

The OSCILLATOR at WORK



By

JOHN F. RIDER

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*Perpetual Trouble Shooter's Manuals, The Cathode Ray
Tube at Work, Servicing Superheterodynes and
other books for the Radio Service Industry*

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FLORENCE
who is still patient

AUTHOR'S FOREWORD

IN READING through the pages of this book you will note a certain form of presentation, and we feel that a few words concerning the contents of this book are quite in order.

This book is not intended as a theoretical text covering the operation of oscillators. It is true that a portion of the book is devoted to what might appear to be a theoretical discussion, but every effort has deliberately been made to omit equations, mathematical references, and graphs of various types, relating to theory. The reason for this is that a large number of very excellent theoretical texts are available, as indicated by the bibliography; hence any material of this type would be unnecessary duplication.

The amount of theory we present is that which we think is required so as to lead up to the chapters devoted to the testing of oscillators utilized in the radio servicing field, and to the application of signal sources employed by men who are associated with the maintenance of radio receiving equipment. Many tens of thousands of signal sources identified as test oscillators or signal generators are in use by radio servicemen and practical texts devoted to the maintenance and to the application of such units have been absent. We hope that the material contained in this book is of such character as to enable the serviceman to realize the fullest return from the investment he made in the equipment intended to furnish the test signals required during his work.

Inasmuch as this book is aimed at one particular branch of the radio industry, and since the nature of the work carried on by the personnel in this field is of a certain specific kind, we have attempted to cover the utility of such equipment to conform with present needs, and have also tried to project the needs into the

near future. However, as you will readily see after perusal of the text, we have omitted—and this omission is deliberate—references to oscillators operating at extremely high frequencies, frequencies which today primarily represent research in certain fields. We are referring to electron-oscillators of the Barkhausen-Kurz and similar types which generate signals at frequencies above 300 megacycles.

In our estimation the most valuable portions of this book are those chapters which are devoted the manner in which a serviceman can test his signal source, how he can identify and remedy defects in it, and to those facts which relate to the application of the device. In fact, we feel it would even have been possible to omit everything else in this book, and still make it of practical value. However, in view of the varieties of oscillating systems and circuit arrangements which are to be found in such signal sources, and inasmuch as the design of such equipment is continually advancing, we felt that theory in explanatory form and a discussion of the various types of signal sources—some of which have not as yet reached their full-fledged growth in this field, but without doubt will reach such growth—is justified. That is why this book contains more than just the practical facts concerning oscillators and signal generators.

This belief is not a refutation of ideas which we have advanced at different times concerning the need for more theoretical knowledge on the part of the serviceman; we still believe that a man can get the most out of his equipment only if he understands how it operates. However, as far as that is concerned, we also believe that it is possible for the individual to comprehend the operation from a theoretical viewpoint without the need of mathematical references.

We wish to express grateful acknowledgment for the cooperation from the various manufacturers whose circuit diagrams are contained herein, and to G. C. B. Rowe, J. Avins, J. H. Potts, and R. Lorenzen, for their cooperation in the preparation of this book.

JOHN F. RIDER

January 18, 1940

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Chapter I

OSCILLATION

IF WE are to ask what an *electric oscillation* is, we must also ask what an *alternating current* is and what an *electric wave* is, because all three are one and the same thing and the terms are therefore interchangeable. We may refer to oscillations at one time, alternating currents at another time, and waves at still other times, depending upon which of the three terms is most convenient to or descriptive of the subject under discussion.

Irrespective of which term is applied, we are in all cases dealing with a definite fundamental, and that is the behavior of an electric current. Consequently, our first concern should be to determine the nature of an alternating current in its most intimate aspects; that is, from the viewpoint of electron motion.

Electron Motion

As in the case of direct current, alternating current is the electron in motion. However, in contrast to the type of electron motion exhibited in the case of direct current, the electron motion exhibited in the case of alternating current is substantially different.

In a direct-current circuit there is a steady and constant drift of electrons in one direction around the circuit. This drift is due to the application of a voltage which likewise is constant in value and polarity.¹

In alternating-current circuits we find a somewhat different situation. Alternating current, identified as a.c., does not at all times flow in the same direction. Instead, it reverses its direction of flow periodically and the reason why this happens is that the

voltage which causes the current to flow alternately reverses its polarity.

There is a widespread belief that in an a-c circuit electrons first drift in one direction *around the circuit*, in a manner similar to that found in d-c circuits, then reverse themselves and drift back in the opposite direction *around the circuit*. It is true that where an alternating voltage is present, the electrons in the circuit move first in one direction and then in the opposite direction, but this is a *to-and-fro* motion or oscillation of the electrons and occurs about a central position but over a *very limited distance*.¹

Alternating Current

If we now interpret the motion of the electron into current, the same designations apply. . . . In other words, alternating current is identified with reference to both frequency and cycles. . . . An alternating current representative of electrons which complete one million cycles per second is an alternating current of one million cycles or one thousand kilocycles or one megacycle. . . . An alternating current which is representative of electrons in motion which complete sixty cycles in a second, has a frequency of sixty cycles, etc.

A graphic representation of alternating current is illustrated in Fig. 1-1, and it is seen that it assumes the shape of a series of *waves*. The upward direction along the vertical axis or the portion above the horizontal (zero) reference line, represents current flow in one direction along the wire and the downward direction along the vertical axis—or the portion below the horizontal (zero) reference line—represents the current flow in the opposite direction. . . . By means of these two axes—the horizontal for time and vertical for amplitude and direction—it is possible to represent the magnitude and direction of an alternating current at any instant of time. . . . In all such representation of alternating current—the magnitude or the value of the current is shown with respect to time. . . .

Referring to Fig. 1-1, you will note that we begin plotting the current at the zero value and moving in the upward direction. . . .

¹ This subject is discussed in greater detail in the "Hour a Day" series under the title "Alternating Currents in Radio Receivers."

The current now starts flowing and in accordance with the time designations in Fig. 1-1, in $1/240$ of a second it has reached its maximum positive value. . . . In the next $1/240$ of a second the current again drops to zero. . . . This completes a half cycle or an *alternation*. . . .

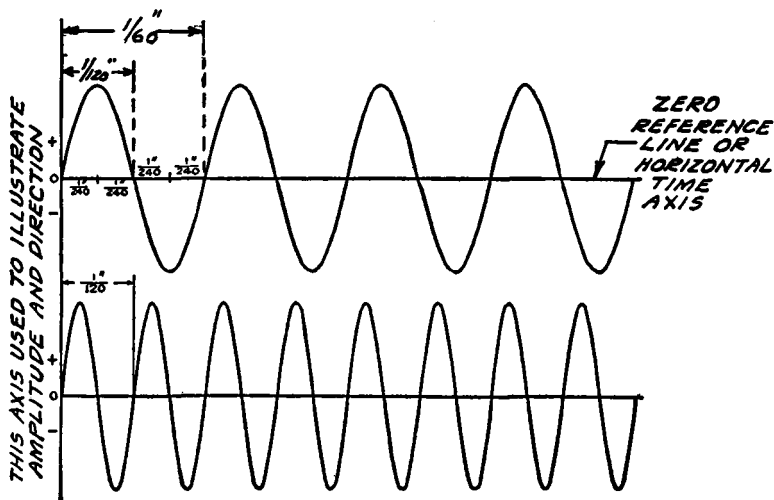


FIG. 1-1, above, FIG. 1-2, below. When a complete cycle of current occurs in $1/60$ second, Fig. 1-1, its frequency is 60 cycles. If a cycle is completed in one-half the time, Fig. 1-2, its frequency is 120 cycles.

The current now reverses its direction of flow as indicated by being shown below the horizontal axis. . . . In $1/240$ of a second the current reaches its maximum negative value which is in the opposite direction. . . . In the next $1/240$ of a second the current again decreases to zero. This completes the second alternation and the entire cycle.

Thus you note that a cycle of current is completed when the current goes through all of its variations of *magnitude* and *direction*. . . . In the example cited the current cycle is completed in $4/240$ of a second or $1/60$ of a second. . . . It therefore stands to reason that in one second, 60 such complete cycles of *current variation* and direction occur and the current is identified as hav-

ing a *frequency of 60 cycles per second*. The illustration in Fig. 1-1 shows four cycles of such a 60-cycle current.

If we continue with another example of either alternating current or voltage, which has a frequency of, say, 120 cycles per second, then this wave will complete one cycle in $1/120$ of a second. . . . Consequently, a wave of this frequency would be represented as shown in Fig. 1-2. If you compare Figs. 1-1 and 1-2, you will find that the shape and character of the two waves are identical, but the one shown in Fig. 1-2 has *twice* the frequency of the wave shown in Fig. 1-1. It completes the cycle of amplitude variation and direction in half of the time required for the current illustrated in Fig. 1-1.

Cycle and Frequency

When referring to alternating currents or voltages, the frequency or periodicity of the cyclic motion is referred to with respect to the number of whole cycles completed in a second. If the voltage shown in Fig. 1-1 completes a cycle of amplitude, polarity and motion respectively, in $1/60$ of a second, 60 complete cycles would take place in one second and the frequency would then be *60 cycles per second*. If the cycle is completed in $1/120$ of a second so that 120 complete cycles of voltage and electron motion occur in a second, then the frequency of the voltage and electron motion is 120 cycles per second. If a million-cycle voltage is applied, the cycle of electron motion is completed in one one-millionth of a second.

While speaking of frequency expressed in cycles per second, use is generally made of prefixes which denote a multiplying factor. The table which follows illustrates the use of these prefixes and the conversion from one quantity to another:

$$\begin{aligned} 1000 \text{ cycles} &= 1 \text{ kilocycle} \\ 1,000,000 \text{ cycles} &= 1,000 \text{ kilocycles (kc)} \\ 1,000 \text{ kilocycles} &= 1 \text{ megacycle (mc)} \end{aligned}$$

Thus a frequency of 1200 cycles can be expressed in the whole quantity or as 1.2 kilocycles per second. A frequency of 1,750,000 cycles can be expressed as 1750 kilocycles or 1.75 megacycles.

A frequency of 20,250,000 cycles can be expressed as 20,250 kc or as 20.25 megacycles.

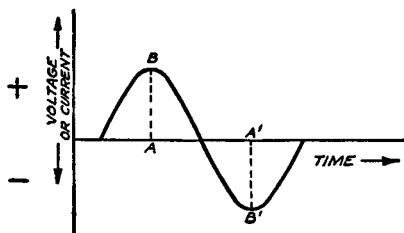
Start of Cycle

By this time it is possible that you may have formed the opinion that a cycle of voltage or current can be represented only by showing the voltage or current starting from zero and ending at zero. . . . Such is not the case. . . . It is possible to start at any point along the cycle, which means at any point along the positive half—or at any point along the negative half. . . .

Peak and Effective Values

Although alternating current varies in magnitude throughout the cycle, there are values which are used as a measure of the current strength. One of these is the so-called *instantaneous peak value* or *maximum value of the current* and is denoted on Fig. 1-3 by the line AB . AB indicates the *positive* peak

FIG. 1-3 The line AB denotes the positive peak value of current or voltage of a wave and $A'B'$ denotes the negative peak value.



value while $A'B'$ represents the *negative* peak value. Under ordinary conditions these two peak values are equal, but as we shall see later on, special cases occur when the positive peak value is not equal to the negative peak value.

Continuing with the discussion of the peak value, it is evident that it is an instantaneous value which under normal conditions occurs but once in each half cycle, or twice during a complete cycle. We know that other values of current exist in the circuit during the other instants, consequently, the instantaneous peak

or maximum value cannot be used as the value representative of all of the instants during which current flows in the circuit.

If we are to use current and voltage, it is necessary to establish a means of comparing values of current and voltage which is based upon *work done*—and whatever this basis of comparison, it then becomes one which is practical, because it embraces all of the instants during which the current varies between zero and maximum. . . . This practical value is known as the *effective* or *rms value*. . . .

The *effective* value is arrived at by determining the amount of heat produced in a resistor by a direct current and an alternating current. It is considered on the basis that *the effective value of an alternating current is one ampere when it generates heat at the same rate as one ampere of direct current flowing through a resistor of given value*.

It is evident from the above that the effective value must embrace all of the instantaneous values and is, in a sense, a composite value. It turns out that the effective value of the current (or voltage) is equal to the square root of the mean value of the instantaneous currents squared, and consequently is referred to as the *root-mean-square* or *rms* value. Stated mathematically, the rms value of an alternating current (or voltage) is equal to .707 times (or about 70% of) the peak value.

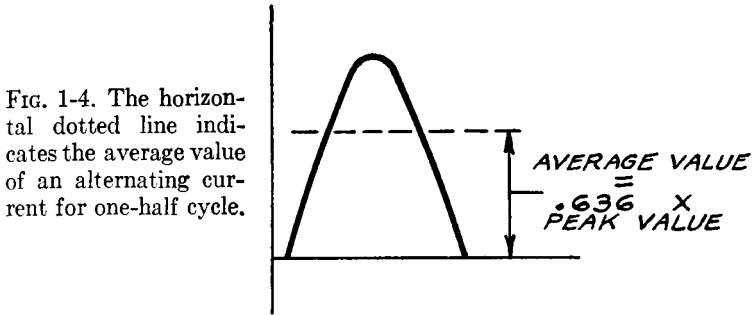
Average Value

There is another composite value of current known as the *average* value. While this is not used to as great an extent as the effective value, it should be included in this text.

The average value of an alternating current over a complete cycle is zero because the current flows just as long in one direction as it does in the other. In other words, as far as average value is concerned, the positive and negative alternations cancel each other. However, the average value *over a half cycle*, as you can see by reference to Fig. 1-4, is certainly quite different from zero. Mathematically, the average value of the current is .636 times the peak value of the current.

Alternating Voltage

A number of facts have been stated concerning alternating current. These same facts apply in practically every case to alternating voltage. In accordance with what has already been said, we can state simply that a cycle of alternating voltage is completed when the voltage goes through all its variations in



value as well as in polarity. Hence, the significance of the cycle as stated for alternating current applies equally well to alternating voltage.

The equations which specify the relation between the peak, effective and average value likewise take on the same form for voltage as they did for current. Thus we have

$$\begin{aligned}
 E_{eff} &= E_m \times .707 \\
 E_m &= E_{eff} \times 1.414 \\
 E_{av} &= E_m \times .636
 \end{aligned}$$

The Sine Wave

The sine wave represents the basis of comparison of all other types of waves. It is a wave of *single frequency*, which in sound is representative of a pure note having no overtones.

One of the characteristics of a sine wave is that the voltage or current wave has identical positive and negative alternations. But the most important characteristic of the sine wave is that the rate of amplitude change of the current or voltage is a mathematical progression based on what is known as the "sine of the angle."

Phase Relation

It is a peculiarity of alternating-current circuits that the current and voltage present in the same circuit may not rise and fall in amplitude and alter polarity in exact unison. In other words, the zero points and peaks do not occur together. When that is the case, the voltage and current are said to be *out of phase*. When the zero values of the voltage and current waves occur together, the voltage and current are said to be *in phase* with each other.

The “in-phase” condition, where the phase angle of the current and voltage waves is the same, is shown in Fig. 1-5. The

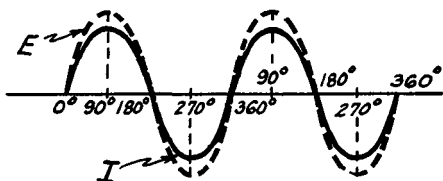


FIG. 1-5. When the zero and peak values of a voltage wave and a current wave occur at the same instant, the two waves are in phase.

current cycle is represented by I and the voltage cycle by E , and though the peak amplitude of I is less than the peak amplitude of E —which is usually the case in oscillatory circuits—it can be seen that the zero and peak values of the two occur simultaneously in both the positive and negative alternations. You will note that the time axis is now divided in terms of degrees instead of fractions of a second. This follows in line with the general interpretation of the structure of a cycle of voltage or current generated by a loop of wire rotating within a steady magnetic field. Since this loop of wire rotates through 360° in order to make a complete revolution and complete one cycle of voltage or current, it is possible to express fractions of the cycle in terms of degrees. Thus a quarter of a cycle represents 90° , a half cycle 180° , three quarters of a cycle, 270° , and a complete cycle 360° . As you can see in Fig. 1-5, the point of 360° for one cycle is the start of the next cycle, and consequently is the 0° position of the second cycle. If current and voltage do not coincide in their positive and negative points, the difference in time between these points is usually expressed in degrees and is known as the *phase angle*.

An "out-of-phase" condition, where the phase angles of the current and voltage do not coincide, is shown in Fig. 1-6. In this case, the current "leads" the voltage (or the voltage "lags" the current) since its zero value or point of origin, relatively speaking, came before that of the voltage. To put it another way, the current starts to flow before the voltage starts to rise for any given cycle. In this particular case the current and voltage are 90° out of phase, as indicated in the drawing.

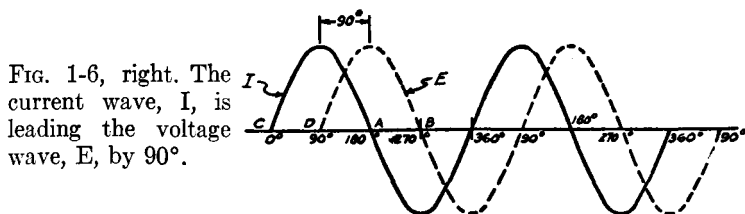


FIG. 1-6, right. The current wave, I , is leading the voltage wave, E , by 90° .

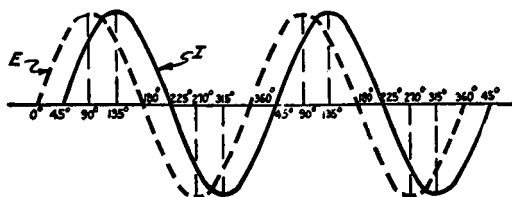


FIG. 1-7, left. The voltage wave, E , is leading the current, I . The phase difference here is 45° .

Another out-of-phase condition is shown in Fig. 1-7. Here the current lags the voltage, (or the voltage leads the current) and for the sake of variety the phase difference is shown as 45° , thought it might be more or less, depending upon circuit characteristics.

The degree of phase difference is naturally dependent upon the relationship between the values of capacity and inductance in the circuit, since one tends to counteract the effect of the other on an alternating current. For instance, if capacity predominates—the current leads the voltage. If inductance predominates, the voltage leads the current. If the circuit is purely resistive, the current and voltage are in phase. On the other hand, if the circuit is pure capacity or pure inductance, the voltage and current will be 90° out of phase. If two waves are 180° out of phase, their zero and maximum values occur at the same instant, but

the directions of change and polarity respectively are opposite. This is shown in Fig. 1-8.

Since phase relationship is purely an expression of time in which we determine the value or polarity of an alternating current and voltage at any given instant or series of instants represented by degrees of an angle, it becomes evident that two or more alternating voltages (or alternating currents) in the same circuit may also have different phase angles.

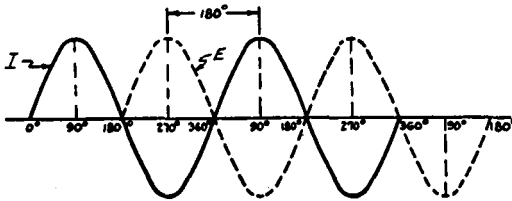


FIG. 1-8. The phase difference between the voltage and current waves here is 180° .

Sources of Alternating Voltage

We mentioned the loop of wire rotating in a constant magnetic field as a source of alternating voltage. However, it is not the only means of generating such a voltage; in fact, it is not discussed any more in this book. Instead we concern ourselves with such generators as vibrating crystals and the vacuum tube—especially the latter.

Chapter II

COMPLEX WAVES

WHILE the sine wave is the basic electrical wave of alternating character, it is necessary for you to understand that there are other classifications of electrical waves which appear in radio nomenclature and which are of interest to us, because they come within the category of alternating-current waves. One in particular is known as the "complex" wave—in contrast to the "sine" wave.

The distinction between a sine wave and a complex wave is two fold: first, the sine wave has but a single frequency, whereas the complex wave consists of a number of different frequencies; second, the variation in amplitude of a sine wave follows a certain definite law as stated in the preceding chapter, whereas the variation in amplitude of a complex wave depends upon a number of different factors: the number of component frequencies, the relative amplitude of these frequencies, and the phase relation between the different voltages or currents which compose the wave.

Any wave which departs in character from a sine wave is a complex wave. The greater the number of frequencies present in the wave, the more complex becomes its nature. The same is true of variations in amplitude and phase relation. Complex waves, in which the amplitude of the odd harmonics is particularly great, assume the form of *square waves*. Close approximations of square waves can be created by simple means shown elsewhere in this book. Another example of an extremely complex wave is one which appears triangular. The square and triangular

waves are shown in Figs. 2-1 and 2-2. Examples of simpler complex waves are shown in Figs. 2-3 to 2-6 inclusive.

Origin of Complex Waves

Space does not permit a full description of complex waves; fortunately such a description is not required for the proper application of an oscillator. However, in view of the fact that complex waves appear as the output of many oscillators, it is necessary to devote some time to a discussion of the subject.

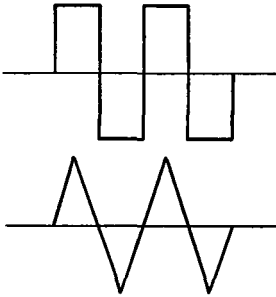


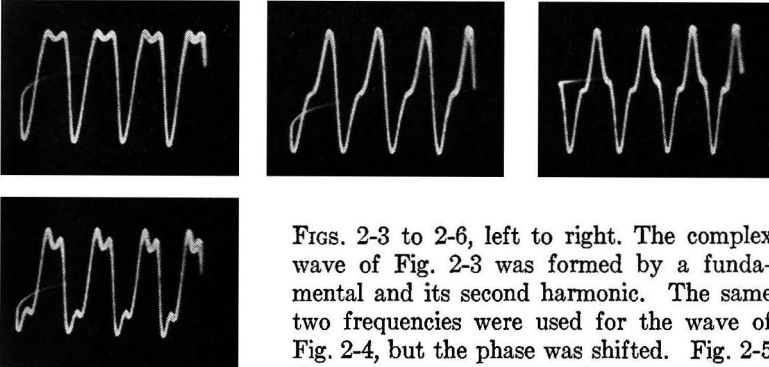
FIG. 2-1, above, FIG. 2-2, below. Complex waves may appear as square or triangular and are composed of a fundamental frequency and a number of its harmonics. Compare with Fig. 11-11.

To start with, it is possible to say that taking oscillators at large the output is invariably a complex wave, but this statement must be qualified to explain that there is such a thing as a complex wave which is accepted as being a sine wave. By this we mean it has become accepted engineering practice to view a complex wave which has one major frequency component of the greatest magnitude and other frequency components of very low magnitude—those, for example, having an amplitude equal to a few percent. of the amplitude of the major component—as a sine wave, because if this wave were viewed on an oscillograph screen, it would look like a sine wave and because the other components are of such small magnitudes. In contrast to those oscillators with an output similar to that described, we find the majority, whether in the r-f, i-f, a-f, or in the ultra-high frequency bands, are productive of output waves which are more complex than those described, in that while there is one major frequency, the other frequencies also present are of a substantial

order. With respect to the relationship of these components one significant detail must be remembered: these components are all multiples in frequency of the major component. In other words, they are harmonics. This subject is described elsewhere in this book.

With respect to types of oscillators with outputs as described above, what has been said is true of virtually every type within the province of the radio industry—at least every type described in this book.

As to reasons for the presence of such a complex output, a number of conditions contribute. In the case of vacuum-tube oscillators the departure from perfect linearity in the vacuum tube and in the operation of other components found in the system, results in what we can call such a distorted output. As to the



FIGS. 2-3 to 2-6, left to right. The complex wave of Fig. 2-3 was formed by a fundamental and its second harmonic. The same two frequencies were used for the wave of Fig. 2-4, but the phase was shifted. Fig. 2-5 shows a complex wave made up of a fundamental and its third harmonic and that of Fig. 2-6, the same two frequencies when out of phase.

other types of oscillators, it is not necessary to discuss reasons why the output contains more than one frequency. Of far greater interest and importance is realization of the fact that such a distorted output exists and the ways and means of securing from this complex output the desired frequency component. Usually this method is the use of a tuned circuit which would respond essentially to its resonant frequency and reject the harmonic frequencies.

Harmonics

Whenever the wave form of an alternating current or voltage differs from a sine wave, then that waveform is complex and contains a fundamental frequency and other related frequencies. Note that this statement stipulates that the frequencies present in a complex wave are related to each other.

As to the definition of a harmonic, *it is a component of a periodic quantity which is an integral multiple of the fundamental frequency.* Let us break this down into simpler sentences. Since the sine wave of any frequency is a periodic quantity as previously stated—and since the complex wave consists of a number of such sine waves, *the complex wave becomes the periodic quantity.* . . . By component is meant a part of the whole, so that any one of the frequencies present in the complex wave is a component frequency—and *any one of the sine waves of voltage or current, which contributes to the composition of the complex wave, would be a component wave.* . . . Now, the harmonic is a component of the complex wave, and *its frequency is an integral multiple of the fundamental frequency.*

In accordance with the above, if the fundamental frequency of a complex wave of voltage or current is 1000 cycles, the harmonic frequencies of this fundamental are 1000×2 , 1000×3 , 1000×4 , 1000×5 , etc.—up to whatever order of harmonic is being considered. Thus, it is possible that a complex wave may consist of a 1000-kc fundamental and harmonics up to the 50th . . .

If we select “f” as the fundamental frequency, a short tabulation of the relation between the fundamental and the harmonics would be:

$$\text{Fundamental frequency} = f$$

$$\text{Second harmonic} = 2f$$

$$\text{Third harmonic} = 3f$$

$$\text{Fourth harmonic} = 4f$$

$$\text{Nth harmonic} = nf$$

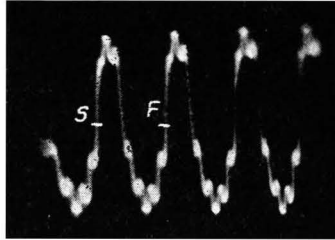
In connection with harmonic frequencies, they may or may not be desired—depending entirely upon the conditions required. In certain instances, the source of voltage generates the funda-

mental as well as the harmonic frequencies and the latter are deliberately removed—whereas in other instances, the unit responsible for the complex wave is deliberately arranged to produce a complex wave very rich in harmonics. As far as harmonics are concerned—they can be produced for all fundamental frequencies—regardless of the numerical value of the fundamental.

The Fundamental Frequency

Speaking about the fundamental frequency of a complex wave, it is, to start with, the lowest frequency—there being no subharmonics below the fundamental. . . . Expressed in another manner, it is the number of times per second that a complete cycle of the complex wave repeats itself. This is true no matter how complex the wave or the number of harmonics contained in the wave. For example, in Fig. 2-7 is shown a complex wave containing a number of harmonics. Four cycles are shown. If

FIG. 2-7. The start and finish of a cycle of this complex wave, whose fundamental frequency is 5000 cycles, are indicated on the oscillogram by S and F respectively. This means that one complete cycle occurs in $1/5000$ second.



the duration of one complete cycle is $1/5000$ th of a second, the fundamental frequency is 5000 cycles.

As to what constitutes a cycle of a complex wave—the definition given for a cycle of a sine wave applies, namely—that period during which the voltage or current passes through all of its variations in amplitude and polarity. This is shown in Fig. 2-7. If you examine the portion designated as being a cycle, you will find that the voltage has passed through all of its variations in direction and magnitude. As a matter of convenience, it is best to employ the zero voltage points as the reference points, representing the limits of the cycle.

From a practical viewpoint, i.e. the application of an oscillator, the exact nature of a complex wave is of very little significance unless during the construction of an oscillator which is intended to produce a substantially sine-wave voltage, we find such a complex wave output. In such event the important detail is the application of some change so as to produce the desired output signal voltage. The relative amplitude of the component frequencies are likewise of little significance. Again the important detail is to make such changes that will reduce these amplitudes. As to the phase relationships existing in such a distorted output voltage, that too is of little importance other than its effect upon the shape of the wave.

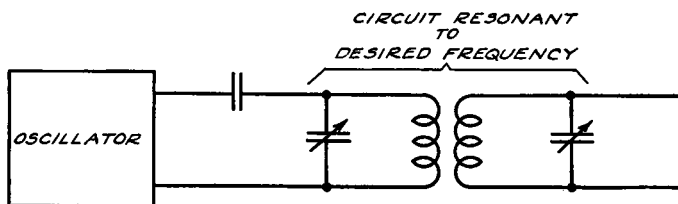


FIG. 2-8. When the output of an oscillator is a complex wave, a single frequency can be selected by means of a tuned circuit as indicated above.

In view of the fact that under some conditions oscillators are deliberately designed to produce a complex wave rich in frequencies, our present interest lies in a simple means of securing or selecting the desired frequency from this group of frequencies. This is done by means of a tuned circuit arranged as in Fig. 2-8 with further elaboration to follow later on in this book. As to the reasons why and the conditions contributing to the generation of such a complex wave and remedies applicable to an oscillator, these are also discussed later on.

Chapter III

HOW AN OSCILLATOR WORKS

Now that we have an understanding of the nature of an electrical oscillation and its characteristic waveform under various conditions, let us turn to the manner in which oscillations are generated.

There are, naturally, a number of different devices capable of producing oscillations or alternating currents. Some of these devices, such as the power plant alternating-current generator, or certain types of vibrating units, are electro-mechanical in operation. A few of these will be dealt with in subsequent chapters, as they are important by reason of the character of the oscillations they develop, and also because they find use in radio work.

However, we are principally interested, in this volume, in the purely electrical types of alternating-current generators, and of these the vacuum-tube oscillator is by far the most important and the most commonly used in radio practice. It has a flexibility of application known to no other type of oscillator, and is capable of producing alternating currents of practically any desired frequency or waveform, which is not true of other oscillators. Moreover, it has an inherent simplicity of electrical structure which permits the generation of substantially pure or complex waves of any order by—in most cases—the mere selection of the proper circuit, values of voltage and values of circuit components.

The Vacuum-Tube Oscillator

A tube can be spoken of as being in an oscillatory state when it is converting d-c power in the plate circuit into a-c power

available from the output circuit. This definition naturally assumes that there is no a-c input of any nature to the tube. In other words, we are interested in drawing a line of demarcation between a power amplifier tube and an oscillator. Actually, the oscillator tube is functioning as a power *converter*.

The ability of a tube to function as an oscillator is predicated upon its ability to amplify. By virtue of this ability, the initial output is greater than the input and the tube can be arranged to supply its own input signal. When this state exists the tube will generate sustained oscillations at a pre-determined frequency, depending upon the constants of the circuit. What we normally term "feedback" in a circuit is the equivalent of feeding a part of the output back to the input. When this feedback, which may be accomplished in as many ways as it is possible to link the output circuit with the input circuit, occurs in a certain manner and when it reaches a certain state, the generation of oscillations will start. When the feedback is decreased beyond a certain level or when other conflicting conditions arise, the generation of oscillations will cease.

With respect to the feedback action, certain requirements exist. The first is that which relates to the phase of the feedback voltage. Every man who at any time has constructed a simple tickler feedback oscillator or built a plate tickler type regenerative detector has experienced the condition that proper regeneration depended on the connections to the tickler coil and the direction of the turns in the tickler winding with respect to the turns in the grid winding. In other words, it is essential that the alternating voltage fed back from the output to the input circuit must be of such phase that it will aid in the grid or input circuit.

The normal action of the amplifying vacuum tube is such that the a-c voltage developed in the plate circuit is approximately 180° out of phase with the a-c voltage fed into the grid circuit. Consequently, in order that oscillation generation start, the voltage fed back from the plate circuit must be 180° out of phase with the a-c voltage existing in the plate circuit. Under such conditions the voltage fed back and the a-c voltage at the grid will be in phase.

Assuming that the phase of the voltage fed back from the plate circuit to the grid circuit is correct, it is necessary that another condition be fulfilled in order that oscillations start. This condition is that the amount of feedback be sufficient to neutralize or offset some of the effective resistance of the tuned circuit in the oscillator system. This is true irrespective of how the inductance constituting the respective circuits is resonated, be it by means of the usual variable condenser or by means of the inter-electrode capacity of the tube. Only after the effective resistance has been neutralized or offset to a certain degree will the circuit start to oscillate or the tube start generating oscillations.

The starting of oscillations in a vacuum-tube circuit is dependent upon the occurrence of an instantaneous change in the operating conditions of the grid or input circuit; that is, the presence of an initial impulse. The very first impulse representing some change in operating conditions results in the application of an instantaneous change in the voltage applied to the grid of the tube. This causes an amplified variation in the plate circuit. Since the plate circuit is coupled back to the grid circuit, a portion of the voltage change developed in the plate circuit is fed back to the grid circuit in the form of voltage variation greater than the original which started the operation. This is repeated until the amount of feedback is sufficient to cause sustained oscillations. In properly designed circuits, this state of sustained oscillation generation is arrived at in very short time, usually after the first few cycles.

The feedback action which takes place in oscillators is like the feedback action which takes place in regenerative detectors and even regenerative radio-frequency amplifiers. The presence of feedback voltage in a circuit is the equivalent of feeding into that circuit negative resistance to offset the positive resistance of that circuit. This positive resistance is not the direct-current resistance of the wires alone, but also includes the a-c resistance offered by the various components of the circuit to the flow of alternating current. In effect it represents those factors which combine to create a loss of power in the circuit or the absorption of power.

The greater the losses in the circuit or the lower the efficiency of the circuit, the greater will be the amount of feedback required

to overcome these losses so that the circuit will be able to oscillate. If the losses, for one reason or another, are excessively great, and the maximum amount of feedback available in the system is not sufficient, the tube will not generate sustained oscillations.

Excessive losses of this type need not be represented in the design of the inductances used in the circuit. The inductances may be perfect, but the associated condensers may be extremely poor and the cumulative effect will be excessive resistance. Of course, the high-loss part of the system may be the coil or perhaps some metallic object adjacent to the coil.

Whether or not a circuit will oscillate depends upon the tube used and the operating potentials. Certain tubes will generate oscillations more readily than others and will do so at lower operating potentials. The potentials involved encompass every element within the tube.

Grid bias voltage has a definite effect upon the ease of starting oscillations automatically. If the initial impulse is of such low order that it cannot cause a variation in the plate circuit, because of the condition that the negative bias is so high that the tube is operating below cut-off, oscillations will not start automatically. It is imperative that the grid bias voltage be of such magnitude as to allow satisfactory operation and still allow the instantaneous impulse previously referred to, to start the process of oscillation generation.

How Oscillations Are Generated

The well-known and much used tickler feedback oscillator circuit shown in Fig. 3-1 lends itself particularly well to our discussion. The operation of this oscillator circuit is predicated upon the elementary theory that whenever the rate of flow of a current through a coil varies, the magnetic field about the coil will also change. If the changing magnetic field threads the turns of another coil, a voltage will be induced in the second coil. As the magnetic field expands with an increase in current flow, the polarity of the induced voltage in the second coil will be in one direction, and as the magnetic field contracts with a decrease in current flow, the voltage induced in the second coil will be in

the *opposite* direction. If the flow of current in the first coil is steady and unvarying, the magnetic field about the coil will be static and will induce no voltage in the second coil.

Now, referring to Fig. 3-1, when the cathode is energized, electrons will commence flowing from cathode to plate, thus setting up a rising plate current which will flow through the tickler coil L . Since the plate current continues to rise as the cathode heats up, an expanding magnetic field is formed around coil L which is coupled inductively to the grid coil L_1 in the grid circuit. As a result of this expanding field, a voltage is induced in L_1 and this voltage appears on the grid.

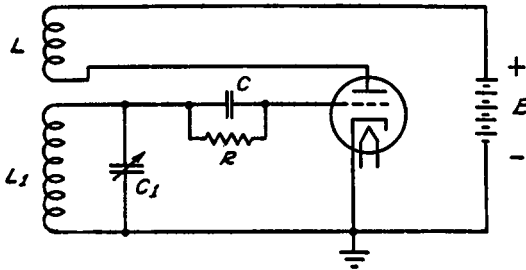


FIG. 3-1. A typical tickler-feedback oscillator circuit in which energy from the plate circuit is fed from the tickler, L , to the grid coil, L_1 . The frequency of oscillation is determined mainly by the values of L_1 and C_1 .

Now, it should be noted from the circuit that there is no C battery to provide a negative bias for the grid. At the very outset, then, the grid is at zero potential, and a tube so operated is highly sensitive to any circuit change. Therefore, the very first minute voltage appearing on the grid will cause an immediate change in the plate current.

If we assume this initial voltage to be positive, then the grid, having this positive value, will cause the plate current to further increase. This, naturally, hastens the expansion of the field about coil L , inducing a still larger positive voltage on the grid through L_1 , and bringing about a still more extensive increase in plate current.

This action will continue until the plate current reaches a maximum when it will taper off and finally become a steady value. Thereupon the field about coil L will become static and no longer induce a voltage in L_1 ; the grid therefore loses its positive value and returns to zero. Since it is now "less positive" than its previous value, the plate current commences to decrease. As it does so, the field about coil L starts contracting and induces a negative voltage in the grid circuit. The grid, now having a negative value, decreases the plate current still further and thus hastens the collapse of the field about coil L . A still larger negative voltage is therefore present on the grid which further reduces the plate current. This continues until the field about coil L is completely collapsed and no longer induces a voltage in L_1 . The grid voltage thereupon returns to zero and now being "less negative" than its previous value will again cause the plate current to increase . . . and so the process repeats itself.

The foregoing may be termed the "initial action" for other actions are taking place simultaneously. The first of these actions to be considered is that which takes place in the tank circuit L_1C_1 . That is to say, the voltage pulses induced in the grid circuit by the feedback action charge the variable condenser C_1 which, when it has reached its capacity, will discharge through the coil, thus setting up an oscillating current that will continue throughout the cycle. The charge and discharge of the condenser takes place first in one direction and then the other, and the resulting alternating current would eventually die out if it were not for the fact that the grid circuit is replenished once each cycle from the output circuit of the tube through the feedback coil L . Therefore the oscillations continue unabated and the waveform of these oscillations, *as well as their frequency*, is governed by the grid circuit. The frequency of the oscillations so produced is dependent upon the resonant frequency of the grid circuit; that is, the frequency to which it is tuned by means of the variable condenser. If the circuit is properly designed, the waveform of the alternating current set up in the grid circuit will approach that of a sine wave.

But the action that takes place has not as yet been completed. There is still the grid bias to be considered. This is developed by

the grid condenser C and the grid leak R in series with the grid. The manner in which these components function to maintain a negative bias on the grid is as follows:

The capacity of condenser C is of such a value as to provide a free path to the grid for the excitation signal and therefore bypasses the high-resistance grid leak R . During positive signal alterations the grid draws current in the same manner as a diode. This direct current flows in the external circuit from cathode, through L_1 and R , to the grid. A voltage drop is therefore developed across resistor R , the end of R nearest the cathode being more positive than the end nearest the grid. The grid is therefore negative with respect to the cathode by an amount equal to the voltage drop across R .

The voltage developed across R during that period when grid current flows charges condenser C . If R and C have the proper values, then C will hold its charge at an appreciably constant value throughout the complete cycle of signal voltage.

Actually, the condenser C discharges across resistor R during each negative signal alternation when no grid current flows, but at a comparatively slow rate. The rate of discharge is dependent upon the capacity of the condenser and the resistance of the grid leak. The higher the capacity of the condenser, the larger the charge it can hold and the longer it will take to lose its charge. The higher the resistance of the grid leak, the longer it will take a condenser of given capacity to discharge across it. We refer to the rate at which discharge will take place as the *time constant* of the condenser-leak combination.

Regulating Action

In an oscillator the time constant is usually made fairly long as compared to the time of one cycle so that the charge on the condenser does not begin to fall off appreciably until a signal cycle is completed.

Since an oscillator is self-excited, it is evident that all voltage and current values are interdependent. That is to say, any increase or decrease in the value of the plate current is immediately reflected in an increase or decrease in the excitation voltage induced in the grid circuit by the feedback action. This is

naturally followed by an increase or decrease in grid bias which opposes a further change in plate current in the same direction. Thus, a rise in the average value of plate current is offset within a few cycles by a subsequent rise in grid bias which reduces the plate current. After a short space of time the circuit again reaches stabilization, each succeeding counter-variation in plate current and grid bias becoming smaller and smaller until constant values are once more assumed. This action can be likened to that of the mechanical governor which regulates the speed of a phonograph turntable.

Another point regarding the regulation of the condenser-leak combination is worth mentioning. If the capacity of condenser C is too large, it will take a considerable time for the negative charge to leak off through resistor R . In such a case, the grid will be made insensitive to a sudden change in average plate current and the tube will cease oscillating and start up again at regular intervals, the starting and stopping rate depending upon the time constant of C - R . This action is often productive of an audio howl because it is the equivalent of modulation of the generated voltage. When the oscillator output is rectified, this start and stop of oscillations appears as an audio signal.

For instance, if the average plate current suddenly decreases, the grid bias will remain the same because of the slow rate at which C discharges. But a decrease in plate current will also decrease the grid excitation voltage. This will result in a further decrease in plate current and the action will continue until the output can no longer supply sufficient feedback voltage to overcome the input circuit losses, at which time oscillations will cease. Oscillation will not start again until the charge on condenser C has leaked off and plate current again rises. This cycle will repeat itself over and over again, as the plate current is never so steady that it can maintain constant output without proper bias regulation.

Wave Form of Output

The wave form output of such an oscillator is dependent upon the extent of regeneration and the grid condenser-grid leak combination. The best wave form—that is the closest approach to a

sine wave with minimum harmonic generation—is attained when the amount of regeneration is critical—just enough to maintain oscillation. An excessive amount of regeneration will cause a departure from a sine wave output. However, there is a definite latitude between the critical amount and a greater amount of regeneration, during which the output will be such as to be accepted

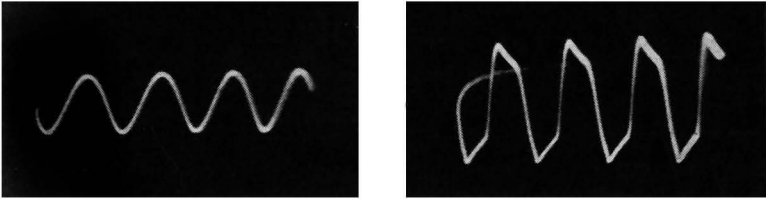


FIG. 3-2, left, FIG. 3-3, right. When there is just sufficient regeneration to maintain oscillations, a sine-wave output can be obtained, Fig. 3-2; but when regeneration is excessive, the output departs from its sine character, Fig. 3-3, showing the presence of harmonics.

as sine. This is of the greatest importance when generating output voltages in the audio-frequency range. At frequencies above the audio range, tuned circuits are involved and the selected single frequency is a sine wave of voltage.

Amplitude and Stability of Output Voltage

The amplitude of the oscillations generated in the circuit depends upon a number of factors: the tube, the operating voltages, the amount of feedback and the electrical efficiency of the components. Critical regeneration provides the purest wave form but not the greatest amplitude. The higher the efficiency—that is, lowest losses—the greater the output. Stability is a function of the constancy of adjustments, operating voltages, tube characteristic and components, and the amount of regeneration. The less the regeneration, the greater will be the stability.

Chapter IV

TRIODE OSCILLATORS

WITH the exception of a very few special types, and regardless of the apparent complexity of the circuits, all vacuum-tube oscillators are ultimately based upon the triode tube. Moreover, although triode oscillators appear in a large number of different disguises, all actually derive from a single type. This is not to say that all oscillator circuits are equally as good for all classes of service; certain circuit arrangements are superior to others in respect to the functions performed.

Just which oscillator circuit is to be considered as fundamental is largely a matter of choice. One becomes another by a rearrangement of the same circuit elements; others are slightly more elaborate, but are limbs on the same family tree.

Sometimes it is not easy to detect the basic resemblance of one oscillator circuit to that of another, or, for that matter, to be sure of the manner in which feedback is obtained. So let us review the circuits most commonly used, and employ the Hartley oscillator as the starting-off point.

The Hartley Circuit

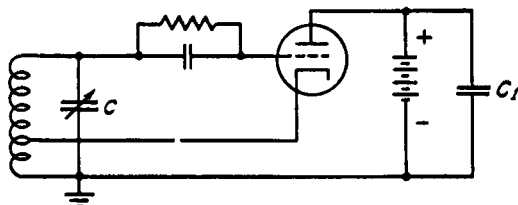
Compare the tickler feedback circuit of Fig. 3-1 in the previous chapter with the Hartley circuit of Fig. 4-1. They are identical but for two exceptions: In the case of Fig. 4-1, the tickler coil has been removed from between the plate of the tube and B plus and put between cathode and B minus; that is, the tickler is in series with the low-potential end of the plate circuit rather than the high-potential end. This being the case, a single coil with a

tap near its lower end for the cathode connection takes the place of two separate coils inductively coupled to each other. The coil in Fig. 4-1 therefore functions as an auto-transformer. As you can see simplification is accomplished.

The degree of voltage feedback induced in the grid-cathode section of the coil is only partly dependent upon the position of the cathode tap, since capacitive coupling exists because the variable condenser C is connected across both coil sections. Since the plate current flows through the plate-cathode section of the coil, the greater the number of turns between cathode and B minus, the larger the feedback voltage developed.

The other exception is that the entire coil forms the resonant circuit tuned by the variable condenser C . The tank circuit is therefore common to both grid and plate. This is characteristic of all Hartley oscillators.

FIG. 4-1. The Hartley oscillator circuit. Note that the tank circuit is common to both the grid and plate circuits of the tube.



It should be noted that in both Fig. 3-1 and Fig. 4-1, the B battery is in series with the feedback coil, or, to put it another way, the plate current flows through the coil. Any oscillator circuit connected in this manner is said to be *series fed*. What this specifically means is that both the d-c plate current and the a-c component of this current flow through a common circuit, there being no other path available.

Since the B battery develops a high internal resistance as its voltage decreases, and will then impede the flow of r.f. and cause a voltage drop, a secondary path for the r.f. is always provided. This is accomplished by connecting across the B battery a fixed condenser of sufficiently high capacity as to have negligible impedance to the flow of r.f. The a-c component of the plate current is therefore *bypassed* around the B battery, taking the path of lowest resistance.

The bypass condenser C_1 in Fig. 4-1 serves this purpose. Having negligible impedance this condenser effectively places the plate of the tube at ground potential insofar as r.f. is concerned. It is interesting to observe that a similar bypass condenser connected across the B battery in the circuit of Fig. 3-1 would serve the same purpose, but would *not* place the plate at r.f. ground potential since the tickler coil is interposed between the plate and the B battery.

Now compare the circuit of Fig. 4-1 with that shown in Fig. 4-2. These circuits are, in a sense, the same, although there is a distinction to be drawn between the two. Both are series-fed

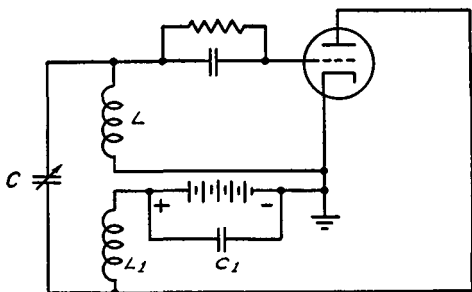


FIG. 4-2. In this Hartley circuit, the tickler coil, L_1 , is in the high-potential end of the plate circuit. Compare with Figs. 3-1 and 4-1.

Hartley circuits, and both have their resonant tank circuits common to plate and grid. But the so-called tickler coil is back in the high-potential end of the plate circuit, where it is in Fig. 3-1.

In order to analyze this circuit properly, it is necessary to consider coils L and L_1 as connected together at their inner terminals, and from that point commonly connected to the cathode, which is grounded. This we may assume without hesitancy because the bypass condenser C_1 offers negligible impedance to the flow of r.f. That being the case, it is equivalent to a direct connection as far as r.f. is concerned; hence the inner terminals of the coils are both at r.f. ground potential.

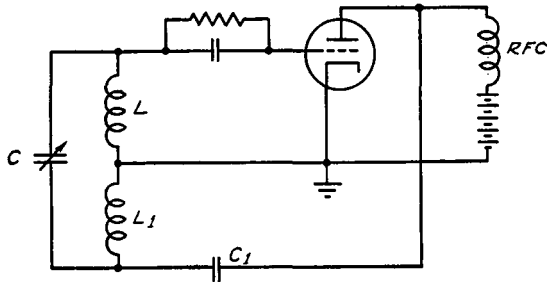
Taken together in this way, coils L and L_1 form an auto-transformer. Since they form the resonant circuit tuned by the variable condenser C , and since L is common only to the grid and L_1 to the plate, then any r-f voltage developed across the plate coil L_1 will also appear in opposite phase across the grid coil L .

Consequently, that portion of the r.f. developed across L will be applied to the grid, and the requirements for the generation of oscillations are satisfied.

Shunt Feed

The Hartley circuit of Fig. 4-3 is identical to that of Fig. 4-2 except that it is *shunt fed* rather than series fed. That is to say, the circuit in which the d-c plate current flows is in shunt with or completely divorced from the circuit in which the a-c component of the plate current flows.

FIG. 4-3. Another Hartley circuit in which the d-c plate current is entirely separated from the circuit in which the a-c component of the plate current flows. This is the shunt-fed Hartley oscillator.



An examination of the circuit will reveal that the d-c plate current can flow only in the plate-to-cathode circuit containing the radio-frequency choke RFC and the B battery. It cannot flow in the plate-to-cathode r-f circuit containing the condenser C_1 and the coil L_1 because C_1 effectively blocks the flow of direct current.

On the other hand, the r.f. cannot flow through the d-c circuit since the r-f choke RFC offers a high impedance to the flow of r.f., but practically none to the direct current since its d-c resistance is negligible. But the r.f. can flow readily in its own circuit since the impedance of condenser C_1 is negligible.

Since C_1 also acts as a blocking condenser, and keeps the plate supply voltage off coil L_1 , this coil and coil L can be connected directly at their inner terminals and returned to the cathode at this point.

It is not necessary in either of these Hartley circuits that the

coils L and L_1 be inductively coupled, though they often are. Since the tank condenser C is common to both coils, the r-f voltage developed in L_1 will appear across the condenser and therefore across coil L .

The Tuned Plate Circuit

The tuned plate circuit, also known as the reversed feedback circuit, shown in Fig. 4-4 may seem at first glance to be similar to the Hartley circuit of Fig. 4-2. However, you will note that the tuning condenser C is shunted across only the plate coil L_1 while in Fig. 4-2 the tuning condenser is across *both* plate and grid coils. Thus, in the tuned plate circuit oscillator, feedback is obtained without the aid of the capacitive coupling through the

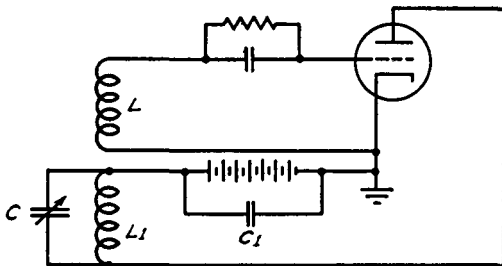


FIG. 4-4. In this oscillator only the plate circuit, L_1 , is tuned by the condenser C . This is known as the tuned plate or reversed feedback circuit.

tuning condenser. Essentially, this circuit is similar in action to the standard tickler feedback circuit shown in Fig. 3-1, but in Fig. 4-4 the grid coil L is the tickler and the plate coil is tuned whereas in Fig. 3-1, the reverse is true.

