zero axis, the result is the difference between the two and is in the direction of the larger.

This process of superimposing the two a-c waves is called "addition" in a-c terminology. Addition, therefore, takes on a new meaning in a-c theory. It means, as described above, the superimposing of the two a-c waves, one upon the other, just as the two waves are superimposed simultaneously in a single conducting circuit.

The graphical method of addition of a-c waves is very laborious, and besides, it is sometimes difficult to picture exactly what is happening. The important point is that the resultant is another sine wave whose amplitude and phase are determined by the amplitudes and phase relationship of the two waves that are added.

There is a much more convenient method of adding sine waves which will also give the amplitude and phase relationships of the resultant waves. This is accomplished by using phasors. In Fig.



Fig. A-10. A phasor diagram of \underline{I}_1 plus \underline{I}_2 to obtain the resultant current \underline{I}_{R^*} . \underline{I}_1 and \underline{I}_2 are drawn proportional to the peak amplitudes of the respective currents and with an angle between them equal to the phase angle between the actual currents.

A-10 is demonstrated how phasors may be used for the addition of sine waves.

First a scale is chosen. For instance, let 1/4 in. be equal to 1 amp. In this problem let us assume <u>I</u>₁ to be 6-amp peak amplitude and <u>I</u>₂ to be 8.5-amp peak amplitude. Also, let us assume that <u>I</u>₂ leads <u>I</u>₁ by 135 deg. A point, labeled O, is used as the reference point from which all the arrows representing phasors are drawn.

The first phasor on the diagram can be drawn anywhere, but for convenience it is shown drawn horizontal and to the right. Since this phasor represents I_1 , which is 6 amp, at 1/4 in. per amp it will be 1 1/2 in. long. Inasmuch as I_2 leads I_1 by 135 deg, the phasor representing I_2 is drawn 135 deg counterclockwise from I_1 . To depict 8.5 amp, it is drawn 2 1/8 in. long.

To add the two phasors, we perform the following steps: At the tip of the phasor \underline{I}_1 , the line \underline{ab} is drawn parallel to \underline{I}_2 . Similarly, at the tip of \underline{I}_2 , the line \underline{cd} is drawn parallel to \underline{I}_1 . These lines

will intersect at the point labeled \underline{g} . An arrow from \underline{O} to \underline{g} will depict the phasor and show its amplitude and position as well as its phase relationship to the two original waves.

Examining this phasor, we see that its length is 1 1/2 in.; hence, the peak amplitude of $\underline{I}_{\underline{R}}$, the resultant current, is 6 amp. It is located 90 deg counterclockwise from \underline{I}_1 , signifying that $\underline{I}_{\underline{R}}$ leads \underline{I}_1 by 90 deg. We note here that the phasor method of addition simplifies the problem of obtaining all the information necessary to completely specify the resultant current.

Subtraction is very similar to addition in both the phasor and the graphic-wave methods. "Subtraction," again, is an a-c term meaning something different from ordinary arithmetical subtraction. Suppose that we have the resultant of two water waves acting simultaneously on the surface of a body of water. Subtraction would mean the removal of one of the waves to obtain the characteristics of the other wave. It can be accomplished by subtracting the wave point by point, analogous to the method of graphic addition. It can also be accomplished by reversing the wave to be subtracted and performing addition.

As an example of phasor subtraction let us assume that another circuit similar to that illustrated in Fig. A-8 is arranged, using a different resistor and different sources. It is adjusted so that I₁ is equal to 10 amp and I_R, the resultant current in the resistor, is equal to 12 amp. I_R leads the current I₁ by 60 deg. Since I₁ and I₂ add together to produce I_R, subtracting I₁ from I_R will yield I₂. The procedure for obtaining I₂ by phasor subtraction is demonstrated in



Fig. A-11. Phasor subtraction of \underline{I}_1 from \underline{I}_R . \underline{I}_1 is reversed and then added to \underline{I}_R to accomplish subtraction.

