MIDLAND RADIO
AND TELEVISION
SCHOOLS
INC.

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

VACUUM TUBE OSCILLATORS

LESSON NO. 1

REVIEW

.......IS ALWAYS IN ORDER

How well do you remember the facts in the last lesson? How about the lesson you studied two weeks ago, and how about that lesson a few months ago on Ohm's Law? Whatever your answers truthfully are, they are entirely dependent upon how well, how thoroughly you applied your attention to those lessons.

Whether or not you feel that you know the fundamental facts contained in all your lessons, you can help yourself very greatly by reviewing them from time to time. Never overlook the value of review. A little thinking on the subject will recall to your mind some of the things you are "weak" on.....then look back and go over that entire lesson, EVEN IF YOU DID GET A GOOD GRADE.

In your spare time, pick up a lesson at random and page through it. When you come to something you don't understand, buckle down and study it.

Tracing diagrams are lots more fun to a Radioman than working crossword puzzles, and you'll learn more practical facts, too. You'll search a long time to find something more interesting than a good, "meaty" circuit. Trace the action of currents and the effects of various parts.

Keep your lessons in order and keep them handy, so that you can refer to them without having to shuffle all through them. Midland furnishes you a great variety of important information, in a large number of lesson-books. Keep them in order, just as you may expect to keep your file in order when you're "on the job".

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JONESPRINTS
KANSAS CITY, MO.

Unit Three

BASIC PRINCIPLES of RADIO TRANSMISSION

"Unit III consists of 15 highly instructive and interesting lessons. First, I start out by giving you the fundamental principles involved in using vacuum tubes as oscillators as applied to Radio Transmitter Circuits. This covers both the old self-excited type of oscillator and the new crystal controlled oscillator. Generation of the radio frequency waves necessary for the propagation of intelligence through space is included in the first lesson of this unit.

"The next part of this study covers Modulation. This is the process of combining voice waves with the radio frequency waves, so that voice and music may be transmitted through the ether. The different classes of amplifiers used in radio transmitters is covered thoroughly. The plate and filament supply sources of large vacuum tubes is also covered. Then, all of this information is combined to form a regular commercial radio transmitter with instructions for the adjustment and operation of this transmitter. The next to the last lesson in this unit covers Radio Laws and Regulations as applied to American broadcasting stations.

"After a student has completed Units I and III, he is then qualified to take his first Government examination if he so desires. While it is advisable to study Units I, II and III in their correct rotation, it is not necessary, if for some good reason you are in a hurry to secure your Government radio operator's license."

Lesson One

VACUUM TUBE OSCILLATORS

"In this lesson I am going to take up one of the most important subjects pertaining to the study of Radio Transmission.

"Radio was very slow in developing until the vacuum tube oscillator was invented. Because of its importance you should study the material in this lesson very carefully. Most of the lessons in this unit will be founded on the fundamental principles set forth in this lesson."

1. **GENERATION OF HIGH-FREQUENCY CURRENTS.** An oscillator may be defined as a device which produces a high-frequency current. Strictly speaking, it is not correct to state that an oscillator is a generator, for it does not generate the high-frequency currents. It is more correct to designate it as a converter. Power is supplied to the oscillator circuit in direct current form. The DC voltage of the B battery or the power supply and the DC current drawn by the oscillator tube together furnish the DC power input to the oscillator. The only source of energy in the circuit is that provided by the power supply. The DC energy supplied to the circuit is converted by it into alternating energy of high-frequency, which may be used for various purposes.

The major use of an oscillator is to produce the high-frequency carrier current required for the transmission of radio signals. Ordinary power frequencies are not radiated by an antenna. To cause appreciable radiation, the frequency of the antenna current must be about 12 kc. or higher. The frequency of the current generated by an alternator depends upon the number of poles and the speed of revolution, increasing either results in an increase in the frequency. High-frequency alternators have been constructed and used in the past, but their inherent disadvantages compared to those of the vacuum tube oscillator have made them obsolete. About the highest frequency generated by mechanical means is 200 kc. Such a machine is the Alexanderson alternator, which was once widely used in long-wave stations. One disadvantage of this type of machine is the tremendous speed required, as may be seen from the fact that the speed of revolution must be 20,000 rpm. to produce a frequency of 100 kc. where there are 600 effective poles.

Other means which have been employed to generate a high-frequency current for communication purposes are the spark coil, the Poulsen Arc, the Goldschmidt Alternator, and medium-frequency alternators used in conjunction with frequency-doubling transformers. As all of these methods have been superseded by the vacuum tube oscillator, which is far more reliable and is capable of producing currents of much higher frequency, the remainder of this lesson will be devoted to a detailed discussion concerning its operation under various conditions.

Besides the necessity of oscillators to produce high-frequency carrier currents, minor uses include the local oscillator required in every superheterodyne receiver, and the signal generator needed for testing and aligning radio receivers. Vacuum tube oscillators have been used to produce frequencies as high as 600 mc. or higher, although frequencies in excess of 100 mc. have as yet found no practical application.

2. CONDITIONS NECESSARY TO ESTABLISH CONTINUOUS OSCILLATIONS. To produce continuous oscillations there is needed, first, an oscillatory circuit consisting of a coil connected across a condenser; second, some means of supplying energy to this oscillatory circuit to prevent the oscillations from dying out. The mere supplying of energy to the oscillatory circuit offers no serious problem, but the requirement that this energy be alternating, have exactly the same frequency as that of the oscillating current, and be perfectly in phase with it, presents a more complicated, though not insurmountable, difficulty.

To state that the energy introduced into the oscillatory circuit must be in phase with the energy already present in this circuit means that when the R.F. voltage across the tuning condenser is rising in such a direction as to increase the instantaneous voltage on the grid, the energy representing the feedback must also increase the voltage on the grid. In a like manner, as the tuned circuit voltage reverses direction and drives the grid to a lower voltage, the feedback voltage must change direction simultaneously.

The problem of furnishing energy at periodic intervals, and in an amount sufficient to compensate for that dissipated in the resistance of the oscillatory circuit, and thus maintain the amplitude of the oscillations at a constant level, is best solved by amplifying a small portion of the energy contained in the tuned circuit and then returning this amplified energy to that circuit in the correct amount and phase relationship. Since the original energy was obtained from the tuned circuit, it will of necessity have the correct frequency; thus, one of the problems is automatically solved, and by adjusting the magnitude and phase of the amplified energy, the two remaining points are taken care of without undue difficulty.

As the vacuum tube represents the only practical means of obtaining amplification, it is natural and logical that it should be used for securing the amplified feedback. Thus the fact that the alternating energy in the plate circuit of a vacuum tube is greater than that present in the grid circuit, or that the tube is capable of producing amplification, is the only reason that it is possible to use a vacuum tube for the production of continuous, high-frequency oscillations.

The influx of energy into the tuned circuit must, as previously stated, be equal to the outgo of energy from this circuit, else the amplitude of the oscillations will change. While the incoming energy has a single source, the extraction of energy from the circuit follows three clearly defined channels. The first of these is the dissipation of power caused by the oscillating current flowing through the resistance of the tuned circuit. The second is represented by the energy taken from the circuit, which, after amplification, will be returned as the feedback. The third is the power derived from the circuit for the purpose of transmitting a radio signal. It is through this third channel that the majority of the power contained in the tuned circuit flows. This is desirable, since the purpose of creating the high-frequency power is to use it for some practical application. Thus, an attempt is made to make the power lost through the first two channels as small as possible.

Just how energy is extracted from the tuned circuit for use in an external circuit, may not at first be clear. It is usually accomplished by coupling a pick-up coil magnetically to the inductance of the tuned circuit. Thus energy is transferred by mutual induction from the tuned circuit to the pick-up coil and thence to the point where it will be used. Although no physical connection exists between the two coils, the interaction of the magnetic fields surrounding them serves to take energy from the tuned circuit and add it to the external circuit. This operation is effected by a phenomenon known as "reflected impedance." A complete discussion concerning this action will be presented in Lesson 4 of this Unit. For the present, it is sufficient that we understand that the presence of the pick-up coil causes the dissipation of high-frequency energy in exactly the same manner that the addition of a resistance to the tuned circuit would do. Thus, it may be said that, in effect, the pick-up coil has reflected a resistance into the tuned circuit.

3. **REVIEW OF THE ACTION OF A HARTLEY OSCILLATOR.** The Hartley oscillator is probably the easiest to understand, and it is for that reason that this oscillator was explained in Lesson 23, Unit 1, to enable you to comprehend the operation of a superheterodyne receiver. The power obtainable from the local oscillator in a superheterodyne receiver is very small, and the efficiency of the conversion of the DC power input to the R.F. power output is of little consequence. This, however, is not true of oscillators used to produce large amounts of R.F. power. In this case, the efficiency is of prime importance and various adjustments are made to secure a reasonable amount of efficiency even at the sacrifice of some power output.

As some of the characteristics of a Hartley oscillator may have slipped your mind, it is advisable to review briefly the action of this circuit. At the outset it must be remembered that the sole purpose of any oscillator is to produce high-frequency power; it is not an amplifier, nor is it a detector, and as such it operates under conditions not found in ordinary amplifiers or detectors.

The first of these conditions is that the grid of the tube is driven positive during a portion of the input cycle, and as a result, some grid current flows. The second condition is that the bias is secured by allowing this grid current to flow through a grid leak, instead of using a C battery or cathode bias. Finally, the normal grid bias is so great that plate current flows only during a small portion of the input cycle.

To review the operation of a Hartley oscillator, the circuit shown in Fig. 1 will be used. This circuit is somewhat different from the one used during the previous explanation. It is at once seen that no plate blocking condenser is used; thus both the RF and DC components of the plate current flow through that part of the tank circuit connected between P and F. This circuit is said to be series fed, while the one shown in Lesson 23 in Fig. 8, is shunt fed, since two separate paths are provided for the two components. The RF component of a shunt-fed oscillator flows from the plate through the blocking condenser to the tank circuit, and thence to the filament, while the DC component flows from the plate, through the RF choke, through the power supply to the filament. In the series-fed oscillator, both components follow the same path except that a bypass condenser (C2) is connected across the power supply to prevent the RF current from flowing through it, as such a procedure is apt to damage the supply.

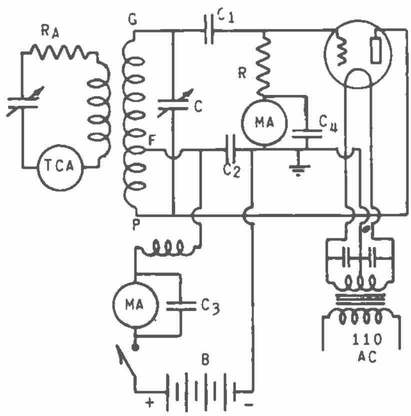


Fig.1 A series-fed Hartley oscillator.

The grid circuit of the Hartley oscillator shown in Fig. 1 is shunt fed, whereas the Hartley oscillator previously studied employed a series fed grid. In Fig. 1, two separate paths are provided for the DC and RF components respectively, of the grid current. The RF component of the grid current will pass through the blocking condenser C1, and the DC component will pass through the grid leak resistor R. Since the RF and DC components have separate paths, the grid circuit is said to be "shunt fed". Both the grid and plate current milliammeters should be by-passed to prevent an RF current flow through them. The grid meter is protected by the condenser C4 and the plate meter by the condenser C3. Both meters are ordinary DC instruments and indicate the average value (DC component of the pulsating high frequency current). If the RF component were permitted to pass through these meters, the fine wire composing the moving coils would be heated excessively and undoubtedly would be burned in two.

The procedure of starting and maintaining the oscillations occurs as follows: The filament voltage is applied and the filament is allowed to reach its operating temperature. The grid is at the same potential as the filament, since no oscillating current has begun to flow in the tank circuit. When the switch in the plate circuit is closed, a rising plate current flows through the tube and through the lower section of the tank. In rising from zero, this current creates an expanding magnetic field about the tank coil, which, in cutting the turns of the coil, induces a voltage across the inductance in such a direction as to oppose the rise of current. Thus the lower end of the coil becomes negative with respect to the top, while the point to which the filament is connected is intermediate in potential.

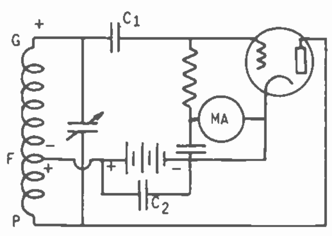


Fig. 2 Illustrating the polarity of the voltages across the tank circuit during one alternation of the oscillating current.

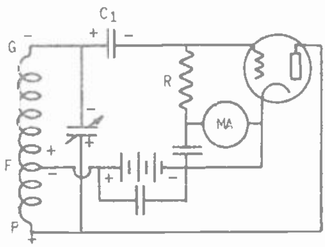
It is not hard to see that a voltage of this polarity will oppose the rise of the plate current, when it is realized that this voltage bucks against the output of the power supply, as may be seen by the diagram in Fig. 2. As this voltage across the tank is increasing, the point G of the tank is becoming positive with respect to the point F, and as a result, the grid voltage is increasing in a positive direction. With the grid growing more positive, the plate current will rise to a value far above that which would have been obtained had the grid been unaffected by the increase in plate current.

It is to be noticed that there are two factors involved which interact directly with each other. One is the plate current and the other is the grid voltage. An increase in plate current causes an increase in grid voltage, which in turn produces a further increase in plate current. It might be thought that this action could continue indefinitely; however, it should be remembered that the actual grid voltage depends not upon the value of the plate current, but upon how rapidly it is increasing. Thus, it is probable that it increases at the most rapid rate during the first few instants of its flow; after which the increase is at a slower rate. Therefore, the grid is at its maximum positive value during the first few instants of the plate current flow, and it begins to decrease its positive voltage as the rise of plate current becomes less rapid. Thus, there is a maximum plate current which will be reached during this first increase.

Throughout this interval when the grid is becoming more positive, the condenser C is being charged. Electrons are removed from its upper plate and added to its lower plate. When the grid voltage reaches the maximum positive value that it can attain during this cycle, the condenser immediately starts to discharge. As the voltage across C is decreasing, the grid voltage is falling from its peak positive value, and the plate current is beginning to decrease. The field produced by the flow of plate current through the lower section of the tank collapses, and the voltage induced across the tank inductance is of such a polarity as to tend to maintain the plate current flow.

Thus, point G is negative with respect to point P, while point F is at an intermediate potential. The voltage produced between the points F and P is in such a direction as to add to the plate voltage and thereby maintain the plate current flow. This is seen by reference to Fig. 3. Since the induced voltage is reducing the grid voltage, the plate current steadily falls despite the effort of the voltage between F and P to keep it at a constant value.

Fig.3 The polarity of the voltages across the tank during the alternation following that shown in Fig. 2.



The plate current continues to decrease and the induced voltage it causes, drives the grid more and more negative. Finally the grid is at such a high negative potential that plate current is unable to flow. During this time, condenser C has discharged and the discharging current in flowing through the tank inductance has produced a voltage across the coil which caused the condenser to charge in the opposite direction. At the present time, therefore it is charged to its maximum voltage, with its top plate negative and its bottom plate positive.

By the increase and decrease of the plate current, an oscillating current has been established in the tank circuit, and if no further energy were added to the tank, the oscillations would gradually die out. The oscillating current would dissipate power in flowing through the resistance of the tank and during each alternation the condenser would be charged to a lower voltage. However, each time that the voltage across the tank drives the grid positive, the plate current will increase, and as the tank voltage reverses polarity, the grid will be driven negative, resulting in a decrease in plate current. Since the increases and decreases of plate current cause induced voltages across the tank which add to the voltages produced by the oscillating current, the energy in the tank circuit remains at a constant level, and the tank condenser is charged to the same peak voltage during succeeding al-

The frequency of the oscillating current will be largely determined by the natural frequency of the tank circuit, which depends upon the value of inductance and capacity composing the tank. We shall learn, however, that there are other factors which influence the oscillation frequency, and the actual frequency of the oscillations is that one at which the voltage fed back from the plate to the grid is of exactly the proper magnitude and phase to enable the tube to supply its own input voltage.

The voltage between G and F is the grid-exciting voltage, while that from F to P is the feedback voltage. Each time that the grid is driven positive, some grid current flows. The grid current, which is a pulsating direct current, has a DC and an AC component. The DC component is unable to flow through the condenser, and must flow through the resistor. The AC component finds a much easier path through the capacitive reactance of the condenser than through the resistor, and therefore the voltage drop produced by the DC component flowing through R is a direct voltage. To obtain these conditions, the capacitive reactance of the condenser to the oscillation frequency must be considerably less than the resistance of the grid leak R.

4. ANTENNAS. While the Hartley oscillator and other oscillators which will be discussed in this lesson are no longer used in commercial transmitters, their principles are so fundamental that they must be thoroughly comprehended before an understanding of modern crystal oscillator circuits is possible.

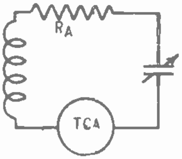
As will be pointed out later in this lesson, it is impossible to obtain appreciable power out of any oscillator circuit unless some circuit is used in conjunction with the oscillator to absorb power from it. This is often the antenna circuit, and power is transferred directly from the oscillator tank to a coupling coil which feeds the antenna. As was stated in Lesson 24 of Unit 1, an antenna has inductance, capacity, and resistance. The inductance is due to the coupling coil and to the antenna itself, since even straight conductors possess some inductance. The capacity is composed of the capacitive effect between the antenna and ground. And the resistance is the natural ohmic resistance of all the conductors in the circuit.

Transmitting antennas are always resonated to the carrier frequency, since the power radiated from an untuned antenna is very small. Lesson 6 of Unit 4 will deal with the method of tuning an actual antenna, and in addition, the design principles of various antennas will be discussed.

To adjust an oscillator for the desired power output and efficiency, while it was coupled to an actual antenna would be undesirable, since considerable interference to other stations might result during the adjustments. It is, therefore common practice to construct an artificial or "dummy" antenna to be coupled to the oscillator during the preliminary adjustments. This dummy antenna consists of a coil, a variable condenser, and a resistor connected in series. The resistor is a high wattage type and could consist of the heating element of an electric heater. The actual value of the resistor will depend upon the expected power output of the oscillator and may be any value from about 5 ohms upward. A circuit diagram of a dummy antenna is shown in Fig. 4.

A thermocouple ammeter is included in this circuit to measure the antenna current. The power absorbed by the dummy may be calculated by squaring the antenna current and multiplying it by the value of the resistor (I^2R law). This assumes that the majority of the resistance of the circuit is contained in the resistor and

Fig. 4 A dummy antenna as it is shown in schematic diagrams.



that the resistance of the coil, the condenser, and the connecting leads is negligible. The oscillator and antenna coils are ordinarily constructed of copper tubing or flat copper strip to insure that their resistances are as low as possible. The tank current and antenna current of medium-powered oscillators are usually several amperes and every precaution must be taken to minimize the loss of power by reducing the resistance of the circuit to a low value.

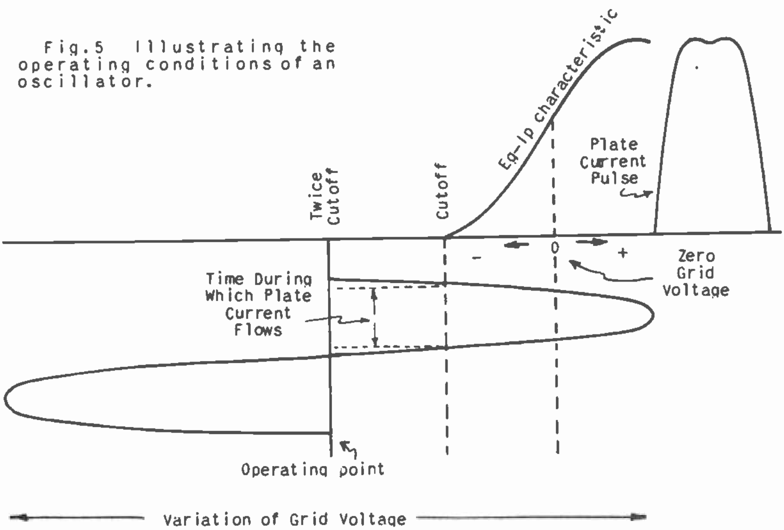
The antenna resistor is, of course, purposely included to provide a means of absorbing power. In actual operation, the power is lost from the antenna by radiation.

5. VOLTAGE AND CURRENT RELATIONS IN AN OSCILLATOR. Before it will be possible to discuss the methods of adjusting an oscillator for the desired power output with a reasonable amount of efficiency, the voltages and currents of the grid and plate circuits must be further investigated. The grid bias obtained by allowing the DC grid current to flow through the grid leak resistor is ordinarily enough to fix the operating point at a position far beyond the cutoff value of the plate current. In fact, the actual grid bias may be anywhere from 1½ to 3 times the cutoff value.

The value of negative grid bias required to completely stop the flow of plate current is known as the cutoff value and its approximate amount may be calculated by the following method: It is known that both the grid and plate voltage control the flow of plate current, but that the grid voltage is far more effective in its controlling action than is the plate voltage. The ratio of the effectiveness of the grid voltage to that of the plate voltage is called the "amplification factor" of the tube. The amplification factor or "mu" of a tube was defined and discussed in a previous lesson, but is reintroduced at this time to refresh your memory. If the mu of a tube were 10, the grid would be 10 times as effective in controlling the plate current as is the plate voltage. Thus an increase of 1 volt on the grid in the positive direction would produce the same plate current increase as would an increase of 10 volts on the plate.

Let us assume that the plate voltage of a tube is 500 volts and that the grid bias is zero. Under this condition, the plate current would be rather high. If the mu of this tube is 10, changing the bias to -50 volts would be equivalent to reducing the plate voltage by 10 × 50 or 500 volts. Such an action would reduce the plate voltage to zero and no plate current would flow. Therefore, the grid bias has been made sufficiently negative to stop the flow of plate current. It is, therefore, evident that the amount of grid bias required to produce cutoff is equal to the plate voltage divided by the amplification factor of the tube. The result of this calculation is only approximate since the assumption was made that the tube's characteristic is a straight line. As no account was taken of the curvature of the tube's characteristic, this value will be slightly less than that needed to stop the flow of plate current.

Fig.5 Illustrating the operating conditions of an oscillator.



Since the tube is operated at a value beyond cutoff, the plate current flows for less than one-half of the grid exciting cycle, as may be seen by reference to Fig. 5. In this figure, the grid bias has a value equal to about twice cutoff, and the plate current consists of a series of pulses of very short duration. Since the plate current flows during less than half of a cycle, the tube is idling the majority of the time. Most oscillators have sufficient grid bias to limit the flow of plate current from 1/3 to 1/6 of the grid voltage cycle.

It is, of course, the alternating voltage produced across the tank circuit from G to F in Fig. 1 that overcomes the bias voltage, drives the grid positive, and allows grid current to flow which produces the bias voltage. If this voltage were not great enough to overcome the bias, the grid would not be driven positive; no grid current would flow and the bias would decrease sufficiently to allow the tank voltage to drive the grid positive. Thus, the bias produced by a grid leak is self-regulating, and the proper operating point is automatically selected. If battery bias were employed, the oscillator would not be self-starting, as the original bias would be so great that no plate current would flow.

In Fig. 5, it was assumed that the plate voltage was constant, and that the time during which the plate current flowed depended only upon the bias voltage, and the amplitude of the grid-exciting voltage. This, however, is untrue since the actual voltage between the plate and filament is also varying at an RF rate. As point G becomes more positive, point P is becoming more negative, and the voltage between F and P is bucking the voltage of the power supply. Thus, the actual voltage on the plate is being reduced. Since point G reaches its maximum positive voltage at the same instant that point P attains its maximum negative voltage, the actual voltages applied to the grid and plate are 180° out of phase. Then when the grid voltage is at its most positive value, the plate voltage reaches the lowest point of its variation. This information is shown in graphical form in Fig. 6.

The line marked filament potential is the zero axis or the line from which all voltages are measured. The voltage of the power supply is represented by a line drawn above the zero axis at a distance equal to the voltage of the power supply. Superimposed on this line is a sinusoidal voltage which represents the RF voltage produced across the points F and P of the tank circuit. It may be seen that the actual voltage on the plate varies from a value almost twice as great as the DC plate voltage to a very low value, not much greater than zero.

The bias voltage is indicated in the diagram by a line drawn below the zero axis at a distance equal to the grid bias. The alternating grid voltage component is shown as a sine wave having the bias line as its axis. Since the grid is driven positive during a small part of the cycle, this sine wave crosses the zero axis and enters the positive territory. It is to be noticed that the maximum grid potential almost reaches the minimum plate potential and, as will be discussed later in this lesson, it is not desirable to allow the maximum positive grid voltage to be greater than 80% of the minimum plate voltage, since these two voltages are attained simultaneously.

The plate current flowing at any instant will be the result of the combined effect of the grid voltage and the plate voltage at that instant, and the shape of the plate current pulse, as well as the duration of time it flows, depends upon the total voltage applied to these two elements. The plate current and grid current that flow as a result of the voltages of Fig. 6, are shown in the same diagram. These graphs are drawn under each other and vertical lines are included so that the relative voltages and currents present at any instant of the cycle may be determined.

The dotted lines drawn through the grid and plate current pulses represent their average values or the values of their DC components. It is these values which are read on the grid and plate meters. The average value of such a current pulse depends

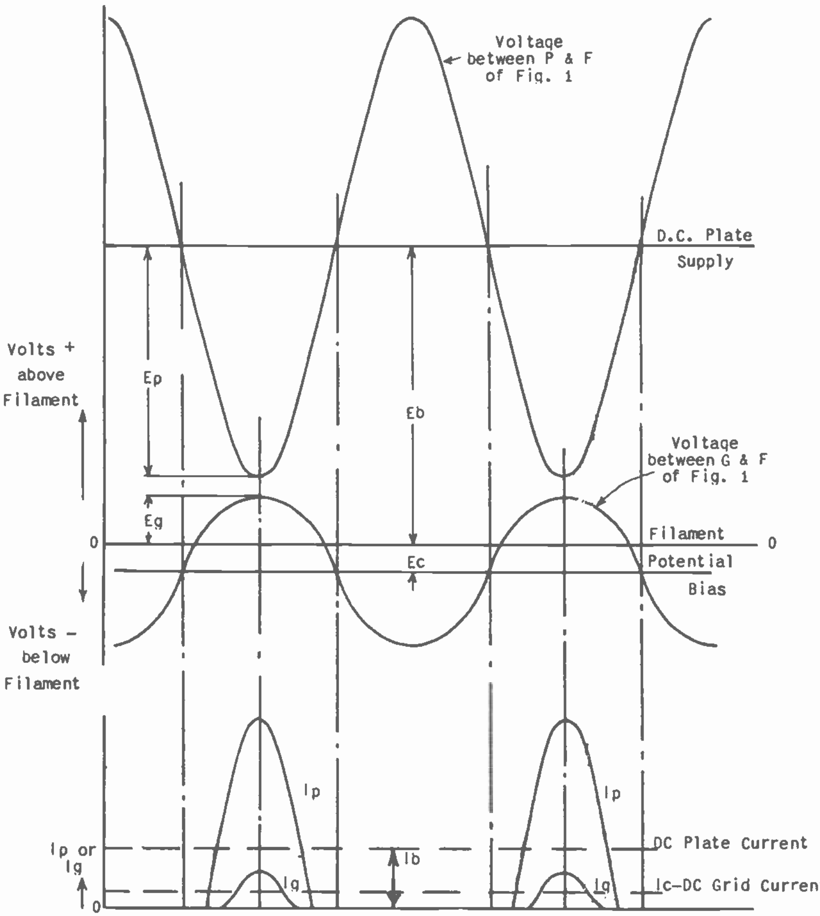


Fig.6 Showing how the plate voltage, grid voltage, plate current, and grid current of an oscillator vary during operation.

upon the maximum current reached, the time during which current flows, and, in general upon the shape of the pulse. The plate current represented in this diagram flows for 120° or 1/3 of a cycle. The grid current flows for even a shorter interval of time, since the grid is positive during a very small part of the cycle.

Let us now summarize the foregoing information. To the plate and to the grid there are applied DC voltages, and alternating voltages of sine wave form. The DC voltage applied to the plate is positive, and that to the grid is negative. The sine wave voltage applied to the plate is 180° out of phase with that applied to the grid. The plate current which flows as a result of these voltages is a pulse of short duration, generally lasting less than 1/3 cycle, and the plate current is ordinarily not of sine wave form. The grid current pulse is similar to that of the plate current, except that it flows for a smaller part of the cycle, and is somewhat more peaked; therefore, it departs even more from true sine wave form.

In the discussion of harmonic frequencies in Lesson 26 of Unit 1, it was pointed out that any complex wave; that is, any wave differing from true sine wave form, was composed of a fundamental frequency and a series of harmonic frequencies. The plate and grid current pulses are complex waves, and may be considered to be composed of a DC component, a fundamental frequency component, a second harmonic component, and other components of higher harmonic frequencies. The fact that the grid current possesses harmonics is of little consequence, but the harmonic frequencies contained in the plate current are of considerable importance.

It is the plate current pulses flowing through the tank circuit that maintain the oscillations at a constant amplitude, and it is now evident that part of the energy fed back is of the fundamental frequency, and the remainder is of harmonic frequencies. It is desirable that the power output of an oscillator consist only of fundamental energy, otherwise some harmonic frequencies will reach the antenna where they will produce sufficient radiation to interfere with other stations whose assigned frequencies are harmonics of the oscillator fundamental.

To produce only fundamental power in the tank circuit requires that the oscillating tank current flow in pure sine waves. Thus, the tank circuit must be able to smooth out the pulses of plate current into pure sine waves, and to accept energy from them only at the fundamental frequency. To accomplish this, the oscillating current must be fairly high, and the power contained in the tank must be large compared to that added or extracted during a cycle. It is for this reason that this circuit is called a tank. It can be thought of as a large reservoir of energy to which a small amount of energy is added, and from which an equally small amount is taken during each cycle.

The action of a tank circuit is very similar to that of a flywheel. In a gasoline engine, the energy is obtained by exploding a mixture of gas vapor and air in a cylinder. The explosion forces down a piston which then transmits its energy by means of a crankshaft to the flywheel. The flywheel does not receive energy at a smooth continuous rate, but in the form of spurts of short duration. Since the flywheel is very heavy, it can, by its motion, store a considerable amount of kinetic energy, which is released in a smooth continuous flow to the point where it will be used. Thus, the flywheel smooths out the pulses of energy it receives, in the same manner that the oscillating tank current smoothes out the pulses of plate current. It is for this reason that the tank current is said to have a "flywheel" effect.

It should be realized that it is impossible to eliminate harmonics completely, since energy is added to the tank only during a small part of one alternation. During the succeeding alternation when no plate current flows, the oscillating current begins to die out, and the peak amplitude of the negative alternation of the oscillating current is slightly less than the positive alternation. Naturally, the power dissipated in the tank between the times that energy is added is extremely small, yet it is sufficient to distort the waveform of the oscillating current very slightly and cause the production of a small amount of harmonics.

Since increasing the grid bias to the point where the plate current consists of pulses which have a high harmonic content is the major cause of the production of harmonics, it might well be asked why this is done. It would seem more reasonable to reduce the grid bias and allow the plate current to flow for 360° or through a complete cycle. In this case, the plate current would be more nearly of sine wave form and the flow of energy into the tank circuit would be much smoother. The reason that this is not done is that such a circuit would be very inefficient. Most oscillators are operated at very high power and the efficiency of the conversion of DC input power to RF output power is of major importance. Before entering the subject of efficiency, it is first necessary to discuss the power relations existing in an oscillator.

6. POWER RELATIONS OF AN OSCILLATOR. The power drawn from the power supply at any instant is the product of the voltage of the power supply and the value of plate current at that instant. When the peak plate current is flowing, the power taken from the supply is very large, while during the time that no plate current is flowing, no power is being drawn from the power supply. The power supply voltage is constant and is usually represented by the symbol (E_B). The value of the plate current, of course, varies from instant to instant, and the average power taken from the power supply is equal to the power supply voltage times the average of the plate current pulse or its DC component. The DC component is represented by the symbol (I_B). The average power which the power supply furnishes is called the DC power input, and expressed as an equation, it is:

$$P.I. = E_B \times I_B \qquad (1)$$

Of the total power input applied to the tube, a part is used to supply energy to the tank circuit, and the remainder is dissipated within the tube in the form of heat. The instantaneous power furnished to the tank is equal to the instantaneous voltage developed across the points F and P of the tank times the instantaneous value of the fundamental component of the plate current pulse. And, in a like manner, the average power furnished to the tank is equal to the RMS voltage between the points F and P times the RMS value of the fundamental component of the plate current pulse. The average power dissipated in the tank, due to all causes, is necessarily equal to the average power put into it, if continuous oscillations are to be maintained. It is possible to calculate the average power furnished to the tank by finding the RMS value of the plate current pulse by a tedious and intricate mathematical computation, but this is unnecessary. The useful power taken from the tank of an oscillator is found by measuring the antenna current by a thermocouple ammeter, and then using the I^2R law to find the power.

$$P.O. = I^2_a \times R_a \qquad (2)$$

Where: Ia is antenna current, and
Ra is antenna resistance,

The efficiency of the power conversion is the ratio of the RF power output to the DC power input expressed in percent. Stated as an equation, this is:

$$\text{Efficiency} = \frac{I^2a \times Ra}{I_B \times E_B} \qquad (3)$$

Where: Ia is the antenna current in amperes,
Ra is the antenna resistance in ohms,
I_B is the DC plate current in amperes,
E_B is the DC plate supply voltage in volts.

The power dissipated within the tube is caused by the plate current flowing through the plate resistance of the tube. So far as the energy source is concerned, the tank circuit between the points F and P is in series with the filament-to-plate resistance of the tube; therefore, the plate current flows through both of these elements, and in so doing dissipates power in each. That furnished to the tank is useful, while that consumed within the tube is wasted. The power lost in the tube manifests itself in the heating of the plate.

In their flight from the filament, the electrons are rapidly accelerated, and by the time they reach the plate, they have acquired tremendous velocities. These high velocities represent considerable kinetic energy, which is immediately converted into heat energy as the electrons strike the surface of the plate. Thus the temperature of the plate may be raised to a dull red heat, and if the power lost in the tube becomes sufficiently great, the plate becomes white hot and the glass walls of the tube soften, resulting in complete ruination of the tube.

The manufacturers of transmitting tubes rate them according to the power that may be safely dissipated at the plates without increasing the temperature to a dangerous value. The actual power input that may be safely applied to a tube depends upon the allowable plate dissipation, and upon the efficiency of power conversion. Oscillators operate at efficiencies ranging from 50%

to 80%, with the lower value being far more common.

As an example, the type 801 tube, when used as an oscillator, has a maximum plate dissipation of 20 watts. If the efficiency is 50%, a total of 40 watts of DC power input may be safely applied. Of this 40 watts, 20 are converted into RF power and are furnished to the tank; the remaining 20 watts appear as heat dissipation at the plate of the tube, causing it to operate at a dull red heat. Nearly all the tubes when operated at the maximum plate dissipation, show some color in their plates. As long as this color is a dull red, or, as it is sometimes called, a cherry red, the tube is operating at a safe temperature.

If the efficiency in the previous example were increased to 60%, the DC power input could also be increased. Under this condition, 60% of the input power is converted to RF power, and 40% of it is the plate dissipation. To find the allowable power input, a percentage problem must be solved. It is necessary to determine the number of which 20 is 40%. This is 20 ÷ 40% = 50 watts. If the efficiency and plate dissipation are known, the allowable DC power input may be found from this equation:

$$\text{Allowable DC power input} = \frac{\text{Plate Dissipation}}{1 - \text{Efficiency}} \qquad (4)$$

Where the efficiency must be expressed as a decimal fraction.

The instantaneous power lost at the plate is equal to the instantaneous plate voltage times the instantaneous plate current. The average plate dissipation is equal to the average instantaneous power loss, averaged throughout a complete cycle. Since the problem of averaging all these instantaneous powers is an extremely complicated mathematical procedure, the power loss is ordinarily calculated by finding the difference between the power input and the power output. When stated as an equation, this is:

$$\text{Power dissipated at plate} = (I_B \times E_B) - (I^2_a \times R_a) \qquad (5)$$

Where the various symbols represent the same quantities as in equation (3). Fig. 7 illustrates how the input power divides between the tank and the plate of the tube during the time that plate current flows. It should be noticed that the power loss at the plate starts to rise as the plate current begins to flow, but that after it reaches a certain value, it remains fairly constant at this value, and then finally falls to zero as the plate current stops. The reason that the plate power loss does not continue to increase as the plate current rises, is that the plate current rises, is that the plate voltage is falling during this time, and thus the power loss is fairly constant since the plate voltage is falling at practically the same rate at which the plate current is rising. The plate voltage falls at the same time that the plate current rises because they are 180° out of phase. The falling of the plate voltage is due to the voltage built up between F and P (Fig. 1), which during this part of the cycle, bucks against the plate supply voltage.

It can be seen that the plate current flows during the time that the plate voltage is fairly low; therefore, the plate power loss is kept at a minimum and the efficiency is increased. It is for this reason that the tube is operated at a high negative bias far beyond cutoff, and the plate current is allowed to flow only during the times that the plate voltage is low. Such an action reduces the plate loss and increases the efficiency. Allowing the plate current to flow only during a short time of each

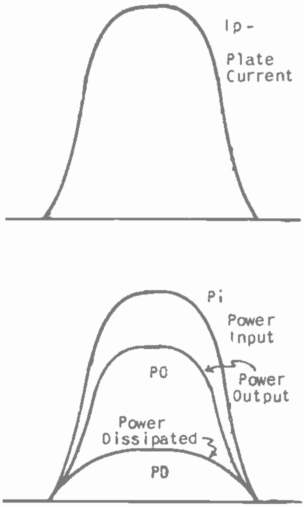


Fig.7 Showing how the power input, power output, and power dissipated vary during the time that plate current flows.

cycle makes the average plate current rather low, and thus the power input to the tube is low. Even though the efficiency under these conditions is quite high, the power output obtained from the oscillator cannot be very great since the power input is so low. It may, therefore, be seen that a compromise must be made between efficiency and power output. High efficiency may be obtained at low power output, or high power output can be secured at low efficiency. Usually the oscillator is adjusted until the desired power output is obtained at the maximum possible efficiency. If this value of efficiency is not 50% or greater, it indicates that an attempt is being made to secure more power from this tube than is economically practical, and that a tube using a higher DC plate voltage and current, and having a greater plate dissipation should be employed.

In adjusting the power input, care must be taken that neither the maximum DC plate voltage or plate current, as recommended by the manufacturer, is exceeded. The maximum allowable plate voltage is determined by the insulation resistance between the plate and the other electrodes of the tube. If the manufacturer's rating for a particular tube is 1000 volts, the actual variation of the plate voltage will be from nearly 0 volts to almost 2000 volts, since the alternating component of the plate voltage is nearly equal to the applied DC voltage. The fact that the plate voltage of the tube will increase to nearly 2000 volts has been taken care of in the manufacturer's calculations, and the tube will be able to withstand this voltage, since it is maintained only for a short interval during each cycle. If an attempt were made to increase the DC plate voltage above its specified rating, the maximum voltage reached during the cycle would be very high, and it is more than probable that the tube would arc from the plate to one of the other electrodes, thereby causing serious damage.

For a similar reason, the maximum DC plate current rating should not be exceeded, as such an action greatly increases the power dissipated within the tube. The maximum peak plate current that can be drawn depends upon the emission capability of the filament. The grid of an oscillator tube is ordinarily driven sufficiently positive to produce saturation current at the comparatively low plate voltage present at this instant. That is, the grid voltage increases until any further increase in grid voltage would produce no further increase in plate current. This is caused by the upper bend of the grid voltage-plate current curve.

With the grid-driving voltage and bias constant, the maximum plate current that flows depends upon the minimum plate potential reached during the cycle. When the DC plate current is increased, it indicates that plate current is flowing for a longer portion of the cycle, and the plate loss increases at a very rapid rate. In fact, tubes having plate dissipation ratings greater than 1000 watts employ water-cooled plates. The subject of water-cooled tubes will be discussed in Lesson 9 of this Unit.

In addition to the power lost at the plate of the tube, there is a power loss in the grid circuit. This power loss consists of two parts; the power dissipated by the DC grid current flowing through the grid leak, and the power lost at the grid itself caused by the electrons striking the grid wires. This power, as well as that used to heat the filament, is usually not taken into account in figuring the efficiency. All the power lost in the grid circuit could be eliminated by not allowing the grid to go positive, but, as previously explained, the use of grid leak bias makes possible the setting of the operating point at a value beyond cutoff with the consequent reduction in plate power loss and increase in efficiency. The advantages to be gained by allowing the grid to go positive far outweigh the disadvantages.

7. FACTORS AFFECTING POWER OUTPUT AND EFFICIENCY. The effect of increasing the size of the grid leak resistor is shown in Fig. 8. It may be seen that a larger resistor increases the grid bias as shown by the solid line. The dotted lines indicate the conditions existing with the smaller grid leak. Since the grid-exciting voltage remains practically constant, the effect is to decrease the amount which the grid goes positive. This decreases the fraction of the cycle during which plate current

flows, and thereby reduces the average plate current. With a smaller plate current pulse, the power furnished to the tank is reduced, and the amplitude of the oscillations decrease. This causes the alternating component of the plate voltage to have a smaller peak value and the minimum plate voltage reached during the cycle is raised. In addition, the smaller plate current

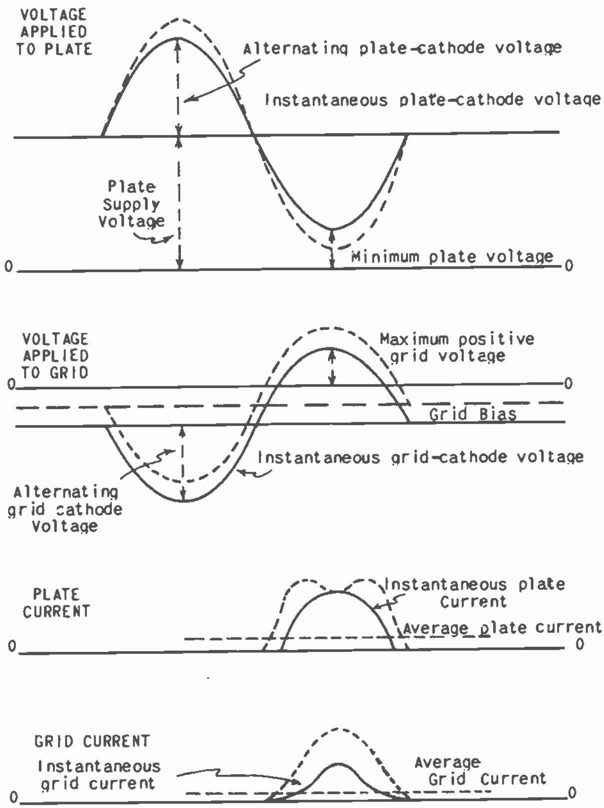


Fig.8 Illustrating the effect of changing the size of the grid-leak resistor. The solid lines are for the larger value of grid leak.

pulse reduces the power input. The overall effect, however, is to increase the efficiency, since the power input decreases at a more rapid rate than the power output. Thus, if the efficiency is not as high as desired, it may be increased by increasing the size of the grid leak. If this action is carried too far, the power output will also begin to decrease, and it is common practice to increase the size of the grid leak just enough until a noticeable drop in the power output occurs.

All the connections to the tank circuit are made by clips which may be fastened to any one of the turns of the tank coil, thus by moving the filament clip toward the plate end of the coil, the number of turns between the grid and filament is increased, while those between the plate and filament are reduced. This increases the grid-exciting voltage and reduces the alternating plate voltage. The increased grid-exciting voltage produces a larger grid current which increases the grid bias. Thus, increasing the grid-exciting voltage does not greatly affect the amount that the grid goes positive, but does increase the grid bias. In fact, the bias voltage is roughly proportional to the grid-exciting voltage, while the amount that the grid goes positive is largely determined by the size of the grid leak.

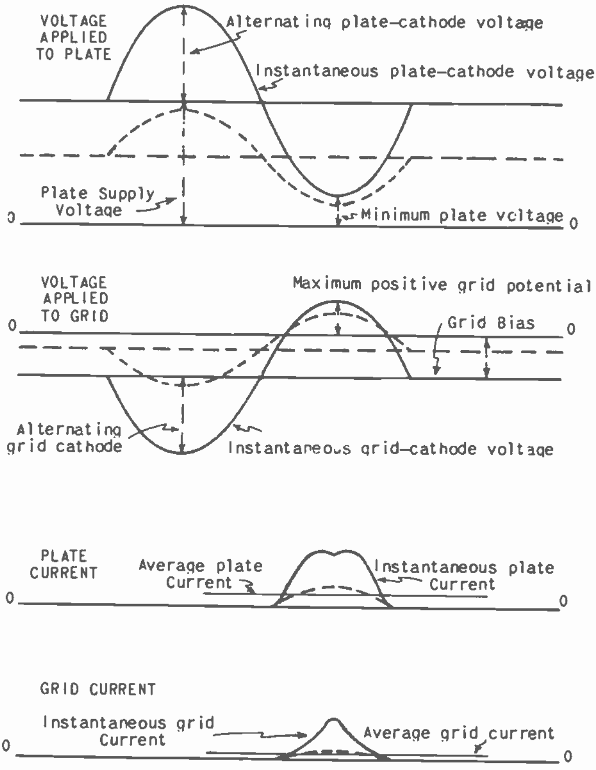


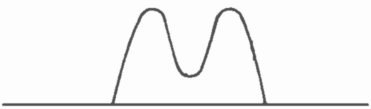
Fig.9 Showing the effect produced by doubling the value of the plate-supply voltage. The solid lines are for the larger value of plate-supply voltage.

Since the foregoing action reduces the alternating plate voltage, the minimum plate potential is raised. This would tend to decrease the efficiency, but, on the other hand, the bias is increased which, in turn, reduces the fraction of the cycle during which plate current flows, and would tend to increase the efficiency. Thus, there are two opposing forces, one tending to increase and the other to decrease the efficiency, and it will be found that there is one position of the filament tap where the efficiency is highest.

Increasing the plate supply voltage increases the time during which the plate current flows, and also increases the amplitude of the oscillating current. This naturally increases the value of the alternating plate voltage, and the minimum plate voltage reached during the cycle remains practically constant. The increase in plate supply voltage raises the DC power input and the RF power output. By increasing the size of the grid leak, the efficiency can again be made high, and the DC plate current brought back to normal. Thus, the higher the plate supply voltage, the greater will be the efficiency at which a given power output can be obtained. It is, of course, assumed that the maximum plate supply voltage as recommended by the manufacturer is not exceeded. The diagrams of Fig. 9 show the changes which occur when the plate supply voltage is doubled and the bias is increased enough to produce the same DC plate current.

Nothing as yet, has been said about how these various factors affect the grid current. Usually, this consideration is not so important, but there is one point which should be given attention. The maximum grid voltage should not exceed minimum plate voltage, for when this occurs, most of the electrons emitted by the filament are attracted to the grid and very few reach the plate. Thus, the highly positive grid robs the plate of electrons and the shape of the plate current pulse assumes that shown in Fig. 10. In addition, the DC grid current becomes very great with a consequent increase in the power lost in the grid circuit.

Fig.10 Illustrating the shape of the plate current pulse when the maximum positive grid voltage exceeds the minimum plate voltage.



This distorted plate current pulse contains an increased amount of harmonic components, and the smoothing action of the oscillating tank current must be great to minimize the radiation of harmonic frequencies.

8. THE LOAD IMPEDANCE. In the study of audio power amplifiers, it was learned that the load impedance should be equal to the plate resistance of the tube to secure the maximum power output, while for maximum undistorted power output, it was necessary to make the load impedance equal to twice the plate resistance. Since distortion is not a serious problem in oscillator circuits, it would seem desirable to use a load impedance of the same value as the plate resistance. In the case of an oscillator tube, however, where the plate current flows for such a short part of the cycle, the plate resistance varies between extremely wide limits. During the time that no plate current is flowing, the plate resis-

tance is infinitely high, and during the peak of the plate current pulse, it is relatively low. Therefore, the load impedance for maximum power output bears no simple relation to the plate resistance of the tube, and it is usually determined by experiment. The oscillator tank circuit is a parallel tuned circuit, and its total shunt impedance, as explained in Lesson 22 of Unit 1, is found by this formula:

$$\text{Shunt Impedance} = \frac{1}{R} \times \frac{L}{C} \qquad (6)$$

Where: R is the resistance of the tuned circuit in ohms,
L is the inductance of the tuned circuit in henries,
C is the capacitance of the tuned circuit in farads.

The tube does not work into the whole parallel tuned circuit, but only into that part of it connected between F and P; therefore, the impedance into which the tube is working is somewhat less than the total shunt impedance of the circuit. From the foregoing equation, it is evident that the shunt impedance is directly proportional to the ratio of L to C, and since there are many different values of L and C which will tune to the desired frequency, it is possible to vary the value of the load impedance by changing the ratio of the inductance to the capacitance.

With a high value of load impedance or a high ratio of L to C, the alternating plate voltage component is large and the minimum plate potential is low. The DC plate current is low, and while the power output is low, the efficiency is high. When a low load impedance is used, the opposite characteristics are true. More will be said about the load impedance and the ratio of L to C in Lesson 7 of this unit.

9. ADJUSTING AN OSCILLATOR. For the present, it is assumed that the oscillator is not coupled to the dummy antenna; that is, it is not loaded, or is not delivering its power output into any load circuit. Under these conditions, the actual power output of the oscillator is low, since there is nothing present to absorb the power. All the power in the tank is dissipated in the resistance of the tank circuit.

To begin the adjustment of the oscillator, the correct filament voltage is first applied, and the filament is allowed to reach its correct operating temperature before the plate voltage is applied. This procedure is always followed in transmitting tubes, since the high plate voltages used (from 500 to 20,000 volts) are liable to produce an arc-over from the plate to the filament if the plate voltage is applied before the filament has had time to warm up. By allowing the filament to reach its operating temperature, a protective space charge is formed around the filament, which eliminates the chance of arc-over.

At the start, the filament tap (F) is usually placed about half-way between the points G and P. A DC meter is connected in series with the grid leak to read the DC grid current, and another DC meter is placed in the plate circuit to read the DC plate current. Care must be taken that both meters are properly by-passed, as the RF current will seriously damage them. The plate voltage is now applied and the tank circuit is tuned to the desired frequency. In low and medium powered oscillators, the tank condenser is made variable, and the resonant frequency of the tank is changed by varying the capacity of this condenser. In very high powered oscillators, the tank condenser is usually fixed and the resonant frequency is changed by varying the value of the inductance. Just how this is done will be explained in Lesson 7 of this unit under the discussion of Class C amplifiers.

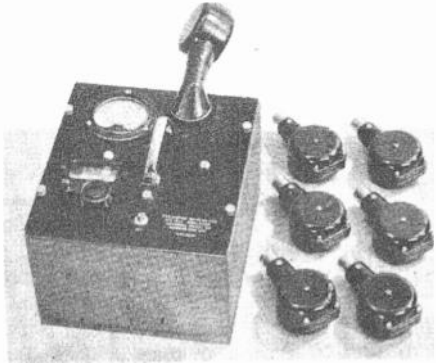


Fig.11 A precision type of wave meter which covers a frequency range from 16 kc. to 50 mc.

To determine when the oscillator is tuned to the correct frequency, some form of frequency-measuring device must be employed. Usually, this is an accurately calibrated wavemeter. The wavemeter may consist of a small coil tuned by a variable condenser. In series with these two components is a very sensitive thermocouple galvanometer. The dial of the variable condenser is divided into equal units, usually from 0 to 100, and a chart is furnished with the instrument which shows the resonant frequency of the wavemeter for any setting of the dial. A photograph of a wavemeter is shown in Fig. 11. It is a precision type meter and is supplied with seven plug-in coils which cover a frequency range between 16 kc. and 50 mc. Instead of using a thermocouple meter to indicate resonance, this instrument employs a rectifier type vacuum-tube voltmeter. Thus, the danger of overloads burning out the indicator is avoided. To use the meter, the coil, which is mounted externally to the case, is *very loosely coupled* to the tank of the oscillator; thus, RF energy is transferred to the wavemeter and, if the resonant frequency of the oscillator is equal to that of the meter, the thermogalvanometer will indicate that a current is flowing in the wavemeter circuit. The variable condenser of the tank is changed until a maximum reading is obtained on the thermogalvanometer, assuming, of course, that the wavemeter has been set for the desired frequency. If, at any period of the adjustment, the needle of the meter is deflected full scale, it is desirable to loosen the coupling between the oscillator and pick-up coil, and thus avoid any possibility of damaging the meter.

With the oscillator tuned to the correct frequency, the filament tap is adjusted until the plate current of the tube is at a minimum. Under this condition, the efficiency is greatest. In case the plate current cannot be reduced to the normal value, the size of the grid leak should be increased. It must, of course, be remembered that too large a grid leak will cause the oscillator to block, and the output will be modulated at a frequency depending on the size of the grid leak. Such a condition is objectionable in high-powered oscillators, and grid leaks in excess of 50,000 ohms should not be used. In making any adjustment on the oscillator, always remember to shut off the plate voltage before touching any part of the circuit, as the voltage applied to the plate is high and a severe shock may result. Even if you were to touch a part of the tank circuit not connected directly to the plate supply, the RF energy in the tank would be liable to produce a painful burn.

10. LOADING THE OSCILLATOR. If the preceding adjustments have been performed correctly, the efficiency of the oscillator should be high, while the plate current, the power input, and power output should be rather low. The first step in loading the oscillator is to couple the dummy antenna loosely to the tank circuit of the oscillator. The antenna is then tuned to the frequency of the oscillator by changing the capacity of its variable condenser. Resonance of the antenna circuit is indicated by a maximum antenna current as read on the thermocouple ammeter. As the antenna circuit is brought into resonance, it is noticed that the plate current of the tube is increasing. This occurs for the following reasons:

The oscillating tank current develops a magnetic field about the tank which cuts through the turns of the antenna coil and induces RF voltages therein. When the antenna is tuned to resonance, these RF voltages force a fairly high current through the antenna circuit, which creates a magnetic field about the antenna coil. This varying magnetic field around the antenna coil cuts through the turns of the tank coil, and induces a voltage in this coil which is in such a direction as to oppose the voltage producing the oscillating tank current. Thus, the effect is the same as though a resistance had been added in series with the tank coil, and it is said that the effect of coupling the antenna to the tank is to couple a resistance into the tank by reflected impedance. The complete discussion of reflected impedance will be given in Lesson 4 of this unit.

This resistance which has been effectively added to the oscillator tank produces several results. The first of these is to reduce the value of the oscillating tank current. With the amplitude of the oscillations reduced, the voltages developed across the tank are lowered; thus the grid excitation voltage is less, and as a result, *the grid current goes down*. Reducing the grid current causes a smaller bias to be produced across the grid leak, and the *DC plate current increases.* Furthermore, the alternating component of the plate voltage is lowered, which raises the minimum plate potential reached during the cycle. Raising the minimum plate voltage always decreases the efficiency, since the peak value attained by the plate current will be greater with a higher minimum plate voltage. The increase in the minimum plate voltage also causes the grid current to be reduced, since, with a higher minimum plate voltage, the plate will be able to attract more of the electrons emitted by the filament and fewer will be diverted to the grid. Thus, the power input as well as the power output of the oscillator *increases* and the efficiency is lowered.

Under this condition, the oscillator is said to be loaded, or to be working into a load. The foregoing actions may be viewed in another manner. With the oscillator unloaded (the antenna not coupled to the tank), the actual shunt impedance of the parallel tuned tank circuit is rather high; much too high to secure appreciable power from the tube. Do not be confused by the fact that although the tube is working into a large load impedance, it is said to be unloaded. The same expression is used with alternators or DC generators. When a high resistance is connected across an alternator, the alternator is able to force only a small current through the resistor, and the power transferred from the alternator to the resistor is low. Therefore, although the actual resistance of the load resistor is high, the alternator is operating at practically no-load, since the power drawn from it is low. By reducing the value of the load resistor, more current can flow through it, and more power will be transferred from the alternator to the resistor. Under this condition, the alternator is said to be loaded, as it is required to produce a greater power output. It should, therefore, be remembered that decreasing the value of the load impedance of any power source always increases the load on that power source.

When the antenna circuit couples a resistance into the tank, the shunt impedance of the tank is decreased, as may be seen from equation (5). This equation has the resistance of the parallel tuned circuit in the denominator, and it is easy to see that an increase in the resistance of the tank circuit will decrease its shunt impedance. When the value of the total shunt impedance of the tank circuit is reduced, the impedance connected between the points F and P is correspondingly lowered and the tube works into a lower load impedance. There is just one value of load impedance which will produce the greatest power output, and, since the value of the load is much greater than this before the antenna is coupled to the tank, the act of coupling the antenna will always result in a larger RF power being developed in the tank circuit, which is then transferred to the antenna.

If the antenna circuit is *coupled too close* to the tank, it may reduce the load impedance to a value so low that the maximum power output will not be obtained. In this case, the antenna current will decrease. It is evident that there is a point of "critical coupling" which will produce the greatest antenna current and the maximum power into the antenna. With less than critical coupling, the power in the antenna circuit will be less than maximum, although the efficiency will be high. With more than critical coupling, the antenna power will be less than maximum, and the efficiency will be low, as is evidenced by the large plate current that will flow.

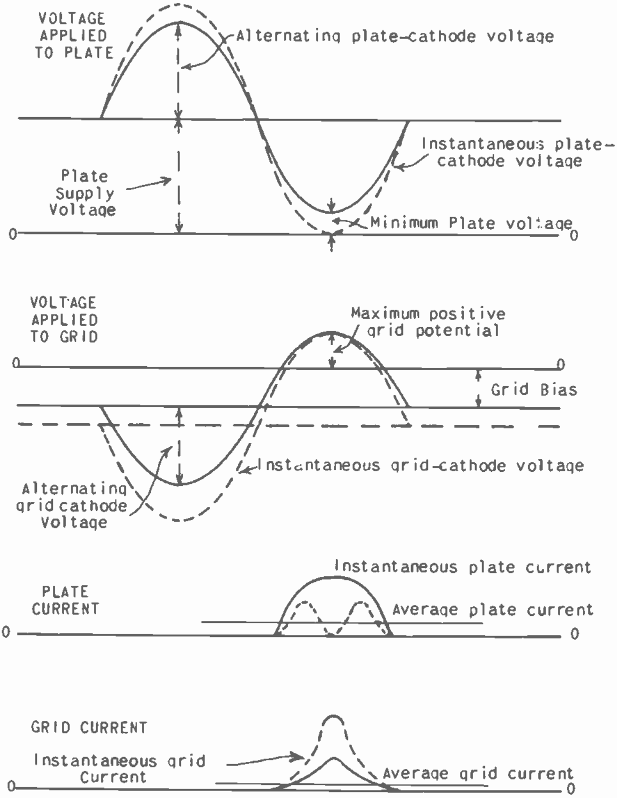


Fig. 12 Showing the effect of loading an oscillator. The solid lines indicate the loaded condition.

The diagrams of Fig. 12 illustrate the effect of loading the oscillator. The dotted curves are for the unloaded condition, while the solid line curves represent the loaded condition. Notice that in the unloaded state the minimum plate voltage is so low that a decided dip is produced in the plate current pulse. This is caused by the fact that the maximum grid potential is nearly equal to the minimum plate potential and the grid is robbing the plate of electrons. Thus the plate current dips and the peak grid current is high. Loading the oscillator raises the minimum plate voltage and removes the dip in the plate current pulse, while at the same time, it reduces the peak grid current and lowers the bias voltage.

To continue with the loading of the oscillator, the antenna circuit has been loosely coupled to the tank, and has been tuned to resonance as indicated by maximum antenna current. The coupling between the antenna and tank is now increased until the current in the antenna is maximum, thus showing that the point of critical coupling has been reached.

During this increase in coupling, the plate current has been steadily rising, and if it exceeds the normal specified value, the coupling should be loosened and the grid bias increased until the antenna power begins to fall off, as evidenced by a decided drop in the antenna current. The plate current can now be brought up to normal by again increasing the coupling. If the desired power output is not obtained at a reasonable efficiency, the efficiency may usually be increased by using a large grid leak, without sacrificing much power output. If the desired power output cannot be obtained by coupling the antenna to the tank, it is necessary to reduce the efficiency by lowering the size of the grid leak, and re-adjusting the position of the filament tap, always remembering not to exceed the maximum plate dissipation of the tube. It is always best to repeat this series of adjustments at least once to make sure that the tube is operating under the very best conditions obtainable. REMEMBER, THE PLATE VOLTAGE MUST BE TURNED OFF WHEN MAKING ANY ADJUSTMENTS.

11. THE SHUNT-FED HARTLEY OSCILLATOR. The shunt-fed Hartley oscillator studied in Lesson 23, Unit 1, operates in exactly the same manner as the series-fed, which we have been discussing. The same procedure for adjusting and loading the shunt-fed oscillator is used, and the only difference in the circuit design is the RF

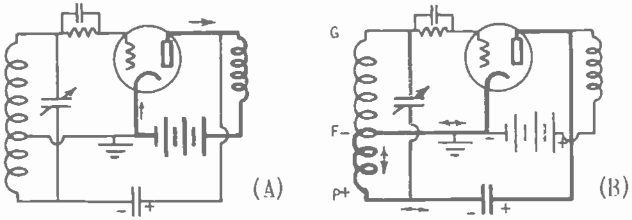


Fig.13 (A) Path followed by the DC component of the plate current in an oscillator.
(B) Path followed by the RF component of the plate current in an oscillator.

choke and the blocking condenser which must be used with the shunt-fed type. Fig. 13 at A shows the path followed by the DC component of the plate current, while B in the same figure illustrates the path taken by the AC component. (In each case, the path taken is shown by the heavy lines). Feedback to the tank circuit is secured by the AC component of the plate current flowing through the blocking condenser, while the DC component flows through the RF choke. Thus, no part of the tank circuit is at a high positive potential with respect to ground. Diagrams illustrating the operating conditions of a correctly adjusted and loaded shunt-fed Hartley oscillator are given in Fig. 14. The blocking condenser has sufficient capacity so that its capacitive reactance to the oscillator frequency is low. Thus, the voltage across it does not change appreciably as the current through it varies. The total DC voltage output of the power supply is applied across this condenser and it must have sufficient dielectric

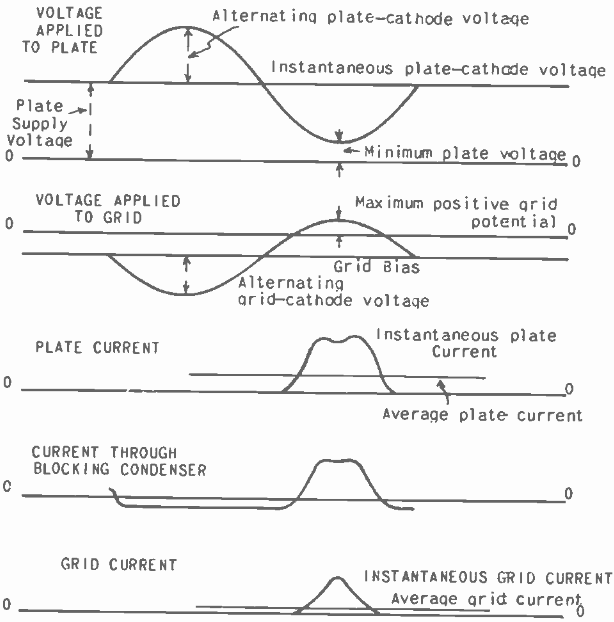


Fig. 14 Illustrating the operating conditions of a correctly adjusted and loaded shunt-fed Hartley oscillator.

strength to withstand this voltage. Since the voltage across the blocking condenser does not change appreciably, it remains charged to the DC plate voltage, and since that part of the tank circuit between F and P (see B in Fig. 13) is in series with the blocking condenser, the total voltage applied to the plate at any instant is the sum of the voltage produced across the two points and the voltage present across the blocking condenser. The RF choke must have enough impedance at the oscillator frequency to effectively prevent the RF component of the plate current from flowing through the power supply.

12. THE COLPITTS OSCILLATOR. Another very common type of oscillator is the Colpitts. A diagram of a Colpitts oscillator is shown in Fig. 15. Notice that the tank inductance is not tapped, but that the filament is connected to the mid-point of

two condensers connected in series. Thus, the tank circuit consists of L1, C1, and C2. C1 and C2 are variable condensers, the rotors of which are connected together and then joined to the mid-point of the filament. The RF component of the plate current flows through C3, the plate blocking condenser, and then through the capacitive reactance of C1 back to the filament. In flowing through the capacitive reactance of C1, it produces a voltage drop which serves to feed energy into the tank circuit. The oscillating tank current flows through C2 and the voltage drop it creates is the grid-exciting voltage.

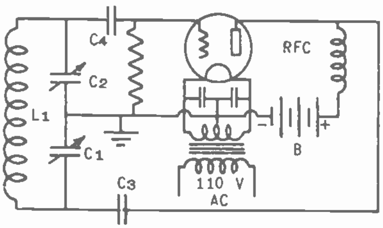


Fig.15 A shunt-fed Colpitts oscillator.

The amount of grid-exciting voltage may be changed by varying the capacity of C2. Increasing the capacity of this condenser will reduce its capacitive reactance, and the voltage drop set up across it will be lower. Thus, the grid-exciting voltage is *reduced*. Decreasing the capacity of this condenser will *increase* the grid-exciting voltage. In a like manner, the alternating component of the plate voltage produced by the oscillating current flowing through the capacitive reactance of C1 may ba changed by varying the capacity of C1. Increasing this capacity will lower the AC plate voltage on the tube and raise the minimum plate potential, while decreasing it will increase the AC plate voltage and lower the minimum plate voltage.

It must be realized that when the capacity of one of these condensers is increased, the other condenser must be decreased a like amount in order to maintain oscillations at the correct frequency. Thus, a change in either condenser necessitates a change in the other. This is a disadvantage not possessed by the Hartley oscillator.

Since the rotor plates of the condensers are at an RF ground potential, hand capacity effects will be negligible. In the Hartley oscillator, neither plate of the tuning condenser is at an RF ground and hand capacity effects may be troublesome.

The grid leak is connected directly to the center tap of the filament circuit instead of across the grid condenser. This is called a shunt-fed grid circuit, since the DC and RF components of the grid current follow separate paths in returning to the filament. The DC component flows through the grid leak to the filament and the RF component flows through C4 and C2 in returning to the filament. This arrangement is necessary since it would be impossible for the DC component to flow through the condenser C2.

In the adjustment of the Colpitts oscillator, the grid excitation and AC plate voltage are varied until the minimum plate current is obtained, remembering, of course, that it is necessary to retune the tank circuit whenever an adjustment is made. The dummy antenna is then coupled to the tank circuit until the desired power output is secured. Control of the feedback voltage is more smoothly obtained in the Colpitts than in the Hartley oscillator.

13. THE TUNED GRID-TUNED PLATE OSCILLATOR. The tuned grid-tuned plate oscillator is usually abbreviated to TGTP, and is sometimes called the Armstrong oscillator after its inventor. An Armstrong oscillator with series-fed plate and grid circuits is illustrated in Fig. 16. At first sight, this might be thought to be an ordinary RF amplifier stage; the source of feedback is not at once evident. In the study of RF amplifiers in Lesson 24, Unit 1, it was learned that an RF amplifier stage using a three-element tube must be neutralized to prevent feedback through the plate-grid capacity of the tube, as such feedback was liable to produce sustained oscillations, and thus interfere with the normal amplifying ability of the RF stage. Advantage is taken of this fact in the design of the Armstrong oscillator.

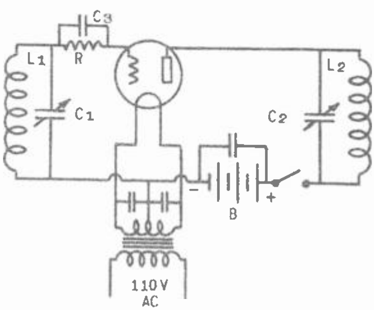


Fig.16 A TGTP, or Armstrong oscillator.

By tuning the plate as well as the grid circuit, the RF voltages developed across the plate tank are much greater than they would be across the primary of an ordinary RF transformer. Since it is these RF voltages which cause an RF current to flow through the plate-grid capacity of the tube, which, in turn, sets up RF voltages across the grid tank, and thereby provides the grid excitation, it is very easy to feed back enough energy from the plate to the grid circuit to produce sustained oscillations.

The operation of this circuit is as follows: Closing the plate voltage switch causes a pulse of plate current to flow from the plate through the plate tank circuit. This small amount of energy delivered to the plate tank creates feeble oscillations in this circuit. Thus, RF voltages are developed across the plate tank which force very small RF currents to flow through the interelectrode capacity of the tube to the grid circuit and through the grid tank. These minute RF currents deliver energy to the grid tank and produce feeble oscillations in it. The oscillating current of the grid tank develops RF voltages across this circuit, which then vary the grid voltage at an RF rate and cause the plate current to flow in the form of pulses which feed additional energy to the plate tank, and the series of events is repeated. Unless both grid and plate tanks are tuned to approximately the same frequency, the RF voltages produced across them will not be of the correct phase to maintain sustained oscillations.

It can be proved mathematically that for the alternating grid and plate voltages to be 180° out of phase (a condition necessary to produce sustained oscillations), the plate tank circuit must present the effect of inductive reactance. This is accomplished by tuning the plate tank to a *frequency slightly higher* than the one desired. It is necessary that the plate tank be slightly inductive so that it will compensate for the phase shift caused by the capacitive reactance of the grid-plate capacity of the tube, and thus make the feedback voltage of the correct phase to maintain the oscillations in the grid tank.

The Armstrong oscillator does not provide any means of adjusting the feedback, and it is evident that the higher the oscillation frequency, the greater will be the magnitude of the feedback voltage, since the capacitive reactance of the interelectrode capacity becomes smaller as the frequency is raised. At low frequencies, it is possible for the interelectrode capacity to have such a high capacitive reactance that the energy fed back from the plate to the grid circuit is insufficient to maintain sustained oscillations. In this case, the amount of feedback may be increased by connecting a condenser of small capacity between the grid and plate electrodes.

Occasionally the grid tuning condenser is omitted. Under this condition, the distributed capacity of the grid coil tunes it rather broadly to a wide band of frequencies, and the frequency of oscillation is determined by the setting of the plate tank condenser. This circuit is often called a TNT oscillator.

The Armstrong oscillator will operate at frequencies much higher than any of the other oscillators described, and, for that reason it has found much favor in ultra-high frequency transmitters.

14. ELECTRON-COUPLED OSCILLATORS. One of the outstanding disadvantages of the self-excited oscillator circuits discussed so far is the necessity of coupling the load directly to the portion of the oscillator which determines the frequency. In the Hartley and Colpitts, the load is coupled directly to the single tank circuit and in the TGTP oscillator the load is coupled to the plate tank circuit. By loading the TGTP plate tank, the frequency will be affected to an undesirable extent. If the load is varied on any one of these three oscillator circuits (by modulation, swinging antenna, or by keying a subsequent stage) the frequency will shift. This is called "frequency shift" or "dynamic instability". A change in plate voltage will also cause considerable change in frequency.

The effect on the generated frequency of changes in load and plate voltage is minimized in the electron-coupled oscillator circuit. This circuit has excellent dynamic frequency stability for a self-excited oscillator; however, it is not considered as stable as the crystal controlled oscillators to be described in the next lesson. The electron-coupled circuit is shown in Fig. 17. The screen grid acts as the plate in a modified Hartley oscillator in which the plate circuit is series fed. The screen grid, control grid, and cathode form the three elements of the triode oscillator. Modification from the true Hartley circuit is due to the position of the B minus or ground connection. In the ECO (electron-coupled oscillator) circuit the bottom of the tank circuit is grounded to DC and RF and the plate of the oscillator section (actually the screen) is grounded to RF; whereas, in the true Hartley circuit, the filament center tap or cathode is grounded to both DC and RF. The grid circuit in Fig. 17 is conventional shunt fed and the screen grid potential is reduced below the actual tetrode plate potential by the screen dropping resistor R2. The RF choke L4 serves to assist in preventing the leakage of RF to ground through the resistor R2 and the power supply.

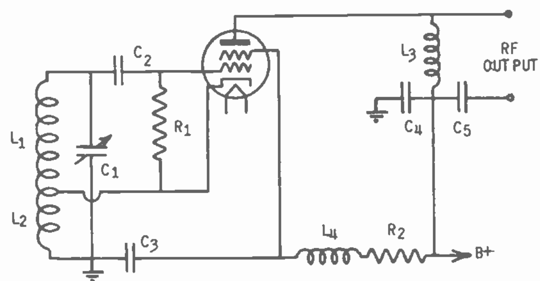


Fig.17 Electron-coupled self-excited oscillator.

With the screen grid at RF ground potential, the plate of the tube is shielded from the elements which generate the oscillations. Since the screen grid is not a solid plate, it will not collect all of the electrons emitted from the cathode. The higher potential on the plate will pull the majority of emitted electrons on through the screen to establish an RF voltage at the plate of the tube. The only coupling between the plate and the oscillating elements is through the electron stream; hence, the name for the circuit, "electron-coupled".

The electron stream coupling is to a large extent unilateral; that is, in one direction only. The instantaneous plate potential depends on the variations of the electron stream; however, the electron stream is affected little by changes of plate potential. This latter statement conforms with previous information concerning the ability of the plate voltage in a tetrode to change the plate current.

To take full advantage of the plate's ability to isolate the oscillating section from load changes in the plate circuit, the screen grid must be operated at RF ground potential and extreme care must be taken in keeping the circuit components well shielded from each other.

In general, the ECO circuit possesses three distinct advantages which make it applicable for various purposes; (1) Frequency stability with variations in load. (2) By properly adjusting the voltages on the screen (oscillator plate) and plate, the frequency may be made independent of small changes in the supply voltage. The constant proportion of screen and plate voltages is maintained by securing both from a voltage divider; thus both voltages will change if the supply voltage varies. (3) The output of the ECO is fairly rich in harmonics.

In Fig. 17 notice that the cathode of the tube is operated above RF ground potential. If a filament type tube is used, the filament voltage should be applied through RF chokes to prevent the filament from being placed at RF ground through the distributed capacity of the filament transformer. With most cathode type tubes, the RF choke is not necessary, since no damage will be done with an RF voltage between the heater and cathode.

Regarding the advantages for the ECO just outlined, it should be understood that the frequency stability is not as good as can be obtained in the crystal controlled oscillator circuits to be described in the next lesson; however, it is superior to the Hartley, Colpitts, and TGTP self-excited oscillators. In some cases, the high harmonic output of the ECO may be undesirable, such as when used to feed an antenna directly or when used in some types of test oscillators for servicing. But, in heterodyne frequency meters (used to measure an unknown frequency), the harmonic output is desirable, and if the ECO is followed by an amplifier stage which employs a tuned tank circuit as the plate load, the harmonic output is of little consequence.

Pentodes can be used as electron-coupled oscillators if: (1) the suppressor has a separate pin connection, and (2) if the internal shield is tied to the suppressor. Certain pentodes, such as the 47 and 2A5 have the suppressor connected internally to the mid-point of the filament or to the cathode. Since the cathode is not at RF ground in the ECO, undesirable electrostatic coupling will exist between the oscillator portion of the tube and the plate, with the suppressor tied to the cathode. To obtain full benefits, the suppressor should be tied to the screen and placed at RF ground potential. Several pentodes employ an internal shield. In some types, including the 6C6, 6D6, and 78, this shield is tied to the cathode internally. Again, undesirable electrostatic coupling will exist between the shield (tied to cathode) and plate, so the tube is unsuitable for use in the ECO circuit. The type 77 tube has the internal shield tied to the suppressor; therefore, it is satisfactory because the suppressor, shield, and screen can all be tied together and placed at RF ground potential through a condenser to form an effective shield between the plate and the oscillator portion of the tube.

The type 802 is a 10 watt pentode for use in transmitter circuits. It has separate base pins for both the suppressor and the internal shield, hence, it can easily be connected for ECO service.

15. DYNATRON OSCILLATORS. The dynatron oscillator is not in popular use as the frequency control unit for a transmitter; however, it is a type of self-excited oscillator adaptable for use in test oscillators and heterodyne frequency meters. This circuit should therefore be associated with the other self-excited oscillators described in this lesson. A circuit diagram of a dynatron oscillator appears in Fig. 18. Tubes satisfactory for use in this circuit include the low power types, 24, 32, and 36. The applications for the dynatron circuit do not require appreciable power output, so no advantage is gained by attempting to use larger screen grid tubes. The oscillator is rather inefficient and the output signal is weak, but the advantages are good frequency stability and high harmonic content in the output signal.

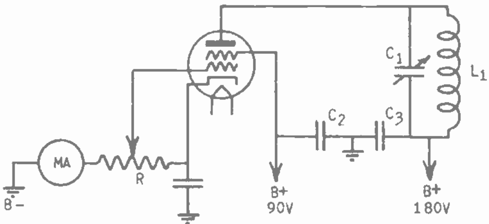


Fig.18 Dynatron self-excited oscillator.

Upon inspecting the plate voltage - plate current characteristic curve for any screen grid tube, it will be found that throughout the region from 0 to 90 volts plate voltage (assuming that the screen voltage is 90), the characteristic curve undergoes abnormal changes. A typical Ep-Ip curve is shown in Fig. 19. The dynatron oscillator operates over the range from B to C on the characteristic curve. Throughout this range, it will be noted that increases in plate voltage cause decreases in plate current. The operating plate voltage for the dynatron oscillator is set at the center of this portion of the characteristic (point O in Fig. 19). The screen voltage is at a higher normal value than the plate voltage, and the peculiar bends in the Ep-Ip curve are caused by the secondary electrons from the plate being attracted to the screen. In Fig. 18, the control grid may be set at a negative potential with respect to the cathode by adjustment of the potentiometer R. Variation of the control grid bias does not change the frequency of the output signal; however, it will vary the amplitude considerably -- decreasing the amplitude with more negative bias, and vice versa. In many dynatron oscillators, the potentiometer R is omitted and the control grid is connected directly to the cathode for the greatest output signal.

When plate voltage is applied to the dynatron oscillator, the instantaneous current surge which passes through the oscillating circuit L1C1 causes a field to be built up around the inductance L1, and the capacity C to become charged. Assuming that the small voltage generated across the oscillating circuit is in such a direction as to increase the plate voltage, from the characteristic curve in Fig. 19 we see that the plate current will decrease. The decrease in plate current will generate a counter voltage across L1 in the opposite direction. As the counter voltage is set up in this direction, the plate voltage will decrease, which causes the plate current to increase. The action

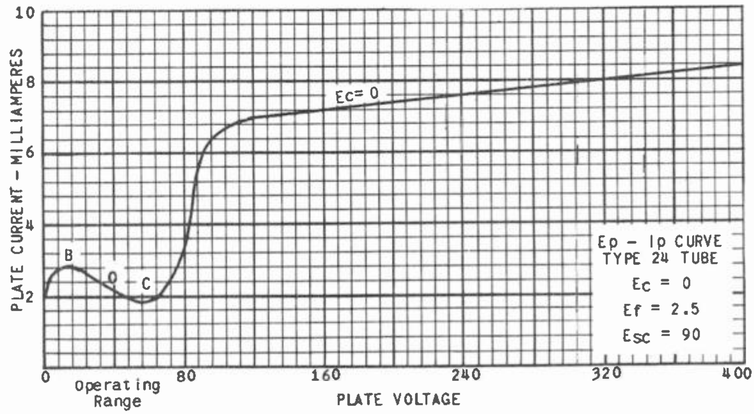


Fig.19 Typical Ep-Ip characteristic curve for a tetrode tube. The negative resistance portion is from B to C.

repeats and builds up to maximum. The normal operating plate voltage is at the center of the down slope on the Ep-Ip curve. Due to the alternating RF voltages set up across L1C1, the instantaneous plate voltage will vary at an RF rate as in other oscillator circuits. The RF plate voltage swing will be confined to the portion of the characteristic curve between B and C in Fig. 19 because it is only over this portion of the curve that changes in plate potential result in an opposite change of plate current.

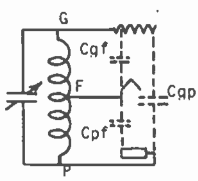
Quite often the action of a dynatron oscillator circuit is explained in terms of "negative resistance". Resistance always means a voltage divided by a current. By "positive resistance" is meant that an increase in voltage will result when the current through a resistance is increased; whereas, the expression "negative resistance" means an increase in voltage results from a decrease in current. From the characteristic curve of the screen grid tube shown in Fig. 19, it will be noticed that throughout the region from B to C, a "negative resistance characteristic" exists between the plate voltage and the plate current; that is, an increase in plate voltage results in a decrease in plate current and vice versa. This negative resistance of the tube's plate circuit is shunted across the oscillating circuit L1C1. The negative resistance of the tube is in opposition to the positive resistance of the oscillating circuit, and when the negative resistance is equal to the positive resistance, they completely cancel each other, resulting in a net resistance in the oscillating circuit equal to zero.

It has previously been explained that when the net (effective) resistance of an oscillating circuit is zero, the oscillating current does not die out, but remains constant in amplitude. Hence, the oscillating current in the tuned circuit L1C1 will be maintained as long as its positive resistance is cancelled by the tube's negative resistance.

The advantages of a dynatron oscillator include the excellent frequency stability, the simplicity of construction, and the ease with which it may be made to function over a wide frequency range. The frequency is affected little by variations in operating voltages and tube characteristics, but the stability is not as good as in a crystal-controlled oscillator. The dynatron makes an ideal oscillator for use in a heterodyne frequency meter (used for making frequency measurements). The output signal has a high harmonic content. The main disadvantage is the fact that the output signal is weak.

16. FREQUENCY STABILITY OF OSCILLATORS. Before discussing the frequency stability of oscillators, it must be realized that the one thing determining the oscillation frequency may be stated as follows: The frequency at which the oscillations occur is the frequency at which the feedback is of the proper phase to make the alternating grid and plate voltages 180° out of phase.

Fig.20 Illustrating how the interelectrode capacities affect the tuning of an oscillator



There are many things which affect the oscillation frequency other than the inductance and capacity values of the tuned circuit. For example, in the Hartley oscillator, the grid-filament capacity is, in effect, connected directly across the grid coil of the tank, while the plate-filament capacity is across the plate coil. In addition, the grid-plate capacity is in parallel with the tank condenser as may be seen in Fig. 20. Thus, all of these interelectrode capacities will be effective in determining the total capacity of the tank circuit.

It has also been determined that the values of the plate resistance, and the grid-to-filament resistance of the tube will affect the oscillation frequency. Thus, changes in the plate supply voltage and changes in the electrode spacing, due to heat expansion, will alter slightly the frequency produced.

None of the oscillators that have been previously discussed will maintain a constant frequency over a very long period of time; that is, any of them will drift slightly in one direction or the other from the desired frequency. Even under the very best conditions, the frequency deviation will be at least 1%, and since the allowable frequency deviation for broadcast transmitters as specified by the Federal Communications Commission is much less than this value, these types of oscillators are not used in broadcast transmitters. Instead, the crystal-controlled oscillator is employed; a complete discussion of which will be given in the following lesson.

The factors which contribute to a shift in the frequencies of oscillators may be listed as follows:

1. Tube characteristics
2. Temperature
3. Vibration
4. External coupling

Among the various things which can produce a change in the tube characteristics are:

1. Changes in plate potential
2. Changes in mean grid potential
3. Changes in filament potential
4. Changes in spacing of the tube's elements

By careful design, it is possible to choose the various circuit elements so that changes in the tube's characteristics will have a minimum effect on the oscillation frequency. The effect of temperature variations is to change the spacing of the tube's elements, thus changing the total capacity associated with the tuned circuit. In extreme cases, temperature control may be employed.

Vibration will cause a variation in the frequency due to changes in the spacing of condenser plates, in the separation between coil turns, or in the spacing of the elements of the tube. Frequency instability due to this cause may be eliminated by proper placement of the oscillator, using some form of shock-absorbing device if necessary to reduce the vibration.

The use of large capacities in the plate tank, thus making the total capacitive effect of the interelectrode capacities small in comparison, will aid in minimizing frequency instability. Finally, the frequency changes produced by coupling the load circuit to the oscillator may be reduced by not attempting to derive the utmost possible power from the oscillator. For maximum frequency stability, the load circuit should be loosely coupled to the oscillator, and, if the power obtained from the oscillator is insufficient to produce the desired results, it should then be amplified by radio frequency power amplifiers.

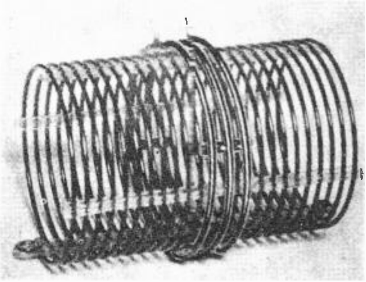
The simplest form of transmitter consists of an oscillator coupled directly to the antenna. Then, by use of a keying circuit, the output of the oscillator may be broken up into dots and dashes, and intelligence may be transmitted by means of the International Morse code. Such an arrangement constitutes a self-excited transmitter, and all the oscillators discussed in this lesson are known as self-excited oscillators.

The frequency stability of the self-excited transmitter is very poor, and this system is no longer used. When the oscillator is of relatively low power, and its output is amplified by one or more power amplifier stages, the transmitter is said to be of the master oscillator-power amplifier type. This rather long expression is abbreviated to M.O.P.A.

Modern broadcast stations, police radio stations, aeronautical stations, and most amateur stations use neither of the foregoing methods. Instead, crystal-controlled transmitters are employed almost exclusively. They consist of a low-powered crystal oscillator, having an output of from 1 to 5 watts, and a series of power amplifier stages which serve the double purpose of increasing the power, and isolating the oscillator from the modulated stage, thus preventing the modulation process from reacting upon the oscillator stage and thereby causing frequency instability.

17. CIRCUIT COMPONENTS. To obtain the best results from any oscillator circuit, certain design conditions should be followed. As previously stated, the tank circuit should have as little resistance as possible, and this minimum resistance is achieved by using copper tubing or flat copper strip to construct the tank coil. (See Fig. 21). By employing a conductor having a large surface area, the apparent increase in resistance of the tank coil

Fig.21 Showing the construction of a tank coil.



to RF currents due to skin effect is minimized. The tank coil must be rigidly mounted, as the open construction offers an opportunity for vibration with its resulting frequency instability. It is usual practice to mount the tank coil on heavy stand-off insulators, and the materials of the insulators used must have good dielectric properties, for they are in the electrostatic field of the tank. That is, a capacity exists between the tank coil and the chassis, and the insulators are a part of the dielectric of this capacity. Thus it might be possible to use a material having good insulating properties and yet be of such a character that its dielectric hysteresis was high. In this case, the losses incurred in the insulator would have to be supplied from the energy contained in the tank circuit, and the effect would be the same as increasing the actual resistance of the tank.

It is also desirable to keep large masses of metal away from the tank since the eddy currents set up in them by the voltages induced from the changing magnetic field, produce power losses which must be supplied by the tank circuit.

All leads carrying RF should be exceptionally short; an unnecessary extra inch in the length of an RF lead will increase the power losses far more than is ordinarily realized. In addition, the leads from the tuning condenser to the tank coil should be amply large, since the tank current is usually several amperes instead of milliamperes.

The tuning condenser must be able to withstand the peak of the RF plate voltage with a measure of safety added. All transmitting condensers have rounded edges and corners, since, for a given separation, the arc-over voltage is much less between sharp points than when the points are rounded.

Since the plate current flows for only a small fractional part of the cycle, the power supply must have exceptionally good voltage regulation. By using a relatively large bleeder current, and mercury vapor tubes, the regulation may be made sufficiently good to satisfy this requirement.

In the shunt-fed circuits, the plate-blocking condenser must have sufficient dielectric strength to withstand the DC plate voltage, and its capacity should be high enough so that it offers very little impedance to the RF plate current component. Any RF choke that is used must have a low DC resistance, while its impedance to RF should be very high.

The grid leak resistor must have a wattage dissipation rating sufficiently high to enable it to safely dissipate the required amount of power and still not become dangerously hot. The power dissipated in this resistor is equal to the DC grid current squared times the value of the resistor in ohms. The capacity of the grid condenser should be somewhat larger than the effective grid-filament capacity of the tube, and should also be large enough to make its reactance to the oscillation frequency considerably less than the resistance of the grid leak. Except for these requirements, its size is not critical.

Nearly all transmitting tubes are of the filament type, and it should be noticed that in each of the diagrams previously given, a condenser is connected from each side of the filament circuit to the center tap of the filament secondary. These condensers are for the purpose of by-passing the RF currents around the secondary winding of the filament transformer. Since both the grid and plate currents contain RF components, and since these currents are returned to the center tap of the filament circuit, they would have to flow through the impedance of the secondary if these condensers were not provided. The impedance of the secondary to RF is rather high, and, for proper operation, a low impedance return path for the RF currents should be furnished. The capacity of these condensers is usually about .002 mfd.

The size of the condenser used to by-pass the power supply in series-fed plate circuits, and the plate-blocking condenser are also ordinarily .002 mfd.

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UNIT NO. 3

CRYSTAL CONTROLLED OSCILLATORS

LESSON NO. 2

FAST TRANSPORTATION

.... TO SUCCESS AND SECURITY

From our office and classroom windows, high up in the Power and Light Building, we can look down upon the Kansas City Municipal airport. It is located just across the Missouri River. Sometimes it seems as though we could throw a stone out on to the field.

Spanning the river is a bridge.

Across this bridge, automobiles and busses travel on their way to and from the airport. Occasionally a side-wheel steamboat moves slowly up the river, making it necessary to open the bridge.

A giant aluminum-clad air liner appears in the distance.

Sunlight flashes from sleek wings as the transport roars over the city and circles for a landing. Its passengers have come from hundreds of miles away in only a few hours. They are aboard that ship because they are in a hurry. The transport companies have sold these passengers 'Fast Transportation'.

Passengers alight from the airplane. They enter taxicabs to be whisked across the river to the doors of buildings they have travelled so far to enter. And they have reached their destination hours ahead of other people travelling to Kansas City on older forms of transportation.

People that travel by air have an urge to get to their destination as quickly as possible. That is why they select a modern mode of travel.

You have an urge to achieve success as quickly as possible. That is why you selected a modern industry, radio.....an industry that is moving ahead with remarkable speed, carrying many young men with it to financial success.

MAKE RADIO CARRY YOU TO SUCCESS TOO. STICK---STUDY AND WIN.

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Lesson Two

CRYSTAL CONTROLLED OSCILLATORS

"The crystal-controlled oscillator is the heart of the modern broadcast transmitter. Without it, it would not be possible for radio transmitters to maintain their assigned frequency, and reception of such stations would be difficult due to interference.

"I shall, in this lesson, describe the principle of the piezo-electric crystal and the methods by which it is used in the more common types of crystal-controlled oscillators. With this knowledge you will be able to progress into the study of other transmitter components, and soon you will have an understanding of a modern broadcast transmitter."

1. **NEED FOR CRYSTAL-CONTROLLED OSCILLATORS.** The fact that the ordinary vacuum tube oscillator is incapable of maintaining oscillations at a constant frequency was discussed in the preceding lesson. It is for this reason that the so-called self-excited oscillator is obsolete insofar as broadcast stations are concerned. It is improbable that the average self-excited oscillator is capable of maintaining the frequency of its oscillations closer than 1 kc. to that of the desired frequency. With careful temperature control this deviation might be reduced to a few tenths of a kilocycle, but even this amount is highly undesirable.

In its regulations, the Federal Communications Commission has established the permissible frequency deviations allowed for the various classes of transmitting stations. Of all types of stations, the broadcast station is required to maintain its actual operating frequency closer to that of its assigned frequency. The maximum frequency deviation permitted a broadcast station is 50 cycles plus or minus its assigned frequency. If it is assumed that the assigned frequency is 1,000 kc., the foregoing statement means that the carrier frequency of the station must not vary beyond the limits of 999,950 cycles to 1,000,050 cycles. This is a variation of 50 parts in a million, or a percentage variation of .005%. Such extremely small frequency deviation is easily possible only by the use of crystal-controlled oscillators.

2. THE PIEZO-ELECTRIC EFFECT. In their many experiments which eventually led to the discovery of radium, Marie Curie and her husband, Pierre, first noticed the piezo-electric effect. This phenomenon manifests itself in the following manner: When a thin plate, cut from a crystal exhibiting piezo-electric properties, is mechanically strained by tension, compression, or twisting, an electric field is produced between its two faces; that is, a voltage is set up between the points where the strain is applied. If the direction of the strain is reversed from tension to compression, the polarity of the voltage changes direction. There are many different crystals which possess this property, but it is most pronounced in the crystals of Rochelle salt, tourmaline, and quartz.

A thin slab of Rochelle salt crystal will, if vigorously twisted, develop a voltage of several hundred volts across its faces. The same amount of strain would produce a potential of a few volts across a quartz crystal. In addition, this phenomenon is reversible-that is, the application of a voltage across the faces of a crystal will produce a mechanical strain within the crystal. If a voltage of a given polarity causes a tension within the crystal, tending to pull its faces outward toward the points where the voltage is applied; the reversal of the polarity of the voltage will cause a compression of the crystal, tending to squeeze it together.

Piezo-electricity derives its name from the Greek work "piezo" which means to press, and it is evident that this phenomenon provides a means of converting mechanical energy directly into electrical energy or vice versa. By setting the crystal into vibration, an alternating voltage is created, and by applying an alternating voltage to the crystal, it is set into vibration.

It has been learned that a tuned circuit possesses a natural frequency of oscillation depending upon the inductance and capacity of the circuit, and if allowed to choose the frequency of its oscillation, it will oscillate at its natural frequency. Such oscillations are called "free oscillations" as distinguished from forced oscillations in which the tuned circuit is caused to oscillate at some frequency other than its natural frequency. In a like manner, it has been discovered that any body has a natural frequency of mechanical oscillation or vibration. A piano string, for example, vibrates at a frequency which depends upon its mass, length and tension. The diaphragm of a headphone has a natural frequency of vibration determined by its mass and stiffness.

A thin slab of crystal also has a natural frequency of vibration, which is also called its "resonant frequency". The vibration of the crystal is opposed by two factors: the mass of the crystal and its stiffness. At some particular frequency, these two oppositions are equal in amount, and since they are opposite in effect, they will cancel each other. At this frequency, the crystal is said to be mechanically resonant.

The application of an alternating voltage to the crystal will cause it to vibrate at the frequency of the applied voltage. If the frequency of the applied voltage is changed until it is equal to the natural frequency of the crystal, the vibration amplitude of the crystal is enormously increased. In a later section of this lesson, we shall see how this vibrating crystal is able to control the oscillation frequency of an oscillator.

3. QUARTZ CRYSTALS. At first thought, it would seem that Rochelle salt would be the ideal material to use in crystal-controlled oscillators. Of all types of crystals, it develops the greatest potential for a given amount of strain. Rochelle salt, however, has several disadvantages which make it undesirable for this purpose. First, it is rather fragile, and the severe vibrations which it would experience would probably cause it to chip and break. Second, its natural vibration frequency is affected by moisture; and third, the amount of variation of its vibration-frequency with changes in temperature is relatively great. Finally, it is not found in the natural state, but must be deposited from saturated solutions of Rochelle salt, which make its manufacture fairly expensive.

Although Rochelle salt crystals are not used for the frequency control of oscillators, they find wide application in the manufacture of crystal phonograph pick-ups, crystal microphones, and crystal speakers. A complete discussion of crystal microphones will be presented in a future lesson.

Tourmaline does not possess the disadvantages of fragility, and susceptibility to moisture, but since it is a semi-precious stone, it is very costly. It has, however, been used considerably in the past few years for controlling the frequency of oscillators at frequencies higher than those at which quartz is practical. Since the majority of all crystal-controlled oscillators employ quartz, this material will now be considered in detail.

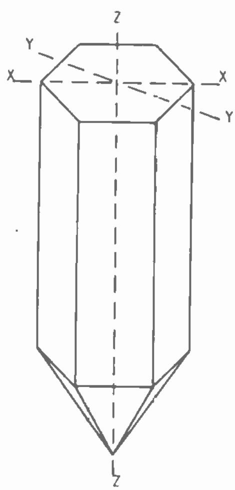


Fig.1 A regular hexagonal prism. Quartz crystals as found in their natural state are more or less of this form.

Not all kinds of quartz are suitable for frequency control. In fact, only that mined in Brazil is ordinarily usable. Quartz crystals in their natural state occur in the form of a rough hexagonal prism. Some specimens are very regular in shape and approach the ideal form shown in Fig. 1. Most of them taper to a point as shown in this drawing.

A piece of quartz as found in the natural state has three principal axes. These are called the X, Y, and Z axes. The Z axis (also called the optical axis because of its ability to affect light rays) extends longitudinally through the center of the crystal: its position may be seen in Fig. 1. The X axes pass through the corners of the hexagon, and are indicated as X_1, X_2,

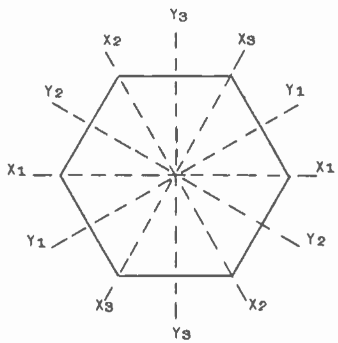


Fig.2 A cross section of a crystal showing the X and Y axes.

and X_3. The Y axes are perpendicular to the faces of the crystal and are designated as Y_1, Y_2, and Y_3. A cross section of the crystal is illustrated in Fig. 2, from which the positions of the various X and Y axes may be clearly noted. The X axes are sometimes called the "electrical axes", while the Y axes are often known as the "mechanical axes."

The crystal does not exhibit any piezo-electric properties along its Z axis, so the following discussion will be limited to the other two axes. If a flat section is cut from the crystal so

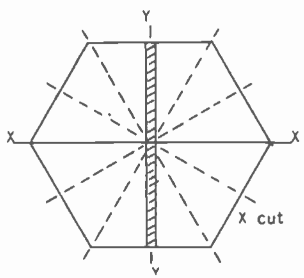


Fig.3 Illustrating how an X-cut crystal is cut from the cross section.

that its flat sides are perpendicular to any one of the X axes, the application of an alternating voltage to the two flat sides (that is along the X axis in Fig. 3.) will cause a mechanical strain along the Y axis from which the crystal was cut. It should be noticed that for every Y axis, there is an X axis perpendicular to it. By cutting the flat section so that its plane coincides with a Y axis, the vibration of the crystal along this Y axis will produce voltages along the X axis, which is perpendicular to the flat faces of the crystal. Such a crystal is said to be an X-cut crystal, since the voltages are produced along an X axis. This flat section is also

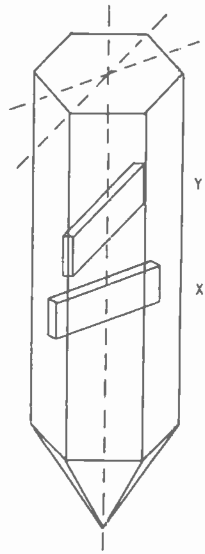


Fig.4 Showing the position of the X and Y-cut crystals with respect to the hexagonal prism.

called a Curie-cut crystal after the investigators who experimented with it. Figs. 3 and 4 show how the original hexagonal crystal would be cut to produce an X-cut crystal.

By cutting the flat section so that its flat faces are perpendicular to a Y axis, a Y-cut crystal is produced. The plane of this crystal is along an X axis, and it vibrates along this axis, but the piezo-electric voltages are created along the Y axis per-

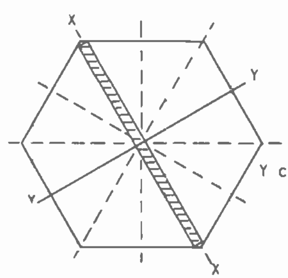


Fig.5 Illustrating how a Y-cut crystal is cut from the cross section.

pendicular to its flat faces. The position of a Y-cut crystal with respect to the original form is shown in Figs. 4 and 5. This type of crystal is sometimes called a 30 degree cut. A cross-section cut out of a hexagonal crystal is shown in Fig. 6. This photograph also shows a block after it has been squared up, and a finished crystal.

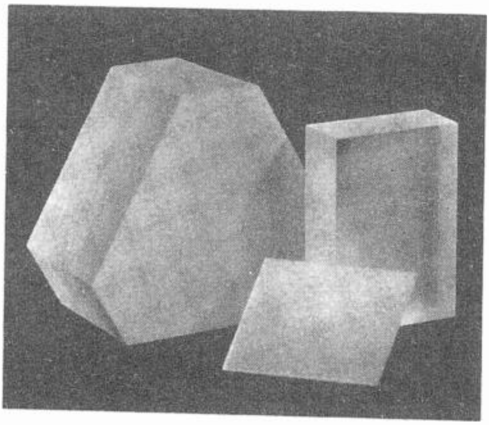


Fig.6 A cross section of a quartz prism, a block after it has been squared up, and a finished crystal.

As will be explained in a subsequent paragraph, the temperature of a quartz crystal has an effect on its resonant vibration frequency. The relation between the temperature and frequency differs considerably for the X and Y-cut crystals. For the X-cut, if the temperature is increased, the resonant vibration frequency decreases. The extent of the frequency variation is about 20 to 25 cycles per million cycles per degree Centigrade. This means that if an X crystal is cut to have a resonant frequency of 1000 Kc. (1,000,000 cycles) for each degree increase of temperature, the resonant frequency will decrease 20 to 25 cycles. Since the frequency goes down when the temperature goes up (and vice versa), the X-cut crystal is said to have a "negative temperature coefficient".

The temperature coefficient of a Y-cut crystal is usually positive and is somewhat larger than that of the X-cut crystal. For most Y-cut crystals, the frequency change will be from 60 to 100 cycles per million cycles for each degree (Centigrade) increase in temperature. The temperature coefficient of a Y-cut crystal is far more difficult to predict than that of an X-cut; in some instances Y-cut crystals have been found to have a negative coefficient. In general, however, the Y-cut is said to have a positive temperature coefficient between the limits as given above.

It is not known exactly why the X and Y-cut crystals have negative and positive (respective) temperature coefficients; however, a crystal may be considered as comparable to a very complex arrangement of coupled circuits, the circuits having different frequencies and with different degrees of coupling. Just as a complex arrangement of coupled electrical circuits will react on one another, so do the various dimensional vibration frequencies of a crystal. These facts are probably responsible for the frequency-temperature properties of quartz crystals.

4. A-CUT CRYSTALS. Since the X-cut crystal has a negative temperature coefficient and the Y-cut a positive, it seems logical that there should be some way of cutting a crystal which would have a practically zero coefficient of temperature. After much experimentation a cutting angle was discovered which would produce this desirable property of a quartz crystal. The general designation for this type of cut is the "A-cut"; however, many manufacturers have attached a different letter to their particular product for identification and perhaps to indicate a slightly different cutting angle and different resultant characteristics.

The expression "low temperature coefficient" is frequently used to identify all the various crystals which have the property of maintaining a nearly constant frequency of vibration with temperature changes. Of all the low-drift crystals available, the so-called "AT cut" is perhaps the most popular. The angle at which the AT plate is cut from the parent hexagon crystal is shown in

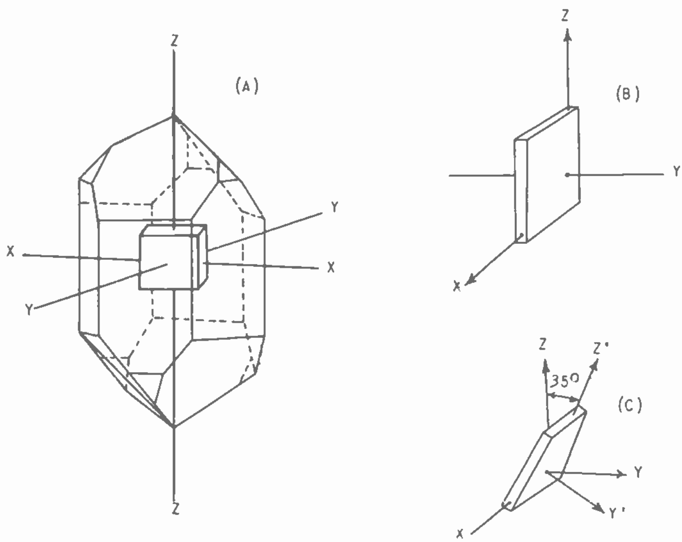


Fig.7 (A) Showing the position of a Y-cut plate in the rough quartz crystal. (B) Same Y-cut plate as in (A) after it has been removed from the rough quartz. (C) Illustrating the change in cut from the Y plate to the A plate. The plate is rotated 35° from the Z axis about the X axis.

Fig. 7. Fig. 7A shows a cross-section of the raw quartz hexagon and the position of a Y-cut crystal at the center. Each axis is identified. In B of this same figure, the Y-cut plate has been removed from the raw quartz and the axes are again identified. C in Fig. 7 shows how the cutting angle is rotated about the X axis

approximately 35 degrees from the Z axis to arrive at the AT cut plate. The 35 degree angle from the Z axis may be changed slightly in either direction without seriously affecting the characteristic low temperature coefficient. When this is done, other desirable properties may be obtained which will be discussed later.

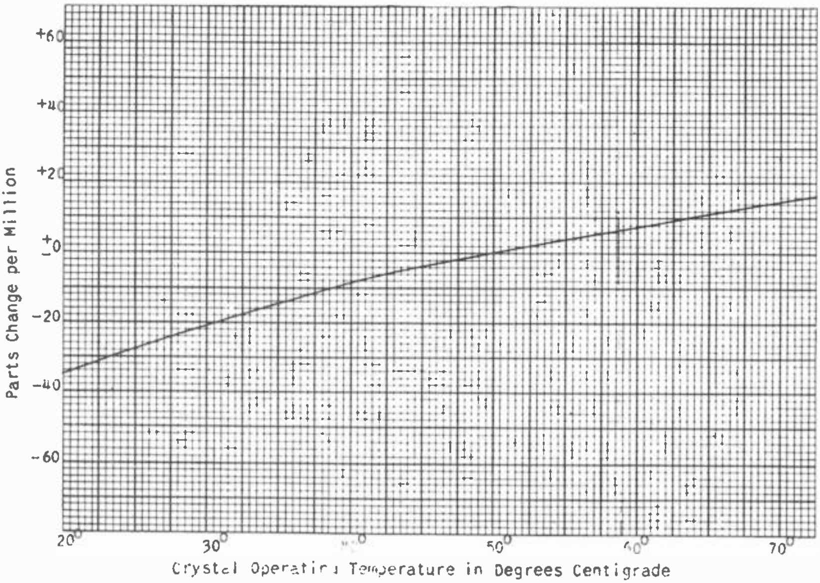


Fig.8 Indicating the effect of temperature on an A-cut crystal.

Fig. 8 shows a temperature-frequency graph for a typical A-cut crystal. Notice that the temperature may be changed over a considerable range without altering the vibration frequency more than a few cycles. Since the primary purpose of a crystal is to control the frequency of a transmitter, it is evident that the A-cut is superior to the X and Y insofar as its ability to hold a given frequency is concerned. When good frequency stability is required, an X or Y-cut crystal must be kept within a heat-proof oven in which the temperature is automatically controlled. An A-cut crystal can provide an equivalent accuracy of frequency control without the expense of the temperature regulating equipment. When the temperature of an A-cut crystal is automatically controlled within one degree or less, the frequency of vibration is extremely accurate. Other advantages of the A-cut crystal will be discussed later.

5. THICKNESS-FREQUENCY RELATION OF QUARTZ CRYSTALS. There is a definite relationship between the thickness of a crystal and its natural frequency of vibration. This relation is different

for each of the three types of cuts. The finished crystal is usually slightly less than one inch square, while its thickness depends on the vibration frequency desired.

For an X-cut crystal, the thickness-frequency relation is expressed by equation:

For an X-cut crystal: $T = \frac{112.6}{F}$

Where: T is the thickness in inches and
F is the frequency in kilocycles.

For a Y-cut crystal: $T = \frac{77}{F}$

Where: T is the thickness in inches and
F is the frequency in kilocycles.

For an A-cut crystal: $T = \frac{66.2}{F}$

Where: T is the thickness in inches and
F is the frequency in kilocycles.

By inspecting the above equations, it is seen that for a given frequency, the X-cut plate is thickest, the Y-cut plate next, and the A-cut plate is thinnest. As a general rule, thin crystal plates are damaged easily by excessive vibration; however, this is no serious disadvantage in the case of the A-cut plate, because the mode of its vibration differs from that of the X and Y in such a way that it will stand considerable abuse.

Using the three formulas just given, it is found that if each of the three types of crystals is ground to have a vibration frequency of 1000 Kc., the X-cut will be .1126 of an inch thick, the Y-cut will be .077 of an inch, and the A-cut will be .0662 of an inch. Crystals have been ground successfully to vibrate at fundamental frequencies as high as 11,000 Kc. The plates are extremely thin, however, and must be handled with caution.

6. COMPARISON OF A CRYSTAL TO A TUNED CIRCUIT. It has been previously stated that a quartz crystal has a natural frequency of vibration, and when caused to vibrate at this frequency, the intensity of its movement is relatively large compared to that produced at other frequencies. Thus it exhibits the phenomenon of resonance, much in the same manner as that of an electrical tuned circuit. Since its resemblance in operation is very similar to a tuned circuit, it seems logical that it could be replaced, insofar as the analysis of its action is concerned, by an equivalent electrical circuit. This was first done by Dr. H. J. Ryan in 1918, and has since been confirmed by several other investigators.

The inductance of a tuned circuit tends to oppose any change in the current flowing through the circuit. Thus, it tends to prevent the current from rising, and after the current has attained its maximum value, the inductance tends to prevent the current from falling. In this action, the inductance is very similar to a large mass or weight. The mass of a body tends to prevent a force from setting the body into motion, and likewise, the mass also tends to prevent the stopping of this motion once it has been established. A quartz crystal, like any other body, possesses mass, and it is the mass of the crystal which attempts to prevent it from being set into vibration.

The capacitance of a tuned circuit stores energy during a part of the cycle, and releases it during the remaining part. A similar characteristic is found in the vibrating crystal; it is the crystal's elasticity. When a force is applied to a crystal, the mass of the crystal tends to prevent its being set into motion, and after the motion has been established, the mass tends to prevent the stopping of the motion. However, as the crystal becomes more and more distorted from its normal shape, its elasticity builds up a force which is proportional to the amount of distortion which the crystal has experienced. This force tends to restore the crystal to its normal shape. The development of this elastic force is analogous to the charging of the condenser of a tuned circuit, while the return of the crystal to normal shape is equivalent to the discharge of the condenser. As the crystal reaches its normal shape, it does not stop its motion, but continues to move, thus distorting itself in the opposite direction. This is due to the mass of the crystal which tends to maintain the motion, much in the same manner that the collapsing magnetic field of a coil tends to charge a condenser in the opposite direction.

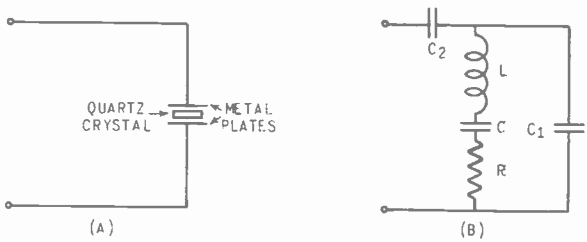


Fig.9 (A) Showing the quartz crystal plate between the metal electrodes which serve as a holder. (B) The equivalent electrical circuit of a quartz crystal mounted in a holder.

The third property of a tuned circuit is its resistance which tends to dissipate the energy of the circuit, and thus prevent the maintenance of sustained oscillations unless additional energy is continually added. The analogy between the vibrating crystal and the tuned circuit can be extended to include this property also. In their to and fro motion, the particles of the vibrating crystal encounter friction which dissipates a part of the energy in the form of heat, thus the amplitudes of the vibration of the crystal will decline unless additional energy is continually supplied.

Since the crystal has the same properties as a tuned circuit, it can, for analytical purposes, be replaced by its electrical equivalent. A diagram of this electrical equivalent is shown in Fig. 9B. When used to control the frequency of an oscillator, the crystal is mounted between two flat metal plates, as shown in Fig. 9A. The condenser C_1 of Fig. 9B represents the capacity existing between these two plates considering the crystal itself as the dielectric. Condenser C represents the elasticity of the crystal; L represents the inductance of the crystal, due to the mass; while R represents the resistance of the crystal caused by its internal losses. The capacity C_2 represents the series capacity between the crystal and its electrodes.

Neglecting capacity C_2 for the present, it will be noticed that the electrical network consisting of L, C, C_1 and R in Fig. 9B has the properties of either a series or a parallel resonant circuit. At some frequency, the reactances of L and C will be numerically equal; hence, a series resonant condition will exist. This will be called the "series resonant" or "natural" frequency of the crystal.

If the frequency is increased slightly above the natural frequency, the reactance of L will rise and the reactance of C decreases. Thus, the circuit consisting of L, C, and R in series has a net inductive reactance. This inductive reactance will balance with the capacitive reactance of C_1 (at a frequency slightly higher than the series resonant frequency) to produce parallel resonance. Hence, when a quartz crystal is placed between the metal plates of a holder, it will have two resonant points, one equivalent to series resonance and one equivalent to parallel resonance. The parallel resonant frequency will be slightly higher than the series resonant frequency.

The capacity C_2 in Fig. 9B is only effective when the holder electrodes are not in direct contact with the crystal faces. This condition will exist in holders of the "variable air-gap" type to be described later on. As the value of the capacity C_2 is decreased, the resonant frequency will rise slightly.

The natural or resonant frequency of an ordinary tuned circuit is determined by its inductance and capacity; that of a piezo-electric crystal is dependent upon its mass and elasticity. Increasing the thickness of a crystal increases its mass and results in a decrease in the natural frequency of vibration. The elasticity along a given axis is constant; therefore, the vibration frequency is determined only by its thickness. However, the elasticity of quartz along its X axis is not the same as along its Y axis; hence, for a given thickness, X and Y-cut crystals do not have the same frequency of vibration. Elasticity of a quartz plate along the angle from which an A-cut is secured is different from that along either the X or Y axis; hence, the A-cut will have a different thickness-frequency coefficient from either the X or Y.

It is possible to express the mass, elasticity, and resistance components of a quartz crystal in terms of the equivalent electrical circuit units of henrys, farads, and ohms. For example, a typical X-cut crystal ground to the proper thickness for a natural vibration frequency of 1100 Kc. possesses the following equivalent electrical characteristics.

L = .33 henry,
C = .065 micro-microfarad,
R = 2700 ohms,
C_1 = 1 micro-microfarad.

(Refer to Fig. 9B)

From the table, it is at once seen that the equivalent electrical inductance of the crystal is quite high, much higher than used in ordinary tuned circuits. In fact, the equivalent inductance of quartz crystals may vary from one-tenth of a henry to 100 henrys, depending upon the manner in which the crystal is cut from the raw quartz, its physical dimensions, and the frequency. The equivalent capacity is much lower than that of an ordinary tuned circuit while the equivalent resistance is very high. The unusually high resistance would seem to be a disadvantage, for it was learned in previous lessons that the higher the resistance in a tuned circuit, the lower the effective gain and selectivity. However, there are other factors to be taken into consideration.

It was learned that the voltage gain of a tuned circuit is equal to the reactance of either the coil or the condenser (they are equal at resonance) divided by the resistance (opposition to the flow of the circulating RF current) of the circuit. Both the voltage gain and the sharpness of selectivity depend upon this ratio and not only upon the resistance of the circuit. Thus, if the reactance-to-resistance ratio is high, the voltage gain and selectivity will be high; while if the ratio is low, the voltage gain will be less and the selectivity curve broadened considerably.

To determine the sharpness of the resonance curve that would result from the equivalent electrical circuit of a crystal, let us first calculate the inductive reactance of the above-mentioned crystal at a frequency of 1100 Kc. This is:

$$XL = 6.28 \times 1{,}100{,}000 \times .33 = 2{,}279{,}640 \text{ ohms}$$

The ratio of the reactance to the resistance is:

$$\frac{X}{R} = \frac{2{,}279{,}640}{2{,}700} = 844$$

When it is realized that with most careful construction, it is practically impossible to design a tuned circuit to have a ratio of reactance to resistance greater than 150 (in the medium frequency ranges), it is clearly evident that the crystal has a much sharper selectivity curve and is particularly adaptable for controlling the frequency of an oscillator.

7. CRYSTAL-CONTROLLED OSCILLATORS. When used to control the frequency of an oscillator, a quartz crystal is placed in a specially designed holder. Since the design of this holder is very important, a separate section of this lesson will be devoted to this subject. For the present, it will be assumed that the crystal is mounted between two parallel, flat, metal plates which make light contact with the surfaces of the crystal. The symbol used

to designate a crystal in a circuit is that shown in Fig. 10.

Since a crystal has all the properties of a very sharply tuned resonant circuit, it may be used to replace the grid tank circuit of an Armstrong oscillator. When this is done, the oscillator appears as shown in Fig. 11. Notice that the grid circuit must be shunt-fed, since it is impossible for the direct current component of the grid current to flow through the crystal. The operation of this oscillator is precisely the same as that of the Armstrong type.

Fig.10 The symbol used to designate a quartz crystal.

With the filament heated, the closing of the plate circuit switch applies plate voltage to the tube, and a pulse of plate current flows through the plate tank, LC. The counter voltage produced across L as the current rises, charges the condenser C, and feeble oscillations are established in this tank circuit. The oscillating current produces RF voltage drops across the plate tank, which then cause an RF current to flow through the interelectrode capacity of the tube to the crystal. If the frequency of the feedback voltage is equal to the natural frequency of the crystal, the crystal vibrates through relatively large amplitudes, and, in so doing, it produces, by piezo-electric effect, RF voltages across its faces, which then serve as the grid excitation voltage. Thus, the amplitude of the oscillations occurring in the plate tank circuit build up until the saturation point of the tube is reached.

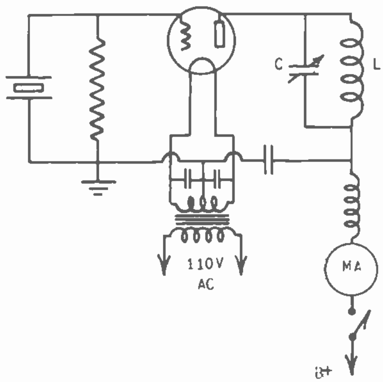


Fig.11 A simple crystal oscillator which is a modification of the TGTP oscillator.

By very slightly detuning the plate tank circuit to a frequency higher than resonance, it is caused to have a small predominance of inductive reactance. This amount of inductive reactance compensates for the capacitive reactance of the interelectrode capacity of the tube and makes the phase of the feedback voltage correct to maintain continuous oscillations. Detuning the plate tank in this manner also increases the stability of the oscillator. If the plate tank is detuned more than a small amount, the feedback voltage is reduced so that it is incapable of maintaining the crystal in vibration, and the oscillations in the plate tank stop for lack of grid excitation.

When the plate tank is detuned enough to cause the oscillations to stop, the oscillating current in the plate tank dies out, and, since there are no RF voltages built up across the plate tank, the actual voltage applied to the plate of the tube remains steady. Under this condition, the tube is not working into any load, and, since there is no grid excitation, the plate current flows all the time and at a steady value. A milliammeter connected in the plate circuit would indicate a fairly large value of plate current.

As the plate tank is tuned through the resonant frequency, a dip in the plate milliammeter is noted. At exact resonance, the meter indicates a plate current of less than half of that which flows when the circuit is not oscillating. By tuning the plate tank to resonance, small oscillations are started which then excite the crystal and cause the building up of the oscillations in the plate tank. When the circuit is oscillating, the actual voltage applied to the plate varies through wide amplitudes, since the RF voltages across the plate tank add to and subtract from the plate supply voltage. In addition, the voltages produced across the crystal drive the grid positive during a part of each cycle, and cause the flow of grid current, which produces a bias voltage in flowing through the grid leak. Since the plate current is now flowing in the form of pulses of short duration, the average plate current as read by the DC milliammeter is much less than before. Therefore, the presence of oscillations in the plate tank is indicated by a dip in the DC plate current.

8. CRYSTALS AND SERIES RESONANCE. A quartz crystal will have minimum impedance at its series resonant frequency and highest impedance at the parallel resonant frequency. At frequencies remote from these, the crystal will act merely as a fixed condenser. A representative reactance curve for a crystal is shown in Fig. 12. At frequencies slightly below the series resonant frequency, the net reactance of the crystal is capacitive and slightly above, the net reactance will be inductive. As indicated in Fig. 12, frequencies slightly below parallel resonance will cause the crystal to appear as an inductance and at frequencies slightly higher than parallel resonance, the crystal will appear as a capacity. The property of a crystal to act as a resonant circuit having a very rapid increase in impedance on either side of series resonance is very useful in certain types of RF filters and for the control of frequency in certain types of oscillators.

One circuit in which the oscillator frequency will be controlled by the series resonant frequency of the crystal is shown in Fig. 13. This circuit is often used in frequency standards and is particularly recommended for use with low-temperature coefficient crystals from 85 Kc. to 150 Kc. An inductance-capacity tank circuit with the crystal connected directly in series constitutes

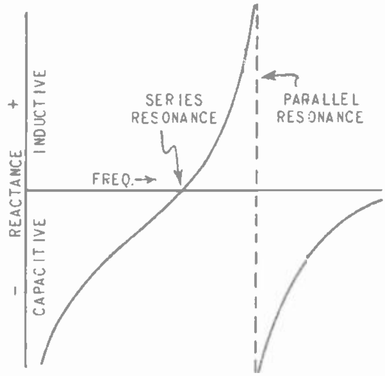


Fig.12 The reactance curve of a quartz crystal.

the portion of the circuit in which the oscillations are generated. This is a modified Colpitt's oscillator. It has a relatively low power output, but very good frequency stability. When the tank is tuned to a frequency at or close to the resonant frequency of the crystal, the crystal will assume control of the generated oscillation since its impedance is lowest at the resonant

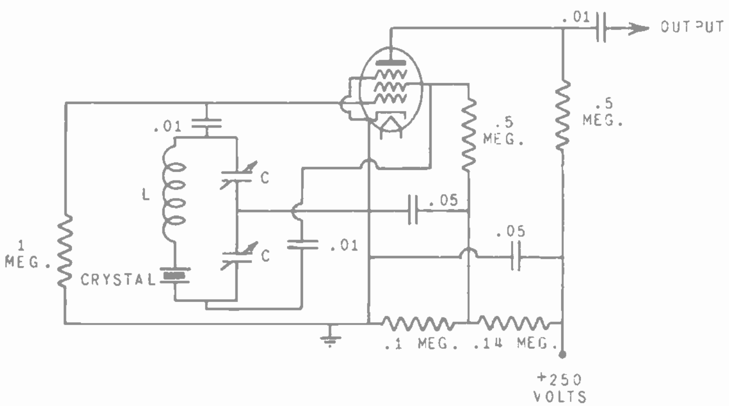


Fig.13 Modified Colpitts oscillator with crystal control. Suitable as a low-frequency standard oscillator.

frequency and rises very rapidly for other frequencies. The crystal will be capable of holding control over the oscillations for a small tuning range of the tank condenser C; beyond which it will serve only as a series condenser in the tank circuit.

The oscillating frequency can be varied over a limited range by changing the capacity of the tuning condenser and the crystal will still maintain control of the oscillations. This will be true as long as the frequency is not shifted over about 8 cycles per 100 Kc. The values shown in the diagram in Fig. 13 should be suitable for frequencies between 50 and 500 Kc. The LC ratio is not extremely critical; however, best frequency stability will be obtained with the LC ratio quite low. If an attempt is made to operate this modified crystal-controlled Colpitt's oscillator much above 500 Kc., it will be difficult for the crystal to maintain control of the oscillations. There will be a strong tendency in the circuit shown in Fig. 13 for self-oscillations at frequencies other than the series resonant frequency of the crystal; the range over which the crystal can assume control of the oscillation is very narrow. The circulating tank current in this circuit must be kept quite low in order that the crystal itself will not be damaged. A small receiver-type pentode tube such as the 6J7 is quite satisfactory for use in this circuit.

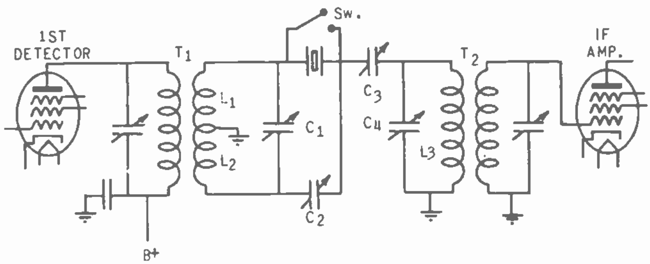


Fig.14 Typical circuit of a crystal filter as used in the IF amplifier of a superheterodyne receiver.

The series resonant properties of a quartz crystal are also used advantageously in modern communications receivers to produce a high degree of selectivity. Special quartz crystals are ground for this purpose and are made to have an extremely high Q[1] of from 9,000 to 16,000. By using a crystal having such a high Q in a filter circuit, the frequency discrimination or selectivity of the IF amplifier in a superheterodyne will be many times better than could be obtained with ordinary tuned circuits. With proper adjustment of a quartz crystal filter circuit, it will be possible to limit the band of frequencies passed through the IF amplifier to 50 cycles.

The theory of the crystal filter circuit shown in Fig. 14 can be explained with the aid of an elementary circuit as outlined in Fig. 15. The transformer T_1 is assumed to have a current through the primary and a voltage induced in the secondary at the intermediate frequency. The components L, C, and R represent the crystal. If the crystal is ground so that its series resonant fre-

[1]The "Q" of a crystal or tank circuit means the ratio of reactance to resistance. The Q will be discussed fully in a later lesson.

quency is equal to the IF, the impedance of the path from L_1 through L, C, and R to L_3 will be very low and maximum IF current will flow in L_3. It is evident from this illustration that the series resonant property of the crystal is the one desired (not parallel resonance) and that maximum signal transfer will occur through the crystal at the IF frequency.

Fig.15 Elementary circuit showing the equivalent series-resonant property of a crystal acting as a filter to pass the fundamental frequency only.

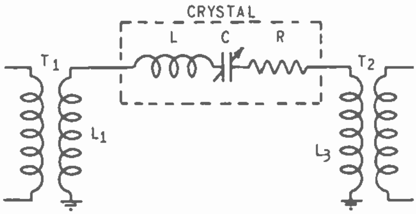
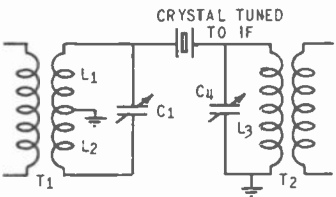


Fig. 16 is changed from Fig. 15 in that L_1 and L_3 are tuned and the secondary of T_1 is grounded at the center. With these changes, the circuit may be made to operate as before if the resonant circuits are properly tuned. The voltage developed across L_1 by the circulating current in the series-tuned secondary of T_1 can pass a high current through the series resonant

Fig.16 Same as Fig. 15, but with L, C, and R replaced with the crystal itself.



crystal and set up an oscillating current in the parallel tuned primary of T_2. If, however, both the secondary of T_1 and the primary of T_2 are tuned to exact resonance with the IF and the crystal, the impedance offered by these circuits to the flow of IF current will be very high and the effect will be to increase the value of the series resistance R in Fig. 15. Increasing the size of this resistance causes the resonance property of the crystal to be damped and the selectivity of the crystal filter circuit will be materially reduced. Therefore, both C_1 and C_4 are adjusted so as to throw their respective circuits slightly off exact resonance. C_1 is slightly increased in capacity so as to allow inductive reactance to predominate in the series resonant secondary of T_1. The oscillating current will then lag the induced voltage and the damping effect of the circuit on the crystal will be destroyed. The capacity of C_4 is slightly decreased so that the inductive reactance of L_3 will be lower than the capacity reactance of C_4 in the parallel tuned circuit. Hence, more current will be passing through L_3 than through C_4 and the net line current through L_3C_4 will be lagging. The tuned primary of

T_2 then presents the effect of an inductance and the damping resistance effect of the circuit on the crystal will be destroyed.

Now in order to obtain maximum IF current flow through the crystal and still avoid the appearance of a high series resistance, the inductive effect of the secondary of T_1 and the primary of T_2 must be eliminated by an equivalent capacitive reactance in series. When series Xc is added, a series resonant circuit with its characteristic low impedance will be established. The condenser C_3 in Fig. 14 serves this purpose. When C_3 is adjusted to the proper capacity, maximum current will flow through the crystal and the oscillating current in the primary of T_2 will be highest. This, in turn, supplies maximum signal on the grid of the IF amplifier.

Again referring to Fig. 16, notice that the crystal in its holder would provide capacity coupling between T_1 and T_2. The capacity across the crystal in this case is not that due to the crystal's elasticity, but that due to the fact that two metal plates are insulated from each other and have a dielectric material between them. This corresponds to C_1 in Fig. 9B. Since the dielectric constant of quartz is quite high, the capacity across the crystal is much greater than would be measured with the same size plates separated by air. Capacity coupling between T_1 and T_2 is undesirable, since interfering signals of a high frequency (and harmonics of the IF) could easily pass through and offset the purpose of the entire crystal filter which is to provide selectivity.

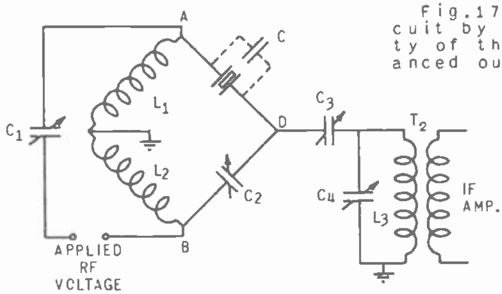


Fig.17 The bridge circuit by which the capacity of the crystal is balanced out.

The capacity coupling can be eliminated by use of a balanced Wheatstone bridge. The secondary of T_1 was center-tapped to make the bridge circuit possible. Fig. 17 shows the bridge circuit clearly; it consists of two inductive arms, L_1 and L_2, and two capacitive arms, C_2 and the capacity across the crystal represented by the dotted C. When properly balanced by varying the capacity of C_2, no RF voltage can appear between point D and ground even though these voltages are developed across the resonant circuit from A to B. It should be understood that the Wheatstone bridge will not balance out the desired signal at the intermediate frequency because the low series impedance of the crystal is far less than the capacity reactance of C and the bridge is completely out of balance at the desired signal frequency.

From Fig. 14 it can be seen that C_2 and C_1 in series are in parallel with the crystal capacity C. This makes it possible to vary the effective capacity across the crystal over a limited range by changing the capacity of C_2. By so doing, the parallel resonant frequency of the crystal can be adjusted a few kilocycles away from the series resonant frequency and set to an adjacent channel station which is causing interference with the desired station. When parallel resonance of the crystal opposes an interfering signal, the signal is eliminated from the remainder of the IF amplifier by the high impedance characteristic of a parallel resonant circuit.

C_2 is often called the "phasing condenser", the "neutralizing condenser", or the "balancing condenser". After learning the operation of the circuit, the application of these expressions should be self-explanatory. The selectivity of the crystal filter circuit can be varied over a limited range to provide sharp CW reception or low-fidelity phone reception by varying C_2, by changing resonance in the primary of T_2, or both. If high-fidelity phone reception is desired, the crystal may be shorted out entirely by the switch Sw in Fig. 14.

9. FACTORS AFFECTING FREQUENCY STABILITY. The important factors which affect the generated frequency of a crystal oscillator are:

1. Changes in voltages applied to the elements of the tube.
2. Variations in the load on the oscillator.
3. Changes in temperature.

In a properly designed transmitter, the first two listed above have relatively little effect. The crystal oscillator should be supplied with plate voltage from a separate power supply having good voltage regulation. The load on the RF output of the oscillator should be light and constant. It is particularly important that the load on the oscillator does not change with modulation in a radiotelephone transmitter. If this should occur, the frequency will change or "flutter" with modulation and frequency instability or "dynamic" instability will result. A following lesson on buffer amplifiers will explain in detail how dynamic instability can be prevented.

Variations in the temperature of the crystal plate, the tube itself, and the circuit components will alter the frequency of the generated oscillations; however, of these three, changes in the temperature of the crystal is the most important. It has previously been stated that the natural frequency of a crystal plate is determined by its mass and elasticity (comparable to inductance and capacity, respectively). After the quartz plate has once been ground to a certain thickness, its mass is fixed, and is not affected by temperature changes. The elasticity, however, is affected by temperature changes and for that reason the temperature of the crystal must be maintained constant if extreme frequency stability is the principal objective.

The temperature coefficients of X and Y-cut crystals have been given previously. It was stated that the coefficient of the X-cut crystal was 20 to 25 parts per million per degree Centigrade and was negative. The coefficient of the Y-cut is 60 to 100 parts per million per degree Centigrade and is positive. We shall solve illustrative examples to show how these coefficients should be used.

Example 1: If the temperature coefficient of an X-cut crystal is -20, how much will the frequency of a 600 Kc. oscillator change if the temperature varies from 55.2 to 53.8 degrees Centigrade?

Solution: Since the coefficient is 20 parts per million, the change in frequency will be 20 cycles for each megacycle. The frequency of the given oscillator is only 600,000 cycles, or .6 megacycle. Thus, the coefficient for a frequency of 600 Kc. will be .6 times 20, or 12 cycles. The frequency will increase 12 cycles for each degree decrease in temperature because the coefficient is negative. The temperature varies from 55.2 to 53.8 degrees, or a decrease of 1.4 degrees. The number of cycles change in the 600 Kc. oscillator frequency will then be 12 cycles (cycles change per degree per 600,000 cycles) times 1.4, or 16.8. The frequency will increase 16.8 cycles above 600,000 cycles since the temperature decreased and the coefficient of the X-cut is negative. The new oscillation frequency will be 600,016.8 cycles per second.

Example 2: If the temperature of a 2,500 Kc. Y-cut plate increases from 50 degrees to 58 degrees, what will be the new oscillation frequency? Assume that the temperature coefficient is positive and is 70 parts per million per degree.

Solution: The fundamental vibration frequency of the Y plate is 2.5 times the reference frequency of 1 megacycle on which the coefficient is based; therefore, the given coefficient should first be multiplied by 2.5. The coefficient then becomes 2.5 times 70, or 175. This is the number of cycles change per degree per 2.5 million cycles. The problem states that the temperature rises 8 degrees, so the total number of cycles change will be 8 times 175, or 1,400 cycles. The coefficient of a Y-cut crystal is positive, so as the temperature rises the frequency will increase, making the new oscillation frequency 1,400 cycles above the former frequency of 2.5 megacycles, or 2,501,400 cycles per second.

The temperature coefficient of A-cut crystals is much less than that of either X or Y-cut plates. Ordinarily the crystal manufacturers do not specify a definite frequency-temperature coefficient for A-cut plates. They are called "low-drift" crystals; in general most A-cut plates have a positive frequency-temperature coefficient of 1 to 4 parts per million per degree Centigrade. This temperature coefficient of an A-cut crystal does not remain constant over a wide range of temperature variations; that is, the curve of frequency versus temperature is not a straight line; in fact, the coefficient may be positive over one part of the temperature range and negative over the other portions.

All of the crystal manufacturers do not use the letter "A" to identify their low-drift crystals. For example, the Bliley Company use the letters C, D, and E to designate low-drift crystals applicable for different frequency ranges and having different minor characteristics. In the event that exact specifications are needed on a crystal of the low-drift type classified by the manufacturer under one of these special type numbers, it is necessary to obtain that information directly from the crystal manufacturer.

The operating temperature of a crystal is dependent upon the surrounding temperature, the amount of heat developed by the crystal while oscillating, and the rate of heat dissipation by the crystal holder. For good frequency stability the crystal holder should be capable of dissipating heat rapidly. In addition, the intensity or amplitude of the crystal's vibration should be kept as low as possible in order that a minimum of heat will be developed. Where a high degree of frequency stability is absolutely essential, the crystal temperature must be controlled automatically in a constant temperature oven. Such ovens will be described in detail in a subsequent section of this lesson.

10. MODE OF VIBRATION. Any quartz crystal has two and sometimes three widely separated possible frequencies of oscillation. This is due to the fact that a vibrating body of this type can be made to vibrate in at least two different manners or "modes". In addition, a crystal plate that has been ground improperly may have one or two additional frequencies close to the thickness frequency. This is very possible when the faces of the crystal are not perfectly flat and parallel to each other. The crystal will oscillate at slightly different frequencies over small portions of the surface.

If a crystal is ground with precision accuracy, the two faces may be made perfectly flat and the sides parallel, thus eliminating the possibility of the plate being set into vibration at a "twin" frequency. With such precision grinding, the cost of the crystal plate is increased considerably; hence, the use of precision crystals is generally restricted to commercial broadcast stations. They are not in popular use for amateur and general communications work.

By properly choosing the mode of vibration, it is possible to manufacture quartz crystals of practical dimensions over a wide frequency range. At present, quartz crystals are available for oscillator frequency control over the range from 16 Kc. to 30 megacycles. X-cut plates are generally used for frequencies from 250 Kc. to about 10,000 Kc. For frequencies lower than 250 Kc., the X-cut plate becomes too large, so to reduce the crystal size to satisfactory dimensions, the crystals are cut as "bars", in which one dimension is considerably greater than the other two dimensions. Such crystals oscillate along their greatest dimension or length. When properly cut, X bars have a negative frequency-temperature coefficient ranging from 6 to 15 cycles per megacycle per degree Centigrade and can be made to vibrate at frequencies as low as 16 Kc.

Y-cut plates can be made to oscillate over the frequency range from 200 Kc. to about 8,000 Kc. In general, the use of X and Y-cut crystals for oscillators generating a frequency in excess of 4,000 Kc. is being rapidly supplanted by A-cut low-temperature coefficient crystals. A-cut plates can be made to vibrate at fundamental frequencies from about 4,000 Kc. to 11,000 Kc. By referring to the thickness-frequency coefficients previously given in this lesson, it is found that the A-cut plate is thinner than either the X or Y for a given vibration frequency. However, the A-cut plate exhibits only one mode of vibration even though its faces may not be exactly plane and parallel. Thus, the A-cut crystal can be used for frequencies in excess of the X or Y-cut plates without danger of becoming cracked or chipped due to internal strains which would be set up by several modes of vibration.

Above 11,000 Kc. A-cut plates become quite thin and are apt to be damaged even though they have but one principal mode of vibration. The upper frequency range has been extended, however, to 30,000 Kc. by grinding A-cut plates in such a way that they can be excited at the third harmonic of their fundamental frequency. Such crystals are very practical but, of course, do not oscillate as freely as the fundamental plates. In the higher frequency ranges above 18,000 Kc., third harmonic crystals have a rather high temperature-frequency coefficient, ranging from 20 cycles per megacycle per degree Centigrade to approximately 43 cycles per megacycle per degree Centigrade.

Some constructors choose to operate the crystal oscillator in the transmitter on a low fundamental frequency, then by the use of frequency doubling circuits, obtain crystal control of a high frequency carrier. This method is employed as an alternate to operating the crystal on a harmonic of its fundamental frequency. In general, better frequency stability can be secured by using a low frequency oscillator followed by doubler stages instead of using a high frequency crystal oscillator employing a third harmonic crystal plate. If simplified construction is essential and frequency stability is of secondary importance, a more economical high frequency transmitter can be constructed by using the third harmonic crystal. A complete discussion of frequency doubling circuits will be given in the following lesson.

Since X and Y-cut plates in general exhibit a "twin" frequency close to the fundamental, trouble sometimes develops wherein the crystal oscillator quickly and abruptly jumps from one frequency to another. This is a very undesirable condition in a crystal oscillator and if encountered must be corrected immediately in order to stabilize the frequency of the transmitter on the channel assigned by the government. The first indication of trouble of this nature on a commercial transmitter should be corrected immediately by replacing the crystal with another that is known to be ground in a more precise manner. If replacement of the crystal is not practical, it may be possible to readjust the crystal oscillator circuit in such a way that future trouble of this nature will not be experienced. One of the most effective adjustments in the crystal oscillator circuit would be to increase the sharpness of selectivity in the tuned plate circuit. This can be done by re-designing the tank circuit so that the inherent resistance in the tank circuit is lower or the reactance of the tank circuit is higher. In either case, if a higher ratio of reactance to resistance is established, the tank circuit will tune more sharply and oscillations at the undesired frequency are not likely to be sustained. Also, if the plate tank circuit is heavily loaded by the succeeding buffer amplifier, reducing the load will assist in increasing the sharpness of selectivity of the crystal's plate tank circuit and thereby assist in preventing the crystal oscillator from jumping frequency.

11. CRYSTAL POWER. When a quartz crystal is "oscillating" it is in reality a mechanical vibrating body. Upon vibrating mechanically, internal stresses are set up within the crystal and heat is developed as a result of the motion. Should the amplitude of vibration become too great, the internal stresses can become sufficient to shatter the crystal and destroy its ability to oscillate. The "shattering" is a physical rupture of the quartz and is caused by the crystal literally tearing itself to pieces under the extreme stresses set up by hard vibration. The rupture appears as a ragged crack or a series of cracks in the crystal. In some cases, especially with the harmonic type crystals, the rupture may occur at a single point, as though it had been punctured by high voltage.

The heat developed by an oscillating crystal is the direct result of frictional losses. Heating is undesirable in that it will cause the temperature of the crystal to change. The temperature change will cause a shift in frequency according to the frequency-temperature coefficient of the crystal. If accurate temperature control of the crystal is not economically practical, the amplitude of the crystal's vibration should be kept low and the holder should have good heat dissipating ability. In some cases the crystal holder may be rested on a large heat dissipating surface such as in direct contact with the metal chassis of the transmitter or in direct contact with a large metal block of copper or aluminum. Large masses of metal will tend to maintain a constant temperature and keep the crystal plate at a constant temperature to prevent frequency changes while the oscillator is warming up or as the crystal exciting current varies.

The amplitude at which a crystal vibrates is directly dependent upon the RF voltage applied to it. The RF voltage applied to a crystal is called the "excitation" and is a direct function of the RF current effectively passing through the crystal. It is conventional to assume that the vibration amplitude is a direct function of the crystal current.

The crystal current will depend upon several factors among which may be listed:

1. Size of the grid leak
2. Plate voltage
3. The LC ratio of the plate tank circuit
4. The load on the crystal oscillator
5. Interelectrode plate to grid capacity

As the size of the grid leak is increased, the grid bias will increase and it will be necessary for the crystal to vibrate at a higher amplitude in order to develop the RF voltages necessary to drive the grid of the tube positive. Thus the RF crystal current will increase with the size of the grid leak until the grid leak is made above about 25,000 ohms. It is not safe to use a grid leak in excess of around 100,000 ohms in a crystal oscillator circuit.

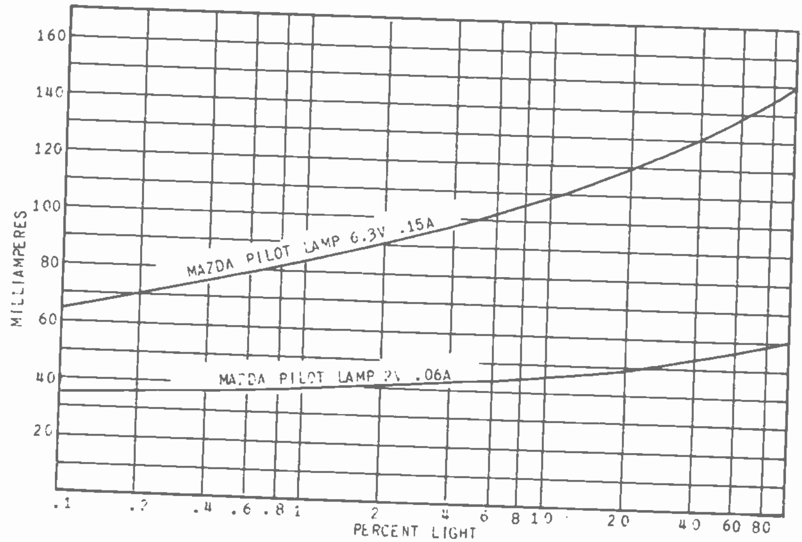


Fig.18 Graph showing the brilliancy of pilot lamps plotted against RF current.

The RF voltages developed across the plate tank circuit will also appear between the plate and cathode of the tube and will cause an RF current to flow through the inter-electrode capacity between plate and control grid. The RF current flowing through the inter-electrode capacity of the tube will return to the cathode through the crystal. As the RF current passes through the crystal, RF voltages will be developed across it to serve as the excitation for the crystal's vibration. The higher the RF crystal current, the greater will be the amplitude of vibration. Crystal current will increase if the RF voltages across the plate tank circuit increase, or if the inter-electrode capacity of the tube is increased. Thus, in a crystal oscillator circuit using a triode with a high plate voltage and a high LC ratio in the plate tank circuit, the crystal current will be very high and the crystal is apt to be damaged. By reducing the inter-electrode capacity of the tube (changing the tube to a tetrode or a pentode), by using a low plate voltage, or by using a low LC ratio in the plate tank circuit, the crystal current can be made sufficiently low that the crystal will not be endangered. The RF voltage across the plate tank circuit will decrease when the oscillator is loaded; hence, loading of the plate tank will also affect the amount of crystal current.

The amount of RF current that a crystal can safely carry will depend upon the fundamental frequency and the type of cut. This information can be secured from the manufacturer when desired. Some crystals are rated over 100 ma.; however, 60 milliamperes is a safe maximum current for most quartz crystals. To measure the crystal current, either a thermo-couple galvanometer may be used, or a low current pilot bulb can be connected directly in series with the crystal. The brilliancy of the pilot bulb will indicate roughly the amount of RF current. The chart in Fig. 18 can be used as a guide; however, the estimation of brilliancy by the human eye is subject to considerable error.

Fig.19 The Bliley VP4 pressure-type crystal holder.

12. CRYSTAL HOLDERS. As previously explained, the series-resonant and parallel-resonant properties of a quartz crystal are manifested when the crystal is placed in an RF field. This will be true regardless of whether the field is produced by an external source of energy or by the feedback of energy from the plate circuit as commonly used in crystal oscillator circuits. The method of producing the necessary field is to place the crystal between two metal electrodes connected to the source of RF potential. The

complete assembly, consisting of the two electrodes and a dust-proof case, is known as a "crystal holder". There are two popular types of crystal holders in use today; one, the pressure-type mounting, and the other, the air-gap mounting.

The pressure-type holder is used in most applications and is best fitted for installations where the crystal is to develop high potentials, or where the mounting will be subjected to external vibration or shock, as would be encountered in mobile or portable applications. In the pressure-type holder, the electrodes are maintained in intimate contact with the crystal faces and slight pressure is exerted by a small spring. Holders to be used with several different crystals are provided with a variable spring pressure so that optimum pressure can be obtained for each particular crystal. Pressure-type holders are suitable for frequencies from 400 Kc. to 30 megacycles. A typical pressure-type holder is shown in Fig. 19.

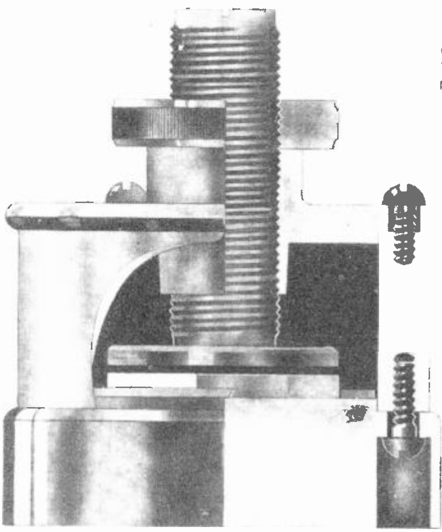


Fig.20 A cut-away drawing illustrating the parts of a variable air-gap crystal holder. (Courtesy Commercial Radio Equipment Co.)

In the air-gap type holder there is an air gap between the crystal and one of the electrodes. Holders of this type are generally provided with a means for varying the size of the gap. This is usually accomplished by attaching one electrode to a threaded screw so that the electrode can be moved in a direction perpendicular to the plane of the crystal face. By adjusting the size of the air gap, the oscillation frequency will be changed slightly. A variable air-gap holder is illustrated in Fig. 20. This cut-away drawing shows how the upper plate is raised or lowered by turning the threaded shaft. After the desired oscillation frequency has been obtained, the upper plate is locked tight by means of the knurled locking nut. The material used in the construction of this holder is a heavy, solid brass base plate and the body is made of Isolantite. The use of a large base plate aids in reducing the temperature variations of the crystal, even though the crystal current varies with different operating voltages and loads on the oscillator circuit. The heavy metal has a fairly high thermal delay; that is, it does not respond quickly to external variations in temperature. In this respect, the heavy metal plate acts very much like the filter of a power supply system; it is effective in smoothing out variations of ambient temperatures. When a holder of this type is used in conjunction with a well designed crystal oven, the temperature variation of the crystal can be held to less than plus or minus .1 degree centigrade. If a low temperature coefficient crystal is employed, this small temperature variation will result in negligible frequency deviation.

Fig.21 A plug-in type of crystal holder. (Courtesy General Radio Co.)

Figs. 21 and 22 show two views of another type of crystal holder provided with an adjustable air gap. In the cover of the crystal is a threaded plug which may be raised or lowered. This plug also serves as the upper electrode for the crystal. Notice the method of mounting the quartz plate. This arrangement prevents random movement of the crystal while vibrating. The cover is made of heavy aluminum and the base is Isolantite. The connections to the electrodes of the holder are made by means of two banana plugs.

The outstanding advantage of a variable air-gap holder is that the oscillating frequency can be varied over a limited range. This is a most convenient feature in applications where the oscillating frequency must be accurately maintained. It is not always possible to grind crystals exactly for a particular frequency, but by the use of a variable air-gap holder, the frequency can be shifted slightly to secure the desired precision. A variable air-gap holder is also advantageous in communications work for the purpose of shifting frequency slightly to avoid bad interference. Variable air-gap holders are suitable for use with crystals from

100 Kc. to 5,000 Kc. Above 5,000 Kc., the crystal performance generally becomes quite erratic and unsatisfactory. The range over which the crystal frequency can be adjusted by means of an air gap varies with frequency and is somewhat dependent upon the amount of circuit capacity in parallel with it. At 4,000 Kc. a typical crystal unit could be varied over a range of about 6 Kc. As the air gap is increased, the effective activity of the crystal is decreased and the RF output from the crystal oscillator will be reduced. If the air gap is made too large, the crystal will refuse to oscillate because of insufficient excitation.

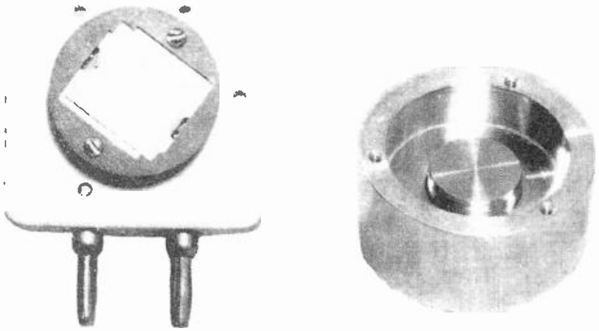


Fig.22 The crystal holder shown in Fig. 21 with the top removed to expose the crystal.

Another type of holder, namely the knife-edge mounting, is used for the bar type crystals in the frequency range from 16 to 275 Kc. Such mountings have few applications in commercial communications transmitters, being restricted mainly to experimental work and low standard-frequency oscillators.

The requirements which must be met by a well-designed crystal holder are:

1. The electrode surfaces must be perfectly parallel and must be free from oil and dirt.
2. The electrodes must be made from a metal that will not corrode.
3. When an air-gap is employed, means must be provided for clamping the movable electrode after the final adjustment is made.
4. Some type of construction is usually necessary to prevent lateral motion of the plate. This may be accomplished by enclosing the plate and electrodes in close-fitting cases of suitable insulating material.
5. The entire assembly must be dust-proof.

Foreign particles on a crystal can cause erratic performances or prohibit oscillation entirely. A crystal will not oscillate if there is any grease, wax, or similar substance on its faces. Such substances are removed during manufacture by a special de-greasing process, but may be deposited by handling the crystal afterwards.

Dust on the faces of the crystal is perhaps the outstanding cause for erratic performance. Dust will cause poor contact between the crystal and its electrodes and often will prevent oscillation entirely. Corona discharges can be developed when particles of dust separate the crystal from its electrodes since points of high potential will naturally appear at each particle. If the crystal is subjected to a rather high excitation, an RF arc can result, which, if sustained, may fracture the crystal.

The crystal holder should be thoroughly washed with carbon tetrachloride before inserting the crystal. Carbon tetrachloride (carbona) is the best cleansing agent; however, other solvents may be used providing they have no dissolved or suspended impurities. Clean soap and water is effective but requires greater care and a more vigorous scrubbing action is necessary. The crystal should be carefully washed and then dried with a clean, lint-free cloth. After cleaning, the fingers should not be allowed to come in contact with the major faces, as the oil from the fingers will offset the cleaning operation. Crystals can be handled by grasping them by the edges or by using a pair of tweezers. The same procedure should be followed when cleaning the electrodes in the crystal holder. However, they are not as fragile, so the operation is considerably simplified.

Care must be exercised when placing a crystal in its holder so as not to chip the corners or break it by placing it in such a position that it will bind in the holder. Where both of the crystal electrodes are separate from the holder assembly, the crystal is merely placed between its two electrodes and inserted in the holder cavity. The edge of the crystal should not protrude beyond the edge of the electrode as chipping might result. It should be noticed that one face of the electrode is finely finished while the other face is rough in comparison. It is necessary that the finely finished face be in contact with the crystal plate.

In the pressure-type crystal mounting, the spring pressure is adjustable by bending the spring until the desired tension is obtained. The optimum pressure should be determined by experimentation for optimum crystal performance.

13. CRYSTAL OVENS AND THERMOSTATS. To maintain the crystal at a constant temperature demands that some means be provided for adding heat to the crystal. It is obviously impossible to maintain the crystal at a temperature at or below room temperature unless refrigeration is provided. This is neither necessary nor desirable. If the constant temperature is to be maintained solely by the addition of heat to the crystal, the fixed temperature must be considerably above room temperature. It is usually about 50° Centigrade.

Viewed as a whole, the crystal oven comprises a heat-insulated chamber or box in which the crystal and its holder rests, an electric heating element for adding heat to the chamber, and a very sensitive thermostat for maintaining the oven temperature at as constant a value as possible. The oven itself may consist of a metal box carefully lined with a heat-insulating material such as Celotex or balsa wood. The thermostat is a very sensitive mercury-filled thermometer into the stem of which there has been fused two wire electrodes separated a short distance from each other. By making the bulb of the thermometer large so that it contains a relatively large amount of mercury, and by using a very small bore in the stem, a small change in temperature will cause a large change in the height of the thread of mercury in the stem. When the thread of mercury makes contact with both of the electrodes in the stem, a circuit is completed from one electrode to the other, which causes current to flow through a relay. The energizing of this relay opens the circuit feeding the heating element and thereby allows the oven to cool.

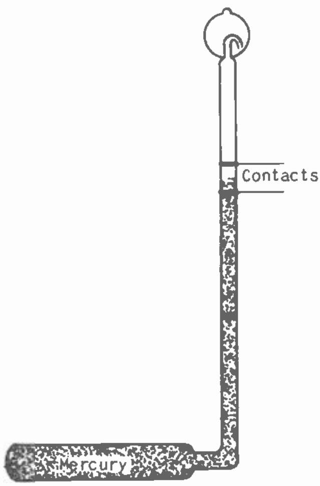


Fig.23 A mercury thermostat.

The appearance of this mercury thermostat is shown in Fig. 23. It consists of a tube bent at right angles. A reservoir of mercury is contained in a bulb at the bottom of the tube, and the temperature at which the mercury makes contact with the electrodes depends upon the amount of mercury in this bulb. The design of the thermostat is such that the temperature at which it operates may be varied in the following manner: There is another bulb at the top of the tube, and extending into this bulb is a small tube with a sharp bend at its extremity. If the thermostat operates at too low a temperature, there is too much mercury in the lower reservoir, and some mercury must be transferred to the upper bulb. This is done by applying heat to the reservoir until the thread of mercury is at the top of the small tube, and a few drops fall into the upper bulb. On the other hand, if the thermostat operates at too high a temperature, there is not enough mercury in the lower reservoir, and some is transferred from the upper bulb by again applying heat to the reservoir until the thread of mercury is at the end of the bent tube. At this point, the thermostat is tilted until the end of this tube dips into the pool of mercury in the upper bulb. The lower reservoir is now suddenly cooled, and this action sucks mercury from the upper bulb into the stem.

The heating element is usually composed of a resistor having a fairly large surface area. Thus, the heating is more uniform throughout the oven and there is less tendency to develop "hot spots". Often, the heater is composed of high-resistance wire woven back and forth in a thin sheet of asbestos.

Another type of thermostat frequently used to hold the temperature of a crystal oven constant is known as the "bi-metallic" thermostat. This device is made by welding together two dissimilar metal strips; the metals having different temperature-expansion coefficients. One of the metals is generally "invar" an alloy made of 36% nickel and 64% steel having a very low linear expansion coefficient. This means that invar can be heated considerably without appreciable increase in length. The other metal used in making a bi-metallic unit can be any hard durable metal; brass is generally employed. The expansion coefficient of brass is about 20 times that of invar.

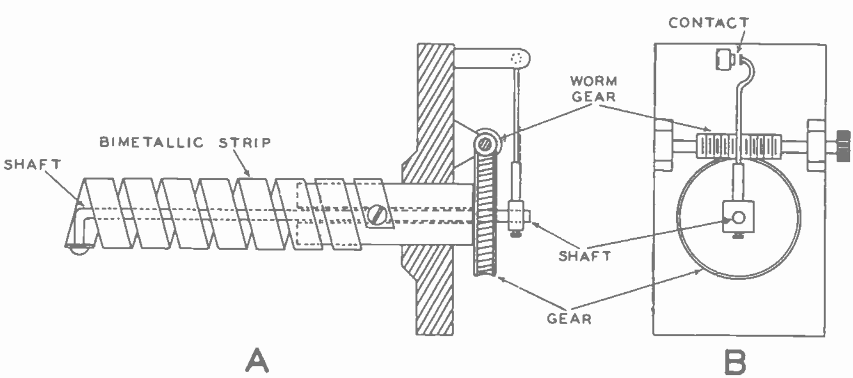


Fig.24 (A) Side view of bi-metallic thermostat. (B) End view of bi-metallic thermostat.

The bi-metallic thermostat is made by coiling the two narrow strips of metal (welded together) into a spiral shaped spring, one end being permanently fastened to the base of the unit. A diagram is shown in Fig. 24A. The metal which has the highest temperature coefficient (brass) is on the inside so that as the temperature rises, the expansion tends to uncoil the spring. As the spring moves, the shaft through the center is rotated. An armature is secured to the end of the shaft, so when the shaft turns in this direction, the contact points on the front of the thermostat are broken. The armature and contact points on the front of the unit are shown in Fig. 24B. Opening the contact points will in turn operate a relay which disconnects power from the heating resistance, so the oven in which the crystal is located cools. As soon as the temperature in the oven begins to fall, the bi-metallic strip will contract and twist the shaft in the opposite direction. At a critical temperature the armature will be moved sufficiently to again close the contacts and heat will be supplied to the oven. The oven temperature at which the contacts on the bi-metallic thermostat close can be adjusted by rotating the knob on the worm gear. It is possible to keep the temperature of the oven constant to a fraction of a degree by using this type of thermostat.

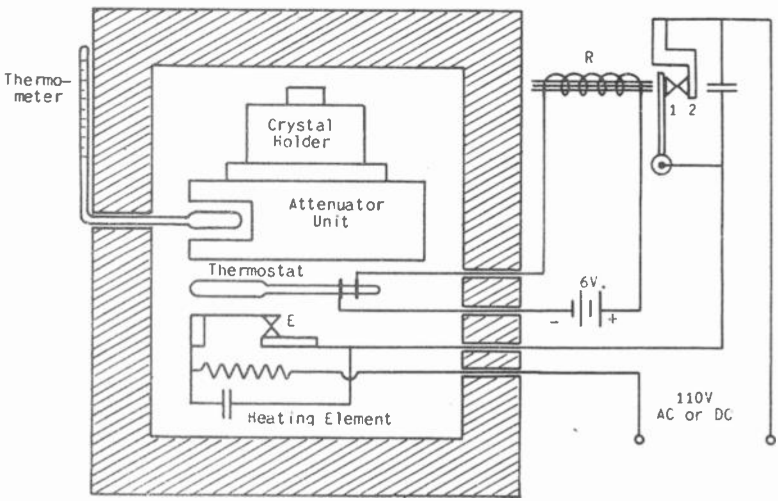


Fig.25 Crystal oven and control circuit. (Courtesy Commercial Radio Equipment Co.)

The two types of thermostats just discussed produce normally opposite results; the bi-metallic type closes its contacts when the temperature falls and the mercury type opens its contacts as the temperature decreases. The relay circuit with which the thermostat is associated must be designed in such a way that it will accommodate the method of thermostat operation. A mechanical relay circuit designed for a mercury thermostat is shown in Fig. 25. The operation of the circuit is as follows: Assuming that the temperature of the oven is below normal, the contacts of the mercury thermostat will be open. Current then flows from the 110-volt AC or DC source, through the heating element, through the contacts marked "E" (this is a protective bi-metallic thermostat), and through the relay contacts 1-2 to the other side of the voltage source. During this time the oven heats. When the temperature of the oven has been raised sufficiently to close the contacts of the mercury thermostat, current will flow from the 6-volt DC source through the relay coil R. Energizing the relay will open contacts 1-2 and the 110-volt heating circuit will be disconnected.

When the oven has cooled, the mercury thermostat contacts are opened, the relay R is de-energized and contacts 1-2 close to complete the 110-volt circuit to the heating resistance. The points of the bi-metallic thermostat (marked E) are for the purpose of preventing the oven from overheating in the event of failure of the mercury thermostat or relay. The bi-metallic thermostat is adjusted to open its contacts at a temperature slightly higher than the normal oven temperature.

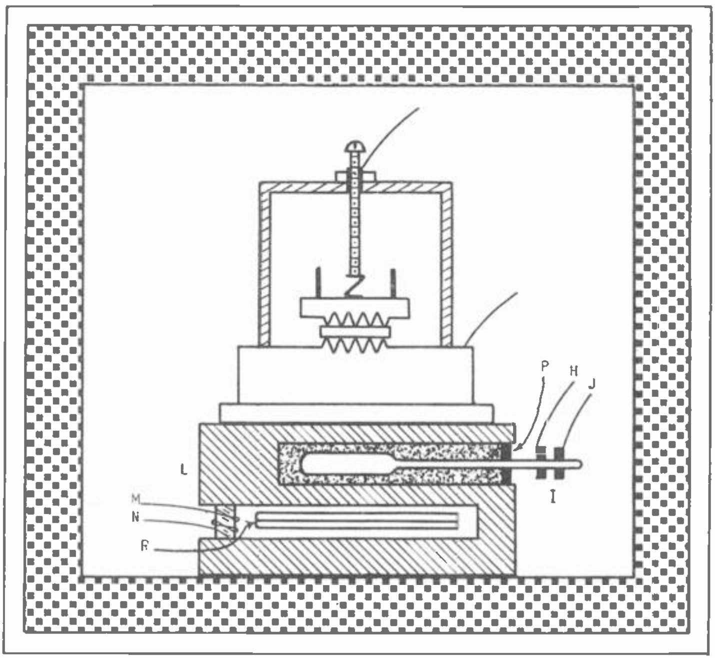


Fig.26 Western Electric Crystal Oven.

The condensers connected across the relay contacts are for the purpose of preventing arcing when the contacts are opened. Due to inductance in the circuit, when the points are opened, there is a tendency for an arc to be formed which would cause the faces of the contacts to be burned. In time the points are then liable to stick. The condenser connected across the contacts reduces the size of the arc and prevents excessive burning of the points.

An illustration of another type of crystal oven and its associated apparatus is shown in Fig. 26. This is the Western

Electric type used in many of the broadcast transmitters designed and built by this company. The crystal plate is clamped between two electrodes, and the entire crystal holder rests upon the heater unit L. This unit contains the thermostat I and the heating element R. Graphite powder P is used to hold the thermostat in place, and the leads H and J are connected to the mercury contacts within the stem. The heater unit L is a copper casting which acts as a thermal filter, allowing the flow of heat to the crystal to take place at a uniform rate. The heating element is carefully insulated from this block, and its external leads are shown at M and N. The entire unit is enclosed in a metal box designed to be mounted in the radio transmitter. All of the leads are brought through the box to a terminal strip where connections are made to the oscillator and to the temperature-control circuits.

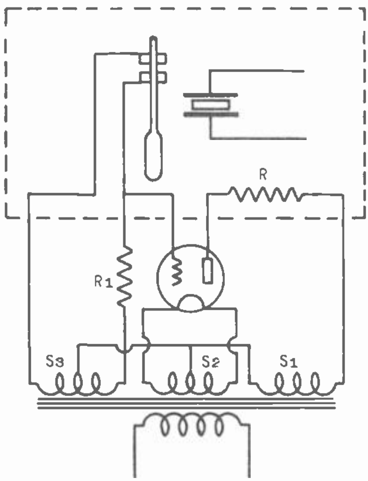


Fig.27 The control circuit used with the oven shown in Fig. 26.

The control circuit used with this particular type of crystal oven is shown in Fig. 27. The tube used in the control circuit is a gaseous-type rectifier which has a low internal voltage drop. It resembles the ordinary three-element tube in that a control grid is used. The control action does not regulate the heat to meet varying conditions, but is an intermittent process. Assume that the mercury column is low and that the circuit through the thermostat is open. With this condition, the grid of the vacuum tube is connected through resistor R_1 to what we shall call the positive side of the transformer S_3. The tube passes current only when the right hand side of the winding S_1 is positive. At the moment that a positive voltage is applied to the plate, the right hand end of winding S_3 is also positive. The center tap of S_3 is connected to the center tap of S_2. Therefore, when the plate of the tube is positive, the grid is also positive. With both plate and grid positive during every other alternation, plate current flows through the tube in the form of pulses lasting one-half cycle. It is noticed that the plate current of the tube flows through the resistor R, which is the heating element of the oven; thus heat is being added to the oven.

After the temperature of the oven becomes high enough to cause the mercury in the thermostat to rise and close the circuit through its contacts, an alternating current flows from one side of the transformer winding S_3, through the thermostat contacts, and through the resistor R_1 to the opposite side of the winding.

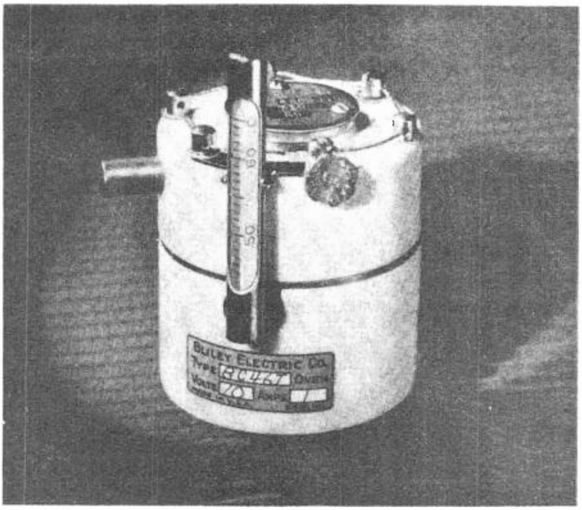


Fig.28 The Bliley BC46T complete crystal unit, including mounting, thermostat, and thermometer.