The Colpitts Circuit

Now compare any of the Hartley circuits with that of the Colpitts oscillator circuit shown in Fig. 4-5. It will be seen that a condenser has been substituted for each of the coils, and a coil for the condenser, or, that the condenser is "tapped" instead of the coil.

The tank is still common to the plate and grid circuits, but feedback is obtained by coupling the plate to condenser C_1 . The

amount of feedback is determined by the ratio of the capacities of C and C_1 .

Note that this circuit is also shunt fed. With the circuit as it is, the bypass condenser C_2 also acts as a blocking condenser. Hence, only the a-c component of the plate current appears in the tank circuit.

FIG. 4-5, right. The Colpitts circuit. Here feedback is accomplished by coupling the plate of the tube to condenser C_1 , the amount of feedback depending on the ratio of C to C_1 .

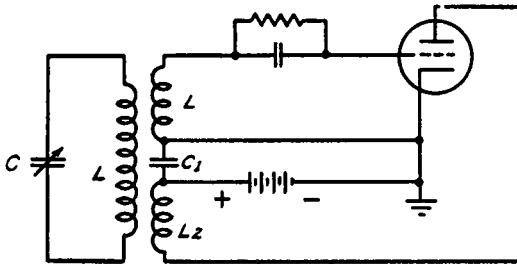
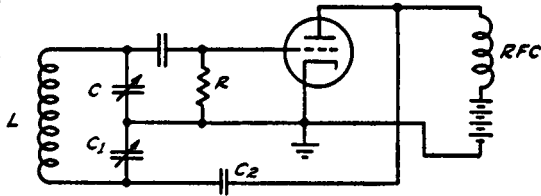


FIG. 4-6, left. The Meissner oscillator. The floating tank circuit, L-C, serves as an inductive feedback coupling unit between L_2 and L_1 , which are coupled to L-C.

It should also be noted that it is necessary to connect the grid leak, R , in shunt with the grid circuit rather than in series with it. In the latter case, there would be no available return circuit to the cathode, as the grid is completely isolated from the cathode insofar as d.c. is concerned. It is necessary, therefore, to connect the grid leak from grid to cathode in order to create a d-c path for the grid current.

The Meissner Circuit

The Meissner oscillator circuit shown in Fig. 4-6 is seen to be similar to the Hartley series fed circuit of Fig. 4-2, with the exception that the tank circuit L-C is "floating." Actually, it

serves as an inductive feedback coupling unit between the plate coil L_2 and the grid coil L_1 which are coupled to it. Moreover, it determines the frequency of the oscillating circuit. The tank is excited by the voltage induced in it by the plate coil; the oscillating current thus set up develops a voltage in L which is induced into the grid coil L_1 .

Tuned-plate—Tuned-grid Circuit

The tuned-plate—tuned-grid oscillator circuit shown in Fig. 4-7 is a bit deceptive in that it looks very much like the series-fed Hartley of Fig. 4-2 opened out, but with the grid and plate coils separately tuned. That is about what it is, but the action

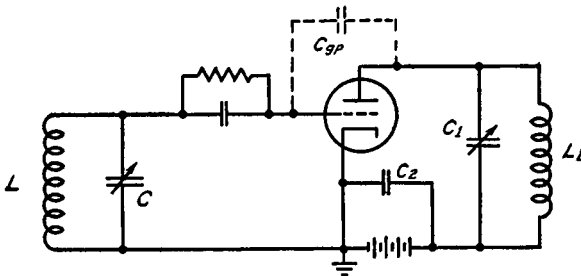


FIG. 4-7. In this tuned-plate—tuned-grid oscillator circuit, feedback is accomplished by means of the coupling effect of the capacity existing between the plate and the grid of the tube, as indicated by the dotted condenser C_{gp} .

of the tuned-plate—tuned-grid oscillator is predicated on the feedback from the output to the input via the inter-element capacity of the tube; that is, the internal grid-plate capacity, represented in the diagram by C_{gp} .

If both tuned circuits are adjusted to resonate at approximately the same frequency, oscillations will be set up in the following manner: The initial plate-current pulses will set up an oscillating current in the tank circuit L_1C_1 . A portion of the r-f voltage thus developed across the tank will be fed back to the grid circuit through the internal grid-plate capacity C_{gp} of the triode. Since the grid tuned circuit LC is adjusted to the same

frequency as L_1C_1 , an r-f voltage will build up in it, the point of high potential being at the grid end. This voltage will appear on the grid and cause a larger change in plate current. The frequency of the plate-current pulses will be the same as the frequency to which the plate tuned circuit is resonated.

Once the frequency of the oscillations is established by the plate tank circuit, all subsequent oscillations will be of the same frequency and the grid and plate tuned circuits will present a high impedance to these r-f currents. Consequently there will be a maximum voltage built up in each of the tuned circuits.

If either of the tank circuits is tuned off resonance to an appreciable degree, the tube will cease oscillating for, in that event, the circuit tuned off resonance will present a low impedance to the r-f current. As a result of this, there will be no appreciable voltage build-up and therefore little if any voltage on the tube element to which the tuned circuit is connected. If it is the grid tuned circuit that is off resonance, there will not be sufficient voltage to excite the grid and cause an appreciable change in plate current; if it is the plate tuned circuit that is off resonance, there will be insufficient feedback voltage to excite the grid tank circuit. In either case, oscillations will die out and will start up again only when both plate and grid tanks are turned to approximately the same resonant frequency.

In actual practice, the frequency of oscillation is controlled principally by the plate tank circuit, and the degree of excitation by the grid tank circuit. The plate and grid tanks need not be in exact resonance with each other. A slight discrepancy is permissible.

Parallel and Push-pull Circuits

Two (or more) tubes of the same type may be paralleled in any oscillator circuit for greater power output, as, for instance, in Fig. 4-8. Naturally, the plate-cathode, grid-cathode and grid-plate capacitances are also paralleled, with the result that each of these inter-element capacitances has twice the value of that for a single tube, thereby increasing the minimum circuit capacitance and decreasing the range of tuning. Moreover, when the

total capacitance is small, variations in tube capacitances cause frequency instability.

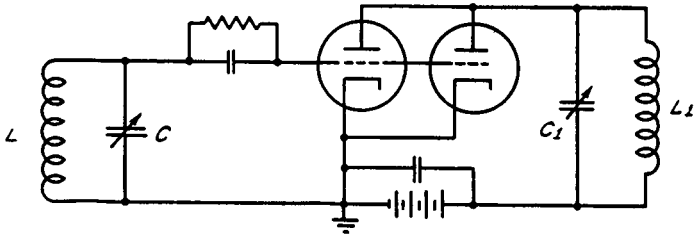


FIG. 4-8. In order to obtain an increase in the power output of an oscillator, two tubes of the same type can be connected in parallel, as shown in this schematic.

The push-pull oscillator circuit shown in Fig. 4-9 is particularly desirable at the higher frequencies as the inter-element capacitances are effectively in series, and therefore have one-half the value of those for a single tube. At ultra-high frequencies, however, the tube lead inductances, which are also in series, tend to overcome the advantages of decreasing input capacitance.

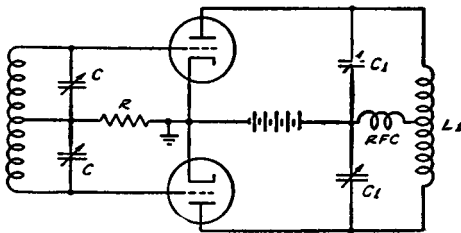


FIG. 4-9. This push-pull oscillator has good frequency stability and a relatively small harmonic content.

The push-pull circuit has greater frequency stability and a decreased harmonic content. The tank coils are center-tapped and tuned by split-stator variable condensers. Therefore the circuits are balanced against ground and both halves of each excitation voltage cycle are positive with respect to this point. Each tube handles an alternation opposite in original polarity to that handled by the other tube so that both alternations are utilized. As a result of this, even-order harmonics are cancelled out in the plate tank circuit.

The push-pull oscillator, and the parallel circuit employing two tubes provide twice the power output obtained from a single tube.

Two-terminal Oscillator Circuits

The two-terminal oscillator shown in Fig. 4-10 is one of the few arrangements that appears to bear no relation to the more common circuits. This is due solely to the fact that an additional triode is used in place of the usual tickler coil.

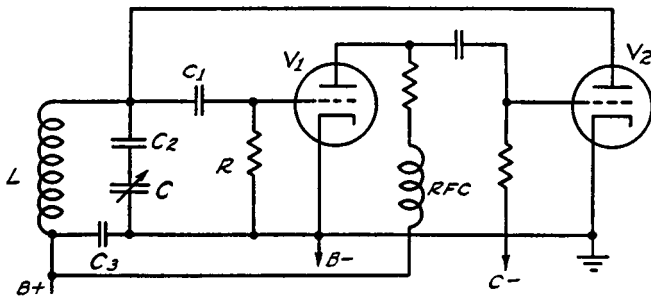


FIG. 4-10. In the two-terminal oscillator circuit a second triode, V_2 , replaces the usual tickler coil in the plate circuit of the oscillator tube, V_1 .

Referring to Fig. 4-10, $L-C-C_2$ form the tank circuit connected to the grid of tube V_1 . C_1 is the grid condenser and R the grid leak. The output of tube V_1 is resistance-capacity coupled to the input of tube V_2 . The output of V_2 is coupled to the input of V_1 through C_1 . Thus, the action of V_2 is analogous to that of a tickler.

The a-c voltage on the plate of V_1 is 180° out of phase with the a-c voltage on the grid. Therefore the a-c voltage on the grid of tube V_2 is 180° out of phase with the grid of V_1 . But the phase is again reversed in tube V_2 so that the feedback voltage from the output of V_2 to the input of V_1 is in phase and the condition for the generation of oscillations is met.

It will be noted that the plates of both tubes are series fed. Tube V_2 receives its plate voltage through the tank coil L . The blocking condensers C_1 and C_3 are therefore necessary to keep the high voltage off the grid of tube V_1 . The purpose of the choke RFC is to keep the out-of-phase a-c in the plate circuit of one tube from appearing in the plate circuit of the other tube.

In the circuit of Fig. 4-11, the plate of tube V_1 is series fed and the plate of tube V_2 shunt fed. The tank circuit is not "hot" in this case, but the blocking condenser C_1 is required in the feedback lead to keep the plate voltage of V_2 off the grid of V_1 .

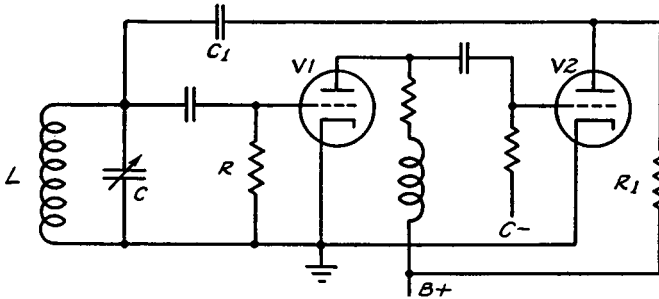


FIG. 4-11. In this two-terminal oscillator the plate of V_1 is series fed and that of V_2 is shunt fed. Although the output of this type of oscillator is rather low, it is fairly constant over a wide frequency range.

A disadvantage of this system is that the plate voltage on tube V_2 is reduced considerably by the drop across the plate load resistor R_1 . However, this circuit provides a wider tuning range than the circuit of Fig. 4-10.

An advantage of these oscillator circuits is that the output voltage is substantially constant over a fairly wide tuning range, although the output is less than that obtained from conventional oscillators. This type of oscillator is therefore restricted to such services where large outputs are not required. The two-terminal feature is advantageous in all-wave superheterodynes, all-wave test oscillators, etc., since it simplifies the switching problem. It will operate at frequencies up to 20 mc.

Grid Bias Methods

The methods by which grid bias can be obtained in oscillator circuits are somewhat limited by the necessity in the majority of cases of having an arrangement that is self-regulating. This rules out the exclusive use of a fixed bias such as may be obtained from a C battery or a power-supply system. The bias on the grid of an oscillator tube must be low enough so that oscillations can start, yet high enough to keep the value of d-c plate current within reason. A fixed bias cannot very well meet both of these requirements at the same time. Moreover, no matter what the value of the bias may be, if it is fixed, there is no way of compensating for any sudden change in plate current. Thus, as we have previously mentioned, if for some reason the plate current dropped below the value where the output was less than that required to overcome the grid circuit losses, then oscillation would cease. It is evident, therefore, that some method of biasing is required that has the ability of adjusting itself to any sudden changes in plate current.

Grid Leak and Condenser

The only adequate self-biasing method that is completely automatic in its regulation is the grid condenser and leak arrangement, the action of which has been previously covered. There are, however, a number of different ways in which this method can be applied.

In some cases it is desirable to connect the grid leak directly from the grid to the cathode, or to ground, as in the instance of the Colpitts oscillator, shown in Fig. 4-5. The shunt connection has an advantage in oscillator circuits where coil changing is a necessity, such as in all-wave superheterodynes, all-wave test oscillators, etc., as indicated in Fig. 4-12(a). It is clear from this figure that if both the grid condenser and leak were in series with the grid, there would be no return circuit from grid to cathode, and consequently no bias, during the interval of switching from one coil to another when the circuit is open. As a result of this, plate current would be excessive for a brief instant until the bias again took hold. The voltage surge thus set up would create

instability for a period of time that, in a superheterodyne receiver, would result in irregular operation until the oscillator again reached stability.

But if the grid leak is connected from grid to cathode, as shown in Fig. 4-12(a), the bias is not removed from the oscillator tube at any time during the switching from one band to another, as there is always a return circuit.

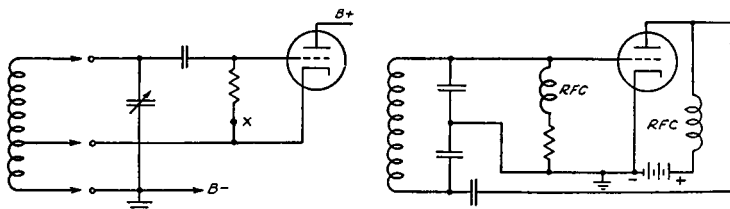


FIG. 4-12(a), left, FIG. 4-12(b), right. In all-wave superheterodyne oscillators, the grid leak is usually connected as shown. The r-f choke (right) reduces losses in power oscillators.

If the parallel grid leak connection is to be used in a power oscillator, it is necessary to connect a radio-frequency choke *RFC* in series with it, as shown in Fig. 4-12(b). The leak in such an oscillator is usually of low value and is capable of absorbing a large amount of the a-c power from the tuned grid circuit. The inclusion of the r-f choke prevents the diversion of the r.f. from the grid through the leak to cathode. In low-power oscillators, such as used in superheterodynes, the value of the grid leak is high enough to prevent any appreciable loading of the tuned circuit.

Cathode Bias

Cathode biasing provides partial regulation since the bias developed is dependent upon the amount of plate current. Bias voltage for an oscillator may be obtained in this manner by inserting a resistor *R* in the cathode circuit of the triode, as illustrated in Fig. 4-13(a). Plate current flowing through this resistor results in a voltage drop across it which makes the grid negative with respect to the cathode by an amount equal to the voltage drop across the resistor.

This method of obtaining grid bias can be explained more adequately by pointing out that the cathode resistor R is in series with the plate resistance of the tube, and forms a part of a voltage divider between cathode and ground. Since the plate is at a positive potential, and the cathode resistor is in series with the plate resistance, it stands to reason that the end of the resistor connected to the cathode will be positive with respect to ground, as will the cathode itself.

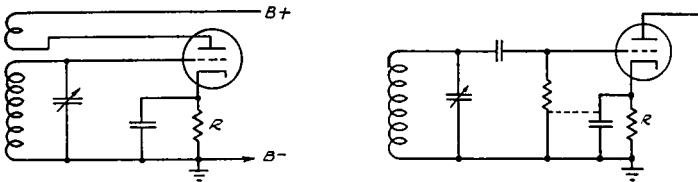


FIG. 4-13(a), left, FIG. 4-13(b), right. The grid bias voltage can be obtained by inserting a resistor, R , in the cathode circuit of the tube. Plate current flowing through R establishes a voltage drop which makes the cathode end of R positive with respect to ground, to which the grid is returned.

Now, the ground or return circuit is at zero potential, and it is to this point that the negative terminal of the B-battery is connected. It is also the point to which the grid return lead connects. But, the cathode is *not* directly connected to the ground or negative terminal; if it were, the grid would be at zero potential *in relation* to the cathode. As it is, the cathode is *positive* with respect to the ground or negative terminal, and therefore the grid is *negative* with respect to the cathode by an amount equal to the voltage drop across the cathode resistor. If the cathode were 5 volts positive above ground, then the grid would be 5 volts negative with respect to the cathode.

It is clear from the foregoing that cathode biasing used in conjunction with an oscillator will provide some regulation. The bias on the grid is zero when the tube is first turned on, and negative bias is developed only as the plate current rises. Any sudden decrease in plate current, for instance, will also reduce the negative bias and permit the plate current to rise again. Moreover, the grid current also flows through the cathode resistor,

and though its effect in altering bias is nowhere near as extensive as a change in plate current, the plate and grid current combined do offer moderate bias voltage regulation.

However, the regulation that can be obtained with cathode biasing is not nearly as effective as the regulation offered by the grid condenser and leak method. Hence, cathode biasing is not commonly used for oscillator control, unless it is employed in conjunction with condenser and leak biasing, as shown in Fig. 4-13(b).

A similar arrangement is often used in a superheterodyne receiver where a dual-purpose tube is used as a combination oscillator and mixer. In such a case, the cathode resistor provides the bias for the mixer section of the tube and the condenser and leak the automatic bias for the oscillator section of the tube. It should be pointed out that in such an arrangement, if the bias developed across the cathode resistor R is to be kept off the grid of the oscillator section of the tube, it is necessary that the grid leak return be directly to the cathode, rather than to ground as shown by the dotted lines in Fig. 4-13(b). If we disregard the bias developed by the condenser-leak combination, then the grid will be zero with respect to the cathode since the grid will have the same positive voltage value as the cathode. The fact that the grid will then be positive with respect to ground will in no way alter the bias developed by the condenser-leak combination.

Measuring Grid Bias

The grid bias developed across a cathode resistor can be measured with an ordinary d-c voltmeter. The reading obtained is the amount of bias on the grid of the tube.

To measure the bias voltage across an oscillator grid leak, a d-c voltmeter having a very high input resistance should be used. One of the electronic type, employing a high resistance in the probe to isolate the cable capacity and inductance from the circuit under test, is desirable, otherwise the loading effect on the oscillator circuit may be sufficient to reduce the voltage developed across the grid leak or even to stop oscillation. If such a voltmeter is not available, the average rectified d-c voltage may be approximately determined by inserting a milliammeter or micro-

ammeter in the grid return circuit at the point marked X in Fig. 4-12(a). The positive terminal of the meter should connect to the cathode and the negative terminal to the oscillator grid leak. The average d-c voltage is approximately given by Ohm's law, the voltage E being equal to the current reading in amperes times the resistance of the grid leak in ohms.

The grid bias reading so obtained gives a valuable indication of the relative magnitude and uniformity of oscillator output. The voltage reading should not vary by a ratio of greater than 2 to 1 as the oscillator tuning condenser is rotated over its range. This voltage reading is normally negative and ranges from 5 to 20 volts, depending upon the type of tube, circuit and component design, and the operating voltages employed in the particular system under test. The voltage limits given do not include high power oscillators such as used in transmitters, but include service oscillators and oscillator systems used in receivers.

Automatic Amplitude Control

Automatic amplitude control, sometimes called automatic oscillation control, is receiving increased recognition because of the benefits it offers. Although the grid leak and condenser method of grid biasing affords a certain amount of automatic control of the grid bias, and therefore of oscillation amplitude, it nevertheless lacks the advantages of true automatic amplitude control.

Automatic amplitude control (aac) as applied to a Hartley oscillator, is shown in Fig. 4-14. A coil L which is connected to a diode rectifier V is inductively coupled to the plate coil of the oscillator. The absorbed power is rectified and the resulting d-c voltage is injected into the grid circuit of the oscillator at points X . Since the amplitude of oscillation is dependent upon the magnitude of the grid bias, it is readily seen that this circuit will tend to maintain the oscillation amplitude constant. For, if the oscillator were to generate an oscillation having a large amplitude, the rectified d-c voltage applied to the oscillator grid would become larger and thereby cut down the oscillation amplitude. Similarly, if the oscillation intensity should decrease, a smaller value of rectified d-c voltage would be applied to the oscillator

grid, resulting in the oscillation amplitude resuming its normal level.

The value of resistor R in the diode rectifier load circuit should be quite high in order to prevent excessive power absorption from the oscillator.

Resistor R_1 and condenser C_1 comprise a filter to eliminate the oscillation frequency from the rectified output applied to the grid of the oscillator. The time constant of the R_1C_1 filter network

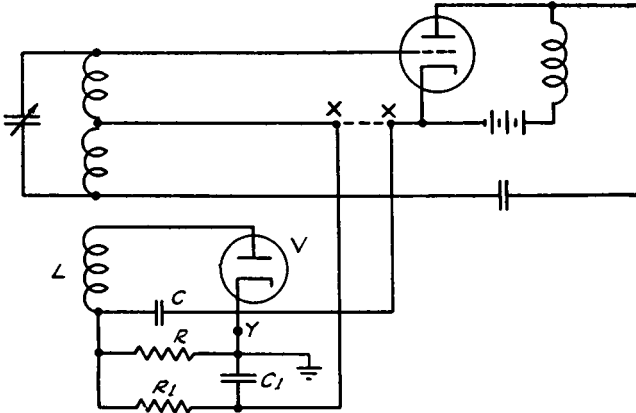


FIG. 4-14. Automatic amplitude control is obtained by injecting the d-c output of the rectifier, V , into the grid circuit of the oscillator tube at the points $X-X$. The input of V is obtained from the coil L , which is inductively coupled to the oscillator plate coil.

is rather critical, for if it is too low the automatic amplitude control action will be inoperative, whereas if it is too high an audio-frequency modulation or motorboating will result. Because of the rather critical nature of this filter, the values of R_1 and C_1 have to be changed if the frequency of the oscillator is changed by any considerable amount.

A delay voltage may be introduced at point Y by biasing the cathode of the diode positively. If this is done the automatic amplitude control circuit will not function until the oscillation amplitude is near the desired operating level.

As can be seen from Fig. 4-15, the automatic amplitude control of an oscillator is essentially the same as automatic volume con-

trol in a radio receiver. In this figure as well as the others, corresponding letters and subscripts designate parts which are functionally equivalent. The filter comprised of R_1 and C_1 in the avc circuit of Fig. 4-15 is utilized to eliminate audio-frequency modulation components from the avc output, whereas in aac application, this filter is employed to eliminate the radio-frequency components from the aac output.

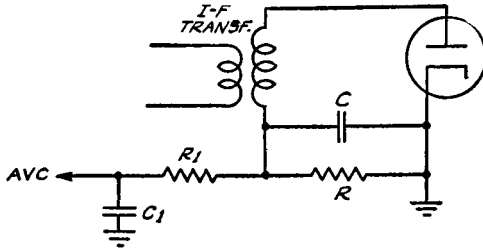
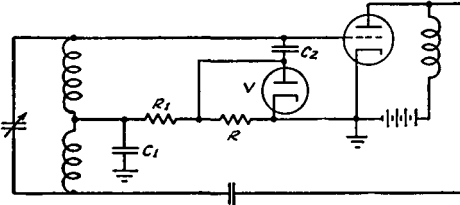


FIG. 4-15, left. Automatic amplitude control in an oscillator is practically the same as avc action in a receiver. Compare this avc circuit with the schematic on the opposite page. The same designating letters are used on parts that function the same.

FIG. 4-16, right. Here aac is obtained by capacitively coupling the rectifier output to the oscillator grid instead of coupling it inductively, as in Fig. 4-14.



Instead of inductively coupling the rectifier circuit to the oscillator, the rectifier may be capacitively coupled through condenser C_2 , as shown in Fig. 4-16. With the exception that capacitive coupling is employed rather than inductive coupling, the similarity between the automatic volume control circuit of Fig. 4-15 and the automatic amplitude control circuit of Fig. 4-16 is very marked.

The use of automatic control results in several advantages. The oscillation amplitude is maintained at a fairly constant level as the frequency of the oscillator is changed. Furthermore, the frequency stability of the oscillator is increased, although it must be kept in mind that the use of aac does not compensate for amplitude variations which are due to changes in the plate voltage of the oscillator.

Oscillator Requirements

It is difficult to speak of oscillator requirements since they depend upon the service to which the oscillator is put. Generally speaking, frequency stability, constant output and purity of waveform are the principal requirements, but there are exceptions.

Frequency stability, however, is most often a requirement, and its desirability in most cases is obvious. An oscillator in a superheterodyne receiver, for instance, must remain stable or the signal output level will vary. If the frequency of the oscillator "drifts" from its proper point, the i-f signal frequency will not correspond with the fixed frequency of the i-f amplifier. In that event the i-f amplifier will be off resonance with the i-f signal and the result will be distortion and reduced signal level due to mistuning.

Frequency drift in an oscillator which determines the carrier frequency of a transmitter will cause much the same difficulty; it will fall out of resonance with the tuned r-f power amplifiers, thus reducing the output and operating off frequency. A high degree of frequency stability is also required in a laboratory oscillator if the calibration is to be relied upon.

Purity of waveform, that is, a minimum of harmonics, may or may not be a requirement. Harmonics may be desirable in a test oscillator so that measurements may be made not only at the fundamental frequency but also at the harmonics of the fundamental. Harmonics may also be desirable in certain types of transmitters. On the other hand, harmonics in the output of a laboratory oscillator might result in erroneous measurements in certain types of tests.

Frequency Stability

There are a number of factors which cause the frequency of an oscillator to vary. If the component parts of the oscillator are not rigidly mounted, mechanical vibration may result in an alteration of the relative position of these parts, with consequent frequency drift.

Temperature changes, in particular, materially influence frequency stability. Not only will a variation in temperature cause the physical size of the circuit elements to change, thereby resulting in a change of frequency, but also such factors as the tube interelectrode capacitance will vary due to an alteration of spacing. Temperature changes may alter the size and impedance of the load circuit, thereby causing a variation in frequency.

Changes in the operating voltages supplied to the tube may cause marked alteration in the oscillator frequency. The frequency of an oscillator responds quite readily to changes in plate voltage.

Numerous methods have been devised to stabilize the frequency of oscillators. Some of these will be discussed in greater detail elsewhere. Some, but not all, methods for attaining frequency stability are given below.

Temperature control to maintain a constant temperature will largely eliminate frequency variations due to the effects of the heating of components. Voltage regulation of the power supply will decrease the possibility of frequency change due to an alteration of operating potentials. A high ratio of capacitance to inductance will also assist in stabilizing the frequency.

In order to minimize the effect of the load upon the oscillation frequency, the load may be electron-coupled to the oscillator. Another method of eliminating the effect of the load upon frequency is to employ a buffer amplifier between the oscillator and the load. The employment of a neutralizing circuit to prevent the load (the circuit from which the output is obtained) from affecting the effective input capacitance of the tube will aid in stabilizing the frequency.

Electro-mechanical devices, such as quartz crystals, magnetostriction rods, or tuning forks, may be employed to stabilize the frequency.

Harmonic production and frequency instability are to some extent correlated phenomena. Accordingly, if the harmonic content of the oscillator is made low the frequency stability will be improved.

Oscillator circuits employing a grid leak and condenser to obtain grid bias will have their frequency stability improved by using a high value of grid leak resistance.

Oscillator circuits may utilize resistance stabilization and inverse feedback in order to secure a high degree of frequency stability. "Tight" coupling between the feedback coil and the tuned circuit improves stability when the feedback voltage is small.

Waveform Purity

Unless special precautions in design are taken, an oscillator will generate not only a fundamental frequency but also harmonics.

Harmonic content may be kept low by keeping the oscillation amplitude small. Class A operation of tubes—that is, operation at the middle of the linear portion of the tube characteristic—will minimize the production of harmonics, although in r-f oscillators, Class B operation (grid bias approximately equal to cutoff) or Class C operation (grid bias appreciably beyond cutoff) may be employed as the harmonics may be filtered out without much difficulty. An increase in the effective load impedance also assists in decreasing harmonic content.

Chapter V

ELECTRON-COUPLED OSCILLATORS

THE principle of operation of an electron-coupled oscillator is identical to that of a triode oscillator insofar as the generation of oscillations is concerned. It differs only in the manner in which the load is coupled to the oscillator circuit.

In the usual oscillator, the load is coupled to the circuit directly, capacitively or inductively, and any change in the value of the load tends to alter the frequency of the oscillator. In an electron-coupled oscillator, the load is coupled to the output by the electron stream.

Fundamental Circuit

The fundamental circuit of an electron-coupled oscillator employing a screen-grid tube, is shown in Fig. 5-1. The oscillating circuit is composed of the control grid, the tuned input circuit, the cathode and the screen grid, the latter functioning as the plate for the oscillator. The load—in this case another tuned circuit—is connected in the regular plate circuit.

In the circuit shown, feedback is through the cathode which is operated above ground potential by tapping it one-fifth to one-third the way up on the grid coil L . Consequently that portion of coil L between the cathode and ground functions as the tickler or feedback coil because it is in series with the cathode which is in the low-potential end of the internal screen-grid circuit. Since the screen grid is at a positive potential, electrons flow from cathode to the screen grid and oscillations are generated just as in the case of the conventional triode oscillator, the frequency of

the oscillation being determined by the value of L and C in the control-grid circuit.

A sufficient number of the electrons strike the screen grid to maintain oscillation; however, the larger number of electrons pass through the screen grid and are attracted by the plate. Since the screen-grid current varies at the oscillation frequency determined by LC , the intensity of the electron stream through the screen grid to the plate will be varied accordingly. As a result of this, oscillations of the same frequency appear in the load L_1C_1 in the plate circuit. The load can be tuned to the same frequency as the oscillator or to one of its harmonics.

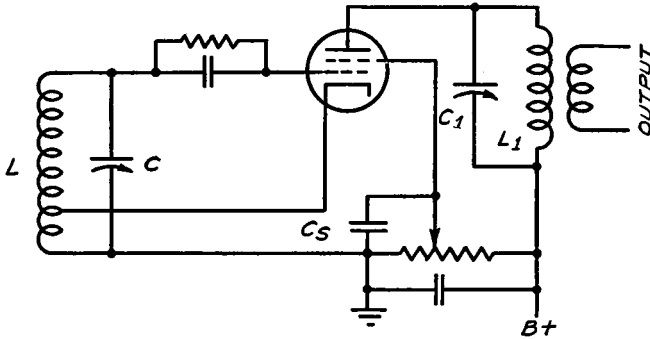


FIG. 5-1. A typical electron-coupled oscillator using a tetrode. Here the screen functions as the plate of the oscillator circuit, while the load is connected to the plate proper.

An oscillator employing a circuit such as the one just described is said to be *electron coupled* because the oscillator circuit and the output or plate circuit are coupled only by a stream of electrons. Since capacitive feedback from plate to grid is largely eliminated, variations in the load circuit do not affect the frequency of oscillations. Electron-coupled oscillators, therefore, have excellent frequency stability as compared to the usual self-excited triode oscillator.

The position of the cathode tap determines the amount of feedback and therefore the degree of excitation, which in some cases may be rather critical. An alternative method is shown in Fig.

5-2 where a separate cathode feedback coil, L_1 , is employed. This is closely coupled to the ground end of the grid winding. The number of turns required will run from one-fifth to one-third the total number of turns on the grid coil. Too many turns will cause excessive excitation of the grid circuit and consequent instability of frequency.

Balancing Frequency Shift

The direction of the alteration of oscillation frequency due to an increase or decrease in plate voltage is *opposite* to the frequency shift occasioned by a corresponding increase or decrease in screen-grid voltage. Accordingly, if the screen-grid voltage be

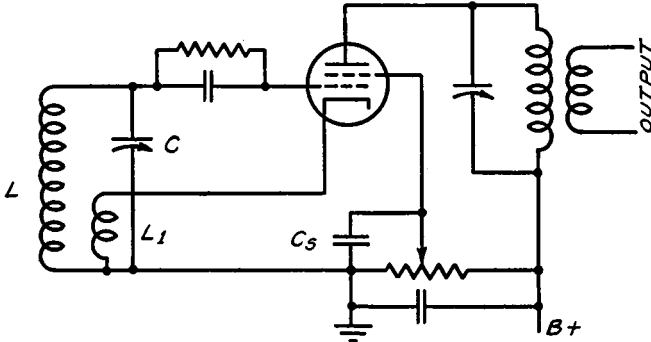


FIG. 5-2. This is an alternative method of feedback, where a separate cathode coil, L_1 , is used, closely coupled to the ground end of the grid coil, L .

taken from a voltage-divider resistance in shunt with the plate-voltage supply, a point can be found on this voltage divider where the frequency shift due to a variation in plate voltage will be balanced by an equal and opposite frequency shift due to the change in screen-grid voltage. This principle is adopted in the circuits shown in Figs. 5-1 and 5-2. The r-f potential of the screen grid is at ground potential by virtue of the fact that the condenser C_s is connected between screen grid and ground. The circuits illustrated are free of frequency variations due to either alterations in the load or in plate-supply voltage.

Pentode E. C. Oscillator

It has been stated that the effects of variations in the load have negligible influence on the oscillator frequency because of the fact that capacitive feedback has been practically eliminated. Actually, a small amount of capacitive feedback is still present so that the oscillator frequency is not completely independent of load variations. By employing a pentode, as shown in Fig. 5-3, such capacitive feedback can be eliminated. The suppressor grid (the

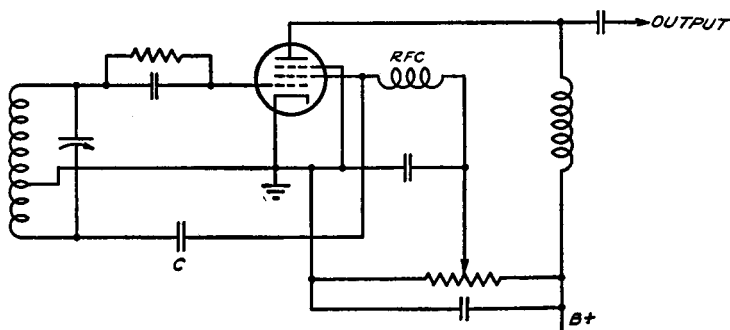


FIG. 5-3. By the use of a pentode, the small amount of capacitive feedback can be eliminated, because the suppressor grid electrostatically shields the plate from the screen grid.

grid nearest the plate) acts as an electrostatic shield which shields the plate from the screen grid. Unlike the circuits of Figs. 5-1 and 5-2, the circuit shown in Fig. 5-3 may be operated with the cathode at ground potential, because the feedback is from the screen grid, through condenser *C*. The suppressor grid is either connected directly to the cathode or it may be operated at a small positive potential. Although this circuit may be said to be free of frequency drift due to variations in the load, it is unfortunately the case that the frequency is not independent of variations in plate-supply voltage, for no ratio of plate voltage to screen voltage can be found which would make this possible.

Neutralization

The circuit shown in Fig. 5-4 employs a neutralizing condenser in order to neutralize capacitive feedback. This circuit is there-

fore free of frequency drift due to load variations. Furthermore, a ratio of plate to screen-grid voltage can be found which

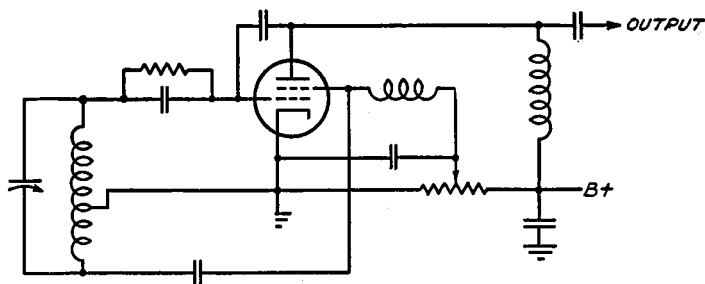


FIG. 5-4. The neutralizing condenser connected between the plate and grid of the tube, neutralizes capacitive feedback and the circuit consequently is free of frequency drift due to variations in the load.

will make the oscillator frequency independent of changes in the plate-supply voltage.

Wave Form

Electron-coupled oscillators, although possessing excellent frequency stability, do not produce a good waveform and the output is accordingly rich in harmonics. For some purposes these harmonics are advantageous. On the other hand, when a good waveform is desired, it is necessary to employ filters in the output circuit to filter out the harmonics. The use of filters is not convenient, particularly when it is desired that the frequency of the oscillator be continuously variable.

Resistance Stabilization

By incorporating *resistance stabilization*, as shown in Fig. 5-5, it is possible to minimize harmonic content. This circuit also utilizes neutralization and an adjustment of the ratio of plate to screen-grid voltages. Accordingly, this circuit not only produces a good waveform, but also has excellent frequency stability since

the frequency is both independent of variations in the load and in plate-supply voltage changes.

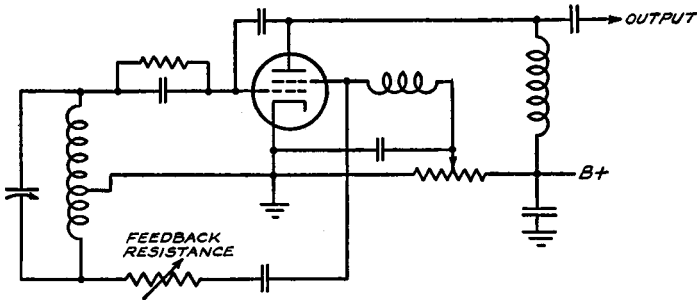


FIG. 5-5. Controlling feedback by means of a variable resistance improves the waveform of this oscillator. This is called resistance stabilization.

Shielding

In all electron-coupled oscillators, adequate shielding of the apparatus should be provided so that no external electromagnetic or electrostatic coupling results between the plate and grid circuits. For maximum frequency stability the plate circuit should be tuned to a harmonic—usually the second—of the grid circuit fundamental frequency. This is done in transmitters but not in service oscillators. Mechanical vibration should be reduced to a minimum in order to prevent the vibration of the tube elements, which would cause internal modulation of the generated signal.

Chapter VI

ULTRA-HIGH-FREQUENCY OSCILLATORS

ALTHOUGH there are no universally accepted lines of demarcation for the various frequency bands, it is found convenient to lump certain bands together and apply a particular designation to them. Thus, the *broadcast band* is generally considered to extend from 500 to 1500 kilocycles (600 to 200 meters). The region from 2000 to 25,000 kilocycles (150 to 12 meters) is accepted as the *short-wave* band. The radio frequencies which have wavelengths from 0.0008 millimeter to 10 meters are frequently known as *quasi-optical waves* because their characteristics are quite similar to light waves. Quasi-optical waves are further subdivided into *ultra-high frequencies* (ultra-short waves) which encompass the range from 30 to 3000 megacycles (10 to 1 meters), and the *micro-waves* which comprise all frequencies greater than 300 megacycles (under 1 meter).

In view of the importance of ultra-high frequencies in television, as well as amplitude-modulated (AM) and frequency-modulated (FM) high-fidelity broadcasting, we shall momentarily digress from the subject of oscillators, and describe some of the characteristics of the ultra-high frequencies.

Unlike the broadcast and short-wave radiations which are subjected to reflection and refraction from the ionized layers in the atmosphere and can, therefore, travel around the earth's curvature, ultra-high-frequency waves are capable only of straight-line propagation; that is, the receiver must be in the optical path of the wave from the transmitter.

Since there is no reflection or refraction of these waves from the upper atmosphere, no interference patterns are set up between

the ground and sky wave, or between several sky waves, and, consequently, no fading of signals. However, any conducting obstacle in the direct path of the wave will result in a pronounced "shadow"; that is, the field strength of the wave will be greatly reduced on the far side of the obstacle.

An important characteristic of ultra-high frequencies is their freedom from atmospheric disturbances. Only man-made static, chiefly from automobile ignition circuits and diathermy apparatus, has an adverse effect on the received signal.

Due to the extremely high frequencies involved, ultra-high-frequency oscillators introduce many difficulties in design which are not of particular importance for oscillators operating at lower frequencies. The arrangement of the apparatus, the length of connecting leads, the design of the tubes used and their operating voltages, play an important part in the resulting frequency of the oscillator. Since a very slight change in the relative position of the various parts or leads will result in a considerable alteration of the oscillator frequency, particular care must be taken to secure mechanical rigidity of the component parts. The internal interelectrode capacitance of the tube may often be greater than the capacitance employed in the external circuit, and will, therefore, limit the maximum attainable frequency. The very small values of inductance and capacitance employed may cause trouble in obtaining suitable impedance values in the tuned circuits.

In the usual oscillator the time taken by an electron to travel from cathode to plate is so short that for most purposes it is considered to be instantaneous. At ultra-high frequencies, however, this assumption no longer holds, for the time taken by an oscillation cycle may be much less than the electron transit time, and this factor must, therefore, be taken into consideration.

Types of UHF Oscillators

In order to overcome the foregoing difficulties, numerous types of oscillators have been designed, of which the following classification lists the more important, together with their frequency ranges.

Oscillators for Ultra-High-Frequencies

I. *Regenerative*

1. Tuned-plate tuned-grid, Colpitts, Hartley, (Up to 120 megacycles)
(10 to 2.5 meters)
2. Gutton-Touly (Up to 150 megacycles)
(Down to 2 meters)
3. Mesny balanced (Up to 200 megacycles)
(Down to 1.5 meters)

Regenerative Oscillators

The tuned-grid tuned-plate oscillator has already been described and will not be further discussed except to say that for UHF work the plate and grid inductances and capacitances are reduced to very small values.

FIG. 6-1, right. The Gutton-Touly uhf oscillator. The inductance consists of a single turn of wire connected between the grid and plate.

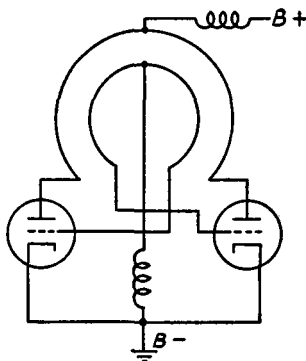
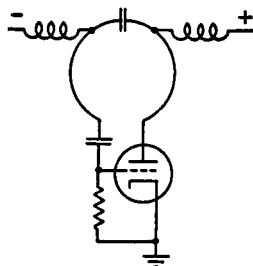


FIG. 6-2, left. The Mesny balanced uhf oscillator. This is similar to that shown above, but the use of two tubes gives greater electrical symmetry.

Mention has already been made of the importance of the circuit connecting leads at the ultra-high frequencies. As the inductance and the capacitance of the tuned circuits are further and further decreased, the inductance and capacitance of the

leads becomes more and more important until they finally exert a preponderating influence. This effect has been utilized in the Gutton-Touly oscillator which consists of a single turn of wire between the grid and plate of a vacuum tube, as shown in Fig. 6-1. A somewhat similar scheme is employed in the Mesny balanced oscillator, as shown in Fig. 6-2, except that two tubes are used in order to obtain greater electrical symmetry.

The frequency limit of regenerative oscillators is approximately 300 megacycles, for beyond that frequency the time of a single high-frequency oscillation is less than the electron transit time. Frequencies beyond the limits of regenerative oscillators are accordingly produced by means of *electron oscillators*.

Chapter VII

NEGATIVE RESISTANCE OSCILLATORS

IN ANY resistive circuit in which an electron flow occurs, there are three factors which must be taken into consideration: the resistance, the voltage across the resistance, and the current through the resistance. These three factors are related to each other in Ohm's law, as follows:

$$R = \frac{E}{I}$$

where

$R = \text{resistance (ohms)}$

$E = \text{voltage (volts)}$

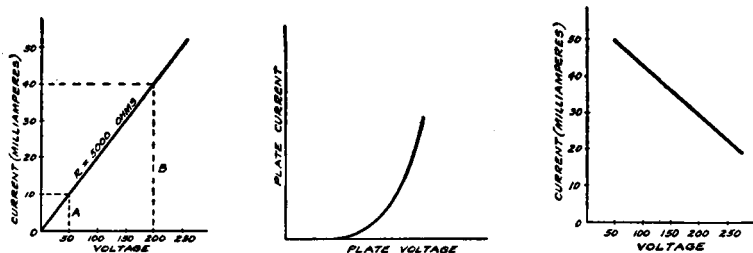
$I = \text{current (amperes)}$

It may be said that for any given value of resistance, an increase in voltage will produce a corresponding increase in current; likewise, if the voltage is decreased the current decreases. This is shown graphically in Fig. 7-1, where the current in milliamperes through a resistance is plotted against the voltage across the resistance. For instance, a potential of 50 volts will produce a current of 10 milliamperes (0.010 ampere), as at *A*. If the voltage is raised to 200 volts an increased current will flow; namely, 40 milliamperes (0.040 ampere), as at *B*.

The voltage-current relationship does not necessarily have to be linear; that is, the line on the graph relating voltage and current need not necessarily be straight. For example, when plate current is plotted for various values of plate voltage applied to a triode, (assuming the grid voltage to remain constant) a characteristic curve such as that shown in Fig. 7-2, is obtained. It will

be observed that as the plate voltage increases the plate current also increases.

Any device in which the current increases when the applied voltage is increased is said to possess *positive resistance* or, for the sake of brevity, *resistance*.



FIGS. 7-1, 7-2, 7-3, left to right. In general, the relation between voltage and current is as shown in Figs. 7-1 and 7-2. In vacuum tubes that are connected in a certain way, the current decreases with an increase in voltage; this is called negative resistance effect.

Now consider the voltage-current characteristic shown in Fig. 7-3. It will be noted that as the applied voltage is increased the current *decreases*. Any device that exhibits this peculiar property is said to possess *negative resistance*. Vacuum tubes connected in certain ways have a negative resistance characteristic over part of their operating range and the electric arc also manifests this property.

Oscillators employing a negative resistance characteristic may be classified as follows:

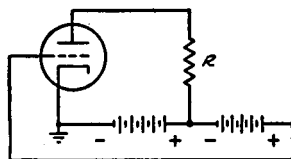
- I. Dynatron oscillator
- II. Negative transconductance oscillator
- III. Electric arc

Of these three types this book discusses types I and II. Type III, the electric arc, receives no consideration because its application is beyond the field of this book.

The Dynatron

If a triode is connected as shown in Fig. 7-4, so that the grid voltage has a higher positive potential than the plate, it will be found to possess a negative-resistance characteristic over a portion of its operating range, as illustrated in Fig. 7-5. A triode connected in this manner is called a *dynatron*. The triode is used here for simple analysis, as tetrodes are generally employed, using the screen grid as the control element. Control grids of most American tubes are not designed to operate at high positive potentials and therefore should not be used.

FIG. 7-4. In this circuit, where the grid has a greater positive potential than the plate, the tube will be found to have a negative-resistance characteristic over a part of its operating range.



The electrons emitted by the cathode are attracted to the positively charged grid, but, since the grid is comprised of a spiral of wire with spacing between turns, many of the electrons pass through these spaces and strike the plate. These electrons which emanate from the cathode are called *primary electrons*. As the positive plate potential is raised, the electrons strike the plate with such violence that their impact dislodges other electrons from the metal plate itself. The electrons thus liberated as a result of impact are called *secondary electrons*.

Now, since the grid has a higher positive potential than the plate, most of the secondary electrons liberated by the plate will be attracted to the grid. And, of course, the number of secondary electrons emitted by the plate increases as the plate voltage is increased. The resulting electronic plate current will be the *difference* between the number of primary electrons which *enter* the plate and the number of secondary electrons *leaving* it. The number of primary electrons emitted by the cathode is almost entirely independent of the plate voltage. On the other hand, since each primary electron may liberate several secondary electrons, the number of secondary electrons issuing from the plate increases rapidly as the plate voltage is increased. It will be

observed from Fig. 7-5 that the number of primary electrons reaching the plate increases as the plate potential is raised from zero to twenty-five volts, as at *A*. Since an increase in plate voltage results in an increase in plate current, this region exhibits a positive resistance characteristic.

So far, so good. But for plate voltages greater than 25 volts, the emission of secondary electrons *from* the plate becomes *greater* than the number of primary electrons *to* the plate, with the result that the *net* electronic plate current to the plate decreases. At point *B*, which corresponds to a plate voltage of 100, the number of primary electrons received by the plate is *equal* to the number of secondary electrons leaving the plate, and the net plate current is zero. For plate voltages from 100 to 175 volts—

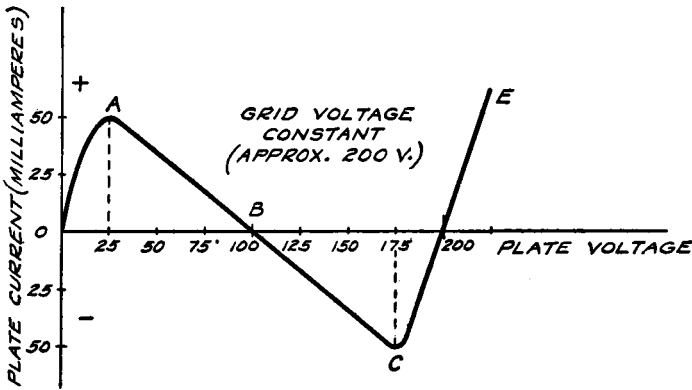


FIG. 7-5. A characteristic curve of a dynatron oscillator, such as that shown in Fig. 7-4. The region from *A* to *C* exhibits negative resistance, as explained in the text.

point *C*, for instance—the number of secondary electrons issuing from the plate *exceeds* the number of primary electrons received by the plate. Under this condition a milliammeter in the external circuit of the tube would read *negative* current; that is, the meter would read backwards!

For plate voltages greater than 175 volts, the grid is not sufficiently positive to attract all of the secondary electrons emanating from the plate, and the plate current consequently rises.

At point *D*, which corresponds to a plate voltage of 200 volts, the number of primary electrons to the plate is again equal to the number of secondary electrons from the plate and the net plate current is zero. Above 200 volts—the region from *D* to *E*, for instance—the plate current rapidly increases and becomes dependent on the primary electronic emission only, for the plate voltage now exceeds the grid voltage. Since the plate current decreases as the plate voltage is increased, the region *A* to *C* exhibits a negative resistance characteristic.

If, instead of the resistance *R* in Fig. 7-4 a coil *L* and condenser *C* are substituted, as shown in Fig. 7-6, the circuit will

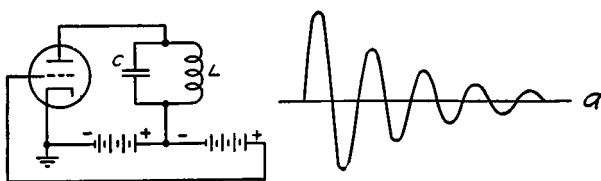


FIG. 7-6, above. In this circuit, inductance and capacity, *L* and *C*, have been substituted for the resistance *R* in Fig. 7-4. The values of *C* and *L* determine the frequency at which this dynatron oscillator will function.

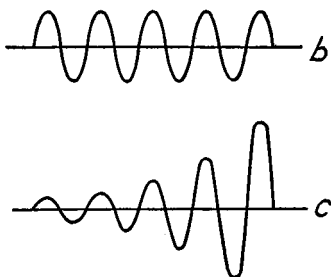


FIG. 7-7, right. When an alternating emf is impressed on a circuit and then removed the amplitude varies as shown according to the nature of the resistance involved, as explained in the text.

oscillate at a frequency dependent on the values of the inductance and capacitance. If an alternating electromotive force is impressed upon a circuit and then removed, the amplitude of the oscillation will decrease, as shown in Fig. 7-7(a), due to the resistance of the circuit. If the circuit possesses zero resistance, any oscillation set up in the circuit will continue indefinitely, as shown at (b) in the same figure.