Fig. A-11. In this case 1/8 in. is made equivalent to 1 amp; hence, I_1 will be 1 1/4 in. long and I_R will be 1 1/2 in. long. The angle between

them is made equal to 60 deg.

The first step in phasor subtraction is to reverse the arrow representing the phasor being subtracted. Thus, \underline{I}_1 is reversed by drawing it dotted and pointing in the opposite direction to that of the original phasor. This reversed \underline{I}_1 is now added by phasor addition to \underline{I}_R . The line $\underline{a}b$ is drawn from the tip of \underline{I}_R parallel to reversed \underline{I}_1 . The line $\underline{c}d$ is drawn at the tip of \underline{I}_1 reversed parallel to \underline{I}_R . These lines intersect at \underline{h} . An arrow drawn from \underline{O} to the point \underline{h} will represent the phasor \underline{I}_2 . Thus, \underline{I}_2 will be 11.25 amp and will lead the current \underline{I}_1 by 106 deg.



Fig. A-12. The phasor solution of Ex. A-2 where two a-c voltages, 100 v and 50 v at a phase angle of 45 deg, are added together.

<u>Example.</u> Two a-c generators are connected in series so that their voltages add. One has a voltage of 100 v and the other has a voltage of 50 v. The phase angle between the two voltages is 45 deg. What is the total voltage across the two?

The phasor solution is shown in Fig. A-12. For this diagram, 1/4 in. is chosen as equivalent to 10 v. The phasor for the 100 v input is drawn horizontal 2 1/2 in. in length, while the phasor representing the 50 v input is drawn at an angle 45 deg leading and 1 1/4 in. long. Drawing the two broken lines parallel to the phasors from the tips of the arrows, we find that the sum is a phasor 3 9/16 in. long. This is equivalent to a voltage of 143 v, since each 1/4 in. represents 10 v. An a-c meter placed across the two generators would measure 143 v.

WRH

Index

A

Adjacent signal interference, 40 in mobile service, 48 Alignment, of detector, 381 of discriminator, 375 equipment for, 374 of i-f stages, 374 of ratio detector, 376 of r-f section, 378 of single-stage oscillator detector, 377 using an a-m signal generator, 373 visual, using an oscilloscope, 379 Amplifier, audio, 180-181, 277 class C, 94 in dual-channel transmitter, 167 for signal generator, 359 feedback, 245 grounded grid, 97, 110 grounded-grid equivalent circuit, 99 grounded plate, 100 i-f, 180, 273 alignment of, 374 intermediate power and multiplier, 118 multiplier, 123 power, 112 for 20 kw, 121 r-f, 180, 273 alignment of, 378 stage gain, 238 Amplitude, 394 Amplitude modulation, 2 adjacent signal interference, 41 in Armstrong modulator, 134 band width, 21 requirements, 18 carrier amplitude, 20 effect of, on f-m discrimination, 191 on ratio detector, 193, 198 100 per cent modulation, 3 overmodulation, 4 phasor diagram, 22 of pulses, 10-11 rejection in an f-m receiver, 208 sideband amplitudes, 20 in signal generators, 360 Amplitude-to-phase-to-frequency modulation, 129 Antenna, 180, 286 broadcast, 292 circular, 294

cloverleaf, 298 combined a-m and f-m, 315 directive, 312 director, 312 dummy, 389 feed, 293 feeding problems, 313 folded dipole, 295 hat, 295 height, 289 impedance matching, 314 loop, 293 magnetic dipole, 293 monopole, 310 pylon, 301 reflector, 312 for relay station, 350 series fed, 315 shunt fed. 315 slot, 300 square loop, 296 super turnstile, 304 turnstile, 293, 303 vertical transmitting, 308 Antenna array, 292 linear, 307 power-gain curves, 309 stacked, 305 cloverleaf, 306 Antenna-stage image ratio, 252 Argument, 25 Armstrong, Major E. H., 4, 129 dual-channel modulator, 163 Artificial line, in f-m detector, 205 Attenuator, bridged T, 359 mutual inductance piston type, 361 Audio amplifier, 270 servicing, 384 Audio-circuit requirements, 240 Audio correction network, 132 Audio signal, 14 Automatic - frequency control, 259 Automatic noise limiter, 272, 276 Automatic volume control, 263, 272 on i-f tubes, 266 in ratio detector, 198