It is now evident that the grid of the tube is connected directly to the left hand end of the winding S_3. But when the right hand side of S_1 is positive so that the tube might draw current, the grid is connected to the left hand side of S_3, which is negative. The negative voltage during this alternation is sufficient to cut off the plate current. During the next alternation, the grid is positive, but the plate is negative, and therefore no plate current flows. Thus, with the thermostat contacts closed, the tube does not draw any plate current, no current flows through the heating element, and the oven cools. The action of the thermostat has been to change the phase relation of the voltage applied to the grid and thus indirectly prevent further heating of R_1.

After a time, the temperature of the oven will fall, due to the loss of heat through the walls. The mercury in the thermostat will drop, thereby opening the thermostat contacts. When this occurs, the tube is returned to its former condition, and again pulses of current flow through the heating element, causing it to add heat to the oven. It might be thought that the crystal would be alternately heated and cooled, but the thermal filter tends to iron out these changes and thereby provide a constant flow of heat to the crystal. The heating period and cooling period will, of course, depend on the design of the oven, but, on the average, it will require about 30 seconds for the oven to heat and approximately one minute for it to cool. The thermostat is sensitive enough to respond to changes of temperature of only .1° C.

In present day commercial ovens designed for broadcast station use, we find both mercury and bi-metallic thermostats. Each has its advantages and disadvantages. The principal advantage of the bi-metallic type is its ruggedness; however, it is not nearly so sensitive as a mercury thermostat. Even though the mercury type is more sensitive and accurate, it cannot be used to break circuits carrying a very high current. To obtain normal life from a mercury thermostat, the maximum current it should break is limited to 10 milliamperes. The contact points of a bi-metallic thermostat are exposed to open air and arcing occurs more readily than in the mercury type where the points are sealed in a glass tube. The points on a bi-metallic unit will require frequent inspection and cleaning whereas the contact points on a mercury thermostat require no attention.

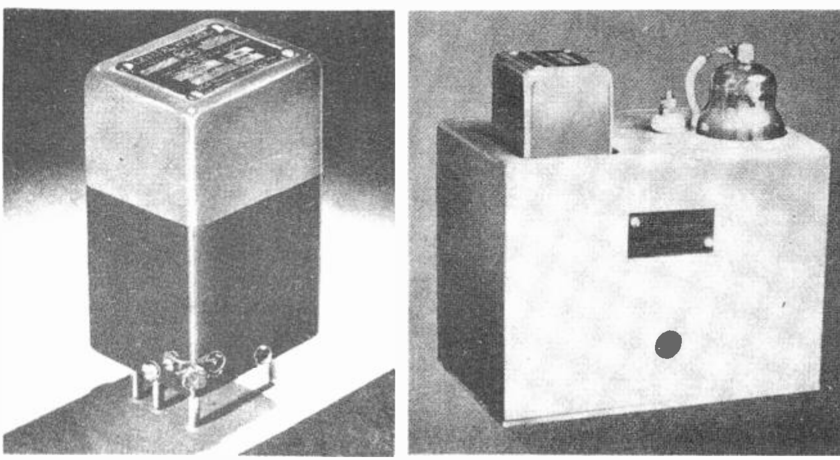


Fig.29 (A) The 129-B RCA Crystal Unit. (B) Photograph of the crystal unit placed in the complete oscillator.

Several manufacturers have recently developed temperature control units of small dimensions to meet the exacting requirements of broadcast services. With the advent of A-cut crystals, it is no longer necessary to control the crystal temperature as accurately as before when the X and Y-cut crystals were used exclusively. The Bliley Company has been successful in designing the BC46T unit to house the crystal, heating element, bi-metallic thermostat, and thermometer in a glazed Isolantite oven mounting capable of holding the crystal temperature constant to within 1° Centigrade over wide ranges of ambient temperatures. A photograph of the unit is shown in Fig. 28. A duralumin thermal-block which also serves as one crystal electrode, is temperature regulated by means of a heater element and a bi-metallic thermostat. The specially designed thermostat has large contact points which will safely carry the heater current of 1 ampere at 10 volts without the use of auxiliary relays. The BC46T unit has been approved by the Federal Communications Commission for use in all U. S. broadcast stations.

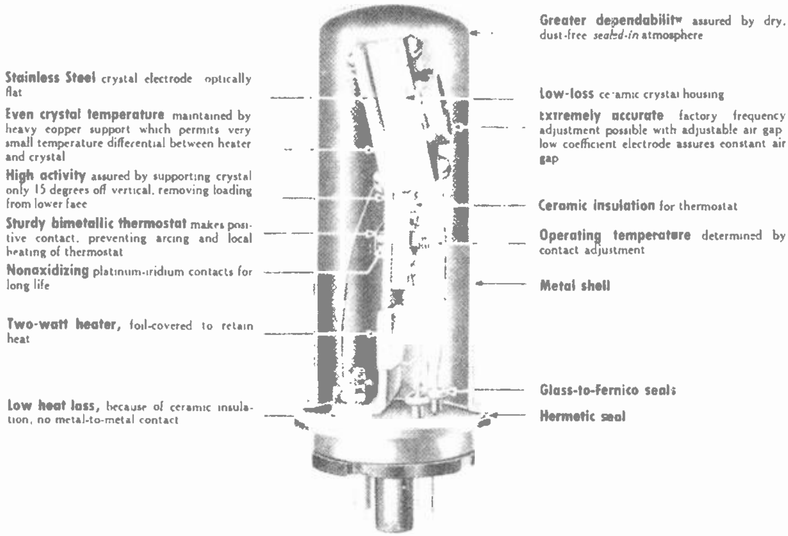


Fig.30 General Electric complete crystal oven. The metal shell is identical to that used for a 6L6 tube and the unit plugs into a conventional octal socket.

Since the advent of practical and consistent A-cut crystals. the trend in broadcast equipment is toward more miniature crystal ovens and oscillator units. Fig. 29A shows a photograph of the new RCA crystal holder and oven. This unit is plugged in a compact complete crystal oscillator unit as shown in Fig. 29B. Fig. 30 is a photograph of the new General Electric oven designed for a similar purpose.

14. CRYSTAL OSCILLATOR CIRCUITS. Crystal-controlled oscillators have as their origin some basic self-excited oscillator circuit. Accurate frequency control of the oscillator is accomplished by connecting a quartz crystal into the circuit in such a manner that the crystal itself becomes the frequency controlling element. In many cases, the crystal-controlled oscillator is merely an adaption of the well known tuned-grid, tuned-plate oscillator, with the quartz crystal substituted for the grid tank. A circuit diagram of such a triode crystal-controlled oscillator was given in a preceding section of this lesson. Another circuit which is equally satisfactory for operating a triode as a crystal-controlled oscillator is shown in Fig. 31.

Oscillator circuits in general have a remarkable self-regulating ability; that is, the circuit values can be varied over wide ranges and the oscillator will continue to function. As long as the component values employed do not prohibit oscillation entirely, the various currents and voltages will distribute themselves for the best performance under these conditions. Naturally, there are certain component values which will give optimum performance and efficiency; however, in most practical cases it is not necessary to pay particular attention to the size of the component parts.

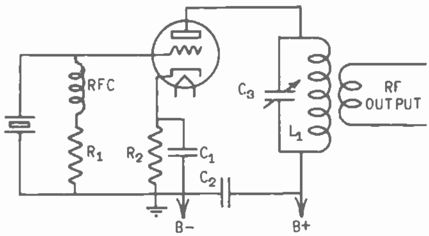


Fig.31 Typical triode crystal oscillator.

Since a crystal-controlled oscillator will also be self-regulating, care must be exercised in the design of the circuit so that excessive vibration amplitude of the quartz crystal will not cause it to be shattered. Excessive crystal excitation could easily result from the selection of certain component values; the oscillator circuit would function and an RF output would be obtained; however, the crystal may be damaged due to excessive excitation. For this reason, a careful choice of circuit values must be exercised, and also the RF output that can be obtained from a crystal-controlled oscillator is limited for the same reason in comparison to the self-excited oscillator.

In addition to factors given previously, the crystal excitation depends on: (1) the amplification factor of the tube; (2) the grid bias; (3) the DC operating potential; (4) the interelectrode feedback; and (5) the activity of the crystal. In order to secure a given power output from a crystal oscillator, the tube with the highest amplification factor will generally require the least crystal excitation (crystal current). This fact is immediately apparent when comparing the performance of pentode crystal oscillators with triode crystal oscillators. Tetrode tubes having a higher amplification factor than triodes will also require less excitation for a given power output. In addition, beam power tubes perform very satisfactorily as crystal oscillators because of their high amplification factor.

In all cases, the feedback through the interelectrode capacity from the plate circuit to the grid circuit in a crystal-controlled oscillator will be in phase with the grid circuit energy and will serve to increase the amplitude of oscillations. In other words, the interelectrode capacity feedback will be regenerative and not degenerative. In a given oscillator circuit, the

RF feedback is determined by the inter-electrode capacity and the RF voltage produced across the plate tank circuit. The LC ratio of the plate tank and the load on the plate tank circuit will both be effective in determining the RF voltage developed across it. A reasonably high LC ratio is desirable in pentode or tetrode oscillators, while a lower ratio is required for triode crystal oscillator tubes. Due to the higher inter-electrode plate-to-grid capacity of the triode and the lower amplification factor, the RF voltage across the plate tank must be limited so that the crystal excitation will not become excessive. If the excitation to the crystal (crystal current) becomes excessive, the amplitude of its vibration may be sufficient to damage the crystal.

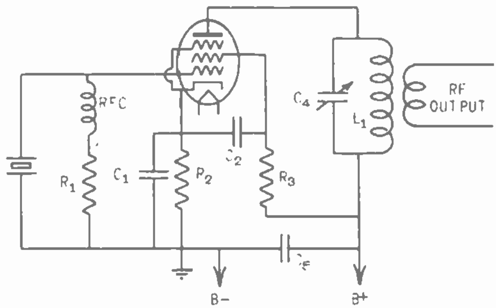


Fig.32 Pentode crystal oscillator.

A typical pentode crystal oscillator circuit is shown in Fig. 32. Due to the low inter-electrode capacity between plate and control grid in pentode tubes, at frequencies below 1000 Kc. the reactance of the inter-electrode capacity becomes too high to sustain oscillations. Some pentode tubes, even when operating at high frequencies, have such low internal capacity that a small amount of external feedback is recommended by the manufacturer. This external feedback can be supplied by use of a small padder condenser connected between plate and control grid or by the capacity that would be provided through short, twisted wires. If the crystal itself is reasonably active, most pentode tubes will oscillate without the addition of external capacity. In screen grid and pentode tubes, proper by-passing of the screen grid is very essential. If the size of the screen by-pass condenser C_2 in Fig. 32 is inadequate, the screen grid will assume an RF potential, thereby greatly increasing the feedback and possibly damaging the crystal unit.

Grid bias on a crystal-controlled oscillator is an important consideration. In general, the higher the bias, the greater will be the crystal current and the power output. However, if the bias is increased beyond a certain limit, the result will be more crystal current, but only a small gain in power output. The grid bias is generally obtained by the use of a grid leak resistor, cathode resistor, or a combination of both. When grid leak bias is employed, an increase in the size of the grid leak will result in an increased crystal current. When only grid leak bias is employed, the crystal will start oscillating under conditions of zero bias with a continually increasing bias as the excitation becomes greater. Crystal current will be greatest when the oscillator is not loaded because the plate tank voltage, grid current, and bias will be highest under this condition.

If only grid leak bias is employed (no cathode bias), the crystal oscillator may be difficult to get started, especially when a small grid leak is employed. If a little cathode bias is employed, it will be found that the crystal will start oscillating under more favorable conditions. The initial cathode bias has a tendency to increase the plate-to-grid feedback and bring about a condition in the grid circuit more conducive to the establishment of oscillations. Too much cathode bias, on the other hand, will have the opposite effect. The correct value of cathode resistor is generally between 200 and 500 ohms for both triodes and pentodes. The readiness with which the crystal oscillator starts oscillating is an important factor in the design of a telegraph transmitter in which the oscillator circuit is keyed.

The size of the grid leak resistor should generally not be higher than around 20,000 ohms for optimum crystal current and power output in triodes, tetrodes, and pentodes. If a low value grid leak resistor is used, an RF choke must be placed in series so that the small resistor will not load the crystal. The minimum bias on tetrodes and pentodes is generally secured with a grid leak in the vicinity of 10,000 ohms. On triodes, especially for high frequency operation, cathode bias alone may be used, and the grid leak can be replaced with a good RF choke.

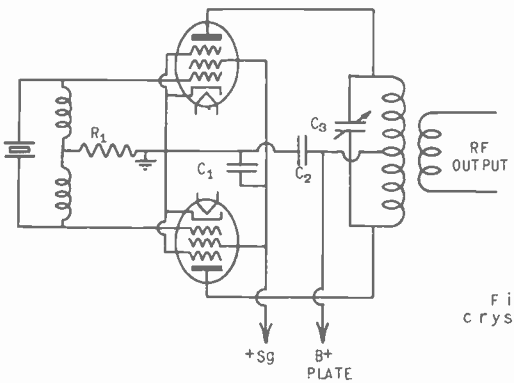


Fig.33 Push-pull pentode crystal oscillator.

Referring to the triode crystal oscillator circuit in Fig. 31, the size of the cathode resistor R_2 should be in the vicinity of 350 ohms and R_1 may be from 0 to 20,000 ohms. The DC plate voltage should be 250 volts or less; a higher plate voltage will cause more inter-electrode feedback from plate to grid circuit and the crystal is apt to be damaged. The LC ratio of the plate tank circuit in Fig. 31 should be comparatively low. The power output of most triode crystal oscillators is relatively low under conditions of safe crystal current. The size of the by-pass condensers C_1 and C_2 should be such that their reactance at the frequency of oscillation is very low.

In selecting the component parts for the pentode oscillator circuit shown in Fig. 32, for most tubes the optimum sizes for R_1 and R_2 will be 20,000 ohms and 350 ohms respectively. The screen voltage has quite an influence on the crystal current and for most tubes should be kept less than 250 volts. Proper by-passing from screen to cathode with a mica condenser is very important. The condenser should be placed directly at the tube socket so that the leads will be as short as possible.

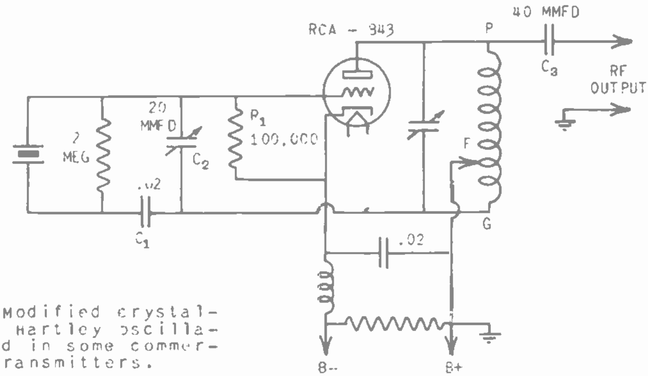


Fig. 34 Modified crystal-controlled Hartley oscillator as used in some commercial RCA transmitters.

Pentode and tetrode tubes should be operated in much the same manner, since the essential difference between these tubes simply lies in the method of suppressing the effects of secondary emission. The high-mu tubes will provide a greater power output for a given crystal current and due to the shielding action of the screen grid, the frequency stability is slightly improved compared to triodes. With the screen at RF ground potential, the plate load impedance and the DC plate voltage may be changed over small ranges without affecting the frequency to any great extent.

A push-pull pentode oscillator circuit is shown in Fig. 33. Oscillators of this type are advantageous only in that the output circuit is balanced and even harmonics are cancelled out. By using tubes requiring a low grid excitation, it is possible to obtain a substantial increase in power output with the push-pull arrangement. Since two tubes in push-pull require approximately twice as much driving power as a single tube of the same type, it is necessary that the crystal vibrate more intensely to drive both tubes to full output. With most tubes, it is necessary to reduce the operating voltages so that the crystal current will be within safe limits under all conditions of performance. The final result will generally be a small increase of output power over the use of a single tube oscillator.

A circuit diagram of the crystal oscillator used in some of the RCA 1 Kw. transmitters is shown in Fig. 34. This is a revised

Hartley oscillator adapted for crystal control.

Referring to Fig. 35, we see a conventional diagram of a series-fed, self-excited Hartley oscillator. One difference between this circuit and those previously shown is that the plate of the tube is operated at DC ground potential and the cathode is at a high negative potential. As long as the cathode of the tube is at a high negative potential with respect to the plate, the operation of the circuit will be the same as though the cathode were at DC ground and the plate positive with respect to ground. In

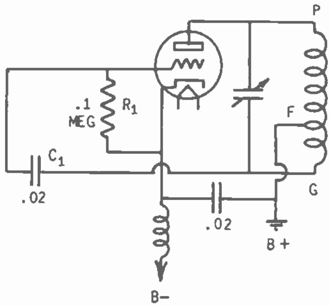


Fig.35 Conventional series-fed Hartley oscillator. Compare with Fig. 34.

Fig. 35, the RF voltages developed across the tank circuit from P to F will serve to vary the instantaneous plate potential, and the RF voltages developed between F and G will be fed through the grid capacity C_1 to the grid of the self-excited oscillator tube. R_1 serves as the grid leak.

In the revised crystal-controlled Hartley oscillator shown in Fig. 34, the crystal is inserted in the grid circuit and the RF voltages developed when it is set into vibration serve as the grid excitation. The RF voltages developed across F and G of the oscillator circuit serve merely to excite the crystal. The crystal's vibration, in turn, develops RF voltages which are applied to the grid of the tube. Thus, the crystal is the frequency-controlling element and the RF voltages developed across F-G serve

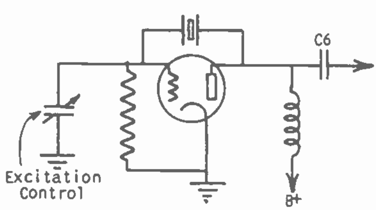
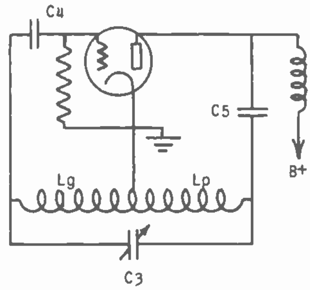


Fig.36 The Pierce oscillator.

as the crystal excitation. It is important that the position of the filament tap F be such that the excitation to the crystal be kept low. If the RF voltage across F-G is too high, the crystal will lose control of the oscillations and the circuit will perform as a self-excited oscillator.

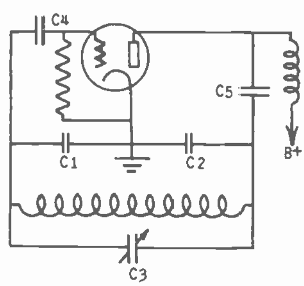
The 20 micro-microfarad condenser C_2 in Fig. 34 is directly in parallel with the crystal. Thus, the frequency of the crystal can be shifted over a range of a few cycles by variation of the capacity of C_2. This is advantageous in commercial broadcast transmitters because it is frequently necessary to set the crystal oscillator exactly on an assigned frequency. The RF output from the oscillator is delivered to the grid circuit of the first buffer amplifier through the 40 mmfd. condenser C_3. By using a very small coupling condenser to the grid of the first amplifier, the load on the crystal oscillator is kept light and the frequency stability is improved.

Fig.37 A shunt-fed Hartley Oscillator.



A type of crystal oscillator circuit which has but recently come into use in broadcast transmitters is the Pierce oscillator, a diagram of which is shown in Fig. 36. A peculiarity which is at once evident is that the crystal is connected between the grid and the plate. Just how this oscillator is able to function will now be explained.

Fig.38 A modification of the Colpitts Oscillator.



Let us first consider Fig. 37. This figure illustrates an oscillator circuit, which, upon inspection, is seen to be a shunt-fed Hartley. It is true that it is drawn in a somewhat different manner than those with which you may be familiar, but it is a Hartley oscillator nevertheless. The tank coil is effectively divided into two parts by the connection of the filament tap. That part of the coil designated as Lg provides grid excitation, whereas Lp is the load into which the tube works. It is necessary that the tank be divided into two parts in this manner so that the alternating grid and plate voltages will be 180° out of phase.

Since the filament center tap is always grounded, grounding some point in the tank circuit, either directly or through an RF by-pass condenser, will divide the tank circuit effectively into two parts. In the Colpitts oscillator, as you know, no part of the tank coil itself is grounded, yet the tank circuit is effectively divided into two parts by grounding the junction of the two condensers which compose the tank capacity.

It would be possible, for example, to construct an oscillator circuit such as the one illustrated in Fig. 38. In this case, two small condensers are connected in series across the tank circuit and their mid-point is grounded. Thus, the voltage drop across C_1 would be the grid excitation voltage, whereas the drop across C_2 would provide the feedback voltage. These two small condensers would add to the total capacity, and the effective capacity of the tank circuit would be the combination of C_1, C_2, and the main tuning condenser C_3. These two condensers could be of any convenient size as long as they were small compared to the tuning condenser.

As curious as it may seem, this circuit would probably oscillate, even though these two condensers were omitted. For example, consider Fig. 39 which illustrates the same circuit without the two small condensers. It is apparent that the combination of the grid-filament capacity and the plate-filament capacity would form a voltage-dividing circuit which would, in effect, ground one point of the tank circuit. In this case the voltage drop across the grid-filament capacity provides the grid excitation, whereas the drop across the plate-filament capacity is the feedback voltage. Notice that the total tank capacity consists of the tuning condenser C_3, and the series combination of C_4, Cgf, Cpg, and C_5.

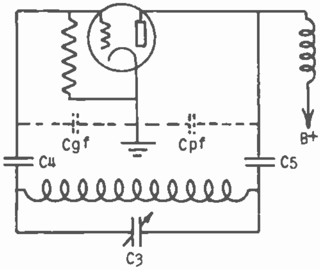
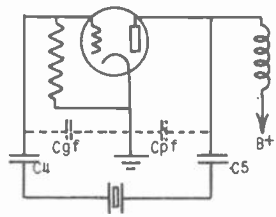


Fig.39 A type of oscillator in which the inter-electrode capacities form the voltage-dividing circuit across the tank.

A piezo-electric crystal has the same properties as a tuned tank circuit, and may be used to replace the tank circuit in this oscillator; that is, a quartz crystal will take the place of L_1C_3. The oscillator then appears as shown in Fig. 40. Again the combination of the inter-electrode capacities form the voltage-dividing circuit so that a part of the voltage across the crystal provides grid excitation, and the remainder is the feedback voltage. Since a direct current cannot pass through the crystal, there is now no real need for the blocking condenser C_5, and it is ordinarily omitted. Similarly, the grid blocking condenser C_4 may be left out of the circuit without causing any appreciable change in its operation.

In this oscillator, the amount of grid excitation will depend upon the relative sizes of the grid-filament and the plate-filament capacities. To provide better control of the excitation voltage, it is customary to connect a small trimmer condenser between the grid and ground. By increasing the capacity of this trimmer condenser, the total capacity between the grid and filament is made larger, its capacitive reactance becomes less, and there is a smaller voltage drop across this capacity thereby reducing the grid excitation voltage. Conversely, reducing the capacity of this trimmer condenser will increase the grid excitation. When all of these changes have been made, the Pierce oscillator appears as shown in Fig. 36.

Fig. 40 A crystal replaces the tuned circuit L_1C_3 in Fig. 38, and the Pierce oscillator results.



There is an RF voltage between the plate of the tube and ground, and this voltage is transferred through the coupling condenser C_6 to the grid of the buffer stage. Notice that there is no tuned circuit associated with the oscillator. Thus, there is nothing to be tuned and the oscillator begins to function as soon as the plate voltage is applied.

The only disadvantage of the Pierce oscillator compared to other crystal oscillator circuits is the fact that it has a comparatively low output. This is of no particular disadvantage in broadcast transmitters, since the primary purpose of the oscillator is to supply a radio frequency voltage having excellent frequency stability, and its low output may be built up with radio frequency power amplifiers.

On the other hand, the Pierce oscillator has several advantages which recommend its use. First, a minimum amount of equipment is needed; since no tuned circuits are used, temperature changes which might alter the frequency of the oscillator will have less effect on the Pierce circuit than they would upon other circuits using more equipment. Second, the Pierce circuit is by far the most active of any of the common types of crystal oscillator circuits. This may be proved by using inferior grades of X-cut crystals. Many crystals which are so poor that they cannot be made to oscillate in any of the common types of circuits will function satisfactorily in the Pierce oscillator. The small trimmer condenser should have a maximum capacity of approximately 150 mmfds., and the RF choke used in the plate circuit should offer a high impedance to the oscillation frequency.

In addition to properly adjusting the feedback capacity C_1, it is necessary to operate the Pierce oscillator circuit at a low plate voltage to limit the RF voltage developed across the plate circuit. A DC plate potential of approximately 150 volts is considered maximum in order that the crystal will not be subjected to excessive strains.

The plate circuit of a Pierce oscillator circuit must present a capacitive reactance to satisfy the conditions for oscillation. A capacitive reactance may be obtained with a detuned tank circuit, an RF choke having a resonant frequency lower than that of the crystal, or a resistance. A pure resistance, of course, has no reactance, and by itself would not satisfy the conditions for oscillation. However, the internal plate-to-filament capacity of the tube is in parallel with the resistance and this provides the necessary capacitive reactance if the plate resistor is not too large. If an RF choke is used, for high frequencies, a choke of 2½ millihenries is generally employed, while a considerably larger inductance is necessary at lower frequencies. Best performance of a Pierce oscillator circuit is generally secured with a combination of grid leak and cathode bias. The grid leak should be limited to around 50,000 ohms if the oscillator is operating above 1500 Kc., while 100,000 ohms is better at the lower frequencies. About 250 ohms cathode bias will generally be sufficient in either a triode or a pentode circuit.

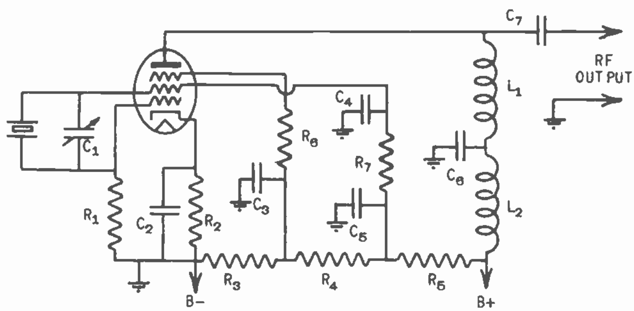


Fig.41 RCA Electron-coupled Pierce Oscillator.

The RCA adaptation of the Pierce oscillator is the electron-coupled Pierce, a diagram of which appears in Fig. 41. In this case, the crystal is connected between the control grid and the screen grid. As in all electron-coupled oscillators, the screen functions as the virtual or working anode of the oscillator section of the tube. The potential of the plate is somewhat greater than that of the screen, and the larger part of the space current is attracted to the more positive plate, very little being diverted to the other positive electrodes within the tube. In Fig. 41 a pentode tube is being used, and to make it function somewhat more efficiently, a small positive voltage is applied to the suppressor grid. In addition, this tube (an RCA 802) has an internal shield which is connected to the suppressor. The plate voltage of this particular circuit is 340 volts; the screen voltage 175 volts, and the plate current 30 Ma.

15. THE MULTIVIBRATOR CIRCUIT. A multivibrator is an oscillating circuit having special advantages applicable in frequency measurement work. It is, in effect, a two-stage, resistance-coupled amplifier with the output circuit coupled back to the input through a proper size capacity. This feedback from the output to the input causes the amplifier to oscillate (motor-boat) at a frequency determined by the time constants of the resistance-capacity combinations in the circuit. Since the oscillations are produced by charging and discharging of the condensers through the resistors, the waveform of the oscillatory current and the waveform of the output voltage are irregular and distorted. The output will be very rich in harmonics; in fact, harmonics up to the 100th can be detected and utilized.

When the multivibrator is oscillating by itself, it possesses no particular advantage since the circuit is relatively unstable. However, if a low voltage from a stable oscillator is injected into the circuit, the conditions are changed considerably. If the frequency of the multivibrator oscillator is made nearly equal to the frequency of the injected voltage, the injected voltage will assume control and cause stable performance of the circuit. The multivibrator frequency is then dependent upon the controlling voltage and is independent of small changes which might occur in the circuit itself. When stabilized in this manner by an external controlling frequency, the multivibrator becomes a very useful instrument and serves as an excellent harmonic generator.

One of the most important properties of the multivibrator is that synchronization of its frequency can also be produced when the frequency of the controlling voltage is harmonically related to the natural frequency of the multivibrator circuit. In this way, the multivibrator may be employed for frequency division or multiplication. Practical application, however, is usually limited to frequency division, since there are more preferable frequency multiplying circuits available. Thus, it would be possible to synchronize the operation of a 10 Kc. multivibrator with a stable crystal-controlled 50 Kc. oscillator. Because of the stability of a multivibrator when synchronized and the fact that its frequency can be determined by the harmonically-related controlling voltage, the multivibrator is widely used to produce a series of standard frequencies from a single, crystal-controlled standard oscillator. Frequency division can be accomplished when the ratio of standard to multivibrator frequency is as high as 40 to 1, but for best stability, it is limited to approximately 10 to 1.

The maximum operating frequency of a multivibrator circuit is in the vicinity of 100 Kc., thus, it is not truly an RF oscillator. However, if a multivibrator is operating on a fundamental frequency of 10 Kc., the higher order of harmonics produced will fall in the radio frequency spectrum. Thus, the 100th harmonic of a 10 Kc. multivibrator will be 1000 Kc., or 1 megacycle. The consecutive harmonics from the second to the 100th will give a usable output signal every 10 Kc. between 20 Kc. and 1 megacycle.

Fig. 42 is a representative multivibrator circuit. Small tubes such as the 56, 6C5, etc., may be used in a circuit of this type, or the newer twin-triode tubes, such as 6N7, 6C8, etc., may be employed. In Fig. 42, the resistors R_1 and R_2 and the coupling condensers C_1 and C_2 are the major frequency-determining components. There is no simple formula that will give the exact frequency of motorboating (oscillating), but for approximation:

$$F = \frac{1000}{R_1C_1 + R_2C_2}$$

Where: F is the frequency in Kc.
R is the resistance in ohms
C is the capacity in microfarads.

Generally R_1 and R_2 are made the same size, and C_1 is made equal to C_2) In Fig. 42, R_1 is shown as a potentiometer and fixed resistor in series. The potentiometer will generally have a value in the vicinity of 5000 ohms. It offers a simple means of injecting the synchronizing or controlling voltage and regulating its amplitude. When substituting in the formula above, R_1 should be the total of the potentiometer and fixed resistor in series.

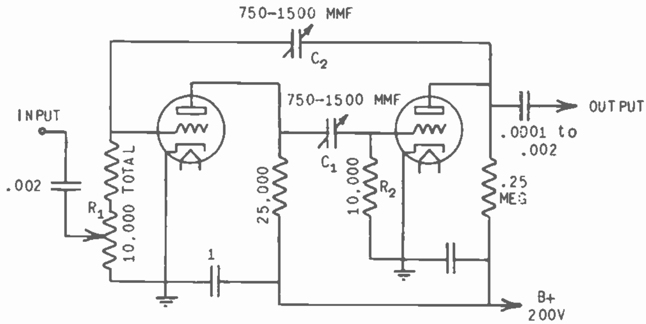


Fig.42 Multivibrator circuit with provisions for injecting synchronizing voltage. Used for dividing frequency.

Either the grid coupling condensers or the grid leak resistors should be made variable so that the multivibrator can be adjusted to the correct frequency. The frequency desired will be a sub-multiple of the injected controlling frequency, generally the 10th. When adjusting the multivibrator, the input controlling signal is reduced to zero and the variable condensers C_1 and C_2 (or R_1 and R_2) are varied simultaneously until the fundamental frequency of the multivibrator's operation is very close to the desired value. As the capacities or resistors are increased, the frequency will decrease, and vice versa. When the multivibrator is set close to the desired frequency, a small voltage from a crystal-controlled higher frequency oscillator may be injected into the circuit. As the input voltage is increased, a point will be noticed where the multivibrator becomes very stable and the output voltage is resolved into a definite frequency. For best performance, the input should be increased just slightly above this point. An excessive increase in synchronizing voltage will cause the multivibrator to jump to another frequency. When the crystal oscillator is assuming full control, the variable circuit elements can be changed over an appreciable range without changing the output frequency of the multivibrator.

For a good, low-frequency standard oscillator to serve as the controlling signal for the multivibrator, a circuit such as is shown in Fig. 13 may be employed. Thus, if the crystal-controlled oscillator of Fig. 13 is operating at a frequency of 50 Kc. and the multivibrator circuit shown in Fig. 42 is synchronized on the fifth sub-multiple of 50 Kc. or 10 Kc., accurate, crystal controlled frequencies will be obtained every 10 Kc. from 10 Kc. to 1 megacycle out of the multivibrator circuit. For frequencies above 1 megacycle, the harmonic output of the standard, crystal-controlled 50 Kc. oscillator may be employed.

With such a great number of standard frequencies available it is possible to determine the frequency of an unknown signal by the beat frequency method. The multivibrator finds its greatest application in the measurement of unknown frequencies by beating the unknown against a known harmonic of the multivibrator. More detailed information on the exact procedure will be given in a future lesson.

FAHRENHEIT TO CENTIGRADE CONVERSION TABLE.

To change a temperature, given in the Fahrenheit scale, tc the Centigrade system, the following formula is used:

$$C = \frac{5}{9} \times (F - 32)$$

Fahrenheit	Centigrade	Fahrenheit	Centigrade
0°	- 17.78°	80°	26.67°
5	- 15.00	85	29.44
10	- 12.22	90	32.22
15	- 9.44	95	35.00
20	- 6.67	100	37.78
25	- 3.89	105	40.56
30	- 1.11	110	43.33
32	0	115	46.11
35	+ 1.67	120	48.89
40	4.44	125	51.67
45	7.22	130	54.44
50	10.00	135	57.22
55	12.78	140	60.00
60	15.56	145	62.78
65	18.33	150	65.56
70	21.11	200	93.99
75	23.89	212	100.00

CENTIGRADE TO FAHRENHEIT CONVERSION TABLE

To change a temperature, given in the Centigrade scale, to the Fahrenheit system, the following formula is used:

$$F = \left(\frac{9}{5} \times C\right) + 32$$

Centigrade	Fahrenheit	Centigrade	Fahrenheit
- 20°	-4.0°	45°	113.0°
- 15	+5.0	50	122.0
- 10	14.0	55	131.0
- 5	23.0	60	140.0
0	32.0	65	149.0
+ 5	41.0	70	158.0
10	50.0	75	167.0
15	59.0	80	176.0
20	68.0	85	185.0
25	77.0	90	194.0
30	86.0	95	203.0
35	95.0	100	212.0
40	104.0		

Notes

(These extra pages are provided for your use in taking special notes)

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MIDLAND RADIO AND TELEVISION SCHOOLS INC.

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

BUFFER AMPLIFIERS

NEUTRALIZATION

LESSON NO. 3

TO BAG DUCKS

.......HAVE PLENTY OF AMMUNITION

Included in the famous tales of fiction's history are those involving that great truth-stretcher, Baron Munchausen. And among the tales of Munchausen, the one which stands out is that of his remarkable duck-hunting trip.

It seems that the Baron went hunting. He concealed himself in the reeds, and waited for the birds to appear. But to his embarassment he found that he had no ammunition, and just at the moment of his discovery, a nice flock of fat ducks settled to the surface of the water before his eyes!

After a moment of thinking, he decided that he would wait for the right moment, and then use a valuable pearl from his ring in place of a bullet. So he carefully tamped down a light quantity of powder, and then placed the pearl in the barrel. After a long wait, the ducks drifted into a straight line. Lining up his sights, he shot. The pearl bullet neatly pierced every head in the line, and the Baron extracted the pearl from the head of the last duck!

There are a lot of fellows who are waiting, with only one little bit of "ammunition" for the right moment when all conditions are exactly right. "Then", they say, "I will make a fine clean-up and be on easy street".

But those fellows are fooling only themselves. They are very busy in the meanwhile, growing long grey beards. The chances of finding everything just exactly right are mighty slim!

How much better it would be for them, like yourself, to have plenty of "ammunition".

You are rapidly coming to the point where you will have a wealth of radio knowledge. You will soon have enough "Ammunition" so that you will not have to wait for just the right opportunity, or all conditions to be just exactly right.

Study each lesson thoroughly; catch each fact and file it in your mind. Then, when you need the information you will have it, and you can "bring home your bag of fine fat birds".

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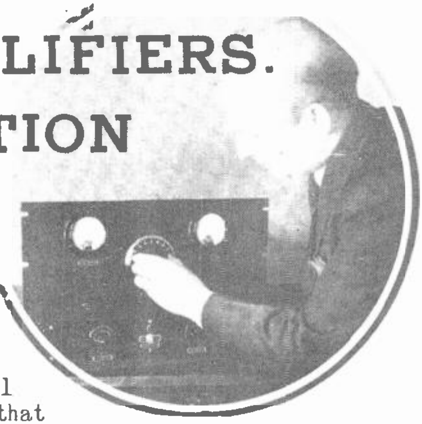
Lesson Three

BUFFER AMPLIFIERS. NEUTRALIZATION



"All of the transmitters used during the earlier days of broadcasting consisted of high-powered, self-excited oscillators, connected directly to an antenna system. It was not until the perfection of neutralization that high-powered, high-efficiency transmitters were made successful. Neutralization plays such an important part in the present day transmitter that it is advisable for the broadcast engineer to be thoroughly acquainted with the fundamental theory of operation and the practical methods of securing it.

"Crystal-controlled transmitters would be practically impossible without buffer amplifiers, so a thorough study of this subject is also quite important in the study of modern broadcast transmitters."

1. GENERAL TRANSMITTER CONSIDERATIONS. In the two preceding lessons, the subject of oscillators was thoroughly discussed. The oscillator is, of course, the heart of the transmitter; it is the device which produces the high-frequency carrier wave which is radiated from the antenna. If voice and music are to be communicated, this carrier wave must have its amplitude varied in direct accordance with the changes of the audio currents generated in the microphone circuit. This operation is called "modulation" and the device that effects it is known as the "modulator".

The simplest possible type of voice-transmitter consists of an oscillator connected to a modulator. The modulated output of the oscillator feeds an antenna which radiates the intelligence in the form of electromagnetic waves. The simplest transmitter is shown in block diagram form in Fig. 1. Except for portable ultra-high-frequency transmitters, this system is obsolete. To obtain a reasonable power output at ordinary broadcast frequencies would require that the oscillator be self-excited. The comparatively poor frequency stability of the self-excited oscillator is well known, and this fact itself would prevent its use in broadcast transmitters owing to the rigid requirements of frequency maintenance specified by the Federal Communications Commission.

With the fact definitely determined that a broadcast transmitter must employ a crystal-controlled oscillator to secure the required frequency stability, the next problem is to discover the most desirable stage in which to modulate the transmitter. Since crystal-controlled oscillators with good frequency stability are incapable of producing more than 1 or 2 watts of R.F. power, it is essential that they be followed by several R.F. power amplifiers,

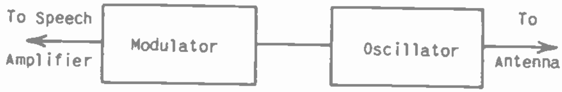


Fig.1 Block diagram of a simple radio-telephone transmitter.

whose purpose is to build up the weak output of the oscillator until it is capable of causing appreciable radiation from the transmitting antenna. It is theoretically possible to modulate the crystal oscillator; that is, cause the amplitude of its oscillations to vary at an audio frequency rate. It is also possible to amplify the unmodulated output of the oscillator with several stages of R.F. power amplifiers, and then modulate the final amplifier stage which feeds the antenna. Between these two extremes, there is the possibility of amplifying the unmodulated carrier through one or two stages, then applying the modulation, and finally amplifying the modulated carrier with one or more stages.

The system of modulating the crystal oscillator is never used. In Lesson 5 of this Unit, it will be shown that to cause 100% modulation requires that the DC plate voltage applied to the modulated stage be varied from twice its normal value to zero. Also, it was learned in Lesson 1 of this Unit, that a change in the plate-supply voltage was one of the causes of frequency variation. Therefore, the attempt to modulate the crystal oscillator would result in very poor frequency stability. Either of the other two systems of modulating a transmitter may be employed; a complete discussion of the relative merits of each will be given in a later lesson.

For the same reason, the stage following a crystal oscillator is never modulated, since changing the plate voltage of this stage would change the load on the oscillator with a resulting variation in frequency. There is at least one stage and sometimes more between the oscillator and the modulated stage. Fig. 2 shows an im-

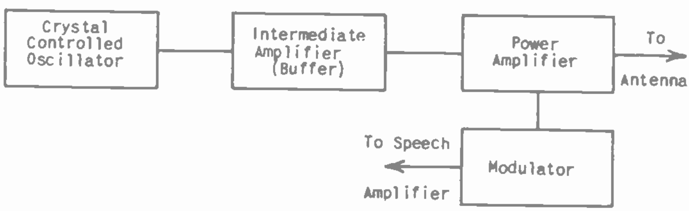


Fig.2 An improved radio-telephone transmitter.

proved type of broadcast transmitter illustrated in block diagram form. It consists of a crystal-controlled oscillator followed by an intermediate amplifier which feeds the power amplifier. The power amplifier is modulated and its output is connected to the transmitting antenna.

A more modern radio broadcast transmitter is illustrated in Fig. 3. The oscillator is crystal-controlled and associated with the oscillator is the temperature-control equipment. The oscillator feeds a buffer amplifier, which in turn is connected to a second amplifier. The output of this second amplifier excites the modulated stage, and following this stage are two power amplifiers

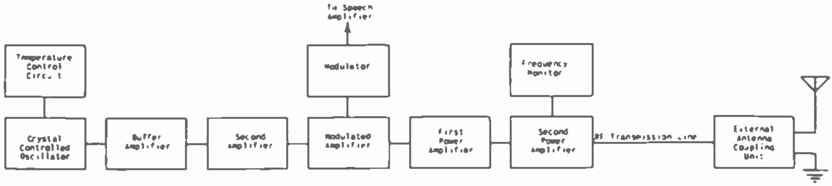


Fig.3 Block diagram of a modern broadcast transmitter.

which amplify the modulated carrier. The speech amplifier builds up the audio frequencies until they are strong enough to cause 100% modulation of the carrier frequency. The frequency monitor is a separate crystal oscillator used to check the transmission frequency. The R.F. transmission line connects the output of the final stage to the antenna. In high-powered transmitters, it is not desirable to have the antenna tower too close to the transmitter house. The external antenna coupling unit is used to connect the R.F. transmission line to the antenna.

2. **THE BUFFER AMPLIFIER.** The stage following the crystal oscillator in any transmitter is called the "buffer amplifier". Its purpose is to prevent the variations produced in the modulated amplifier from reacting on the oscillator and thereby causing a variation of the oscillation frequency. It serves the additional purpose of building up the relatively weak oscillations produced by the oscillator to the point where they are able to excite the modulated stage. Nearly all modern transmitters employ at least two amplifiers between the oscillator and the modulated stage. Both of these amplifiers are properly called buffers, although in some diagrams, the second one is not so labeled.

The diagram shown in Fig. 4 illustrates the schematic of a crystal oscillator feeding a buffer amplifier. The oscillating current flowing in the plate tank circuit of the oscillator sets up R.F. voltages across the tank. The condenser C_3 connected between the lower end of the tank and ground has a very low capacitive reactance to the oscillation frequency; hence, the bottom of the tank is effectively grounded with respect to R.F. By connecting a blocking condenser between the tank and the grid of the buffer to prevent the DC plate voltage from being applied to the grid, the R.F. voltages present across the plate tank circuit may be used to excite the buffer stage. This blocking condenser is the one marked C_4 in the diagram; its capacitive reactance to the oscillation frequency should not be very great, in order that the R.F. voltage present across the tank will not be dropped across it, but will be available for exciting the grid of the buffer.

Bias voltage must be applied to the grid of the buffer to limit its plate current to a safe value. In this diagram, the bias is supplied by a battery, with its negative terminal connected toward the grid, and its positive terminal grounded. Since the center tap

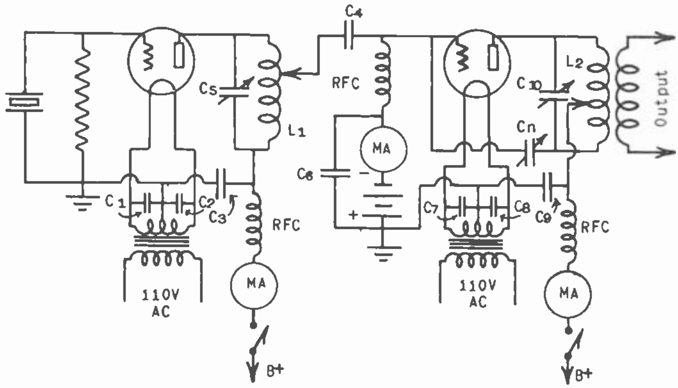


Fig.4 A crystal oscillator capacitively coupled to a buffer amplifier.

of the filament is also connected to ground, this battery causes the grid to be negative with respect to the center of the filament. This battery does not offer an appreciable impedance to the R.F. voltage applied to the grid, and to prevent this voltage from being shorted to ground by the low impedance of the battery, an R.F. choke is connected between the negative terminal of the battery and the grid. This choke should have a high inductive reactance to the transmission frequency, and, at the same time, a very low DC resistance. The buffer amplifier is usually biased to at least cut-off, and sometimes as much as one and one-half times cut-off.

Filament by-pass condensers are used on the buffer amplifier as well as the oscillator; they are shown as C_7 and C_8 in the diagram. They provide a low-impedance path around the filament secondary windings for the R.F. component of the plate current. The load impedance of the buffer tube is a parallel tuned circuit, tuned to the fundamental frequency of the oscillator. The condenser C_n is the neutralizing condenser, about which more will be said in a later section of this lesson. C_3 and C_9 are plate by-pass condensers; they provide a low-impedance path around the power supply for the R.F. variations of the plate current.

This method of coupling the oscillator to the buffer is called "capacitive coupling." When the oscillator uses a shunt-fed plate circuit, the connection to the grid of the buffer is made directly from the oscillator tank; no coupling condenser is needed since the plate tank of the oscillator is isolated from the DC plate-supply voltage by the condenser C_4 shown in Fig. 5. Notice that this circuit requires an R.F. choke in the plate circuit of the oscillator, but none is needed in the grid circuit of the buffer. The grid bias battery is connected to the bottom end of the oscillator tank circuit. In order to prevent the shorting of this C battery, the condenser C_3 is provided; it is connected between the lower end of the oscillator tank and ground. It is necessary to ground the

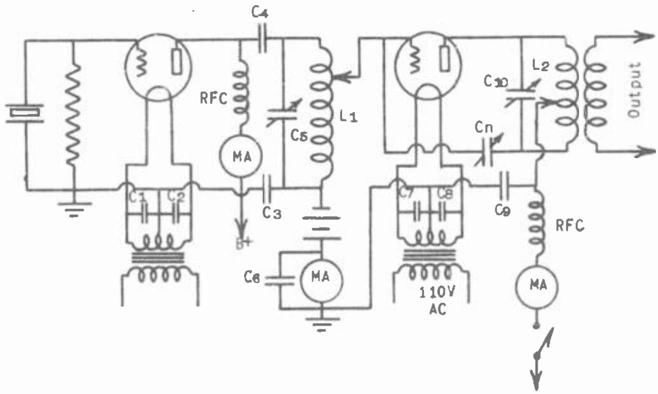


Fig.5 A shunt-fed oscillator and a buffer amplifier.

lower end of this tank circuit with respect to R.F., but this cannot be done directly without grounding the bias voltage of the buffer. The other components of this circuit serve the same purposes as in Fig. 4.

3. NEUTRALIZATION. In Lesson 24 of Unit 1, it was learned that a radio-frequency amplifier employing a three-element tube must be neutralized to prevent self oscillations due to feedback through the interelectrode capacity of the tube. The oscillations are set up by the R.F. voltages developed across the plate load by the varying plate current. These voltages cause R.F. currents to flow through the plate-grid capacity of the tube to the grid circuit, and in flowing through whatever is connected in the grid circuit of the tube, the R.F. currents set up R.F. voltages which then serve as the grid excitation of the tube. Since the three-element tube has the ability to amplify a voltage applied to its grid circuit, the power developed in the plate circuit of the tube is greater than that present in the grid circuit. Therefore, by feeding a part of the R.F. energy of the plate circuit back into the grid circuit, the tube may be made to supply its own input voltage. When this occurs, the tube acts as an oscillator instead of an amplifier.

The fact that it is possible to produce sustained oscillations by the feedback of R.F. energy through the interelectrode capacity of the tube is taken advantage of in the tuned-grid, tuned-plate oscillator. All the oscillators illustrated in this lesson depend upon this principle for their operation. This condition is, then, desirable in oscillators.

On the other hand, this action is very undesirable in amplifiers. The carrier frequency has been generated by the oscillator and great care has been taken to see that this frequency is as stable as possible, by the use of a temperature-controlled, piezo-electric crystal. The purpose of the stages following the oscillator is to amplify this weak output of the oscillator and not to oscillate and thereby introduce other frequencies of lesser frequency stability. Since it is absolutely essential that the buffer amplifier and the remaining amplifiers do not of themselves oscillate, some means must be taken to prevent this occurrence.

The prevention of oscillations due to feedback through the interelectrode capacity is effected by neutralization; that caused by other sources, such as magnetic feedback, must be eliminated by careful construction and placement of the component parts. Since the problem of neutralization is of major importance in the design and operation of transmitters, it is advisable that this subject be considered in detail.

In the discussion of the neutralization of radio-frequency amplifiers for reception purposes, two different methods of neutralization were given. The principle of operation consists of the feeding back from the plate to the grid circuit an R.F. current equal in magnitude and opposite in phase to that fed through the interelectrode capacity of the tube. The amount of current fed back is adjusted by varying the capacity of the neutralizing condenser. If the current fed through the neutralizing condenser is to be 180° out of phase with that fed through the interelectrode capacity, the voltage producing this current must be 180° out of phase with that between the plate end of the tank circuit and ground. Such a voltage may be easily secured by tapping the plate tank at its exact center and feeding the plate voltage into the tank at this

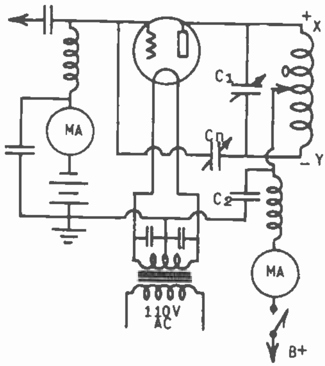


Fig. 6 Showing the connections for plate neutralization.

point. From the point where the plate voltage is applied, an R.F. by-pass condenser is connected to ground, thereby grounding this point with respect to R.F. as is shown in Fig. 6. This is called "plate neutralizing".

It is a well known fact that with the R.F. ground at this point, the voltage between the top of the tank and ground will be 180° out of phase with the voltage between the bottom end of the tank and ground. That is, as point X is becoming more positive, point Y will be growing more negative. Therefore, the voltage between the bottom of the tank and ground serves to force an R.F. current through the neutralizing condenser to the grid circuit of the tube. It is, of course, possible to vary the amount of neutralizing voltage by changing the position of the tap, O. However, it has been definitely determined that the voltage across the two parts of the tank circuit will not be perfectly 180° out of phase unless the plate tank is tapped at its exact center. Thus an imperfect form of neutraliza-

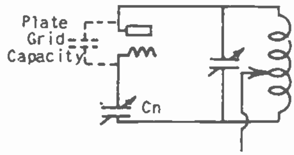


Fig. 7 Showing how the neutralizing condenser affects the tuning.

tion may be obtained when the plate is tapped off-center, but maximum efficiency and stability demand that the neutralization be as perfect as possible. For this reason, all broadcast transmitters (and many amateur transmitters) use center-tapped tank circuits.

It should be noticed that the neutralizing condenser is in series with the grid-plate capacity of the tube, and this series combination is, in effect, in parallel with the plate-tank, tuning condenser. Thus every change of the capacity of the neutralizing condenser will slightly affect the tuning of the tank circuit. This is illustrated in Fig. 7.

Fig. 8 illustrates another method of neutralization. It is called "grid neutralization" or the "Rice" method of neutralization.

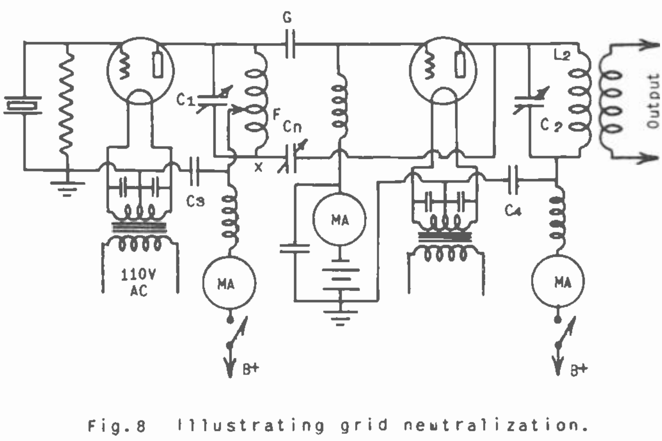


Fig. 8 Illustrating grid neutralization.

In this circuit, the plate voltage of the oscillator is applied not at the bottom of the oscillator plate tank, but at its exact center. Since this point of the tank circuit is grounded with respect to R.F., the excitation voltage for the grid of the buffer is that voltage present between the top of the plate-tank circuit and point F or ground. The neutralizing voltage appears across the lower portion of the oscillator tank from point F to point X. Just how this circuit is able to neutralize the energy fed back through the interelectrode capacity of the tube may be more easily seen by redrawing the circuit in the form of a Wheatstone Bridge. This is done in Fig. 9. The voltage developed across the buffer tank circuit forces an R.F. current through the grid-plate capacity of the tube, and through that part of the oscillator tank circuit between

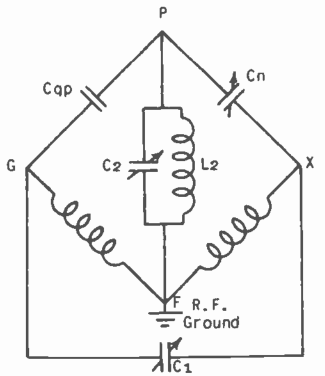


Fig.9 The circuit of Fig. 8, redrawn to show the components of the Wheatstone Bridge.

points G and F. Point F, we must remember, is at R.F. ground potential. In addition, the bottom end of the buffer tank circuit is also at an R.F. ground, since the capacitive reactance of the bypass condenser C_4 is very low to the oscillation frequency. Therefore, this circuit redrawn as a Wheatstone Bridge is equivalent to that of Fig. 8.

The R.F. voltage developed across the plate tank of the buffer also forces an R.F. current through the neutralizing condenser, and through the plate tank of the oscillator from point X to point F. If proper neutralization has been achieved, the voltage set up across the top of the oscillator plate tank from points G to F is equal to that produced from points F to X, and the two voltages are opposite in phase and therefore cancel. Or, looking at the circuit from the standpoint of a Wheatstone Bridge, the application of a voltage between the points P and F will not produce any voltage between the points G and X, if the bridge is balanced.

This method of neutralization has the disadvantage that not all of the voltage developed across the oscillator plate tank is available for grid excitation; half of it is used as the neutralizating voltage. Since the output of the oscillator is, at best, relatively weak; it is sometimes found that with this method, the voltage across the upper part of the oscillator tank is insufficient to properly excite the grid of the buffer. For this reason, grid neutralization is not ordinarily used with buffer amplifiers, but is quite popular for the amplifier stages following the buffer, as is evidenced by the fact that most broadcast transmitters employ this method in the higher-powered stages.

4. SCREEN-GRID BUFFER AMPLIFIERS. The fact that the need for neutralization is easily obviated by the use of screen-grid tubes was thoroughly discussed in Lesson 25 of Unit 1. By using a second grid within the tube between the control grid and plate to which a voltage somewhat less than the plate voltage is applied, and by grounding this grid with respect to R.F. by means of a low-reactance condenser, the effective grid-plate capacity is reduced to a very low value. The screen grid acts as an electrostatic shield between the plate and the control grid. The DC voltage applied to the screen grid varies from one-sixth to one-third of the plate voltage.

The outstanding success attained with screen grid tubes in receivers led to the design and construction of larger screen-grid tubes to be used in transmitting circuits. They are especially desirable in high-frequency transmitters where the capacitive reactance of the grid-plate capacity of a three-element tube is so low that neutralization is difficult. Screen-grid tubes for use in transmitters have a wide variety of plate dissipations, ranging from about 15 watts to 750 watts. In designating the rating of a transmitting screen-grid tube, it is also necessary to specify the allowable screen-grid dissipation.

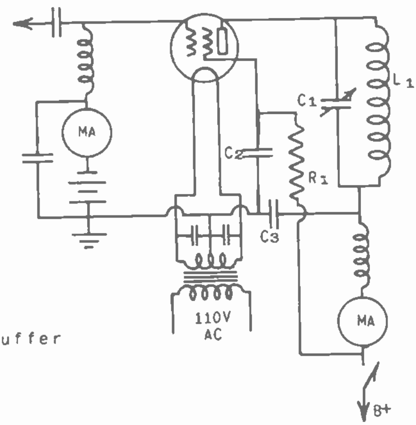


Fig.10 A screen grid buffer amplifier.

A buffer amplifier circuit using a screen-grid tube is shown in Fig. 10. R_1 is the screen-dropping resistor, used to reduce the voltage of the power supply to a value suitable for application to the screen. C_2 is the screen by-pass condenser, while C_3 serves the same purpose for the plate circuit. If care is taken in the placement of the parts, and if proper shielding is used, a screen-grid tube will operate properly in a circuit without neutralization, even at the higher frequencies encountered in short-wave transmitters.

5. INDUCTIVE COUPLING. All of the buffers illustrated thus far in this lesson have been capacitively coupled to the crystal oscillator. It is possible, however, to couple the oscillator to the buffer by transformer action. Fig. 11 shows such a circuit. The current flowing in the tank circuit of the oscillator, L_1C_1

induces R.F. voltages into the grid tank, L_2C_2. This grid tank should be tuned to resonance with the oscillator frequency in order to develop maximum grid excitation voltage for the buffer tube. Resonance of the grid tank is indicated by the flow of max-

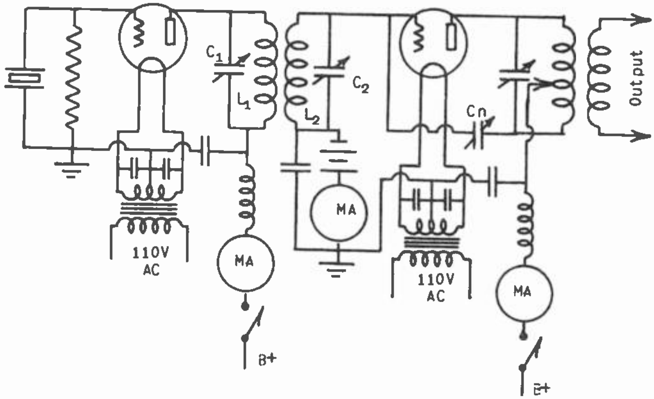


Fig.11 Buffer stage transformer-coupled to crystal oscillator.

imum grid current. The amount of grid excitation is changed by varying the distance between the plate tank coil of the oscillator and the grid tank coil of the buffer. Straight inductive

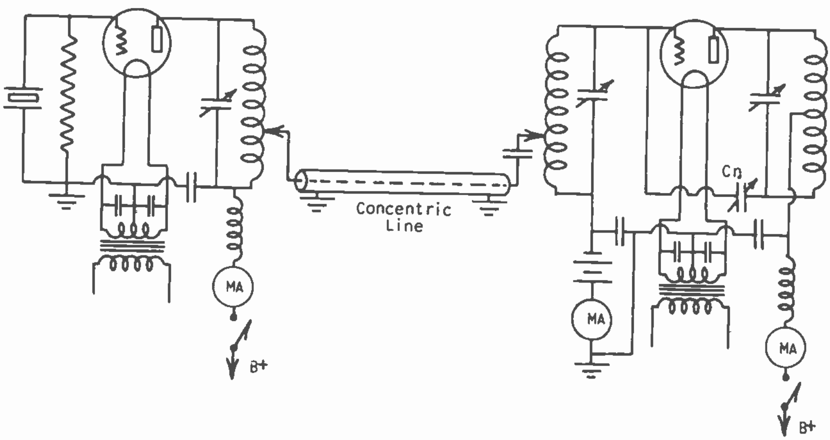


Fig.12 Illustrating the use of a concentric line to couple the oscillator to the buffer.

coupling is, of course, limited to those cases where it is practical to mount the plate coil of a stage close to the grid coil of the following stage. Often it is not convenient to arrange the various parts on the chassis or in the transmitter cabinet so that the two coils may be close together. In this case a modified form of inductive coupling known as link coupling is to be preferred.

Instead of coupling the plate circuit of one stage directly to the grid circuit of the following stage, they are coupled by means of a third circuit called a link. Broadcast transmitters ordinarily employ a concentric line, the ends of which are coupled either inductively or capacitively to the respective stages.

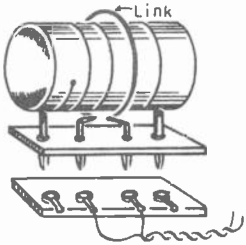


Fig.13 A plug-in coil with the coupling link.

Amateurs, on the other hand, commonly use a twisted-pair line. Either method is satisfactory, but naturally the concentric line has less loss, although it is considerably more expensive.

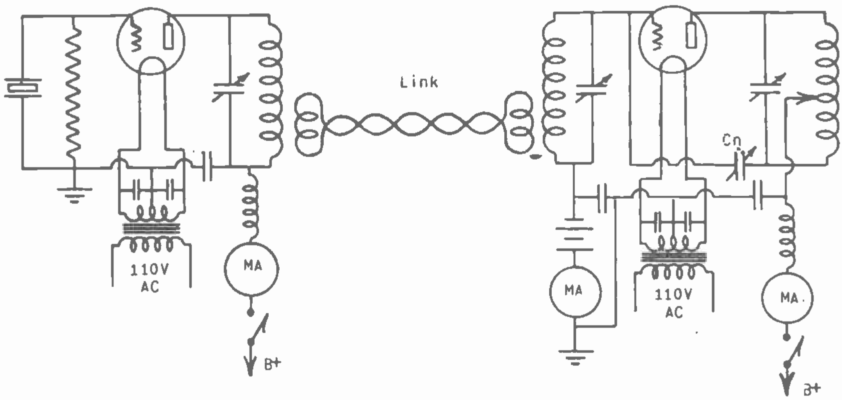


Fig.14 An oscillator link-coupled to a buffer.

Fig. 12 shows two stages capacitively coupled by means of a concentric link. Such an arrangement allows the two stages to be placed on different shelves of a rack or wherever is most convenient. When the link is inductively coupled to the tank circuit, one or two turns of the line are placed around the tank coil as shown in Fig. 13. This coil is of the plug-in type; ordinarily only the high-powered stages use copper tubing or copper strip in the construction of tank coils.

Fig. 14 shows a schematic diagram of a crystal oscillator link-coupled to a buffer stage. Sometimes more than one turn is used in the link circuit, as illustrated in Fig. 15. This figure shows a tank coil made of copper tubing; the link circuit consists of three turns of tubing magnetically coupled to, but electrically insulated from the tank coil. In many instances the link is smaller in diameter than the tank coil and is placed inside the tank. One advantage of this method of construction is that the link may be pivoted so that it will rotate within the tank coil.

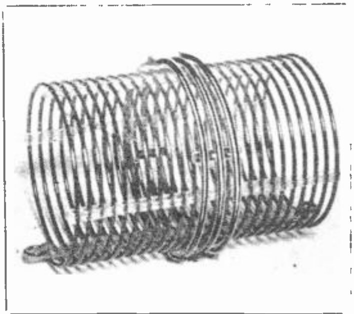


Fig.15 A large tank coil with a coupling link.

When the axis of the link is parallel to the axis of the tank coil, there will be maximum inductive coupling between them; and when the two axes are at right angles, the coupling is minimum. This affords a convenient means of varying continuously and smoothly the amount of power transferred from one stage to another. A tank coil of this type is illustrated in Fig. 16. When inductive coupling is employed, the link circuit should be coupled to the "cold" end of the oscillator and the buffer tanks. By "cold" is meant that end of the tank which is grounded with respect to R.F.; the "hot" end of the tank which is the end nearest

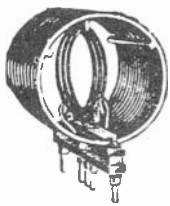


Fig.16 Tank coil with link mounted inside for variable coupling.

the plate or grid of the tube. By coupling the link circuit to the cold end of the tank, the energy which is transferred from the tank to the link, due to capacitive effect between them, is reduced to a minimum. It is desirable that the transfer of energy between them be due only to inductive coupling.

The advantages claimed for link coupling are:

(1) Provides a flexible feed line up to several feet in length, which results in more efficient operation between stages in "rack" type transmitters in which the stages are relatively far apart.

(2) It permits the use of series-feed in both the grid and the plate circuits.

(3) With a given amount of excitation on the grid of the first buffer, the use of link coupling decreases the plate current in the crystal oscillator stage and thereby reduces the R.F. current through the crystal itself.
(4) Eliminates the use of taps on coils with their resulting attendant losses.
(5) Due to the lack of capacitive coupling effect when using link coupling, neutralization is made easier.

Most of these claims result from the fact that link coupling provides a certain amount of impedance matching between the stages. The fact that a tube should work into a certain value of load impedance for its best operation has been discussed in previous lessons. In the following lesson, a detailed discussion of impedance matching will be presented.

6. THE COMPONENT PARTS OF A TRANSMITTER STAGE. In the preceding lesson, the construction of tank coils was discussed. It was learned that high-powered stages always employ a tank coil made of copper tubing or flat copper strip; the purpose being to secure the proper amount of inductance with the least amount of resistance. Stages of lower power use wire-wound coils, which may or may not be of the plug-in type. Except when link coupling is used, all connections are made to the tank coil by means of clips, so that any part of the coil may be used. Wire-wound transmitter tank coils have taps brought out every few turns, so that the amount of coil in use may be varied. With capacitive

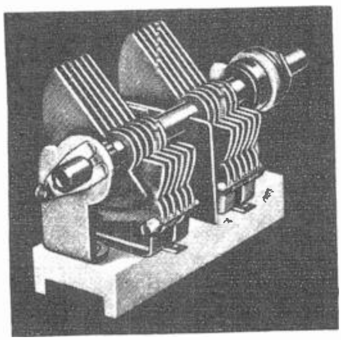


Fig. 17 A split-stator tuning condenser.

coupling, it is desirable that the connection to the grid of the following stage be made with a clip to make the adjustment of the grid excitation easy. By moving the tap toward the top or plate end of the coil, the grid excitation is increased, while moving it toward the lower or ground end of the tank causes a decrease in the excitation. All tank coils should be mounted on stand-off insulators having low dielectric losses, and large masses of metal should be kept away from the field of the coil. Standard practice demands that a space around the tank coil equal at least to one-half the diameter of the coil (and preferably equal to the diameter of the coil) be kept free from metallic objects or other bodies might cause losses in the tank.

Even if the feedback through the interelectrode capacity is perfectly neutralized, the stage will still oscillate if there is magnetic coupling between the plate coil of one stage and that of the preceding stage. For this reason, the two coils should be

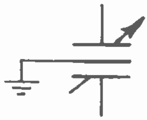


Fig. 18 The symbol used to denote a split-stator tuning condenser.

mounted at right angles to each other to minimize the inductive coupling between them, and for the best results, each stage should be separately shielded and the shield well grounded. The shield is usually composed of sheet copper or aluminum or may be constructed of any non-magnetic material which is a good conductor.

The tuning condensers used in transmitting stages are similar to those employed in receivers, except that they are of a more rigid construction and the plates are more widely spaced to prevent arcing from one plate to another under the influence of the high voltages commonly used in transmitters. A type of condenser

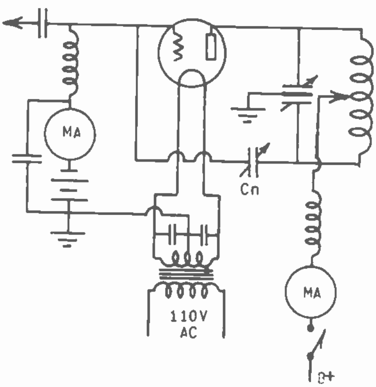


Fig. 19. Plate neutralization using a split-stator tuning condenser.

extensively employed in transmitters which has not as yet received consideration is the split-stator tuning condenser. It consists of two stator sections insulated from each other and two rotor sections mounted on the same rotating shaft, and therefore connected together electrically. This condenser is identical in construction to an ordinary two-gang receiving condenser. An illustration of a split-stator tuning condenser is shown in Fig. 17, while the symbol used to represent it is given in Fig. 18.

A circuit using a split-stator tuning condenser is shown in Fig. 19. In this circuit the split-stator condenser is used to tune the plate tank. The rotor is grounded and the two stators connect to the two ends of the tank coil. In using this type of circuit, the plate voltage is always applied to the exact center of the plate tank coil. Notice that no R.F. by-pass condenser is connected between the point where the plate voltage is applied and ground. It is unnecessary, since the grounding of the rotor of the condenser effectively grounds the tank circuit. Of course, such a circuit as this is used only when it is necessary to split

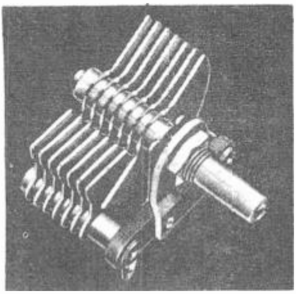


Fig. 20 A neutralizing condenser.

the tank into two sections to obtain neutralizing voltage. The excitation voltage for the succeeding stage is equal to the voltage developed across the upper section of the tuning condenser, while that across the bottom section is used for neutralization. Since the two sections of the tuning condenser are in series, the total tank capacity is equal to one-half of the capacity of each section. Since the exact center of the tank coil is at the same voltage as the center of the condenser, or the rotor, this point of the coil is at ground potential with respect to R.F. It is for this reason that the plate voltage should be applied at this point.

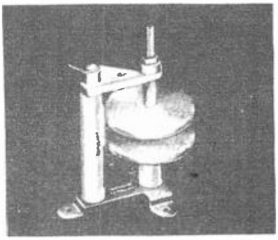


Fig. 21 A neutralizing condenser used for high frequencies.

Neutralizing condensers for transmitters are small variable condensers having a small plate area and consisting of just a few plates. Sometimes they have just three plates; two stator plates and one rotor. Fig. 20 shows an illustration of a neutralizing condenser. High-frequency transmitters often employ neutralizing condensers having just two plates as the one shown in Fig. 21. The capacity is varied by changing the distance between the two plates; the shaft of the upper plate is threaded and this plate may be raised or lowered by means of a screw driver.

All R.F. by-pass condensers used in a transmitting stage have a capacity of .002 mfd. or higher. They should be absolutely non-inductive; and, since mica condensers comply with this requirement better than any other kind, they are nearly always used. The by-pass condensers used in the filament circuit do not need to have a high breakdown rating as they are subjected to only one-half the filament voltage. The plate by-pass condenser must be able to withstand somewhat more than twice the DC plate voltage. Since the voltages used in a transmitter are rather high, mica condensers are best suited for this purpose; because, for a given thickness, mica is able to withstand a higher voltage than any other common dielectric.

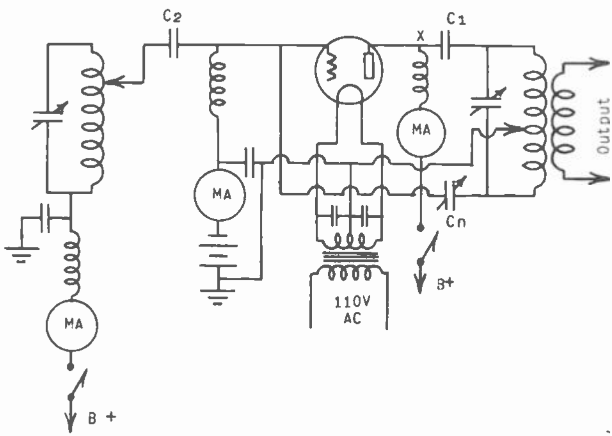


Fig. 22 A buffer amplifier with a shunt-fed plate circuit.

The coupling condenser between the oscillator and the buffer is usually about .0001 mfd. A larger capacity would, of course, have a lower reactance to the oscillator frequency, but it is better to use a condenser of this size so that the oscillator will not be loaded too heavily. A smaller capacity condenser in this position reduces the excitation voltage on the buffer and therefore reduces the power drawn from the oscillator, which tends to make the oscillator more stable. The left plate of this condenser (C_2 in Fig. 22) is connected to the positive side of the plate voltage supply, while the right plate is connected to the negative side of the bias supply of the following tube. Thus, this condenser must be able to withstand the plate voltage of the first tube plus the bias voltage of the tube following.

In this connection, it should also be noticed that the voltage across the neutralizing condenser is equal to the plate voltage plus the bias voltage, and it must be designed to withstand this voltage. The condenser that by-passes the C battery is subjected to the entire bias voltage. Since this voltage will average from 100 to 200 volts, a 600 volt mica condenser will be satisfactory in this position. In the interest of safety, the rating of any condenser used in the transmitter should be somewhat higher than the voltage that is to be applied to it.

The R.F. chokes used in a transmitter operate in conjunction with the R.F. by-pass condensers to separate the DC from the R.F. components of the plate and grid currents. Fig. 22 shows a buffer amplifier with a shunt-fed grid and plate circuit. The R.F. chokes used in this circuit will probably have an inductance of 2 to 20 millihenries, and a DC resistance of from 8 to 75 ohms. Their current-carrying capacity will be determined by the size of the wire with which they are wound. The DC current rating should be large enough to insure that the choke does not run hot. To the DC component of the plate current, the resistance of the choke would offer very little opposition, while the practically infinite resistance of the blocking condenser C_1 would be the same as open circuit to the DC component. Therefore, the DC component flows through the choke and back to the power supply.

Now let us assume that the inductance of the choke is 4 millihenries; the capacity of the condenser is .002 mfd.; and the frequency is 1000 kc. The inductive reactance of the choke at this frequency is 25,120 ohms; and the capacitive reactance of the condenser is 79 ohms. Thus it is easy to see that the choke will offer a very large impedance to the R.F. component of the plate current, while the by-pass condenser offers a negligible impedance.

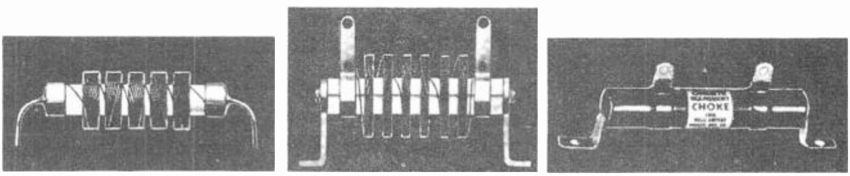


Fig.23 R.F. chokes for transmitter stages.

The same reasoning applies to the grid circuit. The DC component of the grid current finds an easy path through the R.F. choke and the C battery to ground, while the R.F. component flows through the coupling condenser, through the plate tank coil of the preceding stage, and through the plate by-pass condenser to ground. Despite the fact that the R.F. choke is used, it is also necessary to by-pass the C battery to insure that the R.F. current does not flow through it. It should be observed that an R.F. choke does not present an infinite impedance to the R.F. current, and therefore there will be a small amount of R.F. current flowing through it. In the shunt-fed plate circuit, this is very undesirable, since the R.F. current flowing through the choke represents a loss of R.F. power which does not reach the tank circuit. This R.F. current which flows through the choke is called the leakage current, and the fact that any R.F. choke will have some leakage current makes the series-fed plate circuit preferable to the shunt-fed type.

An R.F. choke possesses some distributed capacity. In well-designed chokes, this will be a minimum, but it can never be completely eliminated. This distributed capacity has the same effect as a condenser connected in parallel with the choke. At some particular frequency, this capacitive effect will tune the choke to resonance. At frequencies lower than the resonant frequency of the choke, it will act as an inductance, while at frequencies above the resonant frequency, the choke will act as a condenser, and all its choking action will be lost. For this reason, chokes that are suitable for broadcast frequencies are not generally usable in high-frequency transmitters. Several R.F. chokes are illustrated in Fig. 23.

The R.F. choke should be mounted so that it is not in the field of the tank coil, for if the oscillating tank current induces voltages into the choke, the operation of the transmitter is extremely unstable.

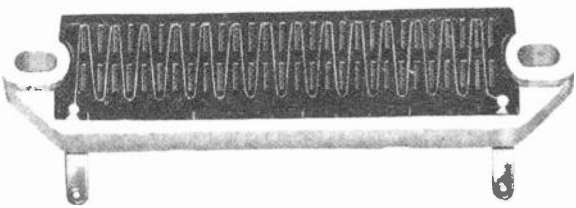


Fig. 24 A high-wattage, non-inductive resistor.

The resistors used in transmitting circuits should usually be of the so-called "non-inductive" type. The ordinary wire-wound resistor is not suitable, for it has considerable inductive reactance at radio frequencies. The carbon type resistor is non-inductive, but it is not capable of dissipating much more than 5 watts. When the wattage to be dissipated is much in excess of this value, a special non-inductive, high-frequency resistor is used. This resistor, which is shown in Fig. 24, consists of high-resistance wire arranged on a rectangular refractory plate in such a manner as to give the least possible values of inductance and distributed capacity. Because of these desirable features, it is often used as the antenna resistance in dummy antennas. By using several of these resistors in series, parallel, or series-parallel combinations, it is possible to secure almost any resistance value with practically any wattage rating.

Before describing the process of neutralization and adjustment of a buffer amplifier, there is one other piece of apparatus which should be discussed, since it is very useful in detecting R.F. voltages. This is the neon lamp shown in Fig. 25. A neon lamp consists of an evacuated glass bulb containing the two flat metal plates placed in the same plane and separated from each other by a very small air gap. The base of the bulb is usually threaded the same as an ordinary electric light bulb, and the plates are connected to the tip and the threaded part of the base.

When a DC voltage is applied to these electrodes, that one connected to the negative terminal of the voltage source will glow with an orange color, provided that the voltage exceeds a certain value. If the voltage source is AC, both electrodes will glow. Thus, this lamp affords an easy means of determining whether a voltage is AC or DC. These lamps are manufactured in sizes from one-fourth watt to 2 watts, and have found a limited use as night lights. However, their major application is the detection of R.F. voltages. The small ones are used in wavemeters for rough frequency measurements.

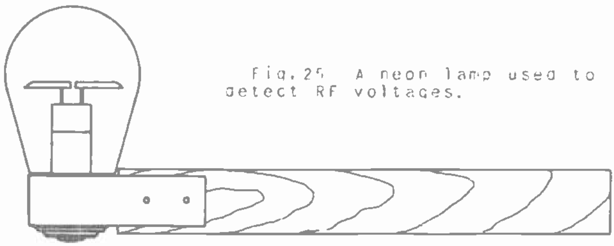


Fig. 25 A neon lamp used to detect RF voltages.

No attempt will be made to describe the operation principle of the neon light, since this is a subject that properly belongs in the study of television. If one of the small size lamps is touched to the plate terminal of the oscillator tank coil, it will glow if oscillation is taking place. Connection need be made to only one of the electrodes to produce this phenomenon. The capacity existing between the other terminal and ground is sufficient to complete the circuit at ordinary radio frequencies. To detect R.F. voltages in the buffer and succeeding stages, the 2 watt size is used. It is fastened to a wooden stick to prevent the user from receiving a painful R.F. burn. When the lamp is touched to the plate end of a tank coil, it glows brightly; and, as it is moved toward the bottom end of the coil, the brilliance of the glow diminishes, indicating smaller R.F. voltages. When it reaches the point where the by-pass condenser is connected, the lamp shows no illumination, since this point is at an R.F. ground. Then, as it is moved on downward to the extreme bottom end of the coil to the point where the neutralizing condenser is connected, it again glows. If the lamp is touched to the plate end of an R.F. choke in a shunt-fed plate circuit, (point X in Fig. 22), it should indicate an R.F. voltage at this point; but if it is moved to the other end of the choke, it should not glow at all. An indication of R.F. voltage at this point means that the choke is ineffective in keeping the R.F. currents out of this circuit and should be replaced. The neon lamp method for indicating R.F. voltages should never be used on a tank circuit that contains more than 100 watts of R.F. power. Furthermore, the lamp should be mounted at the end of a long wooden stick and reasonable precaution should be exercised when using this method for detecting R.F.

voltages. NEVER HOLD THE LAMP DIRECTLY IN YOUR HAND WHEN APPLYING IT TO THE TANK CIRCUIT, FOR SUCH AN ACTION IS LIABLE TO RESULT IN A PAINFUL R.F. BURN OR A SEVERE SHOCK.

7. NEUTRALIZATION AND ADJUSTMENT. The procedure for neutralizing a transmitter stage is somewhat different from that used to neutralize an R.F. stage in a receiver. The circuit shown in Fig. 26 will be referred to in describing the neutralization and adjustment process. The oscillator is a type 843 triode; the first buffer is a type 10 triode; and the final stage is another type 10 tube.

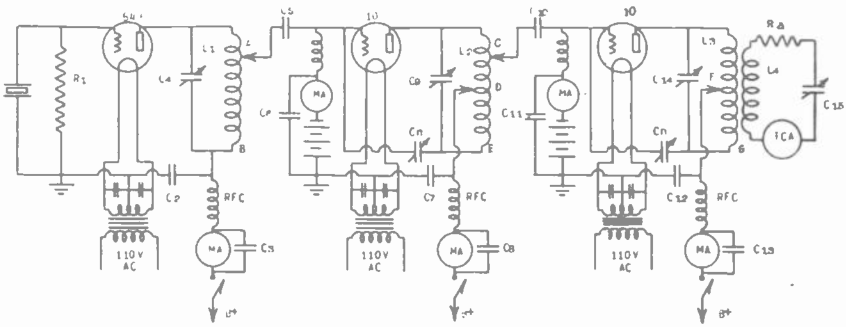
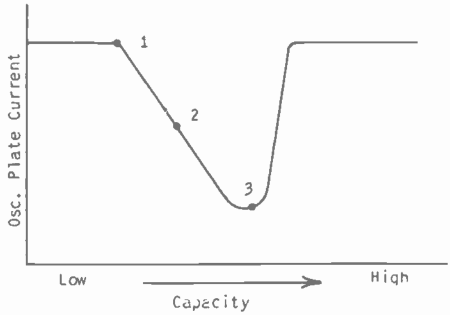


Fig.26 A three-stage transmitter.

Before starting the description of the neutralization process, it is necessary that you be impressed with the fact that no attempt should ever be made to operate a three-element tube in a transmitting circuit until it has been neutralized. If such an attempt is made, the circuit will self-oscillate, and when this occurs, the plate current rises to a dangerously high value, and the plate of the tube becomes white hot, usually resulting in complete ruination of the tube. Therefore, remember never to apply the plate voltage to a three-element transmitting tube until it has been properly neutralized. (The exception to this statement is, of course, the oscillator tube.)

To start the neutralization process, the filament voltage is applied to all three tubes, and the filaments are allowed to reach their operating temperature. The coupling between the oscillator and first buffer is disconnected by removing the excitation tap A from the oscillator plate tank. The oscillator plate voltage switch is next closed; this applies plate voltage to the oscillator. Simultaneous with this action, the plate milliammeter will indicate that plate current is flowing. It will probably have a value ranging from 20 to 30 ma. The oscillator tank condenser is now slowly rotated while the reading of the plate milliammeter is closely watched. When the plate tank is tuned to the resonant frequency of the crystal, the needle of the plate milliammeter will dip, indicating a lower value of plate current. Since, as was discussed in the previous lesson, it is not desirable to operate the crystal oscillator at the minimum reading of the plate milliammeter, the plates of the tuning condenser are unmeshed a trifle more until the plate current reading is slightly above its minimum value. We have learned that the oscillator tank circuit should be tuned to a frequency slightly higher than the resonant frequency of the crystal so that it will have a slight inductive reactance to compensate for the capacitive reactance of the interelectrode capacity. Fig. 27 shows a graph illustrating the relation between the DC plate current of the oscillator and the tank capacity, C_4. The oscillator will be most efficient if operated at point 3, but its stability will be poor, since a slight increase in capacity will cause the oscillations to stop. For this reason, it is advisable to operate the oscillator at point 2. When the proper adjustment has been made, the plate current will be from 12 to 15 ma.

Fig. 27 Showing the relation between the capacity of the tuning condenser and the oscillator plate current.



The oscillator plate voltage is removed and the buffer excitation tap A is connected to the tank coil of the oscillator near the plate end of the coil. After it has been ascertained that the correct bias voltage has been applied to the buffer tube, it is safe to close the oscillator plate voltage switch. The bias voltage should be approximately one and one-half times the cutoff value. You will remember that in the lesson on self-excited oscillators, it was learned that the approximate cutoff bias for a triode is determined by dividing the DC plate voltage by the amplification factor of the tube.

With the crystal oscillator again functioning, it will be noticed that a grid current is flowing in the first buffer stage. Although the type 10 tube is biased to a value beyond cutoff, the excitation voltage supplied by the oscillator plate tank is sufficient to overcome the bias voltage during a part of the R.F. cycle, and the grid of the tube is driven positive for a short time each cycle. When the grid goes positive, it draws current which flows in the form of pulses lasting for a very short time. The average of these pulses or the DC component of the grid current flows through the R.F. choke, through the meter and battery, and back to the filament. The R.F. component flows through the coupling condenser C_5, through the oscillator plate tank, and through the plate by-pass condenser C_2 to ground.

It should also be observed that the oscillator plate current increases slightly when the excitation voltage is applied to the buffer stage. This indicates that the oscillator has been loaded. Whenever a stage draws grid current, power is dissipated in the grid circuit. This power is not supplied by the C battery, because the grid current flows through this battery in the opposite direction from that in which the battery can force current. When a battery delivers power to a circuit, the current flows from the negative terminal of the battery, through the external circuit, and back to the positive terminal of the battery. In this case, however, the grid current flows into the negative terminal of the battery and out of the positive terminal of the battery; therefore, it is not the C battery that is producing the grid current or supplying the power.

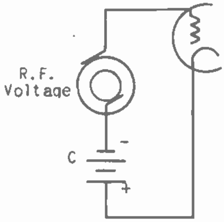


Fig. 28 Illustrating how the excitation voltage may be replaced by an equivalent RF generator.

The power dissipated in the grid circuit must be supplied by the source of excitation voltage which is the tank circuit of the oscillator. Perhaps this may be made clearer by reference to the diagram of Fig. 28. The R.F. voltage present across the tank circuit of the oscillator has been replaced by an alternator whose peak voltage is equal to that produced across the oscillator tank. For the sake of simplicity, we will assume that this alternator is in series with the C battery. The C battery cannot produce a current flow through this circuit, because it is not possible for current to flow from the grid of the tube to the filament. Likewise, when the voltage of the alternator is in such a direction as to make the grid negative with respect to the filament, no grid current can flow. But when the voltage of the alternator changes polarity and reaches its maximum voltage in the opposite direction, the voltage across it is greater than the voltage of the C battery, and as a result, the grid is positive with respect to the filament. When this occurs, the grid draws current and power is dissipated in the grid circuit. This power is dissipated in forcing the grid current through the C battery against the voltage of this battery. Since it is the alternator which causes this current flow, the alternator is the device that furnishes the power, or, in actuality, the power is taken from the tank circuit of the oscillator.

In Lesson 1 of this unit, we learned that the extraction of power from a tank circuit produced the same effect as adding resistance to the tank. Or, in effect, the power drawn from the oscillator has reflected a resistance into the tank circuit of the oscillator. Therefore, the amplitude of the oscillating current decreases and the voltage produced across the tank is reduced. This increases the minimum plate voltage, and as a result, the plate current of the oscillator is increased. Reference to the curves given in Lesson 1, which show the effect of loading an oscillator, should be made at this time, if the foregoing statements are not entirely clear.

Some transmitter manufacturers prefer to operate the buffer amplifier in such a manner that it does not draw any grid current. It is obvious that if no buffer grid current flows, no power will be drawn from the crystal oscillator. This, in itself, is an advantage, since if there is no load on the oscillator, its frequency stability will be considerably better. In this case, the purpose of the oscillator is merely to furnish a voltage of the proper frequency for exciting the buffer. The buffer stage operates as a voltage amplifier, and its power output is rather low; therefore, more power amplifiers will be required. In any event, the grid current of the buffer, assuming that it is allowed to flow, should be low in order that the load on the oscillator is light.

The resonant frequency of the oscillator plate tank may have changed slightly when the excitation voltage is applied to the buffer; therefore, it should be checked at this time. We are now ready to neutralize the buffer stage. The neutralizing condenser should have a minimum capacity which is low compared with the grid-to-plate capacity of the tube, whereas its maximum capacity should be at least twice as great as this interelectrode capacity. This amount of capacity variation will be sufficient to insure that proper neutralization may be secured. The neutralizing condenser should be set to minimum capacity, and the tap D on the buffer tank circuit should be connected to the center of the tank coil, whereas tap C should be disconnected from the tank. With the stage un-neutralized, the R.F. voltage between the grid and ground will force an R.F. current through the interelectrode capacity of the tube, through the buffer tank circuit, and through the plate by-pass condenser C_7 to ground. The R.F. current which flows through the buffer tank will set up oscillations in the tank circuit, if this circuit is tuned to the resonant frequency. Before it is possible to proceed with the neutralization, some means must be provided to indicate whether oscillating current is flowing through the buffer tank or not. In commercial transmitters, the presence of oscillating current in the tank is often indicated by connecting a thermogalvanometer[1] in series with the tank tuning condenser. Sometimes a thermogalvanometer type of wavemeter is link-coupled to the tank coil by winding a few turns of wire around the coil and then connecting a twisted-pair line from these turns to the instrument. This method allows the wavemeter to be placed in a convenient position for easy reading.

[1]A thermogalvanometer is a very sensitive thermo-couple instrument for detecting a small amount of R.F. current.

If a thermogalvanometer or sensitive wavemeter is available, it should be used, because it will indicate when the stage is perfectly neutralized. However, it often happens that such an instrument is not easily obtainable; in which case it is common practice to use a small flashlight bulb connected to two or three turns of insulated wire. The wire may be almost anything on hand, and should be wound into a few loops with a diameter of about 2½ inches. One end of the wire is soldered to the threaded part of the bulb and the other end to the tip. When this flashlight loop is hung over the end of the tank coil or placed upright inside of the tank so that the plane of its loop is parallel to that of the coil's turns, it will glow if there is oscillating current in the tank circuit. The magnetic field set up by the oscillating current cuts through the turns of the loop and induces a voltage therein which causes a current to flow, thereby lighting the bulb. The flashlight bulb may be used only on stages of low power.

With the R.F. indicator in position, the buffer tank tuning condenser is rotated until the indication of oscillating current is maximum. This will occur when a maximum reading is obtained on the thermogalvanometer, or when the flashlight bulb burns the most brilliantly. When this point is reached, the tank is tuned to the resonant frequency of the oscillator. If, while tuning the buffer tank, the thermogalvanometer indicates that it is about to go off scale, or the flashlight bulb glows with more than normal brilliancy, it is advisable to increase the capacity of the neutralizing condenser slightly before making the final tuning of the tank circuit and thus avoid damage to the R.F. indicator.

With the buffer tank tuned to exact resonance, the stage is neutralized by increasing the capacity of the neutralizing condenser until the thermogalvanometer indicates that no R.F. current is flowing in the tank. The capacity of the neutralizing condenser should be increased in small steps, and the resonant frequency of the tank circuit should be checked in between these steps. This is necessary since the setting of the neutralizing condenser affects the resonant frequency of the tank circuit very slightly.

If the neutralizing voltage is correct, the thermogalvanometer should give a reading when the neutralizing capacity is minimum; should be zero when the right amount of capacity for perfect neutralization has been inserted into the circuit; and it should again give a reading when the neutralizing capacity is maximum. Unless the stages are well shielded and there is no magnetic coupling between the oscillator and the buffer tank coils, the thermogalvanometer will not show a zero reading, but will indicate a minimum value when the feedback through the interelectrode capacity has been neutralized.

When perfect neutralization has been attained, the R.F. voltage between the top end of the tank coil and point D, due to the R.F. current fed through the interelectrode capacity, is equal to the R.F. voltage between the point D and the lower end of the coil, due to the R.F. current fed through the neutralizing condenser. Since these two voltages across the plate tank buck each other, there will be no net R.F. voltage across the tank and no oscillating current will flow in the tank circuit. Thus, when the plate voltage is applied, the R.F. voltage built up across the coil will force two R.F. currents of equal value and of opposite phase through the interelectrode capacity and the neutralizing condenser respectively, and they will cancel in the grid circuit.

If a flashlight bulb is used for neutralizing, the capacity of the neutralizing condenser should be increased to the point where the filament of the bulb is no longer red, and the dial reading of the neutralizing condenser at this point should be carefully recorded. Let us say, for example, that this reading is 34. It cannot be assumed that the stage is now completely neutralized, merely because the current flowing through the flashlight bulb is insufficient to cause it to glow. Therefore, the capacity of the neutralizing condenser should be further increased until the glow of the flashlight bulb is again seen very faintly. Again the dial reading of the neutralizing condenser should be recorded; let us say that it is 42. Thus, the flashlight bulb was extinguished for 8 points of the dial (from 34 to 42). It would, therefore, seem reasonable to assume that the exact point of neutralization would occur when the dial reading was half-way between 34 and 42, or at 38. This is a correct conclusion, and the neutralizing condenser should be set half-way between the point at which the light goes out and the point at which it is again visible. This point is found by subtracting the lower reading from the higher and then adding half of the difference between the two readings to the lower value.

There is another method often used to indicate when the neutralization is complete. It is particularly advantageous since it does not require any additional equipment. When a stage is un-neutralized, the tuning of the buffer plate tank through resonance will cause the grid current to fall. This will be indicated by a sharp dip in the grid current meter as the resonant point is passed. One explanation of this phenomenon is as follows: The untuned tank circuit offers a very low impedance to the R.F. current which flows through the interelectrode capacity. Also, the impedance presented by the tank is off tune on the high or the low frequency side of resonance. Since the tank circuit offers a low impedance, a large amount of R.F. current will flow through the interelectrode capacity of the tube. We have learned that only a resistance is able to dissipate power; a reactance cannot do so, since all the power it takes from the voltage source on one alternation is returned to that source on the succeeding alternation. Since the tank circuit is predominantly reactive and is not acting as a resistance, it is unable to dissipate an appreciable amount of power even though the current flowing through it is relatively large. Therefore, it will have but little effect on the voltage source, which we have learned is the tank circuit of the preceding stage.

On the other hand, when the plate tank is tuned to the resonant frequency, it acts like a pure resistance of a fairly high value, and, as such, it is capable of dissipating an appreciable amount of power. Since this tank circuit has a fairly high impedance, the amount of R.F. current flowing through the interelectrode capacity will be small, yet even this small amount of current is able to dissipate a noticeable amount of power in the resonant tank circuit. This additional power used up in the tank circuit must be furnished by the voltage source or by the tank circuit of the oscillator. As previously stated, when power is drawn from a tank circuit, the amplitude of the oscillations decrease, and the voltage developed across the tank is reduced. Thus, it is evident that by tuning the tank of the buffer to resonance, and thereby causing this tank to dissipate power, the load on the oscillator is increased, with the result that the voltage across the oscillator tank is lowered, and the grid excitation voltage is reduced. The smaller excitation voltage will not drive the grid as far positive, and the DC grid current becomes smaller.

When the stage is perfectly neutralized, the R.F. current flowing through the interelectrode capacity cannot cause the tank circuit to oscillate, because the effect of this R.F. current is neutralized by the current fed through the neutralizing condenser. If the buffer tank does not have an oscillating current, it is unable to dissipate power and the load on the oscillator is unaffected by the buffer tank circuit.

The procedure used to neutralize a stage by observing the DC grid current meter of that stage is as follows: The filaments of the oscillator and the buffer are lighted, and the oscillator is put into operation by tuning its tank circuit. Then the excitation tap to the buffer is connected, and with the plate voltage of the buffer off, the buffer tank is tuned to resonance by noting the lowest point in the dip of the grid current meter. With one hand engaged in rocking the buffer tank condenser back and forth through resonance, the other hand slowly increases the capacity of the neutralizing condenser. It will be noted that as the capacity of the neutralizing condenser is increased, the grid meter dips less as the buffer tank is tuned through resonance. When perfect neutralization is attained, the grid meter will not even quiver when the buffer tank is rocked back and forth through the resonant frequency. This method is often used to check the completeness of the neutralization when it is performed by one of the other methods.

After it is reasonably certain that the stage has been properly neutralized, the buffer tank circuit is set at the resonant position, and the thermogalvanometer or flashlight bulb is removed. Failure to do this before applying the plate voltage will cause irreparable damage to either piece of equipment. The burning out of a flashlight bulb is not a serious loss, but the ruination of a thermogalvanometer is quite costly. The thermogalvanometer usually has a maximum scale reading of 100 ma., and since the oscillating current which will flow in the tank circuit after the plate voltage is applied is far in excess of this value, this instrument is sure to be ruined if it is left in the circuit. It is always good practice to resonate the tank circuit of a stage before applying the plate voltage to that stage. As far as the buffer stage is concerned, it is improbable that any damage would result if this were not done, but a high-powered stage would draw so much plate current if the plate voltage were applied with the tank circuit untuned, that the tube would be damaged. For this reason, the dial reading of the tank circuit at resonance should be noted during the neutralization process in order that the tank may be tuned after the neutralization is completed.

When the foregoing points have been checked, it is permissible to close the buffer plate voltage switch. The plate milliammeter will at once show a reading and the next thing to be done is to make sure that the buffer tank is properly tuned. This is accomplished by moving the tank condenser a small amount to either side of its present position. It should be observed that as the tank is detuned, the buffer plate current rises, and that the plate current is minimum when the tank is exactly tuned. Knowing the reason for this will help to fix the tuning process in your mind. At exact resonance, the tank is equivalent to a resistor of high value. In this case, the tube is working into a high load impedance. The large R.F. voltages set up across the tank cause the minimum plate voltage to be very low, and since the plate current flows in pulses which last only during the time that the plate voltage is low, the peak plate current that flows will not be very great. Therefore, the DC component of the plate current will also be low.

When the tank circuit is detuned, its shunt impedance is much less, and the tube works into a lower load impedance. The R.F. voltages across the tank are small and the minimum plate voltage is much greater. This causes the peak plate current to have a greater value, and its DC component, as read on the DC plate milliammeter is correspondingly larger. The action of most transmitter stages is practically the same as that of a self-excited oscillator. Therefore, the curves illustrating the relation between the various voltages and currents which were given in Lesson 1 of this unit apply equally well to most transmitter stages. Reference to these curves will make many of the characteristics of the buffer stage clearer. It should be realized that anything that causes the minimum plate voltage to be raised will produce an increase in the plate current. The similarity between the self-excited oscillator and the buffer stage is due to the fact that both are biased to a point beyond cutoff and both draw grid current.

To make certain that the buffer tank is tuned to the fundamental frequency of the crystal oscillator, a calibrated wavemeter should be loosely coupled to the tank of the oscillator and the dial rotated until the proper indication occurs. Then, without disturbing the setting of the wavemeter, it should be loosely coupled to the buffer tank circuit. If both tanks are tuned to the same frequency, the neon lamp should glow; but if the buffer is tuned to a harmonic frequency of the oscillator, the lamp will remain dark.

The excitation tap to the second buffer stage (C in Fig. 26) should now be connected to the tank of the first buffer. When this is done, the second buffer will draw grid current and the power consumed in this grid circuit will load the first buffer, thereby increasing its plate current. At this point, it is advisable to check the tuning of the first buffer, since applying the excitation voltage to the second buffer is apt to cause slight changes in the tuning of the preceding stage. Both the oscillator and the first buffer should now draw normal plate current as specified in the manufacturer's ratings. If the plate current of either stage is above normal, the ratio of inductance to capacity of both the oscillator and first buffer tanks, the grid excitation of the first buffer, and the grid bias of both stages, should be checked.

The grid bias of the crystal oscillator is supplied by the voltage developed across the grid leak R. Normal values of oscillator grid leaks range from 10,000 to 50,000 ohms; the higher value will give greater efficiency but less power output. The ratio of inductance to capacity of the oscillator tank should be quite high, since this tends to reduce the crystal current, and thus avoids overheating of the crystal. When the excitation tap A is moved to the plate end of the oscillator tank, the first buffer grid excitation is increased, but the load on the oscillator is also made greater. It is not desirable to load the oscillator very heavily, as to do so will lessen its frequency stability. The grid current of the first buffer, which is given in the manufacturer's specifications, is ordinarily from one-eighth to one-fourth of the plate current of that stage. If a further increase in excitation causes the plate current to decrease, the excitation is too great; while if the increased excitation produces an increase in plate current, the excitation is not excessive.

For a given power output from the buffer stage, it will require more grid excitation, when the tube is working into a high load impedance than when into a low impedance. For this reason, the inductance to capacity ratio of the buffer tank should be low, since it is this condition that produces a low load impedance. The first buffer is never adjusted for maximum power output, because this reduces the frequency stability of the oscillator. This, however, is not true of the remaining stages of the transmitter. They are adjusted for the maximum power obtainable at a reasonable efficiency. If the plate current of the first buffer is still too high after the excitation and the load have been adjusted, it may be reduced by increasing the grid bias of this stage. Due to the many factors involved, considerable experimenting may have to be done before sufficient power is obtained to excite the second buffer, without drawing excessive plate current from either the oscillator or first buffer. Never raise the applied plate voltage of either the oscillator or the first buffer above that recommended by the manufacturer in an attempt to increase the power output.

Although there is no simple method of determining the power output of either the oscillator or the first buffer, this matters little if the power output is sufficient to properly excite the grid of the following tube and cause it to draw the correct plate current when the bias voltage, plate voltage, and loading are correct.

When the oscillator and first buffer are drawing the correct plate current and the excitation of the second buffer is sufficient to cause it to draw the proper grid current, the next step is to neutralize the second buffer stage. This may be accomplished by any one of the methods previously outlined. With the second buffer correctly neutralized, and its tank circuit tuned to resonance, plate voltage is applied to this stage. To load the second buffer, the dummy antenna is loosely coupled to the plate tank circuit, and the antenna is then tuned to the resonant frequency as indicated by maximum antenna current. It is now advisable to check the tuning of the first buffer, since putting the final stage into operation will probably cause slight changes in the tuning of the first buffer. When the final stage is tuned to exact resonance as indicated by minimum plate current, the coupling between the antenna and the final tank circuit should be increased until the tube draws normal plate current. If the plate current of the final stage is excessive, loosen the coupling between it and the antenna and proceed to vary the inductance to capacity ratio of the final tank until the correct amount of plate current with a maximum amount of antenna current is obtained. The use of a high inductance to capacity ratio will increase the efficiency and cause a lower tank current, but it will lower the power output. A low inductance to capacity ratio will increase the power output and also the losses; therefore, it reduces the efficiency. A large oscillating tank current will flow, and the losses in the tank circuit will probably be large.

The following rule will help in determining when the inductance to capacity ratio has the correct value: With the antenna coupled to the final stage, tune the plate tank through resonance. If the minimum plate current does not occur at the same point as the maximum antenna current, the inductance to capacity ratio of the second buffer tank is too high. Now tune the plate tank to resonance, couple the antenna to the tank until the antenna current is maximum; then disconnect the antenna. If the plate current of the final stage does not drop to approximately 15% of its former value, the inductance to capacity ratio is too low, and the losses in the tank circuit are excessive.

With the load adjusted to the proper value, calculate the power in the antenna circuit by the I^2R Law, and also the power input to the final stage. From these values, compute the efficiency. If the efficiency is not 50% or greater, an attempt should be made to increase it by using a larger grid bias. Increasing the bias will reduce the power input to the tube without seriously affecting the power output up to a certain point. The plate current may be returned to normal value by increasing the load.

As a final test of the stability of the transmitter, open the oscillator plate voltage switch. This action should cause all the grid current meters to drop to zero, and if battery bias is used, the plate current meters will also indicate zero, since the stages are biased to a value greater than cutoff. However, when grid-bias is employed, the plate current meters will continue to read, because the removal of the grid excitation voltage also removes the grid bias. If any grid currents do flow, it indicates that one of the amplifier stages is self-oscillating. This may be caused either by imperfect neutralization or by magnetic coupling between the tank circuits. Furthermore, as the oscillator plate voltage switch is rapidly opened and closed, all of the meters of the transmitter should return each time to their former readings.

In closing this section on the neutralization and adjustment of buffer amplifiers, we wish to leave this timely warning. IN ORDER TO SAFEGUARD YOUR LIFE, NEVER MAKE ANY ADJUSTMENTS WHATEVER TO ANY PART OF THE TRANSMITTER WITHOUT FIRST MAKING SURE THAT THE PLATE VOLTAGE OF ALL THE STAGES IS REMOVED. HIGH VOLTAGE IS TREACHEROUS AND THE MAN WHO IS SENSIBLY AFRAID OF IT IS NOT LIABLE TO BECOME LAX AND CARELESS IN HIS DEALINGS WITH IT.

8. GRID-LEAK BIAS. Some transmitter designers prefer to obtain the bias for the buffer stages from grid leaks rather than from batteries or grid-bias power supplies. There is considerable economy in this method, since the resistors used for grid leaks are cheaper than batteries, and considerably less expensive than a

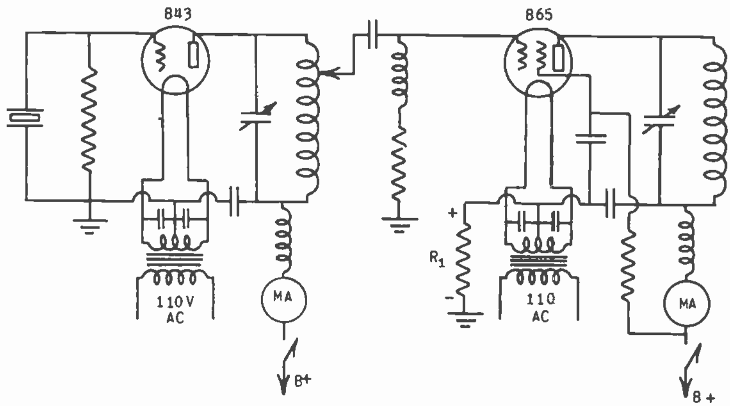


Fig. 29 Illustrating how the bias of a buffer amplifier may be obtained from a grid leak.

grid bias power supply. When this method is used, the bias is proportional to the amount of grid excitation, and is equal to the DC grid current times the value of the resistor.

There is, however, one decided disadvantage in the use of grid-leak bias. If, for some reason, the excitation of a tube should fail, that tube would not draw any grid current, and, as a result, there would be no bias voltage produced across the grid leak. With the high voltages used on transmitter stages, the absence of bias whatsoever would cause the tube to draw a very high plate current, which, in a very few moments, would ruin the tube. For this reason, grid leaks are never used as the sole means of obtaining bias on stages where the plate voltage is very high.

However, the grid leak may be used to obtain a part of the bias; the remaining part being derived from the drop across a resistor connected between the center tap of the filament circuit and ground. Such a circuit is shown in Fig. 29. This diagram illustrates an 843 oscillator coupled to an 865 screen-grid buffer stage. The resistor R_1 between the center tap of the filament of the buffer tube and ground provides just enough bias voltage to prevent the plate current from reaching an excessive value in case the excitation voltage fails and the bias developed by the grid leak is destroyed.

9. THE BUFFER DOUBLER. The difficulty of grinding quartz crystals that will oscillate at frequencies in excess of 7 megacycles was discussed in the preceding lesson. As the frequency increases, the thickness of the quartz plate decreases and, at very high frequencies, the quartz plate would be extremely thin and fragile. Such exceedingly thin plates are not practical for use in crystal-controlled oscillators, because the quartz plate is very liable to crack. The possibility of using tourmaline plates for frequency control at the ultra-high frequencies has been mentioned. Although this material is somewhat more suitable than quartz, its costliness has prevented its extensive use.

The remaining possibility of securing an ultra-high frequency signal which has good stability is to use a crystal oscillator having an output frequency of one-fourth or one-eighth of the desired frequency; operating the first buffer stage so that it will amplify only the second harmonic of the oscillator, and then working the second buffer to amplify the second harmonic frequency of the output of the first buffer. When a buffer is so adjusted as to amplify the second harmonic frequency of the crystal oscillator, it is called a "doubler". If it were desired to obtain a 16 megacycle signal with good frequency stability, it would be possible to use a crystal oscillator having a fundamental frequency of 4000 kc., and then by using two buffer doubler stages, increase the frequency to 16,000 kc. The tank circuit of the first doubler would be tuned to 8000 kc. (the second harmonic of 4000 kc.), while the tank of the second doubler would be tuned to 16,000 kc. (the second harmonic of 8000 kc.).

The fact that it is possible to obtain power from an oscillator at its second harmonic frequency is due to the impossibility of preventing the production of harmonics. The plate current of an oscillator or of a buffer consists of pulses which depart radically from sine wave form. It has been shown that any waveform that is not a pure sine wave will contain harmonic frequencies; therefore, these plate current pulses are rich in harmonics. It is, of course, the purpose of the tuned tank circuit to smooth these pulses into practically pure sine waves, since normally they are very undesirable. The larger the oscillating current in the tank circuit, the better the smoothing action will be, due to the fly-wheel effect of the tank.

In addition, the desirability of preventing the maximum grid voltage from exceeding 80% of the minimum plate voltage has been discussed. A maximum grid voltage of larger value will cause a dip in the plate current pulse, causing it to be even more distorted, and to contain more harmonics.

In frequency-doubling circuits, the harmonics are advantageous, and the oscillator and buffer circuits are adjusted to give a maximum amount of harmonic distortion. Thus the plate voltages are somewhat higher than for normal operating conditions; a larger grid bias is used, and the excitation voltage should be quite high. To prevent a large oscillating tank current, the inductance to capacity ratio of the tank is made considerably higher than for normal conditions.

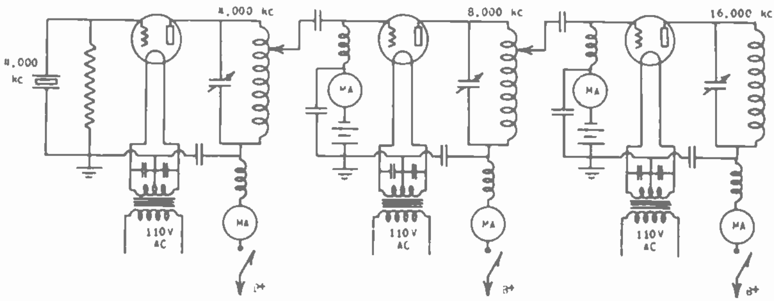


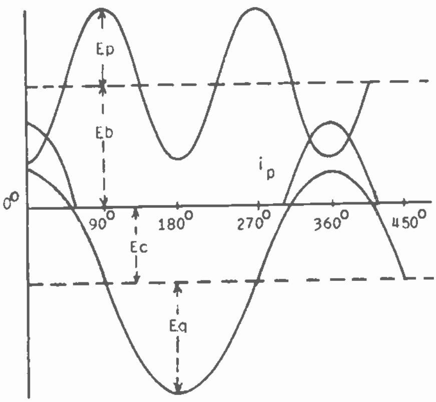
Fig.30 An oscillator and two buffer doublers.

A diagram of a transmitter using a crystal oscillator and two frequency doublers is shown in Fig. 30. Let us say that the crystal has a natural frequency of 4000 kc. The tank circuit of the oscillator is also tuned to this frequency, but its inductance to capacity ratio is high so that the oscillating current will be small. With a low oscillating current, the R.F. voltages built up across the oscillator tank will not be of pure sine wave form, but will be somewhat distorted. Therefore, the excitation voltage of the first buffer contains not only the fundamental frequency, but a considerable portion of harmonics as well. By tuning the plate tank of the first buffer to the second harmonic frequency of the oscillator, it can be made to absorb considerable power from the plate current pulse at this frequency and very little power at any other frequency. Since the buffer tank and the oscillator tank are tuned to two different frequencies, there is no need of neutralizing the buffer, for the feedback produced by the buffer tank is not of the correct frequency to produce grid excitation for this stage, and cause it to self-oscillate.

The power output of the first buffer when it is doubling is not as great as when it is working as a straight amplifier; nor is the efficiency as high. A stage working at 80% efficiency as a straight amplifier, would probably have an efficiency of not more than 70% when operated as a doubler. Thus, when using doublers, more amplifying stages will be needed in the transmitter to produce the same power output, because the power amplification per stage is much less.

The tank circuit of the second buffer doubler is tuned to 16,000 kc., and, since the excitation voltage of this stage has a frequency of 8000 kc., this stage is also operating as a frequency doubler. Its tank circuit absorbs power from the plate current pulses at the second harmonic frequency of the grid excitation. Assuming that 16,000 kc. is the desired radiation frequency, any following stages will operate as straight amplifiers.

Fig.31 The relation between the plate voltage, grid voltage, and plate current of a doubler stage.



In Fig. 31 are shown a set of curves which illustrate the variation of the plate voltage, the grid excitation, and the plate current of a stage acting as a doubler. Notice that the tank circuit receives a pulse of energy just once in two oscillations of the tank circuit. This is illustrated by the fact that a plate current pulse occurs just once during two cycles of the alternating plate voltage, which is produced by the oscillating tank current. Sometimes the output of a crystal oscillator is insufficient to properly excite a buffer stage to be used as a frequency doubler. In this case, it is advisable to amplify the fundamental frequency of the oscillator by one buffer stage, and then use the output of this stage to excite the doubler. It is possible to amplify the third or fourth harmonic frequency of the plate current pulse instead of the second harmonic, but the power obtained is very small.

Another type of doubler circuit is illustrated in Fig. 32. It is called a "push-push" doubler. Notice that the grids are connected in push-pull, whereas the plates are in parallel. The grid tank, L_1C_1 is tuned to the fundamental frequency or the frequency of the driving stage, and the plate tank, L_2C_2 is tuned to the second harmonic frequency. Since the grids of the tubes are excited out of phase, the two tubes will draw plate current pulses alternately. Thus there will be two plate current pulses flowing into the plate tank for each cycle of the excitation voltage. Also the plate tank will receive energy once each cycle of the second harmonic frequency or twice as often as with the simpler single-tube doubler. The power obtainable from this arrangement is about twice that from a single tube, and with somewhat greater efficiency.

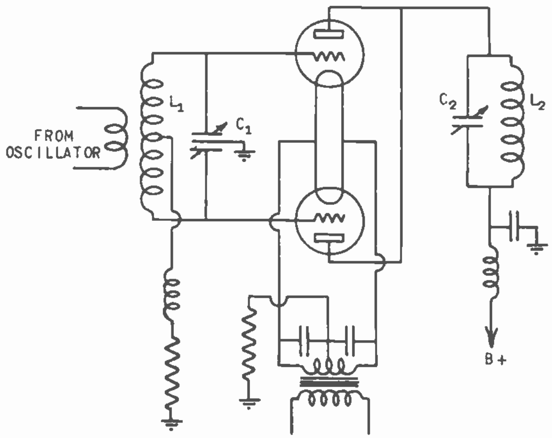


Fig.32 Circuit diagram of push-push doubler.

No neutralization is required since the grid and plate circuits are tuned to different frequencies. It has been determined, however, that self-oscillation may occur unless a split-stator tuning condenser is used in the grid circuit. With a split-stator tuner, the current which would be fed back through the interelectrode capacity of the tube will pass through the upper section of the tuning condenser to ground. Since the capacitive reactance of this section will be fairly low at the harmonic frequency, the feedback current is not able to set up enough voltage in the grid circuit to cause self-oscillation. If the conventional type tuning condenser is employed and the center of the grid tank coil by-passed to ground, then the feedback current will have to flow through the upper half of the grid tank coil to ground, and since the inductive reactance would be higher at the second harmonic frequency, there is a strong possibility that sufficient voltage will be set up across the grid circuit by the feedback current to cause self-oscillation. Thus, a split-stator condenser should be used in the grid circuit.

The push-push doubler may be used to amplify the second, fourth, or any even harmonic of the fundamental excitation frequency, although the efficiency is of course lower when amplifying the higher harmonics. This circuit cannot be used to amplify the third harmonic or any odd harmonic, nor can it be used to amplify the fundamental frequency. Even if the plate tank circuit were tuned to the fundamental frequency, there would be no output if the two tubes are fairly well balanced. This is due to the fact that the plate current pulses of the two tubes flow through the plate tank in the same direction, and would not add energy to the plate tank at the correct instants to maintain an oscillating tank current.

Since the grids are in push-pull, they must be fed by two excitation voltages which are 180° out of phase. The simplest method of obtaining two out-of-phase voltages is to split the grid tank, grounding it at its center with respect to R.F. Thus, the R.F. voltage between the top end of the tank will be 180° out of phase with the voltage between the bottom end of the tank end ground.

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UNIT NO. 3

MATCHING IMPEDANCES

LESSON NO. 4

AN ELECTRIC CITY

.......OPERATED BY AN INVISIBLE FORCE

Very few people realize the marvelous accomplishments that have been achieved in the electrical field. Yet if we were to build a new city, and take full advantage of these accomplishments, it would seem as though we were living in another world.

The highway entering our new city would be brilliantly illuminated by electric lights, eliminating the necessity of glaring headlights and promoting safety. As we neared the city we would call our home on our short wave auto transmitter and converse with our family. If we were particular, we might give a few orders as to how we wanted our steak cooked.

Arriving home, the garage doors would open and close automatically and as we approached the door leading into our home, it would do likewise. Nothing fantastic about this. Just photo-electric cells at work. Our nostrils would be greeted by the savory odor of a steak broiling in our electric oven and we would prepare a cool drink from ice made in our electric refrigerator.

Strolling into the front room we would read the latest news as it was received and printed by impulses from our radio receiver. And as our new city would have a television station, we would enjoy a combination sight and sound program. Reclining on the divan after a hard days work, we wonder how the steak is progressing. Pressing a tiny button we converse with the wife in the kitchen, who reminds us that we must make a phone call to a neighbor. Entering the hallway, the lights are turned on automatically by a photo-cell, and we make our call by radio-telephone much as we would with the ordinary telephone.

Such things are all possible today. And tomorrows accomplishments will make them appear simple. We are living in a truly marvelous age. An age that offers a wealth of opportunity to the young man that is trained to take advantage of them. Midland training prepares you for the opportunities of today and TOMORROW. Let nothing interfere with your progress.

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Lesson Four

MATCHING IMPEDANCES

"To secure maximum efficiency with minimum distortion, it is absolutely essential that vacuum tubes and other associated equipment be properly loaded.

"Correct loading is a process of properly matching impedances between the load and the source. Unless the load is properly matched to its source, the power output realized will be less than the desired value. Since this process of impedance matching is applicable between any source and load, it is at once evident that the information contained in this lesson is vitally important to all phases of Radio and Television study."

1. THE NECESSITY FOR IMPEDANCE MATCHING. At various times throughout previous lessons of this course, you have been confronted with the idea of "impedance matching". The first time this subject arose was in Lesson 28 of Unit 1, wherein the operation of audio power amplifiers was discussed. At that time, it was learned that the maximum possible power output of a tube is obtained when the value of the load impedance is equal to the plate resistance of the tube. The problem of distortion prevents obtaining the maximum power output from a tube, however, and it was found that the maximum power output having an unnoticeable amount of distortion was secured when the load impedance had a value twice as great as the plate resistance of the tube. The actual power in the load circuit under this condition is slightly less than the theoretical maximum.

Unlike voltage amplifiers, a power amplifier must work into a particular value of load impedance, neither too high nor too low, if maximum power is to be secured. In voltage amplifiers, the object is to develop as large an alternating voltage across the load as possible. The larger the value of the load impedance, compared with the plate resistance of the tube, the greater the voltage across the load will be.

A vacuum tube is comparable to an alternator. For purposes of calculation, it is customary to draw an equivalent circuit in which an alternator replaces the vacuum tube. The plate resistance of the tube is represented in this equivalent diagram by a resistor in series with the alternator. This resistor has a value equal to the plate resistance of the tube which the circuit represents.

A voltage amplifier stage is illustrated in Fig. 1. A signal voltage E_s is applied to the grid circuit of this tube. This signal voltage causes the plate current to vary in direct accordance, and the pulsating direct plate current which results, produces a pulsating direct voltage across the load circuit. The DC component of this plate current develops a pure direct voltage across the load which serves no useful purpose. The AC component, however, creates an alternating voltage across the load impedance which is then transferred through the condenser C_1 to the following stage.

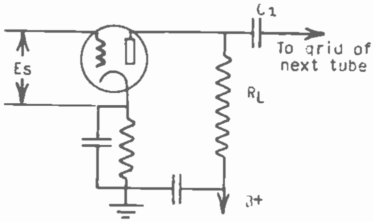


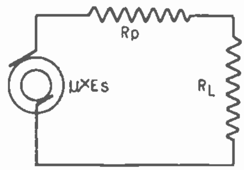
Fig. 1 A voltage amplifier.

The amplification factor of a vacuum tube is a measure of the effectiveness of a plate voltage change in producing a given plate current change, to the change in grid voltage required to produce the same change in plate current. Thus, the signal voltage E_s causes the same plate current change as would a voltage $\mu \times E_s$ if it were applied to the plate of the tube. Although the vacuum tube has both AC and DC voltages applied to its elements, the DC voltages and the DC plate current do not enter into the calculations used to determine the voltage amplification of the tube. For this reason, they are neglected in the equivalent circuit. The peak voltage of the alternator used in the equivalent circuit has the value $\mu \times E_s$ (assuming that E_s is the peak value of the signal voltage applied to the grid circuit of the vacuum tube.)

By inspection of the equivalent circuit shown in Fig. 2, it is seen that the alternator, the plate resistance of the tube and the load impedance are in series. The peak value of the current that flows through this circuit is equal to the voltage of the alternator divided by the total resistance of the circuit. Also, the peak voltage developed across the load is equal to the peak value of this current times the value of the load impedance in ohms. It is easily seen that it is not possible to secure a voltage amplification equal to the amplification factor of the tube, since a part of the voltage in the equivalent circuit is, necessarily, dropped across the plate resistance of the tube. A fundamental principle of series circuits states that the applied voltage will be dropped across the various resistors in direct proportion to the value of the resistors; that is, the largest voltage will be produced across the resistor having the greatest value. If the greater part of the voltage of the alternator is to be developed across the load impedance, the value of the load impedance must be considerably larger than the plate resistance.

Let us now investigate why this condition does not produce maximum power. The peak power in the load impedance is equal to the peak value of the voltage across the load multiplied by the peak value of the current flowing through the load. To develop a large amount of power across the load circuit requires that both the peak voltage across the load and the peak current flowing through the load be as large as possible. When a very large load impedance is used, the voltage across the load will be high, as just explained. Since a high load impedance will produce a comparatively low value of alternating current, the power developed in the load will be only moderate because the product of a high load voltage and a low load current would produce only a moderate amount of power.

Fig. 2 A circuit equivalent to a vacuum tube.



A low value of load impedance would cause a relatively large amount of current to flow through the circuit, since the total resistance of the circuit is less than in the aforementioned case. Should it happen that the load impedance is less than the plate resistance, the greater part of the effective voltage would be dropped across the plate resistance and a smaller portion would be available for the load. Again, the power developed in the load would be relatively small, because the product of a high current through the load and a low voltage across the load would not be very large.

Thus, the problem of securing maximum power into the load is resolved into finding that value of load impedance which produces neither the greatest voltage across the load nor the largest current through the load, but which produces such a value of load voltage and load current that their product is a maximum. To determine the optimum load, let us assign values to the circuit components, and then, using various sized load impedances, calculate the power developed in the load in each case. To simplify the following calculations, it is assumed that the R.M.S. value of the alternator voltage is 100 volts and that the plate resistance is 10,000 ohms. If the first value of load impedance used is 1,000 ohms, the total resistance of the circuit is 11,000 ohms and the current flowing through the circuit as found by Ohm's Law is 9.09 ma. The voltage produced across the load, due to this current, is 9.09 volts, while the power developed in the load impedance is:

$$W = .00909 \times 9.09 \text{ volts}$$

$$= .08 + \text{watts}$$

Let us now calculate the power developed by the alternator and from this value, determine the efficiency. The power developed by the alternator is:

$$W = 100 \text{ volts} \times .00909 \text{ ampere}$$

$$= .90 \text{ watt}$$

The efficiency is, of course, the power developed in the load circuit divided by the total power generated by the alternator. It is:

$$\text{Efficiency} = \frac{.08}{.90}$$

$$= .088$$

$$= 8.8\%$$

By using other values of load impedance, let us construct the following table which shows how the power dissipated in the load and the efficiency vary as the value of load impedance is changed.

LOAD IMPEDANCE	CURRENT	POWER IN LOAD	TOTAL POWER IN CIRCUIT	EFFICIENCY
1,000 ohms	9.09 ma.	.08 W	.90 W	8.8%
5,000	6.6	.22	.66	33⅓%
10,000	5.0	.25	.50	50%
15,000	4.0	.24	.40	60%
20,000	3.3	.22	.33	66⅔%

By plotting the power output and the efficiency against the value of load impedance, we may determine what value of load impedance will produce the greatest power output. These graphs are shown in Fig. 3. It should be noticed that the curve representing the power output rises rapidly at first, reaches a maximum value, and then falls slowly. The efficiency curve rises rapidly as the load impedance is low, and then at a slower rate when the load impedance is higher. It does not, however, reach a maximum value nor does it fall off. It is seen that the power output is maximum when the load impedance is 10,000 ohms, or is equal to the plate resistance, and therein lies the desirability of matching the load impedance to the impedance of the voltage source. Matching impedances, however, should not be construed to mean that the load impedance is always exactly equal to the impedance of the power source. Instead, this statement means that the value of the load impedance is adjusted until the most desirable operating conditions are obtained. Thus, when a load impedance is to be matched to the plate resistance of a vacuum tube, it is not made equal to the plate resistance, but is usually given a value twice as great. This is necessary, as previously explained, to limit the distortion to a point where it is not objectionable.

Notice that the power output into the load drops off rather slowly as the load impedance is increased above that value which gives maximum power. For example, when the load impedance is twice as great as the plate resistance of the tube, the power output is .22 watts compared to .25, the maximum value. Also observe that

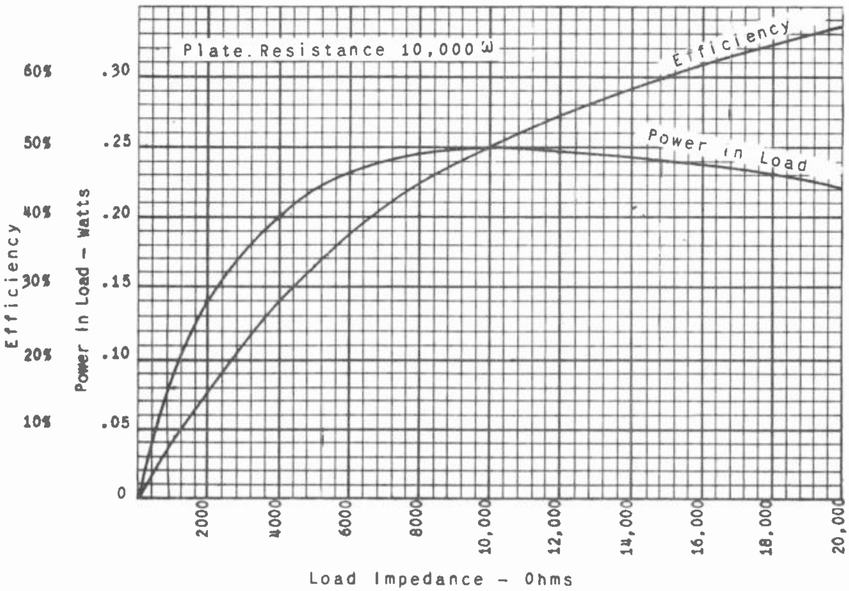


Fig.3 Illustrating the relation between power output and efficiency for various values of load impedance.

the power developed in the load falls off very rapidly as the load impedance is made lower than the value which produces maximum power. At the point of maximum power output, the efficiency is 50%; that is, half of the alternating power developed by the tube is furnished to the load while the other half is dissipated within the tube itself. Under the actual operating conditions when the load impedance is twice as great as the plate resistance, the efficiency increases to 66⅔%. It should be remembered that in this equivalent circuit, we are considering only the alternating power in the calculation of the efficiency. In the actual vacuum tube circuit, DC power is dissipated both in the load resistance and in the plate resistance of the tube. Thus the actual overall efficiency, the alternating power output divided by the total power input to the tube, is considerably less than the values we have calculated. In ordinary power amplifier tubes used in receivers where plate current flows throughout the complete cycle, the efficiency averages 25%. Efficiencies of 50% or greater are possible only when the plate current is allowed to flow during a small part of each cycle. This occurs in oscillators and R.F. amplifiers where the tubes are biased beyond the cutoff value.

In transmitter circuits, it is desirable that the load impedance be greater than the plate resistance of the tube, not because of the distortion which might be produced, but due to the increase in efficiency which is obtained. In commercial power distribution systems, it is desirable that the total load resistance be much larger than the internal resistance of the power source because large amounts of power are involved, and to avoid excessive loss, the efficiency of transmission must be high. On the other hand, where the impedance of a microphone is to be matched to the grid circuit of a vacuum tube amplifier, it is desirable that the load impedance be equal to the power-source impedance because the total power involved is very small and it is necessary to secure all the power possible from the microphone.

It is not always possible to make the load impedance equal to the source impedance. For example, suppose that an audio power amplifier is to furnish power to the voice coil of a dynamic speaker whose impedance is 15 ohms. No vacuum tube has a plate resistance as low as 15 ohms, and thus the power transmitted from the tube to the voice coil would be very low, if some means were not taken to match the two impedances. This action is accomplished by the output transformer whose purpose is to match the impedance of the voice coil to the plate resistance of the tube. Before it is possible to understand how a transformer is able to match a load impedance to a source impedance, it is necessary that the subject of reflected impedance be discussed.

2. REFLECTED IMPEDANCE. A circuit containing an alternator with its internal impedance, a transformer and a load impedance is illustrated in Fig. 4. It is assumed that the resistance of the

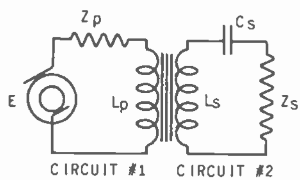


Fig.4 A transformer with a tuned secondary circuit.
$X_{Ls} = X_{Cs}$

transformer windings is negligible. A capacitance Cs is placed in the secondary circuit to neutralize the inductive reactance of the secondary winding. Thus the secondary has no net reactance and the secondary current will be in phase with the voltage causing it to flow. (While this does not illustrate a practical application of an iron-core transformer, its use will be convenient in showing some of the fundamentals of reflected impedance.) The voltage of the alternator will cause a current to flow through its internal impedance and through the primary of the transformer. Let curve A of Fig. 5 represent the current in circuit #1 of Fig. 4. The varying magnetic field produced about the primary of the transformer will induce a voltage into the secondary winding which lags the current of circuit #1 by 90 degrees. The greatest voltage will be induced in the secondary winding at the time that the magnetic field surrounding the primary is changing at its most rapid rate. This, of course, occurs at the time that the current is passing through zero; therefore, the secondary voltage will lag 90 degrees behind the primary current. Curve B of Fig. 5 may then represent the voltage induced into the secondary of the transformer. Since the secondary circuit is purely resistive (has zero net reactance), the secondary current will be in phase with the induced secondary voltage. Thus, curve C in Fig. 5 may represent the secondary current. This current flows through the secondary

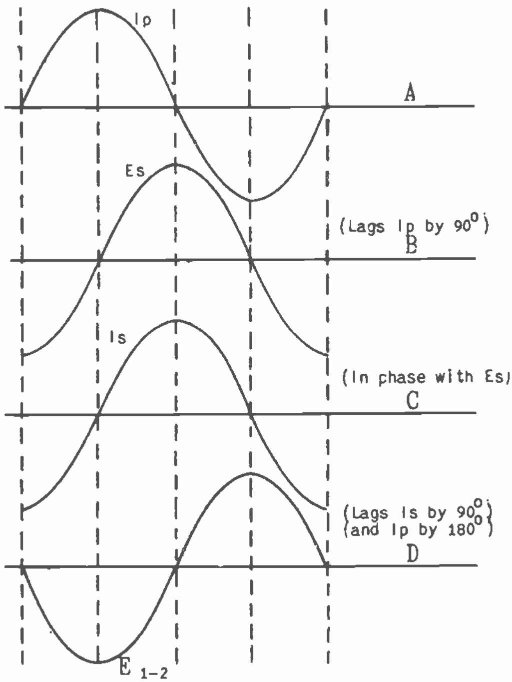
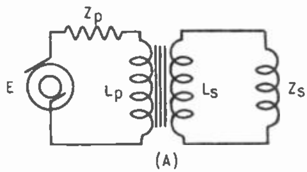


Fig.5 Showing the relation of the currents and voltages of Fig. 4, when the secondary circuit contains only pure resistance.

winding of the transformer and creates a magnetic field around the secondary. In expanding and collapsing, the secondary field will cut through the turns of the primary winding and induce a voltage into the primary which lags the secondary current by 90 degrees. Thus the voltage induced into the primary winding by the secondary current may be represented by curve D of Fig. 5. It is labeled E_{1-2} to signify that it is a voltage induced in circuit #1 due to the current flow in circuit #2.

Notice that the voltage induced in the primary by the secondary current is 180° out of phase with the primary current. Thus, this voltage will oppose the primary current flow at every instant. The primary current is therefore reduced, and the effect is the same as though a resistance had been added in series with the primary. For this reason, it is said that the presence of the secondary circuit has reflected a resistance into the primary circuit. It should, of course, be realized that this reflected resistance has no physical existence, yet this terminology is used because the effect of the secondary on the primary circuit is the same as would be produced by adding a resistance in the primary circuit. It is logical to call this a reflected resistance, since it has no effect upon the phase angle between the primary voltage and the primary current.



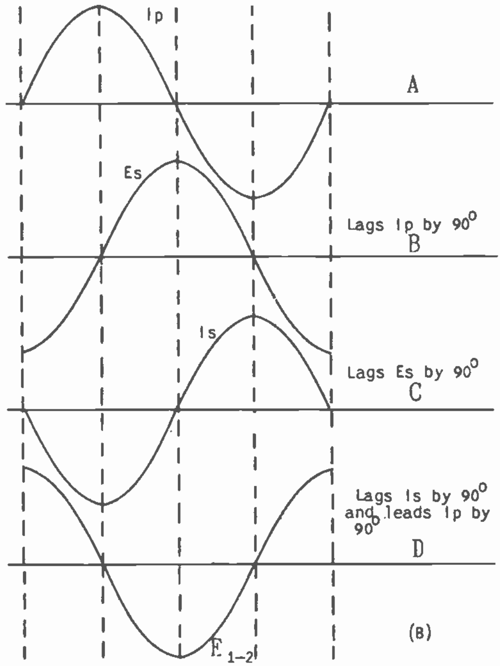
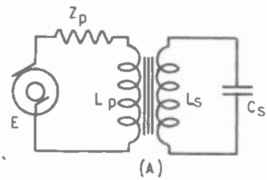


Fig.6 (A) A circuit that has a purely inductive secondary. (B) Illustrating the relations of the currents and voltages of Fig. 6A.

For the sake of argument, let us now assume that the secondary circuit contains only inductance, such as the circuit shown in Fig. 6A. Such a condition would cause the secondary current to lag 90° behind the secondary voltage. If curve A of Fig. 6B represents the primary current, and curve B the secondary voltage, then the secondary current which lags 90° behind the secondary voltage could be represented by curve C. This secondary current will induce a voltage into the primary which lags 90° behind the secondary current; therefore, this voltage may be represented by curve D of Fig. 6B.

By close examination of curves A and D of Fig. 6B, it may be seen that the voltage induced into the primary by the secondary current (curve D) leads the primary current (curve A) by 90°.

There is, then, in the primary circuit a voltage which leads the current by 90°. Such a voltage would also be produced by allowing the primary current to flow through a capacitance.[1] It is, therefore, within the bounds of reason to state that the presence of the secondary has reflected a capacitive reactance into the primary circuit, since the voltage induced into the primary has the same effect as would be produced by connecting a condenser in series with the primary circuit.



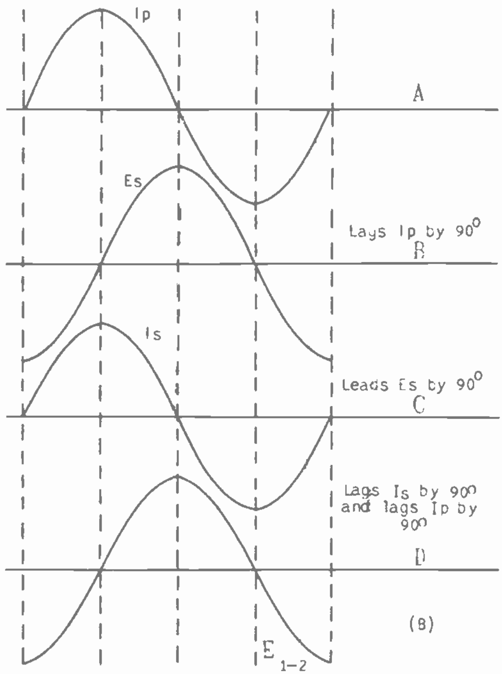


Fig. 7 (A) A circuit containing a predominance of capacitive reactance in the secondary. X_{Cs} is greater than X_{Ls}. (B) Showing the phases of the voltages and currents of Fig. 7A.

To make this explanation complete, let us now assume that the secondary circuit contains no resistance and has a predominance of capacitive reactance, such as the circuit illustrated in Fig. 7A. In this figure, X_{Cs} is greater than X_{Ls}. Curve A of Fig. 7B may represent the primary current, and curve B, the secondary voltage. Since the secondary circuit contains only capacitive reactance and no resistance, the secondary current will lead the secondary voltage by 90°. Thus, curve C of Fig. 7B may portray the secondary current. This current induces a voltage into the primary, which lags it by 90°, and so curve D may represent the voltage induced into the primary by the secondary current.

By inspection of curves A and D of Fig. 7B, it is evident that the voltage induced into the primary by the secondary current

[1] That is, the counter-voltage across a condenser leads the current 90°. See Sec. 12, Lesson 13, Unit 1.

(curve D) lags the primary current (curve A) by 90°. Such a voltage would also be produced in the primary by allowing the primary current to flow through a pure inductance.[1] For this reason, it is permissible to state that the effect of the secondary circuit is to reflect an inductive reactance into the primary circuit.

It should now be apparent that the effect of the secondary circuit upon the primary circuit depends on the nature of the opposition to current flow in the secondary. A pure resistive secondary will reflect a pure resistance; a pure inductive secondary will reflect a pure capacitance; and a pure capacitive secondary will reflect a pure inductance.

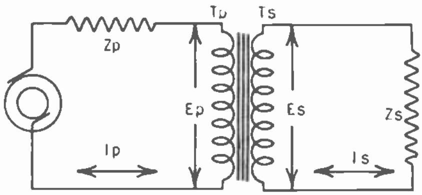


Fig.8 The circuit used to derive the formula for reflected impedance.

When the secondary circuit contains both resistance and inductive reactance, its effect upon the primary circuit is to reflect an impedance into the primary which consists of both resistance and capacitive reactance. Thus, if the relation of inductive reactance to resistance in the secondary were such that the secondary current lagged the secondary voltage by 45°, the voltage induced into the primary by the secondary current would lead the current in the primary by 45°, and the result would be the same as though a resistance and a capacitive reactance were connected in series with the primary.

In a like manner, a secondary containing both resistance and capacitive reactance would reflect an impedance into the primary having the nature of a resistance and an inductive reactance.

It is, of course, impossible for the secondary current to be in phase with the secondary voltage, unless there is no net reactance in the secondary circuit. Since the secondary winding of the transformer is highly inductive, this is possible only when the circuit is tuned so that the capacitive reactance is equal to the inductive reactance. Iron-core transformers such as used for audio and power frequencies are never tuned, and so the secondary circuit always has a predominance of inductive reactance. This causes the secondary current to lag the secondary voltage, and as previously explained, when this condition prevails, the reflected impedance consists of a resistance and a capacitance. If the correct size of transformer is employed, the amount of capacitive reactance reflected from secondary into primary is sufficient to cancel the inductive reactance of the primary winding, whereas the resistance reflected from secondary into primary is of the proper

[1]The counter-voltage across an inductance lags the current 90°. See Sec. 12, Lesson 13, Unit 1.

value to match the impedance of the power source. Thus, the power source works into a pure resistance instead of an inductive reactance as might be supposed. The value of the resistance reflected from the secondary circuit into the primary depends upon the size of the load resistor connected in the secondary circuit and upon the turn ratio of the transformer.

Using the circuit shown in Fig. 8, let us determine what the turn ratio should be to reflect enough resistance into the primary circuit so that the internal impedance of the alternator will be matched to the primary of the transformer. If the correct turn ratio is used, maximum power will be delivered to the load resistance, and the efficiency of power transfer will be 50%. That is, under this condition, half of the total power developed by the alternator will be dissipated in the internal impedance of the alternator, and the other half will be delivered to the load. The following symbols will be used in this discussion:

I_p is the primary current,
I_s is the secondary current,
Z_p is the internal resistance of the alternator,
Z_s is the resistance of the load,
T_p is the number of turns on the primary winding,
T_s is the number of turns on the secondary winding.

The power dissipated in the internal resistance of the alternator is equal to the primary current squared times the alternator's resistance:

$$\text{Power dissipated in alternator} = I_p^2 \times Z_p \qquad (1)$$

The power delivered to the load circuit is equal to the secondary current squared times the resistance of the load:

$$\text{Power delivered to load} = I_s^2 \times Z_s \qquad (2)$$

When maximum power is being transferred, the power in the load will be equal to the power dissipated in the alternator or:

$$I_p^2 \times Z_p = I_s^2 \times Z_s \qquad (3)$$

Dividing equation (3) by Z_p, the following is obtained:

$$I_p^2 = I_s^2 \times \frac{Z_s}{Z_p} \qquad (4)$$

Now if equation (4) is divided by I_s^2, there results:

$$\frac{I_p^2}{I_s^2} = \frac{Z_s}{Z_p} \qquad (5)$$

Taking the square root of both sides of this equation, we obtain:

$$\frac{I_p}{I_s} = \sqrt{\frac{Z_s}{Z_p}} \qquad (6)$$

In Lesson 10, Unit 1, it was learned that the ratio between the primary and secondary currents is equal to the inverse ratio of the turns, or that the primary current is in the same ratio to the secondary current as the secondary turns are to the primary turns. This may be expressed as follows:

$$\frac{I_p}{I_s} = \frac{T_s}{T_p} \tag{7}$$

If the value for I_p/I_s, as given in equation (7), is substituted into equation (6), the following results:

$$\frac{T_s}{T_p} = \sqrt{\frac{Z_s}{Z_p}} \tag{8}$$

From this last equation, it is evident that, for proper operation, the turns ratio should be equal to the square root of the ratio between the secondary and primary impedances.

If equation (8) is squared, it then becomes:

$$\frac{T_s^2}{T_p^2} = \frac{Z_s}{Z_p} \tag{9}$$

This is really a proportion, since it states the equality between two ratios. To solve this proportion for Z_p, a fundamental principle of proportions is used. This principle states that when the product of the means is divided by one extreme, the result gives the other extreme. Performing this operation, this results:

$$Z_p = \frac{T_p^2 \times Z_s}{T_s^2} = Z_s \times \left(\frac{T_p}{T_s}\right)^2 \tag{10}$$

This last equation states that the primary impedance is equal to the load impedance multiplied by the square of the turns ratio. Since maximum power is being developed in the load, the resistance reflected from the secondary into the primary must be equal to the internal impedance of the alternator; therefore, the value of the reflected resistance must be equal to:

$$Z_r = Z_s \times \left(\frac{T_p}{T_s}\right)^2 \tag{11}$$

Where Z_r is the resistance reflected from secondary into primary.

In addition to this reflected resistance, the secondary circuit reflects a capacitive reactance into the primary which is just large enough to cancel the actual inductive reactance of the primary winding. (This is true only if the actual inductive reactances of the primary and secondary windings are large compared to the resistance of the alternator and that of the load.)

Thus, in effect, the alternator is working into a pure resistance equal to its internal resistance. However, the alternator does not dissipate any power in this resistance, since the resistance is only hypothetical and has no physical existence. Actually, all of the power not dissipated in the alternator itself is transferred to the secondary circuit, and is thus available for the load.

With the load connected across the secondary of the transformer, the actual impedance measured across the terminals of the primary with the alternator disconnected will be equal to the internal resistance of the alternator, if the correct turns ratio is used.

For example, suppose that the alternator has an internal impedance of 400 ohms and that the load impedance into which power is to be developed is 100 ohms. If the 100-ohm load resistor were connected directly across the terminals of the alternator, maximum power could not be developed into the load; for, as previously stated, to develop maximum power in a load impedance requires that the value of the load impedance be equal to the internal impedance of the voltage source. The ratio between the impedance of the voltage source and that of the load resistor is 400:100 or 4:1. The transformer used to match the load impedance to the alternator should have a turns ratio equal to the square root of this impedance ratio. The impedance ratio is 4:1 and the square root of this value is 2:1. Thus, the primary winding of the transformer should have twice as many turns as the secondary winding. Using a transformer with such a turns ratio, the impedance reflected from the secondary into the primary will be 400 ohms and maximum power will be transferred from the primary circuit to the secondary circuit. Let us prove that this is the case.

Let us assume that the no-load voltage of the alternator is 200 volts. (This is the voltage across the terminals of the alternator when it is not delivering any current.) The total resistance of the primary circuit is the sum of the internal resistance of the alternator (400 ohms) plus the resistance measured across the terminals of the transformer primary. (This is also 400 ohms, since this is the resistance reflected from secondary to primary. The actual primary inductance is practically cancelled by the capacitance which is reflected back.) Thus the 200 volts generated by the alternator will force through this 800 ohms of resistance a current which has this value:

$$I_p = \frac{200}{800} = .25 \text{ ampere.}$$

The primary current flowing through the reflected impedance of the primary will cause a voltage drop of:

$$E = .25 \times 400 = 100 \text{ volts.}$$

Since the secondary winding has just half as many turns as the primary, the secondary voltage will be one-half of the primary voltage or 50 volts. Therefore the power delivered to the load is:

$$\frac{E^2}{R} = \frac{50^2}{100} = \frac{2500}{100} = 25 \text{ watts.}$$

If maximum power has been transferred to the load, the power in the load should be equal to that dissipated in the internal resistance of the alternator, and the efficiency should be 50%. The power dissipated in the alternator is:

$$W_p = I_p^2 \times Z_p = (.25)^2 \times 400 = 25 \text{ watts.}$$

The total power generated by the alternator is the sum of that dissipated in the alternator and that delivered to the load, or 50 watts. Since the power delivered to the alternator is one-half of the total power developed, the efficiency of transfer is 50%, and the power transferred is maximum. We have thus transferred maximum power to the load, even though the load impedance was not equal to the internal impedance of the voltage source.

The actual resistance of the primary winding is comparatively low, while the actual impedance (inductive reactance) across the primary winding, with the secondary open, is very large. It should be realized that the impedance, looking into the primary of the transformer, or across the primary terminals of the transformer, when the secondary circuit is open is not the same value that would be measured between these terminals when a load resistor is connected across the secondary. The presence of the load resistor across the secondary circuit causes a secondary current to flow and the impedance reflected back into the primary circuit consists of both a resistance and a capacitance. The capacity reflected back is sufficient to cancel the inductive reactance of the primary, while the resistance reflected from secondary to primary is equal to the impedance of the load resistor times the square of the turns ratio from primary to secondary.

The fact that connecting the load resistor across the secondary terminals actually reduces the impedance of the primary is easily demonstrated by connecting the primary of a power transformer across a 110 volt AC supply line, and measuring the primary current that flows when the secondary circuit is open. It will be found that under this condition, the primary current is very low, indicating that the impedance of the primary is very high. When a moderate sized resistor is connected across the secondary of the power transformer and the primary current is again measured, it is found to have increased considerably. An increase in primary current can only mean that the impedance of the primary has been reduced since obviously the supply voltage has not changed. The lower the resistor connected across the secondary circuit, the smaller the reflected resistance will be; and, in turn, the more the impedance of the primary will be reduced and the larger the primary current will become.

Let us consider another example. A power amplifier tube having a plate resistance of 1125 ohms is to deliver power to the voice coil of a dynamic speaker with an impedance of 10 ohms. If the 10-ohm voice coil were connected in the plate circuit of this tube, very little power would be furnished to the voice coil. However, by using a matching transformer, a large amount of power may be transferred from the plate circuit of the tube to the coil. To develop maximum undistorted power, the tube should work into a load impedance twice as great as its plate resistance; that is, into 2250 ohms. Thus the primary should present an impedance of 2250 ohms for the tube to work into, when there is a 10-ohm load connected across the secondary. In this case Z_p is 2250 ohms and Z_s is 10 ohms. Equation (8) gives the relationship between the turns ratio from secondary to primary as compared to the ratio between the load and source impedances. It is also possible to express the turns ratio from primary to secondary as compared to the ratio between the source and load impedances. This would be:

$$\frac{T_p}{T_s} = \sqrt{\frac{Z_p}{Z_s}} \qquad (12)$$

Therefore, either equation (8) or (12) may be used in solving this problem, it being usual to select that equation which gives a ratio greater than one. In this case, equation (12) is more convenient.

$$\frac{T_p}{T_s} = \sqrt{\frac{2250}{10}} = \sqrt{225} = 15$$

Thus, the turns ratio is 15 to 1, or there should be fifteen times as many primary turns as secondary turns. Since the secondary of this transformer has fewer turns than the primary, it is obviously a step-down transformer. The voltage produced across the secondary will be only 1/15 as much as that which appears across the primary. This would seem to be a disadvantage, but it should also be realized that while the transformer steps down the voltage, it also steps up the current that may be drawn from the transformer. Thus the current that flows in the secondary circuit is 15 times as great as that flowing in the primary circuit. An average value for the DC plate current of a power tube would be 30 ma. Under excitation, the current would probably increase to a peak of 50 ma. and fall to a minimum of 10 ma. Thus it is changing 20 ma. in each direction from its no-signal value. From this, it is evident that the peak value of the AC plate current component is 20 ma. If the voice coil were connected in the plate circuit, the 20 ma. (peak value) of alternating current flowing through it could not produce much power in the voice coil. However, by using the matching transformer, the alternating current is stepped up 15 times. Thus, the current flowing through the secondary has a peak value of 300 ma. This amount of current is able to produce appreciable power in the voice coil. A current of 300 ma. peak corresponds to an RMS current of 212.1 ma. Thus, the power developed in the 10-ohm voice coil is:

$$W = I^2R = (.2121)^2 \times 10 = .45 \text{ watt}$$

If the 10-ohm voice coil were connected directly in the plate circuit of the power tube, the RMS current through it would be .707 × 20, or 14.14 ma. This would develop a power of:

$$W = I^2R = (.01414)^2 \times 10 = .002 \text{ watt}$$

It is therefore clear that a much larger amount of power is furnished to the voice coil by using a matching transformer than would be produced if the voice coil were connected into the plate circuit. A magnetic speaker, on the other hand, has a much higher impedance (several thousand ohms), and may be connected directly to the power tube without using a matching transformer. In this case, the impedance of the speaker is close enough in value to the plate resistance of the power tube that a large amount of power is transferred.

In the previous examples, it has been assumed that the matching transformers used did not of themselves introduce any power loss in the circuit; that is, they were ideal transformers. Naturally, such perfect conditions are not obtained in actual practice. Every transformer introduces some losses. These losses are due to several causes. One is known as the "copper loss". This is merely the loss due to the power dissipated in the resistance of the primary and secondary windings. Of course, any transformer will have some resistance because it is composed of conductors, and all conductors possess resistance. Whatever power is dissipated in the resistance of the windings of the transformer is not available to do useful work in the load impedance of the circuit. The copper loss of the primary would be directly proportional to the square of the primary current, and that of the secondary would vary as the square of the secondary current.

A second source of loss is the power dissipated in the core of the transformer. This power loss is due to two factors; the eddy current loss, and the magnetic hysteresis loss. It is realized that the magnetic flux will cut through the core material and induce a voltage therein. This voltage will cause currents to circulate in the core. These eddy currents, as they are called, will produce a power loss which is equal to the square of the current in the core times the core resistance. To minimize this loss, the core resistance is arbitrarily increased. That this will reduce the loss is easily seen by considering what would happen if the core resistance were doubled. With the same induced voltage, the eddy currents would become half as great. As may be determined from the I^2R law, half as much current flowing through twice as much resistance would dissipate just half as much power.

Since the core must be composed of a material having high permeability, the only convenient means of increasing the core resistance is to build the core from thin sheets of transformer iron, known as laminations. Each lamination is lacquered on both sides, and is thus electrically insulated from those adjacent to it. This construction reduces the possible paths for eddy currents, and effectively increases the resistance.

The second core loss, "magnetic hysteresis" is, in many respects, comparable to the dielectric hysteresis loss of an insulator. It is well known that when an insulator is placed in a rapidly varying electrostatic field, such as a dielectric between the two plates of a condenser, a loss will occur. This loss is due to the energy required to strain the orbital electrons of the material first toward one plate and then the other. When the voltage across the condenser falls to zero, the dielectric electrons do not immediately return to their normal positions, and a part of the succeeding alternation is necessary to return them to normal before they can be strained in the opposite direction. This hysteresis or lag of the electrons constitutes a loss which increases rapidly with frequency. In a like manner, when an iron core is subjected to a rapidly varying magnetic flux, the magnetism in the core is not completely lost as the current drops to zero, and a part of the succeeding alternation is required to rid the core of this magnetism before it can be magnetized with opposite polarity. This rapid reversal of the minute molecular magnets produces heat and is, therefore, a power loss, and one which increases rapidly with frequency. Not all types of iron have the same magnetic hysteresis, and this loss is made as small as possible by selecting for the core a material in which hysteresis loss is a minimum.

Assuming that the frequency is constant, it should be apparent that the eddy current loss and hysteresis loss will be constant with a constant flux density in the core. Of course, if the flux density increases, both of the losses will become greater. A larger flux density signifies that the total magnetism in the core is greater on each alternation of the current, and more of the molecular magnets would have to be moved, thereby increasing the hysteresis loss. Likewise, since the flux would be varying through a greater range, the voltages induced in the core would increase, and the eddy current loss would be greater.

When voltage is applied to the primary of an iron-core transformer, and the secondary is open, the primary current will be quite small. This, of course, is due to the fact that the inductive reactance of the primary is large. This no-load current drawn by the primary is often called the magnetizing current, since it creates in the core a certain flux density as determined by the design of the transformer. The flux density developed will depend on the permeability of the core and the total mass of the core; and in addition, the amount of magnetizing current and the number of turns on the primary. By choosing a primary with a large number of turns, the same number of ampere-turns (magnetizing force) may be developed with a smaller no-load primary current. This, of course, will lessen the copper loss of the primary with no load, but will not affect the core losses for a given flux density. Most iron-core transformers will work into resistive loads, and, as mentioned before, the inductive reactance of the secondary winding will be large in comparison to the load resistance. This being so, the secondary current will lag nearly 90° behind the secondary voltage and will, therefore, be nearly 180° out of phase with the primary current. Thus, when a load is placed across the secondary of the transformer, the primary impedance is reduced, and an increased primary current flows. This greater primary current may be considered as having two components; one, the magnetizing current, and the other the load current. Both of these components would tend to magnetize the core, but the secondary current creates a magnetomotive force which is exactly equal and opposite to that developed by the load current of the primary. Thus, the flux density of the core is that due to the magnetizing component of the primary current, and, as a result, the flux density of the core does not change when a load is connected across the secondary of the transformer. An increased load would cause more primary current to flow, but it would also result in more secondary current, thereby keeping the flux density the same, since these two currents create equal but opposite magnetizing effects. The conclusion is that the core losses of an iron-core transformer are independent of changes in the load.

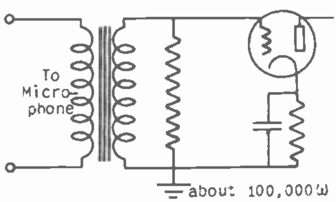
Nearly any well-designed iron-core transformer will be at least 90% efficient. This means that the power available from the secondary will be at least 90% of that furnished to the primary.

Realizing that any transformer will introduce some loss into the circuit, it is apparent that there are occasions when it is inadvisable to use a matching transformer. When the load resistor into which power is to be transferred, is very nearly equal to the internal impedance of the power source, the loss due to the mismatch of the impedances is probably less than the loss which would occur if a matching transformer were used. The loss occasioned by the matching transformer is known as its "insertion" loss. When the ratio between the load impedance and the internal impedance of the power source is not more than 2:1 nor less than 1:2, the presence of a matching transformer would probably cause more loss than the mismatch between the two impedances would produce if no transformer were used. In many such instances, a transformer would not be employed. This, however, is not universally true, because there are cases where a mismatch will not only prevent maximum power from being delivered to a load, but will cause frequency discrimination. In these instances, it is desirable to use a matching transformer to prevent this frequency discrimination, even though the insertion loss of the transformer is greater then the loss due to a mismatch would be.

It has been assumed that the iron-core transformers under discussion had 100% coupling between primary and secondary. This means that all of the flux of the primary links every turn of the secondary and all of the flux due to the secondary current links every turn of the primary. This is approximately true because practically all of the flux is confined to the iron core. In actual transformers, the coupling will be 90% or greater, which indicates that of the total flux developed by primary and secondary, at least 90% will cut through both windings. That part of the magnetic field which does not link both primary and secondary is known as "leakage flux". Formula (8) assumes that there is no leakage flux, and if this is true, and the proper turns ratio is employed; then, the capacitive reactance reflected into the primary is just sufficient to cancel the inductive reactance of the primary, and the resistance reflected is of the proper amount to load the power source. If, however, the coupling is not 100%, then the capacitive reactance is insufficient to cancel the inductive reactance of the primary, and the reflected resistance is somewhat smaller than the optimum value for maximum power transfer. The primary, therefore, has a small net inductive reactance which is known as "leakage reactance", and the power source is not working into a purely resistive load. It should be understood that this condition does not represent a power loss in the same sense as the copper and core losses of the transformer, yet it does prevent the attainment of optimum operating conditions. High quality transformers have very little leakage reactance, whereas, cheaper types may have considerable. The effect is to prevent maximum power transfer and to cause frequency distortion, since the load on the power source would vary considerably with frequency.

3. TRANSFORMER RATINGS. In the manufacturers' specifications of matching transformers, the turns ratio is not ordinarily given. Instead, it is the usual practice to state the impedances into which the primary and secondary should be operated. For example, the transformer used to match a double button carbon microphone to the grid circuit of an amplifier stage would probably be listed as having a primary impedance of 200 ohms and a secondary impedance of 100,000 ohms. It should be understood that this does not mean that the actual impedance of the primary winding is 200 ohms, or that the impedance of the secondary is 100,000 ohms. Instead, the significance of the statement is that when the secondary of the transformer is connected in the grid circuit of a vacuum tube, the actual impedance measured between the terminals of the primary winding would then be 200 ohms and, therefore, should be used with a 200-ohm microphone. If no grid current flows, the actual impedance of the grid circuit is very high; in fact, it is limited only by the capacitive effect between the grid and filament. In order to improve the frequency response, it is usual practice, especially in microphone amplifiers, to connect a resistor of 100,000 ohms or higher across the secondary of the transformer. This action definitely sets the value of the impedance into which the transformer works, and thus makes for better operation. Such a circuit is illustrated in Fig. 9.

Fig.9 Fixing the impedance of the secondary circuit by connecting a high value resistor across the secondary winding.



Transformers used to match power tubes to dynamic speakers will have primary impedances ranging from 1500 to 10,000 ohms or higher, depending upon the type of power tube employed; while the secondary winding of the output transformer is usually tapped so that proper connections can be made to dynamic speakers having voice coils of various common impedances such as 4, 8, 15 ohms, etc.

Although a transformer may have a rating of 10,000 ohms primary and 500 ohms secondary, it should be realized that this same transformer will give equally good results when worked between other impedances having the same ratio, as long as these impedances do not differ too much from the rated values. Satisfactory results would be obtained when this transformer is worked between 20,000 ohms and 1000 ohms, for these values are in the same ratio as 10,000 and 500. It would also be possible to use this transformer between 5000 ohms and 250 ohms. However, an attempt to connect a generator of 100 ohms impedance to the primary of this transformer and a load of 5 ohms to the secondary would be impractical, even though the impedance ratio is correct. Let us determine why this action would not produce efficient results.

In the derivation of the formula relating the turns ratio to the impedance ratio, two assumptions were made. The first was that neither the primary nor secondary winding of a transformer had any resistance, while the second was that the primary and secondary inductive reactances, although having a definite impedance ratio, were themselves very large; that is, practically infinite. If the resistance of the windings is very low compared with the impedances into which they work, and if the inductive reactances of the primary and secondary are respectively very large compared with the source and load impedances, the errors produced by these assumptions are negligible.

This transformer, which has been under discussion was designed to be used between 10,000 and 500 ohms. Therefore, sufficient turns were placed on the primary winding to make the inductive reactance of the primary large in comparison with 10,000 ohms. To be efficient, the primary inductive reactance should be at least 100,000 ohms, or 10 times as large as the source impedance. Since the secondary of this transformer is to work into 500 ohms, it should have an inductive reactance of at least 5000 ohms. With the size of wire commonly used to wind transformers, the primary resistance would probably be 200 ohms or more, while the secondary resistance would amount to at least 10 ohms. To attempt to work this transformer between a source impedance of 100 ohms and a load impedance of 5 ohms would produce inefficient results, because the actual resistance of the primary winding (200 ohms) is about the same size as the impedance of the source (100 ohms). Thus, most of the power developed by the voltage source would be dissipated in the primary resistance and very little would be transferred to the secondary circuit. What little did reach the secondary would not be all available for the load, for a large part of it would be dissipated in the resistance of the secondary winding.

In a like manner, it would not be feasible to operate this transformer between a source impedance of 200,000 ohms and a load impedance of 10,000 ohms, even though these impedances have the correct ratio. In this case, the primary and secondary inductive reactances would not be as large as the source and load impedances, and the equations which were derived on the assumption that the reactances of the transformer were much larger than the impedances between which it is worked would not hold true. Unless these assumptions are correct, the resistance reflected from the secondary to the primary is less than the calculated value; and the capacitive reactance, which is reflected, is insufficient to properly neutralize the inductive reactance of the primary. Thus, the voltage source does not work into a pure resistance, but into a resistance plus an inductive reactance, and maximum power is not transferred to the load.

The actual inductive reactance of the transformer windings will, of course, depend on the frequency of the current flowing through them; however, a standard frequency of 400 cycles is usually used in determining the rating of the transformer. At frequencies higher than this value, the reactance will be greater, and the power transfer will be even more efficient.

The frequency range over which the transformer will operate efficiently is usually included in the manufacturer's ratings. The lower end of the frequency range is determined by that frequency at which the inductive reactances of the primary and secondary are so low compared to the generator and load impedances that very little power is transferred from primary to secondary. On the other hand, the upper end of the frequency range is determined by the distributed capacity which, of course, by-passes the high-frequency current and thus prevents its transfer from one side of the transformer to the other.

It should be understood that the purpose of a transformer is to match a load impedance to a source impedance in order that maximum power may be transferred from the source to the load. Matching transformers are widely used in all audio equipment. For example, a matching transformer is used in a broadcast studio to match the output of the microphone to the grid circuit of the first amplifier stage; while the output of the amplifier is matched to a telephone line by means of another matching transformer. The monitor speaker used to determine the quality of the program must also be matched to the output of the amplifier. The telephone lines that carry the programs from remote points must be matched to the input of the amplifier. Several important points concerning transformers which should be remembered are:

1. The primary impedance of a transformer depends on the secondary load and upon the turns ratio. The proper turns ratio is equal to the square root of the ratio between the load and source impedances.
2. A transformer may be used between other impedances than those for which it is rated, as long as the variation is not too far from the manufacturer's rating, and with the knowledge that the original ratio will still apply.
3. The actual inductive reactance of the primary and secondary windings should be large compared to the impedances into which these windings are to be worked.
4. The losses of a transformer are the copper loss, which varies directly with the square of the currents; and the core loss which is independent of the currents.
5. Coupling less than 100% causes leakage reactance, which is not a power loss, but does prevent the transfer of maximum power.

4. MATCHING IMPEDANCES AT RADIO FREQUENCIES. Matching impedances at radio frequencies differs in several respects from impedance matching at audio frequencies. In the first place, an audio frequency transformer is called upon to provide a correct match over a very large band of frequencies; a band which may well range from 30 to 10,000 cycles, in which case, the ratio between the limiting frequencies is greater than 300 to 1. To be satisfactory over such a range, the transformer must have high quality iron in its core, and low resistance and high inductance windings. The distributed capacity must be low and the coupling should be very tight so that the leakage flux will be minimum.

Ordinary iron-core transformers are not practical at radio frequencies. The core losses, both eddy current and hysteresis, increase very rapidly as the frequency is raised, thereby precluding the use of such transformers in the radio frequency range. There are available so-called iron-core R.F. transformers, in which the core consists of a mixture of clay and finely divided particles of iron. These have proved very satisfactory, because the iron particles increase the permeability of the core and therefore, fewer turns are needed for the windings to produce a given inductance. The fewer turns mean less distributed capacity and greater efficiency which more than compensates for the small core losses. In general, however, iron-core R.F. transformers and coils are in the minority; most R.F. coils in use in receivers and transmitters being of the air-core type.

With air-core coils, 100% coupling between primary and secondary is never possible; in fact, it is not even desirable. In many instances, the coupling may be 1% or even less. It is at once evident that R.F. transformers are decidedly different from A.F. transformers. In an R.F. transformer, it is possible and desirable that the amount of coupling between primary and secondary be small because the transformer operates over a very narrow frequency range and may, therefore, be tuned. When the R.F. is modulated, it consists of a carrier and sideband components; but even so, the ratio between the highest and lowest frequencies is nearly one, and the band is comparatively narrow. It is easily possible to tune either primary or secondary or both, and in such an instance, the requirements for impedance matching are different in some respects from those for iron-core A.F. transformers.

In an air-core transformer, the magnetic flux linking the primary winding is not all effective in inducing a voltage in the secondary; that is, there is a considerable amount of leakage lines or leakage reactance. For this reason, the ratio of the secondary and primary voltages is much less than the actual turn ratio of the transformers. Therefore, the turn ratio of an R.F. transformer is not an indication of the ratio of voltage transformation. Due to the fact that an air-core transformer has considerable leakage reactance, the mutual inductance between the primary and secondary is much less than that of an iron-core transformer.

Equation (11), which gives the amount of reflected resistance for iron-core transformers, could not be used to calculate the reflected impedance of an R.F. transformer, because of the large amount of leakage inductance involved. Instead, another formula must be derived that will give the reflected impedance, and which takes into account the fact that the voltage ratio is less than the turn ratio.

To introduce this new concept, let us review briefly the subject of inductance. We know that by definition, the inductance of a coil is the ability of that coil to induce a voltage within itself when the current through it is changed. Thus, the term "inductance", as normally employed, is merely a shortened form of "self inductance". The unit of measurement of inductance is the henry, and we should remember from a previous lesson, that a coil is said to have an inductance of one henry if there is developed across it a self-induced voltage of one volt, when the current through the coil changes at the rate of one ampere per second.

In a like manner, there exists mutual inductance between the primary and secondary of a transformer, and this mutual inductance is a measure of the ability of the primary to induce a voltage into the secondary, or the ability of the secondary to induce a voltage into the primary. That is to say: the mutual inductance between primary and secondary is equal to that between secondary and primary. This mutual inductance is also measured in henries, and, if there exists a mutual inductance of one henry between two coils, then a current change at the rate of one ampere per second through either winding will induce a voltage of one volt into the other.