If a circuit has *negative* resistance, any oscillation started in the circuit would increase in amplitude indefinitely as indicated at (c). In actual practice, however, negative resistance devices exhibit this characteristic only over a limited range, such as the region from *A* to *C* in Fig. 7-5. The amplitude of the oscillation does not, therefore, become infinitely great, but is of some finite value.

Dynatron oscillators employing triodes, as shown in Fig. 7-6, have been displaced by circuits employing tetrodes, as in Fig. 7-8.

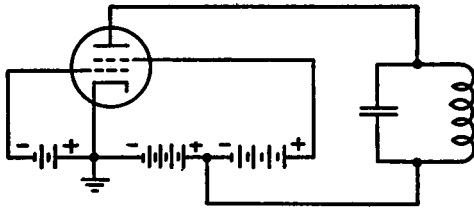


FIG. 7-8. The tetrode is more commonly used in a dynatron oscillator circuit, as it is especially suitable because of its characteristic curve, shown below.

This is advantageous because of the negative resistance properties which a screen-grid tube exhibits over the portion *ABC* of its characteristic, as shown in Fig. 7-9. Grid number two (screen

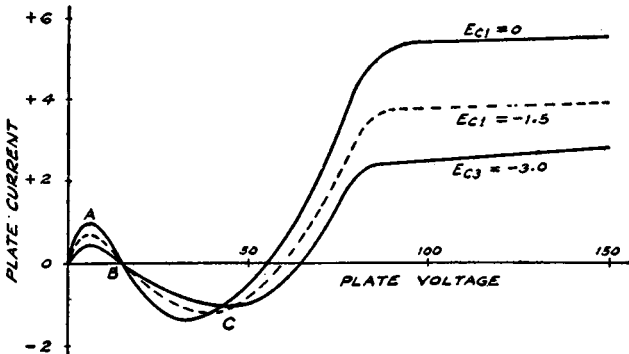


FIG. 7-9. The plate-current—plate-voltage characteristic curve of a tetrode, which exhibits negative-resistance properties over the portion marked *ABC*.

grid) is operated at a higher positive potential than the plate. Grid number one (control grid) controls the number of electrons leaving the cathode and consequently the degree of negative re-

sistance. Electrons emitted by the cathode are attracted to grid number two, pass through the spaces between the wire turns of this grid, and strike the plate. The impact of the electrons results in the liberation of secondary electrons from the plate, and, since the screen grid is at a higher positive potential than the plate, they are attracted to grid number two. When the number of secondary electrons leaving the plate exceeds the number of primary electrons striking it, the tube has negative resistance characteristics; namely, an increase in plate potential results in a decrease in plate current.

The advantages and disadvantages of dynatron oscillators may be summarized as follows:

Advantages:

1. Circuit simplicity.
2. Operable over a wide frequency range (From low audio frequencies to 15 megacycles).
3. Excellent frequency stability; oscillator frequency almost completely independent of changes in operating voltages.
4. Output is an almost pure sine wave. Very low harmonic content.
5. Shifting from one frequency band to another requires the switching of only one connection.

Disadvantages:

1. Not suitable for frequencies above 15 megacycles.
2. Intended mainly for low power output.
3. Dynatron characteristics are not constant due to the depositing of thin films of foreign material on the surface of the plate.
4. Dynatron characteristics change with aging of tube.
5. American vacuum tubes do not possess good dynatron characteristics and show great variation in characteristics, even among tubes of the same type.

Negative Transconductance Oscillator

It is possible to obtain negative-resistance characteristics in vacuum tube circuits which do not employ secondary electron emission effects. For example, the *negative transconductance os-*

illator, as illustrated in Fig. 7-10, does not utilize secondary emission as the basis of its operation, but, rather, employs an electrostatic field to retard or repel the electron stream.

The connections to a pentode are usually such that the No. 1 grid is the control grid, No. 2 grid the screen grid, and No. 3 grid the suppressor grid. The functions of the three grids are quite different in the retarding field tube oscillator. Here No. 3 grid is

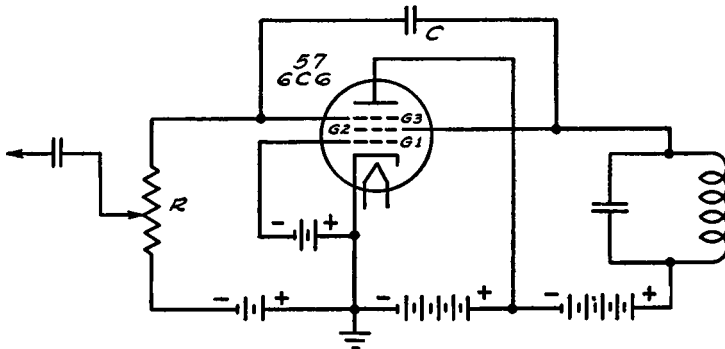


FIG. 7-10. Circuit of a negative transconductance oscillator, which retards the electron stream electrostatically instead of using secondary emission of electrons from the plate.

used as the control grid, No. 2 grid as the anode, and No. 1 grid as a current-regulating grid. Grid No. 3 is negatively biased through a high resistance, R . Grid No. 2 and grid No. 3 are coupled by means of the large condenser, C , so that any variation in the potential on grid No. 2 is transmitted to grid No. 3. The reactance of condenser C must be small as compared to the resistance of R at the lowest frequency employed. The amplitude of the oscillations may be controlled by varying the negative bias on No. 1 grid, the more negative the bias the smaller the amplitude of oscillation. The output signal voltage across R may be fed to the grid of an amplifier tube or other high-impedance circuit, as shown.

The maximum frequency of oscillation that is obtainable with this circuit is in the neighborhood of 15 megacycles, for above this frequency the electron transit time effect becomes an important factor.

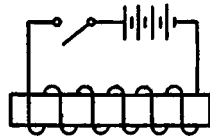
Chapter VIII

ELECTRO-MECHANICAL OSCILLATORS

IF A rod of certain metals is placed within a coil, as shown in Fig. 8-1, and the switch closed so that current flows through the coil, the resulting magnetization of the rod will cause it to increase or decrease in length. This property of changing length when subjected to a magnetic field is called *magnetostriction*.

The principal metals exhibiting magnetostrictive properties are nickel, monel metal, stainless steel, invar, stoic metal, and cer-

FIG. 8-1. When current flows through the coil surrounding the metal rod, the magnetization so set up causes the rod to vary in length. This is called magnetostriction.



tain other alloys. Whether the rod contracts or expands in length does not depend on the direction of the current through the coil, but depends on the metal of which the rod is made. The opposite effect is also obtained; namely, if a magnetostrictive rod is subjected to compression its magnetic properties will change.

If an alternating current is passed through the coil in which the magnetostrictive rod is placed, the rod—assuming that it expands under magnetization—will expand for each positive half cycle, return to normal length when the current is zero, expand for each negative half cycle, and return again to normal length when the current is again zero. It is thus seen that the rod will vibrate at *twice* the frequency of the alternating current supplying the magnetization. In order to make the rod vibrate at the same

frequency as the alternating-current supply, various polarizing devices may be employed, such as, initially magnetizing the rod, or by having a constant direct current flowing through the coil.

This magnetostrictive property may be utilized to control the frequency of an oscillator. Consider, for example, the Hartley oscillator shown in Fig. 8-2. A magnetostrictive rod is placed inside the grid and plate coils. When the oscillator is tuned to the same frequency as the natural frequency of the rod, the rod will maintain the oscillator frequency constant even for considerable capacitance changes in the tuning condenser.

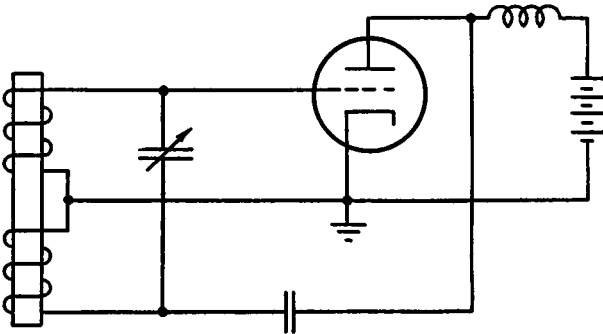


FIG. 8-2. A Hartley oscillator circuit in which a magnetostrictive rod is inserted in the coils for controlling the frequency of the circuit.

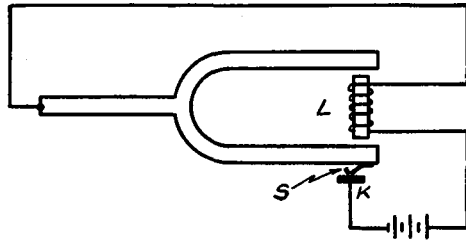
When the oscillator is turned on, the build-up in plate current will result in a magnetic field in the plate coil. This causes a change in the length of the magnetostrictive rod. The change in length of the rod also produces a change in its permeability which, in turn, causes a change in the strength of the magnetic field in the grid coil. The change of magnetic flux in the grid coil results in an alteration of the grid voltage. In consequence, the plate current varies, and the foregoing cycle repeats indefinitely.

The frequency range of magnetostriction oscillators is from 500 cycles to 300 kilocycles. Its frequency stability is one part in thirty thousand when the rod is kept at a constant temperature.

Tuning-Fork Oscillators

Another type of electro-mechanical oscillator is that which employs a tuning fork, such as those used in the musical field. This fork is caused to vibrate by electrical means and in conjunction with other apparatus produces electric oscillations of a predetermined frequency, which is that of the fork itself. Such arrangements are quite popular as sources of audio frequencies.

FIG. 8-3. This type of electro-mechanical oscillator uses a tuning fork, which is made to vibrate by electrical means. The frequency of the unit is determined by the natural frequency of the fork.



A tuning fork may be kept in continuous vibration by means of the circuit shown in Fig. 8-3. The flexible spring, *S*, makes contact with the point *K*. Therefore, a current flows from the battery through the coil *L*. The iron core within the coil becomes magnetized and attracts the prongs of the fork. The contact between *S* and *K* is thereby broken and current from the battery ceases flowing, resulting in the collapse of the magnetic field in the coil. The iron core no longer attracts the prongs of the fork which, having been displaced from their normal position, spring back. Contact is again established and the cycle is repeated.

A tuning fork in which the vibrations are maintained by means of a vacuum tube, is shown in Fig. 8-4. The vibration of the fork prongs causes the magnetic field in the coils *GG* to change, which results in a varying voltage in the grid transformer, *T*. The resultant alternating voltage applied to the grid of the tube is amplified and applied to the coils *PP* which alternately attracts the prongs of the fork. If the phase relations of grid and plate circuits are correct, the vibration of the fork will be maintained.

The frequency of a tuning-fork oscillator is quite stable and when properly designed may be used as a secondary frequency standard. The principal factor which is apt to cause an altera-

tion in the frequency of the fork is temperature. By making the fork of a suitable alloy, whose temperature coefficient is about plus fifteen parts in a million, frequency drift due to temperature variations may be kept very low.

For low frequencies—in the neighborhood of 25 to 1500 cycles—tuning fork-controlled oscillators are superior to crystal-controlled oscillators.

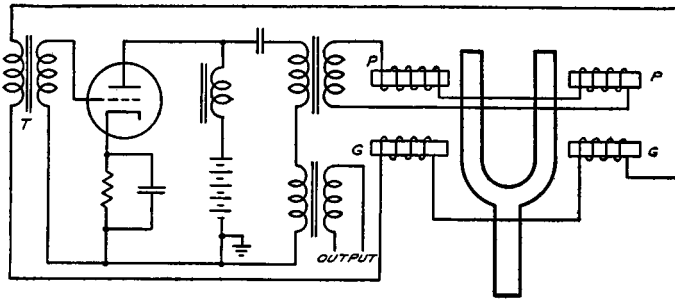


FIG. 8-4. A tuning-fork oscillator in which the vibrations are maintained by means of a vacuum tube.

If frequencies other than the fundamentals of the tuning fork are desired, the electrically maintained tuning fork can be used to control a *multivibrator*, to be explained in the next chapter.

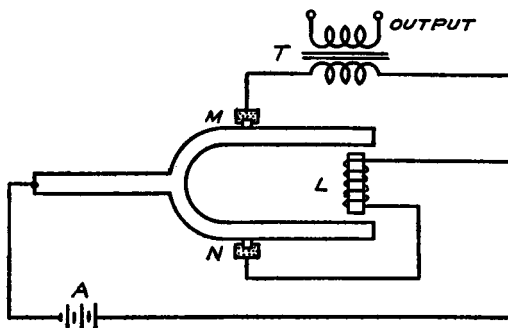


FIG. 8-5. A commercial type of tuning-fork oscillator manufactured by the General Radio Co. M and N are the contacts through which the battery current flows through the fork; L, the magnetizing coil and its core, and T, the output transformer.

A commercial type of tuning-fork oscillator, that made by General Radio, is illustrated schematically in Fig. 8-5.

Crystal-Controlled Oscillators

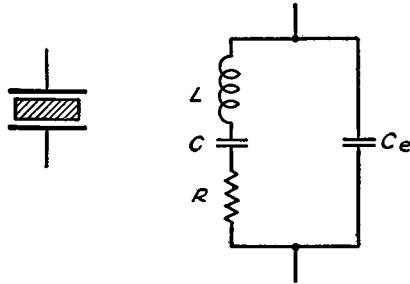
The extensive use of crystals for controlling the frequency of radio transmitters, broadcast stations and laboratory equipment undoubtedly makes this device the most important type of electro-mechanical oscillator. Since quartz crystals are used almost exclusively in crystal-controlled oscillators, we will confine our discussion to them.

Quartz crystals have a property known as the *piezoelectric effect*. When one of these crystals is subjected to mechanical stress, a potential will appear across the faces of the crystal and, conversely, when a voltage is applied across the faces, the crystal will change its shape.

Equivalent Circuit

The schematic drawing for a crystal unit is shown in Fig. 8-6(a). It will be seen that the crystal is sandwiched between

FIG. 8-6(a), left, FIG. 8-6(b), right. The schematic symbol for a crystal unit is shown at the left. At the right is an equivalent circuit of a quartz crystal, which is discussed in the text.



two metal mounting plates. Since this amounts to two metal plates being separated by an excellent insulator having a dielectric constant of approximately 4.5, it might be thought that a quartz crystal unit would act as a condenser. This, however, is not necessarily the case, for, depending upon the frequency applied to the crystal, it may act as an inductance, a resistance, or a capacitance.

The equivalent electrical circuit of a quartz crystal is given in Fig. 8-6(b). It consists of inductance, capacity and resistance. It should not be thought, however, that an ordinary coil, con-

denser, and resistor could be substituted for the crystal, for the ratio of reactance to resistance in the crystal may be more than 100 times greater than that obtainable in an equivalent electrical circuit. Accordingly, the resonance peak of a quartz crystal is many times more sharp than could be achieved by the best coils and condensers. It is this property, coupled with its piezoelectric characteristics, which makes a quartz crystal so valuable in stabilizing the frequency of an oscillator.

Operation of Crystal-controlled Oscillator

The circuit of a typical crystal-controlled oscillator is shown in Fig. 8-7. It is essentially the tuned-plate, tuned-grid circuit except that a quartz crystal X is substituted for the usual coil and condenser in the grid circuit.

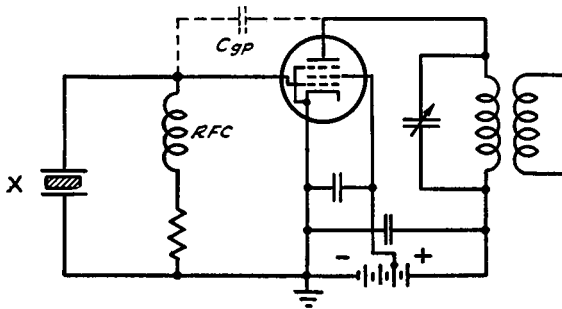


FIG. 8-7. A typical crystal-controlled oscillator circuit, in which the quartz crystal X replaces the usual grid coil and condenser.

Any instantaneous change in the plate current—as when the circuit is placed into operation—results in the transmission of a voltage surge to the grid by way of the internal grid-to-plate capacitance C_{gp} . This voltage is also applied to the crystal and causes it to change its shape and vibrate. Upon the cessation of this voltage surge, the crystal returns to its former shape and position and in so doing develops a potential across its faces. This voltage is applied to the grid and is amplified by the tube. Part of this amplified voltage is fed back to the grid as before,

and therefore to the crystal—and the generation of sustained oscillation starts.

The crystal thus alternately has a voltage applied to it which causes it to change shape, and then, in resuming its shape generates a potential which is impressed upon the grid. The grid is thereby actuated by a voltage corresponding to the natural resonant frequency of the quartz crystal.

It is clear from the foregoing that the crystal functions in the same typical manner as a tank circuit, since it stores energy (in mechanical form) during one-half of the excitation voltage cycle and releases it (in electrical form) during the second half of the cycle. The rate of storage and release is dependent upon the natural resonant frequency of the crystal, and this determines the frequency of the oscillations generated by the tube.

The impedance of a crystal is lowest at its resonant frequency and highest within a comparatively narrow region on either side of the resonant point. Beyond these limits the crystal has no activity and acts as a capacity in the circuit. The resonance peak is many times more sharp than can be achieved by the finest coil and condenser.

A crystal-controlled oscillator is naturally one of fixed frequency. However, the frequency can be varied to a limited extent by means of a crystal holder with variable plate spacing, or by shunting a variable condenser across the crystal. Increasing the capacitance of the condenser results in a lowering of the crystal frequency until a point is reached where the value of the shunt capacitance shorts out the crystal and robs it of excitation voltage. At this point the crystal ceases to oscillate.

Crystal Power and Activity

Crystal power and crystal activity are inter-related factors. The activity of a crystal is not so much the intensity of vibration or oscillation—for this can be raised by increasing the feedback excitation voltage—but rather the sensitivity of the crystal to excitation voltage. Thus, a highly active crystal is more easily excited than a crystal of low activity, which is another way of saying that for a given oscillator output, much less feedback voltage is required for a crystal of high activity.

Since the amplitude of oscillation of a crystal is a direct function of the voltage it develops, or of the excitation applied to it, it might be said that a highly active crystal has high power sensitivity, just as a pentode or beam-power tube is said to have high power sensitivity as compared to a triode power amplifier.

If the excitation voltage applied to an active crystal is excessive, the crystal may vibrate so strenuously as to shatter itself. A crystal of low activity is more rugged in this respect but has poor frequency stability and cannot provide the same power output at high circuit efficiency that is obtainable with an active crystal.

Since there is a limit to the excitation that can be safely applied to the surfaces of an active crystal, the greatest power output can be obtained from oscillator tubes having the highest amplification factor. Pentodes and beam-power tubes require the least amount of excitation for a given power output and are, therefore, to be preferred to triodes in crystal-controlled oscillators. By virtue of the fact that a small degree of excitation is sufficient to drive these tubes to full power output, the required amount of feedback excitation applied to the crystal is also small.

The circuit of a typical crystal-controlled pentode oscillator is shown in Fig. 8-8. Though the grid-to-plate capacitance of a pentode is of a low value, it is sufficient in most cases to provide adequate feedback voltage for crystal excitation. With some pentodes, and with crystals of low activity, it is often necessary to increase the feedback by the addition of a small value of capacitance C connected between plate and control grid. But this condenser should be removed before replacing the crystal of low activity with one of high activity, for the increased excitation might well shatter the new crystal.

Grid Biasing

It will be noted that both grid leak and cathode biasing are used in the circuit of Fig. 8-8. In crystal-controlled oscillators with grid leak bias alone, the crystal ordinarily starts oscillating at zero grid bias and, as the crystal excitation increases, the bias builds up. However, the crystal may be hard starting under such conditions, but if an initial bias is provided by means of a cath-

ode resistor, the plate-to-grid feedback will be increased and the crystal is more readily excited into action.

The addition of cathode bias calls for a corresponding reduction in the grid leak bias, and therefore the value of the grid-leak

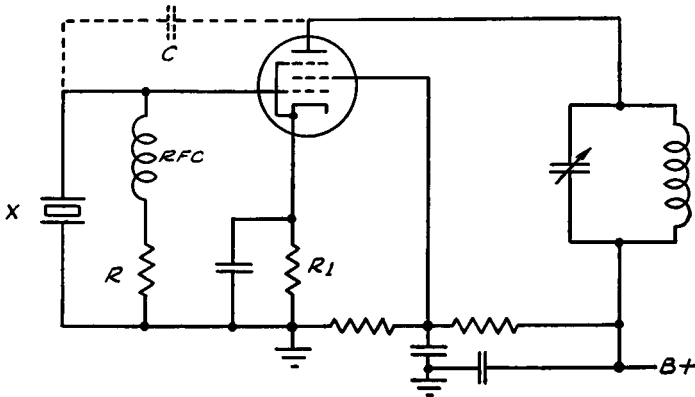


FIG. 8-8. In this crystal-controlled oscillator, both grid-leak and cathode biasing are employed. It is sometimes necessary to increase the feedback by the insertion of a small condenser at C .

resistor R . It is necessary in this case to use an r-f choke RFC in series with R in order to prevent the r.f. from being bypassed.

The Pierce Oscillator

The Pierce oscillator is shown in Fig. 8-9. This is the well-known Colpitts oscillator with the crystal replacing the tank inductance, and is satisfactory only at the lower radio frequencies.

The crystal is connected directly between the plate and the grid of the tube, which may be a triode, a tetrode or a pentode. The variable condenser C is in series with the crystal. There are no tuned circuits involved whatsoever, the frequency of the oscillations generated by the tube being determined by the crystal.

Condenser C is actually in shunt with the plate circuit, and therefore controls the amount of feedback voltage or crystal excitation.

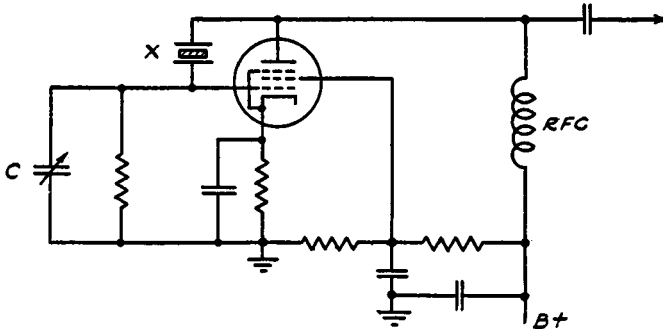


FIG. 8-9. Circuit of the Pierce oscillator, in which the crystal is connected between grid and plate.

Push-Pull Crystal Oscillator

A push-pull crystal-controlled oscillator using pentodes, is shown in Fig. 8-10. Since the output circuit is balanced, the even harmonics are cancelled out, but beyond this the circuit has no particular advantage over a single-tube circuit, except possibly at high radio frequencies.

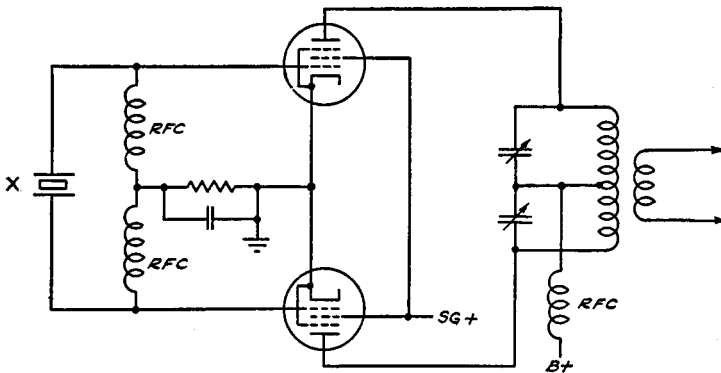


FIG. 8-10. Circuit of a push-pull crystal-controlled oscillator. Due to the balanced output, the even harmonics in the output are eliminated.

In the first place, any circuit using two tubes in push-pull requires twice the driving power of a circuit using a single tube of the same type. Hence, the crystal in the circuit of Fig. 8-10 requires approximately twice the normal excitation voltage in order to drive both of the tubes. This excessive excitation is apt to shatter the crystal. In order to hold the crystal excitation within reasonable limits, it is necessary to use pentodes or similar tubes of high power sensitivity. Even though these tubes require small excitation, excessive crystal excitation is usually required to drive them to their full power output. As a result of this, the safe power output obtainable is only slightly more than can be obtained from a single pentode with the crystal normally excited.

Crystal-Controlled Test Oscillator

The circuit of a simple standard frequency oscillator with high harmonic output and two fundamental ranges, is shown in Fig. 8-11. This arrangement is particularly applicable to radio servicing where high accuracy and frequency stability are required. Of course, in any test oscillator where harmonics are to be used for the purpose of measurement and alignment, a high degree

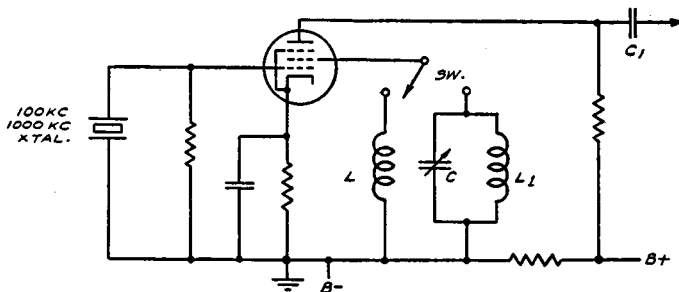


FIG. 8-11. A two-range, standard-frequency test oscillator whose accuracy and frequency stability are assured by the use of the crystal.

of accuracy and frequency stability are absolute necessities. Crystal control meets these demands where ordinary tuned circuits would not.

The circuit shown employs a specially ground Bliley crystal with two vibration modes, which will oscillate at either 100 kc

or 1000 kc. Harmonics of the 100-kc fundamental are good up to 4000 kc. Beyond this, the harmonics of the 1000-kc fundamental are used.

Coil L forms the tank circuit for the 100-kc range. For the 1000-kc range, switch SW is thrown to connect in the tank circuit composed of the coil L_1 and the trimmer condenser C .

It will be noted that the circuit is an electron-coupled affair, with the screen grid functioning as the oscillator plate and the plate circuit as the electron-coupled load. The plate is capacitively coupled to the external load through condenser C_1 .

This oscillator may be modulated by connecting B plus to the output of the power-supply rectifier rather than to the output of the power-supply filter so that the plate and screen voltages will have a comparatively high a-c ripple. The modulation frequency will be 60 cycles if a half-wave rectifier is used in the power supply, and 120 cycles if the rectifier is full-wave.

The RCA Piezo-Electric Calibrator

The schematic, Fig. 8-12, shows the circuit of an RCA crystal oscillator. In this model, a single crystal with two modes of oscillation supplies fundamental frequencies of 250 and 2000 kc. For unmodulated signals, the instrument operates from a 90-volt

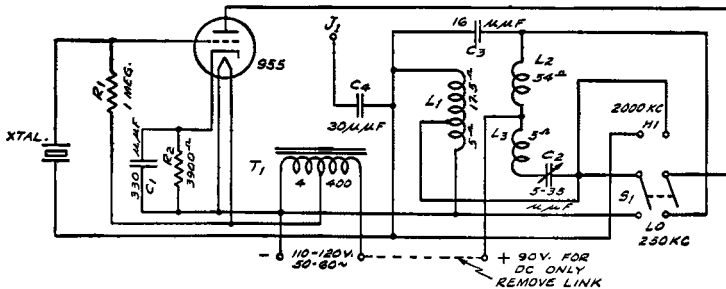


FIG. 8-12. An RCA crystal oscillator which supplies fundamental frequencies of 250 kc and 2000 kc.

d-c voltage source. For 60-cycle modulated signals the instrument is operated directly from the 110-120 volt a-c line supply,

the auto-transformer $T1$ being tapped to supply heater voltage for the 955.

The d.p.-d.t. switch $S1$ may be thrown to connect in either the 2000-kc tuned circuit composed of $L2$ and $L3$, or the 250-kc circuit formed by the tapped coil $L1$. The output signal is coupled to a pin jack, J_1 , by the condenser $C4$.

Power Output at Crystal Frequency

If high power output is required, it must be obtained by means of r-f power amplifiers. The crystal oscillator should be confined to a few watts output, and this power built up in successive amplifier stages. Naturally, the tube in each successive stage must have a power rating greater than that of the tube in the preceding stage.

An oscillator should be lightly loaded in any event, and it should be isolated from a changing load by means of a buffer stage. This holds true even in the case of low power-output requirements. In the latter instance the buffer need not be a tuned stage; it can be resistance-capacity coupled to the oscillator, as it is in certain types of r-f signal generators.

Output Wave Form

Recognizing that the crystal in a crystal-controlled oscillator is used primarily for the purpose of providing a stable output with respect to frequency, it is still necessary to realize that the various types of crystal oscillators and calibrators employed in connection with radio servicing, provide an output which is rich in harmonics. Naturally any one of these harmonics or even the fundamental when selected by means of a tuned circuit will be of undistorted character. But the fact that harmonics of accurate frequency are available is the more important consideration. This is true when this signal is desired as a test signal, as is described later on in this text, or when it is used as a calibrating signal for the checking of some other oscillator. This, too, is described later in this book.

With respect to the crystal-controlled systems shown in this chapter, it should be understood that the tuned circuit in the

plate circuit is normally tuned to resonance with the resonant frequency of the crystal. However, modified circuit arrangements are possible whereby in a single tube an output voltage is secured which is a harmonic frequency of the fundamental of the resonant frequency of the crystal. Elaborations of this circuit are not required at this time in view of the data given in a subsequent chapter.

Chapter IX

RELAXATION OSCILLATORS

IN THE absence of something more definite we define a relaxation oscillator as one in which the ratio of inductance to capacitance is abnormally low or high. In general but not always, relaxation oscillators are used to produce steep front waves which are rich in harmonics. The application of such oscillators is as a source of sweep voltage or linear time base in cathode-ray oscillography, television transmitters and receivers, and as a source of standard calibrating signals.

Saw-tooth Wave

The usual run of relaxation oscillators are utilized to produce a wave which approximates a saw-tooth form, which type of wave is needed as the time base voltage in cathode-ray work. The ideal form of wave is illustrated in Fig. 9-1. It will be seen that during the time interval OP the amplitude of the voltage gradually increases linearly from O to A . Then, in the short time interval PB , the voltage drops suddenly to zero, as shown by line AB . The time interval PB is called the *flyback* or *retrace time*. It is desirable that the flyback time be as brief as possible so that no visible trace is discernible on the fluorescent screen of the cathode-ray oscillograph.

The attainment of such a wave is not a simple matter. Approximations are easily secured, but special circuit arrangements are required so as to obtain the linear increase from A to O and the steep decrease from A to B . The ideal flyback time would be an approach to the zero interval, as indicated by the

dotted line AP . The usual arrangement for securing such saw-tooth waves involves the use of resistors and condensers.

Charge and Discharge of a Condenser

One type of relaxation oscillator operates by charging and discharging a condenser. Accordingly a few words concerning the manner in which a condenser charges and discharges will not be amiss.

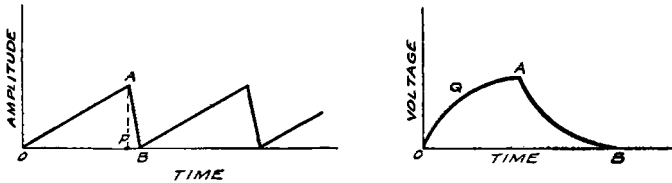


FIG. 9-1, left, FIG. 9-3, right. An ideal saw-tooth wave is shown at the left and the right-hand curve shows the variation of the voltage during the charge and discharge of a condenser with respect to time.

Consider the circuit shown in Fig. 9-2. By throwing the switch to position A , the condenser C will be charged by the current from the battery which flows through the resistor R ; and when the switch is thrown to position B , the condenser C will be discharged through the resistor R . Assuming that the condenser is initially discharged, then, when the switch is closed to make contact with A there will be a sudden surge of current through the

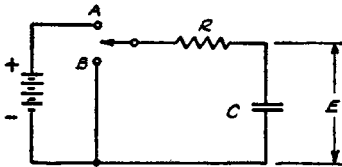


FIG. 9-2. When the switch is thrown to A , the condenser C is charged and when it is thrown to B , the condenser is discharged. How the charge on the condenser varies with time is shown above in Fig. 9-3.

resistor which causes the condenser to become charged. The current flow is rapid at first but gradually tapers off. The voltage E across the condenser likewise increases rapidly at first and then increases less rapidly up to some final value. This rise in condenser voltage is shown as OA in Fig. 9-3.

Now if the switch be thrown to position B , the condenser will rapidly discharge through the resistor, thereby resulting in a

rapid decrease in voltage across the condenser, as shown by AB in Fig. 9-3.

The time required to charge and discharge the condenser depends upon the value of the condenser and resistor. The greater the capacitance and resistance the longer the condenser takes to charge and discharge. Conversely, the smaller the capacitance and resistance the more rapidly the charge and discharge of the condenser takes place. Putting this in terms of frequency, the smaller the capacitance or resistance, the higher the frequency.

It is this charge and discharge of a condenser through a resistor that is the principle of operation of the relaxation oscillator. It will be observed, however, that neither the charge nor discharge shows a linear characteristic and that it is only a poor approximation to a saw-tooth wave. Methods for obtaining greater linearity will be discussed in the appropriate sections. The linearity is desired on the rise from O to A , as the shape of the retrace AB is unimportant providing that the retrace time is so short that it is not visible on the fluorescent screen.

Neon Tube Type

If a voltage is applied to a neon tube and this is gradually increased, the tube will pass negligible current until a voltage known as the *striking* or *ignition potential* is attained. At this point the gas within the tube will become ionized and current will flow through it. As the voltage is lowered below the striking potential the tube will continue passing current until it reaches its *extinction potential*, at which point current will no longer flow. The difference between the striking potential and the extinction potential is called the *overlap*.

The values of striking potential and extinction potential and the overlap vary in different types of neon tubes.

This property is utilized in the neon tube type of relaxation oscillator, which is depicted in Fig. 9-4. A source of voltage is applied to the neon tube N through the resistor R , the resistor serving to limit the current through the neon tube. The condenser C is shunted across the neon tube. The condenser is charged

through the resistor, but before the condenser is able to acquire its full charge, the striking potential of the neon tube is reached. The neon tube therefore becomes a conductor and the condenser discharges through it until its voltage falls below the extinction voltage of the neon tube. Since the neon tube is now no longer a conductor below the extinction voltage, the condenser begins to charge once more until its voltage again reaches the striking potential of the neon tube, whereupon the condenser again dis-

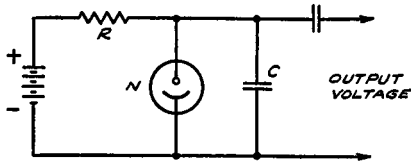
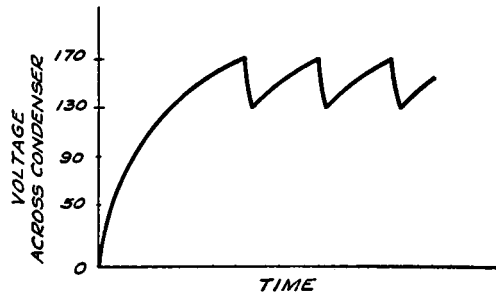


FIG. 9-4, left. A simple neon-tube relaxation oscillator, the action of which is explained above.

FIG. 9-5, right. The charge-discharge characteristic of a relaxation oscillator, such as that of Fig. 9-4.



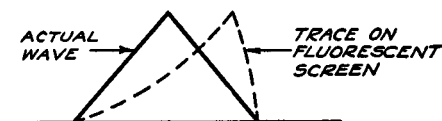
charges. This action continues and results in the typical condenser charge-discharge characteristic, as shown in Fig. 9-5. You will note in Fig. 9-5 that the charge and discharge of the condenser takes place within a range of voltages between 130 and 170 volts. These figures represent the striking and extinction voltages of one particular type of neon tube and are by no means intended as indicating the range of overlap for all neon tubes.

Distortion Due to Non-linearity

If you now compare the charge and discharge curve illustrated in Fig. 9-5 with that of Fig. 9-1, it will be evident that the waveform of the voltage developed by this relaxation oscillator is

not of linear character and if this voltage is used as a time base on, for example, one set of plates of a cathode-ray oscillograph, that a certain amount of distortion will manifest itself in the examined wave impressed on the other set of plates.

FIG. 9-6. Because of the non-linearity character of the neon-tube oscillator, it will distort even a pure wave when used as a time base in an oscillograph.



Thus, if a triangular voltage wave is placed on the *Y* or vertical plates, as shown by the solid curve in Fig. 9-6, and the non-linear output from the relaxation oscillator is used as the sweep voltage on the *X* or horizontal plates, the resulting trace on the fluorescent screen will have the shape of the dashed curve.

Improving Output Voltage Linearity

It is possible to improve the linearity of such a simple relaxation oscillator. This is done by utilizing only a portion of the charging curve of the condenser. For example, if you refer back to Fig. 9-3 you will note that the charging curve of the condenser is denoted by the line *OQA*. Also that the portion of this charging curve, *OQ*, is a closer approach to a linear curve than the portion between *Q* and *A*. Therefore by employing the curve *OQ* and causing the tube to strike at the point *Q* instead of *A*, we obtain a much improved shape of output voltage curve. This is accomplished by using a supply voltage which is high compared to the striking voltage of the neon tube.

The larger the supply voltage as compared to the striking voltage, the smaller the section of the charging curve that is used. There is an improvement in the linearity of the sweep, but, at the same time, there is a decrease in the amplitude of the output voltage. However, even for a very large supply voltage, the output is only approximately linear.

Still further improvement in linearity can be accomplished by the use of a saturated diode as explained in the following paragraphs.

Neon Tube and Saturated Diode

The rate at which the voltage across the condenser C in Fig. 9-4 builds up is dependent upon the rate at which the condenser is charged, and this in turn is dependent on the current through the resistor R to the condenser. If the fixed resistor R be replaced by one which limits the current to a constant value regardless of the voltage across the condenser, then the condenser voltage will build up linearly.

This can be accomplished by using a saturated diode as the resistor. The plate current in a diode increases as the plate voltage is increased until the space charge becomes neutralized, after which point the plate current remains almost constant and increases only slightly as the plate voltage is further increased. The point at which saturation occurs can be regulated by varying the temperature of the cathode. A circuit embodying a saturated diode D acting as a current regulator is shown in Fig. 9-7.

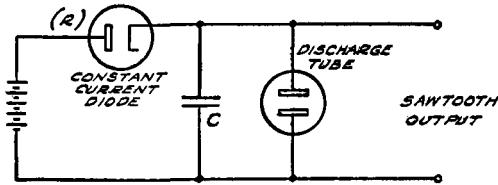


FIG. 9-7. In order to approach an output that is more nearly linear, a diode is used instead of a current-limiting resistor. Compare Fig. 9-4.

Unfortunately, the saturation characteristic of a diode is not linear for low values of plate voltage and, furthermore, the plate current varies considerably for small changes in the cathode temperature. These defects may be overcome by using a tetrode or a pentode as a current-regulating tube, instead of the diode. (see Fig. 9-9.)

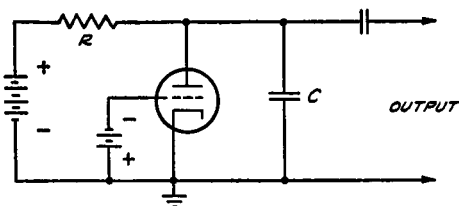
One of the main defects of two-element neon tube circuits is their limited voltage range. The maximum voltage amplitude obtainable is the difference between the striking and extinction potentials, which is approximately 40 volts.

Grid-controlled Discharge Tube

Much larger voltage amplitudes are obtainable by using a *grid-controlled discharge tube*, sometimes called a *thyatron*. This

type of tube is essentially a triode which contains some mercury vapor. The negative bias voltage applied to the grid determines the plate voltage which must be applied in order to cause ionization to occur; the more negative the grid voltage the greater the plate voltage needed to produce ionization. As soon as the plate attains the critical voltage necessary to ionize the gas the grid loses control, and current continues to flow through the tube until the plate voltage is sufficiently lowered.

FIG. 9-8. The use of a mercury-vapor filled triode, the thyratron, in a relaxation oscillator gives an output that is practically linear.



A simple relaxation oscillator circuit which utilizes a thyratron is depicted in Fig. 9-8. The condenser C charges until the tube ionizes, whereupon the condenser discharges through the thyratron until ionization ceases. The grid again gains control and the condenser begins to charge once more, and so on. The amplitude of the voltage output may be varied by changing the negative bias applied to the grid. However, in order to obtain a linear sweep voltage, it is necessary to use low values of grid bias. The greater the negative grid bias the greater the departure from linearity of the output voltage.

Circuit Using 885 Tube

The 885 is a grid-controlled gaseous tube. A typical circuit employing the 885 as a discharge tube, and a pentode as a current-limiting tube to obtain linearity of sweep voltage, is shown in Fig. 9-9.

A remote cut-off pentode is used to keep the condenser charging current constant, instead of the fixed resistor R of Fig. 9-8. The internal impedance of the pentode may be varied by changing either the control grid or the screen grid voltage; in this case the control grid bias. The striking voltage of the 885 gas triode,

and therefore the amplitude of the output voltage, is varied by altering the grid bias of the 885 tube. The frequency of the output voltage is varied by changing the capacitance of condenser *C*. This circuit is capable of producing a maximum frequency of about 10,000 to 20,000 cycles.

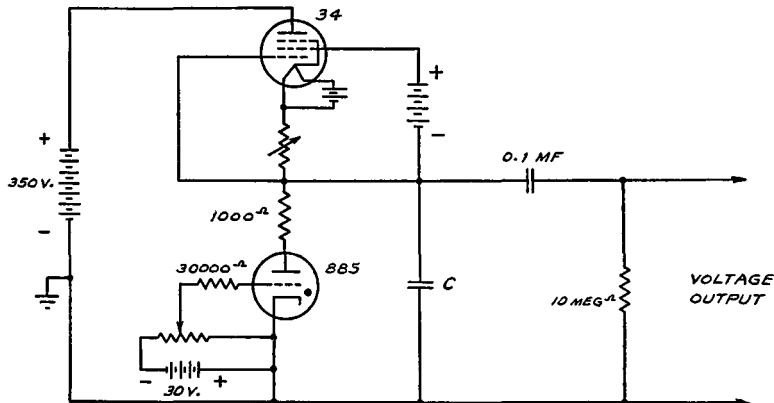


FIG. 9-9. A linear output is obtained from an oscillator using a grid-controlled gaseous discharge tube, the 885, and a type 34 pentode.

Another circuit using a grid-controlled gaseous triode oscillator operating with an amplifier is shown in Fig. 9-10. This circuit is a simplification of the actual arrangement of components used in a commercial cathode-ray oscillograph. Instead of a single condenser, as shown in the schematic at C_2 , in the complete unit a number of these capacities are used to control the frequency, selection being made by means of a switch.

Other Modifications

There are other ways and means of securing a substantially linear wave form from such relaxation oscillators. For example, one method consists of feeding this non-linear voltage into a non-linear amplifier and by operating over a portion of the grid voltage-plate current characteristic of this amplifier, securing a substantially linear output.

In some modern circuits, as those used in television, a conventional vacuum tube is used as the discharge tube instead of the pentode or instead of the gaseous tube. In these systems a signal is fed to the control grid in such a manner that at definite

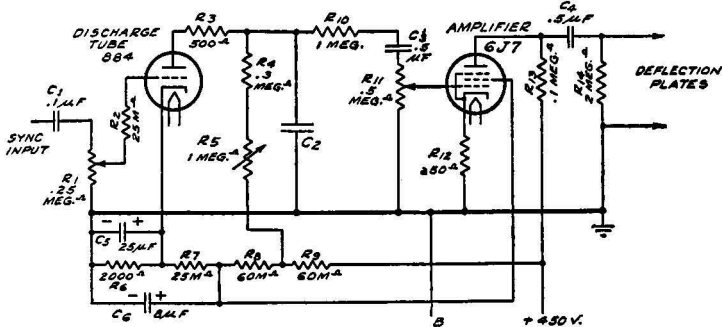
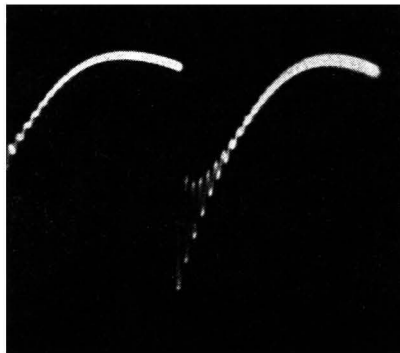


Fig. 9-10. A simplified schematic of the relaxation oscillator used in a commercial cathode-ray oscillograph.

intervals the grid of this tube is driven positive. This impulse is secured from a blocking oscillator or a vacuum-tube impulse generator. When the grid of the control tube is driven positive,

Fig. 9-10(a). The wave train of a blocking oscillator is shown.



the tube draws current and discharges a condenser which is connected across the output circuit of this tube. In circuits of this type it is possible to secure output voltages of frequencies up to about 100,000 cycles.

Still other arrangements are used in connection with high-frequency oscillographs whereby sweep voltages up to several megacycles are available.

Multivibrator

A *multivibrator*, which is a special type of relaxation oscillator, is essentially a two-stage resistance coupled amplifier whose output is fed back into its input circuit, as shown in Fig. 9-11. This is a rather unusual type of oscillator inasmuch as no inductances are employed. The theory of operation is quite different from that of other oscillators and will, therefore, be described in detail. The action may be divided into four separate parts.

Part I:—Due to some momentary unbalance in the circuit the plate current i_2 in the tube *II* increases. The voltage drop i_2R_2 across R_2 increases, in consequence of which the plate voltage E_2 of tube *II* is decreased. This lowering of plate voltage acts through condenser C_1 and makes the grid voltage e_1 of tube *I* more negative. The plate current i_1 of tube *I* decreases, and, due to amplification, this decrease in plate current i_1 is considerably greater than the originating plate current i_2 increase.

Since a smaller plate current i_1 flows, there is a reduced voltage drop i_1R_1 across the resistor R_1 and, accordingly, the plate voltage E_1 across tube *I* is increased. This increase in plate potential E_1 acts through condenser C_2 and makes the grid potential e_2 less negative. Consequently, there is a further increase in the plate current i_2 of tube *II* and this increase, due to amplification, is much larger than the initial increase in i_2 which started the action.

This process is cumulative, each further increase in i_2 resulting in a further decrease in i_1 so that the plate current i_1 of tube *I* is very rapidly reduced to zero because the grid voltage e_1 becomes so highly negative. At the same time, the plate current i_2 of tube *II* increases.

Part II:—Such a state of unbalance does not continue indefinitely and condenser C_1 discharges, thereby making the grid potential e_1 of tube *I* less negative. Accordingly, the plate current i_1 of tube *I* begins to flow again.

Part III: This slight increase in plate current i_1 of tube *I* results in an increased voltage drop $i_1 R_1$ across resistor R_1 , and the plate voltage E_1 across tube *I* accordingly decreases. This drop in plate voltage acts through condenser C_2 and causes the grid potential of e_2 of tube *II* to become more negative, and consequently the plate current i_2 of this tube decreases. The decreased voltage drop $i_2 R_2$ which therefore occurs across resistance R_2 causes the plate voltage E_2 of tube *II* to increase. This increased plate voltage E_2 acts through condenser C_1 and makes the grid voltage e_1 of tube *I* less negative. In consequence, the plate current i_1 of tube *I* increases and this increase in i_1 is, due to amplification, much greater than the initial increase in i_1 which started the operation described in Part III.

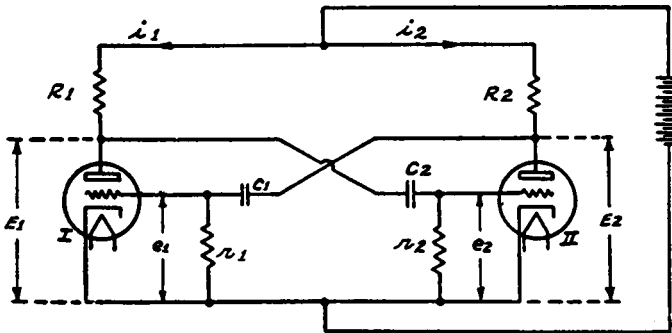


FIG. 9-11. The multivibrator circuit, which is a special form of the relaxation oscillator. Note the absence of inductances.

This process is cumulative, for each further increase in plate current i_1 of tube *I* causes a still further decrease in plate current i_2 of tube *II*. Consequently, the plate current i_2 of tube *II* falls to zero very rapidly due to the high negative grid bias e_2 , while the plate current i_1 of tube *I* simultaneously attains some finite value which is limited by the resistor R_1 .

Part IV: Condenser C_2 discharges, thereby making the grid voltage e_2 less negative. Consequently, the plate current i_2 of tube *II* begins to increase.

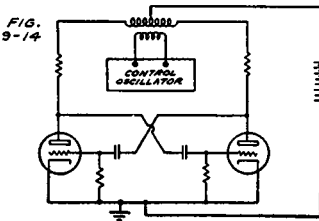
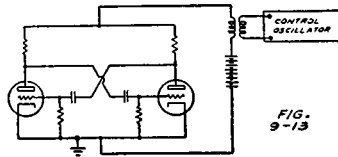
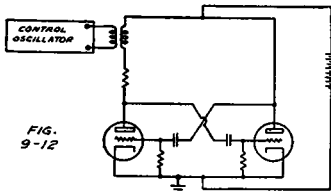
The operation is now back to the point at which it started in

Part I, and the sequence of events described in Parts I, II, III and IV continues as long as the circuit is operative.

The frequency of oscillation of a multivibrator depends mainly upon the discharge times of condensers C_1 and C_2 , which in turn is determined by their capacity and the circuit resistance, but is also slightly affected by the other circuit constants and operating voltages. It is not necessary that the discharge times of the two condensers be the same. The output of the multivibrator is extremely rich in harmonics. It is this feature which makes the multivibrator so useful.

Controlled Multivibrators

If an oscillating voltage is injected into a multivibrator circuit, it will be found that the multivibrator will become synchronously locked with the inserted alternating voltage. The multivibrator



FIGS. 9-12, 9-13, 9-14. Three different methods for injecting an external oscillating frequency into a multivibrator for locking purposes. The frequencies at which these circuits will oscillate with the methods illustrated are explained in the accompanying text.

will lock at some multiple or submultiple of the control frequency. Thus, a multivibrator whose fundamental frequency is 10 kilocycles will, if subjected to the control of a 100-kilocycle oscillator, produce a series of frequencies spaced 10 kilocycles apart throughout the frequency spectrum. Several methods of injecting this external frequency are shown in Figs. 9-12, 9-13 and 9-14.

When the control frequency is injected into only one of the plate circuits, as shown in Fig. 9-12, the multivibrator will oscillate at a frequency $\frac{1}{n}$ times that of the control frequency, where n is either an even or an odd number. If the circuit of Fig. 9-13 is used, where the control frequency is injected in the same phase into both plate circuits, the multivibrator will oscillate at a frequency $\frac{1}{n}$ times that of the control frequency, where n is an even number. On the other hand, if the control frequency is injected in opposite phase into each of the plate circuits, the multivibrator will oscillate at a frequency $\frac{1}{n}$ times that of the control frequency, where n is an odd number.

The multivibrator depicted in Figs. 9-13 and 9-14 will oscillate at $\frac{1}{n}$ times the control frequency, where n is even and odd, respectively, if the multivibrator is symmetrical. If dissymmetry is introduced, for example, by making the discharge time constants of the two multivibrator tubes different, then the multivibrator may lock with any submultiple of the control frequency, whether odd or even.