В

Baffle, 248

405

WR

Balanced modulator, 104, 152 in Armstrong modulator, 136 in frequency control, 75 Band-pass circuit, 233 using signal-resonant circuits, 234 Band width, 16, 181, 245, 273 of amplifier, 218 for broadcasting, 18 comparison between a-m and f-m, 35 control. 274 for frequency modulation, 26 of an i-f amplifier, 233 maximum, 18 requirements of, for phase modulation, 34 Bass boost, 271 Battery, in ratio detector, 195 Bazooka, 168 Beat frequency, in integrated-pulse control, 79 in motor control, 75 Beers, G. L., 202 Bessel functions, 23 curves, 26 in f-m. 31 table, 25 Blocking oscillator, frequency divider, 87 with stabilizing circuit, 88 Broad band, antenna, 323 dipole, 324

С

Cable assembly, installation, 345 Cable diagram, mobile receiver, 340 mobile transmitter, 342 Capacitance, total shunting in an audio amplifier, 243 Capacitive reactance tube, balanced modulator, 57 employing a capacitor, 52 employing an inductance, 54 modulator, 51 Capacitor, temperature compensated, 221 in audio correction network, 132 Carrier, amplitude in f-m signals, 68 component, 363 definition, 1 frequency-stability requirements, 68, 92 Cascade phase-shift transmitter, 173 Cathode follower, in integrated pulse-control circuit. 83 in phase shifter, 145 in transmission-line modulator, 61 as a variable-resistance tube, 63 Cathode-ray tube, used as a phase shifter, 147 Center-frequency indication, 369 Characteristic curve of an f-m discriminator, 192 Characteristic impedance, 60 Class C amplifier, as multiplier, 96 Coaxial transmission line, for antenna, 327 in grounded plate amplifier, 101 Coefficient of coupling, 234 Colpitts oscillator, 204 Comparator circuit, 69 Cone antenna, 324 Constant-deviation modulation system, 356 Control-head mounting, 344 Converter, 180 double superheterodyne, 224, 227 pentagrid, 222, 226 servicing, 387 systems, 223 tube requirements, 268 Correction network, 152, 158 Coupled circuits, 187

Crossover frequency, 247 Current sheet, 302, 304 Cycle, 394

D

De-emphasis, 46, 369, 385 network, 276 Degeneration, in ratio detector, 200 Degree, electrical, 396 De-icer, 293 Detector, 276 employing an oscillator, 201 employing two tuned circuits, 185 fremodyne, 204 f-m. 180 miscellaneous types, 205 single-tuned circuit, 181 Detector ratio, 193 **Detector servicing**, 386 Detuning, effect on band-pass circuits, 234 Deviation checking, 32 Dielectric constant, 287 Differentiation, in integrated pulse control, 80 Diodes, in ratio detector, 201 Dipole, antenna, 319 folded, 325 Direct f-m transmitter, block diagram, 92 Direct method of frequency modulation, 49 Directive pattern (see Radiation pattern) Discriminator, 187, 253, 276, 366 alignment, 375 current type, 367 in frequency control, 70 servicing with S curves, 390 **Dispatching communication system**, 330 Distortion, 387 in i-f amplifier, 231 in ratio detector, 200 in single resonant-circuit detector, 183 Divider, locked in oscillator, 105, 202 Double-tuned circuits, 236 Doubler, 96 Dual-channel modulator, 161 Dual-channel transmitter, 165 Dual limiter, 214 Dual triode, used in tuning indicator, 258