The mutual inductance between two coils depends upon the self-inductance of the primary, the self-inductance of the secondary, the distance between the windings, the angle which their axes form, and the permeability of the core between them. Increasing the inductance of either the primary or secondary or both will raise the mutual inductance between the two coils. Moving the two coils closer together increases the mutual inductance. With other factors constant, the mutual inductance will be maximum when the axes of the two coils are parallel, and minimum when the axes are perpendicular. Finally, the use of an iron core will increase the mutual inductance; but it is not practical at radio frequencies.

The relationship between the mutual inductance, the amount of coupling between the two coils, and the self-inductances of the coils is given in the following equation:

$$M = K\sqrt{L_p \times L_s} \qquad (13)$$

Where: M is the mutual inductance in henries,
L_p is the self-inductance of the primary in henries,
L_s is the self-inductance of the secondary in henries,
K is the coefficient of coupling.

In the above formula, the factor k needs further explanation. If all of the flux cuts through every turn of the primary and secondary, then the coefficient of coupling K is one, or the two coils are said to be 100% coupled. If less than the total flux links the two coils, K is smaller than one, and may be as small as a few thousandths in some R.F. circuits. Thus, the factor K takes into account the distance between the two coils as well as their relative positions. Note that in an iron-core transformer where K is practically one, the mutual inductance would be equal to the square root of the product of the inductances of the primary and the secondary. This is, of course, the maximum value that the mutual inductance can have. As an example of the use of formula (13), suppose that the primary coil has an inductance of 20 microhenries, the secondary an inductance of 40 microhenries, and the coefficient of coupling is .04. Determine the mutual inductance. Before substituting in formula (13), it would be well to point out that the inductance in the formula may be substituted in microhenries if desired; in which case the mutual inductance will be in microhenries. In this problem:

$$M = .04\sqrt{20 \times 40} = .04 \times \sqrt{800} = .04 \times 28.28 = 1.13 \ \mu hy.$$

Formula (13) can be solved for the coefficient of coupling (K); it then becomes:

$$K = \frac{M}{\sqrt{L_p \times L_s}} \qquad (14)$$

This formula is useful in determining the amount of coupling when the self-inductances of the two coils, as well as the mutual inductance between them, is known. For example, suppose that the mutual inductance between two coils is 10 microhenries; the self-inductance of the primary, 20 microhenries; and the self-inductance of the secondary, 30 microhenries. The coefficient of coupling would be:

$$K = \frac{10}{\sqrt{20 \times 30}} = \frac{10}{24.5} = .408$$

Note that the coefficient of coupling does not have a unit because it is merely a ratio.

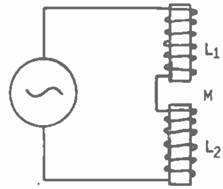
The maximum value that K can have is one, and in this case the two coils are said to be 100% coupled, or there is unity coupling between them. It should be remembered that with this condition, all of the flux encircles all the turns of both primary and secondary. When the coefficient of coupling is one, the mutual inductance is equal to the square root of the product of the primary and secondary inductances.

In one of the early lessons in Unit 1, a formula was given for finding the total inductance of two or more inductances connected in series. This formula stated that the total inductance was equal to the sum of the separate inductances so connected. This is true, however, only when no mutual inductance exists between any of the coils in the circuit. Let us consider the circuit shown in Fig. 10.

If there is no mutual inductance between the two coils in this figure, the total inductance in the circuit is equal to the sum of L_1 and L_2. Let us assume, however, that some mutual inductance does exist between the two coils.

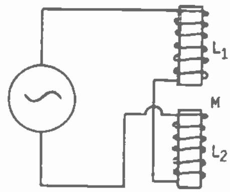
There will be two voltages developed across L_1. One will be due to the self-inductance of L_1, and one to the mutual inductance between L_2 and L_1. Likewise, there will be two voltages set up

Fig.10 Two coils connected in series-aiding.



across L_2 for the same reason. If the mutually-induced voltage in either coil is in phase with the self-induced voltage of that coil, then the two coils are connected in "series-aiding". The total inductance of the top coil is $L_1 + M$, and that of the lower coil is $L_2 + M$. The total inductance of the circuit is the sum of these two inductances which is $L_1 + M + L_2 + M$, or $L_1 + L_2 + 2M$.

Fig.11 Two coils connected in series-opposing.



If the two coils are arranged as shown in Fig. 11, the mutually-induced voltage of either coil is out of phase with the self-induced voltage of that coil. With this condition, the total inductance of the top coil is $L_1 - M$, and that of the bottom coil is $L_2 - M$. Therefore, the total inductance of the circuit is $L_1 + L_2 - 2M$, and in this case the two coils are connected "series-opposing". Thus, the general formula for the total inductance of two coils connected in series is:

$$L_{eff} = L_1 + L_2 \pm 2M \qquad (15)$$

The positive sign before the term 2M is used when the coils are connected series-aiding, and the negative sign when they are connected series-opposing.

Let us consider the case of two coils of equal self-inductance connected in series with unity coupling between them. Since L_1 equals L_2, then the mutual inductance would be:

$$M = 1 \times \sqrt{L^2} = L$$

Where L is the inductance of either coil.

If the two coils are connected series-aiding, the total inductance would be:

$L_{eff} = 2L + 2M$, and since M is equal to L,

$$L_{eff} = 4L$$

Suppose that two 10-henry coils are connected in series-aiding with unity coefficient of coupling. The mutual inductance between the two coils would be 10 henries, and the total inductance in the circuit would be 40 henries. Thus, when two equal inductances are connected in series with unity coupling, and with their magnetic fields adding, the total inductance in the circuit is four times the value of either.

The foregoing principle brings forth an interesting fact. Suppose that one of the 10-henry coils in the preceding example is composed of a thousand turns, and the second coil is formed by merely winding an additional thousand turns on the same form. The total inductance then becomes 40 henries, or the inductance increases four times when the number of turns is doubled. Thus, with all other factors remaining constant, the inductance varies as the square of the number of turns.

If the two coils are connected in series-opposing, the effective inductance is:

$$L_{eff} = 2L - 2M$$

And, since M is equal to L for unity coupling,

$$L_{eff} = 2L - 2L = 0$$

The total inductance of the circuit is zero and this principle is made use of in creating non-inductive coils by winding one-half of the coil in one direction and the other half in the opposite direction.

Now that we have learned some of the principles of mutual inductance, we are ready to consider impedance matching in R.F. circuits. As is already known, the voltage developed across an inductance due to the self-induced voltage is found by multiplying the current by the inductive reactance. In a like manner, the voltage developed in the secondary circuit by the primary current is found by multiplying the primary current by the mutual reactance. The mutual reactance is found in a manner analogous to that used for determining the inductive reactance. It is:

$$X_m = 6.28 \times F \times M \qquad (16)$$

Where: X_m is the mutual reactance in ohms,
F is the frequency in cycles,
M is the mutual inductance in henries.

It may be proved that the impedance reflected from secondary to primary circuit is:

$$Z_r = \frac{X_m^2}{Z_s} \qquad (17)$$

Where: X_m is the mutual reactance between the two circuits, in ohms,
Z_s is the total impedance of the secondary circuit, including that of the secondary coil and the load,
Z_r is the reflected impedance in ohms.

Formula (17) gives the total impedance reflected into the primary circuit, but it does not tell what part of that impedance is resistive and what part is reactive. Of course, the same rules apply to R.F. transformers as to iron-core transformers insofar as the nature of the reflected impedance is concerned. That is, the reflected impedance will be purely resistive if the secondary circuit contains only resistance, whereas it will be purely capacitive if the secondary is inductive, and will be purely inductive if the secondary is capacitive. If the secondary circuit contains both reactance and resistance, then the reflected impedance will be both reactive and resistive, the reactive part being of opposite kind to the secondary reactance. The actual values of the reflected resistance and reflected reactance may be found by the following formulas:

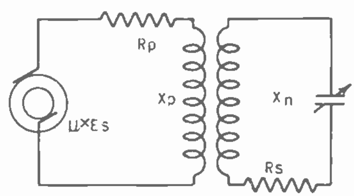
$$X_r = \frac{Z_r \times X_{ns}}{Z_s} \tag{18}$$

$$R_r = \frac{Z_r \times R_s}{Z_s} \tag{19}$$

Where: X_r is the reflected reactance in ohms,
X_{ns} is the net reactance of the secondary,
R_s is the resistance of the secondary,
Z_r is the total reflected impedance determined by formula (17),
Z_s is the total impedance of the secondary circuit.

Let us now investigate how the presence of a tuned secondary affects the current in an untuned primary. This corresponds to an ordinary R.F. transformer such as is used in receivers. The

Fig.12 The circuits used to explain the phenomenon of reflected impedance when the primary is untuned and the secondary is tuned.



circuit shown in Fig. 12 will be referred to during the following discussion. The primary voltage ($\mu \times E_s$) is the effective alternating voltage in the plate circuit. R_p is the plate resistance of the tube; X_p, the inductive reactance of the primary; X_n, the net reactance of the secondary; and R_s the total resistance of the secondary. When the incoming frequency is less than the resonant frequency of the tuned circuit, the secondary has a predominance of capacitive reactance,[1] and the current flowing in the secondary is low. This causes the reflected impedance to consist of an inductive reactance and a resistance. The reflected inductive reactance adds to the normal reactance of the primary and the current flowing in the primary is less than it would be if the secondary were not present.

As the frequency of the incoming signal is increased, the predominance of capacitive reactance in the secondary decreases,

[1] Refer to Sec. 5, Lesson 22, Unit 1.

And the secondary current becomes larger. Due to the smaller amount of net capacitive reactance in the secondary, the reflected impedance contains less inductive reactance, and due to the increased secondary current, it contains more resistance. Although the amount of reflected inductive reactance diminishes, the value of the reflected resistance increases, and, as a result, the total amount of reflected impedance becomes larger. This causes the primary current to fall as the frequency of the applied voltage is raised.

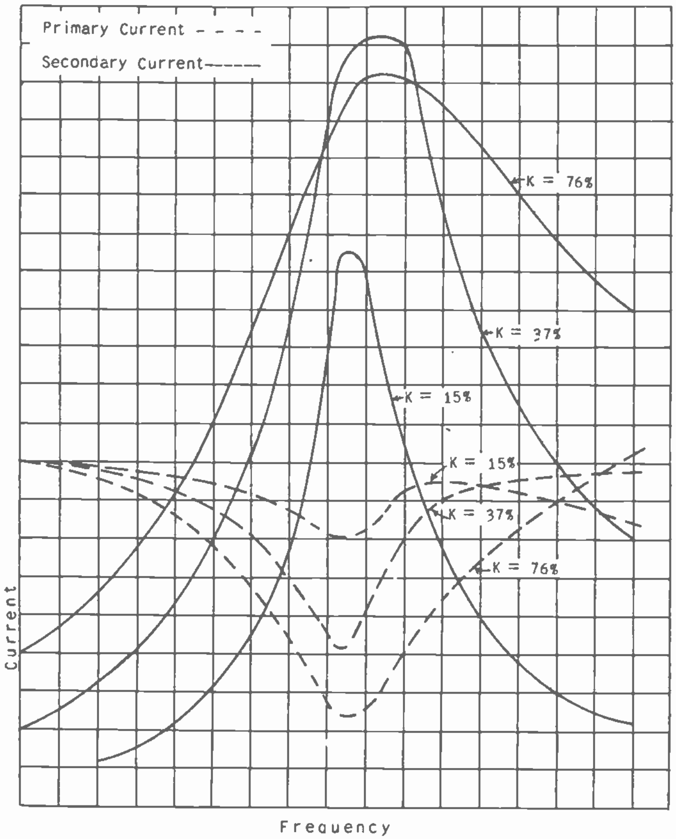


Fig.13 Showing how the currents in the primary and secondary circuits of Fig. 12 vary with frequency and with different amounts of coupling.

When the frequency of the incoming signal has been increased until it is equal to the resonant frequency of the tuned circuit, the net reactance of the secondary is zero, and a relatively large amount of secondary current flows. This produces a reflected impedance of such a phase that it has the nature of a pure resistance, and, due to the large secondary current, the value of the reflected resistance is maximum. At this point, the primary current reaches a minimum value.

Increasing the frequency of the incoming signal above the resonant frequency of the tuned circuit causes the secondary to have a net inductive reactance, and thereby reduces the secondary current. Under this condition, the reflected impedance consists of a capacitive reactance and a resistance. The smaller secondary current produces a lower value of reflected resistance, while the reflected capacitive reactance partially neutralizes the inductive reactance of the primary and, therefore, causes the primary current to increase.

At a frequency slightly above the resonant frequency of the tuned circuit, the reflected capacitive reactance is just sufficient to completely cancel the inductive reactance of the primary, and the phase angle of the primary is zero. At frequencies higher than this value, the reflected capacitive reactance is greater than the inductive reactance of the primary and the primary has a net capacitive reactance which causes it to draw a leading current. Thus, the primary circuit acts like a condenser (draws a leading current), although there is no condenser in the primary circuit.

The curves shown in Fig. 13 illustrate how the primary current and secondary current vary for different degrees of coupling, as the frequency of the applied voltage is changed. It should be noticed that the resonance curve of the secondary has a flatter top and is less peaked when it is closely coupled to the primary, and would, therefore, offer less selectivity. The reason that the secondary resonance curve is not sharp is due to the fact that the voltage induced into the secondary is not constant. The secondary voltage depends directly on the primary current, and the primary current reaches a minimum value at resonance. This causes the secondary voltage to be lowest at the resonant frequency, and prevents the secondary current from reaching as high a value as it otherwise would, even though the impedance of the secondary is minimum. For this reason, relatively loose coupling should be used between the primary and secondary of an R.F. transformer, so that the selectivity will not be seriously affected.

5. COUPLED TUNED CIRCUITS. A special case arises when both primary and secondary circuits are series tuned circuits. (It is assumed that both circuits are tuned to the same frequency.) With very loose coupling, the resonance curve of the primary circuit is practically the same as though the secondary were not present. When the coupling between secondary and primary is increased, a change occurs in the resonance curve of the primary current. With medium coupling, there is a fairly large value of resistance reflected from the secondary circuit into the primary circuit at the resonant frequency. Since the effective resistance of the primary circuit has been increased, the resonance curve will be less peaked and the resonant effect will be less pronounced. This was learned in Lesson 22 of Unit 1.

Let us assume that a constant voltage of variable frequency is applied to the circuit shown in Fig. 14, and from the data obtained, a resonance curve of the primary current is plotted. When the frequency of the applied voltage is less than the resonant frequency of the two circuits, the inductive reactance of

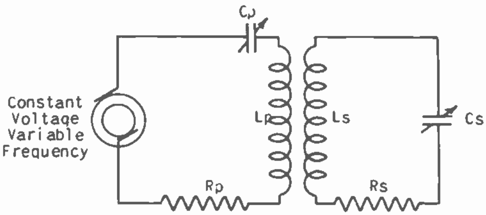


Fig.14 The circuit used to explain reflected impedance between two series-tuned circuits.

both primary and secondary will be less than their corresponding capacitive reactances, and both circuits will have a predominance of capacitive reactance. Therefore, the current flowing in the primary winding will lead the applied voltage, and the current flowing in the secondary winding will lead the voltage produced across the secondary. Since the secondary circuit is capacitive, the impedance it reflects back into the primary will be largely inductive. It will reflect a small amount of resistance, but not as much as when the applied voltage has a frequency equal to the resonant frequency of the two circuits. This inductive reactance that is reflected from the secondary back to primary partially neutralizes the predominance of capacitive reactance existing in the primary circuit, and thereby causes the primary current to be slightly larger than it would be if the secondary were not present. As the frequency of the applied voltage is increased, the predominance of capacitive reactance in the primary decreases, and, at some particular frequency, the amount of inductive reactance coupled from secondary to primary will be sufficient to exactly cancel the excess of capacitive reactance existing in the primary. At this point, then, the current flowing in the primary circuit is limited only by the resistance of the circuit. Thus a resonant point has been reached. Note that this resonant frequency is not the frequency to which the two circuits are tuned, but is a frequency somewhat less than the actual resonant frequency of the circuits.

As the frequency of the applied voltage is increased until it is exactly equal to the resonant frequency of the two circuits, the amount of resistance that is reflected from secondary to primary increases and, at this frequency, it is maximum. At this point, therefore, the current that flows in the primary circuit is rather low due to the large amount of resistance which has been coupled into the primary.

Now as the frequency of the applied voltage is increased above the actual resonant frequency of the two circuits, the inductive reactance of the primary becomes larger than its capacitive reactance and the same is true of the secondary. Under this condition, the predominance of inductive reactance in the primary circuit causes a lagging current and the secondary current also lags the voltage produced across the secondary. The impedance coupled from secondary into primary is capacitive since there is a predominance of inductive reactance in the secondary circuit. The amount of resistance coupled from secondary to primary is fairly low and the capacitive reactance coupled from secondary into primary partially neutralizes the predominance of inductive reactance in the primary. At a particular frequency somewhat above the actual resonant frequency of the two circuits, there is enough capacitive reactance coupled into the primary circuit to exactly neutralize the excess of inductive reactance present. In this case, the primary current is again limited only by its resistance. Thus another actual resonant point has been reached.

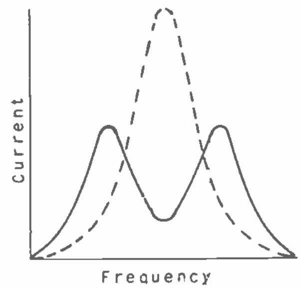


Fig.15 The resonance curve of the primary circuit of Fig. 14, when the secondary is open and closed. The dotted curve represents the primary resonance curve with the secondary open.

The resonance curve of the primary circuit will have the form shown in Fig. 15. Notice that the current flowing in the primary circuit starts from a very low value at frequencies far below resonance, increases to a maximum value at a frequency somewhat below the resonant frequency, reaches a minimum value at the exact resonant frequency, and again rises to another maximum value somewhat above the actual resonant frequency. The dotted curve in the same figure is the resonance curve of the primary when the secondary is open-circuited.

If the coupling between primary and secondary is sufficiently great, the resonance curve of the secondary current will also have two humps, or two resonant points, one above and one below the actual resonant frequency of the secondary circuit considered by itself. In Fig. 16 are shown two sets of curves which illustrate the values of the primary and secondary currents for different amounts of coupling. It should be noticed that when the coupling is very loose, (K is .002), the resonance curve of the primary current is very sharp, practically the same as it would be if the secondary were not present. In this case, the secondary current does not reach a very high value; however, the secondary resonance curve with this coupling is very peaked, indicating a large amount of selectivity. When the coupling is increased until K = .01, the primary resonance curve has begun to show the characteristic twin peaks, or double humps, while the secondary current with this amount of coupling has reached its maximum value. Notice that the secondary resonance curve, when K equals .01, has a flat top. Such a circuit would not be very selective but would be desirable in R.F amplifier stages to prevent sideband cutting.

Also observe that with this amount of coupling, the secondary current is maximum. With either more or less coupling than this value, the secondary current is lower. The amount of coupling that produces a maximum secondary current is called the "critical coupling". In coupling a dummy antenna to the tank circuit of an R.F. power amplifier, it will be found that there is one particular point that gives maximum antenna current. When the antenna coupling coil is either closer or farther from the plate tank circuit, the antenna current drops off.

With coupling greater than .01, the secondary resonance curve also begins to show the double-hump effect. Notice that the secondary resonance curve for K equals .015 would, if used in the I.F. stages of a receiver, produce very good fidelity, as there would not likely be any possibility of sideband cutting. Some modern receivers that employ selectivity controls are so arranged that the coupling between the secondary and primary of one or more of the I.F. transformers may be varied. With loose coupling, the selectivity is maximum because the secondary resonance curve is very peaked; while, by increasing the coupling, a flat-top curve is obtained which lowers the selectivity but greatly increases the fidelity.

The primary and secondary current peaks occur at practically the same frequencies and are very nearly symmetrically located on either side of the resonant frequency of the two circuits. The height of the current peaks become smaller as the coupling is increased, but the two humps always have approximately the same maximum. When the coefficient of coupling is approximately 1½ times the critical coupling, the double humps appear in the secondary current curve. The secondary current curve follows through practically the same stages as does the primary current characteristic, first becoming lower and flatter and then separating into a two-humped curve with a minimum between the humps at the frequency to which the circuits are tuned. The main difference between the primary and secondary resonance curves is that the drop in current between the two humps is less pronounced in the curve of the secondary current. The voltage induced in the secondary has the same general characteristics as the primary current curve and will therefore have two humps when the primary current does. However, the low impedance of the secondary at its resonant frequency tends to counteract the low voltage that is induced at this frequency.

The mutual inductance required for critical coupling is determined by the following formula:

$$M = \frac{\sqrt{R_p \times R_s}}{6.28 \times F} \qquad (20)$$

Where: M is the mutual inductance in henries,
R_p is the resistance of the voltage source,
R_s is the resistance of the secondary circuit,
F is the frequency in cycles.

The coefficient of coupling needed to produce this amount of mutual inductance is:

$$K = \sqrt{\frac{R_p \times R_s}{X_{Lp} \times X_{Ls}}} \qquad (21)$$

Where: R_p is the resistance of the voltage source,
R_s is the resistance of the secondary circuit,
X_{Lp} is the inductive reactance of the primary coil,
X_{Ls} is the inductive reactance of the secondary coil.

The circuit which we have been considering has a series tuned primary and secondary circuit. Such a combination is not ordinarily found in radio receivers or transmitters. In most cases where coupled tuned circuits are employed, the primary consists of a parallel tuned circuit, while the secondary is series tuned. This is the case in tuned I.F. transformers and in the coupling arrangement between the dummy antenna and the plate tank circuit of an R.F. power amplifier. When the primary is a parallel tuned circuit, the primary current is not of as much importance as is the shunt impedance of the parallel tuned circuit. When the shunt impedance is too high, the vacuum tube is unable to furnish appreciable power to the tank, and the same is true when the shunt impedance is less than the plate resistance of the tube.

With the inductance and capacity of the primary circuit constant, the shunt impedance will vary inversely with the amount of resistance reflected into the tank. In Lesson 22 of Unit 1, it was learned that the graph showing the relationship between the shunt impedance and the applied frequency of a parallel tuned circuit had practically the same shape as the resonance curve of an ordinary series tuned circuit; that is, at frequencies lower than resonance, the shunt impedance is low; at the resonant frequency it is maximum; while at frequencies higher than resonance, it is again low. Therefore, the curves shown in Fig. 16 will apply equally well, when the primary is a parallel tuned circuit, if it is remembered that the upper set of curves refer, not to the primary current, but to the shunt impedance of the primary. Since the secondary is always a series tuned circuit, the lower set of curves refer to the actual secondary current.

The coupling between primary and secondary should never be increased beyond the point of critical coupling. If a dummy antenna is coupled too tightly to a tank circuit, maximum antenna current is not obtained, and the double-hump resonance curve which might be produced in the antenna circuit would probably cause instability. It is also interesting to note that at critical

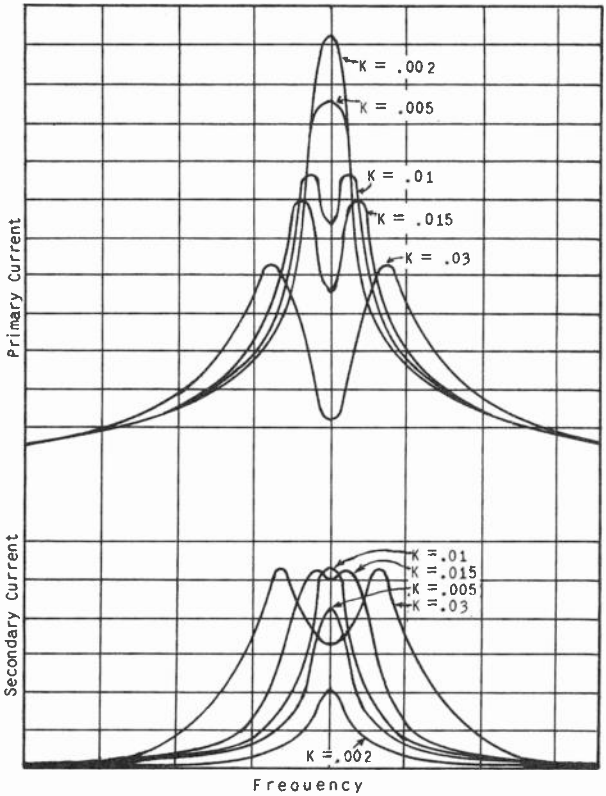


Fig.16 Illustrating the effect of changes in coupling.

coupling, the resistance reflected from the secondary into the primary is sufficient to make the shunt impedance of the primary equal to the plate resistance of the tube. This, of course, is the condition for maximum power transfer, and therefore the secondary current will be maximum and the power transfer to the secondary circuit will be greatest with this amount of coupling.

The important points to remember about reflected impedance in R.F. circuits are as follows:

1. The impedance coupled from secondary into primary depends upon the mutual reactance between the two circuits and the total impedance of the secondary.
2. When the secondary circuit has a predominance of inductive reactance, the reflected impedance consists of a small amount of resistance and a relatively large amount of capacitive reactance.
3. When the secondary circuit has a predominance of capacitive reactance, the reflected impedance consists of a small amount of resistance plus a

relatively large amount of inductive reactance.
4. When the secondary circuit contains only resistance, the reflected impedance is itself a pure resistance.
5. The resonance curve of two coupled tuned circuits will be practically the same as that of either circuit considered alone, if the coupling is very loose; but will exhibit double-peaked resonance curves when the coupling is tight. At the point of critical coupling, maximum power will be transferred to the secondary circuit and the secondary resonance curve will be flat-topped. Also, with this amount of coupling, the resistance reflected from secondary into primary is equal to the primary impedance.

6. EXAMPLES OF IMPEDANCE MATCHING IN R.F. CIRCUITS. Mention has already been made of the use of coupled tuned circuits in I.F. amplifiers to secure band-pass effects. In ordinary T.R.F. amplifiers, where the secondary is tuned to resonance with the desired signal, the oscillating current that flows in the tuned circuit reflects practically a pure resistance into the primary winding of the R.F. transformer, and it is, in effect into this reflected resistance that the vacuum tube works. Although it may, at first sight, appear that the tube is working into a reactance, actually it is working into a resistance. In most receiver R.F. amplifiers which are for the purpose of amplifying voltage only, it is not necessary that the grid circuit of one tube be matched to the plate circuit of the preceding tube. The reflected impedance, however, should be large enough so that a large R.F. voltage is produced across the primary of the R.F. transformer. This *demands* that the reflected resistance be much larger than the plate resistance of the tube.

In most R.F. power amplifiers used in transmitters, some grid current is allowed to flow, and this action absorbs power from the plate circuit of the preceding stage. To transfer the power efficiently, the grid circuit should be matched to the plate circuit of the tube exciting it.

Since it is usually difficult to calculate the inductance of the tank coils, and the coupling between the tank coil and the grid circuit of the succeeding stage, most transmitters are adjusted by experimental methods, rather than attempting to calculate what values the various circuit components should have to give the optimum operating characteristics. Although it may be necessary to make many changes in the transmitter adjustments to secure the desired power output with a reasonable degree of efficiency, these adjustments should not be made in a haphazard manner. Rather, a definite procedure should be followed so that no unnecessary changes will be made.

In regard to the vacuum tubes used in the transmitter, it is usually possible to obtain the correct operating conditions (plate voltage and grid bias) from the manufacturer's specifications. These values given by the manufacturer should be adhered to fairly closely. It may be permissible to use a smaller plate voltage than that recommended by the manufacturer, if a power supply for furnishing the rated plate voltage is not available; however, it should not be expected that the tube will produce as satisfactory results as if the rated plate voltage were employed. Never, under any considerations, use a plate voltage greater than the maximum specified by the manufacturer, nor allow a larger plate current to flow than that given in the tube's ratings.

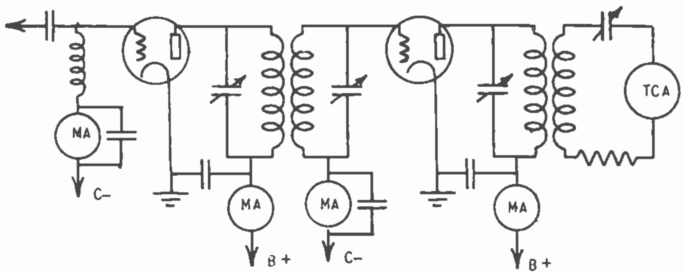


Fig.17 Circuit used to describe the process of reflected impedance in transmitter stages that are inductively coupled.

In Fig. 17 are illustrated two R.F. transmitting amplifiers and a dummy antenna. To simplify the diagram, the neutralizing circuits have been omitted. Assuming that the first stage is correctly neutralized and the proper voltages are applied, when the plate tank circuit is tuned to resonance, it offers a maximum impedance to the R.F. plate current pulses. As the coupling between the grid circuit of the second stage and the tank circuit of the first stage is increased, the impedance reflected from the grid circuit into the plate tank increases. This impedance is a pure resistance, provided that the grid tank is tuned to resonance. By reflecting a pure resistance into the plate tank circuit, the shunt impedance of the tank is reduced. With the stage unloaded, the shunt impedance of the tank should be very large compared to the plate resistance of the tube. Then, as the coupling is increased, the shunt impedance of the tank is reduced, and therefore the tube is able to furnish more power to the tank.

When more than critical coupling is employed, the resistance reflected into the plate tank circuit is so large that the shunt impedance of this tank is reduced to a value below the plate resistance of the tube. Under this condition, the power furnished to the plate tank drops off very rapidly, and most of the power is dissipated within the tube, causing its plate to become red hot. When this occurs, the efficiency is rather low, and since the oscillating current will be small due to the large value of resistance reflected into the tank circuit, the tank will not have sufficient fly-wheel effect to smooth out the plate current pulses and many harmonic frequencies will be transferred to the following stage.

When the loading of the first tube is correct, the second stage should be neutralized, and then its plate voltage may be applied. The plate tank of the second stage is now tuned to resonance as indicated by minimum plate current and is then loaded by coupling the antenna to the plate tank.[1] The antenna is tuned to resonance and the coupling between it and the plate tank is increased until maximum antenna current is attained. If it should happen that the antenna circuit is slightly out of resonance, the current flowing in it will not be exactly in phase with the voltage producing it, and a predominance of either capacitive or inductive reactance will prevail. When this is the case, the reflected impedance will contain either inductive or capacitive reactance. In either case, the reactance that is reflected into the plate tank circuit will detune this circuit. For this reason, it is necessary to again check the tuning of both the plate tank and the antenna. Do not increase the coupling beyond the point that gives maximum antenna current, for this causes the double-humped resonance curve which is undesirable.

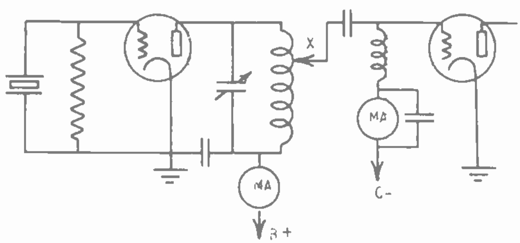


Fig.18 Showing how reflected impedance occurs between the grid circuit of the buffer and the plate circuit of the crystal oscillator.

Fig. 18 shows the essential parts of a buffer amplifier capacitively coupled to a crystal oscillator. With the excitation tap (X) removed, the oscillator plate tank offers a large impedance to the plate current, and the oscillator is unable to furnish appreciable power to this tank circuit. By lighting the filament of the buffer stage, and connecting the excitation tap to the oscillator tank circuit, the impedance of the oscillator plate tank is lowered. Since some grid current flows in the buffer stage, the grid-to-cathode impedance of the buffer is not infinite, but has a very definite value. Therefore, connecting the excitation tap to the oscillator plate tank has the same effect as connecting a resistor across a part of the oscillator tank circuit. A resistor so connected would draw power from the tank and would, therefore, reduce the value of the oscillating current. Thus, the effect is the same as though a resistance had been reflected into the oscillator tank circuit as this would also reduce the oscillating current. The result is that the shunt

[1] A more complete discussion of loading an R.F. power amplifier stage will be given in Lesson 7 of this unit.

impedance of the oscillator tank is lowered, and the oscillator tube is able to furnish more power to the tank circuit, for the load into which the tube is working is more nearly equal to the plate resistance of the oscillator.

If the grid circuit of the buffer amplifier has a fairly low impedance, the excitation tap should be connected down on the oscillator plate tank. In this case, there will be less impedance between the excitation tap and ground, and thus the impedance of the source furnishing power to the grid circuit will more nearly match the impedance of the grid circuit.

In general, a low-mu tube will have a larger grid impedance than will a high-mu tube. Thus, in using a low-mu tube, the excitation tap should be connected near the plate end of the oscillator tank in order to more nearly match the grid circuit to the plate circuit of the preceding tube. This also means, of course, that the low-mu tube requires more grid excitation voltage and, as a result, it is said that it is harder to drive or requires more driving power.

Notes

(These extra pages are provided for your use in taking special notes)

Notes

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UNIT NO. 3

FUNDAMENTAL MODULATION

LESSON NO. 5

THE TRANSMISSION OF THOUGHT

.......FOLLOWED BY THE TRANSMISSION OF SIGHT

For years upon years thousands of people have held a firm belief in the transmission of thought. They are convinced that the mind emitted thought waves that travelled through the air much like radio waves, to be received by a mind in a distant city that was IN TUNE with the sender.

Today, because of the marvelous development of radio IT IS possible to transmit thought..... thousands upon thousands of miles. A man may stand before a microphone in New York City. His mind creates certain thoughts that he wishes to convey to thousands of listeners in far away Europe. His mind actuates the vocal cords and mouth and he gives voice to his thoughts. The varying sounds he creates strike the microphone, and create weak electrical impulses of a varying intensity. These impulses pass into powerful amplifiers.....a maze of grotesquely shaped tubes, condensers, filters and other weird looking contrivances, and emerge greatly amplified to pass to the vertical antenna where they are radiated in the form of so-called "radio waves."

These waves, representing the thoughts of the man in the studio, arrive in Europe almost as quickly as they are transmitted. They strike the antennas of radio receivers TUNED to their frequency, and wonders of wonders, THE THOUGHTS OF THE MAN BEFORE THE MICROPHONE IN NEW YORK CITY HAVE BEEN TRANSMITTED TO OTHER MINDS IN EUROPE IN ALMOST "NOTHING FLAT".

Because of the fact that radio makes possible the transmission of thought, it has become a powerful factor in the world's politics, education and entertainment. And the power of this new force is increasing year after year. Soon, Television will make it possible for speakers to add NEW FORCE to their transmitted thoughts, for they will be able to add gestures and facial expressions to the power of their voice.

Radio, with Television closely following, will control the destinies of millions of people and great nations. You are, indeed, fortunate that you have cast your lot with a new scientific wonder which offers so much NOW and in the future.

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Lesson Five

FUNDAMENTAL MODULATION

"Modulation is defined as the process of impressing voice or music waves on a carrier wave so that the transmission of intelligence without special codes is made possible.

"Without the process of modulation, two-thirds of our present radio progress would have been impossible. Radio would still be confined to the transmission of intelligence by the sending of dots and dashes which represented a code system. All broadcast transmitters are modulated, and the advancing art of television has presented special problems in the design of modulation systems. Thus, it is highly essential that you secure a thorough fundamental knowledge of this process."

1. **TYPES OF MODULATION.** In radio telephony, the transmission of intelligence is accomplished by varying the amplitude of the radio frequency wave, which is generated by the transmitter, at the audio frequency rates corresponding to the audio currents produced by the microphone when the sound waves strike it. This process of varying the amplitude of the radio frequency wave is called "amplitude modulation". Thus a definition for modulation would be as follows: Modulation is the process by which electrical waves produced in the microphone circuit by sound waves are caused to vary the amplitude of the transmitted radio frequency wave in direct accordance with the variations of the sound waves. Actually, this is the definition for amplitude modulation; in addition to which, there is also frequency modulation, in which the frequency of the R.F. wave is varied in accordance with the changes of the sound waves. Frequency modulation has, at present, no commercial applications except as used to align receivers in conjunction with a cathode ray oscilloscope, as was explained in Lesson 4, Unit 2.

Fig. 1 at A shows a graphic representation of a radio frequency carrier wave such as produced by an R.F. oscillator, while at B is illustrated a modulated wave that would be created by superimposing a single audio frequency on this carrier wave.

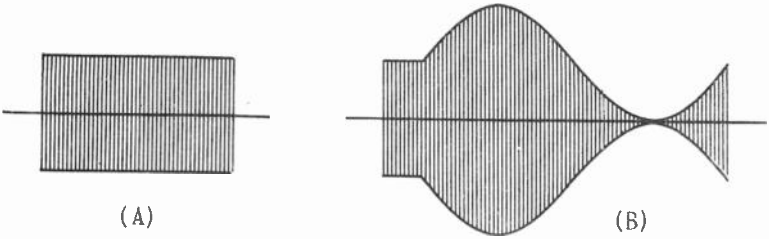


Fig.1 (A) An unmodulated radio-frequency wave.
(B) A modulated RF wave.

There are several ways in which the amplitude of the R.F. carrier wave may be changed. They are:

1. By variation of the plate voltage applied to one of the power amplifiers (called plate modulation).
2. By variation of the grid bias voltage of the power amplifier (known as grid modulation).
3. By variation of the screen grid voltage in a screen grid amplifier (called screen grid modulation).
4. By changing the suppressor grid voltage on a pentode power amplifier. (suppressor modulation).

The conception of a vacuum tube oscillator as a converter of DC power into R.F. power was presented in Lesson 1 of this Unit. (The oscillator should not be looked upon as a generator because it does not function in this manner.) A block diagram representing the three principal parts of such a converting system is shown in Fig. 2; they are the power supply, the converter, and the antenna.

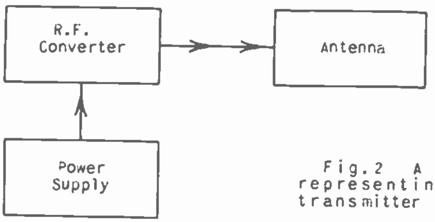


Fig.2 A block diagram representing a simple transmitter

The power in the antenna system is dependent on two things; the power input to the converter, and the efficiency of conversion; and, therefore, may be controlled by either one of two methods. The first is by keeping the efficiency constant and varying the power input to the converter, and the second is by keeping the power input to the converter constant, and varying the efficiency between the converter and the antenna. The first method represents the process known as power or plate modulation, while the second is typical of efficiency modulation, in which the grid bias or screen grid voltage is changed. This lesson will be devoted entirely to power or plate modulation, the subject of efficiency modulation being reserved for Lesson 12 of this Unit.

2. WHY POWER IS NEEDED TO MODULATE A CARRIER WAVE. Since plate modulation is concerned with power, it is felt advisable in this preliminary discussion to review a few fundamental principles of power. Therefore, let us turn our attention to the simple circuit of Fig. 3 at A, which consists of a 10-ohm resistor connected across a 20-

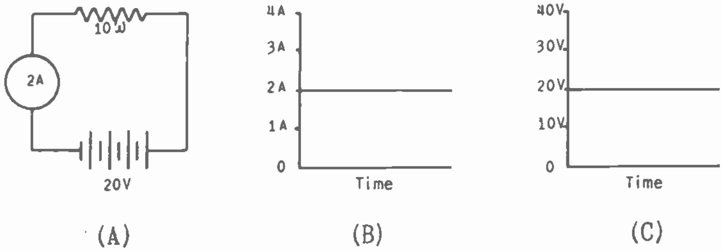


Fig.3 (A) A simple series circuit.
(B) Graph representing the current of A.
(C) Graph illustrating the voltage of A.

volt battery. The current flowing in this circuit is 2 amperes and could be plotted as shown at B in the figure. Likewise, the voltage across the resistance might be represented by the graph at C. The power being consumed in the resistance due to its frictional opposition to the flow of current could be calculated very simply as follows:

$$W = E \times I$$

$$= 20 \times 2$$

$$= 40 \text{ watts}$$

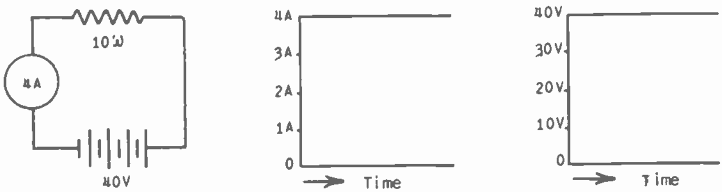


Fig.4 Illustrating the effect of doubling the voltage of the battery supplying power. This circuit is the same as Fig.3, except that the voltage has been increased.

Suppose, however, that the battery voltage is increased to 40 volts, as shown at A in Fig. 4. The current and voltage would now be represented by the graphs of B and C in this figure. The power dissipated in the resistor with this new voltage is:

$$W = 40 \times 4$$

$$= 160 \text{ watts.}$$

Thus, a very fundamental law of electricity has been illustrated. The law is stated by this formula:

$$W = \frac{E^2}{R}$$

In words, this formula states that, with the resistance of a circuit unchanged, a change in the voltage applied to the circuit will cause a change in power proportional to the square of the voltage change. This is true because the current increases in the same ratio as the voltage increase. If the voltage applied to a circuit is doubled, the current also is doubled and the power increases to four times its original value.

To continue this discussion, let us set up a series circuit containing a 20-volt battery, an alternator having a peak voltage of 20 volts, and a 10-ohm resistor. Such a circuit is illustrated at A in Fig. 5. With the alternator idle and assuming that it has

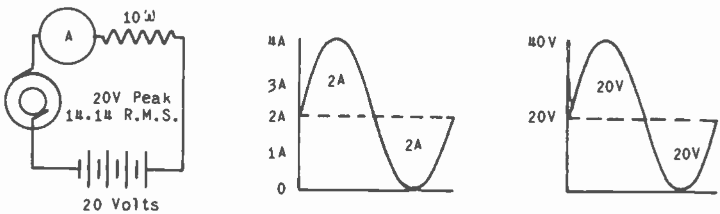


Fig.5 An AC generator supplying power to a simple series circuit. The graphs of the voltage and the current are also shown.

no resistance, the voltage and current in the circuit would be reresented by the dotted lines at B and C in the figure, and the power in the circuit would again be 40 watts. With the alternator running, however, the total effective voltage in the circuit is constantly changing. When the voltage of the alternator is 20 volts and in such a direction as to buck against the voltage of the battery, there is no net voltage present across the resistor, no current flows in the circuit, and the power consumption is zero. When the voltage of the alternator reverses polarity and again reaches its peak value, it adds to the voltage of the battery and the effective voltage across the resistor is 40 volts. This produces a current flow of 4 amperes and a power consumption of 160 watts. Thus, the power in the resistor varies at a periodic rate from 0 to 160 watts.

The extremes of the power variation are not, however, of as much interest as the average power. It is possible to plot the curve representing the power variation in this circuit and by averaging all the instantaneous values, determine the average power consumed in the resistor. This, however, entails considerable unnecessary calculation. It is much simpler to calculate the average power contributed by the alternator and to that add the power delivered by the battery. Thus, considering the AC and DC components separately, it is evident that the DC power furnished by the battery when the alternator is not running is 40 watts as previously calculated. To determine the average AC power, let us remove the battery from the circuit, and calculate the AC power furnished to the resistor by the alternator. The average AC power under this condition will be equal to the R.M.S. value of the current, times the R.M.S. value of the voltage, times the power factor. Since the circuit contains only pure resistance, the power factor will be one. The peak voltage of the alternator is 20 volts, and this would force a peak current of 2 amperes through the resistor. The R.M.S. values are:

$$E_{rms} = 20 \times .707 = 14.14 \text{ volts}$$

$$I_{rms} = 2 \times .707 = 1.414 \text{ amperes.}$$

Thus, the average AC power in the resistor is:

$$W_{ac} = 14.14 \times 1.414 = 20 \text{ watts.}$$

Since the average DC power has already been calculated as 40 watts, it is clear that the total average power in the resistor, when both the battery and the alternator are functioning, is their sum or 60 watts. Thus, the conclusion is as follows: When to a DC voltage there is added an AC voltage having a peak value equal to the DC voltage, the total average power in the circuit increases to one and one-half times its former value.

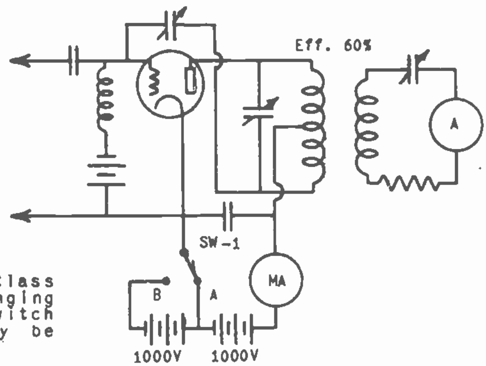


Fig.6 Diagram of a Class C amplifier. By changing the position of the switch the plate voltage may be doubled.

In Fig. 6 is shown a Class C, R.F. amplifier. It is the class C amplifier which is modulated in a transmitter. While we have not yet studied the characteristics of such an amplifier, it will be completely covered in Lesson 7 of this Unit. At the present, it is sufficient to state that such an amplifier is biased beyond cut-off, grid current is allowed to flow, and the plate current flows for less than half of a cycle. Assuming that this amplifier is properly biased, excited and loaded, its efficiency will remain constant, even though its plate voltage is varied. Notice that the plate voltage of this amplifier may be changed from 1000 to 2000 volts by changing the position of SW-1. With 1000 volts applied, assume that the amplifier tube draws 200 ma. plate current; this makes the DC power input equal to 200 watts. An efficiency of 60% will then cause the power in the dummy antenna to be 120 watts. If the switch is thrown from A to B, however, the plate voltage is increased to 2000 volts, and assuming a linear relation between plate voltage and plate current, the plate current will increase to 400 ma. Under this condition the DC power input to the tube increases four times, or to 800 watts. Since the efficiency remains constant, the power in the dummy antenna is found by multiplying the power input by .6 which gives 480 watts. Thus the power in the antenna also increases four times (from 120 to 480 watts.)

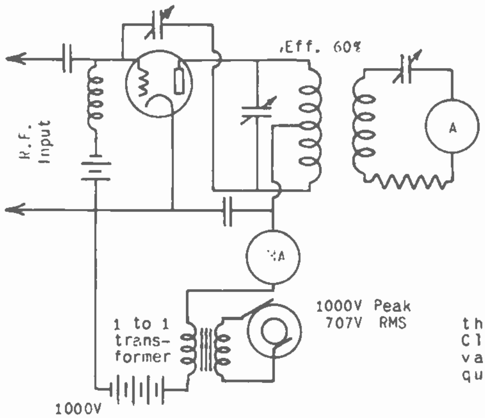


Fig.7 Illustrating how the plate voltage of a Class C amplifier may be varied at an audio frequency rate.

Let us now consider the circuit shown in Fig. 7. In this circuit the conditions are analogous to those existing in Fig. 5. An alternator having a peak voltage of 1000 volts is connected across the primary of a 1:1 ratio transformer. The secondary of this transformer is in series with the 1000-volt battery. Thus with the alternator running there will be developed across the secondary of the transformer a 1000-volt AC voltage which adds to and subtracts from the voltage of the battery. The actual voltage applied to the plate will therefore vary from 0 to 2000 volts, and the plate current will change from 0 to 400 ma. It must be realized that the current to which we are referring is the DC component of the plate current. Actually, the plate current flowing through the tube will vary at an RF rate as well as at the rate of the alternator frequency. In this type of amplifier the plate current flows for only ½ of RF cycle or less, but the DC component of the plate current pulses will vary from 0 to 400 ma. at the frequency of the alternator.

As previously stated, when to a DC voltage there is added an AC voltage having a peak value equal to the DC voltage, the average power in the circuit increases one and one-half times. Therefore, the average power input to the tube increases from 200 to 300 watts, and with the efficiency remaining the same, the average power in the antenna goes up to

180 watts. This 180 watts is the average power in the antenna; in reality, the antenna power is varying from 0 to 480 watts.

Since the DC voltage applied to the plate, the DC plate current and the DC power input to the tube are all caused to vary at the frequency of the alternator, it should be evident that the oscillating tank current, the RF power furnished to the tank, the oscillating antenna current, and the RF power in the antenna are also varying at the rate of the alternator frequency. Since the peak power furnished to the antenna when the alternator is operating is four times the antenna power when the alternator is disconnected, it is clear that the amplitude of the RF current in the antenna, and the amplitude of the RF voltage across the antenna must each have doubled its former value. Thus, if the peak value of the RF antenna current was formerly 1 ampere, the peaks are now varying from 0 to 2 amperes, at the frequency of the alternator. The presence of the alternator has caused the amplitude of the RF carrier current to vary at the frequency of the alternator. This, however, is the definition of modulation. Therefore, assuming that the alternator has a frequency of 60 cycles, it is evident that it has modulated the RF carrier with a 60 cycle frequency.

It should by this time be apparent that the alternator must furnish not only a peak voltage equal to the voltage of the B battery, but it must also furnish a power equal to one-half of the power drawn from the B supply, if the RF carrier is to be modulated in this manner. Through the milliammeter connected in the plate circuit there flows the DC component of the plate current which, due to the alternator, is varying at a 60 cycle rate from 0 to 400 ma. The meter is unable to follow the 60-cycle variations of the current, and therefore reads the average value of 200 ma. or the same value as when the alternator is not running. The actual plate current flowing through the tube has in reality three components; the RF pulses, a 60 cycle component furnished by the alternator, and the DC component which is drawn from the battery. The presence of the alternator produces no effect on the power that is drawn from the B battery. In actual operation the battery, or power supply, will be by-passed by a high-capacity condenser which prevents the variations in current from passing through the battery, and the current drawn from the battery will remain constant at 200 ma. Thus the extra power input to the tube which causes both the plate voltage and plate current to double their values must be supplied by the alternator. In the transmission of speech and music, the output of an audio amplifier instead of the alternator would be connected across the primary of the transformer. Since audio power as well as audio voltages must be developed across the secondary of this transformer, an audio power amplifier must be used.

3. PERCENTAGE OF MODULATION. We have seen that adding an AC voltage with a peak value equal to the DC voltage of the B battery causes the average power in the antenna to increase one and one-half times. Also with this amount of AC voltage, the amplitude of the RF carrier current varies from zero to twice its unmodulated value. Let us now determine what the effect of adding an alter-

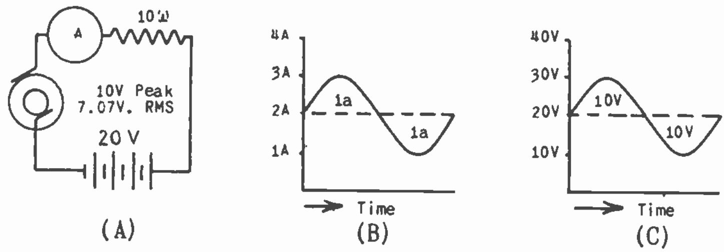


Fig. 8 Showing the effect of a DC and an AC voltage in the same circuit.

nating voltage whose peak value is only one-half of the DC voltage of the battery would be.

The circuit shown at A in Fig. 8 will be used. The peak value of the AC voltage is 10 volts, and the voltage of the battery is 20 volts. The voltage across and the current through the resistor when the alternator is not running are represented by the dotted lines at B and C of Fig. 8. With the alternator operating, the effective voltage in the circuit will vary between 10 and 30 volts, and this voltage will produce a current through the 10-ohm resistor which changes from 1 to 3 amperes. To find the amount of power in the resistor, the AC and DC components will be considered separately. The DC power is 40 watts, or the same as in Fig. 5. The RMS voltage and RMS current of the AC component are:

$$E_{rms} = 10 \times .707 = 7.07 \text{ volts.}$$

$$I_{rms} = 1 \times .707 = .707 \text{ ampere.}$$

Thus the AC power developed by the alternator is:

$$W_{ac} = 7.07 \times .707 = 5 \text{ watts}$$

Therefore the average power dissipated in the resistor is 45 watts.

It is seen that when the peak voltage of the alternator is one-half of the DC voltage of the battery, the power developed by the alternator is ⅛ or 12.5% of the power developed by the battery (5 watts AC power is ⅛ of 40 watts DC power).

Returning now to the tube arrangement shown in Fig. 9, let us use an alternator whose peak voltage is one half of the voltage of the B battery. The battery produces a voltage of 1000 volts, and the peak voltage of the alternator is 500 volts. If the DC plate current with the alternator disconnected is 200 ma., the DC power input is 200 watts. With the alternator connected, the voltage applied to the plate will vary from 500 to 1500 volts, and the plate current will change from 100 to 300 ma. The power supplied by the alternator will be 12.5% of the DC power furnished by the battery, or 25 watts. With a 60% efficiency the power in the antenna when the carrier is unmodulated is 120 watts. Under modulation the total average power input increases to 225 watts, and the average power in the antenna rises to 135 watts. Thus the power in the antenna also increases 12.5% with this amount of modulation. Since both the plate voltage and the plate current vary from ½ to 1½ their unmodulated values, the amplitude of the RF antenna current will also change from ½ to 1½ its unmodulated value. If the peak antenna

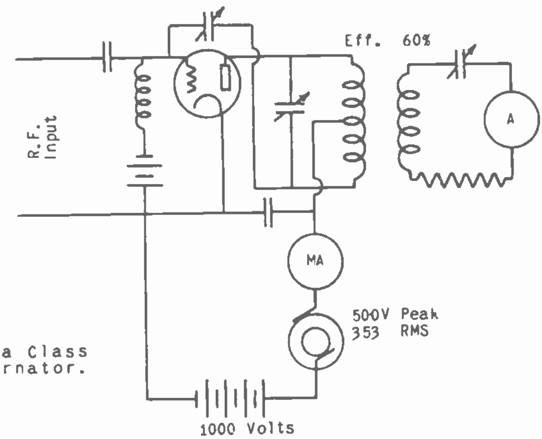


Fig.9 Modulating a Class C stage with an alternator.

current before modulation was 1 ampere, it will, with this amount of modulation, vary from ½ to 1½ amperes.

At A in Fig. 10 is illustrated the unmodulated carrier which is assumed to have a peak value of 1 ampere. At B in the same figure is shown the modulated carrier produced when the peak voltage of the alternator is equal to the voltage of the battery and the AC power applied to the circuit is 50% of the DC power input. At C in this figure is shown the modulated RF carrier produced when the peak AC input voltage is equal to one-half of the applied DC voltage, and the AC power input is 12.5% of the DC power input.

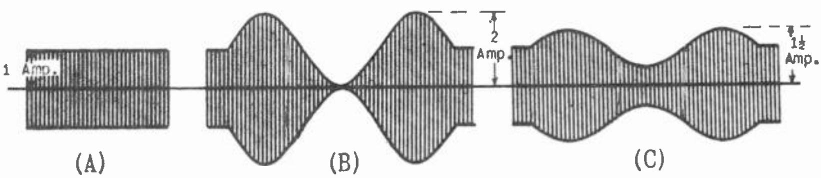


Fig.10 (A) An unmodulated RF wave. (B) A 100% modulated RF wave. (C) A 50% modulated RF wave.

The modulated wave shown at B in Fig. 10 is said to be 100% modulated. Whenever the amplitude of the carrier wave varies from zero to twice its unmodulated value, it is modulated 100%. The amplitude of the carrier at C in this figure varies from ½ to 1½ its unmodulated value, and it is said to be modulated 50%.

It is time that we learned how to determine the percentage of modulation when the variation of the carrier's amplitude is known. Most of the following discussion will refer to Fig. 11. In this figure there is shown a carrier wave, a part of which is modulated. The line joining the peak values of the modulated part of the carrier is known as the modulation envelope. By inspection of this envelope it is seen that it is really the wave form of the audio frequency that is producing the modulation. The following symbols will be used:

- I_c is the peak value of the unmodulated carrier.
- I_M is the greatest value obtained by the carrier during modulation. It is the value measured during the peak of the AF modulation cycle.
- I_{min} is the minimum peak value of carrier current. It occurs during the trough of the modulation cycle.
- I_m is the peak value of the modulation envelope. It is measured from the peak of the unmodulated wave to the greatest peak obtained during the modulation cycle.

The percentage of modulation may be defined as the ratio of the AF contained in the modulated wave to the RF present in the carrier. Since the amount of AF in the modulated wave is proportional to the amplitude of the modulation envelope, it may also be said that the percentage of modulation is equal to the ratio of the amplitude of the envelope to the amplitude of the unmodulated carrier expressed as a percentage. As a formula this is:

$$\text{Per cent of modulation} = \frac{I_m}{I_c} \times 100 \qquad (1)$$

From Fig. 11 it is seen that the total variation of the peaks of the modulated wave is equal to the maximum peak obtained minus the minimum peak reached. This total variation , however, is equal to twice the amplitude of the modulation envelope; therefore, the

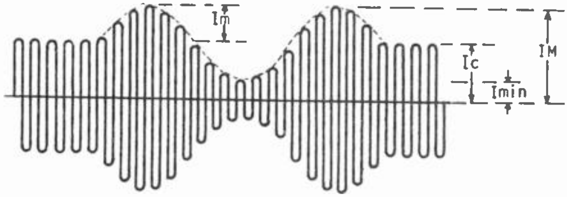


Fig.11 Illustrating how the various current values of a modulated wave are measured.

modulation percentage may be expressed as follows:

$$\text{Per cent of modulation} = \tfrac{1}{2} \times \frac{(I_M - I_{min}) \times 100}{I_c} \qquad (2)$$

Furthermore, it is evident that the amplitude of the modulation envelope is equal to the maximum amplitude obtained minus the amplitude of the unmodulated carrier; therefore, a third method of expressing the modulation is:

$$\text{Per cent of modulation} = \frac{(I_M - I_C) \times 100}{I_C} \qquad (3)$$

The modulated waves shown at B and C in Fig. 10 are modulated 100% and 50% respectively. Let us use these new formulas to prove that this is true. In the wave shown at B, the amplitude of the unmodulated carrier was assumed to be 1 ampere; the maximum peak obtained was 2 amperes; and the minimum peak was zero. By substitution in formula (2), we have:

$$\text{Per cent of modulation} = \frac{\frac{1}{2} \times (2 - 0) \times 100}{1}$$

$$= \frac{\frac{1}{2} \times 2 \times 100}{1} = \frac{1 \times 100}{1} = 100\%.$$

If formula (3) is used the same results will be secured:

$$\text{Per cent of modulation} = \frac{(2 - 1) \times 100}{1} = 100\%$$

In the wave shown at C in Fig. 10, the amplitude of the unmodulated carrier is 1 ampere; the maximum peak is 1½ amperes; and the minimum peak is ½ ampere. Substituting these values in formula (2), we secure:

$$\text{Per cent of modulation} = \frac{\frac{1}{2} \times (1\frac{1}{2} - \frac{1}{2}) \times 100}{1}$$

$$= \frac{\frac{1}{2} \times 1 \times 100}{1} = \frac{1}{2} \times 100 = 50\%$$

Or, if formula (3) is used, the substitution will appear as follows:

$$\text{Per cent of modulation} = \frac{(1\frac{1}{2} - 1) \times 100}{1} = \frac{\frac{1}{2} \times 100}{1} = 50\%$$

That the average power in the antenna increases 1½ times during 100% modulation has been previously mentioned. Also it was found that with 50% modulation the average antenna power increase was 1⅛ times. Thus it is seen that the relation between percentage of modulation and the amount of power increase is not linear, because it requires a 12.5% increase in power to produce 50% modulation, and four times as much, or a 50% increase in power to secure 100% modulation. Let us now determine the relation between the antenna power increase and the percentage of modulation produced.

The average power in the unmodulated carrier is equal to the

RMS value of the carrier current squared times the value of the antenna resistance. Calling the RMS value of the unmodulated carrier I_1, the following is true:

$$P_c = I_1^2 \times R_a \qquad (4)$$

Where: P_c is the average power in the carrier and
R_a is the antenna resistance.

Throughout the previous discussion it was convenient to use peak values instead of RMS values; therefore, let us determine what the power in the antenna due to the carrier is in terms of the peak value of the unmodulated carrier. The peak value of the carrier is equal to the RMS value times 1.414. In Lesson 13 of Unit 1, it was stated that the factor 1.414 was equal to the square root of 2, therefore the peak value is found as follows:

$$I_c = 1.414 \times I_1 = \sqrt{2} \times I_1$$

Squaring both sides of this equation, we have:

$$I_c^2 = 2 \times I_1^2$$

This is true because the square of the square root of 2 is obviously 2. By dividing both sides of this equation by 2 there results:

$$I_1^2 = \frac{I_c^2}{2} \qquad (5)$$

Now by substituting the value of I_1^2 as found in equation (5) into equation (4), we have:

$$P_c = \frac{I_c^2}{2} \times R_a \qquad (6)$$

From this last equation it is seen that the power in any AC sine wave may be found by taking one-half of the square of the peak value and multiplying it by the resistance in the circuit.

In Lesson 22 of Unit 1, it was learned that a modulated wave might be considered as being composed of three AC components: an unmodulated wave having the frequency and amplitude of the carrier; an unmodulated wave known as the upper sideband having a frequency equal to the carrier frequency plus the frequency of the impressed audio, and an amplitude equal to one-half that of the carrier (if the modulation was 100%); and a third unmodulated wave called the lower sideband whose frequency was equal to that of the carrier minus the frequency of the impressed audio and whose amplitude is the same as the other sideband. It may be proved that the amplitude of each of the sidebands will be equal to one-half of the amplitude of the modulation envelope, therefore the amplitude of the sidebands will vary with the percentage of modulation.

The total average power in the modulated wave may be determined by finding the average power contributed by each of these three AC components. The average power in the carrier is given by equation (6); that due to one sideband may be found as follows: Since the amplitude of each of the sidebands is equal to one-half of the amplitude of the modulation envelope, then the sideband amplitude may be expressed in this manner.

$$\text{Amplitude of sideband} = \frac{I_m}{2}$$

Now the average power in this sideband may be found by taking one-half of this amplitude or peak value squared and multiplying it by the antenna resistance.

$$\text{Power in one sideband} = \frac{1}{2} \times \left(\frac{I_m}{2}\right)^2 \times R_a$$

$$= \frac{1}{2} \times \frac{I_m^2}{4} \times R_a$$

$$= \frac{I_m^2}{8} \times R_a$$

Since this is the power in one sideband, that contributed by both of them will be twice this value.

$$P_s = 2 \times \frac{I_m^2}{8} \times R_a = \frac{I_m^2}{4} \times R_a \qquad (7)$$

Where: P_s is the power in the sidebands.

Thus, the total average power in the modulated wave is equal to that furnished by the carrier as given in equation (6) and that contributed by the sidebands as given in equation (7).

$$P_m = \frac{I_c^2}{2} \times R_a + \frac{I_m^2}{4} \times R_a$$

$$= \left(\frac{I_c^2}{2} + \frac{I_m^2}{4}\right) \times R_a \qquad (8)$$

Where: P_m is the total average power in the modulated wave.

Equation (8) gives the total amount of average power in the modulated wave for any percentage of modulation. It would, perhaps, be convenient to know the ratio of the sideband power to that contained in the carrier. This may be found by the following method: Equation (6) is first multiplied through by 2. This gives:

$$2 \times P_c = I_c^2 \times R_a \qquad (9)$$

Next equation (7) is multiplied through by 4, which produces:

$$4 \times P_s = I_m^2 \times R_a \qquad (10)$$

Now equation (10) is divided by equation (9), and this results in:

$$\frac{4 \times P_s}{2 \times P_c} = \frac{I_m^2 \times R_a}{I_c^2 \times R_a} \text{ or } 2 \times \frac{P_s}{P_c} = \frac{I_m^2}{I_c^2} \qquad (11)$$

As shown in equation (1), $I_m/I_c \times 100$ is equal to the percentage of modulation, therefore I_m/I_c is equal to the degree of modulation expressed as a decimal fraction, or:

$$m = \frac{I_m}{I_c}$$

Where: m is the degree of modulation

By squaring both sides of this equation we obtain:

$$m^2 = \frac{I_m^2}{I_c^2}$$

Therefore m^2 may be substituted for the right hand member of equation (11):

$$2 \times \frac{P_s}{P_c} = m^2$$

Dividing this last equation through by 2 there results:

$$\frac{P_s}{P_c} = \frac{m^2}{2} \qquad (12)$$

This equation states that the ratio of the power in the sidebands to the power in the carrier is equal to one-half of the square of the degree of modulation expressed as a decimal.

Equation (12) may be made more useful by multiplying it through by P_c as follows:

$$P_s = \frac{m^2}{2} \times P_c \qquad (13)$$

The power in the sidebands represents the increase in average power during modulation and must be supplied by the output of the alternator or modulator. It is not equal to the power output of the alternator, because the efficiency is less than 100%, however, the sideband power is proportional to the power of the alternator. Likewise, the power in the carrier is proportional to the DC power input to the tube furnished by the B battery. For this reason, equation (13) may be used to determine how much audio power is needed from the modulator to modulate a given amount of DC power input to any degree of modulation. For example, suppose that the DC power input to a tube is 200 watts, and it is desired to modulate the carrier 100%. How much audio power is needed to accomplish this? Substituting 200 watts for P_c, and 1 for m in equation (13), we have:

$$P_s = \frac{1^2}{2} \times 200 = \tfrac{1}{2} \times 200 = 100 \text{ watts of audio power.}$$

If it were desired to modulate this carrier 50%, the required amount of audio would be:

$$P_s = \frac{.5^2}{2} \times 200 = \frac{.25}{2} \times 200 = .125 \times 200 = 25 \text{ watts of audio power}$$

To modulate this same 200 watts, 10% would require:

$$P_s = \frac{.1^2}{2} \times 200 = \frac{.01}{2} \times 200 = .005 \times 200 = 1 \text{ watt of audio power}$$

We should now be able to calculate the amount of audio power necessary to modulate any amount of DC power input to any degree. The several formulas previously given for calculating the percentage of modulation are all correct, but they involve quantities not ordinarily known or easily found. The thermocouple ammeter in the dummy antenna reads effective or R.M.S. current values, while the preceding formulas require that peak values be known. It would, therefore, be desirable to introduce a percentage of modulation formula in terms of R.M.S. values.

The effective current value of the unmodulated carrier is, of course, equal to .707 times the peak value. The effective current value of the modulated carrier, however, does not have this relation to its peak value, but instead, depends on the degree of modulation. This is true because the modulated carrier is not a pure sine wave, but rather is a sine wave with a varying amplitude. This fact need give us no concern if we remember that the effective current value of any wave form, no matter how much it differs from a sine wave, is that value which, when squared and multiplied by the resistance of the circuit, gives the true power dissipated in the circuit. With this in mind, the following formula may be derived. Equation (12) is multiplied by 2. This gives:

$$2 \times \frac{P_s}{P_c} = m^2 \qquad (14)$$

By taking the square root of both sides of equation (14), the following is secured.

$$m = \sqrt{2 \times \frac{P_s}{P_c}} \qquad (15)$$

The power contained in the sidebands is equal to the total power in the modulated wave less the power contributed by the unmodulated carrier. This may be expressed as:

$$P_s = P_m - P_c \qquad (16)$$

We may, therefore, substitute the value of P_s given in equation (16) into equation (15).

$$m = \sqrt{2 \times \frac{(P_m - P_c)}{P_c}} \qquad (17)$$

From equation (4), we know that the power in the unmodulated carrier is:

$$P_c = I_1^2 \times R_a \qquad (18)$$

Where: I_1 is the effective value of the carrier current.

Also, the power in the modulated wave may be found by:

$$P_m = I_2^2 \times R_a \qquad (19)$$

Where: I_2 is the effective current value of the modulated wave.

Substituting for P_c and P_m, the values given in equations (18) and (19) into equation (17), we have:

$$m = \sqrt{\frac{2 \times (I_2^2 \times R_a - I_1^2 \times R_a)}{I_1^2 \times R_a}}$$

$$= \sqrt{\frac{2 \times R_a \times (I_2^2 - I_1^2)}{I_1^2 \times R_a}}$$

$$= \sqrt{\frac{2 \times (I_2^2 - I_1^2)}{I_1^2}}$$

$$= \sqrt{2 \times \left(\frac{I_2^2}{I_1^2} - \frac{I_1^2}{I_1^2}\right)}$$

$$= \sqrt{2 \times \left(\frac{I_2^2}{I_1^2} - 1\right)}$$

Or finally:

$$m = \sqrt{2 \times \left(\frac{I_2}{I_1}\right)^2 - 2} \qquad (20)$$

This equation may be used to determine the degree of modulation and, if the result is multiplied by 100, the modulation is given in per cent. I_1 is the effective value of the unmodulated carrier as read on the thermocouple ammeter, and I_2 is the reading of the thermocouple ammeter during modulation.

The average power in the antenna increases as the square of the effective current during modulation; therefore, the effective current must increase as the square root of the power increase. During 100% modulation, the average power in the antenna increases 1.5 times, and this means that the effective current must have increased the square root of 1.5 times, or 1.225 times. This is the same as saying that with 100% modulation, the reading of the thermocouple ammeter increases 22.5%. For these results to be obtained, the carrier must be modulated by a single audio frequency of sine wave form. The following table will help in the solving of modulation problems without the use of formulas.

Per Cent of modulation	Ratio of average power in modulated wave to power in unmodulated wave	Ratio of effective current of modulated wave to effective current of unmodulated wave	Ratio of AF power output of modulator to DC power input of modulated stage.
100	1.5	1.225	.5
95	1.45	1.204	.45
90	1.4	1.18	.4
85	1.36	1.166	.36
80	1.32	1.148	.32
75	1.276	1.13	.276
70	1.245	1.115	.245
65	1.211	1.1	.211
60	1.18	1.08	.18
55	1.15	1.07	.15
50	1.125	1.06	.125

A graph illustrating the relation between the increase in antenna current and the per cent of modulation is shown in Fig. 12.

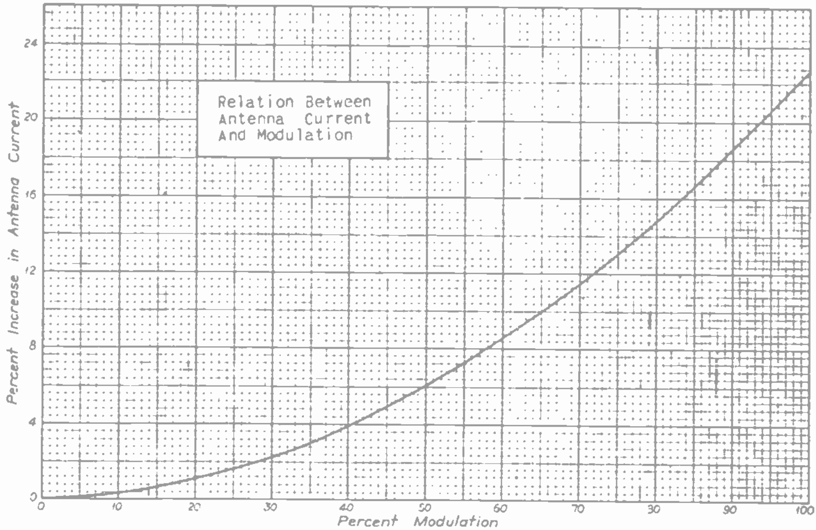


Fig. 12 Chart illustrating the relation between the percent of modulation and the per cent increase in the antenna current.

4. OVERMODULATION. When the carrier is being modulated by speech or music, the percentage of modulation changes from instant to instant. The complex audio wave form representing speech or music is not of constant amplitude, and so the modulated R.F. wave

it produces does not have a constant degree of modulation. It is not possible for the average percentage of modulation to be very large, for this will cause the occasional audio peaks to produce a percentage of modulation greater than 100%. Modulation greater than 100% is called "overmodulation", and when produced by a sine wave audio signal, it creates a wave form such as shown in Fig. 13. Notice that the carrier wave is cut off entirely during a part of the modulation cycle. Naturally, this results in severe audio distortion, and is highly undesirable.

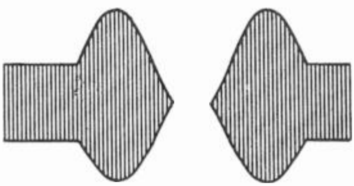


Fig. 13 Showing how overmodulation causes the carrier to be cut off during a part of the modulation cycle.

If the envelope of the modulated wave were a single sine-wave frequency, prevention of overmodulation would be very easy, but with the extreme variation of the amplitude of the envelope produced by speech and music, constant attention is required to prevent its occurrence. When the carrier wave is overmodulated, the envelope contains audio components not present in the original A.F. signal, and spurious side bands are created. This action increases the band width required for the transmission of the modulated carrier and is very liable to cause adjacent channel interference. For this reason, the Federal Communications Commission requires that overmodulation be prevented at all costs, and further than an approved type of overmodulation indicator be in operation at every transmitter. The amount of audio power fed to the modulated stage must be carefully adjusted so that the sharp peaks of the speech and music do not cause overmodulation. Since these peaks occur at relatively infrequent intervals, the average modulation is less than 100%.

Since the modulation due to speech and music is not constant, all modulation measurements are made while a single audio note is producing the modulation. While it is possible to use the effective values of current as read on the thermocouple antenna ammeter to calculate the modulation percentage, there is another method often employed to determine when 100% modulation has been obtained. This method makes use of a small thermocouple galvanometer used in a wave meter circuit. The scale of the meter is divided into 100 equal units, and, therefore, the readings of the meter are proportional to the current squared. For this reason, this instrument is often called a "current squared" meter. Since the power in the antenna varies directly as the square of the antenna current, the reading of this instrument, when it is loosely coupled to the antenna, is proportional to the amount of power contained in the antenna.

To determine when the carrier is 100% modulated, the modulation is shut off and the coupling between the pick-up coil of the instrument and the antenna is varied until the reading of the meter is 60. Then, without changing the coupling, the gain of the modulator is increased until the meter reads 90, and when this occurs, the modulation is 100%. This is true because 100% modulation produces a power increase of one and one-half times, and the increase in the reading of the meter from 60 to 90 is also an increase of one and one-half.

5. THE MODULATOR. Now that the results of modulation have been thoroughly discussed, it is time that we turned our attention to the device that is to develop the required audio power. This stage is called the "modulator", but before attempting to understand the action of the modulator, it is advisable that a few fundamental principles be reviewed.

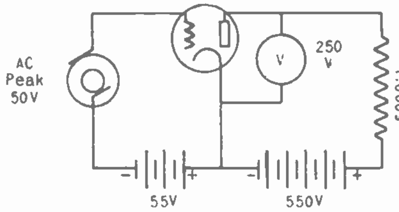


Fig. 14 An AF amplifier circuit.

Let us first consider an ordinary A.F. amplifier working into a resistive load. Such an amplifier is illustrated in Fig. 14. The excitation voltage in this case is represented by an alternator connected in the grid circuit. The bias voltage is -55 volts, and the plate voltage is 250 volts. (These are the voltages measured from grid to cathode, and plate to cathode respectively, when no excitation voltage is applied.) In the plate circuit of the tube, is a 5000-ohm resistor, and it is assumed that, under the influence of these voltages, the tube draws 60 ma. plate current. In flowing through the plate load resistor, this amount of plate current will cause a voltage drop of 300 volts. Therefore, the total voltage of the power supply or B battery must be 550 volts.

The peak voltage of the alternator in the grid circuit is 50 volts, and under excitation, the voltage of the alternator adds to and subtracts from the voltage of the C battery. This causes a number of things to happen. First, the grid voltage varies as shown at C in Fig. 15. The total variation of the grid voltage is from -5 to -105 volts. Assuming that this grid voltage variation causes the plate current to change from 20 to 100 ma., the change in plate current would be represented at B in Fig. 15. As the plate current increases, it produces a larger voltage drop across the load resistor, and there is less voltage available for the plate of the tube. In a like manner, a decrease in plate current is accompanied by a decrease in the voltage dropped across the load resistor, and the actual voltage at the plate of the tube is increased. As a result of this action, the voltage on the plate is caused to vary from 50 to 450 volts.

The important point which we wish to be impressed on you is the fact that the varying plate current and the changing plate voltage are 180° out of phase. As the current reaches its maximum value (100 ma.), the plate voltage is at its lowest value (50 volts); and when the plate current is minimum (20 ma.); the plate voltage is maximum (450 volts).

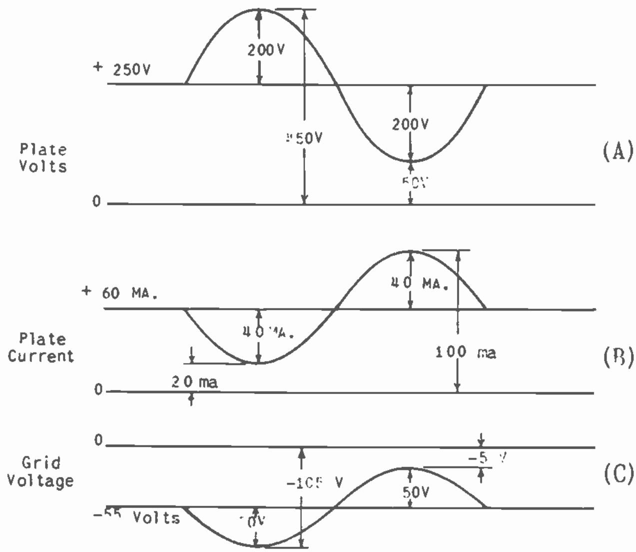


Fig. 15 The relations between the voltages and currents of the circuit shown in Fig. 14.

Let us now determine how the power output is developed. In Fig. 16 at A and B are shown the varying plate voltage and plate current of this tube, while at C, a new curve has been added. This curve represents the voltage changes produced across the load resistor. Notice that the current flowing through the load resistor is in phase with the voltage across it. The current through the load is pulsating direct current whose AC component has a peak value of 40 ma., as may be seen from the figure. Also, the voltage across the load is a pulsating direct voltage whose AC component has a peak voltage of 200 volts. We are not interested in the power dissipated by the DC components of the current through and the voltage across the load, for it does not contain the signal. To find the AC power in the load resistor, the AC voltage and current components must be reduced to RMS values. These are:

$$E_{rms} = 200 \times .707 = 141.4 \text{ volts}$$
$$I_{rms} = .040 \times .707 = .02828$$

Since the power factor in this circuit is one, the average power is calculated as follows:

W = .02828 × 141.4 = 3.998+ watts

Let us consider the phase relations of the voltages of a resistance-coupled amplifier. Such an amplifier is illustrated in Fig. 17, while the curves of Fig. 18 show the phases of the various voltages. During the first alternation, the grid of tube T_1 is made

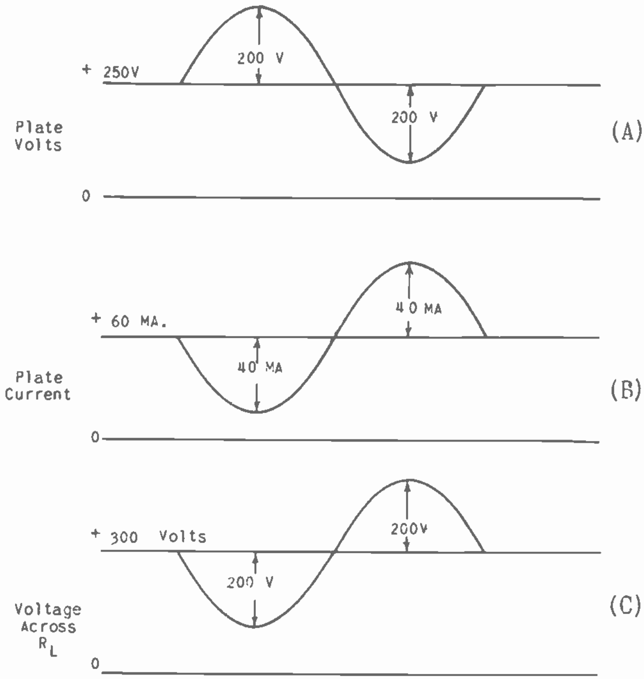
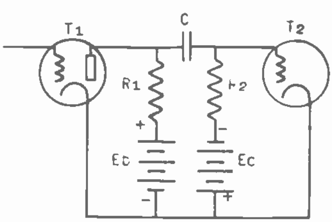


Fig. 16 Showing the phase relation between the plate current, plate voltage, and voltage across the load.

more negative than normal. Since the grid and plate voltages of an amplifier are 180° out of phase, the plate voltage of this tube is increasing above its no-signal value throughout this time. With

Fig. 17 Simplified diagram of a resistance-coupled amplifier.



an increased voltage on the plate of T_1, the left hand plate of the coupling condenser becomes more positive, and the condenser charges to a higher voltage. The only way that this can occur is for electrons to flow from this plate, down through the load resistor and the B battery, up through the C battery of the following stage, through the grid leak R_2, and on to the right hand plate of the coupling condenser. Notice that current flows through the grid leak from bottom to top. This causes the top end of the grid leak to be positive with respect to the bottom end, and the voltage across the grid leak bucks against the voltage of the C battery. Thus, the grid of T_2 is increasing in a positive direction during this interval.

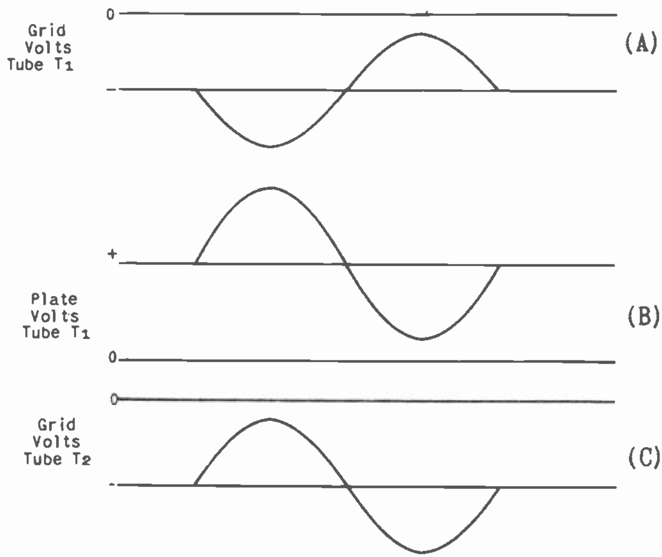


Fig.18 The phase relations of the voltages in the circuit shown in Fig.17.

Throughout the succeeding alternation, the grid of T_1 is driven less negative, and an increased plate current flows. The greater voltage drop produced across the load resistor causes the voltage of the plate of T_1 to become less positive. Thus, the left hand plate of the coupling condenser is at a lower positive voltage, and the condenser must discharge. To accomplish this, electrons flow from the right hand plate of the condenser, down through the grid leak and the C battery, and up through the B battery and the load resistor, to the left hand plate of the condenser. This flow of current through the grid leak produces a voltage drop across it which adds to the voltage of the C battery, and the grid of T_2 is driven more negative.

From this discussion, and by reference to the curves of Fig. 18, it should be apparent that the plate voltage of T_1 and the grid voltage of T_2 are in phase. This is an important point and should be remembered by the student.

A direct-coupled amplifier is no different from a resistance-coupled amplifier insofar as the phase relations of its voltages are concerned. A simplified direct-coupled amplifier circuit is illustrated in Fig. 19. The normal plate current through the resistor R provides bias on tube T_2. It is, of course, necessary that the plate voltage supply for tube T_2 be considerably larger than that used for tube T_1. A signal voltage applied to the grid of tube T_1 causes plate current variations, and the varying plate

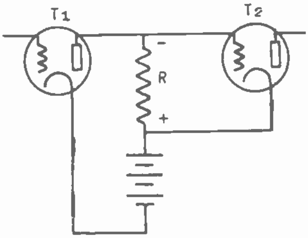


Fig.19 A simplified direct-coupled amplifier.

current produces a changing voltage across the load resistor. It is this changing voltage which serves as the grid excitation for tube T_2. In this case, it is very easy to see that the plate voltage of T_1 is in phase with the grid voltage of T_2, since both elements are connected to the same point.

If a large audio frequency choke coil were substituted for resistor R, the circuit would appear as shown in Fig. 20. However, the plate voltage of T_1 would still be in phase with the grid voltage of T_2. It is true that in this circuit, the voltage set up across the choke would not be in phase with the current through it, but whatever the phase might be, the plate voltage of the first tube would be in phase with the grid voltage of the second.

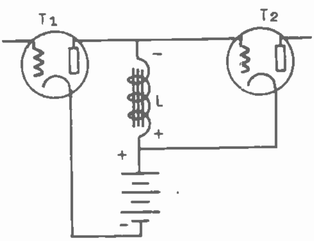


Fig.20 A direct-coupled amplifier using an AF choke for coupling.

Before continuing, it is thought advisable to refresh your memory with the idea of loading. This subject was first discussed in Lesson 1 of this Unit. It was learned that when the load impedance into which a vacuum tube or any other voltage source works is reduced, the tube or voltage source is said to have its load increased, or to be more heavily loaded. Thus, we can state that the load on a vacuum tube varies inversely as the value of the load impedance into which it works.

In Fig. 17, the load on tube T_1 is composed of the resistor R_1, and in parallel with this, there is the coupling condenser C and the grid leak R_2. In the usual voltage amplifying circuit, R_2 and C constitute such a high impedance that even though placed in parallel with R_1, they cause a negligible reduction of the load impedance into which the tube works. Therefore, for all practical purposes, the tube is working into a load equal to the value of the load resistor.

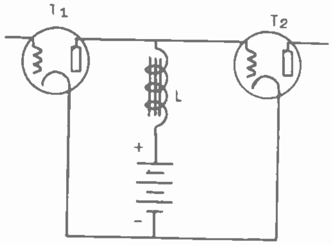


Fig. 21 Showing how the grid of the second tube would draw current and thereby provide a resistance path in parallel with the AF choke.

Now consider Fig. 21. In this circuit, the cathode of the second tube is connected to the negative terminal of the B battery, and the grid of the second tube is at the same positive potential as the plate of the first tube. This would cause the grid of T_2 to draw current, and, since there is a conducting path in parallel with the load impedance, the effect is the same as though a resistor has been connected in parallel with the load impedance. This circuit would then be equivalent to the one shown in Fig. 22. Suppose that the inductance of the choke is 20 henries, and the frequency of the signal voltage applied to the grid of T_1 is 1000 cycles. At this frequency, the inductive reactance of the choke is 123,600 ohms. When the grid of the second tube draws current, the grid-to-cathode resistance of this tube will probably be as low as 25,000 ohms.

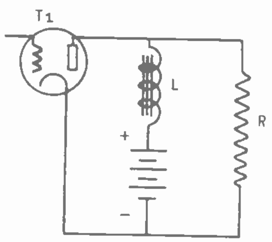


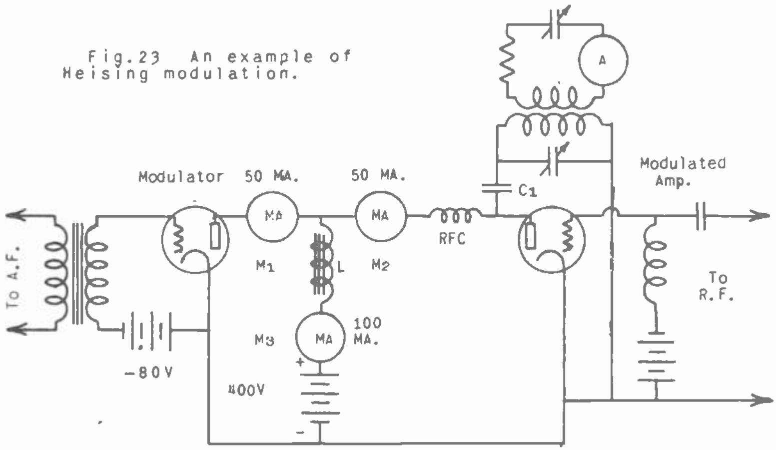
Fig. 22 A circuit equivalent to the one in Fig. 21, when the second tube draws grid current.

It is, therefore, clearly evident that tube T_1 is working into a load impedance somewhat less than 25,000 ohms in value. In fact, the load impedance of T_1 is determined, in this case, not by the reactance of the choke, but by the grid-to-cathode resistance of the second tube. By making the grid of T_2 positive and allowing this tube to draw grid current, the actual load impedance of T_1 is greatly reduced, and as a consequence, it may be stated that the load on T_1 has been increased, or T_1 is more heavily loaded.

The plate current variations of T_1 constitute the audio signal, and, since the impedance of the path through the grid-to-cathode resistance of T_2 is about one-fifth of that through the audio choke, five times as much audio current will flow through this path as will flow through the choke or practically the entire load on the tube T_1 would be represented by the grid-to-cathode resistance of T_2. This is especially true if the audio choke has a high inductance value.

6. HEISING MODULATION. We are now ready to discuss a system of modulation due to R. A. Heising, and named in honor of him. The circuit is shown in Fig. 23. It is assumed that the modulated R.F.

Fig.23 An example of Heising modulation.



stage is excited by a crystal oscillator and buffer stage not shown in the drawing. Notice that the plate voltage for both the Class C stage and the modulator stage is supplied through the iron-core choke L. This choke is known as the "Heising choke" or "modulation choke", and it should have a very high reactance to the lowest audio frequency amplified by the modulator. Assume that the correct operating voltages are applied to both tubes, and that the Class C stage is producing an unmodulated current in the dummy antenna. When an audio signal voltage is applied to the input transformer of the modulator, the grid of the modulator is caused to vary at an audio frequency rate. This causes the plate current drawn by the modulator to change, and this plate current may be considered to have a DC and an A.F. component. The DC component returns to the cathode by the route through the modulation choke and the B battery. Since the inductive reactance of the choke is very high, only a very small part of the A.F. component flows through it. The small amount of A.F. current that does pass through the choke produces

A.F. voltages across the choke which cause the plate voltage of the modulator tube to vary at an A.F. rate. Since the plate of the Class C stage is connected to the same point as the plate of the modulator (there being practically no resistance in the R.F. choke), the voltage on the plate of the modulated stage will also vary at an A.F. rate. The varying A.F. plate voltage on the Class C stage causes the DC current drawn by this stage to vary at an A.F. rate, and the peaks of the R.F. current pulses actually flowing through this tube also vary in amplitude at an A.F. rate. The R.F. component of the modulated stage is fed effectively through the plate blocking condenser C_1, and, since the amplitude of this R.F. component is varying at an A.F. rate, the R.F. energy furnished to the tank circuit also varies at this rate. This causes the oscillating tank current and antenna current to do likewise, and thus the amplitude of the antenna current and the power which would be radiated from an actual transmitting antenna is varied in direct accordance with the variations of the audio signal voltage fed to the input of the modulator.

The DC plate resistance of the Class C stage is directly in parallel with the modulation choke; and, since the DC plate resistance is considerably less than the impedance of the choke, the actual load into which the modulator delivers power is this DC plate resistance of the modulated stage. For this condition to be obtained, the impedance of the modulation choke at the lowest audio frequency to be modulated must be at least three times the DC plate resistance of the Class C stage.

We may consider that the A.F. current generated by the modulator flows through the DC plate resistance of the modulated stage, and, in so doing, causes the voltage applied to the plate of the Class C stage to be varied at an A.F. rate. To vary the plate voltage of the modulated stage requires power because an increase (or decrease) in the plate voltage is accompanied by a corresponding increase (or decrease) in the average plate current. If maximum undistorted power is to be obtained from the modulator stage, the load impedance into which the modulator delivers power must be of the correct value. This problem will be further discussed later in this lesson.

Assume that the plate voltage of the Class C stage (unmodulated) is 400 volts, and the DC plate current is 50 ma. Also, assume that the plate voltage of the modulator is 400 volts, and that it draws 50 ma. plate current when no A.F. signal is applied to its grid. The modulation choke will be designed to have a minimum amount of DC resistance, and the DC voltage developed across the choke due to its resistance will be negligible; therefore, the voltage of the power supply may be considered to be 400 volts.

With modulation, the A.F. signal voltage applied to the grid of the modulator has a peak value of 80 volts. (This is assuming that the grid bias of the modulator is -80 volts, and that the application of a signal voltage of this value will neither drive the grid of this tube positive, or cause it to operate on the lower bend of its E_g-I_p curve.) With this condition, the various voltages and currents of the two tubes are graphically illustrated in Fig. 24. At A is shown the variation of the modulator grid voltage; it is seen to vary from -160 to 0 volts. The modulator plate current is shown at B; it varies from 12.5 ma. to 87.5 ma. C gives the picture of the variation of modulator plate voltage; it varies from 100 to 700 volts.

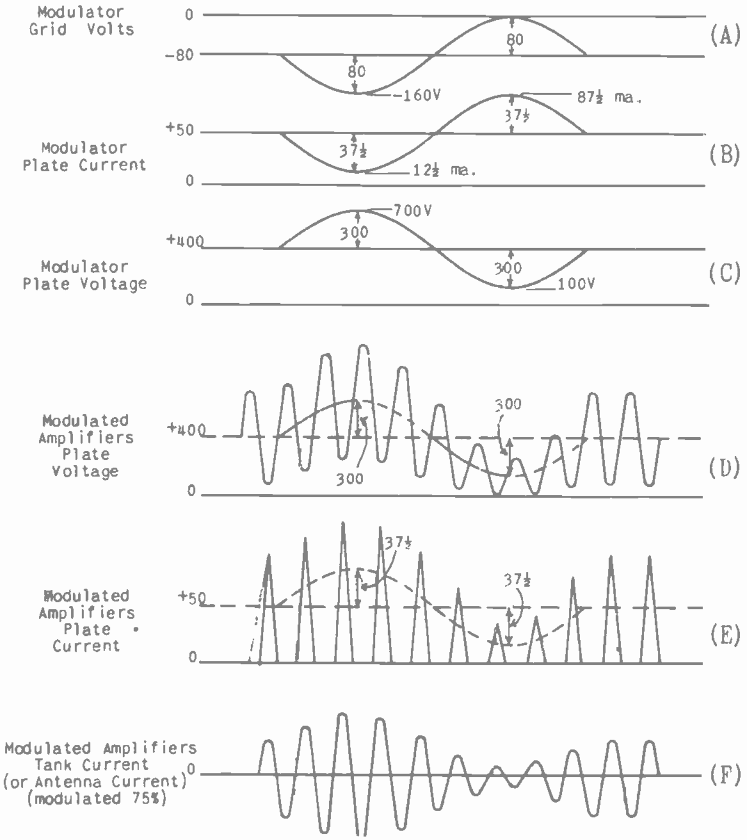


Fig.24 Showing how the currents and voltages of the modulator and modulated stages vary.

Insofar as the power supply is concerned, the modulator and Class C tubes are in parallel, and whatever voltage is applied to the plate of the modulator is necessarily present at the plate of the Class C stage. It has been stated that the plate voltage of the modulator varies at an A.F. rate, and, for this reason, the voltage applied to the plate of the Class C stage must also vary at an A.F. rate in direct accordance with the modulator plate variations. In addition, the plate voltage of the Class C stage is influenced by the R.F. voltages developed across the plate tank circuit. It is, therefore, evident that the plate voltage of the

Class C stage is rather complex, consisting as it does of three components.

Fig. 24 at D shows this complex plate voltage of the Class C stage. The DC plate voltage furnished by the power supply is 400 volts; the A.F. variations cause the voltage applied to the plate to change from 100 to 700 volts and the R.F. voltages present across the tank circuit are usually sufficient to cause the plate voltage to vary from nearly twice the applied value at that instant to a very low value.

In a like manner, the plate current of the modulated stage, shown at E in this figure, consists of three components. These are the R.F. current pulses which last for perhaps one-fifth of an R.F. cycle or less, and which are not of constant amplitude but vary in amplitude at an A.F. rate; the A.F. current component, which is the average value of the R.F. pulses, and, in this case, has a peak value of 37.5 ma.; and the DC component which is the average value of the A.F. component, and has a value of 50 ma. This complex plate current causes the oscillating current in the tank and antenna circuits to have the form shown at F in the figure.

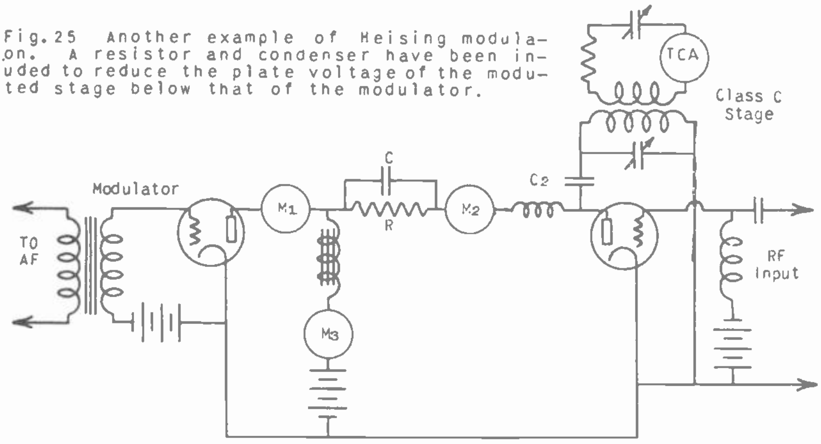
The R.F. component of the plate current of the Class C stage is prevented from flowing through the power supply by means of the R.F. choke. The meters used in this circuit read only DC current values, and should not change during modulation; that is, if the modulator and Class C stage are properly adjusted, their DC plate currents will remain the same whether the modulator is supplying A.F. power or not.

Heising modulation is sometimes called "the constant current" system of modulation. This name results from the fact that the total current drawn from the power supply is practically constant. Assuming that this is so, an increase in the modulator plate current must be accompanied by a corresponding decrease in the plate current drawn by the Class C stage. Further, the plate current of the Class C stage must increase as the plate current of the modulator decreases. That this is true is evident from curves B and E of Fig. 24. When the plate current of the modulator is minimum (12.5 ma.), the voltage applied to the plate of the Class C stage is greatest, and this tube consequently draws its maximum current. Of course, the plate current of the Class C stage is also varying at an R.F. rate, but the R.F. variations are not able to pass the R.F. choke. Likewise, when the plate current of the modulator is maximum (37.5 ma.), the current drawn by the modulated stage is minimum. It is not strictly correct to say that the current through the modulation choke does not change at all, for if there were no current changes through this choke, there could be no A.F. voltages developed across its inductive reactance. However, the impedance of this choke is so large that only a small change in current through it is necessary to develop the A.F. voltages across it which change the plate voltages of the modulator and the Class C stage.

Upon inspection of curve F in Fig. 24, it is seen that the amplitude of the antenna current does not fall to zero, nor is the peak amplitude equal to twice that of the unmodulated carrier. Thus, this wave is not 100% modulated, but represents a degree of modulation of about 75%. To produce 100% modulation, the plate voltage of the Class C stage would have to be reduced to zero during an instant of the A.F. cycle. In this circuit, however, the DC voltage applied to the plate of the Class C stage is the same as that applied to the plate of the modulator. Also, notice that the modulator draws the most plate current when its plate voltage is lowest. Now it is manifestly impossible for the modulator tube to draw a high plate current, or for that matter, any plate current, if its plate voltage is zero. For this reason, it is not possible to produce 100% modulation with this circuit, since the DC voltage applied to the modulator and to the Class C stage are equal.

To modulate the carrier wave to as large a percentage as possible without causing overmodulation is always the principal objective. As the amount of modulation is reduced below 100%, the power contained in the sidebands is rapidly diminished. Unless the carrier is deeply modulated, the strength of the carrier is no indication of how well the program will be received. For example, a 1000 watt carrier modulated 50% is no more effective than a 250 watt carrier modulated 100%; each contain 125 watts of sideband power.

Fig. 25 Another example of Heising modulation. A resistor and condenser have been included to reduce the plate voltage of the modulated stage below that of the modulator.



As pointed out, it is not possible to produce 100% modulation when the plate voltage of the modulator is equal to that of the Class C stage. This condition, however, may be remedied by using the circuit shown in Fig. 25. This circuit is similar to that of Fig. 23, with the exception that a high-wattage resistor shunted by a high-capacity condenser is connected in series with the plate lead of the Class C stage. The purpose of the resistor is to reduce the DC plate voltage applied to the Class C stage below that value applied to the modulator. With this arrangement, the plate voltage on the Class C stage may be reduced to zero for an instant during the A.F. modulation cycle without causing the voltage applied to the plate of the modulator to also fall to zero. The resistor should dissipate only DC power and no A.F. power. For this condition to result, the resistor must be shunted by a condenser whose reactance to the lowest modulation frequency is very small. Ordinarily a condenser of from 5 to 20 mfd. is required.

The resistor used to reduce the DC voltage applied to the plate of the Class C stage must be large enough to cause the actual DC plate voltage of the Class C stage to be equal to the peak alternating plate voltage of the modulator tube. It is shown in the graphs of Fig. 24 that with 400 volts applied to the plate of the modulator, the total variation of the plate voltage is from 100 to 700 volts. This pulsating DC voltage has an AC component whose peak value is 300 volts. Therefore, with 300 volts DC voltage applied to the plate of the Class C stage, the actual variation of the Class C plate voltage is from 0 to 600 volts, or from zero to twice its normal value. Such a variation of the applied plate voltage is a condition necessary for obtaining 100% modulation. By inspecting the curves of the particular tube used in the Class C amplifier, it would be possible to foretell what DC plate current the tube would draw when its plate voltage is 300 volts (assuming that the bias remains constant), and thereby calculate the size of the resistor needed to drop the plate voltage from 400 to 300 volts.

It is to be noted that the audio frequency output of the modulator is supplied to the Class C stage through the R.F. choke. The reactance of this choke, at the highest modulating frequency, should be small compared to the DC plate resistance of the Class C tube. This limits the choke to a value no larger than that necessary to exclude R.F. currents from the modulator tube. This condition is easily achieved when the carrier frequency used is high in comparison to the highest modulation frequencies. But with low carrier frequencies, the R.F. choke required may offer appreciable impedance to the higher modulation frequencies, with resultant frequency distortion.

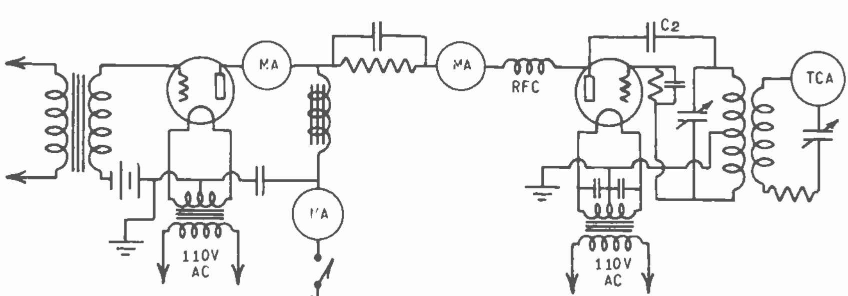


Fig.26 A Hartley oscillator used with Heising modulation.

In addition, the blocking condenser C_2 is effectively in parallel with the DC plate resistance of the Class C tube, since the tank impedance is negligible at audio frequencies. Consequently, the higher modulation frequencies tend to be by-passed through this condenser, instead of flowing through the resistance of the tube, if this condenser has too high a capacity. It should be no larger than necessary to provide a low-impedance path for the R.F. component of the plate current.

A Hartley oscillator coupled to a Heising modulator is shown in Fig. 26. The operation is exactly the same as when a Class C stage is modulated; however, it is difficult to secure 100% modulation, because of the inability of the oscillator to oscillate when its applied plate voltage is reduced to a low value. In addition, the output frequency of the oscillator is somewhat dependent on its plate voltage, and the changes in its applied plate voltage that occur during modulation tend to produce some frequency modulation. For these reasons, modulated oscillators are not used in broadcasting.

7. **ADJUSTING THE LOAD ON THE MODULATOR.** To modulate the Class C stage requires power, and to obtain maximum undistorted power from the modulator necessitates that the load impedance into which the modulator tube works be equal to twice the plate resistance of this tube. The load impedance of the modulator is the DC plate resistance of the Class C stage. This resistance value is determined by dividing the applied DC plate voltage by the DC plate current. It would, therefore, seem possible to adjust the load on the modulator to the correct value by varying the DC plate voltage and plate current of the Class C stage until the DC plate resistance was of the proper value. Quite often this may be done, however, there are other conditions which must also be satisfied.

There are many values of DC plate voltage and plate current which will produce the correct load impedance for the modulator, but there is only one set of values of voltage and current which will satisfy the load impedance consideration, and, at the same time, furnish the proper DC power input to the Class C stage. Furthermore, care must be taken to insure that the DC voltage on the plate of the Class C stage is less than that on the modulator. The DC power input to the Class C stage should be so adjusted that the power output from the modulator is able to produce 100% modulation; that is, it should be exactly equal to twice the modulator's power. The DC power input is found as follows:

$$\text{DC power input to Class C stage} = E_B \times I_B$$

Where: E_B is the applied DC voltage to the plate of the Class C stage and
I_B is the DC plate current of the Class C stage.

Likewise, the DC plate resistance of the Class C stage is:

$$DC\ R_p = \frac{E_B}{I_B}$$

The DC plate voltage and plate current must be so selected that they satisfy both of these equations.

With the correct load impedance of the modulator, and the proper DC power input to the Class C stage known, the necessary DC plate voltage and plate current of the Class C stage may be found by these formulas:

$$I_B = \sqrt{\frac{P_i}{R_L}} \qquad E_B = \sqrt{P_i \times R_L}$$

Where: P_i is the DC power input to the Class C stage, and R_L is the load impedance into which the modulator should work.

For example, assume that the A.F. power output of a modulator is 100 watts, and that the modulator has a plate resistance of 4000 ohms. This 100 watts of A.F. power is capable of modulating 200 watts of DC power input to the Class C stage 100%. Also, if the total 100 watts of A.F. power is to be obtained, the modulator tube must work into a load impedance of 8000 ohms, or twice the value of its plate resistance. We now know the required DC power input, and the necessary value of load impedance for the modulator. The value of DC plate voltage that must be applied to the Class C stage and the DC current that this tube must draw may now be found by using the two preceding formulas.

$$I_B = \sqrt{\frac{200}{8000}} = \sqrt{.025} = .158 \text{ ampere or } 158 \text{ ma.}$$

$$E_B = \sqrt{200 \times 8000} = \sqrt{1,600,000} = 1265 \text{ volts.}$$

Since the DC plate voltage of the Class C stage must be 1265 volts, it is obvious that the DC plate voltage of the modulator must be in excess of this value, probably 1500 volts or more. To cause the Class C stage to draw 158 ma. when its plate voltage is 1265 volts will probably necessitate an adjustment of the bias and excitation voltage on the Class C tube.

Sometimes the results obtained by the use of the foregoing formulas are impractical. It may happen that either the DC plate voltage or plate current, or both, as found by these formulas, exceeds the manufacturer's specifications for the particular tube involved. In this case, a mismatch between the modulator and its load must be tolerated. When it is not practical to match the load exactly to the modulator, it is better to make the load larger than the proper value, rather than smaller. This tends to cause a smaller reduction in the power output of the modulator, and aids in the minimizing of distortion. To make the load impedance of the modulator larger, the DC plate voltage of the Class C stage must be increased, and the DC plate current reduced. If the DC plate voltage found by the formula does not exceed the rated value, it should be used; while, if the DC plate current needed to provide the correct load impedance is in excess of the allowable maximum, a smaller plate current will have to be employed which tends to increase the load impedance of the modulator.

This form of Heising modulation was in popular use in transmitters until approximately 1934, and there are probably many transmitters which still employ this system. Its several disadvantages, however, have led to the development of other types of modulation of improved form. The first disadvantage of this type of modulation is the fact that it is not always possible to match the modulator to its load impedance; a second, is the possibility of saturating the modulation choke. The modulation choke must carry the DC current of both the Class C stage and the modulator stage. This relatively large DC current tends to saturate the core of the choke and thereby causes the effective inductance of the choke to changes in current to be less than its rated value. This is partially overcome by using a choke of unusually large core area. Also, the choke must be wound with fairly large wire so that it may safely carry the required amount of current.

A system of modulation which reduces the tendency of the DC plate current for saturating the modulation choke is shown in Fig. 27. It is commonly called "double-choke" Heising modulation. The choke L_1 carries only the DC current of the modulator, while L_2, has only the DC current of the Class C stage passing through it; thus, there is less possibility of core saturation. The condenser, which couples the output of the modulator to the input of the Class C stage must be large enough to have a low reactance to the lowest modulation frequency amplified by the modulator.

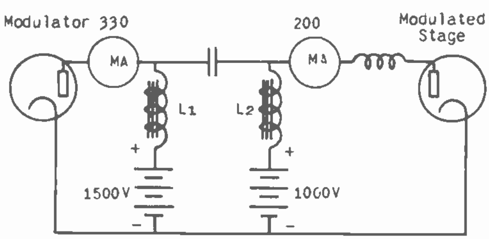


Fig.27 An example of "double - choke" Heising modulation.

The load on the modulator, in this case, consists of the parallel combination of the first choke, the second choke, and the DC plate resistance of the Class C stage. The latter item will constitute 95% of the load, and the same amount of juggling of the DC plate voltage and plate current of the Class C stage will have to be done, in order to match the modulator to its load. This system does, however, possess the advantage of allowing the use of two power supplies, if desired. This is seldom considered an advantage in commercial practice, because one large power supply is usually employed; however, in amateur transmitters, it is an advantage to be able to use two smaller power supplies, rather than one large one, due to economic considerations.

A modulation - coupling system that is occasionally found is illustrated in Fig. 28. This system makes use of a tapped modulation choke, or, as it is usually called, an auto-transformer. An auto-transformer is a transformer having just one winding which is tapped somewhere along its length. If it is to be employed as a step-up transformer, the primary is that part of the winding between one end and the tap, while the entire winding constitutes the secondary. When an alternating voltage is applied between one end of the coil and the tap, the alternating current that flows creates a magnetic field which cuts through the turns of the entire coil, and thus induces a voltage across the whole winding. The ratio of the voltage across the whole coil to the voltage applied between one end of the coil and the tap is equal to the ratio of the total number of turns in the winding to the number of turns between which the primary voltage is applied.

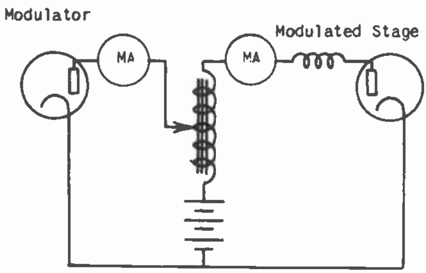


Fig.28 A modulation system using a tapped audio choke.

The auto-transformer allows the same DC plate voltage to be applied to both the modulator and the modulated stage. While it is not possible for the alternating voltage on the plate of the modulator to be large enough to cause the plate voltage of the modulator to vary from zero to twice its normal value, yet this voltage may be stepped up by the auto-transformer until it is of sufficient magnitude to vary the voltage applied to the plate of the Class C stage from zero to twice its DC value, and thus effect 100% modulation.

The load impedance into which the modulator operates is not equal to the DC plate resistance of the Class C stage in this system, but is equal to this resistance value multiplied by the square of the turn ratio from primary to secondary of the auto-transformer. The auto-transformer, like an ordinary transformer, reflects an impedance value from secondary to primary, which is equal to the impedance connected across the secondary (in this case, the DC resistance of the Class C tube) times the square of the turn ratio from primary to secondary. Thus the auto-transformer serves as an impedance-matching device, which places the correct load on the modulator.

About the only disadvantage of the auto-transformer type of coupling between modulator and modulated stage is the fact that no provision is made to eliminate core saturation. A system which retains all the desirable features of the auto-transformer method and still reduces core saturation to a negligible value is shown in Fig. 29. A regulation iron-core transformer replaces the auto-transformer. The current of the modulator is drawn through the primary, while the current of the Class C stage flows through the secondary. By winding the two coils in opposite directions, the

DC flux around the secondary tends to cancel the DC flux surrounding the primary, and thus the net DC flux in the common core is reduced to a value too small to cause magnetic saturation. Unless the ampere-turns of the secondary are equal to the ampere turns of the primary, and there is no magnetic leakage, complete cancellation of the DC fluxes will not be obtained, This, however, is not important as long as the net DC flux remaining in the core is insufficient to saturate it.

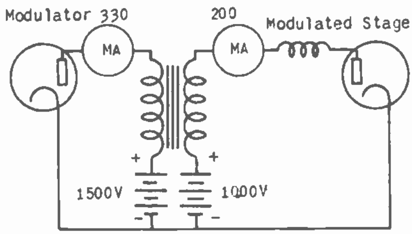


Fig.29 A modulator coupled to the Class C stage by means of a modulation transformer.

The types of modulators which we have been discussing are all Class A audio amplifiers. As you will remember, a Class A amplifier is one in which the plate current flows during the complete cycle. The advantage of a Class A amplifier is its low distortion content. Its major disadvantage is the fact that its efficiency is relatively low. This is of little importance in receivers where small amounts of power are involved, and, for this reason, nearly all amplifiers used in receivers (both R.F. and A.F.) are of the Class A type, in order to reduce distortion to a minimum.

The audio power required in a transmitter to modulate the carrier wave, however, is considerable; and the matter of efficiency is of no small consequence. For example, suppose that a 1000 watt carrier is to be modulated 100%. The efficiency of the Class C stage will be about 60%, and the necessary DC power input to this stage will be 1,666 watts. To modulate this amount of DC power input, requires that the A.F. power output of the modulator be 833 watts. The efficiency of a Class A amplifier will average 20%, and with this efficiency, the DC power input required by the Class A modulator would be 4,165 watts. Of this power applied to the Class A modulator, 80% will have to be dissipated at the plate of the tube. This amounts to 3,333 watts. It is at once evident that the modulator used in such a transmitter would have to be a very rugged, high-powered tube. Practically, such a tube was never used in ordinary transmitters, due to the expense of such a tube and to the cost of the power supply needed to furnish the very high voltage required. Instead, many smaller tubes were connected in parallel, and each of the small tubes furnished its share of the A.F. output. By this method, a large amount of A.F. power could be obtained without using excessive plate voltages. In addition, tubes were often connected in push-pull, or four tubes were used in a parallel, push-pull arrangement. However, as long as the tubes were operated under Class A conditions, their efficiency was low, and a considerable amount of power was wasted.

A diagram of a Class A, push-pull modulator coupled to a Class C stage by a modulation transformer is shown in Fig. 30. The use of a push-pull modulator enables the obtaining of a relatively high A.F. power output without using extremely high plate voltages; also, it is well known that the push-pull connection minimizes audio distortion. The DC ampere-turns on the two halves of the primary of the modulation transformer cancel each other, and therefore are not

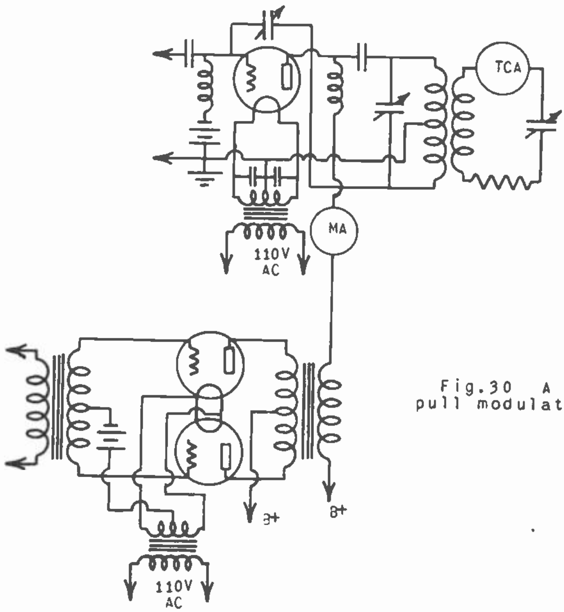


Fig.30 A Class A, push-pull modulator.

available to oppose the ampere turns of the secondary caused by the DC plate current of the Class C stage flowing through it. For this reason, the modulation transformer will require a larger core than that used with one tube, if core saturation is to be avoided.

6. A MODERN MODULATOR AND ITS COUPLING SYSTEM. The low efficiency of a Class A amplifier has led to the development of Class B amplifiers for modulators. The theoretical discussion of Class B amplifiers will be considered in Lesson 6 of this Unit. At the present, it is sufficient to state that a Class B amplifier is one in which plate current flows for approximately 180° or one-half of a cycle. When used as an audio power amplifier, a Class B stage requires two tubes connected in push-pull, and it draws some grid current. The power dissipated in the grid circuit must be supplied by the preceding stage, which is ordinarily a power amplifier operated under Class A conditions. The advantage of a Class B modulator is the fact that a much larger power output may be secured at a higher efficiency. The dis-

advantage is a somewhat greater amount of distortion than that obtained from Class A amplifiers.

A diagram of a circuit using a Class B modulator coupled to a modulated Class C stage is shown in Fig. 31. In appearance, the Class B modulator is very similar to a Class A push-pull amplifier. In actual commercial modulators, there are a few minor changes which will be discussed when Class B amplification is taken up. The load on the Class B modulator is quite critical, and, unless it is carefully matched to the resistance of the Class C stage, the desired power output from the modulator will not be obtained and the amount of distortion will be excessive. For this reason, an impedance matching transformer is always used with Class B modulators.

Also observe the peculiar coupling arrangement between the modulator and the modulated stage. The plate current of the Class C stage is not allowed to flow through the secondary of the modulation transformer, because to do so would increase the distortion of the Class B modulator due to saturation of the core. Instead, the audio voltages built up across the secondary of the modulation transformer are transferred through a high-capacity condenser to the Class C stage. This condenser must have a low reactance at the lowest audio frequency to be amplified. The plate current of the Class C stage is fed through a modulation choke having a very large core area. This choke must be so designed that the core saturation is negligible, and so that its reactance at the lowest modulation frequency is much larger than the DC resistance of the Class C stage.

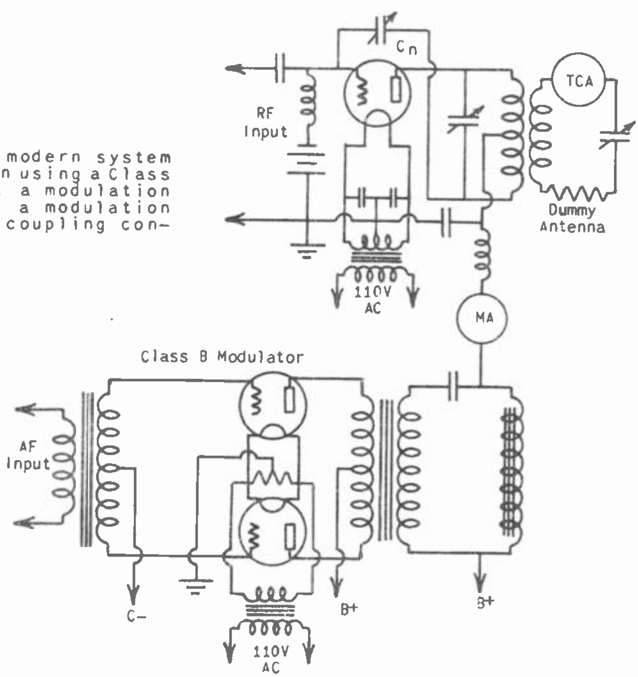


Fig.31 A modern system of modulation using a Class B modulator, a modulation transformer, a modulation choke, and a coupling condenser.

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT NO. 3

CLASS A AND B AMPLIFIERS

LESSON NO. 6

DAVID AND GOLIATH

.......AND YOUR TRAINING

David was a little fellow compared to Goliath.

And when he faced this brawny giant it seemed utterly impossible that he could even hope to remove the huge obstacle that barred his onward path.

Goliath had might. He was immensely powerful. His scowling face and gnashing teeth were something fearful to behold. And David seemed so tiny that Goliath laughed a heartless laugh. For how could this little fellow do him any harm.

But David had something that Goliath did not have. David had a mind that was trained to "think". He was courageous and determined. His muscles were small, but they too were trained. And as they swung the sling a stone was propelled through the air with amazing swiftness.......straight to the spot on Goliath's head where David's eyes were centered.

Goliath fell to the ground with a crash, completely mastered by a smaller, but determined man. The obstacle was removed.

When you first started your Midland training, perhaps it seemed that the lessons ahead of you were as Goliaths in your path. But as each lesson was mastered, Goliath became smaller and smaller. And now you have reduced him to a size that can be easily conquered. You are well on your way to your objective in your chosen profession.

David was impelled to action by the will to win. And he won.

You too have the will to win. If you will continue your march to success with this same spirit,.......

YOU WILL WIN.

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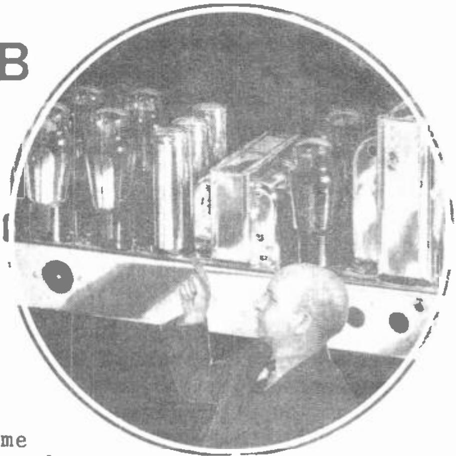
Lesson Six

CLASS A & B AMPLIFIERS



"In the study of radio it is highly essential that you have a very thorough knowledge of the different classes of operation under which a vacuum tube is operated. While you do not need to be materially concerned about some of the formulas used in this work, still you must know how tubes are operated when used in the various classes of service.

"Class A amplification was the only type used in modulation during the early years of radio; however, it was not until Class B modulators were developed that high-powered high-level modulation became economically successful."

1. THE VARIOUS TYPES OF AMPLIFICATION. At times throughout the preceding lessons of this course, reference has been made to the different types of amplifiers. These are Class A, B, and C amplifiers. The distinction between these amplifier types was mentioned briefly in a lesson of Unit 1.

Since a complete understanding of the operation of each type is essential for the student engineer, this lesson will be devoted to the discussion of Class A and B amplifiers, and the following lesson will deal with Class C amplifiers.

Class A amplification is used in practically all receivers in both the audio and R.F. stages. In transmitters, a Class A amplifier is occasionally employed for the modulator stage. Class B audio amplification finds its major use in the modulators of modern transmitters, and in high-powered public address systems. The Class B type of amplification is also employed to amplify radio frequencies, and when so used, is called a "linear amplifier". Linear amplifiers are often found in transmitters; their purpose is to amplify the modulated carrier. Since the requirements for Class B audio amplifiers and linear amplifiers are somewhat different, linear amplifiers will be treated separately. A complete discussion of their operating characteristics will be presented in a later lesson. A Class C amplifier may be used only to amplify an unmodulated radio wave, because the conditions under which it operates would produce distortion in audio frequencies or modulated radio frequencies.

Before continuing this discussion, let us make sure that the major difference between the various classes of amplification is known. The following definitions are practically the same as those given in a preceding lesson.

Class A Amplifier: In a Class A amplifier, the grid bias and alternating signal voltages are so chosen that plate current flows all the time, or for 360° of the input cycle.

Class B Amplifier: In a Class B amplifier, the grid bias is approximately equal to the cutoff value and, as a result, the plate current is practically zero when no signal voltage is applied to the grid. Such a condition causes the plate current to flow during approximately 180° of the grid-exciting cycle.

Class C Amplifier: In the Class C amplifier, the grid bias is appreciably greater than the cutoff value. It may vary from one and one-half times cutoff to twice cutoff or more. With this high bias, no plate current flows unless a grid excitation voltage is applied. Under excitation, the plate current flows in the form of pulses lasting less than one-half cycle, or less than 180°.

In addition to these three general classifications of amplifiers, there are two intermediate types which should be given in order to make this list complete. These are:

Class AB Amplifier: In a Class AB amplifier, the grid bias and excitation voltage are so adjusted that plate current flows for more than one-half cycle, but less than the full cycle; that is, the plate current flows for less than 360°, but more than 180°. A Class AB amplifier is also called a Class A-prime amplifier. This type is sometimes subdivided into two parts, which are known as Class AB_1 and Class AB_2. A Class AB_1 amplifier is one which does not draw grid current; whereas, a Class AB_2 amplifier has sufficient excitation to cause a grid current flow.

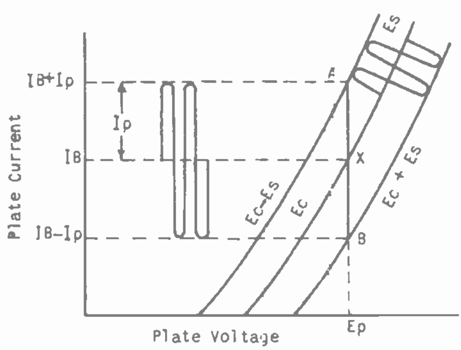
Class BC Amplifier: The Class BC amplifier has operating conditions which are intermediate to the Class B and the Class C amplifier. Plate current flows for less than 180°, but for a longer time than that usually found in Class C amplifiers. Grid-modulated amplifiers are usually operated under Class BC conditions.

2. THE CLASS A SINGLE-ENDED AMPLIFIER. All the amplifiers discussed in this lesson will be for the purpose of supplying a maximum amount of undistorted audio frequency power. The characteristics of voltage amplifiers have been thoroughly covered elsewhere. An elementary treatment of Class A power amplifiers was given in a Unit 1 lesson. At that time, the formula for determining the maximum power output was given. It was learned that maximum power output always results when the load impedance is equal to the plate resistance of the tube. The statement was also made that maximum undistorted power output was secured when the load impedance had a value equal to twice the plate resistance of the tube.

The term "maximum undistorted power output" is somewhat confusing. It might be thought that no distortion occurs when the load impedance is equal to twice the plate resistance. This, however, is untrue. It has been experimentally determined that when the total amount of distortion is 5% or less, it is unnoticeable to all except the musically trained ear. Thus, the term "maximum undistorted power output" has come to mean the amount of power that may be obtained without exceeding this arbitrary value of 5% distortion. Furthermore, plate load impedances of twice the plate resistance do not always produce exactly 5% distortion. Sometimes the distortion is more or less than this value. For this reason, the condition that the load impedance be equal to twice the value of the plate resistance should not be decreed as the criterion from which no deviation is allowable. Rather, this condition should be considered as an approximation, which represents the average ratio between load impedance and plate resistance for a large number of triodes.

There was also presented in a preceding lesson a formula from which the maximum undistorted power output could be calculated, if the amplification factor, grid-exciting voltage and plate resistance were known. Again, it should not be thought that this formula is infallible. It assumes that the grid voltage-plate current characteristic curves are straight parallel lines; a condition which, of course, is not strictly true. The actual amount of power output and distortion is not easily determinable by ordinary mathematical procedure, because the vacuum tube is not a linear device. Instead, a graphical analysis must be resorted to, if approximately correct values are to be obtained, and the major part of this lesson will be devoted to this graphical procedure.

Fig.1 Illustrating the operating path of a triode when there is no load impedance.



In Fig. 1 are illustrated three curves from a family of plate voltage-plate current characteristic curves. These curves are not representative of any particular tube, but might be those of nearly any triode. The three curves represent the relation existing between the plate voltage and plate current for three different values of grid voltage.

Let us assume that the amplifier stage using the tube which these curves represent has an applied plate voltage E_p and a fixed grid bias E_c. If E_c is the grid bias typified by the middle curve of the group and E_p is arbitrarily chosen along the plate voltage axis, then X becomes the operating point of the amplifier. It is that point at which the vertical line representing the plate voltage E_p intersects the characteristic curve corresponding to a bias voltage of E_c. With the plate voltage and grid bias values assumed, the plate current that will flow as a result of these voltages is at once determined. The no-signal plate current I_B is the vertical distance of point X above the horizontal axis.

If this amplifier contains no plate load resistance, the application of a signal voltage to the grid of the tube will not cause the plate voltage to change; there will be no voltage developed across the plate circuit, and all the power applied to the tube will be dissipated at the plate. Let us assume that the signal voltage applied to the grid has a value E_s. This voltage will add to and subtract from the normal grid bias E_c. As the grid becomes less negative, the voltage of the signal is subtracting from the grid bias, and, at its least negative point, the voltage on the grid will be $E_c - E_s$. Assuming that the left-hand characteristic curve corresponds to a grid voltage of $E_c - E_s$, it is evident that the point of operation is shifted from X to A. Since the plate voltage does not change during this interval, point A is found by extending the vertical line corresponding to the applied plate voltage until it crosses the left-hand characteristic curve.

On the succeeding alternation, the signal voltage adds to the grid bias and the grid is driven more negative. At its most negative point, the grid voltage has a value of $E_c + E_s$. If the right-hand characteristic curve corresponds to a grid voltage of this value, it is seen that the point of operation moves to point B. This point is at the intersection of the vertical line representing the applied plate voltage and the right-hand characteristic curve.

Thus, during one cycle of the exciting voltage, the operating point moves from point X to point A, to point B, and back to point X. The plate current that flows when the grid is least negative is determined by drawing a horizontal line from point A to the vertical or plate current axis. This amount of current is indicated by $I_B + I_p$; where I_B is the normal, no-signal plate current, and I_p is the increase in plate current above this value.

On the negative alternation, the minimum plate current is found by drawing a horizontal line from point B to the plate current axis. This plate current is designated by $I_B - I_p$; where these symbols are the same as in the preceding case. (It is assumed that the increase in plate current during the positive alternation is equal to the decrease during the negative alternation, or that the tube is working over the straight portion of its characteristic curve.) The grid-exciting voltage is represented by a sine wave drawn about the middle curve as an axis and extending from the left curve to the right curve. The pulsating plate current is represented by a sine wave drawn about I_B as an axis, and having a peak value equal to I_p.

This example shows how the variation in plate current due to a given grid-exciting voltage may be determined, if a family of plate voltage-plate current curves for the tube is available. In this particular case, where the plate voltage remains constant, it would be easier to use a grid voltage-plate current curve to determine the amount of plate current variation; however, as we shall now see, when the plate voltage varies (as it will when a load resistance is connected in the plate circuit), the family of plate voltage-plate current curves are more convenient for finding the variation of the plate current.

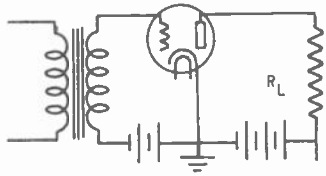


Fig.2 An A.F. amplifier stage with a load resistance.

Fig. 2 shows an amplifier stage with a load resistance connected in the plate circuit. The voltage of the power supply must be greater than the desired plate voltage because a part of this voltage is dropped across the load resistor, due to the no-signal plate current flowing through it. We shall assume that this is the same tube used in the preceding discussion, and that the same plate voltage, grid bias, and grid-exciting voltage are applied.

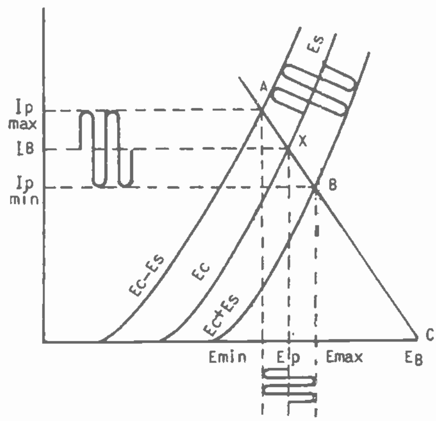


Fig.3 Illustrating the operating path of the tube shown in Fig. 2.

The curves for this tube are illustrated in Fig. 3. The operating point is at X, and the no-signal plate current is I_B. As the grid reaches its least negative point, the plate current increases to its maximum value. The increase in plate current causes a larger voltage to be dropped across the load resistor R_L, and the plate voltage is correspondingly reduced. Let us assume that the minimum plate voltage reached is represented by E_{min}, as shown on the plate voltage axis.

The amount of plate current that flows when the grid is least negative and the plate is at its lowest potential is found by drawing a vertical line from the voltage axis at the point marked E_{min} upward until it intersects the left-hand curve which represents the grid voltage at this time. This point is marked A, and the plate current is found by drawing a horizontal line from this point over to the current axis. Let us call this current I_{max}.

When the grid is driven more negative, the plate current decreases, and, as a result, the voltage dropped across the load resistor is reduced. This causes the voltage available for the plate of the tube to be greater. With the grid voltage most negative, the plate voltage reaches a maximum value, which shall be designated by the point on the voltage axis marked E_{max}. The plate current that flows when the grid is most negative and the plate voltage is maximum is determined by drawing a vertical line from the point E_{max} upward until it intersects the right-hand curve which represents the present grid voltage. This point is designated B, and the plate current at this point is called I_{min}.

Notice that the increase and decrease in the plate current is much less when a load resistor is used, even though the grid-exciting voltage remains the same. The pulsating plate current is represented by a sine wave drawn about I_B as an axis, and extending from I_{max} to I_{min}. The grid-exciting voltage is the sine wave drawn about E_c as an axis; it has the same value as before. The alternating component of the plate voltage is symbolized by the sine wave drawn from E_p to E_{max} and E_{min}.

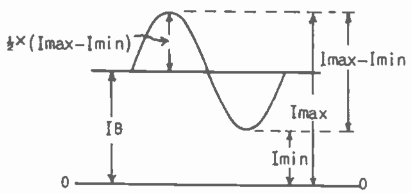
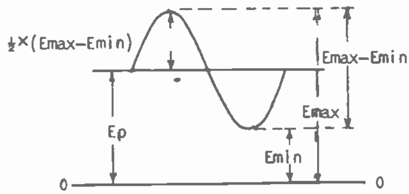


Fig.4 Waveform of the changing plate current of an A.F. amplifier.

The actual operating path of the tube is along the line AXB. The voltage needed from the power supply to produce a plate voltage of E_p when the no-signal plate current flows through the load resistor may be found as follows: As the plate current is reduced, less and less voltage is dropped across the load resistor and more is available for the plate of the tube. If the plate current were reduced to zero, there would be no voltage drop across the load and the actual plate voltage would be the total voltage output of the power supply. Therefore, extend the line AXB until it crosses the plate voltage axis; this occurs at point C. At this point, the plate current is zero, and the total voltage of the power supply is applied to the plate of the tube. This value of voltage E_B, therefore, represents the required voltage output of the power supply to produce a plate voltage of E_p when the normal, no-signal plate current is flowing.

Let us now determine how the power output of this tube would be calculated. The power output is equal to the R.M.S. current flowing through the load multiplied by the R.M.S. voltage across the load. Since it is only the AC power in the load that is of interest, only the AC components of current and voltage will be considered. The total variation of the plate current is from I_{max} to I_{min}. The peak value of the AC component of this current is the variation above or below the no-signal value. If $I_{max}-I_{min}$ represents the total variation of the current, then $\frac{1}{2} \times (I_{max}-I_{min})$ will be the peak value of the AC component. This is illustrated in Fig. 4.

Fig.5 Waveform of the changing plate voltage of an A.F. amplifier.



The changing plate voltage is illustrated in Fig. 5. It is seen that the plate voltage changes from E_{max} to E_{min}, or the total variation in the plate voltage is $E_{max}-E_{min}$. Let us assume that the power supply has a voltage of 1500 volts, and that, with no grid excitation, the applied plate voltage is 1000 volts, 500 volts being dropped across the load resistor. If, during excitation, the plate voltage varies from 800 to 1200 volts, the total variation is 400 volts. Let us now determine how the voltage dropped across the load varies with this change in plate voltage. The voltage across the load at any instant is equal to the voltage of the power supply minus the actual plate voltage at that instant. When the plate voltage fell to 800 volts, the voltage across the load must have been 1500 - 800, or 700 volts. Also, when the plate voltage increased to 1200 volts, the load voltage was 1500 - 1200, or 300 volts. Thus, it is seen that the total variation in the voltage across the load is from 700 to 300 volts, or a change of 400 volts.

From the foregoing, it is evident that the total change in the voltage across the load is also equal to $E_{max} - E_{min}$. The peak value of the alternating voltage across the load is equal to one-half of the total variation, or is $\frac{1}{2} \times (E_{max} - E_{min})$.

We now have the peak value of the voltage across the load and the peak value of the current through the load, which must be changed into R.M.S. values before the power can be calculated. To change peak values into R.M.S. values, the peak value is multiplied by .707. This number, however, is one-half of the square root of 2, or is $\sqrt{2} \div 2$. Therefore, the peak values may be converted into R.M.S. values as follows:

$$I_{rms} = \frac{\sqrt{2}}{2} \times \frac{(I_{max} - I_{min})}{2} = \sqrt{2} \times \frac{(I_{max} - I_{min})}{4}$$

$$E_{rms} = \frac{\sqrt{2}}{2} \times \frac{(E_{max} - E_{min})}{2} = \sqrt{2} \times \frac{(E_{max} - E_{min})}{4}$$

The AC power in the load is obviously the product of these two, or:

$$W = \sqrt{2} \times \frac{(I_{max} - I_{min})}{4} \times \sqrt{2} \times \frac{(E_{max} - E_{min})}{4}$$

$$= 2 \times \frac{(I_{max} - I_{min}) \times (E_{max} - E_{min})}{16}$$

$$= \frac{(I_{max} - I_{min}) \times (E_{max} - E_{min})}{8} \qquad (1)$$

Equation (1) may be used to find the power output of any triode operated as a Class A amplifier. It is always correct because its derivation does not depend on the linearity of the characteristic curves.

The total power furnished by the power supply divides between the load impedance and the plate of the tube. When the tube is not amplifying, a certain amount of DC power is dissipated in the load resistor (assuming that resistance coupling is used), and the remainder of the power is dissipated at the plate of the tube. When transformer coupling is used, the DC resistance of the primary winding is naturally very low, and practically all of the DC power supplied is dissipated within the tube.

With excitation, the plate current will be driven above and below its normal value and, if there is no distortion, the increases in plate current will be equal to the decreases. The instantaneous power taken from the power supply will vary from a maximum value, when the plate current is greatest, to a minimum, when the plate current is lowest. However, if the power supply is by-passed by a large condenser, this condenser will absorb the variations of the current and only the average current will actually be drawn from the supply. With no distortion, this average current is equal to the no-signal current, and it is, therefore, evident that the power supply furnishes the same amount of power whether the tube is amplifying a voltage or not. In a like manner, the DC power dissipated in the load resistor is the same with or without excitation, since the DC or average current through the load does not change with excitation. With excitation, however, there is developed in the load some AC power in addition to the normal DC power. Thus, the total power in the load is greater with excitation than without it, even though the average power input does not change. The only way that this may happen is for the average power dissipated in the tube during the excitation to be less than with no excitation. Under excitation, there is a redistribution of the total power furnished by the supply; more of this power is dissipated in the load resistor, and less goes to heat the plate of the tube. The decrease in average power lost in the tube will be exactly equal to the increase in power in the load resistor.

Let us now consider a typical example. Suppose that the voltage of the power supply is 1500 volts, and that the size of the load resistor is 5000 ohms. If the no-signal plate current is

100 ma., the DC voltage across the load is 500 volts, and the applied plate voltage is 1000 volts. The DC power dissipated in the load is:

$$\text{DC power in load} = .100 \times 500 = 50 \text{ watts.}$$

The DC power dissipated in the tube is:

$$\text{DC power in tube} = .100 \times 1000 = 100 \text{ watts.}$$

And: $$\text{Total power input} = .100 \times 1500 = 150 \text{ watts.}$$

Thus, with no excitation, it is evident that the sum of the DC power in the load and the DC power in the tube is equal to the total power furnished by the supply.

Let us assume that the plate current of this tube, during excitation, varies from 60 to 140 ma. This change in plate current causes the plate voltage to vary from 800 to 1200 volts. (This variation in plate voltage should be checked by the student.) The AC power output may now be found by formula (1).

$$\text{AC power output} = \frac{(.140 - .060) \times (1200 - 800)}{8}$$

$$= \frac{.08 \times 400}{8}$$

$$= 4 \text{ watts.}$$

This is, of course, the average value of the AC power developed in the load resistor. Since the DC power in the load is 50 watts, the total power in the load is now 54 watts. The total power furnished by the supply remains constant at 150 watts; therefore, the average power dissipated at the plate of the tube is now equal to the difference between the total power furnished and that used by the load. This is 150 - 54, or 96 watts. It is then clear that the average power consumed by the tube has reduced 4 watts to allow the power in the load to be increased by 4 watts.

There now results the somewhat amazing fact that a Class A amplifier tube runs cooler during excitation than it does when it is not amplifying a signal voltage! This must be true since the average power dissipated at the plate of the tube is less during excitation than without excitation.

3. LOAD LINES. When an amplifier circuit does not contain a load impedance, it is very easy to determine what plate current variation a given grid voltage swing will produce, because the plate voltage does not change. With a load impedance, however, the plate current variations cannot be found unless the plate voltage change is also known. The tube is said to be in a static condition, when it is not operating into any load impedance, and the plate current variations set up by a specified grid-exciting voltage may be determined by reference to a static grid voltage-plate current characteristic curve for the particular tube under consideration.

When worked into a load impedance, the tube is said to be in a dynamic condition, and dynamic characteristic curves must be used to determine its operating values. The difficulty lies in the fact that each different value of load impedance produces a different type of dynamic characteristic curve. A high value of load impedance causes the characteristic curve to be rather flat and to have a long straight portion. Smaller load impedances produce dynamic curves more nearly like those representing the static condition. Thus, it is evident that each different value of load impedance presents a special case.

It is manifestly impossible for the tube manufacturer to draw dynamic characteristics for each different load impedance which might be used, therefore, some other solution to the problem must be found. When a family of plate voltage-plate current curves are used, the operating path of the tube is over a straight line drawn at some angle to the plate voltage-plate current curves. This line is called the "load line" and it is placed at such an angle that the given grid-excitation voltage produces the correct amount of plate current variation for the particular load impedance used. The problem is to determine how this load line should be drawn.

The load line is a straight line, and is determined when two of its points are found. One point is already known; it is the operating point. Before it is possible to plot the load line, the applied plate voltage and grid bias which are to be used must be known. The other point of the load line must now be determined. To demonstrate the process of plotting a load line, a typical example will be given.

A type 10 tube is to be employed as a Class A audio power amplifier. Reference to the tube manual indicates that the maximum plate voltage that may be used for this class of service is 425 volts. Also, it is stated that the grid bias with this plate voltage should be -40 volts. This grid voltage, however, is that value measured from the grid to the mid-point of an AC operated filament. By definition, the actual grid bias is the voltage from the grid to the negative side of the filament, which would be the stated grid voltage less one-half of the peak filament voltage. For that reason, we will decrease the value of -40 volts by 5 volts (approximately one-half of the peak filament voltage), thus making the actual grid bias -35 volts. With these particular values of voltage, the manufacturer recommends a load impedance of 10,200 ohms.

The load line plotted on the characteristic curves of this tube is illustrated in Fig. 6. The procedure used to draw the load line is as follows: A plate voltage of 425 volts is not shown on the graph; however, it may be estimated very closely by taking a value half-way between 400 volts and 450 volts. From the point corresponding to a plate voltage of 425 volts, a vertical line is drawn. The operating point is the intersection of this vertical line and the characteristic curve representing a grid bias of -35 volts. Notice that this particular characteristic is not given, but it may be assumed to lie half-way between the -30 volt and the -40 volt curves. With the position of the -35 volt curve approximated, the operating point is found to be point X.

By glancing to the left, it is seen that point X corresponds to a plate current of 17 ma.

When there is no excitation, this 17 ma. of static plate current flows through the 10,200 ohm load impedance. The load impedance may be a resistor of this value, or, as is more probable, the stage will be transformer coupled, and the 10,200 ohm load impedance will be the resistance reflected from the secondary to the primary of the coupling transformer. If transformer coupling is used, there will be a negligible DC voltage loss in the

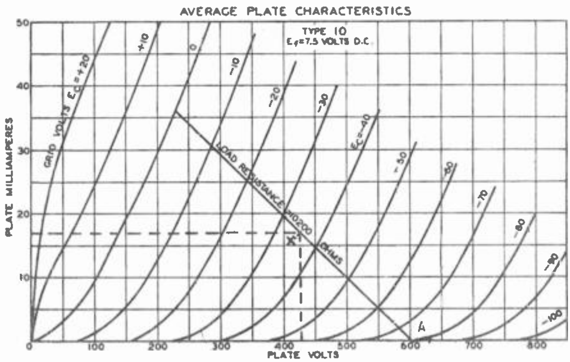


Fig.6 Illustrating the procedure for drawing a load line of a type 10 tube.

primary of the transformer, and the power supply will need to furnish only slightly more than 425 volts. A higher voltage power supply will be required if the stage is resistance coupled, for there will be a considerable DC voltage loss in the load resistance. Insofar as plotting the load line is concerned, the procedure followed is exactly the same no matter what type of coupling is used.

The 17 ma. of no-signal plate current in flowing through the 10,200 ohm load resistor produces a voltage drop of:

$$E = .017 \times 10,200 = 173.4$$

Therefore, the total voltage of the power supply must be the applied plate voltage plus the voltage lost across the load resistor, or:

$$425 + 173.4 = 598.4 \text{ volts.}$$

The total voltage of the power supply will be applied to the plate when there is no voltage drop across the load, or when the plate current is zero.

Again, it is simpler to use 600 volts rather than 598.4 volts. Thus, we shall assume that the plate voltage would be 600 volts when the plate current is zero, and we shall take point A as the second point of the load line. It should, of course, be realized that there will be practically no DC voltage drop across the load, if transformer coupling is employed, yet the load line will be correctly placed, if this assumption is made and the second point of the load line is found in this manner.

With two points of the load line determined, there now remains only the connecting of them with a straight line to produce the desired load line. This line may be extended as far as we wish toward the plate current axis.

The operating path of the tube has been determined, and the next step is to decide what grid voltage swing is to be allowed. Since the grid bias is -35 volts, a peak signal voltage greater than this value may not be used without driving the grid positive, causing grid current to flow, and producing excessive distortion. Therefore, let us limit the peak voltage of the signal to 35 volts. A signal voltage of this value would cause the grid voltage to vary from -35 volts to 0, to -70 volts, and back to -35 volts. We shall, therefore, use only that part of the load line extending from the zero grid voltage characteristic to the -70 volt curve.

The maximum plate current occurs when the grid voltage is zero; it is approximately 36 ma. The minimum plate voltage also occurs at this time; it has a value of about 230 volts. When the grid voltage is -70 volts, the plate current has a minimum value of about 2 ma., whereas the plate voltage at this time is maximum with a value of 580 volts.

The power output developed in the load resistor may be found by formula (1). It is:

$$\text{Power Output} = \frac{(.036 - .002) \times (580 - 230)}{8}$$

$$= \frac{.034 \times 350}{8}$$

$$= \frac{11.9}{8}$$

$$= 1.4875 \text{ watts.}$$

We may, therefore, state that under the given conditions the power output of a type 10 tube is approximately 1.5 watts.

Whether this is the best value of load impedance to use could be determined by plotting other load lines for load impedances having larger and smaller values. What is probably more important than the power output secured with different values of load impedances is the amount of distortion each produces.

As has been stated before, the maximum power will be secured from a tube when it is worked into a load equal to its plate resistance; however, this value of load ordinarily produces an objectionable amount of distortion. In a Unit 1 lesson, it was arbitrarily stated that maximum undistorted power output is obtained when the load impedance is twice the plate resistance of the tube. However, it is found that this is but an average value, and may vary considerably with different tubes. Some tubes may require a load impedance of nearly three times their plate resistance to reduce distortion to a point where it is not noticeable, whereas others may need a load impedance of slightly less than twice their plate resistance.

It is found that increasing the value of the load impedance always reduces the amount of distortion, but also decreases the power output somewhat. Let us see why this is so. In Fig. 7 is shown a family of E_p-I_p curves which may represent any triode. Three load lines are drawn as shown; one for a 1000-ohm load, one for a 2000-ohm load, and one for a 4000-ohm load. The grid bias is -15 volts, and the grid swing is from 0 to -30 volts. Let us first consider the 1000-ohm load line. As the grid voltage changes from 0 to -5 volts, the plate current changes 16 ma. (from 86 to 70 ma.). Also, as the grid voltage changes from -25 to -30 volts (a change of 5 volts), the plate current changes 8 ma. (from 16 to 8 ma.). It is, therefore, seen that the operation of the tube with this load impedance is far from linear. Equal changes

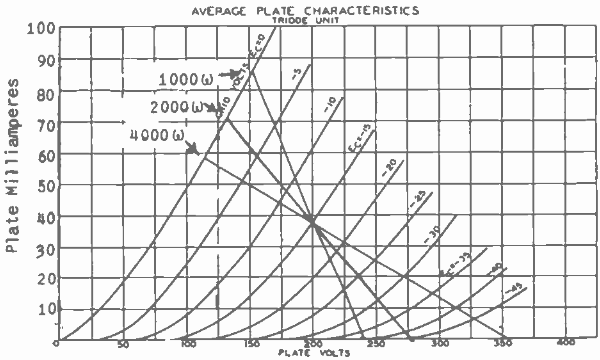


Fig.7 Illustrating the effect of changing the value of the load resistance.

in grid voltage do not produce equal changes in plate current throughout the entire operating range. In one instance a change of 5 volts in grid voltage caused a 16 ma. plate current change, and, at another part of the operating path, the same change in grid voltage produced only an 8 ma. change in plate current. Obviously the distortion produced with this load impedance would be very high, and extremely objectionable.

When the load is increased to 2000 ohms, the conditions are somewhat improved. Changing the grid voltage from 0 to -5 volts causes a change in plate current of 12 ma., whereas changing it from -25 to -30 causes a change of 8 ma. in the plate current. There is still considerable distortion, but the changes are not as unequal as before.

Then when the load is increased to 4000 ohms, the plate current changes are respectively 8 ma. and 6 ma.; and it is clear that the distortion is much reduced. Increasing the load impedance reduces the amount of plate current swing, and prevents the tube from being driven into the curve, non-linear portions of the characteristic curves. It is also clear that a larger load causes the plate voltage swing produced by a given grid swing to be larger. The power output depends upon the product of the plate current swing and the plate voltage swing, and is maximum when this product is greatest. This point occurs when the load impedance is equal to the plate resistance of the tube.

4. CALCULATING THE DISTORTION. After plotting the load line and calculating the power output, it is necessary to find the amount of distortion. A large part of the distortion is that due to the second harmonic, and a formula for determining the percentage of second harmonic distortion is as follows:

$$\text{Percent of second harmonic distortion} = \frac{\frac{1}{2} \times (I_{max} + I_{min}) - I_B}{I_{max} - I_{min}} \times 100 \qquad (2)$$

Where: I_{max} is the maximum value of the current,
I_{min} is the minimum value of the current,
I_B is the no-signal plate current.

Let us use this equation to calculate the percent of second harmonic distortion produced by the type 10 tube under the conditions previously set forth.

$$\%\ \text{distortion} = \frac{\frac{1}{2} \times (.036 + .002) - .017}{.036 - .002} \times 100$$

$$= \frac{.019 - .017}{.034} \times 100$$

$$= .058 \times 100 = 5.8\%$$

It is seen that the amount of distortion is slightly above the limiting value of 5%. Whether or not the signal amplified by this stage would contain excessive distortion would depend on the individual's reaction to it. To some, it would be acceptable, whereas others would consider the distortion objectionable. In this particular example, the distortion is produced by reducing the plate current to too low a value during the time that the grid is most negative.

By using a slightly larger value of load impedance, the dynamic characteristic would have a longer straight portion, and the distortion would be reduced. This would also cause the power output to decrease somewhat, since a larger load impedance would be even farther from the value at which maximum power is secured. The plate resistance of the tube under these operating conditions is 5000 ohms. Thus, a load impedance of 5000 ohms would produce maximum power output, but the distortion would be unacceptable. Let us draw a new load line using a load impedance of 12,000 ohms. The first point of the load line is the operating point of 425 volts plate voltage, 17 ma. plate current, and -35 volts grid bias. When the 17 ma. of plate current flows through the 12,000 ohm load impedance, a voltage drop of .017 × 12,000, or 204 volts occurs. By adding the 425 volts plate voltage to the drop of 204 volts, a maximum plate voltage of 629 volts results. This is the plate voltage that would be applied, if the plate current were zero, and there was no voltage drop across the load. The second point of the load line is the point on the graph representing 629 volts plate voltage and zero plate current. The two points are connected by a straight line and the load line shown in Fig. 8 is produced.

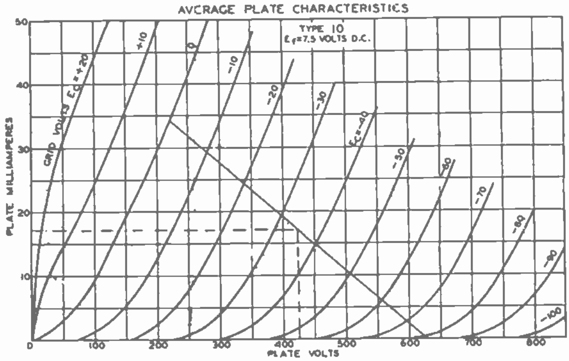


Fig.8 A load line representing a load impedance of 12,000 ohms for a type 10 triode.

If this load line is carefully compared with the one drawn for a load impedance of 10,200 ohms, it is found that it is less vertical and is more nearly horizontal. The greater the load impedance, the more nearly flat the load line will be. A low load impedance, on the other hand, will produce a more nearly vertical load line.

Using the same grid voltage swing, we shall calculate the power output and the percent of distortion for this new value of load impedance. When the grid voltage is zero, the plate current is approximately 34 ma. and the plate voltage is about 220 volts. With a grid voltage of -70 volts, the plate current is 3 ma. and the plate voltage is 590 volts. The power output is therefore:

$$\text{Power output} = \frac{(.034 - .003) \times (590 - 220)}{8}$$

$$= \frac{.031 \times 370}{8}$$

$$= \frac{11.47}{8} = 1.43 \text{ watts.}$$

The percent of harmonic distortion is:

$$\%\ \text{distortion} = \frac{\frac{1}{2} \times (.034 + .003) - .017}{.034 - .003} \times 100$$

$$= \frac{.0185 - .017}{.031} \times 100$$

$$= \frac{.0015}{.031} \times 100 = 4.84\%$$

Although the power output has been reduced slightly, the percent of distortion is now below the limiting value of 5%. The distortion found in this manner is only that due to the second harmonic component. If there are other higher harmonic frequencies contained in the distorted plate current waveform, they will also produce some distortion. Most of the distortion, however, is that caused by the second harmonic frequency; the percent of distortion due to higher harmonics in triodes is negligible. If the percent of second harmonic distortion as found by equation (2) is 5% or less, it is safe to assume that the total distortion will not be noticeable.

The second harmonic frequency is caused by the unequal amplitudes of the positive and the negative alternations of the plate current. If the ratio of the amplitude of the positive alternation to that of the negative alternation is one, there will be no distortion. With a second harmonic distortion of 5%, the ratio of the amplitude of the positive alternation to that of the negative alternation is 11:9. Any ratio greater than this will cause more than 5% distortion.

The procedure to be observed in plotting a load line and determining the power output and distortion is summarized as follows:

1. Obtain a family of plate voltage-plate current curves of the tube to be used.
2. From the manufacturer's ratings, decide what plate voltage and grid bias will be applied.
3. Select a value of load impedance at least twice as great as the plate resistance of the tube.
4. Multiply the no-signal plate current by the value of the load impedance.
5. To the result of step 4, add the applied plate voltage. The sum represents the plate voltage that would be applied to the tube, if the plate current were zero.
6. Plot this value of plate voltage as found in step 5 on the plate voltage axis. It is the second point on the load line.
7. Draw the load line, and determine what grid voltage swing may be allowed. The peak signal voltage cannot be greater than the grid bias without causing the grid to be driven positive.
8. From the graph, determine the maximum and minimum values of plate current and plate voltage produced by the selected signal voltage.
9. Using formula (1), calculate the power output.
10. Find the amount of distortion by formula (2).

If the distortion is excessive, two different procedures may be followed. Using the same load impedance, decrease the value of the signal voltage and recalculate the power output and the distortion. This will reduce the distortion, but may not produce the desired power output. In this case, it will be necessary to draw a new load line which represents a larger value of load impedance. Increasing the load impedance will always reduce the distortion. It should be remembered that maximum power cannot be obtained from the tube unless the applied plate voltage is the maximum recommended by the manufacturer.

5. PUSH-PULL CLASS A AMPLIFIERS. A push-pull amplifier is capable of delivering more than twice the amount of power available from a single-ended amplifier with considerably less distortion. Even under the best of conditions, there will always be a certain amount of unavoidable distortion in a single-ended amplifier, because of the curvature of the characteristic curves.

When the size of the load impedance is the variable factor, maximum power will always be obtained when the load is made equal to the internal resistance of the power source. This is not possible with one-tube amplifiers, because of the excessive distortion which results. On the other hand, this condition for maximum power is easily obtained with the push-pull amplifier, since the characteristics of this type of amplifier are such that the greater part of the distortion is eliminated.

Furthermore, it should not be thought that making the load impedance equal to the internal resistance of the voltage source always gives the maximum power output. This condition produces the maximum power only for this particular power source, and is true only when the internal resistance of the voltage source is fixed and the load impedance is variable. Suppose that the value of load impedance is fixed, whereas the internal resistance of the source is variable. In this case, maximum power will be secured, not when the resistance of the source is made equal to the fixed value of load impedance, but will occur when the resistance of the power source is made as low as possible. For example, a magnetic speaker with a given value of resistance is to be fed by a power amplifier stage. We are at liberty to choose any power tube we wish. It might be thought that maximum power would be delivered to the speaker, if a power tube whose plate resistance was equal to the resistance of the speaker was selected. This, however, is an erroneous conception. The speaker will receive maximum power when the plate resistance of the power tube, with which it is used, is as low as possible. Thus a speaker of 1600 ohms resistance would not receive as much power from a type 45 tube, whose plate resistance is about 1600 ohms, as it would from a type 2A3 tube, whose plate resistance is 800 ohms.

If it so happens that the resistance of both the power source and the load are variable, maximum power will be transferred to the load by making the source impedance as low as possible, and then using a load impedance equal to the resistance of the source.

Viewing a push-pull amplifier as a voltage source, we find that the apparent internal resistance of this source is not equal to the plate resistance of one of the tubes. Instead, it has a resistance of about half the plate resistance. Since the apparent internal resistance is lowered by connecting the tubes in push-pull, it is evident that the power output will increase accordingly.

There are, therefore, two reasons why two tubes connected in push-pull will deliver more than twice as much power as will a single tube of the same type. First the internal resistance of the push-pull amplifier is lower than that of a single tube. Second, it is possible to use a value of load impedance which is equal to the internal resistance of the push-pull combination without causing excessive distortion.

When the two tubes of a push-pull amplifier are correctly matched, and they should be for the best results, the plate currents drawn by the tubes are equal. These two currents flow through the primary of the output transformer in opposite directions, and the field created by one current is equal and opposite to that produced by the other. Thus, with no signal voltage applied, there is no net magnetomotive force available to magnetize the core. When the grids of the tubes are excited, this balance is upset. During one alternation of the signal voltage, the plate current of one tube increases and that of the other decreases. The net magnetizing force, which induces a voltage into the secondary, is at every instant equal to the difference between the two plate currents. The same voltage would be induced into the secondary by a single plate current, which, at every instant, had a value equal to the difference between the plate currents of the two tubes at that instant.

It is thus apparent that it is not the actual values of the two plate currents, but rather their difference, which is of primary interest. The signal voltages on the grids of the two tubes are 180° out of phase. As one grid is growing more negative, the other is becoming less negative. The variation of the plate current of each tube, as well as the difference between the plate currents at any instant, produced by a given grid voltage swing may be shown by using two dynamic grid voltage-plate current characteristic curves. Such an arrangement is illustrated in Fig. 9. The characteristic of tube 2 is inverted and placed below that of tube 1. This is necessary because of the 180° phase difference of the grid voltages. The grid voltage axis of the bottom tube is super-imposed on the grid voltage axis of the top tube. Notice, however, that the point representing zero grid voltage for the bottom tube is at the left of the diagram, and that the plate current scale of this tube is read from top to bottom.

We have assumed that the grid bias of the two tubes is -30 volts; therefore, in constructing this double curve, the point representing -30 volts grid voltage for the bottom tube was placed directly over the -30 volt point of the top tube. The no-signal plate current of each tube is 30 ma. This is found by drawing a vertical line through the -30 volt point and extending it in either direction until it intersects the two characteristic curves. At each point of intersection, horizontal lines are drawn which cross the plate current axes at the 30 ma. points.

The applied signal voltage has a peak value of 20 volts, which causes the grid voltage of each tube to vary from -10 to -50 volts. When the grid voltage of the top tube is -10 volts, that of the bottom tube is -50 volts, and vice versa. To show the excitation voltage in a graphical manner, a vertical line is drawn through the point representing a grid voltage of -50 for the top

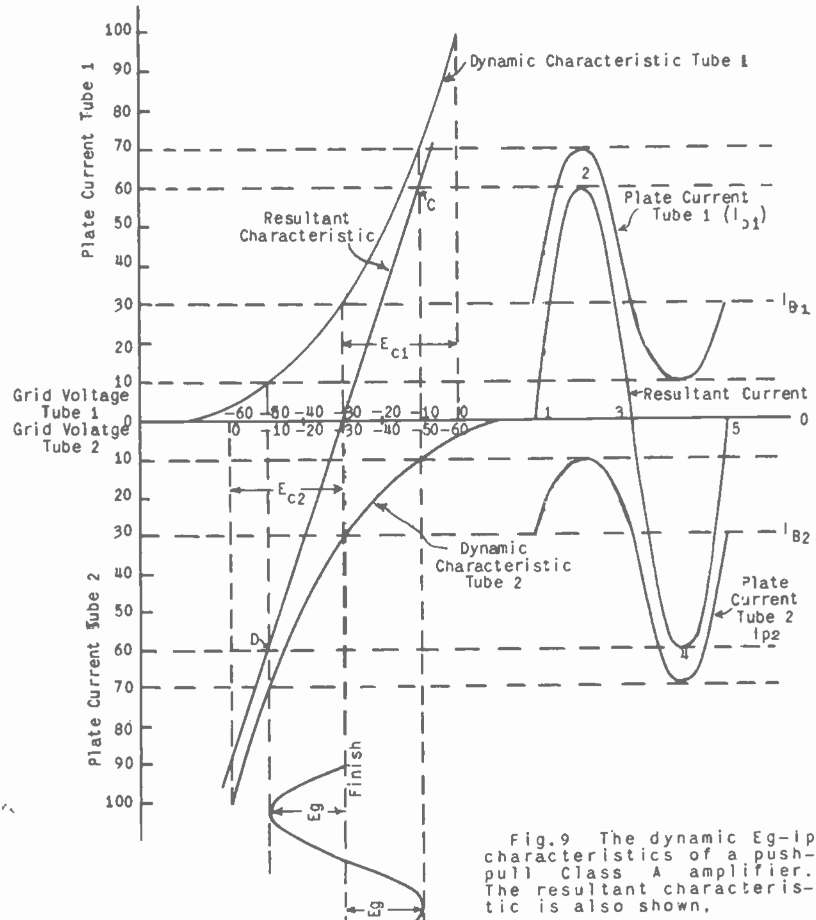


Fig.9 The dynamic Eg-Ip characteristics of a push-pull Class A amplifier. The resultant characteristic is also shown.

tube and -10 for the bottom one. This line represents one extreme of the grid voltage variation. The other extreme is determined by drawing another vertical line through the -10, -50 grid voltage point. Both of these lines are extended until they cross the two characteristic curves.

The actual excitation voltage is shown as a sine wave, drawn within the limits of these vertical lines, and having the -30 volt line as an axis. During the first alternation of the signal voltage, the grid of tube 1 is made less negative, whereas that of tube 2 becomes more negative; thus, the plate current of the top tube increases, and that of the bottom tube decreases. At the peak of this alternation, the grid voltage of the top tube is -10 volts, and that of the bottom tube -50 volts. The plate current drawn by each tube at this time is determined by noting the points where the vertical line representing these values of grid voltage crosses the two characteristic curves, and by drawing a horizontal line through each of these points. These two horizontal lines cross the plate current axis at 70 and 10 ma. respectively. Thus, the plate current of the top tube is 70 ma., and that of the bottom tube 10 ma.

During the next alternation, these conditions are reversed; the plate current of the top tube decreases and that of the bottom tube increases. At the peak of this alternation, the grid voltages of the top and bottom tubes are -50 and -10 volts respectively. These grid voltages cause the top tube to draw 10 ma. of plate current and the bottom tube 70 ma. The plate current of tube 1 varies from 30 ma. (its no-signal value), to a peak value of 70 ma., and down to a minimum value of 10 ma. Therefore, the waveform of this current may be represented by the distorted sine wave marked I_{p_1}. In a similar manner, the current variations of tube 2 may be shown by the distorted sine wave, I_{p_2}.

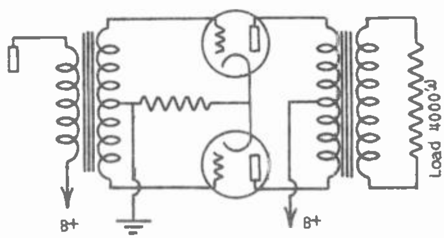
Considering only the current drawn by tube 1, notice how much greater the amplitude of the positive alternation is than that of the negative alternation. If the plate current of a single-ended amplifier were allowed to behave in this manner, the distortion produced would be intolerable. In fact, this curve represents a second harmonic distortion content of about 16%. The plate current of tube 2 also possesses this same amount of distortion, and is exactly the same as that of tube 1, except for the fact that it is 180° out of phase with it.

Taken at face value, this information would seem to indicate that a very distorted voltage would be induced into the secondary of the output transformer. However, when it is realized that it is the difference between these two currents that determines the net magnetizing force available in the core of the transformer, it is at once clear that the secondary voltage will be practically undistorted. The waveform representing the resultant current or difference current could be found by taking various points along each of these distorted waves and subtracting one current value from the other. A simpler method is to first find the resultant or difference characteristic curve. At -30 volts grid voltage, both tubes draw 30 ma. of plate current; the difference is zero, and thus, one point of the resultant characteristic is at -30 volts grid voltage and zero plate current. When tube 1 draws 70 ma., tube 2 is drawing 10 ma.; therefore, another point on the resultant characteristic is at point C. With tube 1 drawing 10 ma., tube 2 draws 70 ma.; and point D becomes a third point of the resultant characteristic.

It will be noticed that these three points lie practically in a straight line, and the error introduced by assuming that the resultant characteristic is the straight line joining these three points is negligible. The nearly straight line characteristic of the resultant curve is the determining factor which causes the secondary voltage to be practically distortionless. Unless the grid bias and load resistance are carefully chosen, the resultant characteristic will not be linear; and although there will not be any second harmonic distortion present in the output, the distortion due to the third harmonic content will be appreciable.

With the resultant characteristic drawn, we are now ready to plot the waveform of the difference current which is effective in producing a secondary voltage. At -30 volts grid voltage, the resultant current is zero. This is indicated by point 1 of the waveform. At the peak of the first alternation, the resultant current is 60 ma., and is represented by point 2. The grid voltage of both tubes now returns to the no-signal value and the resultant characteristic falls to zero (point 3). At the peak of the next alternation, the resultant current is again 60 ma.; however, the current of tube 2 is now greater than that of tube 1, and the difference current must be regarded as flowing in the opposite direction. Therefore, point 4 represents the resultant current at this time. Finally, both grids return to their normal value, and the resultant current decreases to zero (point 5).

Fig.10 A Class A, push-pull amplifier stage.



It is seen that the resultant current has nearly a pure sine waveform, although the current of either tube considered separately is very distorted. Even though the resultant characteristic were not a straight line, the amplitudes of the positive and negative alternations of the resultant current wave would be equal. As long as this condition is fulfilled, the resultant wave can have no second harmonic component. A resultant characteristic which departed from a straight line would affect both alternations of the resultant waveform equally. Both peaks would either be slightly flattened, or both sharpened. In either case, the resultant waveform would have a third harmonic component.

This resultant current will produce the same voltage in the secondary of the transformer as would an alternating current having a peak value of 60 ma., if it were allowed to flow through one-half of the primary winding.