Figs. 9-15, 9-16, 9-17, left to right. Examples of multivibrator output. The wave shape depends on the circuit constants.

The Hallicrafters Crystal Calibrator

An application of the multivibrator in a small crystal calibrator is shown in Fig. 9-18, which represents the Model HT7 Frequency Standard circuit. The 6F6 functions as the crystal oscillator, operating at 1000 kc and 100 kc from a dual-type crystal. The output at these frequencies is fed to a 6L7 harmonic amplifier through the coupling condenser $C7$. An r-f choke in the plate circuit of the 6L7 forms its plate load and the output frequencies

are coupled by *C10* to a 4-range tapped coil which may be tuned to desired harmonics. When this output circuit is tuned to resonance with a harmonic, the output at that frequency is increased; further, its frequency is more definitely identified.

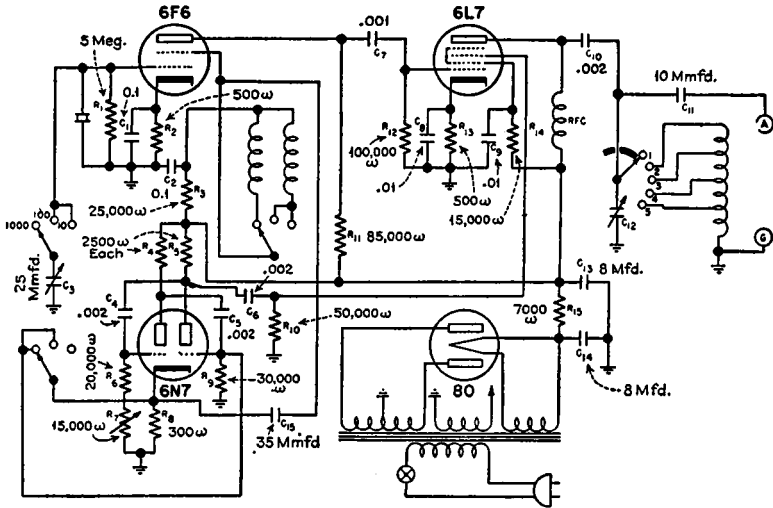


FIG. 9-18. A commercial use of a multivibrator circuit used in the Hallierafters crystal calibrator.

Harmonics at 10-kc intervals are secured by the 6N7 multivibrator, which is coupled to the screen circuit of the 6F6 through *C15*. When the crystal oscillator is functioning at 100 kc, the multivibrator 10th harmonic locks in with the crystal 100-kc fundamental. The output frequency of the multivibrator is thus stabilized at 10 kc and this signal is fed to the signal grid of the 6L7. Since the 100-kc signal is likewise present in the 6L7 circuit, an output at both frequencies is obtained. This enables frequency checking at 10-kc intervals, with a strong 100-kc fundamental serving as a marker.

Chapter X

MODULATION OF OSCILLATORS

WHEN a low-frequency current is superimposed upon a high-frequency current, either the amplitude or the frequency of the latter varies in accordance with the frequency of the former.* An example is the audio frequencies set up in the microphone circuit of a broadcast station, which are superimposed upon the radio-frequency carrier currents flowing in the transmitter's antenna circuit. These audio currents are said to *modulate* the carrier—to make its amplitude or its frequency increase or decrease in accordance with the variations of the *modulating* current. This process of superimposing a current of one frequency upon another of some other frequency is called *modulation*.

In just the same way the output of a radio-frequency oscillator can be modulated by varying its amplitude or its frequency in accordance with some low-frequency current. The commonest form is, of course, amplitude modulation, employed by practically all broadcast stations and as the oscillators used for checking the receivers that pickup signals from these stations must have modulated output of the same type, most of the test oscillators and signal generators you use are equipped with means for supplying an amplitude modulated signal. Other oscillators are available whose output is frequency modulated, but first let us see just exactly what amplitude modulation entails.

*Phase modulation is not considered in this book because it is not a commonplace method.

Amplitude Modulation

In amplitude modulation the carrier is kept constant in *frequency*, while its *amplitude* is varied in accordance with the audio-frequency modulation. Thus, in Fig. 10-1, since there is no audio modulation from *A* to *B*, the carrier has a constant amplitude. In the interval from *B* to *G* an audio-frequency modulation is superimposed upon the carrier frequency and the carrier amplitude varies in accordance with this impressed modulation frequency. The carrier frequency, however, remains unaltered.

Consider the interval from *B* to *G*. The audio modulating frequency has one-half the amplitude of the carrier. For convenience, it will be assumed that the audio modulating signal and the carrier frequency are in phase at each of the lettered points. Then, at *B* the audio signal has zero amplitude and the carrier therefore possesses an amplitude equal to its unmodulated state. At *C* the audio modulation has its maximum positive amplitude and consequently the amplitude of the carrier becomes greater. At *D* the amplitude of the audio wave is again zero and, accordingly, the carrier amplitude is again that of its unmodulated state. At *E* the audio modulation assumes its maximum negative amplitude, which results in a corresponding decrease in the carrier amplitude. At *F* the audio modulation amplitude is zero again, in consequence of which the carrier once more possesses its original unmodulated amplitude. The interval from *F* to *G* repeats the cyclic changes which occurred from *B* to *F*. From *G* to *H* there is no audio modulation and the carrier is seen to have a constant amplitude. Between *B* and *G* the carrier is 50 percent modulated.

In the interval *H* to *N* the audio signal has the same amplitude as the carrier. Since, at point *H* the audio modulating signal has zero amplitude, the carrier height will be that of its unmodulated condition. At *J*, when the audio wave has its maximum positive value, the carrier amplitude will be twice as great as when it is unmodulated. At *K* the amplitude of the audio wave is zero and the carrier width is again that of its unmodulated state. Point *L* represents the maximum negative amplitude of the audio frequency and, since the audio and carrier amplitudes are equal, the carrier amplitude will be reduced to

zero. At *M* the audio wave is again zero and the carrier amplitude assumes that of its unmodulated condition. The interval from *M* to *N* is a repetition of that which occurred from *H* to *M*. In the interval *N* to *P* the carrier is unmodulated. Between *H* and *N* the carrier is 100 percent modulated.

Now assume that the audio amplitude is greater than that of the unmodulated carrier, as depicted in the interval *P* to *W*. At point *P* the carrier possesses its unmodulated amplitude since the

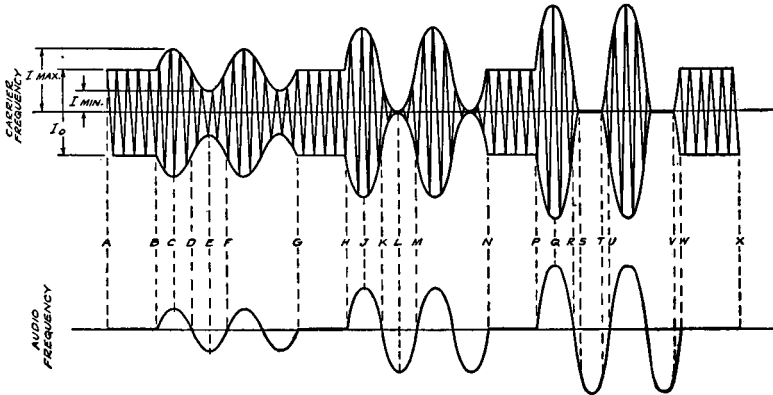


FIG. 10-1. The audio-frequency sine waves of different amplitudes are the modulating waves that cause the amplitude of the high-frequency carrier to vary as shown above.

audio wave at this point is zero. At *Q* the audio wave is at its maximum in the positive direction and the carrier amplitude is accordingly a maximum. At *R* the audio amplitude is zero and that of the carrier is that of the unmodulated state. At *S* the amplitude of the carrier is reduced to zero since the amplitude of the negative half cycle of the audio wave is equal to the amplitude of the carrier. Since the remainder of the negative half cycle of the audio wave exceeds the amplitude of the carrier up to point *T*, there will be no carrier in the interval from *S* to *T*. In the interval *T* to *U* the audio modulation is less than the carrier amplitude and the carrier reappears. From *U* to *W* there is a repetition of the preceding cycle from *P* to *U*.

Just as the interval *B* to *G* was representative of *partial modulation*, and interval *H* to *N* of *complete modulation*, so also the

interval P to W is illustrative of *over-modulation*. In the case of partial and complete modulation, the envelope of the carrier is a faithful replica of the modulating audio frequency. Serious distortion results when the carrier is over-modulated for the modulation envelope of the carrier is no longer a faithful reproduction of the modulating audio frequency, the carrier even being broken up into a series of discrete groups.

Percentage Modulation

According to I.R.E. standards, the percentage modulation can be expressed by the following equation:

$$\% M = \frac{1}{2} \left(\frac{I_{max} - I_{min}}{I_o} \right) \times 100.$$

Where I_{max} = the maximum value of the modulated current

I_{min} = the minimum value of the modulated current

I_o = the average value of the unmodulated current

M = the modulation factor

Let us assume the values of currents in some cases are as follows:

$$I_{max} = 75, I_{min} = 25, \text{ and } I_o = 50$$

Then substituting these values in the above equation, we have

$$\begin{aligned} \% M &= \frac{1}{2} \left(\frac{75 - 25}{50} \right) \times 100 \\ &= \frac{1}{2} \left(\frac{50}{50} \right) \times 100 = 50\% \text{ modulation} \end{aligned}$$

Relation Between Carrier and Sidebands

When the amplitude of a carrier frequency is modulated with a single pure audio frequency, *three* frequencies are produced: (1) the original carrier frequency, (2) a frequency which equals the *sum* of the carrier and the modulating frequencies, and (3) a frequency which is equal to the *difference* between the carrier and the modulating frequencies. These are called the carrier, the upper side frequency and the lower side frequency respectively.

For example, if a 1,000,000-cycle carrier is modulated with a pure 1000-cycle signal, three frequencies will then result: the original 1,000,000-cycle carrier, the upper side frequency of 1,001,000 cycles, and the lower side frequency of 999,000 cycles.

Although in actual practice a carrier is sometimes modulated with a pure single frequency tone, such as in test oscillators, it is much more common for the carrier to be simultaneously modulated with a number of frequencies, that is, with a band of frequencies. The sum and difference frequencies are then called the *upper sideband* and the *lower sideband*, respectively.

Practical Applications

Before describing the various methods of modulation, it might be well to point out that any given arrangement may be applicable to a number of different services. For instance, screen modulation is often used in radiophone transmitters where the modulated r-f power is large, and also in some wireless record players where the modulated r-f power is small. As another instance, suppressor modulation finds wide use in radiophone transmitters where compactness is an essential and a small amount of distortion can be tolerated; and it is also used as a means of modulation in some test oscillators. Moreover, a large percentage of the modulation methods to be described are to be found in the oscillator-mixer circuits of superheterodyne receivers, although we do not as a rule refer to these circuits as modulating systems.

In practically all methods, the oscillator tube can be modulated directly, such as it is in many test oscillators, wireless record players and similar low-power equipment, but in high-power units such as broadcast and radiophone transmitters, where distortion cannot be tolerated and frequency stability is an essential, it is not practical to modulate the oscillator directly. It is necessary in these cases to divorce the oscillator and modulator completely and, instead, modulate the r.f. in an amplifier stage.

Plate Modulation

Perhaps the most famous of all modulation systems is the one invented by R. A. Heising, and used principally in broadcast and

amateur 'phone transmitters. The arrangement is shown in Fig. 10-2.

The oscillator circuit may be of any type, but for illustrative purposes the shunt-fed Hartley is shown. The oscillator and modulator tubes have a common plate supply which passes through the choke coil L . The reactance of this choke for all audio frequencies is high compared to the internal plate resistance of the oscillator and modulator tubes. The modulator tube functions as an a-f power amplifier with the plate circuit of the oscillator tube as its load.

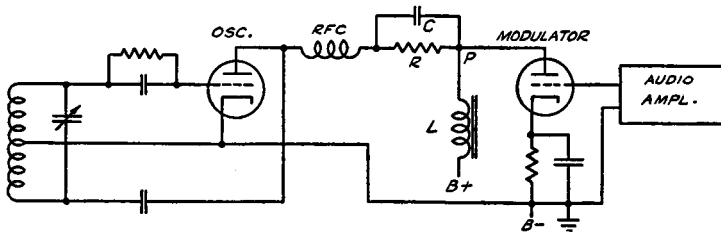


FIG. 10-2. The Heising modulator circuit working into a shunt-fed Hartley oscillator.

As the grid of the modulator tube is made more negative and less negative by the a-f signal voltage, the plate current of the modulator tube alternately decreases and increases. This results in the decay and the building up of the magnetic field about the choke coil which, in turn, results in an increase and decrease in the reactive voltage across the coil. This reactive voltage adds to and subtracts from the fixed plate-voltage supply so that a varying potential appears at point P , and therefore on the plate of the oscillator tube. Since the amplitude of the r.f. generated by an oscillator is proportional to the applied plate voltage, the radio-frequency power in this case is modulated in accordance with the audio-frequency input.

The function of the resistance R is to operate the oscillator tube at a lower plate voltage than the modulator tube in order to obtain a high percentage of modulation without distortion. The condenser C shunting this resistance serves to bypass the audio frequencies. The radio-frequency choke RFC in the plate circuit of the oscillator tube prevents the r.f. from entering the modulator tube circuit.

Since a varying potential on the plate of an oscillator tube results not only in a variation of the amplitude of the r.f. but also to some extent of its frequency, it is necessary, where frequency stability and an absence of frequency and phase modulation are desired, that modulation be accomplished in the plate circuit of an r-f amplifier (in the case of plate modulation) to which the oscillator is coupled directly or through intermediate buffer-amplifiers. This could readily be done in the case of Fig. 10-2 by having the modulator tube feed an r-f amplifier in exactly the same way as it is shown feeding the oscillator.

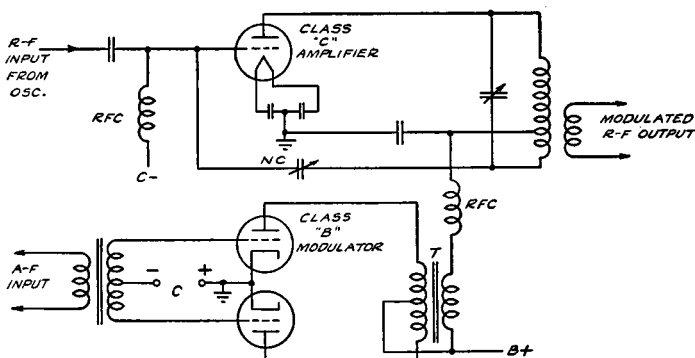


FIG. 10-3. If large amounts of r-f power have to be handled, the Heising Class B modulator is one of the most practical circuits. Note that in this schematic, the oscillator circuit is not shown.

An arrangement more generally used is shown in Fig. 10-3. This is also the Heising constant-current system except that a transformer *T* is used rather than a choke coil. Since the r-f amplifier is independent of the oscillator frequency, a variation in the plate voltage of the amplifier can vary only the amplitude of the r.f. The Heising system is capable of 100 percent modulation without amplitude distortion, but the modulating power required is 50 percent of the d-c power input to the r-f amplifier. A Class B modulator is therefore the most practical, if large r-f power is to be handled.

Grid Modulation

Grid modulation is a method whereby the grid voltage of an oscillator or amplifier tube is made to vary in accordance with the modulating power.

As may be seen in Fig. 10-4, the a-f modulating power is fed into the primary of the transformer *T*, the secondary of which is in series with the fixed grid bias voltage. A condenser *C* is shunted across the transformer secondary to bypass the r.f. The a.f. in the primary of the modulation transformer induces a volt-

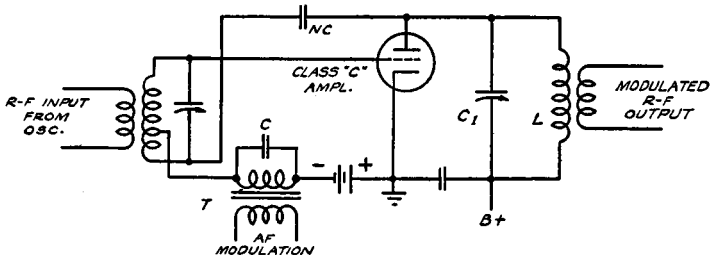


FIG. 10-4. A circuit employing grid modulation. Note that the modulating a-f voltage is introduced into the grid circuit of the amplifier tube from the secondary of *T*, which is in series with the *C* battery.

age in the secondary which alternately adds to and subtracts from the grid bias voltage. This results in alternate decreases and increases in the plate current of the r-f amplifier (or oscil-

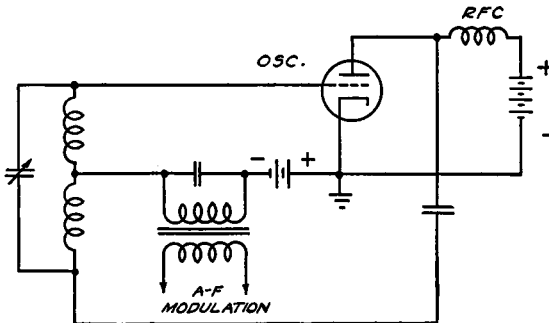


FIG. 10-5. How grid modulation is applied directly to a low-powered oscillator.

lator) and the output carrier is therefore modulated in accordance with the a-f input from the modulator.

Grid modulation as applied directly to a low-powered oscillator is shown in Fig. 10-5. A single modulator tube operated Class A would meet most requirements with a circuit of this sort.

The principal advantage of grid modulation is the relatively small amount of modulating power required, Class A push-pull audio amplifiers being sufficient for modulating comparatively high r-f power. However, greater power output can be obtained with plate modulation.

Cathode Modulation

Cathode modulation as applied to an r-f power amplifier stage is shown in Fig. 10-6. The secondary of the modulation transformer is connected in series with the cathode so that a voltage across this winding will also appear on the cathode. Since the cathode is common to both the grid and plate return circuits, any

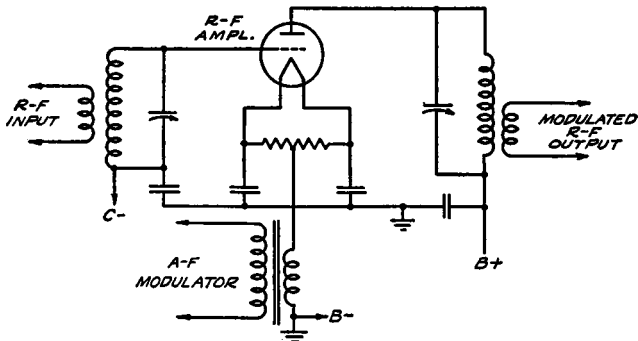


FIG. 10-6. Here the secondary of the modulation transformer is connected in series with the cathode of the amplifier tube; thus any voltage variation appearing in the cathode will affect the grid and the plate.

voltage variation in the cathode circuit will have an effect on the grid as well as the plate. Cathode modulation is therefore a combination of grid and plate modulation.

When the a-f modulation voltage on the cathode is positive, the grid of the amplifier tube is negative with respect to the cathode, just as in any cathode-biased grid circuit. Since the grid is already negative by virtue of the fixed bias, it becomes

more negative than normal when the cathode is positive, and therefore reduces the r-f output of the tube. At the same time the positive voltage on the cathode opposes the positive d-c plate voltage, thereby lowering it. This also reduces the r-f output. Since the positive cathode voltage lowers the r-f output by its effect on both grid and plate circuits, the two actions are in phase.

When the a-f voltage on the cathode is negative, the grid is made less negative with respect to cathode and the r-f output increases. Similarly, the negative voltage on the cathode (with respect to ground) increases the total plate voltage which results in a further increase in r-f output.

This system is capable of 100 percent modulation and requires a modulating power slightly more than for grid modulation—about 10 percent of the d-c plate input power. On the other hand, the r-f output obtained is much greater than that from grid modulation.

Screen Modulation

Since the plate current of a screen-grid tube depends not only upon the control-grid voltage, but also upon the screen voltage, it is possible to apply the modulating voltage to the screen grid, as shown in Fig. 10-7. The modulating voltage induced in the

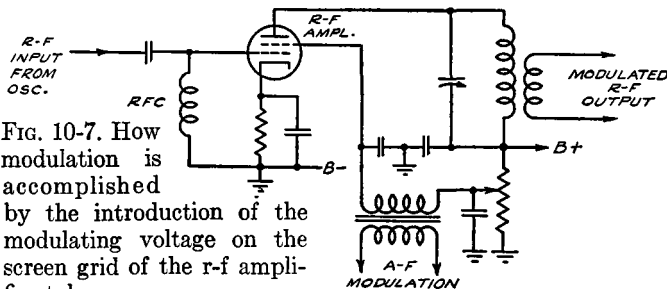


FIG. 10-7. How modulation is accomplished by the introduction of the modulating voltage on the screen grid of the r-f amplifier tube.

secondary of the modulation transformer alternately adds to and subtracts from the constant screen-grid voltage supply. These changes in screen voltage result in a corresponding audio-frequency variation of the plate current, thereby resulting in a modulated carrier output.

A screen-modulated oscillator circuit is shown in Fig. 10-10.

The percentage modulation possible without amplitude distortion is limited to about 60 percent, but only a small amount of modulating power is required for screen-modulated circuits as compared with plate-modulated systems.

Suppressor Modulation

By injecting the modulating voltage into the suppressor-grid circuit of a pentode, as shown in Fig. 10-8, improved modulation capability is obtained. With the proper tubes, modulation up to 100 percent can be reached with only a moderate amount of distortion.

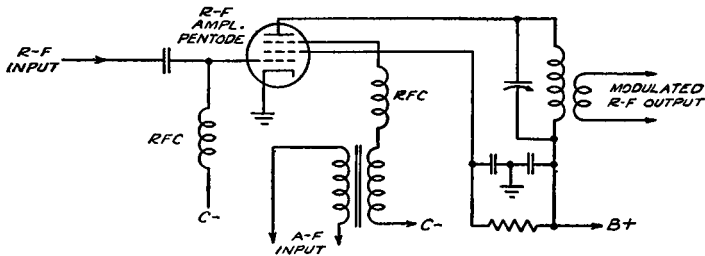


FIG. 10-8. If the modulating voltage is introduced in the suppressor grid, 100 per cent modulation can be reached with moderate distortion.

The suppressor grid is negatively biased. The modulating voltage swings this grid more negative on negative peaks and less negative on positive peaks. Since the r-f output of a pentode can be effectively altered by changing the suppressor-grid bias, the modulating power increases and decreases the r-f output in accordance with the a-f signal.

In this respect the suppressor grid acts similarly to the control grid in grid modulation. But in grid modulation both the r-f energy and the modulating voltage are applied to the control grid, whereas in suppressor modulation the r.f. is applied to the control grid and the modulating voltage to the suppressor grid.

Self Modulation

If the resistance and the capacitance in an oscillator be made high as compared to the normal values used in the grid leak and condenser combination, then the oscillator can be made to modulate itself at an audio-frequency rate. This is due to the fact that as the tube oscillates at a radio-frequency rate a charge builds up on the grid condenser which results in the grid becoming more and more negative. Since the value of the grid leak resistance has been made very high, this negative charge cannot leak off the grid as quickly as it is built up. Consequently the grid becomes more and more negative until the plate current has been reduced to so low a value that the tube stops oscillating.

When the tube stops oscillating, the alternating voltage applied to the grid ceases and the charge on the grid leaks off through the grid leak. The grid then becomes less negative and consequently the plate current rises and the tube again starts oscillating. When the tube oscillates, the negative charge on the grid builds up and the cycle repeats.

The values of the grid leak and condenser are chosen so as to cause the starting and stopping of oscillations at an audio frequency rate, thereby resulting in an audio-frequency-modulated radio-frequency signal.

WIRELESS RECORD PLAYERS

The wireless record player is a good example of an amplitude-modulated device. It is a low-powered transmitter which provides a signal much in the same way as an ordinary broadcast transmitter. However, instead of the modulation being obtained from a studio microphone, the phonograph pickup converts the vibrations on the record into an audio signal and this voltage is used to modulate the oscillator contained in the wireless record player. The modulated signal is radiated to the receiver which is tuned to the frequency of the oscillator section and, as a result, the record is reproduced by the receiver in the conventional manner. Note that the *entire* receiver is used rather than just the audio section. Unlike the phonograph oscillators which preceded wireless record players, there is *no direct connection* to the receiver—hence the name “wireless” record player.

In general, the design of the radiating system in wireless record players is such that the effective range of pickup is approximately fifty feet. Since excessive radiation will cause interference in other radio receivers, the Federal Communications Commission has limited the amount of radiation to a value which is high enough to permit satisfactory operation of the record player in a typical installation, but which at the same time is not so high as to cause interference in neighboring receivers. In general, a long antenna should not be attached to the record player to supplement the built-in antenna, but where the signal input to the receiver is too small, the coupling between the record player and the receiver should be increased. This can be accomplished by running the antenna of the record player closer to the antenna lead-in of the receiver.

Circuit Designs

Most wireless record players use two tubes; one tube as a combination oscillator and modulator, and the other tube as the rectifier. The 6A7 and the 25Z5 (or their newer equivalents) are commonly used to perform these functions. In some cases a single combination tube, such as the 12A7, is used both as an oscillator-modulator and as a rectifier.

The power supply design for the most part follows conventional practice and is not unlike the power supplies used in midget receivers. In some cases a small power transformer is used, but in general the d-c voltage for the oscillator-modulator tube is obtained by means of a half-wave rectification of the line voltage. In the latter case, the voltage for the heaters and the pilot bulb are obtained through a line cord or ballast tube, although in some instances the heaters are in series with the turntable motor windings so that no ballast resistor is required.

Where the 6A7 tube is used as the oscillator-modulator, the first two grids are not used to form the oscillating circuit, as is the case in the pentagrid converter. Instead, the tuned circuit of the oscillator is connected to the signal grid (the #4 grid) and the feedback circuit is through the plate of the tube, which carries the feedback or tickler winding. Essentially, then, as far as the action of the tube as an oscillator is concerned, the tube may be

considered as an ordinary regenerative triode oscillator, in which the signal grid acts as the control grid and in which the feedback is supplied through the plate.

The unmodulated signal produced by this oscillator is modulated by means of the audio signal from the phonograph pickup. The action which takes place here is similar to that which takes place in the pentagrid converter. In the pentagrid converter, the local oscillator signal is produced by the first two grids of the tube, and the incoming signal (applied to the #4 grid) is modulated by means of this local oscillator signal so as to produce the intermediate frequency. In wireless record players, however, the signal to be modulated is the oscillator signal (produced as explained above by connecting a tuned circuit to the #4 grid) and the

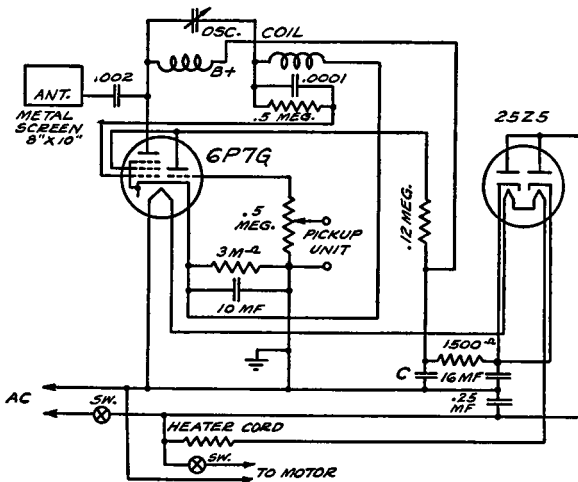


FIG. 10-9. In this schematic of a wireless record player, the 6P7G tube is the oscillator-modulator and the 25Z5 supplies the different rectified voltages.

modulation is accomplished by feeding the audio signal to the #1 grid. Both grid #2 and the screen grid (#3 and #5) are connected to B+.

As a general rule, a crystal pickup is used in wireless record players and, where the type 6A7 tube is used, the pickup is con-

nected to the #1 grid. The output of these pickups is of the order of 1 volt or more, so that with average circuit conditions the percentage of modulation is fairly low. In order to prevent distortion as a result of non-linearity of the modulation, some designs do not apply the full pickup voltage output to the grid, but use a high value of series resistance to form a voltage divider. Where a volume control is used in the record player, this control appears in the input circuit so that it is possible to control the level of the audio signal which is fed to the modulator grid.

The complete schematic diagram of a wireless record player is shown in Fig. 10-9. It employs a 25Z5 rectifier and a 6P7G as the modulator-oscillator. The antenna or radiator is an 8-by-10-inch copper screen tacked to the bottom of the cabinet. It is isolated from the oscillator high-voltage circuit by a .002- μ f mica condenser.

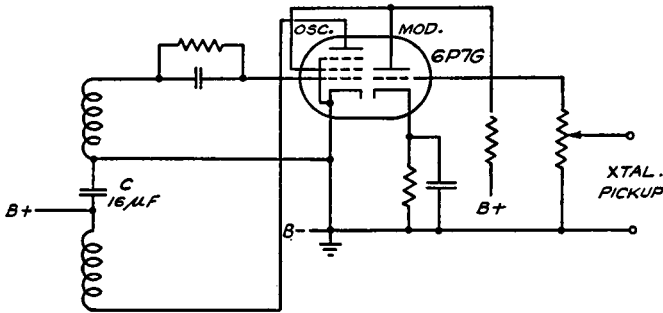


FIG. 10-10. A simplified schematic of Fig. 10-9. The pentode section of the 6P7G is the oscillator and the triode section acts as the modulator.

The modulator-oscillator is a unique arrangement which can be more readily followed from the simplified diagram of Fig. 10-10. The triode section of the 6P7G is used as the modulator and the pentode section as the oscillator.

The oscillator circuit is a series-fed Hartley with the last filter condenser *C* in the rectifier circuit used also as the r-f bypass. The modulator triode is *direct coupled* to the oscillator screen grid.

In this circuit, screen modulation limits modulation percentage to about 70 percent, after which a limiting action takes place due

to the screen-plate voltage relationship. Thus, distortion due to over-modulation is effectively prevented. As a result of this, the record-player volume control can be set at or near maximum and the receiver volume control set near minimum so as to get the best signal-to-noise ratio at the receiver. Usually, sufficient margin is left in the setting of the record-player volume control so that volume can be controlled at the player rather than at the receiver.

FREQUENCY MODULATION

Frequency modulation is quite often an undesirable by-product of amplitude modulation. The presence of frequency modulation in an amplitude-modulated transmitter is highly objectionable since the entire a-m system of transmission and reception is based on the use of a carrier of constant frequency. Frequency modulation in such a system is an indication that the carrier frequency of the transmitter is varying with the a-f modulating voltage. The resultant shift of the carrier from its assigned frequency causes sideband splatter and distortion.

Frequency modulation in an a-m system is caused by a reaction on the oscillator that generates the carrier frequency. Frequency modulation is almost sure to be present if the oscillator is modulated directly, for, in that case, any change in the plate voltage or the load conditions will also alter the frequency of the oscillator, as previously mentioned. For this reason it is customary to isolate the oscillator from the load by the use of buffer amplifier stages, and modulate the final r-f power amplifier stage which is independent of frequency.

Frequency Modulation Method

Frequency modulation is accomplished by varying the carrier frequency at a rate that is equal to the a-f modulating frequency and proportional to the modulating voltage. This method is employed in frequency-modulated (f.m.) transmitters as well as in "wobblers" and other test equipment principally used in conjunction with the cathode-ray oscillograph. By this method the frequency rather than the power of an oscillator is made to vary in accordance with the modulation voltage.

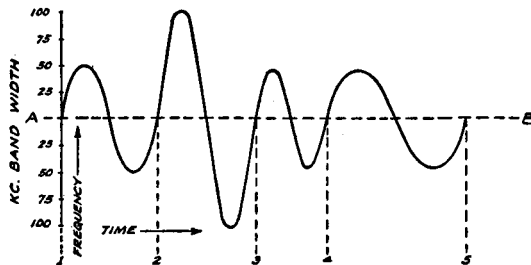
Frequency Modulation in Broadcasting

Until recently, all broadcast stations have used amplitude modulation for program transmission. In this method, as you know, the frequency of the carrier is maintained constant, usually by crystal control, while the amplitude of the carrier is varied by the audio modulation impressed upon it.

A new system of broadcast transmission employing frequency modulation is now being used by a growing number of new stations operating in the ultra-high-frequency band. This system was devised by Major E. H. Armstrong and differs from amplitude modulation in that the carrier *frequency* is varied by the impressed audio modulation, while the carrier *amplitude* is held constant.

There are two functions involved in both a.m. and f.m.; first, varying the carrier at an audio-frequency rate in accordance with the audio modulations, and, second, varying the percentage of modulation in accordance with the variations in amplitude of the audio modulation. In amplitude modulation this is accomplished by adding sidebands to a carrier of constant frequency and by varying the output power. In frequency modulation this is accomplished by varying the degree of frequency swing above and below the mean carrier frequency at the time rate of the modulating frequency.

FIG. 10-11. In a frequency modulated system, the frequency of the audio signal is shown along the horizontal axis and the amplitude of the audio signal is shown along the vertical axis.



The f-m method of transmission is illustrated in Fig. 10-11. Note that the bandwidth is 200 kc rather than 10 kc as it is in standard broadcast practice. The actual assigned carrier frequency is represented by the dashed line A-B. There is a 100-kc band space above and below this point.

Now, amplitude changes in this figure are not represented by actual changes in output power; the carrier power of an f-m station remains constant. Rather, amplitude changes are represented by the extent of the swing in *frequency* above and below line *A-B*. Thus, the first audio-frequency cycle, from 1 to 2, may be said to represent 50-percent modulation, while the second cycle, from 2 to 3, may be said to be the equivalent of 100-percent modulation.

On the other hand, the variations in the audio-frequency components of the carrier are represented by the time rate change in frequency, from left to right in the figure; that is, the horizontal displacement of the audio-frequency cycles in point of time. Thus, the a-f cycle from 3 to 4 is higher in rate of frequency than the cycle from 4 to 5. But both of these cycles are of the same frequency "amplitude," whereas the first two cycles have different frequency "amplitudes" but are of the same (audio) frequency.

Expressed in another manner the frequency of the audio signal is shown along the horizontal axis, whereas the loudness or amplitude of the audio signal is shown along the vertical axis. For example, the cycle between 1-2 and 2-3 might well be a 500-cycle voltage with that between 2-3 being twice as loud as that between 1-2. On the other hand, the signal shown between 3-4 is just slightly weaker than that between 1-2, but is of a higher audio frequency. In turn, the signal between 4-5 is as loud as that between 3-4, but is of a lower frequency than the signal 1-2, 2-3, or 3-4.

The simplest form of frequency modulation system is shown in Fig. 10-12. A condenser microphone *C* is connected across the coil *L* and the condenser C_1 comprising the tank circuit of the oscillator. Sound waves impressed on the diaphragm of the microphone will cause it to vibrate and thus vary the capacity at an audio-frequency rate. Since the capacity of the microphone is in shunt with the tank circuit, any variations in this capacity will alter the frequency of the oscillator at an audio-frequency rate. The variations will be too small for practical purposes, but the arrangement serves to illustrate the basic method of frequency modulation. The louder the audio signal fed into the microphone, the greater the swing in frequency and the higher the frequency of the signal into the microphone, the more rapid the frequency change of the oscillator.

In order to receive f-m signals, it is necessary to provide a means of translating the frequency swings above and below the carrier into actual variations in amplitude. This is accomplished by using a detector similar to the discriminator circuit employed in superheterodynes having automatic frequency control. With this arrangement, described elsewhere, any shift in frequency either side of the i-f alignment frequency upsets the balance of

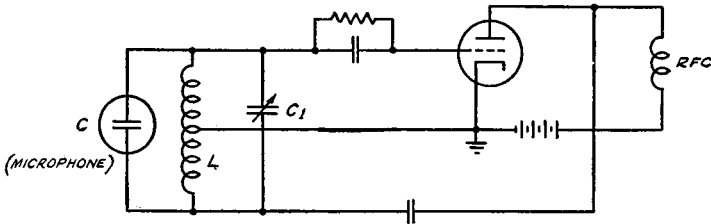


FIG. 10-12. In this simple form of frequency-modulation system, the microphone C is in shunt with the tank circuit, L and C_1 , and modulates the frequency of the oscillator at an audio rate.

the discriminator circuit and produces a voltage difference in the output circuit. The amplitude of the voltage so produced is proportional to the frequency shift. Since variations in amplitude in an f-m transmitter are radiated in the form of frequency shifts, the discriminator circuit is able to translate these shifts into actual variations in amplitude.

Frequency Modulation in Servicing Equipment

In order to examine the resonance curves of tuned circuits on the screen of a cathode-ray oscillograph, it is necessary to employ some form of frequency-modulated oscillator for the purpose of spreading the action into two dimensions by "scanning" or "sweeping" the entire band of frequencies through the tuned circuit which is to be checked.

Motor Driven Frequency Modulated Oscillators

One method of obtaining a frequency-modulated output from a service test oscillator or signal generator is by means of a

motor-driven tuning condenser. A constant voltage is secured from an oscillator at a fixed frequency, determined by the setting of the oscillator tuning condenser. This fixed frequency is really variable, since the tuning condenser is a variable, but, for the sake of clarity, we will classify it as a fixed frequency, because at any one time it is fixed as the setting of the tuning condenser remains untouched. This signal voltage at a certain frequency may be secured, as we know, from a single oscillator, or may be the difference frequency or beat frequency between two oscillators, one of which is fixed tuned. Expressed in another manner, the test voltage may be the voltage at the beat frequency resulting from the heterodyning of one oscillator with another.

The design of these oscillators is such that the output voltage, while not substantially the same at all frequencies, is practically constant for a small band of frequencies, each side of any one setting. To secure such a test signal voltage over a band of frequencies by means of a motor-driven frequency modulator, a small motor-driven variable condenser is connected or arranged in the tuned circuit. This condenser is continuously rotated by a small, fractional horse-power motor. As it rotates, the capacity naturally varies from minimum to maximum and back from maximum to minimum. Since this condenser is electrically connected to the tuning condenser in the oscillator circuit, each variation in capacity varies the total tuning capacity, hence the frequency of the circuit. While the frequency modulator condenser rotates, the frequency of the oscillator is being continually varied. The width of the frequency band thus covered is a function of the circuit capacity change due to the rotation of the frequency modulator condenser and the change in frequency of the oscillator output voltage, as a result of the change in the L-C circuit.

With suitable calibration it is possible to adjust the oscillator condenser to generate a fixed frequency of say 260 kc and to cause, as a result of the capacity change in the frequency modulator condenser, a continuous variation in frequency from 250 kc to 270 kc, thus providing a band width of 20 kc, or a 10-kilocycle variation each side of the mean or periodic frequency. (The mean frequency, which is the setting of the oscillator and which also is the resonant frequency of the circuit under test,

often is referred to as the periodic frequency.) As shown in Fig. 10-13, which is a simple version of what has been said, the continuously rotating variable condenser is that contained in the frequency modulator. The condenser in the oscillator, which is of the continuously variable type, is not geared to a motor or any other form of drive. It is tuned by hand to whatever periodic frequency is desired and is operated in exactly the same manner as if there were no frequency modulator unit and a single frequency signal was desired.

It is possible to arrange the frequency modulator condenser with several sections, so that two ranges of capacity are available, thereby providing two band widths for any one periodic frequency.

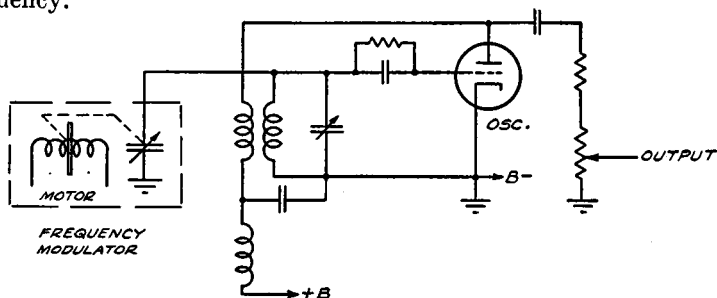


FIG. 10-13. In this frequency-modulated oscillator, the variable condenser on the motor shaft, indicated by the dotted lines, varies the frequency of the system 10 kc above and below the periodic frequency of the system.

The exact band width desired depends upon the nature of the tuned circuit or stage being investigated. A controlling influence is the frequency band width rating of the circuit under investigation. Fortunately, the majority of tuned circuits or stages associated with such devices, are rated at from 5 kilocycles each side of the periodic or resonant frequency to about 10 kilocycles each side of the resonant frequency, so that by arranging a system whereby these two band widths are available, complete coverage is secured, with possibly two adjustments of the frequency modulator capacity range. Whatever the band width of the frequency-modulated system, it should always be at least 25 percent

greater than the rated band width or band pass of the tuned circuit or system being investigated. This is desired so that the base of the resonance curve will be established as a reference point, even when circuit conditions are such that tuning is broader than normal.

In contrast to the single oscillator type of frequency-modulated source of test signal, there is in use the dual oscillator system, operated with a motor-driven frequency modulator condenser, whereby a constant band width is secured. Operation of the system can best be comprehended by referring to the elementary circuit shown in Fig. 10-14 and the following explanation.

The circuit consists of two oscillators, a mixer tube and the motor-driven frequency modulator. One of these oscillators, say Osc. 1, is fixed tuned at some frequency, say 700 kc. Connected across the tuning condenser of this unit is the motor-driven frequency-modulator condenser. The capacity of the rotating condenser is so designed that when in operation, the mean frequency of the oscillator is 700 kc and the frequency limits of the output are 690 kc and 710 kc, a band width of 20 kc or 10 kc each side of the mean frequency. In effect this oscillator is a frequency-modulated r-f oscillator, fixed tuned to 700 kc and frequency modulated over a 20 kc band.

Oscillator 2 is the continuously variable frequency unit. This is the one which employs the wave changing mechanism, so that it may be tuned over the entire i-f, broadcast and high-frequency or short-wave spectrum. The tuning condenser in this unit is operated by hand, just as if it were a conventional oscillator without even the remotest association with any frequency modulating condenser. As is shown in the diagram, the output of the oscillator is fed to the mixer tube through a suitable attenuating system. The 6A7 in this case is used as a mixer tube only. One of the oscillators feeds into one of the control grids, grid number 1, and the other oscillator feeds into the other control grid, grid number 4.

Let us say that oscillator 2, the continuously variable unit, is adjusted to 960 kc. When this is done, two signals are fed into the mixer tube. One of them is of 960 kc and the other is the 700 kc signal frequency modulated over a 20 kc band. These two voltages are mixed and as a result of rectification, there is present

in the plate circuit of the 6A7, a voltage which varies in frequency from 250 kc to 270 kc, or a 260 kc signal, the beat frequency, which is frequency modulated over the 20 kc band, or 10 kc each side of the mean frequency. This frequency modulated signal is produced because one of the signals fed into the mixer is modulated plus and minus 10 kc. We might mention that the

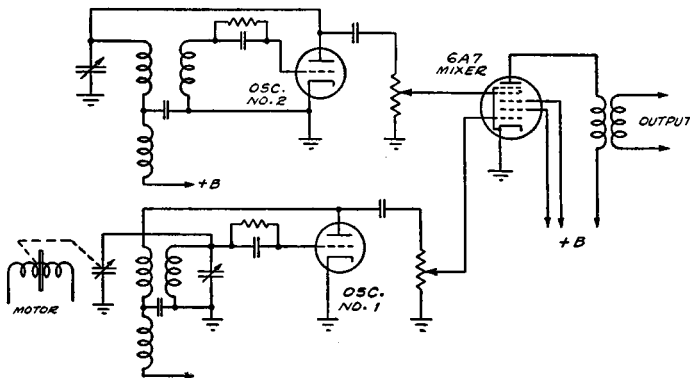


FIG. 10-14. The dual oscillator system for the production of frequency-modulated signals. Oscillator No. 1 feeds a continuously varying signal into the 6A7 mixer, which also receives a fixed frequency signal from No. 2. See explanation in accompanying text.

selection of a 20 kc band width for the frequency modulation of the fixed oscillator, is purely illustrative. It can just as readily be 10 kc or 30 kc.

Impulse from Motor-Driven, Rotating-Condenser Frequency Modulator

The motor-driven rotating condensers, shown in Figs. 10-13 and 10-14, are of two varieties, being used in connection with oscillographs. Means are provided in both instances for the generation of an impulse voltage, which is intended to control the horizontal sweep oscillator in the oscillograph so as to lock the image properly on the screen. In one variety of such a rotating condenser, shown in Fig. 10-15, the shaft of the condenser is coupled to the armature of a small impulse generator. During operation this generator develops an impulse voltage over each

180° of the rotation of the armature. This unit is the RCA model TMV 128A.

In contrast to this type, another is shown in Fig. 10-16, where instead of an impulse generator, a condenser-resistor arrangement

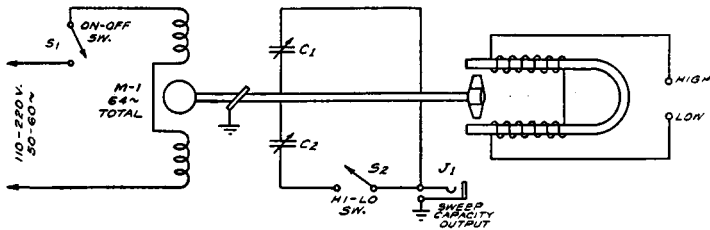


FIG. 10-15. Schematic of the RCA model TMV 128A motor-driven frequency modulator.

is employed. Attached to the rotor of the condenser is a short-circuiting switch consisting of a rotating and stationary contact. The rotating contact makes electrical connection with the stationary contact during 180° of the 360° travel of the rotor of the rotating condenser. A high d-c voltage is fed to condenser C through resistor R , thus charging the condenser. During 180° of the rotation of the condenser, the charge built up across the condenser C is applied to the oscillograph as a pulse. During the other 180° when this condenser is shorted, no pulse is applied to the oscillograph. The result is that the motor-driven, rotating-condenser type of frequency modulator shown in Fig. 10-15 develops a double image on the oscillograph screen, whereas that shown in Fig. 10-16 develops a single image.

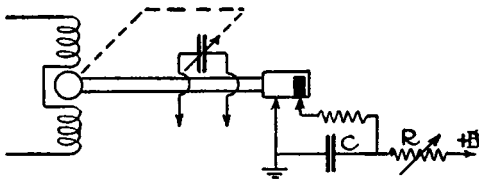


FIG. 10-16. Circuit of the Clough-Brengle frequency modulator, which develops a single image.

As a result of the fixed relation between the frequency modulating condenser capacity and the circuit capacity of the fixed frequency oscillator, the band width of the frequency-modulated

signal is constant, at whatever the design provides. Since the beat signal retains the modulation characteristics of the modulated signal, which is mixed with the sine-wave signal, the output beat voltage, irrespective of frequency, will retain the same modulation characteristics, which in this case means modulation at plus and minus 10 kc. This is how the band width of the frequency modulated output of this dual oscillator arrangement is constant at whatever the design provides.

Electrically Operated Frequency Modulated Oscillators

In contrast to the motor driven condenser type of frequency modulator, several units are in use, wherein frequency modulation is secured by means of an electrical circuit without rotating parts.

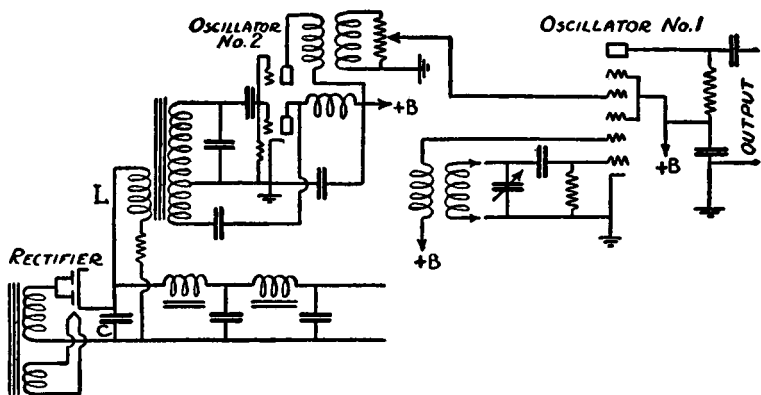


FIG. 10-17. Schematic of the Egert frequency-modulated oscillator, wherein the spreading of the frequency is accomplished by using the variation in permeability of the iron core in the No. 2 oscillator oscillation transformer.

One such system is known as the Egert method, illustrated above. Two oscillators are used to produce the constant band width frequency modulated signal, just as in the motor driven affair. An idea of how this electrically operated type of frequency modulator develops the modulated wave can be gathered from Fig. 10-17 and the following text. The circuit shown is basic and is offered solely for illustrative purposes.

Oscillator number 1 is the variable frequency unit, which covers the various bands. Oscillator number 2 is the fixed frequency oscillator, which is frequency modulated over a constant band width, say 20 kilocycles. Oscillator number 1 uses the oscillator portion of a 6A7 tube. The mixer or pentode portion of this same tube receives the frequency modulated output of oscillator number 2, so that the two signals are mixed and, as a result of the rectifying action within the tube, the plate circuit contains the difference frequency modulated plus and minus 10 kc.

If you examine the wiring diagram, you will also note two significant facts: First the presence of an iron core for the oscillator number 2 coils; second, the use of a rectifier and that a portion of the a-c voltage present across the input condenser is fed into the oscillation transformer used in oscillator number 2.

This iron core and the rectified a-c voltage serve to frequency modulate the output of the fixed frequency oscillator, by employing the variation in permeability of the iron core to change the frequency of the oscillations generated by oscillator number 2. The winding connected to the power supply furnishes a magnetizing current, which varies at the frequency of the a-c voltage across condenser *C*. Since the rectifier is a half-wave system, this magnetizing current varies 60 times per second or at whatever the frequency of the power supply may be. By suitable design of the iron-cored oscillation transformer, operation upon the straight portion of the permeability curve is secured. The normal permeability of the iron core establishes a fixed inductance for the oscillation windings and the oscillator circuit is tuned in the regular manner, except that, since a fixed tuning condenser is used, the resonant frequency is fixed and is of a single value. Now, when an a-c voltage is established across condenser *C*, current will flow through coil *L* and the magnetizing current will change the permeability of the iron core and influence the inductance of the oscillator number 2 winding, thereby changing the frequency of the output voltage from oscillator number 2. Since the magnetizing current secured from the power supply is a varying current, the frequency of oscillator number 2 varies over a range and by designing the magnetizing current supply circuit and the iron core so that the current and the permeability varies between pre-determined limits, the frequency of oscillator num-

ber 2 will also vary between pre-determined limits. In the actual unit, the band width provided is 22 kilocycles, or 11 kilocycles each side of the mean frequency.

Frequency Modulation by Electronic Means

In Fig. 10-18 is shown the basic circuit through which frequency modulation is obtained by employing the dynamic capacitance of a tube to vary the frequency of the oscillator.

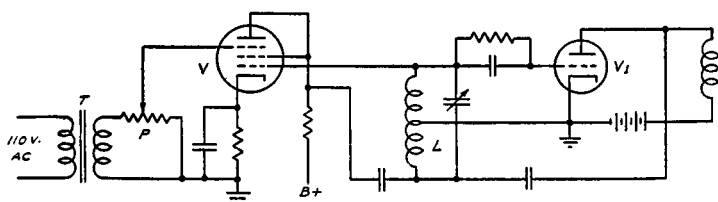


FIG. 10-18. A basic circuit in which frequency modulation is obtained by using the dynamic capacitance of a tube for varying the oscillator frequency.