E

Effective height of a receiving antenna, 320 Effective resistance, 192, 195 Electric field, 286 Electromagnetic wave, 286 reflection and refraction of, 287 Electron disk in phasitron, 139 Equivalent circuit, of an antenna, 319 of mixer network, 165 Exciter, in direct f-m transmitter, 93 input capacitance-phase discriminator type, 113 integrated pulse control, 123 reactance-tube modulator, 105

F

False signals, 332 Federal Communications Commission (FCC), 251, 293, 362 Feedback, 369 in audio amplifier, 245 from electron stream, 202 gain of amplifier with, 246

406

INDEX

in modulator, 125 oscillations caused by, 247 in phasitron modulator, 173 in telephone audio circuits, 347 Fidelity, definition, 15 Field strength, 290 Filter, low pass, 70 Fremodyne, detector, 204 Frequency, 394 effect on phase modulation, 131 Frequency control, motor, 68 Frequency-control system, of maximum excursion, 78 for time intervals, 78 for time-voltage areas, 79 Frequency-control unit for direct f-m transmitter, 93 Frequency deviation, 5 multiplication, 153 in two-channel modulator, 155 Frequency divider circuits, 85 Frequency modulation, adjacent signal interference, 42 advantages, 4 band width, 26, 33 carrier component, 26 becoming zero, 32 of code transmitters, 4 definition, 5 of 100 per cent modulation, 7 factor, 27 frequency deviation, 26-27 overmodulation, 7 from phase modulation, 131 phasor diagram, 33-34 of pulses, 10-12 receiver, 180 sideband frequencies, 28, 31 table of sideband amplitudes, 32 Frequency multipliers in transmitters, 93 Fundamental frequency ranges, 16

G

Gate-type limiter, 212 Graphical addition of phasors, 401 Graphical subtraction of phasors, 402 Grid-circuit limiter, 209 Grid leak, in grid limiter, 211 Ground concept in uM, 103 Ground plane, 286, 306, 311 Ground wave, 288 Grounded-grid amplifier, 220

H

Harmonic allenuator, 112 Harmonic generators, 96 in f-m transmitters, 93 Hartley oscillator, 49, 114, 226 modifiled, 276 used with input capacitance modulator, 59 Heterodyne system for signal generators, 356

I

Image ratio, 251 Improvement constant, 246 Index of refraction, 287 Inductance, of coils in discriminator, 192 as a corrector network, 141 Inductive reactance tube, in balanced modulator, 57

using a capacitor, 55 using an inductance, 56 Inductively coupled 1-m generator, 65 Inductive capacitance modulator, 58 Input impedance of a transmission line, 61 Installation, mobile, 339 Integrated pulse control, of frequency, 77 transmitter, 123 system used in monitor, 366 Interference, adjacent signal, 40 in mobile installation, 345 radio sources, 37 Intermediate frequency, 181 choice of, 231 output voltage, 263 I-f amplifier, design, 237 gain, 266 gain requirements, 233 requirements, 230 servicing, 386 I-f transformer construction, 239 Inverse frequency network, 173 Ionosphere, 288

L

Layout, for mobile installation, 341-344 Lead, in phasors, 397 Limiter, 181, 191, 194, 253, 263, 276, 337 grid voltage, as a tuning indicator, 253 in i-f stage, 266 Durpose, 208 Load-matching inductance in grounded plate amplifier, 103 Locked-in oscillator, 203 Locs resistance, 306 Loudspeaker, cabinet requirements, 248 divider networks, 247 low-pass filter used in phase discriminator, 73

М

m₁, 27 <u>m</u>²_p, 34 Magnetic field, 286 in phasitron, 140 Matching antennas, 323 Matching impedances in a ratio detector, 201 Miller effect, 114 Mistuning in an f-m receiver, 252 Mixer, 273 in integrated pulse control, 80 Mobile communication systems, 330 Mobile f-m transmitter fidelity, 92 Modulating wave, 1 Modulation, from adjacent signal, 42 calibration, 355 definition, 1 effect on carrier in f-m, 69 factor, 131 purpose, 1 (See also specific types) Modulator, Armstrong phase, 133 original Armstrong type, 151 capacitive reactance tube, 52 in direct f-m transmitter, 93 dual channel, 161 phasitron, 137, 172 serrasoid, 149 for signal generator, 356 two-channel type, 155 Monitor, requirements, 362