It is now time that we gave some attention to the load impedance. This impedance is rarely, if ever, a resistance. It is common practice to use an output transformer to couple the push-pull tubes to their load. Such an arrangement is shown in Fig. 10. For simplicity, let us assume that the turns ratio of this transformer is 1:1 or that the total number of turns on the secondary is equal to the total number of turns on the primary. It may be proved that a push-pull amplifier is equivalent to a single tube having a plate resistance equal to about one-half the plate resistance of either of the push-pull tubes, and working into a load equivalent to that looking into one-half of the primary winding. That is, in calculating the power output, it is considered that both of the push-pull tubes work into only one-half of the primary winding. The device which absorbs the power in the circuit shown in Fig. 10 is the resistor connected across the secondary of the output transformer. If this resistor has a value of 4000 ohms, the resistance reflected across the entire primary is also 4000 ohms, since the turn ratio is 1:1. Considering that the tubes work into only one-half of the primary, we see that the actual load on the tubes is less than 4000 ohms. Superficially, it would seem that the resistance reflected across one-half of the primary would be 2000 ohms. We are, however, neglecting some of the principles of reflected impedance.

Suppose that both primary and secondary windings have 40,000 turns. Then, one-half of the primary winding would consist of 20,000 turns. The turns ratio between one-half of the primary and the entire secondary is 1 ÷ 2. The resistance reflected across one-half of the primary is equal to the resistance connected across the secondary multiplied by the square of the turns ratio from one-half of the primary to the secondary. The square of this turns ratio is 1:4. Thus, the resistance looking into one-half of the primary is $\frac{1}{4}$ of 4000 ohms, or 1000 ohms. It should, therefore, be remembered that the actual impedance measured across one-half of the primary is only one-fourth of the impedance across the entire primary. This point must be taken into account in selecting the proper output transformer.

It is now necessary to learn how the power output, and the proper load for a push-pull amplifier is calculated. Since this type of amplifier cancels all even harmonics if it is balanced, it is not necessary to work it into a load equal to twice the internal impedance of the push-pull stage. It is well known that for maximum power output a stage should be worked into a load equal to the internal impedance of the stage; and it might be thought that this would be the optimum value to use for the push-pull stage. It is found, however, that this value of impedance is ordinarily too low for the best results. It is too low, not because of distortion, but because the peak current drawn by the tubes is usually excessive. The smaller the value of the load impedance, the greater the plate current change will be, and consequently the higher the peak plate current will become. With a load equal to the internal impedance, the peak plate current may exceed the emission limit of the filament, and also may produce excessive plate dissipation. Therefore, the proper load impedance to use and the power output which may be expected is found by drawing a load line using the method outlined in the following example.

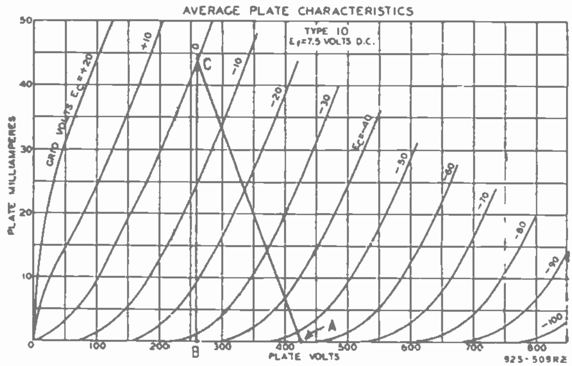


Fig.11 Illustrating the method of constructing a load line for a push-pull stage.

Step 1. From the tube manual determine the maximum plate voltage that may be applied for Class A service. Mark this point on the plate voltage axis. It is point A in Fig. 11. This voltage is known as E_o. In this example, E_o is 425 volts.

Step 2. Multiply E_o by .6, and find this point on the plate voltage axis. In this case, $.6 \times E_o$ is 255 volts and is point B in Fig. 11.

Step 3. Erect a vertical line from point B, extending it until it crosses the zero grid bias characteristic ($E_c = 0$). Extend the zero grid bias characteristic if necessary. This may be done by assuming that the characteristic is a straight line; although in this particular example it is unnecessary to extend this characteristic. In this case the vertical line crosses the zero characteristic at point C.

Step 4. Connect points C and A. The line so formed is the load line of the push-pull stage. Point C is the maximum plate current (I_m), and in this example it is 42 ma.

Step 5. Calculate the power output, using the following formula:

$$P_o = \frac{I_m \times E_o}{5}$$

Solution: $$P_o = \frac{.042 \times 425}{5} = \frac{17.85}{5} = 3.57 \text{ watts.}$$

Step 6. Calculate the plate-to-plate load impedance, using the formula:

$$\text{Plate-to-plate load} = \frac{E_o - (.6 \times E_o)}{I_m} \times 4$$

In this case:

$$\text{Plate-to-plate load} = \frac{425 - 255}{.042} \times 4 = 4046 \times 4 = 16{,}184 \text{ ohms.}$$

The value of load impedance determined by the formula given in step 6 is the smallest that should be employed to insure that the tubes do not heat excessively during the time that the peak current is flowing.

Step 7. Multiply E_o by 1.4, and find this point on the plate voltage axis. In this example: $1.4 \times E_o = 595$ volts. From the plate family, determine how much bias voltage would be necessary to produce plate current cut-off at a voltage equal to $1.4 \times E_o$. It is seen that this is approximately -78 volts. The operating bias voltage should be one-half of this value, or -39 volts. Any greater bias would not allow the plate current to flow throughout the entire grid-exciting cycle, and the stage would not be operating as a Class A amplifier.

6. CLASS B AUDIO AMPLIFIERS. The power output that can be obtained from a Class A amplifier is relatively low. Also, the fact that the no-signal plate current of a Class A amplifier, whether single-ended or push-pull, is fairly high, causes the maximum allowable plate dissipation to be reached before much power output can be obtained. All this is equivalent to saying that a Class A amplifier has low efficiency. When considerable power is desired, the cost of the large tubes and the high-voltage power supply required, mounts rapidly for each additional watt of power output.

To remedy this situation, the Class B audio amplifier was invented. The first step in increasing the efficiency of an amplifier is to reduce the power loss at the plate without excitation. To accomplish this, the no-signal plate current must be reduced by increasing the grid bias. So far as single-ended Class A amplifiers are concerned, it is very necessary that the grid bias be half-way between zero grid voltage and the point where the characteristic starts to curve near its bottom end. This is essential in order that the increases in plate current be equal to the decreases, and distortion thereby be avoided.

The bias of a push-pull amplifier ordinarily may be somewhat greater than that of a single-ended stage since the distortion produced is canceled by the push-pull arrangement.

Since the push-pull type of connection eliminates the original requirement that the increases in plate current be equal to the decreases, it is possible to go even farther and make the grid bias so large that the no-signal current is practically zero. For example, let us assume that the no-signal plate current is actually zero. When an excitation voltage is applied, the tube whose grid is becoming less negative draws a plate current, whereas the other tube which is already biased to cut-off does not have any plate current flow, because its grid is being driven even more negative. Thus, each tube draws plate current for one-half cycle, or 180°.

Assuming that the signal voltage is of sine waveform, it is evident that one of the push-pull tubes furnishes one of the alternations of plate current, and the other tube produces the other alternation. It is not usual to bias the tubes to complete plate current cut-off, because of the excessive third harmonic distortion which would be created. Instead, the applied grid bias is ordinarily that value obtained by dividing the applied plate voltage by the amplification factor. This amount of bias is known as the theoretical cut-off value, and is that point where the E_g-I_p characteristic would intersect the grid voltage axis, if the characteristic were a straight line.

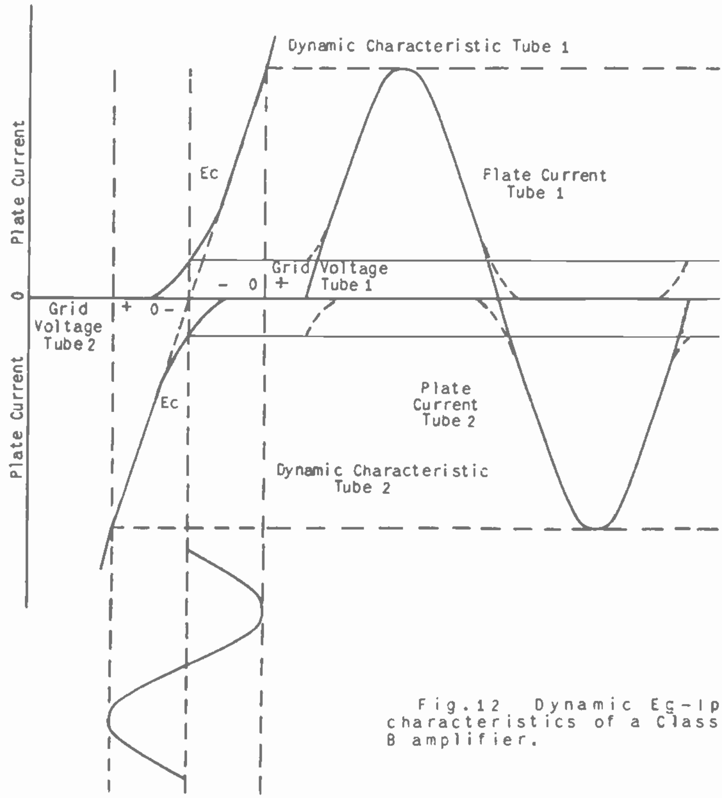


Fig. 12 Dynamic Eg-Ip characteristics of a Class B amplifier.

Reference to Fig. 12 will, perhaps, make this somewhat clearer. This figure shows two dynamic E_g-I_p characteristic curves with that of the bottom tube reversed. Notice that the tubes are so biased that the operating point is at the beginning of the extremely curved portion of the characteristics. The resultant or composite characteristic is the curve which connects the upper characteristic with the lower one. It coincides with the characteristics of the two tubes, except at low values of plate current where it is shown as a dotted line. To prevent distortion, the separate characteristics of the tubes must have fairly long straight portions, because cancellation of the harmonics produced by non-linearity occurs only at low values of plate current when neither tube is cut off. Practically all of the positive alternation of the output wave is furnished by the top tube. When the plate current of the top tube is near its no-signal value, a slight amount of deviation from a pure sine wave occurs. This, however, is cancelled in the output wave because the bottom tube is drawing a very small amount of current at this time, and the output wave which is the difference between the two currents is practically distortionless.

On the negative alternation of the signal voltage, the plate current of the top tube is cut off except for a short time at the beginning and end of this alternation, and the major part of this alternation of the output wave is supplied by the plate current flow of the bottom tube. Since each tube draws plate current for approximately 180°, this arrangement is properly called a Class B amplifier.

In a Class A push-pull amplifier, the total current drawn from the power supply, or the sum of the two plate currents, remains constant, because as one plate current increases, the other plate current decreases an equal amount. This makes a by-pass condenser for the biasing resistor unnecessary. The power supply of a Class B amplifier, on the other hand, furnishes a very small current while no excitation is applied, but needs to supply a peak current which may be many times the no-signal value. If the input voltage is a pure sine wave, meters connected to measure the plate current of each of the Class B tubes will indicate a steady value which is approximately .318 times the peak current of either tube. To be useful, however, an amplifier must amplify speech and music, the waveform of which is very irregular and may have peaks which are many times the average value. With such an excitation voltage, the meters in the plate circuits vary constantly from instant to instant as the speaker's voice rises and falls.

Since the average value of the plate currents and the total current drawn from the power supply varies in an irregular fashion, it is not possible to use cathode resistor bias. Instead, some form of fixed bias must be employed, and means must be available for adjusting the bias of each tube separately, because the bias voltage is quite critical.

Any decrease in the applied plate voltage due to poor regulation of the power supply will cause the peaks of the alternations to be flattened, with the consequent production of third harmonic distortion. Thus, for best operation, the power supply should have mercury-vapor rectifier tubes, and the power supply filter should have a swinging choke. (A complete discussion of the type of power supply to be used with Class B stages and R.F. transmitter stages will be given in a later lesson of this unit.

By eliminating the necessity that the plate current of both tubes flow throughout the entire grid-exciting cycle, the no-signal plate current is made very low, and a very large increase in the efficiency occurs. Although the efficiency may be rather high, it is found that appreciable power cannot be obtained from the Class B amplifier unless the grid-exciting voltage is sufficient to drive both grids positive throughout a part of the grid-excitation cycle. It is at once thought that such a procedure would produce distortion, and, in fact, it does; however, by careful design of the grid circuit, it is possible to reduce this distortion to a value where it is not excessive.

When the grids become positive, grid current flows, and the resulting voltage drop produced across the grid circuit bucks against the signal voltage. Thus, the peak voltages on the grids are less than they otherwise would be, and the peaks of the output waveform are flattened. The first step in the minimizing of the distortion due to this source is to make the actual resistance of the secondary winding of the input transformer as low as possible.

The grid current must flow through the secondary of the input transformer and also through the grid-to-filament resistance of the tubes, and, in so doing, it dissipates power which must be supplied by the preceding stage. Therefore, the stage preceding a Class B amplifier must be a power amplifier rather than a voltage amplifier. It is ordinarily a Class A power amplifier, either single-ended or push-pull, and is known as the "driver stage". The driver tube must have a power rating adequate for the grid requirements, and indeed, should have more than this rating, because a reserve supply of power is necessary, if the distortion in the grid circuit is to be kept within allowable limits.

The load impedance of the driver stage is rather critical for the best operating conditions. Its value is, of course, determined by the turns ratio of the input transformer, and the resistance connected across the secondary of this transformer and therein lies the difficulty. It is simple enough to use a transformer of the correct turns ratio, but the trouble is that the resistance connected across the secondary of this input transformer is not constant in value. In ordinary Class A amplifiers, where the grids do not go positive and no grid current flows, the resistance connected across the secondary of an input transformer is very high, practically infinite. The actual resistance connected across the secondary depends, of course, on the voltage developed across the secondary divided by the current that flows through the secondary winding.

The grids of a Class B amplifier do draw grid current, and as a result, the resistance connected across the secondary of the input transformer consists of the grid-to-filament resistance of the Class B tubes. The grid voltage-grid current curve is not a straight line, but is rather sharply curved at low values of positive grid voltage. This indicates that the relation between the grid voltage and the grid current is not linear, and consequently the grid-to-filament resistance is not constant.

Fig. 13 illustrates a grid voltage-grid current curve of an average triode used in a Class B amplifier. Notice how far this curve departs from a straight line. The grid-to-filament resistance could be found at any point by taking a small change in grid voltage and dividing that change in voltage by the change produced in the grid current. When the grid is highly positive, the curve is very steep and the actual grid-to-filament resistance for a value of approximately 60 volts positive is about 300 ohms. When the grid current is low, the curve is rather flat, and the grid-to-filament resistance is about 2000 ohms at low values of grid voltage. Furthermore, during the time that both grids are negative, no grid current flows, and the grid-to-filament resistance is practically infinite.

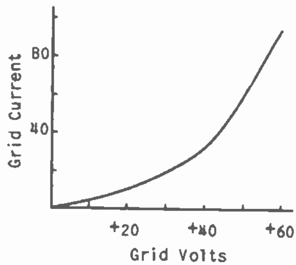


Fig.13 A dynamic grid voltage-grid current curve of an average triode.

All this means that the resistance connected across the secondary of the input transformer may vary from a value as low as 300 ohms to an extremely high value. The load impedance into which the driver tube works is the resistance reflected from the secondary to the primary of the input transformer, and, since the resistance connected across the secondary of this transformer varies through wide limits, it is obvious that the load impedance on the driver tube will also vary considerably. It is well known that the load impedance of the driver must not fall below a value equal to twice the plate impedance of the driver tube, if excessive distortion is to be avoided. Therefore, the load reflected to the driver must not fall below this value, even when the grid impedance of the Class B stage is at its lowest value.

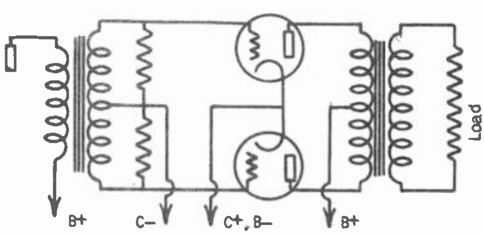


Fig.14 A Class B, audio amplifier with stabilizing resistors.

The correct load impedance of the driver will be at least several thousand ohms, whereas the grid impedance of the Class B stage may fall to a few hundred ohms at its lowest value. It is thus evident that the input transformer must have a step-down ratio from primary to secondary, if the correct load impedance is to be reflected. Although the load resistance of the driver will be the correct value when the grid impedance of the Class B stage is lowest, it will be too high during the time that no grid current is flowing, and the grid impedance is practically infinite. To prevent such wide variations of the driver's load impedance, a resistor is sometimes connected across each half of the secondary winding as shown in Fig. 14. This helps to fix the secondary resistance, because the resistance connected across the secondary cannot rise to a value greater than the value of this resistor. Naturally, this resistance cannot be too low in value, for, in that case, it would draw too much current from the transformer secondary, and cause an excessive voltage drop in the secondary winding; the average value is 50,000 ohms from grid to grid. It does, however, prevent the load on the driver from varying too much, and, therefore, tends to make the driver stage more stable.

It should be realized that only one of the Class B tubes will be drawing grid current at any instant, and, therefore, only one-half of the secondary of the input transformer will be in use at any one time. Thus, when the grid of the top tube is positive and is drawing grid current, the load on the driver tube depends on the effective resistance connected across the top half of the secondary, and the turns ratio between the entire primary and one-half of the secondary.

The necessity of using a step-down input transformer may be viewed in another manner. If the signal voltage is of sine waveform, the voltage across the entire secondary should exactly follow a sine wave in its variations, whether any grid current is flowing or not. This signal voltage, however, must be furnished by a voltage source, which, in this case, is the driver tube, and as we know, all voltage sources contain some internal resistance which causes their voltage regulation to be faulty; that is, the voltage furnished by the source falls when the current is drawn from the source. The internal impedance of the voltage source is the plate resistance of the driver tube, and, since this value is rather high, it is seen that rather poor voltage regulation could be expected. By using a step-down transformer, however, the fairly high plate resistance of the driver tube is transformed into a much lower impedance which appears across the secondary of the input transformer. Thus, the voltages across this secondary appear to be emanating from a voltage source having a much lower internal impedance, and, as a result, the voltage regulation of the grid-exciting voltage is greatly improved. This tends to keep the exciting voltage following a true sine wave, even when the grid current is rather high.

If the step-down ratio is too great, the grid-exciting voltage will not be large enough to properly drive the Class B tubes. The turns ratio of the transformer should, however, step down the voltage until it is just sufficient to give the proper excitation for the power output required.

The fact that self-bias is not practical for Class B tubes retarded the development of this type of amplification for some time. In the past few years, several tubes have been introduced with such a high amplification factor that the plate current is very low, even when the grid voltage is zero. Two such tubes may be employed in a Class B amplifier stage and operated with zero grid bias. It is true that the no-signal plate current is not zero with the so-called zero-bias tubes, but it does have a very low value. It should also be evident that when operated with zero bias, there will be some grid current flowing at all times; that is, the grid of one tube will draw current for one-half cycle when the excitation voltage is such as to make this grid positive, and, as soon as this tube ceases to draw grid current, the other tube will begin to do so. This condition tends to maintain the grid impedance of the Class B stage more nearly constant, because at no time does the grid impedance rise to practically an infinite value. Thus, the load impedance of the driver varies through narrower limits and grid circuit distortion is minimized.

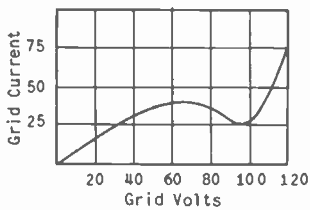


Fig.15 A dynamic Eg-Ig characteristic exhibiting dynatronic qualities.

Grid voltage-grid current characteristics vary considerably, even with tubes of the same type. Some tubes, especially types 203A, 211, 204A, and 849, have a very peculiar grid voltage-grid current curve. Fig. 15 illustrates the type of characteristic representative of these tubes. Increasing the grid voltage from 0 to approximately 50 volts causes an increase in the grid current, but a further increase from 50 to about 95 volts produces a decrease in the grid current. This dip in the curve results from the secondary emission of electrons from the grid. The electrons knocked out of the grid wires join the main electron stream and become part of the plate current. From 50 to 95 volts positive, the grid impedance is negative, and the grid circuit is very liable to act as a dynatron oscillator. The frequency of the oscillations produced are determined by the total inductance and capacity associated with the grid circuit. This includes the inductance of the input secondary winding and the distributed capacity between its turns. Although this type of distortion occurs only on the peaks, it causes a rasping effect which greatly impairs the quality. It is effectively eliminated by placing small resistors of about 40 ohms in each plate lead, directly between the plate connection and the primary of the output transformer, and, by connecting .001 mfd. condensers across each half of the grid circuit. The two resistors connected across the grid circuit also help to eliminate this type of distortion. Fig. 16 illustrates a Class B stage including the components necessary to prevent this dynatronic oscillation.

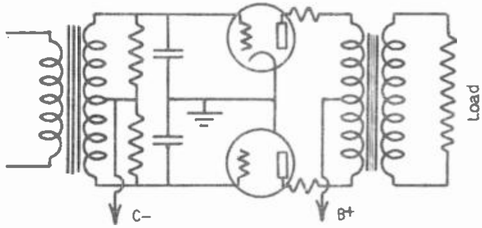


Fig.16 A Class B audio stage with the necessary additions to prevent dynatronic oscillations.

The relation between grid voltage and plate current must be essentially linear over the operating range, if third and higher odd harmonics are to be negligible. The linearity of the dynamic E_g-I_p characteristic curve depends upon the choice of grid bias, plate voltage, and load resistance. During operation, the plates of the Class B tubes will have an alternating voltage component applied to them as the plate currents increase and decrease through the load resistance. Thus, the actual voltage applied to the plates of the tubes will rise and fall. Care must be taken that the maximum positive grid potential is not greater than the minimum voltage applied to the plates. Since the AC voltage on the grids is 180° out of phase with the AC voltage on the plates, the maximum grid voltage occurs at the same time as the minimum plate voltage. Should the maximum grid voltage equal the minimum plate voltage, the grid would rob the plate of electrons which should be attracted by the plate, and, as a result, the linear relation between grid voltage and plate current will no longer exist. In fact, the maximum grid potential should not exceed 50% of the minimum plate voltage, if this type of distortion is to be avoided.

With a given excitation voltage, the minimum plate voltage is determined by the size of the load resistance. As the load resistance is increased, the minimum plate voltage becomes less, and the power output and efficiency both increase. Thus, to obtain the maximum power output with a given grid-exciting voltage, the load resistance should be as large as possible without causing the minimum plate voltage to fall so low that a condition of non-linearity exists. With the load impedance constant, excessive grid excitation has the same effect as excessive load impedance; that is, it causes the maximum grid voltage to approach the minimum plate voltage too closely, and thereby introduces distortion. In a like manner, insufficient grid excitation produces the same result as insufficient load impedance, and therefore results in low power output and low efficiency. The actual value of the load impedance should be somewhat greater than the value which would be used if the tubes were operating under Class A conditions. This is necessary because distortion due to non-linearity of the dynamic E_g-I_p characteristics is cancelled only at low values of plate current when both tubes are conducting.

Fig. 17 illustrates several dynamic E_g-I_p and E_g-I_g characteristic curves for various values of load impedance. These particular curves are for a pair of type 46 tubes used as a Class B amplifier stage. The type 46 is a tetrode and has two grids. By connecting the grid nearest the plate to the plate, the tube is

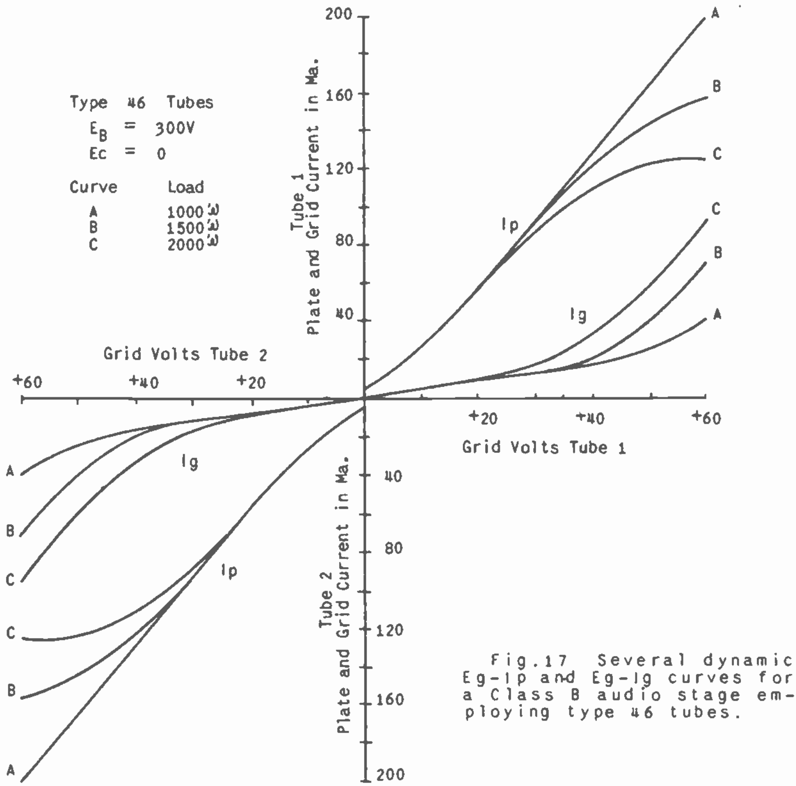


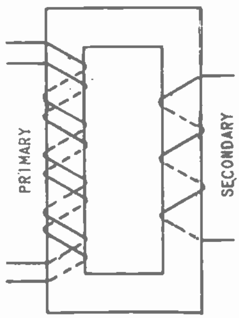
Fig.17 Several dynamic Eg-Ip and Eg-Ig curves for a Class B audio stage employing type 46 tubes.

caused to have an amplification factor of 5.6 and a plate resistance of 2380 ohms. With this connection, it requires a grid bias of -33 volts, and operating as a Class A amplifier, it may be used as the driver stage for a pair of type 46 tubes in Class B. When the two grids of this tube are tied together, it has an entirely different set of characteristics. The amplification factor and plate resistance increase considerably, and the plate current with 400 volts plate voltage and 0 volts grid bias is only 6 ma. With this low value of plate current, the Class B tubes may be operated with zero bias, and with a load impedance of 2000 ohms, and an excitation voltage sufficient to drive the grids to 50 volts positive, the peak plate current is approximately 150 ma., and the power output about 20 watts. With these conditions, the average driving power required is .65 watt.

Notice from the curves of Fig. 17 that very high values of load impedance cause the dynamic characteristic to depart from a straight line. Also, observe that the peak grid current attained, for any given value of excitation, becomes greater as the plate load is increased. When the load resistance is greater than the recommended value, the plate currents will be lower and the grid currents higher than their rated values. Likewise, a load which is less than that recommended will cause the plate currents to be high and the grid currents low. The Class B amplifier actually works into only one-half of the primary of the output transformer, and thus the total plate-to-plate impedance measured across the entire primary winding is four times the actual load impedance.

The output transformer must be large enough to handle the amount of power that is to be transferred to the load. The resistance of the windings and the leakage inductance must be kept at a minimum. Since it takes two windings on the primary of the Class B transformer to simulate a single primary winding of a Class A transformer, it is desirable to interspace these two windings with each other so that they will be balanced with respect to resistance and leakage. A schematic drawing showing how this is done is shown in Fig. 18.

Fig.18 Showing how the two halves of the primary of the output transformer are inter-wound so that they will be balanced with respect to resistance and leakage.



There is no simple method for determining the best load resistance to be used. The optimum value will depend on the power output desired, the grid excitation available, and the allowable peak plate current. For example, a pair of 203-A tubes in Class B audio with a plate voltage of 1000 volts should be worked into a load of 6900 ohms plate-to-plate, and will produce a power output of 200 watts at full excitation. By applying 1250 volts, (the maximum allowable) to the plates of these tubes, they will produce a maximum power output of 260 watts, but must be worked into a load of 9000 ohms plate-to-plate. In each case the plate load used is that value which causes the tubes to dissipate the maximum power permissible. With small tubes, the allowable peak plate current is usually the limiting factor rather than the plate dissipation.

The power output of a Class B audio stage is found by the following formula:

$$P_o = \frac{I_m^2 \times R}{2} \qquad (3)$$

Where: P_o is the power output in watts,
I_m is the peak plate current drawn by either tube,
R is one-fourth of the plate-to-plate load impedance.

Whereas it is possible to draw load lines for Class B audio stages, the procedure is so complicated that ordinarily this is not done. Instead, reference is made to the manufacturer's specifications. The manufacturer gives the proper operating conditions for at least one value of plate voltage (usually the maximum plate voltage allowable), and with this information it is comparatively simple to calculate the power output and proper load impedance for other values of applied plate voltages.

Consider, for example, a pair of 203-A tubes working as a Class B audio stage. From the tube manual, the following information is available.

DC plate voltage...............1250 volts
Max. signal DC plate current...320 ma. (two tubes)
Plate-to-plate load...........9000 ohms
Power output..................260 watts

The first step is to calculate the peak plate current per tube. This may be done by multiplying the maximum signal DC plate current for the two tubes by 1.57.

Peak plate current per tube = 1.57 × 320 = 502 ma.

The load impedance per tube is one-fourth of the plate-to-plate load impedance or is:

$$\frac{9000}{4} = 2250 \text{ ohms.}$$

With a peak plate current of 502 ma. and a load of 2250 ohms, the peak voltage drop across the load is .502 × 2250 = 1130 volts. Therefore, the voltage drop across the tube at this instant is the difference between the applied voltage of 1250 volts and the drop across the load of 1130 volts. This is:

1250 - 1130 = 120 volts.

Having determined the peak plate current and the minimum voltage drop across the tubes, we are able to calculate the power output and optimum load impedance with other values of applied plate voltage. For example, suppose that our power supply is capable of furnishing only 1100 volts. Assuming the same voltage drop across the tube (120 volts), with 1100 volts applied, the voltage drop across the load will be: 1100 - 120 or 980 volts.

The load impedance per tube may be found from this formula:

$$R = \frac{E_1}{I_m}$$

Where: R is the load impedance per tube,
E_1 is the voltage across the load
I_m is the peak plate current per tube.

In our example, the load impedance per tube would be:

$$R = \frac{980}{.502} = 1952 \text{ ohms.}$$

Therefore, the plate-to-plate load impedance would be four times this value or 4 × 1952 = 7808 ohms.

By using formula (3) the power output may be calculated. It is:

$$P_o = \frac{.502^2 \times 1952}{2} = 246 \text{ watts.}$$

Thus, when using a plate voltage of 1100 volts rather than 1250 volts, the power output will be reduced from 260 watts to 246 watts, and the plate-to-plate load impedance should be changed from 9000 ohms to 7800 ohms.

In summarizing the precautions to be observed in designing a Class B stage, the following points are considered to be of major importance.

1. The driver stage should be able to furnish two to three times the actual power required to excite the grids of the Class B tubes.
2. The Class B input transformer must have a sufficient step-down ratio so that the resistance reflected to the primary of the input transformer is never less than the plate resistance of the driver tube, even when the grids are most positive.
3. The load impedance of the Class B stage must be fairly high compared to the plate resistance of the Class B tubes, because distortion due to nonlinearity is not cancelled except at low values of plate current.
4. The load impedance must not be so great that the minimum plate voltage is equal to the maximum grid voltage.
5. The two halves of the circuit must be accurately matched. Each tube furnishes one alternation of the output waveform, and a mismatch between the tubes will cause the two alternations to be different. Since tubes do not retain their original characteristics for an indefinite period, it is necessary that individual bias adjustments be provided unless zero-bias tubes are used. The grid driving voltage and the no-signal plate current

of each tube must be equal so that even harmonics will not be introduced.

6. The plate power supply must have exceptionally good voltage regulation to prevent flattening of the plate current peaks.
7. Means must be taken to prevent dynatronic grid oscillation in case it is present.

Class B amplification affords an economical means of securing relatively large amounts of audio power from small tubes. Its application, however, is not limited to receiving type tubes. Many transmitter tubes may also be operated in Class B amplifiers; the type 849 is representative. Two such tubes operating as a Class B amplifier will deliver nearly 1000 watts of audio power, enough to modulate 2000 watts of DC power input to a Class C stage. The theoretical maximum efficiency of a Class B stage is 78.5%, and efficiencies of 60% or better are ordinarily obtainable without much difficulty. Compared to the 20% efficiency secured with a Class A amplifier, it is evident that considerable power may be obtained from a Class B stage without exceeding the maximum plate dissipation.

DECIMAL EQUIVALENTS

1/64 = .015625	1/4 = .250	1/2 = .500	3/4 = .750
1/32 = .03125	17/64 = .265625	33/64 = .515625	49/64 = .765625
3/64 = .046875	9/32 = .28125	17/32 = .53125	25/32 = .78125
	19/64 = .296875	35/64 = .546875	51/64 = .796875
1/16 = .0625	5/16 = .3125	9/16 = .5626	13/16 = .8125
5/64 = .078125	21/64 = .328125	37/64 = .578125	53/64 = .828125
3/32 = .09375	11/32 = .34375	19/32 = .59375	27/32 = .84375
7/64 = .109375	23/64 = .359375	39/64 = .609375	55/64 = .859375
1/8 = .125	3/8 = .375	5/8 = .625	7/8 = .875
9/64 = .140625	25/64 = .390625	41/64 = .640625	57/64 = .890625
5/32 = .15625	13/32 = .40625	21/32 = .65625	29/32 = .90625
11/64 = .171875	27/64 = .421875	43/64 = .671875	59/64 = .921875
3/16 = .1875	7/16 = .4375	11/16 = .6875	15/16 = .9375
13/64 = .203125	29/64 = .453125	45/64 = .703125	61/64 = .953125
7/32 = .21875	15/32 = .46875	23/32 = .71875	31/32 = .96875
15/64 = .234375	31/64 = .484375	47/64 = .734374	63/64 = .984375

NUMBER - SIZE DRILLS

No.	Dia.	No.	Dia.	No.	Dia.	No.	Dia.	No.	Dia.
1	.2280	17	.1730	33	.1130	49	0730	65	.0350
2	.2210	18	.1695	34	1110	50	.0700	66	.0330
3	.2130	19	.1660	35	.1100	51	.0670	67	.0320
4	.2090	20	1610	36	.1065	52	0635	68	.0310
5	.2055	21	.1590	37	.1040	53	.0595	69	.0292
6	.2040	22	.1570	38	.1015	54	.0550	70	0280
7	.2010	23	.1540	39	.0995	55	.0520	71	.0260
8	.1990	24	.1520	40	.0980	56	.0465	72	.0250
9	.1960	25	.1495	41	.0960	57	.0430	73	.0240
10	.1935	26	.1470	42	.0935	58	.0420	74	.0225
11	.1910	27	.1440	43	.0890	59	.0410	75	.0210
12	.1890	28	.1405	44	.0860	60	.0400	76	.0200
13	.1850	29	.1360	45	.0820	61	.0390	77	.0180
14	.1820	30	.1285	46	.0810	62	.0380	78	.0160
15	.1800	31	.1200	47	.0785	63	.0370	79	.0145
16	.1770	32	.1160	48	.0760	64	.0360	80	.0135

LETTER SIZE DRILLS

Letter	Diam.	Letter	Diam.	Letter	Diam.	Letter	Diam.	Letter	Diam.
A	.234	G	.261	L	.290	Q	.332	V	.377
B	.238	H	.266	M	.295	R	.339	W	.386
C	.242	I	.272	N	.302	S	.348	X	.397
D	.246	J	.277	O	.316	T	.358	Y	.404
E	.250	K	.281	P	.323	U	.368	Z	.413
F	.257								

NOTE: To avoid possibility of error when ordering number and letter size drills, always specify both number and diameter in decimals of an inch. There is not so much confusion regarding drill-gage sizes at present as there has been in the past, but, so long as there is any possibility of confusion between drill and steel-wire gages trouble will be saved the user by specifying both number and decimal equivalent.

DRILL SIZES FOR MACHINE AND WOOD SCREWS

N.C. OR U.S.S. SCREW THREADS

Tap or Screw Size	Number of Threads Per In.	Size of Tap Drill
1/16	64	56
1/8	40	38
3/16	32	22
1/4	20	7
5/16	18	F
3/8	16	5/16
7/16	14	U
1/2	13	27/64
9/16	12	31/64
5/8	11	17/32
11/16	11	19/32
3/4	10	21/32
13/16	10	23/32
7/8	9	49/64
15/16	9	53/64
1	8	7/8

N.C. = American National Coarse Thread Series
U.S.S. = United States Standard Screw Threads

WOOD SCREWS

Gage of Screw	Dia. in Decimals	Body Hole Drill	Lead Hole Drill	C'Bore Drill
0	.058	53	68	32
1	.071	49	57	20
2	.084	44	56	16
3	.097	40	52	4
4	.111	33	52	B
5	.124	1/8	52	F
6	.137	28	47	L
7	.150	24	47	O
8	.163	19	42	S
9	.177	15	42	T
10	.190	10	38	X
11	.202	5	38	7/16
12	.215	7/32	38	29/64
14	.242	D	31	33/64
16	.269	I	28	37/64
18	.294	19/64	23	41/64

Lead holes are seldom used for Nos. 0 and 1 gage screws. In soft wood, lead holes are unnecessary for gages less than No. 6.

MACHINE SCREWS

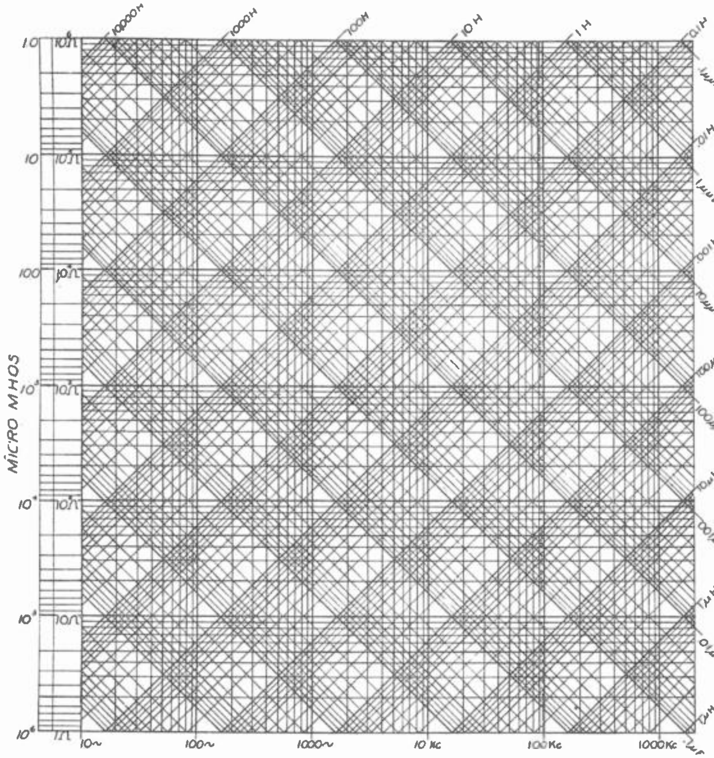
Tap or Screw Size	Threads Per Inch	Tap Drill	Body Drill
0	80 N.F.	3/64	51
1	64 N.C.	53	47
1	72 N.F.	53	47
2	56 N.C.	50	42
2	64 N.F.	50	42
3	48 N.C.	47	37
3	56 N.F.	45	37
4	40 N.C.	43	31
4	48 N.F.	42	31
5	40 N.C.	38	28
5	44 N.F.	37	29
6	32 N.C.	36	27
6	40 N.F.	33	27
8	32 N.C.	29	18
8	36 N.F.	29	18
10	24 N.C.	25	9
10	32 N.F.	21	9
12	24 N.C.	16	2
12	28 N.F.	14	2

N.F. OR S.A.E. STANDARD SCREWS

Tap or Screw Size	Number of Threads Per In.	Size of Tap Drill
1/4	28	3
5/16	24	I
3/8	24	Q
7/16	20	25/64
1/2	20	29/64
9/16	18	33/64
5/8	18	37/64
11/16	16	5/8
3/4	16	11/16
7/8	14	13/16
1	14	15/16
1 1/8	12	1 3/64

N.F. = American National Fine Thread Series
S.A.E. = Society American Engineers

All Tap Drills Allow Approximately 75% Full Thread



KC = KILOCYCLES μF · MICROFARADS H · HENRYS

Ω · OHMS μμF · MICROMICROFARADS μH · MICROHENRYS

The construction and use of this chart is as follows:

The sloping lines from lower left to upper right represent inductances and those from lower right to upper left, capacitances. The main divisions are subdivided on a logarithmic scale, each line being an even multiple of the first. For example, the light line immediately following the dark 1 Henry line represents two times 1 Henry; the next line represents 3 Henrys; etc. It should be noted that when interpolating between minor divisions, the same logarithmic gradation should be used as in the major divisions. To obtain the reactance or admittance of a capacitance or inductance, the sloping line should be followed to the point where it intersects with the vertical line corresponding to the frequency at which the value is desired. The horizontal line which corresponds with this intersection when carried to the left gives the value of reactance or admittance. In a similar manner, if reactance and frequency are known, the inductance or capacitance can be determined by a reverse process.

EXAMPLE TO DETERMINE REACTANCES.

To find the reactance of a coil having an inductance of 1 Henry at a frequency of 1000 cycles, we follow the slanting 1 Henry line until it meets the vertical 1000 cycle line. At the point of intersection, a horizontal line to the left edge of the chart indicates a reactance of 6200 ohms. Similarly, the sloping .1 mfd. line can be followed to the point where it intersects with the vertical 1000 cycle line and a horizontal line from the point of intersection to the left side of the chart indicates a reactance of 1600 ohms.

EXAMPLE TO DETERMINE RESONANCE.

To determine the point of resonance, let us consider the values of inductance and capacitance chosen in the previous example. We follow the sloping 1 Henry line to the point where it intersects with the .1 mfd. line. The vertical line through the point of intersection indicates a resonance frequency of 500 cycles.

EXAMPLE OF PARALLEL CIRCUIT CALCULATIONS.

The mho scale is primarily useful in the solution of parallel circuits. Consider an inductance of 1 Henry in parallel with .1 mfd. at 1000 cycles. The admittance of 1 Henry is found by the chart to be 161 micromhos at 1000 cycles, and the admittance of a .1 mfd. condenser 625 micromhos at the same frequency. In adding admittances of inductances in parallel or condensers in parallel, the admittances are simply added without regard to sign. However, the sign of capacitance and inductance are opposite when used in parallel. Therefore, in this example, the total admittance is 161 minus 625, or minus 464 micromhos. Glancing at the left side of the chart, this is found to correspond to 2160 ohms, which is the total reactance of the combination.

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POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

CLASS C AMPLIFIERS

LESSON NO. 7

TRANSPORTATION....

The steady drone of powerful engines and the beat of steel propellers driving huge, luxurious liners of the air through the air lanes at tremendous speeds.

A peculiar moaning whistle, a swift flash as beautiful, streamlined, stainless steel trains click over the rails for new transcontinental speed records.

COMMUNICATION....

Modern transportation, with its terrific speed, its split-minute schedules and vast number of important messages must have available swift, reliable communications.

Radio communication has solved the problem for the airlines most effectively. The radiotelephone and radiotelegraph are even vastly swifter than the planes they keep operating on schedule.

Telegraph and the 'Morse code' have served the railroads faithfully for years. But time passes. The old 'iron horse' is on the way out. In its stead we have the new 'streamliners' and the application of radio to rail transportation.

In the air..on the land..on the water..transportation is dependent upon radio for communications. This is just one of the many marvelous fields that will be open to you as a Midland Graduate.

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Lesson Seven

CLASS C AMPLIFIERS

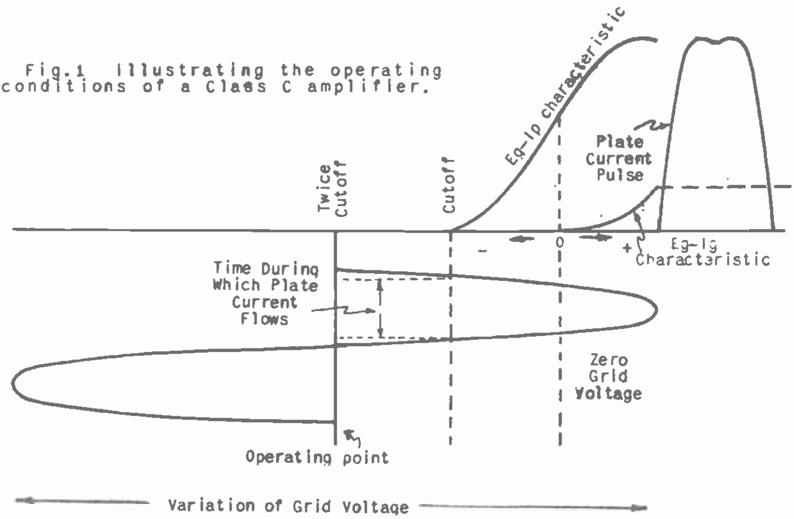
"The study of how a Class C amplifier operates is one of the most interesting subjects which will be encountered by the broadcast engineer. Because of the extreme importance of this material, no student desiring a complete knowledge of this type of amplifier, should hurry through this lesson.

"Whenever plate modulation is employed in a transmitter, Class C amplifiers must be used. Therefore it is advisable that you have a complete understanding of the operation of this type of equipment."

1. **THEORY OF OPERATION.** Strictly speaking, the Class C amplifier is any amplifier stage in which the grid bias is greater than that value required to produce plate current cut-off. This definition corresponds to that given in the Standardization Rules of the Institute of Radio Engineers. It has, however, become common practice to apply the term Class C amplifier only to the modulated stage which uses a grid bias of at least twice cut-off. When the amplifier uses a bias approximately one and one-half times cut-off, it is sometimes called a "Class BC amplifier", indicating that its characteristics are midway between those of a Class B and a Class C amplifier. Sometimes buffer stages are operated under Class BC conditions, and the grid-modulated amplifier (to be discussed in a later lesson) is another example of this type of operation.

With a grid bias of twice cut-off, no plate current flows unless an exciting voltage is applied. Furthermore, since plate current flows for less than 180° of the cycle with excitation, it is evident that the relation between the plate current and the grid voltage is far from linear. A grid exciting voltage of sine wave form produces a plate current wave which in no way resembles a sine wave. Instead, the plate current consists of pulses lasting from one-third to one-sixth of a cycle or less, depending upon the values of the applied voltages and upon the load. The shape of the plate current pulse, whether it is peaked, flat topped, or has a dip, is determined by the dynamic I_p-E_g characteristic. This, in turn, depends on the load. The average operating characteristics of a Class C amplifier are shown in Fig. 1.

Fig.1 Illustrating the operating conditions of a Class C amplifier.



The plate current of a Class C amplifier is not of sine wave form, and, therefore, consists of a fundamental and a series of harmonic frequencies. It would appear that this plate current would cause considerable distortion and that the voltages produced across the load would contain a large number of harmonics. This would be the case, except for the fact that the load is of a particular kind. The load of a Class C amplifier is always a parallel tuned circuit. Such a load has the ability to smooth out the irregularly shaped plate current pulses into practically pure sine waves, or to accept the fundamental frequency and reject the harmonics. This is accomplished by the fly-wheel effect of the tuned circuit. The tuned tank circuit stores a considerable amount of energy, and the plate current pulses merely replenish this energy as it is absorbed by the transmitting antenna. If the adjustment of the load circuit is correct, the voltage across the tank is very nearly sinusoidal and therefore contains very few harmonics.

Since a Class C amplifier must work into a tuned load circuit, it is obviously impossible to use this type of amplifier for the amplification of audio frequencies. The Class C amplifier is a single-frequency amplifier and cannot be used to amplify a modulated R.F. wave. An inspection of Fig. 2 will make this apparent. The grid exciting voltage is a 100% modulated R.F. wave, and the grid bias is twice cut-off. Practically, there are two reasons why the Class C amplifier cannot be used in this manner. The peaks of the modulated R.F. wave vary from zero to twice their unmodulated value. When these peaks are less than that value required to drive the grid voltage above the cut-off value, no plate current flows, and thus the Class C stage amplifies only the higher peaks of the modulated R.F. wave. Such a condition produces distortion of the audio envelope contained in the modulated wave. A second reason is that the dynamic E_g-I_p characteristic is not linear, and, as a

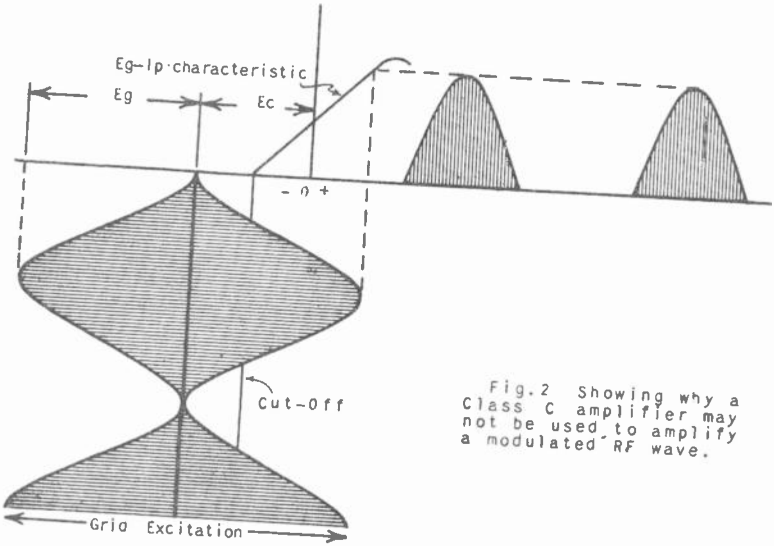


Fig. 2 Showing why a Class C amplifier may not be used to amplify a modulated RF wave.

result, the envelope of the plate current pulses is not the same as the envelope of the grid exciting voltage and distortion would occur, even if plate current cut-off did not take place.

In its appearance in circuit diagrams, the Class C amplifier is exactly like the buffer stage. A buffer amplifier supplying the excitation for a Class C stage is shown in Fig. 3. A dummy antenna is included to absorb the power from the Class C stage. From Fig. 1, it is evident that the grid is driven considerably positive during a part of the grid excitation cycle with the result that a grid current of approximately 25% of the plate current flows. Of course,

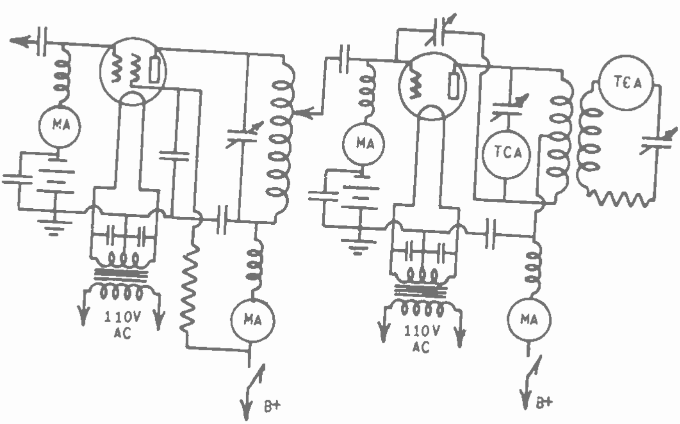


Fig. 3 A buffer stage capacitively coupled to a Class C Amplifier.

the grid current flows for a shorter time than the plate current, because grid current can only flow while the grid is positive.

It should now be evident that the Class C amplifier is very similar to a self-excited oscillator in its operating characteristics. In fact, it may be considered to be a separately-excited oscillator, and all the curves given in Lesson 1 of this Unit apply equally well to the Class C amplifier.

2. **PLATE CIRCUIT EFFICIENCY.** The advantage of the Class C amplifier and the reason it is used instead of a Class A or a Class B amplifier to amplify an unmodulated radio frequency wave is its efficiency. In the preceding lesson, it was discovered that the major reason that Class A amplifiers were not used to obtain large amounts of power was that their efficiency is so low. With an average efficiency of about 20%, not much power can be secured from a Class A amplifier before the plate dissipation rating of the tube is reached. Fully 80% of the power drawn from the power supply is lost in the heating of the plate of the tube. To employ a Class A amplifier for the production of large amounts of power requires a very large tube capable of dissipating at least four times the amount of power output desired. In addition, the plate voltage will have to be very high and a very expensive power supply will be required. It was these disadvantages that led to the development of the Class B amplifier which has a considerably higher efficiency. This greater efficiency is secured by reducing the proportion of the input power which serves only to heat the plate of the tube. Specifically it is accomplished by increasing the bias to the cutoff point so that plate current flows for only half of a cycle. With plate current flowing only half of the time, the plate is able to cool during that part of the cycle when no plate current flows and the average power dissipated at the plate is reduced appreciably.

This method of increasing the plate circuit efficiency is carried one step further in the Class C amplifier. If allowing the plate current to flow only half the time results in a decided increase in the efficiency, then is it not logical to assume that the efficiency will be further increased if the plate current is allowed to flow for only one-third or one-sixth of the time? Such has been found to be the case, and the comparatively high efficiency of a Class C amplifier is due to the fact that plate current flows only during a short time each cycle, and that this flow of plate current occurs at the time that the plate voltage is minimum.

Efficiencies as high as 90% are possible in well designed laboratory Class C amplifiers operating with very low power output. This, however, is very uncommon; the average efficiency obtainable will range from 50 to 70% in commercial Class C amplifiers. It should also be realized that not all of the losses take place at the plate of the tube. Losses occur in the tank circuit due to the oscillating current flowing through the plate tank resistance. The efficiency is found by determining the ratio of the power actually delivered to the antenna to the power input furnished by the power supply. It might be thought that this efficiency could be increased on up to practically 100% by further increasing the grid bias and allowing the plate current to flow during even smaller parts of the cycle. It is true that such an action will reduce the losses at the plate, but the plate current pulses produced have such a high harmonic content that a larger oscillating tank current is required to smooth them into pure sine waves, and the result is that more power is dissipated in the tank itself and the actual amount of power reaching the antenna changes very little. Thus, if the power reaching the antenna is to be relatively free from harmonics, the upper practical limit of efficiency is from 60 to 70%. Increasing the bias beyond the point that produces this amount of efficiency is unwarranted, since the efficiency increases very little and a much higher R.F. grid exciting voltage is necessary.

3. **POWER GAIN.** The power gain of any amplifier circuit is the ratio of the power output to the power required to drive the grid circuit. If 10 watts of grid driving power are required to produce a power output of 90 watts, the power gain would be 9.

The power gain of a Class A amplifier is theoretically infinite. Since the Class A amplifier does not draw any grid current, no power is dissipated in the grid circuit, and no driving power is required. Therefore, no matter how small the output power may be, the power gain is still infinite, since the driving power is zero.

A Class B amplifier, on the other hand, does draw grid current, and therefore requires some grid driving power. The amount of grid driving power needed depends on how far the grids are driven positive, and on the maximum grid current that flows. The maximum grid current, in turn, is determined by both the grid voltage and the plate voltage at the instant that the greatest grid current flows. The greater the maximum grid voltage, the larger the maximum grid current will be, and the smaller the plate voltage at this instant, the larger will be the amount of grid current. It might seem odd that the plate voltage could have any effect on the grid current, but such is the case. With a low plate voltage, a larger proportion of the emitted electrons will be attracted by the grid, whereas a higher plate voltage will be able to attract more of the electrons to the plate which otherwise would become a part of the grid current. Just how low the plate voltage will be when the grid voltage is maximum depends on how large a value of load impedance is used. A high value of load impedance will cause the minimum plate potential to be less, with the result that the maximum grid current will be higher, and the driving power greater. The power gain of a Class B amplifier will range from 40 to 60. Thus, if the power gain is 40, it will require 4 watts of driving power to obtain 160 watts of power output.

Since the grid of a Class C amplifier is driven even more positive than the grids of a Class B stage, it is evident that more driving power will be necessary, and the power gain will be less. On the average, the power gain of a Class C amplifier is approximately 10. An amplifier whose output is 10 watts will ordinarily be sufficient to drive a 100 watt stage.

There are two sources of power loss in the grid circuit. The first is the lòss incurred by allowing the grid current to flow through the grid-to-filament resistance of the tube. This power loss is manifested by the heating of the grid electrode. The second loss is due to the DC grid current flowing through the source of grid bias. If the bias voltage is developed by a grid leak resistor, it is not hard to see that power will be dissipated in this resistor. However, if fixed or battery bias is employed, it is not so easy to see how this loss occurs. From Fig. 3, note that the negative terminal of the C battery is connected toward the grid. When grid current flows, it travels from the filament to the grid, to the negative terminal of the C battery, through the battery from negative to positive, and back to the filament. This grid current, therefore, flows through the battery in the opposite direction to that which the battery would force a current, if a closed circuit were provided. For the grid current to overcome the opposing voltage of the C battery requires power, and this power appears as heat which dries out the battery and materially shortens its life. The actual amount of power dissipated in the C battery is equal to the DC grid current times the battery voltage.

When the grid bias is made larger in an attempt to increase the plate circuit efficiency, the excitation voltage must also be increased to obtain the same power output. This causes the DC grid current to remain the same, and, in flowing through a larger bias voltage source, this grid current will dissipate more power. Thus the grid losses increase and the power gain is reduced, although the plate circuit efficiency is greater. It is, however, false economy to continue to increase the plate circuit efficiency by using larger and larger bias voltages, for such an action causes the grid losses to mount rapidly. As stated before, it would be possible to have a plate efficiency of 90%, but it is very probable that the grid driving power required in such a case would be nearly one-third of the power output. Experiment indicates that a power gain of approximately 10 is the correct operating value, and attempts to further increase the plate circuit efficiency result in more disadvantages than advantages.

4. **METHODS OF OBTAINING GRID BIAS.** Although battery bias is shown in the diagram of Fig. 3, it is never used in commercial transmitters. Battery power is expensive. Although the battery is not required to furnish any current, yet the grid current which flows through it in such a direction as to tend to charge it, creates considerable heat, and thereby destroys the usefulness of the battery in a very short time.

Grid-leak bias is far more economical than battery bias, and is also advantageous in that much less space is required. Also, as we shall learn later, the use of grid leak bias tends to make the modulation of a Class C amplifier more linear. The outstanding disadvantage of this type of bias is that bias voltage is present only as long as there is some grid excitation. Should the grid excitation fail, the bias would be lost, and there is a possibility that the plate current would be excessive, causing damage to the tube.

To avoid possible damage to the tube (especially if the mu is low), it is better to use a combination of grid leak and self-bias. Self bias sufficient to prevent excessive plate current may be provided by a resistor of adequate size from filament center tap or cathode to ground. Also, a separate bias pack may be used for protection.

It is probable that the bias will have to be changed a number of times during the adjustment of the amplifier, and a resistor with a sliding tap makes these changes somewhat more simple.

5. GRID EXCITATION. Normally, very little adjustment of the grid excitation is necessary or desirable in a transmitter amplifier stage. It does not often happen that there is an excess of grid excitation available. Ordinarily, the maximum excitation obtainable from the preceding buffer stage is used, and any adjustment of the grid circuit is made by varying the grid bias. *A tube with a high amplification factor, however, has a relatively low grid impedance, and does not require as much excitation voltage to produce a given power output, and, in this case, it may be that the maximum excitation available is excessive.* An attempt should be made to match the grid impedance of the Class C stage to the load impedance of the preceding driver stage. With low or medium-mu tubes, a sufficient match is obtained by placing the excitation tap at the plate end of the preceding tank circuit. A high-mu tube, on the other hand, has such a low value of grid impedance that it is necessary to tap down on the preceding tank coil to secure a fair impedance match.

It should be evident that adjusting the grid excitation is merely attempting to match the grid circuit of the Class C stage to the plate circuit of the driver stage; a high-impedance grid requires a high excitation voltage and a low-impedance grid needs a low excitation voltage.

Normally, the grid excitation is sufficient to produce plate current saturation: that is, the grid is driven enough positive that at its peak the tube is working near the upper curved portion of the dynamic E_g-I_p characteristic. Thus, a simple method of determining when the excitation is correct is to increase the grid excitation until a further increase does not produce an increase in the DC plate current. Increasing the excitation beyond this point results in a decrease in the DC plate current, because the maximum grid voltage reached is approximately equal to the minimum instantaneous plate voltage at this time, and the grid attracts a disproportionate share of the emitted electrons.

If the preceding stage does not draw normal rated plate current, it is more than probable that the grid excitation of the Class C stage is insufficient, whereas an excessive plate current in the driver stage may indicate that the excitation of the Class C stage is too great. The fact that the plate current of the driver stage is not normal, however, is no sure indication that the grid excitation of the Class C amplifier is not of the proper value. It may so happen that the plate impedance of the driver is so far removed in value from the grid impedance of the driven stage that an approximate impedance match is not possible by simply adjusting the position of the grid excitation tap.

If the driver tube does not draw normal plate current even when the excitation tap is at the plate end of the tank coil, it is probably that the load impedance of the driver stage is too high. It may be made less by using a smaller inductance and a larger capacitance in the tank, or by tapping the plate lead of the driver down on the tank coil as shown in Fig. 4.

Capacitive coupling will give just as good results as any other method of coupling, but it is somewhat more difficult to adjust for a proper impedance match than is link coupling. In commercial transmitters, laboratory facilities are available for determining the proper capacity of the coupling condenser and the position of the excitation tap. Amateurs, on the other hand, nearly always use link coupling, since it makes impedance matching more simple. If the excitation is insufficient, the coupling link is moved toward the plate end of the driver tank or more turns are added to the coupling link. Excessive excitation is reduced by reversing this procedure.

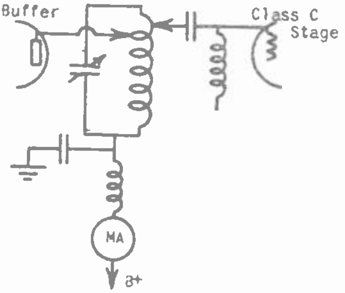


Fig. 4 Tapping the buffer plate down on its tank coil so that it will work into the correct load impedance.

6. **PLATE VOLTAGE.** *(When properly adjusted, the power output of a Class C amplifier varies directly as the square of the plate voltage, and therein is seen the desirability of using a high plate voltage to obtain large amounts of R.F. power.)* Every tube, of course, has a maximum plate voltage rating which should not be exceeded. This maximum plate voltage is determined by the insulation resistance between the plate connection and the connections to the other elements of the tube, and by the gas content of the tube. In high-powered tubes, the insulation resistance is made large by so designing the tube that the connection to the plate is as far from the other connections as possible. Thus, it is common practice to construct transmitting tubes with the filament and grid connections at the bottom of the tube and with the plate connection brought out to a top cap. In addition to raising the maximum allowable plate voltage, this method of construction also minimizes the plate-to-grid capacity and tends to make neutralization easier.

Naturally, if the plate voltage is too high, arc-over can take place directly from the plate surface to the grid (or filament). By placing the plate relatively far from the other electrodes, a greater plate voltage may be used, but such design increases the plate resistance of the tube which is undesirable. The actual amount of voltage required to produce an arc-over from the plate to the other electrodes depends also on the air pressure between them. With a given distance between them, the arcing voltage is quite high at normal atmospheric pressure. It is not, however, possible to operate the tube in normal air pressure; a partial vacuum must be provided. As the pressure within the tube is reduced, the voltage necessary to produce an arc - over decreases at first. At some particular value of air pressure, which will depend on the spacing of the electrodes, the arcing voltage is least. A continued decrease in the air pressure raises the value of the required arcing voltage. All vacuum tubes are evacuated until the best possible vacuum is obtained. It should not be thought, however, that the air pressure in a vacuum tube is zero; this is far from the truth. In any vacuum tube there are literally billions of air molecules remaining after the best possible vacuum is secured. It is the presence of the air molecules which prevent the application of a plate voltage in excess of that specified by the manufacturer.

The actual vacuum obtained will probably vary slightly even with tubes of the same type; some will have a larger and some a smaller gas content than the average. The manufacturer so chooses the maximum plate voltage that any of the tubes which he markets will be able to withstand it. That is, he guarantees his tubes for any plate voltage up to this maximum rating. It is probable that a few specific tubes having a gas content slightly lower than the average would be able to withstand a little higher voltage, but these would be exceptions. On the other hand, any tube not able to withstand the rated maximum would be rejected by the inspection department. It is never wise under any conditions to apply a plate voltage in excess of the rated maximum. An arc-over may occur, and, if it does, the tube is made worthless. Transmitting tubes are far too expensive to be experimented with in this manner.

The higher the plate voltage, the easier a tube is to drive to a desired power output with a given plate efficiency.

With a given grid excitation and power output, the plate efficiency will be raised by increasing the plate voltage.

The actual power output is the antenna current squared times the antenna resistance. The power delivered to the tank circuit is the R.M.S. value of the fundamental component of the plate current times the R.M.S. value of the R.F. voltage built up across the tank. With an increased plate voltage, the R.F. voltages produced across the tank circuit will be correspondingly larger, and the power output will be increased. The minimum plate voltage will be practically the same as with the lower plate voltage. By increasing the grid bias to correspond to the increased value of plate voltage, the plate current that flows can be made practically the same as with the lower plate voltage. Thus, the power dissipated at the plate remains practically constant although the power output increases, and therefore, the efficiency must have been raised when the higher plate voltage was applied. Increasing the plate voltage is the only means of increasing the plate efficiency which does not make the grid circuit losses greater.

7. Q AND L/C RATIO. The gain of a tuned circuit was discussed in Lesson 22 of Unit 1. At that time, it was learned that the gain of a tuned circuit is equal to the reactance of either the coil or the condenser divided by the resistance. This ratio of reactance to resistance is very important and is used so often in speaking of transmitting circuits that it is given a special symbol. This symbol is Q, and is used in referring to this ratio of reactance to resistance, whether the ratio pertains to a coil, a condenser, a tuned circuit, or any other type of complicated circuit. The Q of a coil is the ratio of the coil's reactance to its resistance. It is sometimes called the "figure of merit" of the coil. The Q of a condenser is also the reactance of the condenser divided by the resistance. Since the resistance of a condenser is ordinarily considerably less than the resistance of a coil, it is evident that condensers will have higher Q's than will coils.

All tuned circuits are operated at the resonant frequency, in which case the inductive reactance is equal to the capacitive reactance. Therefore, the Q of a tuned circuit is equal to the reactance of either the coil or the condenser divided by the total resistance contained in the tuned circuit.

A Class C amplifier works into a parallel tuned circuit. When the stage is unloaded; that is, when the dummy antenna is not coupled to the tank circuit, the Q of this tank is fairly high, perhaps as high as 100. Coupling the antenna to the tank causes a resistance to be reflected from the antenna into the tank circuit. Since the apparent resistance of the tank is increased and its reactance is unchanged, it is clear that loading the stage has caused the Q of the tank to be reduced.

When the Class C stage is unloaded, the shunt impedance of the parallel tuned circuit into which it works is very high, much too high to secure an appreciable amount of power from the amplifier. Loading the stage causes resistance to be reflected into the tank, which thereby reduces the value of its shunt impedance and allows the tube to develop more power. Since the process of loading the amplifier reduces both the Q and the shunt impedance of the tank, it might be suspected that some relation exists between the Q and the shunt impedance. Let us prove that this is true.

As stated in several previous lessons, the shunt impedance of a parallel tuned circuit is found by this equation:

$$\text{Shunt impedance} = \frac{1}{R} \times \frac{L}{C} \qquad (1)$$

Where: R is the resistance of the tuned circuit in ohms
L is the inductance of the tuned circuit in henries
C is the capacitance of the tuned circuit in farads.

Let us take the fraction L/C and multiply both numerator and denominator by 6.28 × F. Equation (1) then becomes:

$$\text{Shunt impedance} = \frac{1}{R} \times \frac{6.28 \times F \times L}{6.28 \times F \times C}$$

$$= \frac{1}{R} \times 6.28 \times F \times L \times \frac{1}{6.28 \times F \times C}$$

The expression 6.28 × F × L is equal to the inductive reactance X_L, and 1/ (6.28 × F × C) is equal to the capacitive reactance X_C. Therefore, this last equation may be written as follows:

$$\text{Shunt impedance} = \frac{1}{R} \times X_L \times X_C$$

At resonance, X_L is equal to X_C and the equation for shunt impedance is:

$$\text{Shunt impedance} = \frac{X^2}{R} \qquad (2)$$

Where: X is the reactance of either the coil or the condenser.

Equation (2) may be written in this manner:

$$\text{Shunt impedance} = \frac{X}{R} \times X$$

X divided by R, however, is the definition of the Q of the circuit, and so another formula for the shunt impedance is:

$$\text{Shunt impedance} = Q \times X \qquad (3)$$

From this last equation, it is apparent that the shunt impedance of the tank circuit depends directly on the Q of the tank, and from equation (1), it is seen that the shunt impedance also depends on the ratio of L to C. Reducing either the L/C ratio or the Q will cause the shunt impedance to be lowered.

There is no direct relation between the L/C ratio and the Q of a tank circuit. Increasing the L/C ratio increases the reactance of the circuit, but also tends to increase the resistance. The reactance, however, increases at a faster rate than the resistance, and so some increase in the circuit Q results from an increase in the L/C ratio.

8. **THE LOAD CIRCUIT.** As previously stated, the load of a Class C amplifier is a parallel tuned circuit. According to the theory of tuned circuits, the shunt impedance of a parallel tuned circuit is maximum at the resonant frequency. We will first consider that the amplifier is unloaded and that the tank circuit is not in tune. With the tank circuit untuned, it offers a compara-

tively small impedance to the plate current pulses, and the fundamental component of the plate current is unable to develop an appreciable voltage across the tank. The fraction of the cycle that plate current flows and the peak plate current reached depends, among other things, on the minimum plate voltage. Any circuit change which reduces the minimum plate voltage will cause the plate current to flow for a shorter time and will make the peak plate current less. This, in turn, will cause both the fundamental component and the DC component of the plate current pulses to be smaller.

Since the R.F. voltage developed across the tank is comparatively low when the tank circuit is out of tune, it is evident that the actual variation of the plate voltage above and below the DC value will likewise be small. Thus, the minimum plate voltage is rather high, and the DC plate current is large. By tuning the tank to resonance, the circuit conditions are changed considerably. At resonance, the tank offers a fairly high impedance to the plate current pulses, and the fundamental component of the plate current is able to build up a peak voltage across the tank, which is very nearly equal to the DC plate voltage. With this condition, the plate voltage varies from nearly twice the DC value to a very low value not far above zero. This very low minimum plate voltage naturally reduces the DC plate current.

The DC plate current will be minimum, when the minimum plate voltage is lowest. This will occur when the voltage built up across the tank is maximum, or when the shunt impedance of the tank is the greatest. A maximum shunt impedance is secured at the resonant frequency, and it is, therefore, apparent that the tank will be correctly tuned, when the DC plate current is minimum.

In the preceding discussion, it was assumed that the L/C ratio remained constant. If we now increase the L/C ratio and again tune the tank to resonance, we will find that the DC plate current is lower. By increasing the L/C ratio, we cause the tank circuit to have a greater shunt impedance at resonance, and, consequently, a lower DC plate current flows. Increasing the L/C ratio raises the plate efficiency for the following reason: A high L/C ratio causes the fraction of the cycle during which plate current flows, to be very small, and, consequently, the tube has a longer interval during each cycle for cooling. This causes the average power dissipated at the plate to be reduced considerably. Of course, a high L/C ratio also produces a lower DC plate current with the result that the power input is also reduced; however, the reduction of the plate loss is greater than that of the power input, and a larger percentage of the power input is delivered to the tank, which raises the plate efficiency.

With the amplifier unloaded, the DC plate current is very low, the power output is small, the efficiency is high, and the shunt impedance of the tank is very high. For example, suppose that an amplifier, when properly adjusted and loaded, delivers 100 watts to a dummy antenna with an efficiency of 50%. This means that 100 watts of power would be dissipated by the tube itself, and the power input would be 200 watts. If the DC plate voltage is 1000 volts, the DC plate current would have to be 200 ma. to furnish this power input. When the dummy antenna is disconnected, the DC plate current should fall to approximately 10% of its loaded value, or to 20 ma. The power input is therefore 20 watts when the stage is unloaded. The efficiency, however, will increase considerably, perhaps to 70%. Thus, 14 watts are furnished to the tank and 6 watts are dissipated at the plate.

It is evident that that part of the input power not lost at the plate must be dissipated in the tank circuit, if the dummy antenna is disconnected. Thus, in the preceding example, 14 watts must be so dissipated. Power can be dissipated only by a resistance, therefore, these 14 watts must be converted in heat by the oscillating tank current flowing through the resistance of the tank. It is probable that the R.F. resistance of the tank is about 10 ohms, and so to find the amount of tank current that must flow in order to dissipate 14 watts, we may use the I^2R law.

$$I^2 \times 10 = 14$$

$$I^2 = \frac{14}{10} = 1.4$$

$$I = \sqrt{1.4} = 1.18 \text{ amperes}$$

Thus, it is clear that the oscillating tank current will be fairly high compared to the DC plate current when the stage is unloaded.

When the antenna is coupled to the tank, a number of things happen. The oscillating tank current induces R.F. voltages into the antenna coil which cause an antenna current to flow. Tuning the antenna circuit to resonance with these induced voltages produces a large antenna current which is in phase with the voltage causing it. Through the phenomenon of reflected impedance, a pure resistance is coupled back into the tank circuit, and, since this coupled resistance is effectively in series with the tank, it causes the oscillating current to be reduced. The coupled resistance also produces other changes. From equation (1), it is seen that the shunt impedance of the tank is inversely proportional to the resistance contained in the tank. Thus, the act of coupling resistance into the tank serves to reduce the shunt impedance, and allows the tube to work into a load into which it can furnish more power. With a lower load impedance, the R.F. voltages developed across the tank are smaller and the minimum plate voltage increases. This, in turn, increases both the fundamental and the DC components of the plate current. The increase in the DC plate current signifies that the power input to the tube has increased, whereas the larger value of fundamental component indicates that more power is being supplied to the tank. Since the minimum plate voltage is raised, it is evident that the plate current reaches a higher peak and flows for a longer time, with the result that the plate losses mount and the efficiency is less.

The total resistance in the tank, when the stage is loaded is the tank's inherent resistance and the coupled resistance. The oscillating current in flowing through the tank's own resistance creates a power loss in the tank circuit, and the power developed by the tank current in flowing through the coupled resistance is actually the power which is transferred to the antenna circuit. It is, therefore, desirable that the actual resistance of the tank be low compared to the coupled resistance, and that the oscillating current not be excessive, if tank circuit losses are to be minimized.

The circuit adjustments for maximum power output are not consistent with those for high efficiency. A high load impedance, a large L/C ratio, a low DC plate current, and a small oscillating tank current are all synonymous with high efficiency. High efficiency is obtainable only at low power outputs. To obtain maximum power, or even an appreciable amount of power, it is necessary that the shunt impedance of the tank be reduced from its unloaded value, and there are two ways in which this may be done. First, it would be possible to make the shunt impedance of the tank low, even when unloaded, by using a small L/C ratio. However, the shunt impedance would be further reduced when the antenna was coupled to the tank, and it is more than probable that the efficiency when loaded would be very small.

To obtain the desired power output with a reasonable amount of efficiency, it is necessary to use a fairly large L/C ratio. If, however, this ratio is too great, it may happen that the shunt impedance of the tank is still too large, even when the stage is loaded, to produce the necessary power output. If this is the case, some efficiency will have to be sacrificed, and the L/C ratio made correspondingly lower.

The actual value of the shunt impedance bears no simple relation to the actual plate resistance of the tube. The plate resistance of a tube is, as we know, a small change in plate voltage divided by the small change produced in the plate current. In Class A amplifiers, where the operating path is confined to the linear portion of the E_p-I_p curve, the plate resistance is essentially constant throughout the operating range. A Class B amplifier operates on both the straight and the curved portion of the characteristic with the result that the plate resistance may vary considerably. Finally, the operation of a Class C amplifier is such that no plate current flows for the major part of the cycle. Obviously, during this time, a small change in plate voltage would produce no change in the plate current and the plate resistance may be said to be infinite. As plate current starts to flow, the plate resistance decreases from a very high value to a minimum which it maintains throughout the straight part of the characteristic, and then increases as the upper curved part of the characteristic is reached. The average value of the plate resistance throughout a cycle depends upon how long plate current flows.

It would seem from the previous discussion that the L/C ratio of the tank should be as high as will allow the required power output to be developed in the antenna circuit. There are, however, other factors to be taken into account. The greater the L/C ratio, the smaller the oscillating tank current will be when the stage is loaded. This would seem to be desirable since it reduces the tank circuit losses; however, it must be remembered that energy is supplied to the tank in the form of pulses of short duration. In the interval between these pulses, the total energy in the tank decreases, due to dissipation of power both in the tank and in the antenna. Unless the energy taken from the tank during each cycle is small compared to the total energy contained in the tank, the oscillating current will not be of pure sine wave form, and harmonic frequencies will be transferred to the antenna. The minimizing of harmonics is not only desirable, it is essential to conform with the rulings of the Federal Communications Commission.

There are harmonic suppressors available which may be connected between the tank and the antenna which reduce the harmonics and allow only the fundamental to reach the antenna, but it is also necessary that the production of harmonics in the tank circuit be minimized as much as possible. If the harmonics in the tank circuit are to be low, the oscillating current must be of practically pure sine wave form, and the energy stored in the tank must be appreciably larger than the energy taken out each cycle. The actual energy stored in the tank is directly proportional to the value of the oscillating tank current, and it is, therefore, apparent that the oscillating current must not fall to too low a value when the stage is loaded. This is even more important for phone that it is for "CW" transmissions.

It has been experimentally determined that the energy stored in the tank must be at least twice as great as the energy taken out per cycle to reduce the harmonics to a value where they are not troublesome. To make this ratio 2, it has been discovered that the Q of the loaded tank should be at least 12. This value is for phone operation; a somewhat smaller Q is permissible for "CW" transmissions.

As previously stated, there is no simple relation between the L/C ratio and the Q of the tank circuit. Increasing the L/C ratio will increase the Q of the unloaded tank, but will probably cause the Q of the loaded tank circuit to be less. Thus, decreasing the L/C ratio will cause the tank current to be larger when the stage is loaded, will increase the Q, and reduce the harmonics.

To determine what values of inductance and capacity should be used to give a fair degree of efficiency, and to keep harmonic production to a minimum, the following formulas may be employed.

$$L = \frac{E^2}{16 \times \pi^2 \times F \times P} \qquad (4)$$

$$C = \frac{4 \times P}{E^2 \times F} \qquad (5)$$

Where:
E is the peak voltage developed across the tank
F is the frequency in cycles
P is the power output in watts
π is 3.1416

The peak voltage across the tank circuit is not known, but a fair approximation may be made by assuming that it is .8 of the applied DC voltage. These formulas are not critical; they merely serve as a starting point for determining what values of inductance and capacity will give the best operating conditions.

As a final test for the correct L/C ratio, the two following rules should be applied.

1. If the value of the DC plate current, when the stage is unloaded, is more than 15% of the DC plate current when the stage is properly loaded, the L/C ratio is probably too low. This assumes that the tank circuit is well designed and has a minimum of losses.
2. If, while tuning the tank circuit, the minimum plate current does not occur at the same point as the maximum antenna current, the L/C ratio is too high.

Fundamentally, there are three types of plate tank circuits. These three types are illustrated in Fig. 5. The one shown at A in this figure is an unsplit tank and grid neutralization is employed. The other two use plate neutralization and require a split tank. It may be proved that the neutralization cannot be perfect unless the tank circuit is tapped at its exact center. It is possible to achieve an imperfect sort of neutralization by connecting the neutralization tap near the bottom of the tank, but the voltages across the two parts of the tank will not be exactly 180° out of phase, and while the stage may not self-oscillate, it will be sufficiently regenerative to impair the linearity of the modulation.

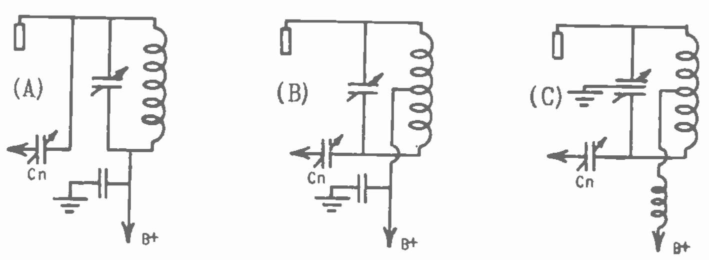


Fig. 5 The three fundamental types of plate tanks.

Assuming that the tank circuit is split at its exact center, it is evident that the required neutralizing capacity will be equal to the plate-to-grid capacity of the tube. This is obvious, since the R.F. voltage which could cause feedback is equal to the neutralizing voltage.

It is possible to tap the plate tank at its exact center, either by tapping the coil, or by using a split stator tuning condenser. In either case, it is apparent that the impedance connected between the plate of the tube and the point where the tap is connected is less than the total shunt impedance of the tank circuit. Therefore, the tube is working into only a part of the tank circuit. The shunt impedance between two points on a parallel tank circuit is directly proportional to the square of the turns connected between those two points. Therefore, the shunt impedance measured across one-half of the tank circuit is equal to one-fourth of the total shunt impedance.

In order that the tube shall work into the correct load impedance, it is necessary that the total shunt impedance of the split tank be four times the desired load impedance. Thus, when plate neutralization is employed, the shunt impedance of the split tank must be four times that of the unsplit tank used with grid neutralization. To make the shunt impedance of the split tank equal to four times that of the unsplit type, requires that the reactances of the coil and the condenser be twice as large. For a given power output and plate voltage, the peak R.F. voltage built up across the split tank will be twice as great as across the unsplit type. The dielectric losses of the tank circuit vary directly as the square of the R.F. voltage across the tank; thus, it is seen that the dielectric losses of the split tank will be four times as large as those of the unsplit tank. On the other hand, the resistance losses of the unsplit tank will be four times as large as those of the split tank, because the oscillating tank current is twice as large.

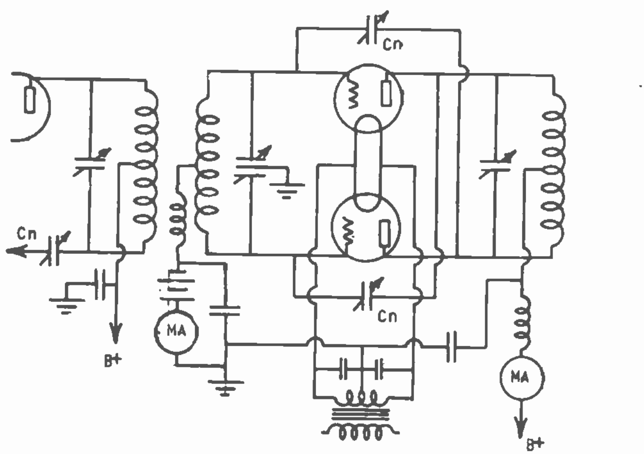


Fig.6 A push-pull Class C amplifier.

A push-pull Class C amplifier is illustrated in Fig. 6. It also requires a split tank circuit. In this case, the tank coil is tapped, but a split stator tuning condenser would do just as well. Perhaps the major point of interest in the push-pull circuit is the method by which neutralization is obtained. The voltage built up across the lower half of the tuning coil serves as the neutralizing voltage for the top tube, and that across the upper half of the coil serves the same purpose for the bottom tube. In neutralizing a push-pull stage, both neutralizing condensers are advanced in small steps until there is no indication of R.F. current in the tank circuit.

It is desirable that the capacity of each be approximately equal after the final adjustment is made.

The push-pull Class C amplifier has another advantage in that the tank circuit is furnished a pulse of energy during every alternation, instead of just once during the cycle. This tends to keep the energy stored in the tank circuit at a more constant value, and thereby reduces even harmonics. For this reason, the oscillating tank current need not be as large and a smaller Q is permissible. In general, the Q of a push-pull stage need be only 60% of that of a single-ended stage.

In order to excite the grids of the push-pull amplifier, two R.F. grid-exciting voltages differing in phase by 180° must be provided. The simplest method of accomplishing this is to use inductive or link coupling between the buffer and the push-pull stage. This is the method employed in the diagram shown in Fig. 6. It is necessary to split the grid-tank circuit, as well as the plate tank, and the bias voltage is fed into the center of the grid tank. This is by far the most common means. There is, however, another method which is occasionally used. It is illustrated in Fig. 7. The preceding buffer tank circuit is split into two equal parts, and the R.F. voltage across each half is capacitively coupled to the grids of the push-pull amplifier.

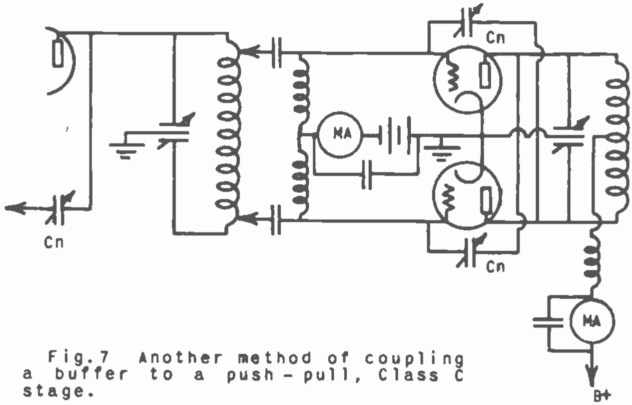


Fig.7 Another method of coupling a buffer to a push-pull, Class C stage.

The larger tank circuits use fixed rather than variable condensers; tuning is accomplished by varying the inductance of the tank. Coarse adjustments are made by changing the position of variable taps which determine how much of the coil is in use. Fine adjustment of the tuning is obtained in the following manner. Mounted within the tank coil so that it may be rotated is a disc of sheet copper. When the plane of this disc is parallel to the magnetic lines of force, it has but little effect on the inductance of the tank. By rotating the disc so that it is at right angles to the magnetic lines of force, voltages are induced into the disc and eddy currents flow through it. Since the disc has a small amount of inductance, it will reflect a small capacitive reactance into the tank circuit and thereby provide a means of tuning the tank circuit. The amount of capacitive reactance reflected into the tank may be varied by changing the position of the disc with respect to the tank coil. This method is used because variable tuning condensers for very large voltages are extremely expensive and rather bulky.

9. THE PLATE AND GRID CURRENT PULSES. By means of a complex mathematical procedure, it is possible to predict the operation of a Class C amplifier having a given set of applied voltages. This is accomplished with a family of plate voltage-plate current curves in much the same manner as was done with the Class A amplifier. However, the fact that the plate current flows in pulses, and the tube operates over its entire dynamic characteristic curve, including both curved portions, makes this computation so involved that it is purely of theoretical interest. Since this procedure requires a knowledge of trigonometry and integral calculus, it will not be discussed, although a few of the conclusions reached are of sufficient interest to be included.

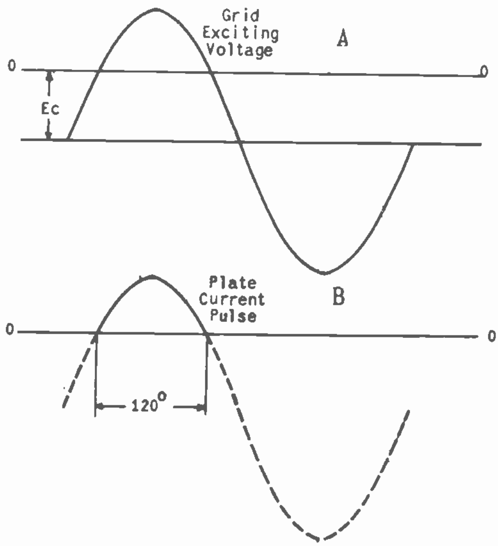


Fig.8 Showing that the plate-current pulse would be the upper part of a sine wave, if the dynamic Eg-Ip characteristic were linear.

The time, in electrical degrees, during which plate current flows is called the "operating angle", Thus, if the amplifier is so adjusted that plate current flows for one-third of the cycle, the operating angle is said to be 120° (⅓ × 360°). Similarly a plate current that flows for one-sixth of the cycle would have an operating angle of 60° (⅙ × 360°). There are five different factors which determine what the operating angle will be. They are the DC plate voltage, the grid bias, the amplification factor of the tube, the peak value of the grid exciting voltage, and the peak value of the voltage developed across the tank.

If the dynamic E_g-I_p characteristic were a straight line, the plate current would vary directly with the grid exciting voltage. The grid exciting voltage is a pure sine wave, and the plate current pulse, even though it flows for less than one-half cycle, would be the upper portion of a sine wave. This is illustrated in Fig. 3. Curve A represents the grid exciting voltage, and curve B the plate current that would flow, assuming that the operating angle is 120°. In curve B, a complete sine wave is drawn to show that the plate current pulse is part of a sine wave.

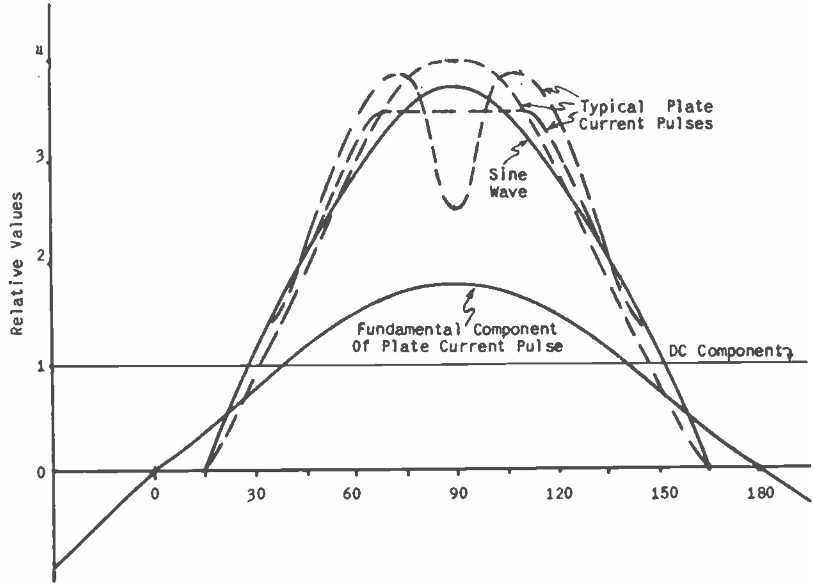


Fig.9 Several typical plate current pulses with the upper part of a sine wave for comparison. The DC, and fundamental components are also shown.

Actually, however, the dynamic E_g-I_p characteristic is not a straight line, but is considerably curved. Throughout the extreme curved portions, the plate current is not directly proportional to the grid exciting voltage, but is more nearly proportional to the square of the grid voltage. This causes the shape of the plate current pulse to be somewhat different from sine wave form. Typical plate current pulses as actually measured on various transmitting tubes are shown as dashed lines in Fig. 9. The solid line curve is the upper portion of a sine wave and is included for comparison. In this case the operating angle is 150°.

Although each of the pulses is slightly different, it is seen that each have approximately the same peak value and, since they all flow for the same length of time, it is evident that each will have the same DC component. This DC component is shown in the figure as a horizontal line. The vertical axis of this figure is not calibrated in milliamperes, but instead is marked off in relative values. By so doing, it is possible to see that the peak value of the pulses is approximately 3.7 times as great as the DC component. This, of course, will only be true when the operating angle is 150°.

By reducing the operating angle to 60°, it has been determined that the ratio of the peak plate current to the DC component is approximately 9. (The operating angle may, of course, be reduced by increasing the grid bias or by using a greater L/C ratio.) Since the operating angles of practically all Class C amplifiers will fall between these two limits, it is apparent that the peak value of the plate current may range from 3.7 to 9 times the DC plate current.

Every vacuum tube has a maximum plate current which may be drawn from it, and this peak plate current depends on the electron emitting ability of the filament. It is not wise to attempt to draw the maximum allowable plate current, for this greatly shortens the tube's life. By reading the DC plate current on a meter, and by knowing that the peak of the plate current pulse will rarely be more than 9 times as great, it is possible to determine whether the peak current drawn by the tube is within the allowable limit.

In Fig. 9 there is also illustrated the fundamental component of these plate current pulses. It is seen to have a peak value of about 1.7 times the DC component. When the operating angle is reduced to 60°, the ratio of the peak value of the fundamental to the DC component is about 1.95.

Since grid current can flow only when the grid is positive, it is evident that the grid current flows for a shorter length of time than the plate current. The operating angle of the grid current depends on the grid bias and the grid excitation voltage. On the average, it will range from 40° to 130°, but will always be less than the operating angle of the plate current. The dynamic E_g-I_g characteristic is much less linear than the E_g-I_p characteristic, with the result that the grid current pulse departs considerably from sine wave form. When the operating angle is 130°, the ratio of the peak grid current to the DC grid current is approximately 5.4, whereas, an operating angle of 40° causes this ratio to increase to about 16.

10. THE PLATE-MODULATED CLASS C AMPLIFIER. The plate-modulated Class C amplifier presents special problems not found in the Class C amplifier used for "CW". In the first place, the applied DC voltage must be somewhat lower. This is due to the fact that the actual voltage applied to the plate doubles its value during the crest of the audio cycle when 100% modulation is used. For example, a manufacturer may state that the maximum DC plate voltage is 1250 volts for a particular tube, when it is to be used in a "CW" Class C amplifier. During operation, the voltage on the plate will probably vary from twice this value to a very low minimum at an R.F. rate, due to the R.F. voltages set up across the tank circuit. Thus, the actual maximum plate voltage may be as high as 2300 or 2400 volts. This, however, has been taken into account by the tube manufacturer, and the maximum DC voltage rating which he assigns to the tube is such that the tube will be able to withstand this peak voltage.

The maximum DC plate voltage rating, when the tube is used in a plate-modulated Class C amplifier, is ordinarily two-thirds of the unmodulated value, and in some cases, it may be slightly higher. When the stage is modulated, the plate voltage on the tube varies at both an A.F. and an R.F. rate. The A.F. variation, due to the modulation, causes the applied voltage to change from zero to twice its average value. In addition, the R.F. voltages developed across the tank also double their value at the crest of the modulation cycle. Let us suppose that the applied voltage is 1000 volts, and that the peak voltage set up across the tank circuit is .8 of this value, or 800 volts. In this case, the peak plate voltage would be 1800 volts. At the crest of the modulation cycle (with 100% modulation), the applied voltage increases to 2000 volts. Also, the peak R.F. voltage across the tank doubles its value and becomes 1600 volts. Therefore, the peak plate voltage at this time is 3600 volts. This plate voltage variation is shown in Fig. 10. The solid line sine wave represents the audio voltage which is superimposed on the DC plate voltage. Each horizontal line represents a different voltage, the value of which is found on the vertical axis.

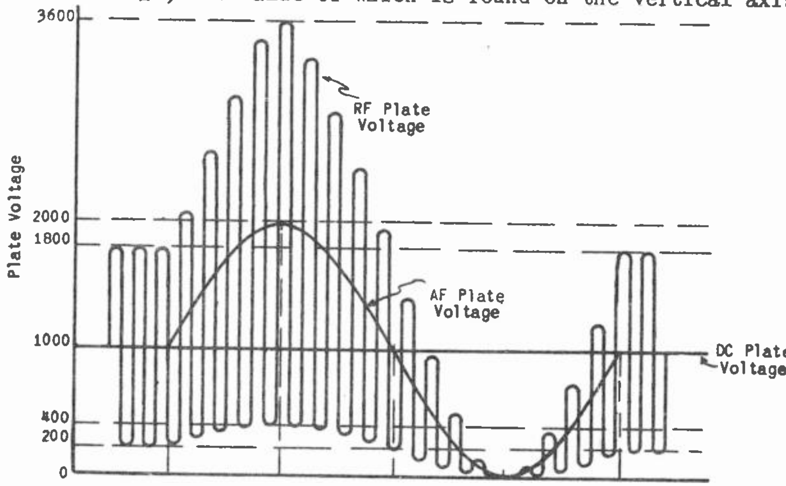


Fig. 10 The variation of the plate voltage of a Class C amplifier during 100% modulation.

From point A to point B, the stage is unmodulated; the applied plate voltage is 1000 volts, and the R.F. voltage variation is from 200 to 1800 volts. At point C, the modulation is at its peak; the applied voltage at this time is 2000 volts, and the R.F. voltage variation is from 400 to 3600 volts. It is, therefore, evident that the peak voltage on the plate of a tube when it is 100% modulated may be nearly four times the actual DC plate voltage, and therein lies the necessity of reducing the DC plate voltage below that value which may be used with a CW Class C amplifier. The maximum voltage ratings of the manufacturer are conservative enough that the tube will be safe if the applied DC plate voltage is reduced to about two-thirds of the value used for CW.

If the modulation is to be linear, the peak value of the fundamental plate current component must also double its value at the modulation crest. The value of the load impedance is the same whether the stage is unmodulated or modulated, and when the fundamental component of the plate current doubles its value, it produces twice as large a voltage drop across the tank circuit. With the voltage drop across the tank twice as great, it is evident that the tank current will also become twice its unmodulated value, and the antenna current will likewise double. Although the peak value of the fundamental component must increase two times at the modulation crest, it is not necessary that the peaks of the plate current pulses increase this much. Let us see why this is so.

When the stage is unmodulated, the grid bias is equal to twice cut-off, the operating angle is small, and the ratio of the peak plate current to the peak of the fundamental component is high. At the crest of the audio cycle, the applied plate voltage doubles, and the grid bias remains practically constant. Therefore, the operating angle increases or the plate current flows for a longer part of the cycle. This causes the ratio of the peak plate current to the peak of the fundamental component to be somewhat less, and it is, therefore, not necessary that the peak plate current double its value. If the plate current flowed for the same length of time at the crest of the audio cycle as it did when unmodulated, it would be necessary for the peak plate current to double, but due to the fact that the operating angle increases, a fundamental component of double peak value may be obtained when the peak current increases to about 1.6 times its unmodulated value.

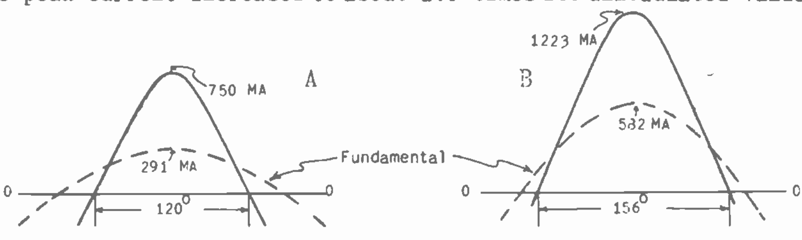


Fig. 11 A. The plate current pulse and fundamental component with no modulation.
B. The plate current pulse and fundamental component at the peak of 100% modulation.

This fact is clearly illustrated in Fig. 11. At A is shown a plate current pulse that flows for 120°. It is assumed to have a peak value of 750 ma., and represents the current that flows when the stage is unmodulated. With this operating angle, the ratio of the peak to the fundamental is 2.57, and the peak value of the fundamental is, therefore, 291 ma. The fundamental component is shown as a dotted line. At B is shown the plate current that flows at the crest of the modulation cycle; it has a peak value of 1223 ma. In addition, the operating angle has increased to 156°, and the ratio of the peak to the fundamental is now 2.1. This causes the fundamental, shown in dotted lines to have a peak value of 582 ma., or just double its former value. In this case, the peak value of the plate current pulse increased 1.63 times (from 750 ma. to 1233 ma.).

For linear modulation, the relation between the plate voltage and the oscillating tank current must be such that the tank current doubles its value at the time that the applied plate voltage doubles. The tank current, however, is directly proportional to the fundamental component of the plate current and it is, therefore, necessary that the fundamental current also double at this time. This fact is usually stated by saying that the power output of a Class C amplifier should be directly proportional to the square of the plate voltage. It is obvious that doubling both the plate voltage and the fundamental plate current will cause the power output to increase four times, and this is also one of the requirements of linear modulation.

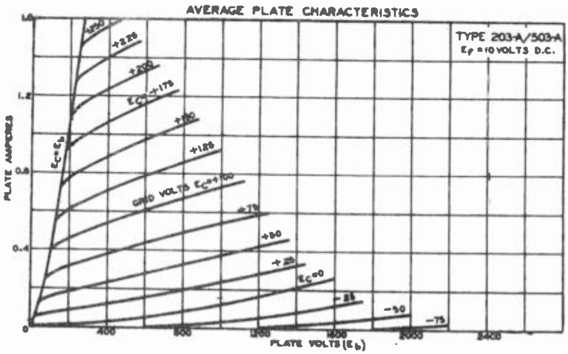


Fig.12 A family of Ep-Ip curves to demonstrate the necessity of a large excitation voltage.

As previously stated, it is not necessary that the peak plate current increase two times in order that the fundamental current do so. The peak plate current flows at the time of minimum plate voltage. By referring to Fig. 10, it is seen that in this particular case the minimum plate voltage increases from 200 to 400 volts. At first thought, it would appear that doubling the minimum plate voltage would cause the peak plate current to also double. This, however, is not true. An inspection of Fig. 12 will make this clear. This figure shows a family of plate voltage-plate current curves for a type 203-A transmitting tube. The grid excitation voltage does not change during modulation, and if the grid bias remains constant, it is evident that the maximum positive grid voltage will also remain at the same value. First, let us assume that the maximum positive grid voltage is 125 volts. When the minimum plate voltage is 200 volts, the peak plate current will be .6 ampere or 600 ma. Now when the minimum plate voltage increases to 400 volts, the peak plate current increases to only 700 ma. This is an increase of only 1.16 times.

We have, however, stated that the peak plate current must increase 1.63 times, in order that the fundamental current may double. Of course, the actual amount that the peak plate current must increase for the fundamental current to double depends on the operating angle when the stage is unmodulated. In this case, however, it is 1.63 times, and we found that increasing the minimum plate voltage from 200 to 400 volts produced an increase of 1.16 times, if the maximum positive grid voltage was 125 volts. Thus, the maximum plate current does not increase enough to allow the fundamental plate current to double, and, as a result, the modulation will not be linear. Such a condition will cause the tank current and antenna current to have a wave form similar to that shown in Fig. 13. Notice that the positive peaks of the modulation envelope are clipped; that is, the peaks of the R.F. cycles do not rise to twice their unmodulated value. This is most undesirable, since it constitutes distortion of the impressed audio wave. Let us see what is to be done about this condition.

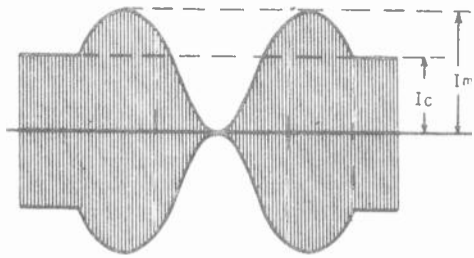


Fig.13 The type of modulated wave produced when the excitation of the Class C stage is insufficient.

Suppose that the excitation voltage of the Class C stage is increased enough so that, with the same grid bias, the maximum positive grid voltage is 200 volts. When this is true, the maximum positive grid potential is equal to the minimum plate voltage when the stage is unmodulated; a condition not entirely desirable, since it tends to cause the plate current pulse to have a dip in it, and therefore causes R.F. harmonics. However, if the oscillating tank current is sufficiently large, it will be able to smooth out the pulses into practically pure sine waves. With a maximum positive grid voltage of 200 volts, the peak plate current that flows when the minimum plate voltage is 200 volts is 1100 ma. Then at the crest of the modulation cycle, when the minimum plate voltage increases to 400 volts, the peak plate current becomes 1250 ma. Therefore, the peak plate current increases 1.23 times.

This is still not a sufficient increase to cause the fundamental plate current to double, but it is a greater increase than was obtained when the maximum grid voltage was 125 volts. Thus, is seen the necessity of using a large grid excitation voltage. In fact, if the grid excitation voltage is not large, the relation between the plate voltage and the fundamental plate current will not be such that doubling the voltage will double the fundamental current.

We have, however, still not obtained a sufficient increase in the peak plate current to cause the fundamental current to double its value. The peak plate current is 1100 ma. when the maximum positive grid voltage is 200 volts and the minimum plate potential is also 200 volts. In order for the fundamental plate current to double, the peak plate current must increase 1.63 times or to 1773 ma. Although this value of plate current is not shown on the graph of Fig. 11, it may be calculated that the positive grid voltage necessary to make the plate current 1773 ma., when the minimum plate voltage is 400 volts, is about 260 volts. Thus, the maximum grid voltage must increase from 200 to 260 volts from the unmodulated condition to the crest of the modulation cycle.

The grid excitation voltage, however, remains constant, and the only way in which the maximum positive grid voltage may increase at the modulation peak is for the grid bias to be less at this time. This is automatically taken care of if part of the grid bias is secured by means of a grid leak. The minimum plate voltage increases from 200 to 400 volts, and the peak plate current increases; the peak grid current, however, decreases. Raising the minimum plate voltage will always reduce the grid current, because more of the emitted electrons are drawn to the plate and there are not as many available to be attracted to the grid. Since the peak grid current decreases, the DC component of the grid current will also decrease, and, as a result, the grid bias voltage produced across the grid leak will be reduced.

Had we assumed that the plate current flowed for a shorter time than 120°, when the stage was unmodulated, we would have found that the peak plate current would not have needed to increase 1.63 times in order for the fundamental plate current to double. In fact, it is possible that the peak plate current would need to increase only 1.2 times, if the unmodulated operating angle were small enough. Using a large value of excitation voltage, it is probable that this amount of peak plate current increase could be caused, even if a fixed grid bias were used and the maximum positive grid potential did not change. In any event, it is necessary that the unmodulated grid bias be at least twice cut-off so that the operating angle will be small when the stage is unmodulated.

It should now be evident that there are many factors which can cause or prevent linear modulation. To avoid extensive calculation and yet be able to determine whether the modulation will be linear or not, it is often advisable to perform the following experiment: The Class C amplifier is first adjusted for the proper operating conditions with no modulation. A thermocouple ammeter is placed in the tank circuit to read the tank current, and the value of tank current and DC plate current are recorded. Then, the applied plate voltage is increased to twice its unmodulated value, and the tank current and DC plate current are again read. Naturally, extreme caution must be observed in doing this experiment, because the plate losses on the tube, when the plate voltage is doubled, will be excessive. The high plate voltage should be applied only long enough to take the readings, and, if the plate of the tube becomes excessively red, the plate voltage must be immediately disconnected..

From these two sets of readings, the tank current, and the DC plate current should be plotted against the plate voltage. If the tank current doubled its value, when the plate voltage was doubled, the modulation will be linear. It is also desirable that the DC plate current approximately double in value when the plate voltage is doubled. The reason for this is as follows: The load into which the modulator stage is working is equal to the DC plate resistance of the Class C tube. This is applied plate voltage divided by the DC plate current. Unless the relation between the applied plate voltage and the DC plate current is fairly linear, the DC plate resistance of the tube will not be the same at the crest of the modulation cycle as it is with no modulation, and the load on the modulator will vary, which may cause the modulator to produce some distortion.

Assuming that the DC plate current does double its value at the crest of the modulation cycle, it will, of course, fall to zero as the applied plate voltage drops to zero, and the average value of the DC plate current (the value which the DC plate milliammeter will read) will be equal to the DC plate current when the stage is unmodulated. Therefore, one of the tests of the linearity of the modulation is that the DC plate current not change during modulation.

	Carrier	Crest
Applied DC plate voltage	1000 volts	2000 volts
DC plate current	175 ma.	375 ma.
Power input	175 watts	750 watts
Maximum plate current	805 ma.	1310 ma.
Operating angle of plate current	120°	156°
Fundamental plate current (peak value)	313 ma.	626 ma.
Voltage across tank (peak value)	840 volts	1680 volts
Minimum plate voltage	160 volts	320 volts
Tank Current	2.9 amperes	5.8 amperes
Power output	131 watts	524 watts
Efficiency	75%	70%
Grid Bias	-200 volts	-140 volts
Excitation voltage (peak value)	360 volts	360 volts
Maximum positive grid voltage	160 volts	220 volts
DC grid current	36.6 ma.	25.6 ma.
Maximum grid current	220 ma.	133 ma.
Operating angle of grid current	113°	134°
Grid leak resistor	5460 ohms	5460 ohms
Grid driving power	11.9 watts	8.3 watts
Plate loss	44 watts	225 watts

At the peak of the audio cycle when the DC plate current has momentarily doubled, the power input will be four times as large as when the stage is unmodulated. Also, the power output will be four times as great. The average increase of the power output, however, will be one and one-half times, and, since the average efficiency is constant, the average power input must also increase one and one-half times during 100% modulation. The average power input, as we know, includes both the DC power input and the audio power input. The DC power input, which is the DC plate current times the voltage of the power supply, does not change; therefore, the 50% increase in the average power input must come from the modulator.

In addition, the plate dissipation increases four times at the peak of the modulation cycle, and the average power lost at the plate is one and one-half times larger. Care must be taken that the tube is not dissipating the maximum power allowable before the modulation is applied, because the plate losses will increase 50% with

100% modulation.

As stated at the beginning of this section, it is possible to calculate the performance of a Class C amplifier by means of a complicated mathematical procedure. Such a procedure was carried out for a type 203-A transmitting tube, and the following results were obtained. The two columns of the preceding table represent conditions existing at no modulation and at the crest of the modulation cycle respectively.

Note that the efficiency at the crest of the audio cycle is slightly less than the efficiency when unmodulated. This is due to the fact that the power input at the crest was somewhat greater than four times that of the carrier. The trouble was that the DC plate current increased slightly more than twice, and, as a result, the plate losses went up over five times instead of four. However, the results represent average operating conditions.

11. ADJUSTING THE CLASS C AMPLIFIER. In Fig. 14 there is illustrated a small transmitter. It includes the crystal oscillator, the buffer, the Class C amplifier, and the modulator stage. To give the student a comprehensive idea of the proper procedure for setting this transmitter into operation, each stage will be adjusted separately.

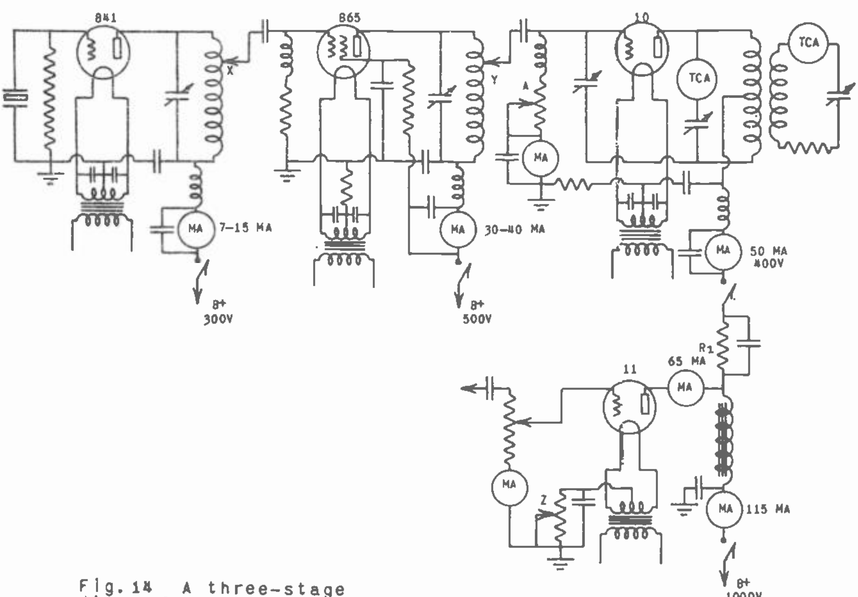


Fig. 14 A three-stage radio telephone transmitter.

The crystal oscillator stage uses a type 841; the buffer amplifier a type 865 screen-grid tube; the Class C stage a type 10 triode; and the modulator a type 211 tube. First, make sure that all plate voltage switches are open, and then light the filaments of all the tubes. Disconnect the excitation taps X and Y, and apply plate voltage to the oscillator. Tune the oscillator until the dip occurs in the plate current, and then to make the oscillator more stable, its plate tank circuit should be tuned to a slightly higher frequency. The oscillator should not be operated at the exact minimum of the plate current, but the point of operation cannot be far from this minimum or the oscillator will not function. With a plate voltage of approximately 300 volts on the oscillator, the plate current will range from 7 to 15 ma. depending on the L/C ratio employed.

The excitation tap X should now be connected to the plate tank of the oscillator. (IT IS, OF COURSE, UNDERSTOOD THAT ALL PLATE VOLTAGES ARE TO BE DISCONNECTED WHEN ANY CHANGE IS MADE.) Since the 365 is fairly easy to drive, it is probable that the tap X will not need to be connected at the extreme top end of the oscillator tank. The plate voltage of the buffer stage is applied, and this stage is tuned to resonance, as indicated by minimum plate current. Grid bias for the buffer is secured by means of a grid leak and a filament resistor. The excitation tap Y is connected to the plate tank of the buffer, and the grid excitation of the buffer, and the L/C ratio of its tank are changed until the grid current of the Class C stage is about 15 ma. When properly adjusted, the plate current of the buffer will be about 35 ma.

It is now wise to use a wave meter to make sure that the buffer is tuned to the same frequency as the oscillator, and is not acting as a doubler stage. The main objective is to supply sufficient grid excitation to the Class C stage without causing either the oscillator or buffer to be overtaxed.

Tune the tank circuit of the Class C stage to resonance; this point will be indicated by a dip in the grid current, if the stage is not neutralized. Now proceed to neutralize this stage, using any of the methods discussed in Lesson 3 of this Unit. Perhaps the simplest method would be to rotate the neutralizing condenser until the grid current does not change as the plate tank is tuned through resonance.

Turn the gain control of the modulator to such a position that no A.F. signal is being fed into the grid of this stage, and then apply plate voltage to the modulator. Vary the grid bias on the modulator by changing the position of the shorting tap Z, until the modulator is drawing exactly 65 ma. With this amount of plate current and a 1000 volts plate voltage, the load on the modulator should be 3000 ohms. This load is, of course, the DC plate resistance of the Class C tube; therefore, the plate voltage and plate current of the Class C stage must be so adjusted that it presents this impedance to the modulator.

The type 10 tube has a plate dissipation rating of 15 watts, and assuming 50% efficiency, it would be possible to use an input power of 30 watts. However, since the tube is to be modulated, it cannot be worked up to its rated capacity, for it would then heat excessively during the peaks of modulation. In this case, we will use A DC power input of 20 watts. To modulate this 20 watts of DC power input requires that the modulator deliver 10 watts of undistorted power. According to the tube manual, the type 11 tube can deliver 12 watts, and therefore, it will be able to fnlly modulate the Class C stage.

The next step is to determine what values of plate voltage and plate current the Class C stage should have so that its power input will be 20 watts, and its DC plate resistance 8000 ohms. These values may be found by using the formulas given in Section 5 Lesson 5 of this Unit. They are:

$$I_B = \sqrt{\frac{P_i}{R_L}} = \sqrt{\frac{20}{8000}} = \sqrt{.0025} = .05 \text{ ampere or } 50 \text{ ma.}$$

$$E_B = \sqrt{P_i \times R_L} = \sqrt{20 \times 8000} = \sqrt{160{,}000} = 400 \text{ volts.}$$

Therefore, the plate voltage of the Class C stage should be 400 volts, and it should be loaded until it is drawing exactly 50 ma. The plate voltage of the modulator is 1000 volts, and this voltage is reduced for application to the Class C stage by means of the resistor R_1. The required reduction in voltage is 600 volts, and the current flowing through R_1 is 50 ma.; therefore, R_1 must have a value of 12,000 ohms, and must be able to dissipate the 30 watts of power which this 50 ma. at 600 volts will produce.

With the tank circuit of the Class C stage tuned to resonance, it is safe to apply the plate voltage to this stage. (Assuming, of course, that the neutralization process has been carried to completion.) After applying the plate voltage, make sure that the Class C stage is properly tuned by varying the capacity of the tuning condenser until minimum plate current flows, and then check the frequency of the oscillating current with the wave meter. If the L/C ratio of the tank circuit is rather high, the minimum plate current should be quite low. This indicates that the efficiency is high. If this is not the case, the L/C ratio of the tank circuit is too low, and must be increased. However, whenever a change in the L/C ratio is made, the stage must be reneutralized.

The grid bias should be at least twice cut-off, and the grid excitation should be great enough so that the DC grid current is from 15% to 25% of the plate current of the stage when loaded.

The dummy antenna is coupled to the tank circuit of the Class C stage and is tuned to resonance as indicated by maximum antenna current. The coupling between the antenna and the tank is increased until the Class C stage draws exactly 50 ma. It will be necessary to retune the tank circuit whenever the antenna coupling is changed. If it is not possible to cause the plate current to increase to 50 ma., even when the coupling between the antenna and the tank is at the critical value, the stage is too efficient, and some efficiency must be sacrificed by reducing the grid bias.

When the coupling between the antenna and the tank is such that the plate current is exactly 50 ma., the amount of power in the antenna should be calculated. This is done by squaring the antenna current and multiplying it by the antenna resistance. If the power in the antenna is 10 watts or more, the efficiency is at least 50%. If less than 10 watts are in the antenna circuit, the efficiency is too low and must be increased. This is done by increasing the grid bias. With each increase in grid bias, the plate current is brought back to normal by increasing the antenna coupling. This process should be carried on until it is just possible to cause the tube to draw 50 ma. when the coupling is slightly less than the critical value. At this point, the desired power output should be in the antenna and the efficiency should be 50% or better.

Do not attempt to secure unusually high efficiency, because it is only possible by using very large values of grid bias, and grid excitation, and such an adjustment causes the grid losses to be excessive.

As a final check, tune the tank circuit through resonance and determine whether minimum plate current occurs at the same point as maximum antenna current. If it does not, the L/C ratio of the tank circuit is too high and excessive R.F. harmonics will be produced. Should it happen that the L/C ratio is too high, it must be reduced, the stage must be reneutralized, and the adjustment procedure as outlined above must be repeated.

To check the stability of the transmitter, open and close the oscillator plate voltage switch. When the switch is open, the grid current of the Class C stage and the antenna current should fall to zero. If they do not, the Class C stage is imperfectly neutralized and is self-oscillating. The plate currents of the buffer and Class C stage will not drop to zero when the oscillator plate-voltage switch is opened, because these stages use grid-leak bias. When the switch is open, the excitation voltage is removed, and no grid current flows; therefore, the grid bias on both stages is limited to that provided by the filament bias resistors, which is ordinarily just enough to prevent the plate currents from becoming excessive when the grid excitation is removed. If the transmitter is stable, all of the meters will return to the same position each time the oscillator plate-voltage switch is closed.

As previously stated, the modulation will not be linear unless the grid excitation is fairly great. There is no simple direct method of checking the grid excitation, but it is assumed to be correct if the DC grid current is from 15% to 25% of the plate current.

After all adjustments on the Class C stage have been made, modulation is applied by connecting the output of an audio oscillator to the grid circuit of the modulator. The gain control of the modulator is then adjusted until the power in the antenna circuit has increased one and one-half times, or until the antenna current increases 22.5%. This indicates that the Class C stage is 100% modulated. The modulator is operating under Class A conditions and should not draw any grid current; therefore, a low range milliammeter is connected in the grid circuit of the modulator to make sure that the signal voltage is not great enough to drive the grid positive and cause the flow of grid current. The plate current of the Class C stage should not change during modulation.

12. CHECKING THE ADJUSTMENT OF THE TRANSMITTER WITH A CATHODE RAY OSCILLOSCOPE. It should be realized that there are five factors which together determine the correct adjustment of a plate-modulated transmitter. These are the grid bias, the grid excitation, the L/C ratio, the antenna coupling, and the load on the modulator. Unless the value of each one of these factors in relation to the others is correct, the best operating conditions will not be obtained.

To determine the power input and power output is relatively simple, and when these values are known, the plate circuit efficiency, and the plate power loss is easily calculated. Likewise, the presence of excessive grid circuit losses will be evidenced by high grid current and, in some cases, by heating of the grid itself. The determination of the linearity of modulation, however, is something that cannot be read from meters. Naturally, if the DC plate current changes noticeably during modulation, its probable cause is due to non-linearity of the modulation, but the change in plate current gives no evidence as to the possible source of the trouble. It might be caused by insufficient grid excitation on the Class C stage, imperfect neutralization, or a mismatch between the modulator and its load. Anyone of these factors will cause the modulation to be non-linear, and each will produce a different kind of distortion in the wave form of modulated R.F. wave. Since the wave form will be affected, the best method of discovering the existence of non-linear modulation and of tracing the trouble to its source is by observance of the wave form with a cathode ray oscilloscope.

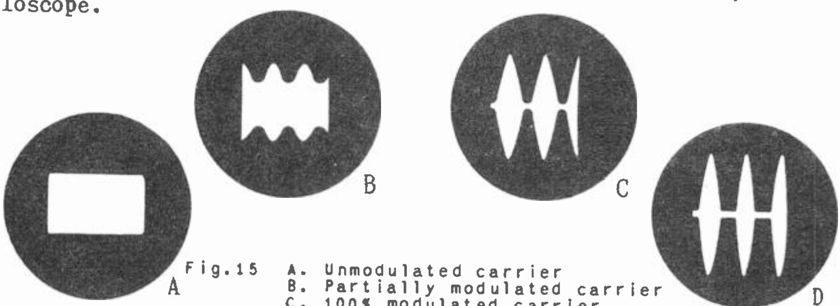


Fig.15 A. Unmodulated carrier
B. Partially modulated carrier
C. 100% modulated carrier
D. Overmodulation.

There are two general methods of using the cathode ray oscilloscope for determining the linearity of the modulation. In the first method, the modulated R.F. wave is applied to the vertical deflecting plates and the linear sweep circuit is so adjusted that the audio envelope is stationary on the screen. The best method of applying the modulated R.F. voltage to the vertical plates is to connect them to a small pick-up coil which is placed in the field of the Class C plate tank circuit. A twisted pair of wires should be used to connect the pick-up coil with the oscilloscope. Since there are always stray fields, both electrostatic and electromagnetic, in the vicinity of a transmitter, the placement of the oscilloscope requires some care in order that spurious voltages will not be induced in the test leads and cause the image on the screen to becomes valueless for linearity testing.

When no audio signal is fed to the modulator, the image on the screen will be a rectangle as shown at A in Fig. 15. With an audio signal of sine wave form applied to the grid of the modulator, and the gain control advanced slightly, the image takes on the form shown at B in this figure. This is a partially modulated R.F. wave. With a further advance of the gain control, the R.F. wave is 100% modulated, and the image appears as at C. Such an image indicates that all adjustments are correct and the modulation is linear.

Any further advance of the gain control will cause over-modulation, and will produce the image shown at D. Notice that, in this case, the modulated R.F. wave is zero for an appreciable part of the modulation cycle. The distortion of the audio envelope is considerable, and a shift in the average value of the rectified carrier has taken place. This is called "carrier shift". It must be avoided.

The results of non-linear modulation are best shown by using a different connection of the oscilloscope. The modulated R.F. voltage is fed to the vertical plates as before. However, instead of using the linear sweep voltage on the horizontal plates, the A.F. voltage output of the modulator is applied to them. This audio voltage may be taken from across the modulation choke, however, it is probable that the total voltage across the choke will be too much for the oscilloscope to handle, and the DC voltage at the top of the choke might be great enough to break down the condenser connected internally between the ungrounded horizontal plate and the binding post. For this reason, it is suggested that the scheme shown in Fig. 16 be employed. In this case, a condenser, a fixed resistor, and a potentiometer in series, are connected across the modulation choke. The values of these components are not critical, but the condenser must be able to withstand the peak voltage of the modulator, and its reactance should be low compared to the total resistance. By using a fairly high value of resistance, a lower capacity may be employed. If the total resistance is several hundred thousand ohms, the capacity of the condenser may be low as .25 mfd.

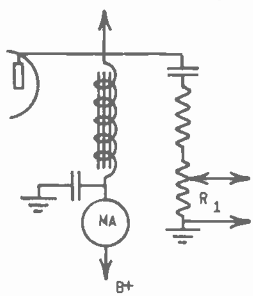


Fig.16 A method of applying modulation voltage to horizontal plate of oscilloscope.

The sliding arm of the potentiometer is connected to the ungrounded horizontal plate of the oscilloscope and the other plate is connected to the bottom of R_1. By varying the position of the sliding arm, the voltage applied to the horizontal plates may be changed. It is necessary that the total resistance of this auxiliary circuit be large compared to the reactance of the condenser so that the voltage applied to the horizontal plates will be practically in phase with the voltage across the modulation choke.

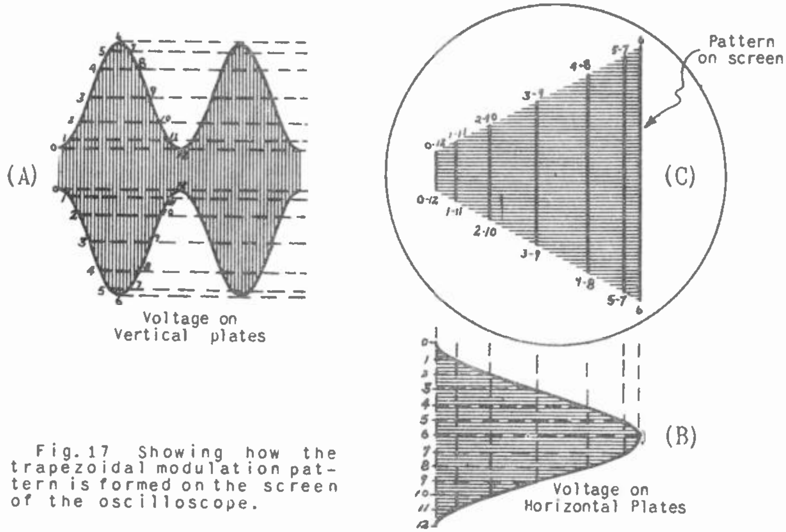


Fig.17 Showing how the trapezoidal modulation pattern is formed on the screen of the oscilloscope.

If the Class C amplifier is modulated less than 100% and the modulation is linear, a trapezoidal pattern will be seen on the screen. Why this pattern is formed may be seen by reference to Fig. 17. At A is shown the modulated R.F. wave that is applied to the vertical plates and at B is shown the audio voltage representing the output of the modulator. The combined result of these two voltages will cause the spot to trace the trapezoidal pattern shown at C in this figure. This may be verified by projecting various points from the wave form of each voltage on to the pattern at C.

Fig.18 100% modulation

Thus, at point 0 the position of the spot would be the intersection of the horizontal line drawn from point 0 on the modulated R.F. wave, and the vertical line drawn from point 0 on the audio wave.

This pattern is called a trapezoid. A trapezoid is a four-sided figure which has two opposite sides parallel. If the modulation is linear, all four sides of this figure will be straight lines. The per cent of modulation may also be calculated from this pattern. To do this, the two parallel sides of the figure are measured in any convenient units. The degree of modulation is the length of the long side minus the length of the short side, all divided by the sum of the lengths of these two sides. To change this result into per cent, it is multiplied by 100.

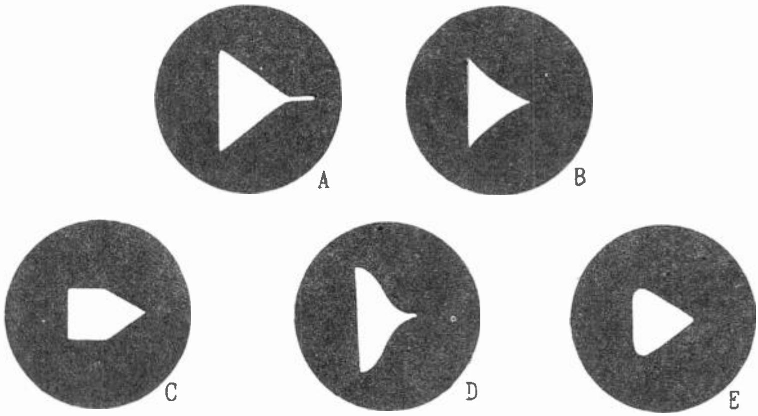


Fig. 19 A. Overmodulation
B. Regeneration in Class C stage
C. Insufficient excitation
D. Insufficient excitation and overmodulation
E. Poor regulation of power supply.

Let us now suppose that the per cent of modulation is increased until the carrier is modulated 100%. When this occurs, the short side of the trapezoidal pattern decreases until it is merely a point and the image becomes triangular. (See Fig. 18.) To make sure that the modulator itself is not producing any distortion, it is advisable to observe the wave form of the input voltage applied to the grid of the modulator, and also the output voltage to determine whether their wave forms are the same. This should be done by using a linear sweep voltage on the horizontal plates. If the output voltage differs in wave form from the input voltage, the trouble may be due to a mismatch between the modulator and its load.

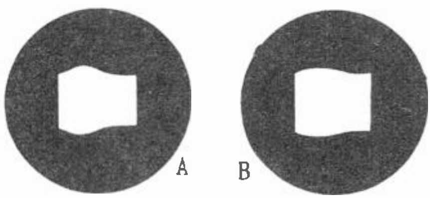


Fig. 20 A. Unmodulated carrier when power supply filter contains only one section. The filter choke was magnetically saturated. B. Same carrier with larger choke which does not saturate.

Various other forms of distortion are shown in Fig. 19. The pattern at A shows the result of overmodulation, that at B is due to regeneration in the Class C stage due to imperfect neutralization or caused by stray magnetic coupling between plate and grid circuits. C illustrated the pattern produced when the grid excitation is insufficient, and D shows the result of overmodulation and insufficient excitation. The pattern at E is caused by poor regulation of the power supply.

Unless the plate power supply of the modulator and Class C stage is well filtered, the R.F. carrier wave will have some 120 cycle hum modulation. Using a linear sweep voltage on the horizontal plates and the R.F. carrier on the vertical plates, the pattern shown at A in Fig. 20 was produced. In this case, the filter section of the power supply consisted on one section and the choke was magnetically saturated. By replacing this choke with another one of larger core area, which would not saturate, a better pattern results as shown at B in this figure. Two sections of filter were sufficient to eliminate entirely this source of distortion.

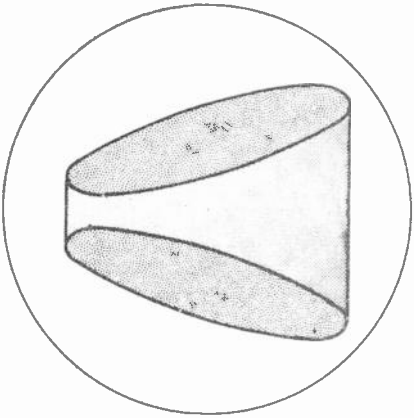


Fig.21 Modulation pattern produced when the audio voltage was of the improper phase.

To produce the trapezoidal patterns previously referred to, it is necessary that the audio voltage applied to the horizontal plates be taken from the output of the modulator. If an attempt is made to secure this audio voltage from any of the voltage amplifiers feeding the modulator, it will be found that its phase is incorrect, and the pattern produced will be similar to that shown in Fig. 21. In this case, the transmitter was correctly adjusted and was 100% modulated, but the phase of the audio voltage was not correct to produce a trapezoidal pattern.

TRANSMITTING TUBE CHART

TYPE NO.	DESCRIPTION CATH	DESCRIPTION BASE	FILAMENT VOLTS	FILAMENT AMPS	CAPACITANCES MICRO-MICRO FARADS C_{GP}	C_{GF}	C_{PF}	APPLICATION	RATED VOLTAGES E_P	E_G	SC. GRID	SUP GRID	RATED MA. I_P	I_G	SC GRID	POWER PL. DIS.	OUT PUT	DRIVER POWER WATTS	GRID LEAK OHMS
TRIODES																			
10	THOR FIL.	MED. 4 PIN	7.5	1.25	7.0	4.0	3.0	CLASS C AMP-OSC.	450	-100			60	10		15	12	4.0	10,000
								CLASS B MOD ●	425	-30	TWO TUBES		8				35		
UNITED HV12	THOR. FIL.	JUMBO 4 PIN	10.0	4.0	13.0	7.0	5.0	CLASS C AMP-OSC.	2000				250	60		200	500		
								CLASS C MOD-AMP	1750				200	60		125	250		
RK18	THOR FIL.	MED. 4 PIN	7.5	1.25	5.0	5.8	2.0	CLASS C AMPLIFIER	1000	-150			85	15		40	50	3.0	15,000
T20*	THOR FIL.	MED. 4 PIN	7.5	1.75	4.0			CLASS C AMP-OSC.	750	-100			75	17		20	42	5.0	6,000
								CL. B MOD 2 TUBES●	800	-40			20				70		
RK24	OXIDE FIL.	SMALL 4 PIN	2.0	0.12	5.5	3.5	3.0	5 METER PORT OSC	180	-45			20			1.5	1.2		
HV27	THOR. FIL.	JUMBO 4 PIN	10.0	4.0	14.0	7.0	4.5	CLASS C AMP-OSC	2000				250	60		200	500		
								CLASS C MOD-AMP	1750				200	80		125	250		
								CLASS B MOD ●	2000										
RK30	THOR. FIL.	MED. 4 PIN	7.5	3.25	2.5	2.7	1.0	CLASS C AMPLIFIER	1250	-175			70	15		35	65	4.0	10,000
								CLASS C MOD-AMP	1000	-200			70	15		35	50	4.0	10,000
								CL. B LINEAR AMP.	1000	-55			42			35	14		
RK31	THOR. FIL.	MED. 4 PIN	7.5	3.0				CLASS C AMPLIFIER	1000	-50			85	15		55	50	5.0	3,000
								CLASS B MOD ●	1250	0	TWO TUBES		50				140		
RK32*	THOR. FIL.	MED. 4 PIN	7.5	3.25	3.0	2.0	1.0	CLASS C AMPLIFIER	1250	-250			100	25		50	75	7.5	10,000
								CLASS C MOD AMP	1000	-185			100	25		50	50	7.5	7500
RK34*	OXIDE CATH.	MED. 7 PIN	6.3	0.8	2.7	4.2	2.1	CLASS C AMP-OSC.	300	-36			80	18		10	14		2000
								CLASS B MOD ●	300	-15			16				12		
RK35*	THOR. FIL.	MED. 4 PIN	7.5	3.25	2.7	3.5	0.4	TELEPHONY											
								CLASS C AMP OSC	1000	-320			100	20		85	65	7.0	20,000
								TELEGRAPHY											
								CLASS C AMP OSC	1500	-400			100	20		55	115	7.0	20,000
35T*	THOR. FIL.	MED. 4 PIN	5.0	4.0	1.9	2.5	0.3	TELEPHONY											
								CLASS C AMP-OSC	1500	-180			100	20		55	120	4.0	5000
								TELEGRAPHY											
								CLASS C AMP-OSC	1500	-180			125	20		55	170	4.0	5000
	A MAXIMUM OF 2000 VOLTS CAN BE APPLIED TO THE 35T. PLATE DISSIPATION SHOULD NOT EXCEED 70 WATTS, 150 MA PLATE CURRENT FOR CW, 125 FOR PHONE							CLASS B MOD ●	1500				155				235		
RK36*	THOR FIL.	MED 4 PIN	5.0	8.0	5.0	4.5	1.0	CLASS C AMP-OSC	3000	-360			165	35		100	370	15	
								CLASS C AMPLIFIER TELEPHONY	2000	-360			150	30		100	200	15	15,000
RK37*	THOR FIL.	MED 4 PIN	7.5	3.25	3.2	3.5	0.2	CLASS C AMP-OSC	1250	-90			90	20		35	78	3.2	5000
								CLASS B MOD ●	1250	-32	TWO TUBES		32	29			125	2.8	
								CL. B LINEAR AMP	1250	-45			43			35	19	1.8	
								CL. B LINEAR AMP	PEAK OUTPUT 76 WATTS										
RK38*	THOR FIL.	MED 4 PIN	5.0	8.0	4.8	4.6	0.9	CLASS C AMP-OSC.	2000	-200			150	30		100	225	11.0	15,000
								CLASS B MOD ●	2000	-52	TWO TUBES		36	39			330	5.8	
								CL. B LINEAR AMP	2000	-100			75			115	55	7.0	
								CL. B LINEAR AMP	PEAK OUTPUT 220 WATTS										
T55*	THOR FIL.	MED 4 PIN	7.5	2.75	3.75	4.0	1.5	CLASS C AMPLIFIER	1500	-200	40		150	25		55	168	15	8000
								OSCILLATOR	1250	-200	40		125			55	66		8000
F100	TUNGS FIL.	SPECIAL	11.0	25.0	10.0	4.0	2.0	CLASS C AMPLIFIER	2000	-300			500			500	800		10,000
F108A	TUNGS FIL.	JUMBO 4 PIN	10.0	11.0	7.0	3.0	2.0	CLASS C AMPLIFIER	3000	-350			200			175	400		15,000
HF100*	THOR FIL.	MED 4 PIN	10.0	2.0	4.5	3.5	1.4	OSCILLATOR	1200	-225			130	30		60	96		7500
								CL. B LINEAR AMP	1500	-55			75	1.5		80	42	PEAK 3	
								CLASS C AMPLIFIER	1500	-200			150	18		55	170	6.0	10,000
								CL. C AMP TELEPHONY	1250	-250			110	21		35	105	8.0	10,000
								CLASS B MOD ●	1500	-52	TWO TUBES		50				260	2.0	
100TL*	THOR FIL.	MED 4 PIN	5.0	6.5	2.3	2.0	0.4	CLASS C AMPLIFIER	3000	-600			135	30		105	300		20,000
								TELEPHONY OR	2000	-400			150	30		75	225		13,000
								TELEGRAPHY	1000	-200			200	30		60	120		7000
100TH*	THOR FIL.	MED 4 PIN	5.0	6.5	2.0	2.2	0.3	CLASS B MOD ●	1250	0							260		
								CLASS C AMPLIFIER	3000	-210			135	45		105	300		5,000
								TELEPHONY OR	2000	-140			150	45		75	225		3000
								TELEGRAPHY	1000	-70			200	45		80	120		1500
ZB120		JUMBO 4 PIN	10.0	2.0	5.2	5.3	3.2	CLASS B MOD ●	1250	0	TWO TUBES		95				245	4.0	
								CLASS B MOD ●	1500	-9	TWO TUBES		50				300	5.0	
								CL. B LINEAR AMP	1250	0	TELEGRAPHY		150	21		187	120	1.2	
	A SELF BIAS RESISTOR OF 750 OHMS SHOULD BE INCLUDED IN THE FIXED BIAS FOR THE CLASS C MOD-AMP							CL. B LINEAR AMP	1250	0	TELEPHONY		95	8		74	45	1.5	
								CLASS C AMPLIFIER	1250	-135	TELEGRAPHY		180	25		55	145	5.5	6000
								CLASS C AMPLIFIER	1000	-400			120	21		25	95	5.0	7000
								CL. C GRID MOD AMP	1250	-75			90	7		70	42	1.6	750
HK154*	THOR FIL.	MED 4 PIN	5.0	6.5	5.9	4.3	1.1	CLASS C AMPLIFIER	1000	-380			75	20		48	127	10.0	20,000
								CL. C AMPLIFIER -1-	1250	-460			70	20		50	162	12.0	23,000
								CL. C AMPLIFIER -2-	1500	-590			167	20		50	200	15.0	30,000
					1- MAXIMUM FOR TELEPHONY			CL. B LINEAR AMP	1500	-285			52			50	28	5.0	
					2- MAXIMUM FOR TELEGRAPHY			CLASS B MOD ●	1500	-265	TWO TUBES		40			95	250	10.0	
								CLASS B MOD ●	1000	-155	TWO TUBES		60			100	200	10.0	
T155*	THOR TUNGS	JUMBO 4 PIN	10.0	4.0	8.0	2.5	1.0	CLASS C AMP-OSC	3000	-250			300	60		155	450		4200
T200*	THOR FIL.	JUMBO 4 PIN	10-11	4.0	7.0	5.0	3.0	CLASS C AMP-OSC	2500	-300			350	80		200	500		3750
								CLASS C MOD AMP	2000	-300			350	80		200	500		3750
C200* HF200	THOR FIL.	JUMBO 4 PIN	10.5	3.4	5.8	5.2	1.2	CLASS C AMPLIFIER	2500	-300	TELEGRAPHY		200	18		120	380	8.0	17,000
								CLASS C AMP	1750	-300	TELEPHONY		200	50		80	270	14.0	10,000
								CLASS B MOD ●	2500	-150	TWO TUBES		60				600	8.0	
C201	THOR FIL.	JUMBO 4 PIN	10.0	3.25	9.0	6.0	1.8	CLASS B MOD	1250	-45	TWO TUBES		160			75	250		
								CL. B. LINEAR	1250	-45			110			92	46		
								CLASS C AMPLIFIER	1250	-200	TELEPHONY		165	50		71	135		4000
								CLASS C AMPLIFIER	1250	-125	TELEGRAPHY		165	25		71	135		5000

*RECOMMENDED VALUES UP TO 56-60 MEGACYCLES—HIGH EFFICIENCY OBTAINED FROM THESE TUBES AT 56-60 MEGACYCLES.

● STATIC PLATE CURRENT IS GIVEN UNDER "IP" FOR TWO TUBES.

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TRANSMITTING TUBE CHART

TYPE NO.	DESCRIPTION CATH	DESCRIPTION BASE	FILAMENT VOLTS	FILAMENT AMPS	CAPACITANCES MICRO-MICRO FARADS C_{gp}	C_{gf}	C_{pf}	APPLICATION	RATED VOLTAGES E_p	E_g	SG	SUP	RATED MA I_p	I_g	SG	POWER PLATE DISS	POWER OUTPUT	DRIVER POWER WATTS	GRID LEAK OHMS
TRIODES																			
C202	THOR FIL	JUMBO 4 PIN	10.0	3.25	8.0	5.5	1.0	CLASS B MOD	1250	-100	TWO TUBES		160			75	250		
								CL B LINEAR AMP	1250	-100			110			92	46		
								CLASS C AMPLIFIER	1000	-260	TELEPHONY		165	50		55	110		5000
								CLASS C AMPLIFIER	1250	-200	TELEGRAPHY		165	25		71	135		10,400
203B	THOR-TUNGS	JUMBO 4 PIN	10.0	3.85	14.0	6.0	3.0	CLASS B MOD ●	1000	-35	TWO TUBES		40				200		
HD203A	THOR TUNGS	JUMBO 4 PIN	10.0	4.0	12.0	7.0	5.0	CLASS C AMP-OSC	2000		TELEGRAPHY		250	60		150	300		3000
								CLASS C AMP-OSC	1750	-180			250	60		150	300		5000
								CLASS B MOD ●	1750	-675	TWO TUBES		36				500		
203A 303A C203A	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	14.5	6.5	5.5	CLASS C AMPLIFIER	1250	-125	TELEGRAPHY		150	25		100	130	7.0	5000
								CLASS C AMPLIFIER	1000	-135	TELEPHONY		150	50		100	100	14.0	3000
								CL B LINEAR AMP	1250	-45			106			100	425		
HD203C HD211C	THOR TUNGS	JUMBO 4 PIN	10.0	4.0	9.0 9.0	6.0 5.0	4.0 4.0	CLASS C AMP-OSC	2000	-200	TELEGRAPHY		250	60			250		3333
								CLASS C AMP-OSC	1750	-200			250	60			250		3333
204A 304A 504A	THOR TUNGS	SPECIAL	11.0	3.85	15.0	12.5	2.3	CLASS C AMPLIFIER	2000	-175			250	50		250	350		5000
304B *	THOR TUNGS	MED 4 PIN	7.5	3.25	2.5	2.0	0.7	CLASS B MOD ●	1250	-110	TWO TUBES		40				140	10.0	
								CLASS C AMPLIFIER	1250	-180	TELEGRAPHY		100				85		
								CLASS C AMPLIFIER	1000	-180	TELEPHONY		100	25			65		7500
211 311	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	14.5	6.0	5.5	CLASS C AMPLIFIER	1250	-225	TELEGRAPHY		150	18		100	130	7.0	10,000
								CLASS C AMPLIFIER	1000	-260			150	35		100	100	14.0	5000
								CL B LINEAR AMP	1250	-100			106			100	425		
211C 311C	THOR TUNGS	JUMBO 4 PIN	10.0	3.10	9.0	6.0	5.0	SAME AS 311											
316A **	THOR TUNGS	NO BASE	2.0	3.65	1.6	1.2	0.8	CLASS C AMPLIFIER	450		TELEGRAPHY		80	12			7.5	SPECIAL ULTRA	
								CLASS C AMPLIFIER	400		TELEPHONY		80	12			6.5	HIGH FREQ TUBE	
242C	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	13.0	6.1	4.7	CLASS B MOD ●	1250	-80	TWO TUBES		50			100	200	25	
								CL B LINEAR AMP	1250	-90			120				50		
								CLASS C MOD AMP	1000	-160			150	50			100		5200
250TL *	THOR FIL	JUMBO 4 PIN	5.0	10.5	3.5	3.0	0.5	CLASS C AMPLIFIER	3000	-600			330	45		240	750		13,500
								CLASS C AMPLIFIER	2000	-400			350	45		200	500		9000
								CLASS C AMPLIFIER	1000	-200			300	45		100	200		4500
250TH *	THOR FIL	JUMBO 4 PIN	5.0	10.5	3.3	3.5	0.3	CLASS C AMPLIFIER	3000	-210			350	55		240	750		3800
								CLASS C AMPLIFIER	2000	-140			350	55		200	500		2550
								CLASS C AMPLIFIER	1000	-70			300	55		100	200		1500
								CLASS B MOD ●	1400	0	TWO TUBES						575	APPROX.	
261A 361A	SEE 211C-311C																		
276A 376A	THOR FIL	JUMBO 4 PIN	10.0	3.0				SEE 211C-311C											
C300 HF300 *	THOR FIL	JUMBO 4 PIN	11.5	4.0	6.5	6.0	1.4	CLASS C AMP OSC	3000	-400	TELEGRAPHY		250	28		150	600	16.0	14,300
								CLASS C AMPLIFIER	2000	-300	TELEPHONY		250	36		115	385	17.0	8500
								CL B LINEAR AMP	2500	-100			120	0.5		195	105	6.0	
								CLASS B MOD ●	2000	-72	TWO TUBES		60				650	14.0	
300T *	THOR TUNGS	JUMBO 4 PIN	7.5	12.0	4.0	4.0	0.6	CLASS C AMPLIFIER	2500	-400			300	60		190	560	60.0	6700
								CL B LINEAR AMP	2500	-150			200			300	200		
HK354D * HK354	THOR TUNGS	JUMBO 4 PIN	5.0	10.0	3.8	4.5	1.1	CLASS C AMPLIFIER	3000	-275			150	27		150	300		10,000
								CLASS B AMPLIFIER	3000	-275	TWO TUBES		390				665	56.0	
500T *	THOR TUNGS	SPECIAL	7.5	20.0	4.5	6.0	0.8	CLASS C AMPLIFIER	2000	-400			450	100		250	650		4000
								CLASS C AMPLIFIER	3000	-600			450	100		350	1000		6000
								CLASS C AMPLIFIER	4000	-800			450	100		450	1350		8000
756	THOR TUNGS	MED 4 PIN	7.5	2.0	8.0	3.5	2.7	CLASS C AMP-OSC	850	-75	TELEGRAPHY		110	20		34	60		3750
								CLASS C AMP-OSC	750	-75	TELEPHONY		110	20		34	48		3750
								CLASS B AMPLIFIER	850	-30	TWO TUBES		20				100		
800 *	THOR-TUNGS	MED 4 PIN	7.5	3.25	2.5	2.75	1.0	SEE RK50											
801	THOR TUNGS	MED 4 PIN	7.5	1.25	6.0	4.5	1.5	CLASS C AMP-OSC	600	-150			65	15		20	25	4.0	10,000
								CLASS C AMP-OSC	500	-190			55	15		20	18	4.5	10,800
								CL B LINEAR AMP	600	-75			45			20	7.5		
								GRID BIAS MOD-AMP	600				50	20		20	10	2.0	
805 905	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	6.5	8.5	10.5	CLASS C AMP OSC	1500	-105			200	40		85	215	8.5	2625
								CLASS C AMPLIFIER	1250	-160			160	60		60	140	16.0	2650
								CL B LINEAR AMP	1500	-10			115	15		115	375	7.5	
								CL B AMPLIFIER ●	1250	0	TWO TUBES		148				300	7.0	
806	THOR TUNGS	JUMBO 4 PIN	5.0	10.0	3.4	6.1	[illegible]	CLASS C AMP OSC	3000	-600			195	25		135	450	20.0	24,000
								CLASS C AMPLIFIER	2500	-600			195	40		97	390	32.0	15,000
								CL B LINEAR AMP	3000	-240			70	0		140	70	5.0	
								CLASS B MOD ●	3000	-240	TWO TUBES		20				660	10.0	
808 *	THOR TUNGS	MED 4 PIN	7.5	4.0	5.0	5.0	0.2	CLASS C AMP-OSC	1500	-200			125	30		47.5	140	9.5	6,700
								CLASS C AMPLIFIER	1250	-225			100	32		20	105	10.5	7,000
								CL B LINEAR AMP	1500	-35			45	1.0			22	2	
								CLASS B MOD ●	1250	-15	TWO TUBES		40				[illegible]	7.8	
814	THOR TUNGS	JUMBO 4 PIN	10.0	4.0	13.0	7.0	3.5	CLASS C AMP-OSC	2000	-400			300	75		200	400	3.00	5200
822	THOR TUNGS	JUMBO 4 PIN	10.0	4.0	14.0	8.0	6.0	CLASS C AMP-OSC	2000	-220			300	60		200	400		3700
								CLASS B MOD	2000	-90	TWO TUBES		50				500	5.00	
825	THOR TUNGS	MED 4 PIN	7.5	2.0	7.0	3.0	2.7	CLASS C AMP-OSC	750	-180			110	25		40	50		7200
								CLASS B AMPLIFIER	850	-675	TWO TUBES		50				82		
830 930	THOR TUNGS	MED 4 PIN	10.0	2.15	9.9	4.9	2.2	CLASS C AMPLIFIER	750	-180			110	18		40	55	7.0	10,000
								GRID BIAS MOD AMP	1000	-200			50	20		40	15	3.0	
830B 930B	THOR TUNGS	MED 4 PIN	10.0	2.0	11.0	5.0	1.8	CLASS C AMP OSC	1000	-110	TELEGRAPHY		140	30		50	90	7.0	3670
								CLASS C AMP OSC	800	-150	TELEPHONY		95	20		26	50	5.0	7500
								CL B LINEAR AMP	1000	-35			65	6.0			28	6.0	
								CL B AMPLIFIER ●	1000	-35	TWO TUBES		20				175	6.0	
834 *	SEE RK52																		
841 941	THOR TUNGS	MED 4 PIN	7.5	1.25	7.0	4.0	3.0	CLASS C AMP OSC	450	-32			50	12.5		8.5	14	1.25	2560
								CLASS C AMPLIFIER	350	-36			50	18		6.0	11.5	1.75	2000
								CL B AMPLIFIER ●	425	-5	TWO TUBES		13				28		
841A	THOR TUNGS	MED 4 PIN	10.0	2.0	9.0	3.5	2.5	CLASS C AMP-OSC	1250	-180			150	30			85		6000
								CLASS C AMPLIFIER	1000	-180			150	30					6000
838	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	8.0	6.5	5.0	CLASS C AMP-OSC	1250	-80			150	30		100	130	6.0	3000
								CLASS C AMPLIFIER	1000	-35			150	80		100	100	16.0	3000
								CLASS B MOD ●	1250	0	TWO TUBES		74			100	130		

*RECOMMENDED VALUES UP TO 56-60 MEGACYCLES—HIGH EFFICIENCY OBTAINED FROM THESE TUBES AT 56-60 MEGACYCLES.
● STATIC PLATE CURRENT IS GIVEN UNDER "IP" FOR TWO TUBES.

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TRANSMITTING TUBE CHART

TYPE NO	DESCRIPTION CATH.	DESCRIPTION BASE	FILAMENT VOLTS	FILAMENT AMPS	CAPACITANCES MICRO-MICRO FARADS Cgp	Cgf	Cpf	APPLICATION	RATED VOLTAGES Ep	Eg	SG	SUP	RATED MA Ip	Ig	SG	POWER PL DISS	POWER OUT-PUT	DRIVER POWER WATTS	GRID LEAK OHMS
TRIODES																			
849 949	THOR TUNGS	SPECIAL	11.0	5.0	33.5	17.0	3.0	CLASS C AMPLIFIER	2000	-200			300	40		400	450		5000
								CLASS B MOD ⊗	2500	-130	TWO TUBES		20			400	500	7	
852 952	THOR TUNGS	MED 4 PIN	10.0	3.25	2.6	1.9	1.0	CLASS C AMP-OSC	3000	-600			85	15		100	165	12	10,000
								CLASS C AMPLIFIER	2000	-500			67	30		100	75	23	10,000
								CL. B LINEAR AMP	3000	-250			43			100	40		
831	THOR-TUNG	SPECIAL	11.0	10.0	4.0	3.8	1.4	CLASS C AMPLIFIER	3500	-400			275	40		400	590	30	10,000
TETRODES PENTODES																			
RK20 RK20A	THOR TUNGS	MED 5 PIN	7.5	3.0 3.25	0.12	11.0	10.0	CLASS C AMPLIFIER	1250	-100	300	0	80	7-10	37	40	64	1.0	15,000
								CLASS C AMPLIFIER	1250	-100	300	+45	92	7-10	32	40	80	1.0	15,000
								SUPP. MOD. AMP	1250	-100	300	-45	43	7-10	36	40	18	1.0	15,000
RK23 RK25 RK25B	OXIDE CATH	MED 7 PIN	2.5 6.3	2.0 0.8	0.2	10.0	10.0	CLASS C AMP-OSC	500	-90	200	0	50	6-8	40	10	18	0.8	15,000
								CLASS C AMP-OSC	500	-90	200	+45	55	6-8	35	10	24	0.8	15,000
								SUPP MOD. AMP	500	-90	200	-45	32	6-8	40	10	5.5	0.8	15,000
RK28	THOR TUNGS	JUMBO 5 PIN	10.0	5.0	0.2	15.5	5.5	CLASS C AMPLIFIER	2000	-100	400	0	120	10-12	75	125	160	1.8	10,000
								CLASS C AMPLIFIER	2000	-100	400	+45	140	10-12	60	125	200	1.8	10,000
								SUPP. MOD. AMP	2000	-100	400	-45	80	10-12	85	125	60	2.7	10,000
RK39 RK41	OXIDE CATH	MED 5 PIN	6.3 2.5	0.9 2.4	0.2	13.0	10.5	CLASS C AMP-OSC	500	-60	250		95	3.0	12	42	35	0.26	10,000
								CLASS C AMPLIFIER	400	-50	250		95	2.5	8	33	25	0.18	10,000
								CLASS B RF AMP	500	-30	250		75	0.3	3		11		
305A*	THOR TUNGS	MED 4 PIN	10.0	3.1	0.14	10.5	5.4	CLASS C AMP-OSC	1000	-270	200		125				85		
								CLASS C AMPLIFIER	800	-270	200		125				70		
								CLASS B RF AMP	1000	-135	200		90			60	30		
306A	THOR TUNGS	MED 5 PIN	2.75	2.0	0.35	15.0	13.0	CLASS C AMPLIFIER PLATE & SCREEN MODULATED	300	-50	180		36	3.0	15		7		10,000
307A	THOR TUNGS	MED 5 PIN	5.5	1.0	0.35	15.0	12.0	SUPP MOD CL C AMP	500	-35	200	-50	40	1.5	20		6		
								CLASS C AMP-OSC	500	-35	250	0	60	14	13		20		
802	OXIDE CATH	MED 7 PIN	6.3	0.95	0.15	12.0	8.5	CLASS C AMPLIFIER	500	-100	250	+40	45	2.0	12	10	16	0.25	15,000
								SUPP. MOD. AMP	500	-90	200	-45	22	4.5	28	10	8.5	0.5	15000
								CL. B LINEAR AMP	500	-28	200	0	25	0	7	10	3.5	0.18	
803	THOR TUNGS	JUMBO 5 PIN	10.0	3.25	0.15	15.5	28.5	CLASS C AMPLIFIER	2000		500	-30	160	16	42	125	210	1.6	5000
								SUPP MOD AMP	2000		500	-50	80	1.5	55	125	53	1.6	5000
								CL. B LINEAR AMP	2000		500	-40	80	3.0	15	125	54	1.5	
804	THOR TUNGS	MED 5 PIN	7.5	3.0	0.01	16.0	14.6	CLASS C AMPLIFIER	1250	-100	300	0	80	7.0	33	40	64	0.9	15,000
								CLASS C AMPLIFIER	1250	-100	300	+45	92	7.0	27	40	80	0.9	15,000
								SUPP MOD AMP	1250	-100	300	-50	48	7.0	35.5	40	21	0.85	15,000
807	THOR TUNGS	MED 5 PIN	6.3	0.9	0.2	11.6	5.6	CLASS C AMP-OSC	400	-50	250		95	2.5	9		25	0.18	20,000
								CLASS C MOD-AMP	325	-75	270		80	1.5	9		17	0.15	50,000
								CL. B LINEAR AMP	400	-25	250		75	0	4		9	0.25	
						TWO TUBES		CL. AB MOD ⊗	400	-25	300		102	20			60	0.35	
850	THOR TUNGS	JUMBO 4 PIN	10.0	3.25	0.2	17.0	26.0	CLASS C MOD-AMP	1000	-100	140		125	45		60	65	10	5000
								CLASS C AMP-OSC	1250	-150	175		160	35		70	130	10	4300
860	THOR TUNGS	MED 4 PIN	10.0	3.25	0.08	7.75	7.5	CLASS C AMPLIFIER	3000	-150	300		85	15		100	165	7.0	10,000
								CLASS C MOD-AMP	2000	-225	300		67	30		69	75	15.0	10,000
								CL. B LINEAR AMP	3500	-150	300		43			100	40		
861	THOR-TUNGS	SPECIAL	11.0	10.0	0.10	17.0	13.0	CLASS C AMPLIFIER	3500	-250	600		275	30		400	590	25	8000
865		MED 4 PIN	7.5	2.0	0.10	8.5	8.5	CLASS C AMP-OSC	750	-80	125		40	5.5		14	16	1.0	14,000
								CLASS B MOD-AMP	500	-120	125		40	9.0			10	2.5	13,500
								CL. B LINEAR AMP	750	-30	125		22				45		

*RECOMMENDED VALUES UP TO 56-60 MEGACYCLES—HIGH EFFICIENCY OBTAINED FROM THESE TUBES AT 56-60 MEGACYCLES

⊗STATIC PLATE CURRENT IS GIVEN UNDER "IP" FOR TWO TUBES.

RECEIVING TUBES USED IN TRANSMITTERS

TYPE NO	CATH.	BASE	VOLTS	AMPS	Cgp	Cgf	Cpf	APPLICATION	Ep	Eg	SG	SUP	Ip	Ig	SG	PL DISS	OUT-PUT	DRIVER POWER WATTS	GRID LEAK OHMS
TRIODES																			
2A3	OXIDE	MED. 4 PIN	2.5	2.5	13.0	9.0	4.0	CLASS AB-PP-MOD	300	-62	TWO TUBES		80				15		750
								CLASS C AMP-OSC	400	-180			100	10		15	25	5.0	50,000
6A3	OXIDE	MED. 4PIN	6.3	1.0	16.0	7.0	3.5	SAME AS 2A3											
6A6	OXIDE	MED. 7PIN	6.3	0.8				SEE RK54	PUSH PULL TRIODE										
6E6 #	OXIDE	MED. 7PIN	6.3	0.6				SEE RK34	PUSH PULL TRIODE										
6N7G 6N7	OXIDE	SMALL OCTAL 8 PIN	6.3	0.8				SEE RK 34	PUSH PULL TRIODE										
12A	THOR FIL	MED 4PIN	5.0	0.25	8.0	3.5	2.5	CLASS C AMP-OSC.	250	-90			80			9	11		20,000
19 #	THOR FIL	SMALL 6 PIN	2.0	0.26	PUSH PULL TRIODE			CLASS C AMP-OSC	135	-50			27	10		1.6	2.0		5000
								CLASS B MOD	135				100	MAX.			2.1		
45 #	OXIDE	MED 4PIN	2.5	1.5	6.5	3.6	3.0	CLASS C AMP-OSC	400	-200			40			10	10	5.0	50,000
53	OXIDE	MED 7PIN	2.5	2.0	PUSH PULL TRIODE			SEE RK54											
71A #	THOR. FIL	MED 4PIN	5.0	0.25	6.6	5.2	2.8	SEE 12A											
46	OXIDE	MED 5 PIN	2.5	1.75	AS A TRIODE			CLASS C AMPLIFIER	400	-50			40	30		10	10	5.0	20,000
								CLASS B MOD	400	0	TWO TUBES		200	MAX		20	20		
955 ●	OXIDE	SPECIAL	6.3	0.15	ACORN-FOR ULTRA HIGH FREQ.			CLASS C AMP-OSC	180	-55			7.0	1.5			0.5		20,000

SATISFACTORY OPERATION AS MODULATED OSCILLATOR CAN BE HAD ON 56-60 MC. BAND
TYPES 19 AND 6E6 IDEAL FOR PORTABLE AND PORTABLE MOBILE USE RESPECTIVELY ON 56-60 MC.
● WILL OSCILLATE AT THE VALUES DOWN TO 1 METER - BELOW THIS POINT REDUCE RATINGS

TETRODE-PENTODE TYPES

TYPE NO	CATH.	BASE	VOLTS	AMPS	Cgp	Cgf	Cpf	APPLICATION	Ep	Eg	SG	SUP	Ip	Ig	SG	PL DISS	OUT-PUT	DRIVER POWER WATTS	GRID LEAK OHMS
2A5	OXIDE	MED 6PIN	2.5	1.75				CLASS C AMP-OSC.	400	-50	100		30	10		5.0	7.0		5000
6F6 6F6G	OXIDE	SMALL OCTAL 7 PIN	6.3	0.7				SAME AS 2A5											
6L6 # 6L6G	OXIDE	SMALL OCTAL 7 PIN	6.3	0.9				CLASS C AMP-OSC	450	-120	300		100	40	80	10	25		50,000
						TWO TUBES		CLASS AB MOD ●	400	-25	300		100				60		
42	OXIDE	MED 6PIN	6.3	0.7				SAME AS 2A5											
47	OXIDE	MED 5PIN	2.5	1.75	1.2	8.6	13.0	OSC. DOUBLER	350		100	50	40	30		7.0	7.0		50,000
59	OXIDE	MED 7 PIN	2.5	2.0				OSC. DOUBLER	SAME AS 47										
								CLASS B MOD	400	0	TWO TUBES		200	MAX		20	20		

● STATIC PLATE CURRENT GIVEN

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POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

MOTORS AND GENERATORS

LESSON NO. 8

LITTLE SECONDS

.....but they tick away years.

Tick-tick-tick. It does not take very long for a few seconds to be 'ticked' off by your watch. And when we think of time as the passing of millions of seconds it may seem as though we have a never-ending supply of it.

But the 'ticking' goes on endlessly. While we work, play, and sleep, seconds become hours and grow into days and years. And the older a person becomes, the swifter time seems to pass.

Unfortunately, some young men do not realize that during early life they MUST equip themselves to make money in substantial amounts during the middle years of life......the productive years. As a result, they are forced to enjoy a small income and a 'merry-go-round' ride. But when the 'merry-go-round' stops, they are just about where they started out. And from that point on they slip and skid down the reclining years of their life.

The fact that you are devoting your time to training is an excellent indication that YOU recognize the need of equipping yourself to make money. During the years that your less ambitious fellow men are going round and round on the 'merry-go-round' YOU will be forging ahead steadily, turning seconds, minutes, and hours into MONEY.

Let nothing turn you from the straight road you are following. It leads through an enjoyable, financially secure life for YOU and YOUR loved ones.

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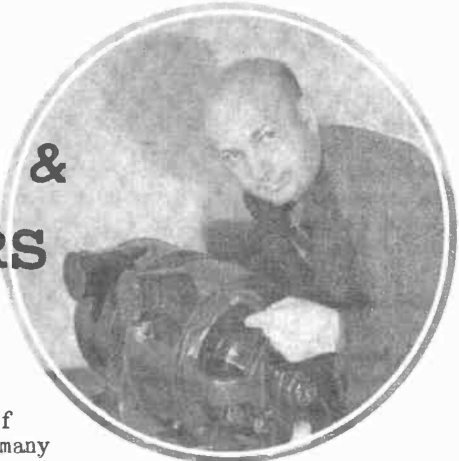
Lesson Eight



DC MOTORS & GENERATORS

"While it is true that practically all modern broadcast equipment is now designed to operate directly from an alternating current source of supply, still, there exists many transmitters which use some form of motor-generator unit. Therefore, every broadcast engineer who can consider himself properly qualified for his job must have a thorough understanding of both DC motors and generators.

"While the subject is too long to be completely covered in this lesson, it is the purpose of this material to provide you with sufficient information so that you will understand the care and operation of this type of equipment."

1. THE NEED OF MOTORS AND GENERATORS. The modern broadcast transmitter is completely AC operated. Only in isolated cases is any type of battery power employed. This, however, has not always been true. It has not been many years since large banks of storage batteries and their associated charging equipment were seen in practically every broadcast station. The first step toward the elimination of batteries was provided by the increasing popularity of motor-generator sets. The motor-generator set consists of a motor designed to operate from the power line which feeds the station, and a generator whose voltage and current capacity depend upon whether it is to supply plate voltage or filament voltage.

Until recently, it has not been practical to use alternating current for the filaments of large transmitting tubes. This was due to the inability of designing these large tubes so that AC hum would not be introduced into the carrier wave, thereby causing "hum modulation". In such instances, it was common practice to employ motor-generator sets which converted the AC line voltage into DC voltage of the proper value to be applied to the filaments. One of the most important discoveries of recent years is that of inverse feedback, the fundamental principles of which were described in Lesson 23 of Unit 1. This same system is now being applied to modern broadcast transmitters. A portion of the modulated R.F. output is rectified and fed back to the input of the audio system.

As a result, distortion and hum modulation are cancelled and the output of the transmitter is relatively hum free. This system makes possible the use of AC filament voltages and all transmitters now being constructed employ step-down transformers for filament supply.

The perfection of the mercury-vapor rectifier tube has led to the further development of high-voltage rectifier circuits, and all transmitters built in the past few years employ tube rectifiers for plate-supply voltage. There are, however, many transmitters still in use which use one or more motor-generator sets, and the subject of motors and generators is of sufficient importance for us to inquire into their operating characteristics.

In this lesson we shall cover the main principles of DC motors and generators in such a manner that the student will be able to distinguish the different types of machines, make minor repairs and adjustments, and be able to maintain such equipment in good operating condition. It is, of course, impossible to exhaust the subject in a lesson of this type, nor is it desirable. The ability to design and construct motors and generators is properly left for the electrical engineer.

2. THE ELECTRICAL GENERATOR. *(A) Definition.* To the average mind, the electrical generator is a machine which "makes" electricity. This, we know to be a popular misconception of the operation of a generator. If we define an electrical current as the drift of electrons in a given direction through a conductor, produced by an electromotive force, we see that it is impossible to "make" electricity in the sense that it is created. Nor is the term "generator" well advised; for generation also implies the idea of creation. Rather, the electrical generator should be thought of as a converter in that it changes or converts mechanical energy into electrical energy. In fact, modern physics is based on the law of the conservation of energy. This law states, in effect, that the total energy in the universe is constant, and that it is not possible to either create or destroy energy. It is, however, possible to change energy from one form to another, and this is the purpose of the electrical generator.

Electrical generators have their foundation in the law of induction. As given in Lesson 10 of Unit 1, this important law states that whenever there is relative motion between a magnetic field and a conductor in such a manner that the lines of force are cut by the conductor, a voltage is induced into the conductor. In Lesson 13 of Unit 1, there was presented a discussion of how it is possible to induce a voltage of sine wave form in a single-turn coil, when it is rotated through a steady magnetic field, To refresh the memory of the student, a brief review of this action will be given.