The 110-volt a-c stepdown transformer T supplies a 60-cycle timing voltage to the suppressor grid of the pentode tube V . The control grid and the plate of this tube are connected across the tank coil L in the oscillator circuit. Therefore, the frequency generated by the oscillator tube V_1 is determined by the inductance of coil L and the grid-to-plate capacitance of tube V which shunts the coil.

Now, the grid-to-plate capacitance of the pentode is reflected across its own input and is amplified to an extent determined by the operating potentials, so that the dynamic capacity existing in shunt with coil L is greater than the static grid-plate capacity. This dynamic capacity may be increased and decreased by varying the potential on the suppressor grid.

The frequency generated by the oscillator with zero potential on the suppressor grid of the pentode is, naturally, the mean frequency, for an alternating potential on the suppressor grid will vary the frequency of the oscillator above and below this value. Since a 60-cycle voltage is applied to the suppressor grid of the pentode, the frequency of the oscillator varies above and below

the mean frequency 60 times a second, or, each cycle of frequency variation above and below the mean frequency will occupy one-sixtieth of a second.

The magnitude of the dynamic capacitance of the pentode, and therefore the extent of the frequency shift of the oscillator, is dependent upon the peak voltage impressed on the suppressor grid. Consequently, the degree of frequency variation in the oscillator circuit may be controlled by the potentiometer P which varies the a-c voltage impressed on the suppressor grid of the pentode.

If the oscillator is well designed, the output amplitude will remain reasonably constant over comparatively wide frequency variations.

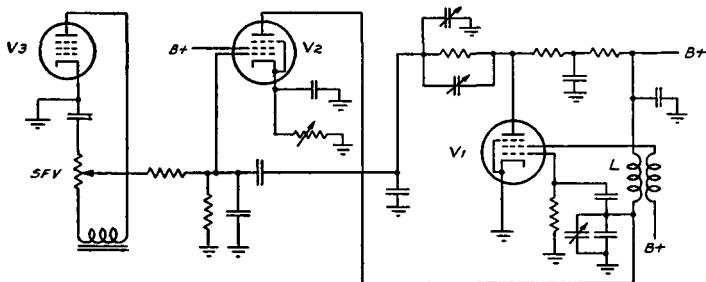


FIG. 10-19. A basic circuit in which the apparent inductance of a tube is used to provide the variation in frequency to obtain a frequency-modulated output.

Another electronic method of frequency-modulating a test oscillator, and similar to the one described above, uses the apparent inductance of a tube to provide the frequency variation. This is the same as the arrangement employed in many automatic frequency control systems, and the circuit is basically the same as that shown in Fig. 10-19.*

In this system the apparent inductance of a tube is varied by means of an alternating voltage impressed on the control grid of the tube. The plate impedance is caused to vary in frequency in the same manner as an inductance. This fictitious inductance

* For a complete discussion of such systems see "Automatic Frequency Control Systems" by John F. Rider.

being placed in shunt with the tuned circuit of an oscillator, varies the frequency of this oscillator by varying the total effective inductance acting in the oscillator circuit.

In Fig. 10-19 coil L is the tuned grid winding of the oscillator tube V_1 . If you check through the circuit, you will find that the plate-cathode circuit is connected across the oscillator grid coil L . Also you will see that the control grid of tube V_2 is arranged to receive a bias from the sweep voltage generator tube V_3 . This voltage acting upon the control grid of V_2 varies the apparent inductance (plate-cathode impedance) over a range of values determined by the magnitude of control voltage applied to V_2 from V_3 . Thus by varying the control, SVV , we can vary the range of frequency variation or band width of the oscillations developed in V_1 .

The A-F Frequency Modulator

The a-f frequency modulator is a device which supplies a continuously variable band of audio frequencies. In effect it is like the various r-f and i-f frequency modulators discussed earlier in this chapter, except that the mean frequency is zero and frequency modulation results in a final output which varies over 10,000 or 15,000 cycles each side of the zero beat or mean frequency. The ideal unit of this character would be one which varies over the aforementioned band, but only on one side of the zero beat setting with no signal whatsoever on the other side of the zero beat.

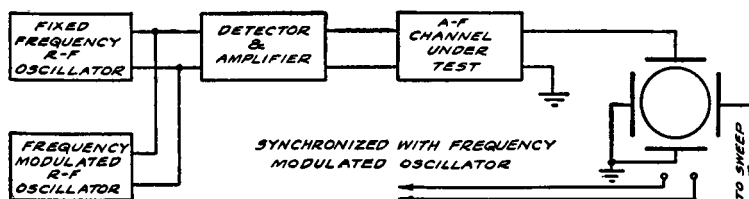


FIG. 10-20. Arrangement of the apparatus for the determination of overall a-f response curves.

The a-f frequency modulator herein illustrated is an experi-

mental model and to the best of our knowledge no similar piece of commercial equipment is available. Frequency modulation is accomplished by means of a rotating condenser which varies the frequency of one of two fixed-frequency oscillators. The other oscillator is, as has been said, of a fixed frequency. The device is illustrated graphically in Fig. 10-20. Since there is nothing special about the circuit, we do not deem it necessary to show the schematic of the respective oscillators. These two oscillators are operated at approximately 200 kc. One is frequency modulated with a motor-driven condenser so as to supply frequency modulation over a 10-kc band each side of zero beat.

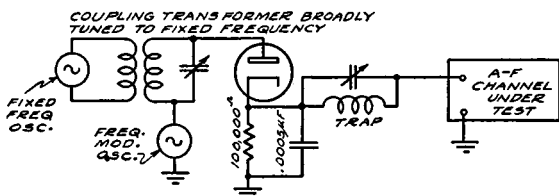


Fig. 10-21. The trap circuit in this audio frequency-modulation unit keeps the i.f. from the a-f channel being tested.

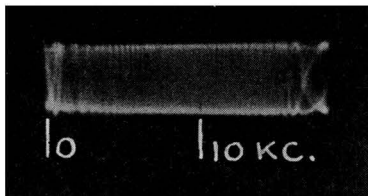
Because of the a. c. present in the frequency-modulated unit, the supplementary diagram, shown in Fig. 10-21 is offered. The frequency-modulated oscillator is grounded. In series with its output is the secondary of an i-f transformer tuned to about 200 kc or whatever the frequency being used. To this winding is coupled the primary winding, which is connected to the other oscillator. The detector is a diode tube with a 100,000 ohm load resistor, shunted with a .0005- μ f condenser. In the output voltage supply circuit is a trap tuned roughly to the frequency of the two oscillators, namely about 200 kc. This is used to keep i.f. out of the a-f signal. The rest is the usual routine.

The operating parameters were so selected that the output beat signal was linear over the 10,000-cycle frequency sweep. This is not difficult of accomplishment, since the variation of 10,000 cycles at 200,000 cycles is sufficiently small so that very little departure from the actual output at 200,000 cycles occurs. The oscillogram indicating the linearity of the output voltage over the

stated frequency range is shown in Fig. 10-22. The type of frequency modulation used is that which furnishes a double image, so that two types of image will be shown. One type is that which occurs when the timing frequency is the same as the speed of the frequency modulating condenser rotor, and the other type is that which occurs when the timing frequency is twice the speed of the frequency modulating condenser. A single image pattern can be obtained by making the sweep frequency inoperative during one half of the complete arc of travel of the frequency modulating condenser.

When the synchronization adjustment is such that the timing (sweep) frequency is equal to the frequency modulating condenser rotor speed, only one half of the entire pattern is used, as indicated. The other half of the trace is a repetition of the first in reverse order. Incidentally, due to lack of linearity of the sweep at the low frequency employed, that is between 20 and 25 cycles, the trace is not equally divided. The second trace is somewhat more crowded than the first.

FIG. 10-22. An oscillogram of the output of the audio frequency modulator. Note the linearity over the band from 0 to 10 kc.



By properly synchronizing the timing (sweep) voltage and the rotation of the frequency modulating condenser, the zero beat or lowest frequency setting can be set at the outside limits of the pattern, so that frequency variation in the used half of the complete trace is in the normal progression from the lowest to the highest, reading from left to right. The limitation of image size, only one half of the complete trace being of utility, is not a limitation of the system, but is due essentially to the design of the device we were using. Further elaboration of design can remove one half of the trace, leaving the entire screen available for the remaining trace.

Chapter XI

AUDIO-FREQUENCY OSCILLATORS

SIGNAL sources intended to operate over the audio band are known as audio oscillators and cover roughly from 10 cycles to 20,000 cycles. In a few instances, units are designed to develop frequencies as high as 40 kc. Still other units that are identified as audio-frequency oscillators cover not only the audio band, but extend into the supersonic range up to approximately 100 kc. In contrast with radio-frequency signal sources the name "signal generator" is seldom if ever used in connection with audio signal sources. Not that any rigid rule controls this, but it has become accepted practice to use the word oscillator.

Requirements

Requirements of audio oscillators are: good frequency stability, accurately calibrated frequency, good waveform, and uniformity of output over the frequency range. Without attempting to set specific standards for such types of units, it can be stated that investigation of audio oscillators of the vacuum tube type available on the market in the various categories interpreted on a qualitative basis, shows that a good oscillator of the vacuum-tube type is accurate in frequency to about 2 or 3%, usually less than the higher figure. As to harmonic content representative of the wave form, those units in the upper brackets have an overall harmonic content between .2 and 1%, whereas the less expensive units have an overall harmonic content between 5 and 10%. Audio oscillators operated in conjunction with an audio amplifier or which contain an audio amplifier, have a somewhat higher harmonic

content than those operated without, because the amplifier introduces a slight amount of distortion. The need for uniform output over the whole frequency range, or at least that portion of the range which is used for checking a-f amplifiers or portions of an audio system, can not be stressed too strongly. The absence of such uniformity leads to incorrect observations or measurements when checking such systems, unless the output is measured and adjusted at each test frequency.

Square Wave Generators

Much attention has been focused of late, in view of television development wherein phase distortion is an important consideration, upon amplifiers which are supposed to be of such design as to pass square waves. This means that there is a need for audio oscillators capable of producing square waves. Thus it can be said that the audio signal source of the future must not only meet all the requirements mentioned above but also must be capable of producing a square wave and in line with applications which appear upon the horizon, it must extend its lower limit of frequency to less than 10 cycles. Whether or not the audio signal source of the future will produce both sine and square waves, since the former is a single tone and the latter represents an infinite series of harmonics, is problematical. It is very likely that such square waves will be produced by possibly clipping the peaks of waves which contain a fundamental and a number of odd harmonics.

It has been found possible to develop such square waves by clipping in a series of stages the major portion of a sine audio wave of sufficient magnitude so that what was left represented substantially vertical sides. (Examples of such clipping action will be found elsewhere in this chapter.) Such square waves serve well as a means of checking phase distortion when and where such information is required. The presence of phase distortion in the amplifier being checked tends to slope the vertical sides of such square waves and also tends to slope the horizontal line representing the top and bottom of the wave.

Fundamental Frequency Oscillators

As you can readily understand, a-f oscillators must bear some resemblance to r-f oscillators, as both represent oscillating systems, so that it is impossible to segregate a-f oscillators into a type which has no equal in any frequency range with respect to circuit structure. By and large, a-f oscillators differ from r-f oscillators in the constants of the components. However, it is possible to state that in contrast with r-f oscillators, wherein the harmonic output of the oscillator is often used as the test signal, such is seldom the case in a-f oscillators wherein the fundamental signal developed by the oscillator is the one used as a test signal. The manner in which the output signal frequencies are produced in vacuum-tube a-f oscillators can be separated into two types: (1) the fundamental frequency oscillator, wherein the oscillating circuits are resonated to the frequency available at the output, and (2) a unit wherein the oscillating circuits are producing two different signals of different frequency, which, when mixed, produce the final signal available at the output.

The fundamental a-f oscillator is comparable to an r-f oscillator, except that much larger values of inductance and capacitance are necessarily employed. So large, in fact, that, in the case of inductance, except in high-priced oscillators where air-core coils are employed, iron-core coils must be used in order to conserve space. The cores are provided with air gaps to decrease the possibility of saturation which would result in poor wave-form. The capacitances required to produce audio frequencies are also so large that the use of variable condensers are out of the question. Consequently, it is necessary to use a number of fixed condensers in conjunction with a switching system, and cover the frequency range in steps.

In order to minimize harmonic distortion in such fundamental frequency a-f oscillators, the amplitude of oscillations is kept fairly low. To secure the required output, supplementary amplifiers are employed. Frequency stability is accomplished in two ways. Generally speaking, resistance stabilization is one of these methods and is used in the majority of the systems. The second method is by the use of a buffer amplifier between the oscillating system and the supplementary amplifier or the load. A typical a-f oscillator with all the frills omitted, showing resistance stabli-

zation and a buffer amplifier, is illustrated in Fig. 11-1. The resistance stabilizing circuit consists of the feedback resistance R in series with the fixed condenser C connected between the plate and input circuit of the oscillator tube. This variable resistance is of such design that it possesses a fixed minimum value. The

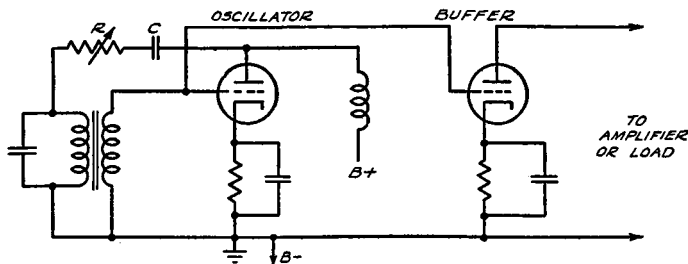


FIG. 11-1. A simple a-f oscillator with resistance stabilization and a buffer amplifier. The oscillator plate voltage is applied through an a-f choke.

value of the feedback resistance is such that critical adjustment of regeneration is possible. This resistance stabilizing circuit not only provides the feedback path, but likewise tends to isolate the tuned circuit from the tube element voltage variations, tube changes, etc. The buffer amplifier in turn effectively isolates the oscillating system and tube from the load.

Low impedance tubes are usually used for such fundamental audio oscillators. In determining the value of the resistance, a number of conditions must be recognized such as the amplification constant of the tube, the plate resistance of the tube at whatever values of operating voltages are used, and also the resistance of the tuned circuit. Speaking in generalities, as a result of experimental work with a number of L, C, and Q values wherein the frequency range covered was from approximately 35 cycles to 10,000 cycles, the circuit resistance varied between 3000 and 250,000 ohms, and the feedback resistance range extended from a minimum of approximately 8000 ohms to a maximum of 750,000 ohms. This minimum value can be either a fixed resistor or a tap on a variable resistor that covers the complete range.

As to an equation which will express the critical value of this feedback resistance, the following can be used:

$$\text{Critical Resistance} = R_L (\mu - 1) - R_P$$

where R_L = tuned circuit resistance

μ = amplification constant of the tube

R_P = plate resistance

The circuit of a typical fundamental frequency type of a-f test oscillator is shown in Fig. 11-2. It consists of an oscillator, a buffer, an intermediate voltage amplifier stage, and a push-pull power amplifier. The various frequency ranges are obtained in steps, by means of a multi-section selector switch which not only connects in the proper fixed coils and condensers for a given range, but also the correct value of feedback resistance for each range. The variable condenser, C , which is

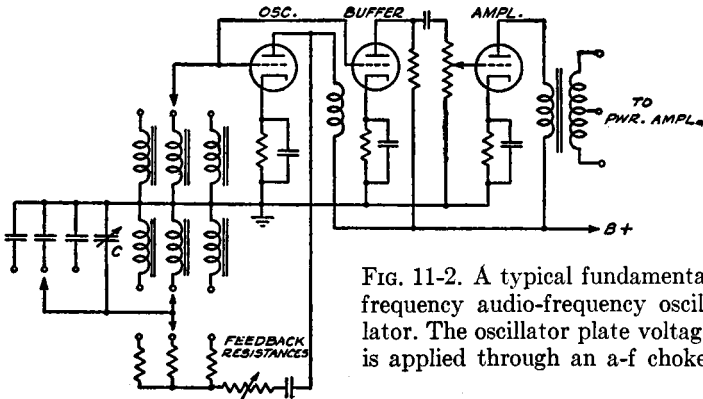


FIG. 11-2. A typical fundamental frequency audio-frequency oscillator. The oscillator plate voltage is applied through an a-f choke.

automatically placed in shunt with whichever coil-condenser combination is in circuit, enables intermediate values of frequency to be obtained over any of the fixed ranges. (Only three are shown for convenience's sake.) As you can see, tuning is accomplished in the plate circuit.

A resistance-stabilized audio-frequency oscillator of this type produces an unusually good waveform, and the amplitude of oscillation is practically constant over the entire frequency range. The upper frequency limit is about 100 kilocycles.

A simple Hartley oscillator circuit employing resistance stabilization is shown in Fig. 11-3. Here both grid and plate coils are shunted by the tuning condenser. Although not so shown, these coils can be of the iron-core variety and the variable condenser

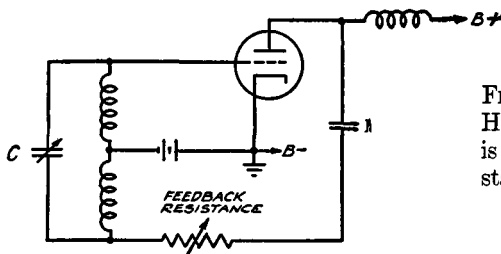
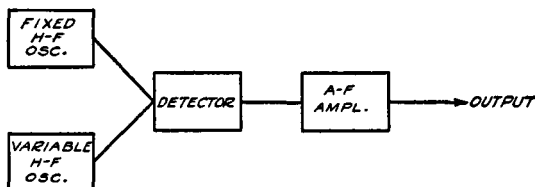


FIG. 11-3, left. A simple Hartley oscillator in which is incorporated resistance stabilization.

FIG. 11-4, right. Block diagram of a beat-frequency a-f oscillator.



can be shunted by a series of switch-controlled fixed condensers. Such a circuit operates well as a low-frequency oscillator.

Beat-frequency oscillators

The principle of operation of a *beat-frequency oscillator*, as its name implies, is based upon the use of *two* high-frequency oscillators, the outputs of which differ somewhat in their respective frequencies, and are combined so as to produce a *difference* or *beat* frequency. For instance, if one oscillator is set at 100,000 cycles and the other at 95,000 cycles, the resultant audible beat note when the two frequencies are combined will be 5000 cycles.

For convenience, one of the oscillators is maintained at a fixed frequency while the other oscillator is varied to produce the beat frequency. A block diagram showing this operation is given in Fig. 11-4. The outputs of the two oscillators are fed into a detector, where they are combined and rectified. The beat frequency is obtained in the output of the detector circuit. The

undesired high frequencies are filtered out so that only the audible beat frequency remains.

The main requirements of a beat-frequency oscillator are a wide frequency range, purity of waveform, frequency stability, and constant output over the frequency range.

Unlike the fundamental frequency type of audio oscillator, both oscillators in a beat-frequency unit operate at high frequencies. Consequently, air-core coils fit all requirements and a single variable air-condenser is used to obtain a continuous change in frequency over the entire audio-frequency range. Due to the small variation in frequency required, it is possible to secure an output which is of practically constant amplitude over the whole range.

Frequency locking

Unless special precautions are taken to avoid it, the frequency of the weaker oscillator will tend to lock into synchronism with the frequency of the stronger oscillator when the frequency of the former closely approaches the frequency of the latter. Under these circumstances the two oscillators are in a condition of zero-beat and, consequently, no audio-frequency output is obtained.

The prevention of such frequency locking requires the complete electrical isolation of the two oscillators so that they cannot interact upon each other. Adequate shielding of the two oscillators, particularly the weaker one, will minimize interaction due to stray fields. Nevertheless, the outputs of the two oscillators must be combined and, in order to prevent interaction at this point a buffer amplifier is usually interposed between the weaker oscillator and the point at which the two frequencies are mixed. The function of the buffer amplifier in this case is to prevent the stronger oscillation in the mixing circuit from getting to the output circuit of the weak oscillator.

Frequency stability

Great care must be taken that neither of the high-frequency oscillators vary in frequency, for this would result in a change in the pitch of the beat frequency. Assuming that one of the oscil-

lators is operating at 100,000 cycles and the other at 99,000 cycles, a 0.1% change in frequency by the first would vary the beat frequency by 100 cycles one way or the other, making the resultant beat 900 cycles or 1100 cycles rather than the expected 1000 cycles. The discrepancy would be even more pronounced at beat frequencies below 1000 cycles.

Changes in temperature due to vacuum tubes and other heat-producing apparatus is probably the major factor which causes frequency instability. The oscillator coils and condensers should possess a high inherent stability which will enable them to be affected only slightly by temperature changes. In addition, these coils and condensers should be located at a reasonable distance from heat producing apparatus.

If the two oscillators are made as nearly identical as possible, changes in temperature and operating potentials will affect both oscillators about equally, so that the frequency change in one will be offset by the corresponding frequency change in the other, since both frequency variations will tend to be in the same direction, thereby maintaining the beat frequency constant.

A symmetrical placing of the components and rigid mounting will further assist in minimizing frequency instability. Ordinarily, both oscillators are of the electron-coupled type, which have comparatively high frequency stability.

Choice of fundamental frequencies

The frequencies of the two oscillators also have a significant influence upon frequency stability. If the fundamental frequencies are made low, frequency stability is attainable by virtue of the fact that an undesired change in frequency of either oscillator will result in only a small variation of the resulting beat frequency. On the other hand, the lower the fundamental frequencies, the greater the percentage difference between them when they beat to produce a high audio frequency. Consequently, the effects of temperature and other factors will influence both oscillators differently, and since the frequency change in one will not be compensated by the frequency change in the other, there will be a certain amount of frequency instability.

When a high frequency is employed, the ratio of inductance to capacitance, when using commercially available condensers, can be made small and the frequency stability is thereby increased. This, however, is offset by the fact that a small change in the fundamental frequency will result in a large change in the beat frequency. General practice would seem to indicate that a fundamental frequency between 100 and 350 kilocycles offers an excellent compromise between the various conflicting factors.

Tuning condenser

In a beat-frequency oscillator the entire audio-frequency range may be easily covered with a single variable air condenser, the range covered depending upon the condenser chosen.

In order that readings will be well spaced at the lower frequencies, the condenser plates are shaped to give a logarithmic scale. Beyond the audio-frequency range, the plate shape may be designed to give either a logarithmic or linear scale.

Elimination of harmonics

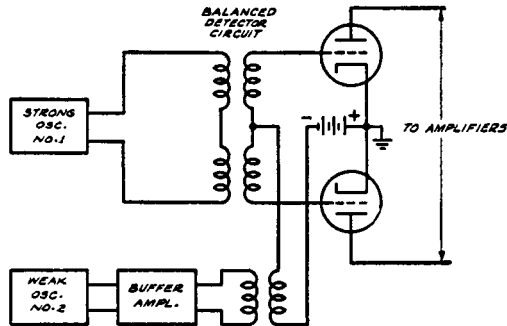
If both of the fundamental oscillators contain harmonics, the detector output will be a beat frequency which also contains harmonics. In order to obtain a harmonic-free beat-frequency output, at least one, and for convenience usually the fixed frequency oscillator, is thoroughly filtered.

A number of methods can be used for such filtering. One of these consists of arranging an electron-coupled oscillator for the fixed-frequency unit wherein the output of the oscillator is coupled to the mixer by means of a tuned circuit in the plate system and which circuit is resonated to the fundamental frequency being produced by that oscillator. It stands to reason that similar coupling of the fixed-frequency oscillator, if it is of conventional design, can be used wherein the resonant circuit, which is acting as a filter, serves as the coupling link between the fixed-frequency oscillator and the mixer. The fact that harmonics are being produced by the variable-frequency oscillator is of no moment, because these harmonics can not beat against the harmonics pro-

duced in the oscillating circuit of the fixed-frequency oscillator. The trap prevents the harmonics of the fixed-frequency oscillator from reaching the mixer.

Still another method of minimizing harmonics is to employ a balanced pushpull system of detection, whereby even harmonic distortion is eliminated from the output. Since the largest offender in such distortion production is the second harmonic component, such a balanced detector, shown in Fig. 11-5, is very effective.

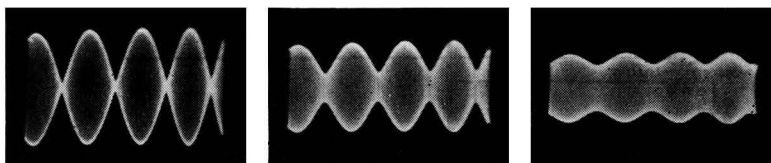
FIG. 11-5. A beat-frequency a-f oscillator in which is incorporated a balanced detector circuit for reducing the second-harmonic content of the output.



Every effort to make the weaker oscillator of the purest wave form will aid in the production of a final output signal which is substantially free from harmonics. As to detector distortion, best operation is secured by arranging a linear detector, wherein the signal from the strong oscillator is from ten to twenty times as great as the output of the weaker oscillator. Under these circumstances, the amplitude of the beat-frequency output is determined by the weaker oscillation and since distortion introduced during detection is influenced by the percentage modulation, distortion is minimized by such a ratio between the two signals. Since the output of the detector is proportional to the amplitude of the weaker signal, there exists a definite limit to which this weaker signal can be reduced in order to minimize the harmonic content.

In Figs. 11-6, 11-7, and 11-8 are shown oscillograms illustrating the manner in which the ratio between the fixed and variable oscillator voltages determines the extent of distortion in the beat signal. Note that a 1-to-1 ratio in the amplitude of these two

signals, which are mixed to produce the final audio beats, shows a definite "V" form of distortion in the troughs. This is reduced as the ratio of the signals is increased to 2-to-1 and when the ratio is 5-to-1, a sine wave envelope is present. These oscillo-



FIGS. 11-6, 11-7, 11-8, left to right. These oscillograms illustrate the manner in which the ratio between the fixed and variable oscillator voltages determines the distortion in the beat-frequency output. Ratios, from left to right, are 1-to-1, 2-to-1, and 5-to-1.

grams illustrate the two signals at the input of the mixer. The rectified output from the mixer, which is the audio beat, is proportional to and in wave shape is like the envelope.

RCA Type 68-B Beat Frequency Oscillator

The schematic of this laboratory instrument is shown in Fig. 11-9. Two r-f oscillators, one of which functions at a fixed frequency of 180 kc and one which is operated over a frequency range of from 179.98 to 163 kc, feed into a push-pull connected balanced modulator circuit to produce an a-f output range of from 20 to 17,000 cycles. This type of detector circuit reduces distortion and enables easier filtration.

The detector output is resistance-coupled to the output stage, consisting of four 6C5G's in push-pull parallel. The output transformer secondary is tapped to provide various output impedances which may be balanced to ground by using the center-tap.

The power supply incorporates a 6J7 which is used in conjunction with a 45-triode for voltage regulation. Further stabilization is secured by the 874 voltage regulator. A 6E5 tuning eye is used for checking the frequency calibration against the supply line frequency.

Distortion below 100 cycles is rated at 0.3% and above 100 cycles at 0.2%. Scale accuracy is within 1% above 100 cycles and within 1 cycle below 100 cycles. The maximum output voltage is 25 volts r-m-s, and the output deviation from that at 1000 cycles is within plus or minus 10% at output impedances of 250 or 500 ohms. At 5000 ohms, it is within 7.5%. The maximum power output is 0.125 watt.

RCA Model 154 Beat-Frequency Oscillator

A representative beat-type a-f oscillator used in the service field is shown in Fig. 11-10. A 6J7 is used as the fixed-frequency oscillator operating at 350 kc. The 6J7 is connected as a tetrode in that the screen and suppressor grids are joined and form the plate of an electron-coupled oscillator.

The same type of circuit is employed in the variable-frequency oscillator. The output of this oscillator is coupled to the control grid of the 6C5 triode detector. A coil, L_4 , in series with the cathode of the detector, feeds the electron-coupled output of the fixed-frequency oscillator to the detector. The resulting difference frequency is developed across the plate load R_9 of the 6C5. The coils L_7 and L_8 with their associated condensers, C_{16} , C_{17} and C_{18} , form a two-section filter to eliminate harmonics of the fixed-frequency and variable-frequency oscillators which are present in the plate circuit.

The desired beat frequency is fed to the 6J5 amplifier tube. R_{11} serves as an output control by varying the audio voltage applied to the control grid of the 6J5 amplifier.

Calibration of the beat oscillator is secured by using a neon tube which is coupled to the plate of the 6J5 amplifier by C_{20} . Calibration is effected at either 60 or 120 cycles by applying a.c. from one half of the power transformer secondary through the voltage-divider network composed of R_{16} , R_{17} and R_{18} to the neon tube S_3 in series with R_{19} , which limits the current in the neon tube. As zero beat is approached the lamp will flash at the difference frequency, which will become lower and lower as the variable-frequency oscillator is adjusted to produce the proper beat. The initial calibration is effected by varying the small air condenser, C_{11} , which is connected in parallel with the main

tuning condenser and therefore varies the variable-oscillator frequency.

Clipper Action

In Figs. 11-11 and 11-12, are oscillograms illustrating square waves produced by "clipping" sine waves of large amplitude. The wave of Fig. 11-11 was obtained by properly adjusting a diode

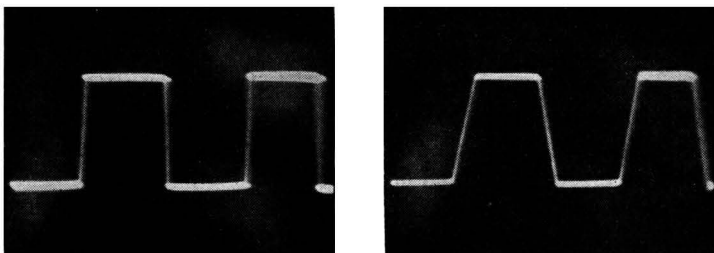


FIG. 11-11, left, FIG. 11-12, right. The oscillogram at the left shows a "clipped" sine wave when the diode was properly adjusted and the right-hand wave is that produced with insufficient input signal amplitude.

clipper and the wave form shown in Fig. 11-12 is a type of wave produced by clipping with a diode when the sine-wave input signal to the diode was of insufficient amplitude. A pentode may also be used as a clipper.

Square law detection has been used in the past, but since it has been replaced by the linear type of detection, we do not find it necessary to discuss the former in this book.

In line with the subject of distortion, it is necessary to include the amplifier. Distortion produced by the amplifier can be minimized by operating the tubes well below maximum and by the use of inverse feedback where it is found to be necessary.

Calibration

As a general rule a-f oscillators are equipped with means for calibration. That is to say a definite calibrating point is established upon the dial and some means is provided whereby a cali-

brating signal may be injected into the circuit so as to beat against the output signal developed in the mixer. In beat-frequency oscillators an indicator in the form of a meter, a tuning eye, or a neon tube is then used as a visual indicator to facilitate adjustment of the calibrating trimmer condenser provided for such purposes.

This calibrating trimmer is adjusted for zero beat between the calibrating frequency and the frequency being generated by the oscillator. In some cases such calibration is done without the use of an injected calibrating frequency. Instead the tuning dial is set at zero frequency and the calibrating trimmer is adjusted for accurate zero-beat indication upon the meter, the neon tube or whatever form of indicator is employed.

Chapter XII

RADIO-FREQUENCY SIGNAL SOURCES

AS FAR as radio-frequency signal sources are concerned, two classifications of equipment exist: test oscillators and signal generators. Recognizing a distinction between these devices, a similarity in requirements does exist. Because these requirements relate to signal sources as a whole, we feel that it is better to speak in terms of what such signal sources should possess rather than elaborate upon the existing capabilities of test oscillators and signal generators.

Devices identified as signal generators should first of all contain a reference level indicator so that it is possible to identify definitely, with of course reasonable tolerance, the output signal level and by the same token, be able to set the instrument to a certain predetermined signal level. Such devices should provide with respect to the future for frequency modulation, broad-band modulation up to at least several megacycles and square wave modulation. The a-f modulation need not necessarily be of variable percentage, particularly in view of the standardization of receiver testing at 30% modulation. However, if a signal generator is provided with variable percentage modulation, it naturally increases the utility of the device.

From the viewpoint of utility, those devices identified as signal generators should cover the frequency range in fundamentals and such control must be maintained so that if a sine-wave modulation is desired, this sine character is retained on all the r-f bands. The need for the use of fundamental frequencies lies in the demand for a calibrated output at all frequencies within the range of the instrument.

Test Oscillators

Test oscillators, which in general represent a classification of instruments usually lower in cost than signal generators, differ from signal generators in one basic respect: a reference level indicator is not employed. This can not help but result in output calibrations which are not as accurate as those of the signal generator. We recognize that such control systems as automatic amplitude control exist and if such systems are incorporated in a test oscillator and really do keep the carrier level constant on all frequencies within the range of the device, then it is possible to secure a substantially accurate signal output in accordance with the attenuator calibrations without the use of reference level indicator. Experimental verification of the feasibility of such automatic amplitude control shows that it can be done over a number of bands, but we are not in a position at the time of this writing to state definitely that it can be done with the same accuracy over a range, let us say, of 100 kc to 130 or 150 mc. Test oscillators equipped with automatic amplitude control should, like signal generators, employ fundamental frequencies throughout the range of the instrument.

As to percentage of modulation, test oscillators do not require variable percentages, but standardization of some percentage of modulation should be effected. Experimental investigations of a number of such instruments available today show a percentage of modulation fixed at various values between 30% and 60%. The reason we mention this percentage of modulation is that receiver manufacturers are including in their service notes receiver sensitivity data expressed in terms of 30% modulation and if the signal source used differs from this percentage of modulation, test results will be misleading, unless proper correction is employed in the calculations for any difference in percentage modulation between the reference data furnished by the manufacturer and that provided in the signal source.

It is important to remember in connection with variable percentage of modulation that this can be accomplished only when one of two conditions exist. First, when it is possible to set both audio and carrier levels on all frequencies within the range of the instrument; second, when the carrier level is held constant on all frequencies within the range of the instrument and the modu-

lating signal is always of a constant level. If external modulation is used and a certain percentage of modulation is intended, then both the audio voltage and the carrier voltage levels must be known and under control.

Accuracy of Frequency Calibration

It is difficult to identify the accuracy of frequency calibration required in both signal generators and test oscillators. Naturally the most accurate frequency calibration possible is desired in all signal sources, but we have learned to accept as the result of practice an accuracy of from $\frac{1}{2}$ to 1% in signal generators and from 2 to 3% in devices identified as test oscillators. This by no means signifies that a test oscillator should not be more accurate than 2 or 3%. Every reduction in such tolerance is advantageous.

Another factor associated with accuracy of frequency calibration is permanency of such calibrations which include frequency stability. In this respect, too, we have learned to accept a difference between signal generators and test oscillators purely on the basis of price. This, however, does not mean that we can accept either temporary drift or a permanent change in calibration in the test oscillator. Since even simple test oscillators can be held stable to within .05% and in many cases even better than that over a period of hours with as high as 30% change in line voltage, stability of frequency calibration should be possessed by the test oscillator as well as the signal generator.

Accuracy of Output Voltage

The accuracy of output voltages obtained from an r-f signal source is a broad subject. A great deal depends on four factors: (1) leakage; (2) the design of the attenuator network; (3) the highest voltage available from the output; and (4) the frequency of the signal. As is to be expected, signal generators, if they really are representative of such instruments, are usually more accurate than test oscillators, but even in signal generators we find definite variations. The highest order of accuracy is usually found over the range of lower frequencies, i.e. from the lower limit of the i-f band to possibly several megacycles. An idea of the

accuracy to be obtained can be gleaned from the fact that one of the very good and standard signal generators found in many laboratories, is rated at frequencies up to a few megacycles at about 3%; that from about 5 to 15 mc the rating varies from around 6 to 14%.

In the case mentioned the frequencies are all fundamentals. In test oscillators wherein the design of the attenuator system is not as accurate as in signal generators, it would not be surprising to find accuracies from 20 to 30%. At higher frequencies—say from 20 to 150 mc—the usual run of signal sources intended for radio service work are to be found within a range between 20 and 50%.

The use of harmonics still further precludes the possibility of a high order of accuracy in output voltage ratings, because by-passing effects become more pronounced as the order of the harmonic increases. Calibration of an attenuator on a fundamental will not be found to apply with the same order of accuracy on the harmonics.

As to the minimum signal output from a signal source—either a test oscillator or a signal generator—depends a great deal on the intended application. Personally—and this is strictly our own viewpoint—we do not see the need for extremely low values of signal output for signal sources employed in the service field. Our reasons for this statement are twofold: (1) the instrument is seldom if ever used in a locality where the noise level is sufficiently low to permit testing with a signal input from $\frac{1}{2}$ to 1 or 2 microvolts. (In fact unless a shielded room is used, which is very seldom found in a service shop, the noise level is several times the amount mentioned and sometimes 10 or 20 times this amount.) (2) the leakage in signal sources intended for the service field is seldom if ever less than 1 or 2 microvolts on the broadcast band and higher on higher bands.

Because of all this we feel that if the signal source is rated at a minimum of approximately 5 microvolts over the intermediate to high-frequency bands, up to perhaps 20 or 30 mc—that is perfectly satisfactory. Of course, if the signal source is controllable to some output less than this figure, it most certainly broadens its field of application. At frequencies above 20 or 30 mc, a minimum output of from 25 to 50 microvolts is usable in the radio service field because of the much higher noise level usually experienced and because of the lower sensitivity of the receivers.

As to the maximum output, this, too, depends on the intended application and is definitely reflected in the cost. The higher the signal output, the greater is the problem of attenuation and shielding and if this high output is to be attenuated to a very low level, it stands to reason that the cost of the instrument is increased. Leakage is also increased as the signal output is increased and if an attempt is made to hold down this leakage, the cost of the device must necessarily rise. As to instruments intended for the service field, considering existing applications of the device, an output of from 50,000 to 100,000 microvolts (.05 to 0.1 volt) up to frequencies approximating 70 or 80 mc is ample. In the i-f bands, an output up to 1 volt is a convenience in aligning.

Types of Signal Sources

By types of signal sources, we mean the three classifications of such r-f signal sources. They are

- I. Fundamental-frequency types.
- II. Fundamental and harmonic frequency types.
- III. Beat-frequency types.

Signal generators as a rule are of the fundamental-frequency type, i.e. the entire frequency range of the instrument is composed of fundamental frequencies. Such an arrangement has the advantage of providing more accurate output voltages over the entire frequency range than the type of signal source which uses fundamentals for one portion of the frequency range of the instrument and harmonics for the remainder of the instrument's range. The first classification also provides for greater output at the higher frequencies than is generally available in the fundamental-harmonic type of signal source. However, it is true that greater attenuation of the higher frequencies is possible in the harmonic type of signal source than in the fundamental type, if only because the signal itself, secured from the oscillating system, is much weaker. It is to be remembered that such harmonics are not constant with respect to the fundamental amplitude over the various ranges so that attenuator settings do not really represent values of signal output.

As to accuracy of frequency calibration, the fundamental-harmonic type would tend to provide greater accuracy than the strictly fundamental type, because better control of accuracy is possible at lower frequencies than at the higher frequencies for which calibration would be required in the fundamental type of oscillator. This, of course, assumes that calibration of the fundamental frequencies in the fundamental-harmonic type of oscillator is correct. This is quite important because the error in calibration on the fundamental frequency is reflected in the harmonic frequency. If the tolerance employed when calibrating the fundamental frequency is too great, it might make such a difference on one of the higher harmonics as to make the harmonic signal unsuitable for use. For example, an accuracy of 3% at 10 mc means 300 kc. The same accuracy on the 4th harmonic of that signal would mean a deviation from the correct 40 mc of 1.2 mc.

Another difference between the two types of signal sources is that an error at one calibration in the fundamental type of oscillator influences but one frequency, whereas in the fundamental-harmonic type of oscillator, an error in the fundamental calibration influences all the harmonic frequencies. Since any fundamental type of signal source can be used as a fundamental-harmonic type by simply using the harmonic of the desired fundamental, the fundamental type of oscillator seems superior. This takes into consideration the fact that greater frequency stability is available in the fundamental-harmonic type where the harmonic is used, as against the fundamental of the same frequency in the fundamental type, because in the final analysis the same order of stability is available in the fundamental type of signal source by employing a harmonic of the fundamental frequency, i.e. if the stability of the high-frequency signal is not as great as the harmonic of the other type of signal source.

The beat-frequency type of oscillator serves its purpose best when frequency modulation is required, although it must be remembered that it is possible to frequency-modulate both the fundamental and the fundamental-harmonic types of oscillators. In the frequency-modulated oscillator it is also possible to use a harmonic of the fundamental which is being frequency modulated and in that way secure frequency modulation on a higher frequency than is the normal range of the instrument. There

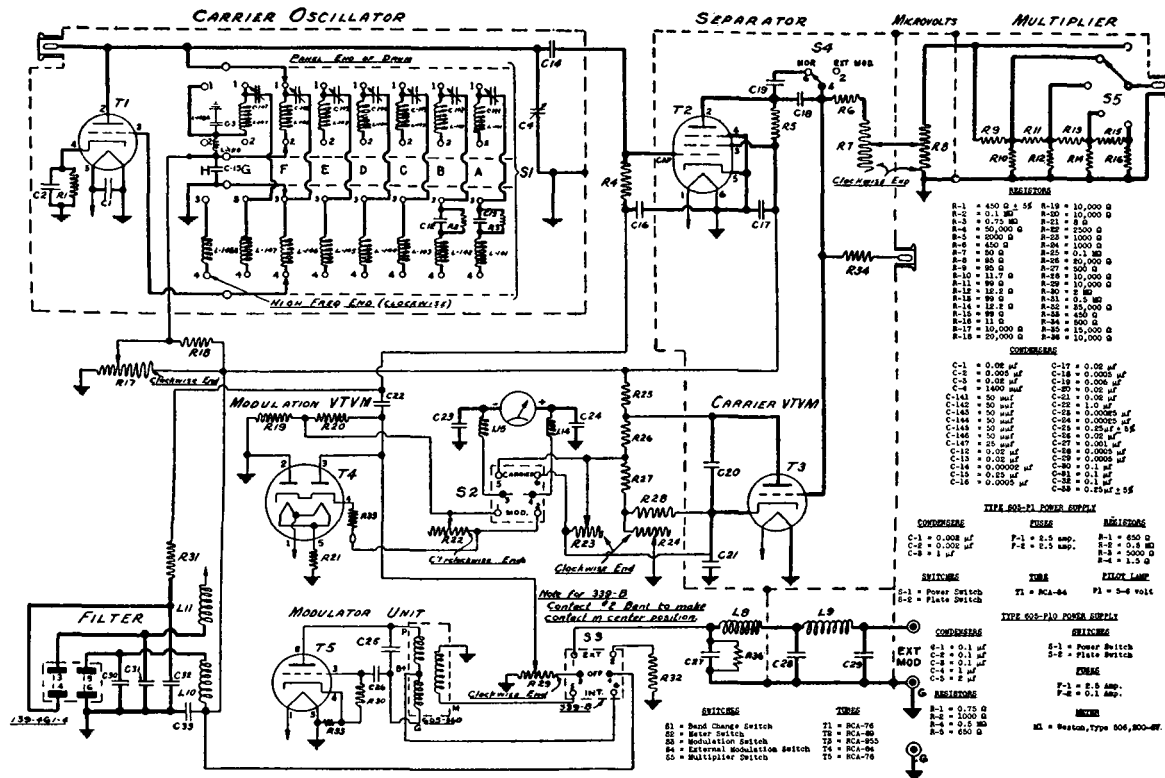


FIG. 12-1. General Radio Type 605-B Standard-Signal Generator.

Courtesy of General Radio Co.

seems to be no advantage in the beat-frequency oscillator for the production of amplitude modulated r-f and u-h-f signals.

Combination R-F, A-F Signal Sources

An example of the beat-frequency type of r-f oscillator which differs from that type of instrument used for frequency modulation is the r-f signal source wherein one oscillator tube is used as a fundamental-frequency oscillator and the audio oscillator tube, which serves to produce the modulating voltage, also is arranged to function as a fixed i-f oscillator and to heterodyne the variable-frequency r-f oscillator. A mixer tube provided for the purpose produces a final output which is of audio character and which is continuously variable between zero and an upper limit between 10,000 and 15,000 cycles.

Audio Output

Practically all r-f signal sources intended for use in the radio service field provide an audio output, usually at a fixed frequency of 400 cycles. This is the most commonly used modulating frequency employed in the oscillator or signal generator, whichever is the designation of the instrument. In the majority of cases this audio voltage is a close approach to a sine character, and varies in output level according to the design of the instrument in question. In a few instances provision is made for variable control of this audio signal level; this is an advantage because it enables the application of the audio signal to tube circuits where high-level audio signals will cause overload. When such variable audio level voltage is available, the output is sometimes calibrated, although the maximum voltage usually is specified. As to this maximum level, there seems to be no particular advantage in a high figure over a low figure. 3 or 4 volts of audio appears just as satisfactory as 10 or 15 volts, particularly in view of recommended receiver test procedure, wherein the required audio signal across the first a-f tube in receivers varies between .05 and .1 volt. It so happens that the maximum available a-f output voltage is usually determined by the level needed to modulate the r-f carrier properly within the signal source.

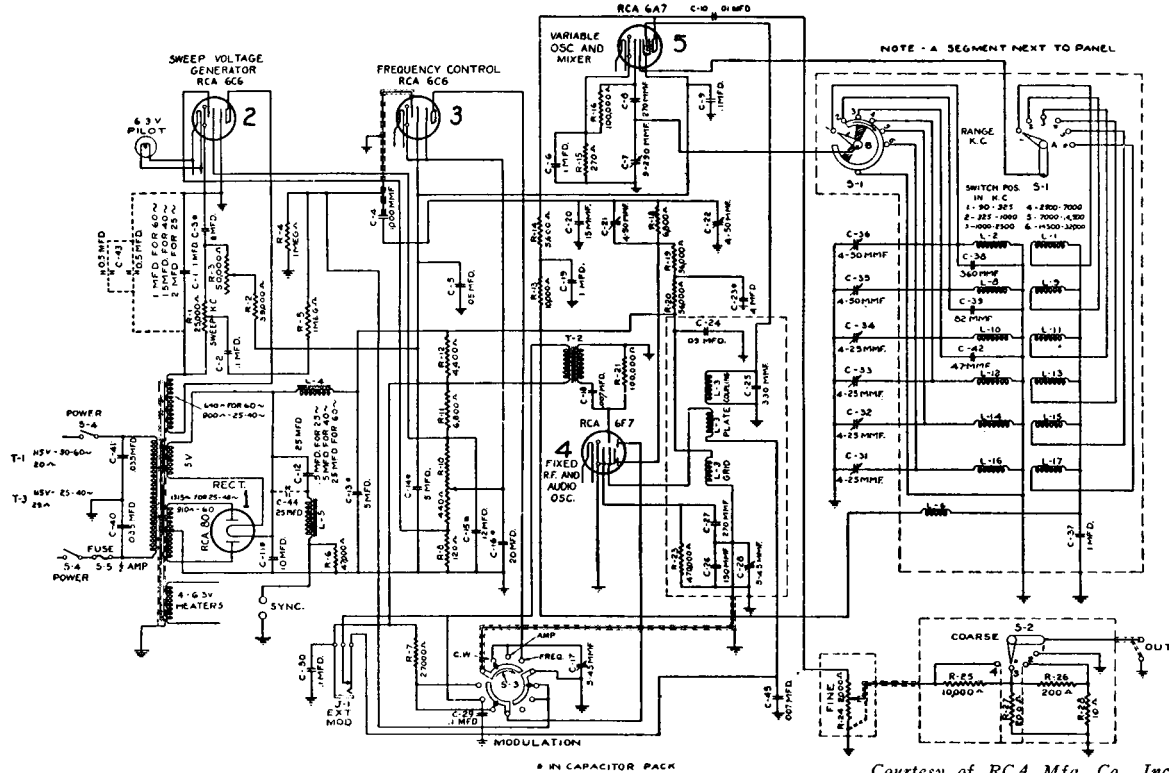


FIG. 12-2. RCA Model 150 Test Oscillator.

Courtesy of RCA Mfg. Co., Inc.

Shielding

The subject of shielding is of particular interest from the angle of stray signal leakage. Granting that the function of this book is not to furnish constructional material, reference to shielding requirements cannot help but be of aid in selecting instruments. In this respect it should be understood that the statements made herein are not necessarily rigid requirements, but represent accepted practices.

Shielding is very much more effective on the lower frequencies than on the higher frequencies. In fact, the connection to ground at one end of a tuning condenser shaft does not necessarily place all positions of that condenser shaft at ground potential for ultra-high frequencies—say, between 50 and 150 megacycles. Because of this, most effective design calls for an insulated shaft between the condenser rotor and the junction to the dial, so that the entire condenser assembly is behind the panel, and no metal junction to the circuits carrying high-frequency currents is located on the outside of the panel, which, naturally, is of metal. Unfortunately, however, very few such insulated shafts are used in signal sources intended for the service field.

High-frequency signal sources should be provided with all shielded output cables, or shielded output terminals.

Recognizing that a test oscillator or signal generator is, in the final analysis, just a signal source and that the cost determines the design, there are nevertheless certain facts which must be appreciated. This is particularly true if the unit is intended to furnish ultra-high frequency signals, although it also applies to signals of lower frequencies if best results are to be obtained.

Starting with the case of cabinet and panel, the most perfect junction possible between the cabinet and the panel should be used. Stray signal leakage will take place between the cabinet and the panel unless the metal junction between these two is good. This calls for panel mounting all around the edge of the panel, so as to avoid buckling of the panel and the space between the panel and the cabinet. Perhaps it might not add to the appearance of the unit when taken apart, but junctions of the metal panel and the cabinet by means of spring wiper contacts throughout the length can well supplement fastening screws.

Large-size holes in the panel are to be avoided as much as possible—for that matter, all holes—because each such hole is a possible point of leakage. We must of course accept the holes needed for the various control shafts. The more perfect the design within the cabinet, the greater the latitude allowed in the panel design. As to the design within the cabinet, rigid rules cannot be set, but there are certain standard practices which seem quite important and necessary.

As much as possible of the oscillating circuits should be contained within a shielded compartment—particularly the r-f oscillator coils and the tuning condenser, if possible. Coil cans should not be open or have holes in the walls. The attenuator system must be kept apart from the oscillating circuits and separately shielded. Likewise, in signal generators equipped with reference-level indicators, only that portion of the reference-level circuit which receives energy from the oscillating system must be close to the oscillating system. The remainder of the circuit must be shielded from the oscillating system.

Sections of attenuator networks should likewise be shielded from each other because they operate at widely different power levels. Capacity coupling between sections must be a minimum.

Ideal design of such r-f signal sources calls for the use of a r-f choke and bypass condenser at all points where hot leads leave shielded compartments, but since such design greatly influences the cost, they are more likely to appear in signal generators than in test oscillators.

Calibration

All signal sources, regardless of design, require a periodic check of the condition of calibration. The more precise the application of the unit, the greater the need for accurate checking of the calibration. A number of signal generators intended for use in the service field contain means for checking the calibration of the instrument. In some, a crystal is included within the device, or the device contains some means whereby a test signal from a crystal calibrator or a broadcast station can be fed into the signal source and heterodyned against the signal generated within the signal source.