INDEX

station, 363 Monitoring audio quality, 367 Motor, two phase, for frequency control, 108 Motor control, 105 of carrier frequency, 73 in transmitter, 104 Mounting, vertical type, advantages of, 111 Multiplication table, 24 Multiplicers, 98 in transmitter, 111 in two-channel modulator, 160 Multivibrator frequency divider, 85 in integrated pulse control, 80

N

Negative feedback, in r-f amplifier, 219 Neutralization, in class C amplifier, 95 of grounded grid circuit, 99 of grounded plate amplifier, 102 Noise, characteristics compared between a-m and f-m, 47 from a carbon resistor, 38 in converter and r-f amplifier, 217 impulse, 37, 39 impulse-reduction factor, 45 level, 245 man made, 38 in mobile services, 48 pickup, 40 radio sources, 37 random, 37 random reduction factor, 46 reduction in frequency modulation, 44 tube, 38

0

One-way mobile system, 330 Oscillator, 337, 359 beat frequency, 276 circuits, 222 detector alignment, 377 frequency generated by, 49 frequency temperature stability, 268 frequency variation, 49 harmonic operation, 269 requirements, 221 servicing, 387 Oscilloscope for transmitter testing, 107

P

Pass band for i-f amplifier, 230 Peak amplitude, 394 Pentode tube in r-f amplifier, 218 Permeability, 287 Permeability tuning, 269 Phase deviation, of phasitron, 143 of serrasoid, 149 Phase difference, 397 Phase discriminator, in an exciter unit, 114 used for frequency control, 71 Phase distortion, 266 in sound waves, 15 Phase factor, 34 Phase modulation, 7 band-width requirements, 34 compared with frequency modulation, 8, 129 definition, 7 instantaneous phase shift, 9 100 per cent modulation, 9 overmodulation, 9 to produce f-m, 131

by resistance variation, 142 Phase modulator, 159 cathode-ray tube type, 147 Phase shift through transformer coupling, 105 Phase shifter, in Armstrong modulator, 153 bridge, 144 constant impedance, 146 in integrated pulse control method, 80 using series R-C circuit, 144 Phasitron, 137 Phasor, 394 addition and subtraction, 399 Phasor diagram, 397 for an a-m wave, 22 for Armstrong modulator, 134 for coupled circuits, 187 for discriminator, 189 for ratio detector, 197 for a three-phase generator, 171 used to analyze phase discriminator operation, 72 Pitch control, 276 Placement of parts in an f-m tuner, 261 Plane wave, 286 Plug-in units in exciter, 127 Polarization, 292 of antenna, 180 of waves, 287 Power amplifier in direct f-m transmitter, 94 Power signal-to-noise ratio, 39 Power supply, dynamotor, 338 servicing, 384 vibrator, 334, 338 Pre-emphasis, 114, 123 Propagation of electromagnetic waves, 286 in f-m bands, 289 Pulse, amplitude, 10 control of frequency, 77 counter detector, 205 counter divider, 89 counters, 82 counting circuits in integrated pulse control, 79 generator, 82 modulation, 9 peak voltage, 10 position modulation, 11 time modulation, 10-11

multiplex transmission, 11

Q

Q, of coils in discriminator, 192 as a resonant circuit curve parameter, 182 of tuned circuit in an amplifier, 218 Quadrature circuit, 203

R

Radiation pattern, of an antenna, 321 for reflector and director, 313 Radiation resistance, 296, 302 R-f frequency amplifier, reasons for, 217 circuits, 219 selectivity, 218 servicing, 387 R-f sidebands, 17 R-f tube choice, 269 Range prediction, 291 Ratio detector, 193 alignment, 376 full wave, 200 receiver, 277 Reactance, as an f-m detector, 206