(B) Generation of a Voltage in a Loop of Wire. Fig. 1 illustrates a single loop of wire rotating in a magnetic field. To identify the two sides of the loop, one is colored black and the other is white. The magnetic field poles are not shown, but the field itself is represented by lines of force threading through the

loop from left to right. One end of the loop is connected to the shaft and the other to a slip ring insulated from the shaft. The current which flows in the loop is conducted to the external circuit by brushes which make sliding contacts with the slip ring and the shaft.

We shall consider the first position of the loop as shown at A in this figure. The white conductor is at the top and the black conductor is at the bottom of the figure; the motion of the loop is clockwise. When in this position, a small movement of the loop will not cause any induced voltage, because its two sides are moving parallel to the lines of force and are not cutting them. As the loop rotates farther to the right, it begins to cut magnetic lines and a small voltage is induced in both the black and white sides of the loop. Since the white side is cutting the lines from top to bottom and the black side from bottom to top, the voltage induced in one side will be opposite in direction to that induced in the other, and the combined effect of the two voltages will be such as to cause a current to circulate through the loop and the external circuit.

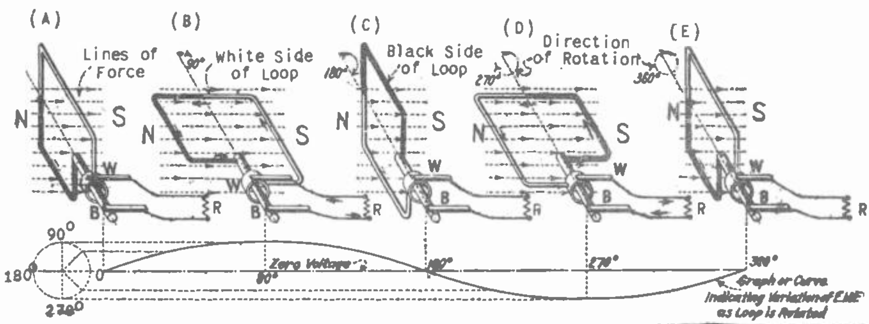


Fig. 1. Showing how a single loop of wire can generate an alternating voltage.

As the loop continues to move to the right, its sides cut an increasingly larger number of lines of force for a given amount of rotation. Thus, considering A as the zero position of the loop and B as the 90° position, it is evident that the sides of the loop will cut only a few lines as the loop moves from 0° to 1°, because it is moving nearly parallel to the lines of force; whereas when it moves from its 89° position to its 90° position, it will cut a large number of lines, because it is moving nearly perpendicular to the lines of force. The magnitude of the induced voltage depends on the number of lines of force cut in a unit length of time. Since it requires the same amount of time for the loop to move from 0° to 1° as it does from 89° to 90°, it is clear that a much larger voltage will be induced in the loop when it is near its 90° position than hen it is close to its 0° position. Thus, at the instant shown at A in the figure, the voltage is zero; and at B the voltage is maximum.

The loop now rotates from 90° to 180°, and in so doing, its sides cut fewer lines of force for each degree of rotation. This causes the induced voltage to decrease from its maximum value and finally to become zero as the loop reaches 180°. When the loop passes 180° and continues on toward 270°, its sides begin to cut the lines of force in the reverse order; that is, the white side is cutting them from bottom to top and the black side from top to bottom, or in the reverse direction as during the first half of the revolution. This causes the voltage induced in either side of the loop to reverse, and the current flowing through the loop and the external circuit also changes direction.

In rotating from 180° to 270°, the loop cuts a larger number of lines of force for each degree of rotation and a higher voltage is induced in the loop. At 270° this voltage is again maximum, but is opposite in direction from that induced at the 90° position. Finally, from 270° to 360°, the sides of the loop cut fewer lines of force and the voltage decreases becoming zero as 360° is reached.

Thus, by rotating a loop of wire in a magnetic field, an alternating voltage has been produced; one complete revolution of the loop constituting one cycle of voltage.

(C) Direction of the Induced Voltage. According to Lenz's Law, the direction of the induced voltage must be such that the current it causes to flow will create a magnetic field which opposes the motion of the loop. Let us examine Fig. 2. This figure

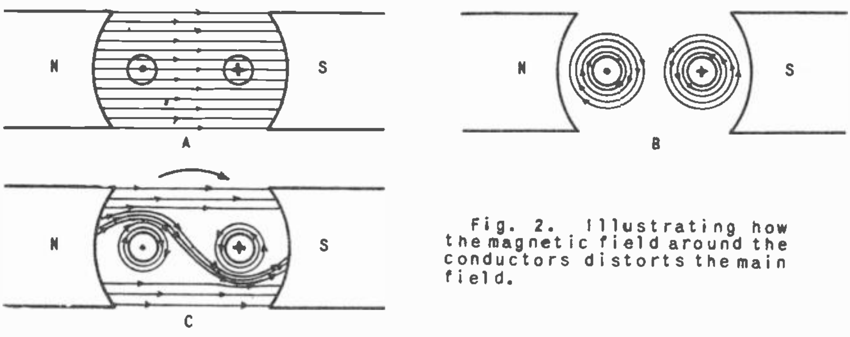


Fig. 2. Illustrating how the magnetic field around the conductors distorts the main field.

shows the field poles of the generator and the cross sections of the sides of the loop. (The loop is assumed to be at its 90° position.) At A in this figure is shown the magnetic field of the two field poles extending from left to right or from north to south. Let us now assume that the current in the right side of the loop (the white conductor) flows away from the reader. This is represented by placing a small cross on the end of this conductor. The magnetic field produced by this current would, according to the left hand rule, flow in a counter-clockwise direction about this side of the loop. With the direction of current in this side of the loop assumed, it is obvious that the current in the opposite or left side must flow toward the reader. Its direction is represented by placing a small dot on the end of the left conductor. Likewise, the magnetic field around the left conductor would consist of concentric circles and the direction of flow would be clockwise. The magnetic field produced by the current in the loop could therefore be represented by B in this figure. (The lines of force due to the field poles have been omitted in this diagram so that the field around the sides of the loop could be visualized more readily.)

Thus, there are two separate magnetic fields between the field poles; that due to the steady magnetic field of the field poles, and that due to the current flowing through the loop. By superimposing these two fields, a combined field is produced which is somewhat distorted. This resultant field is shown at C in this figure. It is not possible for magnetic lines of force to cross each other, and so the resultant field is due to the combined effects of both of the separate fields, and is strengthened at some points where the two fields add and weakened at other points where the fields buck. For example, around the right conductor, the field due to the current flowing through this conductor is from right to left above the conductor, and is therefore bucking against the main field due to the field poles. This causes the resultant field to be weaker above the right side of the loop, and is so represented by fewer lines of force at this spot in the resultant field. Conversely, at the bottom of this side of the loop, the field due to the current is from left to right. This is in the same direction as the main field, and the resultant field is stronger at this point, as indicated by a greater number of lines of force below the right conductor in the diagram illustrating the resultant field.

Of course, this same distortion of the resultant field also occurs around the left side of the loop with the result that the field is strengthened above the left conductor and weakened below it.

Now let us see whether we have assumed the correct direction for the current flow in the loop of wire. The loop rotates clockwise and if the direction of current flow is correct, the resultant field will be such as to tend to prevent the rotation. As learned in Lesson 9 of Unit 1, magnetic lines of force act like stretched rubber bands in that they always attempt to shorten their length. With this thought in mind, it is not hard to see that there will be a force on the right side of the loop tending to cause it to move upward, and a force on the left side tending to cause a downward movement. Thus, the force exerted by the magnetic field is such as to tend to cause the loop of wire to rotate counter-clockwise, or to oppose the motion originally given to the loop; therefore, Lenz's Law has been satisfied and the direction of current as assumed is the direction that current will actually flow.

The information given in the previous paragraph is very important. It illustrates in a graphical manner why it is not possible to create electrical energy. When the external circuit is open, no current flows in the loop, although voltages are induced in it as before. With no current flowing through the loop, no magnetic field is set up around it, and no distortion of the main field occurs. Thus, the generator is not furnishing any electrical power; there is no magnetic force tending to prevent the rotation of the loop; and the only energy required to rotate the loop is the small amount necessary to overcome the friction of the bearings.

With a large resistance connected in the external circuit, a small current flows, and a slight distortion of the main field takes place. The generator supplies a small amount of electrical power to the external circuit; a small magnetic force tends to oppose the rotation of the loop; and extra mechanical energy is needed to cause the loop to rotate against the opposition of the magnetic force. As the resistance in the external circuit is reduced, more current flows, more power is supplied to this resistance, the distortion of the main field increases, the opposing magnetic force becomes greater, and a larger amount of mechanical energy is required to cause the loop to rotate. For this reason, it is not possible to secure electrical energy from a generator without expending a corresponding amount of mechanical energy. In fact, if one horse power of electrical energy is being drawn from the generator, somewhat more than this amount of mechanical energy must be supplied, because the efficiency of the generator is not 100%.

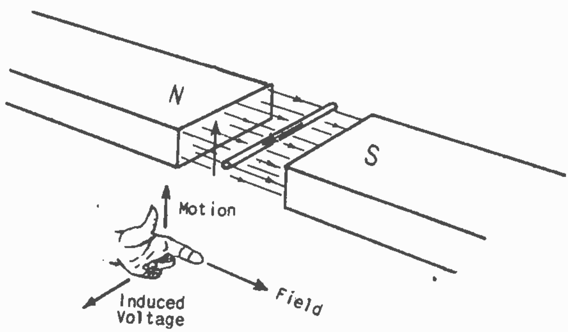


Fig. 3. The left hand rule for determining the direction of the induced voltage.

There is a very simple method of determining the direction that current will flow in a conductor when it is moved through a magnetic field. It is known as the "left hand rule", and is used in this manner: The thumb, forefinger, and middle finger of the left hand are extended so that each points in a direction at right angles to the other two, as illustrated in Fig. 3. If the thumb is pointed in the direction of the motion, and the forefinger in the direction of the magnetic flux from the field poles, then the middle finger will point in the direction of the induced electromotive force, or the direction in which current will flow if the external circuit is closed. Trying this rule on the right side of the loop shown in Fig. 2, we extend the left forefinger from left to right (in the direction of the main flux); the thumb downward (in the direction of the motion); and the middle finger will then extend away from us indicating that the induced electromotive force in this side of the loop is such as to cause the current to flow into the paper.

3. THE DC GENERATOR. It is evident that the simple generator shown in Fig. 1 will create an alternating voltage and cause an alternating current to flow through the external circuit. This is due to the fact that the voltage induced in the two sides of the loop changes direction as the loop passes through its 180° position. If it were possible to reverse the connection of the loop to the external circuit at the instant that the loop passes through 180°, the current in the external circuit would flow in the same direction during the last half of the revolution as during the first half, even though the voltages induced in the sides of the loop are

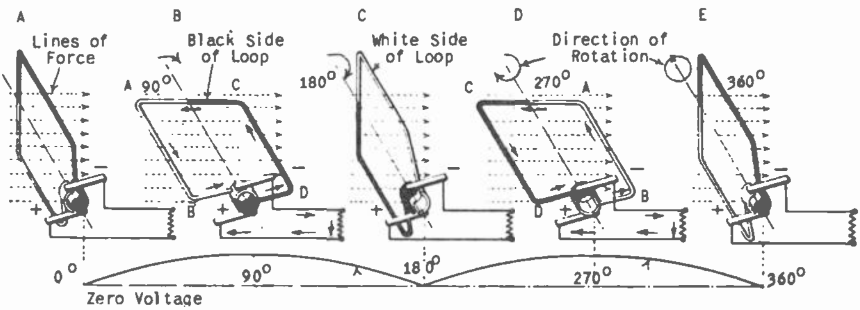


Fig. 4. Showing how the commutator causes a unidirectional current to flow in the external circuit.

reversed. This action is accomplished by a device called a "commutator". Fig. 4 shows the same loop of wire fitted with a commutator instead of slip rings. In this case, the commutator consists of a metallic cylinder on which the brushes make sliding contact. This cylinder is divided into two parts, electrically insulated from each other. Each half of the commutator connects to one side of the loop.

At A in this figure, the loop is in its 0° position and the induced voltage is zero. The brushes which connect to the external circuit are so arranged that each air gap which separates the two halves of the commutator is directly below a brush, when the loop is in this position. Thus each brush makes contact with both halves of the commutator and the loop is short circuited. As the loop begins to rotate to its 90° position, the air gaps are passed and the top brush makes contact only with that half of the commutator connected to the side of the loop marked AB, whereas the other side of the loop, CD, is connected to the bottom brush.

During this half of the revolution, the conductor AB is moving up through the field, and the other conductor CD is moving downward. According to the left hand rule, the voltage induced in AB is toward the reader, and that induced in CD is away from the reader. Thus, current will flow from AB to the top brush, through the external circuit, and back to the bottom brush to the conductor CD. The wave form of the current flowing in the external circuit is represented by a half sine wave during this part of the revolution.

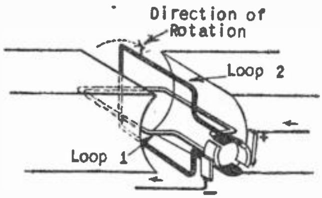


Fig. 5. Two loops of wire placed at right angles to each other.

When the loop reaches the 180° position shown at C, the induced voltage has dropped to zero, the air gaps in the commutator are again directly beneath the brushes, and the loop is short-circuited. As the loop continues to rotate, the air gaps are passed, the top brush now makes contact with that half of the commutator connected to the conductor CD, and the bottom brush is connected to the conductor AB. During this half of the revolution, the conductor CD is moving up through the field, and the conductor AB is moving downward. This causes the voltage induced in CD to be toward the reader, and that in AB to be in the opposite direction. Thus, current flows from CD to the top brush, through the external circuit in the same direction as before, and back to the bottom brush and the conductor AB.

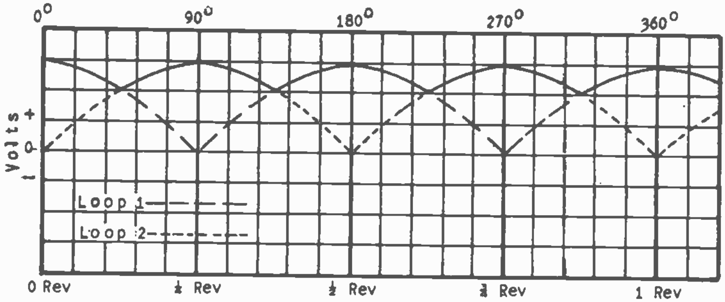


Fig. 6. The wave form of the voltages generated by the loops shown in Fig. 5.

It is, therefore, evident that the action of the commutator is to make the current flow through the external circuit in the same direction even when the current flowing through the loop reverses. The wave form of the current flowing in the external circuit is also illustrated in this diagram. It is practically the same as that produced by a full wave vacuum tube rectifier, and is by no means a pure direct current.

To produce a practically continuous E.M.F., it is necessary that more than one loop of wire be revolved through the magnetic field. Consider the diagram shown in Fig. 5. Here two loops of wire placed at right angles to each other are rotated in the magnetic field. There are four conductors which cut the magnetic

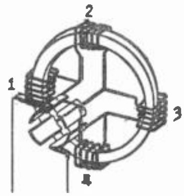


Fig. 7. A closed-coil generator having four coils.

field, and there must be four separate sections to the commutator. In the position shown, loop 1 is cutting lines of force at the fastest rate and consequently has a maximum voltage induced in it which causes a fairly large current to flow from the brushes which are making contact only with this loop. At this time, no voltage is induced into loop 2. As the two loops rotate to the right, the voltage induced into loop 1 decreases and that induced into loop 2 increases. At the end of one-eighth of a revolution, both loops are cutting the field at the same rate and the voltage induced in each is the same. The voltage induced in loop 2, however, has, as yet, served no useful purpose because its ends have not been connected to the brushes. A further rotation of the two loops will cause the voltage induced in loop 2 to be greater than that induced in loop 1, and the brushes are so arranged that at this point they leave the sections of the commutator connected to loop 1 and begin to make contact with those connected to loop 2.

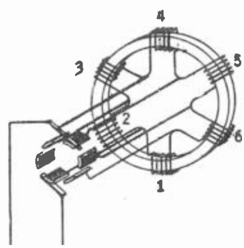


Fig. 8. A six-coil generator of the closed-coil type.

From the foregoing, it is clear that the voltage across the brushes never falls to zero, even though the voltages induced in the two loops are zero twice during every revolution. As soon as the voltage in the loop connected to the brushes is equal to that induced in the other loop, the brushes pass the air gap between the sections, and the voltage induced in the second loop which is still increasing forces current through the external circuit. The wave form of the voltages induced in the two loops are shown by the dotted line curves of Fig. 6. The actual voltage present across the brushes is the solid line curve which connects the tops of all these alternations. Thus, although the voltage across the brushes is not a pure DC voltage, nevertheless, it is more nearly constant

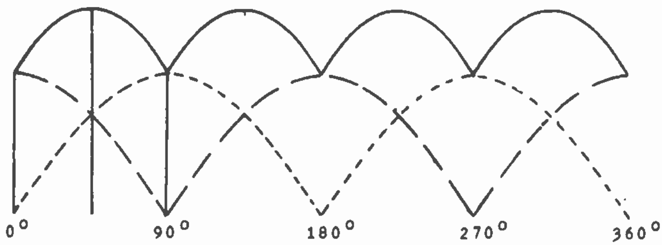


Fig. 9. The output voltage of the generator shown in Fig. 7.

in this case than when a single loop is used, and by rotating enough conductors so placed that at every instant the voltage induced in two of them is a maximum, it would be theoretically possible to produce a pure direct voltage.

Actually, of course, it is never possible to produce a pure DC with any commercial generator; there will always be some commutator ripple, as it is called, present in the output wave form, although the more conductors that are rotated, the more nearly constant the voltage will be. Thus, when the output of a DC generator

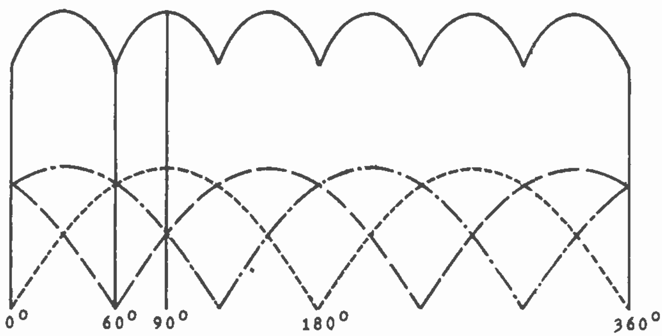


Fig. 10. The output voltage of the generator shown in Fig. 8.

is used to supply plate voltage for vacuum tubes, it is essential that at least a one-section filter be employed to smooth the output of the generator into pure DC.

The coils or loops shown in Fig. 5 are called "open coils" and the winding is said to be an "open-coil winding". Each coil is open circuited except during the time that it is connected to the brushes. Such an arrangement is rarely used in commercial machines; instead the so-called closed-coil winding is employed. In order to strengthen the field of the generator as much as possible, the conductors which cut through the lines of force are wound on a laminated, soft-iron core. The presence of this core reduces the reluctance of the magnetic path and concentrates the lines of force. This core, together with the coils which are wound on it, is known as the "armature" of the generator. In a later section of this lesson, we shall discuss the construction of armatures used on commercial generators.

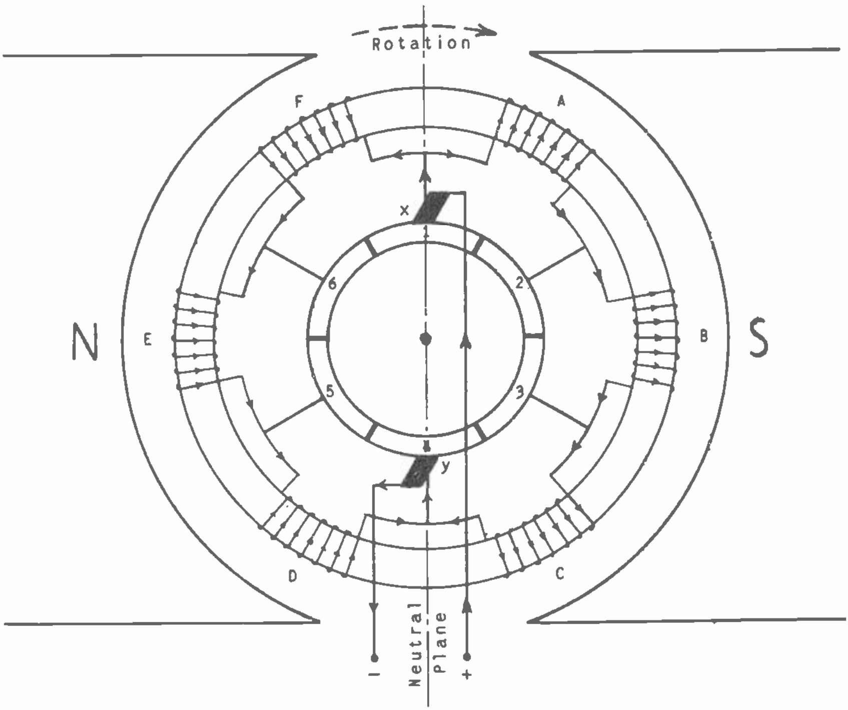


Fig. 11. A six-coil generator with a ring-wound armature.

Nearly all commercial DC generators have what is known as a "closed-coil winding". A four coil generator wound in this manner is shown diagrammatically in Fig. 7. Notice that all of the four coils are in series, and that the four junctions between the coils are each connected to a segment of the commutator. Thus every coil is in the circuit at all times, and the total voltage across the brushes is equal to the sum of the voltages induced in each coil. A six coil generator of the same type is illustrated in Fig. 8. The brush voltages of each of these two machines are shown in Figs. 9 and 10 respectively. The closed coil type of winding produces a somewhat greater E.M.F. at the brushes than does the open coil type. This is due to the fact that all the coils are in series and the total voltage across the brushes is the sum of that induced in each pair of coils.

Just how the closed coil generator delivers a voltage to the brushes is illustrated in Fig. 11. This machine is a six-coil generator with a closed-coil winding. Each coil consists of six turns. Coils B and E are cutting through the field at the greatest rate and therefore have the maximum amount of induced voltage. Let us assume that the voltage induced in each of these two coils is 50 volts. Coils A and C will have a lower induced voltage, because they are not moving perpendicular to the lines of force; we shall assume that the voltage induced in each is 25 volts. Likewise both coils D and F will also have an induced voltage equal to 25 volts.

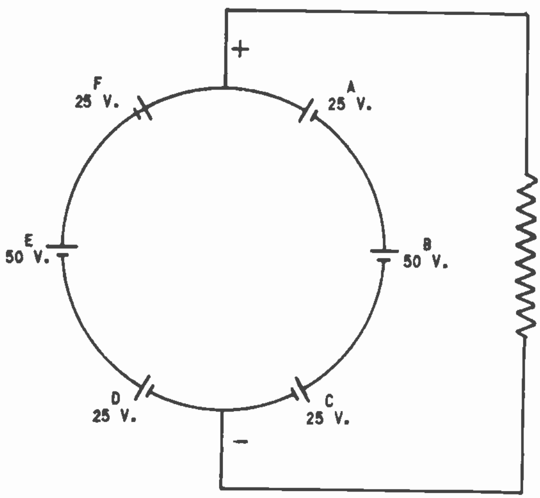


Fig. 12. Illustrating how the voltage of the generator of Fig. 11 is developed.

The total voltage between brush X and brush Y is the sum of the voltages induced in coils A, B, and C, or is 25 + 50 + 25, or 100 volts. The sum of the voltages induced in coils D, E, and F is also 100 volts, but these coils are effectively in parallel with the first three coils as far as the external circuit is concerued, and they do not add to the total voltage. Perhaps this may be made somewhat clearer if each coil is represented by a battery whose voltage is equal to the voltage induced in that particular coil. Such a circuit is shown in Fig. 12. Here there are six batteries connected in a series-parallel arrangement. Batteries A, B, and C in series are connected in parallel with batteries D, E, and F, which are also in series. The total voltage across the load resistor is equal to 100 volts, although any one battery supplies only one-half of the total load current.

The same is true of the generator shown in Fig. 11. There are two separate paths through the armature winding, and any one coil furnishes only one-half of the total current flowing to the brushes. Current flows out of the brush marked Y, through the external circuit and back to the brush labeled X. From this brush, the current flows to segment 1 of the commutator to the junction of the coils F and A, at which point it divides; one-half of the total current flowing through coils A, B, and C to segment 4 of the commutator, and the other half through coils D, E, and F to segment 4. From segment 4 of the commutator, both currents combine and the total current leaving brush Y is the same as the current entering brush X.

The generator of Fig. 11 has a ring-wound armature. Most commercial machines, however, are of the drum-wound type. The distinction between these two is shown by the diagram of Fig. 13.

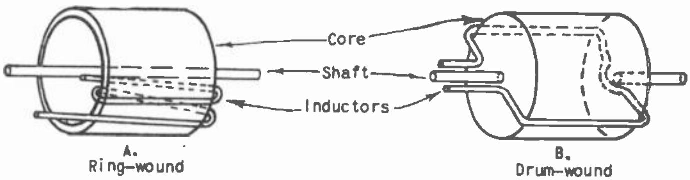


Fig. 13. Showing the difference between a ring-wound armature and a drum-wound armature.

At A is the ring-wound type and at B is shown the drum-wound variety. The ring-wound armature is wound on a hollow, soft-iron cylinder. The cylinder is not solid but is built up of laminations. The turns of the coils are wound around this iron cylinder and there are conductors on the inside as well as the outside of the cylinder. There is, however, practically no magnetic flux on the inside of the cylinder, because the cylinder acts as a magnetic shield. Therefore, only one side of each turn or coil cuts through the lines of force, and those sides lying on the inside are ineffective in producing a voltage. In the drum-wound armature, on the other hand, all of the conductors lie on the outside of the armature core, and are thus effective in producing a part of the brush voltage. The principle of operation of the two types is the same, but due to the advantage of the drum-wound type in requiring a smaller amount of copper for the same generated voltage, and the fact that it is much easier to wind, has made this type of armature the more popular, and it is used almost entirely.

4. COMMUTATION. The principle of commutation has already been discussed briefly. In this section we shall inquire into the requirements for satisfactory commutation, and a later section will deal with the construction of commutators. For this explanation, we shall use the generator shown in Fig. 11, but the process will be clearer if we consider that the armature is cut open at the center of segment 1, and is laid out in a straight line. When this is done, the six coils and their commutator sections appear as shown

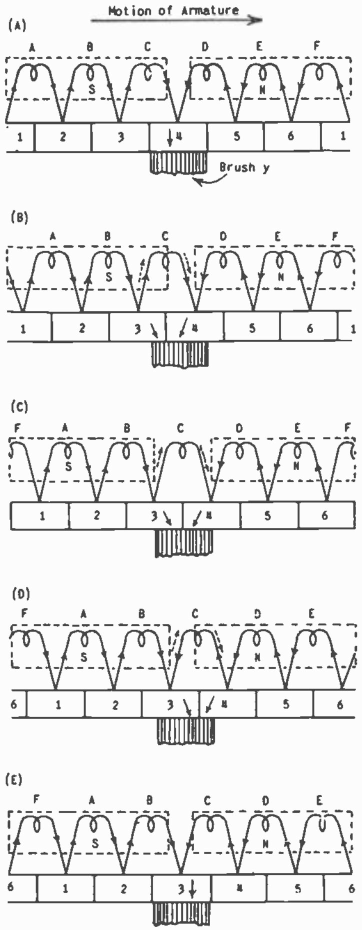


Fig. 14. Illustrating the process of commutation.

at A in Fig. 14. In the diagram of the generator, the brush Y is making contact with segment 4, and it is so shown in this second figure.

At the instant shown, the current in coils A, B, and C is flowing to the right toward segment 4 of the commutator, whereas that in coils D, E, and F is flowing to the left toward this segment. The coil to be commutated is coil C. The field poles are represented by dotted-line rectangles; thus a south pole lies behind coils A, B, and C, and a north pole behind coils D, E, and F. The direction of motion of the armature is from left to right.

We have assumed that the width of the brush is equal to the width of one of the commutator sections; this may or may not be true. The brushes are ordinarily made of carbon and the resistance of the contact that a brush makes with a commutator segment depends entirely on the surface area of that brush that is in contact with the segment. At B, the armature has moved on until one-quarter of the brush is making contact with segment 3, and three-fourths of its area is making contact with segment 4. In this position, all of the current from the right side of the armature will flow directly into segment 4, whereas the current coming from the left side of the armature will divide; one-half flowing directly through segment 3 to the brush, and the other half flowing through coil C and then through segment 4 to the brush. Thus, the current flowing through C is only half as much as that through the other coils.

As the armature continues to rotate, it reaches the position shown at C. In this position, one-half of the brush makes contact with segment 3, and one-half with segment 4. Coil C is effectively shorted out, since the contact resistance of the brush is considerably less than that of the coil, when the brush is making equal contact with both segments. Therefore, the current flow in coil C stops; all the current from the left side of the armature flows through segment 3 to the brush, whereas that from the right side flows through segment 4.

The next position of the armature is illustrated at D. In this case, three-quarters of the brush makes contact with segment 3 and one-quarter with segment 4. The current from the left of the armature flows into segment 3, and that from the right side divides, part flowing directly into segment 4, and part flowing through coil C to segment 3. Finally, the armature reaches the position shown at E. The brush has left segment 4 and is making full contact with segment 3. The current from coils F, A, and B flows from left to right into segment 3, and that from coils C, D, and E flows from right to left into this same segment.

The important point to observe is that during the time that coil C is commutated, the current through it must completely reverse its direction and must build up to a value in the opposite direction equal to the value it had before the coil was commutated. At A, the current through coil C is from left to right and is of the same value as that in the rest of the coils; but at E, the current in coil C is from right to left and is also of the same value. Therefore, it is clear that the current in coil C must reverse direction very rapidly when the armature is rotating at high speed. Naturally, since there are two brushes, there are two coils being commutated at the same time. In this illustration, coil F was commutated at the same time as coil C, but the second brush was left out of the drawing to simplify the explanation.

To illustrate just how fast the current through the commutated coil must change, let us suppose that we have a machine rotating 1800 r.p.m. This speed corresponds to 30 revolutions per second, or one revolution will require $\frac{1}{30}$ of a second. Now suppose that the armature has 90 coils, which will require 90 commutator segments. If the width of the brush is equal to that of a commutator segment, the time during which any coil is shorted out, is the time needed for the mica insulation between two segments to travel the width of the brush. Since there are also 90 pieces of mica insulation, the time required for one piece of mica to travel the width of the brush would be $\frac{1}{90}$ of the time of one revolution, or $\frac{1}{2700}$ of a second. This is obviously a very short interval of time. If the current in the coil before it is commutated is 5 amperes, this current must fall to zero and rise to 5 amperes in the opposite direction during this small time. The current in the coil must change a total of 10 amperes in $\frac{1}{2700}$ of a second, or the rate of change of the current is 27,000 amperes per second. Even though the coil may have very little self-inductance, it is apparent that a fairly high induced voltage will accompany this rapid current change.

As the current in the coil starts to decrease, an induced voltage is set up which is in such a direction as to tend to prevent the current from falling. This induced voltage is represented at B by the dotted arrows drawn alongside the commutated coil. This voltage persists even when the current in the coil has dropped to zero as is shown by the arrows at C. Then as the current through the coil begins to build up in the opposite direction, this same induced voltage is present, since it now tends to prevent the rising of the current. Unless the current in the coil is able to rise to its final value by the time that the brush has left the segment which it is moving away from, severe sparking between the trailing edge of the brush and the commutator segment will occur. This sparking quickly causes the commutator segments to become badly pitted, and produces uneven wearing away of both the commutator and brushes.

The appearance of a badly pitted commutator caused by excessive sparking is illustrated in Fig. 15. It is evident that there should be no voltage across a coil when it is being commutated, because it is this voltage which causes the sparking. If the coil had no self-

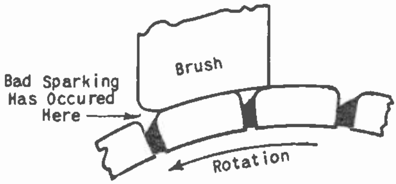


Fig. 15. Showing the effect of severe sparking.

inductance, this condition could be satisfied by commutating the coil at the time that it was not cutting any lines of force. Sometimes, the inductance of the armature coils is low enough that the induced voltage may be neglected, and in this case, the brushes are set on the so-called neutral plane so that the coils will be commutated when they are not cutting flux. This neutral plane is illustrated in Fig. 11.

Often, the induced voltage in the commutated coil is great enough that it cannot be neglected. In such a case, the only solution is to induce a voltage into the coil equal and opposite to that produced by the self-inductance of the coil. The net voltage in

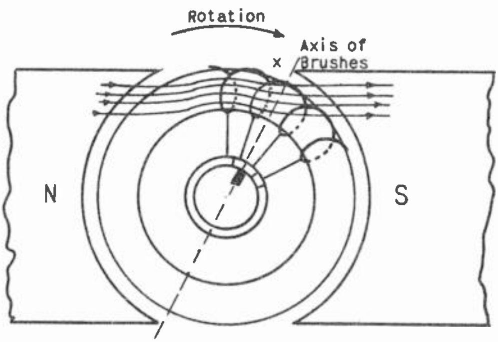


Fig. 16. Shifting the position of the brushes in the direction to reduce sparking.

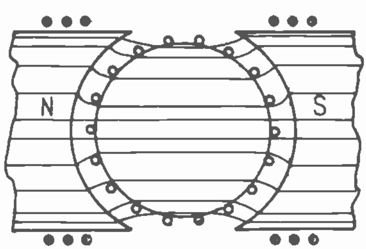
the coil will then be zero, and sparking will not take place. Notice from Fig. 14 that the induced voltage of coil C, which is is being commutated, is opposite in direction to the voltage that is present in coils D, E, and F. Therefore, if the brushes are so placed that the coil is cutting a small amount of the flux in the same direction as the coils D, E, and F, it is evident that the voltage induced in coil C, due to the fact that it is cutting flux, may be made equal to the self-induced voltage of this coil, and there will be no net voltage in the coil to cause sparking.

This condition is satisfied by advancing the position of the brushes in the direction of rotation until the sparking is minimum. This condition is illustrated in Fig. 16. The angle through which the brushes are shifted is not constant but depends on the load placed on the generator. For example, suppose that the load on the generator is increased until the current in each coil is 10 amperes instead of 5 amperes. With the increased load, the current through the commutated coil must change a total of 20 amperes during the commutation process and in the same length of time. Since the amount of current change is twice as great, it is evident that the self-induced voltage will also double its value. To neutralize the effect of this self-induced voltage, the coil must be cutting more flux during commutation, and the brushes will have to be shifted further in the direction of rotation for this to be accomplished.

We shall learn that this is not the only reason that shifting of the brushes is necessary for sparkless commutation. In fact, the major reason for the brush shift is due to another factor which we shall now consider in detail.

5. ARMATURE REACTION. When the armature of a generator is carrying current, it produces a magnetic field which alters both the distribution and the magnitude of the field which would be produced by the field poles themselves. This action of the armature in creating this change in the main field is called the "armature reaction".

Fig. 17. The distribution of the magnetic field when the armature is not carrying current.



The effect of the armature reaction is two-fold. It causes a distortion in the distribution of the lines of force, and may also cause an actual weakening of this field. Fig. 17 shows the undistorted field produced by the field poles when the armature is not carrying current. This field is of practically uniform intensity, and extends in nearly straight lines from left to right. It should be noticed that the air gap between the pole shoes and the armature is purposely small so that the field strength will be as large as possible. The armature shown in this diagram is of the drum-wound type.

When current is drawn from the armature, a field is created around the armature winding as shown in Fig. 18. In this figure, the lines of force of the main field have been omitted to simplify the diagram. Notice that this field, due to the armature current, is at right angles to the main field.

When the armature is rotating and delivering current, both of these magnetomotive forces exist simultaneously with the result that the resultant field will be similar to that shown in Fig. 19. It is seen that the armature reaction causes the field to be rotated in the direction of rotation. The field around the upper tip of the north pole is strengthened, whereas, that around the upper tip

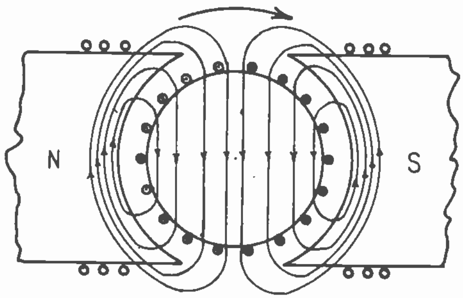


Fig. 18. The field due to the armature current.

of the south pole is correspondingly weakened. Furthermore, the neutral plane is no longer perpendicular to the field poles, but is also shifted in the direction of the rotation.

If the brushes are placed on the neutral plane of the field when the armature is not carrying current, it is evident that sparking will occur when the machine is loaded. Perhaps this will be made clearer by again referring to Fig. 19. If the brushes are

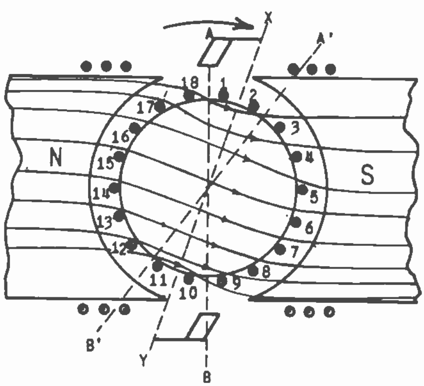


Fig. 19. The field distribution due to the combined effect of the field flux and the armature flux.

placed on the line AB, and the coils are commutated as they pass this line, they will still be cutting flux at this time. Also, the voltage induced in a coil as it passes this point is in the same direction as the self-induced voltage of the coil; therefore, instead of neutralizing the self-induced voltage and producing no net voltage in the coil at the time it is commutated, the voltage produced by the flux aids the self-induced voltage and causes the sparking to be very severe.

This can be remedied by advancing the brushes until they again lie on the neutral plane, which has now been shifted in the direction of the rotation to xy. Of course, if the coils have much self inductance, the brushes should be advanced beyond this new neutral plane to A'B' so that they will be cutting some flux in the right direction to neutralize the self-induced voltage. This, then, is the principal reason of shifting the brushes forward, in the direction of rotation, in a direct current generator. The armature reaction twists the field, and to avoid sparking, the brushes must be moved to a different position to correspond with the new direction of the field.

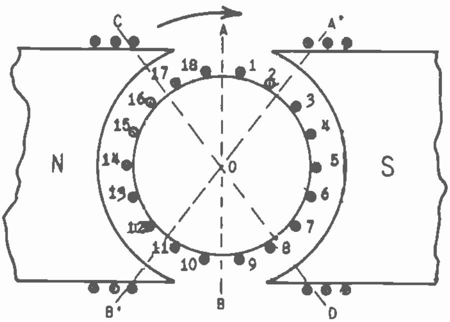


Fig. 20. Illustrating the effect on the armature current distribution caused by shifting the brushes.

Shifting the brushes forward produces a change in the current distribution of the armature. This is illustrated in Fig. 20. In Fig. 19, where the brushes were on the line AB, conductors 1 and 2 were carrying current away from the reader, whereas they are now carrying current toward the reader. Likewise, conductors 10 and 11, which were carrying current toward the reader, are now carrying current away from the reader.

We may now consider the armature conductors in two groups. Considering the conductors 17 and 11 as one turn, conductors 18 and 10 as another turn, etc., we find that the conductors lying in the angle COA' combined with those in the angle B'OD, make up a set of turns whose magnetomotive force is directly opposed to that of the main field, with the result that the main field is weakened. The turns included in this angle, which is twice the angle through which the brushes have been shifted, are called the "demagnetizing turns" or the "back-ampere turns". On the other hand, those conductors lying in the angle A'OD combined with those in angle COB', constitute a set of turns whose magnetomotive force is at right angles to the main field, and these turns are called the "cross-magnetizing" turns.

All of these steps follow in logical order. First, the field due to the armature causes a twisting of the main field; second, the brushes are shifted forward to prevent sparking; third, the current distribution of the armature is changed, with the result that the demagnetizing turns weaken the main field. It should be evident that shifting the brushes in the direction opposite to that of the rotation would cause the current in the demagnetizing turns to reverse, and the magnetomotive force of these turns would then assist or strengthen the main field. This, however, would cause such a large amount of sparking that the commutator would soon be ruined.

6. COMMUTATING POLES. The necessity of shifting the brushes forward has been covered at great length in the preceding sections, Furthermore, it should be apparent that the amount of brush shift will depend directly upon how much the generator is loaded or how much current is being drawn from the armature. The greater the load on the machine, the more the main field will be twisted due to armature reaction and the larger the angle through which the brushes must be shifted. If the load on the generator is of a variable nature, a constant changing of the brush positions will be necessary.

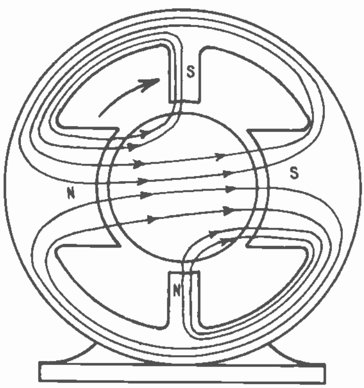


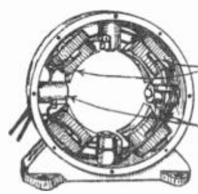
Fig. 21. The field distribution of a two-pole generator employing commutating poles.

To obviate this difficulty, most modern machines use commutating poles. These are small poles placed midway between the main field poles, and are excited by windings which are in series with the armature. The resultant field pattern of a generator using commutating poles is shown in Fig. 21. These small poles are sometimes called "interpoles", and there are as many of them used as there are main field poles. They must produce enough flux to neutralize the cross magnetization around the commutated coil, and also induce sufficient voltage into the coil to counter balance the effect of the self-induced voltage. If they are properly designed, the coil can be commutated when it is on the normal neutral plane, and practically no sparking will occur.

To determine the correct polarity of the commutating pole, it should be remembered that to produce sparkless commutation, it is necessary to advance the brushes forward so that the coil being commutated is in the pole fringe of the next succeeding main pole; therefore, the commutating pole should have the same polarity as the next pole under which the coil will pass. Since the windings of these poles are in series with the armature, their field will be stronger as the generator is more heavily loaded, and the E.M.F. induced into the short-circuited coil becomes larger. This is necessary because a heavier load magnifies the twisting of the main field and a larger voltage is necessary (from the commutating pole) to neutralize the voltage produced as the coil cuts the main flux. Thus, throughout a limited range, the tendency of a varying load to produce sparking at the commutator is cancelled, and shifting of the brushes is unnecessary. It should, of course, be realized that the automatic compensation provided by the commutating poles is limited, and if the generator is either underloaded or overloaded, sparking will take place.

7. CLASSIFICATION OF DC GENERATORS. DC generators may be classified by any one of several different methods. They may be classified according to the number of poles they have, according to whether the field is separately excited or self-excited, or according to the way in which the field is connected.

Fig. 22. A four-pole generator fitted with commutating poles.



All of the generators which we have been discussing are bipolar; that is, they have just two main field poles. Ordinarily, however, a DC generator will have more than two and may have four, six, eight, or as high as twenty-four field poles. They must, of course, have an even number of poles and the poles are so arranged around the armature that they alternate in polarity. The field structure of a four-pole generator fitted with interpoles is shown in Fig. 22.

For a given size of machine, the greater the number of field poles, the higher the output voltage will be with a given speed of revolution. Therefore, the choice of the number of field poles will depend upon the source of mechanical power which is to drive the generator. When a generator is driven by a reciprocating steam engine, it has a large number of field poles because the speed of snch an engine is comparatively slow. On the other hand, generators to be driven by steam turbines, which travel at high speed, will have only a small number of poles, perhaps 4, or 6. Nearly all small generators of less than 5 horsepower capacity are bipolar.

If the field coils of a generator are excited by batteries or from some external source, the generator is said to be separately excited. Separately excited generators find but little application and are mainly used for electroplating.

The field coils of all commercial machines are self-excited; that is, the current for the field winding is derived from the voltage produced by the generator. Perhaps it is not clear how the voltage produced by the generator may be used to supply the field, since the field must be present before this voltage can be produced. There is, however, some residual magnetism present in the field structure of any generator that is in constant use, and this initial field causes a small voltage to be produced in the armature which further excites the field and causes a larger armature voltage. Thus, it is evident that the field builds up as the rotation of the armature is begun, and this building up process continues until the generator is producing its normal voltage output. Sometimes, when machines are shipped from one place to another, the consequent jarring destroys the residual magnetism, and in order to cause the generator to build up, the field must be excited from an external DC source until the machine, after running for a few minutes, has replaced the residual magnetism.

There are three separate methods of connecting the field winding with respect to the armature winding. Machines using these three methods are called series, shunt, and compound generators. Each type of connection causes the machine to have a different operating characteristic; therefore, each type of generator will now be considered separately.

8. THE SERIES-WOUND DC GENERATOR. In the series-wound generator, the field winding is connected in series with the armature. The field winding consists of relatively few turns of heavy wire. It is necessary that large wire be used, because all of the cur-

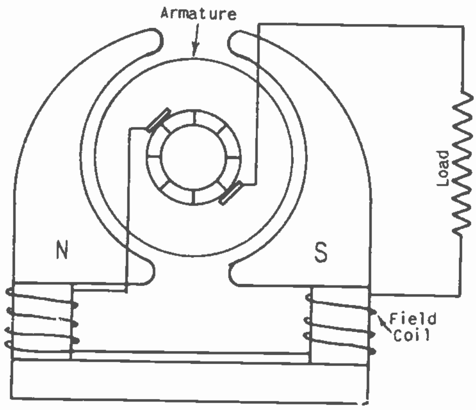


Fig. 23. Showing the connection of a series generator.

rent drawn from the generator must pass through the field coils. Since the current through the field winding is ordinarily rather large, it is not necessary to wind many turns on the field coil to obtain the desired field strength. Fig. 23 illustrates a series generator; the armature, load resistor, and the two field coils are all connected in series.

To understand the operating characteristic of any type of generator, its external characteristic must be plotted. If the load on a generator, running at constant speed, is increased, the terminal voltage, or voltage available for the external circuit decreases due to the increased voltage drop in the resistance of the armature windings and to the greater armature reaction. The curve which relates the terminal voltage to the generator's load is called the "external characteristic" of the generator. Such a curve for a typical series generator is shown in Fig. 24.

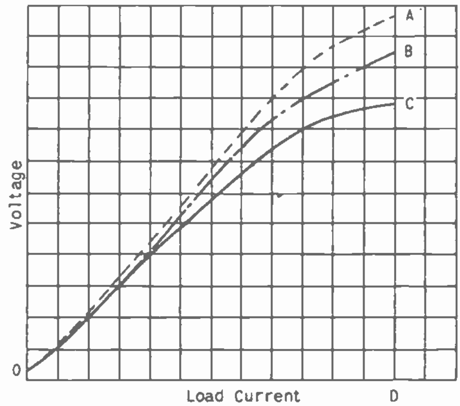


Fig. 24. External characteristic curve of a series generator.

When there is no load on the generator, it is not delivering any current, no current flows through the field, and the output voltage is limited to that which can be produced by the residual magnetism of the field. As the load on the generator increases, more current is drawn from the armature, and since this current must flow through the field windings, the strength of the field increases and causes a larger output voltage to be produced. Thus, increasing the load on a series generator, increases the terminal voltage of the generator. This building up of the output voltage with increased loads cannot continue indefinitely, because the field cores finally become saturated, and a further increase in the load is no longer able to increase the field strength. When this point is reached, a further increase in the load will cause a decrease in the output voltage, due to the greater losses in the armature.

The machine is so designed that the maximum load which may be safely drawn from the armature is nearly sufficient to cause saturation of the field.

Let us now inspect the external characteristic curve of the series generator which is shown in Fig. 24. It is noticed that there are three curves on this graph. The solid line curve is the actual characteristic of the generator and it shows what the terminal voltage will be for different values of load current. The curve OA is the curve which would be produced, if there were no losses in the generator due to resistance and armature reaction. Point D represents the full load current, and the distance CD is the actual output voltage at this value of load. If there were no losses in the generator, the full load voltage would be equal to the distance

AD. Thus, the distance AC is the total voltage drop within the machine at full load due to all causes. If there were no resistance losses in either the armature or the field, the only loss would be that caused by armature reaction, and the curve OB would represent the characteristic of the machine. In this case, the distance BD would represent the output voltage at full load. Thus, the distance AB represents the voltage loss due to armature reaction, and the distance BC that due to resistance losses.

Notice that the characteristic curve is linear for a part of its length, but begins to flatten out as larger load currents are drawn from the generator. If the load were increased beyond the full load value (assuming that the generator does not overheat and is able to withstand a small overload), the output voltage would soon begin to drop due to the increased losses within the machine.

The wide change in output voltage which would occur with varying loads has prevented the series generator from obtaining much popularity. In fact, its present-day applications are limited to its use as a "series booster", whereby it is used to raise the voltage on long power lines which have suffered a voltage drop due to the resistance of the lines.

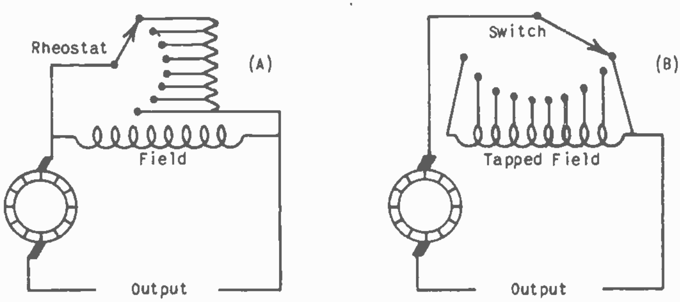


Fig. 25. Methods of controlling the output voltage of a series generator.

If it is desired to vary the output voltage while maintaining the load current constant, either of the two methods shown in Fig. 25 may be employed. In the first case, a rheostat is connected in parallel with the series field, and by changing the position of the rheostat, the amount of current flowing through the field and the field excitation may be changed. The second method uses a variable field winding, any part of which may be connected in the circuit by changing the position of a tapped switch. If either the amount of current flowing through the field or the number of turns in use is reduced, the output voltage will fall.

9. THE SHUNT-WOUND GENERATOR. In the shunt-wound generator, the field winding is connected in shunt or in parallel with the armature. The field is connected directly across the output terminals of the machine, and thus it draws a part of the load current. The current drawn by the field should be small so that the field

itself will not place much load on the machine. In order that the proper field strength may be obtained, the field must be wound with many turns of wire which may be of a small size, since the current in the field will not be large. Fig. 26 shows a shunt generator supplying a load.

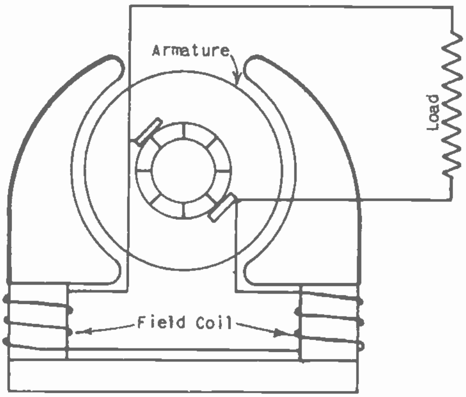


Fig. 26. Connections of a shunt generator.

When there is no load placed on the generator, the current flowing through the armature is limited to that required to supply the field winding, and the losses due to resistance and armature reaction are low. Consequently, the terminal voltage is high.

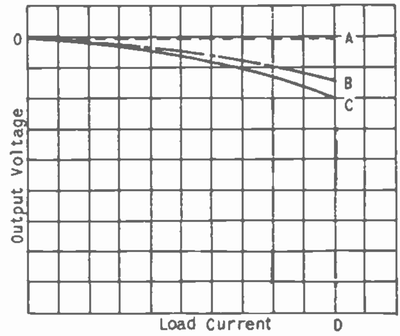


Fig. 27. External characteristic of a shunt generator.

When a small load is connected to the generator, the armature current increases, and a larger voltage drop occurs within the armature. This causes the terminal voltage to decrease, and since the voltage across the field is equal to the terminal voltage, it is evident that the field excitation is also decreased, which further reduces the terminal voltage.

The external characteristic of a shunt-wound generator is illustrated in Fig. 27. The solid line curve OC is the actual variation of the terminal voltage for various loads up to the maximum.

The characteristic of a shunt-wound generator is such that a much more constant voltage is obtained with varying loads than is the case of a series generator. The horizontal line OA is the characteristic which would be produced, if the machine had no losses, whereas the curve OB is the characteristic, assuming that the armature windings have no resistance. Thus, it is clear that the distance AC represents the total voltage loss at full load, BC the voltage loss due to armature resistance, and AB the voltage loss due to armature reaction, and to the fact the field current is decreased as the terminal voltage is reduced.

The output voltage of a shunt generator may be varied by connecting a rheostat in series with the field winding and changing the position of the rheostat. As more resistance is cut out, the field current increases, and the terminal voltage is raised. Of course, the rheostat cannot raise the voltage at full load, but it can reduce the voltage at lighter loads and thus tend to keep the output voltage constant with varying loads. Such an arrangement for controlling output voltage is shown in Fig. 28.

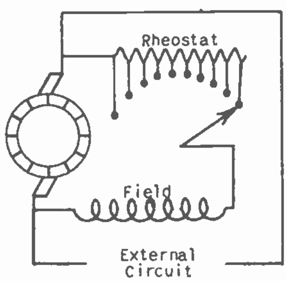


Fig. 28. Method of controlling the output voltage of a shunt generator.

The shunt-wound generator has a much more uniform characteristic than the series generator, and it is used almost entirely in radio installations, where the load does not vary through wide limits. In commercial installations requiring a generator to deliver varying amounts of power, the compound-wound generator is ordinarily employed.

10. THE COMPOUND-WOUND GENERATOR. The disadvantage of the shunt generator is that its terminal voltage falls with increased loads. This fall in voltage is due to three causes: the resistance drop in the armature; the armature reaction, and the decreased field current. If there were some way to cause the field current to increase enough to produce sufficient additional field strength to compensate for the resistance and armature reaction losses, the external characteristic of the generator would be a horizontal line and the voltage regulation would be zero. A voltage regulation of zero means that the full load voltage is equal to the no-load voltage. The voltage regulation of a generator is exactly the same as the voltage regulation of a power supply rectifier circuit, and the same formula is used. The smaller the per cent of voltage regulation, the more nearly constant the output voltage will be

with changing loads.

It is possible to construct a generator in which the output voltage is essentially constant at any load up to the maximum allowable. Such a machine is called a compound-wound generator, and

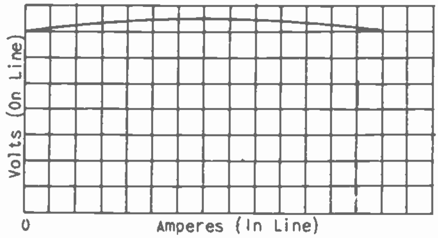


Fig. 29. External characteristic of a compound generator.

its field consists of two separate windings. Actually, it is a shunt-wound generator which has a few extra turns of heavy wire on its field poles which are connected in series with the armature. Such a machine will have the characteristics of both a series and a shunt-wound generator. The shunt field provides the correct amount of no-load field flux, whereas the series field, which carries the same current as the external circuit, increases the flux by a predetermined amount.

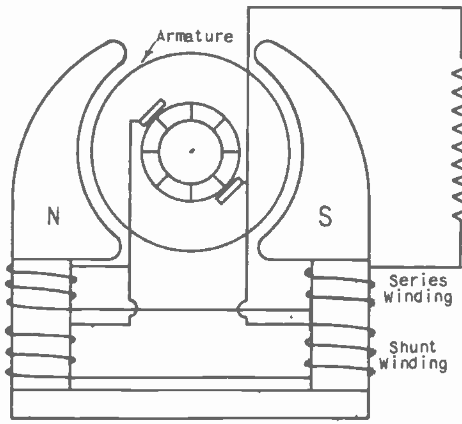


Fig. 30. Connections of a compound generator.

If the series field increases the field flux more than enough to compensate for the losses of the generator, the terminal voltage will rise slightly with an increase in the load, and the machine is said to be over-compounded. If the series field provides just enough flux to make the terminal voltage at full load equal to the no-load voltage, the generator is flat-compounded. Finally, if the increase in flux produced by the series field is not sufficient to overcome the losses of the machine, it is said to be under-compounded. The external characteristic of a flat-compounded generator is shown in Fig. 29. Fig. 30 illustrates a compound generator furnishing current to a load.

Many commercial DC generators are of the compound type, and nearly all of them are over-compounded slightly. Usually the actual load of a generator is located some distance away from the machine itself, and there is some voltage loss in the connecting wires. By

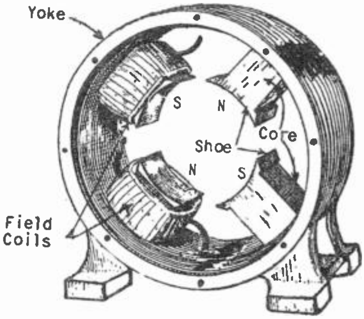


Fig. 31. Showing the construction of the field of a four-pole generator.

over-compounding the generator, the full-load terminal voltage will be somewhat greater than the no-load voltage, but the voltage across the load itself is practically constant due to the increased resistance loss in the connecting wires at full load.

The output of a compound - wound generator may be varied by a rheostat in series with the shunt field, or a rheostat in parallel with the series field; the former method, however, is nearly always used.

11. DETAILS OF CONSTRUCTION. *(A) The Field.* The field of a four-pole DC generator is shown in Fig. 31. The cylindrical frame

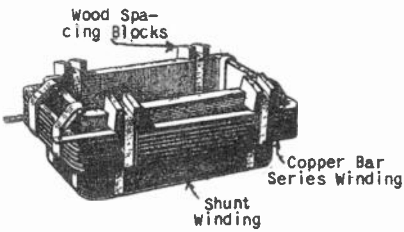
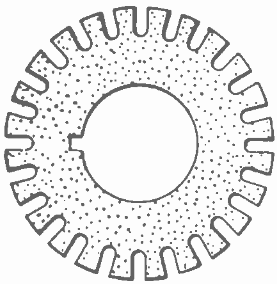


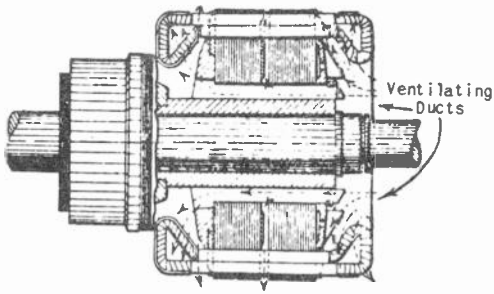
Fig. 32. The construction of a field coil.

is called the yoke and is made of cast iron. The field cores, as well as the pole shoes are built up of soft iron laminations. This is necessary to reduce eddy-current losses,[1] since the interaction of the main field and the field around the armature induces voltages in the field cores, which tend to overheat the machine. By using laminated cores, the effective lengths of the eddy - current paths are reduced and the heating is minimized.

[1] Refer to Section 7, Lesson 24, Unit 1.



A

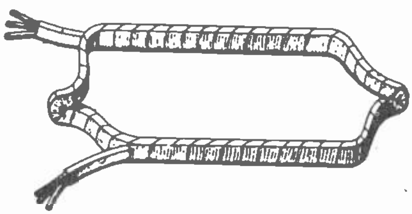


B

Fig. 33. (A) An armature core lamination. (B) An assembled armature.

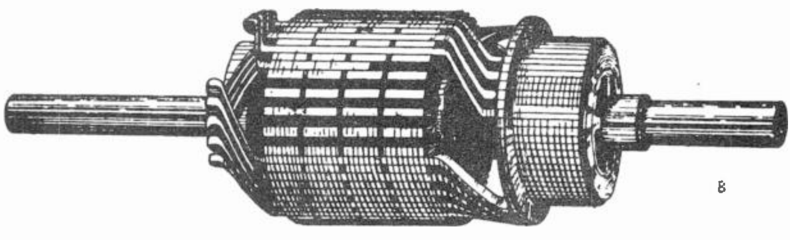
The field coils are wound with insulated copper wire or copper strap. Fig. 32 illustrates a field coil having a shunt and series winding. Often air spaces are left between layers or turns to facilitate ventilation.

(B) The Armature. The core of the armature should be of such a material that it will decrease the reluctance of the magnetic circuit. Also, since it will have voltages induced in it, it should not be solid, but should be built up of laminations. The laminations are composed of soft iron. At A in Fig. 33 is illustrated an armature core lamination. It is provided with slots into which the conductors are placed. There is less danger of the arma-



A

Fig. 34. (A) A form-wound armature coil. (B) A partly wound armature.



B

ture winding being torn apart by centrifugal force when the conductors are placed in slots instead of being wound on the surface of the core. At B in this figure is shown an assembled armature. Notice that ventilating ducts are provided.

Most armatures are wound with "form-wound" coils. Since all of the coils of a particular armature are of a given size and shape, they may be wound on a form by automatic machinery, and then placed in the slots of the armature core. Fig. 34 at A shows a form-wound armature coil, and a partly wound armature is illustrated at B in this figure.

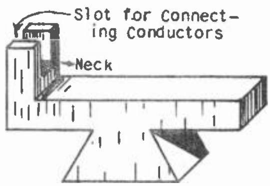


Fig. 35. A commutator segment.

(C) *The Commutator.* The commutator is the weakest component of any DC generator both electrically and mechanically. The commutator segments are composed of L-shaped copper bars; which, when placed together, form a hollow cylinder. One of these bars is shown in Fig. 35. The bars are insulated from each other and from the frame on which they are mounted by strips of mica. The mica must be of a good grade and should have the same wearing qualities as the copper; otherwise, the brushes will wear down the copper faster than they do the mica, and cause the mica insulation to project above the copper, with the result that excessive sparking occurs. To prevent this from happening, it is common practice to cut away the mica about $\frac{1}{16}$ inch below the surface of the copper. An undercut commutator of this type is illustrated in Fig. 36.

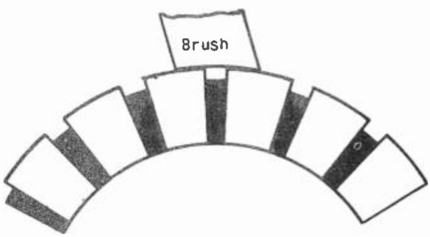


Fig. 36. An undercut commutator.

Care must be taken in the design of the machine that the voltage between any two adjacent commutator bars is not excessive. The mica insulation between the bars is sufficient to withstand several thousand volts, but a leakage path is liable to be formed over the surface of the mica. If oil gets on the commutator, the sparking of the brushes may carbonize it and form a layer of carbon over the surface of the mica.

Fig. 37 shows a sectional view of a commutator. The copper bars are held in place by a cast-iron spider. Pieces of mica insulate the bars from the spider. After the commutator has been assembled, it is placed in a lathe and turned until it is perfectly cylindrical.

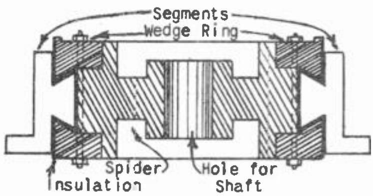


Fig. 37. A sectional view of a commutator.

(D) The Brushes. The brushes are the devices that make contact with the commutator segments and carry the current to the external circuit. They are generally made of carbon blocks, but may be made of copper leaves or copper gauze. All machines generating over 100 volts use carbon brushes. The copper brushes are more flexible, and if they pass over a rough spot on the commutator, only a part of the contact surface will be lifted from the segment. They cannot, however, be used with high-voltage machines without causing excessive sparking. On the other hand, the carbon brush is not flexible at all, and if it strikes a rough spot, the whole contact surface will be lifted from the commutator. To minimize trouble due to this source, very large machines employ several

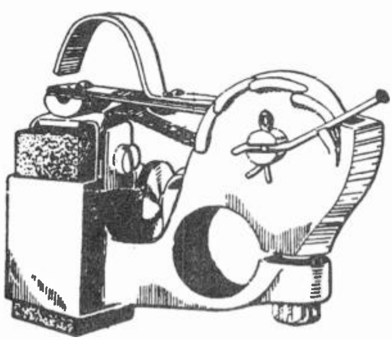


Fig. 38. A brush holder.

small carbon brushes connected in parallel, and each brush capable of separate movement. As one of the small brushes is lifted from the commutator, no open circuit results, because the other small brushes are still making contact.

With continued use, the brushes wear away, and some means must be provided to feed the brush continually toward the commutator The contact resistance between the brushes and the commutator depends somewhat on the pressure, and it is desirable to maintain this pressure constant. Therefore, the brushes are mounted in a brush holder in such a manner that a tensed spring provides the proper pressure at all times. An illustration of a brush holder appears in Fig. 38.

12. THE DIRECT-CURRENT MOTOR. Fundamentally, the DC motor is no different from the DC generator. In one case, the armature is rotated through a magnetic field, and current is taken from the brushes; whereas, in the other case, current is fed to the armature through the brushes, and the armature rotates through the magnetic field.

The motor is able to convert electrical energy into mechanical energy. In order to produce rotation of the armature when current is supplied to the motor, a twisting or rotating force must be created. This twisting force results from the interaction of the field flux and the armature flux. It will be remembered that the resultant field composed of these two separate fields produces a force on the armature conductors in such a direction as to tend to oppose the mechanical motion given to the armature, and that this is the reason that mechanical energy is required to rotate the armature in order to produce electrical energy in the DC generator. The same action takes place in the motor, except that the current in the armature is supplied from an external source, and instead of rotating the armature by mechanical force, the magnetic force which is produced is allowed to create the rotation.

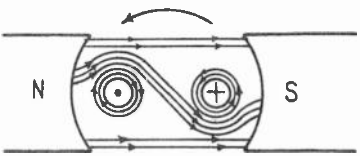


Fig. 39. Showing how the torque of a motor is produced.

This twisting or rotating magnetic force is known as the "torque" of the motor. By definition, a torque is any force which tends to cause the rotation of a body about an axis. By referring to Fig. 39, we can see how this torque is produced. This illustration shows a distorted field and the ends of two armature conductors. It is assumed that the current in the left hand conductor is moving toward the reader and that in the right conductor away from the reader. With the direction of current flow in each conductor assumed, it is evident that the field will be distorted as shown. Since a conductor always tends to move from the stronger to the weaker part of the field, it is clear that the left conductor will be forced downward and the right conductor upward. This, then, tends to cause a rotation of the armature in a counter-clock wise direction.

There is a rule for determining what direction the armature of a motor will rotate when the direction of the main field and that of the current flowing in a conductor are known. This rule is very similar to the one used to find the direction of an induced voltage in an armature conductor of a generator. To use this rule, the thumb, forefinger, and middle finger of the *right* hand are extended at right angles to each other.[1] The forefinger is extended in the direction of the field flux, and the middle finger in the direction

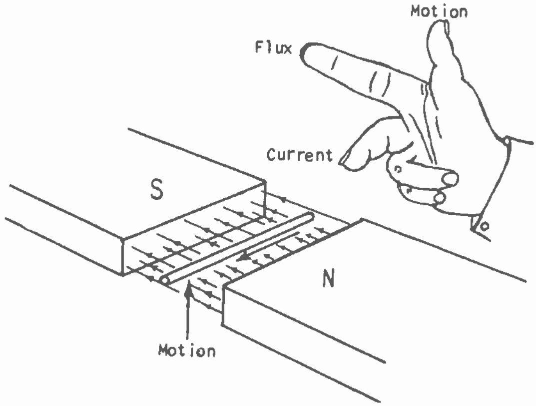
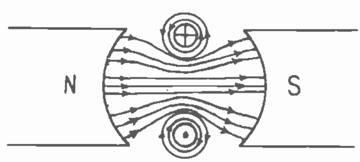


Fig. 40. The right hand rule for motors.

of the current in the armature conductor; the thumb will then point in the direction of rotation. An application of this rule is illustrated in Fig. 40.

When the ends of the two sides of the loop of wire are in the position shown in Fig. 39, the torque is a maximum; the force on the left conductor and that on the right conductor are in such directions that they add together in tending to cause rotation. However, when the loop is in the position shown in Fig. 41, the force on the top conductor is straight upward and that on the bottom conductor is straight downward. Since these two forces lie in the same straight line, they cancel each other and there is no net force to cause rotation. Thus, the torque on the armature in this position is zero. After a conductor reaches the topmost part of its rotation, the current through it must be reversed so that the direction of the magnetic force exerted on that conductor will change. For example, the force on the left conductor shown in Fig. 39 will always be downward as long as the current flows through it in the direction shown. However, when this conductor reaches its lowest

Fig. 41. Illustrating that there is no torque produced when the loop is in this position.



[1] It must be remembered that the left hand rule for generators and the right hand rule for motors are based on the newer idea of current flow from negative to positive.

position, the magnetic force must change so that the conductor will be urged upward and the same direction of rotation will be maintained. Likewise, the force on the right conductor, which in Fig. 39 is upward, must be reversed and this conductor must be urged downward after it has reached its topmost position. Therefore, the currents flowing through the two conductors should be reversed at the instant that their plane reaches a vertical position.

This reversing of the current through the armature conductors is accomplished by a commutator. Since there are many conductors spaced around the armature, the total torque is practically constant, because there will always be a pair of conductors so placed that their torque is maximum. Nearly all DC motors employ commutating poles so that sparkless commutation may be accomplished. In case commutating poles are not used, it is necessary to shift the position of the brushes to prevent sparking. Armature reaction occurs in a motor, as well as in a generator, and it produces a twisting of the main field. However, in the case of a motor, the field is twisted in the direction opposite to that of the rotation of the armature, with the result that the brushes must be shifted backward or against the direction of rotation to produce sparkless commutation. Likewise the commutating pole of a motor must be of the same polarity as the field pole under which the coil has just passed.

13. THE COUNTER E.M.F. OF A MOTOR. It has been stated that a generator creates a torque which opposes the mechanical energy applied to the generator. This action is sometimes called "the motor effect" of a generator. Likewise, a motor produces an action known as the "generator effect" of a motor. The armature conductors of a motor cut through the flux of the field poles, and voltages are induced in these conductors. By using the left hand rule, it may be proved that the direction of the induced voltages is opposite to that in which the current flows through the conductors. Thus, these induced voltages oppose the external electromotive force which is supplying current to the motor, and cause the current which flows through the armature winding to be less than it otherwise would be. The sum total of all these induced voltages is known as the "counter electromotive force" of a motor and is abbreviated to C.E.M.F. Sometimes this is called the back E.M.F. of the motor.

When there is no mechanical load on the motor, the C.E.M.F. is practically equal to the impressed E.M.F., and very little current flows through the armature. In this case, the torque is low, but the speed of the motor is normal because there is no load to drive. Placing a load on the motor tends to reduce its speed, which causes the C.E.M.F. to decrease. This, in turn, allows more current to flow through the armature, and the torque increases enough so that the actual decrease in the speed of the motor, when the load is applied is small.

In some applications, it is desirable to have a motor maintain as nearly constant speed as possible even with varying loads. The ability of a motor to accomplish this is known as its speed regulation. The difference between the speed at full load and the speed at no load is the regulation, but it is ordinarily expressed as a per cent of the no-load speed. Therefore, the formula for finding the speed regulation is:

$$\text{Speed regulation} = \frac{\text{Full-load speed - No-load speed}}{\text{No-load speed}} \times 100$$

The speed of a motor at any load is such that the sum of the C.E.M.F. and the armature resistance loss is equal to the impressed E.M.F. At first thought, it might appear that the C.E.M.F. is detrimental, however, this is not true. The C.E.M.F. prevents the motor from drawing a large current when the load is small, and automatically allows the armature current to increase with an increased load. The actual resistance of an armature winding is usually less than 1 ohm. Suppose that a motor under full load draws an armature current of 50 amperes and that the impressed voltage is 500 volts. If the armature resistance is exactly 1 ohm, the resistance loss is:

$$E = 1 \times 50 = 50 \text{ volts}$$

Therefore, the remainder of the impressed voltage must be balanced by the C.E.M.F., which in this case would be 450 volts. This C.E.M.F. will be developed only when the armature is in motion. If the full 500 volts are applied to the motor to start it, the armature current will be excessive. The current that would flow would be limited only by the resistance of the armature, and would be:

$$I = \frac{500}{1}$$

$$= 500 \text{ amperes}$$

Thus, the initial current that would flow would be ten times as large as the full load current and would probably ruin the motor.

In all but the smallest of motors, a starting resistance is used. To start the motor, all of this resistance is connected in series with the armature and the current through the armature is limited to a safe value. As the motor gains speed, this starting resistance is cut out, and when the motor has reached its operating speed, the resistance is disconnected entirely. As long as the armature is rotating at normal speed, the impressed voltage of 500 volts cannot cause an excessive current, because the C.E.M.F. bucks against the impressed voltage, and the net voltage across the armature is just enough to produce the proper current.

14. THE SHUNT MOTOR. DC motors are divided into three classes according to the way in which their field winding is connected with respect to the armature. These three types are the shunt motor, the series motor, and the compound motor. Of all three types, the shunt motor has the best speed regulation, and so it will be considered first.

A schematic diagram of a shunt motor is shown in Fig. 42. The field winding is composed of many turns of fine wire, and its resistance is such as to limit the field current to the proper value when the motor is connected to the supply line of the voltage for which it was designed. Assuming that the supply line has good voltage regulation, it is evident that the current flowing through the field winding and the strength of the field flux will be practically constant whether the motor is running at no-load or full-load.

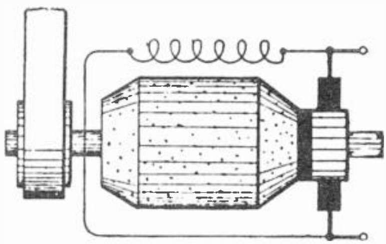


Fig. 42. Schematic diagram of a shunt motor.

The shunt motor has a very definite no - load speed which is determined by the strength of the field flux, the number of poles, the number of armature conductors, and the resistance of the armature winding. At no - load, the C.E.M.F. of a well - designed shunt motor is about 1% less than the impressed E.M.F. Thus, the net voltage across the armature is very small, and the armature current is exceptionally low. An increase in the load causes the speed to decrease momentarily, which reduces the C.E.M.F., and allows more current to flow through the armature, thereby producing more torque. When a state of equilibrium has again been established, it is found that the actual decrease in the motor's speed is very small. In fact, the full - load speed will be approximately 5% less than the no-load speed.

Often it is desirable to vary the speed at which a shunt motor will operate with a given load. The speed of a shunt motor at any load is equal to that speed at which the sum of the C.E.M.F. and the armature voltage drop is equal to the impressed E.M.F. If resistance is inserted in the armature circuit by an external rheostat, the total armature voltage drop will increase, and in order for the sum of the armature voltage drop and the C.E.M.F. to still equal the applied voltage, it is evident that the C.E.M.F. must decrease. The value of the C.E.M.F. depends on the strength of the field flux, and upon the speed of rotation. Since the field flux remains constant, the only way that the C.E.M.F. may decrease is for the speed of the motor to be reduced. Thus, increasing the resistance of the armature circuit reduces the speed of the motor, and lowering the resistance of the armature circuit increases the speed of the motor.

Let us now consider a 110 volt shunt motor whose full-load current is 40 amperes and whose armature resistance is .2 ohm. At full-load the armature voltage drop would be 40 × .2, or 8 volts, and the C.E.M.F, would be 110-8, or 102 volts. At no-load, the armature current of this motor would be not less than 2 amperes, and the armature voltage drop would be 2 × .2, or .4 volt, making the C.E.M.F. equal to 110 - ,4, or 109.6 volts. The C.E.M.F. of the motor changes only 7.6 volts as the armature current varies from 2 to 40 amperes; therefore, it is evident that a small change in the C.E.M.F. will produce a large change in the armature current.

Suppose that the motor discussed in the preceding paragraph is running at some load between the no-load and the full-load values. What will happen if a resistance is inserted in series with the field winding? Will the speed of the motor increase or decrease? Superficially, it would seem that the speed would decrease, but let us determine what actually takes place. Inserting resistance in the field circuit causes the field current to decrease with a consequent reduction in the field flux. As soon as the field flux is weakened, the C.E.M.F. decreases and the armature current increases in accordance. With a larger armature current, a greater torque is produced, but since the load on the motor has not changed, the effect of this greater torque is to increase the speed of the motor. The final speed of the motor after this weakening of the field will be such that the sum of the armature voltage drop and the C.E.M.F. equals the impressed E.M.F. It has been shown that even a large increase in the armature current does not produce a very large amount of armature voltage drop, and so it is apparent that the C.E.M.F. of the motor is decreased only slightly. Although the C.E.M.F. decreases only slightly, the reduction of the field strength may have been considerable; therefore, it is necessary that the armature rotate at a greater speed to produce nearly the same C.E.M.F. with the lowered field strength. Thus, the conclusion is that weakening the field of a shunt motor causes an increase in its speed.

This method is often used to control the speed of a shunt motor; however, an increase of about 30% in speed by field weakening is about the limit. If the attempt is made to increase the speed more than this amount by this method, vicious sparking occurs at the commutator. As the field strength is lowered, the resultant field becomes more and more distorted, and unless the motor employs commutating poles, it is not possible to secure sparkless commutation.

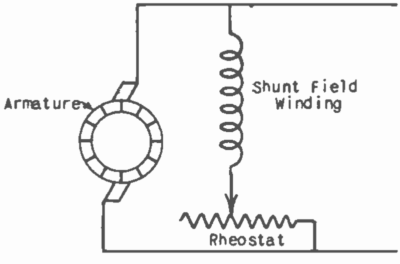


Fig. 43. Controlling the speed of a shunt motor.

When the motor employs commutating poles, its speed may be regulated either by a rheostat in series with the armature or with the field. However, it is clear that the armature current will be much larger than the field current, and a rheostat used in the armature circuit will need to dissipate considerable power, causing the motor to be somewhat inefficient. For this reason, the speed of a shunt motor is usually controlled by placing a rheostat in series with the field winding as shown in Fig. 43. It must always be remembered that *increasing* the resistance of the field circuit causes an *increase* in the motor speed, whereas *reducing* the resistance of the field circuit produces a *reduction* in speed. This is not hard to remember if it is understood that weakening the field reduces the C.E.M.F., and for the C.E.M.F. to be of practically the same value as before, requires that the armature rotate faster,

Although it is possible to cause some decrease in the speed of a shunt motor by increasing the current through the field coils, this method has its limitations. As soon as the current through the field becomes great enough to saturate the field cores, any further increase in the field current will not produce an increase in the field strength, and will not cause a decrease in the speed.

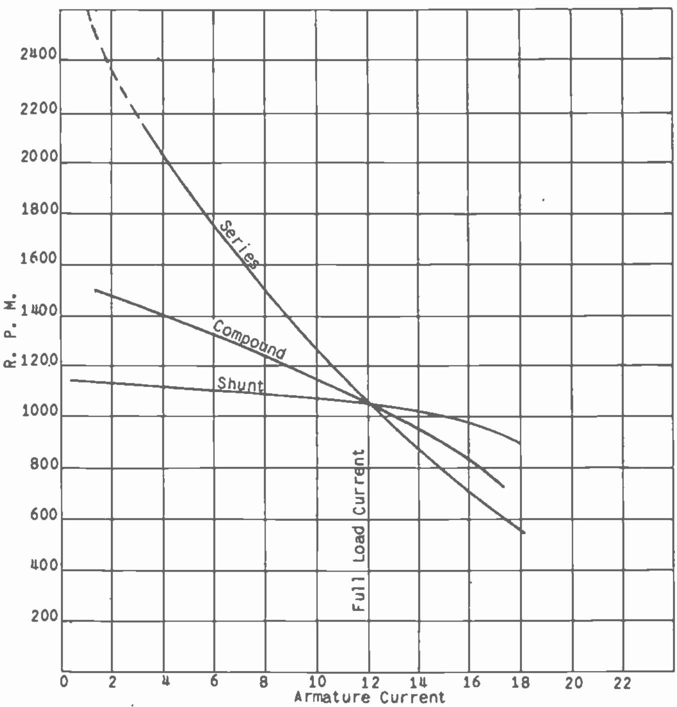


Fig. 44. External characteristics of the three types of motors.

On the other hand, care must be taken to see that the field circuit of a shunt motor is never accidentally disconnected while the motor is running. Should this occur, the field strength would be limited to the residual magnetism of the field core, and the motor would race in an attempt to produce a C.E.M.F. equal to the impressed voltage with this very weak field. The speed of the motor would probably become great enough so that the centrifugal force set up would tear the armature apart.

To compare the three types of motors, their speeds at various load currents are plotted, as shown in Fig. 44. Notice that the curve for the shunt motor is practically flat, indicating that it has good speed regulation.

15. THE SERIES MOTOR. A schematic diagram of a series motor is illustrated in Fig. 45. The field winding is in series with the armature, and is wound with heavy copper wire or copper strap so that it will be able to carry the armature current. The current drawn by the armature depends on the speed of the motor and upon the load as previously explained; therefore, it is clear that the field strength also depends on the load and upon the speed of the motor. The torque of any motor varies directly as the armature current, and also as the field strength. In the shunt motor, where the field strength is constant, the torque is directly proportional to the armature current. However, increasing the armature current of a series motor also increases the field strength in direct proportion; hence, the torque of a series motor varies directly as the square of the armature current.

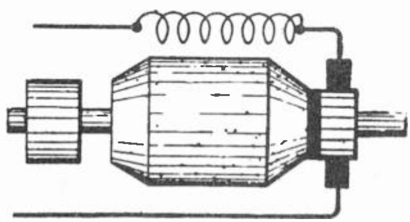


Fig. 45. Schematic diagram of a series motor.

Assume that the motor is operating at a light load. The C.E.M.F. is nearly equal to the impressed voltage and the armature current is small. This causes the torque to be small. When the load on the motor is increased, the speed of rotation decreases considerably, the C.E.M.F. is reduced, and a larger current flows through the armature and the series field. This action greatly increases the torque, and the motor is able to drive the heavy load, although at a considerably lower speed.

For a given increase in load, the series motor will reduce its speed much more than the shunt motor. In the shunt motor, the speed decreases just enough so that the C.E.M.F., produced with the *same* field strength, is slightly lowered, thus allowing more armature current to flow, and produce a greater torque. It is evident that the C.E.M.F. must be reduced so that a larger torque will be created. As the speed of a series motor starts to fall under a heavier load, it decreases the C.E.M.F., and allows more armature current to flow, but it also increases the field strength, which tends to generate a larger C.E.M.F. Thus it is necessary that the speed of the motor decrease considerably so that the C.E.M.F. will be reduced the required amount, in spite of the increase in the field flux.

For two motors of the same size, the series type has a much greater starting torque than the shunt type. Even when a starting resistance is used to start the motor, the current that is allowed to flow through the armature is somewhat larger than that which it will draw at full load. Therefore, the starting torque will be larger than the running torque. The starting torque of the shunt motor is directly proportional to the armature current, whereas the starting torque of a series motor is proportional to the square of the armature current. Thus, if both motors draw the same current when started, the torque of the series motor will be four times as great as that of the shunt type.

From Fig. 44, it is seen that a moderate increase in the load current of a series motor produces a considerable decrease in speed. This action is more pronounced at light loads than at heavy loads. When the motor is already loaded fairly heavily, an increase in the load increases the armature and the field current, but the field is near the saturation point when the load is heavy, and a further increase in the field current does not produce an appreciable change in the field strength. Therefore, when heavily loaded, the action of a series motor is more nearly that of a shunt motor, and an increase in load does not produce as large a decrease in the speed.

Series motors differ from shunt motors in that they have no definite no-load speed. When a series motor is unloaded, it continues to increase its speed, until the centrifugal force becomes great enough to tear the armature conductors from the core and literally destroy the armature. Very small series motors of one-tenth horsepower or under usually have sufficient friction in their bearings to keep the speed down to a safe limit. The racing of a series motor may be understood by considering its behavior when unloaded. As the current is sent into the armature and field coils, the motor starts up, and since there is no opposing torque, the speed tends to increase until the C.E.M.F. equals the impressed voltage. The increasing C.E.M.F., however, decreases the current in both the armature and the field, and as the field weakens, it requires a higher speed to set up the desired C.E.M.F. The field continues to become weaker and the speed increases in an attempt to set up a C.E.M.F. equal to the impressed voltage, until the large centrifugal force that is created wrecks the machine. For this reason, a series motor must never have its load removed, and it may be used only in applications where its load is permanently attached, such as in a fan, a railway car, an electric crane, etc. Furthermore, the load should be geared to the motor and not belted. There is too much danger of the belt slipping off and causing damage to the motor before the current can be disconnected.

16. THE COMPOUND MOTOR. The compound motor like the compound generator has two field windings, one in series and one in shunt with the armature. A schematic diagram of a compound motor is shown in Fig. 46. There are two types of compound motors, the cumulative-compound, and the differential-compound. Since the cumulative-compound type is by far the most important, it will be considered first. In the cumulative-compound motor, the series field aids the shunt field, and the motor has a greater starting torque than the shunt motor, but not as large as the series motor. Its speed regulation is also intermediate between the series and shunt motors. As the load increases, the motor slows down, the armature current and current through the series field increases and a greater driving torque is provided. Its actual speed regulation will depend on the degree of compounding. If the shunt field is greater than the series field (and it usually is), the speed regulation will be only a little poorer than the straight shunt motor, and the torque will be only a little greater than this type of motor. Most cumulative-

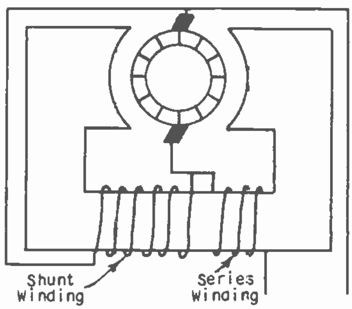


Fig. 46. Schematic diagram of a cumulative-compound motor.

compound motors are so designed that at full load, the series field provides 40% of the maximum field strength. However, sometimes motors of this type are so designed that the series field is much the stronger. In this case, the motor has practically the same characteristics as a series motor, and just enough shunt field is used to keep the motor from racing when operating at no-load. The characteristic of a cumulative-compound motor is shown in Fig. 44.

The other type of compound motor is the differential-compound, or, as it is more usually called, the differential motor. (See Fig. 47.) In this type, the polarity of the series field is such that it bucks against the field due to the shunt winding. The shunt winding, however, is always of greater strength, and its polarity determines the direction in which the machine will rotate. By proper design, it is possible to make the speed of this motor practically constant from no load to full load. As the load is increased, there is a tendency for the motor to slow down, thereby reducing the C.E.M.F. This allows more current to flow through the armature and the series field, and since the series field bucks against the main field, the main field is weakened. Weakening the main field, however, tends to cause an increase in the speed, and if the number of series-field turns be properly chosen, the effects of the increased load and the weakened field will neutralize each other and the speed will remain unchanged.

Although the speed regulation of a differential motor may be made very good, its starting torque is very low, even lower than that of a shunt motor. Since the speed regulation of a shunt motor is close enough for nearly all applications that require a constant speed, the differential motor finds but little use.

The differential motor has two distinct disadvantages. The first of these is that the motor tends to race if the machine is overloaded. When the load is very great, the current through the series field is large and the field strength of this winding is nearly sufficient to cancel that due to the shunt winding. This causes the resultant field to be very weak, and the motor tends to race. The second disadvantage is that the motor is liable to start up in the wrong direction when the current is applied. The shunt winding has a high inductance and the current through it builds up

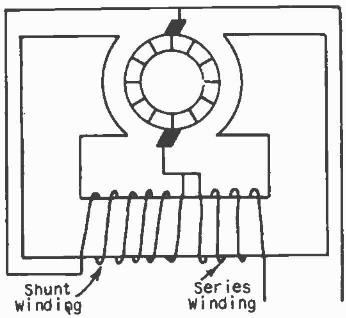


Fig. 47. Schematic diagram of a differential-compound motor.

slowly, whereas the inductance of the series winding is low and the current through it and its field strength build up practically instantaneously. At the instant that the voltage is applied, the field flux of the series winding may be larger than that of the shunt winding, with the result that the resultant field is of opposite polarity, since the two fluxes oppose each other. If this occurs, the motor starts to rotate in the wrong direction. To avoid this, it is common practice to short circuit the series field when starting a differential motor.

17. THE STARTING BOX. The necessity of introducing resistance into the armature circuit when the motor is started has previously been mentioned. It has been shown that there is no C.E.M.F. present until the motor begins to rotate, and that the armature current would be large enough to damage the motor if the full line-voltage were applied to start it. The starting resistance is usually called a "starting box", and is constructed in many different ways. Perhaps the most useful type is the one shown in Fig. 48. There should be sufficient resistance in the rheostat to limit the starting current to 150% of its full-load value.

We shall assume that this starting box is to be used to start a shunt motor. When the lever H is moved to contact 1, current flows from L_1 to point D, where it divides. That part that flows through the armature flows through the resistance from point 8 to point 1, and through the lever to L_2. The other part of the current flows through the field, passes through the electromagnet M to

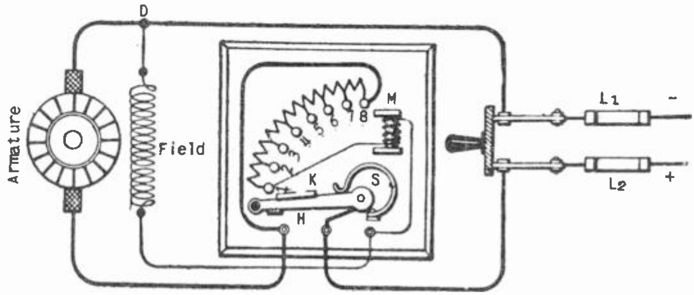


Fig. 48. Diagram of a starting box.

point 1, and back to the positive side of the line. The current through the armature is fairly heavy, and the motor begins to rotate. As it gains speed, it builds up a C.E.M.F. and the armature current is reduced. The field current at this time is maximum. As the lever is moved to the right, the armature current increases and the field current decreases; both actions tend to increase the speed of the motor. From 15 to 30 seconds should be consumed in starting the motor, depending on its size.

If the lever is moved to the right too rapidly, the impressed voltage increases much faster than the C.E.M.F., and the motor is liable to be damaged by the excessive armature current. On the other hand, too much time should not be used to start the motor, because the resistance of the rheostat is not designed to carry the armature current for an appreciable length of time, and the resistance may be burned out. It should be understood that this starting box is not to be used as a method of controlling the speed of the motor.

The lever is provided with a spring S, which tends to keep the lever at the left end of its movement. This lever, however, is fitted with a soft iron keeper K, and when the lever is on contact 8, the attractive force of the magnet M holds it in this position as long as this magnet is energized. The purpose of this magnet is two-fold. Suppose that for some reason the line voltage fails. If it were not for this magnet and the spring connected to the lever, the lever would remain on contact 8. Then when voltage was again established on the line, the total line voltage would be impressed across the motor, which having stopped, would now draw excessive current. With the magnet in the circuit, however, an interruption of the line voltage or the opening of the motor switch will de-energize the magnet, and allow the lever to spring back to its starting position. This action precludes any possibility of damage to the motor due to line voltage interruptions. Sometimes this magnet is connected directly across the line, and its current does not flow through the field; however, as we shall now see, it is desirable that the energizing current flow through the field winding. This magnet is called a "no-voltage" release.

It has been stated that the field of a shunt motor must never be disconnected while the machine is in operation, because such an action causes the motor to race with consequent damage to its armature. The starting box which we have been discussing is so designed that the motor is disconnected from the line if an open circuit occurs in the field winding. If the field circuit develops an open circuit, the magnet M is de-energized and the lever flies back to the left disconnecting the motor from the line. Thus, the magnet serves as a "no-field" release also.

The motor should never be stopped by attempting to pull the lever back to the left, as this will cause unnecessary sparking of the contact points on the starting box. Instead, the motor switch should be opened and the lever of the starting box will automatically return to its starting position as the holding magnet is de-energized.

Some starting boxes are equipped with overload releases which will disconnect the motor from the line if the load current becomes too great. These, however, are not essential, since the line fuses and circuit breaker will protect the motor, if it is overloaded.

18. APPLICATIONS OF MOTORS. The choice of a motor for a certain class of work depends on whether the load will be variable, and upon whether much change in the speed may be tolerated. The shunt motor is used in applications where a practically constant speed is required, regardless of load variation. Thus, it is suitable for driving machine tools, blowers, fans, line shafting, etc.

Compound motors are used where sudden applications of heavy loads will be experienced such as driving punches, shears, etc. Also, compound motors are suitable in cases where a large starting torque is required, but where less speed variation under load can be tolerated than a series motor would give.

Series motors are particularly suitable for railway and hoisting service, where a large starting torque is necessary and speed variation under load is not important. When used in such applications, they are very efficient, for as the load increases, they promptly slow down and produce a tremendous torque for driving the load. Of course, they must never be used where there is any danger of the motor becoming unloaded.

19. THE DYNAMOTOR. There is one other type of machine which should be discussed before we take up the subject of alternating current motors and generators. It is the dynamotor. A dynamotor is a machine whose armature has two separate windings. Current is fed to one winding and causes the armature to rotate. The other winding of the armature cuts through the field flux and a voltage is induced in it, which may be applied to an external circuit by a commutator or a pair of slip rings. Thus, the dynamotor is a

combination of a motor and a generator. There is just one field winding, and one armature core, and each end of the armature is fitted with either a commutator or a pair of slip rings. If the armature has two commutators, it will change direct current from one voltage to another. Some machines are designed to be driven by a 6 volt DC source, and will deliver several hundred volts of direct voltage. Perhaps the most common type is that one which is driven by a 6 volt DC source and furnishes 110 volts DC for operating equipment designed for this voltage. The ratio of the voltage output to the voltage input is, of course, fixed at the time of the machine's construction and is not variable. If the machine is to be driven by an alternating voltage and is to furnish

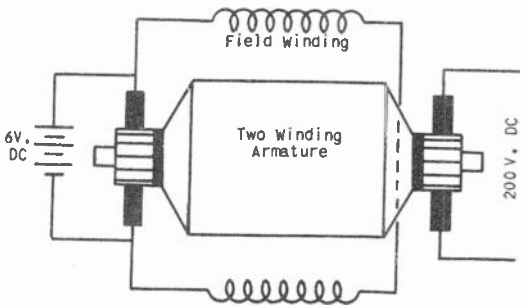


Fig. 49. Schematic diagram of a dynamotor.

a direct voltage, it will have a pair of slip rings on one end of the shaft and a commutator on the other. Naturally, there is nothing to be gained by using a machine to change an alternating current from one voltage to another, since this operation can be accomplished more efficiently by a transformer. A schematic diagram of a dynamotor is shown in Fig. 49.

We should now have a fairly comprehensive idea of the construction and operation of direct-current motors and generators. The following lesson (Lesson 8A) will be devoted to the subject of alternating current machinery.

Notes

(These extra pages are provided for your use in taking special notes)

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT NO. 3

A.C. MOTORS AND GENERATORS

LESSON NO. 8A

QUICK SUCCESS

.....at one big jump.

A long time ago when we were in our carly teens, we remember reading about a big strong man who devoted the majority of his time telling stories about his heroic deeds and accomplishments. He was always happiest when he was the center of attraction, and for a time managed to uphold the fables that had been built around him.

One day the town in which this man lived was electrified by the news that gold had been discovered in the nearby mountains. Almost the entire male population made hurried preparations to rush to the new gold fields and stake their claims. Departing from the town in a body, they rode through the rough and torturous mountains until they finally were halted by a deep gorge. Scouts were sent out to discover a way through the gorge.

Several hours later the scouts returned and one had good news. He had discovered a natural trail that could be traveled with extreme caution. But valuable time would be lost. The object of our story, the big strong man, had also made a discovery. He had found a place where the walls of the gorge came close together...close enough for him to leap across because he was so strong. He would let his foolish neighbors travel the safer way. And he would reach the gold fields first...by making one big jump.

Abandoning his horse and running swiftly across the ground, he leaped into the air. Down and down he went, ending his leap in a fatal crash. On the following day his more patient and sensible townsfolk reached the gold field and staked their claims. They knew that success was rarely attained in "one big jump".

Your success in radio also calls for patience and common sense. You must master your studies thoroughly. When you do this you will be capable of earning more money. Lesson by lesson you are traveling toward your "gold field". And every lesson that you complete today, can prove to be shining nuggets of gold in the future.

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Lesson Eight-A

AC MOTORS & GENERATORS

"The complete study of AC operated motors and generators would require many volumes and a great deal of unnecessary time. Since the average radio technician needs to know only the fundamentals of operation and maintenance, it will be the purpose of this lesson to supply that need.

"We will discuss those types of motors and generators most commonly encountered in radio stations."

1. THE ALTERNATOR. The common name for the alternating current generator is the alternator. As described in the preceding lesson, an alternating voltage is induced in a loop of wire when it is rotated in a steady magnetic field. In fact, the armature conductors of all DC generators have alternating voltages induced therein, but this voltage is changed to a direct voltage by the action of the commutator. Therefore, it would appear that the alternator is simply a DC generator fitted with slip rings instead of a commutator. While it is true that such a machine would function as an alternator, the actual construction of commercial alternators differs in other details as well.

To produce a voltage requires that there be relative movement between the armature and the field. In the DC generator, this is accomplished by a stationary field and a rotating armature. The alternator, on the other hand, ordinarily has a stationary armature and a rotating field. Such a machine is illustrated in Fig. 1. The armature is wound on the inside of a circular steel frame, and the field poles (in this case 26) are mounted on a steel ring which is attached to the shaft, and rotate within the armature frame, thereby causing the field flux to cut the armature conductors. The appearance of a 100-pole rotating field is shown in Fig. 2. The field must be excited by a DC voltage which is fed to the field by two slip rings which are distinguishable in this figure.

Alternators are never self-excited; the DC field voltage must

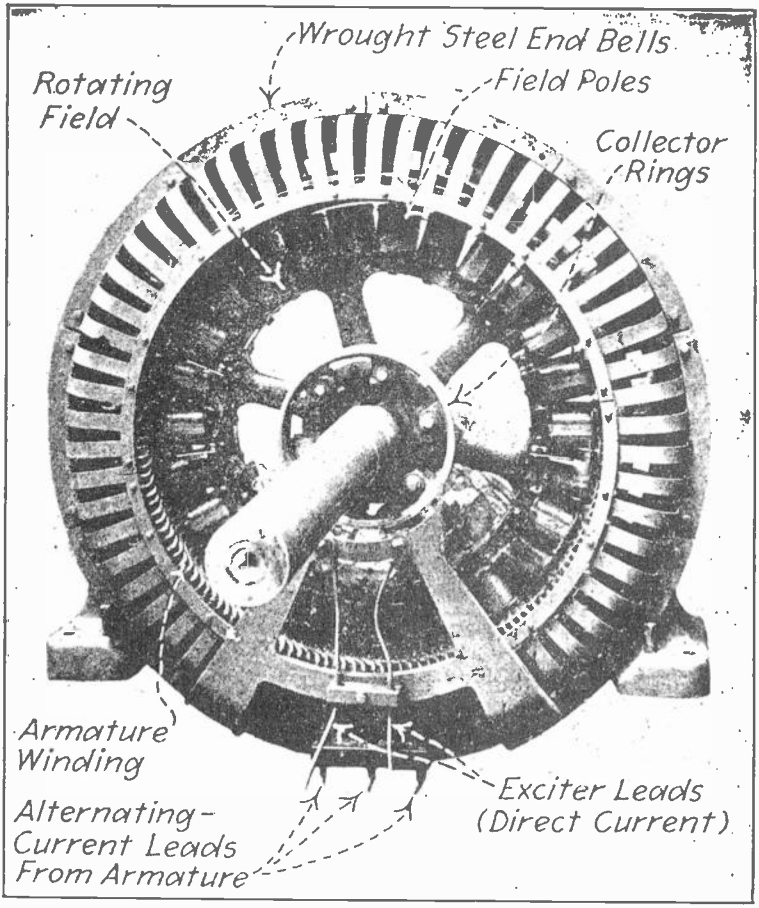


Fig.1 An alternator having a stationary armature and a rotating field.

be furnished by an external source. In large installations, each alternator is equipped with a small DC generator mounted on the same shaft, whose only purpose is to furnish the DC field voltage for the alternator. These small DC generators are called "exciters". Occasionally, each alternator does not have its own exciter, but instead, one large DC generator is used to excite several alternators. An alternator having a capacity of 1000 kw. requires an exciter of approximately 25 kw. capacity.

The number of field poles that an alternator will have depends upon the speed at which the moving field is rotated and the fre-

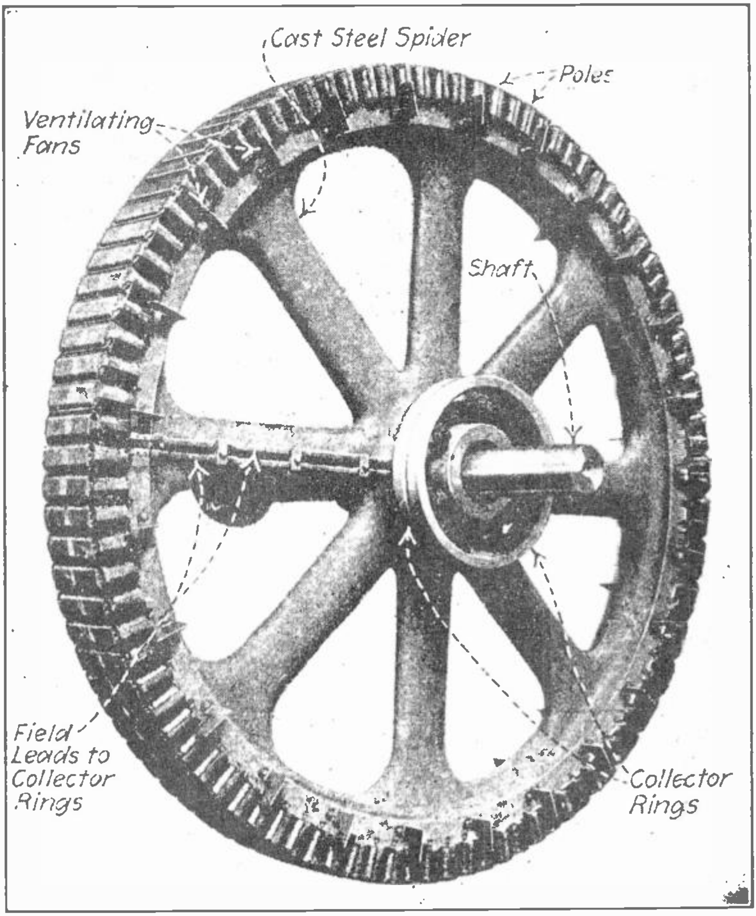


Fig.2 A 100-pole rotating field of an alternator.

quency of the desired voltage. In the foregoing lesson, it was shown that one cycle of alternating voltage results when a loop of wire makes one revolution between one pair of poles. If the machine had four poles (two pair), two cycles would result from one complete revolution. In fact, whenever a conductor passes from one pole to the next succeeding pole, one alternation is generated. Thus, there will be as many cycles of alternating voltage generated for each revolution as there are pairs of field poles. The frequency, however, is stated in cycles per second; and is found by knowing how many cycles are generated per revolution and at how many revolutions per second the machine is rotating. There is a very simple formula for determining the frequency of the generated voltage, when the number of poles and the speed of the machine are known. This formula is:

$$\text{Frequency in cps} = \frac{N \times S}{60}$$

Where: N is the number of pairs of poles, and
S is the speed in revolutions per minute.

For example, suppose that a machine having 8 poles is rotated at a speed of 900 r.p.m. What is the frequency? This problem is easily solved by substitution in the above formula. Since there are 8 poles, there must be 4 pairs of poles, and the solution is:

$$\text{Frequency} = \frac{4 \times 900}{60}$$

$$= \frac{3600}{60}$$

$$= 60 \text{ Cycles per second.}$$

If a machine is to be driven by a high-speed turbine, it will have only a few poles, perhaps 2 or 4; whereas, an alternator driven by a comparatively slow-speed reciprocating steam engine will have a large number of poles.

Perhaps you are wondering why the commercial alternator uses a rotating field and a stationary armature. Alternators are usually built in large sizes capable of supplying far more power, at higher voltages, than is possible with a DC generator. The armature winding may generate a voltage as high as 13,000 volts, and such high voltages require very careful insulation. Naturally, it is easier to insulate a stationary winding than it is one which is revolving. A stationary armature requires no slip rings and the leads from the armature can be continuously insulated from the machine to the switchboard. With a rotating armature, slip rings are necessary, and these rings are difficult to insulate at high voltages. Furthermore, the DC voltage applied to the rotating field is seldom more than 250 volts, and insulation at this voltage is no particular problem. It is for these reasons that practically all alternators have stationary armatures and rotating fields.

2. THE POLYPHASE CIRCUIT. The type of alternator which we have been discussing is called a "single-phase" machine. This means that it generates just one alternating voltage. Perhaps most of the alternating current circuits with which you are familiar are single-phase circuits. Small AC motors and all lighting require single-phase power. However, as we shall now see, single-phase alternators and single-phase AC motors are the exception rather than the rule.

The study of polyphase circuits is quite complex, and we shall not attempt to explain more than a few fundamental details, since a complete understanding of such circuits requires a knowledge of higher mathematics. It is, however, necessary that you learn something about them, because many power supplies for transmitters use polyphase power.

In general, polyphase circuits may be divided into two classes; the two-phase, and the three-phase systems. Since three phase alternators and motors are far more common than the two-phase variety, this discussion will be limited to the three-phase type. The term "three-phase" as applied to alternating current circuits characterizes the combination of three circuits energized by three alternating E. M. F.'s differing in phase by 120°, or one third of a cycle. First, let us assume that we have three separate alternators all joined to the same driving shaft, and each generating the same alternating voltage. We shall assume that the field of each machine consists of two poles which are stationary, and that each has a revolving armature. The three alternating voltages generated by the three separate machines may be all in phase, or each voltage may be out of phase with the other two by a certain number of degrees, depending on the position of the armature coils of one machine with respect to those of the others. If these three alternators are to represent a three-phase system, each voltage will be 120° out of phase with the other two.

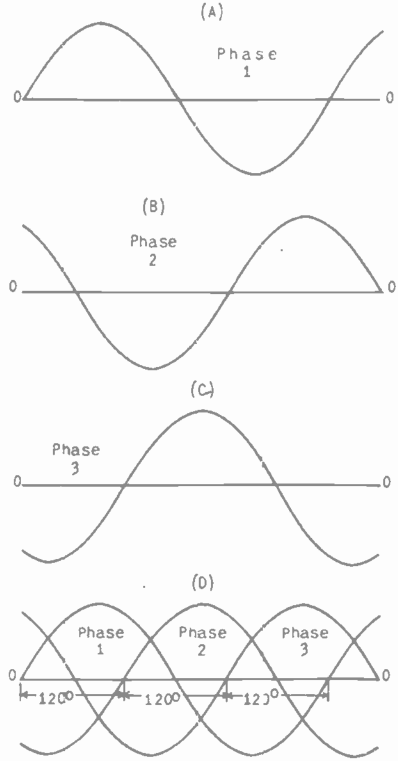


Fig.3 Illustrating the three voltages of a three-phase system. At A, B, and C the voltages are shown separately, while D illustrates the three wave forms drawn on the same axis.

The first voltage is represented by A in Fig. 3; the second by B; and the third, by C. The combination of all the voltages shown on a single axis is given at D in this figure. Each alternator is said to be generating one phase of the three-phase system.

Such a system, using three alternators, is naturally not economical. It is much more convenient to use one alternator which has three separate windings on its armature spaced equally distant apart. Let us suppose that this armature is composed of three coils as shown in Fig. 4. S_1 represents the beginning of coil 1, and F_1 the finish of this coil. At the instant shown, the voltage induced in coil 1 is maximum. If we consider that the positive direction of measuring voltages is from the start of one coil to the finish of the same coil, it is evident that the voltage induced in coil 2 is 120° out of phase with that in coil 1, because the start of this coil is spaced 120° from that of the first coil. Likewise, the voltage induced in coil 3 is 120° out of phase with that induced

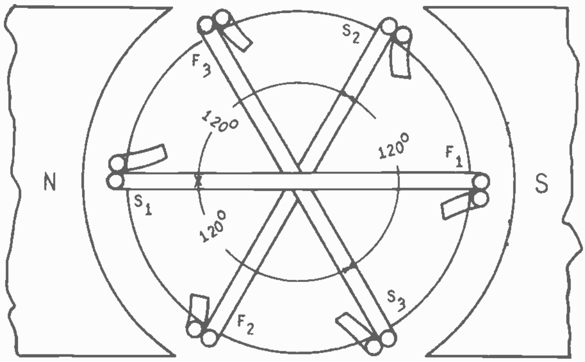


Fig.4 A diagrammatic view of a three-phase alternator, showing the three windings spaced 120° apart.

in coil 2. At the instant shown, the voltages induced in coils 2 and 3 are in the same direction, because both of their starting points are passing a south pole. Also, the voltage induced in coil 1 is opposite to that induced in the other two coils, because the beginning of this coil is passing a north pole. The wave form of these three alternating voltages is shown in Fig. 5. It is assumed that coil 1 generates phase 1; coil 2, phase 2; etc. The vertical line drawn through these three sine waves indicates the instant represented by the position of the alternator.

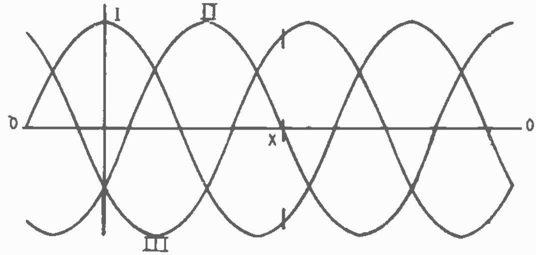


Fig.5 The wave form of the three voltages generated by the alternator shown in Fig. 4. The vertical line shows the instantaneous voltages generated by each of the three windings.

There are several ways in which these three alternating voltages may be used. If the three-phase alternator is to drive a three-phase motor, it would be possible to connect each end of these three coils to a slip ring; use six brushes for the six slip rings; and connect the alternator to the motor with six wires.

This would be represented diagrammatically by Fig. 6. The three coils on the left, spaced 120° apart, represent the three windings of the alternator, and the three coils at the right represent the three windings on the armature of the motor.

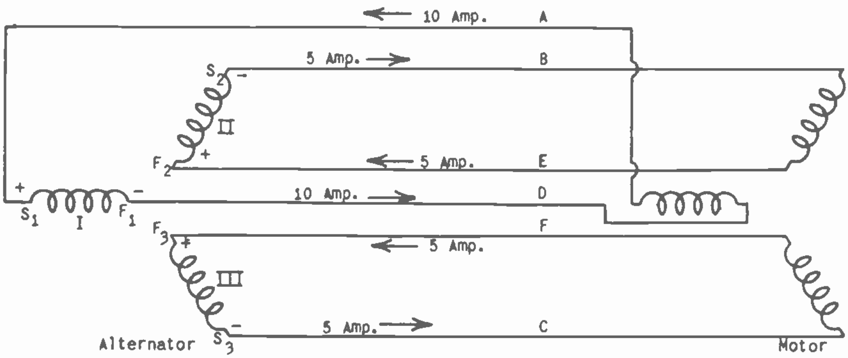


Fig.6 Showing one method of connecting the three windings of a three-phase alternator to its load. In this case, six line wires are needed.

Such an arrangement, however, is never used in actual practice. Instead, only three wires are used to connect the alternator with the motor. Let us see how this is possible. At the instant shown by the vertical line in Fig. 5, the voltage across phase I is maximum in a positive direction. At this same instant, the voltages of phases II and III are negative, and are equal to each other. Furthermore, it may be proved that the sum of the instantaneous voltages of phases II and III is equal and opposite to the voltage of phase I. This is also true at any instant of the cycle. The algebraic sum of the three instantaneous voltages is always zero. Perhaps this can be made more clear by choosing an instant when the voltage of one phase is zero, such as point X in Fig. 5. At this time, the voltage of phase II is zero, and the voltages of phases I and III are equal and opposite; therefore, their algebraic sum is zero.

Let us now return to Fig. 6. We shall assume that the maximum current drawn from each phase is 10 amperes. Again considering the instant shown by the vertical line in Fig. 5, the current in phase I is maximum and positive; therefore, let us assume that the current in line A is 10 amperes and in the direction shown by the arrow. It is, of course, evident that the current in line D, which connects to the opposite side of this phase is also 10 amperes and in the direction as shown. The current in phase II is opposite in direction to that of phase I and is equal to 5 amperes. Therefore, the currents in lines B and E are 5 amperes and in the directions shown. Also, the current in phase III is equal to that of phase II and in the same direction; and so lines C and F each carry 5 amperes.

In order to eliminate some of these six line wires, we shall connect the points F_1, F_2, and F_3 together, and also connect the coils of the motor winding together as shown in Fig. 7. There are now just four line wires extending from the alternator to the motor; the line N replaces the three wires D, E, and F. Since one

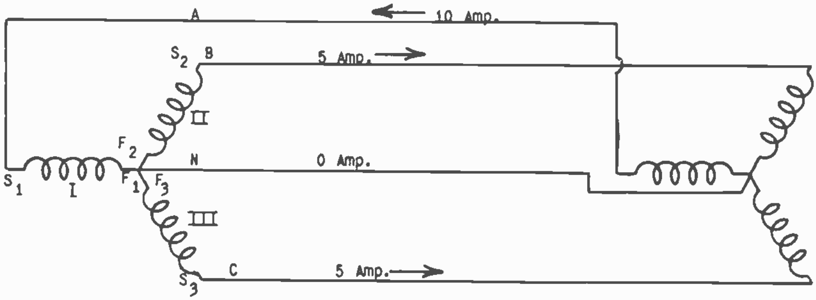


Fig.7 The Y-connection of the three windings of a three-phase alternator. Only four line wires are used.

wire is used where three were connected before, it is clear that line wire N must carry the combined currents that previously flowed in lines D, E, and F. Knowing this fact, let us determine what current will flow in line N. Line D was carrying 10 amperes away from the alternator; line E was carrying 5 amperes toward the alternator; and line F was carrying 5 amperes toward the alternator.

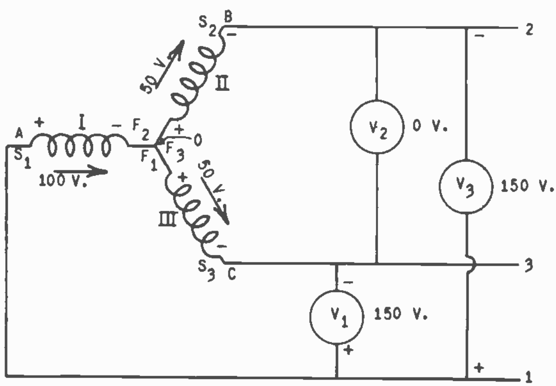


Fig.8 A schematic diagram of a Y-connected alternator.

Thus, of the currents in these three wires, 10 amperes was flowing toward the alternator, and an equal amount was flowing away from the alternator. Therefore, the sum of the three currents of these three lines is zero, and the current flowing in line N is likewise zero. Line N is called the "neutral" wire, and since there is no current flowing in it, it is usually omitted, and only three wires connect the alternator with its load. As long as the maximum current drawn from each phase is equal, the three-phase system is said to be balanced, and the neutral wire is not used.

When the three windings of a three-phase alternator are connected together in this manner, the alternator is said to be Y-connected or star-connected. Let us now consider Fig. 8, which shows the schematic diagram of a Y-connected alternator. Coils I, II, and III represent the three windings of the alternator, and we shall assume that the maximum voltage generated in each winding is 100 volts. At the instant shown, the voltage of phase I is maximum and is in the direction shown by the arrow. At this time, the voltages in phases II and III are each 50 volts, and in the opposite direction. The voltage between lines 1 and 2 at this instant is seen to be 150 volts, or the sum of the voltages in coils I and II. Likewise, the voltage between lines 1 and 3 is also 150 volts. On the other hand, the voltage between lines 2 and 3 is zero, because in going from line 2 to line 3, we travel from point B to point O against the direction of the arrow; but from point O to point C, we travel in the direction of the arrow. Therefore, the voltage in coil II and that in coil III oppose each other, so that there is no difference in potential between lines 2 and 3. Or, we may consider that the voltage in coil II from B to O is equal to the voltage in coil III from C to O, and thus point B is at the same potential as point C.

It is evident in the preceding case that the voltage between the line wires may be as high as 150 volts, even though the maximum voltage generated in each armature winding is only 100 volts. However, can the voltage between two line wires ever be greater than 150 volts, under this condition? Let us investigate this problem. The maximum voltage between a pair of the line wires will occur when the sum of the instantaneous voltages across two of the windings is maximum. It may be proved that this will be the case when the voltage in one winding is zero.

Let us now consider Fig. 9. At A is shown the same Y-connected alternator. At B is seen the three sine waves representing the three voltages differing in phase by 120°. We shall select the point designated by the vertical line. At this instant, the voltage generated in phase III is zero. Also at this time, the voltage generated in phase I is positive and not far from its maximum value, whereas the voltage in phase II is negative and equal in value to that of phase I. By using trigonometry, it may be proved that the voltage in phase 1 and in phase 2 is 86.6 volts at this time. Therefore, let us transfer these values to the diagram at A. An arrow is drawn from A to O beside coil I, assuming that this is the positive direction of measuring voltages. This voltage is labeled 86.6 volts. The voltage in coil 2 is also 86.6 volts, but this voltage is in the negative direction so that the arrow is drawn from O to B. There is no voltage in coil 3.

The voltage between the lines 1 and 2 is, therefore, seen to be 86.6 + 86.6 or 173.2 volts, since these voltages add together in traveling from line 1 to line 2. Therefore V1 will read 173.2 volts. The voltage between lines 1 and 3 will be due only to the voltage of coil I and V_3 will read 86.6 volts. Likewise, the voltage between lines 2 and 3 is due only to the voltage of coil II, and V_2 will read 86.6 volts.

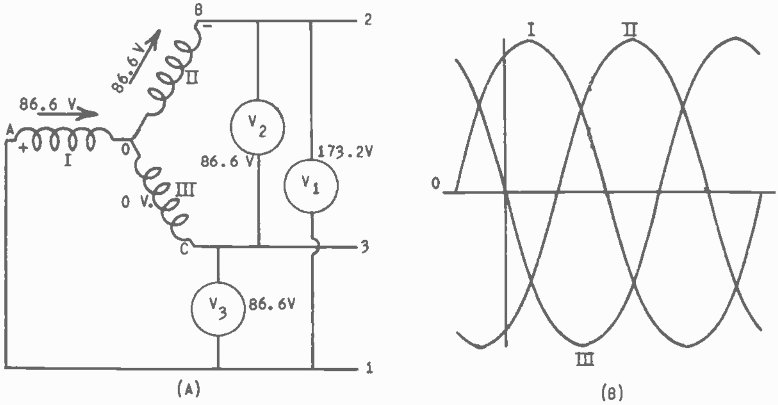


Fig. 9 Illustrating that the voltage across two of the line wires is greater than the voltage generated by each winding.

Thus, it is seen that the maximum voltage between any two of the line wires is 173.2 volts, when the maximum voltage generated in each phase of the armature winding is 100 volts. The ratio of 173.2 volts to 100 volts is 1.732 to 1, and since 1.732 is equal to the square root of three ($\sqrt{3}$), it is clear that the maximum line voltage is equal to $\sqrt{3}$ times the maximum phase voltage. We have been dealing with maximum values, but the same relation holds true for R.M.S. values as well. The R.M.S. voltage measured between

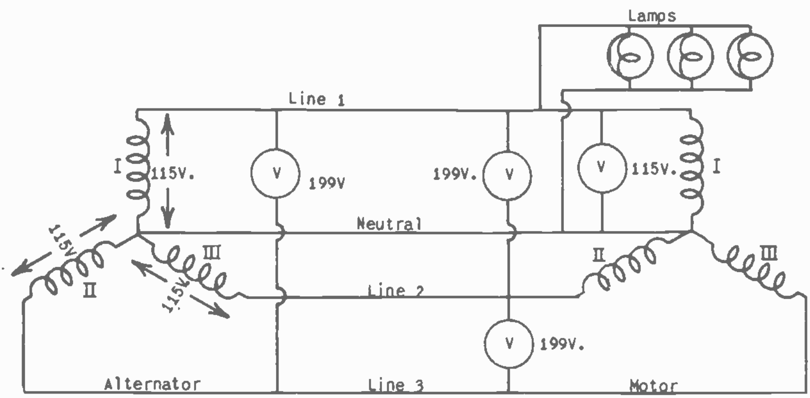


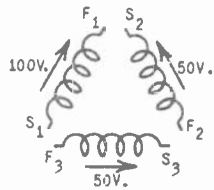
Fig. 10 Showing a three-phase alternator supplying power to a three-phase motor, as well as to a bank of lamps.

any two of the line wires is equal to the $\sqrt{3}$ times the R.M.S. voltage that would be measured across any one of the armature windings.

It is realized that the foregoing explanation is far from thorough, but, as previously stated, it is impossible to explain the action of three-phase circuits in a complete manner, without resorting to the use of trigonometry and vector analysis.

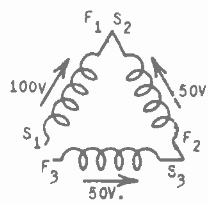
Lines 1 and 2 constitute one phase; lines 1 and 3, the second phase; and lines 2 and 3, the third phase. Unless, the three-phase alternator is connected to a three-phase motor, or to some device that draws the same current from each phase, the system will not be balanced, and excessive currents will flow through the armature

Fig.11 The three windings of the three-phase alternator, and the instantaneous voltages generated in each.



windings. However, if the neutral wire is connected to the armature, and is carried along with the three line wires, it is possible to operate a three-phase system in an unbalanced condition. For example, consider Fig. 10. The voltage across each coil of the alternator is 115 volts R.M.S.; therefore, the voltage across any two of the three line wires (that is across any phase) is: 1.732 × 115, or 199 volts. All three of the phases are applied to the motor in the diagram, and, in addition, phase 1 is supplying a bank of lamps, which require 115 volts. The voltage between any one of the line wires and the neutral wire is 115 volts, and the lamps are connected between line 1 and the neutral. In this case, the current drawn from phase 1 is greater than that drawn from each of the other two phases, and the neutral line is obliged to carry some current.

Fig.12 Illustrating that the net voltage around the delta-connected alternator is zero.



There is another method of connecting the three phases of an alternator to its load using only three line wires. Fig. 11 shows the same three armature windings, which we have been considering. The voltages induced in the three coils is the same as at the instant considered in Fig. 8. However, instead of connecting the finishing points of all these three windings together, we shall connect F_1 (the finish point of coil 1) to S_2 (the start of coil 2);

and F_2, to S_3. These connections are shown in Fig. 12. Now let us determine what the resultant voltage measured between S_1 and F_3 will be. In traveling from S_1 to F_3, we go through coil 1 with a voltage of 100 volts, and, in passing through coil 2, we travel against a voltage of 50 volts. Likewise, through coil 3, we are traveling against a voltage of 50 volts. Thus, the net voltage between S_1 and F_3 is zero, and S_1 and F_3 are at the same potential. Therefore, let us connect S_1 and F_3 together to form the schematic shown in Fig. 13. Perhaps your first thought is that current will circulate around this closed circuit, but such is not the case. At any instant, such as the one described, the voltage tending to force a current clockwise around this loop is equal to the voltage tending to force a current in a counter-clockwise direction around the loop, and, as a result, there is no circulating current.

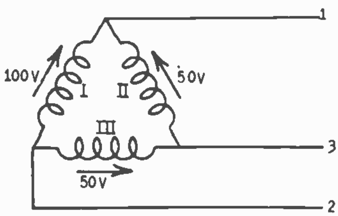


Fig.13 A diagram of a delta-connected, three phase alternator.

This type of connection is known as the Δ or delta connection. After the three armature windings are connected as shown, the line wires are connected to each corner of the triangle. It is evident that the voltage between lines 1 and 2 is the same as the voltage across the coil I; also, the voltage across lines 1 and 3 is equal to the voltage across coil II; and the voltage across lines 2 and 3 is equal to the voltage across coil III. Therefore, in the delta-connected alternator, the line voltages are equal to the coil or phase voltages. If the R.M.S. voltage of each armature winding is 100 volts, the R.M.S. voltages between any two of the line wires will also be 100 volts.

On the other hand, the currents flowing in the line wires are not of the same value as the currents flowing in the armature coils. It is apparent that line 1 is fed from both coils I and II at certain instants. Likewise, line 2 is fed from coils I and III, and line 3 from coils II and III. By a method similar to the one used to prove that the line voltage of a Y-connected machine is $\sqrt{3}$ times the voltage in each armature phase, it is possible to prove that the R.M.S. current in each line wire of a delta-connected, balanced system is $\sqrt{3}$ times the R.M.S. current in each armature winding. Thus the Y-connection increases the line voltages, whereas the delta connection tends to increase the line currents. If the same amount of line current flows in either case, it is clear that each armature winding will have to supply a smaller current, if the machine is delta-connected.

At present there are two systems in use by power companies for supplying three-phase power. In one system, the three windings on the secondary of the pole transformer are delta-connected, and the voltage (R.M.S.) across each phase is 220 volts. Such a system is convenient for many electrical machines requiring three-phase power, since nearly all of them employ 220 volts. With this system, however, it is not possible to secure 110 volts for lighting or for small motors which need this voltage. For this reason, many power companies are now using three-phase transformers in which the secondaries are Y-connected, and the neutral wire enters the premises of the consumer along with the three line wires. The R.M.S. voltage across each secondary, with this system, is 115 volts, and this voltage may be secured between any one of the line wires and the neutral. Between any two of the line wires, the R.M.S. voltage is 199 volts, and this voltage may be used to supply equipment requiring three-phase power. Such a system is desirable for transmitter installations, because both voltages will be required.

In order that the student shall know what should be learned from the foregoing discussion of three-phase circuits, a summary will now be given.

1. Nearly all alternators are designed to generate three alternating voltages differing in phase by 120°. To do this, requires that the armature have three, equidistantly spaced windings.
2. There are two standard ways to connect these three windings together so that the alternator may supply power to its load. These are the Y-connection and the delta-connection.
3. In the Y-connection, the neutral wire is not used unless the three phases are unbalanced. The voltage between any two of the three line wires is equal to $\sqrt{3}$ times the phase voltage. The current flowing in any of the lines is equal to the current in the armature windings.
4. In the delta-connection, the line voltages are equal to the phase voltages, but the line currents are equal to $\sqrt{3}$ times the phase currents. The delta-connection is not ordinarily used in unbalanced systems.
5. Nearly all alternators are Y-connected; motors may be either Y or delta-connected. Thus, a Y-connected alternator might furnish current for a delta-connected motor.
6. There are three important reasons for using polyphase systems. These are: (A) Polyphase apparatus is smaller and weighs less than single-phase apparatus of the same capacity, and is usually less complicated. (B) Polyphase machinery has better operating characteristics than single-phase machinery. (C) To transmit a given amount of power, at a certain voltage and efficiency, requires only three-quarters as much copper for the line as would a single-phase system.

3. TYPES OF ALTERNATING CURRENT MOTORS. There are several types of alternating current motors. Listed according to their popularity and usefulness, they are the induction motor, the series motor, the repulsion motor, and the synchronous motor. Also, each of these motors may be either single-phase or three-phase. The following discussion of the various types will be limited to a brief description of each. Enough information will be given to enable the student to recognize the various types and to know for what applications each is best suited. It is neither possible nor desirable to present a complete treatise of each type, for not only are some of the principles very complex, but it is not necessary that the student of radio engineering have a thorough knowledge of that subject.

The student should, however, know some of the operating characteristics of each type of motor, and he must be able to follow a definite routine for maintaining them in their proper operating conditions.

4. THE INDUCTION MOTOR. *(A) The three-phase type.* The induction motor depends for its operation upon the production of a rotating field. This does not mean that the field structure or

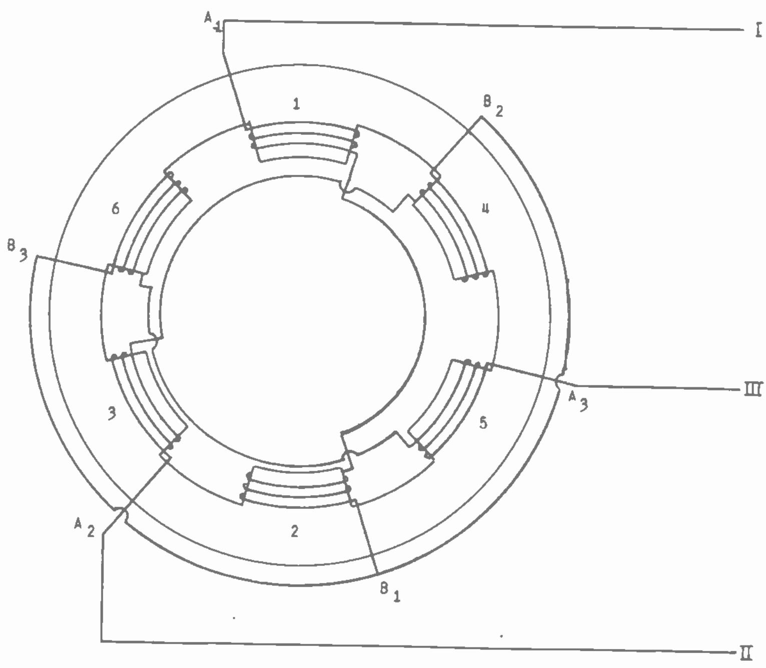


Fig.14 The field structure of a three-phase induction motor. I, II, and III are the three line wires supplying the motor. 1 to 6 inclusive are the field poles. There are two poles per phase.

field winding actually rotates through space. Instead, the field winding is stationary, but the direction of the magnetic flux rotates from instant to instant as the current through the field winding alternates. We shall first consider a three-phase motor, since it is by far the easier to understand. In Fig. 14 there is illustrated the stationary field structure of an induction motor having six poles equidistantly spaced around the periphery of the frame.

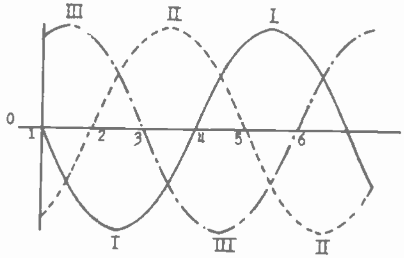


Fig.15 The three voltages applied to the induction motor shown in Fig. 14.

Each pair of these poles is fed from one phase of a three-phase alternator. Thus, poles 1 and 2 are fed by phase I, the terminals of which are A_1 and B_1. Likewise, poles 3 and 4 are fed by phase II with terminals A_2 and B_2; and poles 5 and 6 by phase III with terminals A_3 and B_3. Notice that the terminals B_1, B_2, and B_3 are connected together; therefore, the motor is Y-connected.

We shall assume that when the phase of any voltage is positive; that is, when its wave form is above the zero reference line in Fig. 15, the current of that phase enters its A terminal of the motor and leaves by way of its B terminal. Thus, when phase I is positive, pole 1 is a north pole, and pole 2 a south pole. Likewise, when phase II is positive, 3 is a north pole and 4 a south pole; also,

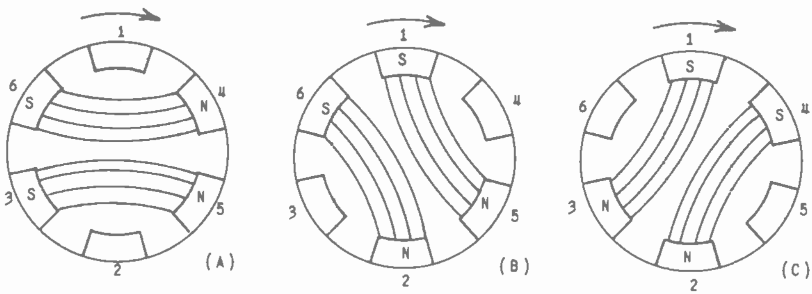


Fig.16 Showing how a revolving field is produced in a three-phase induction motor.

when phase III is positive, 5 is a north pole and 6 a south pole. When any of the phases are negative, the two poles corresponding to that phase reverse polarity.

Now referring to Fig. 15, let us consider the instant marked 1. At this time, the current of phase I is zero, whereas that of phases II and III are equal and opposite. Phase III is positive, so that pole 5 is a north pole and pole 6 a south pole. Also, phase II is negative, so that 3 is a south pole and 4 a north pole. Therefore, the field flux at this instant will have the distribution shown at A in Fig. 16..

At the time represented by 2 on the graph in Fig. 15, phase II is zero, phase III is positive, and phase I is negative. Thus, poles 5 and 6 are respectively north and south, and poles 1 and 2 are respectively south and north, which causes the flux distribution to assume the form shown at B in Fig. 16. Notice that the direction of the field flux in the second case has been rotated to the right.

At the instant designated by 3 on the graph in Fig. 15, phase III is zero, whereas phases II and I are respectively positive and negative. This causes poles 3 and 4 to be north and south respectively and produces a north pole at pole 2 and a south pole at pole 1. In this case, the flux distribution is as shown at C in Fig. 16. Notice that the direction of the flux has been rotated even farther to the right than in the second case. At the first instant, the direction of the flux was horizontal and toward the left; at the second instant, it was upward and toward the left; and at the last instant, it was upward and toward the right. Thus, it is evident that the field flux is rotating around the field frame in a clockwise direction. If we were to consider further instants, we would find that the flux would continue to rotate around the field frame, at a speed which depended on the frequency of the applied voltage, and upon the number of poles. The motor, which we have been considering, is called a "two-pole" motor, since there are just two poles per phase. In this case, the rotating flux makes one revolution for each cycle of the applied voltage. A four-pole motor, on the other hand, would have a total of 12 poles, and the rotating flux would experience one revolution during the time that the alternating applied voltage passed through two cycles. Therefore, the larger the number of poles per phase, the slower will be the speed of rotation of the revolving flux.

We now have the first requisite of an induction motor, the rotating field flux. In the preceding case, we have represented the stationary field frame as having six projecting poles; this, however, is never the case in actual commercial machines. Instead of having projecting poles, the field frame of an induction motor has a flat surface in which slots are cut. In these slots are placed the field conductors, much in the same manner in which the stationary armature of an alternator is wound. Thus, the flux does not emanate from definite pole pieces, but, instead, various areas of this field winding become alternately north and south. The number of poles possessed by the machine depends on the type of winding, and upon how close together the field conductors lie. If it happens that there are but two pole areas at any instant, the machine is a two-pole motor. This corresponds to the motor which we have been discussing, which has two poles per phase, or six poles in all. It should be realized that these two pole areas move progressively around the surface of the field frame, thereby producing the rotating flux. It is, of course, possible to wind the field so that there will be four pole areas at any instant, in which case the machine is said to have four poles. When a machine has definite pole pieces, it is said to have salient poles; the induction motor does not have salient poles.

Instead of calling the winding which produces the rotating flux the field, it is more commonly known as the "stator". Likewise, the revolving part of the motor is called the "rotor" instead of the "armature". There are, in general, two types of rotors; one is called the "squirrel-cage" type. This type consists of a laminated, soft-iron core in which slots are cut. The winding is composed of heavy copper bars embedded in these slots, and the bars are all short-circuited at either end by being soldered to heavy copper rings, punched with holes to slip over the ends of the bar winding. This type of rotor has no brushes, slip rings or commutators, and the principle of operation may be explained as follows:

As the field flux rotates, it induces voltages in these copper bars and fairly heavy currents flow through the rotor. The interaction of the rotating field flux and the flux created by the rotor currents produces a torque which causes the rotor to revolve in the same direction as the rotating field. The speed at which the field flux rotates is called the "synchronous speed". The speed of the rotor must always be less than the synchronous speed in order for a torque to be produced. It is obvious that if the rotor and the field flux were rotating at the same speed, the copper bars would not be cutting through the flux and no torque would be developed. When the induction motor is operating under no load, the rotor speed is nearly equal to the synchronous speed, there being just enough difference so that the bars will cut a small amount of flux sufficient to produce the small torque needed to overcome the friction of the bearings and other losses which the machine has. In this case, the current drawn by the stator is fairly small.

As the load on the motor is increased, the rotor tends to slow down. This causes the copper bars to cut more flux, and produces enough extra torque to drive the increased load. Furthermore, since the rotor conductors are cutting more flux, the rotor current will be larger, and by the phenomenon of reflected impedance, this action tends to load the stator more heavily, and the stator winding consequently draws more line current.

From this explanation, it is apparent that the induction motor is very similar to a transformer. The stator corresponds to the primary, and the rotor to the secondary. The alternating voltage is applied to the stator, and the rotating field sets up voltages in the rotor. As the field flux rotates, it drags the rotor after it. Also, just as increasing the load on the secondary of a transformer causes the primary to draw more current, so does increasing the mechanical load on the rotor cause it to slow down, to cut more flux, to create a larger rotor current, and finally to cause the stator to draw more line current.

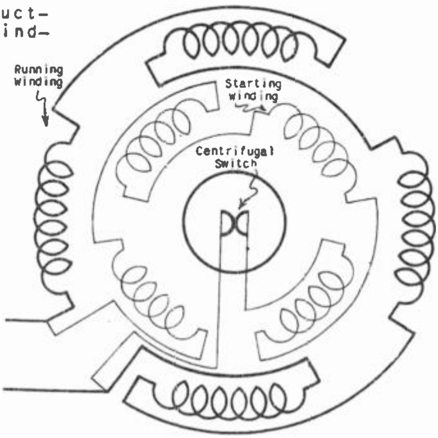
The other type of rotor is called a "wound-rotor"; it is very similar to the armature of a revolving-armature type alternator. There are ordinarily three windings connected in a Y-arrangement. The three free ends are brought out to three slip rings on which brushes make contact. It is thus seen that the rotor windings do not constitute closed circuits until the brushes are connected together externally. The purpose of this method is to allow more or less resistance to be cut into or out of the rotor circuits by means of a special rheostat. It is evident that when an induction motor is started, the rotor conductors will cut considerable flux until the rotor reaches its running speed. Therefore, the rotor currents, and stator current as well, are apt to be excessive. By introducing resistance into the rotor circuit when the motor is started, the rotor current is held to a safe value, which in turn prevents the stator current from becoming excessive. Nearly all induction motors under 5 hp. use squirrel-cage rotors, whereas most larger motors are of the wound-rotor variety.

The characteristics of an induction motor are very similar to a DC shunt-wound motor. The induction motor tends to run at a fairly constant speed, and therefore has good speed regulation. Various attempts have been made to control the speed of an induction motor, but all of them are either inefficient or very complicated and costly. Varying the rotor circuit resistance will produce a change in the speed (more resistance, slower speed), but this method is very inefficient, and results in poor speed regulation, and reduced horsepower output. Another method employs a series of switches to which the stator windings are connected, and by operating these switches the number of poles of the motor may be changed. This causes the synchronous speed to vary, and likewise produces a change in the rotor speed. The motor might be so wound that throwing these switches one way would cause it to have six poles, and throwing the switches the other way would increase the poles to eight. Such an arrangement is efficient, but requires rather complicated connections.

(B) The Single-Phase Type. Single-phase induction motors are used only in the smaller sizes, or in localities where three-phase power is not available. They are considerably less efficient than the three-phase type, and, for a given horsepower output, they are larger and more expensive. Furthermore, they are not self-starting. The field of a single-phase motor periodically changes direction as the current through its windings alternates, but this field does not rotate, and consequently the motor has no starting torque. If, however, the rotor of such a motor is brought up to nearly synchronous speed by some external means; the rotor currents, acting in conjunction with the current through the single-phase stator, do produce a rotating flux which creates the torque necessary to drive a load. It is, therefore, essential that some means be provided for starting the motor.

There are several methods which may be used to start a single-phase induction motor, but the most common for motors of moderate size is that known as the "split-phase" method. Motors of this type have, in addition to the regular field winding, a starting winding, the magnetic field of which is spaced 90 electrical degrees from the main field. (See Fig. 17.) Furthermore, this starting winding is composed of many turns of comparatively small wire, and possesses considerable inductance. The main field winding, on the other hand, has very little inductance. Thus, the current and the flux of the main field are nearly in phase with the line voltage, whereas the current and flux of the starting winding lag almost 90° behind the line voltage. The combination of these two fields which are spaced 90° from each other, and which also are practically 90° out of phase, produces a resultant flux which rotates around the field frame. Thus, a starting torque is created, and the motor becomes self-starting. If it so happens that the inductance of the starting winding is not large enough, it may be increased by connecting an external inductance in series with it.

Fig. 17 A single-phase induction motor, with a starting winding.



After the machine has reached its operating speed, the starting winding should be disconnected, because it is a source of power loss, and also, its winding is not designed to carry current for an appreciable length of time. This action is usually performed automatically by means of a centrifugal switch which opens the starting winding after the motor has reached a certain speed.

Small single-phase induction motors, such as sometimes used in electric fans or other devices requiring equivalent power, employ a somewhat different method for developing the starting torque, Such machines make use of shading coils. A four-pole, single-phase, induction motor using shading coils is illustrated in Fig. 18. Machines of this type have projecting or salient poles, and surrounding approximately half of each pole face is a short-circuited turn of heavy copper. As the current through the field winding begins to rise, the field flux increases, and induces a voltage in this copper ring which causes a fairly large current to circulate in this one-turn coil. The current in this coil produces a magnetic field which directly opposes the main field flux which is threading through the turn, and, as a result, an unequal distribution of field flux is present on the surface of the field pole. That part of the pole face surrounded by the turn is weakened, and most of the flux emanates from the other half of the face. Fig. 18 shows the field distribution as the current is increasing.

As the current in the field winding reaches its maximum value. it is no longer changing (or rate of change at this time is zero).

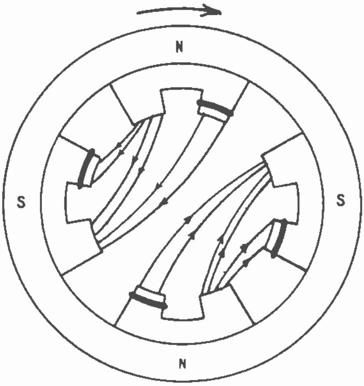


Fig. 18 Showing the field distribution of a shaded-pole when the current through the field is increasing.

and the voltage induced in the copper turn drops to zero. This causes the current to stop flowing in the turn and the magnetic field due to this current becomes zero, so that there is no longer any opposition to the main field flux. When this occurs, the flux distribution is practically uniform over the surface of the field pole.

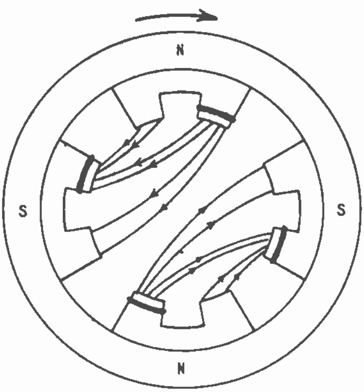


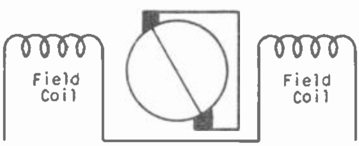
Fig. 19 The field distribution of a shaded-pole motor when the current is decreasing.

Now, as the field current begins to fall after having reached its maximum value, the voltage induced in the copper turn reverses polarity (in accordance with Lenz Law), and the magnetic field created by the current flowing in this turn aids the main field flux. This again results in an unequal distribution of the flux surrounding the pole face, and most of the flux emanates from that part of the face surrounded by the copper turn. (See Fig. 19.) Thus, the result is that the flux rises to a maximum in the unshaded portions of the pole face before it reaches its maximum in the shaded portions, and there is a progressive shift in the field from the unshaded to the shaded portions of the poles. This shift in the field causes the rotor conductors to be cut by flux, and the effect is the same as though a rotating flux had been produced. The fact that the rotor conductors are cut by the shifting flux causes voltages and currents to be created in the rotor, and a starting torque is created. The shading coils are left in the circuit after the motor has reached its operating speed, because the loss they introduce is of little importance in the small motors that use this method of starting.

There is another fairly common method used to start single-phase induction motors. In this arrangement, the motor is started as a repulsion motor, and is automatically converted to an induction motor after the operating speed is attained. It is called a "repulsion-induction" motor, and will be briefly discussed in the section on repulsion motors which immediately follows.

5. THE REPULSION MOTOR. In principle, the repulsion motor is very similar to the DC series motor. It has a commutator and two or more brushes. The armature current, however, is not supplied conductively from the line, but is furnished by the inductive relation between the stationary field and the armature. Fig. 20 shows a schematic diagram of a repulsion motor. Notice that

Fig. 20 A schematic diagram of a repulsion motor.



the two brushes are connected together, and that the plane of the brushes is at an acute angle with the plane of the field. As the current in the field winding alternates, it induces voltages in the armature conductors, and since the armature is effectively divided into two halves by the brushes, the currents in all the conductors in one half of the armature will flow in the same direction and those in the other half in the opposite direction.

The magnetic field set up by the armature current reacts with the main field and produces a torque which is in the same direction that the plane of the brushes have been shifted from the plane of the main field. As the direction of the main field changes, the induced armature voltages reverse, the armature current changes direction, and the torque is in the same direction as before. Therefore, the torque is uni-directional and pulsating. Furthermore, it may be proved that the magnitude of the torque depends on the angle between the plane of the brushes and the plane of the field flux, and that the speed of the motor may be changed by shifting the position of the brushes.

The losses of the repulsion motor are fairly great unless special compensating windings are employed. For this reason, most small motors of this type are started as repulsion motors and run as induction motors. This is accomplished by a centrifugal switch, which, when the motor reaches a certain speed, short-circuits all the commutator segments and lifts the brushes from the commutator.

6. THE SERIES AC MOTOR. The series AC motor has exactly the same principle of operation as the series DC motor. It has a commutator and brushes, and the field winding and armature are connected in series. It may be proved that any DC series motor will operate on alternating current, because the field current and armature current reverse direction simultaneously, with the result that both the main field flux and the armature flux change at the same time, and the torque has the same direction in either case. In fact, many small motors used to operate household appliances are of the series type, and are called "universal motors", since they work equally well on AC or DC.

Large AC series motors, although they operate on the same principle as the DC type, are usually somewhat different in construction. In the DC type, the field flux is steady, and the field frame is composed of one piece of cast iron. AC series motors, on the other hand, have a constantly varying field flux, and in order to prevent excessive eddy current losses, the entire field frame must be built up of laminations. In addition, it has been found that the losses of an AC series motor are lower when it is constructed without projecting poles, and this is also true of the repulsion motor. Furthermore, large AC series motors ordinarily have a compensating winding whose purpose is to minimize the losses. For these reasons, the AC series motor is more expensive than a DC series motor of the same capacity.

7. THE SYNCHRONOUS MOTOR. The synchronous motor is nothing more than an alternator running as a motor instead of a generator. It has a stationary armature and a revolving field frame. The field is excited with direct current by brushes which make contact with slip rings, and an alternating voltage is applied to the armature. Neither the single-phase nor three-phase type has any starting torque, and must be started by some external means. The synchronous motor has just one outstanding characteristic. It runs exactly at synchronous speed, some submultiple thereof, or not at all. The speed at which it will operate is the same speed at which the field would have to be revolved if the machine were working as an alternator, to produce an alternating voltage of the same frequency that is being applied to it. For example, if a given alternator must be driven at 1200 r.p.m. to produce a frequency of 60 cycles, then that alternator will operate as a synchronous motor running at 1200 r.p.m. when a 60 cycle voltage is applied to its armature.

One other feature of the synchronous motor is that it can be made to draw a leading current from the supply line by over-exciting its field. The induction motor, on the other hand, ordinarily draws a lagging current, and if there are a lot of induction motors on the same line, the power factor of that line is liable to be poor. To counteract this poor power factor, the power company often attaches a synchronous motor to the same line, and over - excites its field so that it will draw a leading current. By this method, the power factor may be brought up to nearly one, and the power company benefits, even though the synchronous motor may not be needed for any other purpose. When used in this manner, the synchronous motor is called a "synchronous condenser". Many times the power company will offer a premium in reduced rates to a large industrial plant using many induction motors, if they will install a synchronous motor to raise the power factor. This is of advantage to both the consumer and the power company, because the consumer gains nothing by having a low power factor.

8. COMPARISON OF DIFFERENT AC MOTORS. Of all the various types of AC motors, the most widely used, except for applications requiring very little mechanical power, is the three-phase induction motor. This motor is fairly simple in construction, is rugged, and is fairly efficient. It is essentially a constant - speed motor, and its speed cannot be easily varied without causing loss of efficiency. Thus, this type of motor is most suited for applications requiring a practically constant speed, and would not be advisable for railway traction work, where varying speeds are required.

The single - phase induction motor is somewhat less efficient than the three-phase type, but nevertheless finds wide application where three-phase power is not available. It, too, is a constant speed motor.

As regards efficiency, the series AC motor and the repulsion motor are practically in the same class. Both are less efficient than the induction motor, and both suffer from commutation troubles. They both, however, may be used at varying speeds, and find some application where a varying speed is desirable. A few transportation companies have installed AC series motors in their cars, but the fact that the DC series motor is more efficient, has caused most companies to cling to the DC type.

The synchronous motor finds its widest application in power factor correcting, although it is sometimes used where an absolutely constant speed is essential. Thus, the television scanning disc, is driven by this type of motor.

9. THE CARE OF MOTORS AND GENERATORS. In general, only two requirements are necessary for maintaining any motor or generator installation in proper operation. These are cleanliness and correct lubrication. Motors and generators do not need a great deal of care, but they do require attention at regular and periodic intervals. The best method of maintaining an installation is to establish a definite routine for cleaning and oiling.

The machines should always be installed in a location where vibration is at a minimum. Vibration is liable to cause the bearings to heat excessively, and is almost sure to cause sparking due to the inability of the brushes to make perfect contact with the vibrating commutator.

Nothing will cause a machine to develop trouble quicker than an accumulation of dirt. Greasy dirt on the commutator is sure to cause excessive sparking, and allowing dirt to clog the ventilation ducts will prevent the free circulation of air, thereby causing the machine to overheat. The machine should be cleaned with a duster or a clean cloth at frequent intervals.

On a DC machine, the commutator requires more attention than any other part. Furthermore, the condition of the commutator is a good indicator of the general condition of the machine, since the commutator gets more wear than any other part of the machine, and nearly any type of trouble is liable to affect the commutator. When the machine is functioning properly, the commutator should have a deep chocolate color, especially if carbon brushes are used; and it should acquire a highly polished surface.

Oil and grease on the commutator may have two different effects. The layer of oil may partially insulate the brushes from the segments, and thereby produce sparking, and the sparking, in turn, may partially carbonize the oil particles producing a leakage path from one segment to another.

Excessive sparking will always roughen the surface of the commutator, if it is allowed to continue; and the rough surface furthers the production of sparking. This rough surface may be smoothed by using a fine grade of sandpaper. Never, under any conditions, use emery paper or emery cloth for this purpose. The particles of the emery powder are conductors and they are liable to become wedged in the spaces between the segments and cause short circuits. Furthermore, the particles of emery dust may become embedded in the contact surface of the brushes where they will scratch the commutator surface. If the commutator is in very bad shape, it may be necessary to turn it down to a new level of smoothness in a lathe.

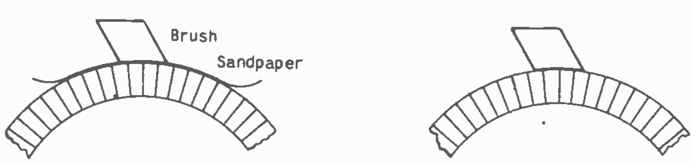


Fig.21 Illustrating the method of using sandpaper to fit the brush to the curvature of the commutator.

The next most frequent source of trouble is due to some defect or bad adjustment of the brushes. If the brush pressure is too low, a slight arc will be produced between the brushes and the commutator and this condition is aggravated by the roughening of the commutator which results. Adjust the tension of the brushes until a good contact results when the machine is operating. When new brushes are to be installed, their contact surfaces must be made sufficiently concave so that they will fit closely to the convex surface of the commutator. The correct curvature may be secured by using a strip of sandpaper slightly wider than the width of the brush, and by holding this piece of sandpaper on the surface of the commutator with the rough side up as shown in Fig. 21. With the brushes adjusted for proper tension, and with the paper held flat against the surface of the commutator, it is worked back and forth until the surface of the brush has the correct curvature. Unless the paper is firmly pressed against the commutator, the brush fits properly only at its center, as shown in Fig. 22. Thus only a small part of the brush's surface actually makes contact with the commutator, and excessive heating will be developed at this point. After the commutator has been smoothed and the brushes

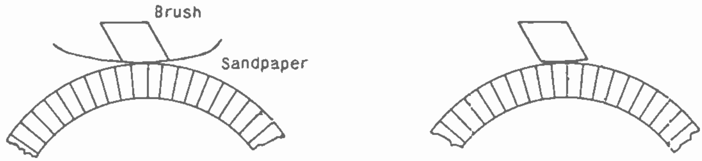


Fig.22 Showing how the wrong curvature may result unless the sandpaper is held properly.

fitted, all copper and carbon dust should be blown away with a hand bellows before the machine is again put into operation.

Unless the machine is equipped with commutating poles, the axis of the brushes must be shifted every time a change in the load occurs. Never change the position of the brushes on a machine fitted with commutating poles, for if sparking occurs, it is due to some other source than wrong position of the brushes.

Now we come to lubrication, and the first thing to be said on this subject is that oil, even the best grade obtainable, is far cheaper than repairing and replacing bearings. It is certainly poor economy to attempt to save by using an inferior grade of oil at the expense of the bearings. Whenever there are moving parts, there is friction, and it is the purpose of lubrication to reduce this friction to a minimum. The oil forms a thin film over the moving part and prevents its actual contact with the stationary part. After a time, this oil film becomes so thin it can no longer prevent contact, and friction occurs.

Some machines are fitted with oil rings and oil wells. When such is the case, these oil rings should be inspected frequently to make sure that they are loose, are dipping freely into the wells, and are supplying oil to the shaft in the proper manner. Other bearings are fitted with grease cups which should be given a quarter-turn once a week. The use of too much oil is almost as bad as too little oil. If the oil cups are filled too full, the oil will be splashed over the commutator and the armature windings.

Great care must be taken to keep dirt and grit out of the bearings. This is especially true of ball bearings, and roller bearings; and only grease which is known to be free from foreign particles should ever be used.

Heating of the bearings may be due to lack of oil, to a poor grade of oil, or to dirt in the bearing. When a hot bearing occurs, a definite procedure should be followed. First, do not shut off the machine. The heating of the bearing has caused the shaft to expand, and if the machine is stopped, the bearing will contract around the rotor shaft, and the shaft is said to "freeze" in the bearing. Should this happen, the shaft is held fast in the bearing and cannot rotate. The shaft must then be pried out, and this is the work of a skillful mechanic if the shaft is not to be ruined, Therefore, the first thing to remember is to keep the machine turning. If it is convenient, the machine may be slowed down somewhat. The next step is to flush out the bearing with a good grade of fairly heavy cylinder oil, while at the same time cold water or a rag dipped in ice water is applied to the shaft. Care must be taken that the water does not get on to the box. When the bearing has cooled, the machine may be stopped, but if there is any likelihood of the shaft having been scratched or scored, the machine should be taken down and the shaft inspected before it is again put into use.

Induction motors, of course, are not bothered with commutator troubles, and as long as they are kept free from dirt, and well lubricated, they will function properly.

Some of the more common troubles which occur in motors and generators are listed in the following summary:

A. Motor fails to start.
   1. Supply line dead.
   2. Fuses blown.
   3. Field or armature circuit open.
   4. Too much resistance in the field rheostat.
   5. Load on motor too heavy.

B. Generator fails to build up.
   1. Field circuit open.
   2. High resistance in the field rheostat.
   3. Commutator dirty.
   4. Brushes in the wrong position.
   5. No residual magnetism.
   6. Field connected incorrectly to the armature.
   7. Speed low.

C. Sparking at the commutator.
   1. Rough surface.
   2. Brushes in the wrong position.
   3. Insufficient brush tension.
   4. Poorly fitted brushes.
   5. Armature overloaded
   6. Vibration of the machine.
   7. Armature coil short-circuited.
   8. Armature coil open-circuited.
   9. Unequal brush spacing.

D. Heating.

1. Machine overloaded.
2. Ventilating ducts closed.
3. Insufficient lubrication.
4. Field strength below normal.
5. Any of the factors which also cause sparking at the commutator.

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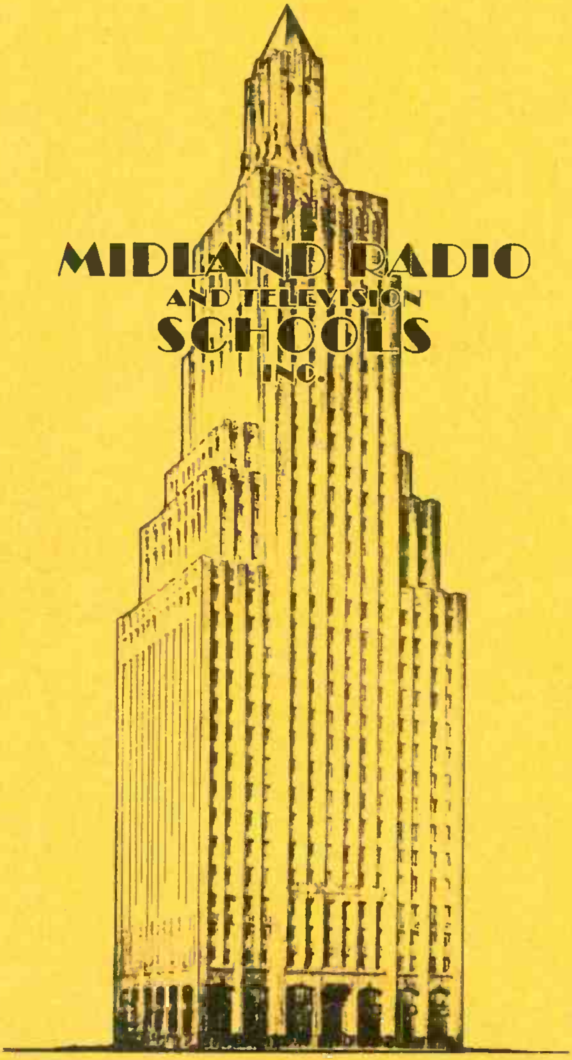
Midland Radio & Television Schools, Inc.

MIDLAND RADIO AND TELEVISION SCHOOLS INC.



POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

POWER TUBES

PLATE AND FILAMENT SUPPLIES

LESSON NO. 9

FROM CURIOSITY

.....TO MAMMOTH INDUSTRY.

The writer being an amateur wireless operator at the time of radio's advent, has had many occasions to compare the public's attitude toward radio then, and their attitude today.

When radio first blossomed forth, I was one of many amateurs who became crystal set manufacturers on a small scale. We sold the sets for $25.00 and charged as much as $10.00 to install an antenna. It is amusing to recall that the people who bought the little sets, considered that radio had reached it's peak, and that it would never become anything of consequence.

On one occasion I was attending a bazaar for the benefit of something or other in Chicago. In one room of the building, an enterprising young man had a crystal set in operation and was charging twenty-five cents for a 'listen'. A long line of people waited for an opportunity to pay the quarter and 'listen'. And while waiting, I heard many of them remark that it was a 'trick'....'I know good and well you can't send music through the air'......'Why of course you can't, it's absurd....etc., etc.

While the people of today have become accustomed to expecting the 'impossible', many still maintain the same attitude toward television, that was so apparent when radio broadcasting first became a reality.

Every new development of a startling, scientific nature will probably be met with ridicule, skepticism, and doubt. But as the new development becomes an established industry, the very people who ridiculed will make good use of the new development. And in the majority of cased, they will even forget that any doubt ever existed in their minds.

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Lesson Nine

POWER TUBES. PLATE & FILAMENT SUPPLIES

"Without the modern high-powered tubes available today, we would not have the vast radio system which not only entertains us in our homes but carries on world commerce through the air. In this lesson, we are giving you fundamental information concerning the construction and operation of high-powered tubes.

"The second part of this lesson is devoted to modern means of securing plate and filament supplies for high-powered transmitting tubes. I am sure you will find this material interesting and helpful."

PART I POWER TUBES

1. **TYPES OF POWER TUBES.** It is the purpose of the radio transmitter to develop large amounts of radio - frequency power. The oscillator itself produces comparatively little power, since its major purpose is to determine the frequency of transmission, and it is better able to perform this duty when the power demands made on it are small. The first buffer serves the dual purpose of amplifying the output of the oscillator and isolating it from the variations taking place in the modulated stage. It may or may not be operated as a power amplifier. In some cases, it is merely a voltage amplifier which increases the R.F. voltage variations occurring across the plate tank circuit of the oscillator. When so used, it places no load on the oscillator, and thus makes for better frequency stability. Even in this case, there is more R.F. power in the plate tank circuit of the buffer than there is present in the output of the oscillator. On the other hand, the buffer is sometimes allowed to draw a small amount of grid current, and is operated with the idea of increasing the power output of the oscillator. This is permissible if the load placed on the oscillator is light.

The stages following the first buffer are definitely power amplifiers; their sole purpose is to increase the amount of R.F. power until it is sufficient to produce considerable radiation from the transmitting antenna. It is, therefore, evident that practically all of the tubes in a radio transmitter (with the exception of the rectifiers) are power amplifiers, and the first part of this lesson will deal with their construction and care.

2. **THE EMITTER.** The emitter may be called the heart of any vacuum tube; it is the source of the electrons composing the plate current. Unless it is of rugged construction and is easily able to supply the needed electrons, the life of the tube will be short. Various materials which are used for emitters are tungsten, oxide-coated tungsten, and thoriated tungsten. Of these three, the oxide-coated type has the highest thermionic efficiency; that is, for a given amount of emission, it requires the least amount of filament-heating power. However, neither the oxide-coated nor thoriated types are used with tubes requiring a high plate voltage. The difficulty lies in the fact that it is impossible to eliminate completely all traces of gas in a vacuum tube. Under the influence of high plate voltages, the primary electrons acquire tremendous velocities and collide with the residual gas molecules with sufficient force to dislodge one or more electrons from them. This action changes the gas molecules into positive ions, which are attracted by the negative filament. The positive ions strike the filament with considerable force, actually stripping the oxide coating or thorium layer from its surface. Thus the filament gradually loses its emitting ability, and the life of the tube is comparatively short. For this reason, many tubes using plate voltages much in excess of 1000 volts employ pure tungsten filaments.

The dimensions of the filament are determined by the peak space current which includes the plate current, grid current, and, in addition, the screen current, if the tube is a tetrode. When tungsten is used for the filament, it is designed so that the peak current is practically equal to the full emission of the filament. Thoriated and oxide-coated filaments require that the peak emission be several times as large as the maximum space current in order to allow for deterioration of the emission surface.

Considerable power is required to heat the filaments of the larger sized power tubes, in fact, some of the water-cooled tubes use several thousand watts of filament power. The filament voltage is never very large, being only 33 volts for the largest tube manufactured; however, the current drawn by this filament is 207 amperes, and the leads must be of ample size. This filament power is not ordinarily taken into account in calculating the efficiency of the tube. Since the resistance of the filament, when cold, is much less than after it has reached its operating temperature, it is customary with large tubes to place a resistance in series with the filament when the filament voltage is first applied. Without this starting resistance, the large current that would otherwise flow might damage the lead wires passing through the glass seal.

3. **VOLTAGE REQUIREMENTS.** In addition to the normal DC plate voltage, the voltage between the plate of the tube and the filament consists of an R.F. component with a peak value nearly equal to the applied plate voltage. Furthermore, if the tube is used in a modulated stage, there is also an A.F. component of plate voltage, which,

if the degree of modulation is 100%, is equal in peak value to the applied DC plate voltage. Thus the applied DC plate voltage may be only 1000 volts, but it is probable that the peak voltage existing between the plate and filament will be nearly 4000 volts. (It is assumed that the stage is 100% modulated, and that the applied voltage varies from 0 to 2000 volts. In this case, the R.F. voltage across the plate tank circuit will have a peak value of nearly 2000 volts, during the time that the applied voltage is 2000 volts, and the peak plate voltage will approach 4000 volts.)

At the higher frequencies, the R.F. component of the plate voltage produces large dielectric losses in the glass envelope which is in its electrostatic field. The losses are largely due to the high temperature of the glass during normal operation, and increase in amount as the frequency is raised. These losses often result in excessive heating of particular spots of the envelope which may soften the glass and destroy the vacuum. Even though these effects are minimized as much as possible by special design, it is usually necessary to lower the plate voltage somewhat when operating the tube at very high frequencies.

4. **HEAT DISSIPATION.** The power supplied to a vacuum tube, other than the filament power, is determined by the product of the average plate current, and the voltage of the power supply. This power is limited by the maximum permissible plate voltage and the allowable peak plate current. Of this total power, a part is converted into R.F. energy which drives other tubes or is radiated from an antenna. The remainder is dissipated in the tube and its circuits in the form of heat. The amount actually dissipated in the tank circuits and the connecting leads is small compared to that which must be converted into heat within the tube. The major part of this internal dissipation takes place at the plate of the tube, which reaches a relatively high temperature during normal operation. The plate must be capable of radiating this heat to the walls of the tube and thence to the surrounding air without either the plate or the glass envelope becoming excessively hot. To facilitate this heat radiation, the plate is usually blackened, because a dull black surface is able to radiate heat better than a smooth polished surface. Some manufacturers make the plates of carbon, since it possesses a large radiation ability.

There is also a power loss at the grid of the tube which is somewhat smaller than the plate loss, but which must be taken into account because it is not permissible to allow the grid to become red hot, as this would cause it to emit electrons.

Another source of power loss in a vacuum tube is that due to the circulating R.F. grid and plate currents. Nearly all tube manufacturers specify how much circulating R.F. current is permissible without causing overheating. This circulating current is not the ordinary grid or plate currents, nor their R.F. components; but is the charging current taken by the interelectrode capacities of the tube. The plate tank circuit is connected between the plate and the filament and therefore causes an R.F. voltage to be present between these two points. This R.F. voltage causes an R.F. current to circulate between the plate and the filament, charging the capacity existing between them. In a like manner, a circulating R.F. current flows between the grid and filament electrodes to charge the grid-to-filament capacity. At low or medium frequencies, these circulating currents are of no consequence. With high frequencies, however, the interelectrode capacitances have such low capacitive reactances that the circulating current is considerable and large electrode leads must be used to carry the R.F. current without excessive heating. The maximum allowable circulating R.F. current is, for most tubes, several amperes.

5. CONSTRUCTION OF POWER TUBES. The problem which presents the greatest difficulty in the manufacture of high-powered tubes is the production of the vacuum. The air originally in the tube may be easily pumped out with mechanical and diffusion suction pumps; however, there are many gas molecules which adhere to the metal electrodes and the inner walls of the glass envelope. This gas is said to be occluded by the metal and glass parts of the tube. The gas molecules enter the pores of the glass and the metal and are extremely difficult to remove. In fact, it is possible to remove them only by heating the parts which have absorbed the gas.

The pumping procedure is performed while the tube is in a special oven where the temperature is just below the softening point of the glass envelope. It is necessary to heat the elements of the tube to temperatures above that which they will experience during normal operation, or else additional gas will be liberated by the metal parts while the tube is in operation. Therefore, before the tube is removed from the exhausting pump, the metal electrodes are brought to a red heat by inducing a high - frequency voltage into them, which heats the metal parts without directly heating the glass. When it is realized that a tube is connected to the exhausting pump for approximately 24 hours to remove the absorbed gas, the difficulty in obtaining the desired vacuum may be readily understood.

There are only a few materials which are suitable for the grid and plate electrodes. These include tungsten, molybdenum, and tantalum, while carbon is often used for the plate. The grids of most tubes are made of tungsten, because of its ability to withstand high temperatures, however, molybdenum is rapidly gaining in favor. Plates are generally constructed of molybdenum, although carbon plates are featured by some manufacturers because of their greater heat - radiating ability. A few manufacturers employ tantalum exclusively for the grids and plates of their tubes. It has the added advantage that it liberates almost all the occluded gas at a bright-red heat, and tends to absorb the residual gases at lower temperatures so that the tube will be essentially free from gas during its normal life. Furthermore, it is softer than tungsten and may be easily welded to form solid joints having low resistance; a feature which is especially desirable for high-frequency operation.