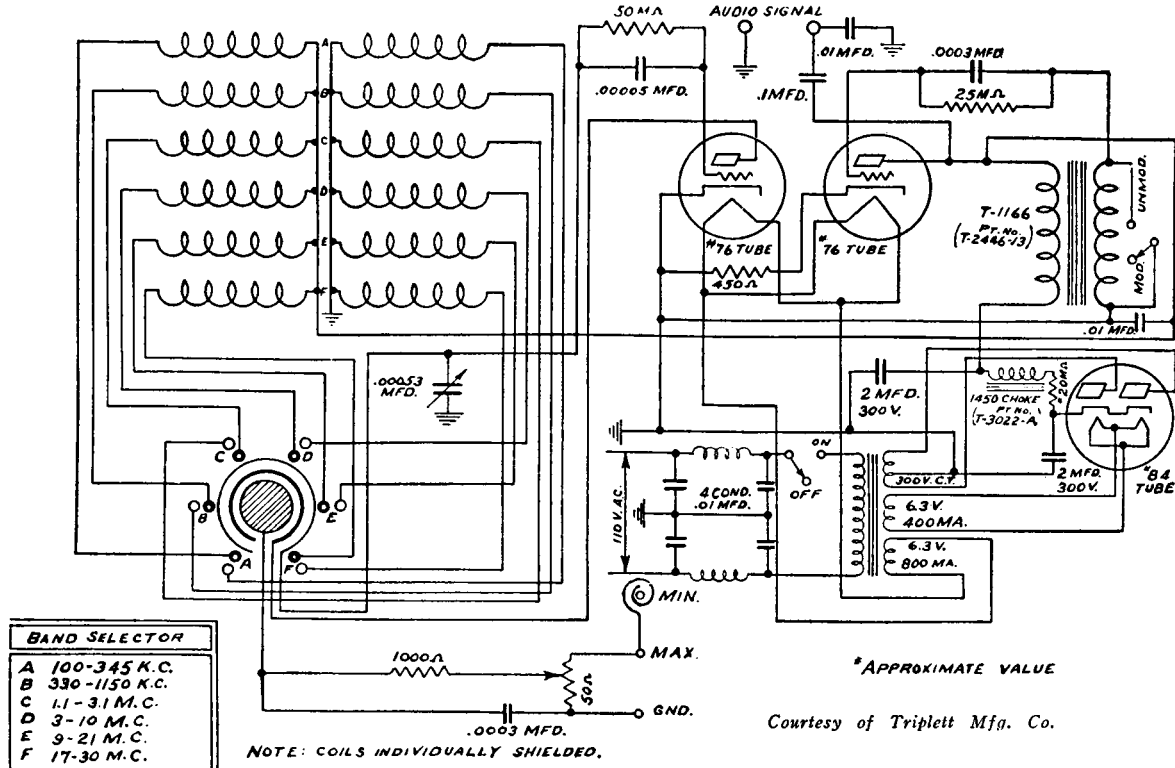


Fig. 12-4. The Triplet Model 1232 Test Oscillator. Frequency range from 100 kc to 30 mc in six bands on fundamental frequencies. Internal modulation at 400 cycles. Provision for supplying an unmodulated signal. Audio signal available for external use.

What has been said does not, however, mean that such calibration arrangements must be included in all signal sources. It is a definite convenience which can be interpreted in two financial values, but there are numerous other methods of checking the calibration of either test oscillators or signal generators.

Commercial Signal Sources

Examples of commercial r-f oscillators and signal sources which find use in the service field are shown herewith, and also a typical laboratory type of standard signal generator.

The General Radio Type 605-B Standard-Signal Generator

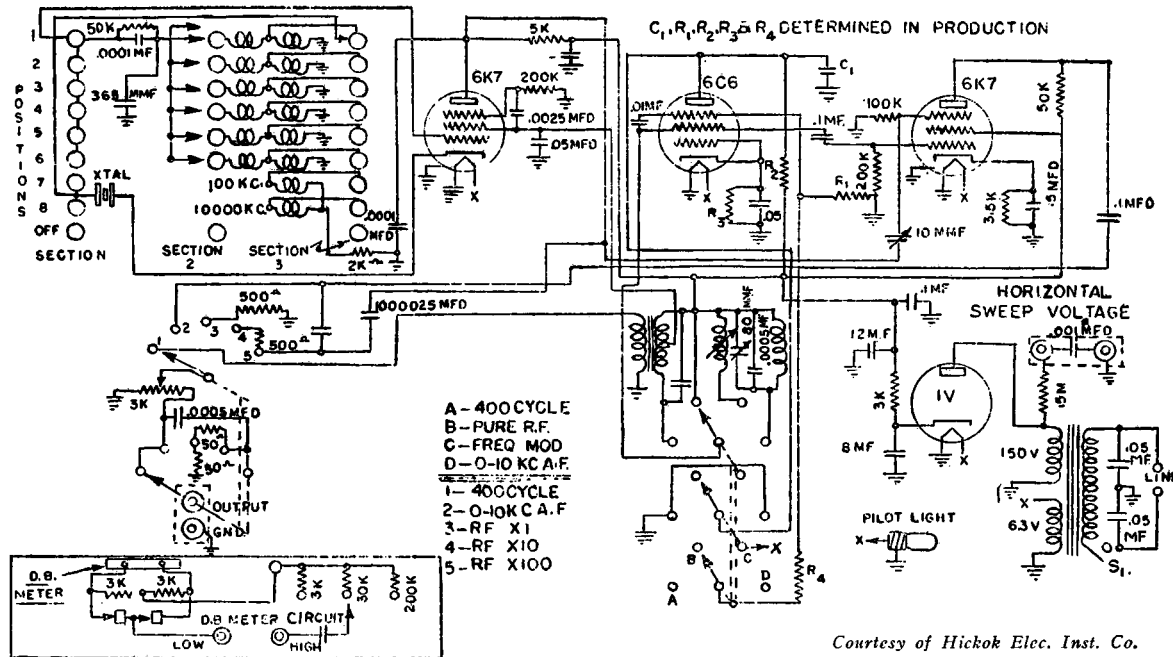
The circuit diagram of this laboratory-type signal generator is shown in Fig. 12-1. The frequency range of 9.5 kc to 30 mc is covered in 7 bands and is direct-reading from a calibrated dial. An additional range of 30 to 50 mc, not direct reading, is also provided.

As shown, the r-f oscillator *T1* operates in a reversed feedback circuit and the r-f voltage is coupled to the control grid of the separator tube *T2* by *C14*. The separator tube is an amplifier which is used to isolate the r-f oscillator from the attenuator circuit so that the attenuator load will not react on the carrier frequency. The resistance ladder attenuator is coupled to the plate circuit of the separator tube by *C18* and *C19*.

Modulation at 400 cycles is provided by the a-f oscillator *T5*, which uses a Hartley circuit. The modulation percentage is continuously variable up to 80% and is determined by the modulation voltage applied to the control grid of the separator tube. This a-f voltage is measured by the type 84 full-wave rectifier *T4* and a d-c meter. External modulation may be applied and the modulation characteristic is flat to within 1 db. from 30 to 15,000 cycles.

A type 955 acorn triode is used as a v-t voltmeter to measure the r-f signal voltage and is connected across the output load of the separator tube.

The frequency calibration is rated accurate to 1% up to 10 mc and to 1.5% between 10 and 30 mc. The output voltage range



Courtesy of Hickok Elec. Inst. Co.

Fig. 12-5. The Hickok Models 170 and 180 Test Oscillator. Frequency range 100 kc to 30 mc in six bands on fundamental frequencies. Provides amplitude and frequency modulation. 100 to 10,000-cycle variable a-f modulation. Crystal-controlled 100 kc and 1000 kc output available.

extends from 0.5 microvolt to 0.1.volt and is rated accurate to 3% below 3 mc; to 5% from 3 to 10 mc, and to 10% from 10 to 30 mc, (or to 0.1, 0.2 or 0.4 microvolt respectively for very low output signal voltages).

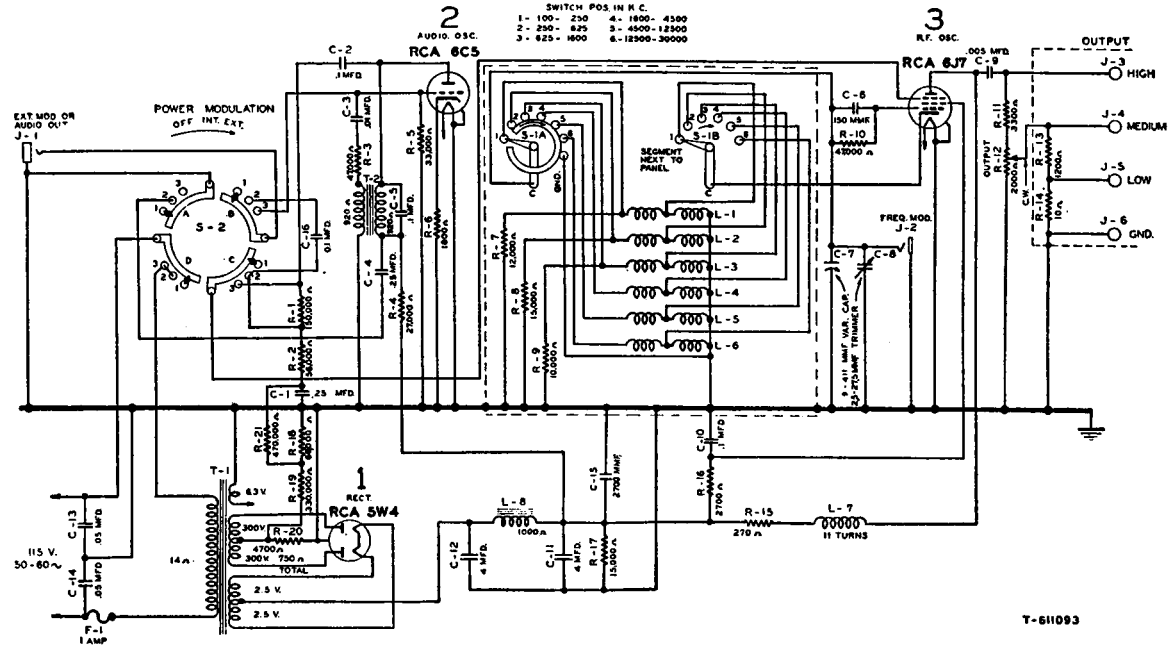
The a-c operated power supply may be replaced by a control panel for battery operation.

The RCA Model 150 Test Oscillator

This test oscillator provides either amplitude or frequency-modulated signals from 90 kc to 32 mc in six ranges. As shown in Fig. 12-2, the beat-frequency type of circuit is employed, the pentode section of the 6F7 tube forming an electron-coupled oscillator that generates the normally fixed frequency of 800 kc which is fed to the signal grid of the 6A7 mixer-oscillator. The first and second grids of the 6A7 are utilized in a tickler-feedback circuit to form the variable-frequency oscillator. The two signals are mixed electronically in the 6A7 and the resulting amplitude-modulated output signal developed at the 6A7 plate, is coupled through *C10* to the output attenuator. The plate load for the 6A7 is the r-f choke, *L6*. This test oscillator utilizes the difference frequency thus produced for frequencies up to 7 mc; above that the sum frequency is employed.

For 400-cycle amplitude modulation the triode section of the 6F7 is employed. The iron-core transformer *T2* is used in a tickler-feedback circuit to produce a 400-cycle note that plate-modulates the 800-kc fixed-frequency oscillator. When this modulated signal is fed to the 6A7 mixer-oscillator, the output frequency is also amplitude modulated. For external modulation, the triode section of the 6F7 is not used. The external modulation voltage is placed in series with the plate of the 800-kc fixed-frequency oscillator and plate modulation at the applied audio frequency this results.

Frequency modulation is more complex. In this test oscillator, the frequency-control tube, which serves as a variable inductance, is a 6C6. The 800-kc signal voltage from the plate of the pentode section of the 6F7 is fed through a phase-shifting network composed of *R18* and the condensers *C20*, *C21* and *C22*, to the control grid of the frequency-control 6C6 tube. The plate of this 6C6 is



Courtesy of RCA Mfg. Co., Inc.

FIG. 12-6. The RCA Model 153 Test Oscillator. Frequency range 100 kc to 30 mc in six bands on fundamental frequencies. Internal modulation at 400 cycles. Provision for external amplitude or frequency modulation and use of internal audio voltage externally.

coupled to the pentode control grid of the 6F7. Since the cathodes of both tubes are at ground r-f potential, the apparent inductance of the frequency-control tube is effectively across the tuned grid circuit of the 800-kc oscillator. Any variations in the apparent inductance of the control tube will shift the operating frequency of the 800-kc oscillator.

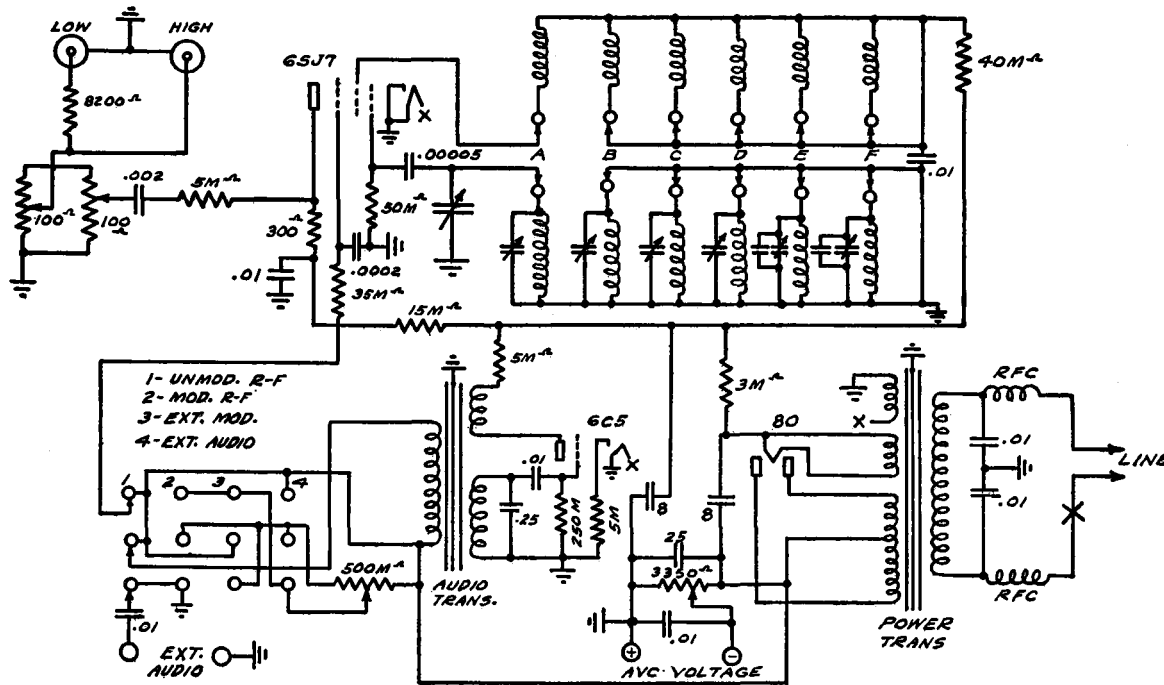
The variation in apparent inductance of the frequency-control tube is effected by varying its control voltage. This is accomplished by the sweep-voltage generator tube, the 6C6. The plate voltage of the sweep-voltage generator tube is modulated at 60 cycles and the operating voltages are chosen so that a pyramidal output wave results, which varies at a 60-cycle rate. This output voltage is developed across $R1$ and is fed to the frequency-control tube grid. The adjustment of $R1$ controls the amount of sweep voltage fed to the frequency-control tube grid, and, therefore, the amount of frequency shift which results from this change of grid potential. Since the control voltage thus applied varies at a 60-cycle rate, the frequency change in the 800-kc oscillator varies at the same rate. This frequency change is limited to a maximum of 20 kc above and below the normal 800-kc frequency. The total band width is variable from 1 to 40 kc.

If we now apply this varying-frequency signal to the 6A7 mixer-oscillator, it will combine with the signal produced in the 6A7 oscillator circuit and provide a frequency-modulated signal in the output circuit. This type of frequency-modulated signal is employed in aligning with the cathode-ray oscillograph. The curves resulting are of the double-image type and are discussed in further detail in the chapter devoted to receiver testing.

The RCA Signalyst

This is a 10-band r-f signal generator covering a range extending from 100 kc to 120,000 kc on fundamental frequencies. As shown in Fig. 12-3, the instrument comprises an r-f oscillator, buffer-amplifier, tube voltmeter and a combination heterodyne detector, a-f amplifier-audio oscillator.

The r-f oscillator circuit is unique in that a tickler-feedback circuit is employed for all ranges except the three highest frequency bands. For the latter, a Colpitts circuit is utilized. Thus, on the first seven bands, the two sections of the gang condenser



Courtesy of Precision Apparatus Co.

FIG. 12-7. Circuit of the Precision Series E-200 Test Oscillator. Frequency 110 kc to 72,000 kc in six bands. Fundamental frequencies up to 18 mc. Variable modulation control. Provision for external modulation. Internal 400-cycle audio voltage may be used externally. Supplies avc substitution voltage.

C1 are connected in parallel across the grid coil, while for the remaining bands the sections are in series. No coil is shown for the highest frequency range, since the inductance of the leads is sufficient to enable oscillation at such high frequencies.

The r-f output is coupled to the #3 grid of the 6SA7 buffer-amplifier and external modulation up to 5 mc may be fed to the #1 grid of this tube. The triode section of the 6F7 generates the 400-cycle internal a-f modulating voltage and the pentode section functions as a heterodyne detector for calibration purposes. When so used, the triode section operates as an a-f amplifier.

The signal voltage developed in the plate circuit of the buffer-amplifier is capacity-coupled to a shielded resistance ladder attenuator and to the output measuring system, composed of a 6H6 rectifier and d-c meter.

The maximum output voltage is 50,000 microvolts on the low range and 0.3 volts on the high range. The calibrated attenuator provides a continuous output range which is direct-reading in microvolts to a minimum of one microvolt. The frequency calibration is rated accurate to within 1%.

Chapter XIII

TESTING AND SERVICING TEST OSCILLATORS

THERE are many troubles which may develop in commercial signal sources designed for use in servicing receivers, most of which may be corrected quite simply if the operation of the instrument is thoroughly understood. There are fewer parts, fewer circuits in the average test oscillator than in even the simplest radio receiver sold commercially, so it should not ordinarily be necessary for any serviceman to deprive himself of the use of the instrument by returning it to the factory if its performance becomes unsatisfactory. The necessary tests can be made with apparatus which you have on hand and in most cases the cause of defective operation can be quickly isolated and corrected.

It is understood, of course, that instruments which become unsatisfactory during the warranty period should be returned to the manufacturer, through the usual channels, for servicing. The difficulties with which we are primarily concerned are those which arise after a long period of use.

Frequency Instability

Every oscillator which is tunable over a wide frequency range is more or less subject to frequency instability, due to the difficulties involved in obtaining extreme accuracy and stability in such circuits. This trouble is normally present to some extent in even the best laboratory signal generators and, from a practical standpoint, has received far greater attention in the service field than is warranted. Since the average man seldom, if ever,

resets a service-type signal source, be it a test oscillator or a signal generator, to within 0.1% of the dial-calibrated frequency, it is illogical to condemn a signal source intended for the service field if its frequency instability is found to be of the order of .075 to .1%. The more expensive the signal source, the more stable should it be.

Frequency instability may be either electrical or mechanical in nature. Electrically, a certain amount of frequency drift is due to slight changes in the components of the oscillating circuit during the warming-up period. After stable temperature has been reached, frequency changes from this source should be negligible.

A change in line voltage in a-c operated test oscillators normally has but slight effect on the oscillator frequency. This is shown by the data tabulated below, which represent measurements made on a typical service-type test oscillator employing an electron-coupled oscillator circuit. Before varying the line voltage, the test oscillator was allowed to operate for about one hour, so that normal warming-up drift could be eliminated as a source of frequency change.

| <i>Freq. Range</i> | <i>Osc. Freq.</i> <i>(at 115 v.)</i> | <i>Line Voltage</i> | <i>Freq. Change</i> | <i>% Change</i> |
|--------------------|---|---------------------|---------------------|-----------------|
| 625-1550 kc | 750 kc | 100 | 75 cycles | .01 |
| “ “ | “ “ | 130 | 135 “ | .018 |
| “ “ | 1500 kc | 100 | 65 cycles | .004 |
| “ “ | “ “ | 130 | 295 “ | .02 |
| 1550-4500 kc | 1750 kc | 100 | 105 cycles | .006 |
| “ “ | “ “ | 130 | 215 “ | .012 |

After three hours operation, the change in frequency with change in line voltage was again checked at 750 kc and found to be less than .002% when the line voltage was varied from 100 to 130 volts. A period of 5 minutes was allowed to elapse in all cases listed above before the frequency change was measured.

Tests made at frequencies as high as 4 mc showed that frequency changes as a result of line voltage changes are entirely negligible. Over a range from 100 to 130 volts the maximum deviation was of the order of .05%.

Testing For Oscillator Drift

When abnormal conditions cause the test oscillator frequency to drift to a noticeable degree, it will be apparent in the application of the instrument to receiver alignment. If it is found that the test oscillator must be continually returned to the alignment frequency, when checking a receiver which performs normally, then it is evident that some defective condition is present in the oscillator.

However, if a superheterodyne receiver is used for the test, it is always possible that the oscillator in the receiver is drifting and not the test oscillator. This may be checked by tuning the receiver to a standard broadcast station and noting if retuning is necessary during a period of operation. A receiver employing a tuning eye, or other indicator, should be used for the purpose if *ave* is incorporated in the set.

To check the degree of drift, it is possible to eliminate the set oscillator of the receiver as a cause by using the test oscillator to beat against the broadcast signal or the signal from another oscillator, preferably of the crystal-controlled type, which is known to be stable.

A simple method of making this test, which also gives an accurate check on the test oscillator frequency, is shown in Fig. 13-1. A receiver, preferably one employing a tuning eye, is set

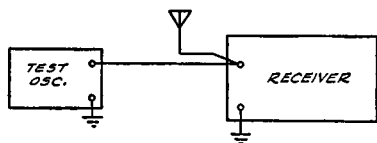


FIG. 13-1. Frequency drift may be checked by feeding the test oscillator signal to a broadcast receiver and zero-beating this signal against a broadcast signal.

up and connected to an antenna. A broadcast station is then tuned in. The test oscillator is placed in operation and allowed to warm up for a half hour or so. Then it is connected to the receiver and adjusted to the same frequency as that of the broadcast station.

As the test oscillator frequency approaches that of the broadcast station a high-pitched squeal will be heard which will decrease in pitch as the oscillator frequency more closely approaches that of the broadcast signal. Eventually a point will be reached

where this beat note becomes inaudible. This is the *zero-beat* condition and tuning the test oscillator ever so slightly either side of this point will make the note again audible. When the frequency of the test oscillator is within a few cycles of that of the broadcast signal, the tuning eye on the receiver will open and close at a rate corresponding to the frequency difference. By calibrating the test oscillator must be continually retuned to the alignment calibration within 50 parts in a million, since that is the rated accuracy of the broadcast signal frequency, at the time of writing.

Once this zero-beat condition has been reached, it is a good idea to make certain that the test oscillator is not beating with the set oscillator rather than the broadcast signal. If the beat is between the broadcast signal and the test oscillator, which is the desired condition, then detuning the receiver slightly will not cause the beat note to change in pitch, though it may become weaker. If the beat is between the set oscillator and the test oscillator, the beat note will change when the receiver tuning is changed.

If the oscillator frequency drifts abnormally, the beat note will reappear and rapidly increase in pitch until it becomes inaudible, after the zero-beat condition has been momentarily obtained. Then it will be necessary to retune the test oscillator until the zero-beat indication is again obtained. By noting on the test oscillator dial calibration the difference in kilocycles between the point at which zero beat was originally obtained and that to which it was necessary to tune to restore the zero-beat indication, the degree of frequency drift may be determined. This is a rough method but is suitable for the purpose. In the laboratory data presented, the frequency drift was measured by comparison with a known audio frequency, obtained from a beat-frequency oscillator, using an oscillograph for frequency comparison.

Mechanical Frequency Instability

Frequency instability which is mechanical in nature can be detected by checking the test oscillator frequency in the manner described above. Then, after the point on the dial calibration is noted at which zero-beat is obtained, the test oscillator is tuned to each end of its tuning range and back to the original

point. If it is found that the point on the dial calibration at which the oscillator must be reset differs materially from that of the original point to which it was set, then the frequency instability is mechanical, assuming that the previous electrical test has shown no abnormal drift.

Some of the possible causes of mechanical frequency instability are defective tuning condenser bearings, dial drive backlash, warping of the chassis or panel due to abuse, a loose lock screw on the condenser shaft or loose connections in any portion of the oscillator system. A general inspection and tightening of all screws and nuts is often all that is necessary to correct such conditions.

Correcting Electrical Frequency Instability

Erratic frequency changes, after the warming-up period, may be caused by abnormal changes in the characteristics of the oscillator tube or in the components of the oscillator circuit. The tube is best checked by substitution. Uniformity of oscillation over the frequency range may be checked by measuring the d-c voltage developed across the oscillator grid leak; this is best done with an electronic voltmeter, fitted with an isolating resistor at the test probe point. Alternatively, the alternating voltage across the tank circuit may be measured while the tuning condenser is varied over its range.

Trimmer condensers are subject to capacity change due to moisture absorption, and a loose trimmer screw may cause considerable frequency instability. These may be checked by temporarily disconnecting the suspected condenser from the circuit and rechecking for drift in the manner previously described.

Resistors may change in value during operation; these may be checked while warm, removing the oscillator tube so that no conduction current influences the ohmmeter reading. Small variations between the warm and cold readings of carbon resistors of the order of 10%, are normal but wide variations are possible causes of unsatisfactory performance.

Checking Frequency Calibration

Most test oscillators are equipped with trimmer condensers for recalibration purposes. Over the standard broadcast band, the test oscillator may be checked against broadcast stations and it is usually possible to find some station near an alignment frequency of the test oscillator. At higher frequencies, this is more difficult and it is desirable to have a crystal oscillator or some other type which supplies a highly-stable output for calibration purposes.

Crystal oscillators which supply frequencies of 100 kc and 1000 kc are most convenient for this purpose. Harmonics of these frequencies may be used for calibration purposes well beyond the range of the average all-wave receiver.

An all-wave receiver which has previously been carefully aligned is desirable for calibration purposes, and is used in the same manner as described for checking frequency instability. Instead of using an antenna and a broadcast signal, the signal from the crystal oscillator becomes our standard. Harmonics of the 100-kc fundamental will occur at 200 kc, 300 kc and every 100 kc higher, decreasing in strength at the higher frequencies. Harmonics of the 1000-kc crystal will occur 2000-kc, 3000-kc, etc.

It is best to start checking the test oscillator frequency on the standard broadcast band. After alignment has been done on this band, others may be more easily checked.

To check alignment at 1500 kc, place the 1000-kc crystal in operation and pick up its signal on the receiver. This signal should preferably be unmodulated and its presence will be indicated by closing of the tuning eye—if used on the receiver—or by a decrease in the receiver noise level due to AVC action when the signal is present.

After this point has been checked, switch to the 100-kc crystal. Its 10th harmonic should coincide precisely with the 1000-kc point to which the receiver was tuned. Now tune the receiver to the 11th, 12th and succeeding harmonics until the 15th has been reached, counting each in turn. This is done so that we may be certain we are tuned to the 15th and not the 14th or 16th harmonics, which will be only 100 kc separated from the desired frequency. Thus we do not have to rely on the frequency calibration of the receiver.

Once the receiver is tuned to the 15th harmonic, the test oscillator is tuned to zero beat in the manner previously described. If the dial calibration is off frequency, the pointer may be reset and the band trimmer adjusted until zero beat is restored. Other points may now be checked by tuning the test oscillator and receiver to lower frequencies in successive steps of 100 kc and checking the dial calibration of the receiver and test oscillator against the crystal harmonics.

Low-frequency ranges, beyond the bands included in the receiver, may be checked by beating harmonics of the test oscillator against those of the crystal oscillator. If the test oscillator is tuned to 100 kc, for instance, its 6th harmonic may be used to zero-beat against the corresponding 6th harmonic of the 100 kc crystal. At 200 kc, the test oscillator 3rd harmonic will likewise beat against the crystal 6th harmonic.

This method may be used for intermediate points, such as 150 kc. If the test oscillator is adjusted to this frequency, its 4th harmonic will beat with the 6th harmonic of the crystal oscillator. When the 700-kc 7th harmonic of the crystal oscillator is tuned in on the receiver, a new series of low-frequency test points are available, the 6th harmonic of the test oscillator fundamental of 116.67 kc, the 5th of the 140 kc fundamental, and so on may be used to provide additional points at which the test oscillator fundamental frequencies may be checked. The common i-f point of 450 kc will provide a second harmonic at 900 kc to beat with the 9th harmonic of the 100-kc crystal.

After the calibration of the low-frequency and broadcast bands have been checked, the next higher frequency band may be calibrated. Using the 3-mc point, the 1000-kc crystal may be employed, its 3rd harmonic beating with the 3-mc fundamental of the test oscillator. When frequencies of the order of 10 mc are to be checked, it is advisable to check the order of the harmonic to make certain that the proper one is being used. Confusion may arise due to the image response of the superheterodyne receiver, which will provide two frequencies differing from the desired frequency by twice the intermediate frequency employed in the receiver. Thus, if the receiver is tuned to 10 mc and its intermediate frequency is 450 kc, a response will be obtained from the 9th harmonic, or 9 mc, when the receiver tuning is

changed to 9.9 mc—only 100 kc removed from the desired point.

To avoid this difficulty, after tuning the receiver to 10 mc, tune the test oscillator to its previously-checked calibration point of 5 mc and note if its second harmonic provides a beat with the 10-mc signal. If the 9th and not the 10th harmonic of the crystal is present, then the beat will not be present.

Checking of higher frequency points should be done in the same manner. While it is a great convenience to have at least two crystal-calibrated frequencies to check against, by carefully following the method described above a single crystal-controlled frequency may be employed.

It is well to check for frequency shift in the oscillator due to the application of modulation. When using any test oscillator with variable percentage modulation, it may be found that a change in the carrier frequency will result when the modulation percentage is high. If this is appreciable, it will be detected by connecting an output meter to the receiver and noting if retuning of the receiver is necessary in order to reach a peak as the modulation percentage is increased. If such is the case, then low percentage modulation should be used when checking the calibration of a receiver against the test oscillator. However, it is advisable not to depend upon any test oscillator for extreme accuracy in frequency checking, if other means, such as broadcast stations or crystal oscillators, are available.

Eliminating Harmonics in Beat-Type R-F Signal Generators

In test oscillators which are frequency-modulated for oscillograph alignment work, two oscillators are frequently employed, the resulting beat between them producing the desired signal frequency. Over a portion of the range the difference frequency is used, while at the higher frequencies, the sum frequency is usually adopted. Thus, to produce a 1000-kc signal, one oscillator may be operating at 800 kc while the other is functioning at 1800 kc. The difference frequency which thus results is 1000 kc.

This type of test oscillator is most desirable when frequency modulation is being used, since it permits frequency modulation over a uniform band at any desired alignment frequency. How-

ever, when aligning at higher frequencies on short-wave bands with amplitude modulation, some confusion in identifying the proper frequency may be caused by harmonics of the fixed oscillator appearing near the desired beat frequency. This difficulty may be overcome by rendering the fixed oscillator inoperative when amplitude modulation is being used and by modulating the variable frequency oscillator directly. When this is done, it will be necessary to add a frequency equal to that of the fixed frequency oscillator to the dial reading over bands where the difference beat frequency is employed and to subtract the same amount from the dial reading at frequencies where the sum frequency is used. Instructions for making the change may be usually obtained from the manufacturer.

Modulation Troubles

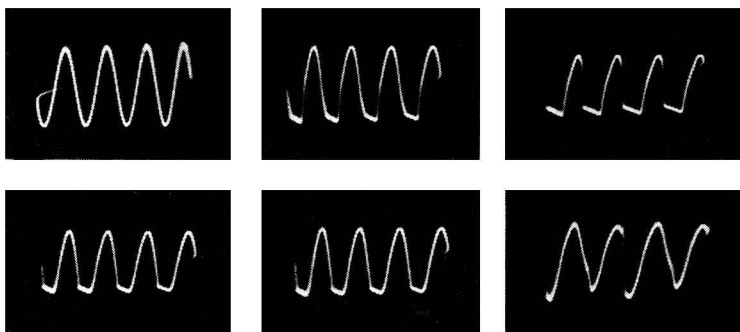
In the past little attention has been paid to the waveform of the a-f modulating oscillator in commercial test oscillators. And, if the instrument is used solely for alignment purposes with an output meter, considerable distortion may be tolerated without affecting its usefulness, provided the percentage modulation is not too high. However, when such oscillators are used in conjunction with a cathode-ray oscillograph for distortion checking, a waveform of better characteristics is required. If the a-f waveform is bad, then faults in the receiver circuit which cause distortion are more difficult to recognize.

Most of the better service test oscillators now on the market provide an a-f oscillator circuit which normally has very acceptable waveform. If distortion develops due to variations in circuit components, it will be immediately apparent because the character of the tone will vary. If the normal frequency is 400 cycles, a decrease in the capacity of the condensers which shunt the a-f tuned circuit, will cause the frequency to increase. A shorted turn in the oscillator tuning coil will greatly reduce its inductance and consequently the resonant frequency will increase or the oscillator will become inoperative. If the oscillator grid leak increases in value, the loading effect on the tuned circuit will be reduced and the oscillator frequency will be lowered. A

defective audio-oscillator tube also will cause a frequency change; this should be checked first by substitution.

Effect on Oscillator Operation of Grid Leak and Grid Condenser Variations

The oscillograms shown in Figs. 13-2(a) to 13-2(f) inclusive illustrate, in a service-type test oscillator, the effects of variations in the values of the r-f oscillator grid leak and its grid condenser on the waveform of the a-f component. The test oscillator, the circuit of which is shown in Fig. 13-3, employed a tickler-feedback circuit and plate modulation was used. The normal values of the grid leak $R1$ and grid condenser $C1$ were 50,000 ohms and 100 $\mu\mu\text{f}$ respectively.



Figs. 13-2 (a) to 13-2 (f), left to right. How variations in the values of the r-f oscillator grid leak and condenser affect the waveform of the a-f component.

The normal waveform of the 400-cycle component is shown in Fig. 13-2(a) and is seen to be substantially a sine wave. In Fig. 13-2(b), the effect of an increase in the grid leak resistance to 2 megohms is shown. This could result in service due to a partial open in the grid leak circuit. This wave, you will note, is badly distorted. Fig. 13-2 (c) illustrates the waveform secured when the grid leak is 2 megohms and the grid condenser capacity is reduced to 11 $\mu\mu\text{f}$. In these two oscillograms, the grid condenser was shunted by the grid leak.

In Fig. 13-2(d), the grid leak is likewise 2 megohms and the grid condenser $11 \mu\mu\text{f}$, but the grid leak is returned directly to the cathode. Note that the wave is substantially the same as that of Fig. 13-2(c), showing that these effects may be obtained no matter which form of grid leak connection is used.

In Fig. 13-2(e), the normal grid condenser of $100 \mu\mu\text{f}$ is used in combination with a 20,000-ohm grid resistor.

These oscillograms illustrate that a bad a-f waveform is not necessarily the result of defects in the a-f oscillator circuit, since all these conditions represent effects of changing constants in the r-f oscillator circuit.

R-F Voltage Variations

An increase in the value of the grid resistor R_1 in Fig. 13-3 causes a decrease in the r-f voltage E_1 across the tuned circuit. With the normal grid leak of 50,000 ohms, E_1 varied from 11 to

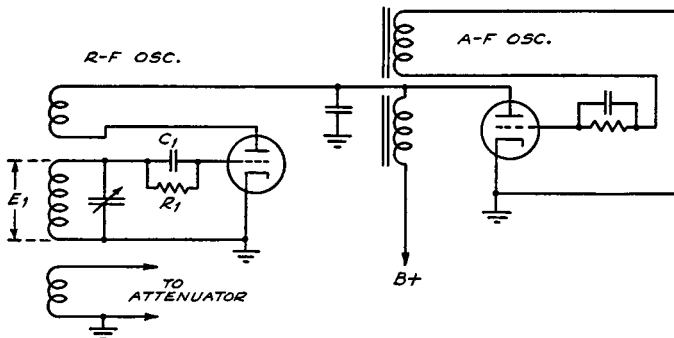


FIG. 13-3. The circuit of the test oscillator used for these tests. The r-f oscillator is plate-modulated.

13.7 volts as the oscillator was tuned from 200 kc to 400 kc. Over the same range, when the grid leak was increased to 2 megohms, the oscillator voltage varied from 5.6 to 7.9 volts. The same effect was noted when the grid condenser C_1 decreased in value. Using $10 \mu\mu\text{f}$ instead of $100 \mu\mu\text{f}$, the oscillator output varied from 8.8 to 8.9 volts over the same tuning range. Note that the uni-

formity of output was considerably improved with the smaller value of grid condenser, though the r-f output dropped somewhat.

However, this applies only when the grid condenser is shunted by *R1*. When the grid resistor returns to the cathode directly, the oscillator will not function at such low frequencies with a grid condenser smaller than 20 μf .

Increasing the size of the grid condenser to .001 μf or larger caused increased distortion of the modulated wave, though minor variations in capacity had no noticeable effect. A shorted grid condenser caused a considerable decrease in r-f output but had little effect on the waveshape of the a-f component.

If the oscillator grid leak becomes open-circuited, or increases to a value in excess of 2 megohms, the oscillator is likely to motorboat.

Leakage

All signal generators and test oscillators are subject to some leakage, particularly on the high-frequency bands. This makes it difficult to obtain sufficient attenuation of the signal to align the receiver by customary methods, since it is obvious that the attenuator cannot lower the signal below the level of the signal which is radiated directly to the receiver. In some instances, a balancing effect may be obtained by operation of the attenuator; this is considered on page 173.

To check for excessive leakage in a test oscillator that formerly gave no trouble, simply connect a wire to the antenna post of a sensitive receiver and with the test oscillator in operation, hold the wire near joints in the test oscillator cabinet, around the dial, and note the relative strength of the signal at each point. If the radiation is severe, it may be necessary to ground the avc system in the receiver to detect differences in signal strength. If the signal is much greater at one point than at another, examine the cabinet contacts and make certain that a clean joint is obtained, since corrosion will affect the shielding efficiency. Often, due to accidental knocks, the shield case will not fit properly and signal leakage will occur through the resulting gap.

Modulation Percentage

The percentage modulation depends upon the ratio of the modulating voltage to that of the carrier signal voltage. Since the 400-cycle modulation voltage remains fixed, while the carrier voltage often varies to a considerable degree as the r-f oscillator frequency is varied, it follows that the ratio of carrier-signal voltage to the modulating voltage does not remain fixed. Therefore, in most test oscillators, the modulation percentage will vary at different carrier frequencies. This is of no particular importance insofar as general service tests are concerned, provided the modulation percentage does not rise too high.

However, if the modulation percentage is 30 at one carrier frequency and the carrier voltage increases greatly when tuned to another frequency, the modulation percentage will be greatly reduced. Under such conditions, it will be difficult to align a receiver in the conventional manner because when the carrier signal level is made low enough so that the avc system in the receiver is ineffective, the a-f output at the point to which the output meter is connected may be so low that the a-f signal will not override the noise and hum levels.

On the other hand, if the carrier voltage becomes lower while the modulating voltage remains constant, the modulation percentage may increase to the point where serious distortion results, thereby creating a "broad" signal, due to frequency modulation, or one with a double peak which likewise affects alignment. Distortion in the modulated signal interferes with distortion checking in a-f circuits.

If the test oscillator is fitted with a control to vary the modulation percentage, it is of course possible to overcome these troubles by simply readjusting the control. A method of determining the modulation percentage by means of a cathode ray oscillograph has been described in Chapter X. Another, using the v-t voltmeter, is described in the Appendix.

Attenuator Troubles

The difficulties which occur in test oscillator attenuators as a result of long-continued use are similar to those which happen to

similar components used in radio receivers. For instance, we have a variable control, often a simple potentiometer, which functions in a manner similar to a volume control in a receiver. Volume controls can become noisy, open-circuited or change in resistance value. Where switches are employed, we are likely to have troubles due to wear which causes poor contact tension or corrosion, which results in erratic operation.

A variable control which does not give a smooth variation in test oscillator output is best replaced. If a multi-point switch is used for the multiplying range, erratic operation resulting from corrosion can often be corrected by cleaning the contacts with carbon tetrachloride.

Failure to secure proper attenuation at very low signal levels is often due to trouble in other portions of the circuit, rather than the attenuator. If the grounding points of the oscillator circuit develop high-resistance joints, if the shields and the cabinet do not have good, clean contacts, as was previously described, leakage of the oscillator signal will be increased and an increased potential difference at the oscillator frequency may develop between the attenuator ground point and that of the oscillator. This results in a condition whereby minimum attenuation may be secured at some point other than the minimum setting of the attenuator because of a difference in phase between the signal in the shield and that at the output terminals of the attenuator. This trouble will exist to a certain extent in all attenuators, particularly at very high frequencies, but if the degree of trouble is greater than was originally present in the instrument, servicing along the lines indicated will be required.

Checking Attenuator Calibration

Many of the higher-grade test oscillators are fitted with ladder attenuators, which multiply the output signal strength by factors of 1, 10, 100, etc. A variable control, which feeds the ladder attenuator, is often calibrated in a continuous range from 1 to 10.

Due to reactive and other effects at high frequencies, these attenuation ratios may vary at different frequencies. This is particularly the case on short-wave bands, where an accuracy

within 25 percent of the calibrated value is considered sufficiently good for measurements of laboratory precision.

A simple method of calibrating such attenuators, and others which have an output impedance in the neighborhood of 10 ohms, is shown in Fig. 13-4. This method utilizes a radio receiver—one of the older types which employs a cathode voltage control of the r-f and i-f tubes to vary its sensitivity. The set should not employ *avc*; if *avc* is incorporated, the *avc* bus should be grounded. If this causes instability, a C-bias of a few volts may be inserted in the grid return leads of one or more of the tubes until stability is restored.

To make this calibration, the knob of the receiver sensitivity control is removed and a large white card is fastened over the control shaft. A pointer-type knob is now substituted for the ordinary knob. An output meter is connected to any desired portion of the audio system, or to the second detector, if a tuned v-t voltmeter is used. This is desirable, particularly if the receiver is none too sensitive, because of the high sensitivity of this type of voltmeter.

A calibrating attenuator, which may be made up of carbon or other non-inductive resistors, having the values indicated on the diagram is then installed in a shielded box with taps brought out to insulated pin jacks, as shown. The 90-ohm section, *R2*, should be kept at least an inch away from the shield box to avoid shunt capacity effects. When a signal voltage is impressed across *R1*, the output voltage across *R3* should be one-tenth that across *R1*. (If the output resistance of the test oscillator is 10 ohms, *R1* may be omitted.)

To make the test, the receiver and test oscillator are tuned to resonance at a frequency of about 600 kc and the test oscillator output is adjusted until a reference reading is obtained on the output meter when the sensitivity control of the receiver is at maximum and the antenna post is connected—through the dummy antenna—to point *A*. The position of the sensitivity control is marked (1) on the white card at the pointer tip. The output meter reading should be noted; this is the reference indication.

Without changing the test oscillator adjustments, move the lead at point *A* on the attenuator to point *B*. The output meter reading should increase. Now turn the sensitivity control until

the reference indication on the output meter is restored. The receiver gain is now 1/10 that which was originally obtained, so a signal 10 times as strong is required to produce the same output. This pointer position is accordingly marked (10) on the white card.

Now reconnect the dummy antenna lead to point A, leaving the sensitivity control pointer at 10. Increase the test oscillator output until the original reference indication on the output meter is restored. Then again move the dummy antenna lead to point B, reducing the sensitivity control pointer setting until the output meter reference reading is again re-established. Mark this last point 100.

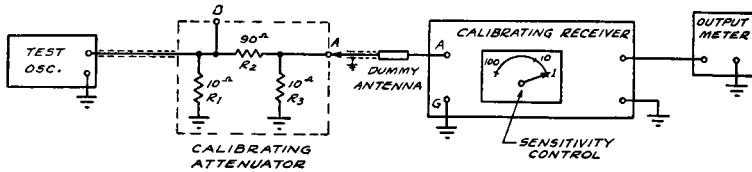


FIG. 13-4. How attenuators can be calibrated by using a receiver and a simple calibrating attenuator.

We now have calibrated our receiver so that it will give gain ratios of 1/10 and 1/100 of maximum. To utilize this calibration, connect the test oscillator directly to the dummy antenna, removing the calibrating attenuator from the circuit. Set the sensitivity control pointer at (1) and, with the test oscillator multiplier at X1, adjust the variable control until the initial reference reading of the output meter is obtained.

Now set the calibrated receiver sensitivity control at (10). The receiver is now 1/10 as sensitive and will accordingly require 10 times the signal voltage to produce the initial reference output meter reading. Switch the test oscillator multiplier control to the X10 position. The initial output meter reading should now be restored. If not, then the multiplying factor may be adjusted by changing the setting of the variable control on the test oscillator until the proper factor is established; this should be noted for future reference. The X100 point may be checked in the same manner.

This calibration will apply only to one frequency and gain tests should be made on this basis.

The variable control on the test oscillator may be calibrated over a range of one to ten by feeding the output of the test oscillator into a tuned v-t voltmeter with a calibrated level control. The v-t voltmeter level control is set at 10 and the test oscillator output control is adjusted until a reference level reading is obtained on the tuned voltmeter indicator. This point is marked on the test oscillator control calibration. Additional points, from 1 to 10, are obtained in the same manner. These calibrations may be made without using the isolating probe furnished with such instruments if the output of the test oscillator is insufficient to enable calibration with the usual probe. Multiplier ranges, at signal levels above 10 millivolts may be checked when an isolating probe is used, by direct connection to the signal-tracing instrument.

If no signal-tracing instrument is at hand, the calibration of the variable control on the test oscillator can be made by making up $R3$ of the calibrating attenuator with ten 1-ohm resistors, bringing out taps at each junction. The receiver may be calibrated at these additional points, each representing steps of 1 instead of 10, in the same manner as has been described above.

Checking Frequency-Modulated Test Oscillators

Frequency-modulated oscillators are described in detail in Chapter XII. Those of the electronic type, employing a fixed and variable frequency oscillator, are more complex than those which employ a motor-driven variable condenser in shunt with the main tuning condenser.

When either of these types is used as an amplitude-modulated oscillator, the problems which arise in servicing them do not differ materially from those considered in the foregoing paragraphs. When frequency modulation is used, they may have faults which are not obvious in use but which will affect their ability to enable proper alignment of receiver circuits.

In the electronic type, the fixed-frequency oscillator usually functions at some frequency near the middle of the standard broadcast band and its operating frequency and stability may be

checked by simply tuning a receiver to its frequency and beating a harmonic of a crystal oscillator or a broadcast station against it in the manner described for testing frequency stability for amplitude-modulated test oscillators. This test may be made without removing the instrument from its case. The fixed-frequency oscillator is readily identified since variation of the main tuning condenser will cause no change in its frequency. The initial test should be made without frequency modulation.

In an electronic frequency-modulated oscillator, an incorrect curve can result from changes in the adjustment of the phase-angle compensation network. Instructions for correcting such troubles are given in the manuals which accompany each type of instrument.

In motor-driven frequency-modulated oscillators sputtering noises which are apparent in the speaker output of the receiver being tested, and which also affect the resonance curve, result from defective contacts of the rotating condenser. These should be cleaned and tightened.

Troubles which occur in the sweep-voltage generator can be located by checking the voltages, components and the sweep voltage generator tube.

Other circuits of frequency-modulated oscillators are checked in the same manner as described for amplitude-modulated types.

Servicing Audio Oscillators

In audio oscillators of the fundamental-frequency type, unsatisfactory operation is most often due to defective switch contacts, since these points are continually exposed to, and affected by, accumulations of dirt, corrosion and wear on the contacts. Such troubles frequently make the contact resistance so high that erratic operation or complete failure to operate at one or more of the fixed frequencies results.

Periodic cleaning of all contacts with carbon tetrachloride, tightening of contact arm tension if necessary, and storing of the instrument in a cabinet when not in use will take care of many of the normal causes of poor operation.

Since beat-frequency audio oscillators do not employ switches for frequency selection, this source of trouble is avoided. How-

ever, since audio frequencies are obtained by tuning a variable-frequency oscillator so that its frequency approaches closely that of a fixed-frequency oscillator, any undesired influences which affect the frequency of either oscillator will cause frequency instability. In practice, frequency instability is generally due to changes in tube characteristics, wide fluctuations in line voltage (of line-operated instruments) and mechanical troubles, such as wear on tuning condenser bearings, loosening of screws which anchor components to the chassis and defective grounding contacts. All these troubles which cause small variations in the circuit capacity across tuned circuits affect the frequency, and at low frequencies the percentage variation in frequency may be considerable.

It should not be expected that any beat-frequency oscillator will operate without some frequency variation when it is first put in operation. Stable operating conditions are not obtained until all components, particularly the tubes, reach a constant temperature, because heat affects the characteristics of every tube, every resistor and every coil and condenser in the receiver. To a certain extent, this is taken into consideration in the design of the instrument. Components of similar characteristics are used and the layout of the fixed-frequency and variable-frequency sections is made so that each section will be subject to the same influences, as nearly as possible. When this is done, then if any slight frequency drift occurs in the fixed-frequency oscillator approximately the same degree of frequency drift will also occur in the other oscillator, *in the same direction*. When this is the case, then the resulting beat-frequency remains constant at any frequency setting. For instance, if the fixed-frequency oscillator normally operates at 300,000 cycles and the variable-frequency oscillator is tuned to 300,030 cycles, the beat-frequency is the difference between 300,030 and 300,000 or 30 cycles. Now, if both oscillators drift 10 cycles in the same direction, the resulting oscillator frequencies will become 300,040 and 300,010, so the resulting beat frequency still remains 30 cycles.

If the characteristics of the tubes used as the variable and fixed frequency oscillators are not similar, then some instability may be contributed, particularly at low frequencies. If it becomes necessary, after a period of use, to replace one of the oscil-

lator tubes, it is a good idea to replace both, choosing a pair of tubes which gives most stable operation in the beat-frequency oscillator.

In audio oscillators of the fundamental-frequency type, the influence of tube capacity variations on the oscillator frequency is negligible, since the shunt capacity in such oscillator circuits is always very high with respect to the tube capacitance.

Checking the Frequency of B-F Oscillators

Many beat-frequency audio oscillators are equipped with a means for checking the dial calibration against the output frequency by comparison with a known frequency, such as the 60-cycle line, if the instrument is a-c operated. For battery-operated instruments, and for other beat-frequency oscillators not so equipped, a simple method of calibrating the instrument against the 60-cycle line is shown in Fig. 13-5. As indicated, a pair of

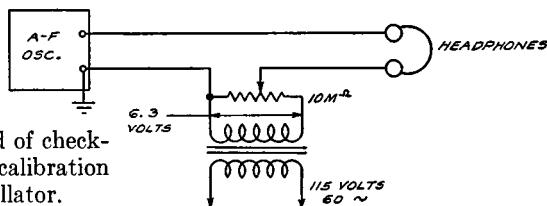


FIG. 13-5. A method of checking the 60-cycle calibration point of an a-f oscillator.

headphones is used to note the beat between the oscillator output and the 60-cycle voltage taken from the secondary of a filament transformer.

In operation, the dial pointer is set at 60 cycles and the calibration control of the a-f oscillator is adjusted until a beat is obtained between the 60-cycle note from the oscillator and that from the a-c line. As adjustment of the control approaches the proper point, the signal in the phones will increase and decrease slowly at the difference frequency between the oscillator and the a-c line. Adjustment is correct when moving the calibration control either side of the proper point causes an increase in frequency of the beat heard in the phones. During this process of adjustment, it may be necessary to adjust the output levels from the oscillator and line by varying the 10,000-ohm potentiometer and

the oscillator output control until the intensity of the oscillator signal is the same as that from the 60-cycle line.