INDEX

simulated, 52 Reactance tube, 50 circuit of constant-deviation type, 358 modulator balanced, 56 modulator-exciter using motor control, 105 modulator in signal generator, 356 transmitter using motor control, 103 Receiver, all purpose, 272 combined a-m and f-m, 263 f-m, 180 f-m compared to a-m, 373 mobile, 332 for relay station, 351 requirements for f-m, 251 Receiving antenna, 323 fundamental characteristics, 319 installation, 326 requirements for f-m, 322 Reciprocity, 322 Reflection of electromagnetic waves, 287 Refraction of electromagnetic waves, 288 Relay, converter, 353 Relay equipment, 348 Relay transmitter, 350 Resistance-capacitance coupled amplifier, 240 equivalent circuits, 241 gain, 242 Resistance-capacitance coupled oscillator, 62 Resonant circuit, as an f-m detector, 181 high Q, 218 response curve, 182 Ring modulator, frequency divider, 88

S

S curves, 389 Saturation limiter, 208 Scale in phasor diagram, 401 Selective call, 332, 338, 347 Selectivity, characteristic, 266 r-f, 251 Send-receive switch, 277 Sensitivity, 266 figure, 251 Separation between peaks in an f-m discriminator, 192 Serrasold modulator, 149 Servicing, 373 audio amplifier, 384 detector and i-f stages, 386 preliminary steps, 381 r-f stages, oscillator and first detector, 387 with S curves, 389 Sideband frequencies, 17 in amplitude modulation, 18 in Armstrong modulator, 133 in frequency modulation, 69 Sideband generator, 152 Signal, 1 audio, 1 Signal generator, a-m used in alignment, 374 requirements, 355 Signal-to-noise ratio, 39, 217, 251, 290 measurement, 40 Signal voltage at receiver, 39 Sine wave, 394 Single resonant circuits used in band pass, 235 Sky wave, 288 Sleeve on antenna, 311-312 Slug tuning, 201 Sound, 14 Sound-wave components, 14 Speaker, coaxial, 247

Spectrum analyzer, 369 Spurious response from converter, 218 Spurious signal rejection, 217 Squelch circuit, 333, 337 Stable gain, 266 Stability, of an oscillator, 222 in signal generator, 355 Stabilizing circuit in blocking oscillator, 88 Static, 38 Static, 38 Staticn-wagon installation, 345 Supersked carrier modulator, 136 Switch, used on tuning indicator, 254 Synchronizing voltage, in blocking oscillator, 88 in multivibrator, 86

т

Tank circuit, in class C amplifier, 95 Telephone service, 346 Three-phase voltage generator, 171 Three-way mobile system, 331 Threshold point, 214 Time constant for grid limiter, 211 Tone control, 277 Tracking, in an oscillator, 221 Transformer, mechanical design, 266 Transmission lines, 313 matching, 315 to antenna, 320 modulator, 60 Transmitter, installation, 174 mobile, 334 relay, 350 (See also specific types) Tube, GL-2H21 phasitron, 141 7C24 for grounded-grid applications, 111 special heptode for detector, 202 Tuner, 260-262 Tuning characteristic of ratio-detector receiver, 278 Tuning eye, 253 **Tuning indicators**, 252 contactor type, 255 meter, 277 Tweeter, 247 **Two-channel multiplication**, 154 Two-channel transmitter, 158 Two-way mobile system, 331

U

U antenna, 326 Ultrahigh frequency, 180 Under-the-rug antenna, 328

۷

V antenna, 326 Vacuum-tube voltmeter used in alignment, 374 Variable capacitor, motor driven, 356 tuning, 269 Vertical construction, 171 Vertically polarized radiation, nullified, 300 Voltage-gain stability, 266 Voltmeter, used as tuning indicator, 253

W

Weather, effect on antenna, 293 Window receiving antenna, 328 Woofer, 247

409

WRH

WRH