In general, the main differences between power tubes and receiving tubes are the very high voltages that the tubes can stand, the large space currents that they are able to produce, and their ability to dissipate large power losses. Several representative

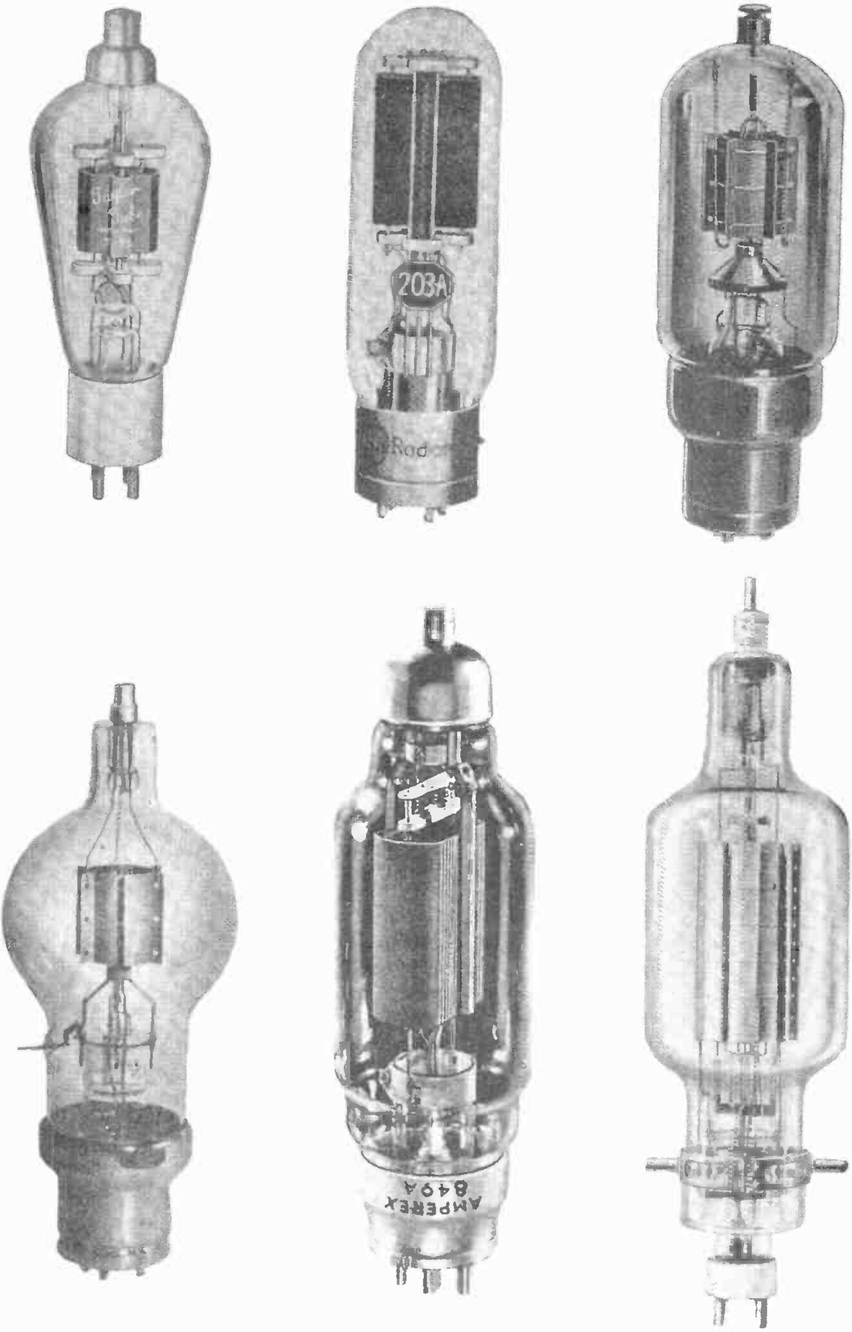
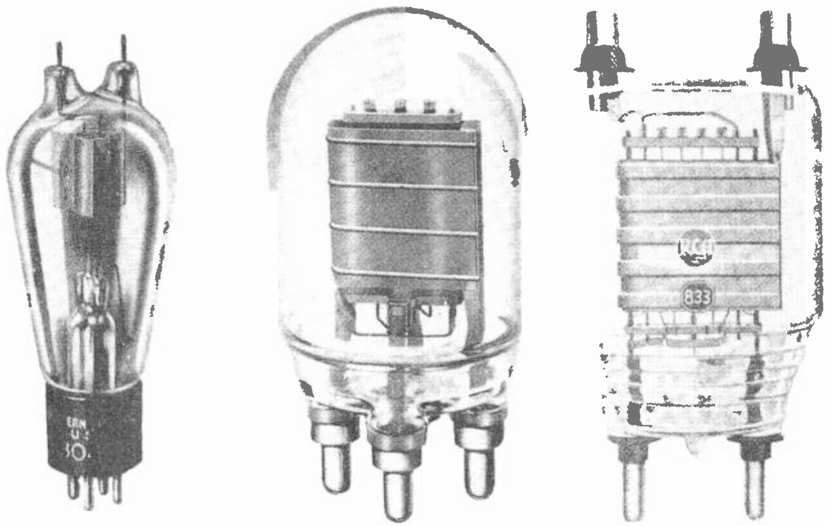


Fig. 1 Representative air-cooled power tubes.

air-cooled power tubes are illustrated in Fig. 1. Notice that the size of the glass bulb is proportional to the power rating. Furthermore, tubes designed for high voltages are so arranged that the plate connection enters the tube at a point remote from the grid and filament connection. This increases the insulation resistance between the leads and, in addition, tends to reduce the interelectrode capacities.

Power tubes are rated on a basis determined by their ability to dissipate the internal power losses. Thus a 50-watt tube is one capable of dissipating 50 watts at its plate without becoming overheated. With an efficiency of 50%, such a tube is capable of producing 50 watts of R.F. power. Since most ratings are fairly conservative, it is wholly possible that this tube could produce an output power of at least 100 watts, indicating an efficiency of 67%.

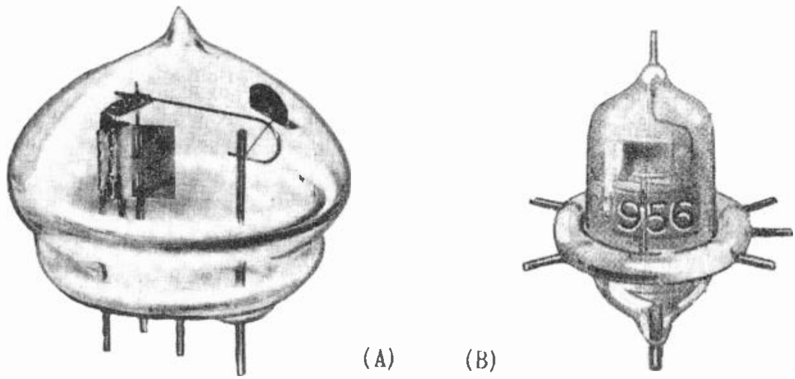


(A) (B) (C)

Fig. 2 Air-cooled tubes designed for short waves.

Many of the common power tubes are not suitable for high-frequency operation because of their large interelectrode capacities. For this reason, special high-frequency tubes have been designed to minimize these capacities as much as possible. Such tubes employ electrodes of small surface area, widely spaced, and often each lead enters the glass at a point remote from all others. Several tubes of this type are illustrated in Fig. 2. The one at A is a Western Electric type 304B and has a plate dissipation of 50 watts. Notice how the grid and plate leads are separated in this tube in order to minimize the interelectrode capacities. The tube shown at B is also a Western Electric tube and its type number is WL-461. It has a plate dissipation of 160 watts. The one at C is an RCA type 333, having a plate dissipation of 300 watts. The lead wires in the last two tubes are especially large and rugged as they must be to handle the large R.F. currents.

For ultra-high frequencies, even these tubes are unsuitable. It has been determined that the only feasible method of reducing the interelectrode capacities is to reduce the physical size of the tube itself. This idea has been given extensive consideration for the past few years, and the result has been several extremely small tubes particularly adapted for ultra-high frequency operation. One such tube is the Western Electric type 316A shown at A, Fig. 3. It is sometimes called the "door knob" tube from its general resemblance to this familiar object. Its comparatively small size may be judged from the fact that its longest dimension is less than 3".



(A) (B)

Fig.3 Ultra-high frequency tubes

Naturally, such a small tube cannot dissipate as much power at its plate as some of the larger tubes, but it will operate at frequencies much higher than most power tubes. It has a maximum plate dissipation of 30 watts, and will deliver 3.5 watts of R.F. power at 300 megacycles. Its upper frequency limit is 750 megacycles.

Another type of ultra-high frequency tube is illustrated at B in Fig. 3. It is the so-called "acorn tube" manufactured by RCA. Its longest dimension is about 1¾", and it is manufactured in both triode and pentode types. (This illustration is larger than actual size.) With an upper frequency limit of about .5 meter, it has become popular for ultra-short wave oscillators and receivers.

6. WATER-COOLED TUBES. The total power to be dissipated in an air-cooled tube consists of the plate losses, the grid losses, and the heat generated by the filament. Tubes having a rating of less than 2000 watts can dissipate this energy through the walls

of the glass envelope without increasing its temperature to a dangerous point. However, the amount of energy that each square inch of the glass envelope may radiate safely is comparatively low, and larger tubes would need to be enclosed in glass bulbs of prohibitive size. Sometimes a forced draft created by a fan is necessary to prevent the glass from reaching the softening point.

Since a large proportion of this heat dissipation is produced at the plate of the tube, this problem has been partially solved by using water-cooled plates. A diagram of a tube of this type is illustrated in Fig. 4. The plate consists of a copper cylinder joined to the glass envelope by an air tight metal-glass seal. The copper plate is then inserted into a water jacket through which water circulates freely. Thus, the copper cylinder serves as both the plate and a part of the wall of the tube. Since it is in direct contact with the cooling water, many kilowatts of energy may be dissipated before a dangerous temperature is reached.

The major problem in the design of water-cooled tubes has been the metal-glass seal. Naturally, a type of glass must be used which has the same coefficient of expansion as the copper plate, otherwise the seal would break during temperature changes. Much research has been conducted on this particular problem, and it was not until a new type of glass was developed that water-cooled tubes became practical.

The largest water-cooled tube commercially manufactured is the RCA type 862. It has a maximum plate dissipation of 100,000 watts, and is over 5 feet in length. Although water cooling the plate of

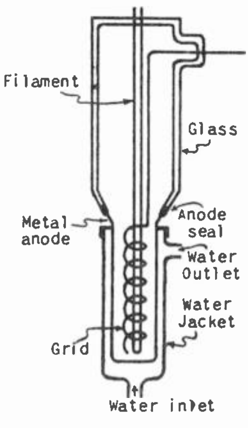
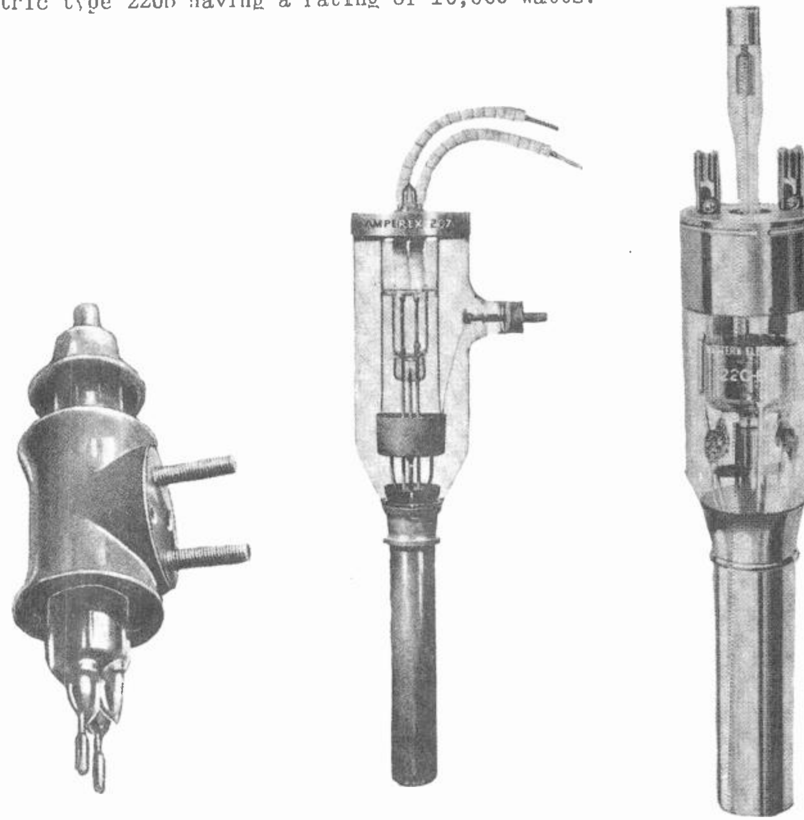


Fig.4 Construction of a water-cooled tube.

a tube helps the dissipation of heat energy, the losses of the grids of large tubes is so great that the grids sometimes operate at a red heat. Tubes with water-cooled grids have been designed, although they are not now in commercial use.

Three different types of water-cooled tubes are illustrated in Fig. 5. The tube shown at A is a comparatively new tube designed by RCA. It is especially adapted for high-frequency operation and will deliver an R.F. power of 1200 watts at 1¼ meters. It is a type No. 888. The one at B is an Amperex type 207 capable of dissipating 7,500 watts at its plate. The one at C is a Western Electric type 220B having a rating of 10,000 watts.



(A) (B) (C)

Fig.5 Representative water-cooled tubes

The plate connection is usually made to the water jacket itself, and the shoulder on the copper plate must make good contact with the jacket. Furthermore, the jacket must provide the proper clearance around the plate to allow for the free circulation of the cooling water. It is common practice to ground the filaments of all vacuum tubes; this makes the plate highly positive with respect to ground. In a water-cooled tube, the plate is in direct contact with the cooling water and a leakage path to ground is provided through the conductivity of the water. Naturally, the hose used to carry the water to the jacket must be a non-conductor, and either rubber hose or Isolantite pipe is used for this purpose. The inlet and outlet hose should each be at least 15 feet long so that the conductivity between the plate and ground will be very low. The conductivity of the water will depend upon how much mineral matter it has in solution. Although tap water may sometimes be used, distilled water is desirable. A photograph of a water-cooled tube in its jacket is shown in Fig. 6.

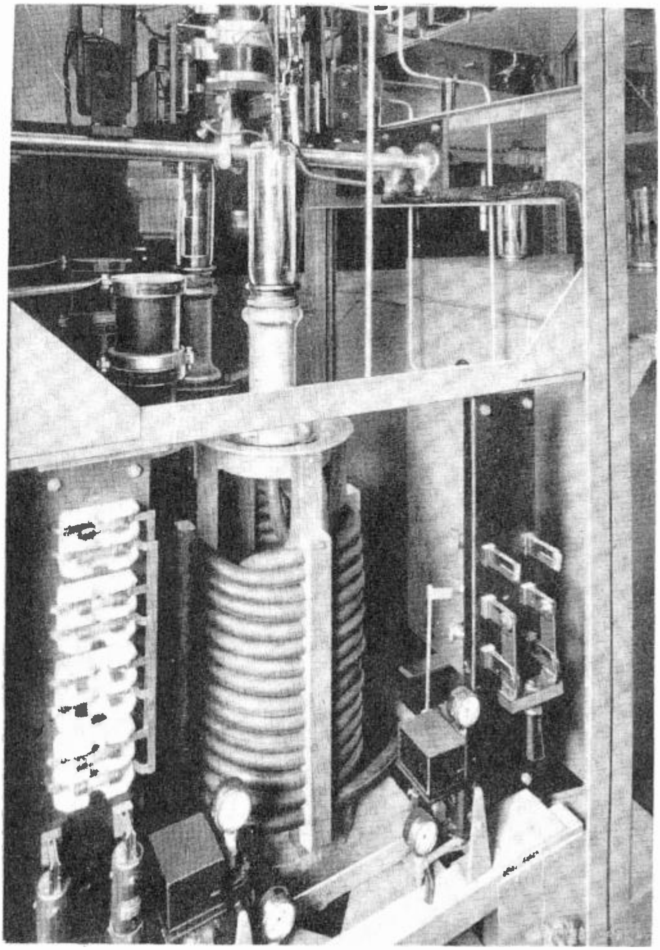


Fig. 6 Photograph of a water-cooled tube in its jacket. The hose carrying the cooling water is clearly visible.

As the water flows from the jacket, it has a fairly high temperature, perhaps as high as 50° C. Since it is desirable to use distilled water, the same water may be used over and over again, if it is cooled in some manner. One type of cooling arrangement is illustrated in Fig. 7. A reservoir is provided and a centrifugal pump forces water from the reservoir through a radiator to the cooling jackets, and thence back to the reservoir. A large blower fan creates a draft through the radiator which cools the water. The cooling arrangement shown in the figure is used with a 5 kw. and a 50 kw. tube. The water is circulated at the rate of

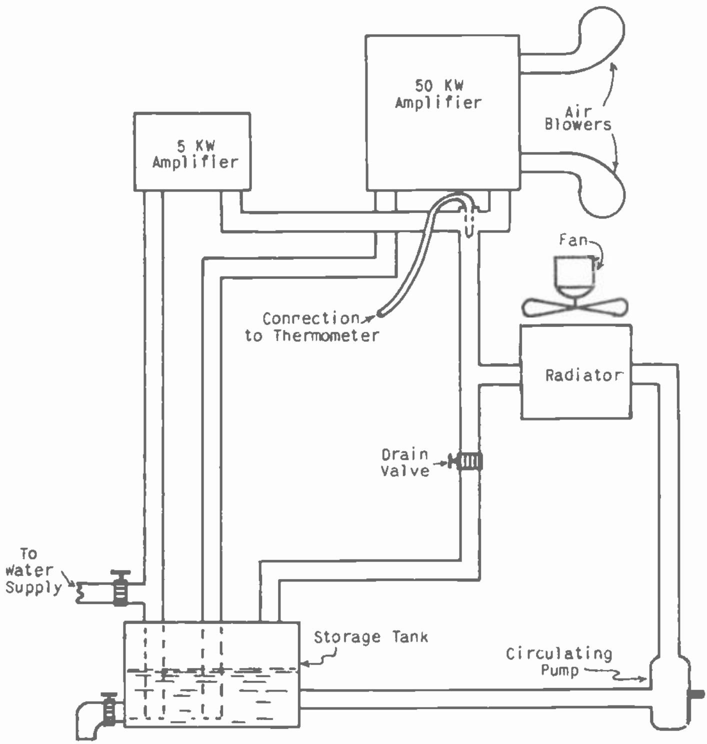


Fig.7 Diagram of a water-cooling system.

50 gallons per minute, and the blower fan forces 15,000 cu. ft. of air through the radiator per minute. In addition, the larger tube is equipped with two blower fans which aid in keeping the temperature of the tube below the danger point. A dial thermometer is provided to indicate the temperature of the water as it comes from the water jackets. Included in this arrangement are protective relays which disconnect all voltages from the tubes in case the temperature of the water leaving the jackets exceeds a predetermined value, usually 70° C; or in case the circulation of the water is interrupted.

Water-cooled tubes are very expensive, and reasonable care should be taken in handling them. In case city water is used for cooling, it is probable that a scale will form on the plate of the tube after it has been in operation for some time. This scale should be removed for it obstructs the free circulation of the cooling water. The tube must be taken from its water jacket, and the scale removed by rinsing the plate with weak hydrochloric acid. When the plate is again clean, it must be thoroughly washed with pure water to remove all traces of the cleansing acid. A water-tight gasket is used to seal the top of the water jacket, and it should be firmly bolted into place. Do not, however, use any adhesive cement to make the seal water-tight, for the tube may be damaged when it is again dislodged from its jacket.

Inspect the hose connections at regular intervals to make sure that they are water-tight and are free from corrosion. The moving parts of the water jacket should be coated with a light film of oil to prevent corrosion.

Fig.8 A water-cooled tube with cooling fins attached.

Within the past year there has been developed a 5 kw. transmitter in which all the tubes are air-cooled. This is a new transmitter built by RCA, and the final stage and modulator employ regular water-cooled tubes; however, by a rather ingenious arrangement, it is not necessary to use water for cooling them. This is made possible by silver-soldering a large number of copper fins to the metal plate and blowing air through them. Fig. 8 shows how one of these tubes appears after the addition of the cooling fins. This method has proved highly successful, and the modified tubes are now available from RCA. The tubes, fins and all, are placed into porcelain cups and a large flexible hose connects the cup to the blower fan as shown in Fig. 9. A dust screen is placed over the blower to prevent the collection of dust on the fins. Several advantages are claimed for this system. First, there can be no accumulation

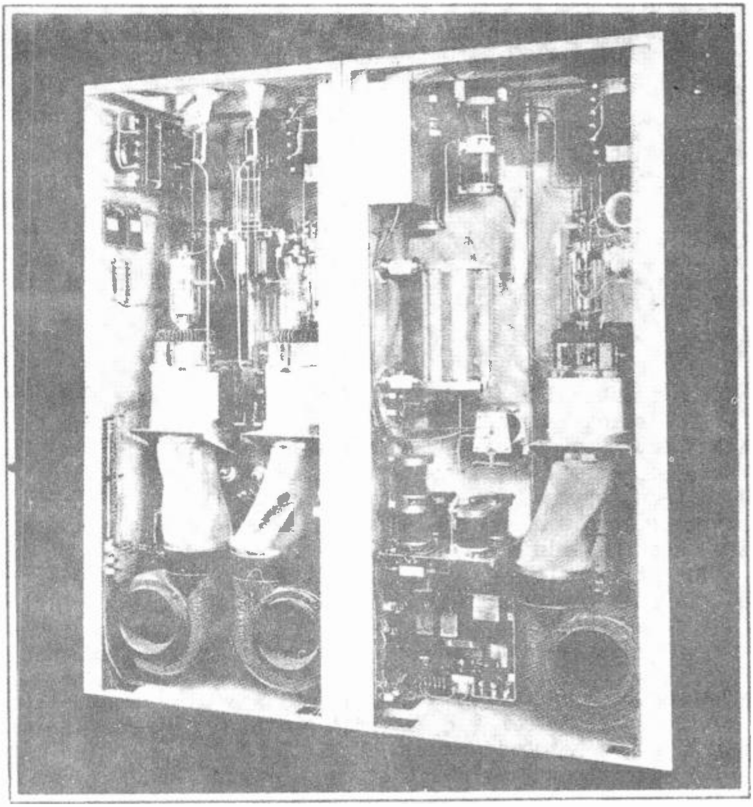


Fig.9 Photograph of a transmitter employing water-cooled tubes with cooling fins.

of scale on the copper plates of the tubes; second, the tubes are more evenly cooled, and the metal-glass seal is cooled as well as the plate; and third, the fact that the radiation surface is much larger tends to prevent the formation of hot spots.

7. **EMISSION TEST FOR POWER TUBES.** After a power tube has been in operation for some time, it naturally loses a part of its emitting ability. It is, of course, desirable to know the condition of a tube at any time so that a replacement may be made before the tube breaks down altogether and causes loss of time on the air. A record is kept of the life of each tube; this record includes the date of purchase of the tube and the number of hours it has been in service. After a tube has had 1,000 hours of service, it should be tested in some manner to ascertain whether it should be kept in the transmitter or should be retired from service. The best way of testing a power tube is to replace it by one known to be good.

Since a complete set of spare tubes are always on hand at any transmitter, this is a simple method. If the operation of the transmitter is much improved with the new tube, it should be left in the transmitter, and the old tube kept as an extra spare.

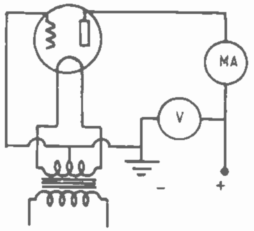


Fig. 10 Circuit used to determine emission qualities of a power tube.

Often a tube will appear to be operating properly, but due to age, its emission is gradually becoming weaker, and it may suddenly fail without any warning. To eliminate this possibility, it is desirable to perform an emission test on a power tube at least every 1,000 hours of service. To perform this test, the tube is connected in a circuit as shown in Fig. 10. The proper filament voltage is applied, and the grid is connected to the center tap of the filament. The plate voltage is next set at a low value and the plate current is read and recorded. The plate voltage is now increased in equal steps and the plate current for each voltage is recorded. A graph of plate voltage and plate current is plotted, and from this graph the condition of the tube is determined. Since all of the power applied to the tube must be dissipated at the plate, it will not be possible to increase the plate voltage to the normal operating value without exceeding the plate dissipation rating. The plate voltage must not be increased beyond the point at which the plate of the tube becomes a dull red.

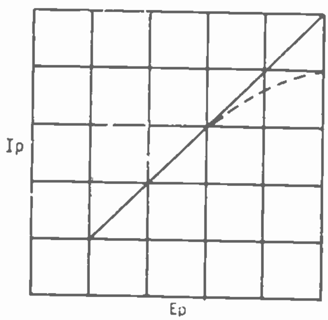


Fig. 11 Emission characteristic of a power tube. The solid line indicates a good tube; the dotted line, one in which the emission is faulty.

If the plate voltage - plate current curve is practically a straight line, the emission of the filament is normal; however, if the plate current begins to drop off with increases in plate voltage, as shown by the dotted line in Fig. 11, it is an indication that the tube has lost a part of its emitting ability and should be kept only as a spare.

PART II PLATE AND FILAMENT SUPPLIES

8. **THE FILAMENT SUPPLY.** Among the sources which are or have been used to supply filament heating power may be mentioned batteries, direct-current generators, and commercial AC supply lines used with step-down transformers. There is little doubt that from a noise-free standpoint, the DC filament sources are far the best. DC generators are sometimes used in transmitting stations to eliminate the possibility of AC hum; storage batteries are used in automobile and aircraft receivers; and primary batteries are employed for portable equipment. However, the large economy which results from the use of alternating-current power has caused it to replace gradually the other forms of filament power wherever possible.

There are two types of AC hum which may develop when 60 cycle supply lines are used to heat the filaments of vacuum tubes. The first is a 60 cycle fundamental hum frequency, and the second is a 120 cycle second harmonic frequency of the supply line. Let us first consider how these hums may develop in cathode-type tubes. The heater current creates alternating electrostatic and magnetic fields and the electrostatic fields from the unshielded portions of the heater will affect the plate current in the same manner that a voltage on the control grid does. This field can cause both 60 and 120 cycle variations of the plate current. The 60 cycle hum results from the difference in potential between the heater and the cathode. If a part of the unshielded heater is only slightly positive with respect to the cathode, it will attract some of the primary electrons which would otherwise form a part of the plate current. Since the voltage between the cathode and the heater alternates, this effect causes the plate current to vary in accordance with the alternating heater voltage. By careful shielding of the heater and by making sure that the heater is never positive with respect to the cathode, this type of interference may be easily eliminated. The same procedure effectively reduces the second-harmonic hum which results from the same source.

As will be explained later, the magnetic field set up around the heater wire may also influence the plate current. This effect is minimized by constructing the heater in the form of a double spiral, thereby causing its external magnetic field to be very low. Another method is to use a high-voltage, low-current heater, which would naturally have a smaller magnetic field. Another source of hum in cathode-type tubes is that due to leakage and capacitive coupling between the heater and the cathode. The heater voltage causes an alternating current to flow in the cathode circuit, through the grid-bias resistance, where it influences the grid voltage and, in turn, the plate current. The capacitive coupling is eliminated by shielding the heater wires and by careful placement of the lead-in wires through the glass base of the tube. The leakage between the heater and the cathode ordinarily results from a thin film of the "getter" material being deposited on the glass insulation between the lead wires. The logical means of preventing this occurrence is to arrange the getter cup so that none of the getter material will be deposited on the heater wires as they pass through the base of the tube. With reasonable care in construction, cathode-type tubes may be manufactured having a hum level too low to be objectionable.

There are several causes of AC hum in filament-type tubes, and either 60 or 120 cycles variations may be introduced into the normal plate current changes. The first of these causes is eliminated by returning the plate and grid circuits to a point which has the same potential as the mid-point of the filament. This, as explained in Lesson 16 of Unit 1, is accomplished either by returning both circuits to the center tap of the filament secondary winding or to the mid-point of a center-tapped resistor connected across the filament circuit. Such an action eliminates the fundamental frequency component of the hum voltage, but there still remains a residual hum due to the second-harmonic component. This hum may be due to the fact that the temperature of the filament is varying at a 120 cycle rate as the filament current increases and decreases. The obvious remedy is to make the filament large so that it will have a high thermal capacity, and will not cool appreciably during the time that the filament current is zero.

Another source of this double-frequency hum in filament-type tubes is produced by the varying magnetic field surrounding the filament. A magnetic field has no effect upon electrons unless they are in motion; however, the electrons are liberated from the emission surface with appreciable velocity and are thus influenced by the magnetic field. The effect of a magnetic field is to deflect the electrons in a direction at right angles to the magnetic lines of force. As an electron leaves the filament, it experiences two forces; the first is the electrostatic field between the plate and the filament which tends to draw it toward the plate; and the second is the magnetic field about the filament which tends to deflect it either upward or downward, depending on the direction of the field at that instant. The result is that the electron takes a long spiral path in reaching the plate, and the plate does not collect as many electrons in a unit of time as it would if the magnetic field were not present. Therefore, the effect of the magnetic field is to reduce the plate current. This tends to make the plate current slightly greater when the filament current is zero and has no magnetic field.

If all parts of the filament were at the same potential, the plate would attract an equal number of electrons from each part. This, however, is not the case, for when the filament current is maximum, the total voltage drop in the filament is equal to the peak of the applied AC filament voltage. At this instant, the plate draws more electrons from that part of the filament which is most negative, and fewer electrons from the end which is least negative. The increase in electrons which are drawn from the most negative end is greater than the decrease in electrons drawn from the least negative end, and therefore, the plate current is slightly larger when there is a voltage drop in the filament or when the filament current is maximum. This is due to the fact that the number of electrons drawn from any part of the filament is proportional to the square of the voltage between that part of the filament and the plate. This effect is minimized by arranging the filament in an inverted V or W construction. In this manner, the two ends of the filament are close together and there is an interchange of electrons between the most negative end and the least negative end, which tends to balance out this hum.

The hum produced by the magnetic field and that caused by the voltage drop in the filament are 180° out of phase and tend to cancel each other. That, due to the magnetic field, tends to make the plate current larger when the filament current is zero, whereas that, due to the voltage drop, tends to make the plate current larger when the filament current is maximum. By using a low filament voltage, and by careful consideration of the construction of the filament, these two effects may be made to cancel.

Except for very large water-cooled tubes which require an enormous amount of filament current, the filaments of nearly all vacuum tubes, both receiving and transmitting, employ commercial alternating current used in conjunction with step-down transformers. When the filament current demands are very large, there is a possibility of introducing hum in the carrier wave, and often it is better to use a DC generator driven by an AC motor. The output of the generator should be at least 10% greater than the needed filament voltage, and it may then be controlled by varying the amount of resistance in the field circuit of the generator, which is ordinarily of the shunt-wound type. Fig. 12 shows an arrangement in which a DC generator is used to provide filament voltage for several tubes.

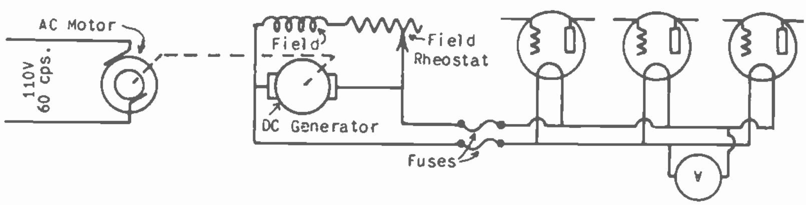


Fig.12 Using a DC generator to supply filament voltage for power tubes.

When a DC voltage is used on the filaments, the connections to the filaments should be reversed after every few hundred hours of operation. The negative leg of the filament will supply more of the emission current than the positive leg, and the wear on the filament will be unequal. By reversing the polarity of the filament voltage at regular intervals, the wear on the filament is equalized, and the tubes will have a longer life.

Maximum tube life demands that the filament voltage be maintained as near to the rated value as possible. For this reason, some means of adjusting the filament voltage should be provided. If the tubes receive their filament voltage from a step-down transformer, a tapped auto transformer may be inserted in the primary side of the transformer to vary the applied voltage.

As previously stated, the filament voltage should be applied gradually to the filaments of very large power tubes, in order to avoid a large current surge through the cold filaments. The initial filament voltage should not be more than half the rated value. After the filaments have warmed for a few seconds, the voltage may be increased to its rated value. It is only after the filaments have reached their normal operating temperature that the plate voltage may be safely applied. An arc-over is liable to occur if the plate voltage is applied before the filaments have had time to develop a protective space charge around themselves. It is well to develop the habit of allowing 30 seconds or so to pass after the filament voltage is applied before the plate voltage switch is closed.

9. THERMIONIC RECTIFIER TUBES. The sources available for supplying the high voltage needed for the plate supply include 110 volts DC supply lines, motor-generator sets, storage battery banks, copper-oxide rectifiers, and tube rectifier circuits operating from the alternating current supply line. DC supply lines are used only where this is the only voltage available; vibrators are employed in automobile and aircraft to a limited extent where only a low-voltage storage battery is available; high-voltage storage batteries are used on board ship and in a few radio stations; copper oxide rectifiers experience a very limited use for low-voltage tubes. Of all the sources of plate power, the high-voltage vacuum tube rectifier so far surpasses all the other methods that it is used wherever possible. Tube rectifiers may be divided into two general classes, depending on whether the tubes are of the high-vacuum (thermionic) variety or of the mercury-vapor type. This particular section of this lesson is devoted to the high-vacuum or thermionic type.

The rectifier tubes discussed in Lesson 16 of Unit 1 were of the thermionic type. The half-wave tube consists of an emitter, which in small tubes may be either a cathode or a filament, and an anode or plate which surrounds the emitter. The full-wave tube is nothing more than two half-wave tubes contained in the same glass envelope. The full-wave tube is not manufactured in the larger sizes, being confined mostly to rectifiers used for receivers.

The two-element vacuum tube is able to act as a rectifier, since current may pass only from the heated filament to the positive plate and not in the reverse direction. There are two important characteristics which determine the ability of any two-element tube to act as a rectifier. These are the peak plate current and the maximum peak inverse voltage (abbreviated M.P.I.V.). The peak plate current is the maximum current that may be drawn through the rectifier tube. It determines, but is not equal to, the maximum DC current that the power supply can furnish. The peak plate current will be greater than the current drawn from the supply, because a part of this current is used to charge the filter condensers. The peak plate current of any rectifier tube is determined by the emission capability of the filament or emitter. The maximum peak inverse voltage is the greatest voltage which may be safely applied between the plate and the filament, with the plate negative with respect to the emitter, without causing an arc-over between the two during the time that the tube is not rectifying.

The maximum peak inverse voltage rating determines how much alternating voltage may be applied between the plate and the filament, and, in turn, governs the DC voltage output of the power supply. The maximum peak inverse voltage will always be at least equal to the DC output voltage, and in some rectifier circuits, it may be as much as 3 times the DC voltage output.

The construction and the materials used for the component parts of a high-voltage, thermionic rectifier tube do not differ materially from the construction of an ordinary high-voltage power tube. The size of the filament used is dependent upon the peak plate current desired, and the spacing of the electrodes and the degree of vacuum obtained determine the maximum peak inverse voltage. The losses of the tube include the power used to heat the filament, and the average plate loss. Both of these power losses appear in the form of heat which must be dissipated through the glass walls of the tube. The very large thermionic rectifier tubes employ water-cooled plates as do large power tubes.

The voltage drop across the rectifier tube when it is conducting depends on the value of the current passing through the tube at that instant. It reaches a maximum of approximately 20% of the DC output voltage when the tube is passing its peak plate current. The fact that the voltage drop across the tube changes as the current drawn from the power supply varies tends to cause the power supply to have poor voltage regulation. It is this disadvantage which has led to the development of mercury-vapor rectifier tubes, which do not possess this disadvantage.

Thermionic rectifier tubes have been built with peak plate currents of 7.5 amperes, and able to withstand inverse voltages of 100,000 volts. However, except for very high direct-current voltages greater than those which can be conveniently obtained with mercury vapor tubes, and for low voltages such as used in receiver power supplies, the thermionic rectifier has been replaced in nearly every instance by the mercury-vapor tube. The fact that the thermionic tube is more rugged and requires less attention than the mercury-vapor type has caused its continued use in small power supplies.

Before beginning the discussion of the mercury-vapor tube, let us determine why the thermionic tube does have a variable voltage drop which causes poor voltage regulation. At low or medium plate voltages, the filament emits more electrons than the plate can attract. This causes the formation of a space charge about the filament which increases until a point of equilibrium is established. When this occurs, the plate current is independent of the filament emission and is determined only by the plate voltage. To cause an increase in the plate current, the plate voltage must increase; therefore, when more current is drawn from the power supply, the voltage drop across the tube increases. At very high plate voltages, the plate attracts all the electrons emitted by the filament, and the peak plate current is determined by the saturation point of the tube. Any further increase in the plate voltage would not cause an increase in the plate current because the full filament emission is already being used to form the plate current. The effect of the space charge is to neutralize the electrostatic field of the plate near the filament, and to reduce the amount of plate current that will flow with a given applied voltage.

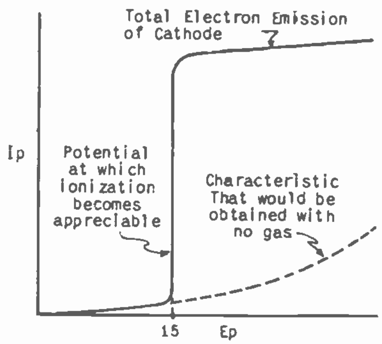
10. **THE MERCURY-VAPOR, RECTIFIER TUBE.** It has been stated that the poor voltage regulation of the thermionic-type rectifier tube is due to the formation of the negative space charge around the filament which limits the amount of plate current that can flow. If it were possible to neutralize this space charge in some manner, the plate would be able to draw the full filament emission at a very low plate voltage. This is accomplished in the mercury-vapor tube, as will now be explained. The mercury-vapor tube has a very large heavy filament, and a comparatively small plate situated a relatively great distance from the filament. After the tube has been thoroughly evacuated, a small amount of liquid mercury is introduced into the glass envelope. Due to the high degree of vacuum, a part of this mercury vaporizes and the gaseous mercury-vapor completely fills the tube. Of course, the pressure of the gas is still very low compared to normal atmospheric pressure; but the presence of the mercury-vapor makes the gas pressure within the tube considerably greater than that in an ordinary high-vacuum tube. Off hand, it would seem that the presence of these mercury-vapor molecules would retard or obstruct the flow of the electrons between the filament and the plate. However, as we shall now see, the mercury vapor partially neutralizes the space charge surrounding the filament, and makes the plate current practically independent of the plate voltage.

After they have been emitted by the filament, the electrons are attracted by the positive plate. In their flight, they collide with the molecules of mercury-vapor, and if they have attained sufficient velocity at the time of the collision, they will dislodge one or more of the planetary electrons of the mercury-vapor molecules. These dislodged electrons join the main electron stream and may, in turn, dislodge other electrons from mercury-vapor molecules. It would seem that the total number of electrons reaching the plate would be much greater than the number which leave the filament, due to this pyramiding process. This, however, is not true. The mercury-vapor molecules which have lost electrons are positively charged and are therefore positive ions. Being positive, these ions are attracted by the negative filament, but, having a much greater mass than the electrons, they move at a comparatively slow speed toward the filament. Some of the ions will be neutralized by attaching electrons to themselves from the space charge; others will strike the filament and gain their lost electrons directly from it. Thus, for each electron which arrives at the plate, there will be one electron leaving the filament. Many of the electrons reaching the plate are not the same ones which left the filament; they are electrons produced by the ionization of the mercury vapor. Likewise, many of the electrons leaving the filament never reach the plate, because they are used to neutralize the positive mercury-vapor ions. The ionization process will continue until there are as many ions being formed as there are ions being neutralized, or until a point of equilibrium is reached.

Perhaps the advantage of this process is not immediately apparent. It accrues from the fact that the positive mercury-vapor ions almost completely neutralize the negative space charge surrounding the filament which allows the plate to draw the entire filament emission when the plate is no more than 15 volts positive with respect to the filament.

Unless the difference in potential between the plate and the filament is at least 10.4 volts, there is no ionization. There is a certain minimum velocity which the electrons must acquire before they possess sufficient kinetic energy to dislodge electrons from the mercury-vapor molecules. It has been determined that this amount of kinetic energy is obtained by the electrons when they have fallen through a difference of potential of 10.4 volts. This number 10.4 is called the ionization potential of mercury; other gases have different ionization potentials.

Fig. 13 A plate voltage-plate current characteristic of a mercury-vapor tube.



If we were to plot a plate voltage-plate current characteristic of a mercury-vapor tube, we should find the curve similar to the one shown in Fig. 13. At very low plate voltages, the characteristic is the same as a thermionic tube. When the plate voltage reaches 10.4 volts, the characteristic departs from the normal curve for a thermionic tube, and the plate current is somewhat greater. There is not, as yet, much ionization; because the electrons almost reach the plate before they obtain sufficient energy to produce ionization. At a plate potential of 15 volts, the ionization becomes appreciable, and the plate current rises to a large value. Further increases in plate voltage cause but little increase in the plate current, and the characteristic is practically flat up to the maximum emission of the filament. It is at once apparent that the plate current is nearly independent of the plate voltage after the ionization becomes appreciable. In fact, the voltage drop across the tube remains constant at 15 volts from a very low value of plate current up to nearly the maximum that the tube may safely withstand. This is indicated by the vertical part of the characteristic curve shown in Fig. 13.

When the current drawn from a power supply using mercury-vapor tubes is low or moderate, the ionization of the tubes is small.

As the current drawn from the supply is increased, more plate current is drawn through the tubes, the ionization increases, and the tube drop remains constant. Thus, the effect of the increased plate current, which would cause an increased voltage drop across a thermionic rectifier tube, is neutralized by the production of more mercury-vapor ions.

Naturally, the constant voltage drop of the nercury-vapor tube, makes the voltage regulation of the power supply much improved over that obtainable with thermionic-type tubes. It is, perhaps, wondered why the thermionic tube is used at all, especially since the mercury-vapor tube provides so much better voltage regulation. As has been mentioned before, the peak inverse voltage which even the largest sized mercury-vapor tube can withstand is not as great as may be applied to some of the larger thermionic tubes. For this reason, power supplies designed for extremely high output voltages still employ thermionic tubes. For output voltages in excess of approximately 20,000 volts, mercury-vapor tubes may not be employed.

Since the voltage drop across a mercury-vapor tube is very low, it is permissible to use oxide-coated filaments in mercury-vapor tubes with a consequent saving in filament power. Furthermore, even the largest sizes of mercury-vapor tubes are air-cooled, because the low plate voltage, when conducting, prevents the development of excessive heat.

In receiver power supplies, the thermionic tube is preferred to the mercury-vapor type, because of its greater ruggedness. The thermionic tube is able to withstand abuses and overloads which would completely ruin a mercury-vapor tube. Let us determine why this is so. When the peak plate current of a thermionic tube is exceeded, due perhaps, to a short in the power-supply filter, the plate of the tube becomes red hot. This is caused primarily by the large voltage drop which is produced across the tube. The electrons attain enormous velocities, and strike the plate with such force that the energy, which is converted into heat, raises the temperature of the plate to incandescence. Of course, if the short is allowed to remain for any length of time, the filament of the tube will burn in two, or the power transformer will be ruined due to the excessive current. A momentary short, however, will produce no permanent damage, and the tube will again function when the short is removed. Also, if the peak inverse voltage is exceeded and an arc is formed; it is possible that the tube will again be serviceable, provided that the excess voltage is removed before the arc has done much damage. This is not true of a mercury-vapor tube; even a slight overload of current or voltage will render the tube permanently valueless.

In the previous description of the operation of a mercury-vapor tube, it was learned that the positive ions which were formed drifted toward the filament. When the drop across the tube is normal (15 volts), the ions are almost completely neutralized by the space charge or else strike the filament so lightly as to cause no damage. It has been determined that the bombardment of the filament by the positive ions will cause no damage, provided that the drop across the tube is less than 22 volts. With a larger tube drop, the ions strike the filament with destructive force, literally stripping the emission surface from it. As the peak plate current is exceeded, the voltage drop across the tube rises rapidly, and since the destructive action of the filament bombardment is practically instantaneous, even a momentary overload is ruinous to the tube.

The maximum peak inverse voltage which a mercury-vapor tube can withstand is somewhat less than a thermionic tube of the same electrode spacing. This is due primarily to the presence of the mercury vapor. During the inverse part of the cycle, when the plate is negative with respect to the filament, there is no attraction for the electrons emitted by the filament, and, as a result, there is no ionization of the mercury-vapor molecules. If, however, the maximum peak inverse voltage is exceeded, the few free electrons in the space between the filament and plate are attracted so strongly by the high voltage toward the positive filament that they do produce some ionization. The positive ions formed are attracted by the negative plate and attempt to gain their needed electrons from it. The first ones arriving at the plate crowd around it, and tend to repel those at their rear. Some of the ions which arrive subsequently shoot around the plate and strike the glass walls of the tube with considerable force, often puncturing the glass and ruining the tube. The greater the pressure of the mercury vapor, the more mercury-vapor molecules and free electrons there will be in the space between the filament and the plate, and the easier it will be for this destructive action to occur. The reason that the thermionic tube has such a high maximum peak inverse voltage is due to the fact that the gas pressure in such a tube is exceedingly low, and very high voltages are required to accelerate the few free electrons to the point where they will ionize some of the residual gas molecules, thereby causing an arc-over. In addition, it is not possible to cause much ionization because of the relative scarcity of gas molecules. On the other hand, the pressure of the vapor in a mercury-vapor tube is at least 100,000 times as great as in a thermionic tube, and much smaller inverse voltages will cause arc-over. This phenomenon is often called "flash-back".

A further precaution which must be observed in the operation of mercury-vapor tubes is to allow the filament to reach full operating temperature before the alternating voltage is applied to the plate. Should the filament and plate voltage be applied simultaneously, before the protective space charge has had time to develop around the filament, the positive ions formed would bombard the filament with such destructive force that it is improbable that the filament would be able to furnish an appreciable emission thereafter. For this reason, great care must be taken to allow the filament temperature to rise to its normal value before applying the plate voltage. The amount of time which should elapse before the plate-voltage switch is closed is dependent on the size of the tube. For the smallest tube (the type 83), the delay need be only 1 or 2 seconds; in fact, little damage is done to this particular tube if the filament and plate voltages are applied simultaneously, because the filament reaches its operating temperature almost immediately. The type 371 should have a delay period of at least 10 seconds;

types 866 and 872, a delay of 30 seconds; and the type 857, which is the largest tube manufactured, and has a cathode in addition to a heater, should have its plate voltage delayed at least 10 minutes after the filament voltage is applied. Usually this delay period is automatically taken care of by time-delay relays which do not close the plate circuit of the mercury-vapor tubes until a definite length of time has elapsed after the filament voltage has been applied. Should you ever be in doubt as to whether such a relay is incorporated in a transmitter, use due caution in applying the plate voltage to the rectifier tubes. All transmitters not using time-delay relays will have separate filament and plate switches, and make sure that the filament switch is the first one thrown.

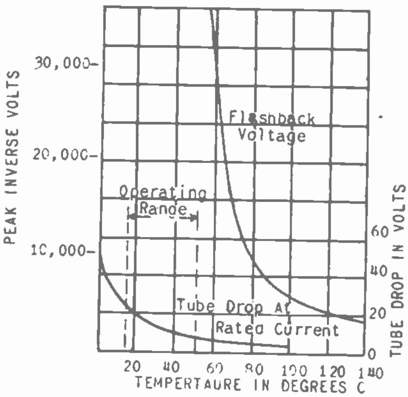


Fig. 14 Illustrating the relation between temperature and tube drop and temperature and peak inverse voltage of a mercury-vapor tube.

The pressure of the vapor in a mercury-vapor tube determines to a large extent just how well the tube will operate. As has been stated, when the pressure is too high, the maximum allowable peak inverse voltage is greatly reduced and there is danger of flashback occurring. On the other hand, when the pressure is too low, the ionization will be incomplete, the voltage drop across the tube will exceed the critical value of 22 volts, and the bombardment produced will cause stripping of the filament. The pressure of the mercury vapor is largely dependent on its temperature. Thus, mercury-vapor tubes have a critical range of operating temperatures, and if the temperature is allowed to become either too low or too high, damage will result to the tube. Fig. 14 shows how the voltage drop across the tube and the maximum peak inverse voltage are affected by temperature. The temperature in this case is the air temperature measured a few inches from the base of the tube just outside of the glass envelope; it is sometimes called the "ambient" temperature. Notice that the normal operating range is from 15° C to approximately 50° C. From the manufacturer's specifications, it will be found that the maximum peak inverse voltage of the type 857 is 10,000 volts, when the operating temperature is maintained within the range of 15° to 50° C.; but that the maximum peak inverse voltage may be increased to 22,000 volts, provided that the temperature range is limited from 30° to 40° C. To keep the temperature within this range requires the use of blower fans.

Mercury-vapor tubes should be shielded or placed so that they will not be influenced by R.F. fields. The effect of an R.F. field is to maintain the mercury vapor in a partial state of ionization during the inverse part of the cycle, thereby causing the operation of the tube to be quite erratic, and possibly damaging the tube due to flash-back. The distinctive features of a mercury-vapor tube are the low filament voltage (never over 5 volts), the fact that oxide-coated filaments are used, and the relatively small plate. Since the voltage between the plate and any part of the filament must not exceed the critical value of 22 volts, the filament voltage is never greater than 5 volts.

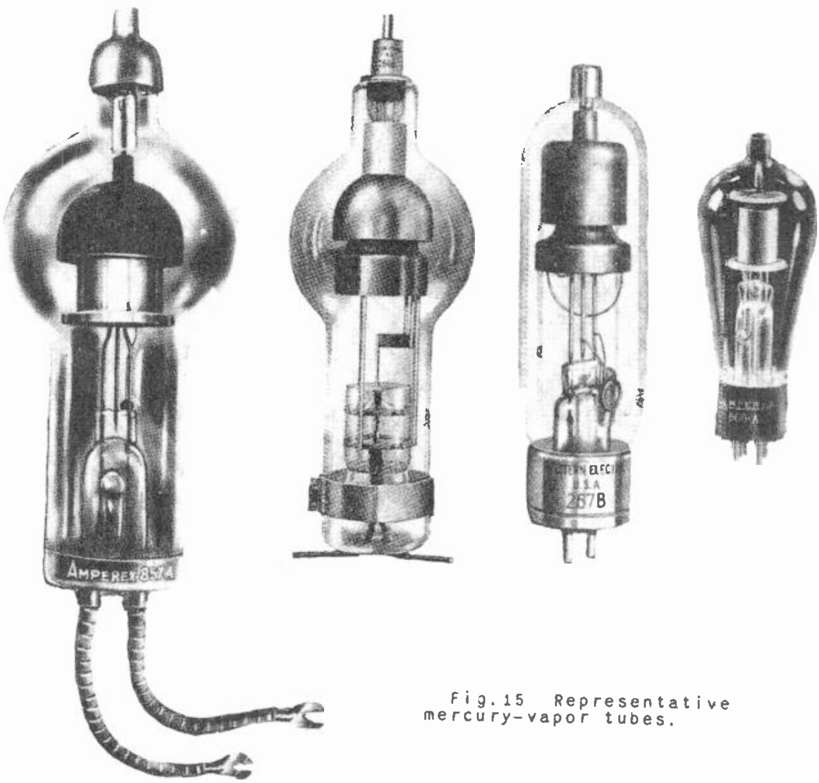


Fig. 15 Representative mercury-vapor tubes.

Care should be taken to make sure that the filament voltage of a mercury-vapor tube is of the correct value. Much of the damage to these tubes results from improper filament voltages. If the voltage applied to the filament is too low, the ionization will not be very large and the space charge around the filament will not be sufficient to protect it from the ion bombardment caused by the excessive tube drop. A filament voltage above the rated value will cause needless evaporation of the filament material, and will tend to heat the tube, raising the pressure of the mercury-vapor and reducing the maximum peak inverse voltage.

In removing a mercury-vapor tube from its socket always keep the tube in an upright position. If the tube is tilted, the excess mercury will run down the walls of the tube and splash over on the plate and filament electrodes. Then when the plate voltage is applied, there is considerable danger of arc-over. When installing a new tube, always let the filament heat for at least three-quarters of an hour, before applying the plate voltage, to make sure that the heat drives the excess mercury into the bottom of the tube.

Several different types of mercury-vapor tubes are illustrated in Fig. 15. The smaller types have ribbon-like filaments and small disc plates. The larger sizes employ plates in the form of a cup which fits down over the cathode assembly. The cathode assembly includes a heater, and vanes or discs coated with the emitting oxide. The whole is surrounded by one or more shields polished to prevent the radiation of heat. The cathode assembly is illustrated in Fig. 16.

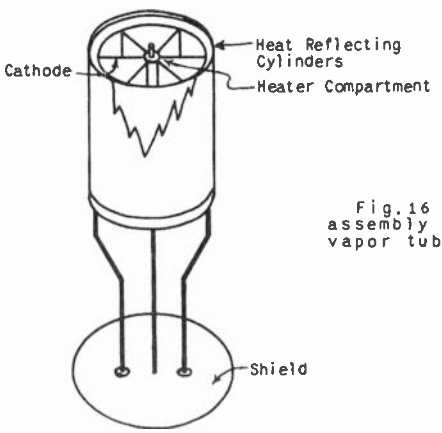


Fig. 16 Showing the cathode assembly of a large mercury-vapor tube.

In the comparison of mercury-vapor rectifier tubes with those of the thermionic type, we find that the mercury-vapor tube has the higher efficiency, better regulation, low filament power, and low first cost. Among its disadvantages may be listed, a limited inverse-voltage rating, a tendency to flash-back, subject to the effect of R.F. fields, and the possibility of damage to the emitter as a result of momentary overloads. In conclusion, it is found that the mercury-vapor tube is the most desirable for transmitters, and the thermionic tube, the better for receivers.

11. RECTIFIER CIRCUITS. Rectifier circuits may be divided into two general classes; those for single-phase power, and those intended to be used with three-phase systems. We shall first consider those to be used with single-phase power. The simplest of all rectifier circuits is, of course, the half-wave circuit. Illustrated at A in Fig. 17, it is seen to be exactly the same circuit which was discussed in Lesson 16 of Unit 1. Its operation is

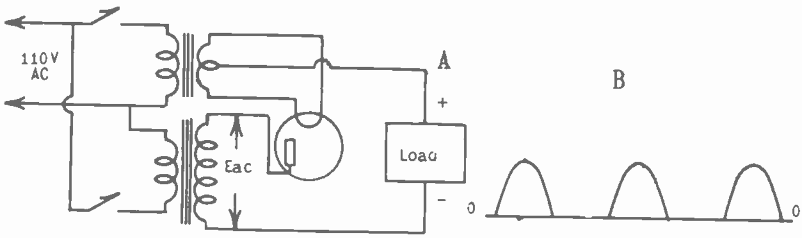


Fig.17 A simple half-wave rectifier circuit and the wave form of its output voltage.

very easy to understand, and is doubtlessly clear to every student. The high-voltage winding of the transformer is in series with the tube and the load. Current may flow through the load circuit only when the plate of the rectifier is positive with respect to its filament. Naturally, a filter system (not shown) must be included to eliminate the AC component of the output of the rectifier. Notice that the filament voltage is supplied from a separate transformer; this is nearly always true in transmitter circuits. The plate transformer, rectifier filament transformer, and other filament transformers are ordinarily separate pieces of equipment. Since the half-wave rectifier is but rarely used in transmitters, it will not be discussed further. Its output voltage is shown at B in the figure.

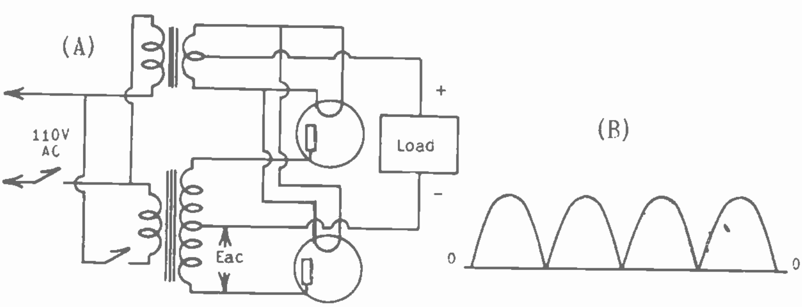


Fig.18 A full-wave rectifier circuit and the wave form of its output voltage.

The next type of rectifier circuit to be discussed is the full-wave type, which employs two tubes, as shown at A in Fig. 18. This is also exactly the same circuit as was given in Lesson 16, Unit 1. Two tubes, however, are nearly always used, since the full-wave tube is manufactured in only the smaller sizes. The output voltage which this circuit supplies to the input of the filter system is shown at B in this figure. Since current flows nearly all the time from the output of this rectifier, the output current has a larger DC component and filtering is much easier than in the half-wave circuit. The ripple voltage in this case has a frequency of 120 cycles, compared to the 60 cycle ripple of the half-wave rectifier.

The maximum peak inverse voltage applied to the tube in the half-wave rectifier circuit is the peak AC voltage developed across the high-voltage winding, plus the voltage developed across the load. In the full-wave circuit employing two tubes, the peak inverse voltage across either tube is equal to the peak voltage built up across the entire high-voltage winding and not just across that half which supplies this tube.

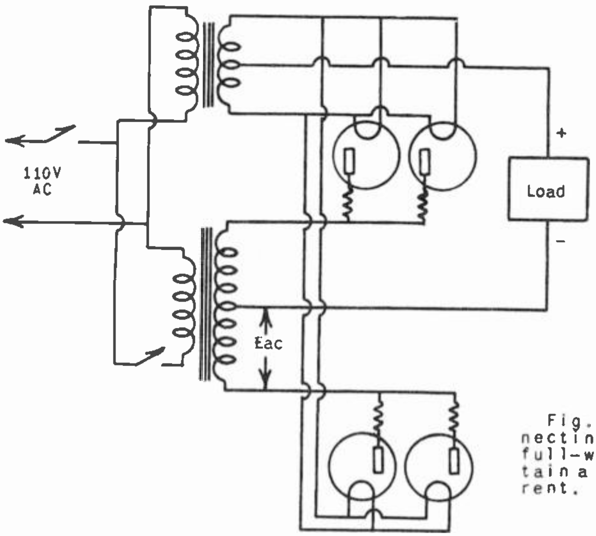


Fig. 19 Method of connecting four tubes in a full-wave circuit to obtain a larger output current.

When two half-wave tubes connected in a full-wave circuit are unable to supply sufficient current, four such tubes may be used as illustrated in Fig. 19. This circuit consists of two sets of parallel-connected tubes, and the current drawn through each tube is one-half of that furnished to the filter system. The filament transformer must be capable of supplying twice as much current as in the preceding circuit. Unless the resistance of each tube of a parallel-connected pair is the same, the current drawn through each will not divide equally, and one tube will carry the majority of the current. To prevent this occurrence, small equalizing resistors are connected in series with each plate lead to provide stability and an equal division of the current. Stability will be obtained if the drop across these resistors is approximately 6 volts. Thus, if the power supply is to furnish 240 ma., half this, or 120 ma. will flow through each tube. To provide a voltage drop of 6 volts across the equalizing resistors, they must each have a value of:

$$R = \frac{E}{I} = \frac{6}{.120} = 50 \text{ ohms.}$$

The greatest disadvantage of the two foregoing full-wave rectifier circuits is the fact that only one-half of the high-voltage winding is effective in producing a DC output voltage at any one instant. The DC output voltage will be the average value of the voltage applied to the input of the filter (assuming a choke-input filter), less the drop in the filter chokes. Neglecting the voltage drop in the tube and the transformer winding, the peak voltage applied to the input of the filter is equal to the peak voltage developed across one-half of the high-voltage secondary. To supply 300 volts DC output voltage requires that the total voltage developed across the high-voltage winding of the transformer in Fig. 18 be between 600 and 700 volts. Furthermore, the amount of voltage that may be applied to the plates of the rectifier tubes is limited by the fact that the peak inverse voltage is equal to the peak voltage set up across the entire secondary winding.

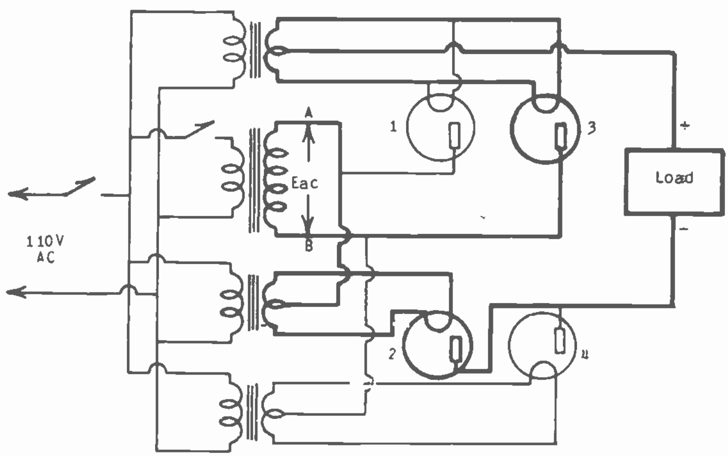


Fig.20 A full-wave bridge circuit employing four tubes.

A method whereby a much larger DC output voltage may be obtained, without using larger sized tubes or a transformer developing a greater voltage, is illustrated in Fig. 20. This is called a bridge circuit and it employs four half-wave tubes. Notice that it eliminates the necessity of tapping the high-voltage winding with the result that the total voltage across this winding is effective in producing rectification. Let us trace the path of the current flow when point A is negative with respect to point B. Current flows from point A to the filament of tube 2. Tube 1 cannot pass current because its plate is negative with respect to its filament. Tube 2, however, is able to pass current, since its filament is connected to point A, which is negative. Therefore, the current flows through tube 2, through the load, and back to the filaments of tubes 1 and 3, which are connected in parallel. Tube 3 may pass current, because its plate is connected to point B, which is positive; therefore, current flows through tube 3 to point B, and has thus completed its circuit. The path taken by the current during this alternation is illustrated by the heavy lines in Fig. 20.

During the next alternation, point B is negative with respect to point A. Tube 2 cannot pass current since its filament is connected to point A, which is positive, and tube 3 is likewise unable to pass current because its plate is connected to point B, which is negative. The polarity of the voltages across tubes 1 and 4, however, is such that both these tubes are able to conduct. Therefore, current flows from point B through tube 4, through the load, and through tube 1 to point A. At any instant, there are two tubes passing current. The two tubes which are conducting are in series, and although the total current which may be drawn from this circuit is no greater than the simpler full-wave circuit of Fig. 18, the output voltage delivered to the filter system is twice that obtained with the center-tapped transformer circuit. The peak in-

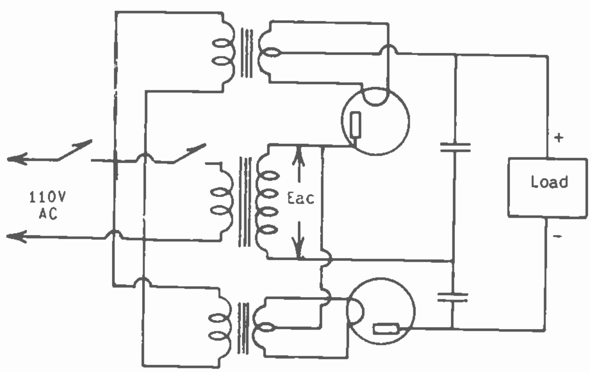


Fig. 21 A voltage doubler circuit.

verse voltage across any tube when it is not conducting is equal to the peak voltage developed across the high-voltage winding. With given size tubes and a given transformer, the bridge circuit provides a method of securing twice as much output voltage as the center-tapped system without increasing the peak inverse voltage across any tube.

The disadvantages of the bridge rectifier are: First, the fact that four tubes are required instead of two; second, the fact that three separate filament windings for the rectifier tubes must be available. From Fig. 20, it is seen that the filaments of tubes 2 and 4 are connected to opposite ends of the high-voltage winding; therefore, one filament winding could not be used to supply filament voltage to both of these tubes without shorting out the high-voltage secondary. Furthermore, the filaments of tubes 1 and 3, although connected together, are at a different potential than the filaments of either tubes 2 or 4; and so three separate filament secondaries must be used.

One other type of single-phase rectifier circuit that is occasionally used is the voltage doubler illustrated in Fig. 21. The operation of this circuit was explained in Lesson 16, Unit 1 and will not be repeated; however, it is to be observed that this circuit uses two tubes instead of the ordinary voltage doubler tube (25Z5, etc.). Also, a plate transformer and two separate filament windings are employed in this circuit. The DC output voltage is approximately twice the R.M.S. value of the alternating voltage developed across the high-voltage winding.

12. **THREE-PHASE RECTIFIER CIRCUITS.** When three-phase power is available, it is nearly always used for rectifiers having an output voltage in excess of 1,000 volts. It is much more efficient, and the output from such a circuit requires far less filtering to remove the AC component. As explained in the foregoing lesson, there are two general methods of connecting three-phase circuits;

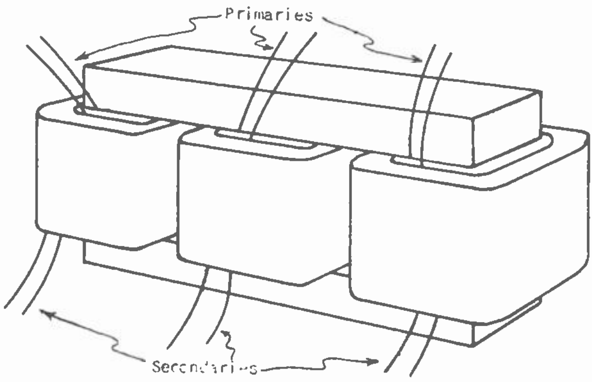


Fig. 22 A three-phase transformer

they are the Y and the delta connections. Only the plates of the rectifier tubes receive three-phase power; the filaments, since they are used only to supply emission, operate just as well from a single-phase line. Therefore, the filament transformers will be single-phase and will be connected to one phase of the three-phase system. The plate transformer is ordinarily a specially designed three-phase type, although three single-phase transformers may be used with a somewhat lowered efficiency.

The three-phase transformer has six separate windings (three primaries and three secondaries) wound on three legs of a common iron core as shown in Fig. 22. With this arrangement, the primaries may be Y-connected, and the secondaries delta-connected, or various other combinations may be used. The wave-forms of the three voltages applied to the primaries are shown in Fig. 23. If

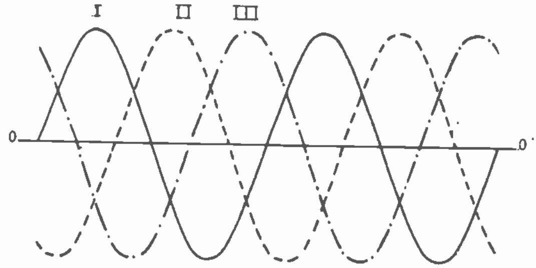


Fig. 23 Wave form of the three voltages applied to the primaries of a three-phase transformer.

the transformer is used in a three-phase half-wave rectifier, only the positive alternations are effective in producing an output voltage, and the wave form of this voltage is as illustrated in Fig. 24. The light lines in this diagram indicate the voltage pulses from each of three tubes, whereas the heavy line joining the tops of the pulses is the total output voltage devliered to the filter circuit. Notice that the output voltage never falls to zero, and is far less pulsating than the output of any single-phase rectifier. Furthermore, the frequency of this pulsating voltage is three times the fundamental supply frequency, or 180 cycles. The higher the pulsating frequency, the easier it is for the filter system to remove the AC component, leaving only the DC component.

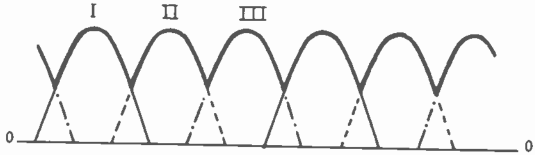


Fig. 24 The output of a three-phase rectifier.

Now that we have seen the advantages of using a three-phase rectifier, let us continue with the discussion of the practical operation of a circuit of this type. There are four basic methods of connecting the primaries and secondaries of the plate transformer. These are shown in Fig. 25. At A, both primaries and secondaries are delta connected; at B, both are Y-connected. At C in this figure, the primaries are Y-connected and the secondaries, delta connected; whereas at D, the primaries have the delta connection and the secondaries the Y connection. Let us, for argument, assume that the primaries and secondaries have the same number of turns, and that the voltage between any two of the three supply lines is 100 volts. Considering A of Fig. 25, we see that each primary would have 100 volts applied to it, there would be developed 100 volts across each secondary, and the voltage between any two of the wires leading from the secondaries would be 100 volts.

In the circuit shown at B, the output voltage between any two leads would also be 100 volts, although the actual voltages across the separate primaries and secondaries would be less than this value, due to the type of connection. In the preceding lesson, it was stated that the voltage between any two line wires in the Y connec-

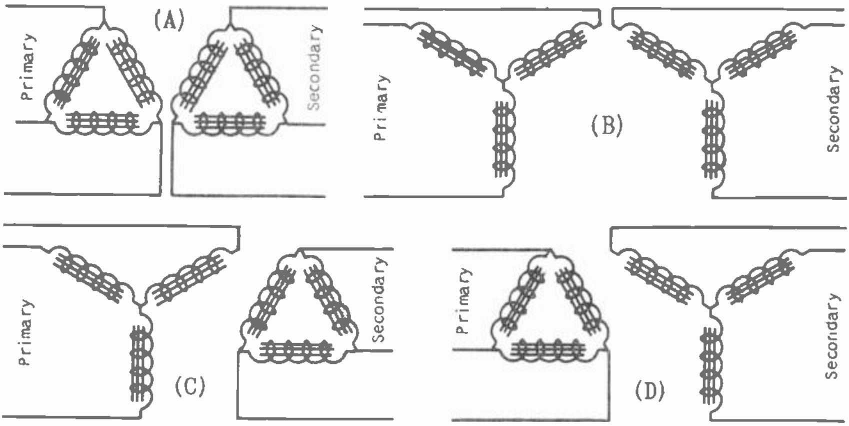


Fig.25 Four methods of connecting a three-phase transformer.

tion was equal to 1.73 times the voltage across any phase coil. Therefore, the voltage across each primary and each secondary would be $\frac{1}{1.73}$ of 100, or 57.8 volts. In a like manner, the voltage across each primary of C, Fig. 25 would be 57.8 volts, and since the secondaries have the same number of turns as the primaries, the voltage across each secondary would likewise be 57.8 volts. However, since the secondaries are delta-connected, the voltage between any two of the secondary lead wires would be only 57.8 volts, and it is seen that the voltage has been stepped down because of the type of connection.

Finally, the voltage across each primary of D in this figure would be 100 volts, and the voltage across each secondary would also be 100 volts. Since the secondaries are Y-connected, the voltage between any two lead wires of the secondary would be 1.73 times 100 or 173 volts. Thus, the voltage has been stepped up with this type of connection.

As it is desirable to have a large secondary voltage. the connection shown at D (known as the delta-Y connection) is nearly always used. With this method, the secondary voltage is equal to 1.73 times the primary voltage times the turn ratio between the secondaries and the primaries. Drawing the schematic diagram in this manner helps the student to visualize the increase in the secondary voltage produced by the type of connection; however, the diagram shown in Fig. 26, which is equivalent to D of Fig. 25, is of more aid in making the connections to the three-phase transformer.

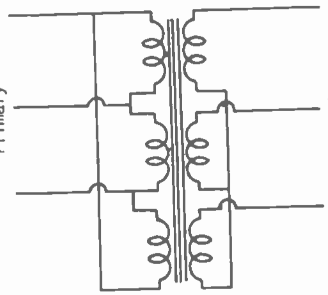


Fig.26 Showing how the transformer is wired for the Delta-Y connection.

A three-phase transformer should always be used in preference to three single-phase transformers, since there is less tendency for core saturation. The current flowing through the secondaries is pulsating DC, and it is easy to saturate the cores if three separate transformers are used. In the three-phase transformer, the secondaries are so wound that the DC magnetization produced is cancelled out.

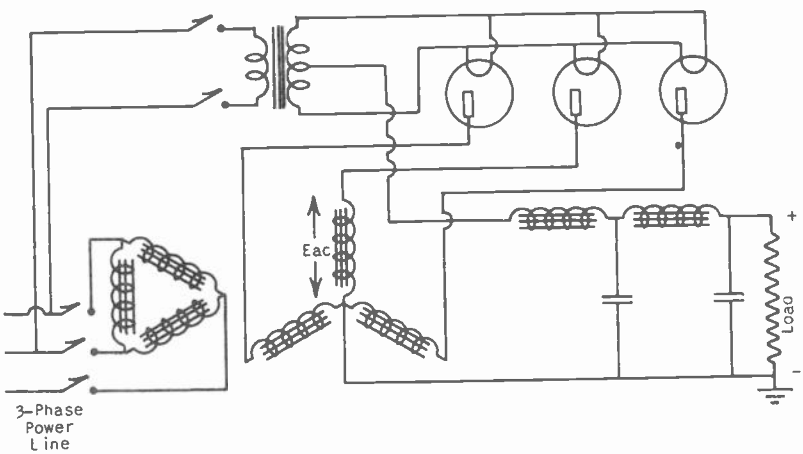


Fig.27 A three-phase, half-wave rectifier circuit.

The three popular types of three-phase rectifier circuits are the half-wave, the half-wave double-Y, and the full-wave. The half-wave circuit is illustrated in Fig. 27. It requires three tubes, each plate of which is connected to one leg of the three-phase secondary. The filaments of the three tubes are connected in parallel and filament voltage is supplied by one filament transformer. The positive terminal of the circuit is the center tap of this filament winding, and the negative terminal is the center of the

Y-connected secondary. The three alternating voltages which are applied to the plates of the tubes are 120° out of phase and have the wave forms which were shown in Fig. 23. Each tube draws current for one-third of the time, however, these periods overlap so that at some instants there are two tubes passing current. The output voltage that results from this circuit is that illustrated in Fig. 24. Since the conduction periods of the tubes overlap, the output voltage never falls to zero and the frequency of the pulsations is 180 cps. This circuit is essentially nothing more than three simple half-wave rectifiers with each tube supplied with an alternating voltage differing in phase by 120° from the voltages furnished to the other two tubes.

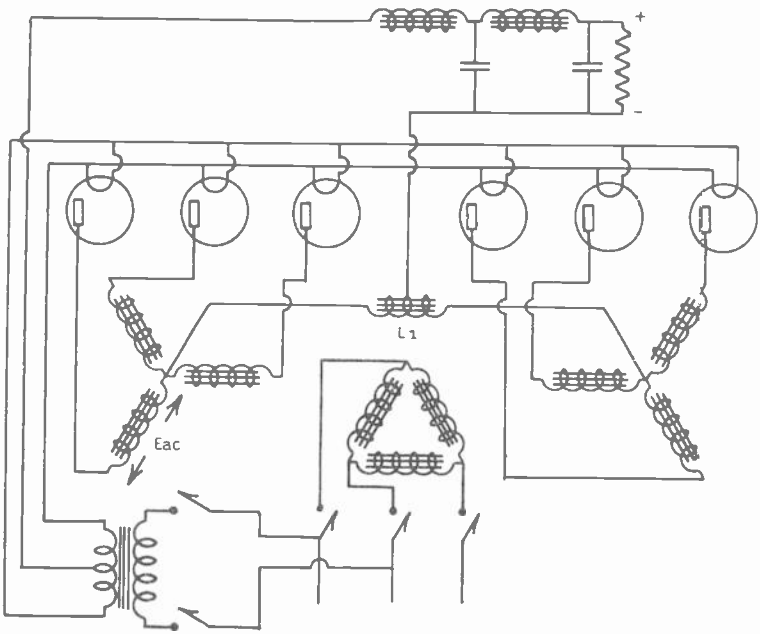


Fig. 28 A three-phase, half-wave, double-Y rectifier circuit.

The half-wave double-Y rectifier circuit is shown in Fig. 28. It is practically the same as two three-phase, half-wave rectifier circuits connected in parallel. It employs six tubes and the voltages developed across one Y-connected secondary are 180° out of phase with those present across the other secondary. Thus, the output of one three-phase unit is maximum at the time that the output of the other unit is minimum. The centers of the two secondaries are connected together by an interphase reactor or balance coil (L_1) which allows each unit to work independently of the other. Each tube supplies current for one-third of the time, but since the two units are 180° out of phase, the output voltage is that shown in Fig. 29. The light lines show the output voltage of each separate tube, whereas the heavy line joining the tops of the waves represents the total output to the filter. From this figure, it is seen

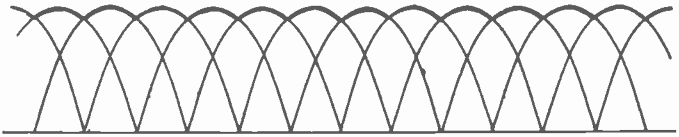
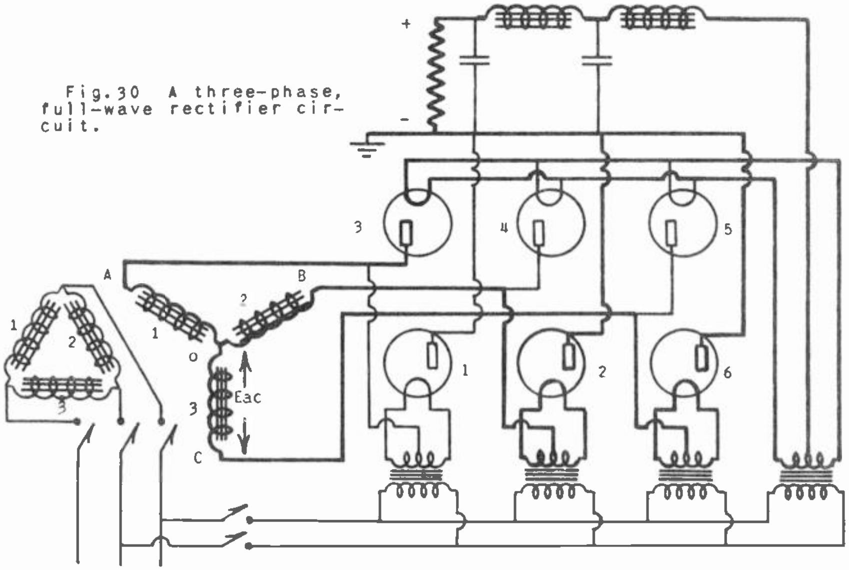


Fig. 29 The output voltage of a three-phase, half-wave, double-Y rectifier. The solid line at the top is the voltage output.

that the output voltage is less pulsating than in the case of the ordinary half-wave circuit, and that the frequency of the pulsations is six times the fundamental frequency or 360 cycles. Naturally, such an output voltage is easier to filter. This circuit will supply twice as much DC load current as will the half wave circuit without exceeding the maximum peak current of any tube. The DC output voltage is the same as the half-wave circuit.

Fig. 30 A three-phase, full-wave rectifier circuit.



Since the output current flows through the two halves of the balance coil in opposite directions, there is no tendency for the core of this coil to saturate. This balance coil acts as the first filter choke.

The last type of rectifier circuit which we shall consider is the three-phase, full-wave shown in Fig. 30. Like the half-wave, double-Y, it uses six tubes, but has only one Y-connected secondary. Actually, it is a modification of the bridge type of circuit adapted for three-phase power. It is called a full-wave circuit, since it takes advantage of both the positive and negative alternations of the three-phase voltage. Let us consider its action in detail. In Fig. 31 are reproduced the wave forms of the three alternating voltages which are developed across the three legs of the secondary windings. These three voltages are labeled 1, 2, and 3 to correspond with the numbers marked beside the legs of the secondary. We shall consider the instant designated by the vertical line drawn through these wave forms.

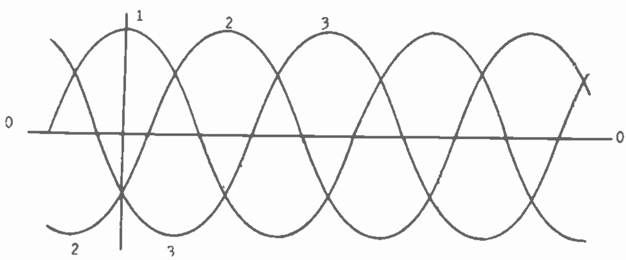


Fig. 31 The wave form of the three voltages appearing across the secondaries of the three-phase transformer.

At this time, the voltage of phase 1 is maximum in a positive direction; that of phase 2 is negative and is decreasing to zero; whereas that of phase 3 is negative, equal to the voltage of phase 2, and is increasing to its maximum value. This makes point A positive with respect to point O; and points B and C negative with respect to point O. Points B and C are connected respectively to the filaments of tubes 2 and 6, and these tubes are therefore able to pass current, since their filaments are negative. Point A, on the other hand, is connected to the plate of tube 3, making this plate positive with respect to its filament, and enabling this tube to conduct. At this instant, therefore, current will flow from the negative terminal of the load, through the load to the filament of tube 3, through this tube, through coil 1 from A to O, at which point the current divides. From point O, the current flows through coil 2, from O to B, to the filament of tube 2, through this tube, back to the negative terminal of the load. Also, current flows from point O, through coil 3, from O to C, to the filament of tube 6, and through this tube to the negative terminal of the load. (The heavier lines indicate the path of current at this time.) Thus, there are three tubes passing current at this instant, and there will always be three tubes passing current except for the instant when the voltage of one phase is zero.

It is evident that current flows through all three legs of the secondary, even when the voltage across a leg is negative. Therefore, this circuit takes advantage of both the positive and negative alternations, and is a full-wave circuit. The output voltage has the wave form illustrated in Fig. 32. The three wave forms are numbered to correspond with the three alternating voltages of Fig. 31. The output voltage corresponding to the instant previously discussed is shown by the vertical line drawn through these waves. Notice that at this instant the output voltage is maximum; and that it falls to a minimum value when the voltage of one phase drops to zero, and there are but two tubes conducting.

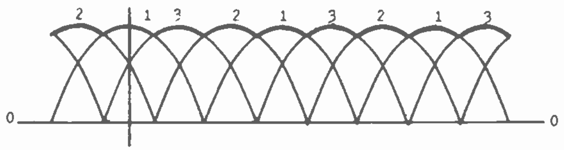


Fig.32 Showing the output voltage of the three-phase, full-wave rectifier.

The actual output voltage is the heavy line joining the tops of the waves illustrated in Fig. 32. It has a pulsation frequency of 360 cycles (assuming a line frequency of 60 cycles), and is comparatively easy to filter. It is the same output wave as that produced by the half-wave, double-Y circuit. The maximum DC current which may be drawn from a full-wave circuit is not as large as that available from a half-wave, double-Y, because the tubes in a full-wave circuit are not connected in parallel. However, the maximum DC output voltage is nearly twice that of either the half-wave or the half-wave, double-Y. This is due to the fact that full-wave rectification is used and the combined voltages across two legs of the secondary is available for forcing current through the load.

The outstanding disadvantage of the full-wave circuit is the fact that four separate filament windings must be provided, one of which must be able to supply three tubes, and the others one tube each.

13. DESIGNING THE POWER SUPPLY. In designing a power supply, the two things which must be known are the DC voltage required and the maximum DC current to be drawn from the supply. The next step is to decide which circuit is to be used and what type of tubes are to be employed. This decision will, of course, be influenced by the DC voltage desired and the DC current which will be needed. When the circuit and tubes have been selected, we must determine how much AC voltage will have to be applied to the plates of the tubes to produce the required DC output voltage, and whether this voltage will cause the maximum peak inverse voltage to be exceeded. Also, we must learn whether the tubes selected are capable of supplying the DC current without exceeding the maximum peak plate current. To make the problem as simple as possible, the following table has been included:

The first column of this table gives the type of circuit and whether it is single-phase or three-phase. The second column shows the figure number of that circuit in this lesson. The third column tells the maximum voltage which may be developed across the high-voltage winding without exceeding the maximum peak inverse voltage.

Since this voltage is measured in different ways for different circuits, it is well to refer to the figures which indicate between what two points this voltage is measured. The fourth column indicates the maximum DC voltage applied to the input of the filter system, and the last column tells how much current may be drawn from the power supply without exceeding the peak plate current of the tube.

	TYPE OF CIRCUIT	FIG. NO.	RMS VOLTAGE ACROSS TRANS-FORMER. (E_{ac}) SEE FIGS.	MAXIMUM DC OUTPUT VOLTAGE (E_{dc})	MAXIMUM DC CURRENT I_{dc}
Single-Phase	Half-wave	17	.353 × MPIV.	.45 × E_{ac}	.318 × I_m
Single-Phase	Full-wave	18	.353 × MPIV.	.9 × E_{ac}	.636 × I_m
Single-Phase	Full-wave parallel	19	.353 × MPIV.	.9 × E_{ac}	1.33 × I_m
Single-Phase	Full-wave bridge	20	.707 × MPIV.	.9 × E_{ac}	.636 × I_m
3-Phase	Half-wave	27	.408 × MPIV.	1.17 × E_{ac}	.827 × I_m
3-Phase	Half-wave Double - Y	28	.408 × MPIV.	1.17 × E_{ac}	1.91 × I_m
3-Phase	Full-wave	30	.408 × MPIV.	2.34 × E_{ac}	.955 × I_m

Conditions Assumed: 1. Sine-wave supply. 2. Balanced Phase voltages. 3. Zero tube drop. 4. Pure resistance load. 5. No filter used. MPIV is maximum peak inverse voltage. I_m is peak plate current.

For example, let us consider the type 872-A mercury-vapor tube. A table giving the maximum peak inverse voltages and peak plate currents of the more common of these tubes will be found at the end of this lesson. From this table, it is seen that the maximum peak inverse voltagé is 10,000 volts, and the peak plate current is 2.5 amperes. We will assume that tubes of this type are to be used in a three-phase, half-wave rectifier. Let us now calculate what voltage may be applied to the tubes, what DC voltage will be produced, and what current may be drawn from the circuit. From the table, it is seen that the R.M.S., AC voltage which may be developed across each leg of the three-phase secondary is .408 times the maximum peak inverse voltage. This is:

$$E_{ac} = .408 \times 10{,}000$$
$$= 4{,}080 \text{ volts}$$

With this AC voltage applied, the maximum DC voltage which will be applied to the filter is 1.17 × E_{ac} or:

$$E_{dc} = 1.17 \times 4{,}080$$
$$= 4{,}774 \text{ volts.}$$

The maximum DC current which may be drawn from the supply is .827 × I_m or;

$$I_{dc} = .827 \times 2.5$$
$$= 2.07 \text{ amperes.}$$

If we wished to design a power supply capable of furnishing 4,000 volts DC at 1.5 amperes, we would proceed as follows. To allow for the voltage drops in the filter, the tubes and the transformer windings, we shall add 10% to the required output voltage, making it 4,400 volts. Using the figures for the three-phase half-wave circuit, we see that the DC output voltage is 1.17 times the AC input voltage; therefore, the AC input voltage must be $\frac{1}{1.17}$ times the DC output voltage, and for an output voltage of 4,400 volts this is:

$$\frac{1}{1.17} \times 4,400 = 3,761 \text{ volts}$$

This is the R.M.S. value of the AC voltage which should be developed across each leg of the secondary winding. Since this circuit is capable of furnishing an output voltage of 4,774 volts without exceeding the maximum peak inverse voltage, we know that we are not attempting to derive too much voltage from this type of tube. Furthermore, the maximum DC current which this tube and circuit are able to furnish is 2.07 amperes, and since the maximum current which we shall draw is only 1.5 amperes, we are again assured that we are not abusing any of the tubes in the circuit.

14. **THE CHOKE-INPUT FILTER CIRCUIT.** As is well known, the output voltage of a rectifier circuit contains a DC and an AC component. It is the purpose of the filter to allow the DC voltage component to build up a DC voltage across the load which is as

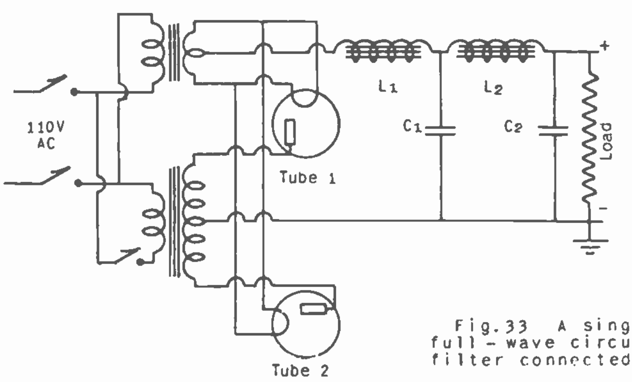


Fig.33 A single-phase, full-wave circuit with a filter connected.

nearly equal to the DC input voltage as possible, and to prevent the AC component of the input voltages from developing an appreciable voltage across the load. If the filter were perfect in its action, the DC voltage present across its output would be equal to the DC voltage applied to its input, and there would be no AC voltage developed across the output of the filter.

A detailed analysis of what takes place in the filter circuit is very complicated and requires a knowledge of higher mathematics. We shall attempt to describe the process of filtration in words rather than resorting to a large number of complex mathematical manipulations. Let us first consider the single-phase, full wave circuit shown in Fig. 33. The wave form of the voltage which an ideal rectifier of this type supplies to the input of the filter system is that shown in Fig. 34. Stating that the rectifier is ideal, means that we are assuming that there is no voltage drop across the tubes or across the plate-transformer secondary. This assumption is permissible since these factors are but modifying influences which do not seriously affect the accuracy of the analysis.

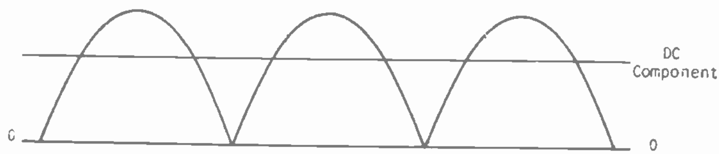


Fig. 34 The output voltage of the circuit shown in Fig. 33. The horizontal line represents the DC voltage component.

The voltage wave of Fig. 34 has a DC component represented by the horizontal line drawn through the wave. There are a very large number of AC components, consisting of a fundamental having a frequency of 120 cycles, and many even harmonic frequencies such as 240 cycles, 480 cycles, etc. The fundamental frequency of 120 cycles has, by far, the largest amplitude, and, if we are able to suppress this frequency effectively, we can be assured that the harmonics will not cause any trouble.

As will be pointed out later, mercury-vapor tubes must be used with a choke-input filter; a condenser-input is in no way practical. Considering the circuit shown in Fig. 33, it should be evident that if the first choke (L_1) had an infinite reactance, it would not be possible for the AC voltage component to produce a current flow through this choke, and the current actually flowing in it would be a pure direct current having a value determined by the DC voltage component and the resistance of the filter and load. Although chokes do not have an infinite reactance, it is clear that the larger the inductance of this first choke, the more nearly constant the current flowing through it will be.

As the voltage output of the rectifier increases, it tends to increase the current flowing through this choke. This causes the choke to produce an induced voltage of such a polarity as to tend to prevent the current from rising. In a like manner, when the voltage output of the rectifier falls, the voltage induced across the choke is in such a direction as to tend to keep the current flowing. The DC component of the output voltage is, of course, the average value of the wave form of Fig. 34. This DC voltage will tend to maintain a constant current through the choke, whereas, the AC component will cause the current flowing in the choke to increase and decrease above and below this average value. If the choke has sufficient inductance, there will be some current flowing through it at all times; that is, the induced voltage across the choke will cause a current to flow from the rectifier, even when the AC voltage across the high-voltage winding is zero. With a moderately large inductance, the current flowing through the first choke will have the wave form shown at A in Fig. 35. The pulses of current passed by the individual tubes will be those illustrated at B and C in this figure.

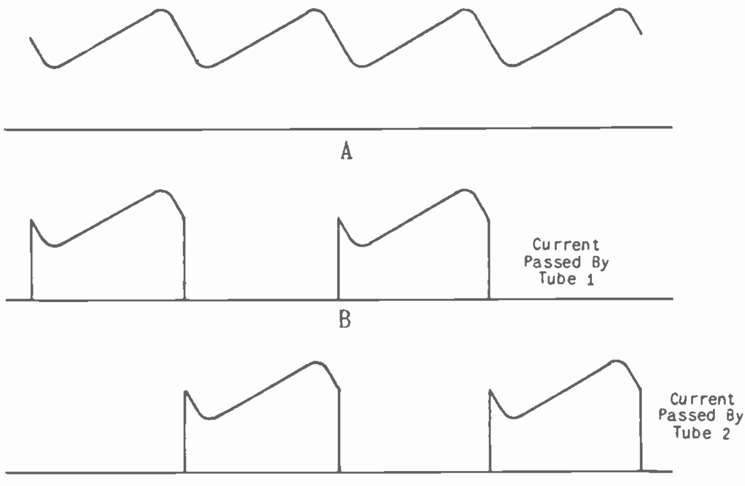


Fig.35 A. Current drawn by the first choke when it has a moderately large inductance.
B. Current drawn by first tube.
C. Current drawn by second tube.

It is this current shown at A in Fig. 35 which charges the first condenser; and since this current does not have a large AC component, it is evident that the voltage across the first condenser will be even more nearly constant than the current through the choke and will have an average value equal to the DC voltage component of the output of the rectifier. In a like manner, the reactance of the second choke will further smooth out the pulsations, and the voltage across the second condenser will be substantially constant.

As the inductance of the first choke is decreased, the current flowing through it becomes more varying, and finally a point is reached where the rectifier does not pass current throughout the entire cycle. The value of the inductance just large enough to cause current to flow from the rectifier at all times is called the "critical inductance". When the fundamental AC voltage component is 120 cycles and the rectifier is of the full-wave type, the value of this critical inductance may be found from this formula.

$$L_1 = \frac{R_{eff.}}{1130} \qquad (1)$$

L_1 is the critical inductance value, and Reff. is the effective resistance of the power supply including the load and the resistances of the chokes. Thus, the effective resistance is very nearly equal to the DC output voltage divided by the DC output current, Suppose, for example, that the output voltage is 1,000 volts and the output current 500 ma. The effective resistance would be:

$$R_{eff.} = \frac{1000}{.500}$$

$$= 2000 \text{ ohms.}$$

In this case, the critical inductance as found by the foregoing formula is:

$$L_1 = \frac{2000}{1130}$$

$$= 2 \text{ henries (approximately)}$$

It is desirable that the rectifier pass current all the time, and the inductance of the first choke should be at least as great as the critical value. When the first choke has an inductance less than the critical value, the current flowing through it has the form shown in Fig. 36. It is seen that there is an appreciable in-

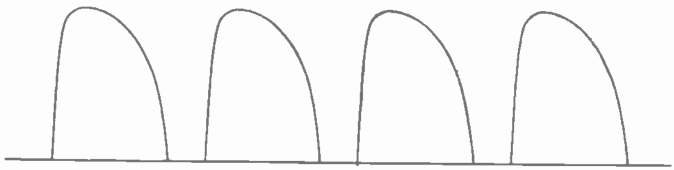


Fig. 36 Current passing through first choke when its inductance is less than the critical value.

terval during which no current passes through the rectifier or the first choke, with the result that the voltage regulation of the power supply will be very poor. In addition, the peak current drawn through the tubes will be considerably greater than the average, and the maximum current that may be drawn from the power supply without exceeding the allowable peak plate current will be considerably reduced.

15. **THE CONDENSER-INPUT FILTER CIRCUIT.** Let us now determine why the condenser-input filter is unsuitable for mercury-vapor tubes and why it has such a poor voltage regulation. A rectifier with a filter of this type is shown in Fig. 37. We shall again assume that the voltage output of the rectifier has the idealized form shown in Fig. 34. As this voltage rises to its peak value, the input condenser (C_1) charges to the peak value of this voltage. The output voltage of the rectifier now falls to zero, but the voltage across the condenser falls relatively slowly, since it must discharge through the comparatively high-impedance path provided by the fil-

ter and the load. The output voltage again begins to increase, and as soon as its value is greater than the voltage now present across the condenser, the condenser again begins to take a charge which

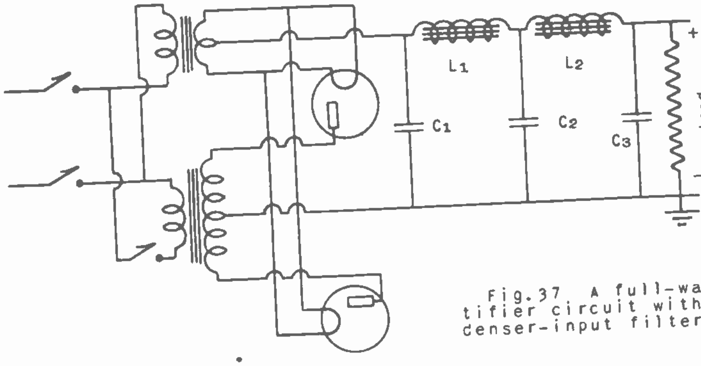


Fig.37 A full-wave rectifier circuit with a condenser-input filter.

continues until the output voltage has reached its peak. Therefore, the voltage across the first condenser has the form shown at A in Fig. 38. This voltage is shown superimposed on the output voltage of the rectifier.

Let us now consider Fig. 39. This figure shows a single tube, the high-voltage winding, and the first condenser and load; the other components have been omitted for simplification. At the instant designated by the vertical line A-A' in Fig. 38, the output voltage of the rectifier is at its peak and the condenser has like-

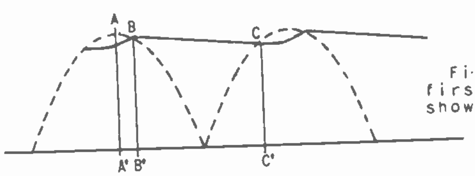


Fig.38A. Voltage across the first condenser of the circuit shown in Fig. 37.

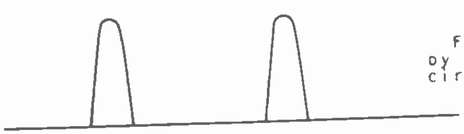
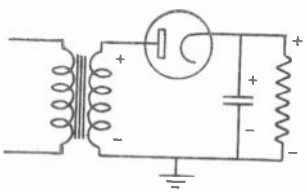


Fig.38B. The current passed by the rectifier of this same circuit.

wise charged to the peak voltage. The voltage output of the rectifier now begins to fall and the condenser starts to discharge through the load. At the time indicated by point B, the voltage output of the rectifier is less than the voltage across the condenser, and the tube no longer passes current. It is evident that the tube will conduct only so long as its plate is positive with respect to its filament. The voltage across the high-voltage winding tends to make the plate positive with respect to ground, but the voltage across the condenser makes the filament positive with respect to ground. Thus, when the instantaneous voltage across the transformer winding is less than the voltage across the condenser at a given time, the filament of the rectifier tube will be more positive than the plate and the tube will not conduct.

And so, no current passes through the tube during the time elapsed between points B and C. The voltage output of the rectifier again rises and as soon as it becomes greater than the voltage to which the condenser is charged; the tube again conducts and the condenser charges to the peak voltage. Thus the current passed by the rectifier tubes in a circuit using a condenser-input filter consists of pulses lasting but a small part of the input cycle, as shown at B in Fig. 38. The peaks attained by these current pulses may be very large, since the charging current that flows when the output voltage is greater than the condenser voltage is limited only by the resistance of the transformer winding, and that of the tube. Furthermore, the average value of this current is very small compared to the peak value, and not much DC current may be drawn from the power supply without exceeding the peak plate current of the tubes.

Fig. 39 Illustrating how the voltage across the first condenser opposes the voltage developed across the high-voltage winding.



Naturally, the DC output voltage when the supply is lightly loaded will be nearly equal to the peak voltage across the condenser, since the condenser will discharge but little during the time that the rectifier is not passing current. As the load is increased and more current is drawn from the supply, the condenser will discharge more during these intervals, and the average DC output voltage will be less. This is the reason that the condenser-input filter produces a greater output voltage and has poorer voltage regulation that the choke-input filter. It is simply not practical to use a condenser-input filter with mercury-vapor tubes, since the peak plate current is sure to be exceeded.

When the input inductance of a choke-input filter has too low a value (less than the critical value), the operation of the rectifier and filter approaches the conditions prevailing with the condenser-input filter, and the regulation is poor. Although the current through the first choke will not fall to zero when the critical inductance is used, much better results are obtained when the first choke has an inductance of twice the critical value. This amount of inductance is usually called the "optimum value".

16. THE SWINGING CHOKE. From formula (1), which gives the critical inductance value, it is seen that this value increases as the effective resistance of the power supply is increased. Naturally, the effective resistance is low when the supply is furnishing its maximum current, and the inductance of the choke will be at least as great as the critical value. When the current drawn from the supply is small, the effective resistance will be large and it is possible that the inductance of the first choke will be less than the critical value. Suppose, for example, that the full-load current of a 1,000-volt power supply is 200 ma. The effective resistance of the supply under this condition is 1,000 divided by .2 or 5,000 ohms. The critical inductance value is:

$$L_1 = \frac{5000}{1130}$$

$$= 4.4 \text{ henries.}$$

If this supply furnishes current for a Class B modulator, the load placed on the supply will vary through considerable limits. There will be instants when nearly the full load is required, and others when no demand is made upon the supply for current. As stated before, the inductance of the first choke may be less than the critical value when the current drawn from the supply is very small, and to prevent the current from falling to too low a value, it is common practice to connect a bleeder resistor across the output of the rectifier so that there will be some load on the supply at all times.

Let us assume that the bleeder current is to be 40 ma. With a 1,000 volt output, this will require a bleeder resistor of 25,000 ohms. When the only current drawn from the supply is the bleeder current, the effective resistance will be approximately 25,000 ohms. The critical inductance value will then be:

$$L_1 = \frac{25,000}{1130}$$

$$= 22 \text{ henries.}$$

Thus, it is clear that the critical inductance value varies from approximately 4 henries to 22 henries from full-load to minimum load. Also, the optimum values range from 8 to 44 henries.

It would, therefore, appear that the inductance of the first choke would have to be at least 44 henries. Before stating definitely that it should have this value, let us inquire further into how this value is measured. A choke may be rated as having an inductance of 30 henries, but unless the amount of DC current flowing through it when it had this value is known, this statement is meaningless. Increasing the DC current which flows through the choke, increases the DC magnetization of the iron core, and tends to change its permeability. Thus, the inductance of the choke decreases as the value of the DC current flowing through it is raised. As the amount of magnetism in the core approaches the saturation value, the inductance drops very rapidly. If the choke is used for smoothing action, its ability to oppose the AC component of the current is decidedly lowered when a large DC current flows through it.

To minimize this effect which the DC current has on the inductance of the choke, an air gap is left in the iron core. Most of the reluctance of the magnetic circuit is that due to the air gap, and since it is not possible to saturate the gap with magnetism, the reluctance of the magnetic circuit as well as its permeability changes a smaller amount when a direct current flows through the choke. This means that the inductance of the choke will not vary as much when the DC current flowing through it is changed. The disadvantage of this method is that the presence of the air gap reduces the overall permeability and the inductance, and to design the choke to have a large inductance requires an unusually large iron core and many turns of wire. Chokes used in small power supplies where the current drain is low will not need a very large air gap, and the size of the core required to produce the desired inductance will not be very great. Chokes which must pass a high value of DC current, however, will require a sizeable air-gap to minimize saturation, and the dimensions of the iron core necessary to produce the desired inductance will be quite large

The second choke of the filter system must maintain a large inductance value even when the full-load current of the supply is drawn through it. If the inductance of this choke were reduced to a low value by the direct current, its sooothing action would be greatly diminished, and the output from the power supply would contain a large ripple voltage. Such a choke is called a "smoothing choke" and is ordinarily quite large.

The first choke of the filter should have an inductance at least as great as the optimum value, and this value changes as the amount of direct current drawn through it varies. (The optimum value is usually greater than that necessary to suppress hum). It would be possible to design a choke having an inductance as large as the highest optimum value (that is, the value of the optimum inductance when the only current drawn from the supply is the bleeder current); and then incorporate a large enough air-gap so that the inductance would not be reduced appreciably when the full-load current was flowing through it, but such a procedure would not be economical. The choke would be very expensive and most of the time the inductance would be greater than is needed. A better solution to the problem is the swinging choke.

A swinging choke is a choke designed with an air gap of such a size that its inductance varies through a predetermined range as the direct current flowing through it changes from minimum to maximum. The air gap need be very small because it is desirable that the inductance of the choke be lowered when a large DC current flows through it. With a small air gap, it is possible to produce a fairly large inductance, when the current flowing through the choke is small, without using an iron core of unusually large dimensions and with considerable less wire on the coil. In the preceding problem, it was determined that the optimum inductance value ranges from 8 to 44 henries. Naturally, a 44-henry choke so designed that its inductance is reduced only slightly when full-load current flows through it would be ideal, but would be very expensive. A swinging choke, whose inductance varied from 8 to 44 henries as the current through it changed from 200 to 40 ma., would serve the purpose just as well and would be considerably cheaper.

The minimum and maximum inductance values and the currents with which these values are obtained are included in the ratings of a swinging choke by the manufacturer along with the DC resistance of the choke. It is desirable that the DC resistance of all chokes be very low, so that there will not be much DC voltage drop across them when the full-load current is drawn from the supply. A high-resistance choke will make the voltage regulation poorer and will dissipate unnecessary power. Furthermore, the insulation resistance between the coil and the core must be sufficient to withstand the peak voltage output of the rectifier. The plate transformer must likewise have low resistance in order to avoid excessive losses. Often a grounded electrostatic shield is placed between primary and secondary to prevent electrical disturbances originating in the power line from being transferred to the output through the capacity existing between primary and secondary.

17. CALCULATING THE RIPPLE VOLTAGE. Naturally, it is desirable to know just how effective the filter is in eliminating the ripple voltage. There are several formulas which give the percentage of ripple voltage in the output of the rectifier. Let us

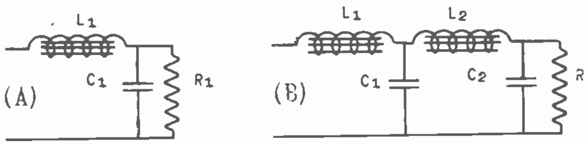


Fig. 40 Two types of filter circuits.

consider Fig. 40. Here are shown two types of choke-input filters. The one at A consists of a single section. The formula for determining the percentage of ripple voltage is:

$$\% \text{ of ripple voltage} = \frac{176}{L_1 \times C_1} \tag{2}$$

Where: L_1 is in henries, and C_1 in mfds.

This formula is only true when the rectifier is a single-phase, full-wave type and the fundamental frequency of the ripple is 120 cycles. When the fundamental frequency of the ripple is 180 cycles, (as it is when the rectifier is of the three-phase, half-wave type), the percentage of ripple is only 44% of the value calculated by the foregoing formula. A three-phase, full-wave or a three-phase, half-wave, double-Y produces a fundamental ripple frequency of 360 cycles, and the percentage of ripple is 11% of the value calculated from formula (2).

The filter shown at B in Fig. 40 consists of two sections and, of course, produces a lower ripple voltage. For a ripple frequency of 120 cycles, the percentage of ripple is:

$$\% \text{ of ripple voltage} = \frac{310}{L_1 \times L_2 \times C_1 \times C_2} \qquad (3)$$

A ripple frequency of 180 cycles produces a ripple percentage of approximately 20% of the value calculated by this formula. With a full-wave, three-phase rectifier, and a ripple frequency of 360 cycles, the ripple voltage is about 1.25% of the value given by formula (3).

It is seen that the filtering action increases rapidly as the number of filter elements is increased, and that with a given size filter, the percentage of ripple voltage from a three-phase rectifier is much less than from a single-phase type. When the power supply is to be used with a CW transmitter, a ripple voltage of 5% or less is satisfactory for the high-powered stages, but should be at least as low as 1% for the low-powered stages. For radiotelephony, the ripple should not be larger than .25%.

There is one other precaution which must be observed in the design of ripple filters. This is to make sure that none of the filter elements produce a series-resonant circuit having a resonant frequency equal to the fundamental ripple frequency. When this occurs, there is a tendency for the filter to become unstable. Should the first choke and condenser be resonant at 120 cycles, the peak current drawn from the rectifier would be tremendous, and the tubes would be damaged. The resonant frequency of these two components should preferably be 20 cycles or less. As the load on the supply is increased, the DC current flowing through the first choke will cause it to saturate and will reduce its inductance, thereby raising the resonant frequency. Care should be taken that the inductance of the first choke is not reduced to such a value as to cause resonance at the ripple frequency.

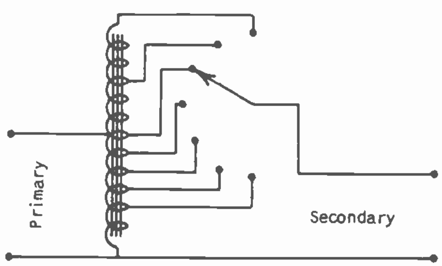


Fig. 41 A simple type of voltage regulator.

18. VOLTAGE REGULATORS. To compensate for poor voltage regulation due to variations in the line voltage, the larger transmitting stations employ automatic voltage regulators. The simplest type of voltage regulator is the one shown in Fig. 41. It is not automatic, but must be controlled manually. It consists of an auto transformer provided with a large number of taps which connect

to the various contacts of a multi-point rotary switch. One side of the line is connected to the bottom of the auto transformer and is also one terminal of the regulated voltage from the output of the device. The other side of the line is connected to a tap on the transformer. The other terminal of the regulated voltage is connected to the contact arm of the rotary switch. The primary of this auto transformer is that part of the coil across which the line voltage is applied. The secondary consists of the entire transformer winding. Since there are more turns in the secondary than in the primary, the total voltage across the coil is somewhat greater than the line voltage. By moving the sliding arm any part of the voltage built up across the secondary may be applied to the transmitter.

When the line voltage is of the correct value, the arm of the switch would be set so that it made contact with the point to which the line voltage lead is connected. In this case, the transformer has a 1:1 ratio, and the secondary voltage (in use) is the same as the primary or line voltage. If the line voltage falls below normal, the sliding arm would be moved to the right, thereby increasing the turn ratio between secondary and primary, until the secondary voltage is again equal to the normal line voltage. In a like manner, a high line voltage would be reduced by moving the arm to a point below the one to which the line wire is connected, thereby using the device as a step down transformer.

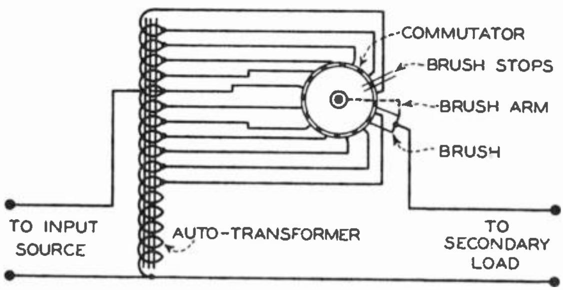


Fig. 42 An improved form of voltage regulator.

One disadvantage of the tap-changing switch is that the circuit is interrupted when the arm is moved from one contact to another. This sets up line surges, causing arcing at the contacts and produces interference in radio receivers. A method which eliminates this difficulty is the one shown in Fig. 42. This voltage regulator also consists of an auto transformer with taps at closely adjacent turns. Each tap is connected to an insulated segment of a commutator which is of practically the same design as is used with direct current motors. A carbon brush makes contact with the commutator, and one line of the regulated voltage is connected to the brush. In the manually controlled regulators, the position of the brush is moved by a hand wheel mounted on the top of the device. By using a brush of such a size that it will bridge two segments of the commutator, continuous control of the output voltage is obtained without the circuit being opened, and with short-circuit currents limited to a negligible value. Furthermore, arcing is prevented by proper design of the brush resistance and by allowing only a small voltage between segments.

Some of the larger regulators are controlled by a motor. Two push buttons are provided: one to increase voltage and one to decrease it. To prevent damage to the regulator, mechanical stops and electrical limiting switches are employed. Thus, it is not possible for the motor to force the mechanism past its maximum or minimum positions.

The automatic types employ a very sensitive voltmeter connected across the output of the regulator. This meter has two contacts, one slightly above the normal voltage and one just below it. As the line voltage increases above normal, the secondary voltage likewise increases, and the needle of the voltmeter makes contact with the upper contact point. This closes a relay circuit, and the relay attracts its armature, thereby starting the motor in such

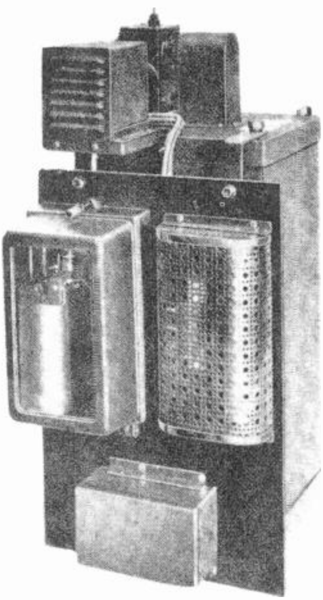


Fig. 43 Photograph of an automatic voltage regulator.

a direction as to decrease the voltage across the secondary. This voltage decreases until the voltmeter needle no longer contacts the upper point, the relay open circuits, and the motors stops. A decrease in voltage below normal causes the needle of the meter to contact the lower contact point, and by means of a relay, the motor is rotated in the opposite direction, thereby increasing the voltage. The regulator will maintain the output voltage within 1% of a predetermined value, being designed so that a 5% change in line voltage is corrected in from 3 to 5 seconds. A photograph of an automatic voltage regulator is shown in Fig. 43. This particular model is for single-phase power, although three-phase models are available. The constructional features of a motor-controlled, non-automatic type is shown in Fig. 44.

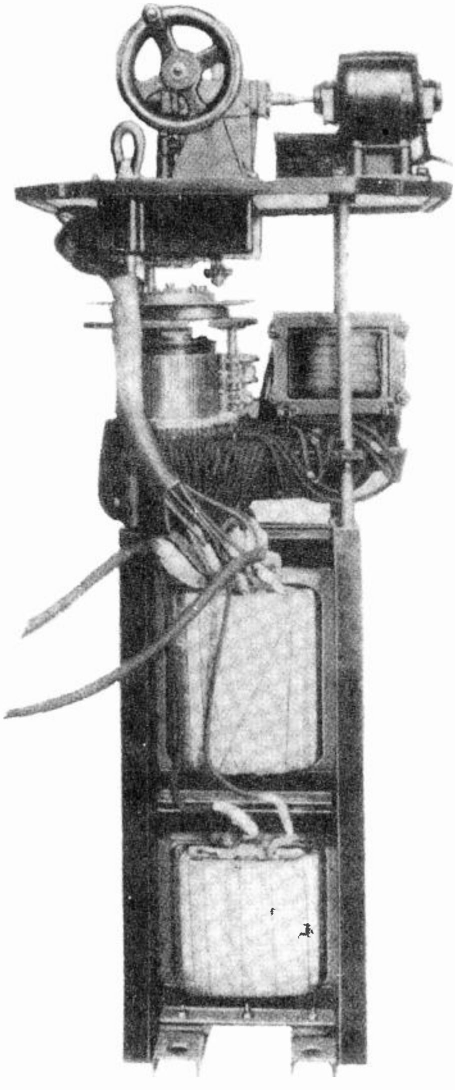


Fig.44 The constructional features of a voltage regulator.

CHARACTERISTICS OF THERMIONIC-TYPE RECTIFIER TUBES.

TYPE	MAXIMUM ALLOWABLE PEAK PLATE CURRENT AMPERES	MAXIMUM SAFE INVERSE VOLTAGE.	FILAMENT VOLTAGE	FILAMENT CURRENT AMPERES.
80	.150	1,400	5	2
5Z3	.300	1,400	5	3
83V	.250	1,100	5	2
84	.075	1,000	6.3	.5
81	.250	2,000	7.5	1.25
217A	.600	3,500	10.	3.25
836	1.000	5,000	2.5	5.
214	7.500	50,000	22.	52.0

CHARACTERISTICS OF TYPICAL MERCURY-VAPOR TUBES.

TYPE	MAXIMUM ALLOWABLE PEAK PLATE CURRENT AMPERES	MAXIMUM SAFE INVERSE VOLTAGE	FILAMENT VOLTAGE	FILAMENT CURRENT AMPERES.
871	.3	5,000	2.5	2.0
866	.6	7,500	2.5	5.0
866A	.6	10,000	2.5	5.0
872	2.5	7,500	5.0	10.0
872A	2.5	10,000	5.0	6.75
869A	5.0	20,000	5.0	18.0
870	450.0	16,000	5.0	65.0
857	20.0	22,000	5.0	30.0

Note: These are all maximum ratings, and in some cases may require careful temperature control. Consult the manufacturer's specifications for temperature limits of each rating.

Notes

(These extra pages are provided for your use in taking special notes)

Notes

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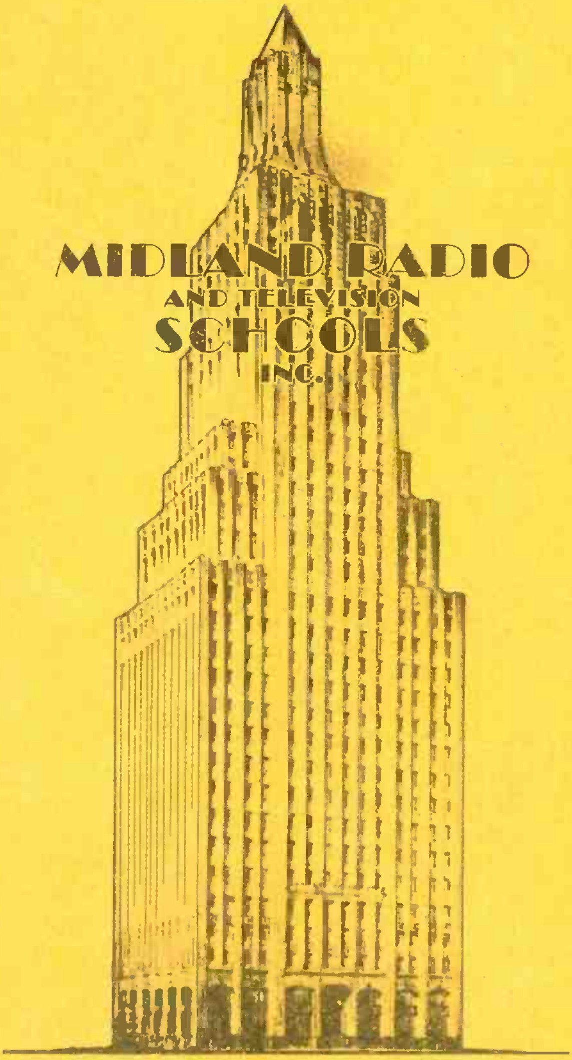

POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

CLASS B R. F. LINEAR AMPLIFIERS

LESSON NO. 10

FORTUNE MAY BE YOURS

.....today's students are tomorrow's inventors.

Radio ears that are far more sensitive than human ears now protect almost every large city from surprise air attack.

Long before the pulsing throb of powerful engines propelling bomb-laden enemy aircraft through the air becomes audible to human ears, the gigantic horns....many of them that comprise the radio ear, pick up the vibration, amplify the sounds and....tne enemy is located.

The many adaptions of radio to peace and war are truly amazing. No doubt there are many new developments awaiting us next year, and the year after, that will make us wonder what is coming next. For example, if the radio ear can be used to detect aircraft, perhaps this same device, highly perfected and compact may be used to apprehend criminals in the future.

Let us use our imagination......A detective, shadowing bank robbers, receives a tip that they are at a certain hotel. He loiters about the lobby, stopping now and then, apparently just another traveling man passing away a few hours before train time.

Groups of people are seated or standing around the lobby conversing in low tones. As the detective is passing within twenty feet of two men, he suddenly walks toward them, gun in hand. And the men prove to be the wanted bank robbers. How did the detective know? He heard their whispered conversation.

Under the detective's coat was a compact device....an electric ear that picked up the vibrations created by the whispering robbers, amplified them, and then made them audible by means of a vibrator placed against a certain place on the detective's body where it would be hidden from view.

Such a device may be created along with thousands of other new inventions. And you, Mr. Student, may become the inventor of one or more Radio or Television devices that can bring wealth and world-wide fame.

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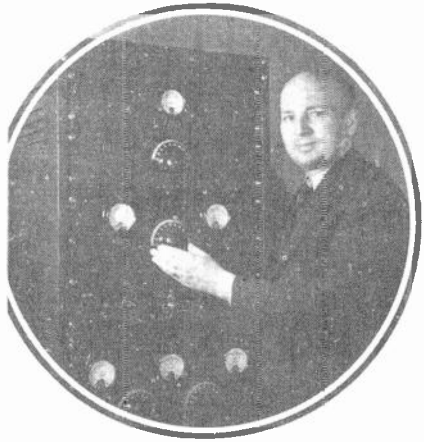
Lesson Ten

CLASS B R.F. LINEAR AMPLIFIERS



"Until high - powered modulator stages were developed the only way in which high power could be obtained from a transmitter was through the use of a linear amplifier.

"Straight linear stages are not used a great deal in new transmitters. However, there are a great number of transmitters now in use which use linear amplifiers.

1. HIGH AND LOW LEVEL MODULATION. Prior to this time, it has been assumed that the output of the modulated Class C stage was fed directly into the transmitting antenna. Quite often this is the case. Such an arrangement is called "high-level modulation", and has advantages and disadvantages which will now be discussed. The Class C amplifier is very efficient and it is desirable that the higher powered stages be operated under those conditions which produce maximum efficiency. Suppose, for example, that the transmitter must develop 10,000 watts in the radiation antenna. With an efficiency of 60% or higher (a not uncommon value for a Class C amplifier), the DC power input to the Class C final stage would be 16,666 watts. Assuming that high-level modulation is employed; that is, that this final stage is modulated, the audio power that would be required from the modulator is 8,333 watts for 100% modulation.

The foregoing illustration discloses both an advantage and a disadvantage of high-level modulation. The fact that the Class C stage is itself very efficient is a decided advantage. On the other hand, the necessity of producing over 8,000 watts of audio power is a disadvantage, since audio amplifiers, even when operated under Class B conditions, have a relatively low efficiency. If a Class A modulator with a probable efficiency of 25% were used, the DC power input to the modulator would need to be over 30,000 watts; thereby, making the overall efficiency of the transmitter relatively low. With the comparatively recent advent of Class B modulators, this condition has been improved. A well designed Class B modulator will have an efficiency of 50%, or greater, and the amount of DC power input to the modulator stage is considerably reduced.

Before the Class B modulator became popular, nearly all high-powered transmitters employed low-level modulation. In low-level modulation, the final or output stage is not modulated, but serves merely to amplify the modulated wave producdd in some preceding stage of the transmitter. When plate modulation is used, the modulated stage is operated under Class C conditions and is followed by one or more Class B, R.F. linear amplifiers. As explained in Lesson 7 of this Unit, it is not possible for a Class C stage to amplify a modulated R.F. wave. Such an attempt produces distortion of the modulation envelope, because the Class C stage is biased to a value equal to twice cut-off and plate current would not flow during the trough of the modulation cycle. To amplify a modulated wave requires a linear amplifier, one in which the tank current is directly proportional to the grid-exciting voltage. The grid-exciting voltage is a modulated R.F. wave and, if the wave is modulated 100%, the amplitude of the R.F. cycles will vary from zero to twice the unmodulated value. At the time the grid-exciting voltage is zero (that is during the trough of the modulation cycle), the radio frequency tank current must likewise be zero. As the amplitude of the alternating grid voltage increases, the tank current must increase in direct proportion, because a linear relationship must exist between the tank current and the alternating grid voltage. In the Class C amplifier, the tube does not begin to draw plate current until the alternating voltage in the grid circuit is sufficient to make the instantaneous grid voltage greater than the cut-off value. Thus, it is seen that a linear amplifier must not be biased beyond the cut-off point, if the troughs of the modulated waves are to cause a change in the plate tank current.

So far as satisfying the condition of linearity is concerned, the Class A amplifier would be ideal. In a Class A amplifier, the plate current is an exact reproduction of the grid-exciting voltage. Naturally, the plate tank current would likewise be directly proportional to the alternating voltage in the grid circuit. Practical considerations, however, demand that the efficiency of the linear amplifier be as large as possible, and the poor efficiency of a Class A amplifier makes it undesirable for this application.

The procedure for increasing the efficiency of any vacuum tube amplifier is to reduce the plate power input when the tube is resting, or is not amplifying a signal voltage. This, of course, is accomplished by increasing the grid bias so that the no-signal plate current will be very low. By making the bias voltage equal to the cut-off value, it is possible to increase the efficiency above that of a Class A amplifier and still obtain linear operating conditions.

The advantage of using low-level modulation combined with linear amplifiers, is the fact that a low-powered stage may be modulated and a smaller amount of audio power will be required. Thus, there is a considerable saving since high-powered audio amplifiers will not be necessary. Opposed to this is the fact that the average efficiency of a linear amplifier is rarely greater than 33%. To develop an R.F. power output of 10,000 watts, the Class B linear amplifier would need a plate input power in excess of 30,000 watts.

In addition, the factors determining the linearity of the amplification are quite critical and linear amplifiers are noted for their difficulty of adjustment.

Many transmitters having an output greater than 1,000 watts employ linear amplifiers. This, however, is not always the case, for high-level modulation is used at present with a 500-kilowatt transmitter. Smaller transmitters of 1,000 watts capacity may employ either system.

2. VOLTAGE AND CURRENT RELATIONS IN THE CLASS B LINEAR AMPLIFIER. In practice, the Class B amplifier is not biased to the actual cut-off point, because the curvature of the lower end of the grid-voltage plate-current characteristic curve introduces distortion and causes the operating conditions of the amplifier to depart from true linearity. Usually the bias value selected is that value obtained by dividing the applied DC plate voltage by the amplification factor of the tube. It is the grid voltage at which plate current cut-off would occur if the Eg-Ip characteristic were a straight line.

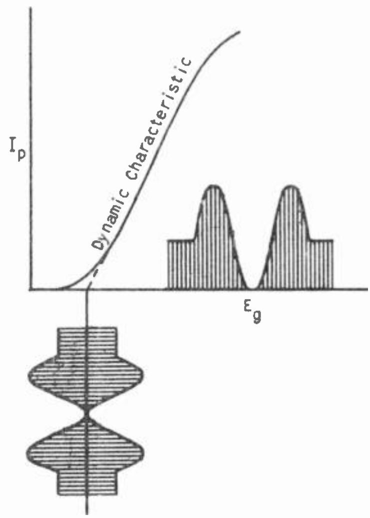


Fig. 1 Illustrating the operating condition of a linear amplifier.

Fig. 1 illustrates a dynamic Eg-Ip characteristic and shows how the grid bias would be selected for linear operation. Since the tube is biased practically to cut-off, plate current flows for approximately 180° of the grid-exciting cycle. The plate current itself is, of course, not a true reproduction of the alternating grid voltage, since no plate current flows during the negative alternation of the grid-exciting cycle. However, the load impedance of a linear amplifier is always a tuned circuit, and this tuned circuit has the ability of smoothing out the plate current pulses into practically pure sine waves. The fundamental frequency component of the plate current pulses creates an R.F. voltage across the plate tank circuit and causes a relatively large tank current to flow. Fig. 2 shows the wave form of the grid-exciting voltage, the actual plate current, the varying plate voltage, and the fundamental frequency component of the plate current.

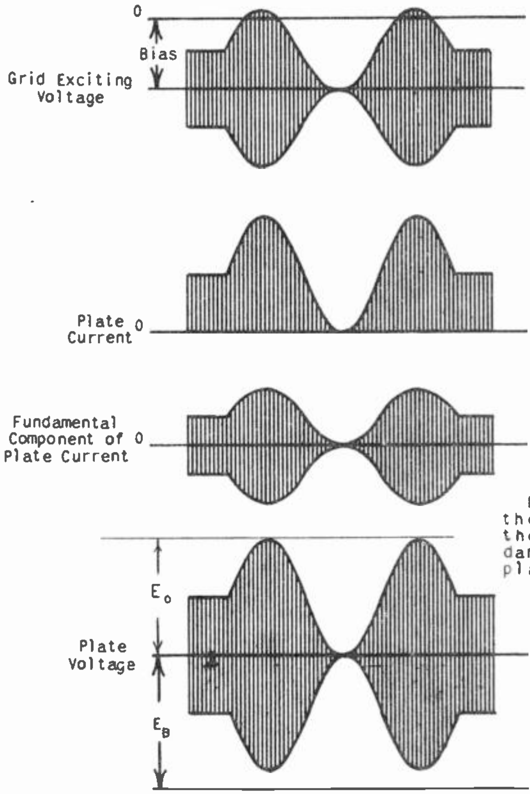


Fig. 2 The wave forms of the grid-exciting voltage, the plate current, its fundamental component, and the plate voltage.

One important difference between the linear amplifier and the Class C amplifier is illustrated in Fig. 1. The amplitude of the grid-exciting voltage is not constant, but varies at an audio rate determined by the output of the modulator. As the amplitude of the grid-exciting voltage increases from its unmodulated value to the greatest value obtained during the modulation cycle, the radio frequency tank current must increase in the same ratio. For example, if the grid-exciting voltage is 100% modulated, the greatest amplitude will be twice the unmodulated value and the tank current must likewise double the value it has with no modulation. For the tank current to double, the peak value of the fundamental plate current component must likewise double. Such a condition will only be obtained when the plate current flows for 180°, or more, and the plate current pulse is an exact reproduction of a half sine wave.

When the linear amplifier is improperly adjusted, the dynamic Eg-Ip curve is not a straight line and the plate current is not directly proportional to the grid voltage. This causes the plate current pulse to be flattened, and the peak of the fundamental component does not bear the same ratio to the peak value of the plate current pulse. This condition is illustrated in Fig. 3. When the

Fig. 3 Showing the effect of non-linear operation.

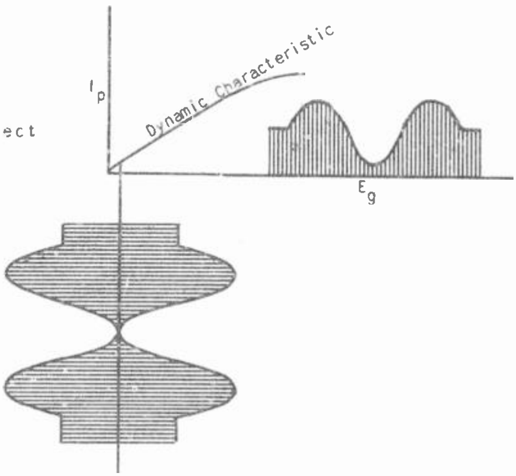
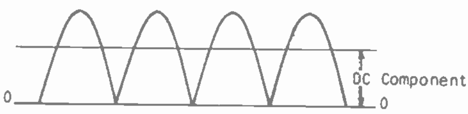


plate current pulses are half sine waves, the peak of the fundamental plate current component is equal to one-half of the peak of the plate current pulse, and a linear condition exists.

3. POWER RELATIONS OF A CLASS B AMPLIFIER. We shall first consider the conditions existing when the grid-exciting voltage is unmodulated and has the greatest possible value which will still allow the amplifier to operate in a linear manner. The DC power

Fig. 4 Illustrating the DC component of the output of a full-wave rectifier.



input applied to the plate circuit is equal to the DC plate current component multiplied by the plate-supply voltage. This is:

$$P_i = I_b \times E_B \qquad (1)$$

Where: P_i is the power input in watts,
I_b is the DC plate current in amperes,
E_B is the plate-supply voltage in volts.

The DC plate current component is the average value of the plate current pulses averaged throughout one complete cycle of the grid-exciting voltage. As we already know, the average value of an alternating current is equal to .637 times the peak value. As explained in Lesson 13 of Unit 1, this is the average of all the instantaneous values throughout one alternation. If the plate current wave form were the same as the output of a full wave rectifier, as shown in Fig. 4, the DC component would still be .637 times the peak value of the alternation. In the linear amplifier, however, the plate current flows for only one-half cycle, or only half as

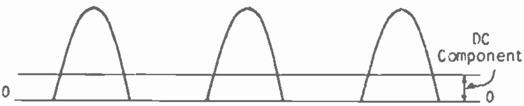


Fig.5 Showing the relation between the plate current pulses of a linear amplifier and their DC component.

long as in the full wave rectifier. For this reason, the DC component of the plate current of a linear amplifier is ½ of .637, or .318 times the peak value of the plate current pulses, as shown in Fig. 5. Expressed as an equation, this becomes:

$$I_B = .318 \times I_m \tag{2}$$

Where: I_B is the DC plate current, and
I_m is the peak value of the plate current pulse.

By substituting the value of I_B as given in equation (2), into equation (1), there is obtained:

$$P_i = .318 \times I_m \times E_B \tag{3}$$

Let us now determine how the power output would be calculated. When the operating angle is 180°, and the plate current pulse is a half sine wave, the peak value of the fundamental plate current

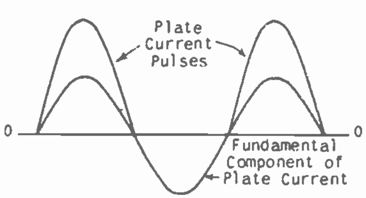


Fig.6 The relation between the plate current pulses and the fundamental plate current component.

component is one-half of the peak of the plate current pulse, as illustrated in Fig. 6. This is:

$$I_o = .5 \times I_m \tag{4}$$

Where: I_o is the peak value of the fundamental plate current component, and
I_m is the peak of the plate current pulse.

This peak value of the fundamental plate current component in flowing through the plate tank circuit develops an R.F. voltage which may be designated as E_o. The power in the plate tank circuit is equal to the R.M.S. value of the fundamental plate current component times the R.M.S. voltage developed across the tank circuit. The R.M.S. current value is:

$$I_{rms} = .707 \times I_o \tag{5}$$

Substituting the value of I_o as given in equation (4) into equation (5), there results:

$$I_{rms} = .707 \times .5 \times I_m$$

$$= .3535 \times I_m \tag{6}$$

The R.M.S. value of the R.F. voltage developed across the tank circuit is:

$$E_{rms} = .707 \times E_o \tag{7}$$

Where: E_o is the peak R.F. voltage developed across the tank circuit.

Equation (6) gives the R.M.S. value of the fundamental plate current and equation (7) the R.M.S. value of the R.F. voltage across the tank. The product of these two represents the average R.F. power in the tank circuit. This is:

$$P_o = .3535 \times I_m \times .707 \times E_o$$

$$= .25 \times I_m \times E_o \tag{8}$$

The efficiency is the power output divided by the power input. This is equation (8) divided by equation (3). When this operation is performed, the following results:

$$\text{Efficiency} = \frac{P_o}{P_i}$$

$$= \frac{.25 \times I_m \times E_o}{.318 \times I_m \times E_B}$$

The I_m's will cancel and the final result is:

$$\text{Efficiency} = .786 \times \frac{E_o}{E_B} \tag{9}$$

E_o, you will remember, is the peak R.F. voltage developed across the plate tank circuit, whereas E_B is the plate-supply voltage. The actual plate voltage at any instant is the sum of the plate-supply voltage and the instantaneous R.F. voltage across the tank circuit. In the study of the Class C amplifier, it was learned that the peak R.F. voltage developed across the tank circuit was very nearly equal to the supply voltage, in which case, the actual plate voltage varied from nearly twice its DC value to a very low value not far from zero. If it were possible to make the peak R.F. voltage across the tank circuit equal to the plate-supply voltage, then E_o would be equal to E_B and the efficiency as given in equation (9) would be equal to 78.6%. This, however, is the theoretical maximum which can never be obtained in actual practice. In the first place, it is not possible to make the peak tank voltage equal to the plate-supply voltage, for, in that case, the actual plate voltage would be zero during an instant of the cycle. As has been previously learned, the grid-exciting voltage and the varying plate voltage are 180° out of phase and the maximum positive grid voltage should not exceed approximately 80% of the minimum plate voltage. When the minimum plate voltage is too low, the tube becomes saturated, the grid robs the plate of electrons, and the plate current pulse has a dip indicating that the dynamic Eg-Ip characteristic is not linear and that the tank current will not be directly proportional to the grid-exciting voltage. In actual practice, the peak tank voltage is usually limited to a value of from .85 to .9 times the plate-supply voltage, and, allowing for a 5% loss in the tuned output circuit, the maximum possible efficiency will range from 63 to 67%.

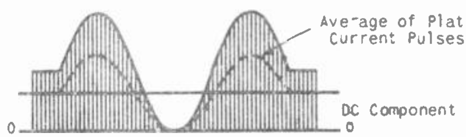
This high value of efficiency is obtainable only when the peak R.F. voltage across the tank is nearly equal to the plate-supply voltage, or when the grid-exciting voltage is the maximum that may be applied without causing tube saturation and departure from a linear condition. When the grid-exciting voltage is 100% modulated, the peaks of the R.F. cycles will be twice as great during the peak of modulation, as they are when the carrier is unmodulated. To achieve linearity, it is necessary that the circuit be so adjusted that it will be linear even during the modulation peak. For this reason, it is necessary that the unmodulated grid-exciting voltage be just half as large as that value obtained during the modulation peak.

When properly adjusted, the unmodulated grid-exciting voltage is half of that value which will cause tube saturation. Assuming that the amplification is linear, the amplitudes of the plate current pulses will be half as large as during the modulation peak, the peak value of the fundamental plate current component will likewise be half as great, and the R.F. voltage developed across the plate tank circuit will also be reduced to one-half of its former value. This means that E_o will now be less than half as large as E_B, and from equation (9), it becomes evident that the efficiency with the unmodulated condition will be just half as great as it is during the peaks of modulation. It has been stated that maximum possible efficiency is approximately 67%. With the carrier unmodulated, the average efficiency will approximate 33%. This means that the efficiency of the linear amplifier is relatively low during the period when the transmitter is not modulated.

Although the efficiency will increase to a maximum of 67% during the peaks of modulation, the average efficiency throughout a modulation cycle will be barely 50%. This, of course, assumes that the carrier wave is 100% modulated. As stated in Lesson 7, 100% modulation is obtained only during the peaks of the audio signal. The average modulation will not be much greater than 30 to 40%. When such is the case, the average efficiency of the Class B linear amplifier will be only slightly greater than that obtained with the unmodulated carrier. This is the greatest single disadvantage of the linear amplifier.

During modulation, the plate current consists of pulses which are half sine waves and which vary in amplitude at the modulation frequency. The fundamental component of the plate current likewise varies in amplitude at the modulation frequency, causing the R.F. voltage produced across the tank circuit and the tank current to vary in exact accordance. Thus, the wave form of the current flowing in the antenna is exactly the same as the varying voltage applied to the grid of the linear amplifier. The average value of the plate current pulses varies at the frequency of the modulation. This is illustrated in Fig. 7. The average value of this

Fig.7 Showing how the DC plate current varies at an audio frequency rate.



audio frequency is a pure direct current having the same value as when the carrier is unmodulated. Therefore, the reading of the DC plate current meter should not change during modulation. Since the DC plate current does remain constant, the average power input to the linear amplifier is the same whether the grid-exciting voltage is modulated or not.

We have learned that the power output of an amplifier, during the peak of the modulation cycle, is four times as great as when the carrier is unmodulated. Let us see how this is accomplished. As the grid - exciting voltage increases from its unmodulated value to the peak obtained during 100% modulation, the fundamental plate current component doubles its value. This causes the peak R.F. voltage developed across the tank to likewise double. According to equation (9), doubling the tank voltage (E_o) will double the efficiency of the amplifier. In addition, the average value of the plate current increases from its unmodulated value until it is twice as large. Therefore, at the peak of the modulation cycle, the plate power input, which is equal to the average plate current times the plate-supply voltage, becomes twice as great. The fact that both the power input and the efficiency increase twofold during the modulation peak causes the R.F. power output at this time to be four times as great as it is with no modulation.

Suppose, for example, that the plate - supply voltage is 1000 volts, the DC plate current 100 ma., the efficiency (when unmodulated) 33%, and that the grid-exciting voltage is to be 100% modulated.

The DC power input is:

$$P_i = E_b \times I_b$$

$$= 1000 \times .100$$

$$= 100 \text{ watts}$$

With an efficiency of 33%, the R.F. power output is:

$$P_o = .33 \times 100$$

$$= 33 \text{ watts}$$

At the peak of the modulation cycle, the R.F. power output will increase four times, or to 132 watts. To accomplish this, the efficiency will increase to 66% and the instantaneous DC power input will become 200 watts. This will give an R.F. power output of:

$$P_o = .66 \times 200$$

$$= 132 \text{ watts}$$

The power dissipated at the plate of the tube, when the carrier is unmodulated, is the difference between the R.F. output power and the DC input. It is:

$$\text{Power loss at plate} = 100 - 33$$

$$= 67 \text{ watts}$$

At the instant of peak modulation, the power dissipated at the plate of the tube is:

$$\text{Power dissipated} = 200 - 132$$

$$= 68 \text{ watts.}$$

It is evident that the plate dissipation increases but very little as the grid-exciting voltage becomes larger. With the plate dissipation practically constant and the power input momentarily doubling, it is seen that the efficiency of the stage increases considerably at the modulation peak.

So far, we have considered only instantaneous values. Let us now determine how the average efficiency varies with modulation. Although the instantaneous DC power input doubles its value during the peak of modulation, the average power input throughout a modulation cycle has the same value as with no modulation. This, of course, is only true if the modulation envelope is undistorted. With 100% modulation, the amplitudes of the R.F. cycles must double their value during the modulation peak and become zero at the trough. With 100% modulation, the power contained in the modulation wave is $1\frac{1}{2}$ times as great as that in the unmodulated carrier.[1] Since the average power output of a linear amplifier increases $1\frac{1}{2}$ times with 100% modulation, and the average DC power input does not change, the average power dissipated at the plate of the tube must be less with modulation than without. This, of course, is the same as saying that the average efficiency is greater with modulation. Although the efficiency doubles its value at the peak of the modulation cycle, the average efficiency is only $1\frac{1}{2}$ times as great as it is with an unmodulated carrier.

Let us now apply this theory to the preceding problem. The power output (unmodulated) was 33 watts. Although the peak power increases to 132 watts, the average power throughout a modulation cycle (with 100% modulation) increases $1\frac{1}{2}$ times, or to 49.5 watts. The average efficiency also increases to $1\frac{1}{2}$ times the value it had with no modulation, and becomes 49.5%. Since the average DC power input is the same as with no modulation, the average power output is:

$$\text{Average power output (modulated)} = .495 \times 100$$

$$= 49.5 \text{ watts.}$$

The average power dissipated at the plate during 100% modulation is the difference between the DC power input and the average power output. This is:

$$\text{Average plate dissipation} = 100 - 49.5$$

$$= 50.5 \text{ watts.}$$

This is in comparison with the 67 watts dissipated when the stage was unmodulated.

From the foregoing discussion, there results the somewhat amazing fact that the linear amplifier tube will run cooler when the grid-exciting voltage is modulated than when it consists of an unmodulated carrier. The power dissipation occurring at the modulation peak will probably be greater than it is when the carrier is unmodulated. However, the increase in plate dissipation as the grid-exciting voltage varies from the carrier to the peak value is not as large as the decrease in the plate loss, which occurs as the grid-exciting voltage falls to its minimum value. Thus, the average dissipation throughout the modulation cycle is less than that taking place at the carrier level. It is permissible to adjust the amplifier so that the plate loss with full-exciting voltage is slightly greater than the rated value, since the exciting voltage is at this peak value only a small fraction of the time when the signal is a modulated wave.

4. DESIGN OF CLASS B LINEAR AMPLIFIERS. There are two important differences between the Class B, R.F. linear amplifier and

[1] Refer to Page 17, Lesson 5, Unit 3.

the Class C amplifier. The Class B amplifier is biased practically to cut-off and the grid excitation is so adjusted that grid current flows only at the peak of a 100% modulated signal and never in amounts exceeding 2 or 3 milliamperes. On the other hand, the Class C amplifier is biased to at least twice cut-off, the grid exciting voltage is unmodulated and is ordinarily sufficient to produce saturation. If the Class C stage is modulated, the operating angle is not constant, but increases and decreases as the applied plate voltage changes with the modulation frequency. For the Class B amplifier to be linear, it is essential that the operating angle not change throughout the modulation cycle. Plate current should flow in pulses lasting for 180 electrical degrees, irrespective of the amplitude of the grid-exciting voltage. To make certain that such is the case and that linear operating conditions will be obtained, it is absolutely necessary that the DC grid current be very low.

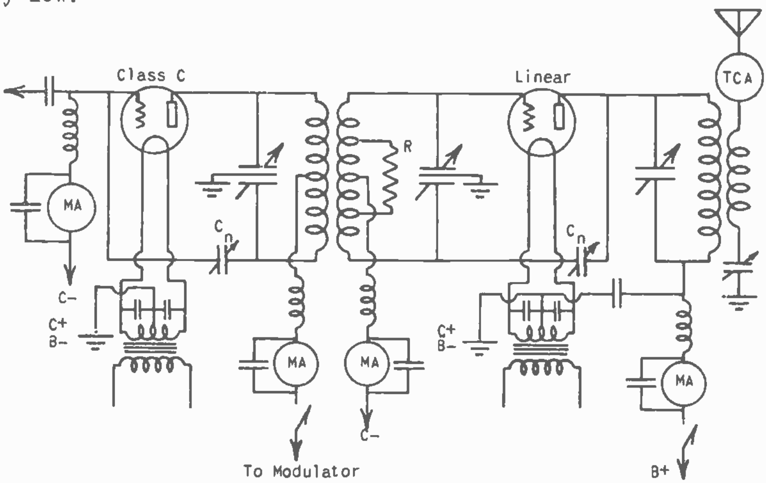


Fig.8 A plate modulated Class C stage coupled to a single-ended linear amplifier.

Since the linear amplifier does not draw grid current except at the peak of a 100% modulated signal, it is evident that the amplifier draws but very little power from the preceding stage. The preceding stage is ordinarily a plate modulated Class C amplifier, and since the linear amplifier does not load the Class C stage very heavily, additional loading is usually necessary. Fig. 8 shows a plate modulated Class C amplifier coupled to a linear amplifier. Notice that inductive coupling is used. The grid circuit of a linear amplifier is ordinarily tuned and either inductive or link coupling is employed between the Class C stage and the linear amplifier. The resistor R which is tapped across the grid tank circuit is called a "swamping resistor". Its purpose is to dissipate some of the power output of the Class C stage and thereby cause the load on the Class C stage to be constant and to have the correct value to produce a maximum power output with reasonable efficiency. The amount of power dissipated in this resistor depends upon the voltage across it, which, in turn, is determined by the number of turns of the grid tank across which it is clipped. In amateur practice, this resistor is often a combination of one or more electric lamps.

A tube to be used as a linear amplifier should have a high plate dissipation rating, a low or medium amplification factor, a low plate resistance, and a high transconductance. The plate dissipation should be high because the efficiency is ordinarily very low and the power output is dependent primarily upon the amount of power that the tube may safely dissipate. A tube with a low mu is preferable, as this will make the cut-off bias value rather high. With a high bias, the grid-excitation voltage may be greater without causing the linear amplifier to draw excessive grid current. High transconductance allows the tube to operate with a minimum of excitation and at the highest efficiency.

To make sure that the operating angle of the plate current remains at 180°, it is necessary that the bias voltage be as constant as possible. Thus, it is not possible to use grid-leak bias, since the excitation voltage would cause grid current to flow only on occasional peaks. A grid-bias power supply with good voltage regulation is the best answer to this problem.

It may be proved that the apparent internal resistance of a vacuum tube working as a Class B linear amplifier is equal to twice the tube's rated plate resistance. With this fact in mind, it is clear that the tube should work into a load impedance equal to twice its rated plate resistance to develop a maximum power output. The design of the load circuit determines to a large extent the linearity of the amplifier. A high value of load impedance tends to straighten out the dynamic E_g-I_p characteristic and thereby creates a more linear operating condition. On the other hand, a given grid-exciting voltage will develop a larger R.F. voltage across a high load impedance than across a low one. When the load impedance is too large, the R.F. tank voltage developed by the unmodulated carrier is greater than it should be, and as the grid-exciting voltage increases on a modulation peak, it is not possible for the R.F. tank voltage to double its value. This is occasioned by the fact that the plate current pulses are not able to double their amplitude as the tube's saturation point is reached before the grid-exciting voltage attains its greatest peak value. This departure from linearity causes distortion of the modulation envelope and introduces audio harmonics which are highly objectionable.

A load impedance of too low a value will cause the dynamic E_g-I_p characteristic to be curved, which produces non-linear operation. It is this critical adjustment of the load impedance and grid-exciting voltage which makes a linear operating condition difficult to obtain. This is not to be understood that the so-called linear amplifier cannot be made to operate under a linear condition, but it does mean that the adjustment of such a stage is far more difficult than that of a Class C stage.

5. ADJUSTMENT OF THE LINEAR AMPLIFIER. In the following dis-

cussion, reference will frequently be made to the diagram shown in Fig. 9. This diagram includes a plate-modulated Class C stage

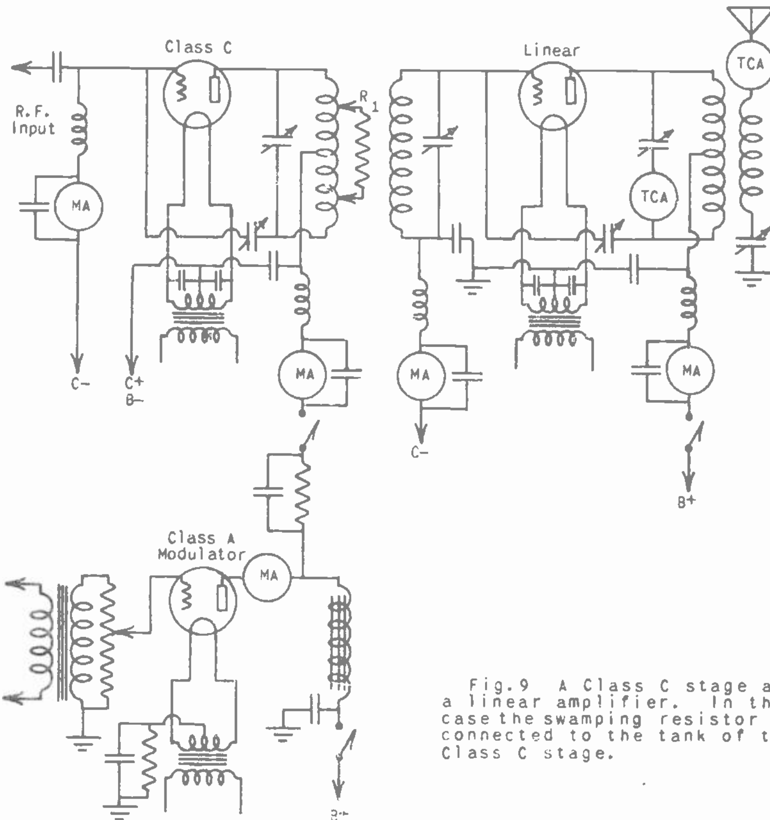


Fig. 9 A Class C stage and a linear amplifier. In this case the swamping resistor is connected to the tank of the Class C stage.

coupled to a Class B linear amplifier. The first step is to apply the filament voltage to both tubes, after which the Class C stage is neutralized using any one of the methods described in Lesson 3 of this Unit. Before any adjustments whatsoever are made on the linear amplifier, the Class C stage must be operating as nearly perfect as possible. For this reason, we may say that the first step in adjusting a linear amplifier is to make certain that the preceding Class C stage is operating as it should. When the Class C stage has been neutralized, the plate voltage should be applied and various adjustments made until its efficiency is quite high. Since the Class C stage is not delivering its power into an antenna, it is loaded in this case by connecting the non-inductive resistor R_1 across a part of the plate tank circuit. The greater the number of turns between the two ends of the resistor. the larger the voltage across it will be and the more power it will dissipate. Therefore, the connections to the resistor are varied until the Class C stage draws normal plate current. Perhaps the LC ratio of the Class C stage will also have to be changed before the tube produces a reasonable power output without exceeding its dissipation rating.

The next step is to secure an audio oscillator whose output is a pure sine wave. The output of the audio oscillator is built up by a suitable voltage amplifier and is then fed to the grid circuit of the modulator. At this stage, a cathode ray oscilloscope is a great help. The oscilloscope should be connected to the tank circuit of the Class C stage, as explained in Lesson 7 of this Unit, and the gain of the modulator should be increased until 100% modulation is indicated. In addition, note whether the modulation is linear. This perhaps can be determined more easily by using the familiar trapezoid pattern on the scope.

With the Class C stage functioning properly, it is now time that we turned our attention to the linear amplifier. When properly adjusted, the linear amplifier draws grid current only during the peaks of 100% modulation. When the excitation voltage is at the carrier level, no grid current flows, the tube is not working near its saturation point, and the efficieacy is low. To make sure that linearity will be obtained, it is desirable that the rough adjustments be made at the saturation point, because that is a limiting factor to linearity. To cause the tube to operate at the saturation point, the unmodulated grid-exciting voltage should be twice as great as normal operating conditions will require. That is, the stage should be adjusted while the grid-exciting voltage is unmodulated, and the excitation voltage should be sufficient to produce saturation. Ordinarily, however, this large amount of excitation voltage is not available from the Class C stage.

For this reason, a better method is to reduce the grid bias to one-half of that value which normally gives plate-current cutoff. Naturally, the plate voltage cannot yet be applied, because the linear amplifier has not been neutralized. With the bias value set to one-half of the value determined by dividing the normal applied plate voltage by the amplification factor of the tube, the grid tank circuit is tuned for maximum grid current and the coupling between this tank and that of the Class C stage is adjusted to produce a small flow of grid current. The stage is then neutralized by one of the methods outlined in Lesson 3 of this Unit.

Neutralization of the Class B linear amplifier is very important and great care should be taken with this process. Whereas a slight misadjustment of the neutralizing condenser of the Class C stage would probably cause no harm, it is absolutely essential that the linear amplifier be perfectly neutralized. Failure to observe this important point will often result in non-linear operating conditions. It is not sufficient that the neutralization be only complete enough to prevent oscillation, but it must be so perfect that no trace of R.F. current in the plate tank of the linear stage is indicated by the best possible R.F. indicator available. When the neutralization is slightly imperfect, it often happens that the stage will oscillate during the peaks of the modulation cycle. This action will absorb power from the tank circuit and will not allow the tank current to increase during the modulation peaks, as much as it should for linear operation.

When it is certain that the linear stage is perfectly neutralized, a plate voltage equal to one-half of that which will be used during normal operation should be applied to the stage. With both the plate voltage and grid bias reduced to one-half value, the amplifier will be biased to the cut-off point, and the normal unmodulated output of the Class C stage should be sufficient to cause saturation.

The plate tank is tuned to resonance, as indicated by minimum plate current, and the grid excitation (unmodulated) should now be applied and increased until the saturation point is reached, as indicated by non-linearity between the excitation voltage and the tank current of the linear amplifier. (For best results, a thermocouple ammeter should be connected in the tank circuit of the linear stage so that the presence of non-linearity may be more easily detected.) The next step is to adjust the plate load. The antenna is coupled to the tank circuit of the linear amplifier and is tuned to resonance. Then the plate loading is varied by changing the antenna coupling and the LC ratio of the tank circuit until the optimum tank current is obtained. It has been determined by a complex mathematical procedure that the optimum tank current is given by this equation:

$$I_T = .707 \times \frac{E_B \times (\mu + 1)}{X_C \times (\mu + 2)} \qquad (10)$$

Where: E_B is the normal applied DC plate voltage,
X_C is the capacitive reactance of the tank circuit,
μ is the amplification factor of the tube, and
I_T is the tank current of the linear amplifier.

The power input to the linear stage should now be calculated by multiplying the DC plate current obtained with the optimum value of tank current, by the normal plate voltage which will be applied to the stage. If this power input is excessive; that is, if it is greater than should be used without causing the tube to overheat, the excitation should be reduced slightly until the correct value is obtained. In any event, it is desirable to reduce the excitation a little below that point which produces saturation, because the relationship between the tank current is not linear in the vicinity of saturation.

Without changing the excitation voltage, the grid bias and plate voltage should now be set at their normal values, and the Class C stage should be 100% modulated. By connecting the oscilloscope to the antenna, the linearity of the Class B stage may be determined. The plate current should not vary appreciably as the modulation is changed from 0 to 100%. If the oscilloscope shows that the modulation increases without any irregularities or flattening of the peaks, as the gain control of the modulator is varied from zero to the point which produces 100% modulation, and if the tubes do not operate too hot when the excitation voltage is unmodulated, the adjustments are satisfactory. It should, however, be realized that the values secured in actual tuning may vary as much as 5% from the calculated values, and the final adjustment must always be experimental. The chances that the amplifier will be linear and that the power output, efficiency, and plate dissipation will be correct after the first adjustments are somewhat remote.

6. DOWNWARD MODULATION. When the carrier wave of a transmitter is 100% modulated, the power increase in the antenna should be 50%; that is, the modulated wave contains 1½ times as much power as the unmodulated carrier. Also, the effective value of the antenna current should increase 22.5%. This is often used as a test of 100% modulation, but should be used only in conjunction with an oscilloscope, because it is possible for the antenna current to increase 22.5%, but the modulation envelope is so distorted that overmodulation occurs. Let us examine Fig. 10, which will make

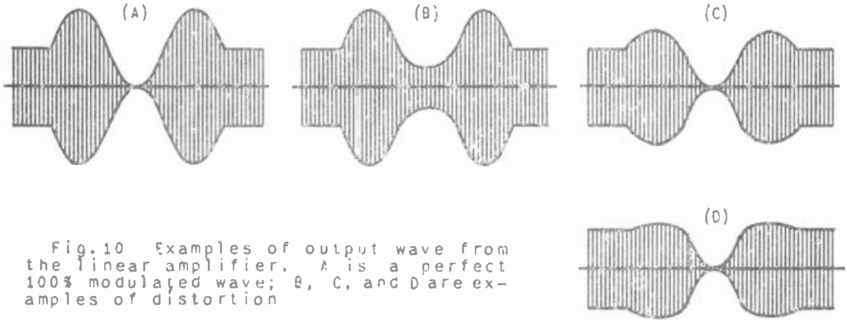


Fig.10 Examples of output wave from the linear amplifier. A is a perfect 100% modulated wave; B, C, and D are examples of distortion

this clear. At A is shown an undistorted 100% modulated carrier, and an antenna current of this wave form would show an increase of 22.5% above the unmodulated value. Suppose, however, that the Class B amplifier is not correctly adjusted for linear operating conditions. Perhaps the amplitudes of the R.F. cycles increase above the carrier level more than they decrease below it. In this case, the modulated wave would have the form shown at B in this figure. It is evident that the effective value of such a wave would be greater than the undistorted wave shown at A, and it is probable that the increase in antenna current would be greater than 22.5% when the correct amount of audio power to produce 100% modulation was obtained from the modulator.

On the other hand, it would be possible for the linear amplifier to be so adjusted that the tank current or antenna current had the form shown at C in this same figure. This wave form indicates that the increase in the amplitudes of the R.F. cycles above the carrier level are not as great as the decreases below it. If the grid-exciting voltage is too great, the peak R.F. tank voltage will be too high and it will be impossible for the tank voltage to double its value during the peaks of 100% modulation, thereby, causing the positive alternations of the modulation envelope to be clipped. If the positive alternations are only slightly clipped, the antenna current will increase during modulation, but will not increase 22.5%, even when the audio voltage is sufficient to cause the carrier wave to fall to zero during the troughs of the modulation cycle. If the misadjustment of the linear amplifier is serious, it may happen that the effective value of the modulated wave will be less than the effective value of the unmodulated carrier. A wave form of this type is illustrated at D in Fig. 10. It is seen that the amplitudes of the R.F. cycles increase but little above the carrier level, although they decrease to zero during the troughs. If the wave form of the antenna current is like this, the antenna meter will show a decreased antenna current during modulation, and will continue to decrease as the percentage of modulation is increased.

This phenomenon is called "downward modulation", and as may be seen, it results from improper adjustment of the linear amplifier, and sometimes, of the Class C stage. In fact, anything which would cause the Class B stage to depart from linear operating conditions may be the cause of downward modulation.

Perhaps the major cause of downward modulation is excessive grid excitation of the linear amplifier. This may be reduced by loosening the coupling between the Class C plate tank and the grid tank of the linear, or by increasing the load on the Class C stage by connecting the swamping resistor across more turns of the tank coil. Should it happen that the load on the linear amplifier is too small, it is probable that the R.F. tank voltage is excessive and will not be able to double its value at the peaks of the modulation cycle; this is also a cause of downward modulation. Therefore, if the downward modulation persists even when the grid-exciting voltage has been reduced, the LC ratio of the tank circuit should be reduced or the amount of coupling between the plate tank circuit of the linear and the antenna should be increased. Either action will lower the shunt impedance of the tank circuit and cause the R.F. voltages developed across it to be lower. If the efficiency at the carrier level is too great, it will be impossible for the efficiency to double its value during the peaks of 100% modulation.

A second cause of downward modulation is imperfect neutralization. As stated previously, it is possible for the amplifier to be sufficiently neutralized so that it will not oscillate when the grid-exciting voltage is at the carrier level, but does oscillate at the peaks of the modulation cycle. This process absorbs power from the tank circuit and prevents the tank current from increasing enough to preserve linearity.

It is also possible for the linear amplifier to oscillate at frequencies other than the fundamental frequency of the grid-exciting voltage. These oscillations are called "parasitic oscillations" and may occur in any stage, but are more often found in linear amplifiers. The following lesson will be devoted to a thorough discussion of "parasitics"; for the present, it is sufficient to know that they often exist, and must be eliminated. The frequency of these spurious oscillations may be anything from a low audio frequency to an ultra-high radio frequency. Sometimes they are caused by tank circuits formed by R.F. chokes and R.F. by-pass condensers. The various methods used to prevent the occurrence of parasitic oscillations will be described in the following lesson. Parasitic oscillations are a third source of downward modulation; they cannot be eliminated by neutralization, and since they draw power from the output circuit of the linear stage, they prevent the tank current from increasing in a linear manner.

As previously stated, in most cases the first adjustment will not be satisfactory. To enable you to know what to do, the following summary is included.

1. Amplifier modulates down - Lower excitation by connecting swamping resistor across more turns of the tank coil. Lower load impedance by coupling antenna closer to plate tank circuit.
2. Amplifier modulates upward, but less than 100% - Same procedure as for downward modulation.
3. Modulates properly but plate dissipation is excessive - Reduce excitation by increasing load on Class C stage. Increase load impedance by reducing antenna coupling.
4. Not enough power output - Increase excitation by coupling grid tank closer to the plate tank of Class C stage, and/or reducing load on Class C stage. Increase antenna coupling.

Before closing this section, it would, perhaps, be well to give further consideration to the problem of neutralization. Improper neutralization is a very common cause of downward modulation, and non-linear operating conditions. Ordinarily, it is not extremely difficult to neutralize the feed back through the interelectrode capacity of the tube. Knowing the proper procedure, and exercising a little care in this process will enable anyone to neutralize the feed back due to this source. After this is accomplished, there still remains the electromagnetic coupling between the grid and plate coils of the linear amplifier. The feed back caused by this type of coupling cannot be eliminated or balanced out with the neutralizing condenser. To eliminate coupling between these two tank circuits, they should be placed as far from each other as is practically possible, the two coils should be at right angles to each other, and the Class C stage should be separately shielded from the linear stage, which should itself be well shielded. The grid tank coil of the linear stage must, of course, be placed within the shield containing the Class C amplifier so that R.F. energy may be fed from the Class C stage to the linear. Usually the antenna coupling coil is placed within the shield holding the linear amplifier. To be effective, both of these shields must be well grounded, and since it is usually necessary to build the shield of several separate pieces of metal, care must be taken to make sure that each piece makes a good electrical connection with the other pieces with which it is in contact. The idea is to make the overall conductivity of the shield as high as possible so that its shielding effect will be most efficient.

7. THE PUSH-PULL LINEAR AMPLIFIER. Any type of R.F. amplifier may be operated as a push-pull stage. This statement applies whether the stage is operating as an oscillator, a buffer, a modulated Class C amplifier or a linear stage. There are several advantages of the push - pull arrangement. The first of these is the cancellation of all even harmonics, and the second is the fact that a larger amount of power may be obtained without resorting to the use of larger, more expensive tubes requiring a higher plate voltage and consequently a more expensive power supply. Self-excited oscillators connected in push-pull have been discussed previously, and the use of push-pull stages for the Class C modulated amplifier and the Class B linear amplifier in modern broadcast transmitters is very common. In amateur work, the crystal oscillator is sometimes a push - pull stage, and many communications transmitters for the ultra-high frequencies use the push-pull arrangement in every stage.

The reason that the push - pull connection eliminates the even harmonics is that the plate currents of the two tubes are 180° out of phase. Assume that the stage under consideration is a Class B linear with an operating angle of 180°. (This means that the plate current of each tube flows for one-half cycle.) In a single-ended stage, the plate tank circuit would receive energy only during the time that plate current flows, or just one - half of the time. Due to the damping effect produced by the resistance of the tank circuit and the resistance which the load circuit couples into it, the amplitude of the oscillations tend to decrease during the times that the tank is not receiving energy, and the negative alternation

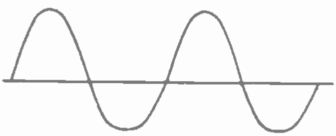


Fig. 11 Illustrating how RF harmonics may be produced in single ended stages.

of the tank current has a slightly smaller amplitude than that of the positive alternation. This tends to produce the somewhat distorted wave shown in exaggerated form in Fig. 11. As we have already learned in preceding lessons, when the two alternations of a wave form are not symmetrical, the wave contains even harmonics. The greater the energy contained in the tank circuit, or the larger the oscillating tank current, the smaller will be the percentage of these even harmonics.

With the push - pull arrangement, the tank circuit receives a pulse of energy during every alternation instead of just once per cycle, and the negative alternation of the tank current has the same amplitude as the positive alternation. Thus, the wave form produced is symmetrical, and can contain no even harmonic components.

The push-pull stage does nothing toward the elimination of third, or odd harmonics, but these are not ordinarily as troublesome as is the second harmonic. Since the Federal Communications Commission has a rather strict ruling regarding the radiation of harmonics, nearly all final stages of broadcast transmitters use the push-pull arrangement.

The two tubes of a push-pull stage must work into the same load impedance and instead of using two plate tank circuits, one tank tapped at its center is employed. The total impedance of the tank must be four times that which would be used for a single tube. When the tank is split in this manner, the plate voltage is fed into the center of the tank and this point is grounded with respect to R.F. with an R.F. by-pass condenser. Thus, considering Fig. 12,

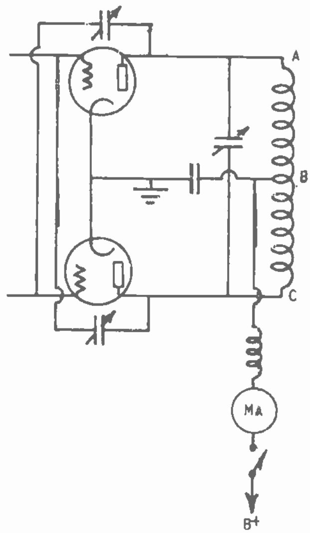


Fig. 12 Showing how neutralization of push-pull stages is accomplished.

the top tube works into the top half of the tank between points A and B, and the neutralizing voltage for this tube is that produced across the bottom half of the tank from points B to C. The load of the bottom tube is the lower half of the tank, and the neutralizing voltage for this tube is the voltage developed across the top part of the tank circuit. Thus, neutralization of a push-pull stage is comparatively simple, and the neutralizing condensers are connected from the grids of the tubes to the plates of the opposite tubes.