When adjustment of the calibration control has been done in this manner, check the setting by cutting off the a-c line calibration signal and, with the phones still connected, note if slightly increasing the main tuning dial frequency setting of the b-f oscillator causes a corresponding increase in frequency in the output signal. It is possible to adjust the calibration the wrong side of

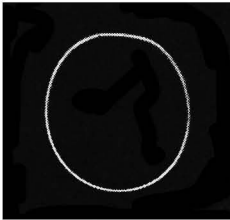


FIG. 13-6(a)

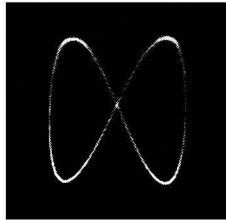


FIG. 13-6(b)

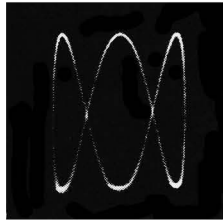


FIG. 13-6(c)

The ratio of the frequencies in Fig. 13-6(a) is 1:1; in (b) 2:1; and in (c), 3:1.

zero beat, in which case the output frequency will decrease as the dial setting is slightly increased. If this effect occurs, readjust the calibration control by finding the second point at which the desired beat may be obtained and repeating the adjustments described above.

Often tuning forks are used for calibration purposes. This is done by tuning the a-f oscillator to the frequency of the tuning fork and listening to the output signal with the phones. The tuning fork is then struck and the handle is placed against the case of one of the phones. The calibration control is then adjusted until the output signal heard increases and decreases slowly, due to the beat between the fork frequency and that of the oscillator. Adjustment is then continued until the signal rises and falls very slowly as the calibration control is adjusted either side of the point where no signal is heard. The output frequency of the oscillator is then substantially the same as that of the tuning fork.

Many other methods of checking calibration are available, employing vibrating reeds, neon tubes, etc., but these are generally

incorporated in the b-f oscillator and instructions concerning their operation accompany the instrument.

Checking the Frequency of A-F Oscillators

In audio oscillators of the fundamental-frequency type, several frequencies for each range are usually obtained by shunting condensers of different values across a single coil. In checking low

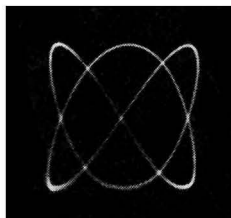


FIG. 13-6(d)

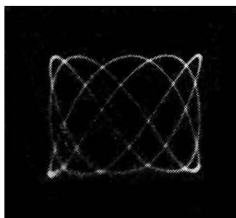


FIG. 13-6(e)

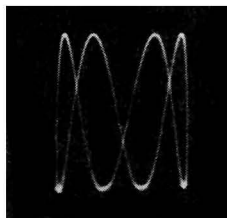
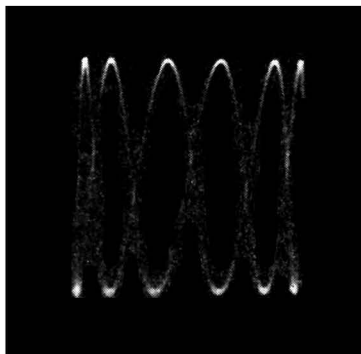


FIG. 13-6(f)

The ratio of the frequencies in Fig. 13-6(d) is 3:2, in (e), 5:3; in (f), 4:1, and in (g) at the right, 6:1.

The Lissajous figures in this series of oscillograms were formed by applying frequencies of the ratios noted to the horizontal and vertical plates of a cathode-ray tube. To determine the ratio count the horizontal and vertical peaks.



frequencies, an output frequency of 60 cycles can be checked in the manner described above. For other frequencies, Lissajous patterns, as obtained with the cathode ray oscillograph, will enable frequency checking at a variety of points, or an assortment of tuning forks may be used.

Frequency calibration by comparison with known frequencies with the cathode ray oscillograph is illustrated by the Lissajous figures shown in Figs. 13-6(a) to 13-6(h). These are made by

feeding the known frequencies to the horizontal plates of the cathode-ray tube and the unknown to the vertical plates.

When the ratio of the known frequency to the unknown is 1 to 1, an oscillogram as shown in Fig. 6(a) results. This is only one of the figures which represent a 1 to 1 ratio, since the pattern normally varies from a slanting line to an ellipse, depending upon the phase relations of the two frequencies. In practice, the forms illustrated will result during the process of calibration.

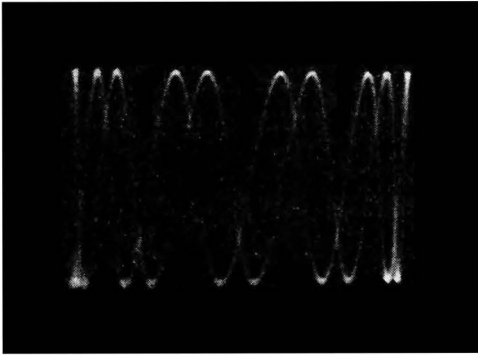


FIG. 13-6(h). The ratio of the frequency applied to the horizontal plates to that applied to the vertical plates is 10 to 1.

If the known frequency is 6 times the unknown, a figure similar to that shown in Fig. 13-6(g) will result. Thus, if the known frequency is 600 cycles and the ratio is 6 to 1, as shown, the unknown frequency is found to be 100 cycles. Other frequencies are determined similarly.

Chapter XIV

RECEIVER CHECKING WITH A TEST OSCILLATOR

SINCE the performance of any radio receiver is based on the manner in which it functions when a signal is being received, it follows that the only conclusive test of the operation of each signal-carrying circuit is one which is made when the signal is present. To make such a test, we may feed the signal to the receiver antenna and ground terminals and examine, with special signal-tracing instruments, the strength and character of the signal in any portion of the receiver. Alternatively, we may feed the signal to any desired portion of the receiver and note the signal strength required to produce the desired output signal level. The former method is known as signal tracing and is described in detail in the author's book, "Servicing By Signal Tracing." The latter method, in which the test oscillator plays the leading part, is called the test oscillator-output meter method, and will be discussed in this chapter.

To understand how the test oscillator-output meter method of receiver checking is applied, let us assume that we have before us an inoperative receiver, the skeleton circuit of which is represented by the diagram, Fig. 14-1. We have already checked the power supply "B" voltage and found it normal. Our attention must therefore be devoted to those circuits and tubes in the receiver which act upon the signal.

Now, in normal operation, the signal will be picked up by the antenna and fed to the input of the receiver. Thence it will progress through the r-f stage, first detector, i-f stage, second detector and from the a-f stages to the speaker. Each of these stages is linked or coupled to the following stage and, if any one

of these stages fails to function, then the signal cannot continue along its normal path to the speaker. If the output stage fails to pass the signal, then it matters not how well the preceding stages are functioning, no signal will be produced by the speaker. Our first test, then, is to check the operation of the output stage.

In operation, an audio signal should appear at the grid of the second a-f tube. Then, if the second a-f tube and the output circuit are functioning properly, a signal will be available across the speaker voice coil. So we connect an output meter across the speaker voice coil and feed an audio signal from our test oscillator—or from a separate audio oscillator if our test oscillator has no provision for supplying a 400-cycle signal for external use—to point 1. If the output stage is operating, we should obtain an indication on our output meter; also, the speaker should reproduce the audio signal. If the output meter gives a reading but there is no speaker response, then we know the trouble is in the speaker. If there is no output meter reading, then the trouble is in some component which affects the operation of the output stage. If we obtain both an output meter reading and speaker operation, we know that the trouble is not in the output stage.

Since we have now eliminated the output stage as the source of trouble, let us now feed our test signal to point 2, the first a-f grid. Again, since this is an a-f stage, we must use an a-f signal. If there is no response in the speaker, then the source of trouble is localized in some component which affects the operation of the first a-f stage. If we hear the signal in the speaker, then we have proved that the a-f section of the receiver is in operating condition.

Our next step is to check the operation of the second detector. We can best do this by feeding a signal to the grid of the i-f tube at point 3. Now, since the i-f stage responds only to signals at the intermediate frequency to which it is tuned, we must adjust our test oscillator to this signal frequency. Let us assume that we require a 465-kc signal. This signal must be modulated, since we want to check the operation of the detector and also because an unmodulated signal will not pass through the a-f amplifier and provide a reading on our output meter and a signal in the speaker. This signal must also be strong, since it is amplified only by the single i-f tube and circuit before detection.

Again we note if the speaker responds when the signal is applied. If not, then we have localized the trouble to some component which affects the performance of the second detector or i-f stage. If the signal is heard, then we may again proceed with our tests.

At point 4, we may again feed a signal at the intermediate frequency to the first detector. The fact that a signal fed to this point is reproduced by the speaker eliminates the first detector as a source of trouble—but not the set oscillator. Since this is a mixer tube, our next test is to feed a signal at the frequency to which the receiver dial is tuned to the same point 4. If no signal is reproduced by the speaker, then we know that the first detector

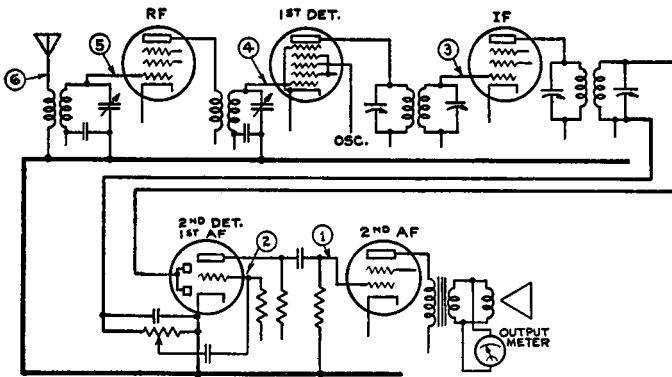


FIG. 14-1. A skeleton schematic of a superheterodyne on which the numbered points of test are indicated.

is not performing its function of converting the r-f signal to an i-f signal. Since our first test of this circuit showed that its performance was normal at the intermediate frequency, the defect is now localized in the oscillator section. On the other hand, if the first detector performs normally when either the intermediate frequency or the r-f signal is applied, then the trouble is localized in the last remaining circuit, the antenna coil.

This, then, is the general procedure for checking inoperative receivers and the same order of test is observed when checking for other defects by this method.

This has been a simple case, because we have concerned ourselves only with the ability of each stage to pass a signal. This is all that is necessary when the set is inoperative, because we know that if each stage passed the signal to the succeeding stage, it would eventually appear at the speaker and the set would therefore operate. Since we need not know the strength of the signal which we feed to the various stages, the test oscillator attenuator need not be calibrated. Neither are we concerned with the signal waveform, modulation percentage or frequency calibration. For other tests, however, we shall see that the requirements of our test oscillator are more exacting.

Now let us assume that the receiver which is under test is not inoperative, but low in sensitivity. Since each stage contributes to the strength of the output signal at the speaker, we know that at least one such stage is not functioning at full efficiency. Perhaps the operating voltages are incorrect, perhaps some of the tuned circuits require alignment . . . these are common causes of low sensitivity when the tubes test normal and may be located and corrected by usual test procedures. But even when the operating voltages are correct, and the circuits are properly aligned, other troubles may cause poor sensitivity. If an audio coupling or an r-f bypass condenser becomes open-circuited, if a transformer has become affected by moisture, if leakage develops in a socket or in wiring . . . these are only a few of the conditions which will affect the amplification of a stage which are not revealed by ordinary test methods.

The only conclusive test which embraces all factors which contribute to the proper amplification of the signal is a stage gain test.

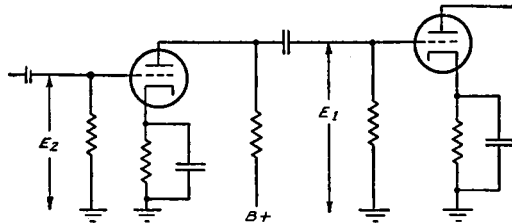
How are we to tell what the gain in any particular stage of a given receiver should be? This is a question you may well ask. In the past, manufacturers have not supplied such data, largely because most servicemen have not been equipped to measure stage gain. Now that suitable equipment is available and is becoming more and more widely used, such information is being supplied, as shown at the end of this chapter. In addition, you will find a Table of Average Gain-Per-Stage Values which represents the average of several thousand receivers now on the market. It is not necessary that you make a precise measurement

of stage gain in trouble shooting, because if the complaint is weak reception there is usually a loss, rather than a gain, in the defective stage. There is such a reserve of sensitivity in most receivers that small variations in stage gain will pass unnoticed insofar as the customer is concerned. However, to achieve the very best results, the stage gain should fall within the limits given in the Table.

How Stage Gain Is Determined

The gain in any stage is the ratio of the signal voltage at the output of the stage to that which is applied to its input. For example, let us consider the typical audio amplifying stage shown in Fig. 14-2. The signal voltage we apply to the grid of the first

FIG. 14-2. A typical a-f amplifier stage in which the gain is found by dividing E_1 by E_2 .



tube we shall call E_2 ; the resulting amplified signal which is passed on to the next stage is designated as E_1 . The ratio of the output signal voltage E_1 to the applied input signal voltage E_2 , or E_1/E_2 represents the gain of the first stage. If E_2 is one volt and E_1 is 10 volts, then the stage gain, E_1/E_2 , is 10. If the input signal voltage were 0.1 volt, and the output signal voltage were 1 volt, the ratio would still be 10 and the gain would therefore be the same. In fact, we are not concerned with the actual value of the input voltage so long as it is not so great as to overload either the input or output tube; all we need to know is the ratio of the output signal voltage to that which is applied to the stage input. This is particularly important in r-f and i-f circuits, for it means that we do not need to know the precise signal voltage in microvolts provided we can determine how much stronger the signal is at one point than at another.

Equipment Requirements for Stage Gain Measurements

At this point it will be apparent that we need some means of indicating relative signal levels if we are to measure stage gain. Insofar as audio circuits are concerned, this requirement involves no particular difficulty since we may use a simple external attenuator in conjunction with our signal source, if no calibrated control is included in the instrument. Also, we may measure signal voltages directly with a sensitive v-t voltmeter.

In r-f circuits, however, direct measurements of normal operating signal levels are impractical unless special signal-tracing instruments are used and if such equipment is at hand, it is better to make all gain measurements with it. If we confine ourselves to the test oscillator and output meter, then the test oscillator must be equipped with an accurately-calibrated attenuator.

The output meter may be of the copper-oxide rectifier type, and should be sufficiently sensitive to measure accurately alternating voltages of one to two volts. If a sensitive v-t voltmeter, capable of reading voltages of .05 to .1 volt, is used, then the a-f attenuator need not be calibrated. Such an output meter need cover only the audio range.

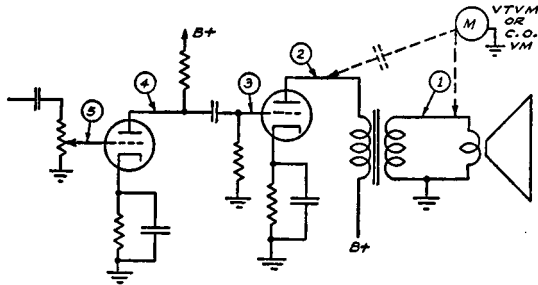
Tests in audio circuits are most conveniently made with either a beat-frequency a-f oscillator or one of the fundamental-frequency type. When such equipment is available, it is possible to check at many frequencies in the audio range thus determining the fidelity of the a-f amplifier. Speaker rattles and other troubles which may be particularly evident at one audio frequency but not at another, are revealed by checking over a wide range of audio frequencies.

Measuring Audio Stage Gain

Since stage gain measurements by the test oscillator-output meter method normally start at the output stage, let us first consider the a-f amplifier circuit shown in Fig. 14-3, in which a triode output stage is preceded by a single-stage triode amplifier. We may assume that this amplifier represents a portion of a receiver which is under test for weak reception.

The gain in the output stage is a power gain rather than a voltage gain, since the output voltage which appears across the voice coil is normally lower than that which is applied to the grid of the output tube. While the signal voltage at the grid is amplified and there is a voltage gain across the primary of the output transformer, the step-down ratio of the output transformer is normally greater than the voltage step-up which results from the amplification secured with the tube. In an output triode, the signal amplification is approximately equal to two-thirds of the amplification factor of the tube. Since output triodes have low amplification factors, of the order of 3 to 5, this means that the voltage amplification is only about 2 to 3. With output pentodes,

FIG. 14-3. By connecting a suitable voltmeter as indicated, the performance of the a-f amplifying stages can be easily checked.



the voltage gain is normally higher, ranging from 10 to 15, but even then, since the output transformer step-down ratio is often greater than 30 to 1, an overall stage loss results in terms of voltage measurements.

To check the voltage gain from grid to plate of the output tube, we may connect our output meter, through a blocking condenser, from point 2 to ground. Then we may apply a known audio voltage from point 3 to ground. The ratio of the audio voltage reading on the output meter to the audio voltage at the grid is the voltage gain due to the tube and its plate load. The output meter for this type of measurement should preferably be a v-t voltmeter, since a copper-oxide type of meter of the 1000-ohms-per-volt variety will load the circuit too much unless it is used on a high range. If the output meter reading is 30 volts and the input signal voltage is 10, the gain is 30/10 or 3.

The normal point of connection of the output meter is across the speaker voice coil, from point 1 to ground if one terminal of the voice coil is grounded. In the case given above, if the step-down ratio of the output transformer is 30 to 1, then a 30-volt signal across the primary of the output transformer will produce only a 1-volt signal across the voice coil. So we see in this example, that a signal input of 10 volts to the output stage will result in a 1-volt output. The stage gain may be expressed as this ratio, 1/10.

If a pentode were employed in the output stage, we should expect from three to five times the stage gain represented by the preceding figures; this accounts for the wide use of pentodes in modern receiver output stages. Typical values of voltage gain from grid to plate are given for various output tubes in the Table of Average Gain-Per-Stage Values on page 211.

From a trouble-shooting standpoint, the value of this test is that we are able to check the performance of the stage and its components in a very simple manner. For instance, shorted turns in the primary of the output transformer would affect not only the frequency response of the stage but also the gain; yet this trouble would not affect the operating voltages. The same applies to shorts in the secondary circuit. Moreover, troubles of this nature are not revealed by usual test methods.

Now let us check the first a-f stage. The output meter may remain connected across the speaker voice coil but the audio signal source should now be connected from point 5 to ground. If a bias cell or other form of fixed grid bias is applied to this tube, we should use a blocking condenser (about 0.1 mf) between the grid of the tube and the a-f oscillator to prevent shorting out the bias.

In our output stage test, we assumed that a 10-volt audio signal at point 3 produced a 1-volt signal across the speaker voice coil. If the first a-f stage is contributing any gain, then a smaller signal voltage applied to point 5 should produce the same output across the voice coil. Actually, if the tube is a high- μ triode, we should expect a gain of 30 to 60 in the first a-f stage. This means that a signal of .25 volt or slightly less should give us the same output signal across the voice coil as a 10-volt signal at point 3, when a triode output tube is used. If a much stronger

signal is required, then we can look for some fault in the components of the circuit between points 3 and 5.

In making this test, the volume control should of course be set at maximum to prevent grounding out the test signal.

This is a stage-by-stage test of the audio amplifier. It is often convenient for trouble localization to check the audio amplifier as a unit by a single test and manufacturers are now supplying data for this purpose. You will note on page 211 the average audio gain figures for the a-f amplifier section of the G-E Models H-600, etc. These data show that a .05-volt, 400-cycle signal across the volume control should produce $\frac{1}{2}$ watt output across the speaker voice coil . . . about 1.3 volts. The test is made in exactly the manner described for the circuit of Fig. 14-3; we need only apply a .05-volt signal to the grid of the 1st a-f tube and note if the output meter reading, when connected across the voice coil, is 1.3 volts. If appreciably less, then the amplifier gain is subnormal. By the same token, if the gain is very much greater, then we may look for regenerative effects or abnormal tubes.

If our audio oscillator is provided with no means for determining when the output voltage is .05 volt, we may use a simple voltage divider as described in the Appendix.

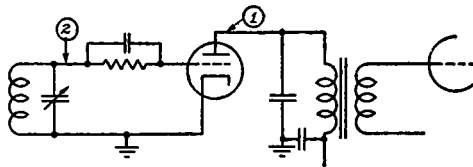
Measuring Gain in Detector Circuits

The gain in a detector circuit is expressed as follows:

$$\text{Gain in Detector} = \frac{\text{Audio output voltage across detector load}}{\text{Input carrier voltage} \times \text{Degree of modulation}}$$

In Fig. 14-4, which shows a grid-leak triode detector, the detector load is the first audio transformer and the audio output voltage is

FIG. 14-4. The test points in a grid-leak triode detector circuit are indicated, the gain being found by using the above equation.



that which is developed across the primary of this transformer when a modulated r-f or i-f signal is present in the grid circuit of the detector.

Note that in the above equation the degree of modulation is specified. If the percentage modulation is 100, then the degree of modulation is 1. The gain in this instance is simply the ratio of the audio output voltage to the input carrier voltage. If 30% modulation is used, the degree of modulation is 0.3; therefore less audio output will result for a given carrier voltage input than in the preceding case, as is shown by the equation. For instance, if a 1-volt carrier signal, modulated 100%, were present across the input circuit of the detector and the resulting audio voltage across the detector load were 10, the detector gain would be 10, but if the 1-volt carrier signal were modulated 30% and the detector gain is 10, the output voltage across the detector load would be only 3 volts.

In Fig. 14-5, a plate power detector circuit is shown. The gain of this circuit is determined in the same manner as that of the grid-leak detector described above.

To measure the gain in these two detector circuits, we cannot proceed as we did for checking audio gain, since this would require our feeding the audio signal first to the a-f grid following the detector and then feeding a modulated r-f or i-f signal to the detector grid, noting the input voltage in each instance which produces the same output meter reading across the voice coil. By making proper compensation for the percentage modulation of the input signal we would then be able to find the gain from grid to grid, but this would also include any gain or loss in the a-f transformer.

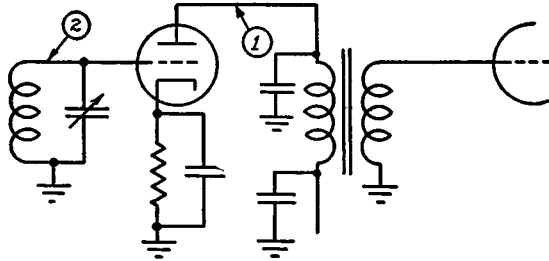
Further, the maximum output signal level of the test oscillator must be sufficiently great to produce a readable deflection on the output meter, when the signal is fed to the detector grid. For 30% modulation, this means that the output signal, with an a-f amplifier of average gain, would have to be from 150,000 to over 300,000 microvolts in order to give a test output of the order of $\frac{1}{2}$ watt.

These difficulties may be overcome by feeding the modulated signal to a preceding r-f or i-f stage and then measuring the a-f signal from point 1 to ground in either Fig. 14-4 or 14-5. A v-t voltmeter will be required for this test, since the loading effect of an ordinary copper-oxide meter is too great to permit a true read-

ing. Since a d-c voltage is present at the plate of the tube, a blocking condenser of $.01\mu\text{f}$ or larger should be used.

The modulated r-f signal voltage may then be read by transferring the v-t voltmeter connections to point 1 and ground. If the percentage modulation is known, the detector gain can then be determined. Biased pentode detectors will normally give a gain of about 50.

FIG. 14-5. Biased detector circuits are checked by measuring the gain between points 2 and 1 as described in the text.



A simple method of checking detector circuits is to feed an audio signal, in place of a modulated r-f signal, to the input of the detector. This enables the detector to be checked as an a-f amplifier. To make this test, the input tuned circuit must be disconnected from the detector to avoid grounding the a-f signal. The audio test oscillator is then connected in place of the tuned circuit. No blocking condenser should be used so the grid may return to ground through the oscillator attenuator. The efficiency of the detector can be determined by comparing the gain measured as an a-f amplifier with that obtained when it operates as a detector. Normally we should expect the gain as an a-f amplifier to be about 25% greater than its gain as a detector. If the gain as an amplifier is much greater than 25%, then either the voltages are incorrect, a bypass condenser is not functioning, or the tube is not a good detector. The last-named possibility should be checked first.

Checking Diode Detector Circuits

In a diode detector circuit there can be no gain, since a diode does not amplify. Actually, there is a loss ranging from 5 to 15 percent in rectification. The efficiency is greater for strong sig-

nals, of the order of 15 to 20 volts, and there is less distortion than for weak signals.

It is impractical to check diode detector circuits by injecting the signal in the manner described for the preceding tests, though direct tests of such circuits may be made with signal-tracing instruments. No matter how the test oscillator is connected to the diode circuit, the correct operating conditions cannot be secured without changing the detector circuit.

If, however, we combine our test of the diode circuit with that of the last i-f stage, these difficulties are avoided. Let us refer to Fig. 14-6, which is similar to Fig. 14-1. When the receiver volume control is at a maximum, the ratio of the audio signal voltage required to produce a given output meter reading when the a-f oscillator is connected from point 2 to ground, to the i-f signal voltage applied to point 3 which produces the same output—assuming 100% modulation—is the gain between these two points. This gain includes the last i-f stage and will be slightly reduced due to the loss in the diode detector stage. Also, since we shall normally use 30% modulation instead of 100% modulation, the ratio between the a-f signal voltage at point 2 to the i-f signal voltage at point 3 will be only 3/10 of the actual gain between these two points.

Let us again consider the typical receiver data for the G-E Model H-600. We note that the average gain of the i-f stage feeding the diode detector is 87, and that an a-f signal of .05 volt—which corresponds to 50,000 microvolts—fed to the a-f amplifier should produce $\frac{1}{2}$ watt output across the speaker voice coil. We may consider this as representing the gain of the corresponding sections of the receiver represented by the diagram Fig. 14-6. Then to determine what i-f signal voltage is required at point 3 to produce a 50,000-microvolt signal at point 2, we must first divide 50,000 by 87, which gives us 574 microvolts. This corresponds to the i-f signal voltage, modulated 100% at 400 cycles, required at point 3 to produce $\frac{1}{2}$ watt output across the receiver voice coil, assuming linear detection and no detection losses. Since our signal is modulated only 30% or 3/10ths of 100%, we must increase the signal level to produce the same a-f output. To find the i-f signal voltage required, we multiply 574 by 10/3, giving

us a required signal of approximately 1900 microvolts, modulated 30% at 400 cycles. Allowing 15% for detection losses, we obtain a final result of 2200 microvolts, 30% modulated at 400 cycles.

It is apparent from this discussion that our r-f test oscillator must be provided with an attenuator which is calibrated in microvolts if measurements of this type are to be made precisely. Also the modulation percentage should be known. While we may obtain a rough idea of the performance of the stage by comparing the signal level required to produce normal output in another normal receiver of similar design, to eliminate guesswork a really good signal generator or other means of measuring stage gain is

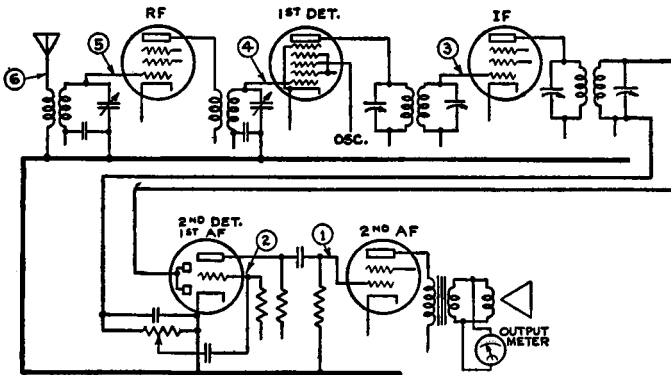


FIG. 14-6. A skeleton diagram of a superheterodyne receiver. The numbered points represent stage test points.

essential. Not that we need to make precise measurements in trouble-shooting, since a serious defect in the stage which in itself would necessitate a service job would normally cause a stage loss rather than a stage gain, but to make certain that the repaired receiver is rendering top-notch performance, we need to be certain that each stage is operating at maximum efficiency.

Since the gain of the tuned stages in a receiver is reduced when the circuits are misaligned, it is well to check the alignment during the stage gain test. The test oscillator will normally be connected through a .01-mf blocking condenser to the last i-f grid, and to ground, in conventional alignment procedure, just as for

the gain test described above. So the trimmers of the last i-f transformer should be adjusted for maximum output indication on the output meter before making the next gain test.

If the last i-f stage uses a broad-band, over-coupled transformer, a resistor approximately 20,000 ohms in value should be shunted across the last i-f transformer secondary while peaking the primary. This resistor will serve to eliminate the normal double peak which otherwise occurs. After the primary is peaked, the resistor may be removed and the secondary can then be peaked in the usual manner.

Checking the Converter Stage

Let us now connect the test oscillator, through a blocking condenser, to the grid of the converter or mixer—point 4 in Fig. 14-6. The grid lead which normally connects to this point should not be removed, since the impedance of the circuit, even when it is not tuned to resonance with the signal frequency, is normally much higher than that of the test oscillator attenuator which shunts it.

Our first test of this stage is to feed a signal at the intermediate frequency to the mixer grid, just as we did at point 3. Referring again to the data for the G-E Model H-600, we find the gain of this stage should be 28 at 455 kc. That is, if a 2200-microvolt signal, modulated 30% at 400 cycles, gave $\frac{1}{2}$ watt output across the speaker voice coil when fed to point 4, then a signal $\frac{1}{28}$ as strong should give the same output when fed to point 3. In microvolts, this means $\frac{1}{28}$ of 2200 microvolts or about 80 microvolts.

We can now peak the first i-f transformer trimmers for maximum output to make certain that the stage gain is a maximum. If the gain is low, we shall require considerably more than 80 microvolts input to obtain the proper output meter reading. Since this test is being made at the intermediate frequency, the trouble is not due to misalignment of the set oscillator, and is localized to components which affect only the mixer performance as an i-f amplifier. Perhaps the trouble is in the tube itself; this may be checked by substitution if it tests satisfactorily in a tube checker.

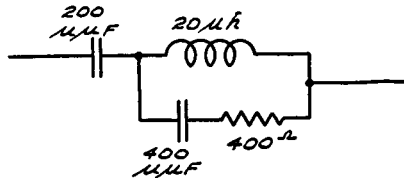
Other troubles, which do not affect operating voltages are, shorted turns in the i-f transformer windings or trimmers (which will be revealed by the alignment check since it will then be impossible to peak the circuits), open bypass condensers in screen, cathode or plate supply circuits and changes in the coupling between the i-f transformer coils.

If this test shows normal performance, we may leave the test oscillator connected to the same point and tune the receiver to 1400 kc. Then adjust the test oscillator also to 1400 kc. If the set oscillator is properly aligned, a 455-kc signal should now pass through the i-f amplifier as for the preceding test. If the set oscillator is perfectly aligned, and is giving proper oscillator signal voltage, the input signal from the test oscillator at 1400 kc need be only slightly greater than that at 455 kc to produce the same output. In average receivers, the loss when operating as a converter is about 10% as compared with the amplification at the intermediate frequency under converter operating potentials.

Measuring R-F Stage Gain

The r-f gain in the receiver circuit which we have been checking is represented by the ratio of the signal level required at point 5 to produce a given output meter reading to that which must be applied to point 6, the antenna post of the receiver, to produce the same output.

FIG. 14-6(a). The schematic of a standard dummy antenna used for all-wave superheterodyne testing.



In normal operation, the receiver is connected to an antenna, so in our tests a dummy antenna should be inserted between the test oscillator and point 6 to simulate normal operating conditions. The dummy antenna should be of the type specified by the manufacturer for gain tests; if none is specified, a standard all-wave dummy antenna is preferable for checking all-wave re-

ceivers. For tests on the broadcast band, where no dummy antenna is specified, a .0002- μ f condenser may be used; for short-wave bands, a 400-ohm resistor will be required. When feeding the signal to r-f stage points other than the antenna input, a .0002 or larger condenser should be used; this is done to avoid grounding the normal avc bias applied to the grid.

In some receiver data, it is specified that the grid connection from the tuned circuit should be disconnected and a resistance shunted from grid to ground of the circuit under test. In such cases, follow the manufacturer's specifications.

In very sensitive receivers, it will be found difficult to measure antenna coil gain by the usual method of applying a modulated signal in turn to the input and output terminals of the stage because of the high noise level. If a sensitive v-t voltmeter is used, the antenna transformer gain may be directly measured. A set-up for making such a measurement is shown in Fig. 14-7.

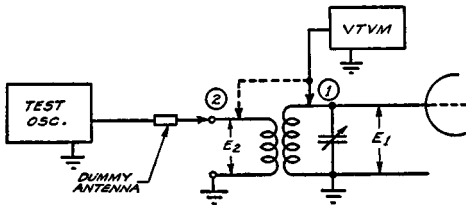


FIG. 14-7. How the gain in an antenna transformer can be determined, the v-t voltmeter being connected as shown.

An interesting fact is that this measurement may be made without turning on the receiver, since the gain is entirely that of the transformer and no tubes are involved.

To make this test, tune the receiver to some frequency within the range of the band used and feed a signal of the same frequency to the antenna post as shown. Connect the v-t voltmeter to point 1. If the receiver is not turned on, no blocking condenser will be required between the v-t voltmeter and point 1; otherwise, it will be advisable to use the condenser to avoid grounding the bias and causing grid current due to overload.

Note the reading of the v-t voltmeter when the circuit is tuned to give maximum output with the voltmeter connected. The v-t voltmeter will normally detune the circuit somewhat, and retuning will therefore be necessary.

Next, transfer the voltmeter lead from point 1 to point 2, and again slightly retune the circuit until a maximum reading at point 2 is obtained. The ratio of the reading at point 1 to that at point 2, $E1/E2$, is the antenna stage gain.

We may use high signal levels in such a measurement, and unless a tuned v-t voltmeter is employed or a test oscillator with at least 0.3-volt output is used, it will be difficult to get a reading. Many test oscillators have sufficient output, if not at all frequencies, at least at some frequency in the broadcast band.

Antenna stage gains range from 2 to 10 in home receivers and considerably higher in auto receivers. For the latter, a special dummy antenna is usually required and this is specified in the service notes for alignment of such receivers.

Measuring R-F Gain at High Frequencies

In checking r-f gain on short-wave bands, the attenuator calibration of most test oscillators and even of signal generators

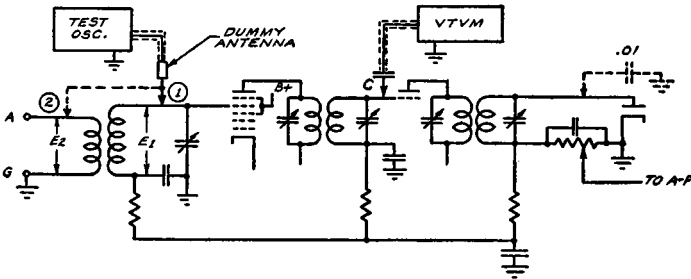


FIG. 14-8. By using a test set-up like this, the r-f gain can be found without having recourse to the oscillator attenuator.

becomes inaccurate. This is particularly true at low signal levels, since the leakage in the signal generator may then be comparable to the output signal level.

If a tuned v-t voltmeter, with an attenuator which is accurate over the i-f range, is employed as an output meter, it is possible to make gain measurements without relying upon the calibration of the test oscillator attenuator.

Referring to Fig. 14-8, the tuned v-t voltmeter is shown connected to the input of the last i-f stage. The blocking condenser *C* is usually incorporated in the probe of such signal-tracing instruments and consists of an air-dielectric condenser of approximately 1 or 2 μ f, so the loading and detuning effect on the circuit under test is negligible. The .01- μ f condenser across the detector is used to detune this stage and thereby prevent AVC action from affecting the gain measurements when a relatively strong signal is used.

To measure the r-f or antenna stage gain by this method, the test oscillator is connected through a dummy antenna first to point 1 and is tuned to the same frequency as that to which the receiver is tuned. The v-t voltmeter is adjusted to the intermediate frequency and its attenuator is set for maximum sensitivity. The test oscillator attenuator is then adjusted until a signal sufficiently strong to give a reference indication on the v-t voltmeter is secured. Then, without changing the setting of the test oscillator attenuator, the signal is fed to point 2. The reading of the v-t voltmeter should increase due to the gain in the r-f circuit under test. By readjusting the v-t voltmeter attenuator, the increase in signal level may be measured.

A tuned v-t voltmeter is required because such instruments have high sensitivity, enabling readings of the order of a few millivolts so that appreciable gain may be measured without overloading the grid of the i-f tube. If an ordinary v-t voltmeter were employed, it would be necessary to connect across the input circuit to the diode and render the AVC circuit inoperative in some other manner. Even then the amount of gain which could be measured would be limited.

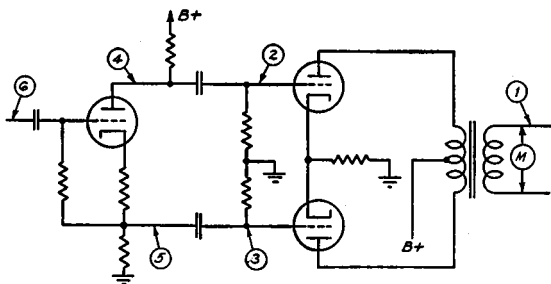
Measuring Gain in a Phase Inverter Stage

A typical phase inverter stage is shown in Fig. 14-9. In this circuit, if a signal voltage is applied to the input circuit, from point 6 to ground, the amplified signal will appear across the plate and cathode load resistors. When this stage is operating properly, the signal voltage from point 4 to ground should equal that from point 5 to ground, but these two voltages will be 180°

out of phase. This same relation applies to point 2 with respect to point 3.

The simplest method of checking this circuit is to apply a 400-cycle a-f signal from the audio oscillator from point 6 to ground and measure the signal voltage from points 4 and 5 to ground; likewise from points 2 and 3 to ground. Equality of all these voltages indicates that the circuits are properly balanced, though it does not necessarily follow that proper phase relations exist. However, by noting if the signal voltage remains constant from points 2 and 3 to ground as the audio oscillator frequency is varied over the normal operating range of the phase inverter stage, a good idea of the performance of the stage is obtained.

FIG. 14-9. Test points for finding the gain in a phase-inverter stage.



When the audio voltage across the two push-pull grids is exactly twice that across a single grid and ground, the voltages are 180° out of phase.

Testing of the output stage may be done by feeding the audio signal alternately to points 2 and 3 and noting the deflection of the output meter, which is connected across point 1 and ground. To check the power output, the audio signal must be fed simultaneously to points 2 and 3. For this purpose, the audio oscillator output transformer must be center-tapped. Then the center tap is grounded and each "high" side is connected to one of the push-pull grids. If the audio oscillator is not so equipped, the signal may be fed to the input of the phase inverter stage, point 6, and the same results will be obtained when the phase inverter stage is operating properly.

Localizing Distortion in A-F Amplifiers

In checking for waveform distortion, the function of the test oscillator is to supply a signal source while the presence of distortion is indicated by the output device. For this purpose, the cathode-ray oscillograph is preferable, when the signal voltage is from a sine-wave source. Alternatively, an amplifier and phones may be used, as supplied in signal-tracing apparatus. Then it is desirable that recorded speech or music rather than a sine-wave source, be used as a test signal since the ear does not detect small amounts of distortion in single-frequency tones.

Referring to Fig. 14-3, we may feed an audio signal to point 5 and, using the cathode-ray oscillograph, connect the vertical amplifier across point 5 and ground. If the a-f amplifier represents that of a receiver employing a diode detector, the detector should be removed to avoid loading the positive half of the audio wave and thus introducing distortion. Alternatively, the a-f amplifier may be disconnected from the diode load.

After noting the waveform of the signal being applied to the amplifier, the oscillograph is connected to point 4. The gain control on the oscillograph should be readjusted until an image of approximately the same size as that originally obtained at point 5 is secured. If the wave is perceptibly distorted at point 4, then the trouble is localized in the components of the first a-f stage.

The test is continued by noting the waveform of the signal when the oscillograph is connected at points 3, 2 and 1, in turn. If the signal becomes distorted at any of these test points, the trouble is then localized.

Care should be taken to keep the signal level low enough so that no stage is overloaded. If the output tube is overloaded, this will cause grid current and may be detected by connecting an electronic voltmeter from point 3 to ground. When the signal voltage is too high, the grid current from this cause will result in a voltage at point 3, which is negative with respect to ground.

Localizing Distortion in R-F and I-F Amplifiers

Testing for distortion in r-f and i-f amplifiers is done in the same manner as for gain measurements. The oscillograph, or other distortion-indicating instrument, is connected in any por-

tion of the a-f system or across the diode load. The modulated signal is fed to the various stages in turn and the demodulated output waveform is noted on the oscillograph. Any departure from the normal waveform when the signal is fed to a given stage is evidence of distortion originating in the particular section under test.

The signal levels used for distortion checking should be higher than for gain measurements. In some instances, distortion in i-f circuits occurs as a result of overload. Overload may result from defective avc action or from misalignment of the last i-f stage, other conditions being normal. It is possible that the output of the test oscillator will be insufficient to produce overload when the signal is fed directly to the stage affected. In such cases, the avc action and receiver alignment should be checked separately.

The avc action may be tested by checking the avc voltage across the diode load and noting if it increases as the signal input to the receiver is increased. A curve showing the normal increase in avc voltage with increased signal input is shown in Fig. 14-10.

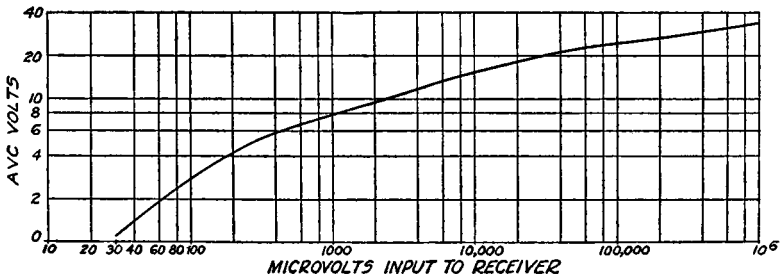


FIG. 14-10. How the avc voltage varies with respect to the signal input to the receiver.

Testing of this type is subject to limitations, as mentioned above, which are not present when using signal-tracing instruments.

It is essential that the waveform of the modulated output of the test oscillator be good if the demodulated output is to represent only distortion added by the receiver circuits.

Adjusting 10-kc Filters

The 10-kc filter used in many high-fidelity receivers may be properly adjusted by means of an audio oscillator or a modulated r-f signal. In Fig. 14-11 the filter network employed in the Stromberg-Carlson models 255-B and 255-L is shown. When the circuit composed of L_1 and C_1 is tuned to resonance with a 10-kc

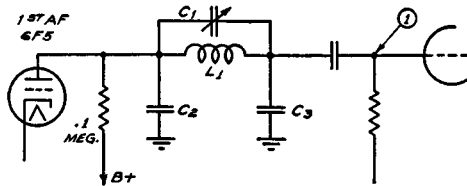


FIG. 14-11. The correct adjustment of C_1 is when a minimum reading is observed on a v-t voltmeter connected across point 1 and ground.

signal it offers a high impedance to that frequency. Consequently, little signal voltage of that frequency will then be applied to the following grid. This action is similar to that of a parallel wave trap connected in series with an antenna and receiver input.

To adjust this filter, feed a 10-kc signal to the grid of the 6F5, then adjust C_1 until a v-t voltmeter connected from point 1 to ground shows *minimum* reading. This point may be noted by varying the oscillator frequency above and below 10 kc and observing if the signal voltage, increases each side of resonance. This is the correct condition for proper adjustment.

Checking Inverse Feedback Networks

The characteristics of an inverse feedback network may be quite simply checked with an audio oscillator and v-t voltmeter. In Fig. 14-12 the inverse feedback network is shown in heavy lines, the signal voltage developed across the voice coil being fed back to the cathode of the second a-f tube through a 180-ohm resistor and 2- μ f condenser to a 50-ohm resistor which forms a portion of the cathode bias for the 6F6 tube. This circuit is used in the G.E. type E-155 receiver.

The test is made when the a-f amplifier is not operating. The lead to the 180-ohm resistor is disconnected at the voice coil and connected to an audio oscillator. The ground terminal of the oscillator is connected to the amplifier ground. By feeding audio

signals of various frequencies to this network and measuring the signal voltage developed across the 50-ohm resistor while the audio oscillator output voltage is maintained constant, the feedback voltage applied to the 6F6 tube can be determined.

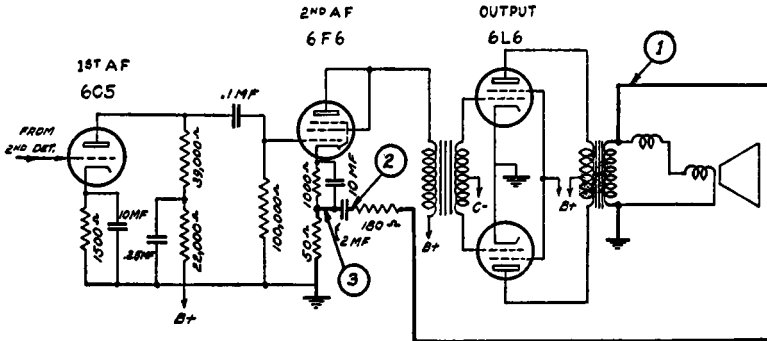


FIG. 14-12. The inverse feedback voltage applied to the 6F6 tube can be found by measuring the signal across the 50-ohm resistor when different audio signals are fed to the network.

In this circuit the audio voltage from point 3 to ground should be about 3 percent of that applied at point 1 at a frequency of 50 cycles. This feedback voltage will increase to approximately 20 percent as the frequency is increased to 5000 cycles. The characteristics of other inverse feedback networks may be checked in the same manner by simply disconnecting them from the circuit. This is, of course, a static test and not a dynamic one, but will serve to give a very good idea of the degree of feedback which will be obtained under actual operating conditions.

Automatic Volume Expansion

In radio broadcasting, it is customary to monitor the audio modulation of the transmitter so as to secure the greatest efficiency of operation. This means that the very soft passages of an orchestral selection are raised in volume and the very loud ones are reduced. This process of monitoring is known as volume compression. The result is that the overall volume range of the transmitted program as received in the home is not as great as

that which is actually covered by the orchestra. However, the distortion which results from over-modulation and the noise which is present with very weak modulation are avoided.

Volume expansion circuits compensate for this reduction in volume range at the transmitter by automatically raising volume in accordance with the average volume level. An ideal volume expander would be one in which the degree of expansion would exactly offset the degree of compression at the transmitter so that the resulting sound output from the receiver speaker varies over the same intensity range as the signal reaching the studio microphone. This cannot be obtained, since the degree of monitoring at the studio varies. Best results are obtained by using volume expansion in record reproduction, since the amount of volume compression does not vary so greatly during recording.

There are various ways of securing volume expansion, but nearly all are based on some method of varying the gain of an audio amplifier tube so that the stage gain is increased as the signal strength increases and decreased as the signal strength decreases. Therefore the test procedure is based upon applying an audio signal of known voltage and noting the resulting output after passing through the amplifier, while the expander circuit is in operation. Then the audio signal level at the input is increased a known amount. If the expander circuit is functioning, the proportionate increase in signal level at the amplifier output will be much greater than would result without expansion.

An example of an automatic volume expansion circuit as incorporated in the audio section of a radio receiver is shown in Fig. 14-13. The fundamental principle of its operation is illustrated in the small insert (a) at the right. A 100,000-ohm resistor in series with the plate resistance of the 6K7G expander tube forms a voltage divider across the input to the power amplifier stage. As a result of this connection the grid of the output stage receives only the voltage which is developed across the plate resistance of the 6K7G, represented in the small diagram by a resistor within a circle. Automatic volume expansion is achieved by automatically varying the value of the plate resistance in accordance with the average audio level. During periods when the audio, or modulation, level is high, the bias on this tube is likewise high and consequently the plate resistance of the 6K7G is likewise

voltage across the 470,000-ohm resistor in the plate circuit. This rectified voltage is filtered by the 1-meg ohm resistor and .2- and .3- μ f condensers before being applied to the suppressor grid of the 6K7G expander. As the signal level applied to the 6J7G increases, the rectified voltage across the plate resistor of the tube increases, thereby increasing the negative d-c bias on the 6K7G suppressor grid. The plate resistance of the 6K7G expander tube increases with this increase of suppressor grid bias and in accordance with the relative increase in strength of the audio signal voltage at the input. This effect adds to the normal increase which results from an increase in signal level so that the signal output of the amplifier is greater than it would be without expansion. The double-pole double-throw switch makes operation of the expander circuit optional.

In testing this circuit, either a modulated r-f signal can be fed to the receiver input or an a-f oscillator may be used and the signal may be fed to grid of the 6Q7G. The output meter may be connected across the speaker voice coil.

The audio signal voltage at the 6Q7G should be low, less than 1 volt, and the resulting output at the voice coil should be noted. Then double the input voltage. If the expander is functioning, the output voltage should more than double. The degree of increase is a measure of the amount of expansion.

Testing Automatic Bass Amplifier Circuits

At low volume levels, it is a characteristic of the human ear that low notes in a musical reproduction seem to be relatively weaker than high notes. Yet, when the overall volume level is raised, the proper proportionate strength of high and low notes is established. Automatic bass control is a means of automatically providing increased bass response at low volume levels to correct for this psychological condition and proportionately less increase as the overall volume level is raised. In addition, a manual control is provided which enables the listener to vary the amount of bass amplification in accordance with his own preferences. While the latter adjustment is in effect a bass tone control, its mode of operation is essentially different from the conventional tone control.

The manner in which automatic bass control action functions may be understood by analyzing the partial schematic, Fig. 14-14. A portion of the output of the 6F6G driver stage is coupled to the diode plates of the 6R7G and this audio signal is rectified. The resulting d-c voltage across the 500,000-ohm diode load is filtered through the 1.0 meg resistor and .45- μ f condenser and fed over to the 6K7G bass amplifier tube. In this way the d-c bias on the 6K7G is made to vary in accordance with the strength of the signal at the output of the driver stage—in other words, in accordance with the output level of the receiver. When the output level is low, the bias voltage on the 6K7G will likewise be low,

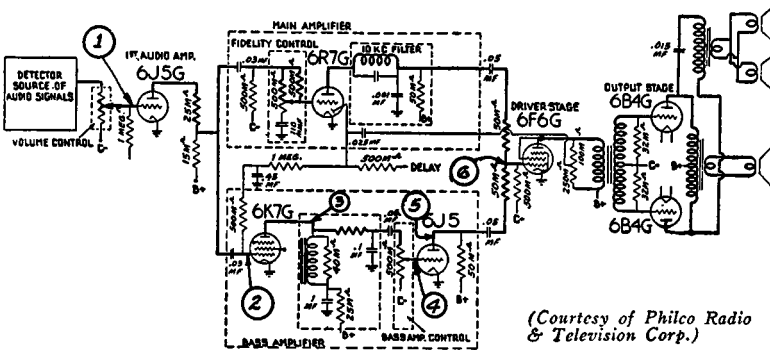


FIG. 14-14. The bass amplifier circuit of the Philco Model 37-690. How the gain is determined is described on the following page.

and the gain of the bass amplifier will be greatest; when the output level is high, the bias will be high and therefore the gain of the 6K7G will be reduced, so the amplification of the bass channel will accordingly be less. A delay action is incorporated in the automatic bass control circuit by applying a negative bias voltage in the diode circuit. For this reason, the diode will not begin to rectify and thus furnish a voltage to control the gain of the amplifier until the receiver output level reaches a certain value.