A push-pull, modulated Class C stage inductively coupled to a push-pull Class B linear is illustrated in Fig. 13. Note that the grid tank coil of the linear amplifier is split into two equal parts and that each part is coupled to one end of the plate tank of the Class C stage. Often these two coils are mounted on a rod forming the axis of the Class C tank coil, and they may be moved closer to or farther from the Class C tank to vary the coupling and the excitation to the linear stage. There are two fixed condensers connected in series with the neutralizing condensers. This is often done in high-powered stages, because the high voltage which is across the neutralizing condensers is liable to break them down,

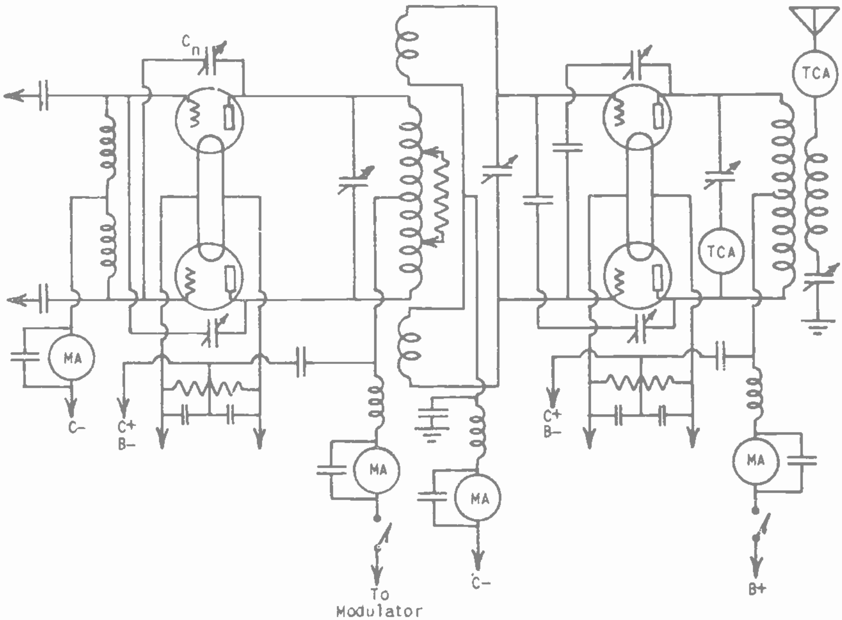


Fig. 13 A push-pull Class C stage coupled to a push-pull linear amplifier.

unless special expensive, high - voltage condensers are used. By connecting these fixed condensers in series with the neutralizing condensers, the voltage is divided, and the breakdown rating of the neutralizing condensers need not be as large.

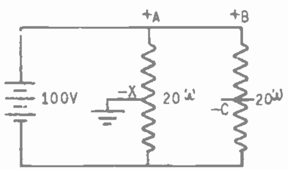
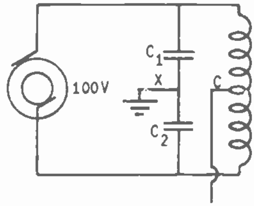


Fig. 14 Illustrating how a point may be at ground potential although it is not connected directly to ground.

Although Fig. 13 shows the center of the tank coils grounded with respect to R.F., it is more common to use split - stator tuning condensers with the rotor grounded. This method seems to give slightly better results, and makes neutralization a little easier. When the rotor of a split-stator condenser is grounded, the center point of the tank coil is also at an R.F. ground potential. Sometimes it is a little hard for the student to see why this is so; perhaps the following explanation will make this clear. In Fig.

14 are shown two resistors connected in parallel and placed across a battery. The center tap of one of the resistors is grounded. Suppose that the voltage of the battery is 100 volts, and the value

Fig. 15 Showing how grounding the rotor of a split-stator tuning condenser causes the center of the tank coil to be at an RF ground.



of each resistor is 20 ohms. Thus each resistor will draw 5 amperes from the battery, and the voltage drop across the top half of the grounded resistor is 5 × 10, or 50 volts. Therefore, point A is 50 volts positive with respect to point X or ground. In a like manner, there will be a 50 volt drop across the top half of the second resistor, or point B will be 50 volts positive with respect to point C. Points A and B are at the same potential; point C is 50 volts negative with respect to the positive side of the battery, and point X or ground is also 50 volts negative with respect to the positive terminal of the battery. Therefore, points C and X are at the same potential, or in effect, point C is grounded.

Now consider Fig. 15. This figure shows a coil and two series-connected condensers placed across an alternator. If C_1 has the same capacity as C_2, then the voltage drop across each of the condensers is equal, and if the voltage of the alternator is 100 volts, there will be a 50-volt drop across each condenser. Likewise, if the coil is tapped at its exact center, there will be a 50-volt drop across each half of the coil, and point C will be at the same potential as point X, the center tap of the condenser branch of the circuit.

In a split-stator tank circuit, the rotor is grounded; there is the same voltage between each of the stators and ground; and if the tank coil is center-tapped, its mid-point will also be at ground potential.

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POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

PARASITIC OSCILLATION

LESSON NO. 11

by ... -frequency choke and an R.F. ... will form a tank circuit resonant at some low radio-frequency. Should it happen that similar tank circuits exist in both the plate and grid circuits of the stage, it is quite possible that the amplifier will act as a self-excited TGTP oscillator, producing power at the spurious frequency to which these unintended tank circuits are resonant. This action absorbs power from the amplifier and prevents the attainment of the best operating conditions.

A "TAYLOR MADE"
.....Director of Radio.

About seven years ago a young man came to Jerry Taylor for his training in radio. This young man ... mind that he was going to ... radio. ...

tifier tube from the grid to the filament. The plate of the rectifier tube is connected to the grid of the amplifier and the filament of the rectifier tube is connected to the filament of the amplifier. Such an arrangement is shown in Fig. 3. When the grid goes positive, current will flow through the rectifier tube and it, together with the resistance R_1 which is connected in series, provides an additional load in the grid circuit. Thus, the grid circuit has additional positive resistance to over-balance the negative resistance of the input of the amplifier tube at this time. If the total positive resistance in the grid circuit is greater than the negative resistance of the tube's input, dynatronic oscillations cannot occur.

It should be realized that these dynatronic oscillations are not limited to the low audio frequency to which the grid filter system is resonant. It is also possible for the stage to oscillate at the resonant frequency of the tank circuit L_3, C_5, shown in Fig. 1. Since the resonant frequency of this circuit is the fundamental frequency, it is possible for the stage to oscillate at the fundamental frequency, even when neutralization of the feedback capacity is complete.

3. LOW RADIO FREQUENCIES. The next type of parasitic oscillations which we shall consider have frequencies ranging from one-third to one-fifth of the fundamental frequency. The tank circuits are formed by R.F. chokes and by-pass condensers. Thus, in Fig. 4, the R.F. choke L_1, and the by-pass condensers C_1 and C_2

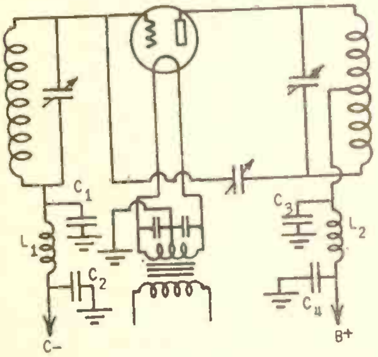


Fig. 4 A circuit which may produce parasitic oscillations due to resonating of RF chokes and RF by-pass condensers.

form the grid tank circuit, whereas L_2, C_3, and C_4 constitute the plate tank. To the low-frequency oscillations, the stage acts as a TGTP oscillator. Fig. 5 shows the same stage redrawn to illustrate that this is the case. The inductance of the grid and plate coils L_3 and L_4 is not great enough to interfere with these low-frequency oscillations, and the neutralizing condenser merely provides another path for feedback from plate to grid circuit to maintain the oscillations.

Naturally, these parasitic oscillations will not exist unless the resonant frequency of the accidental grid tank circuit is the same as that of the plate circuit. Such a condition is liable to prevail if the same size chokes and by-pass condensers are used in both the grid and plate circuits. The two general methods of eliminating any type of oscillation consist of detuning and damping. In this case, it is easier to detune one of these accidental tank circuits by changing the size of one of the R.F. chokes. With this type of oscillation, the large R.F. current that circulates through the by-pass condensers is very liable to ruin them.

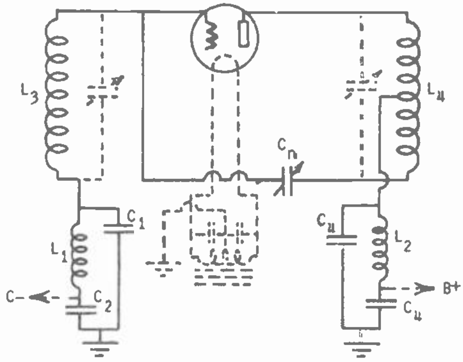


Fig.5 Showing the parasitic tank circuits of the amplifier illustrated in Fig. 4. The solid lines indicate the path of the parasitic oscillations.

It has been determined that the best method of stopping oscillations of this type is to tune the accidental grid tank circuit to a higher frequency than the corresponding plate tank. This particular range of parasitics is one that is often overlooked, and is liable to cause considerable trouble.

Perhaps you are wondering just how the presence of parasitic oscillations may be detected. The usual method is to bias the tube to cut-off, apply normal plate voltage and no excitation. If there are no parasitic oscillations being produced, there will be no plate current flow, because of the cut-off bias. Parasitics, however, will be manifested by a rather large plate current; a neon tube will glow when touched to the plate of the tube; and tuning the plate tank circuit will have but relatively little effect upon the oscillations. Naturally, if the tube is self-oscillating at the fundamental frequency, the same symptoms will be noted. Therefore, make sure that the stage is well neutralized, and use a wave meter to determine whether the frequency of the oscillations being generated are of the fundamental frequency or of some parasitic frequency. Also, if there are parasitics, it is not ordinarily possible to obtain a plate current dip as the plate tank circuit is tuned to the resonant frequency with fundamental excitation voltage applied. A neon tube will indicate that the whole of the plate tank circuit is "hot"; the point usually cold, since it is by-passed to ground, is not at an R.F. ground potential. Sometimes it will be necessary to apply some fundamental excitation voltage to start the parasitics, after which they will continue when the excitation voltage is removed.

Perhaps the greatest difficulty in detecting the condition causing parasitic oscillations is the fact that they are not always sustained. They may be intermittent in character; appearing at irregular intervals. All transmitters are protected by overload relays, and when a parasitic oscillation appears, it is most liable to trip the overload relay, throwing the transmitter off the air. Perhaps the spurious oscillation will again return when the relay is again set, or it may not reappear for several days. Conditions such as these are particularly annoying, and very difficult to eliminate. Especially bad are the types of parasitics which affect the quality or fidelity of the transmitter. The fact that any lead or circuit component in a stage may be part of a spurious tank circuit, makes the finding of these accidental tanks quite a problem. There is no known mathematical procedure for determining what circuit components should be employed to insure that parasitic oscillations will not be generated.

4. HIGH-FREQUENCY OSCILLATIONS. Perhaps the most troublesome type of parasitic oscillations are those having frequencies many times the fundamental. Usually the tank coils consist merely of the inductance of the grid and plate leads, whereas the tank capacities may be the interelectrode capacities of the tube. When a high-powered tube is used as a single-ended R.F. amplifier, this type of oscillation often results.

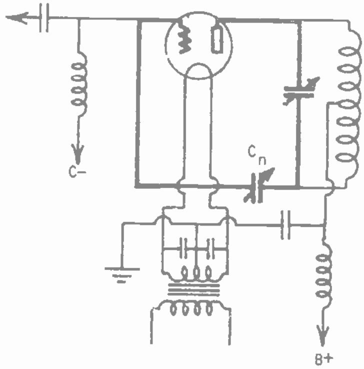


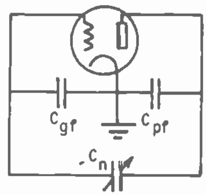
Fig.6 A plate-neutralized, single-ended stage which may have parasitic oscillations, The parasitic tank is drawn with heavy lines

Fig. 6 shows an apparently innocent-looking, plate neutralized R.F. amplifier. At first sight, it seems improbable that this amplifier would be the source of parasitics; yet when the stage had been neutralized, and the bias and plate voltages were applied, the plate current rose to an abnormally high value even though there was no exciting voltage and the bias was sufficient to produce plate current cut-off. A neon tube indicated that the plate end of the tank circuit was hot, and tuning the stage had practically no effect on the plate current. By resorting to a wave meter, it was discovered that the oscillations were of an ultra-high frequency character. This being the case, it was concluded that the tank coil could not be part of the spurious tank circuit, since it would have a very high inductive reactance at this ultra-high frequency.

Thus, it was evident that the tank coil was acting merely as an R.F. choke to these oscillations. This was proved by removing the tank coil and feeding the plate voltage to the tube through an R.F. choke; the parasitic oscillations continued at the same intensity.

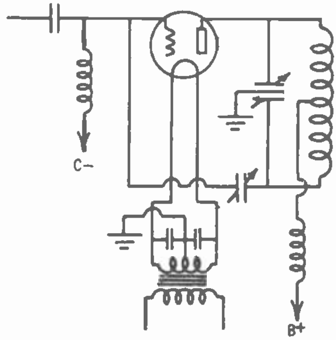
The oscillating tank circuit in this case is that part of Fig. 6 which is drawn with heavy lines. At the ultra-high frequency of these oscillations, the tank tuning condenser has a very low reactance and acts practically as a short circuit to the parasitic

Fig.7 A simplified drawing of the parasitic tank circuit of Fig. 6.



oscillations. Perhaps it is not clear just what composes this tank circuit. Let us refer to Fig. 7 which shows the separate parts of this spurious tank. The capacity consists of the neutralizing condenser, the grid-to-filament interelectrode capacity and the capacity between the plate and the filament. The inductance of the tank consists of the inductance of the grid and plate leads. Since both the capacity and the inductance in this circuit are very low, the parasitic oscillations will have a very high frequency. It is seen that with the circuit redrawn in this manner, it is practically the same as a Colpitts oscillator.

Fig.8 Illustrating how one type of parasitic oscillations may be eliminated by using a split-stator tuning condenser.



The simplest method of eliminating this type of parasitic is to use a split-stator tuning condenser. The same stage with this addition is illustrated in Fig. 8. With this arrangement, the impedance between the plate and ground is largely dependent on the reactance of one section of the split-stator condenser. At the higher frequencies, each section will have a very low reactance, and the plate will practically be at an R.F. ground at ultra-high frequencies. Therefore, the parasitic tank circuit will be shorted to ground, and parasitic oscillations are not liable to occur.

Capacitive coupling between stages sometimes results in the production of parasitic oscillations. Another case, in which the spurious tank circuits are not immediately apparent is the one shown in Fig. 9. The circuit is, in effect, a TGTP oscillator; the heavy lines indicate the parasitic tanks. The circuit redrawn

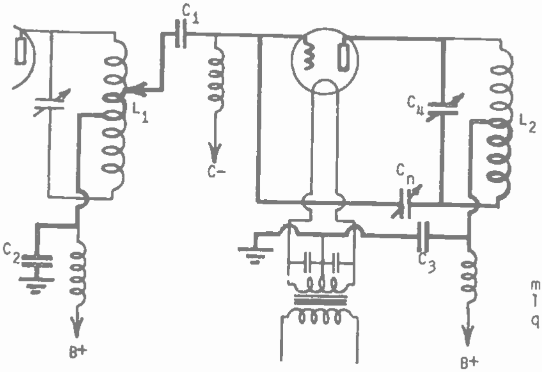


Fig. 9 A circuit which may act as a TGTP oscillator to the parasitic frequency.

to show the tanks more clearly is given in Fig. 10. The tuning condensers consist of the interelectrode capacities and the stray capacities due to the wiring. It is seen that the neutralizing condenser is furnishing feedback. In this case, the condensers C_1, C_2, C_3 and C_4 have such low reactances at the parasitic frequency that they constitute R.F. shorts, and are not included in the simplified diagram.

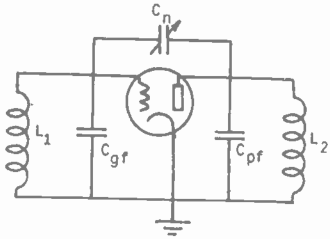
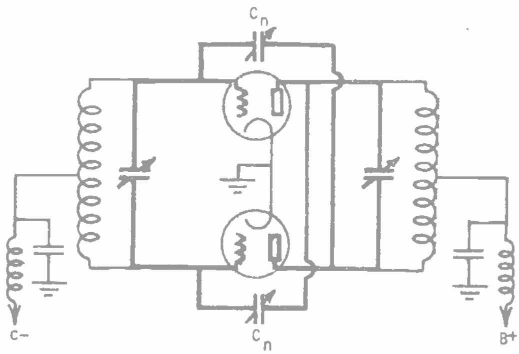


Fig. 10 The circuit of Fig. 9 redrawn to show the parasitic tank circuits.

About the only method of eliminating this type of oscillation is to move the tap on L_1 to the plate end of the coil and to change the sizes of some of the by-pass condensers.

Parasitic oscillations are especially apt to occur in push-pull amplifier stages such as the one illustrated in Fig. 11. The frequency of these oscillations is very high, being determined by the inductance of the grid and plate leads and the stray capacities including the interelectrode capacities of the tube. At the high frequencies, the tuning condensers of the grid and plate circuits act as by-pass condensers, and the circuit performs as a TGTP oscillator as shown in Fig. 12. It might be thought that the neutralizing condensers would prevent sufficient feedback from occurring to maintain sustained oscillations; however, at the frequency of the parasitic oscillations the inductive reactance of the leads to these condensers is considerable, and a phase shift occurs. The currents fed from the plate circuit to the grid circuit would normally be out of phase with the R.F. fed through the interelectrode capacities, and neutralization would take place. But with this

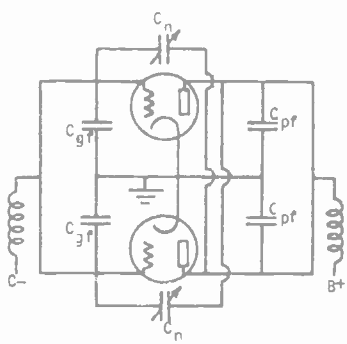
Fig.11 A push-pull stage containing parasitic tank circuits indicated by the heavy lines.



phase shift, produced by the inductance of the leads, the current fed through the neutralizing condensers is not out of phase by the time it reaches the grid circuit, and regeneration instead of neutralization results.

The best remedy for this sort of trouble is to use a split-stator tuning condenser in either the grid or plate circuit, making the lead from the grounded rotor to the filament as short as possible.

Fig.12 A simplified drawing of Fig. 11, showing the parasitic tanks.



When tubes are operated in parallel, an ultra-high frequency oscillation often results. Consider Fig. 13 which illustrates two tubes connected in parallel. The spurious tank circuits are indicated by the heavy lines. They consist of the inductance of the grid and plate leads, the interelectrode capacities of the tube, and the stray capacities of the associated circuit. It is evident that this circuit is acting as a TGTP push-pull oscillator, with the normal grid and plate coils serving as R.F. chokes at the parasitic frequency. This type of oscillation may be detected by noting the point of low R.F. potential at the mid-point of the connecting leads with a neon bulb. These are the points marked A and B in Fig. 13. The stopping of this type of parasitic is comparatively

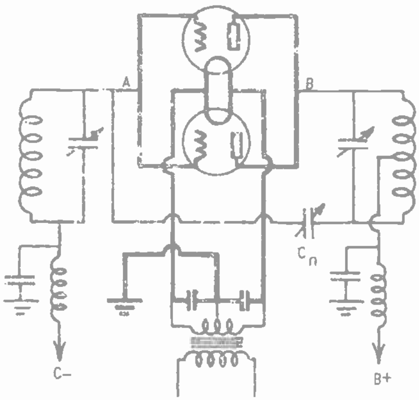


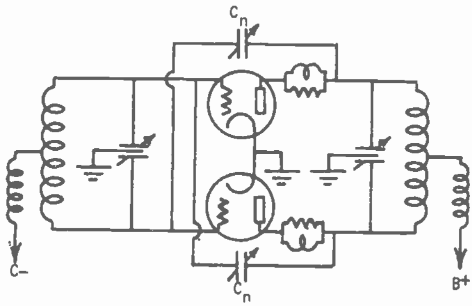
Fig.13 The parasitic tank circuits of two tubes operated in parallel.

simple. One method is to unbalance the circuit slightly by making the grid leads considerably shorter than the plate leads. This unsymmetrical wiring tends to make the resonant frequency of the spurious grid tank higher than that of the corresponding plate tank. Sometimes it is necessary to connect a small choke of 2 to 4 turns with a diameter of ½" in the plate or grid leads to one tube and not to the other.

The parasitic oscillations occurring in linear amplifiers seem to be more difficult to eliminate than those appearing in other types of amplifiers. Often when nothing else is effective, it is common practice to connect small resistors of from 25 to 100 ohms in each plate lead. This added resistance serves to dampen the parasitic oscillations, and makes the resistance of the spurious tanks so great that oscillations cannot be maintained. The only disadvantage of this method is the fact that the plate current must flow through the parasitic resistors and power is wasted making the amplifier harder to drive. To choke back the parasitic oscillations without interfering seriously with the normal flow of plate current, small parasitic chokes are connected in parallel with the resistors. The chokes consist of several turns of wire wound around the resistor. With this arrangement, the normal plate current is by-passed through the chokes which offer practically no opposition to the fundamental frequency. On the other hand, these same chokes offer a high opposition to the parasitic frequency and the parasitic oscillations take the easier path through the resistors. Naturally, the resistors must be absolutely non-inductive, or the damping action of the resistors will be reduced. Perhaps the best type of non-inductive low-valued resistor is one formed of a carbon rod, and this type is nearly always used. A push-pull linear amplifier with these additions is illustrated in Fig. 14.

The highest band of frequencies at which parasitic oscillations may occur is several hundred megacycles. Such unusually high frequencies require tank circuits of very little inductance and capacity. It is thought that these tanks are formed by the internal leads to the grid and plate electrodes of the tubes and the capacities between them. This is especially true in very large tubes where these leads are necessarily long. Oscillations of this band will be revealed by the transmitter kicking off the air by the tripping of the overload relay, although apparently the functioning of the tubes and circuits is correct.

Fig. 14 Showing the use of parasitic resistors and chokes.



Nearly all the parasitic oscillations of an intermittent character are produced by the erratic grid characteristics of the tubes used. These oscillations are very rapid in their action, and may be distinguished from others that overload the power supply, by the fact that they act as if a direct short circuit were instantaneously connected between the plate and filament of the tube. Ordinarily no external effects are apparent.

5. PARASITIC OSCILLATIONS IN AUDIO STAGES. Some audio amplifiers, especially Class B stages, are subject to parasitic oscillations. Usually, these may be detected by connecting the output of the amplifier to the vertical plates of a cathode ray oscilloscope. Often the parasitics occur only during the peaks of the audio grid-exciting voltage and produce a trace on the screen of the oscilloscope similar to the one shown in Fig. 15. These parasitics will have frequencies ranging from 20 to 100 kc. and are caused by the leakage inductance resonating with the stray capacities of the circuit. The first step in the elimination of this parasitic is to connect small condensers from each grid to the filament or perhaps from each plate to the filament. If they still persist, the next step is to connect small damping resistors in the plate leads of the tubes. These resistors should have values from 20 to 100 ohms, and should be connected directly to the plate terminals of the tubes. The final step in their elimination is to connect resistors of from 5000 to 50,000 ohms from grid to grid or from plate to plate.

This type of parasitic oscillation is usually found in Class B modulators and is often present when no signal is being fed to the modulator. The ultra-audible parasitics which are generated will then modulate the carrier wave of the transmitter and will cause side bands spread from 20 to 100 kc. each side of the carrier frequency. If, when testing for parasitics, it is found that

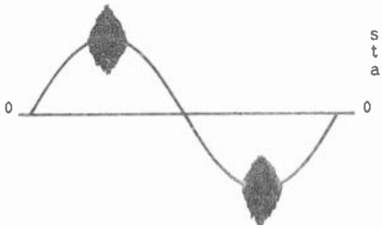


Fig.15 Illustrating how parasitic oscillations may occur at the peaks of the audio signal in a Class B audio amplifier.

they exist the same number of kilocycles above the carrier as below it, it is more than likely that they are being generated in the audio stages.

PART 2. HARMONIC RADIATION

6. HOW HARMONICS ARE PRODUCED. In Lesson 26 of Unit 1, it was shown that any wave differing from a true sine wave could be considered as being composed of a pure sine wave fundamental frequency and many sine wave harmonic frequencies of greater or lesser amplitude and of various phase. Thus, when the output of a vacuum tube amplifier has a wave form differing in any respect from that of the input, the amplifier has introduced harmonic frequencies.

Nearly all previous discussion of harmonic frequencies has been concerned with distortion in audio amplifiers. At present, however, we are more interested in the generation and elimination of R.F. harmonics. If the current flowing in the transmitting antenna has a pure sine wave form, then the transmitter is harmonic free. Naturally the amplitude of the R.F. cycles of the antenna current will vary with the audio modulation, but unless each R.F. cycle itself is a pure sine wave, there will be harmonic radiation.

Harmonic radiation is especially undesirable since it causes interference to other stations. A station operating on an assigned frequency of 1000 kc. would cause interference to stations having frequencies of 2000, 3000, 4000 kc., etc. For this reason, the Federal Communications Commission have caused rigid rules concerning harmonic radiation to be enacted. To prevent the radiation of harmonics, two methods are employed. The first is to prevent their production as far as possible; and the second is to use trap circuits to keep those which are generated from reaching the antenna.

Harmonic generation in radio-frequency power amplifiers is caused by the mode of operation. To achieve efficiency, power amplifiers are worked either under Class B or Class C conditions. In either case, the plate current is distorted; that is, it does not have a pure sine wave form. If the mode of operation is Class B, the plate current flows as half-sine wave pulses, and with Class

C operation, the plate current wave form consists of non-sinusoidal pulses lasting from 60° to 120°. Thus, the plate current of either type of amplifier is very distorted and is rich in R.F. harmonics. In fact, the only reason it is possible to operate the amplifier in this manner and allow the plate current to become so distorted, is that the load is a tuned tank circuit. By using a tank circuit, it is possible to smooth out these pulses into a wave form which is not essentially different from a pure sine wave. The tank circuit is similar to a reservoir of energy from which power is taken and power is added. If the tank circuit had absolutely no resistance, the oscillating tank current would be undamped and each alternation would have the same amplitude. With this condition, the tank current would be truly sinusoidal.

Such a situation is only theoretical, for all tank circuits have some resistance which includes the inherent resistance of the tank as well as that resistance which the load circuit couples into the tank. This resistance tends to cause the oscillating current to be damped, and it is the influx of energy produced by the plate current pulses which overcomes this tendency toward damping. If the energy were supplied to the tank circuit continuously, the damping would be cancelled completely and the oscillating current would be a pure sine wave. In the usual single-ended amplifier, however, the energy is supplied to the tank only during a part of one alternation. Between the times that the energy is supplied to the tank, a slight amount of damping occurs. This causes the negative alternation of the oscillating tank current to have a slightly smaller amplitude than the positive alternation and the wave form is not perfectly sinusoidal. Thus, a single-ended Class B or C amplifier will always generate some harmonics.

The amount of harmonics generated depends on the ratio of the energy stored in the tank to the amount lost per cycle. Unless twice as much energy is stored in the charged tuning condenser during each cycle as is lost to the load circuit in this interval, the per cent of harmonics will be excessive. As explained in Lesson 7 of this Unit, this ratio is determined by the Q of the tank circuit or upon the L/C ratio. The Q of the loaded tank circuit should be at least 12. The amount of inductance and capacity which the tank circuit should contain to produce a Q of this value will depend on the frequency, the applied voltage, and the expected power output; the formulas given in Lesson 7 for the correct amount of inductance and capacity will produce a circuit Q of approximately 12.

When the Q of the loaded tank circuit is comparatively high, there is a large oscillating tank current flowing, the energy stored is large, and the harmonics generated are at a minimum. Too large an L/C ratio will cause the tank current to be low, and the tank will not be able to perform its job of smoothing out the plate current pulses as well, thereby producing an output wave with a high harmonic content. Thus, it is seen that the first prerequisite for low harmonic production is a comparatively large oscillating tank current produced by using a large tank capacity.

When two tubes are operated in push-pull, a slightly different condition exists. The plate currents of the two tubes are 180° out of phase and the tank circuit receives a pulse of energy once during each alternation or twice as often as in the case of the single-ended amplifier. This causes the damping to be reduced, and the negative alternation of the tank current has the same amplitude as the positive alternation. Therefore, the output wave cannot contain any even harmonics. This is a decided advantage, and nearly all of the final stages of transmitters use the push-pull arrangement.

The use of split-stator tuning condensers in both grid and plate circuits is of great help in reducing harmonics. The harmonics are, of course, of higher frequency than the fundamental, and since the rotor of the tuning condenser is grounded, the capacitive reactance between the plate and ground becomes less the higher the frequency. Thus, the harmonics tend to be by-passed to ground. When the ordinary type of tuning condenser is employed, the center of the tank coil is by-passed to ground by an R.F. by-pass condenser. In this case, however, the reactance between the plate and ground is determined by the inductive reactance of one-half of the tank coil. As this inductive reactance becomes larger the higher the frequency, there is no easy path for the harmonics to take in their route to ground, and so they develop appreciable voltages across the tank coil.

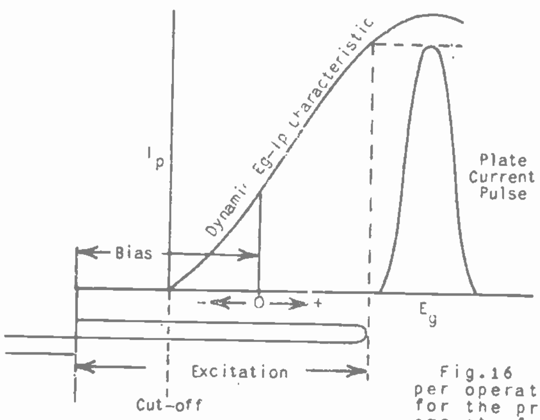


Fig.16 Illustrating the proper operation of a Class C stage for the production of a minimum amount of harmonics.

Capacitive coupling tends to allow more harmonics to be transferred from one stage of the transmitter to another than does inductive coupling. At the harmonic frequencies, the coupling condensers have lower capacitive reactances and offer less opposition. On the other hand, inductive coupling suppresses the harmonics to some extent. The inductances have greater reactances at the harmonic frequencies and therefore offer more opposition to them.

No stage of the transmitter should be over-excited, because this also produces excessive harmonic generation. With proper excitation, the plate current pulse of a Class C stage does not depart materially from the shape of the upper portion of a sine wave. Such a pulse contains practically no odd harmonics, and if the stage is push-pull, a large amount of the even harmonic content will be balanced out. Over-excitation, however, causes the operating path to extend into the upper curved portion of the dynamic E_g-I_p characteristic, and the plate current pulse becomes very distorted. Fig. 16 shows how the excitation should be adjusted so that the odd harmonics will not be formed. It is seen that the excitation is just sufficient to drive the operating path up to the beginning of the upper curved portion of the characteristic.

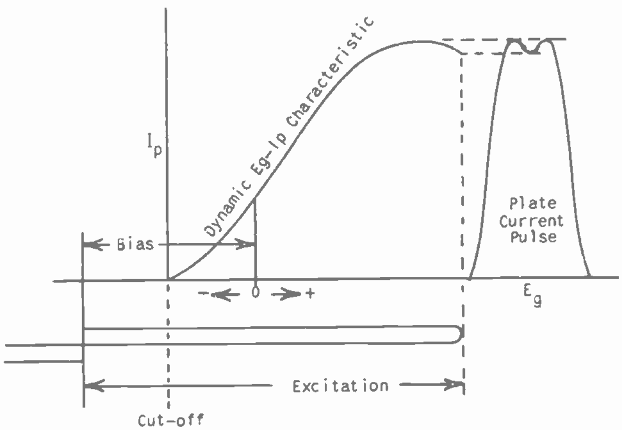


Fig.17 Showing how over-excitation of a Class C stage distorts the plate current pulses which produces third and odd harmonics.

One method of determining when the grid excitation is of the proper value is to increase it slightly. If the increased grid excitation produces an increase in the DC plate current, the excitation is not excessive; however, should it happen that the plate current decreases with increased excitation, it is evident that too much excitation is being used. The excitation should be increased just to the point where further increases do not produce noticeable increases in the DC plate current. This is the saturation point of the tube and is the point at which most Class C amplifiers are worked.

When the stage is over-excited, the plate current pulse assumes the form shown in Fig. 17. This pulse contains many odd harmonic components which will not be cancelled by a push-pull stage.

7. THE SUPPRESSION OF HARMONICS. Now that we have seen how the generation of harmonic frequencies may be reduced, we must turn our attention to the methods employed to prevent the radiation of those which cannot be eliminated. Fig. 18 shows a push-

pull R.F. amplifier which is the final stage of a transmitter. As explained in the preceding section of this lesson, the second and even harmonic frequencies are cancelled with this type of connection. The fundamental component of the plate currents of the tubes are 180° out of phase and, as is usual with push-pull amplifiers of all kinds, they add to make the total power output at the fundamental frequency greater than that which one tube alone could supply. The

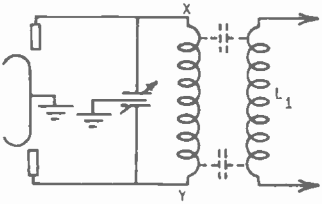


Fig.18 Illustrating how the second harmonics in the load of a push-pull stage may be transferred to the antenna by capacity coupling.

second harmonic components of the two currents, however, are in phase, and are cancelled in the load circuit. Since the second harmonic components are in phase, this means that points X and Y of the tank circuit will be at the same R.F. potential so far as the second harmonics are concerned.

L_1 is the antenna coupling coil; it may couple the final tank either directly to the antenna or to an R.F. transmission line, the other end of which is coupled to the antenna. Since L_1 is fairly close to the final tank coil, it is possible for the second harmonic frequencies to be transferred to this coupling coil by means of the capacity existing between the tank coil and the coupler. If this occurs, much of the desirable feature of the push-pull stage in eliminating the second harmonic frequencies will be lost. To remedy this situation, it is necessary that all R.F. energy which is transferred from the final tank circuit to the coupling coil, be through the medium of the mutual inductance existing between them, and that none of it be transferred by capacitive effect. To

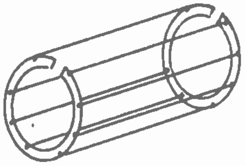


Fig.19 An electrostatic shield used to reduce the capacity between the tank coil of a final stage and the antenna-coupling coil.

move the coupler farther from the tank would reduce the capacitive effect, but would also cause the mutual inductance between the two coils to be lowered. A method by which the capacity between the tank coil and the coupler may be reduced to a very low value without affecting the mutual inductance between the two coils is to place an electrostatic shield between them. This shield follows the same principle as the screen grid in a tetrode vacuum tube.

Usually it consists of a number of parallel rod conductors arranged in the form of a hollow cylinder. The ends of these conductors fit into end rings which are also conductors and aid in the shielding effect. These end rings must be provided with an air gap so that there will be no continuous current path through them. Were it not for the air gaps, current would circulate in these end rings, causing a power loss, and, in addition, the device would act as an electromagnetic shield as well as an electrostatic shield. The general construction of such a shield is shown in Fig. 19. Ordinarily

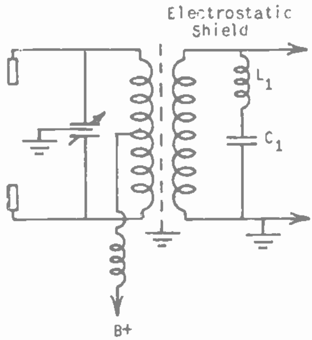


Fig. 20 A series tuned circuit (L_1C_1) tuned to the undesired harmonic.

the coupling coil is made smaller in diameter than the tank coil, and is arranged to fit inside the tank. This electrostatic shield also fits into the tank coil and is of such a size that it may be placed between the two coils. To be effective, the shield must be well grounded, and must be constructed so that there are no closed paths through which eddy currents may flow. If eddy currents flow, the device tends to shield the magnetic field of the tank from the coupling coil. This type of shield is often called a "Faraday screen".

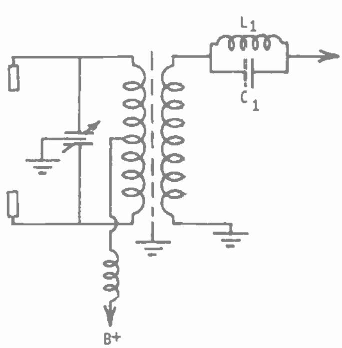


Fig. 21 A parallel tuned circuit used to suppress harmonic frequencies.

Other methods of suppressing harmonic frequencies include trap circuits connected between the final tank circuit and the antenna. One such type is illustrated in Fig. 20. It consists of a series resonant circuit tuned to the harmonic which it is desired to suppress. To the harmonic frequency, the resonant circuit offers very little opposition, and it is shorted to ground and prevented from reaching the antenna. This system is only partially effective as there is a tendency for some of the fundamental frequency to be bypassed also. Ordinarily, only the second and third harmonic frequencies are of sufficient magnitude to cause any trouble, and two such circuits, each tuned to one of these harmonics will greatly reduce the amount of harmonic radiation from the antenna.

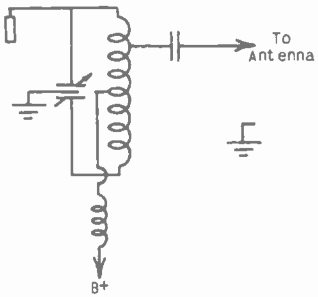


Fig.22 One method of capacitively coupling a final stage to an antenna.

Another type of trap circuit consists of a parallel tuned tank circuit connected as shown in Fig. 21. This circuit is also tuned to the troublesome harmonic, and since it offers a high opposition to currents of its resonant frequency, the amount of harmonic antenna current is considerably lowered. Furthermore, since the trap circuit has negligible opposition at the fundamental frequency, the normal antenna current is unaffected.

It is a well known fact that for maximum transfer of energy, the impedance of the load should be equal to the impedance of the source. Thus, to transfer maximum power from the tank circuit of the final stage to the antenna demands that the impedance of the antenna be matched to the tank circuit of the final amplifier. The impedance of most broadcast antennas, when they are correctly tuned

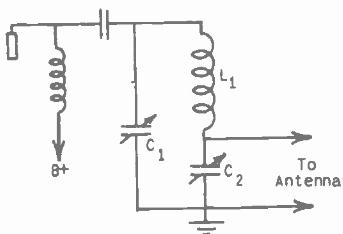


Fig.23 Another method of coupling a final tank to an antenna, which also provides harmonic suppression.

to the fundamental frequency, is fairly low at the point at which they are coupled to the final stage compared to the impedance of the final tank circuit. Therefore, some matching device must be used so that an impedance match may be obtained. This may consist of the R.F. transformer composed of the final tank coil and the coupling coil, or the final tank may be capacitively coupled to the antenna as shown in Fig. 22. If capacitive coupling is used, a better arrangement is the one shown in Fig. 23. In this case, the final tank circuit consists of inductance L_1 and capacities C_1 and C_2. By adjusting the capacity of C_2, and then tuning the tank to resonance with C_1, the impedance across which the antenna is connected may be made equal to the impedance of the source. If, for example, the antenna has an impedance of 20 ohms, the reactance of C_2 would be made 20 ohms at the fundamental frequency, and the capacity of C_1 would be varied until the tank circuit was resonant at the fundamental frequency. At the harmonic frequencies, the reactance of C_2 would be less than this value, and the antenna would not be correctly matched to the tank circuit, thereby preventing much harmonic power from being transferred to the antenna. This is the same as saying that most of the harmonics would be by-passed to ground through the low-reactance of the condenser C_2.

In many instances, the antenna is not situated directly at the transmitter building, but is located some distance away, often as far as several hundred yards. This is desirable for two reasons; first, the transmitter should not be placed directly in the strong R.F. field surrounding the antenna; and second, the presence of the

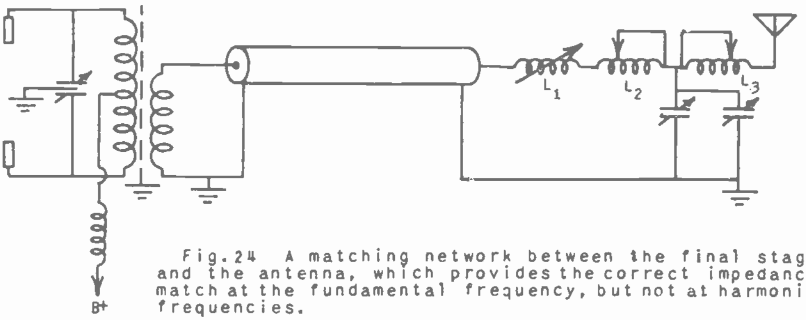


Fig.24 A matching network between the final stage and the antenna, which provides the correct impedance match at the fundamental frequency, but not at harmonic frequencies.

transmitter near the antenna causes the radiation pattern of the antenna to be changed. Ordinarily, the transmitter is connected to the antenna by means of an R.F. transmission line. Sometimes this transmission line is a concentric cable buried underground, and often it is merely a two-wire line strung on poles. In either case, the line must be so terminated that there will be no radiation from the line itself. This demands that the final stage of the transmitter be matched to the line and that the line be matched to the antenna.

The line is matched to the final stage by a coupling coil of the proper number of turns, and is matched to the antenna by means of a somewhat complicated matching circuit consisting of parallel capacitors and series inductors. An arrangement of this sort is illustrated in Fig. 24.

This network of inductors and condensers forms a correct match between the antenna and the transmission line, and, at the same time, offers a high opposition to harmonic frequencies. L_2 is a tapped coil by which rough adjustments in the inductance is made, whereas L_1 is a coil whose inductance is continuously variable, thereby, providing a means of securing fine adjustments of the inductance. Two parallel condensers are provided so that there will be more flexibility in the adjustment of the capacity of this circuit. In reality, this circuit acts simply as a low-pass filter having a cut-off frequency slightly above the resonant frequency of the antenna. The inductance L_3 is in the antenna circuit itself and provides a means of tuning the antenna to resonance with the fundamental frequency of the antenna. Sometimes the impedance matching and harmonic suppression network takes the form shown in Fig. 25.

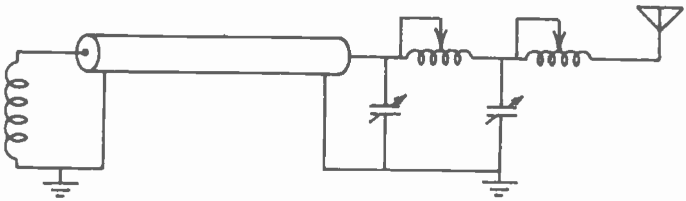


Fig.25 Another type of impedance-matching network.

One other point which must be considered in the suppression of harmonic frequencies is to make sure that there are no metallic objects in the immediate field of the antenna which have a natural frequency equal to a harmonic of the fundamental. Thus, it is common practice to break up the guy wires supporting the antenna into unequal lengths with strain insulators so that none of the guy wires or sections thereof will be resonant at a harmonic of the fundamental frequency. Should this occur, the offending conductor will absorb energy from the antenna field and reradiate it at a harmonic frequency.

8. SUMMARY OF HARMONIC FREQUENCY SUPPRESSION. The following points should be remembered when it becomes necessary to suppress undesirable harmonics:

1. Use tank circuits having a Q of at least 12 so that the oscillating current will be large and will be better able to smooth the plate current pulses into pure sine waves.
2. Do not over-excite Class C stages, as this causes a dip in the plate current pulses and produces odd harmonics.
3. Use push-pull stages so that even harmonics will be cancelled. The push-pull tubes should be well balanced.
4. Employ electrostatic shields between the final tank circuit and the antenna coupling coil to prevent harmonics from being capacitively coupled to the antenna.

5. Prefer inductive coupling to capacitive coupling for harmonic reduction.
6. Split tank circuits should always use split-stator tuning condensers to lower the harmonic content.
7. Use trap circuits or impedance matching networks having the property of offering high opposition to harmonic frequencies.
8. Break up guy wires and other conductors in the field of the antenna into unequal lengths so that they will not reradiate harmonics.

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT NO. 3

GRID MODULATION

LESSON NO. 12

WHEN DETERMINATION POPS UP

.....look out!

Determination is a compelling force that drives human beings through to their objective in spite of obstacles that may stand in the way. Some of us have more determination than others. Those who are lacking in determination should and can develop it.

I well remember a certain student who came to me to study radio several years ago. He was a little, white haired fellow with a happy smile and an easy going way. There was no indication that he had a very large supply of determination. However, he was a good student, and he stuck to what he started just as you are doing.

When his training was completed this little fellow seemed satisfied to just drift along with no particular objective in view. He did not worry about the future. And he made no serious attempts to get a worth while job in radio. While we all liked him very much, we felt that his future was most questionable.

Then one day an abrupt change came over our little fellow. He had made up his mind that he was going to get a job. Determination just radiated from him. And in a few days he disappeared....but not for long. The postman brought me a letter in which he told me that he was employed by a large broadcasting station.

Later on I discovered that he had gone to Chicago and straight to the headquarters of one of the two big broadcasting chains. He applied for a job. But he did not ASK for a job. He sold himself so thoroughly that he was put to work without delay. And....today that young man is making good.

This little true story proves conclusively that size does not count when the will to win is the dominant power. Success is available to all who want it....and are willing to work for it.

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Lesson Twelve

GRID MODULATION

"Since Grid Modulation is being employed by one large radio manufacturer almost exclusively, it is highly important that every radio engineer become thoroughly conversant with both the theory of operation and the method of adjustment.

"The second part of this lesson is devoted to a practical description of a 100 Watt transmitter using this type of modulation".

1. REVIEW OF MODULATION. There are, in general, two types of modulation systems; these are power modulation and efficiency modulation. Representative of the first type is plate modulation discussed at some length in Lesson 5 of this Unit. A brief review of the highlights of plate modulation will be of aid in distinguishing the principal differences between power and efficiency modulation.

In any modulation system, the peak output current (the tank current of the modulated stage or the antenna current) increases to twice its carrier-level amplitude during the peaks of 100% modulation. Furthermore, the negative peaks of the audio signal cause the instantaneous output current to fall to zero. The preceding statements apply equally well to the output voltage, which may be considered as being the R.F. voltage across the tank circuit or that present in the antenna. With this statement in mind, it is apparent that the power contained in the antenna varies from the carrier level to four times this value at the modulation peaks and then to zero.

Obviously, any power present in the antenna must have been derived from the plate voltage supply system of the transmitter. The plate supply furnishes power in DC form, and the R.F. and A.F. stages of the transmitter merely cause this power to be liberated at R.F. and A.F. rates. When the power in the antenna increases to four times its unmodulated value, either the power supply system of the transmitter is furnishing four times as much power, the efficiency of the power conversion has increased four times, or both the power input and the efficiency have increased enough to allow the peak output to quadruple.

In plate modulation, both DC and A.F. power are furnished to the input of the modulated R.F. stage. The DC power is derived from the plate power supply which furnishes plate voltage for the modulated stage. The A.F. power is obtained from an audio-frequency power amplifier. Since the A.F. power, itself, originates from the power supply system, it is correct to state that all of the power contained in the antenna is furnished by the plate-power supply system of the transmitter.

A plate-modulated amplifier stage is always worked under Class C conditions, since the efficiency of this mode of operation is very high. A well-designed Class C stage will have an efficiency ranging from 60% to 70%, and, furthermore, this efficiency does not vary during modulation. If the efficiency remains constant and the power output increases four times with modulation peaks, it is evident that the power input to the plate-modulated stage must likewise increase four times. This is accomplished by the DC power input to the modulated stage instantaneously doubling its carrier-level value and the A.F. power input from the modulator likewise increasing until it is equal to twice the power of the unmodulated carrier. With both the DC power input and the A.F. power input doubling, it is apparent that the total power input during this peak will be four times as great as when the carrier is unmodulated, and with a constant efficiency, the power in the antenna will also quadruple.

Thus, plate modulation is properly called "power modulation", because the power input to the modulated stage is varied at an A.F. rate. Although the DC power input doubles at the modulation peak, the average power furnished by the DC supply throughout a modulation cycle does not change from its unmodulated value. The average power in the antenna, however, increases one and one-half times with 100% modulation, and therefore it is clear that the total average power input to the modulated stage must likewise increase one and one-half times with complete modulation. This is accomplished by the modulator supplying an average A.F. power input equal to one-half of the average power furnished by the DC supply.

Now let us consider the principles of efficiency modulation. The first example that we have had of this type of modulation system was that given in Lesson 10 of this Unit on linear amplifiers. A linear amplifier is nothing more than an efficiency-modulated stage, as will now be disclosed. Again, with 100% modulation, the peak power output is four times the unmodulated power output. The DC plate current varies from twice its average value to zero, indicating that the DC power input doubles its value at the modulation peaks. There is, however, no A.F. power input supplied to the plate circuit of the linear amplifier and the only way that the power output may quadruple is for the efficiency to change with modulation. That the efficiency does change has previously been demonstrated. It was shown in Lesson 10 that the efficiency of a linear amplifier with complete modulation is twice as great as when the grid-exciting voltage is unmodulated. Therefore, at the modulation peaks, the DC power input instantaneously doubles, the efficiency likewise increases two times, and this causes the R.F. power output at this instant to quadruple. Although the DC power input varies from instant to instant, the average power input does not change with modulation, and since the average power in the modulated wave is one and one-half times as great as the unmodulated carrier, it is apparent that the average efficiency increases one and one-half times with 100% modulation.

Thus, to effect efficiency modulation, it is necessary for the DC power input and the efficiency to vary with the audio signal. Now that this has been determined, let us see if there are other ways in which this result may be accomplished. The efficiency of an amplifier depends, of course, upon the ratio of the power output to the power input. With efficiency modulation, the power input varies from its unmodulated value to twice this value and then to zero. It is now necessary to find some means of causing the efficiency to vary likewise and at an audio rate. With the power input determined, the only way that the efficiency may increase is for the power output to increase at a greater rate than the power input. With a given load the power output varies directly as the square of the amplitude of the fundamental component of the plate current. In the linear amplifier, the fundamental plate current component was caused to vary by keeping the operating angle constant at 180° and varying the amplitude of the grid - exciting voltage. With a constant operating angle, the DC plate current increases at the same rate as the fundamental plate current component, and this action causes the power input to double when the fundamental component doubles. Also, the power output quadruples at this time, indicating that the efficiency has increased twofold.

Another method by which the efficiency may be changed is to vary the operating angle. This may be accomplished by causing the grid bias to vary at an audio frequency rate. The grid - exciting voltage is constant and the operating angle decreases and increases as the bias voltage becomes greater and less. When the bias voltage is low, the plate current pulses are large and the plate voltage swing is maximum. This causes the efficiency to be great. At this instant, the fundamental plate current component is maximum, and the power output reaches its peak value. This is the principle of grid-bias modulation.

2. GRID MODULATION. The grid-modulated amplifier is, in some respects, very similar to a Class B linear amplifier. The main differences are that in the linear amplifier the grid bias is constant and the amplitude of the grid - exciting voltage varies at an audio frequency rate. On the other hand, the grid-modulated amplifier has a constant grid-exciting voltage and a grid bias which is varied at an A.F. rate. Shifting the bias of the amplifier causes the operating point to be changed and in turn varies the efficiency of conversion. Fig. 1 shows how the bias voltage is determined for grid modulation. The bias voltage consists of a constant DC voltage on which is superimposed an audio voltage of the proper value.

With no modulation the bias voltage is sufficient to bias the tube to approximately one and one-half times cut-off. With complete modulation this bias will vary from twice cut-off to the cut-off value at an A.F. rate. Therefore, a grid-modulated amplifier has its mode of operation changed from instant to instant. At the modulation peaks, the bias is equal to cut-off, and the stage is working as a Class B amplifier. At the carrier level, the bias is one and one-half times cut-off and the mode of operation is between Class B and Class C. During the troughs of the modulated wave, the bias increases to twice cut-off, and the mode of operation becomes Class C.

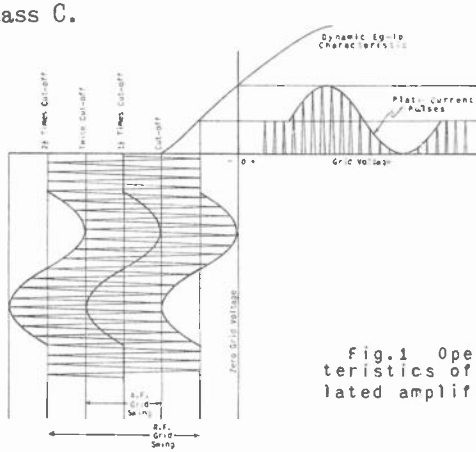


Fig.1 Operating characteristics of a grid-modulated amplifier

Thus, the grid-modulated amplifier changes from an under-excited Class C amplifier to a Class B amplifier. In broadcast operation, grid current is not allowed to flow except for a few milliamperes on occasional 100% modulation peaks. The effect of grid current is to cause the load on the A.F. amplifier to be variable which is liable to result in amplitude distortion of the audio signal. In amateur practice, where high quality is not of paramount importance, the grid-modulated amplifier is biased to approximately twice cut-off, and a sufficient grid-exciting voltage is applied to drive the grid considerably positive. Although the quality suffers by this action, a much larger power output and a higher plate-circuit efficiency may be obtained.

When the grid-modulated amplifier does not draw any grid current, no power is required to excite the tube, and the audio amplifier which causes the bias to vary at an A.F. rate need not supply any power. Practically, however, when the grid-modulated stage does not draw any grid current, the input impedance of the stage is somewhat variable and is dependent upon the leakage current between the grid and filament. To make sure that the correct load is reflected back into the audio amplifier, it is common practice to connect a resistor across the secondary of the modulation transformer so that the secondary impedance will be fixed. A grid-modulated stage is illustrated in Fig. 2. T_1 is the modulation transformer, the primary of which is connected directly in the plate circuit of the modulator. C_1 is an R.F. by-pass condenser used to return the R.F. to the filament. R_1 is the loading resistor which fixes the impedance of the secondary and assures that the proper load will be placed on the modulator. This resistor will vary from about 5,000 ohms to 100,000 ohms.

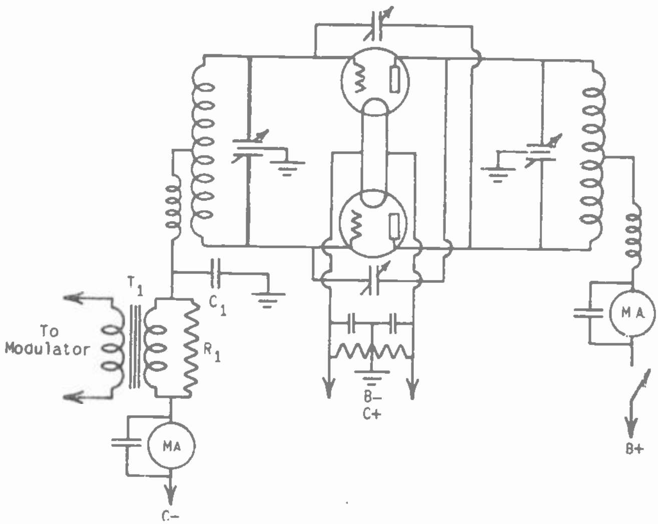


Fig.2 A grid modulated, push-pull amplifier.

This resistor which is placed across the secondary of the modulation transformer will dissipate some audio power, and, therefore, the modulator will have to supply a small amount of A.F. power to the modulated stage. The amount of power which must be supplied will, of course, be determined by the size of the loading resistor, a resistor of low value dissipating more power than one whose value is high.

Tubes suitable for grid-modulated amplifiers, as well as for linear amplifiers should be able to withstand very high plate voltages and should have comparatively low amplification factors. Such tubes will require a very large grid bias to produce plate current cut-off, and a large grid-exciting voltage may be applied without driving the grid more than a few volts positive and thereby distorting the modulation. It has been determined that to obtain a large power output the pulses of plate current at the crest of the modulation cycle should be large. The only way that this may be accomplished is for the grid of the tube to be driven positive, but as previously explained, the resulting grid current places a variable load on both the modulating and exciting voltages and causes the modulation to be distorted. It is possible by careful design to make the exciting and modulating voltage sources have good voltage regulation (that is a low internal impedance), and by driving the grid only moderately positive, the distortion can be kept low enough so that such a stage may be used in a transmitter for police radio or aviation radio purposes. For broadcasting, however, it is essential that the distortion be kept at an absolute minimum, and grid-modulated amplifiers used in broadcast transmitters rarely draw grid current except on occasional modulation peaks. Operation without grid current usually produces low efficiency, often not more than 20 to 25%. Usually the power output even at the peaks of the modulation is not sufficient to bring the peak plate dissipation of the tube up to the rated value, and in order to obtain a high degree of linearity, it is essential that some power output be sacrificed.

3. ADJUSTING A GRID-MODULATED AMPLIFIER. Let us suppose that we are going to adjust a grid-modulated amplifier for broadcast operation. With this type of operation, we shall allow perhaps 1 or 2 milliamperes grid current to flow at the crest of the 100% modulated signal. To make sure that the audio voltage and the grid-exciting voltage are low enough to prevent excessive grid current flow, it is necessary to adjust the amplifier first for the crest operating conditions. The first step, therefore, is to neutralize the amplifier by any one of the methods previously given, and then apply a grid bias approximately equal to cutoff. With the full plate voltage applied, the R.F. excitation voltage is now increased until 1 or 2 ma. of grid current are flowing. The load impedance is now varied by changing the L/C ratio of the tank and the coupling of the antenna or of the grid circuit of the following stage, keeping in mind that a high load impedance will tend to make the modulation more linear but will limit the power output that may be obtained. Perhaps considerable juggling of the excitation and the load will be necessary before the correct conditions are secured. A high load impedance will cause the R.F. voltage produced across the tank to be large, the plate current pulses will be small, the efficiency will be large, and, although the dynamic grid voltage-plate current characteristic will have a longer straight portion, indicating that the modulation will be linear, the power output which may be secured will not be very great. Of course, if the load impedance is too large, the efficiency, when unmodulated, will be so great that it will not be possible for it to double at the peak of the modulation cycle. This condition will produce non-linearity and is remedied by using a smaller L/C ratio or by coupling the antenna closer to the tank circuit. Thus, a compromise between efficiency and power output will be necessary, and the final adjustment of the load impedance will be the one which gives the required amount of power output with as little distortion of the modulation envelope as possible.

The bias voltage is now increased to one and one-half times cut-off, and no further adjustment of the excitation should be made. An audio voltage of sine wave form, such as is produced by an audio oscillator, is fed to a small speech amplifier, and the output of the amplifier is coupled to the modulation transformer in the grid circuit of the modulated stage. The audio voltage developed across the secondary of this transformer should be great enough to swing the bias voltage from twice cut-off to the cut-off value. Ordinarily, this audio voltage should have a peak value approximately equal to $E_B/2\mu$.

The plate current as read by a DC meter will increase slightly with modulation, although the increase should not be very great. Let us determine why the DC plate current does increase with modulation when this is not true of either the plate-modulated amplifier or the linear amplifier. In the linear amplifier the operating angle is constant. At the peak of a 100% modulated signal, the tank current doubles its value, which means that the peak value of the fundamental plate current component has doubled. With a constant operating angle of 180°, the peak of the plate current pulse is two times the peak of the fundamental component, and the DC component is .637 times the peak of the fundamental component. These ratios hold true as long as the operating angle remains unchanged. Therefore, when the fundamental plate current doubles, the DC component likewise doubles, and the average value of the DC component throughout an audio cycle is still equal to the DC component when the stage is unmodulated. The result is that the DC plate current meter shows no change in its reading when modulation is applied.

In the grid-modulated amplifier, the conditions are considerably different. The bias changes at an audio frequency rate, and the operating angle changes with each variation of the bias. Let us suppose that the operating angle when the stage is unmodulated and the bias is approximately one and one-half times cut-off is 130°. With this operating angle, the peak of the plate current pulse is 2.39 times the peak of the fundamental component, and the DC current is .568 times the peak of the fundamental component. Let us assume that the peak of the fundamental component is 200 ma. when the stage is unmodulated. With an operating angle of 130°, the peak of the plate current pulse will be 2.39 × 200 or 478 ma. Also, the DC component will be .568 × 200 or 113.6 ma. The stage is now modulated 100%; and the peak of the fundamental component must double and become 400 ma. The operating angle changes from 130° to 180°, and the peak of the plate current pulse is now 2 × 400 or 800 ma. Likewise, the DC component becomes .637 × 400 or 254.8 ma. The following table gives a summary of this variation.

	Operating Angle	Fundamental Component	Peak Plate Current	DC Plate current
Unmodulated	130°	200 ma.	478 ma.	113.6 ma.
Peak Of Modulation	180°	400 ma.	800 ma.	254.8 ma.

Fig. 3 shows the wave form of these pulses. A is the condition with no modulation, and B represents the conditions existing at the modulation peak. The peak of the fundamental component doubles its value during the peaks of 100% modulation, and thus the R.F. voltage across the tank likewise doubles indicating that the power output has quadrupled. Notice that the peak of the plate current pulse does not need to double (it increases from 478 to

800 ma.) in order that the power output will be four times as great. Also, note that the DC plate current more than doubles (it increases

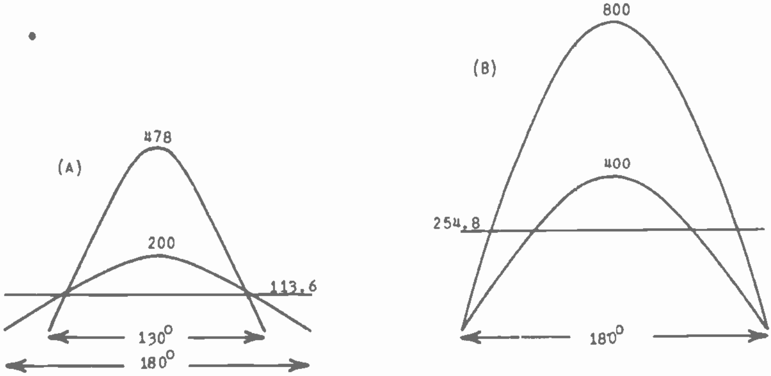


Fig.3 Illustrating the plate current pulses of a grid-modulated amplifier. A is the unmodulated condition, and B is at the crest of the modulation cycle.

from 113.6 to 254.8 ma.) at the peak of the modulation. Therefore, during 100% modulation, the DC plate current varies from zero when the modulated wave is at its trough to 254.8 ma. when it is at its peak. The average value of the DC component throughout a modulation cycle will therefore be one-half of the peak value or 127.4 ma. Thus, the reading of the plate current meter will change from 113.6 ma. with no modulation to 127.4 ma. with complete modulation. Fig. 4 shows the relation between the R.F. current, plate current, grid current, screen-grid current and audio bias voltage. These curves are for an 804 pentode.

The ratios used to determine the peak of the plate current pulse and the DC component were derived by a mathematical procedure too complicated to be included in this text. The foregoing example does not represent any particular amplifier; it merely serves to prove that the DC plate current will not be as constant as it is in a linear amplifier. Sometimes the plate current will increase less than in this example; rarely will it increase more. When the plate current does increase with modulation, the power input more than doubles its value at the modulation peaks, and the efficiency does not quite double. However, both will increase enough for the power output at this time to quadruple.

As is the case with a linear amplifier, the average amount of plate dissipation is less when the signal is modulated than when it is unmodulated. Also, the bias supply must have good regulation so that the DC bias voltage will not change when the grid draws a few milliamperes of current at the peaks of modulation. Since the DC grid current, if any, is quite variable, it is necessary to load the preceding stage artificially by means of a swamping resistor.

When adjusting a commercial transmitter employing grid-bias modulation, the manufacturer's recommendations should be read very carefully and followed explicitly. Naturally, the manufacturer knows his product better than anyone else, and has at his command laboratory equipment with which he can determine quite accurately the constants which should be used for proper operation. Such laboratory equipment is rarely found in the average radio station, and it is best that the operator rely on the manufacturer's figures.

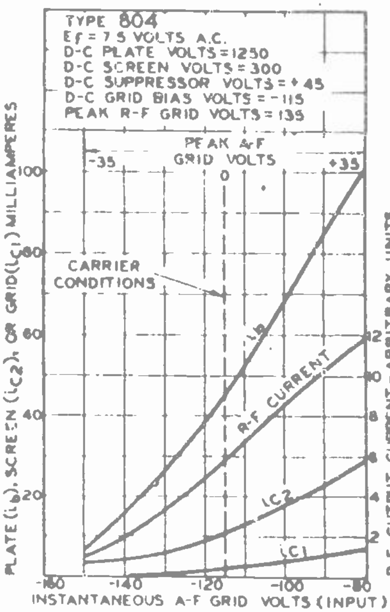


Fig.4 Grid modulation characteristics of a type 804 pentode. Pentodes, as well as triodes may be used as grid-modulated amplifiers and have the added advantage of not requiring neutralization.

When a composite transmitter is to be adjusted, a cathode ray oscilloscope is a distinct advantage. By its use, the linearity of the grid-modulated amplifier can be determined with various adjustments and power outputs. If, however, such an instrument is not available, the engineer must depend on the meter readings to determine the operating condition of the transmitter.

The constants of the modulation transformer will depend upon how heavily the secondary is to be loaded, and upon what impedance the primary is to work from. Most commercial transmitters have output transformers on their speech equipment whose secondary is matched to a 500-ohm line. This is desirable if the audio output of the speech amplifier is to be carried any great distance. When this is the case, the primary of the modulation transformer must be able to match the 500-ohm line. Suppose, for example, that the DC bias voltage of the grid-modulated amplifier is 150 volts, and that this value is one and one-half times cut-off. The audio voltage must be able to drive the bias voltage up to the cut-off value and should have a value of 50 volts. (If 150 volts is one and one-half times cut-off, then two-thirds of this, or 100 volts, would be the cut-off value, and the audio would need to have a peak voltage of 50 volts to change the bias from 150 to 100 volts.) Also, assume that the secondary of the modulation transformer is loaded with a resistor of 50,000 ohms. The speech equipment must deliver enough power so that the peak voltage across this 50,000-ohm resistor is 50 volts. Let us determine how much power this represents. The power in a resistor is equal to E^2/R, and from this formula we can find how much power it takes to produce a voltage of 50 volts across 50,000 ohms. This is:

$$\text{Power} = \frac{50^2}{50,000}$$

$$= \frac{2500}{50,000}$$

$$= .05 \text{ watt or } 50 \text{ milliwatts.}$$

The foregoing example illustrates very clearly the one desirable feature of grid modulation. The speech equipment need furnish only 50 milliwatts peak power to the modulated stage. To continue with the design of the transformer, it is seen that the secondary must work into 50,000 ohms, and that the primary works from 500 ohms. This is an impedance ratio from secondary to primary of 100. The turn ratio from secondary to primary should be the square root of this value or should be 10. Thus, the secondary must have ten times as many turns as the primary.

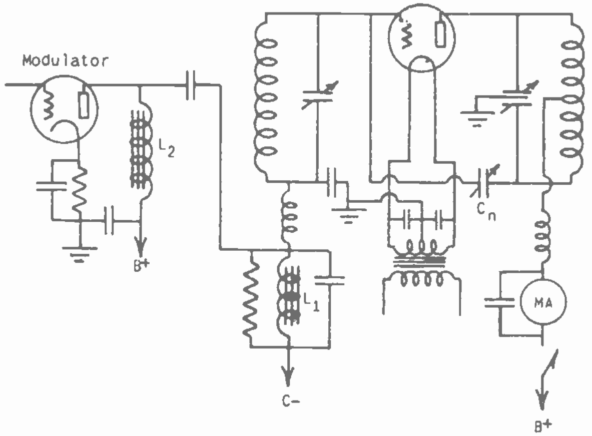


Fig.5 A modulator impedance coupled to the grid-modulated stage.

Sometimes the modulation transformer also serves as the output transformer of the speech amplifier. In that case, the primary impedance must be such that it will match correctly the output tube of the speech amplifier. If there is a sufficient reserve of audio power available from the speech amplifier, it is usually better to load the secondary of the modulation transformer more heavily by placing a smaller load resistor across the secondary. Even if the resistor is reduced to 5,000 ohms, the power that needs to be furnished by the speech amplifier is only .5 watt. In this case, the impedance ratio of the transformer would be 10 to 1 and the turn ratio should be slightly greater than 3 to 1, if the modulator feeds a 500-ohm line.

One other system of coupling the output of the speech amplifier to the modulated stage is shown in Fig. 5. In this case, a modulation choke replaces the modulation transformer, and the output of the modulator is capacitively coupled to the modulation choke. The plate voltage of the modulator is fed through an iron-core choke L_2. This choke should be large in comparison to the modulation choke so that very little audio current will be lost through it.

Since grid modulation has both outstanding advantages and disadvantages, whether it is used or not will depend somewhat upon the transmitter-design engineer's personal convictions. With grid modulation, the audio power required is practically negligible, but the stage must be operated very inefficiently to avoid excessive distortion. Plate modulation is highly efficient, but requires a somewhat elaborate audio setup. The overall efficiency of the two systems is approximately the same. Suppose, for example, that 1000 watts of R.F. power are desired in the antenna. If plate modulation is employed, the efficiency would be approximately 60% and the DC power imput would need to be 1,666 watts. The A.F. power required for 100% modulation would be 833 watts, and assuming that the modulator is a Class B stage with an efficiency of 50%, the DC input to the modulator would be 1,666 watts. Therefore, the total DC power input to the final stage and the modulator would be 3,332 watts. Now suppose that a grid-modulated stage is to be used to produce this 1,000 watts of R.F. output. Let us assume an efficiency of 30%, and with this efficiency, the DC power input to the modulated stage would be 3,333 watts, or the same as was required with the plate-modulated system. Of course, some A.F. power would be required and it is possible that this efficiency, as assumed, is slightly high; however, we have not taken into account the DC power input which the driver and voltage amplifier stages of the plate-modulated stage would need, so that it is probable that the total power input would be very nearly the same in either case.

4. SUPPRESSOR MODULATION. Since the tetrode and pentode types of tubes have become popular in transmitter apparatus, two other types of modulation systems have come into existence. These are screen-grid modulation and suppressor modulation. At present, screen-grid modulation is confined to amateur practice, however, suppressor modulation is being used in certain types of high-frequency transmitters for aviation purposes, and is now coming into use for broadcast transmitters.

The inherent distortion which arises from the use of suppressor modulation limited its use in broadcast transmitters until the introduction of inverse feedback. With inverse feedback, a large part of the distortion is cancelled, and suppressor modulation is quite satisfactory.[1]

[1] Inverse feedback as applied to transmitters will be explained in a subsequent lesson.

Suppressor modulation makes use of a pentode adjusted to operate as a Class C amplifier. The modulation is effected by applying a negative DC bias voltage to the suppressor grid, and then superimposing on this voltage, the audio voltage from the modulator. A stage connected to operate in this manner is illustrated in Fig. 6. The control grid is supplied with an unmodulated R.F. voltage, and the plate with an unmodulated DC voltage. The negative bias voltage for the suppressor grid should be obtained from a bias source having good regulation, since the suppressor is ordinarily driven positive at the crest of the audio cycle and thus draws some current. The screen voltage should be secured from the plate-supply using a series screen-dropping resistor.

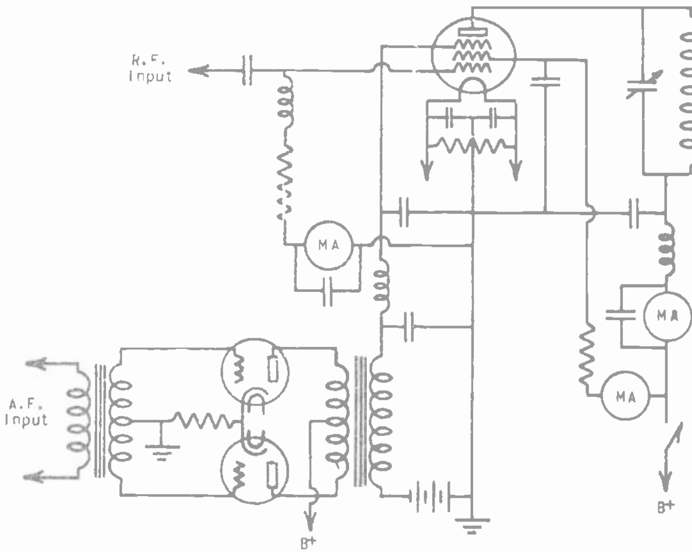


Fig.6 A suppressor modulated amplifier.

Since the control grid usually draws current during a part of the excitation cycle, the control grid bias voltage may be obtained from a grid leak or by any of the other means commonly used for bias voltage. If cathode bias is used, the biasing resistor should be by-passed both for R.F. and A.F.

Suppressor modulation takes advantage of the fact that the minimum plate voltage at which plate current begins to flow is proportional to the potential of the suppressor grid. If a large load impedance is used, the relation between the output voltage and the suppressor voltage will be nearly linear. The fact that it is not entirely linear is illustrated by the curves of Fig. 7. These curves show the relation between the suppressor voltage, the plate current, the screen current, the control-grid current, the suppressor current, and the output current. Notice that the curve representing the output current is fairly linear throughout a portion of its length, but that at high values of negative suppressor voltage, a departure from linearity results. For this reason, the distortion becomes very large when an attempt to secure high percentage of modulation is made.

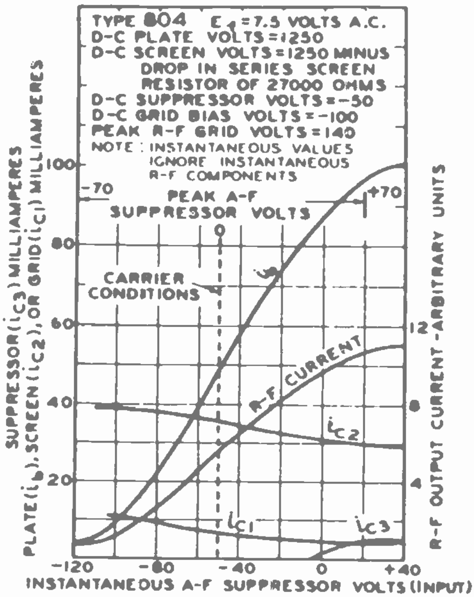


Fig.7 Suppressor modulation characteristics of a type 804 transmitting tube.

The procedure for adjusting a suppressor-modulated amplifier is to apply the correct control-grid bias voltage, screen voltage, plate voltage and R.F. excitation to cause the plate current to assume the value specified by the manufacturer for this type of operation. The next step is to adjust the load impedance until most of the total space current is drawn by the plate when the suppressor grid is slightly positive, but not enough positive to draw more than a little current. The final adjustment of the load impedance should be the one of the highest value which produces the foregoing condition, since the linearity of modulation depends largely upon the use of a large load impedance. This condition where the suppressor is slightly positive corresponds to the crest of the modulation cycle. The suppressor voltage is now made negative to determine what value of voltage will produce plate current cut-off. Half way between these two values is the operating bias value which should then be applied to the suppressor. The peak audio voltage required for 100% modulation would then be one-half of the difference between these two extremes of suppressor voltage, however, as previously explained, 100% modulation is not possible without introducing an excessive amount of distortion.

The adjustment of a suppressor-modulated amplifier is somewhat simpler than that of a grid-modulated stage, because the bias on the suppressor grid can be adjusted independently of the bias and excitation of the control grid. In grid-modulated amplifiers, all of these factors are interdependent and the adjustment for optimum conditions is somewhat critical.

Like grid modulation, suppressor modulation possesses the advantage of requiring very little audio power for complete modulation. The suppressor is operated with sufficient negative bias so that, under carrier conditions, the R.F. output voltage and current are just half of the values that will be reached at the crest of the A.F. signal. Since the suppressor grid will be driven positive during a portion of the audio cycle, the modulator must be capable of supplying the power which will be dissipated when the suppressor draws current. In fact, the modulator must be able to supply more than just the amount of power that will be dissipated. It should have a sufficient reserve of power so that it is capable of producing the power that will be dissipated without introducing serious distortion of the audio signal.

At the crest of the audio signal, the power output of a suppressor-modulated amplifier is approximately the same as would be obtained from the same tube operating as a Class C amplifier, and thus, the power obtained at the carrier level is one-fourth of that produced at the crest of a 100% modulated wave. The plate efficiency at the carrier level is approximately the same as that of a grid-modulated amplifier operated with grid current. This varies from 30 to 35%. This efficiency increases at the crest of the modulation cycle to about 66%. The overall efficiency averaged throughout an audio cycle is, however, lower than that of a grid-modulated stage, since the screen grid losses mount rapidly during the troughs of the modulated wave.

In summary, we may state that suppressor modulation has the advantage of requiring low audio power, and of being relatively easy of adjustment. To its disadvantage, there must be recorded its comparatively large amount of distortion which has limited its use.

PART TWO

The following is a description of a modern, 100-watt, completely AC operated, transmitter. It is the Western Electric No. D-98653. Grid modulation is used. In addition to being a complete 100-watt transmitter, it may be employed as the exciter unit of a 1000-watt transmitter. A thorough study of the description and operating instructions of this transmitter will enable the student to gain a knowledge of this particular transmitter, as well as aiding him in learning the maintenance and operating procedure of broadcast transmitters in general. Fig. 8 shows a complete diagram of this transmitter.

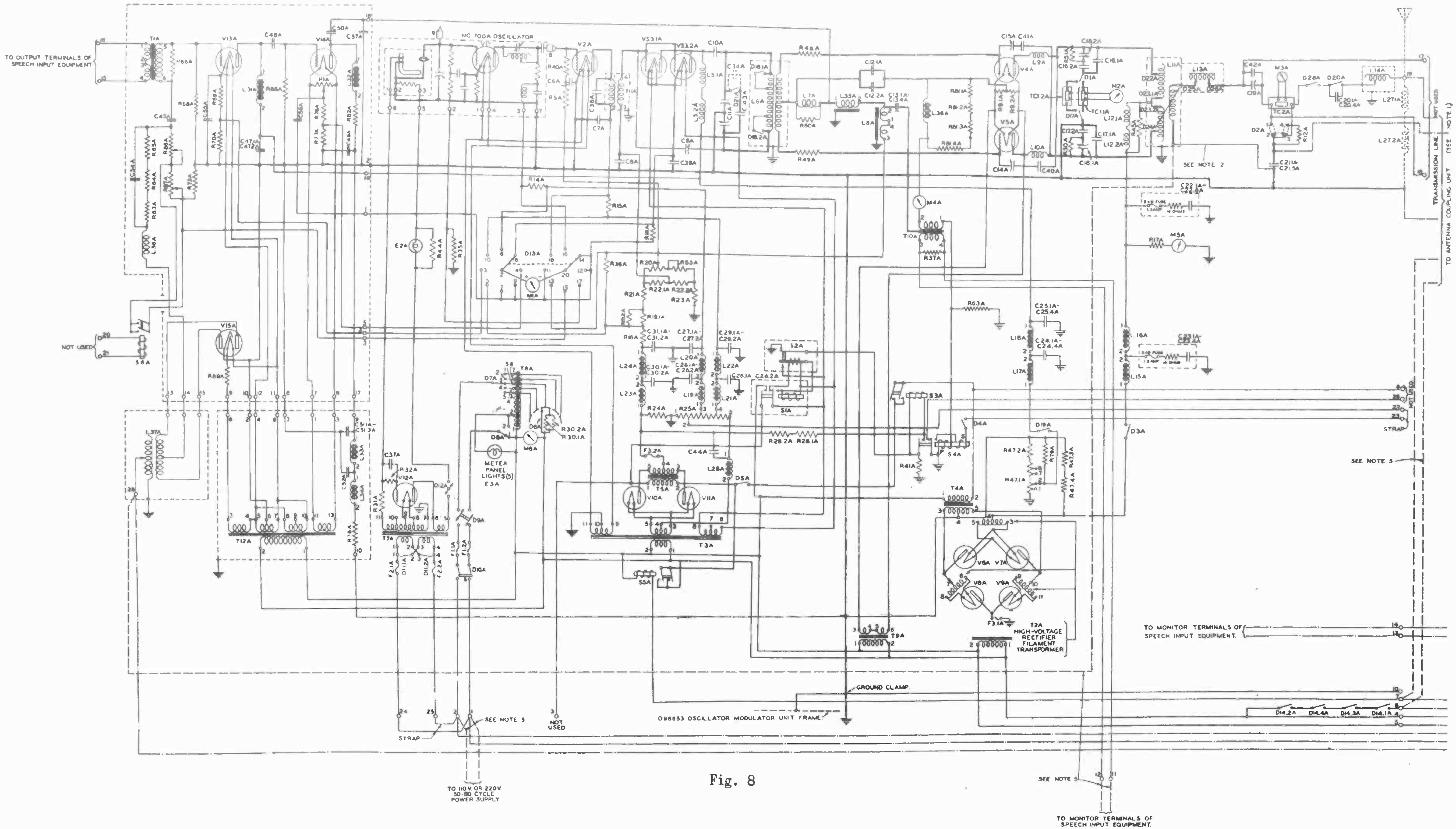

TO OUTPUT TERMINALS OF SPEECH INPUT EQUIPMENT
NOT USED
NO. 700A OSCILLATOR
METER PANEL LIGHTS (5) E3A
TO 110V OR 220V. 50-60 CYCLE POWER SUPPLY
STRAP
SEE NOTE 5
NOT USED
GROUND CLAMP
D98653 OSCILLATOR MODULATOR UNIT FRAME
T2A HIGH-VOLTAGE RECTIFIER FILAMENT TRANSFORMER
TO MONITOR TERMINALS OF SPEECH INPUT EQUIPMENT
SEE NOTE 2
TRANSMISSION LINE
TO ANTENNA COUPLING UNIT (SEE NOTE 1)
TO MONITOR TERMINALS OF SPEECH INPUT EQUIPMENT

Fig. 8

1. INTRODUCTION. The Western Electric No. D-98653 Oscillator-Modulator is a complete 100 watt radio broadcasting transmitter. This equipment is completely AC operated, without rotating machinery or batteries, and is entirely enclosed in a steel cabinet with no outside apparatus except the speech input equipment, the monitoring arrangements and antenna connections. This equipment features 100 per cent modulation, thorough protection for the operating personnel from high voltages, precision frequency stability, careful suppression of radio frequency harmonics and a stabilized feedback circuit to reduce harmonic distortion and noise.

The apparatus requires 1670 watts, 110/220 volts, single phase power. The power factor is 0.85.

In general, the oscillator-modulator consists of a crystal controlled oscillator followed by three stages of amplification. The last stage is a balanced amplifier in which modulation is accomplished.

For ordinary operation and maintenance, complete access to all parts of the equipment is had through the doors in the unit. These doors are provided with safety switches for the protection of the operating personnel.

GENERAL DESCRIPTION

Oscillator

The carrier frequency of the oscillator-modulator is controlled by a No. 700A (quartz crystal controlled) Oscillator. This oscillator is adjusted as a unit to the operating frequency and will maintain its calibration well within ± 50 cycles.

The oscillator unit contains a quartz crystal, crystal heater, thermostat, a vacuum tube V1Y and associated circuits. It is mounted in the slide rail assembly in the upper compartment of the oscillator-modulator. Power connections to the oscillator are made by means of spring contacts which engage when the Oscillator is properly inserted. The radio frequency output is obtained from terminal 8 located at the right rear corner of the Oscillator and is connected to the first amplifier input circuit by means of the connector detail provided for this purpose. The quartz plate is maintained at a constant temperature by means of a heater controlled by a mercury thermostat, and power is available at all times for the crystal heater circuit after the transmitter is connected to the source of supply, irrespective of the operation of the transmitter.

In the temperature control circuit, power is supplied to the heater through a three-element rectifier tube V12A acting as a relay. The grid voltage of this tube is in phase with the plate voltage when the crystal temperature is low, and current then flows into the heater. When the crystal temperature reaches the proper value, the contacts of the mercury thermostat close, applying an out-of-phase voltage to the grid of the rectifier tube, and no current flows into the heater. A single transformer T7A supplies all power for this circuit.

First Amplifier

The first amplifier V2A isolates the oscillator from the succeeding stages of amplification. The grid-bias is obtained from potentiometers R20A and R53A having adjustments marked "R.F. OUTPUT CONTROL," "COARSE" and "FINE," respectively. These adjustments are used to control the output of the transmitter, as the output of this amplifier may be smoothly varied by reducing the bias from a value so far beyond cut-off that no output is obtained to a value that gives the desired output. This amplifier is located in the compartment to the right of the Oscillator.

Second Amplifier

The second amplifier stage employs two tubes, V3.1A and V3.2A in parallel and is coupled to the first amplifier by the untuned radio frequency transformer T11A and to the third amplifier by a tuned radio frequency transformer L6A. This amplifier is located in the compartment with the first amplifier.

Third Amplifier

The third amplifier stage employs two tubes V4A and V5A in a balanced circuit and it is here that modulation takes place. It is effected by what is known as "grid-bias" modulation. The grids of the tubes are biased to considerably beyond cut-off and the radio frequency voltage is applied to the two grids out of phase, as in any push-pull amplifier. The audio frequency voltage is applied to both grids in parallel and is effectively in series with the DC grid-bias voltage. Thus, the resulting grid-bias voltage is varied in accordance with the audio frequency modulating voltage, which accounts for the name "GRID-BIAS" modulation. By changing the bias in this manner, the radio frequency output voltage is varied between zero and twice the normal value, which constitutes complete modulation. The output impedance of this amplifier is such that the relationship between the output and input voltages is essentially linear and the audio distortion is thereby kept at a low value. The negative bias voltage for the modulator-amplifier tube is supplied through inductances L8A, L35A, L7A and the secondary of L6A.

Output Circuit

When the Oscillator-Modulator is used alone, the output of the third amplifier is coupled to the antenna through an antenna coupling circuit for the suppression of radio frequency harmonics. However, when the Oscillator-Modulator is used in conjunction with an amplifier, the antenna coupling circuit is slightly modified and becomes the power amplifier input circuit. The modifications necessary in the output circuit of the Oscillator-Modulator to adapt it for use with Western Electric amplifiers are described in the Amplifier instruction bulletins.

Monitoring Circuit

Monitoring in the Oscillator - Modulator is accomplished by means of an audio transformer T10A connected in the high voltage return lead of the third amplifier. The grids of the vacuum tubes in this stage are biased so far beyond cut-off that no audio frequency power flows in their plate circuits until the radio frequency is applied and modulation effected. The audio frequency component of the rectified carrier power then appears in this circuit and the output of the monitoring transformer T10A is a true reproduction of the program at the output of the Oscillator-Modulator. When the Oscillator-Modulator is used in conjunction with Western Electric Amplifiers, monitoring is accomplished in the amplifier and the monitor in the Oscillator-Modulator is not used.

Audio-Amplifier Unit

The amplifier which provides the necessary audio frequency modulating voltage has two impedance coupled stages employing tubes V13A and V14A. The input to this amplifier is the vector difference of two voltages, one the signal from the microphone across resistance R66A, the other a portion of this signal from the output stage, demodulated by a full wave rectifier V15A and applied across resistance R86A and R87A. This constitutes the "loop" feedback feature of the equipment.

In addition to "loop" feedback, "cathode" feedback is also incorporated. This consists of an impedance composed of R81A, L36A, the primary of T10A and S4A connected in the common plate return circuit of tubes V4A and V5A. Since these tubes are not operated as "Class A" radio frequency amplifiers, the audio frequency component of the rectified carrier power appears in the common plate return. A corresponding voltage appears across the cathode circuit impedance and hence is impressed on the grids of the tubes, in series but out-of-phase with the audio input. This constitutes the "cathode" feedback.

Power Supply Circuits

The Oscillator-Modulator is completely AC operated. The filaments of all vacuum tubes are heated by alternating current; grid-bias and plate voltages are supplied by mercury vapor rectifiers. The unit is arranged to be operated from either a 110 or 220 volt single phase, 50 cycle or 60 cycle power supply, but when operated in conjunction with Western Electric amplifiers, it is always operated from one phase of the 220 volt, three-phase supply.

Bias-Plate Rectifier--The bias-plate rectifier supplies grid-bias voltages for all the radio frequency amplifiers in the unit and plate voltages for the oscillator and first amplifier. This rectifier employs two mercury vapor tubes, V10A and V11A in a conventional full-wave rectifier circuit.

High Voltage Rectifier--The high voltage rectifier employs four mercury vapor tubes V6A, V7A, V8A, and V9A in a single phase full-wave, "bridge type" rectifier circuit. This rectifier supplies the plate voltage to the second and third amplifiers and audio amplifier unit.

Power Control and Protection Circuits

The power control and protection circuits control the sequence of power application to the various circuits and protect the equipment from possible damage in case of failure of any piece of apparatus.

The starting and protection circuits of both the Oscillator-Modulator and any amplifier units may be interlocked and the complete equipment started or stopped by operating the "MASTER CONTROL" switch (D9A) located on the Oscillator-Modulator unit. The "MASTER CONTROL" switch (D9A) actuates the starting circuit which applies power to the circuits of the Oscillator-Modulator (and amplifier) in the correct sequence, introducing such delays as are necessary for the protection of the equipment.

The auto-transformer T8A which supplies power to all circuits of the Oscillator-Modulator is provided with a tap switch "POWER VOLTAGE CONTROL" (D6A) which allows the operator to compensate for variations in the local line voltage during operation of the transmitter.

All power and high voltage circuits are adequately fused. The plate circuit of the third amplifier is protected against accidental overload by means of an overload relay S4A. An overload operates this relay which immediately removes the high voltage. When normal conditions have been restored, the plate voltage may be reapplied by momentarily depressing "OVERLOAD RESET" button (D4A) which opens the holding circuit of overload relay S4A.

The operating personnel is protected against accidental contact with high voltage circuits by means of door switches D14A on each of the four doors to the unit. These switches are series connected and may also be interconnected with similar door switches on any associated amplifier. The opening of any door will immediately remove all high voltage from the apparatus.

INSTALLATION

Antenna

The Oscillator-Modulator can be operated with an antenna of any resistance and reactance but where antennae of less than 10 or more than 90 ohms are encountered, engineering advice should be requested. The antenna lead-in is connected to terminal 19 located on top of the unit. Copper tubing is recommended for this purpose. The equipment will also operate into a concentric transmission line.

PRELIMINARY ADJUSTMENTS

Power Supply Circuits

Before placing fuses in any cutouts or applying power to the equipment, connect link switches D8A, D11.1A and D11.2A as follows:

Link Switch	110-Volt Operation	220-Volt Operation
D8A	Position 2 (right-hand)	Position 1 (left-hand)
D11.1A	Connect terminals 1 and 2	Connect terminals 2 and 3
D11.2A	Connect terminals 3 and 4	Connect terminals 2 and 3

Connect link switch D7A in accordance with the following table for the local line voltage:

Local Line Voltage	Position of Tap Switch D7A
98-102 or 197-205	5
103-107 or 206-215	4
108-112 or 216-225	3
113-117 or 226-235	2
118-125 or 236-249	1

For 220-volt operation, place two 15-ampere fuses in cutouts F1.1A and F1.2A. For 110-volt operation, use two 30-ampere fuses. Place 1-ampere Western Union Telegraph Fuses in cutouts F3.1A and F3.2A. After the links and fuses have been placed as directed, set the "POWER VOLTAGE CONTROL" switch (D6A) in position 3 (normal), and apply power to terminals 1 and 2 of the equipment. The closing of switches D10A and D9A should cause meter M6A to indicate 220 ± 5 volts.

If the line voltage is not known exactly, switch D7A may be properly connected in the following manner: Set the "POWER VOLTAGE CONTROL" switch in position 3, set link switch D8A for 110 or 220-volt operation as required, and then vary D7A until meter M6A indicates 220 ± 5 volts.

CAUTION: THE MAIN POWER SUPPLY SWITCH D10A SHOULD ALWAYS BE OPENED WHENEVER D7A IS ADJUSTED.

Upon completion of these adjustments, insert six No. 258B Vacuum Tubes[1] in sockets VS6A, VS7A, VS8A, VS9A, VS10A and VS11A and connect the flexible plate leads to the corresponding anode caps. Insert five General Electric 5-watt 120-volt Edison (candelabra base) Mazda lamps in the meter panel light sockets and also General Electric Type 4, 18-volt, 2-watt lamp in the crystal heater indicator lamp socket ES2A which is located behind the colored bezel on the front panel.

Oscillator Heater Circuit

To prepare the slide-rail assembly for the Oscillator, the latches on the front of the runners should be opened so that the unit may be inserted in position. The spring contacts on both the oscillator unit and transmitter terminal strips should be examined to see that they are not bent out of alignment and that good contact between the terminals on the Oscillator and the terminal strip is insured. When the oscillator unit has been properly inserted, close the two latches and secure them with the screws provided. Terminal 8 of the Oscillator should be connected by means of the connector detail provided to the similarly mounted terminal projecting through the shield which covers the base of the vacuum tube sockets VS2A, VS3.1A and VS3.2A.

Before inserting fuses F2.1A and F2.2A, disconnect the power supply from the transmitter at the service entrance. See that snap switch D12A is in the "OFF" position. Place 2-ampere D&W Fuses in cutouts F2.2A.

Insert a No. 287A Vacuum Tube in socket VS12A and a No. 271A

[1] The filaments of new mercury vapor tubes should be heated at least 15 minutes before the high voltage is applied. This pre-heating removes any particles of mercury adhering to the sides or elements of the tubes after shipment or handling, thus minimizing the possibility of flash-overs. (See section on "Maintenance.")

Vacuum Tube in socket V1Y of the Oscillator. The filament of the No. 287A Vacuum Tube should begin to heat as soon as the service entrance switch is closed, since this circuit is energized at all times independent of the transmitter power switches. When the filament of the No. 287A Vacuum Tube has been heating for about 15 minutes, switch D12A should be placed in the "ON" position. With the Oscillator in place, the indicator lamp (E2A) will light, indicating that the heater of the Oscillator is receiving current. The vacuum tube V12A will also indicate that heater current is flowing by the presence of a characteristic glow. If the vacuum tube indicates that the heater circuit is functioning, but the indicator lamp E2A does not light, the lamp may be defective and should be replaced. The lamp E2A will remain lighted approximately 45 to 75 minutes and will then operate intermittently, remaining on about 30 seconds and off about 30 seconds. Snap switch D12A should never be opened except when adjusting the contacts on the Oscillator Unit or when installing a new No. 287A Vacuum Tube.

This equipment requires little care once it is in operation as there are no mechanical relays in the system. However, the indicator lamp E2A should be observed from time to time to ascertain that the heater circuit is functioning correctly. Sufficient time should be allowed for this observation to permit a complete cycle of operation.

Adjustment of Power Control Circuits

Adjustment of Time Delay Relay, S2A.--When the "MASTER CONTROL" switch is closed, the heater winding of relay S2A is energized and after about 45 to 75 seconds the relay should close its front contacts and energize relay S1A which will lock up and at the same time open the heater circuit of S2A, thus allowing the armature of S2A to cool and return to its "normal" or "open" position.

The operation of S2A should be checked very carefully to see that the *back contacts open before the front contacts close.* After relay S1A has operated, the back contacts of S2A should close in from two to thirty seconds.

NOTE: When looking at relay S2A from the front, the "back" contact spring is the right-hand spring and the "front" contact spring is the left-hand spring. The "armature" carries the heater winding and controls the two center springs which are mechanically linked together. Adjustments on this relay should be made with a Western Electric No. 259 Tool.

The "front" contact spring of the relay may be bent to adjust the operating time on the heating cycle. To increase this time, the "front" contact spring should be bent away from the armature. To decrease this time, the spring should be bent towards the armature.

If it becomes necessary to adjust the operating time on the heating cycle, the time interval on the cooling cycle should also be checked. The interval may be regulated by bending the "back" contact spring. Bending this spring towards the armature results in a decreased time interval. To increase the time interval, the "back" contact spring should be bent away from the armature. A reasonable amount of time should be allowed between tests so that the relay winding and spring will have time to return to room temperature.

Grid-Bias Control Relay, S1A. Relay S1A, which is mounted on the panel in the lower compartment operates in conjunction with the heater relay S2A. It has two distinct functions, (a) to break the heater circuit of relay S2A, (b) to complete the circuit to the bias-plate rectifier when the door switch relay S5A is closed. This relay should ordinarily require no adjustment. However, the armature must be free to move and its contacts must be kept clean.

High Voltage Control Relay, S3A. Relay S3A has two contacts connected in parallel in order to carry safely the current required by transformer T4A. These contacts should close simultaneously and the armature should move freely during operation. This relay should require no adjustment.

Plate Overload Relay, S4A. Relay S4A is adjusted at the factory so that it will operate between 100 milliamperes minimum and 450 milliamperes maximum. If it becomes necessary to adjust this relay, all switches except D12A should be in the "Off" position and a source of DC voltage (about 2 volts) should be applied in series with a suitable milliammeter and variable resistance from terminal 1 of transformer T10A to ground. The tension of the contact springs may then be adjusted with the No. 259 Tool until the relay operates at the correct current value. Relay S4A should be tested with the relay cover in place.

Door Switch Relay, S5A. This relay should require no adjustment but should function smoothly and have clean, non-sticking contacts. In general, all AC relays of the double contact type must have the same pressure on each contact so that audible chatter and hum may be suppressed.

Control Relay, S6A. This relay is not used when the Oscillator-Modulator is used alone.

Starting Circuit Sequence. When all relays have been adjusted, the complete starting circuit of the transmitter should be checked. With the "HIGH VOLTAGE" switch in the "Off" position, operate the "MASTER CONTROL" switch. The operation of this switch energizes auto-transformer T8A and the following operations should result immediately:

(a) Filaments of all installed vacuum tubes are lighted.
(b) The meter panel is illuminated.
(c) Meter M6A indicates.
(d) Door switch relay S5A is energized if all doors are closed.

CAUTION: IF THE NEW MERCURY VAPOR TUBES HAVE NOT BEEN PREVIOUSLY PREHEATED, THEY SHOULD BE AT THIS POINT, BY LEAVING THE DOORS OF THE TRANSMITTER UNIT OPEN UNTIL THE FILAMENTS HAVE RECEIVED THE REQUIRED HEATING.

After a delay of about 45 seconds, the armature of S2A operates, causing relay S1A to be energized. S1A then locks up, disconnecting the heater winding of S2A and completing the primary circuit of transformer T5A. This transformer applies voltage to the bias-plate rectifier. Transformer T5A will not be energized, however, if relay S5A has failed to operate due to an open door switch.

After relay S1A has operated, disconnecting the heater winding of S2A, there is a small time delay during which the bi-metallic element of S2A is cooling. At the completion of this cooling cycle, the back contacts of relay S2A close and complete the circuit of relay S3A which is operated by the grid-bias voltage. Relay S3A will not operate unless the grid-bias potential is sufficiently high to protect the power tubes against abnormal plate current. The operation of relay S3A will complete the primary circuit to high voltage transformer T4A when the "HIGH VOLTAGE" switch is "On".

Oscillator Unit

The heater circuit of the Oscillator should have been in operation for at least one hour before any tuning adjustments are made on the transmitter. The operation of the Oscillator should now be checked as follows:

Place "TEST METER" switch (D13A)[1] on the "OSC. PLATE X 10" and then on the "OSC. GRID" positions, in order to determine that the observed readings of meter M1A fall within limits specified in Table III.

TABLE III - - TYPICAL METER READINGS

Power Supply Voltage-M6A	220 ± 5 Volts
1st Amplifier Plate Current-M1A	3-20 Milliamperes
Oscillator Plate Current-M1A	6-12 Milliamperes
2nd Amplifier Plate Current-M1A	10-30 Milliamperes
Oscillator Grid Current-M1A	0.05-1.0 Milliamperes
3rd Amplifier Grid Current-M1A	0
Feedback Current-M1A	7 ma.
1st. A.F. Plate-M1A	25 ma. ± 5 ma.
2nd A.F. Plate-M1A	45 ma. ± 5 ma.
3rd Amplifier Plate Voltage-M5A	3000 ± 100 Volts
3rd Amplifier Plate Current-M4A	130 ± 2 Milliamperes[2]
3rd Amplifier Output Current-M2A	0.8-1.1 Amperes
Antenna Current (Mesh)-M3A[3]	0.8-1.2 Amperes
Antenna Current-M3A[4]	

First and Second Amplifier Tuning

Resonance is obtained in all circuits except the antenna circuit and the first amplifier by an adjustment for minimum DC plate current of the tube whose tuned output circuit is being adjusted. In tuning the second or third amplifier, care should be taken that neither stage is tuned to the second harmonic of the fundamental frequency. Should two points of minimum plate current be found with coarse adjustment as specified, the one at which the capaci-

[1] The readings of the "TEST METER" M1A are to be multiplied by 1, 10, or 20 as indicated by the dial plate of the associated switch D13A. In the "OSC. PLATE X 10" position of switch D13A, the readings of meter M1A are to be multiplied by 10.

[2] When using the indirect method of measuring antenna power, it will be necessary to adjust for a plate current of 150 milliamperes.

[3] Link switch D2A must be placed in the 2-3 position in order to read mesh current.

[4] The antenna current will equal the square root of the quotient of the operating power in watts divided by the effective antenna resistance at the operating frequency.

tance of the variable condenser is maximum is the correct adjustment. This will correspond to the point of higher dial reading and higher output current.

The first amplifier requires no tuning adjustment as its plate circuit is coupled to the following stage by means of an untuned radio frequency transformer. The radio frequency output of this stage, and thus the output of the transmitter, is controlled by the "R.F. OUTPUT" control.

Before tuning the second amplifier, set the neutralizing condenser C9A so that its plates are about one-sixth engaged. This condenser is adjusted by means of a screw-driver through the lower hole located directly above the "R.F. OUTPUT" control. This opening in the panel is normally concealed by a pivoted cover. Open link switch D3A and see that link switch D19A is closed. This allows the plate voltage to be applied to the second amplifier but not to the third amplifier.

With the "R.F. OUTPUT" control in the minimum position place "HIGH VOLTAGE" switch in the "On" position and set the "TEST METER" switch on the "2nd AMPLIFIER PLATE X 20" position. Adjust the "R.F. OUTPUT" control until the "TEST METER" M1A reading increases approximately 10 milliamperes. Now vary the "2ND AMP. TUNING"[1] condenser C11A until the "TEST METER" indicates a minimum. If difficulty is experienced in tuning this stage, a slight readjustment of condenser C9A should be made. After obtaining an approximate adjustment, but before the final tuning of the second and third amplifiers, these stages must be neutralized.

Second and Third Amplifier Neutralization

In the neutralizing procedure, the third amplifier is neutralized first. Place the "HIGH VOLTAGE" switch in the "Off" position, and open the doors.[2] Leave the link switch D3A open and link switch D19A closed. Using a screw-driver, set the neutralizing condenser C14A-C15A so that their plates are a little more than one-third engaged. Connect the sensitive thermocouple TC1.1A in the closed circuit by means of the links provided. *The correct polarity must be maintained when the meter leads are transferred to this sensitive thermocouple.* Set the "R.F. OUTPUT" control and the "3RD AMP. OUTPUT" coupling control at minimum.

Close the "HIGH VOLTAGE" switch and then slowly vary "3RD AMP. TUNING" condenser C18A in conjunction with the "R.F. OUTPUT" control until "3RD AMP. OUTPUT CURRENT" meter M2A indicates a maximum. If the reading indicated by this meter becomes excessive, adjust neutralizing condensers C14A-C15A until a reading of approximately 1 ampere is indicated. Continue by increasing the "R.F. OUTPUT" control to its maximum, meanwhile adjusting condensers C14A-C15A to keep the "3RD AMP. OUTPUT CURRENT" at approximately 1 ampere. Check the tuning by readjusting both the "2ND AMP. TUNING" conden-

[1] All radio frequency controls located on the front panel with the exception of "R.F. OUTPUT" control are adjusted by means of a special spanner wrench (Western Electric No. 704A Tool). This tool is included with the equipment.

[2] When opening the doors of the equipment to adjust or handle any of the apparatus, always ascertain that the door switch relay has operated by noting that the "3RD AMP. PLATE VOLTAGE" meter reads zero.

ser and "3RD AMP. TUNING" condenser until the "3RD AMP. OUTPUT CURRENT" Meter indicates a maximum, and then adjust neutralizing condensers C14A-C15A until the "3RD AMP. OUTPUT CURRENT" meter indicates approximately zero current. When this has been accomplished, the third amplifier is neutralized.

The second amplifier should now be neutralized as follows: Place the "HIGH VOLTAGE" switch in the "Off" position and reduce the "R.F. OUTPUT" control to minimum.

CAUTION: WHILE THE SENSITIVE THERMOCOUPLE TC1.1A IS IN CIRCUIT, THE DOORS OF THE UNIT SHOULD NEVER BE OPENED UNTIL THE "HIGH VOLTAGE" SWITCH IS "OFF" AND THE VOLTAGE INDICATION OF THE "3RD AMP. PLATE VOLTAGE" METER M5A IS LESS THAN 500 VOLTS. FAILURE TO OBSERVE THIS PRECAUTION RESULTS IN A TRANSIENT SURGE WHICH MAY BURN OUT THE SENSITIVE THERMOCOUPLE.

Observing the foregoing caution, open link switch D19A and close link switch D3A. Using a screw-driver, operate the potentiometer P1A to either the extreme right or left, first having taken note of its original position. This will apply sufficient modulation to give a readable deflection of the "3RD AMP. OUTPUT CURRENT" meter while neutralizing. Operate the "HIGH VOLTAGE" switch. Vary the "R.F. OUTPUT" control and neutralizing condenser C9A until the "3RD AMP. OUTPUT CURRENT" meter indicates approximately 1 ampere. Check the setting of the "3RD AMP. TUNING" condenser and the 2ND AMP. TUNING" condenser for a maximum reading of the "3RD AMP. OUTPUT CURRENT" meter. Then vary neutralizing condenser C9A for a minimum reading of this meter, meanwhile advancing the "R.F. OUTPUT" control. When the "R.F. OUTPUT" control is at a maximum and neutralizing condenser C9A is so adjusted that there is little or no current indication by the "3RD AMP. OUTPUT CURRENT" meter, the second amplifier is neutralized. Open the "HIGH VOLTAGE" switch and after the reading on M5A has decreased below 500 volts, open the doors and close link switch D19A. Leave all other controls in position so that the process of tuning the third amplifier output circuits will be simplified.

Reconnect thermocouple TC1.2A and transfer the meter leads to it, maintaining the correct polarity. Restore potentiometer P1A to its original position.

Third Amplifier Tuning and Second Mesh Adjustment

Place link switch D2A in the 3-4 position. Open D28A and temporarily short circuit the antenna coupling condenser by placing link switch D20A across the studs occupied by C21A. Set the "R.F. OUTPUT" control to minimum, and apply the high voltage.

After setting "3RD AMP. OUTPUT COUPLING" at a dial reading between 5 and 10, increase the "R.F. OUTPUT" control until the "3RD AMP. OUTPUT CURRENT" meter indicates approximately 1 ampere. Check the tuning of the *second* amplifier for minimum plate current. Again adjust the "R.F. OUTPUT" control until the "3RD AMP. OUTPUT CURRENT" meter indicates about 1.5 amperes. Then vary "ANTENNA TUNING" condenser C19A until the "3RD AMP. OUTPUT CURRENT" Meter indicates a minimum. Adjust "3RD AMP. OUTPUT COUPLING" in conjunction with the "R.F. OUTPUT" control until "ANTENNA CURRENT" meter M3A indicates 1 ampere and "3RD AMP. PLATE CURRENT" meter M4A reads 130 milliamperes.[1] The "3RD AMP. OUTPUT CURRENT" meter then should read between 0.8 and 1.1 amperes. If the reading of the "3RD AMP. OUTPUT CURRENT" meter is low, it is an indication that the primary of L11A has too many active turns; if the reading is high, the primary of L11A has too few active turns. It will then be necessary to decrease or increase, by a turn or two, the number of turns on each half of the coil and repeat the above tuning procedure. The reading of the "3RD AMP. OUTPUT CURRENT" meter should not exceed 1.1 amperes under normal conditions.

At this point a check should be made on the position of the "R.F. OUTPUT" control. If the position of the "COARSE" control is above "80" or below "65" on the scale when the meter readings are within the limits given, the plate voltage on the second amplifier should be increased or decreased respectively by moving the tap connection on R47.1A to the next higher or lower numbered tap. This process should be repeated until the required meter readings are obtained with the "COARSE" control of the "R.F. OUTPUT" between "65" AND "80" On the scale. If it is impossible to secure an adjustment within these limits, the tap on R47.1A should be selected which most nearly meets the limits specified.

Record "3RD AMP. OUTPUT CURRENT" and "3RD AMP. PLATE CURRENT."

After the transmitter is tuned as above, switch D13A should be placed in the "FEEDBACK CURRENT" position and the rotor of L37A adjusted until meter M1A indicates 7 milliamperes. Each adjustment of L37A must be made with the "HIGH VOLTAGE" switch in the "Off" position. After each change of L37A, it may be necessary to make a minor change in adjustment of the second mesh circuit by means of the "ANTENNA TUNING CONDENSER" for a minimum on meter M2A. With the feedback current properly adjusted, the audio input level for complete modulation with speech is +4 db.

The filament center tap potentiometer P1A is provided to adjust for a minimum noise level from the audio amplifier. This minimum noise level condition can be determined by listening to the monitor output or by noise level measurements and P1A should be left at the point giving minimum noise.

Antenna Tuning

Open the "HIGH VOLTAGE" switch and reduce the "R.F. OUTPUT" control to zero. Remove the coupling condenser C21A from its normal position and connect it to the studs normally used for the antenna series condenser C20A. Care should be taken that the adjustment of the "ANTENNA TUNING" condenser is not changed. Remove link switch D2A from the circuit, leave link switch D20A open and close D28A. Apply high voltage, *reduce coupling to zero* and gradually increase "R.F. OUTPUT" control until the "3RD AMP. OUTPUT CURRENT" meter indicates about 1.5 amperes. Adjust the "3RD AMP.

[1] When using the indirect method of measuring antenna power, it will be necessary to adjust for a plate current of 150 milliamperes.

TUNING" condenser for a minimum indication of the "3RD AMP. PLATE CURRENT" meter. Increase the coupling slightly and vary the taps on L14A until antenna resonance is indicated by a minimum indication of the "3RD AMP. OUTPUT CURRENT" meter. In case resonance cannot be obtained by varying the taps on L14A,[1] it will be necessary to employ a series condenser C20A in the antenna circuit. This is done by temporarily connecting one of the C20A condensers (whose reactance does not exceed 200 ohms for the carrier frequency employed) in series with the external antenna lead-in. When resonance has been obtained, the equipment is shut down, coupling condenser C21A is returned to its normal position, and link switch D2A is put in the 2-3 position. If the series condenser was found necessary it should be placed in its normal position (C20A) otherwise the C20A studs should be short-circuited by means of link switch D20A.

Operate the unit and vary the "R.F. OUTPUT" control until the "3RD AMP. OUTPUT CURRENT" meter indicates about 1.5 amperes. Adjust the "ANTENNA TUNING" condenser for a minimum indication of the "3RD AMP. OUTPUT CURRENT" meter and then adjust "R.F. OUTPUT" control in conjunction with "3RD AMP. OUTPUT COUPLING" until the "3RD AMP. PLATE CURRENT" meter and the "3RD AMP. OUTPUT CURRENT" meter indicate those readings obtained in the section under "THIRD AMPLIFIER TUNING AND SECOND MESH ADJUSTMENT." Check the tuning of the third amplifier by varying the "3RD AMP. TUNING" condenser for a minimum reading of the "3RD AMP. PLATE CURRENT" meter. If this point is not found at or near the former setting of the "3RD AMP. TUNING" condenser, it is an indication that the antenna circuit is not exactly tuned and the above tuning procedure must be repeated. If the reading of the "ANTENNA CURRENT" meter exceeds 1.2 amperes, decrease the capacity of the coupling condenser C21A. Increase this capacity if the current is less than 0.8 ampere. All meter readings should fall within the limits shown in Table III.

OPERATING PROCEDURE

Starting

Before starting, ascertain that the antenna grounding switch is in the "OPERATE" position. In the normal operation of the transmitter, full-automatic starting should be used. The "HIGH VOLTAGE" switch is always left in the "On" position, and the unit is started by operating the "MASTER CONTROL" switch. However, where two or more stations are sharing time on the same frequency and a minimum starting time is desirable after one station signs off, it is advantageous to use semi-automatic starting. This consists in placing the "HIGH VOLTAGE" switch in the "Off" position and operating the "MASTER CONTROL" switch. This may be done several minutes before the preceding station signs off, thus allowing the time delay circuit to function and the vacuum tubes to reach normal operating temperature. The operator can then start instantly by operating the "HIGH VOLTAGE" switch.

[1] In the case of an inefficient antenna, it may be necessary to employ additional series inductance in the antenna circuit in order to obtain resonance. In such cases, the additional inductance may be mounted external to the unit and connected in series with the antenna lead-in.

As soon as the equipment is in operation, all meter readings should be checked and any necessary adjustments made.

Stopping

To stop the equipment, operate the "MASTER CONTROL" switch. Ground the antenna. When stopping for a brief interval, it is sufficient to operate only the "HIGH VOLTAGE" switch. This de-energizes the high voltage rectifier and eliminates the time delay when restarting.

Crystal Temperature Control Circuit

The temperature control circuit will require occasional replacements of the relay tube and indicator lamp which are in service continuously. However, *because of the vital function of the relay tube in maintaining the correct oscillator frequency, it is imperative that the tube employed be in an operative condition. Normal operation of this tube is indicated by the periodic flashing of the indicator lamp located on the front of the unit and any irregularity in the operation of this lamp should be promptly investigated.*

The crystal heater circuit should be in operation at least 4 hours before the station is put on the air.

Modulation

The audio input system to the Oscillator-Modulator is arranged to operate from a 500-ohm circuit and requires a speech input level of +4 db. for complete modulation.

During operation, the third amplifier grid current should occasionally be checked. Any grid current during the program, with the exception of occasional pulses, is an indication of over-modulation. *The "TEST METER" switch D13A should not be left on the "3RD AMPLIFIER GRID" position.*

Monitoring

When the Oscillator-Modulator is operated as previously described, the monitoring output level at terminals 11 and 12 is approximately +10 db. The output of the monitoring transformer T10A should normally be terminated in 500 ohms, as otherwise the quality of the transmitted program will be impaired. This is provided for by resistance R37A, which is connected across the output terminals of transformer T10A.

If a monitoring device of 500 ± 100 ohms is used, it should be connected directly to the output of the monitoring circuit (terminals 11 and 12) and resistance R37A should be disconnected. If a monitoring device of other than 500 ± 100 ohms is used, it will be necessary to provide a suitable output transformer to connect between the output of T10A and the monitoring device. This transformer must be of such ratio that T10A is effectively terminated in 500 ohms. Resistance R37A should be disconnected.

If no monitoring device is connected to the output of the transformer T10A, resistance R37A must be left connected.

Overloading

Should the overload relay operate during a program, it is usually sufficient to press the "OVERLOAD RESET" button D4A. If the overload relay continues to operate each time the "OVERLOAD

RESET" button is pressed, trouble in the Third Amplifier output circuit is indicated and it should be determined in accordance with the procedure outlined under "Location of Trouble."

Use of "ANTENNA CURRENT" Meter

The "ANTENNA CURRENT" meter may be connected in either the second mesh circuit or the antenna circuit by means of link switch D2A. To connect the meter in the second mesh circuit, link switch D2A should be placed in the (2-3) position. To connect the meter in the antenna circuit, the link should be placed in the (1-2) position. The antenna coupling circuit is so arranged that when the correct value of coupling capacity C21A is used, and the antenna is connected and properly tuned, the resistance introduced into the second mesh circuit by means of the coupling capacity is always 100 ohms. Thus, with the "ANTENNA CURRENT" meter connected in this circuit, one ampere will always be indicated for 100 watts output ($I^2R=100$) regardless of the actual resistance of the physical antenna.

The permanent use of the "ANTENNA CURRENT" meter in the antenna circuit (link switch D2A in the 1-2 position) is not essential for correct operation. However, if it is desired to employ the meter in this circuit and it is found that the antenna resistance is less than 40 ohms it will be necessary to provide a meter of suitable range for the particular antenna involved. While it is possible to substitute the new meter for the "ANTENNA CURRENT" position. it is advisable to install it external to the unit.

Load Resistance

A load resistance of 100 ohms (R12A) is provided in the Oscillator-Modulator which may be connected in the second mesh circuit by placing link switch D2A in the 3-4 position. This resistance duplicates the resistance introduced by the antenna into the second mesh circuit by means of the antenna coupling capacity C21A and allows the unit to be operated for test purposes under actual load conditions without causing interference to other stations assigned to the same frequency. When this load resistance is used, the antenna should be disconnected by removing the antenna series condenser C20A and the short-circuiting link D20A (when used) from the circuit. Meter readings taken with the load resistance in circuit can always be duplicated with the antenna connected and are always an assurance that the equipment is delivering its full rated power to the antenna.

Meter Readings

All meter readings should be checked periodically during operation. If the "POWER SUPPLY VOLTAGE" should vary appreciably from 220 volts, it should be adjusted to the proper value by means of "POWER VOLTAGE CONTROL" switch.

Check frequently the oscillator grid and antenna current readings. Should the antenna current vary slightly while warming up, it should be adjusted to the correct value by means of the "R.F. OUTPUT" control. This adjustment should be made if possible without modulation. Otherwise, care should be exercised that the adjustment is made during low points of modulation.

MAINTENANCE

General

The Oscillator-Modulator must be kept free from dust and dirt. Compressed air is recommended for cleaning the apparatus inside the enclosure, but a soft clean cloth may be used with good results. Waste or oily cloth should never be used.

All nuts, bolts and screws should be examined occasionally and any loose ones tightened. Examine all electrical connections and tighten any loose contacts. Trouble can often be prevented by such precautions.

Cleaning Air Condensers

The exposed variable air condensers should be cleaned at least onee a week with compressed air or its equivalent. The presence of any dust or dirt on the plates may result in the condenser arcing thereby taking the station off the air. A small bellows and a clean dry cloth can be used to advantage for this purpose where a high pressure air system is not available.

Vacuum Tubes

In order to obtain both maximum life and satisfactory performance, it is important that vacuum tubes be operated within the voltage limits previously specified in this lesson.

As far as possible, the operator should anticipate tube failures and make the required tube replacements. Tube failures may be guarded against to some extent by keeping a careful record of the length of time the tubes have been in service and by observing from time to time the condition of the tube elements. Sagging or warped elements will, of course, increase the probability of tube trouble and such tubes should be replaced when discovered.

It is essential that the filaments of new mercury vapor tubes be heated at least fifteen minutes before the high voltage is applied. This is done in order to remove any particles of mercury adhering to the sides or elements of the tubes which might result in flashovers. It is therefore suggested that spare rectifier tubes be prepared for service in advance by placing them in the Oscillator-Modulator when not in use and giving the filaments the necessary pre-heating with the "HIGH VOLTAGE" switch in the "Off" position, and the doors of the unit opened. This precedure should be repeated at least once a month. Spare rectifier tubes thus preheated should be kept in an upright position until they are required.

It is also recommended that the No. 287A Vacuum Tube employed in the heater circuit be operated without plate voltage for about 15 minutes before placing the tube in actual service. This can be done by opening snap switch D12A.

Vacuum Tube Sockets

It is essential that the contacts of all tube sockets be kept clean and smooth at all times. Care should be exercised to see that the third amplifier tubes V4A and V5A are not subjected to any mechanical strains when placed in the sockets.

Relays

The contacts of all relays should be inspected and carefully cleaned once a month. Dust collects on the contacts in spite of relay covers and sometimes causes a failure in the operation of an important circuit. Relay contacts in the protective and high voltage circuits should receive special attention. Crocus cloth can be used to advantage in cleaning relay contacts. Badly pitted power relay contacts may be carefully smoothed with a fine file.

Carrier Frequency Deviation

In order that there may be no appreciable deviation of the carrier frequency from the assigned value, periodic observations should be made of the crystal heater indicator light located on the front panel. The intermittent operation of this light indicates whether or not the No. 287A Vacuum Tube is supplying heater current to the Oscillator. A defective No. 287A Vacuum Tube should be replaced immediately while a defective indicator lamp should be replaced as soon as possible without interrupting the program. The failure of this lamp does not prevent the operation of the crystal heater circuit and the intermittent flashing of the No. 287A Vacuum Tube will serve temporarily to indicate whether or not heater current is being supplied. (See section on "Operating Procedure.")

LOCATION OF TROUBLE

General

If this equipment is regularly and carefully maintained, very little trouble will be experienced. The operator should endeavor to become familiar with the circuits, their functions and the location of apparatus as quickly as possible. A detailed schematic diagram of the complete unit is given in this lesson.

In case of trouble in any of the control or protection circuits, the operator should remember that these circuits are interlocked so that the failure of one piece of apparatus often prevents other pieces from functioning. For example, should the bias-plate rectifier fail to operate correctly, the high voltage relay S3A cannot operate to energize the high voltage transformer T4A. In case a piece of apparatus fails to function, the operator should first ascertain that the previous interlocking circuits have operated, and then examine the piece of apparatus for defects. Relay contacts should be cleaned regularly as outlined under "Maintenance."

Trouble in the radio frequency circuits is usually caused by the improper adjustment of the circuits. The first step in the case of trouble in these circuits should be to see that all adjustments are in accordance with those described in this lesson as well as with those recorded in the station log.

It is not practical to attempt to describe every possible cause of trouble, but the following paragraphs give some of the more possible ones. It should be remembered that these are only suggestions.

Power Supply Circuits

Bias Voltage Failure--This condition (unless a rectifier tube fails) is probably due to the failure of relay S5A or a faulty door switch. If relay S5A does not operate when doors are closed, examine each door switch for an open circuit. Assuming that relay S5A operated correctly, and yet the rectifier does not function, the trouble might be due to improper operation of relay S1A. If the bias-plate rectifier is functioning as indicated by a blue glow in tubes V10A and V11A, but the grid-bias and plate voltages are incorrect, the load distribution of the potentiometer may be incorrect. This may be brought about by a breakdown to ground or open circuit condition of any equipment energized by this rectifier. In general, if the bias potentials are above normal, the fault is due to excessive loading on the positive side of the rectifier, and vice versa for excessive positive potentials.

High Voltage Failure--If the filaments of the high voltage rectifier are energized, the trouble may be due to failure of some of the preceding interlocking circuits or the failure of relay S3A. The last mentioned trouble may be due to previous operation of overload relay S4A which would short-circuit the winding of relay S3A. Pressing "OVERLOAD RESET" button D4A will ascertain if this is the case.

Temperature Control Circuit Failure. If the filament of the No. 287A Vacuum Tube fails to heat (assuming that the tube is in good condition), examine fuses in cut-outs F2.1A, F2.2A. Check setting of link switches D11.1A-D11.2A for correct voltage settings. (See section on "Adjustment of Apparatus.") Check for line voltage by closing switches D9A and D10A and see if meter M6A reads. When voltage has been checked, and the No. 287A Vacuum Tube heats properly, and yet the heater circuit does not function, look for bad connections at the terminal strip at the back of the Oscillator Unit. The No. 287A Vacuum Tube will light up with a characteristic glow when heater current is flowing. Snap switch D12A might be examined for a possible open circuit.

Oscillator

Check all voltages and the heater circuit. If the voltages are correct, try a new No. 271A Vacuum Tube. If trouble is found to be in the Oscillator, it must be returned to the nearest distributor for repairs.

Plate Circuits

If the output circuit of an amplifier is not tuned to resonance, the plate current of that amplifier will be excessive. Improper tuning of the amplifier is apt to result in excessive plate current due to incorrect plate circuit impedance. If plate current increases suddenly while tuning, and varies erratically with tuning condenser variation, this stage or a previous stage is no doubt oscillating at some parasitic frequency. A periodic variation of output plate current may be caused by a beat note brought about by "singing" in two successive stages. This condition exists whenever the equipment is not properly neutralized and proper neutralization will usually correct such trouble.

Balance of Third Amplifier Tubes

A careful balance of the load should be maintained between the two No. 212E Vacuum Tubes in the third amplifier stage. In this connection, it is to be understood that two vacuum tubes may be poorly matched and yet each be electrically perfect. Poor quality may be due to mismatched vacuum tubes in the third amplifier stage.

Balance is indicated by the relative color of the third amplifier vacuum tube plates when the transmitter is operating under load, Equalization of the load handled by these tubes is accomplished by loosening the set screw of the condenser C18.1A adjusting its capacity with reference to C18.2A until proper balance is indicated. The tuning of the output circuit should be rechecked after each such adjustment.

Filament Circuits

If all filaments (including meter lights) should not light when the "Master Control" switch D9A is closed, the main power is off or switch D10A is in the open position. Check main power bus by noting if heater circuit indicator lamp E2A is functioning. A single tube filament not lighted indicates either a faulty socket connection, or a burned out filament. The filaments of the second amplifier tubes, V3.1A and V3.2A, are connected in series and the failure of either filament will deprive the other tube of filament current. Group filament trouble probably is caused by a faulty transformer or an open circuit. If possible, determine the trouble which caused the failure before any replacement is made.

Excessive Noise Level.

This condition may be brought about by improper tuning of the radio frequency stages. Faulty filter sections, resulting from broken down condensers or damaged inductances, also will cause an increase in noise level. Arcing in any radio frequency circuit or poor filament contacts will produce erratic noises in the output. In general, the type of noise often will indicate the probable source of the trouble. Improper adjustment of P1A will result in excessive noise. (See Page 25)

VACUUM TUBES USED IN THE OSCILLATOR-MODULATOR NO. D-98653

V1Y V2A V3.1A V3.2A	Western Electric No. 271A Vacuum Tubes
V4A, V5A	Western Electric No. 212E Vacuum Tubes
V6A V7A V8A V9A V10A V11A	Western Electric No. 258B (Rectifier) Vacuum Tubes
V12A	Western Electric No. 287A Vacuum Tube
V13A	Western Electric No. 271A Tube
V14A	Western Electric No. 242A Tube
V15A	RCA 84 Tube

NOTES ON SCHEMATIC (FIG 8)

Note 1--For (100W) R.T.E. with the antenna connected directly to terminal No. 19, short-circuit terminals Nos. 6 and 7 and omit ESL-608320 entirely.

Note 2--This connection is to be removed for the (100W) R.T.E.

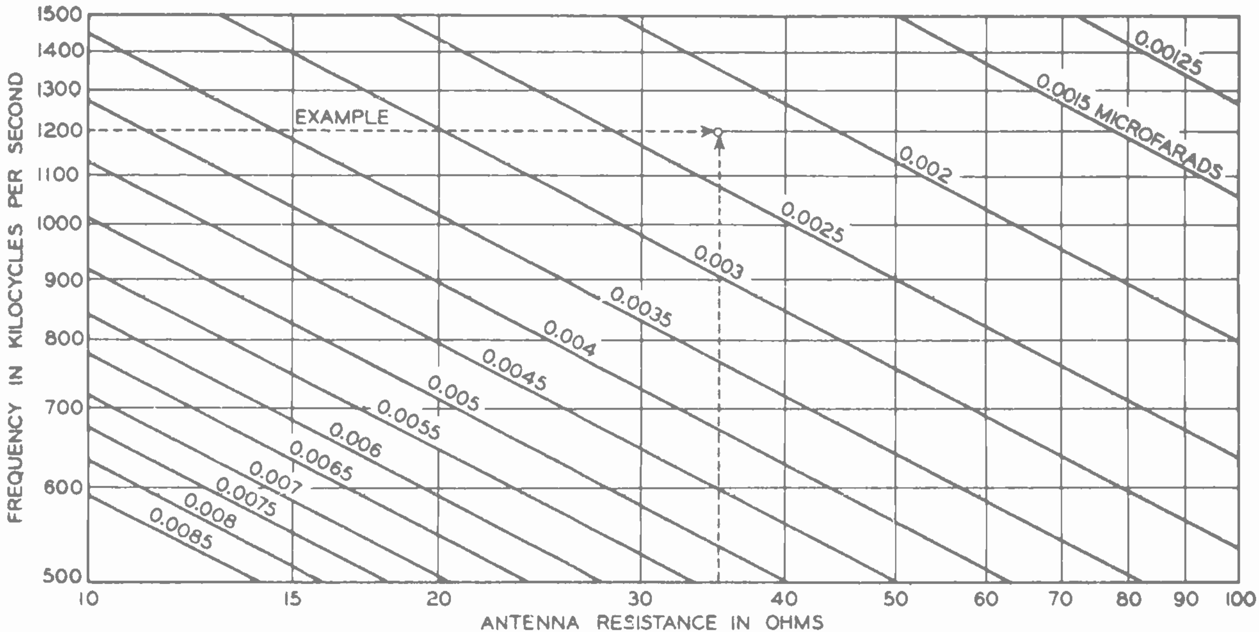
Note 3--All interconnecting wires and wires external to the equipment are shown thus --•--•--• These wires are not used in the (100W) R.T.E.

Note 4--When the D-98653 Oscillator Modulator Unit is followed a 71B Amplifier, remove condensers C21.1A to C21.5A, inclusive.

Note 5--These connections are to be used only in the (100W) R.T.E.

Notes

(These extra pages are provided for your use in taking special notes)



Example: For an antenna of 35 ohms at a frequency of 1,200 kilocycles, the exact value of coupling capacity would be 0.00225 MF. Use the nearest larger value obtainable with the condensers furnished, i.e., 0.0025MF.

Tuning Chart for Antenna Coupling Capacity, C21A

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT NO. 3

RADIO LAWS

LESSON NO. 13

RADIO LAW

.....guardian of the air.

Imagine that you have been called to attend a meeting of several hundred people. The purpose of the meeting is to decide important issues having bearing upon life in your community. Speakers are to address the throng. Entertainment of a varied nature and an orchestra are included in the program.

You arrive at the meeting place and find confusion and a conglomeration of noises. All the speakers are shouting, waving their arms, and competing with one another. The orchestra is blaring---the entertainers are going through their acts---people are arguing and fighting. It is evident to you that absolutely nothing is being accomplished.

Taking matters into your own hands, you quickly organize a small but efficient group of level-headed men. You instruct them to quiet each noisy group, and, as they depart to carry out your orders, you mount the speaker's stand.

Gradually confusion and noise give way to order and quiet. You address the throng and lay down the law, explaining that results can be accomplished only through mutual cooperation. You give each speaker an opportunity to deliver his talk. You intersperse music and entertainment with business. When the meeting is over, definite beneficial results have been obtained.

Were it not for Radio Laws, broadcasting and all forms of radiotelephone and radiotelegraph communications would present a picture of utter confusion. The fact that there ARE radio laws that are enforced has made possible an orderly growth of the great radio and communications industry. To take full advantage of the opportunities in the radio industry, YOU MUST KNOW THE LAWS. Master them thoroughly from beginning to end.

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Lesson Thirteen

RADIO LAWS

The need for government regulation of radio broadcasting was realized soon after communication with wireless electrical equipment became known to the general public. The early experiments by such famous men as Marconi, Hertz, and Edison paved the way for the rise of a gigantic communication system. Without regulation or control by the governments of the world, it would have been impossible for this communication system to provide mankind with all the advantages it has to offer.

The initial regulatory body to convene for the purpose of discussing problems pertaining to radio communication was the First International Radiotelegraphic Conference, held in Berlin on August 4, 1903. Regulations agreeable to the nations represented were drafted, and it was suggested that each nation establish some method of controlling radio communications within its own boundaries. Great Britain passed the Wireless Telegraph Act on August 15, 1904. The United States government passed an act on June 24, 1910 requiring certain passenger ships to carry wireless equipment and operators. Following these pioneer steps, all governments in the world became conscious of the necessity for internal radio regulations and cooperative regulations between nations. The World War succeeding this period prevented world-wide agreement on laws governing radio; however, each country enforced rigid governmental control of its own wireless facilities.

Following the World War, the United States adopted various rules of an inconsequential nature until 1927 when the Federal Radio Commission was formed. This Commission immediately drafted and enforced the Radio Act of 1927, an Act superseding all previous regulations and for the purpose of maintaining government control over the allocation and use of frequencies for radio communication in the United States.

On July 11, 1934, the Federal Communications Commission was organized to succeed the Federal Radio Commission and also to regulate wire telephony and telegraphy. This body is now in existence and consists of a group of Commissioners to whom is entrusted the control of radio and wire facilities in the United States. The Federal Communications Commission immediately passed the Communications Act of 1934 to succeed the Radio Act of 1927. The Act of 1934 is still in effect (as amended); it is necessary that every radio operator be familiar with its contents.

On February 1, 1938, representatives from 71 countries convened at Cairo, Egypt for the purpose of revising previous international rules adopted at Madrid in 1932. The International Telecommunications Conference (Cairo) adopted regulations concerning radio practice and procedure for the purpose of binding all important nations of the world into common agreement on the allocation of frequencies and rules governing the use of radio on the high seas. Since the United States was a member of this conference and approved the rules adopted, the contents of the treaty are binding on the United States.

The student must become familiar with the Communications Act of 1934, the Cairo Conference Regulations, and the rules and regulations imposed by the Federal Communications Commission. Later in your training we shall point out the more important parts of this lesson; but for the present, read it through carefully at least twice. Only the important extracts are given; you will find that some of the section, article, and rule numbers are omitted. There will be no examination covering this material.

STANDARDS OF GOOD ENGINEERING PRACTICE CONCERNING STANDARD BROADCAST STATIONS (550-1600 KC.)

APPROVED JUNE 23, 1939; EFFECTIVE AUGUST 1, 1939.

1. INTRODUCTION

There are presented herein the Standards of Good Engineering Practice giving interpretations and further considerations concerning the Rules and Regulations of the Communications Commission governing standard broadcast stations. While the Rules and Regulations form the basis of good engineering practice, these standards may go beyond the Rules and Regulations and set up engineering principles for consideration of various allocation problems. These standards have been approved by the Commission and thus are considered as reflecting the opinion of the Commission in all matters involved.

The Rules and Regulations contain references to these standards, however, as further standards may be issued after the Rules and Regulations are published, the absence of such reference does not relieve the responsibility of meeting the requirements specified herein. The Standards of Good Engineering Practice are collected in this publication for the convenience of all considering broadcast station operation and problems.

The Standards of Good Engineering Practice set forth herein are those deemed necessary for the construction and operation of standard broadcast stations to meet the requirements of technical regulations and for operation in public interest along technical lines not specifically enunciated in the Regulations. These standards are based on the best engineering data available from evidence supplied in formal and informal hearings and extensive surveys conducted in the field by the Commission's personnel. Numerous informal conferences have been held with radio engineers, manufacturers of radio equipment and others for the guidance of the Commission in the formulation of these standards.

These standards are complete in themselves and supersede any previous announcements or policies which may have been enunciated by the Commission on engineering matters concerning standard broadcast stations.

While these standards provide for flexibility and set forth the conditions under which they are applicable, it is not expected that material deviation therefrom as to fundamental principles will be recognized unless full information is submitted as to the reasonableness of such departure and the need therefor.

These Standards of Good Engineering Practice will necessarily change as progress is made in the art and accordingly it will be necessary to make revisions from time to time. The Commission will accumulate and analyze engineering data available as to the progress of the art so that its standards may be kept current with the developments.

2. FURTHER REQUIREMENTS FOR POWER MEASUREMENTS

Section 3.54 states that the antenna input power determined by direct measurement is the square of the antenna current times the antenna resistance at the point where the current is measured and at the operating frequency, and sets forth certain requirements relative to the determination of the resistance and measurement of the antenna current.

The Commission does not specify any particular method of making antenna resistance measurements. Measurements made by any standard method will be accepted provided satisfactory evidence is submitted in accordance with the following as to the procedure used, accuracy of the instruments and qualifications of the engineer conducting the measurements.

The resistance variation method, substitution method and bridge method are acceptable methods for measuring the total antenna resistance and the following general instructions are given as a guide.

The apparatus required is as follows:

(a) Radio frequency generator to cover the frequency range necessary, power 50 watts or required power when using bridge method.

(b) Wavemeter for broadcasting frequency, accuracy 0.25%.

(c) Decade resistor having steps of units, tens, and hundreds ohms resistance, or equivalent, accuracy 1.0%.

(d) Radio frequency galvanometer or milliammeter of approved type, accuracy 2.0%.

(e) Approved tuning condenser of approximately 0.001 Mfd. capacity and tuning inductance of approximately 60 Mh.

(f) Or suitable bridge if this method is used.

The broadcast transmitter is not usually satisfactory for use as the source of radio frequencies. The maximum power dissipated in the antenna while making measurements should not be over 10% of the power available from the radio frequency generator.

An accurate determination of the antenna resistance can only be made by taking a series of measurements each for a different frequency. From 10 to 12 resistance measurements covering a band 50 to 60 kc. wide with the operating frequency near the middle of the band must be made to give data from which accurate results may be obtained. The values measured should be plotted with frequency as abscissa and resistance in ohms as ordinate and a smooth curve drawn. The point on the ordinate where this curve intersects the operating frequency gives the value of the antenna resistance.

In order to comply with the provisions of Sec. 3.54 the following data should be submitted in duplicate to the Commission in affidavit form, accompanied by duplicate copies of FCC Form 306 properly executed:

1. Complete data taken.

2. The graph drawn.

3. Description of method used to take readings (include schematic circuit diagrams of the measurement circuit and of the antenna system showing point of measurement and location in circuit of both regular and remote antenna ammeters).

4. Manufacturer's name of each calibrated instrument used and manufacturer's rated accuracy.

5. Accuracy, date and by whom each instrument was last calibrated.

6. Qualifications of engineer making measurements.

Licensees of broadcast stations authorized to employ directional antenna systems desiring to determine the operating power by direct measurement of the antenna power shall determine the resistance by the following method:

Measure the resistance at the point of common radio frequency input to the directional antenna system. The following conditions and procedure shall obtain:

(a) The antenna shall be finally adjusted for the required pattern.

(b) The reactance at the operating frequency and at the point of measurement shall be adjusted to zero or as near thereto as practical.

(c) Suitable radio frequency bridge or other method shall be employed to determine the resistance and reactance at the point of common radio frequency input in the same manner as set forth above for a single antenna.

(d) Resistance and reactance measurements at approximately 5, 10, 15, and 20 kc. on each side of the operating frequency shall be made. The values measured shall be plotted and the resistance at the operating frequency determined in the same manner as for a single element antenna.

(e) A permanently installed antenna ammeter shall be placed in each element of the system as well as at the point of measurement of resistance with the remote reading ammeters located in the transmitter room. The application for authority to determine the power by the direct method shall specify not only the current at the point of resistance measurement for the authorized input power (I^2R in accordance with Section 3.54 and (f) below), but also the current of each element of the system when adjusted for the required pattern and for the authorized operating power.

(f) The license for a station of power of 5 kw. or under, which employs a directional antenna and determines the power by the direct method, will specify the antenna resistance as

TABLE A.[1]—*Power rating of vacuum tubes for high-level modulation or plate modulation in the last radio stage*

Power rating (watts)	Amperex	Collins	De Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
50				35T 50T		GL-276-A	HK-54		808	RK-32 RK-35 RK-37 RK-47 RK-51 RK-52	35T		211D 211E 248A 276A		T55 841A
75	HF-100 203-A 211 838 852 860	C-203A C-211	503-A 511 552 560		F-303-A F-311-A F-352-A	GL-203-A GL-211 GL-242-C GL-835 GL-838 GL-860	154	203-A 211 852 860	203-A 211 838 850 852 860	RK-58		303-A 311 361-A 938 952	242A 242B 242C 250A 261A 284A 295A	WL-195 WL-196 WL-203-A WL-211 WL-469 WL-838 WL-850 WL-860	203A 211 211C
100				100TH 100TL 75T	F-102-A F-108-A		254			RK-36 RK-38 RK-48	100TH			WL-460 WL-468	
125	HF-200 203-H 211-C 211-D 211-H 805	C-200 C-201 C-211-D		150-T	F-123-A	GL-803 GL-805 GL-810			803 805 810	RK-57 RK-28A		905		WL-463 WL-803 WL-805 WL-810	T-155 HD-203A
250	204-A HF-300 212-E	C-204-A C-300	504-A 561 571		F-204-A F-212-E F-331-A	GL-204-A GL-806 GL-861	354 HK354A HK354C HK354D HK354E HK354F	204-A 212-D 831 861	204A 806 831 861	RK-63		304-A 312-E	212-D 212-E	WL-204-A WL-806 WL-861	T200 204A 814 822
350	270-A 849		549	250TH 250TL 300T	F-100-A F-849	GL-849		849	849		250TH	949	270-A	WL-849	
500	251-A			450TL 450TH		GL-833	255 654		833		450TH		251-A 357-A	WL-833	

TABLE A.—*Power rating of vacuum tubes for high-level modulation or plate modulation in the last radio stage*—Continued

Power rating (watts)	Amperex	Collins	De Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
750	279-A 849-A 849-H 851	C-849-A C-849-H	551	500T	F-121-A F-351-A	GL-851	2054-A	851	851			951	279A	WL-851	
1,000	846			750TL 750TH	F-128-A F-346-A	GL-846	1554	846	846					WL-846	
2,500	228A 1652 HF3000		520-B 520-M	1500T	F-328-A F-328-B F-348 F-3652-A	GL-207 GL-889	3054	820-B	520B 1652				228A		
5,000	207 220-C 848 343A 863 891 892 892R		507 548 563	2000T	F-126-A F-307-A F-320-A F-320-B F-343-A F-348-A F-363-A F-892 F-892-R	GL-891 GL-892 GL-892-R		207 848 863	207 848 863 891 892 892R				220B 220C 343-A 220CA 343-AA	WL-207 WL-891 WL-892 WL-892-R	
10,000	232-B 342A				F-101-B F-110-A F-110-X F-116-A F-131-A F-332-A F-332-B F-332-C F-342-A F-358-A	GL-880 GL-893			858 893-R				232A 232B 342A	WL-858 WL-893 WL-893-R	
25,000					F-117-B F-124-A F-124-R										
40,000					F-862 F-898	GL-862 GL-898			862 898				298-A	WL-862 WL-898	
100,000													320-A		

TABLE B 1—*Power rating of vacuum tubes for low-level modulation or last radio stage operating as linear power amplifier*

Power rating (watts)	Amperex	Collins	De-Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
25				75T		GL-203-A	154		203A	RK-32 RK-35 RK-37 RK-47 RK-51				WL-203-A	203A 211 211C 841A
50	HF-200 203-H 211-H 242-C			100TH 100TL 150-T	F-123-A	GL-242-C GL-803 GL-810	HK354 HK354A HK354C HK354D HK354E HK354F		803 806 810	RK-36 RK-38 RK-57 RK-58 RK-28A RK-48	100TH		242B 242C	WL-803 WL-810	HD203-A
75	HF300 212-E 204-A		504-A		F-204-A F-212-E	GL-204-A		204-A 212-D	204A	RK-63		304-A 312-E	212D 212E	WL-204-A	T200 204A 814 822
125	270-A 849		549	250TH 250TL 300T 450TL 450TH	F-100-A F-849	GL-833 GL-849	654	849	833 833-A 849		250TH 450TH	949	270A	WL-833 WL-849	
250	251-A 849-A 849-H 851	C-849-A C-849-H	551	500-T	F-121-A F-128-A F-351-A	GL-851	255 1554	851	851			951	251A	WL-851	
350				750TL 750TH											
500	279-A 846			1000UHF 1500T	F-346-A	GL-846	2054-A 3054	846	846				279A	WL-846	
1,000	228-A 1652 ZB3200		520-B 520-M	2000T	F-328-A F-328-B F-3652-A			820-B	520B 1652				228A		
1,250						GL-889									

TABLE B.—*Power rating of vacuum tubes for low-level modulation or last radio stage operating as linear power amplifier*—Continued

Power rating (watts)	Amperex	Collins	De-Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
2,500	207 220-C 343-A 863 892 892-R		507 569		F-126-A F-131-A F-307-A F-320-A F-320-B F-343-A F-363-A F-892 F-892-R	GL-207 GL-892 GL-892-R		207 863	207 863 892				220-B 220-C 220CA 343-A 343-AA	WL-207 WL-892	
5,000					F-358-A	GL-858			858					WL-858	
8,500	232-B 342-A				F-101-B F-110-A F-110-X F-116-A F-131-A F-332-A F-332-B F-332-C F-342-A								232-A 232-B 342-A		
10,000						GL-880 GL-893								WL-893 WL-893-R	
12,500					F-117-B F-124-A										
25,000					F-862 F-898	GL-862 GL-898			862 898				298-A	WL-862 WL-898	
75,000													320-A		

Table BC.[1]—*Power rating of vacuum tubes for low-level modulation in the last radio stage operating as linear power amplifier where efficiency approaches that of Class C operation*

Power rating (watts)	Amperex	Collins	De-Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
250													357-A		
2,500													220-CA		

Table C.[1]—*Power rating of vacuum tubes for grid modulation in the last radio stage (operating efficiency 25 percent)*

Power rating (watts)	Amperex	Collins	De-Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
25				75T											
50	212-E 241-B 270-A			250TH 250TL	F-212-E		354 HK354 A&C			RK-63	250TH		212-E 270-A		
100				300T			654								
125				450TL 500T 750TL			255				450TH				
250				1000UHF			1554								
330				1500T											
500				2000T			3054								

TABLE D.[1]—*Power rating of vacuum tubes for grid modulation in the last radio stage (operating efficiency 35 percent)*

Power rating (watts)	Amperex	Collins	De Forest	Eitel Mc-Cullough	Federal Telegraph	General Electric Co.	Heintz & Kaufman	Hygrade Sylvania	RCA Mfg. Co.	Raytheon Production Corp.	Sheldon Electric Co.	United Electronics	Western Electric	Westinghouse Electric & Manufacturing Co.	Taylor Tubes, Inc.
250	849-A 849-H														
1,000					F-328-A								228-A		
2,500	343-R 892 892-R	C-892			F-307-A F-892 F-892R				892						

[1] These tables apply only to tube ratings for use in the last radio stage of broadcast transmitters and may not be applicable to any other service.

If in an application to the Commission a vacuum tube of a type number and power rating not given in the foregoing tables is specified for operation in the last radio stage, it may be accepted provided there is also submitted to and approved by the Commission the manufacturer's rating of the vacuum tube for the system of modulation or class of service contemplated. These data must be supplied by the manufacturer. (See section 3.42 and "Requirements for Approval of Power Rating of Vacuum Tubes.")

92.5% of that determined at the point of common input in accordance with the above. The resistance specified for stations of a power over 5 kw. will be 95% of that determined at the point of common input.

3. POWER RATING OF VACUUM TUBES

Section 3.42 requires that the maximum rated carrier power of a standard broadcast transmitter shall be determined as the sum of the applicable power ratings of the vacuum tubes employed in the last radio stage. The approved power ratings of vacuum tubes for operation in the last radio stages of broadcast transmitters are fixed as set out in tables A, B, BC, C, and D.

4. REQUIREMENTS FOR APPROVAL OF POWER RATING OF VACUUM TUBES

Section 3.42 (c) requires that only vacuum tubes of approved rating be employed in the last radio stages of standard broadcast transmitters and that such approved ratings will be given only upon submission of the ratings by the manufacturer.

These ratings shall be supplied under oath in the following table form:

TABLE (A, B, C, OR D) - CLASS OF OPERATION

(State whether for plate modulation in the last radio stage, low level modulation, or grid modulation in the last radio stage. See section 3.52)

TYPE OF TUBE	Epn	Epmn	Ipn	Ipmn	RL	Ec	POWER RATING

The value of Epn and Ipn are those values that are recommended by the manufacturers for operation at the power rating specified. The values of Epmn and Ipmn are the maximum continuous operating values that the manufacturers will stand by their guarantee on the tube specified when used in the class of service above set out.

The power ratings given above are to apply when the following conditions and limitations as to operation prevail.

(1) The vacuum tubes are to be used in the last radio stage of standard broadcast transmitters.

(2) On the broadcast frequencies of 550 to 1600 kc., inclusive.

(3) The percentage of amplitude distortion or harmonic generation by the entire transmitter (from microphone terminals to antenna outputs) is not to be over 10 percent at 100 percent modulation. The radio harmonics are not to exceed the amount considered in accordance with good engineering practice. (See section 3.46 and "Construction, General Operation and Safety of Life Requirements" to comply with section 3.46)

(4) The ventilation, cooling, general condition of circuits with respect to tuning, impedance match, and maintenance are to be those encountered in the average broadcast station and in accordance with good engineering practice. The operation, general maintenance, and adjustments are to be those ordinarily encountered in a broadcast station.

(5) The regulation of the power supply is to be that which has been found to exist throughout the United States where broadcast stations are located or may be located.

(6) Table A should give the power rating for use in transmitter employing plate modulation in last radio stage under the above specified conditions. (See section 3.52, Table A)

(7) Table B should give the rating for use in transmitter employing low level modulation under the above specified conditions. (See section 3.52, Table B)

8. Tables C and D should give the power rating for use in transmitters employing grid modulation in the last radio stage under the above specified conditions and plate efficiencies of 25 and 35 percent, respectively. (Request for approval of ratings for grid modulation in the last radio stage should specify whether the operating efficiency is to be 25 percent or 35 percent. No tube will be approved for both efficiencies. See Section 3.52, Table C.)

9. The operating power of the transmitter is to be determined by sections 3.51, 3.52, 3.53, and 3.54.

10. Due consideration will be given to the general standard of rating so that tubes manufactured by different firms but having practically the same absolute characteristics will have approximately the same power rating.

11. The power rating should be an even power step as recognized by the Commission's plan of allocation (100 watts, 250 watts 500 watts, 1000 watts, 5 kw., 10 kw., 25 kw., 50 kw.), or other ratings established in "Power Rating of Vacuum Tubes". If any other rating is desired, it will be necessary to support the request with satisfactory reasons therefor.

5. PLATE EFFICIENCY OF LAST RADIO STAGE

Section 3.53 requires that in computing the operating power of standard broadcast stations by the indirect method the efficiency factors specified in section 3.52 shall apply in all cases and no distinction will be recognized due to the operating power being less than the maximum rated carrier power. (See section 3.52, Table A)

In compliance with this rule, standard broadcast stations permitted to determine the operating power by the indirect method in accordance with section 3.51 (b) and to employ greater daytime power than nighttime power shall maintain the same operating efficiency for both daytime and nighttime operation. (When different last radio stages with different systems of modulation are employed for the two powers, the same principle shall apply; that is, the power shall be determined by the plate input times the proper efficiency factor and the antenna current shall change in proportion to the square root of the change in power within 5 percent.)

To determine whether this condition obtains, the following procedure should be used:

The apparent antenna resistance should be computed from the daytime (highest power) operating constants and then the nighttime power in the antenna determined from the I^2R, using the apparent resistance previously determined. If this computed antenna power agrees with the nighttime operating power determined by the indirect method within plus or minus 5 percent, the station is considered as complying with the requirement of maintaining the same operating efficiency. In case the antenna current is subject to variation due to weather or other conditions, an attempt should be made to arrive at an average value for the purpose of the computations referred to herein.

6. OPERATING POWER TOLERANCE

Section 3.57 requires that except in case of emergency beyond the control of the licensee, the operating power of each standard broadcast station shall be maintained within the prescribed limits of the licensed power.

Each station shall be operated at all times as near the authorized power as practicable. However, in order to provide for variations in the power supply or other factors affecting the operating power which would necessitate continual adjustment to keep the operating power exactly the same as the authorized power, the operating power may be permitted to vary from 5 percent above to 10 percent below the authorized power for periods of short duration.

In addition to maintaining the operating power within the above limitations, broadcast stations employing directional antenna systems shall maintain the ratio of the antenna currents in the elements of the system within 5 percent of that specified by the terms of the license or other instrument of authorization.

7. CONSTRUCTION, GENERAL OPERATION, AND SAFETY OF LIFE REQUIREMENTS

Section 3.46 requires that the transmitter proper and associated transmitting equipment of each broadcast station shall be designed, constructed, and operated in accordance with the standards of good engineering practice in addition to the specific requirements of the Rules and Regulations of the Commission.

The specifications deemed necessary to meet the requirements of the Rules and Regulations and good engineering practice with respect to design, construction, and operation of standard broadcast stations are set forth below. These specifications will be changed from time to time as the state of the art and the need arises for modified or additional specifications.

A. DESIGN. The general design of standard broadcast transmitting equipment [main studio microphone (including telephone lines, if used, as to performance only) to antenna output] shall be in accordance with the following specifications. For the points not specifically covered below, the principles set out shall be followed:

The equipment shall be so designed that:

1. The maximum rated carrier power (determined by section 3.42) is in accordance with the requirements of section 3.41,

2. The equipment is capable of satisfactory operation at the authorized operating power or the proposed operating power with modulation of at least 85 or 95 percent with no more distortion than given in (3) below.

3. The total audio frequency distortion from microphone terminals, including microphone amplifier, to antenna output does not exceed 5 percent harmonics (voltage measurements of arithmetical sum or r.s.s.) when modulated from 0 to 84 percent, and not over 7.5 percent harmonics (voltage measurements of arithmetical sum or r.s.s.) when modulating 85 percent to 95 percent (distortion shall be measured with modulating frequencies of 50, 100, 400, 1000, 5000, and 7500 cycles up to tenth harmonic or 16000 cycles, or any intermediate frequency that readings on the frequencies indicate is desirable.)

4. The audio frequency transmitting characteristics of the equipment from the microphone terminals (including microphone amplifier unless microphone frequency correction is included; in which event proper allowance shall be made accordingly) to the antenna output does not depart more than 2 decibels from that at 1000 cycles between 100 and 5000 cycles.

5. The carrier shift (current) at any percentage of modulation does not exceed 5 percent.

6. The carrier hum and extraneous noise (exclusive of microphone and studio noises) level (unweighted r.s.s.) is at least 50 decibels below 100 percent modulation for the frequency band of 150 to 5000 cycles and at least 40 decibels down outside this range.

7. The transmitter shall be equipped with suitable indicating instruments in accordance with the requirements of section 3.58 and any other instruments necessary for the proper adjustment and operation of the equipment.

8. Adequate provision is made for varying the transmitter power output between sufficient limits to compensate for excessive variations in line voltage, or other factors which may affect the power output.

9. The transmitter is equipped with automatic frequency control equipment capable of maintaining the operating frequency within the limit specified by section 3.59.

a. The maximum temperature variation at the crystal from the normal operating temperature shall not be greater than:

1. Plus or minus 0.1° C. when an X or Y cut crystal is employed, or

2. Plus or minus 1.0° C. when low temperature coefficient crystal is employed. (See "Use of Low Temperature Coefficient Crystals by Broadcast Stations")

b. Unless otherwise authorized, a thermometer shall be installed in such manner that the temperature at the crystal can be accurately measured within 0.05° C. for X or Y cut crystal or 0.5° for low temperature coefficient crystal.

c. It is preferable that the tank circuit of the oscillator tube be installed in the temperature controlled chamber.

10. Means are provided for connection and continuous operation of approved modulation monitor and approved frequency monitor.

a. The radio frequency energy for operation of the approved frequency monitor shall be obtained from a radio frequency stage prior to the modulated stage unless the monitor is of such design as to permit satisfactory operation when otherwise connected and the monitor circuits shall be such that the carrier is not heterodyned thereby.

11. Adequate margin is provided in all component parts to avoid overheating at the maximum rated power output.

B. CONSTRUCTION. In general, the transmitter shall be constructed either on racks and panels or in totally enclosed frames (the final stages of high power transmitters may be assembled in open frames provided the equipment is enclosed by a protective fence), protected as required by article 810 of the National Electrical Code and as set forth below:

1. Means shall be provided for making all tuning adjustments, requiring voltages in excess of 350 volts to be applied to the circuit, from the front of the panels with all access doors closed.

2. Proper bleeder resistors or other automatic means shall be installed across all the condenser banks to remove any charge which may remain after the high voltage circuit is opened (in certain instances the plate circuit of the tubes may provide such protection; however, individual approval of such shall be obtained by the manufacturer in case of standard equipment, and the licensee in case of composite equipment).

3. All plate supply and other high voltage equipment, including transformers, filters, rectifiers and motor generators, shall be protected so as to prevent injury to operating personnel.

a. Commutator guards shall be provided on all high voltage rotating machinery (coupling guards on motor generators, although desirable, are not required).

b. Power equipment and control panels of the transmitter shall meet the above requirements (exposed 220 volt AC switching equipment on the front of the power control panels is not recommended; however, it is not prohibited).

c. Power equipment located at a broadcast station but not directly associated with the transmitter (not purchased as part of same), such as power distribution panels, control equipment on indoor or outdoor stations and the substations associated therewith, are not under the jurisdiction of the Commission; therefore, section 3.46 does not apply.

d. It is not necessary to protect the equipment in the antenna tuning house and the base of the antenna with screens and interlocks, provided the doors to the tuning house and antenna base are fenced and locked at all times, with the keys in the possession of the operator on duty at the transmitter. Ungrounded fencing or wires should be effectively grounded, either directly or through proper static leaks. Lightning protection for the antenna system is not specifically required, but should be installed.

e. The antenna, antenna lead-in, counterpoise (if used), etc., shall be installed so as not to present a hazard. The antenna may be located close by or at a distance from the transmitter building. A properly designed and terminated transmission line should be used between the transmitter and the antenna when located at a distance.

4. Metering Equipment. a. All instruments having more than 1000 volts potential to ground on the movement shall be protected by a cage or cover in addition to the regular case. (Some instruments are designed by the manufacturer to operate safely with voltages in excess of 1000 volts on

the movement. If it can be shown by the manufacturer's rating that the instrument will operate safely at the applied potential, additional protection is not necessary).

b. In case the plate voltmeter is located on the low potential side of the multiplier resistor with one terminal of the instrument at or less than 1000 volts above ground, no protective case is required. However, it is good practice to protect voltmeters subject to more than 5000 volts with suitable over-voltage protective devices across the instrument terminals in case the winding opens.

c. The antenna ammeters (both regular and remote and any other radio frequency instrument which it is necessary for the operator to read) shall be so installed as to be easily and accurately read without the operator having to risk contact with circuits carrying high potential radio frequency energy.

C. WIRING AND SHIELDING. 1. The transmitter panels or units shall be wired in accordance with standard switchboard practice, either with insulated leads properly cabled and supported or with rigid bus bar properly insulated and protected.

2. Wiring between units of the transmitter, with the exception of circuits carrying radio frequency energy, shall be installed in conduits or approved fiber or metal raceways to protect it from mechanical injury.

3. Circuits carrying low level radio frequency energy between units shall be either concentric tube, two wire balanced lines, or properly shielded to prevent the pickup of modulated radio frequency energy from the output circuits.

4. Each stage (including the oscillator) preceding the modulated stage shall be properly shielded and filtered to prevent unintentional feedback from any circuit following the modulated stage (an exception to this requirement may be made in the case of high level modulated transmitters of approved manufacture which have been properly engineered to prevent reaction).

5. The crystal chamber, together with the conductor or conductors to the oscillator circuit shall be totally shielded.

6. The monitors and the radio frequency lines to the transmitter shall be thoroughly shielded.

D. INSTALLATION. 1. The installation shall be made in suitable quarters.

2. Since an operator must be on duty during operation, suitable facilities for his welfare and comfort shall be provided.

E. SPARE TUBES. A spare tube of every type employed in the transmitter and frequency and modulation monitors shall be kept on hand. When more than one tube of any type are employed, the following table determines the number of spares of that type required:

NUMBER OF EACH TYPE EMPLOYED	SPARES REQUIRED
1 or 2	1
3 to 5	2
6 to 8	3
9 or more	4

F. OPERATION. In addition to the specific requirements of the rules governing standard broadcast stations, the following operating requirements shall be observed:

1. The maximum percentage of modulation shall be maintained at as high level as practicable without causing undue audio frequency harmonics, which shall not be in excess of 10 percent when operating with 85 percent modulation.

2. Spurious emissions, including radio frequency harmonics and audio frequency harmonics, shall be maintained at as low a level as practicable at all times in accordance with good engineering practice.

3. In the event interference is caused to other stations by modulating frequencies in excess of 7500 cycles or spurious emissions, including radio frequency harmonics and audio frequency harmonics outside the band plus or minus 7500 cycles of the authorized carrier frequency, the licensee shall install equipment or make adjustments which limit the emissions to within this band or to such an extent above 7500 cycles as to reduce the interference to where it is no longer objectionable.

4. The operating power shall be maintained within the limits of 5 percent above and 10 percent below the authorized operating power and shall be maintained as near as practicable to the authorized operating power.

5. Licensees of broadcast stations employing directional antenna systems shall maintain the ratio of the currents in the elements of the array within 5 percent of that specified by the terms of the license or other instrument of authorization.

6. In case of excessive shift in operating frequency during warm-up periods, the crystal oscillator shall be operated continuously. The automatic temperature control circuits should be operated continuously under all circumstances.

G. STUDIO EQUIPMENT. The studio equipment shall be subject to all the above requirements where applicable except as follows:

1. If it is properly covered by an underwriter's certificate, it will be considered as satisfying the safety requirements.

2. Section 8191 of article 810 of the National Electrical Code shall apply for voltages only when in excess of 500 volts.

No specific requirements are made relative to the design and accoustical treatment. However, the studios and particularly the main studio should be in accordance with the standard practice for the class of station concerned, keeping the noise level as low as reasonably possible.

8. INDICATING INSTRUMENTS PURSUANT TO SECTION 3.58

Section 3.58 requires that each standard broadcast station shall be equipped with suitable indicating instruments of accepted accuracy to measure the antenna current, direct plate circuit voltage, and the direct plate curcuit current of the last radio stage.

The following requirements and specifications shall apply to indicating instruments used by standard broadcast stations in compliance with this rule:

A. Instruments indicating the plate current or plate voltage of the last radio stage (linear scale instruments), shall meet the following specifications:

1. Length of scale shall be not less than 2.3 inches.

2. Accuracy shall be at least 2 percent of the full scale reading.

3. The maximum rating of the meter shall be such that it does not read off scale during modulation.

4. Scale shall have at least 40 divisions.

5. Full scale reading shall not be greater than five times the minimum normal indication.

B. Instruments indicating the antenna current shall meet the following specifications:

1. Instruments having logarithmic or square law scales.

a. Shall meet the same requirements as 1, 2, and 3 above for linear scale instruments.

b. Full scale reading shall not be greater than three times the minimum normal indication.

c. No scale division above one-third full scale reading (in amperes) shall be greater than one-thirtieth of the full scale reading. (Example: An ammeter meeting requirement

(a) above having full scale reading of 6 amperes is acceptable for reading currents from 2 to 6 amperes, provided no scale division between 2 and 6 amperes is greater than one-thirtieth of 6 amperes, or 0.2 ampere.)

2. Radio frequency instruments having expanded scales.

a. Shall meet same requirements as 1, 2, and 3 for linear scale instruments.

b. Full scale reading shall not be greater than five times the minimum normal indication.

c. No scale division above one-fifth full scale reading (in amperes) shall be greater than one-fiftieth of the full scale reading. (Example: An ammeter meeting the requirement (a) above is acceptable for indicating currents from 1 to 5 amperes, provided no division between 1 and 5 amperes is greater than one-fiftieth of 5 amperes, 0.1 ampere.)

d. Manufacturers of instruments of the expanded scale type must submit data to the Commission showing that these instruments have acceptable expanded scales, and the type number of these instruments must include suitable designation.

3. Remote reading antenna ammeters may be employed and the indications logged as the antenna current in accordance with the following:

a. Remote reading antenna ammeters may be provided by:

1. Inserting second thermocouple directly in the antenna circuit with remote leads to the indicating instrument.

2. Inductive coupling to thermocouple or other device for providing direct current to indicating instrument.

3. Capacity coupling to thermocouple or other device for providing direct current to indicating instrument.

4. Current transformer connected to second thermocouple or other device for providing direct current to indicating instrument.

5. Using transmission line current meter at transmitter as remote reading ammeter. See paragraph (h) below.

6. Using indications of phase monitor for determining the ratio of antenna currents in the case of directional antennas, provided the indicating instruments in the unit are connected directly in the current sampling circuits with no other shunt circuits of any nature.

b. A thermocouple type ammeter meeting the above requirements shall be permanently installed in the antenna circuit. (This thermocouple ammeter may be so connected that it is short circuited or open circuited when not actually being read. If open circuited, a make-before-break switch must be employed.)

c. The remote ammeter shall be connected at the same point in the antenna circuit as the thermocouple ammeter or shall be so connected and calibrated as to read in amperes within 2 percent of this meter over the entire range above one-third or one-fifth full scale. See sections B 1 (c) and B 2 (c) above respectively.

d. The regular antenna ammeter shall be above the coupling to the remote meter in the antenna circuit so it does not read the current to ground through the remote meter.

e. All remote meters shall meet the same requirements as the regular antenna ammeter with respect to scale accuracy, etc.

f. Calibration shall be checked against the regular meter at least once a week.

g. All remote meters shall be provided with shielding or filters as necessary to prevent any feedback from the antenna to the transmitter.

h. In the case of shunt excited antennas, the transmission line current meter at the transmitter may be considered as the remote antenna ammeter provided the transmission line is terminated directly into the excitation circuit feed line, which shall employ series tuning only (no shunt circuits of any type shall be employed), and insofar as qualified and equipped instrument repair service. In either case the instrument must be resealed with the symbol or trade mark of the repair service and a certificate of calibration supplied therewith.

i. Remote reading antenna ammeters employing vacuum tube rectifiers are acceptable, provided:

1. The indicating instruments shall meet all the above requirements for linear scale instruments.

2. Data are submitted under oath showing the unit has an overall accuracy of at least 2 percent of the full scale reading.

3. The installation, calibration, and checking are in accordance with the above requirements.

j. In the event there is any question as to the method of providing or the accuracy of the remote meter, the burden of proof of satisfactory performance shall be upon the licensee and the manufacturer of the equipment.

C. Stations determining power by the indirect method may log the transmission line current in lieu of the antenna current provided the instrument meets the above requirements for antenna ammeters, and further provided that the ratio between the transmission line current and the antenna current is entered each time in the log. In case the station is authorized for the same operating power for both day and nighttime operation, this ratio shall be checked at least once daily. Stations which are authorized to operate with nighttime power different from the daytime power shall check the ratio for each power at least once daily.

D. No instruments indicating the plate current or plate voltage of the last radio stage, the antenna current or the transmission line current when logged, in lieu of the antenna current shall be changed or replaced without written authority of the Commission, except by instruments of the same make, type, maximum scale readings, and accuracy. Requests for authority to change an instrument may be made by letter or telegram giving the manufacturer's name, type number, serial number and full scale reading of the proposed instrument and the values of current or voltage the instrument will be employed to indicate. Requests for temporary authority to operate without an instrument or with a substitute instrument may be made by letter or telegram stating the necessity therefor and the period involved.

E. No instrument, the seal of which has been broken, or the accuracy of which is questionable, shall be employed. Any instrument which was not originally sealed by the manufacturer that has been opened shall not be used until it has been recalibrated and sealed in accordance with the following: Repairs and recalibration of instruments shall be made by the manufacturer, by an authorized instrument repair service of the manufacturer or by some other properly qualified and equipped instrument repair service. In either case the instrument must be resealed with the symbol or trade mark of the repair service and a certificate of calibration supplied therewith.

F. Since it is usually impractical to measure the actual antenna current of a shunt excited antenna system, the current measured at the input of the excitation circuit feed line is accepted as the antenna current.