In testing this circuit, we may use either an r-f test oscillator with variable-frequency modulation, feeding the signal to any portion of the receiver ahead of the second detector, or an a-f oscillator may be used and the signal applied to the grid of the first a-f amplifier at point 1 and ground. The output meter

should be connected across a resistor equal to the voice coil impedance at 400 cycles and shunted, in place of the voice coil across the output transformer secondary. This is required when tests are made at low frequencies, particularly if the speaker is not mounted on a baffle.

The gain of the amplifier, without bass amplification, is determined by setting the bass amplifier control at minimum and noting the output meter reading for a constant signal input voltage at different frequencies. The effect of the bass amplifier will of course be most evident at low volume and at low frequencies. In the event that the bass amplification is subnormal or absent, the gain of the 6K7G bass-amplifier tube and circuit may be checked by connecting a v-t voltmeter to point 3 and comparing the signal voltage at this point with that at point 2. The gain should increase when the input signal is lowered. For instance, if the measured gain of this stage is 2 with a signal input of 1 volt at 100 cycles, the gain may increase to 10 or more when the signal input at the same frequency is reduced to 0.1 volt. These gain figures do not represent the actual values, which are not available, but serve merely as an illustration.

AVERAGE GAIN-PER-STAGE VALUES

The figures on gain-per-stage listed below are based on the assumption that the receiver avc system is not operating. AVC action will reduce considerably the r-f, mixer, and i-f stage gains. For comparison purposes, a weak signal should be used, or the avc circuit temporarily shorted out.

In the a-f section, for resistance-coupled amplifiers, the lower gain figures represent average gains for ac-dc receivers while the higher gains apply to a-c operated receivers.

Radio-Frequency Section:

| | |
|--|-----------|
| Antenna to grid of first tube..... | 2 to 10 |
| Antenna to grid of first tube—auto radios..... | 10 to 50 |
| R-F amplifier—superheterodynes..... | 10 to 40 |
| R-F amplifier—tuned r-f receivers..... | 40 to 100 |

Mixer Section:

| | |
|--|----------|
| Converter grid to i-f grid (1 stage i-f amp.)..... | 30 to 60 |
| Converter grid to i-f grid (2 stage i-f amp.)..... | 5 to 30 |

Intermediate-Frequency Amplifier Section:

| | |
|------------------------------------|-----------|
| I-F stage (1 stage i-f amp.)..... | 40 to 150 |
| I-F stage (2 stage i-f amp.) | 5 to 30 |

Biased Detector:

Pentodes (Types 57, 6C8, 6J7, etc.).

A 1.0-volt rms signal (20% mod.) at the grid will produce approximately 10 volts rms of a.f. at the plate. Higher modulation percentages will produce correspondingly higher a-f voltages. Thus 40 percent modulation will produce 20 volts rms of a.f.

Note:—AVC voltage may run as high as 40 volts, depending upon the strength of the input signal and the number of tubes under control. In general the greater the number of tubes controlled, the lower the avc voltage for a given input signal.

Audio-Frequency Section:

Medium- μ triodes, resistance-coupled:

| | |
|------------------------------------|----------|
| Types 6N7, 6C8 (each section)..... | 20 to 25 |
|------------------------------------|----------|

High- μ triodes, resistance-coupled:

| | |
|------------------------------------|----------|
| Types 75, 2A6, 6F5, 6SQ7, etc..... | 30 to 60 |
|------------------------------------|----------|

Output Pentodes:

| | |
|---|---------|
| Types 6F6, 2A5, 47, 6V6, 6L6, 6K6, grid-to-plate gain)... | 8 to 20 |
|---|---------|

Output Triodes:

| | |
|-------------------------------|--------|
| Types 2A3, 45, 71A, 6A5G..... | 2 to 5 |
|-------------------------------|--------|

Service notes on many recent models of receivers manufactured by the General Electric Company contain valuable gain-per-stage data as well as other information applicable to signal-tracing tests.

For Models H-600, 601, 610, 611, 620 and 621, the following information is furnished in the G-E Service Notes:

- | | |
|--|----------------|
| (1) Stage gains | Gain |
| Antenna to Converter Grid | 2.7 at 1000 kc |
| Converter Grid to 6SK7 Grid | 28 at 455 kc |
| 6SK7 Grid to 6SQ7 Diode Plate | 87 at 455 kc |
| (2) Audio Gain | |
| .05 volts, 400-cycle signal across volume control with control set to maximum will give approximately ½ watt output at speaker. Voice coil impedance equals 3½ ohms at 400 cycles. | |
| (3) DC voltage developed across oscillator grid leak averages 13 volts. | |

The stage gain ratings given above are average; variations of plus 10 percent or minus 20 percent are normal, according to the service notes.

In the above data the gain from converter grid to the 6SK7

grid represents the gain of the converter tube and the i-f transformer at the intermediate frequency, 455 kc. This is not the conversion gain but will indicate the performance of the converter as an i-f amplifier.

The watts output may be determined by a voltage reading of the output meter when the speaker voice coil impedance is known. The required voltage reading for a given output in watts is then found by the formula, $E = \sqrt{RW}$. For example, in the above receiver data, the voice coil impedance, R , is $3\frac{1}{2}$ ohms at 400 cycles; the desired watts output, W , is $\frac{1}{2}$ watt. The voltage, E , which will be across the voice coil when the power output is $\frac{1}{2}$ watt, is equal to the square root of ($3\frac{1}{2}$ times $\frac{1}{2}$) . . . about 1.32 volts. This output voltage should preferably be measured with a resistance, equal in value to the voice-coil impedance, shunted across the output transformer secondary in place of the voice coil.

Chapter XV

OSCILLATORS IN SUPERHETERODYNE RECEIVERS

THE purpose of the oscillator in a superheterodyne receiver is to provide a signal which will beat with an incoming broadcast signal and produce in a mixer tube an intermediate frequency signal which is otherwise similar to that of the broadcast signal. In some sets, the oscillator signal is developed by a separate tube and circuit which is coupled to a mixer tube; in others, the functions of the oscillator and mixer are combined within a single tube.

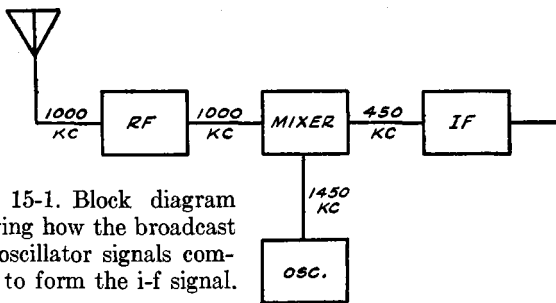


FIG. 15-1. Block diagram showing how the broadcast and oscillator signals combine to form the i-f signal.

The arrangement employed to produce the i-f signal is shown in Fig. 15-1. This block diagram represents equivalent sections of a superheterodyne receiver which is tuned to a broadcast signal of 1000 kc. The 1000-kc signal is amplified by the r-f stage and applied to the input circuit of the mixer stage. The set oscillator supplies a 1450-kc signal which combines with the 1000-kc incoming broadcast signal in the mixer to produce an intermediate

frequency which represents the difference between 1450 and 1000 or 450 kc. The i-f amplifier is tuned to the 450-kc signal.

A 450-kc signal would likewise result if the oscillator operated at 550-kc instead of 1450 kc, since the difference between 1000 and 550 is also 450. In practice, though, most receiver oscillator circuits are designed to function at a frequency which is higher than that of the incoming broadcast signal.

Note that the method of operation of a superheterodyne receiver, as outlined above, is similar in principle to that of the beat-frequency r-f and a-f oscillators described in Chapter XI. In fact, oscillators of these types are derived from fundamental superheterodyne theory.

Image Frequency

We have described briefly how the desired incoming broadcast signal is changed in frequency so that it may be amplified by an i-f amplifier. It is also possible for undesired signals to be converted to the intermediate frequency and therefore to be amplified by the i-f stages, thus causing interference. For instance, if the receiver were tuned to a weak 10,000-kc incoming signal, the oscillator would normally function at 10,450 kc to produce the desired 450-kc i-f signal. But if a strong 10,900-kc signal were likewise present in the antenna circuit, it might arrive at the mixer along with the desired 10,000-kc signal. Since the oscillator is functioning at 10,450 kc, it will beat not only with the 10,000-kc signal but also with the 10,900 kc signal and change the frequency of each to 450 kc, because the difference between 10,900 and 10,450 is the same as that between 10,000 and 10,450. Both the desired and the undesired signals will then be present in the i-f amplifier and interference will result.

This type of undesired response is known as image-frequency response. The image frequency always differs from the desired frequency by twice the intermediate frequency. Whether the image frequency is higher or lower than the desired frequency to which the receiver is tuned depends upon the relation of the oscillator frequency to that of the incoming signal. In most cases, the oscillator frequency is higher than the desired signal frequency. Then the image frequency must also be higher, if the i-f

signal is to result. In some receivers, particularly on higher frequency bands, the oscillator frequency is lower than that to which the receiver is tuned. Then the image frequency also must be lower in frequency. Thus, if the receiver were tuned to 20,000 kc and the oscillator frequency is 19,550 kc, the i-f signal is 450 kc, and the image frequency becomes 19,100 kc.

Image-frequency response is particularly troublesome at very high frequencies, since then the percentage frequency difference between the image frequency and the desired frequency is small, if the i.f. is of the order of 450 kc. Under such conditions the selectivity which can be attained by the r-f amplifier is not sufficient to exclude strong undesired signals when the receiver is tuned to a weak signal. The difficulty becomes more pronounced as the i.f. is lowered; that is why an i.f. of the order of 175 kc is not used in all-wave receivers. If it were, then the image frequency would differ by only 350 kc from the frequency to which the receiver is tuned, whereas if the i.f. is 450 kc, the image frequency is 900 kc removed from the desired signal frequency. Thus, in the latter case, better rejection of the undesired signal by the tuned r-f circuits is obtained.

It might seem that a still higher intermediate frequency might be employed to advantage and this is true when receivers are designed solely for u-h-f reception. Intermediate frequencies in such superheterodynes are often of the order of 3 mc or higher. For all-wave receivers, such a high intermediate frequency is undesirable because it would be necessary to omit a band of frequencies near that of the i.f., since signals in that vicinity would be picked up directly without the heterodyning oscillator. Further, lower intermediate frequencies give greater i.f. selectivity and gain.

Recognition of the normal image response of superheterodynes is of particular importance in aligning all-wave receivers on short-wave bands and will be discussed further in this chapter.

Frequency Stability

The frequency-stability requirements in a superheterodyne oscillator vary with the type of receiver employed. In receivers designed solely for reception on the standard broadcast band, the

requirements are not so exacting as for receivers of the communication type which employ narrow-band i-f amplifiers. Often such receivers are designed to be used with crystal filters, which are designed to pass only a very small band of frequencies. In such cases, a very slight drift in oscillator frequency may necessitate retuning to retain the signal. The oscillator circuits of such receivers require more careful design, particularly for short-wave C.W. reception, than do those for the standard broadcast band.

Automatic frequency control is used in many receivers to compensate for oscillator frequency drift and for mistuning. This is discussed in detail in the book "Automatic Frequency Control Systems," by John F. Rider.

Oscillator Tracking

A wide variety of oscillator circuits is to be found in broadcast receivers, but actually these are simply minor variations of the fundamental types described in Chapter IV. Of these circuits, most sets use either the Hartley or the tickler-feedback types. A representative example of the latter type of oscillator as used in superheterodynes is shown in Fig. 15-2. The modification

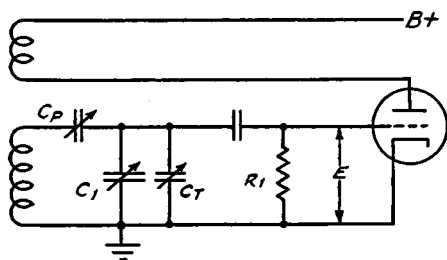


FIG. 15-2. The padder and trimmer condensers, C_P and C_T , in this superheterodyne oscillator circuit are used to obtain proper tracking.

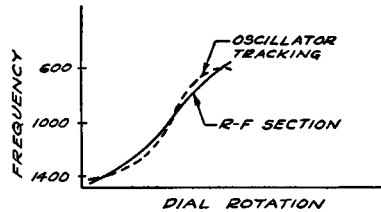
necessary to adapt this circuit to use in a receiver is the inclusion of a means to provide proper oscillator tracking.

When the oscillator in the receiver is perfectly adjusted it will function at a frequency which differs from the incoming signal frequency by an amount equal to the intermediate frequency employed in the receiver i-f amplifier.

When all sections of the gang condenser which is used to tune the r-f and oscillator coils are similar, as is usually the case, it

is necessary to use a padding condenser and a trimmer condenser as shown in Fig. 15-2 to secure proper tracking. Even then, the precise frequency difference is not obtained at all points over the tuning range. When properly adjusted, the tracking will be precise at a low frequency point on the band at which the padding condenser C_P is adjusted, and also at a high frequency point where the trimmer C_T is adjusted. At some frequency between the trimming and padding points, the exact frequency difference can likewise be obtained. A curve showing how this frequency difference varies over the tuning range is shown in Fig. 15-3. As

FIG. 15-3. The dotted line shows how the oscillator tracking varies over the tuning range.



you will note, if the padder is adjusted at 600 kc and the trimmer at 1400 kc, the frequency difference increases at points between 1400 kc and 1000 kc and decreases slightly between 1000 kc and 600 kc. At 1000 kc the dotted line representing the oscillator tracking crosses that of the r-f section; this is termed the “cross-over” point.

Superheterodyne Oscillator Signal Voltage

For proper operation, the oscillator signal voltage applied to the mixer circuit should be far greater than the strongest incoming broadcast signal. The actual value of the oscillator voltage is dependent upon the type of mixer circuit employed. If a pentode mixer is used, which is usually designed to operate at a grid bias of -10 volts, the *peak* oscillator signal voltage applied to the mixer grid or cathode circuit should not exceed -9 volts. For any other value of mixer grid bias, the peak oscillator voltage applied similarly should be 1 volt less than the grid bias.

These limitations as to the magnitude of the oscillator signal voltage so applied arise from the fact that, if the oscillator volt-

age exceeds that of the grid bias, the mixer will draw grid current and load the r-f tuned circuit, resulting in reduced gain and distortion.

In more modern broadcast receivers, mixers of the pentagrid type, such as the 6L7, are often employed. In such mixers, maximum efficiency of operation is obtained when the peak oscillator signal voltage is from 12 to 18 volts, depending upon the operating voltages of the 6L7. Voltages somewhat in excess of this minimum do no harm, but if the peak oscillator signal voltage is too low, the i-f signal output will be subnormal. In some all-wave receivers, which operate up to ultra-high frequency bands, only a portion of the total oscillator signal voltage available is normally employed. While some conversion gain is thereby sacrificed, better overall performance is obtained.

Methods of measuring the oscillator signal voltage are described further on in this chapter.

Superheterodyne Oscillator Circuits

The oscillator voltage may be coupled to the mixer circuit inductively, capacitively, directly or electronically. An example of inductive coupling is shown in Fig. 15-4. In this circuit, which

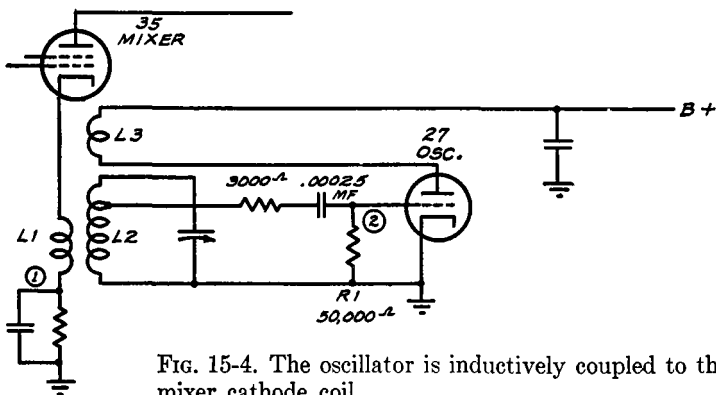


FIG. 15-4. The oscillator is inductively coupled to the mixer cathode coil.

is utilized in the Colonial 47 and 48 receivers, the oscillator signal voltage is coupled to the mixer circuit by reason of the mutual

inductance between the coils L_1 and L_2 . The oscillator itself employs a tickler-feedback circuit. The purpose of the 3000-ohm resistor in series with the oscillator grid condenser is to minimize harmonic production.

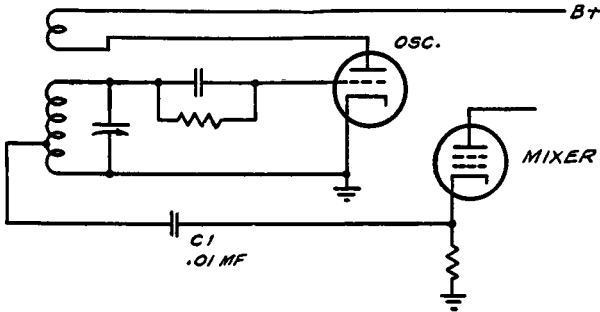


FIG. 15-5. The oscillator signal is coupled to the mixer circuit by the condenser C_1 .

An example of capacitive coupling to the mixer is given in Fig. 15-5. As you will note the condenser C_1 connects to a tapped portion of the tank circuit inductance, and the oscillator signal voltage is developed across the cathode resistor of the mixer tube.

In Fig. 15-6 the oscillator circuit used in the RCA Model C9-4 is shown. In this circuit the injector grid of the 6L7 mixer is

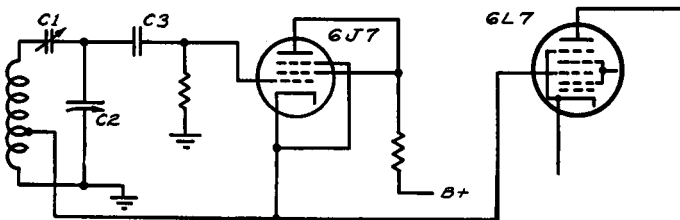


FIG. 15-6. The oscillator circuit used in the RCA Model C9-4.

directly coupled to the cathode of the 6J7 oscillator. Since a Hartley circuit is employed in this oscillator, the voltage fed to the 6L7 is that which is developed between the cathode tap and ground of the oscillator circuit.

Another example of direct coupling is shown in Fig. 15-7, in which the 6K8 combination triode-hexode converter tube is employed. In this circuit the triode section of the tube is used as the oscillator, which operates in a tickler-feedback circuit. Note that the plate supply voltage for the oscillator does not pass directly through the tickler coil, but is shunt-fed through a 15,000-ohm resistor. Since a blocking condenser is employed between the tickler coil and the triode plate, no d.c. is present in this circuit and the tickler coil may therefore be grounded.

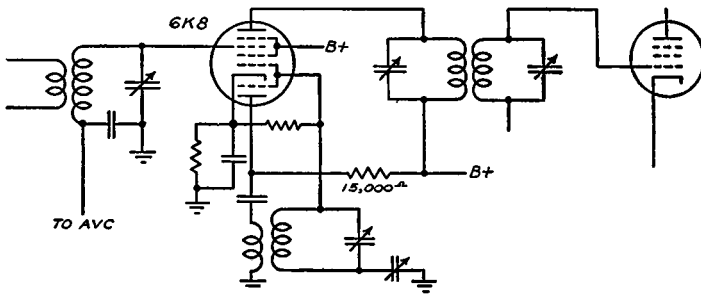


FIG. 15-7. The oscillator and mixer functions are combined in the 6K8 triode-hexode converter.

The triode grid and G_1 of the hexode are directly coupled as shown. The oscillator voltage which therefore appears on G_1 of the mixer is internally shielded from the signal control grid, and the intermediate frequency is developed in the plate circuit of the hexode section by reason of the electronic coupling within this portion of the tube. This circuit is used in the Travler Model 436-M, as well as other types put out by this company.

In Fig. 15-8 the circuit of the RCA Models 5X5I and 5X5W is shown. This is a representative application of the new single-ended 12SA7 pentagrid converter. The oscillator circuit is a Hartley type, and the oscillator voltage is electronically coupled to the signal voltage.

Testing Superheterodyne Oscillators

To check an oscillator for operation we may employ several methods. The simplest one is to measure the rectified voltage

developed across the oscillator grid resistor. If the oscillator is functioning, a d-c voltage will be developed at the grid which is negative with respect to the point to which the grid returns. This voltage will vary to some degree over the band, but normally this variation will not be greater than a ratio of 2 to 1. A Table showing the average d-c voltage across the oscillator grid leak will be found in the Appendix. These are typical values which will provide some guidance as to the magnitude of this voltage, but variations of 50% or more may be found in some receivers. If the voltage drops to zero or becomes slightly positive, it is an indication that the oscillator is not functioning. If the voltage is higher than the values given and a pentagrid mixer is employed, satisfactory operation will still be secured.

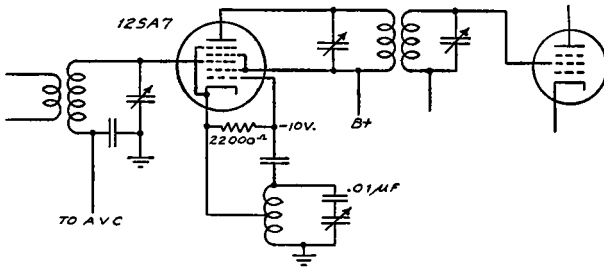


FIG. 15-8. The 12SA7 pentagrid converter is used as both oscillator and mixer.

It is also possible to check for operation by a current measurement. A 1-ma meter may be inserted in series with the oscillator grid resistor at its return point to cathode or ground. The current reading which thus results may be converted to represent the voltage reading which would be obtained if an electronic voltmeter were connected across the oscillator grid leak. This is done by Ohm's law, $E=IR$. Thus, if the oscillator grid leak is 50,000 ohms and the meter reads 0.2 ma, the average rectified voltage across the oscillator grid leak is $50,000 \times 0.0002$ or 10 volts. The reading thus obtained may be used for comparison with the Table data in the Appendix.

A meter may also be used in the series with the plate-supply voltage lead to the oscillator tube to indicate oscillation. When so connected, the plate current is shown. If the oscillator is

functioning, touching its grid will cause an appreciable change in plate current; if not, there will be little or no change in current reading.

Uniformity of oscillation may be checked by rotating the gang condenser over the tuning range and watching the change in current reading. If the oscillator becomes inoperative, the grid current will drop to zero. The grid current normally will vary slightly over the operating range but in normal operation the variation should not exceed 2 to 1.

Sudden dips in current or voltage as the oscillator frequency is varied are caused by absorption, due to the proximity of other circuits which have resonant frequencies corresponding to the frequency to which the oscillator is tuned. It is very important not to disarrange the oscillator circuit wiring during repairing or testing, since some of the oscillator energy may thus be coupled to another circuit from which it is normally isolated, resulting in dips, frequency shift and occasionally stoppage of oscillation.

While stoppage of oscillation will prevent the receiver from functioning as a superheterodyne, it is still possible that signals may be received. A strong local signal near the low-frequency end of the standard broadcast band may reach the receiver with sufficient intensity to pass through the i-f amplifier directly. This is particularly the case with small midgets, which employ no r-f stage, and in which only a single i-f stage is employed, tuned to about 465 kc. In such cases, the signal may be received at any point on the dial.

It is possible that the oscillator may be functioning but at a frequency remote from the proper point, so no signal at the intermediate frequency will result. It may also happen, if a separate oscillator is employed, that an open or short circuit in the coupling to the mixer may result in an extremely weak signal, even when the oscillator is functioning at normal frequency and output.

Troubles of these types are best checked with signal-tracing instruments, which are designed for these specific tests. However, it is possible to check the oscillator frequency by using an auxiliary all-wave receiver, coupling the antenna lead loosely to the receiver oscillator under test, and rotating the all-wave receiver tuning condenser over each band until the oscillator signal

is picked up. A tuning indicator of some type is a desirable accessory for the all-wave receiver used for this purpose, since the oscillator signal is unmodulated, but a thump will be heard in the speaker if the radiated oscillator signal is strong enough.

It is also possible to inject an unmodulated signal from a test oscillator in the mixer circuit and thus restore operation. This serves as an indirect check on the oscillator frequency and its coupling to the mixer. The receiver under test should be tuned to a strong local signal. If tuned to 600 kc and the i.f is 450 kc, the test oscillator signal frequency should be set at 1050 kc.

Hum and Other Troubles

Other oscillator troubles may be present in which additional information may be required-regarding the characteristics of the oscillator signal. It is possible that the oscillator signal may not be a pure, unmodulated wave such as is required for perfect operation. If the filtration of the voltage supply for oscillator operation is inadequate, hum may appear on the oscillator signal. This hum will then modulate any r-f signal present in the mixer, though the hum may not appear in the speaker unless the r-f signal is tuned in. Where there is reason to suspect this condition, the hum level of the d-c voltage supplied to the oscillator should be checked. If this is abnormally high, additional filtering should be introduced and its effect on the output of the receiver noted. If the modulation of hum disappears, then it can be assumed that it was caused by insufficient filtering of the plate supply. In this connection, proper by-passing of the heater is important, especially at the higher frequencies.

If the receiver blocks when tuned to a strong local station or when too high a signal is fed into it from the test oscillator, a variation in power-supply voltage may cause fluctuation of the oscillator voltage and consequently of the oscillator frequency. Occasionally this trouble also results in "motor-boating." The method of test described above will show whether the trouble is tied in with the local oscillator or elsewhere.

Interference Due to Oscillator Harmonics

Since image response may cause interference in superheterodyne operation, similar effects which result from oscillator harmonics

beating with undesired signals are occasionally loosely referred to as images. It should be understood that image frequency response has nothing to do with oscillator harmonics.

Many types of undesired responses occur as a result of harmonics of the oscillator beating with fundamental or harmonic frequencies of other signals in the mixer circuit. For instance, if the local oscillator functions at 1450 kc, its second harmonic at 2900 kc will beat with either a 2450-kc or a 3350-kc signal and produce an i-f signal of 450 kc, in addition to that produced from the normal 1000-kc signal to which the receiver is tuned. Such undesired signals are more remote in frequency than the normal image frequencies and normally give little trouble since they are greatly attenuated by the r-f tuned circuits.

Harmonics of an incoming signal may be developed in the mixer and, beating with harmonics of the oscillator, produce an interfering i-f signal. Thus, if the set is tuned to a 10,000-kc signal and a 10,225-kc signal is likewise present in the antenna circuit, both will be amplified and present in the mixer circuit. If the receiver i.f. is 450-kc, the set oscillator will be tuned to 10,450 kc, thus producing a 450-kc beat with the incoming 10-mc signal. However, a second harmonic of the 10,225-kc interfering signal will occur at 20,450 kc as a result of plate circuit rectification, which will beat with the oscillator second harmonic of 20,900 kc to produce a 450-kc i-f signal. Though the second-harmonic signal may normally be very weak when the two incoming signals are of similar intensity, if the interfering signal is much stronger, the harmonic beat which thus results may be very troublesome.

The correction of such troubles involves reducing the avc voltage applied to the mixer tube so that this type of rectification does not result or, alternatively, by reducing the intensity of the signal by using a different antenna.

Checking the Image Repeat Point in Aligning

In aligning superheterodyne receivers on short-wave bands, care must be taken that the oscillator is not aligned to the image frequency instead of the desired frequency. This may be avoided by a simple test, as described below.

If the oscillator is designed to operate at a frequency which is higher than that of the incoming frequency, the image repeat point will occur at a frequency equal to the signal frequency plus twice the intermediate frequency. Thus, if the i.f. is 450 kc, when the receiver is properly aligned at 18,000 kc the oscillator will be functioning at 18,450 kc and the image repeat point will occur at 18,900 kc.

In normal alignment procedure, a test oscillator signal of 18,000 kc will be fed to the receiver and the h-f oscillator trimmer adjusted until maximum response is indicated on an output meter. To check the image repeat point, the test oscillator frequency is adjusted to 18,900 kc *without* readjusting the receiver tuning. A response should be obtained on the output meter if the oscillator has been properly adjusted to a frequency higher than that of the incoming signal.

In some receivers, the oscillator frequency on short-wave bands is designed to operate below the incoming signal frequency. When this is the case, it will be so stated in the service notes for the receiver. Under these conditions, alignment to an 18,000-kc signal is made as described above, but to check the image repeat point, the test oscillator frequency is shifted to 17,100 kc if the i.f. is 450 kc. A response should be obtained at this point.

Failure to obtain a signal response at the image repeat point in either of the above cases indicates that the oscillator has been aligned to the image repeat point rather than the proper frequency.

The general remarks regarding trouble-shooting in oscillators are equally applicable to pentagrid converters. In some receivers, AVC action causes the frequency of the oscillator section of the converter to change. This is particularly troublesome at the high-frequency end of the tuning range. When a strong signal is being tuned in, especially on short-wave bands, the point of resonance is not the same when the receiver is tuned from a higher-frequency point on the dial as it is when the tuning is approached from a lower-frequency point. Often it is necessary to tune back and forth around the normal point of resonance before the station can be tuned in for maximum response. Then when the receiver is tuned a little beyond this point, the signal may drop out completely and retuning in the same manner may

have to be repeated. This annoying condition occurs because a strong signal creates a high ave voltage and the high ave voltage increases the bias on the converter input grid, thereby decreasing the mutual conductance of the converter tube. This changes the tube load on the oscillator tuning circuit and causes a frequency shift which is particularly large when the shunt capacity of the tuning system is a minimum, as it is at the high-frequency end of the tuning range.

The correction of this trouble requires some modification of the ave so that less control voltage is applied to the converter grid or reduction of the signal strength of the offending station by changing the antenna location or otherwise decreasing the signal pickup. Triode-hexode converters, such as the 6K8, are substantially free from this trouble due to modifications in the design of the tube, but triode-heptodes, such as the 6J8, will be subject to this trouble unless precautions were taken in the design of the receiver to minimize such effects.

Heterodyne Oscillators

Superheterodyne receivers which are used for communication purposes are usually supplied with an auxiliary oscillator for use in the reception of continuous-wave telegraphic signals. Since these signals are unmodulated, a heterodyne oscillator is necessary to render them audible.

The usual practice is to use an oscillator which may be adjusted to a frequency which is within 1000 cycles or less of the i.f. employed in the receiver. When this oscillator signal is fed to the second detector of the receiver, it beats with the incoming i-f signal to produce an audio signal equal to the frequency difference between the i.f. and that of the heterodyning oscillator. Thus, if the i.f. is 456 kc and the heterodyning oscillator is adjusted to 457 kc, a 1000-cycle beat results which is amplified by the a-f stages, and reproduced by the speaker.

This type of oscillator is also useful in finding weak broadcast signals during periods when the station is in operation but no program is being broadcast. The heterodyne oscillator will then beat with the unmodulated carrier arriving at the second detector and may be tuned in before the broadcast program starts.

APPENDIX

Laboratory Methods of Testing Broadcast Receivers

The principal performance characteristics of a radio receiver are its sensitivity, selectivity and fidelity. In order that measurements of receivers of various types may be compared, the general procedures for making tests of these fundamental characteristics have been standardized by the I.R.E. and adopted by receiver test laboratories.

High-grade equipment must be employed if these tests are to have meaning. Further, considerable experience in making such tests is required in order to interpret unexpected results.

The Standard Dummy Antenna

All receiver measurements of sensitivity, selectivity and fidelity are made with a standard dummy antenna connected between the signal generator "high" output lead and the antenna post of the receiver. Preferably, this dummy antenna is placed close to the receiver.

The standard dummy antenna, connected for all-wave receiver testing, is shown in Fig. 16-1. The minimum impedance of this dummy antenna

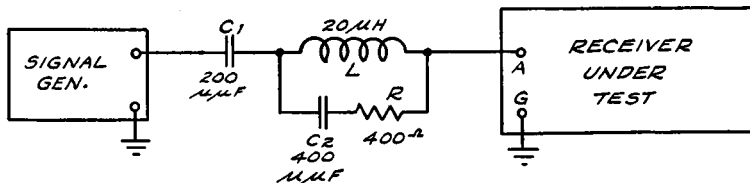


Fig. 16-1. The constants of the standard dummy antenna are shown.

network occurs at about 2.2 megacycles and amounts to 220 ohms. The output impedance of the signal generator should be negligible compared with this value or should be deducted from the respective constants of the dummy antenna.

Shielded leads should be employed in connecting the standard signal generator to the receiver through the dummy antenna. These leads should be short so that negligible voltage drop results.

Receiver Output Measuring Devices

For electrical test purposes, the standard load is a pure resistance, equal to the impedance of the receiver voice coil at 400 cycles and connected across the output circuit in place of the speaker voice coil. This dummy load should be of sufficient power handling capacity to carry the maximum power output of the receiver without changing in resistance.

The power delivered to the standard dummy load may be determined by measurements with copper-oxide rectifier-type voltmeters, v-t voltmeters or thermocouple ammeters. The meter calibration may be in r-m-s values for current, voltage or power, or in decibels.

Normal Test Output

For receivers capable of supplying at least one watt maximum undistorted power output, the normal test output is 0.5 watt in the standard output load. Receivers which supply 0.1 watt undistorted output, but not 1 watt, are tested at an output power of .05 watt in the standard dummy load.

Maximum Undistorted Output

The maximum undistorted output is arbitrarily taken as the least power output at which the total apparent power at harmonic frequencies is equal to one percent of the apparent power at the fundamental frequency. This is equal to a total r-m-s voltage reading at harmonic frequencies of 10 percent of the fundamental frequency, if measured across a pure resistance. The output power, in terms of a-c volts across a pure resistance load in ohms, is given by the formula $P = E^2/R$, where P is the output power and E is the measured audio voltage across the output load R .

Measuring Receiver Sensitivity

To make a receiver sensitivity test, the receiver is set up and the tubes are checked, making certain that those employed have average characteristics. The output meter is connected to the dummy load and the signal generator is connected through the dummy antenna to the receiver input, as shown in Fig. 16-2. If the receiver is operated from the power lines, the line voltage should be measured. If from batteries, then the battery voltages should be measured. For automobile receivers, the recommended A supply voltage is 6.6.

The receiver is placed in operation and the signal generator tuned to the test frequency. Tests should be made at three or more frequencies in each

band. An r-f signal, modulated 30 percent at 400 cycles, is then fed to the receiver. The receiver is tuned to this signal and the dial calibration is checked against the actual frequency being fed to the receiver.

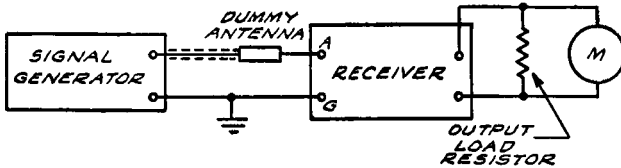


FIG. 16-2. Layout for sensitivity and selectivity tests of receivers.

The signal fed to the receiver should be initially weak, then gradually increased until the output meter reading indicates that the audio power in the dummy load resistor is 0.5 watt. The input signal in microvolts, as indicated by the calibrated signal generator attenuator, is the sensitivity rating of the receiver when all controls are adjusted for maximum sensitivity. Background noise, if present, will be indicated on the output meter when no signal is being fed to the receiver. If the reading exceeds 10 percent of that which is obtained at normal test power output, it may be compensated for by the method described above.

Measurement of Receiver Noise

Since the noise voltage which reaches the speaker is dependent upon the selectivity of the receiver circuits, the measurement of its amplitude is expressed in terms of the noise which passes through these circuits and not of the actual noise level present in the input circuit.

This measurement is made by feeding a 400-cycle modulated r-f signal to the receiver input and noting the output meter reading across the output transformer secondary dummy load. The modulation is then switched off and the output meter reading again taken. The latter reading gives the noise component alone while the former includes both noise and the 400-cycle signal voltage. The signal voltage E may be computed by the formula, $E = \sqrt{E_s^2 - E_n^2}$ where E_s equals the signal voltage in the presence of a noise voltage and E_n is the noise voltage which is present when the carrier is unmodulated. In making this test the carrier signal voltage should be from three to ten times the noise voltage.

The magnitude of the noise which reaches the output of the receiver is expressed as the *equivalent-noise-side-band input* (ENSI). This is determined by the following formula:

$$E = mE_s(E_n/E_s)$$

in which E is the ENSI, m is the modulation factor (.3 for 30 percent modulation) which is assumed to be at 400 cycles, E_s is the 400-cycle output

voltage and E_r is the output voltage from noise modulation with no 400-cycle modulation present.

Since the noise which originates in a receiver is the limiting factor in highly sensitive receivers, this measurement is important. It is possible for one receiver to be greater in sensitivity than another, yet to be less useful because its internal noise is greater at the same signal input.

Measurement of Selectivity

The selectivity of a radio receiver is expressed as the ratio of the signal strength required to produce normal test output at a frequency which differs from that to which the receiver is tuned, to the signal strength required to produce the same output at the frequency to which the receiver is tuned.

To make a selectivity curve, the receiver is adjusted for maximum sensitivity and tuned to the test frequency, usually 1000 kc in the standard broadcast band if only one point is being checked. The signal generator is connected as for sensitivity measurements and a modulated signal is fed to the receiver at the resonant frequency. The signal level is adjusted until the receiver output meter indicates normal test output, usually 0.5 watt.

The signal generator is now detuned 5 or 10 kc from the frequency to which the receiver is tuned and the test signal increased until the output meter again indicates 0.5 watt output. The ratio of the input signal off resonance to that at resonance for the same output is a measure of the selectivity of the receiver. Each ratio is computed for frequencies above and below the resonant frequency in steps of 5 or 10 kc, until the ratio exceeds 10,000 times or until the input voltage required exceeds one volt, whichever requires the least amount of detuning of the signal generator from the resonant frequency of the receiver.

Measurement of Electric Fidelity

The electric fidelity test shows the audio output of the receiver at various modulation frequencies, as applied to the r-f carrier test signal. It provides a check of the overall response of the receiver, which does not include the speaker. Measurements which include the speaker are termed acoustic fidelity tests and have far greater significance when properly made.

In making the electric fidelity test, the receiver is tuned to 1000 kc as for selectivity measurements. An input signal of 5000 microvolts at 1000 kc, modulated 30 percent at 400 cycles, is fed to the receiver as shown in Fig. 16-3 and the output is measured either across a dummy load or, if the speaker is mounted in its normal baffle or cabinet, across the speaker voice coil while the chassis is in place.

With a constant signal input, the receiver volume control is now adjusted until normal test output at 400 cycles is secured. The modulating frequency is now varied from 30 to 10,000 cycles, keeping the modulation at 30 percent, and the variation in output level as the modulating frequency

is varied is noted. The ratio of the response at a frequency other than 400 cycles to that at 400 cycles is plotted, as shown in the typical fidelity curve, Fig. 16-4. The frequency scale, as shown, should be logarithmically divided.

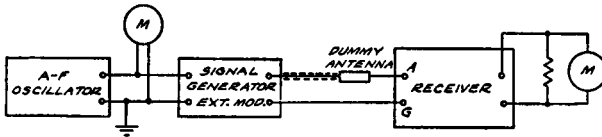


FIG. 16-3. The test setup for measurement of receiver electrical fidelity.

The output levels may be plotted in db, as shown, or in voltage or current ratios. In the latter case, the response scale should also be logarithmically divided.

If overloading occurs at any modulating frequency, the tests should be repeated at a lower reference level. Tests should be made at various settings of any controls which affect the receiver fidelity.

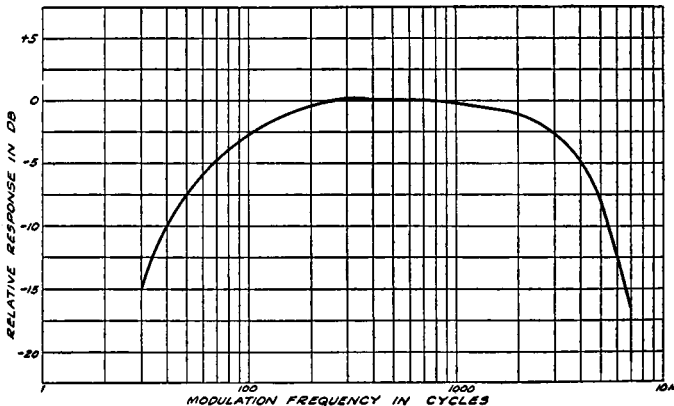


FIG. 16-4. The electrical fidelity curve of a receiver.

Measuring Automatic Volume Control Action

The operation of automatic volume control is shown by measuring the receiver output at input signals of from one microvolt to one volt. To make this test, the receiver is tuned to an input signal of 1000 kc, modulated

30 percent at 400 cycles, and the output meter reading is adjusted by means of the a-f volume control until the output does not exceed one-half the maximum power output, at any input level up to one volt. The output readings are recorded in decibel variations from the maximum output obtained with the initial volume control adjustment at a one-volt signal input.

If QAVC is used in the receiver, tests should be made with and without this feature in operation.

Two-Signal Interference Tests

While selectivity curves provide valuable data as to the band-pass characteristics of a receiver, the ability to discriminate between a desired and an undesired signal is best indicated when both signals are present. This is because the dynamic operating conditions of the receiver are different when it is tuned to a signal and therefore its selectivity characteristic is affected. The type of test to be described duplicates actual operating conditions in service, in that both the desired and undesired signals are heard and measured.

Two signal generators are required. These are preferably connected in parallel, each with its own dummy antenna in series with the receiver input. This manner of connection permits grounding both signal generators directly, whereas a series connection of the generators necessitates the use of inductive type attenuators or coupling transformers.

To standardize the measurement of cross-talk interference, the point of measurement is taken as the greatest interference input which may be fed to the receiver without the modulation output power exceeding one one-thousandth of the output from the desired modulation. Both signals are of the same percentage modulation at 400 cycles.

The receiver is tuned to a standard test frequency, such as 1000 kc, and one signal generator is adjusted to the same frequency and the output set to give a standard input signal, such as 50 microvolts ("distant-signal" voltage) 5 millivolts ("mean-signal" voltage) or .1 volt ("local-signal" voltage). After the output level is noted, the modulation is switched off. The second signal generator is tuned to frequencies up to 100 kc each side of the frequency to which the receiver is tuned, in steps of 10 kc, and the modulated signal voltage which gives interference test output of .001 of normal test output is noted and plotted.

This type of test is the only one which shows the true selectivity of a receiver equipped with avc when a strong signal is being received.

Data regarding the tests described above are based on I.R.E. reports.

Measuring Percentage Modulation with a V-T Voltmeter

Any v-t voltmeter which reads *peak* a-c voltages may be used to check modulation percentage. The measurement is made by noting the voltmeter reading which results when the modulated signal is applied and comparing this reading with that obtained when the unmodulated wave is measured. The percentage modulation is then determined by the formula

$$\frac{E1 - E2}{E2} \times 100 = \text{Percentage Modulation}$$

In the above formula, $E1$ is the reading obtained when the modulated signal is measured and $E2$ is that which results when the modulation is removed. Thus, if $E1$ is 13 volts and $E2$ is 10 volts, the percentage modulation is $\frac{3}{10}$ times 100 or 30%. This method assumes linear, distortionless modulation.

Adapting Electronic Voltmeters for Modulation Measurement

Electronic voltmeters may be adapted to read peak a-c voltages, and therefore percentage modulation by the above method, by using a diode

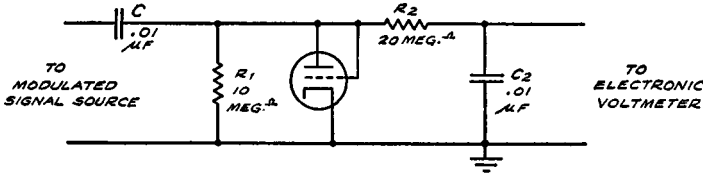


FIG. 16-5. A simple diode rectifier enables a-c measurements with an electronic voltmeter.

rectifier in the circuit shown in Fig. 16-5. Any diode, or as triode connected as a diode as shown, may be used. In our tests, a 955 was employed.

The input may be connected close to the r-f source and the output, which is filtered by $R2$ and $C2$, may be connected to the electronic voltmeter by a shielded cable. The meter may be calibrated at 60 cycles against any a-c meter. With the tube specified, this calibration will hold for frequencies up to several megacycles.

Measuring Output Tube Plate Loads

In Fig. 16-6, a method for measuring the plate load presented by an output transformer is shown. The measurement is made by connecting a v-t voltmeter as indicated on the diagram, and noting the signal-voltage reading which results when an a-f voltage is applied to the output tube grid. Another measurement is then made, without changing the input signal voltage, with a high-impedance choke connected in parallel with a variable resistor substituted for the output transformer load. The resistor R is varied until the v-t voltmeter reading is the same as that first obtained.

The resistance R is then disconnected from the choke, and its value measured with an ohmmeter. The measured resistance represents the equivalent output transformer primary winding impedance. The receiver speaker should be connected while making these tests, or alternatively, a

resistance equal to that of the voice coil impedance should be shunted across the output transformer secondary winding.

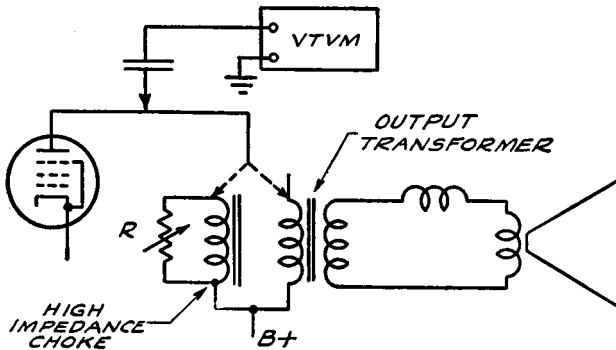


FIG. 16-6. A method of measuring primary load impedance of an output transformer.

Measuring Motor Speeds

A method of using a beat-frequency a-f oscillator in conjunction with a stroboscope disc to measure the speed of rotating devices as shown in Fig. 16-7.

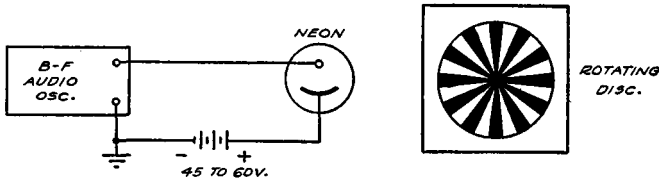


FIG. 16-7. Setup for the measurement of motor speeds.

To make this measurement, a stroboscope disc is fastened to the rotating device and the frequency of the a-f oscillator shown in the diagram is adjusted until the black segments appear stationary. Since the ignition voltage of the neon tube is higher than the usual output voltage of the average oscillator, an a-f amplifier or a series B battery (as shown) may be used to lower the a-c voltage required to light the tube. The neon tube selected should be preferably of the $\frac{1}{4}$ -watt type, with no built-in resistor.

The speed of rotation is determined by the formula:

$$\text{Speed in r.p.m.} = \frac{60 \times \text{Audio Frequency}}{\text{Number of black segments}}$$

For example, in the diagram the stroboscope disc has 8 black segments. If these segments appear stationary when the oscillator frequency is adjusted to 240 cycles, then the speed is determined as follows:

$$\text{Speed} = \frac{60 \times 240}{8} = 1800 \text{ r.p.m.}$$

A larger neon tube may be used by connecting the a-f oscillator to an amplifier with several watts output. The neon tube may then be connected through a blocking condenser to the plate of the output tube and to ground. If the output of the amplifier is sufficiently great, no series battery will be required.

Measuring Audio Amplifier Characteristics

A setup for checking the gain and electrical fidelity of an audio amplifier is shown in Fig. 16-8. The output of the a-f oscillator is measured at a level which will be conveniently indicated by the a-c voltmeter, and is then

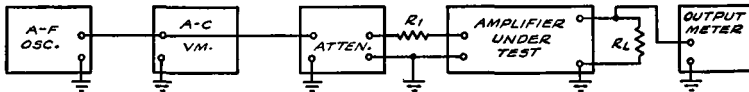


FIG. 16-8. The setup for measuring the overall gain of an a-f amplifier.

connected to an attenuator or a voltage divider before being fed to the amplifier under test. The resistor R_1 is used to represent the impedance of the microphone or other device with which the amplifier input circuit is designed to operate.

The load resistance should correspond similarly to the impedance of the load into which the amplifier is designed to work. For instance, if the output circuit is designed to work into a 500-ohm line, a 500-ohm resistor should be used.

When the input and output impedances are similar, the overall gain of the amplifier is expressed by the formula:

$$\text{Gain in db} = 20 \log_{10} E_2/E_1$$

When R_L and R_1 are dissimilar, this formula should read

$$\text{Gain in db} = 20 \log_{10} E_2/E_1 + 10 \log_{10} R_1/R_L$$

In the above formulas, E_2 represents the audio output voltage measured across R_L and E_1 represents the audio voltage in series with R_1 .

The electrical fidelity is measured by noting the output meter readings resulting when frequencies from 30 to 10,000 cycles are fed to the amplifier. The reference level is taken at 400 cycles, and at an output level well within the maximum undistorted output of the amplifier under test.

An Audio Voltage Attenuator

In Fig. 16-9, a simple audio voltage divider is shown which will enable tests of the gain of a-f amplifiers when the a-f signal source is not supplied with a calibrated attenuator.

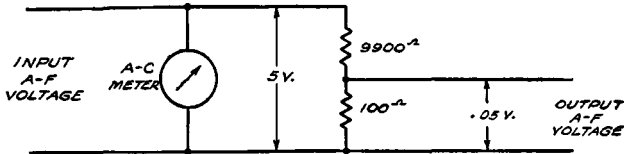


FIG. 16-9. A simple a-f attenuator.

The a-c meter, which may be a copper-oxide type, is connected across the a-f voltage source. The output control provided in the audio oscillator is adjusted until the meter reads 5 volts. The output voltage across the 100-ohm resistor will then be .05 volts, which represents the value of input signal voltage required for many receiver a-f amplifiers to provide $\frac{1}{2}$ watt a-f output.

This is a 100 to 1 voltage divider, so the output voltage will always be $\frac{1}{100}$ that applied to the input. Any other desired ratio may be obtained. In this circuit the 100-ohm resistor represents $\frac{1}{100}$ of the total attenuator resistance, 10,000 ohms. By using a larger or smaller value of output resistance, the ratio may be increased or decreased. Thus, if 1000 ohms is used, the ratio is 10 to 1. In all cases, the total resistance should remain the same.

Average D-C Voltage Across Oscillator Grid Leak

The following tabulation gives the average d-c voltage developed across the oscillator grid leak in representative modern all-wave and broadcast-band receivers at various operating frequencies. The tabulated voltages are measured with a vacuum-tube voltmeter and isolating probe, and therefore represent the actual voltages at the point of test. Ordinary 1000 ohms-per-volt meters are unsuitable for such voltage measurements, due to their loading effect upon the circuit under test.

These are average value in typical receivers; variations of 50 percent are likely to be encountered in other receivers, due to differing design characteristics. Normal variations in tube characteristics will likewise affect the resulting oscillator d-c voltage.

AC-DC Superheterodynes

| | | | | | |
|---------------------------------|-----|------|------|------|--------|
| Frequency (in kc) | 600 | 1000 | 1400 | 6000 | 18,000 |
| E_{osc} (d-c volts) | 11 | 11.5 | 13 | 10.5 | 5.8 |

A-C Superheterodynes

| | | | | | |
|---------------------------------|----|----|----|----|----|
| E_{osc} (d-c volts) | 19 | 19 | 19 | 12 | 14 |
|---------------------------------|----|----|----|----|----|

Automobile Superheterodynes

| | | | |
|---------------------------------|----|----|----|
| E_{osc} (d-c volts) | 19 | 18 | 19 |
|---------------------------------|----|----|----|

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