G. Recording instruments may be employed in addition to the indicating instruments to record the antenna current and the direct plate current and direct plate voltage of the last radio stage provided that they do not affect the operation of the circuits or accuracy of the indicating instruments. If the records are to be used in any proceedings before the Commission as representation of operation with respect to plate or antenna current and plate voltage only, the accuracy must be the equivalent of the indicating instruments and the calibration shall be checked at such intervals as to insure the retention of the accuracy.

H. The function of each instrument shall be clearly and permanently shown on the instrument itself or on the panel immediately adjacent thereto.

PART 3 - RULES GOVERNING STANDARD AND HIGH-FREQUENCY BROADCAST STATIONS

SUBPART A - RULES GOVERNING STANDARD BROADCAST STATIONS

DEFINITIONS

3.1 STANDARD BROADCAST STATION. The term "standard broadcast station" means a station licensed for the transmission of radiotelephone emissions primarily intended to be received by the general public and operated on a channel in the band 550-1600 kilocycles, inclusive.

3.2 STANDARD BROADCAST BAND. The term "standard broadcast band" means the band of frequencies extending from 550-1600 kilocycles, inclusive, both 550 kilocycles and 1600 kilocycles being the carrier frequencies of broadcast channels.

3.3 STANDARD BROADCAST CHANNEL. The term "standard broadcast channel" means the band of frequencies occupied by the carrier and two sidebands of a broadcast signal with the carrier frequency at the center. Channels shall be designated by their assigned carrier frequencies. Carrier frequencies assigned to standard broadcast stations shall begin at 550 kilocycles and be in successive steps of 10 kilocycles.

3.4 DOMINANT STATION. The term "dominant station" means a class I station, as hereinafter defined, operating on a clear channel.

3.5 SECONDARY STATION. The term "secondary station" means any station except a class I station operating on a clear channel.

3.6 DAYTIME. The term "daytime" means that period of time between local sunrise and local sunset.

3.7 NIGHTTIME. The term "nighttime" means that period of time between local sunset and 12 midnight local standard time.

3.8 SUNRISE AND SUNSET. The terms "sunrise and sunset" mean, for each particular location and during any particular month, the average time of sunrise and sunset as specified in the license of a broadcast station. (For tabulation of average sunrise and sunset times for each month at various points in the United States, see "Average Sunrise and Sunset Times.")

3.9 BROADCAST DAY. The term "broadcast day" means that period of time between local sunrise and 12 midnight local standard time.

3.10 EXPERIMENTAL PERIOD. The term "experimental period" means that time between 12 midnight and local sunrise. This period may be used for experimental purposes in testing and maintaining apparatus by the licensee of any standard broadcast station on its assigned frequency and with its authorized power, provided no interference is caused to other stations maintaining a regular operating schedule within such period. No station licensed for "daytime" or "specified hours" of operation may broadcast any regular or scheduled program during this period.

3.11 SERVICE AREAS. (a) The term "primary service area" of a broadcast station means the area in which the ground wave is not subject to objectionable interference or objectionable fading.

(b) The term "secondary service area" of a broadcast station means the area served by the sky wave and not subject to objectionable interference. The signal is subject to intermittent variations in intensity.

(c) The term "intermittent service area" of a broadcast station means the area receiving service from the ground wave but beyond the primary service area and subject to some interference and fading.

3.12 MAIN STUDIO. The term "main studio" means, as to any station, the studio from which the majority of its local programs originate, and-or from which a majority of its station announcements are made of programs originating at remote points.

3.13 PORTABLE TRANSMITTER. The term "portable transmitter" means a transmitter so constructed that it may be moved about conveniently from plate to place, and is in fact so moved about from time to time, but not ordinarily used while in motion. In the standard broadcast band, such a transmitter is used in making field intensity measurements for locating a transmitter site for a standard broadcast station. A portable broadcast station will not be licensed in the standard broadcast band for regular transmission of programs intended to be received by the public.

3.14 AUXILIARY TRANSMITTER. The term "auxiliary transmitter" means a transmitter maintained only for transmitting the regular programs of a station in case of failure of the main transmitter.

3.15 COMBINED AUDIO HARMONICS. The term "combined audio harmonics" means the arithmetical sum of the amplitudes of all the separate harmonic components. Root sum square harmonic readings may be accepted under conditions prescribed by the Commission.

3.16 EFFECTIVE FIELD. The term "effective field" or "effective field intensity" is the root-mean-square (RMS) value of the inverse distance fields at a distance of 1 mile from the antenna in all directions in the horizontal plane.

EQUIPMENT

3.41 MAXIMUM RATED CARRIER POWER; TOLERANCES. The maximum rated carrier power of a standard broadcast transmitter shall not be less than the authorized power nor shall it be greater than the value specified in the following table:

CLASS OF STATION	MAXIMUM POWER AUTHORIZED TO STATION	MAXIMUM RATED CARRIER POWER PERMITTED TO BE INSTALLED*
		Watts
Class IV.....	100 or 250 Watts........	250
Class III....	500 or 1,000 Watts......	1,000
	5,000 Watts.............	5,000
Class II.....	250, 500, or 1,000 Watts	1,000
	5,000 or 10,000 Watts...	10,000
	25,000 or 50,000 Watts..	50,000
Class I......	10,000 Watts............	10,000
	25,000 or 50,000 Watts..	50,000

*The maximum rated carrier power must be distinguished from the operating power. (See sections 2.18 and 2.19)

3.42 MAXIMUM RATED CARRIER POWER; HOW DETERMINED. The maximum rated carrier power of a standard broadcast transmitter shall be determined as the sum of the applicable power ratings of the vacuum tubes employed in the last radio stage.

(a) The power rating of vacuum tubes shall apply to transmitters employing the different classes of operation or systems of modulation as specified in Power Rating of Vacuum Tubes prescribed by the Commission.

(b) If the maximum rated carrier power of any broadcast transmitter, as determined by paragraph (a) of this section, does not give an exact rating as recognized in the Commission's plan of allocation, the nearest rating thereto shall apply to such transmitter.

(c) Authority will not be granted to employ, in the last radio stage of a standard broadcast transmitter, vacuum tubes from a manufacturer or of a type number not listed until the manufacturer's rating for the class of operation or system of modulation is submitted to and approved by the Commission. These data must be supplied by the manufacturer in accordance with Requirements for the Approval of the Power Rating of Vacuum Tubes, prescribed by the Commission.

3.43 CHANGES IN EQUIPMENT; AUTHORITY FOR. No licensee shall change, in the last radio stage; the number of vacuum tubes, to vacuum tubes of different power rating or class of operation; nor shall it change the system of modulation without the authority of the Commission.

3.44 OTHER CHANGES IN EQUIPMENT. Other changes except as provided for in these rules or Standards of Good Engineering Practice, prescribed by the Commission, which do not affect the maximum power rating or operating power of the transmitter or the operation or precision of the frequency control equipment may be made at any time without authority of the Commission, but in the next succeeding application for renewal of license such changes which affect the information already on file shall be shown in full.

3.45 RADIATING SYSTEM. (a) All applicants for new, additional, or different broadcast facilities and all licensees requesting authority to move the transmitter of an existing station shall specify a radiating system the efficiency of which complies with the requirements of good engineering practice for the class and power of the station. (Also see Use of Common Antenna by Standard Broadcast Stations or Another Radio Station.)

(b) The Commission will publish from time to time specifications deemed necessary to meet the requirements of good engineering practice. (See Minimum Antenna Heights or Field Intensity Requirements and Field Intensity Measurements in Allocation, sec. A.)

(c) No broadcast station licensee shall change the physical height of the transmitting antenna, or supporting structures, or make any changes in the radiating system which will measurably alter the radiation patterns, except upon written application to, and authority from, the Commission.

(d) The antenna and-or supporting structure shall be painted and illuminated in accordance with the specifications supplied by the Commission pursuant to section 303 (q) of the Communications Act of 1934, as amended. (See Standard Lamps and Paints)

(e) The simultaneous use of a common antenna or antenna structure by two standard broadcast stations or by a standard broadcast station and a station of any other class or service will not be authorized unless both stations are licensed to the same licensee. (See Use of Common Antenna by Standard Broadcast Stations or Another Radio Station.)

3.46 TRANSMITTER. (a) The transmitter proper and associated transmitting equipment of each broadcast station shall be designed, constructed, and operated in accordance with the standards of good engineering practice in all phases not otherwise specifically included in these regulations.

(b) The transmitter shall be wired and shielded in accordance with good engineering practice and shall be provided with safety features in accordance with the specifications of article 810 of the current National Electrical Code as approved by the American Standards Association.

(c) The station equipment shall be so operated, tuned, and adjusted that emissions are not radiated outside the authorized band which cause or which, in accordance with the Standards of Good Engineering practice, are considered as being capable of causing interference to the communications of other stations. The spurious emissions, including radio frequency harmonics and audio frequency harmonics, shall be maintained at as low level as required by good engineering practice. The audio distortion, audio frequency range, carrier hum, noise level, and other essential phases of the operation which control the external effects shall at all times conform to the requirements of good engineering practice.

(d) Whenever, in this section, the term "good engineering practice" is used, the specifications deemed necessary to meet the requirements thereof will be published from time to time. (See Construction, General Operation and Safety of Life Requirements.)

TECHNICAL OPERATION

3.51 OPERATING POWER; HOW DETERMINED. The operating power of each standard broadcast station shall be determined by:

(a) Direct measurement of the antenna power in accordance with section 3.54.

(1) Each new standard broadcast station.

(2) Each existing standard broadcast station after June 1, 1941.

(b) Indirect measurement by means of the plate input power to the last radio stage on a temporary basis in accordance with sections 3.52 and 3.53.

(1) In the case of existing standard broadcast stations and pending compliance with paragraph (a) (2) of this section;

(2) In case of an emergency where the licensed antenna has been damaged or destroyed by storm or other cause beyond the control of the licensee or pending completion of authorized changes in the antenna system.

(c) Upon making any change in the antenna system, or in the antenna current measuring instruments, or any other change which may change the characteristics of the antenna, the licensee shall immediately make a new determination of the antenna resistance (see section 3.54) and shall submit application for authority to determine power by the direct method on the basis of the new measurements.

3.52 OPERATING POWER; INDIRECT MEASUREMENT. The operating power determined by indirect measurement from the plate input power of the last radio stage is the product of the plate voltage (Ep), the total plate current of the last radio stage (Ip), and the proper factor (F) given in the following tables; that is:

OPERATING POWER = Ep × Ip × F

A. Factor to be used for stations employing plate modulation in the last radio stage.[1]

Maximum Rated Carrier Power Of Transmitter[2]	Factor (F) to be used in determining the operating power from the plate input power
100 - 1,000 Watts	0.70
5,000 and over Watts	.80

B. Factor to be used for stations of all powers using low-level modulation.[1]

Class of Power Amplifier In The Last Radio Stage	Factor (F) to be used in determining the operating power from the plate input power
Class B	.35
Class BC[3]	.65

C. Factors to be used for stations of all powers employing grid modulation in the last radio stage.[1]

Type of Tube In The Last Radio Stage	Factor (F) to be used in determining the operating power from the plate input power
Table C[1]	0.25
Table D[1]	.35

[1] See Power Rating of Vacuum Tubes.

[2] The maximum rated carrier power must be distinguished from the operating power. (See Sections 2.18 and 2.19).

[3] All linear amplifier operation where efficiency approaches that of Class C operation.

3.53 APPLICATION OF EFFICIENCY FACTORS. In computing operating power by indirect measurement the above factors shall apply in all cases, and no distinction will be recognized due to the operating power being less than the maximum rated carrier power. (See Plate Efficiency of Last Radio Stage.)

3.54 OPERATING POWER; DIRECT MEASUREMENT. The antenna input power determined by direct measurement is the square of the antenna current times the antenna resistance at the point where the current is measured and at the operating frequency. Direct measurement of the antenna input power will be accepted as the operating power of the station, provided the data on the antenna resistance measurements are submitted under oath giving detailed description of the method used and the data taken. The antenna current shall be measured by an ammeter of accepted accuracy. These data must be submitted to and approved by the Commission before any licensee will be authorized to operate by this method of power determination. The antenna ammeter shall not be changed to one of different type, maximum reading, or accuracy without the authority of the Commission. If any change is made in the antenna system or any change made which may affect the antenna system, the method of determining operating power shall be changed immediately to the indirect method. (See Further Requirements for Direct Measurements of Power.)

3.55 MODULATION. (a) A licensee of a broadcast station will not be authorized to operate a transmitter unless it is capable of delivering satisfactorily the authorized power with a modulation of at least 85 percent. When the transmitter is operated with 85 percent modulation, not over 10 percent combined audio frequency harmonics shall be generated by the transmitter.

(b) All broadcast stations shall have in operation a modulation monitor approved by the Commission.

(c) The operating percentage of modulation of all stations shall be maintained as high as possible consistent with good quality of transmission and good broadcast practice and in no case less than 85 percent on peaks of frequent recurrence during any selection which normally is transmitted at the highest level of the program under consideration.

(d) The Commission will, from time to time, publish the specifications, requirements for approval, and a list of approved modulation monitors. (See Approved Modulation Monitors and also Requirements for Approval of Modulation Monitors.)

3.56 MODULATION; DATA REQUIRED. A licensee of a broadcast station claiming a greater percentage of modulation than the fundamental design indicates can be procured shall submit full data showing the antenna input power by direct measurement and complete information, either oscillograms or other acceptable data, to show that a modulation of 85 percent or more, with not over 10 percent combined audio harmonics, can be obtained with the transmitter operated at the maximum authorized power.

3.57 OPERATING POWER; MAINTENANCE OF. The licensee of a broadcast station shall maintain the operating power of the station within the prescribed limits of the licensed power at all times except that in an emergency when, due to causes beyond the control of the licensee, it becomes impossible to operate with the full licensed power, the station may be operated at reduced power for a period of not to exceed 10 days, provided that the Commission and the Inspector in Charge shall be notified in writing immediately after the emergency develops. (See Operating Power Tolerance.)

3.58 INDICATING INSTRUMENTS. Each broadcast station shall be equipped with suitable indicating instruments of accepted accuracy to measure the antenna current, direct plate circuit voltage, and the direct plate circuit current of the last radio stage. These indicating instruments shall not be changed or replaced, without authority of the Commission, except by instruments of the same type, maximum scale reading, and accuracy. (See Indicating Instruments Pursuant to Section 3.58)

3.59 FREQUENCY TOLERANCE. The operating frequency of each broadcast station shall be maintained within 50 cycles of the assigned frequency until January 1, 1940, and thereafter the frequency of each new station or each station where a new transmitter be installed shall be maintained within 20 cycles of the assigned frequency, and after January 1, 1942, the frequency of all stations shall be maintained within 20 cycles of the assigned frequency.

3.60 FREQUENCY MONITOR. The licensee of each standard broadcast station shall have in operation at the transmitter a frequency monitor independent of the frequency control of the transmitter. The frequency monitor shall be approved by the Commission. It shall have a stability and accuracy of at least 5 parts per million. (See Approved Frequency Monitors and also Requirements for Approval of Frequency Monitors.)

3.61 NEW EQUIPMENT; RESTRICTIONS. The commission will authorize the installation of new transmitting equipment in a broadcast station or changes in the frequency control of an existing transmitter only if such equipment is so designed that there is reasonable assurance that the transmitter is capable of maintaining automatically the assigned frequency within the limits specified in section 3.59.

3.62 AUTOMATIC FREQUENCY CONTROL EQUIPMENT; AUTHORIZATION REQUIRED. New automatic frequency control equipment and changes in existing automatic frequency control equipment that may affect the precision of frequency control or the operation of the transmitter shall be installed only upon authorization from the Commission. (See Approved Equipment)

3.63 AUXILIARY TRANSMITTER. Upon showing that a need exists for the use of an auxiliary transmitter in addition to the regular transmitter of a broadcast station, a license therefor may be issued provided that:

(a) An auxiliary transmitter may be installed either at the same location as the main transmitter or at another location.

(b) A licensed operator shall be in control whenever an auxiliary transmitter is placed in operation.

(c) The auxiliary transmitter shall be maintained so that it may be put into immediate operation at any time for the following purposes:

(1) The transmission of the regular programs upon the failure of the main transmitter.

(2) The transmission of regular programs during maintenance or modification work on the main transmitter, necessitating discontinuance of its operation for a period not to exceed five days.

(3) Upon request by a duly authorized representative of the Commission.

(d) The auxiliary transmitter shall be tested at least once each week to determine that it is in proper operating condition and that it is adjusted to the proper frequency, except that in case of operation in accordance with paragraph (c) of this section during any week, the test in that week may be omitted provided the operation under paragraph (c) is satisfactory. A record shall be kept of the time and result of each test operating under paragraph (c). Tests shall be conducted only between midnight and 9 AM, local standard time.

(e) The auxiliary transmitter shall be equipped with satisfactory control equipment which will enable the maintenance of the frequency emitted by the station within the limits prescribed by these regulations.

(f) An auxiliary transmitter which is licensed at a geographical location different from that of the main transmitter shall be equipped with a frequency control which will automatically hold the frequency within the limits prescribed by these regulations without any manual adjustment during operation or when it is being put into operation.

(g) The operating power of an auxiliary transmitter may be less than the authorized power, but in no event shall it be greater than such power.

3.64 DUPLICATE MAIN TRANSMITTERS. The licensee of a standard broadcast station may be licensed for duplicate main transmitters provided that a technical need for such duplicate transmitters is shown and that the following conditions are met:

(a) Both transmitters are located at the same place.

(b) The transmitters have the same power rating.

(c) The external effects from both transmitters is substantially the same as to frequency stability, reliability of operation, radio harmonics and other spurious emissions, audio frequency range and audio harmonic generation in the transmitter.

OPERATION

3.71 MINIMUM OPERATING SCHEDULE. Except Sundays, the licensee of each standard broadcast station shall maintain a minimum operating schedule of two-thirds of the total hours that it is authorized to operate between 6 AM and 6 PM, local standard time, and two-thirds of the total hours it is authorized to operate between 6 PM and midnight, local standard time, except that in an emergency when, due to causes beyond the control of the licensee, it becomes impossible to continue operating, the station may cease operation for a period of not to exceed 10 days, provided that the Commission and the Inspector in Charge shall be notified in writing immediately after the emergency develops.

3.72 OPERATION DURING EXPERIMENTAL PERIOD. The licensee of each standard broadcast station shall operate or refrain from operating its station during the experimental period as directed by the Commission in order to facilitate frequency measurement or for the determination of interference. (Stations involved in the after-midnight frequency monitoring programs are notified of their operating and silent schedule.)

SUBPART B - RULES GOVERNING HIGH-FREQUENCY BROADCAST STATIONS

DEFINITIONS

3.201 HIGH-FREQUENCY BROADCAST STATION. The term "high-frequency broadcast station" means a station licensed primarily for the transmission of radiotelephone emissions intended to be received by the general public and operated on a channel in the frequency broadcast band.

3.202 HIGH-FREQUENCY BROADCAST BAND. The term "high-frequency broadcast band" means the band of frequencies extending from 43,000 to 50,000 kilocycles, both inclusive.

3.203 FREQUENCY MODULATION. The term "frequency modulation" means a system of modulation of a radio signal in which the frequency of the carrier wave is varied in accordance with the signal to be transmitted while the amplitude of the carrier remains constant.

3.204 CENTER FREQUENCY. The term "center frequency" means the frequency of the carrier wave with no modulation (with modulation, the instantaneous operating frequency swings above and below the center frequency. The operating frequency with no modulation shall be the center frequency within the frequency tolerance.)

3.205 HIGH-FREQUENCY BROADCAST CHANNEL. The term "high-frequency broadcast channel" means a band of frequencies 200 kilocycles wide and is designated by its center frequency. Channels for high-frequency broadcast stations begin at 43,100 kilocycles and continue in successive steps of 200 kilocycles to and including the frequency 49,900 kilocycles.

3.206 SERVICE AREA. The term "service area" of a high-frequency broadcast station means the area in which the signal is not subject to objectionable interference or objectionable fading. (High-frequency broadcast stations are considered to have only one service area; for determination of such area see Standards of Good Engineering Practice for High-Frequency Broadcast Stations.)

3.207 ANTENNA FIELD GAIN. The term "antenna field gain" of a high-frequency broadcast antenna means the ratio of the effective free space field intensity produced at one mile in the horizontal plane expressed in millivolts per meter for 1 kilowatt antenna input power to 137.6.

3.208 FREE SPACE FIELD INTENSITY. The term "free space field intensity" means the field intensity that would exist at a point in the absence of waves reflected from the earth or other reflecting objects.

3.209 FREQUENCY SWING. The term "frequency swing" is used only with respect to frequency modulation and means the instantaneous departure of the carrier frequency from the center frequency resulting from modulation.

3.210 MULTIPLEX TRANSMISSION. The term "multiplex transmission" means the simultaneous transmission of two or more signals by means of a common carrier wave. (Multiplex transmission as applied to high-frequency broadcast stations means the transmission of facsimile or other aural signals in addition to the regular broadcast signals.)

3.211 PERCENTAGE MODULATION. The term "percentage modulation" with respect to frequency modulation means the ratio of the actual frequency swing to the frequency swing required for 100 percent modulation expressed in percentage. (For high-frequency broadcast stations, a frequency swing of 75 kilocycles is standard for 100 percent modulation.)

3.212 EXPERIMENTAL PERIOD. The term "experimental period" means that period of time between 12 midnight and sunrise. This period may be used for experimental purposes in testing and maintaining apparatus by the licensee of any high-frequency broadcast station, on its assigned frequency and with its authorized power, provided no interference is caused to other stations maintaining a regular operating schedule within such period.

3.213 MAIN STUDIO. The term "main studio" means, as to any station, the studio from which the majority of its local programs originate, and/or from which a majority of its station announcements are made of programs originating at remote points.

ALLOCATION OF FACILITIES

3.221 BASIS OF LICENSING HIGH-FREQUENCY BROADCAST STATIONS. High-frequency broadcast stations shall be licensed to serve a specified area in square miles. The contour bounding the service area and the radii of the contour shall be determined in accordance with the Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

3.222 SERVICE AREAS; DEFINITIONS. For the purpose of determining the areas to be served by high-frequency broadcast stations, the following definitions apply:

(a) "Basic trade areas" and "limited trade areas" consist of areas the boundaries of which are determined by the Commission on the basis of showings made in applications as to retail trading areas or consumer trading areas and from Government data. Each basic trade area includes one "principal city". The boundaries of the basic trade areas are adjoining and the aggregate of all such areas is the total area of the United States. Each "limited trade area" includes one city. The boundaries of limited trade areas are not necessarily adjoining. Such areas may include portions of other limited trade areas and may extend into more than one basic trade area.

(b) "Principal city" means the largest city or the city or cities designated as "principal city" by the Commission, within a basic trade area. "City" means any city, town, or borough in a basic trade area except the principal city. Each "city" has a limited trade area.

(c) "Rural area" means all land area outside incorporated towns or cities with population greater than 2500 and where the density of population is less than 150 per square mile. Incorporated towns or cities with population from 2500 to 5000 without a high-frequency broadcast station and not adjacent to larger cities may be considered rural area.

3.223 SERVICE AREAS; ESTABLISHED. The Commission in considering applications for high-frequency broadcast stations will establish service areas. Such stations will be licensed to serve areas having the following characteristics:

(a) An area comprising a limited trade area and a city. The station shall render good service to the city and its service area shall conform generally with the limited trade area.

(b) An area comprising a basic trade area and a principal city. The station shall render good service to the principal city and its service area shall conform generally with the basic trade area; Provided, however, that the station may be licensed to serve temporarily an area less than the basic trade area, subject to the following conditions:

(1) That an applicant for authority to serve temporarily less than the basic trade area show substantial reason for relaxation of the requirement to serve the basic trade area and for specification of the proposed service area;

(2) That the area to be served include as much of the basic trade area as reasonably may be required in the public interest to be served and in no event less than the principal city and the metropolitan district (as defined by the U.S. Bureau of Census) in which it is located;

(3) That such an applicant show compliance with section 3.225 (b), where applicable and section 3.227 (a), except that such sections shall apply only in relation to other stations established under this proviso;

(4) That the Commission may condition the granting of any application for renewal of license of such station upon the rendering of service by such station to an area conforming generally with the basic trade area.

(c) An area of at least 15,000 square miles comprising primarily a large rural area, and particularly that part of basic trade areas which cannot be served by stations assigned basic trade areas due to economic and technical limitations. The service area may include one or more principal city or cities, provided that in rendering service to such cities the service to rural areas which the station is designated to serve is not impaired. The transmitter of such a station shall be located in such a manner that the service area;

(1) Shall extend into two or more basic trade areas;

(2) Shall not conform generally with a basic trade area;

(3) Shall not merely extend beyond a basic trade area.

(d) An area having substantially different characteristics (social, cultural, or economic) from those areas specified in sub-sections (a), (b), and (c) of this section where, by reason of special conditions, it is shown that a need (which cannot be supplied by a station serving areas under sub-sections (a), (b), or (c) of this section) for the proposed service both program and technical exists which makes the establishment of the service area in the public interest, convenience, or necessity. The Commission will give particular consideration in this connection to competitive advantages which such stations would have over other stations established under other provisions.

(e) In case it is not economically and technically feasible for a station assigned a basic or limited trade area to serve substantially all such area, the Commission will establish the service area on the basis of conditions which obtain in the trade area.

(f) In case an applicant proposes a change in an established service area, the applicant shall make a full showing as to need for such change and the effect on other stations serving the area.

3.224 TIME OF OPERATION. All high-frequency broadcast stations shall be licensed for unlimited time operation.

3.225 SHOWING REQUIRED. Authorization for a new high-frequency broadcast station or increase in facilities of an existing station will be issued only after a satisfactory showing has been made in regard to the following matters:

(a) That the area which the applicant proposes to serve has the characteristics of an area described in section 3.223 hereof.

(b) Where a service area has been established in which one or more existing high-frequency broadcast stations are in operation, that the contours of any new station proposed to serve such area will compare with those of the existing station or stations as nearly as possible, or that the service area already established should be modified.

(c) That objectionable interference will not be caused to existing stations or that if interference will be caused, the need for the proposed service outweigh the need for the service which will be lost by reason of such interference.

(d) That the proposed station will not suffer interference to such an extent that its service would be reduced to an unsatisfactory degree. (For determining objectionable interference, see Standards of Good Engineering Practice for High-Frequency Broadcast Stations.)

(e) That the technical equipment proposed, the location of the transmitter, and other technical phases of operation comply with the regulations governing the same, and the requirements of good engineering practice. (See technical regulations herein and Standards of Good Engineering Practice for High-Frequency Broadcast Stations.)

(f) That the applicant is financially qualified to construct and operate the proposed station; and, if the proposed station is to serve substantially the same area as an existing station, that applicant will be able to compete effectively with the existing station or stations.

(g) That the program service will include a portion of programs particularly adapted to a service utilizing the full fidelity capability of the system, as set forth in the standards of Good Engineering Practice for High-Frequency Broadcast Stations.

(h) That the proposed assignment will tend to effect a fair, efficient, and equitable distribution of radio service among the several states and communities.

(i) That the applicant is legally qualified, is of good character, and possesses other qualifications sufficient to provide a satisfactory public service.

(j) That the facilities sought are subject to assignment as requested under existing international agreements and the Rules and Regulations of the Commission.

(k) That the public interest, convenience, and necessity will be served through the operation under the proposed assignment.

3.226 CHANNEL ASSIGNMENTS. The channels set forth below with the indicated center frequencies are available for assignment to high-frequency broadcast stations to serve the areas provided in section 3.223;

(a) An applicant for a station to serve an area specified in section 3.223 (a) or (b), to be located in a principal city or city which has a population less than 25,000 (city only) shall apply for one of the following channels:

48900	49500
49100	49700
49300	49900

(b) An applicant for a station to serve an area specified in section 3.223 (a) or (b), to be located in a principal city or city which has a population greater than 25,000 (city only) shall apply for one of the following channels:

44500	45900	47500
44700	46100	47700
44900	46300	47900
45100	46500	48100
45300	46700	48300
45500	46900	48500
45700	47100	48700
	47300	

(c) An applicant for a station to serve primarily a large rural area, specified in section 3.223 (c) or an area specified in section 3.223 (d) shall apply for one of the following channels;

43100	43900
43300	44100
43500	44300
43700	

(d) Notwithstanding the provisions of subsection (a) of this section, an applicant for a station to serve an area specified in section 3.223 (a), to be located in a city having a population greater than 25,000, in or adjacent to any metropolitan district having a population greater than 1,000,000, may apply for one of the following channels:

49100
49500
49900

3.227 SPECIAL PROVISIONS CONCERNING ASSIGNMENTS. (a) Stations located in the same city shall have substantially the same service area.

(b) High-frequency broadcast stations shall use frequency modulation exclusively.

(c) Stations serving a substantial part of the same area shall not be assigned adjacent channels.

(d) One channel only will be assigned to a station.

3.228 FACSIMILE BROADCASTING AND MULTIPLEX TRANSMISSION. The Commission may grant authority to a high-frequency broadcast station for the multiplex transmission of facsimile and aural broadcast programs provided the facsimile transmission is incidental to the aural broadcast and does not either reduce the quality of or the frequency swing required for the transmission of the aural program. The frequency swing for the modulation of the aural program should be maintained at 75 kc. and the facsimile signal added thereto. No transmission outside the authorized band of 200 kc. shall result from such multiplex operation nor shall interference be caused to other stations operating on adjacent channels. The transmission of multiplex signals may also be authorized on an experimental basis in accordance with section 3.32, subpart A.

3.229 PROOF OF PERFORMANCE REQUIRED. Within 1 year of the date of first regular operation of a high-frequency broadcast station, continuous field intensity records along several radials shall be submitted to the Commission which will establish the actual field contours, and from which operating constants required to deliver service to the area specified in the license are determined. The Commission may grant extensions of time upon showing of reasonable need therefor.

3.230 MULTIPLE OWNERSHIP. (a) No person (including all persons under common control) shall, directly or indirectly, own, operate, or control more than one high-frequency broadcast station that would serve substantially the same service area as another high-frequency broadcast station owned, operated, or controlled by such person.

(b) No person (including all persons under common control) shall, directly or indirectly, own, operate, or control more than one high-frequency broadcast station except upon a showing:

(1) That such ownership, operation, or control would foster competition among high-frequency broadcast stations or provide a high-frequency broadcasting service distinct and separate from existing services, and,

(2) That such ownership, operation, or control would not result in the concentration of control of high-frequency broadcasting facilities in a manner inconsistent with public interest, convenience, or necessity; provided, however, that the Commission will consider the ownership, operation, or control of more than six high-frequency broadcast stations to constitute the concentration of control of high-frequency broadcasting facilities in a manner inconsistent with public interest, convenience, or necessity.

3.231 NORMAL LICENSE PERIOD. All high-frequency broadcast station licenses will be issued so as to expire at the hour of 3 AM, eastern standard time, and will be issued for a normal license period of 1 year, expiring as follows:

(a) For stations operating on the frequencies 48900, 49100, 49300, 49500, 49700, and 49900; April 1.

(b) For stations operating on the frequencies 44500, 44700, 44900, 45100, 45300, 45500, 45700, 45900, 46100, 46300, and 46500; May 1.

(c) For stations operating on the frequencies 46700, 46900, 47100, 47300, 47500, 47700, 47900, 48100, 48300, 48500, and 48700; June 1.

(d) For stations operating on the frequencies 43100, 43300, 43500, 43700, 43900, 44100, and 44300; July 1.

EQUIPMENT

3.241 MAXIMUM POWER RATING. The commission will not authorize the installation of a transmitter having a maximum rated power more than twice the operating power of the station.

3.242 MAXIMUM RATED CARRIER POWER; HOW DETERMINED. (a) The maximum rated carrier power of a standard transmitter shall be determined by the manufacturer's rating of the equipment.

(b) The maximum rated carrier power of a composite transmitter shall be determined by the sum of the applicable commercial ratings of the vacuum tubes employed in the last radio stage.

3.243 FREQUENCY MONITOR. The licensee of each high-frequency broadcast station shall have in operation at the transmitter a frequency monitor independent of the frequency control of the transmitter. It shall have a stability of 20 parts per million. For detailed requirements thereof see Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

3.244 MODULATION MONITOR. The licensee of each high-frequency broadcast station shall have in operation at the transmitter an approved modulation monitor. For detailed requirements thereof see Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

3.245 REQUIRED TRANSMITTER PERFORMANCE. (a) The external performance of high-frequency broadcast transmitters shall be within the minimum requirements prescribed by the Commission contained in the Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

(b) The transmitter center frequency shall be controlled directly by automatic means which do not depend on inductances and capacities for inherent stability.

(c) The transmitter shall be wired and shielded in accordance with good engineering practice and shall be provided with safety features in accordance with the specifications of article 810 of the current National Electric Code as approved by the American Standards Association.

3.246 INDICATING INSTRUMENTS. The direct plate circuit current and voltage shall be measured by instruments having an acceptable accuracy. (See Standards of Good Engineering Practice for High-Frequency Broadcast Stations.)

3.247 AUXILIARY AND DUPLICATE TRANSMITTERS. See sections 3.63 and 3.64 for provisions governing the use of auxiliary and duplicate transmitters at high-frequency broadcast stations.

3.248 CHANGES IN EQUIPMENT AND ANTENNA SYSTEM. Licensees of high-frequency broadcast stations shall observe the following provisions with regard to changes in equipment and antenna system:

(a) No changes in equipment shall be made:

(1) That would result in the emission of signals outside of the authorized channel.

(2) That would result in the external performance of the transmitter being in disagreement with that prescribed in the Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

(b) Specific authority, upon filing formal application therefor, is required for a change in service area or for any of the following changes:

(1) Changes involving an increase in the maximum power rating of the transmitter.

(2) A replacement of the transmitter as a whole.

(3) Change in the location of the transmitter antenna.

(4) Change in antenna system, including transmission line, which would result in a measurable change in service or which would affect the determination of the operating power by the direct method. If any change is made in the antenna system or any change made which may affect the antenna system, the method of determining operating power shall be changed immediately to the indirect method.

(5) Change in location of main studio to outside of the borders of the city, state, district, territory, or possession.

(6) Change in the power delivered to the antenna.

(c) Specific authority, upon filing informal request therefor, is required for the following change in equipment and antenna:

(1) Change in the indicating instruments installed to measure the antenna current or transmission line, direct plate circuit voltage, and the direct current of the last radio stage, except by instruments of the same type, maximum scale reading and accuracy.

(2) Minor changes in the antenna system and/or transmission line which would not result in an increase of service area.

(3) Changes in the location of the main studio except as provided for in subparagraph (b) (5).

(d) Other changes, except as above provided for in this section or in Standards of Good Engineering Practice for High-Frequency Broadcast Stations prescribed by the Commission may be made at any time without the authority of the Commission, provided that the Commission shall be promptly notified thereof, and such changes shall be shown in the next application for renewal of license.

TECHNICAL OPERATION

3.251 OPERATING POWER; HOW DETERMINED. The operating power, and the requirements for maintenance thereof, of each high-frequency broadcast station shall be determined by the Standards of Good Engineering Practice for High-Frequency Broadcast Stations.

3.252 MODULATION. The percentage of modulation of all stations shall be maintained as high as possible consistent with good quality of transmission and good broadcast practice and in no case less than 85 percent on peaks of frequent recurrence during any selection which normally is transmitted at the highest level of the program under consideration.

3.253 FREQUENCY TOLERANCE. The operating frequency without modulation of each broadcast station shall be maintained within 2000 cycles of the assigned center frequency.

OPERATION

3.261 MINIMUM OPERATING SCHEDULE; SERVICE. (a) Except Sundays, the licensee of each high-frequency broadcast station shall maintain a regular daily operating schedule which shall consist of at least 3 hours of operation during the period 6 AM to 6 PM, local standard time and 3 hours of operation during the period 6 PM to midnight, local standard time. In an emergency, however, when due to causes beyond the control of the licensee, it becomes impossible to continue operating, the station may cease operation for a period not to exceed 10 days, provided that the Commission and the Inspector in charge of the radio district in which the station is located shall be notified in writing immediately after the emergency develops.

(b) Such stations shall devote a minimum of 1 hour each day during the period 6 AM to 6 PM, and 1 hour each day during the period 6 PM to midnight, to programs not duplicated simultaneously as primary service in the same area by any standard broadcast station or by any high-frequency broadcast station. During said 1 hour periods, a service utilizing the full fidelity capability of the system, as set forth in the Standards of Good Engineering Practice for High-Frequency Broadcast Stations, shall be rendered. However, the Commission may, upon request accompanied by a showing of reasons therefor, grant exemption from the foregoing requirements, in whole or in part, for periods not in excess of 3 months.

(c) In addition to the foregoing minimum requirements, the Commission will consider, in determining whether public interest, convenience, and necessity has been or will be served by the operation of the station, the extent to which the station has made or will make use of the facility to develop a distinct and separate service from that otherwise available in the service area.

SUBPART C - GENERAL RULES APPLICABLE TO BOTH STANDARD AND HIGH-FREQUENCY BROADCAST STATIONS

3.401 STATION LICENSE; POSTING OF. The station license and any other instrument of authorization or individual order concerning construction of the equipment or the manner of operation of the station shall be posted in a conspicuous place in the room in which the transmitter is located in such manner that all terms thereof are visible and the license of the station operator shall be posted in the same manner. (See secs. 2.51 and 2.52)

3.402 LICENSED OPERATOR REQUIRED. The licensee of each station shall have a licensed operator or operators of the grade specified by the Commission on duty during all periods of actual operation of the transmitter at the place where the transmitting equipment is located. (See sec. 2.53.)

3.403 LICENSED OPERATOR; OTHER DUTIES. The licensed operator on duty and in charge of a standard broadcast transmitter may, at the discretion of the licensee, be employed for other duties or for the operation of another radio station or stations in accordance with the class of operator's license which he holds and by the rules and regulations governing such other stations; Provided, however, that such duties shall in nowise interfere with the proper operation of the standard broadcast transmitter.

3.404 LOGS. The licensee of each broadcast station shall maintain program and operating logs and shall require entries to be made as follows:

(a) In the program log:

(1) An entry of the time each station identification announcement (call letters and location) is made.

(2) An entry briefly describing each program broadcast, such as "music", "drama", "speech", etc., together with the name or title thereof, and the sponsor's name, with the time of the beginning and ending of the complete program. If a mechanical record is used, the entry shall show the exact nature thereof, such as "record", "transcription", etc., and the time it is announced as a mechanical record. If a speech is made by a political candidate, the name and political affiliations of such speaker shall be entered.

(3) An entry showing that each sponsored program broadcast has been announced as sponsored, paid for, or furnished by the sponsor.

(b) In the operating log:

(1) An entry of the time the station begins to supply power to the antenna, and the time it stops.

(2) An entry of the time the program begins and ends.

(3) An entry of each interruption of the carrier wave, its cause, and duration.

(4) An entry of the following each 30 minutes:

(i) Operating constants of last radio stage (total plate current and plate voltage).

(ii) Antenna current.

(iii) Frequency monitor reading.

(iv) Temperature of crystal control chamber if thermometer is used.

(5) Log of experimental operation during experimental period. (If regular operation is maintained during this period, the above logs shall be kept.)

(i) A log must be kept of all operation during the experimental period. If the entries required above are not applicable thereto, then the entries shall be made so as to fully describe the operation.

3.405 LOGS, RETENTION OF. Logs of standard or high-frequency broadcast stations shall be retained by the licensee for a period of 2 years, except when required to be retained for a longer period in accordance with the provisions of section 2.54.

3.406 STATION IDENTIFICATIONS. (a) A licensee of a standard or high-frequency broadcast station shall make station identification announcement (call letters and location) at the beginning and ending of each time of operation and during operation on the hour and half-hour as provided below:

(b) Such identification announcement need not be made on the hour when to make such announcement would interrupt a single consecutive speech, play, religious service, symphony concert, or operatic production of longer duration than 30 minutes. In such cases, the identification announcement shall be made at the beginning of the program, at the first interruption of the entertainment continuity, and at the conclusion of the program.

(c) Such identification announcement need not be made on the half-hour when to make such announcement would interrupt a single consecutive speech, play, religious service, symphony concert, or operatic production. In such cases an identification announcement shall be made at the first interruption of the entertainment continuity, and at the conclusion of the program; Provided, that an announcement within 5 minutes of the half-hour will waive the requirements of other identification announcements.

(d) In the case of variety show programs, baseball game broadcasts, or similar programs of longer duration than 30 minutes, the identification announcement shall be made within 5 minutes of the hour and half-hour.

(e) In case of all other programs (except as provided in paragraphs (b) and (c) of this section) the identification announcement shall be made within 2 minutes of the hour and half-hour.

(f) In making the identification announcement the call letters shall be given only on the channel of the station identified thereby.

3.407 MECHANICAL RECORDS. Each broadcast program consisting of a mechanical record or a series of mechanical records shall be announced in the manner and to the extent set out below:

(a) A mechanical record or a series thereof, of longer duration than 30 minutes, shall be identified by appropriate announcement at the beginning of the program, at each 30 minute interval, and at the conclusion of the program; Provided, however, that the identifying announcement at each 30 minute interval is not required in case of a mechanical record consisting of a single, continuous, uninterrupted speech, play, religious service, symphony concert, or operatic production of longer duration than 30 minutes.

(b) A mechanical record, or a series thereof, of a longer duration than 5 minutes, and not in excess of 30 minutes, shall be identified by an appropriate announcement at the beginning and end of the program;

(c) A single mechanical record of a duration not in excess of 5 minutes shall be identified by appropriate announcement immediately preceding the use thereof;

(d) In case a mechanical record is used for background music, sound effects, station identification, program identification (theme music of short duration), or identification of the sponsorship of the program proper, no announcement of the mechanical record is required.

(e) The identifying announcement shall accurately describe the type of mechanical record used, i. e., where an electrical transcription is used it shall be announced as a "transcription" or an "electrical transcription", or as "transcribed" or "electrically transcribed", and where a phonograph record is used it shall be announced as a "record".

3.408 REBROADCAST. (a) The term "rebroadcast" means reception by radio of the program of a radio station, and the simultaneous or subsequent retransmission of such program by a broadcast station.

(b) The licensee of a standard or high-frequency broadcast station may, without further authority of the Commission, rebroadcast the program of a United States standard or high-frequency broadcast station, provided the Commission is notified of the call letters of each station rebroadcast and the licensee certifies that express authority has been received from the licensee of the station originating the program.

(c) The licensee of a standard or high-frequency broadcast station may, without further authority of the Commission, rebroadcast on a noncommercial basis a noncommercial program of an international broadcast station, provided the Commission is notified of the call letters of each station rebroadcast and the licensee certifies that express authority has been received from the licensee of the station originating the program.

(d) No licensee of a standard broadcast station shall broadcast the program of any other class of United States radio station without written authority having first been obtained from the Commission upon application accompanied by written consent or certification of consent of the licensee of the station originating the program.

(e) In case of a program rebroadcast by several standard broadcast stations, such as a chain rebroadcast, the person legally responsible for distributing the program or the network facilities may obtain the necessary authorization for the entire rebroadcast both from the Commission and from the person or licensee of the station originating the program.

Attention is directed to section 325 (b) of the Communications Act of 1934, which reads as follows: No person shall be permitted to locate, use, or maintain a radio broadcast studio or other place or apparatus from which or whereby sound waves are converted into electrical energy, or mechanical or physical reproduction of sound waves produced, and caused to be transmitted or delivered to a radio station in a foreign country for the purpose of being broadcast from any radio station there having a power output of sufficient intensity, and/or being so located geographically that its emissions may be received consistently in the United States, without first obtaining a permit from the Commission upon proper application therefor.

BROADCASTS BY CANDIDATES FOR PUBLIC OFFICE

3.421 GENERAL REQUIREMENTS. No station licensee is required to permit the use of its facilities by any legally qualified candidate for public office, but if any licensee shall permit any such candidate to use its facilities, it shall afford equal opportunities to all other such candidates for that office to use such facilities, provided that such licensee shall have no power of censorship over the material broadcast by any such candidate.

3.422 DEFINITIONS. The following definitions shall apply for the purposes of section 3.421:

(a) "A legally qualified candidate" means any person who has met all the requirements prescribed by local, state, or federal authority as a candidate for the office which he seeks, whether it be municipal, county, state, or national, to be determinec according to the applicable local laws.

(b) "Other candidates for that office" means all other legally qualified candidates for the same public office.

3.423 RATES AND PRACTICES. The rates, if any, charged all such candidates for the same office shall be uniform and shall not be rebated by any means, directly or indirectly; no licensee shall make any discrimination in charges, practices, regulations, facilities, or services for or in connection with the service rendered pursuant to these rules, or make or give any preference to any candidate for public office or subject any such candidate to any prejudice or disadvantage; nor shall any licensee make any contract or other agreement which shall have the effect of permitting any legally qualified candidate for any public office to broadcast to the exclusion of other legally qualified candidates for the same public office.

3.424 RECORDS; INSPECTION. Every licensee shall keep and permit public inspection of a complete record of all requests for broadcast time made by or on behalf of candidates for public office, together with an appropriate notation showing the disposition made by the licensee of such requests, and the charges made, if any, if request is granted.

PART 4 - RULES GOVERNING BROADCAST SERVICES OTHER THAN STANDARD BROADCAST

The rules given in this section apply to the following broadcast stations:

RELAY BROADCAST STATIONS
STL BROADCAST STATIONS
INTERNATIONAL BROADCAST STATIONS
EXPERIMENTAL TELEVISION STATIONS
FACSIMILE BROADCAST STATIONS
NONCOMMERCIAL EDUCATIONAL BROADCAST STATIONS
DEVELOPMENTAL BROADCAST STATIONS

IN GENERAL

4.1 FREQUENCY TOLERANCE. The operating frequency of the broadcast stations as listed below shall be maintained within plus or minus the percentage of the assigned frequency as given in table I.

TABLE I

CLASS OF STATION	FREQUENCY TOLERANCE
Relay Broadcast Station	
(a) 1622 to 2830 Kc.	0.04 Percent
(b) 30000 to 40000 Kc. and above	10 Watts or less, 0.1 Percent. Above 10 watts, 0.05 Percent.
STL Broadcast Station	0.01 Percent
International Broadcast Station	0.005 Percent*
Television Broadcast Station	0.01 Percent
Facsimile Broadcast Station	0.05 Percent or less as required.
High-Frequency Broadcast Station	0.01 Percent
Noncommercial Educational Broadcast Station	Do
Developmental Broadcast Station	0.05 Percent or less as required.

*Tolerance may be 0.01 percent on equipment installed prior to January 1, 1940, and until January 1, 1941, when all international stations shall maintain frequency within 0.005 percent of the assigned frequency.

4.2 FREQUENCY MONITORS. (a) The licensee of each broadcast station listed in section 4.1 except relay broadcast stations, shall operate at the transmitter a frequency monitor independent of the frequency control of the transmitter.

(b) The frequency monitor shall be designed and constructed in accordance with good engineering practice and shall have an accuracy sufficient to determine that the operating frequency is within one-half of the allowed tolerance.

(c) The licensee of each relay broadcast station shall provide the necessary means for determining that the frequency of the station is within the allowed tolerance.

(d) The frequency of all stations listed in section 4.1 shall be checked at each time of beginning operation and as often thereafter as necessary to maintain the frequency within the allowed tolerance.

4.3 LICENSE PERIOD; RENEWAL. (a) Licenses for the following classes of broadcast stations will be normally issued for a period of one year expiring as follows:

CLASS OF STATION	DATE OF EXPIRATION
Relay Broadcast Station	
(a) 1622 to 2830 Kc.	Oct. 1
(b) 30000 to 40000 Kc. and above	Dec. 1
STL Broadcast Station	Apr. 1
International Broadcast Station	Nov. 1
Television Broadcast Station	Feb. 1
Facsimile Broadcast Station	Mar. 1
High-Frequency Broadcast Station	Apr. 1
Noncommercial Educational Broadcast Station	May 1
Developmental Broadcast Station	Do.

(b) Each licensee shall submit the application for renewal of license at least 60 days prior to the expiration date (section 1.360).

(c) A supplemental report shall be submitted with each application for renewal of license of a station licensed experimentally in accordance with the regulations governing each class of station.

4.4 REQUIREMENTS, LIMITATIONS, AND RESTRICTIONS. (a) No station licensed experimentally will be assigned for exclusive use of any frequency. In case interference would be caused by simultaneous operation of stations licensed experimentally, such licensees shall endeavor to arrange satisfactory time division. If such agreement cannot be reached, the Commission will determine and specify the time division.

(b) The Commission may from time to time require that a station licensed experimentally conduct such experiments that are deemed desirable and reasonable for the development of the service.

(c) The program of research and experimentation as offered by an applicant in compliance with the requirements for obtaining a license for an experimental station shall be adhered to in the main, unless the licensee is authorized to do otherwise by the Commission.

(d) A licensee of an experimental station is not required to adhere to a regular schedule of operation but shall actively conduct a program of research and experimentation or transmission of programs; provided, however, licensees of developmental broadcast stations which are licensed to conduct special intermittent experiments, such as to develop and test commercial broadcast equipment, are required to operate only when there is a need therefor.

(e) A supplementary statement shall be filed with and made a part of each application for construction permit for any broadcast station on an experimental basis which specifies any frequency above 300000 kilocycles or in the bands 162000 to 168000, 210000 to 216000, and 264000 to 270000 kilocycles except television, confirming the applicant's understanding:

(1) That all operation upon the frequency is experimental only,

(2) That the frequency may not be the best suited to the particular experimental work to be carried on, and,

(3) That the frequency may not be allocated for the service that may be developed experimentally.

4.5 STATION RECORDS. (a) The licensee of each class of broadcast station listed in section 4.1 shall maintain adequate records of the operation, including:

(1) Hours of operation.

(2) Program transmitted.

(3) Frequency check.

(4) Pertinent remarks concerning transmission.

(5) In case of relay station, an entry giving point of program origination and receiver location shall be included.

(6) Research and experimentation conducted in case of an experimental station.

(7) And any additional information specified in the regulations governing each class of station or for completing the supplemental report as required.

(5) The above information shall be made available upon request by authorized Commission representatives.

4.6 EQUIPMENT CHANGES. The licensee of each class of broadcast station listed in section 4.1 may make any changes in the equipment that are deemed desirable or necessary, provided:

(a) That the operating frequency is not permitted to deviate more than the allowed tolerance;

(b) That the emissions are not permitted outside the authorized band;

(c) That the power output complies with the license and the regulations governing the same; and

(d) That the transmitter as a whole or output power rating of the transmitter is not changed.

4.7 EMISSION AUTHORIZED. All classes of broadcast licenses authorize A3 emission only unless otherwise specified in the license. In case A1, A2, A4, A5, or special emission are necessary or helpful in carrying on any phases of experimentation, application setting out fully the needs shall be made to, and authority therefor received from, the Commission.

4.8 ADDITIONAL ORDERS, AS NEEDED. In case all the general rules and regulations and the specific rules governing each class of broadcast station do not cover all phases of operation or experimentation with respect to external effects, the Commission may make supplemental or additional orders in each case as deemed necessary for operation in the public interest, convenience, and/or necessity.

4.9 OPERATION. A licensed operator shall be on duty and in charge of the transmitter of each broadcast station listed in section 4.1. In no case will remote control operation be authorized. A transmitter is not considered as being operated by remote control when the following conditions prevail:

(a) Continuous reading indicating instruments are before the operator as follows:

(1) Frequency deviation meter.

(2) Percentage modulation indicator.

(3) Spurious emission check (receiver).

(4) Last radio stage plate voltage.

(5) Last radio stage total plate current.

(6) Output or antenna current.

(b) The operator has off-and-on control of the power to the last radio stage.

(c) The operator can reach the transmitter proper in not more than 5 minutes to make any changes or adjustments necessary to maintain proper operation.

4.10 REBROADCASTS. (a) The licensee of an international or noncommercial educational broadcast station may, without further authority of the Commission, rebroadcast the program of a United States standard broadcast station, provided the Commission is notified of the call letters of each station rebroadcast and the licensee certified that express authority has been received from the licensee of the station originating the program. (See sections 4.43 and 4.132 (c) concerning commercial announcements).

(b) No licensee of an international or noncommercial educational broadcast station shall rebroadcast the program of any other class of United States radio station without written authority having first been obtained from the Commission.

(c) The licensee of a noncommercial educational broadcast station may, without further authority of the Commission, rebroadcast the noncommercial programs of a standard broadcast station or an international broadcast station, provided the Commission is notified of the call letters of each station rebroadcast and the licensee certifies that express authority has been received from the licensee of the station originating the program.

(d) No licensee of any other class of broadcast station listed in section 4.1 (television, facsimile, high frequency, or developmental) shall rebroadcast the program of any radio station without written authority first having been obtained from the Commission.

(e) A licensee of an international broadcast station may authorize the rebroadcast of its programs by any station outside the limits of the North American Continent without permission from the Commission: Provided, that the station rebroadcasting the programs cannot be received consistently in the United States.

(f) An application for authority to rebroadcast the program of any radio station shall be accompanied by written consent or certification of consent of the licensee of the station originating the program.

4.11 EQUIPMENT AND PROGRAM TESTS. (a) A licensee of a broadcast station listed in section 4.1 shall conduct equipment tests in accordance with section 2.42 and program tests in accordance with section 2.43.

(b) In case the transmitter and associated equipment are on hand in complete form and an application for license was filed and granted with the application for construction permit, then the notification of equipment tests and program tests as required by paragraph (a) of this section need not be made.

4.12 STATION AND OPERATOR LICENSES; POSTING OF. (a) The station license and any other instrument of authorization or individual order concerning the construction of the equipment or manner of operation of the station shall be posted so that all terms thereof are visible in a conspicuous place in the room in which the transmitter is located, provided:

(1) If the transmitter operator is located at a distance from the transmitter pursuant to section 4.9, the station license shall be posted in the above-described manner at the operating position.

(2) If the station is licensed for portable-mobile operation, the station license or a photo copy thereof shall be affixed to the equipment or kept in the possession of the operator on duty at the transmitter. If a photo copy is used, the original license shall be available for inspection by an authorized government representative.

(b) The license of each station operator(s) shall be conspicuously posted at the operating position, provided:

(1) If the station at which the operator is on duty is licensed for portable-mobile operation, the operator's license may be kept in his personal possession.

RELAY BROADCAST STATIONS

4.21 DEFINED. The term "relay broadcast station" means a station licensed to transmit from points where wire facilities are not available, programs for broadcast by one or more broadcast stations, or orders concerning such programs.

4.22 LICENSING AND AUTHORIZATIONS. (a) A license for a relay broadcast station will be issued only to the licensee of a standard broadcast station; Provided, however, in cases where it is impractical, impossible, or prohibited by laws or regulations for the licensee of a standard broadcast station to install, operate, or maintain the necessary equipment under its legal control, the Commission may grant special temporary authority for each event to another person to operate as a relay broadcast station equipment already licensed for another service, or equipment which may be installed under section 319 (b) of the Communications Act of 1934 without a construction permit and provided further:

(b) The Commission may license a special relay broadcast station to the licensee of another class of broadcast station provided a need therefor is shown and the relay station will be used only for relaying of programs for broadcast by such broadcast station.

(c) The license of a relay broadcast station authorizes the transmission of commercial or sustaining programs, or orders concerning such programs, to be broadcast by its standard broadcast station and other broadcast stations transmitting the same programs simultaneously or a chain program to the network with which the licensee is regularly affiliated. The license of a relay station does not authorize transmission of programs to be broadcast solely by other broadcast stations not aforementioned.

(d) In case a licensee has two or more standard broadcast stations located in different cities, it shall, in applying for a new relay station or for renewal of license of an existing relay station, designate the standard broadcast station or stations in conjunction with which the relay station is to be operated principally, and it shall not thereafter operate the relay station in conjunction with another of its standard broadcast stations located in a different city for more than a total of 10 days in any 30 day period.

(e) Each application for temporary authority to operate a relay broadcast station from a person other than a licensee of a standard broadcast station shall be accompanied by an application for authority to broadcast the program from the licensee of the standard broadcast station proposing the broadcast.

(f) An application for special temporary authority to operate another class of station as a relay broadcast station shall specify a group of frequencies allocated in section 4.23; Provided, however, in case of events of national interest and importance which cannot be transmitted successfully to the nearest available wire facilities on these frequencies, other frequencies under the jurisdiction of the Commission may be requested, if it is shown that the operation thereon will not cause interference to established stations.

(g) An application for special temporary authority to operate on frequencies not allocated by section 4.23 or to operate another class of station as a relay broadcast station must be received by the Commission not less than 10 days prior to the actual event to be broadcast, and shall contain complete information concerning the frequencies requested, and the license of the station to be used. In case of emergencies, which shall be fully explained in the application, the Commission may waive the 10 day requirement specified herein.

4.23 FREQUENCY ASSIGNMENT AND OPERATION. (a) The following groups of frequencies are allocated for assignment to relay broadcast stations:

GROUP A	GROUP B	GROUP C	GROUP D	GROUP E	GROUP F	GROUP G	GROUP H	GROUP I
Kc.	Kc.	Kc.	Kc.	Kc.	Kc.	Kc.	Kc.	Kc.
1622	1606	1646	30,820	31,220	31,620	33,380	132,260	133,030
2058	2022	2090	33,740	35,620	35,260	35,020	134,080	134,850
2150	2102	2190	35,820	37,020	37,340	37,620	135,480	136,810
2790	2758	2830	37,980	39,260	39,620	39,820	135,760	138,630

GROUP J

Any 4 frequencies above 300,000 kc.
excluding band 400,000 to 401,000 kc.

(b) One of the above groups only, including all four frequencies, will be assigned each station. The first application from any metropolitan area for the frequencies in groups A, B, or C shall specify group A; the second group B; and the third group C; the fourth group A again, etc., and likewise for frequencies in groups D, E, F, or G, first application group D; second E; third F; etc. Outstanding assignments not following this order will not be changed unless a need therefor develops. Additional applicants shall specify the next unassigned group in sequence or any other group if it appears interference will be avoided thereby.

(c) A station may be licensed for group H when a need for frequencies of this order may be shown.

(d) Group I will be licensed to stations to operate with frequency modulation only when need for such operation and frequencies of this order may be shown.

(e) Any four specific frequencies under group J will be assigned on experimental operation only and an applicant may apply for the four frequencies which appear most suitable for the experimental work to be conducted.

(f) The licensee of a station on group J shall carry on research and experimentation for the advancement of the relay broadcast art and development of these ultra-high frequencies for relay broadcast services. An application for authority to operate a station on frequencies in group J shall include a statement concerning the research and experiments to be conducted. The research and experiments shall indicate reasonable promise of substantial contribution to the development of the program relay services.

(g) A license authorizes operation on only one of the four assigned frequencies at any one time. In case it is desired to transmit programs and spoken orders concerning such programs simultaneously, two licenses are required though each may specify the same group of frequencies.

4.24 FREQUENCY SELECTION TO AVOID INTERFERENCE. In case two or more stations are licensed for the same group of frequencies in the same area and in case simultaneous operation is contemplated, the licensees shall endeavor to select frequencies to avoid interference. If a mutual agreement to this effect cannot be reached the Commission shall be notified and it will specify the frequencies on which each station is to be operated.

4.25 POWER LIMITATIONS. (a) A relay broadcast station assigned frequencies in groups A, B, C, and J will be licensed to operate with a power output not in excess of that necessary to transmit the program and orders satisfactorily to the receivers and shall not be operated with a power greater than licensed.

(b) A relay broadcast station assigned frequencies in groups D, E, F, and G will not be authorized to install equipment or licensed for an output power in excess of 100 watts; provided that before using any frequency in these groups with a power in excess of 25 watts, tests shall be made by the licensee to insure that no objectionable interference will result to the service of any government station, and provided, further, that if the use of any frequency may cause interference then the power shall be reduced to 25 watts or another frequency in the licensed group selected which will not cause objectionable interference.

(c) A relay broadcast station assigned frequencies in groups H and I will be licensed to operate with a power output not in excess of that necessary to transmit the program and orders satisfactorily to the receivers and shall not be operated with a power greater than that licensed. In event interference may be caused to stations on adjacent channels, licensees shall endeavor to make arrangements to reduce power to a point where interference will not be objectionable. If a satisfactory arrangement cannot be agreed upon the Commission will determine and specify the maximum power or conditions of operation of each such station.

4.26 SUPPLEMENTAL REPORT WITH RENEWAL APPLICATION. The licensee of a relay broadcast station assigned frequencies under group J shall submit a supplemental report with and made a part of each application for renewal of license as follows:

(a) Number of hours operated for experimental purposes.

(b) Developments carried on in the relay broadcast service.

(c) Propagation characteristics of the frequencies assigned in regard to relay broadcast service.

(d) All other developments or major changes in equipment.

(e) Any other pertinent developments.

RULES GOVERNING STL BROADCAST STATIONS

4.31 DEFINED. The term "STL broadcast station" (studio transmitter link) means a station used to transmit programs from the main studio to the transmitter of a high-frequency broadcast station, or an international broadcast station.

4.32 LICENSING REQUIREMENTS. An STL broadcast station will be licensed only to the licensee of a high-frequency broadcast station or of an international broadcast station. Only one STL broadcast station will be authorized in connection with the license for any high-frequency broadcast station. Not more than two STL broadcast stations will be authorized in connection with the license for any international broadcast station. Each such STL station shall be at a fixed location.

4.33 SERVICE. The license of an STL broadcast station authorizes the transmission of program material, including commercial programs, from the main studio to the transmitter of the high-frequency broadcast station or international broadcast station in connection with which it is authorized.

4.34 FREQUENCY ASSIGNMENT AND OPERATION. (a) The following frequencies are allocated for assignment to STL broadcast stations upon an experimental basis:

330,400 Kc.	333,400 Kc.	336,400 Kc.	339,400 Kc.	342,400 Kc.
331,000	334,000	337,000	340,000	343,000
331,600	334,600	337,600	340,600	343,600
332,200	335,200	338,200	341,200	
332,800	335,800	338,800	341,800	

(b) STL broadcast stations will be authorized to employ frequency modulation only.

(c) The maximum frequency swing employed by STL broadcast stations shall not be in excess of 200 kilocycles.

(d) The licensee of each STL broadcast station shall install and operate a directional antenna designed so that the gain in power toward the receiver shall be 10 (field gain 3.16) times the free space field from a doublet (137.6 mv/m for 1 kw. at one mile). In all other directions 30° or more off the line to receiver, the power gain shall not exceed 1/4 the free space field gain from a doublet.

4.35 POWER. STL broadcast stations will be licensed with a power output not in excess of that necessary to render a satisfactory service.

4.36 REQUIRED EXPERIMENTATION. The licensee of each STL broadcast station is required to conduct experimentation with regard to the following:

(a) Design of equipment and power required to render a satisfactory service.

(b) Design and adjustment of directional transmitting antennas.

(c) Design and location of receiving antennas.

4.37 SUPPLEMENTAL REPORT WITH RENEWAL APPLICATION. A supplemental report shall be filed with and made a part of each application for renewal of application and shall include statements as to the following items.

(a) Total hours of operation.

(b) Continuity of service, causes and duration of any interruptions.

(c) Power required to deliver satisfactory signal at receiver.

(d) Data on design, adjustments, and operation of directional receiving and transmitting antennas.

(e) Interference to service resulting from other stations or other sources.

(f) Cost of transmitter and receiver installation and expense of operation.

(g) Overall fidelity of equipment, frequency and amplitude.

INTERNATIONAL BROADCAST STATIONS

4.41 DEFINED. The term "international broadcast station" means a station licensed for the transmission of broadcast programs for international public reception. (Frequencies for these stations are allocated from bands assigned, between 6000 and 26,600 kilocycles, for broadcasting by international agreement.)

4.42 LICENSING REQUIREMENTS; NECESSARY SHOWING. A license for an international broadcast station will be issued only after a satisfactory showing has been made in regard to the following, among others:

(a) That there is a need for the international broadcast service proposed to be rendered.

(b) That the necessary program sources are available to the applicant to render an effective international service.

(c) That the technical facilities are available on which the proposed service can be rendered without causing interference to established international stations having prior registration and occupancy in conformity with existing international conventions or regulations on the frequency requested.

(d) That directive antennas and other technical facilities will be employed to deliver maximum signals to the country or countries for which the service is designed.

(e) That the production of the program service and the technical operation of the proposed station will be conducted by qualified persons.

(f) That the applicant is technically and financially qualified and possesses adequate technical facilities to carry forward the service proposed.

(g) That the public interest, convenience, and necessity will be served through the operation of the proposed station.

4.43 SERVICE; COMMERCIAL OR SPONSORED PROGRAMS. (a) A licensee of an international broadcast station shall render only an international broadcast service which will reflect the culture of this country and which will promote international good will, understanding, and cooperation. Any program solely intended for, and directed to an audience in the continental United States does not meet the requirements for this service.

(b) Such international broadcast service may include commercial or sponsored programs; Provided, that -

(1) Commercial program continuities give no more than the name of the sponsor of the program and the name and general character of the commodity, utility or service, or attraction advertised.

(2) In case of advertising a commodity, the commodity is regularly sold or is being promoted for sale on the open market in the foreign country or countries to which the program is directed in accordance with paragraph (c) of this section.

(3) In case of advertising an American utility or service to prospective tourists or visitors to the United States, the advertisement continuity is particularly directed to such persons in the foreign country or countries where they reside and to which the program is directed in accordance with paragraph (c) of this section.

(4) In case of advertising an international attraction (such as a world fair, resort, spa, etc.) to prospective tourists or visitors to the United States, the oral continuity concerning such attraction is consistent with the purpose and intent of this section.

(5) In case of any other type of advertising, such advertising is directed to the foreign country or countries and to which the program is directed in accordance with paragraph (c) of this section and is consistent with the purpose and intent of this section.

(c) The areas or zones established to be served by international broadcast stations are the foreign countries of the world, and directive antennas shall be employed to direct the signals to specific countries. The antenna shall be so designed and operated that the signal (field intensity) toward the specific foreign country or countries served shall be at least 3.16 times the average effective signal from the station (power gain of 10).

(d) An international broadcast station may transmit the program of a standard broadcast station or network system; Provided, the conditions in paragraph (b) of this section in regard to any commercial continuities are observed and when station identifications are made, only the call letter designation of the international station is given on its assigned frequency; And provided further, that in the case of chain broadcasting, the program is not carried simultaneously by another international station (except another station owned by the same licensee operated on a frequency in a different group to obtain continuity of signal service), the signals from which are directed to the same foreign country or countries.

(e) Station identification, program announcements, and oral continuity shall be made with international significance (language particularly) which is designed for the foreign country or countries for which the service is primarily intended.

(f) (1) Each licensee of an international broadcast station shall make verbatim mechanical records of all international programs transmitted.

(2) The mechanical records, and such manuscripts, transcripts, and translations of international broadcast programs as are made shall be kept by the licensee for a period of two years after the date of broadcast and shall be furnished the Commission or be available for inspection by representatives of the Commission upon request.

(3) If the broadcast is in a language other than English, the licensee shall furnish the Commission upon request such record and scripts together with complete translations in English.

4.44 FREQUENCY ASSIGNMENT. (a) The following groups of frequencies are allocated for assignment to international broadcast stations:

Group A	Group B	Group C	Group D	Group E	Group F	Group G	Group H
Kc.	Kc.	Kc.	Kc.	Kc.	Kc.	Kc.	Kc.
6020	9510	11,710	15,110	15,250	17,760	21,460	25,600
6040	9530	11,750	15,150	15,270	17,780	21,480	25,625
6060	9570	11,770	15,170	15,290	17,800	21,520	25,650
6080	9590	11,790	15,190	15,310	17,830	21,540	25,675
6100	9650	11,810	15,210	15,330		21,570	25,700
6140	9670	11,830	15,230			21,590	25,725
6170		11,850				21,610	25,750
6190		11,870				21,630	25,775
		11,890				21,650	25,800
							25,825
							25,850

(b) A separate license and call letter designation will be issued for each frequency except that where frequencies in two or more groups are required to maintain a particular international broadcast service to certain foreign country or countries, one frequency from each of the groups required may be authorized by one license and one call letter designation. In such cases these frequencies shall not be used consecutively during a day as required and they shall not be used simultaneously either on the same transmitter or different transmitters.

(c) Not more than one frequency in any one group in paragraph (a) of this section will be assigned to a station. Any frequency assigned to an international broadcast station shall also be available, during hours when such frequency is not regularly used by such station or when no objectionable interference would be caused to the service rendered by any existing international broadcast station, for assignment to other international broadcast stations.

4.45 POWER REQUIREMENT. No international broadcast station will be authorized to install equipment or licensed for operation with a power less than 50 kilowatts. (Effective as applying to existing stations July 1, 1941).

4.46 SUPPLEMENTAL REPORT WITH RENEWAL APPLICATION. A supplemental report shall be filed with and made a part of each application for renewal of license and shall include statements of the following:

(a) The number of hours operated on each frequency.

(b) A list of programs transmitted of special international interest.

(c) Outline of reports of reception and interference and conclusions with regard to propagation characteristics of the frequency assigned.

4.47 FREQUENCY CONTROL. The transmitter of each international broadcast station shall be equipped with automatic frequency control apparatus so designed and constructed that it is capable of maintaining the operating frequency within plus or minus 0.005 percent of the assigned frequency.

VISUAL BROADCAST SERVICE

4.61. DEFINED. The term "visual broadcast service" means a service rendered by stations broadcasting images for general public reception. There are two classes of stations recognized in the visual broadcast service, namely: Television broadcast stations and facsimile broadcast stations.

EXPERIMENTAL TELEVISION BROADCAST STATIONS

4.71 DEFINED. (a) The term "experimental television broadcast station" means a station licensed for experimental transmission of transient visual images of moving or fixed objects for simultaneous reception and reproduction by the general public. (The transmission of synchronized sound (aural broadcast) is considered an essential phase of television broadcast and one license will authorize both visual and aural broadcast.)

(b) Under these rules for experimental television broadcast stations, the Commission will authorize experimental television relay broadcast stations for transmitting from points where suitable wire facilities are not available, programs for broadcast by one or more television broadcast stations. Such authorization will be granted only to the licensee of a television broadcast station.

4.72 PURPOSE. A license for an experimental television broadcast station will be issued for the purpose of carrying on research and experimentation for the advancement of television broadcasting which may include tests of equipment, training of personnel, and experimental programs as are necessary for the experimentation.

4.73 LICENSING REQUIREMENTS, NECESSARY SHOWING. A license for a television broadcast station will be issued only after a satisfactory showing has been made in regard to the following:

(1) That the applicant has a definite program of research and experimentation in the technical phases of television broadcasting, which indicates reasonable promise of substantial contributions to the developments of the television art.

(2) That upon the authorization of the proposed station, the applicant can and will proceed immediately with its program of research and experimentation.

(3) That the transmission of signals by radio is essential to the proposed program of research and experimentation.

(4) That the program of research and experimentation will be conducted by qualified personnel.

(5) That the applicant is legally, financially, technically, and otherwise qualified to carry forward the program.

(6) That public interest, convenience, or necessity will be served through the operation of the proposed station.

4.74 CHARGES. No charges either direct or indirect shall be made by the licensee of an experimental television station for the production or transmission of either aural or visual programs transmitted by such station except that this section shall not apply to the transmission of commercial programs by an experimental television relay broadcast station for retransmission by a television broadcast station.

4.75 ANNOUNCEMENTS. (a) Station Identification - A licensee of a television broadcast station shall make station identification announcement aurally and visually (call letters and location) at the beginning and ending of each time of operation and during operation on the hour.

(b) At the time station identification announcements are made, there shall be added the following

"This is a special television broadcast made by the authority of the Federal Communications Commission for experimental purposes."

4.76 OPERATING REQUIREMENTS. (a) Each licensee of a television broadcast station shall diligently prosecute its program of research from the time its station is authorized.

(b) Each licensee of a television station will from time to time make such changes in its operations as may be directed by the Commission for the purpose of promoting worthwhile experimentation and improvement in the art of television broadcasting.

4.77 FREQUENCY ASSIGNMENT. (a) The following groups of channels are available for assignment to television broadcast stations licensed experimentally:

GROUP A Channel No.		GROUP B Channel No.		GROUP C
1	50,000 - 56,000 Kc.	8	162,000 - 168,000 Kc.	Any 6000 Kc. band above 300,000 Kc. excluding band 400,000 to 401,000 Kc.
2	60,000 - 66,000 Kc.	9	180,000 - 186,000 Kc.	
3	66,000 - 72,000 Kc.	10	186,000 - 192,000 Kc.	
4	78,000 - 84,000 Kc.	11	204,000 - 210,000 Kc.	
5	84,000 - 90,000 Kc.	12	210,000 - 216,000 Kc.	
6	96,000 -102,000 Kc.	13	230,000 - 236,000 Kc.	
7	102,000 - 108,000 Kc.	14	236,000 - 242,000 Kc.	
		15	258,000 - 264,000 Kc.	
		16	264,000 - 270,000 Kc.	
		17	282,000 - 288,000 Kc.	
		18	288,000 - 294,000 Kc.	

(b) No experimental television broadcast station will be authorized to use more than one channel in Group A except for good cause shown. Both aural and visual carriers with sidebands for modulation are authorized but no emission shall result outside the authorized channel.

(c) No persons (including all persons under common control) shall control directly or indirectly, two or more experimental television broadcast stations (other than television relay broadcast stations) unless a showing is made that the character of the programs of research require a licensing of two or more separate stations.

(d) A license for an experimental television broadcast station will be issued only on the condition that no objectionable interference will result from the transmissions of the station to be the regular program transmissions of television broadcast stations. It shall at all times be the duty of the licensee of an experimental television broadcast station to ascertain that no interference will result from the transmission of its station. With regard to interference with the transmissions of an experimental television broadcast station or the experimental or test transmissions of a television broadcast station, the licensees shall make arrangements for operations to avoid interference.

(e) Channels in Groups B and C may be assigned to experimental television stations to serve auxiliary purposes such as television relay stations. No mobile or portable station will be licensed for the purpose of transmitting television programs to the public directly.

4.78 POWER. The operating power of a television station shall be adequate for but not in excess of that necessary to carry forward the program of research and in no case in excess of the power specified in its license.

4.79 REPORTS. (a) A report shall be filed with each application for renewal of station license which shall include a statement of each of the following:

(1) Number of hours operated.

(2) Full data on research and experimentation conducted including the type of transmitting and studio equipment used and their mode of operation.

(3) Data on expense of research and operation during the period covered.

(4) Power employed, field intensity measurements and visual and aural observations and the types of instruments and receivers utilized to determine the service area of station and the efficiency of respective types of transmissions.

(5) Estimated degree of public participation in reception and the results of observations as to the effectiveness of types of transmission.

(6) Conclusions, tentative and final.

(7) Program for further developments in television broadcasting.

(8) All developments and major changes in equipment.

(9) Any other pertinent developments.

(b) Special or progress reports shall be submitted from time to time as the Commission shall direct.

FACSIMILE BROADCAST STATIONS

4.91 DEFINED. The term "facsimile broadcast station" means a station licensed to transmit images of still objects for record reception by the general public.

4.92 LICENSING REQUIREMENTS. A license for a facsimile broadcast station will be issued only after a satisfactory showing has been made in regard to the following among others:

(a) That the applicant has a program of research and experimentation which indicates reasonable promise of substantial contribution to the development of the facsimile broadcast service.

(b) That sufficient facsimile recorders will be distributed to accomplish the experimental program proposed.

(c) That the program of research and experimentation will be conducted by qualified engineers.

(d) That the applicant is legally and financially qualified and possesses adequate technical facilities to carry forward the program.

(e) That the public interest, convenience, and/or necessity will be served through the operation of the proposed station.

4.93 CHARGES PROHIBITED; RESTRICTIONS. (a) A licensee of a facsimile broadcast station shall not make any charge, directly or indirectly, for the transmission of programs.

(b) No licensee of any standard broadcast station or network shall make any additional charge, directly or indirectly, for the transmission of some phase of the programs by a facsimile broadcast station, nor shall commercial accounts be solicited by any licensee of a standard broadcast station or network, or others acting in their behalf, upon representation that images concerning that commercial program will be transmitted by a facsimile station.

4.94 FREQUENCY ASSIGNMENT. (a) The following groups of frequencies are allocated for assignment to facsimile broadcast stations which will be licensed experimentally only:

GROUP A	GROUP B	GROUP C	GROUP D
Kc.	Kc.	Kc.	
25,025	43,540	116,110	Any frequency above 300,000 kc. excluding band 400,000 kc. to 401,000 kc.
25,050	43,580	116,230	
25,075	43,620	116,350	
25,100	43,660	116,470	
25,125	43,700		
25,150	43,740		
25,175	43,780		
25,200	43,820		
25,225	43,860		
25,250	43,900		
	43,940		

(b) Other broadcast or experimental frequencies may be assigned for the operation of facsimile broadcast stations on an experimental basis provided a sufficient need therefor is shown and no interference will be caused to established radio stations.

(c) One frequency only will be assigned to a facsimile station from the groups in paragraph (a) of this section. More than one frequency may be assigned under provisions of paragraph (b) of this section if a need therefor is shown.

(d) Each applicant shall specify the maximum modulating frequencies proposed to be employed.

(e) The operating frequency of a facsimile broadcast station shall be maintained in accordance with the frequency tolerance given in section 4.1; Provided, however, where a lesser tolerance is necessary to prevent interference, the Commission will specify the tolerance.

(f) A facsimile broadcast station authorized to operate on frequencies regularly allocated to other stations or services shall be required to abide by all rules governing the stations regularly operating thereon, which are applicable to facsimile broadcast stations and are not in conflict with sections 4.1 to 4.11, inclusive, of these rules.

4.95 POWER. The operating power of a facsimile broadcast station shall not be in excess of that necessary to carry forward the program of research; Provided, however, not more than 1000 watts will be authorized on a frequency in group A. The operating power may be maintained at the maximum rating or less, as the conditions of operation may require.

4.96 SUPPLEMENTAL REPORT WITH RENEWAL APPLICATION. A supplemental report shall be filed with and made a part of each application for renewal of license and shall include statements of the following:

(a) Number of hours operated for transmission of facsimile programs.

(b) Comprehensive report of research and experimentation conducted.

(c) Conclusions and program for further developments of the facsimile broadcast service.

(d) All developments and major changes in equipment.

(e) Any other pertinent developments.

NONCOMMERCIAL EDUCATIONAL BROADCAST STATIONS

4.131 DEFINED. The term "noncommercial educational broadcast station" means a station licensed to an organized nonprofit educational agency for the advancement of its educational work and for the transmission of educational and entertainment programs to the general public.

4.132 OPERATION AND SERVICE. The operation of, and the service furnished by, noncommercial educational broadcast stations shall be governed by the following regulations:

(a) A noncommercial educational broadcast station will be licensed only to an organized nonprofit educational agency and upon a showing that the station will be used for the advancement of the agency's educational program particularly with regard to use in an educational system consisting of several units.

(b) Each station may transmit programs directed to specific schools in the system for use in connection with the regular courses as well as routine and administrative material pertaining to the school system and may transmit educational and entertainment programs to the general public.

(c) Each station shall furnish a nonprofit and noncommercial broadcast service. No sponsored or commercial program shall be transmitted nor shall commercial announcements of any character be made. A station shall not transmit the programs of other classes of broadcast stations unless all commercial announcements and commercial references in the continuity are eliminated.

4.133 POWER. The operating power of noncommercial educational broadcast stations shall be not less than 10C watts or greater than 1000 watts unless a definite need for greater power is shown.

4.134 FREQUENCY CONTROL. The transmitter of each noncommercial educational broadcast station shall be equipped with automatic frequency control apparatus so designed and constructed that it is capable of maintaining the operating frequency within plus or minus 0.01 percent of the assigned frequency.

4.135 OPERATING SCHEDULE. Noncommercial educational broadcast stations are not required to operate on any definite schedule or minimum hours.

4.136 EQUIPMENT REQUIREMENTS. The transmitting equipment, installation, and operation as well as the location of the transmitter shall be in conformity with the requirements of good engineering practice as released from time to time by the Commission.

4.137 FREQUENCIES. (a) The following frequencies are allocated for assignment to noncommercial educational broadcast stations.

KILOCYCLES

42,100	42,700
42,300	42,900
42,500	

(b) Stations serving the same area will not be assigned adjacent frequencies.

(c) Frequency modulation shall be employed exclusively unless it is shown that there is a special need for the use of amplitude modulation.

(d) Only one frequency will be assigned to a station.

DEVELOPMENTAL BROADCAST STATIONS

4.151 DEFINED. The term "developmental broadcast station" means a station licensed to carry on development and research for the advancement of broadcast services along lines other than those prescribed by other broadcast rules or a combination of closely related developments that can be better carried on under one license.

4.152 LICENSING REQUIREMENTS; NECESSARY SHOWING. (a) Licences for developmental broadcast stations will be issued only after a satisfactory showing has been made in regard to the following, among others:

(1) That the applicant has a program of research and development which cannot be successfully carried on under any one of the classes of broadcast stations already allocated, or is distinctive from those classes, or combination of closely related developments that involve different phases of broadcasting which can be pursued better under one license.

(2) That the program of research has reasonable promise of substantial contribution to the development of broadcasting, or is along lines not already thoroughly investigated.

(3) That the program of research and experimentation will be conducted by qualified persons.

(4) That the applicant is legally and financially qualified and possesses adequate technical facilities to carry forward the program.

(5) That the public interest, convenience, and necessity will be served through the operation of the proposed station.

(b) A separate developmental broadcast station license will be issued for each major development proposed to be carried forward. When it is desired to carry on several independent developments, it will be necessary to make satisfactory showing and obtain a license for each.

4.153 PROGRAM SERVICE; CHARGES PROHIBITED; ANNOUNCEMENTS. (a) A licensee of developmental broadcast stations shall broadcast programs when they are necessary to the experiments being conducted. No regular program service shall be broadcast unless specifically authorized by the license.

(b) A licensee of a developmental broadcast station shall not make any charge, directly or indirectly, for the transmission of programs, but may transmit the programs of a standard broadcast station or network including commercial programs, if the call letter designation when identifying the developmental broadcast station is given on its assigned frequency only and the statement is made over the developmental broadcast station that the program of a broadcast station or network (identify by call letters or name of network) is being broadcast in connection with the developmental work. In case of the rebroadcast of the program of any broadcast station, section 4.10 applies.

4.154 FREQUENCY ASSIGNMENT. (a) The following frequencies are allotted for assignment to developmental broadcast stations:

1614	12,855 } 12,862.5 12,870	116,050
2396 } 2398 2400	17,300 } 17,310 17,320	116,250
3490 } 3492.5 3495	23,100	116,450
4795 } 4797.5 4800	30,660	116,850
6420 } 6425 6430	31,020	117,050
9130 } 9135 9140	31,140	117,250
	31,180	117,650
	31,540	118,050
	33,340	118,250
	33,460	118,450
	33,620	118,650
	35,060	118,850
	35,460	156,525
	37,060	156,975
	37,140	157,425
	37,540	157,725
	39,140	158,175
	39,460	159,075
	39,540	160,425
		161,325
		161,775
		162,000 to 168,000
		210,000 to 216,000
		264,000 to 270,000
		300,000 to 400,000
		401,000 and above.

(b) A license will be issued for more than one of these frequencies upon a satisfactory showing that there is need therefor.

(c) The frequencies suited to the purpose and in which there appears to be the least or no interference to established stations shall be selected.

(d) In cases of important experimentation which cannot be conducted successfully on the frequencies allocated in paragraph (a) of this section, the Commission may authorize developmental broadcast stations to operate on any frequency allocated for broadcast stations or any frequencies allocated for other services under the jurisdiction of the Commission upon satisfactory showing that such frequencies can be used without causing interference to established services.

4.155 FREQUENCY TOLERANCE. (a) The operating frequency of a developmental broadcast station shall be maintained in accordance with the frequency tolerance given in section 4.1; Provided, however, where lesser tolerance is necessary to prevent interference the Commission will specify the tolerance.

(b) The operating power of a developmental broadcast station shall not be in excess of that necessary to carry on the program of research. The operating power may be obtained at the maximum rating or less, as the conditions of operation may require.

4.156 SUPPLEMENTAL REPORT WITH RENEWAL APPLICATION. A supplemental report shall be filed with and made a part of each application for renewal of license and shall include statements of the following, among others:

(a) The number of hours operated.

(b) Comprehensive report on research and experiments conducted.

(c) Conclusions and program for further development of the broadcast service.

(d) All developments and major changes in equipment.

(e) Any other pertinent developments.

4.157 FREQUENCY RESTRICTIONS. A developmental broadcast station authorized to operate on frequencies regularly allocated to other stations or services, shall be required to abide by all rules governing the stations operating regularly thereon which are applicable to developmental broadcast stations and are not in conflict with sections 4.1 to 4.11, inclusive, and sections 4.151 to 4.156 inclusive, of these rules.

PART 9 - RULES AND REGULATIONS GOVERNING AVIATION SERVICES

DEFINITIONS

9.1 AVIATION SERVICE. The term "aviation service" means a radiocommunication or special service carried on by aircraft stations, airport control stations, aeronautical stations, aeronautical fixed stations, instrument landing stations, and flying school stations.

9.2 PUBLIC AVIATION SERVICE means a radiocommunication service open to public correspondence (paid or toll messages) to provide public communications to, from, and between aircraft in flight.

9.3 AIRCRAFT STATION means a radio station on board any aircraft (either heavier-than-air or lighter-than-air) other than public service aircraft station.

9.4 SCHEDULED AIRCRAFT means an aircraft regularly flying a fixed route.

9.5 NONSCHEDULED AIRCRAFT means other than scheduled aircraft.

9.6 AERONAUTICAL STATION means a station used primarily for radiocommunications with aircraft stations, but which may also carry on a limited fixed service with other aeronautical stations in connection with the handling of communications relating to the safety of life and property in the air.

9.7 AERONAUTICAL FIXED STATION means a station used in the fixed service for the handling of point-to-point communications relating solely to actual aviation needs.

9.8 CHAIN OF STATIONS in the aviation service is a series of coordinated stations operating on frequencies allocated to aviation services in accordance with an approved plan.

9.9 AIRPORT CONTROL STATION means a station provided for furnishing communications limited to actual aviation needs between an airport control tower and aircraft stations in the immediate vicinity of the airport (approximately within 30 miles distance or 10 minutes flight of the airport).

9.10 FLYING SCHOOL STATION means a station used for communications pertaining to instruction to students or pilots while in flight.

9.11 INSTRUMENT LANDING STATION is a special service station for facilitating the landing of aircraft.

9.12 RADIO MARKER STATION means a station marking a definite location on the ground as an aid to air navigation.

9.13 PUBLIC SERVICE AIRCRAFT STATION means a station licensed on board an aircraft for the purpose of carrying on public aviation service.

9.14 PUBLIC SERVICE AERONAUTICAL STATION means a land station licensed for communicating with public service aircraft stations for the purpose of carrying on a public aviation service.

GENERAL REGULATIONS

LICENSE

9.21 LICENSE PERIODS. The license period for all stations in the aviation service shall be for 1 year unless otherwise stated in the instrument of authorization. The date of expiration of license for all classes of stations operating in the aviation service, unless otherwise specified, shall be as follows:

(a) For stations in the aviation service, other than aircraft stations and for all stations in Alaska, the first day of March of each year.

(b) For scheduled aircraft stations in the aviation service other than in Alaska, the first day of April of each year.

(c) For non-scheduled aircraft stations in the aviation service other than in Alaska, the first day of August of each year.

(d) For all classes of stations in the aviation service in Alaska, the first day of January of each year.

9.22 POSTING STATION LICENSES. The station licenses of stations in the aviation service shall be conspicuously posted at the place where the control operator is located except that in aircraft stations the license may be posted or kept at any convenient easily accessible location in the aircraft.

9.23 POSTING OPERATOR LICENSES. The original license of each station operator shall be conspicuously posted at the place he is on duty, or in the case of mobile units either the license or verification card must be kept in his personal possession.

TESTS

9.31 EQUIPMENT AND SERVICE TESTS AUTHORIZED FOR AERONAUTICAL AND AERONAUTICAL FIXED STATIONS. Equipment and service tests as authorized in sections 2.42 and 2.43 may be conducted provided that the necessary precautions are taken to avoid interference.

9.32 ROUTINE TESTS. The licenses of all classes of stations in the aviation service are authorized to make such routine tests as may be required for the proper maintenance of the station provided that precautions are taken to avoid interference with any station. Tests on 3105 and 6210 kilocycles using a regular antenna system can be made only at such times when no interference will be caused, and, if in range of an airport control station or Civil Aeronautics Authority station, only after permission is secured from such stations before commencing the tests.

LOGS

9.41 INFORMATION REQUIREO IN STATION LOGS. All stations in the aviation service except aircraft stations must keep an adequate log showing; (1) hours of operation; (2) frequencies used; (3) stations with which communication was held; and (4) signature of operator(s) on duty.

9.42 STATION LOGS PUBLIC AVIATION SERVICE. In addition to all the requirements in section 9.41 above, all stations (both public service aircraft station and public service aeronautical station) in the public aviation service must keep a file of all record communications handled and a list of radiotelephone contacts established.

9.43 REQUIRED RETENTION PERIOD. The logs in the aviation service, other than public aviation service, may be destroyed after a period of 3 months except in those circumstances where retention of the logs for a longer period is specifically provided for in other rules.

INSPECTIONS

9.51 AVAILABILITY FOR INSPECTIONS. All classes of stations in the aviation service shall be made available for inspection upon request of an authorized representative of the Federal Government.

9.52 RESPONSIBILITY OF LICENSEE. It is the responsibility of the licensees of aircraft radio stations to submit their stations for inspection by a representative of the Commission at least once during the license period.

COMMUNICATIONS

9.61 METHODS OF IDENTIFICATION. The aircraft name, company number, trip number, official registry number, or other identification approved by the Commission may be used in lieu of the call letters; provided that adequate records are maintained to permit ready identification of individual aircraft. Also the name of the city or airport in which other classes of stations are located may be used in lieu of the call letters of the station when using telephony. In the case of stations using telegraphic emissions, the call letters designated in the license shall be used at the end of each sequence of communication to one or more stations.

9.62 PERMISSIBLE COMMUNICATIONS. All stations in the aviation service, except those stations licensed for public aviation service, shall transmit only communications relating to and necessary for aircraft operation and the protection of life and property in the air.

9.63 PRIORITY OF AVIATION COMMUNICATIONS. (a) The regular routine communications of stations in the aviation service are essential to the safe operation of aircraft and shall have priority over the public aviation service stations.

(b) The radio operator in charge of the aircraft station shall suspend operations of aviation public service stations when such operations will delay or interfere with messages pertaining to safety of life and property or when ordered to do so by the captain of the aircraft.

(c) The operation of public aviation service stations shall in no way interfere with the radiocommunications of the aviation service.

(d) In cases where the aviation public service aircraft station license is issued to cover auxiliary equipment of the regular aircraft station, public communications shall be restricted to the extent necessary for the safe operation of aircraft as determined by the person in charge of the aircraft.

FREQUENCIES

9.71 AIRPORT CONTROL FREQUENCIES. (a) 130,400 kc.

(b) 129,200 kc., 129,800, 131,000, 131,600 kc. (These frequencies to be used, in the order named, in the event that the geographical location of airport stations is such as to render the use of the frequency 130,400 kc. impracticable.)

(c) 129,000 kc., 129,400, 129,600, 130,000, 130,200, 130,600, 130,800, 131,200, 131,400, 131,800 kc. (These frequencies will be assigned, when necessary, on the basis of an individual study of the circumstances surrounding each case.

(d) 278 kilocycles; This frequency is available for assignment in lieu of a high frequency, except that its use at some airports, after certain dates, must be supplemented by a service on one of the high frequencies, in accordance with the following schedule:

(1) Airports having fifteen or more scheduled aircraft landings daily, after January 1, 1942.

(2) Airports having six or more but less than fifteen scheduled aircraft landings daily, after July 1, 1942.

(3) All other airports after January 1, 1943, provided, however, that upon application therefor the Commission may exempt any station from the high-frequency service requirement when it appears that in the preservation of life and property in the air such service is not required at that station.

9.72 MISCELLANEOUS CALLING AND WORKING FREQUENCIES. 333 Kc.: General calling frequency for aircraft stations operating outside the North American continent on trans-oceanic flights.

375 Kc.: International directing-finding frequency for use outside the continental United States.

457 Kc.: Working frequency exclusively for aircraft on sea flights desiring an intermediate frequency.

500 Kc.: International calling and distress frequency for ships and aircraft over the seas.

1638 Kc.: Air navigation frequency, available for aeronautical stations; scheduled and nonscheduled aircraft.

3105 and 6210 Kc.: National and international aircraft calling and working frequencies primarily for use by non-scheduled aircraft. The use of these frequencies is restricted to communications pertaining solely to aircraft operation and the protection of life and property.

3120 Kc.: National aircraft working frequency primarily for use by non-scheduled aircraft. The use of this frequency is restricted to communications pertaining solely to aircraft operation and the protection of life and property.

140,100 Kc.: National calling and working frequency available to aircraft for general communication purposes. The use of this frequency is restricted to communications pertaining solely to aircraft operation and the protection of life and property.

MISCELLANEOUS MARITIME FREQUENCIES. Calling and working frequencies of ship stations may also be assigned to aircraft stations for the purpose of communicating with coastal stations, or ship stations, when aircraft are in flight over the seas; available for A1, A2, and A3 emission in conformity with Part 8 of the

Rules Governing Ship Service; provided the Commission is satisfied in each case that undue interference will not be caused to the service of ship or coastal stations.

9.73 FREQUENCIES AVAILABLE FOR ASSIGNMENT TO CHAIN SYSTEMS. The frequencies allocated to the several chains are as follows:

(a) NORTHERN TRANSCONTINENTAL CHAIN AND FEEDERS (RED). Available for aircraft and aeronautical stations:

3147.5	3372.5	5572.5	8240
3162.5	3467.5	5582.5	12330
3172.5	5122.5	5592.5	140800
3182.5	5162.5	5662.5	140940
3322.5	5172.5	5697.5	141080
		5825	141220

Available for aeronautical fixed stations:

12330

(b) MIDTRANSCONTINENTAL CHAIN AND FEEDERS (BLUE). Available for aeronautical and aircraft stations:

2906	4110	4967.5	10125
3062.5	4937.5	5692.5	141360
3072.5	4947.5	6510	141500
3088	4952.5	6520	141640
			141780

Available for aeronautical fixed stations:

2732	6510	8015	10855
4110	6520	10125	

(c) SOUTHERN TRANSCONTINENTAL CHAIN AND FEEDERS (BROWN). Available for aeronautical and aircraft stations:

2946	3432.5	5612.5	6550
3137.5	4732.5	5622.5	7700
3222.5	5365	5632.5	10080
3232.5	5390	5652.5	141920
3242.5	5480	5672.5	142060
3257.5	5602.5	5887.5	142200
			142340

Available for aeronautical fixed stations:

2612	4730	5425	7700
2998	5255	6550	10080
3050	5365	6820	10190
3290			18360
4690			

(d) EASTERN CONTINENTAL CHAIN AND FEEDERS (GREEN). Available for aeronautical and aircraft stations:

2608	2986	5707.5	11960
2898	4122.5	6795	140240
2922	4335	6805	140380
2946	4742.5	8565	140520
	5652.5		140660

Available for aeronautical fixed stations:

2608	4735	5310	8130
2748	4740	6795	10855
3290	4745	6805	11960
4115			
4335			

(e) NORTHWESTERN CONTINENTAL CHAIN AND FEEDERS (PURPLE). Available for aeronautical and aircraft stations:

2994	4917.5	5887.5	142900
3005	5275	6490	143040
3127.5	5377.5		143180

Available for aeronautical fixed stations:

2644	5220	6490	10965
	5275	8700	

(f) MIDCONTINENTAL CHAIN AND FEEDERS (YELLOW). Available for aeronautical and aircraft stations:

3447.5	4650	5215	142480
3457.5	5032.5	5682.5	142620
3485	5042.5	8070	142760

Available for aeronautical fixed stations:

2636 2640 4650 5215 8070 9200 11910

(g) INTERCONTINENTAL CHAIN AND FEEDERS (ORANGE). Frequencies available for traffic control over the international routes as follows:

2870 Kc. for traffic control over Inter-American Route (1) and Transpacific Route (2).

2912 Kc. for traffic control over Europe-North America Route (3) and Arctic Route (4).

I. Available for aeronautical and aircraft stations on the routes designated:

1. INTER-AMERICAN ROUTE: Available for aeronautical and aircraft stations:

3082.5	6583	8225	17257
5405	6590	8233	17274
5692.5	6597	11381	23301
6557	8217	11394	23324

Available for aeronautical fixed stations:

2648	5375	9310	10955
2980	5945	10535	16240
			16290

2. TRANSPACIFIC ROUTE: Available for aeronautical and aircraft stations:

2976	6570	8577	17319
5165	6577	11356	17336
6557	8561	11369	23346
6563	8569	12824	23369

Available for aeronautical fixed stations:

2964	5925	8720	16280
4060	8120	12180	23025

3. EUROPE-NORTH AMERICA ROUTE: Available for aeronautical and aircraft stations:

2912	6570	8554	17288
3285	6577	11306	17350
6543	8538	11319	17367
6563	8546	12776	23211
		12788	23234

Available for aeronautical fixed stations:

2980	5920	11470	16250
4055	8120	12165	16440
5375	8720	12180	

4. EUROPE-ARCTIC ROUTE: Available for aeronautical and aircraft stations:

1674	6530	6550	11344
3285	6537	6557	17288
6523	6543	8485	23256
		11331	23279

Available for aeronautical fixed stations:

1722 2648 8720 10955

II. Available for aeronautical and aircraft stations on routes 1, 2, 3, and 4 listed above.

143460 143600 143740

9.74 LIGHTER-THAN-AIR-CRAFT FREQUENCIES. The following additional frequencies may be assigned to lighter-than-air-craft and to aeronautical stations serving lighter-than-air-craft:

2930 6615 11910

9.75 RADIO MARKER STATION FREQUENCY:

75000

9.76 INSTRUMENT LANDING FREQUENCIES. Glide path and localizer.

93500 93900 94300 109500 109900 110300

9.77 FLYING SCHOOL FREQUENCIES:

33420 33660 37860 39060

9.78 PUBLIC AVIATION SERVICE FREQUENCIES. The frequencies 143320 Kc. and 143460 Kc. are available experimentally for the development of a domestic public aviation radiotelephone service. The frequencies available to ship telegraph and ship telephone stations are also available to public service aircraft stations on intercontinental or transoceanic air routes for the handling of public correspondence in the same manner and to the same extent that they are available to ships of the United States and under restrictions hereinafter provided.

TOLERANCE

9.81 FREQUENCY TOLERANCE. The frequency tolerance of stations in the aviation service shall be as follows:

Class of Station	Frequency Band	Percent Tolerance: Equipment before Jan. 1, 1940	Percent Tolerance: Authorized after Jan. 1, 1940
Aeronautical..................	....	0.04	0.02
Aeronautical, fixed...........	Below 6 megacycles.......	.03	.01
	Above 6 megacycles.......	.02	.01
Aircraft using frequencies assigned aeronautical stations.	....	.04	.02
Aircraft frequencies available to ships....................	....	.05	.05
Aircraft using other frequencies	....	.04	.02
Airport control stations.......	278 kilocycles...........	.10	.02
Airport control stations using other frequencies............	....	....	.02
All stations in aviation service other than airport control stations using frequencies above 30 megacycles.....	....	....	.02
All airport control stations using frequencies above 30 megacycles..................	129000 - 132000 kilocycles	....	.01

9.82 MEASUREMENT PROCEDURE. The licensee of each station shall provide for measurement of the station frequency, or frequencies, regularly used in accordance with instructions issued from time to time by the Commission and establish procedure for regular checking. These measurements of station frequency shall be made by means independent of the frequency control of the transmitter and shall be of such an accuracy that the limit of error is within the frequency tolerance allowed the station.

EMISSION

9.83 TYPE OF EMISSION. Stations in the aviation service shall use amplitude modulation and type A1, A2, and A3 emission, as may be appropriate.

AIRCRAFT STATIONS

9.91 AIRCRAFT STATIONS. Communications by an aircraft station shall be limited to the necessities of safe aircraft navigation. Normally contacts with airport control stations shall not be attempted unless the aircraft is within the control area of the airport (approximately within 30 miles distance or 10 minutes flight).

AERONAUTICAL AND AERONAUTICAL FIXED STATIONS

9.101 SERVICE AERONAUTICAL STATION. Aeronautical stations shall provide non-public service without discrimination to all scheduled aircraft the owners of which make cooperative arrangements for the operation and maintenance of the aeronautical stations which are to furnish such service and for shared liability in the operation of stations. In addition, this class of station shall provide reasonable and fair service to non-scheduled aircraft in accordance with the provisions of these rules.

9.102 SERVICE AERONAUTICAL FIXED STATIONS. Aeronautical fixed stations are authorized primarily for the handling of communications in connection with and relating solely to the actual aviation needs of the licensees, and then only where frequencies are allocated to a chain and cooperative participants upon the basis of equality.

9.103 POWER. Aeronautical or aeronautical fixed stations will not be licensed to use more than 1 kilowatt on the frequencies above 1500 kilocycles unless on proper showing the Commission shall authorize a greater power, in which event the operating frequency must be maintained within 0.02 percent of the assigned frequency and suitable filters must be embodied in the equipment to limit the frequency band of emission to 5 kilocycles.

9.104 EMERGENCY SERVICE. The licensee of an aeronautical fixed station shall be required to transmit, without charge or discrimination, all necessary messages in times of public emergency which involve the safety of life or property.

AIRPORT CONTROL STATIONS

9.111 RECEIVING WATCH ON 3105 KILOCYCLES. The licensee of an airport control station shall without discrimination provide non-public service for any and all aircraft. Such licensee shall maintain a continuous listening watch on the aircraft calling and working frequency 3105 kilocycles, and also be prepared to render a non-public communication service, during all hours of the day and night; Provided, however, that upon application therefor the Commission may exempt any station from the requirements of this provision when it appears that in the preservation of life and property in the air the maintenance of a continuous watch by such station is not required.

9.112 AIRPORT FACILITIES. Only one airport control station will be licensed to operate at an airport.

9.113 SERVICE TO BE RENDERED. Communications of an airport control station shall be limited to the necessities of safe operation of aircraft using the airport facilities or operating within the airport control area (approximately within 30 miles distance or 10 minutes flight from the airport) and in all cases such stations shall be in a position to render, and shall render, all airport control services.

9.114 COMMUNICATIONS must not be attempted with aircraft beyond the control area of the airport.

9.115 INTERFERENCE. The operation of airport control stations in adjacent airport areas shall be on a non-interference basis only. In case of disagreement between adjacent areas, the Commission will specify the arrangements necessary to eliminate interference.

9.116 POWER. (a) Airport control stations using 278 kilocycles will not be licensed to use more than 15 watts power for type A3 emission.

(b) Localizer transmitters authorized to use the frequency 278 kilocycles may use power in excess of 15 watts provided that the power is limited so as not to produce a field strength of more than 1500 microvolts per meter at one mile from the transmitter location, in the direction of the maximum field.

(c) The power of airport control stations operating on other frequencies shall be 100 watts into an antenna system equivalent to two-crossed dipoles placed at a height from 50 to 80 feet above ground, or 50 watts into an antenna system equivalent to a two-stack, two-crossed dipole or a single loop antenna placed at the same height.

FLYING SCHOOL STATIONS

9.121 ELIGIBILITY FOR STATION LICENSE. Radiocommunication facilities for flying schools may be assigned only to bona fide flying schools and soaring societies.

9.122 LIMITATIONS OF INSTRUCTIONAL FACILITIES. Assignments will be limited to one station to an airport location for one or more flying schools.

9.123 COORDINATED USE OF INSTRUCTIONAL FACILITIES. Where more than one flying school operates from an airport location, coordinated use of a single instructional frequency shall be arranged, placed in the form of a signed agreement, and filed with the Commission.

9.124 USE OF FLYING SCHOOL FREQUENCY. All aircraft engaged in instructional flying in the vicinity of an airport shall transmit only on the flying school frequency assigned to that airport location.

9.125 SUPERVISION BY AIRPORT CONTROL OPERATOR. At any airport at which an airport control station or control tower is in operation, the airport control operator must be given a remote microphone connection to the transmitter operating on the flying school frequency for the transmission of orders or instructions of an emergency nature to students in flight within the control area of the airport (approximately within 30 miles distance or 10 minutes flight of the airport).

9.126 POWER. The power output of flying school stations shall not be more than 50 watts nor less than 15 watts for land stations and not more than 20 watts for aircraft stations.

9.127 FREQUENCY ASSIGNMENTS NON-EXCLUSIVE. No frequency available to a station engaged in instructional flying will be assigned exclusively to any applicant. All stations in this service are required to coordinate operation so as to avoid interference and make the most effective use of assignments.

9.128 PRIVATE SERVICE PROHIBITED. The use of flying school frequencies for other than instruction purposes and promotion of safety of life and property is strictly prohibited.

INSTRUMENT LANDING STATIONS

9.141 BASIS OF GRANT OF FACILITIES. Instrument landing service will not be authorized unless (1) the applicant meets all requirements specified by the Civil Aeronautics Authority and the Federal Communications Commission for the type of installation proposed; (2) the applicant executes a specific agreement to relinquish to the Civil Aeronautics Authority the use of any Government frequencies involved on demand; and (3) the applicant executes a specific agreement to release the facilities and remove his equipment if and when such release or removal may be required by the Government.

PUBLIC AVIATION SERVICE

9.151 STATIONS LICENSED FOR PUBLIC AVIATION SERVICE. Only those stations in the aviation service licensed for public aviation service shall carry on public correspondence, i.e. a paid or toll message service in the sense in which these terms are generally understood. (No separate or additional authorization is necessary for licensed coastal or ship stations to communicate with or handle public messages to or from an aircraft in flight over the sea).

9.152 EXTENT OF SERVICE. All stations licensed in the public aviation service shall provide such service without discrimination to any other station similarly licensed.

9.153 SEPARATION OF LICENSES. Each license or other instrument of authorization for aviation public service shall be separate from the aviation service licenses and shall designate the specific frequencies which the station is authorized to use to carry on public correspondence.

9.154 REQUIREMENT FOR A PUBLIC AVIATION SERVICE STATION. Upon showing that a need exists for public aviation service, a license or other instrument of authorization may be issued for a station for public correspondence provided that a continuous effective listening watch is maintained on the frequency or frequencies used for the aviation safety service messages while public service messages are being handled; and that the installation and system of operation will permit instantaneous interruption of public aviation communications to transmit or receive aviation service messages.

PART 10 - RULES GOVERNING EMERGENCY RADIO SERVICE

DEFINITIONS

10.1 EMERGENCY SERVICE. The term "emergency service" means a radiocommunication service carried on for emergency purposes.

10.2 MUNICIPAL POLICE STATION. The term "municipal police station" means a station used by a municipal or county police department for emergency radiotelephone service with mobile police units.

10.3 STATE POLICE STATION. The term "state police station" means a station used by a state police department for emergency radiotelephone service with mobile police units.

10.4 INTERZONE POLICE STATION. The term "interzone police station" means a station used by a police department for radiotelegraph communication (a) with similarly licensed stations in adjacent zones or with the nearest interzone police station, in case there is no similarly licensed station in the adjacent zone; (b) with stations within the zone, and (c) with mobile police units equipped for radiotelegraph reception.

10.5 ZONE POLICE STATION. The term "zone police station" means a station used by police departments for radiotelegraph communication: (a) with stations within the zone; (b) with mobile police units equipped for radiotelegraph reception; and (c) with stations in adjacent zones, provided, in each case, express permission of the interzone stations in control of communications is obtained in accordance with the operating procedure prescribed by the Commission.

10.6 MARINE FIRE STATION. The term "marine fire station" means a station used for intercommunication between municipal fire departments and fireboats.

10.7 SPECIAL EMERGENCY STATION. The term "special emergency station" means a station used for communications in emergencies in lieu of normal means of communication.

10.8 FORESTRY STATION. The term "forestry station" means a station used for communications necessary for the prevention and suppression of forest fires.

GENERAL RULES GOVERNING EMERGENCY RADIO SERVICES

ELIGIBILITY FOR LICENSE

10.21 POLICE STATIONS. Authorization for the various classes of police stations will be issued only to instrumentalities of government.

10.22 MARINE FIRE STATIONS. Authorizations for marine fire stations will be issued only to municipalities.

10.23 SPECIAL EMERGENCY STATIONS. Authorizations for special emergency stations will be issued only to: (a) organizations established for relief purposes in emergencies and which have a disaster communication plan; (b) to persons having establishments in remote locations which cannot be reached by other means of communication; (c) to public utilities.

10.24 FORESTRY STATIONS. Authorizations for forestry stations will be issued to municipal, state, or private organizations which are legally responsible for the protection of forest areas.

APPLICATIONS

10.31 INDIVIDUAL AND BLANKET APPLICATIONS. Individual applications for instruments of authorization shall be submitted for each station to be located at a fixed location. Blanket applications for authorizations for identical mobile, portable-mobile, or low-powered portable transmitters, submitted by a single applicant to cover equipment to be used in a single coordinated communication system will be accepted. A blanket application may be submitted by a single applicant for a license or modification of license, covering both the land transmitter and mobile, portable-mobile, or low-powered portable transmitters used in a single coordinated communication system.

FREQUENCIES

10.41 STATE AND MUNICIPAL POLICE STATIONS. The following frequencies are allocated for use by state and municipal police stations:

1610*	1690*	2382	2458
1626*	1698*	2390*	2466
1634*	1706*	2406	2474
1642*	1714	2414	2482
1658	1722	2422	2490
1666	1730	2430	
1674	2326	2442	
1682	2366*	2450	

*Subject to the condition that no interference is caused to Canadian stations.

10.42 STATE AND MUNICIPAL ADDITIONAL UNLIMITED POWER. (a) The following additional frequencies are allocated for use by land and portable municipal and state police stations without limitation as to power.

GROUP A

30700	31900	33940	37500	39900
31100	33100	35500	39100	

(b) The following additional frequencies are allocated for use by land and portable municipal and state police stations operating with power not in excees of 250 watts:

GROUP B

31500	35900	37900
33500	37100	39500

(c) Notwithstanding the provisions of (a) and (b) of this section, municipalities and states may be authorized to operate mobile and portable-mobile stations on the frequency, or frequencies, assigned to their land station(s). An instrumentality of government operating mobile units only may be authorized to use a frequency from group A or group B of this section assigned an adjacent instrumentality of government, provided a copy of the agreement entered into between the two for the exchange of service is filed with the Commission.

(d) Municipalities and states desiring more than one land frequency shall, in making application, show a proper need therefor.

(e) Municipalities desiring frequencies for use by portable stations of 1 watt power or less, portable-mobile stations, or mobile stations different from those which may be allocated under section 10.42 (a) and (b), may be authorized to use the following frequencies:

GROUP C

30580	31780	33780	35220	37780
30980	33220	35100	37220	39380

(f) States desiring frequencies for use by portable-mobile or mobile stations different from those which may be allocated under section 10.42 (a) and (b), may be authorized to use the following frequencies. These frequencies are also available to states for portable stations of 1 watt power or less:

GROUP D

35780	37380	39180	39780

(g) The number of frequencies which may be assigned to any one municipality or state for either land, portable, or mobile stations will be governed pursuant to announced policies of the Commission.

10.43 SPECIAL ALLOCATION. The frequency 190 kilocycles is allocated for use by state police stations for radiotelegraph communication for emergency use in the event of failure of police wire communication systems.

10.44 ZONE AND INTERZONE. The following frequencies are allocated for zone and interzone police stations:

(a) For interzone communication subject to the condition that no interference is caused to international service (available to interzone police stations and to zone police stations designated as alternate interzone control stations):

5135 Kc. working	7480 Kc. day only
5140 Kc. working	7805 Kc. day only
5195 Kc. calling	7935 Kc. day only

(b) For zone communication (available to interzone and zone police stations):

2804 Kc. calling   2808 Kc. working   2812 Kc. working

(c) Calling frequencies herein allocated may be used for the transmission of operating signals and a single short radiotelegram provided no interference is caused to call signals.

10.45 MARINE FIRE STATIONS. The following frequencies are allocated for use by marine fire stations:

1630   35580   37740

10.46 SPECIAL EMERGENCY STATIONS. The following frequencies are allocated to special emergency stations:

(a) For portable stations with a maximum power of 1 watt, portable-mobile stations, and mobile stations:

31740	33820	37180	39340
33060	35140	37820	

(b) For fixed, land, and portable stations without limitation as to power:

31460   39660   39860

(c) For fixed, land, and portable stations with a maximum power of 1000 watts:

2726 Kc. (A3 Emission)   3190 Kc. (A1 Emission)

(d) For fixed, land, and portable stations of public utilities, using A3 emission, with a maximum power of 50 watts:

2292 Kc.   4637.5 Kc. day only

(e) Notwithstanding the provisions of (b), (c), and (d) of this section, authorizations may be issued covering the operation of mobile and portable-mobile stations on the frequency, or frequencies, assigned to licensees of fixed or land stations.

10.47 FORESTRY STATIONS. The following frequencies are allocated to forestry stations: (a)

30940	39740	35940	31580	31940
35740	31340	39940	37460	39420

(b) Maximum power 50 watts (subject to the condition that no interference is caused to Canadian stations).

2212   2236   2244

(c) Maximum power 500 watts (subject to the condition that no interference is caused to Canadian stations).

2226

10.48 ASSIGNED FREQUENCIES NONEXCLUSIVE. No frequency available to a station in the emergency service will be assigned exclusively to any applicant. All stations in this service are required to coordinate operation so as to avoid interference and make the most effective use of the frequencies assigned.

OPERATING SPECIFICATIONS

10.61 PERCENT OF TOLERANCE. The frequency tolerance of stations in the emergency service shall be as follows:

	EQUIPMENT AUTHORIZED	
	Before Oct. 1, 1938	After Oct. 1, 1938
	Percent	Percent
Fixed stations on frequencies below 30000 Kc.	0.03	0.01
Land stations on frequencies below 30000 Kc.	.04	.02
Portable and mobile stations on frequencies below 30000 Kc.	.04	.02
Fixed and land stations on frequencies above 30000 Kc.	.05	.02
Portable and mobile stations on frequencies above 30000 Kc.	.05	.03
Portable and mobile stations of 1 watt power or less on frequencies above 30000 Kc.	.1	.1

10.62 MODULATION LIMITS. The transmitters of stations in the emergency services using A3 emission shall be modulated not less than 85 percent nor more than 100 percent on peaks.

FREQUENCY MEASUREMENT

10.66 MEASUREMENT PROCEDURE. The licensee of each station shall provide for measurement of the frequency of the transmitter(s) and establish procedure for checking it regularly. These measurements of frequency shall be made by means independent of the frequency control of the transmitter and shall be of such an accuracy that the limit of error is within the frequency tolerance allowed the transmitter.

TESTS

10.71 EQUIPMENT AND SERVICE TESTS. Equipment and service tests as authorized in sections 2.42 and 2.43 may be conducted provided that the necessary precautions are taken to avoid interference. The equipment tests authorized by section 2.42 may be conducted only during daylight hours on frequencies below 6000 Kc.

10.72 ROUTINE TESTS. The licensees of all classes of stations in the emergency service are authorized to make such routine tests as may be required for the proper maintenance of the station and communication network, provided that precautions are taken to avoid interference with any station in the particular service involved.

LICENSES

10.81 PERIOD. The license period for all stations in the emergency service shall be for 1 year unless otherwise stated in the instrument of authorization. The date of expiration of license for all classes of stations operating in the emergency service, unless otherwise specified, shall be the 1st day of May of each year.

10.82 POSTING FIXED STATION LICENSES. The station licenses of stations in this service, operated at fixed locations, shall be conspicuously posted at the place where the control operator is located.

10.83 POSTING PORTABLE OR MOBILE STATION LICENSES. The licenses of portable and mobile stations, if separately issued, shall be readily available for inspection by authorized Government representatives. Either the original authorization or a photo copy of that document shall be available at the portable or mobile station involved.

10.84 OPERATOR LICENSE. The original license of each station operator shall be conspicuously posted at the place he is on duty, or, in the case of portable or mobile units, be kept in his personal possession.

LOGS

10.101 CONTENTS. Each licensee shall maintain adequate records of the operation of the station including: (a) hours of operation; (b) nature and time of each communication; (c) frequency measurements; (d) name of operator on duty at the transmitter. In the cases of groups of stations, either land or land and mobile, operating as a single coordinated communication system controlled from a single point, a single log may be maintained at a central location, provided that such log records the required information with respect to all stations in the network.

INSPECTIONS

10.111 INSPECTION BY COMMISSION'S REPRESENTATIVE. All classes of stations in the emergency service shall be made available for inspection upon request of a representative of the Commission. However, if such station is actually engaged in an emergency which should not be interrupted, the Commission's representative may suspend the inspection and require the station to be made available for inspection immediately after conclusion of the emergency.

MUNICIPAL POLICE STATIONS

10.121 POWER. The maximum power to be assigned for the use of frequencies below 30,000 kc. by municipal police stations will be based on the latest official population figures of the Department of Commerce for the area to be served in accordance with the following table:

POPULATION	POWER - WATTS
Under 100,000	50
100,000 - 200,000	100
200,000 - 300,000	150
300,000 - 400,000	200
400,000 - 500,000	250
500,000 - 600,000	300
600,000 - 700,000	400
Over 700,000	500

10.122 ADDITIONAL POWER. In the event that the amount of power allocated above is insufficient to afford reliable coverage over the desired service area, the Commission may authorize the use of additional stations of the same or less power, or upon proper showing being made, may authorize such additional power as may be necessary, but not to exceed 500 watts; Provided, however, that municipal police stations authorized to serve an entire county under the provisions of section 10.123, may be licensed to employ a maximum power of 1,000 watts between 1 hour after local sunrise and 1 hour before local sunset, on condition that the applicant files with the application an agreement, entered into with other licensees operating on the same frequency and in the same area to which the frequency is assigned, including a statement giving their consent to the use of such increased power; and that such agreement shall provide for notification to the Commission 60 days prior to termination thereof.

10.123 COOPERATIVE SERVICE. An application for an authorization for a municipal police station to serve two or more municipalities shall be supported by sworn copies of agreements made between the proposed licensee and the contiguous municipalities. Such agreements shall show that the applicant is required to furnish emergency police radio service to the contiguous municipalities and that the contiguous municipalities agree to accept such service and not to request individual authority to operate municipal police radio transmitting stations, and that such agreements shall provide for notification to the Commission 60 days prior to termination thereof.

10.124 COOPERATIVE USE OF FREQUENCIES. The frequencies allocated to municipal police stations are assigned for use within specified geographical boundaries and all licensees within those boundaries shall cooperate in the use of the assigned frequency.

10.125 SERVICE WHICH MAY BE RENDERED. Municipal police stations, although licensed primarily for communication with mobile police units, may transmit emergency messages to other mobile units such as fire department vehicles, private ambulances, and repair units of public utilities, in those cases which require cooperation or coordination with police activities. In addition, such stations may communicate among themselves provided: (1) that no interference is caused to the mobile service; and (2) that communication is limited to places between which, by reason of their close proximity, the use of police radiotelegraph stations is impracticable. Municipal police stations shall not engage in point-to-point radiocommunication beyond the good service range of the transmitting station. The transmission or handling of messages requiring radiotelephone relay or the relaying of such messages is prohibited; Provided, however, that after proper showing and in unusual circumstances the Commission may in specific instances authorize communication routes involving such relays. Point-to-point communication between stations in the same local telephone exchange area is likewise prohibited unless the messages to be transmitted are of immediate importance to mobile units.

STATE POLICE STATIONS

10.151 POWER. The maximum power to be assigned for the use of state police stations shall be 5,000 watts during the period from sunrise to sunset and 1,000 watts from sunset to sunrise.

10.152 SERVICE WHICH MAY BE RENDERED. State police stations, although licensed primarily for communication with mobile police units may transmit emergency messages to other mobile units such as fire department vehicles, private ambulances and repair units of public utilities, in those cases which require cooperation or coordination with police activities. In addition, such stations may communicate among themselves provided: (1) that no interference is caused to the mobile service; and (2) that communication is limited to places between which, by reason of their close proximity, the use of police radiotelegraph stations is impracticable. State police stations shall not engage in point-to-point radiocommunication beyond the good service range of the transmitting station. The transmission or handling of messages requiring radiotelephone relay or the relaying of such messages is prohibited; Provided, however, that after proper showing and in unusual circumstances the Commission may in specific instances authorize communication routes involving such relays. Point-to-point communication between stations in the same local telephone exchange area is likewise prohibited unless the messages to be transmitted are of immediate importance to mobile units.

INTERZONE POLICE STATIONS

10.171 ONE STATION PER ZONE. Authorizations for interzone police stations will not be issued for more than one station within a zone.

10.172 ELIGIBILITY FOR LICENSE. In general only the licensees of state and municipal police stations may be granted authorizations to operate interzone police stations.

10.173 EQUIPMENT. Authorizations for interzone police stations may be granted specifying equipment authorized for use by municipal or state police stations provided that the radiotelegraph use of such equipment is on a secondary basis, and that the equipment is so designed that the frequency can be changed without delay.

10.174 POWER. The maximum power to be assigned for the use of interzone police stations shall be 500 watts.

10.175 SERVICE WHICH MAY BE RENDERED. Interzone police stations shall be operated only for the transmission of dispatches of an emergency nature relating to police business between police agencies, using the operating procedure prescribed by the Commission.

10.176 OPERATOR REGULATIONS. The records and method of operation of interzone police stations shall be maintained and conducted in accordance with the operating procedure prescribed by the Commission.

ZONE POLICE STATIONS

10.191 ELIGIBILITY FOR LICENSES. In general only the licensees of state and municipal police stations may be granted authorizations to operate zone police stations.

10.192 EQUIPMENT. Authorizations for zone police stations may be granted specifying equipment authorized for use by municipal or state police stations provided that the radiotelegraph use of such equipment is on a secondary basis, and that the equipment is so designed that the frequency can be changed without delay.

10.193 POWER. The maximum power to be assigned for the use of zone police stations shall be 500 watts.

10.194 SERVICE WHICH MAY BE RENDERED. Zone police stations shall be operated only for the transmission of dispatches of an emergency nature relating to police business between police agencies, using the operating procedure prescribed by the Commission.

10.195 OPERATOR REGULATIONS. The records and method of operation of zone police stations shall be maintained and conducted in accordance with the operating procedure prescribed by the Commission.

10.196 ALTERNATE ZONE CONTROL STATIONS. Zone police stations may be designated to act as alternate zone control stations for the interzone stations designated for the zone concerned; in which event, such zone police stations shall be eligible to be assigned all of the frequencies available for interzone police stations.

MARINE FIRE STATIONS

10.211 POWER. The maximum power to be assigned for the use of marine fire stations will be 500 watts.

10.212 SERVICE WHICH MAY BE RENDERED. Marine fire stations are licensed primarily for intercommunication between fire headquarters and fireboats. However, they may transmit emergency messages to police boats or other marine units in cases which require cooperation or coordination with police or fire department activities.

SPECIAL EMERGENCY STATIONS

10.231 SCOPE OF SERVICE. (a) Special emergency stations may be used only during an emergency jeopardizing life, public safety, or important property: (1) for essential communications arising from the emergency; and (2) for emergency transmission from one point to another between which normal communication facilities do not exist, are not usable, or are temporarily disrupted or inadequate.

(b) The use of special emergency stations for the handling of routine or nonemergency communications is strictly prohibited.

(c) Within the scope of service given in paragraph (a) the licensee of a special emergency station shall make the communication facilities of such station available to any member of the public.

(d) Special emergency stations, except those of communications common carriers utilized temporarily to restore normal public communication service disrupted by an emergency, shall not operate as common carriers of communications for hire. However, licensees of such stations may accept contributions to capital and operating expenses from others who, under the Commission's rules, would be eligible to stations of their own, for the cooperative use of the stations on a cost-sharing basis; Provided, that contracts for such cooperative use are submitted to the Commission 30 days prior to the effective date thereof and that said contracts are not disapproved by the Commission.

10.232 SELECTIVE CALLING SYSTEM. Notwithstanding the provisions of section 10.46 (c) and (d), types A1 or A2 emission may be used on 2726, 2292, and 4637.5 kc. for the sole purpose of establishing a selective calling system.

10.233 TESTS. Special emergency stations may also conduct routine tests not exceeding 2 minutes in each half-hour, or, where more extended tests are required, they may not exceed a total of 4 hours per week.

10.234 AVOIDANCE OF INTERFERENCE. Special emergency stations shall take all reasonable precautions, including listening tests, to avoid any possible interference to the service of another station.

FORESTRY STATIONS

10.251 SCOPE OF SERVICE. Forestry stations, although licensed primarily for communication with mobile forest fire fighting units, may transmit emergency messages to other mobile units such as fire department vehicles, private ambulances, and mobile police units in those cases which require cooperation or coordination with forestry service activities. In addition, such stations may communicate among themselves, provided: (1) no interference is caused to mobile service; and (2) only those communications are transmitted which are necessary for the operation of forestry service.

Notes

(These extra pages are provided for your use in taking special notes)

The text of this lesson was compiled and edited by the following members of the staff:

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POWER & LIGHT BUILDING, KANSAS CITY, MISSOURI

UNIT NO. 3

MAINTENANCE OF BATTERIES

LESSON NO. 14

BATTERIES

.....power waiting for release.

When we look at the black case that houses the ordinary storage battery, we simply see a case instead of the amazing little power house that is within. Few of us give any thought to the wonderful chemical action that stores energy for future use. But when we take a heavy piece of metal and short circuit the two terminals and are greeted with a snapping, blinding flash and perhaps burn our fingers, we realize that the little black case is not as simple as it appears.

Batteries that are allowed to remain idle for long periods of time lose their energy and their usefulness. The same is true of human beings. For in many ways the battery resembles the human body. Both store up energy. Both can serve many useful purposes. And both deteriorate when they are allowed to become idle.

As a Midland student you are storing up knowledge. This, the battery cannot do. Each lesson that you complete brings you closer to the time when your human battery will be fully charged. When that time arrives, you will release your knowledge to employers who will pay you money for it. You will then receive your reward for the time that you are devoting to your training now.

Knowledge and the energy necessary to put it to good use have been the secret of success for ages. Today, more than ever before, success depends upon these two factors. Keep this thought before you constantly. Store up knowledge.....store up energy and you will have a driving force that is far more powerful than the greatest batteries ever built.

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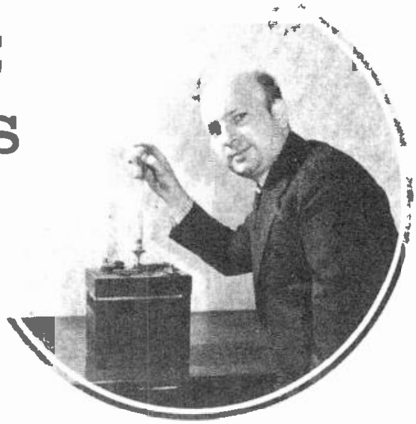
Lesson Fourteen

MAINTENANCE OF BATTERIES



"The use of storage batteries in connection with radio transmitting equipment is rapidly being abandoned. However, it is still highly essential that every radio operator have a thorough knowledge of their care and maintenance because of the large number of installations still using this type of equipment.

"You should become thoroughly conversant with the material in this lesson if it is necessary to memorize the more important parts".

1. REVIEW. As explained in Lesson 15 of Unit 1, secondary cells such as the "lead-acid" and the "nickel-iron-alkaline" types may be recharged when the chemical compounds producing the E.M.F. have been depleted. The process consists of connecting a source of DC voltage to the battery, and allowing the current which this voltage produces to reconvert the chemicals into their original form.

When the lead-acid battery is charged, the active material of the positive plate is lead peroxide (PbO_2) and that of the negative plate is pure lead (Pb). Upon discharge, both of these materials are changed to lead sulphate ($PbSO_4$). The chemical equation which gives the conversion process is:

$$Pb + PbO_2 + 2H_2SO_4 \underset{\text{Charge}}{\overset{\text{Discharge}}{\rightleftarrows}} 2PbSO_4 + 2H_2O$$

The electrolyte is, of course, a dilute solution of sulphuric acid, and, as is stated in the foregoing equation, a part of the sulphuric acid is used to change the active materials into lead sulphate. Sulphuric acid is considerably heavier than water, and so the specific gravity of the electrolyte is somewhat greater than one. When the cell is fully charged, the sulphuric acid content of the electrolyte is fairly great and the specific gravity of the electrolyte as measured with a hydrometer is high (approximately

1.275). As the discharge process of the battery progresses, much of the sulphuric acid is removed from the solution, and, as a result, the specific gravity of the electrolyte becomes less, reaching 1.150 when the battery is fully discharged.

It should be realized that the lead-acid cell is not completely devoid of electrical energy at the time that its specific gravity has fallen to 1.150, but that serious injury will result to the battery if a further discharge is attempted.

The output voltage of a fully charged lead-acid cell is 2.1 volts, and the cell should be recharged when its output voltage has dropped to 1.75 volts. Unless the output voltage of the cell is measured with a low-resistance voltmeter, a true indication of the condition of the cell will not be obtained. When a high-resistance voltmeter, such as is commonly employed in Radio-Television work, is used to measure the voltage of a cell, no load is placed on the cell, and very little current is drawn from it. A voltmeter of this type would probably show a reading of 2.1 volts, even when the cell was nearly discharged. Thus, a low-resistance voltmeter should be provided for storage battery installations, because it places a normal load on the cell, and measures its output voltage under actual working conditions.

The charging source must be DC; it cannot be AC, because one alternation of the voltage would tend to charge the cell and the succeeding alternation would produce a discharge. It is not necessary, however, that the charging voltage be pure DC; in most cases, in fact, it is a pulsating DC voltage. Therefore, the first requisite for maintaining a storage battery installation is to have a source of DC voltage available. If the supply lines are AC, some form of rectification will be needed. A later section of this lesson will deal with the various methods used to produce rectification.

The charging voltage must be so connected to the cell that the current it produces will flow through the cell in the opposite direction of that to which the cell forces current when under discharge. This demands that the positive terminal of the voltage source be connected to the positive terminal of the cell, and the negative terminal of the voltage to the negative terminal of the cell. At times, there may be some doubt as to the polarity of the voltage source. There are several simple methods for determining which terminal is which. Undoubtedly the simplest is to use a DC voltmeter. Should it happen that a voltmeter is not available, another easy method is to place the two ends of the line in a glass of water to which a small amount of acid has been added. When the charging voltage is applied, small bubbles will arise from each of the submerged terminals. There will, however, be nearly twice as many bubbles surrounding the negative terminal as appear at the positive terminal. To avoid any possibility of short-circuiting the line, and to prevent the flow of an excessive current, it is best to connect some resistance in series; a 100-watt lamp will suffice. Another method of determining polarity is to use a small pocket compass, applying the principle of the left-hand rule given in Lesson 9 of Unit 1.

2. CHARGING METHODS. There are, in general, two methods used for charging storage batteries. These are the constant-current method and the constant-potential method. In addition to the source of DC voltage, there must be available an ammeter for measuring the charging current, and some means of regulating the value of this current. In the constant-current method of charging, the same value of charging current is used throughout the charging procedure. This method is often known as "series-charging", because all of the cells to be charged are connected in series. The DC charging voltage to be used will depend upon the number of cells so connected. It will vary from 2.3 to 2.65 volts per cell. Naturally, it must be greater than the output voltage of the cells, or they will discharge through the voltage source. The current regulator may be a heavy duty rheostat, or may be composed of a bank of lamps connected in parallel. Fig. 1 shows an arrangement for constant-current charging. When a battery is taken out of the circuit, the switch corresponding to

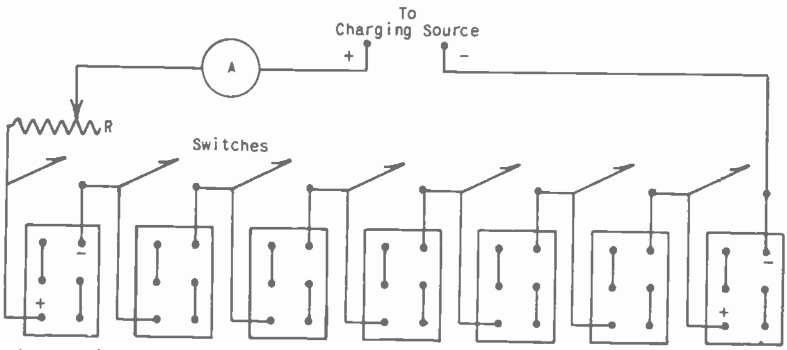


Fig. 1 (Method of connecting batteries when they are to be charged by the constant-current system.) When a battery is taken out of the circuit, the switch corresponding to that position is closed so that the circuit will be completed.

that position is closed, thereby completing the circuit. The current is regulated by the lamp bank. Each time a lamp is connected in parallel, the total resistance of the bank is reduced, and more charging current flows to the series-connected cells.

The charging current depends on the voltage of the charging source, the resistance of the lamp bank, and the voltage of the cells. Since the voltage of the cells opposes the charging voltage, the net voltage in the circuit available to force a current to flow is equal to the difference between these two voltages. Thus, the charging current will be:

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

Where: E is the voltage of the charging source
E_c is the voltage of the series-connected cells
R is the resistance of the current-regulating device.

As the charging process continues, the voltage across the terminals of the cells will increase. This will cause the voltage opposing the charging voltage to be larger, and the charging current will fall. To compensate for this action, the resistance of the current regulator is reduced from time to time so that practically a constant current is delivered to the cells throughout the charging process. The amount of charging current which may be employed depends on the ampere-hour capacity of the battery, and upon its state of discharge. In fact, a battery may be charged at any rate which does not produce appreciable gassing or boiling of the electrolyte. Thus, a battery which is fully discharged may be charged initially at a high rate. After a part of the charge has been put back into the battery, however, the high rate of charging current will be excessive, and will serve merely to overheat the battery, producing gassing of the electrolyte. Since it is not safe to allow the temperature of the electrolyte to rise above 110° F., the amount of charging current used in the constant - current system of charging may not be very great. It should never exceed the finishing rate. When the battery is fully charged, even a small charging rate will produce some gassing of the electrolyte; however, if the charging current is reduced to a low value, the small amount of gassing which results will produce no harmful effects. This safe rate which will not injure the battery, even when it is fully charged, is called the "finishing rate".

The major disadvantage of the constant-current method of charging is the fact that the charging current must be necessarily low. For this reason, a considerable time is required for charging a battery. The charge should be continued until the specific gravity readings of the electrolyte remain constant for three successive readings taken at fifteen minute intervals. The final voltage will be about 2.5 volts per cell, but will depend on the temperature and condition of the battery.

A system which eliminates the disadvantage of the constant-current charging method is known as the modified constant - current method of charging. In this system the charge is started at a high rate, and is reduced as soon as gassing occurs. The charging rate will probably be reduced several times throughout the charging procedure, it being necessary to insert additional resistance each time gassing takes place.

The so - called normal charging rate is that rate which would charge the battery in a period of 8 hours, assuming that gassing does not occur. Naturally, this normal charging rate cannot be maintained throughout the entire charge without injuring the battery. The finishing charge would completely charge the battery in from 12 to 15 hours, depending on the type of cell. Thus, in the modified constant - current charging method, the initial rate might be the normal rate (or higher), and be reduced in successive steps until the finishing rate was reached.

The main source of power aboard ship is ordinarily a 110-volt direct current generator, and the output of this generator not only supplies lighting and runs small motors, but also furnishes power for the ships transmitter. To provide power in case of failure of the main source of supply, every vessel carries a storage battery as an emergency power source for operating the transmitter. This storage battery consists of 60 cells, and thus has an output of about 120 volts. When the battery is not in use, it is connected to the 110-volt line through charging resistances which provide just enough charging current to maintain the battery in a charged condition. Naturally, it is not possible to charge all of the cells in series, because the total output voltage would be more than the charging voltage. Therefore, the battery is divided into two banks of 30 cells each and these two banks are connected in parallel. Fig. 2 illustrates the switching method which allows the two banks to be charged in parallel, but to be discharged in series.

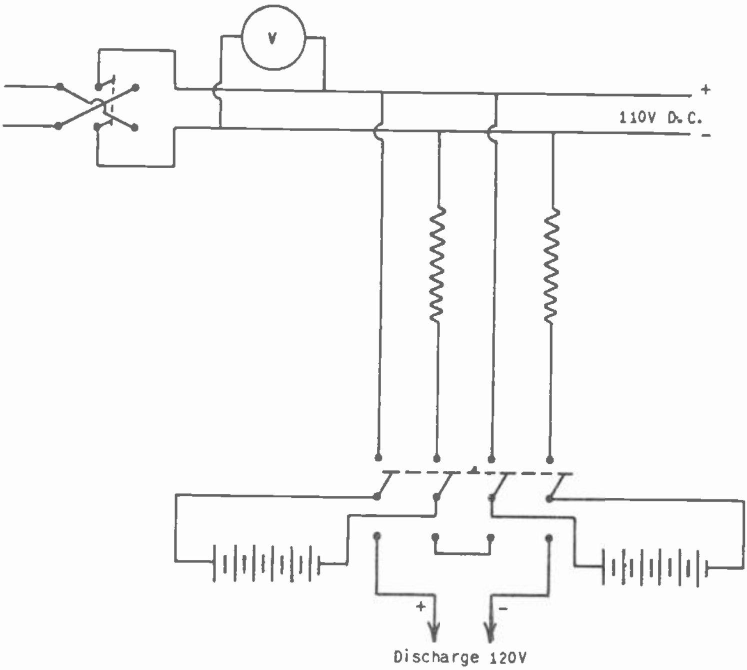


Fig. 2 Illustrating how two banks of batteries may be charged in parallel and discharged in series.

In addition to the constant-current method of charging, there is also the constant potential method. In this method, the voltage of the charging source is kept as near constant as possible. A shunt-wound generator equipped with a voltage regulator is often used as the charging source. The charging voltage should be between 2.32 and 2.4 volts multiplied by the number of cells to be charged. The charge is started at a high rate, and as the voltage of the cells increase, the charging current is reduced, thereby providing a charge which automatically tapers off to a low finishing rate as the charge progresses. Under laboratory conditions, it has been found possible to charge a battery fully by this method in a little over three hours. A charging resistance whose value will depend on the charging voltage and the type of battery to be charged is connected in series with the source. When the system has once been set into operation, and the correct charging resistances have been determined, it requires but little attention, and can charge a battery in about 3 hours. Many automobile batteries are charged by this method. In this case, the charging source has

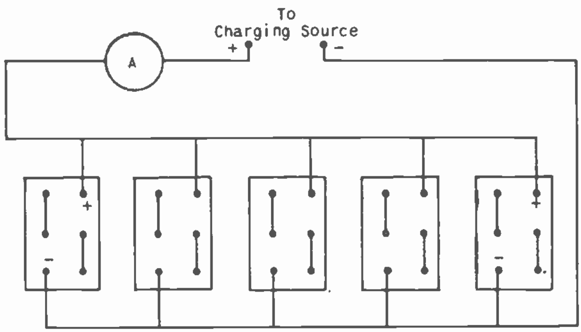


Fig.3 Arrangement of batteries for constant-voltage system of charging.

a voltage of approximately 7.5 volts, and has the advantage that each battery takes the current it requires and a large battery may be charged in the same time as a small one. The batteries are connected in parallel as illustrated in Fig. 3.

3. TRICKLE CHARGING. A trickle charge is a charge at a very low rate. As explained in Lesson 15 of Unit 1, every type of battery is subject to the effects of local action. Local action is due to the presence of impurities in the plates, which produce local currents within the battery tending to cause its discharge. As a result, the plates of a lead-acid battery gradually change to lead sulphate even when the battery is not in use. For this reason, it is necessary to keep a storage battery on charge so that it will be ready for use when it is needed. Naturally, the charging rate need not be very great, since its only purpose is to overcome the effect of local action. It has been determined that the loss of a cell per day due to local action is from .5 to 2% of its rated ampere-hour capacity, depending on the temperature and condition of the cell. Therefore, the charging rate for the trickle charge need not be more than a few per cent of the normal 3 hours charging rate. When a storage battery is to be put on trickle charge, several days may be necessary to determine the correct charging rate. It is, of course, understood that the battery is fully charged when it is put on trickle charge. Furthermore, the specific gravity of the electrolyte should be recorded. At regular intervals thereafter, the distilled water should be replenished, and a specific gravity reading taken. If the specific gravity of the electrolyte

continues to rise or the presence of appreciable gassing is manifested, the trickle charging rate is too high. On the other hand, if the specific gravity falls gradually, the charging rate is insufficient. It should be adjusted so that it is just sufficient to maintain the battery in a fully charged condition. Great care is necessary in adjusting the charging rate of a battery having a pasted plate construction. Continual over-charging will cause gassing, producing small whirlpools in the electrolyte which tend to cause a shedding of the active material from the plates.

4. BOOSTING AND EQUALIZING. Some batteries are maintained merely as an emergency source of power. In this case, they are kept on trickle charge when not in use, so that they will be ready for service when an emergency arises. Other installations demand that a battery give more or less constant service. In some cases, two separate storage batteries are provided and are used alternately, one being charged while the other is in use. Such batteries are said to be cycle-charged. Cycled batteries are normally allowed to reach a certain state of discharge before being recharged. Usually, if time will permit, they are charged at the normal 8 hour rate, and are allowed to become fully charged before being placed on discharge. In some instances, however, it may be necessary to use a battery practically continuously, charging it at odd times when convenient. In this case, a boosting charge is given. The boosting charge is given at a high rate over a comparatively short period of time. Its purpose is not ordinarily to charge the battery fully, but merely to replace some of the charge which has been taken out so that the battery will give further service. The time of the boosting charge will depend on the time available for charging the battery; it may be as long as several hours, or it may be for only a half hour. It is given at the highest rate that will not cause excessive gassing or raise the temperature of the battery above 110° F. If a battery has been completely discharged, it should be charged at once and not allowed to stand in a discharged condition. Part of a charge is better than none, although to charge the battery fully when it is completely discharged will require somewhat longer than normal.

A boosting charge does not particularly harm a battery, but a battery charged in this manner should have an equalizing charge occasionally. The purpose of the equalizing charge is to bring to the normal condition any of the cells which tend to be low. It is merely a continuation of the normal charge at about half of the finishing rate until all of the cells in the battery gas freely and uniformly, and until three specific gravity readings taken at half-hour intervals show no increase. At least once a month the specific gravity of all cells should be taken, and should then be adjusted to the correct value as specified by the manufacturer. If the specific gravity is too high, some of the electrolyte should be drawn off and replaced with distilled water. If too low, due to leakage or spilling, it should be brought back to normal by drawing off some of the electrolyte and replacing with a new electrolyte of 1.400 specific gravity. This, however, should only be done at the end of an equalizing charge. All batteries should receive equalizing charges at least once a month.

5. SULPHATION AND ITS REMEDIES. Sulphation is a very important phenomenon which is liable to occur in the operation of a lead-acid storage battery installation. In order that the student will know its causes, effects, and remedies, a special section of this lesson will be devoted to this subject.

When a battery is discharged, the active materials of its plates are converted into lead sulphate, this being a natural process which is essential if the battery is to supply current. If the battery is recharged immediately, the lead sulphate is readily converted into the original active materials, and the battery is ready for another discharge. It is neither necessary nor desirable that each charge be carried to completion. It is sufficient that each charge bring the specific gravity within a few points of maximum, provided that an equalizing charge is given at regular intervals. If, however, the battery is allowed to remain in a discharged condition or is continually under-charged, the lead sulphate enters the pores of the active materials, and becomes very dense and hard. A battery in this condition is said to be sulphated.

When a battery is sulphated, the sulphate is very dense, and absorbs a charge only with great difficulty. The cause of this condition is nearly always some form of abuse such as:

1. Standing discharged.
2. Habitual under-charging.
3. Neglecting evidence of trouble in individual cells.
4. Adding electrolyte or acid to raise specific gravity, instead of bringing it out of the plates by proper charging.

It is the negative plates which suffer most from sulphation. When sulphated, the active material of the negative plates is light in color, is granular and is easily disintegrated.

Sulphated positive plates are not damaged to such an extent by sulphation and, unless badly buckled, may be restored to operating condition, although the sulphation will shorten their useful life.

When a battery is suspected of being sulphated, it should be charged fully and then discharged at its normal rate. If it fails to give its rated capacity, it is probably sulphated and should be given the following treatment: First recharge the battery in the regular manner, and when it is considered to be fully charged, read and record the specific gravity of each cell, and the temperature of several cells. Next place the battery on charge, maintaining the charging rate as near to one-half the finishing rate as possible. Should the temperature reach 110° F., reduce the rate or interrupt the charge temporarily to allow the battery to cool.

Continue the charge, taking hydrometer readings at intervals of from 3 to 5 hours, to determine whether the specific gravity is rising. When there has been no rise in any cell during a period of 10 hours, the charge may be terminated. The level of the electrolyte should be maintained constant, by adding distilled water after each hydrometer reading. If the specific gravity of any cell rises above normal, draw off some of the electrolyte and replace with distilled water. A continued rise above normal indicates that some acid has been added during some previous operation of the battery, and is probably the cause of the sulphation.

Ordinarily, the foregoing process will remove the hardened sulphate, but in extreme cases it will not. As a last resort, it is possible to use the so-called water treatment. This process consists of reducing the specific gravity of the electrolyte by drawing some off and adding distilled water until it is between approximately 1.050 and 1.100. Next, the battery is charged at about one-half of the finishing rate to both a 20-hour gravity and a 20-hour voltage maximum. If the gravity rises above 1.150 during this process, some of the electrolyte should be removed and replaced with distilled water, until the gravity is between 1.050 and 1.100. This process is repeated until the maximum specific gravity obtained by charging is below 1.150. The specific gravity is now increased by adding electrolyte and charging at the finishing rate or normal rate until the specific gravity is slightly below the operating value. After a cell has undergone the water treatment, the specific gravity can never be brought back to normal.

6. EFFECT OF TEMPERATURE ON OPERATION OF BATTERIES. The temperature of a battery cell should never exceed 110° F. High temperatures have the effect of shortening the life of the wooden separators. The sulphuric acid has the tendency of carbonizing the wood of the separators, and this condition is aggravated by temperatures above 110° F. Continued operation at temperatures above this limit will very likely necessitate the replacement of the separators before the battery itself has become so deteriorated that it is no longer useful.

Some impurities are always present in the materials used in the construction of any battery, and these foreign particles cause the local action which discharges the battery even though it is idle. Furthermore, this action is greatly accelerated by high temperatures. Low temperatures, on the other hand, cause a temporary reduction in the capacity of the battery. This effect, however, is only temporary, and the battery returns to its normal condition when the temperature is increased. The only other effect that low temperature may have on a battery is the danger of the electrolyte freezing in case water is added and is not allowed to become thoroughly mixed with the acid. A fully charged cell is in no danger of freezing.

7. GAS AND VENTILATION. While a battery is being charged, some of the water contained in the electrolyte is broken down into its elements hydrogen and oxygen. These gases escape into the surrounding air, and unless they are thoroughly mixed with large quantities of air, they form a highly explosive mixture. If they are ignited by an electric spark or by an open flame, they will explode

violently. Therefore, the battery compartment should be well ventilated, and free circulation of air must be provided so that the generated gases may become mixed with the air and be no longer explosive. Furthermore, no open flame, cigar, match, or any other form of fire should be brought near the battery while it is being charged or shortly thereafter. If the battery is kept in a closed room, rigid rules regarding smoking, etc., must be enforced.

8. EFFECT OF WRONG CHARGING POLARITY. It is most important that the correct charging polarity be observed. The positive side of the charging source must be connected to the positive terminal of the battery, and the negative side of the source to the negative terminal of the battery. The effect of the charge is to reverse the chemical action which took place during discharge. The lead sulphate on the positive plate must be reconverted to lead peroxide, and that of the negative plate to pure lead. Should it happen that the wrong charging polarity was employed, the same chemical action which took place during discharge would be continued, and the battery will be further discharged. A prolongation of this charge for only a short time will cause the plates to buckle, and will result in ruination of the battery.

When a battery has been connected to the charging source with wrong polarity, the first step in the attempt to restore it to a normal condition is to discharge the battery completely. This is done by connecting a heavy wire across the terminals and allowing it to remain as long as current flows. Next, the battery is charged at a reduced rate in the proper direction until the specific gravity reaches a maximum and is constant over a five hour period. This will probably require 40 hours or more, depending on the type of battery. In certain cases, the outside negative plate may need straightening; in fact, charging the battery in the reversed condition for any length of time will cause the negatives to become useless.

9. MAINTENANCE OF A LEAD-ACID BATTERY INSTALLATION. It should be realized that each battery installation presents a particular problem in maintenance. The specific instructions supplied by the battery manufacturer should always be studied and followed explicitly. There are, however, certain general points which should be observed in any battery installation. Listed in tabular form, the following are the points to be remembered:

1. Keep open flames away from battery at all times during operation.
2. Replace spilled electrolyte before charging.
3. Keep level of electrolyte one-half inch above tops of separators by adding pure distilled water.
4. Never allow battery to remain in a discharged condition.
5. Avoid continual under or over-charging.
6. Keep on trickle charge, if possible, when not in use.
7. Keep tops of cells clean. Remove corrosion from terminals, and coat lightly with vaseline.

8. A separate battery room with walls, floor, and ceiling coated with an acid-resistant paint is desirable.
9. Always provide plenty of ventilation when batteries are charging.
10. If it becomes necessary to replace electrolyte, use only glass or earthenware dishes and utensils in mixing electrolyte.
11. Always pour the acid slowly into the water and never pour the water into the acid.
12. Never use rejuvenating mixtures.
13. Do not allow the temperature of the cells to rise above 110° F.
14. Take frequent voltage and hydrometer readings.
15. Keep vent caps in place and do not allow foreign materials to get into cells.
16. Give an equalizing charge at least once a month.
17. Establish a routine of battery care.

All of the points in this table are merely the application of common - sense rules; two of them, however, may require a little explanation. Number 11 states that water should never be poured into sulphuric acid. When sulphuric acid and water are first mixed, a fairly high temperature results from the chemical action. This high temperature causes the mixture to boil and fume, and may cause a small explosion. When the water is poured into the acid, the mixture becomes violently hot and serious damage is liable to result, On the other hand, if the acid is poured slowly into the water, the large quantity of water provides sufficient cooling to prevent any damage.

Point Number 12 states that rejuvenating mixtures should never be used. There are on the market certain powders and jellies which are claimed to renew the life of a battery. All such treatments are useless and result only in the permanent ruination of the battery.

As stated in the table, a special battery room is very desirable for permanent battery installations. This room should have good ventilation and should be well lighted, although direct sunlight should not be allowed to fall on the cells. The windows may be painted with a light paint and artificial illumination employed. Naturally, all forms of fire must be prohibited.

Air conditioning is desirable so that the temperature of the room will be maintained nearly constant at 70° F., and an exhaust fan must be provided to carry off the gases generated by the cells. Very little, if any, exposed ironwork should be in the room, since the fumes will cause rapid corrosion of the iron. If radiators or such are necessary, they should be placed high above the floor (although not above any of the cells), and should be coated with an acid-resistant paint.

The floor of the room should not be made of wood, but should be of tile with the joints filled with asphalt. Lead drain pipes should be used, and all acid that is spilled should be neutralized by applying a small amount of baking soda. All conductors should be either lead covered or of bare copper cable. If bare copper is used, it must be kept coated with vaseline. If a battery switchboard is necessary, it must be located in another room.

10. THE AMPERE-HOUR METER. The ampere-hour meter is a device which is used extensively in storage battery installations to measure the total number of ampere-hours which have been taken from a battery during discharge or put into a battery during charge. It consists of a mercury motor having a flat copper disc which revolves in a pool of mercury. Also, the disc passes through the intense magnetic field produced by two permanent magnets. The disc is highly damped and its speed of rotation is directly proportional to the amount of current passing through the mechanism. Attached to the disc is a train of gears which actuate a pointer which moves over a dial calibrated in ampere hours. The meter is permanently connected in the battery circuit, and indicates at any time the number of ampere hours which have been taken out of a battery. Furthermore, when the battery is being charged, the indicating needle travels in the opposite direction and shows when the battery is fully charged. When the movable hand reaches the full-charge position, a contact is made which actuates a relay, thereby disconnecting the charging voltage. This action protects the battery from overcharge. Also, there is usually associated with the meter a circuit breaker which disconnects the battery in case the charging voltage fails or its output voltage falls below that of the battery. This prevents the battery from discharging through the charging source. The ampere-hour meter may be used either with the lead-acid type of battery or the Edison battery.

11. CHARGING THE EDISON BATTERY. The methods of charging the Edison nickel-iron-alkaline cell are practically the same as those for the lead-acid cell. Due to the more rigid construction of this type of battery, and to the fact that it will withstand much more abuse, the charging rates are not as critical. The constant-current method of charging is often used and is the same as for the lead-acid type, except that it is not necessary to reduce the normal charging rate when gassing begins. The number of ampere-hours of charge should exceed the ampere-hours of discharge by at least 20 or 30%, but the important thing is to continue the charge until the maximum cell voltage has been reached and maintained for at least 15 minutes. The maximum voltage at the normal charging rate should be between 1.8 and 1.9 volts, but will depend on the temperature. When the cell is placed on discharge, its output voltage under load will approximate 1.37 volts.

Since the electrolyte of an Edison cell remains practically constant at a specific gravity of 1.200 at 60° F. whether the cell is charged or not, a hydrometer reading gives no indication of the state of charge of the cell. Frequent hydrometer readings should be taken, however, and when the specific gravity falls below 1.160, the electrolyte should be replaced.

The Edison cell should be recharged when its voltage falls to .9 volt, not because any injury will result to the cell by further discharge, but because a voltage below this value is ordinarily not usable in the normal applications to which Edison batteries are put. In fact, the cell will not be damaged if it is discharged to zero voltage and allowed to remain in a discharged condition indefinitely.

Low temperatures reduce the capacity of the Edison cell much in the same manner that they affect the lead-acid cell, and the effect is only temporary, the cell returning to normal when the temperature is raised. High temperatures, on the other hand, cause a deterioration of the negative plates, and the cell temperature of an Edison battery should never exceed 115° F.

A low-resistance voltmeter is often used to determine the state of charge of an Edison cell, although in large installations, the ampere-hour meter finds wide application.

When Edison batteries are charged at low rates, they tend to become sluggish and will show loss of capacity if sufficient excess of charge over discharge is not used. For this reason, the Edison Storage Battery Co. recommends the following formula for determining the proper current for the trickle charge:

$$\text{Current} = \frac{(\text{Amp-hr. capacity}) \times .16 + (\text{Amp-hr. used per day} \times 1.1)}{24}$$

The Edison battery may be charged by the constant-voltage method, in which case the charging voltage must be constant and equal to 1.7 times the number of cells connected in series. The initial charging current is about 200% of the normal charging rate, but averages about normal during the complete charge.

Boosting charges are often used to replace part of a charge in an Edison battery. The charging rate is high, but care must be taken to see that the cell temperature does not exceed 115° F. The rates recommended for boosting charges under normal conditions are:

5 minutes at 5 times normal rate.
15 minutes at 4 times normal rate.
30 minutes at 3 times normal rate.
60 minutes at 2 times normal rate.

The Edison cell is quite rugged and if not abused will give excellent service. The level of the electrolyte must be kept above the plates by refilling with pure distilled water. The life of the Edison cell is about 2,000 cycles where each cycle consists of a discharge to 1 volt followed by a complete charge. The actual life, of course, depends on many things, such as; the completeness of the cycles of charge and discharge, the number of cycles per year, and the type of service. The cell is guaranteed for a period of 6½ years for locomotive service to 16 years for emergency lighting service. Thus, the Edison battery has a far longer life than any lead-acid battery.

The Edison cells find their widest application for electrically propelled street and industrial trucks, locomotives, railway car lighting, and other places where a light weight battery requiring a minimum amount of care is desirable. Edison batteries are not used for automobile starting, because of their high internal resistance.

Maintenance of an Edison battery installation requires that the cells and trays be kept clean, and that no foreign substances, especially acid, is allowed to fall into the cell. As long as the level of the electrolyte is kept above the tops of the plates, the electrolyte is replaced when its specific gravity has fallen so low that it is no longer usable, and the cell is not charged at so great a rate to cause the temperature to rise above the specified limit, the battery will give very good service.

No damage will result to an Edison battery if it is inadvertently charged with the wrong polarity. The battery must be discharged and then given a long charge with the charging voltage correctly connected. Care must be taken to see that the battery room is well ventilated, because Edison cells, like the lead-acid type, give off large quantities of hydrogen gas while they are being charged, and this gas is highly explosive.

12. COMPARISON OF THE TWO TYPES OF BATTERIES.

A. The advantages of the lead-acid battery.
1. It has a low internal resistance, and will deliver any current up to several hundred amperes without injury to the battery.
2. It is somewhat more reasonable in cost than the Edison battery.
3. It has a greater efficiency than the Edison battery.
4. Its output voltage is more stable under load.
5. It has a higher output voltage per cell than the Edison type.

B. The disadvantages of the lead-acid battery.
1. It is very bulky and heavy.
2. The acid electrolyte causes corrosion of the terminals.
3. There is danger of ruining the battery due to under or over charging or allowing the battery to remain in an uncharged condition.
4. The active materials are liable to shed from the plates.
5. The ampere-hour capacity depends upon the rate of discharge.

C. The advantages of the Edison battery.
1. It is considerably lighter in weight than the lead-acid type.
2. It is mechanically stronger and able to withstand more abuse.
3. The active materials cannot be shed from the plates because they are enclosed in steel tubes or pockets.

4. Since the electrolyte is not acidic, it is not necessary to place the Edison battery in a specially constructed room to avoid corrosion, etc.
5. The ampere-hour capacity is approximately the same for all rates of discharge.
6. The life of the Edison battery is considerably longer than that of the lead-acid type.

D. The disadvantages of the Edison battery.
   1. The initial cost of the Edison battery is from three to five times that of the lead-acid type.
   2. It has a high internal resistance.
   3. The efficiency of the Edison battery is less than that of the lead-acid type.
   4. The output voltage per cell is lower.
   5. The output voltage under load is more unstable.

13. CHARGING EQUIPMENT. Equipment for charging storage batteries is of two types depending upon whether the storage battery installation is of large size or not. Large installations ordinarily employ a direct current generator to supply the needed DC charging voltage. Although the cost is greater than a rectifier type of charger, the maintenance is less, and the efficiency greater. Aboard ship the charging voltage is derived from a direct current generator.

To arrange conveniently the necessary resistances, switches, and various circuits associated with a storage battery installation, a control switchboard is nearly always employed. One of these boards is shown diagrammatically in Fig. 4. It is an Exide emergency switchboard and is manufactured by the Electric Storage Battery Company. It is capable of performing the following functions:

1. Connects batteries in series for discharge.
2. Connects batteries into two parallel groups for charge.
3. Provides voltmeter for following readings:
   (a) ship's line voltage and polarity,
   (b) batteries on discharge,
   (c) group A on charge,
   (d) group B on charge.
4. Opens or closes ship's line, or reverses polarity connection thereto.
5. Provides switches for controlling certain emergency lights.
6. Provides ampere-hour meter to indicate state of battery charge.
7. Provides circuit breaker fitted with overload release, low-voltage release, and automatic trip operated by the ampere-hour meter.

When the plug switch of the voltmeter is in the lower left hand receptacle, it is connected across the points marked 2-2 and

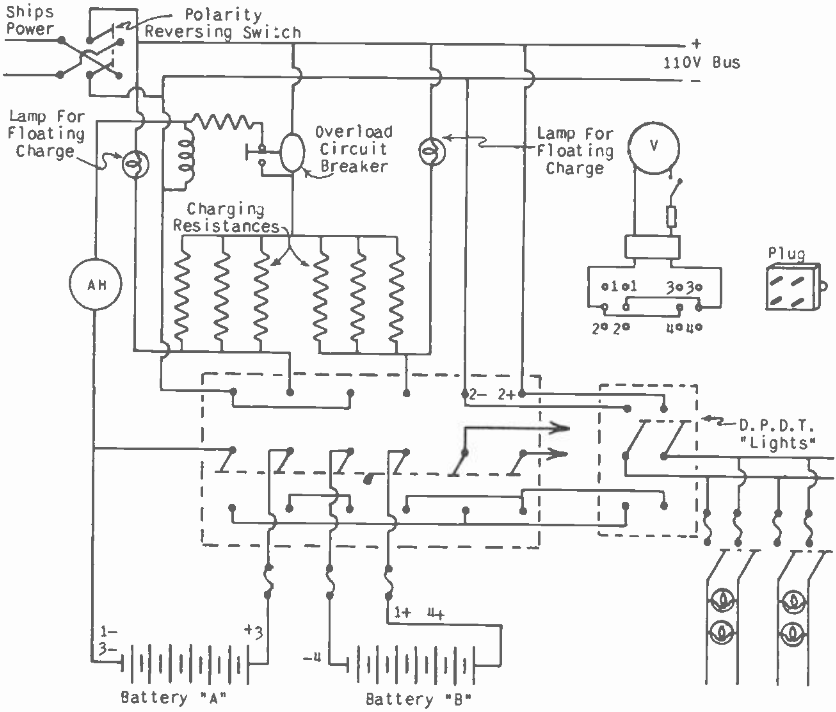


Fig.4 A schematic diagram of the Exide battery switchboard.

it reads the ship's line voltage. In the upper left-hand receptacle, the meter reads the voltage of the batteries on discharge and is connected to the points marked 1-1. In positions 3 - 3 and 4 - 4 it reads the voltage of batteries A and B respectively. A push button is included in the voltmeter circuit so that inductive surges of the transmitter will not damage the meter.

When the large six pole double throw switch is in the "up" position, the batteries are connected in parallel and are charged through the charging resistances from the ship's line. When this switch is thrown downward, the batteries are connected in series and may be used for emergency power. If this switch is in the "up" position, and the circuit breaker is closed, the batteries are charged at the normal rate. With the circuit breaker open, however, the two lamps provide sufficient resistance to cause the battery to be floated on the line, the charging rate being the same as that of a trickle charge.

The ampere-hour meter associated with the switchboard indicates at all times the state of charge of the battery. It is fitted with a contact which opens the circuit breaker and reduces the charging rate from the normal value to the trickle charge rate when the battery is fully charged.

The Edison battery switchboard is illustrated in Fig. 5, It performs practically the same functions as the Exide switchboard previously described, with the exception that an ammeter is included which may be connected across the various shunts shown in the drawing.

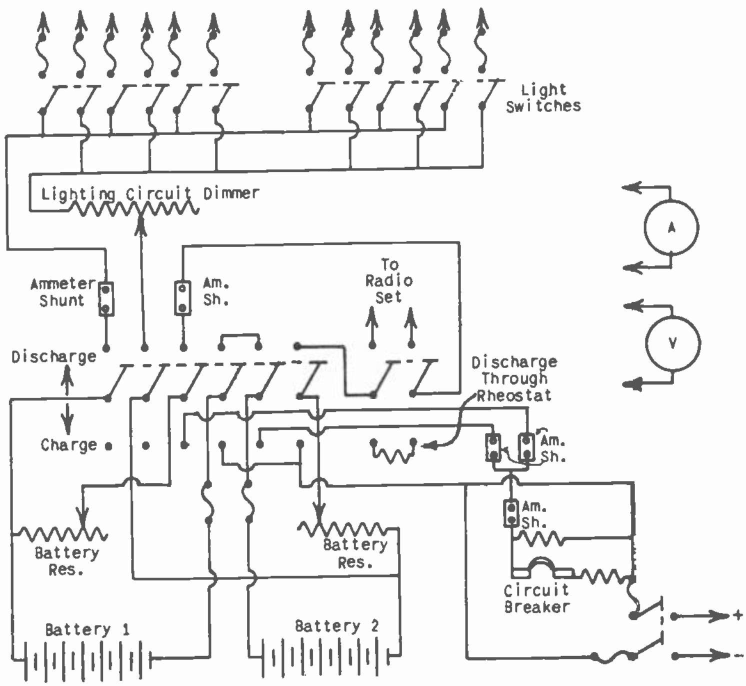


Fig.5 A schematic diagram of the Edison battery switchboard.

It has been determined that Edison batteries when new will have better capacities if they are given plenty of work. Furthermore, regardless of the age of the battery, sluggishness or low capacity will result from continual low-rate discharge or idleness. It is for this reason that a special rheostat is incorporated in the Edison battery switchboard. The normal procedure is to discharge the battery completely through the rheostat and then short-circuit it for two hours. This action is followed by an overcharge which is a charge at the normal rate for a somewhat longer period than is ordinarily given. The length of the overcharge will vary from 8 to 15 hours, depending on the type of cell. The switchboard includes a voltmeter, fuses, circuit-breaker, and lighting-circuit dimmer.

When direct current is not available, and it is not practical to use a DC generator as a charging source, it is necessary to employ a rectifier to convert the alternating current into pulsating direct current. There are two different types of rectifiers; they are the copper oxide type and the gaseous tube type. The copper oxide battery charger has a transformer which reduces the 110 volts AC to about 9 volts AC. This is then applied to a copper oxide rectifier arranged in a bridge circuit. A schematic diagram of this arrangement is illustrated in Fig. 6.

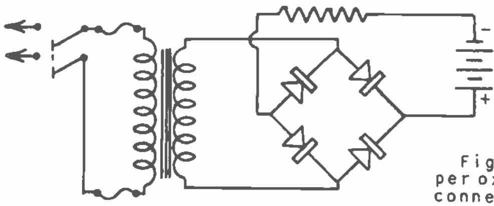


Fig.6 Showing how a copper oxide battery charger is connected.

Perhaps the most popular type of charger is that using a gaseous tube. The tube is very similar to an ordinary half wave mercury vapor tube, but has the difference that it will pass several amperes of current at a low voltage. The circuit used may be either half wave or full wave, two tubes being required for the full-wave charger. The tubes themselves contain either argon or mercury vapor under low pressure. When the plate is made approximately 25 volts positive with respect to the filament, the emitted electrons produce intense ionization, and a large rectified current flows through

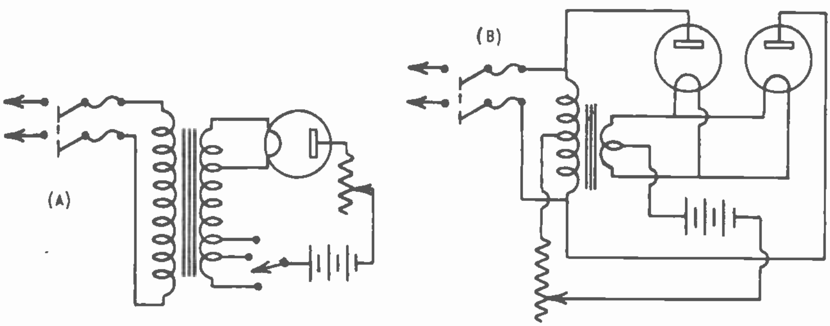


Fig.7 (A) A half-wave tube type battery charger. (B) A full-wave tube type battery charger.

the tube. Two types of circuits often used are shown in Fig. 7. The one at A is a half-wave rectifier and uses one tube, whereas that at B employs two tubes in a full-wave circuit.

An illustration of a Tungar charging bulb is given in Fig. 8. The name "Tungar" is the trade name used by the General Electric Company which manufactures Tungar charging equipment of all sizes.

Rectifier tubes for battery charging equipment are rated by their manufacturers according to their current-carrying ability.

The standard sizes range from .5 ampere to 15 amperes. The important things to be known about a charging bulb are its filament voltage, filament current, pick-up voltage, arc voltage, and maximum peak inverse voltage. Filament voltages on the average range from 1.8 volts to 2.5 volts, whereas the filament current is the prime factor in determining how much current the tube will pass.

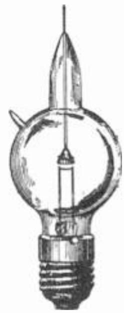


Fig.8 Showing the construction of a Tungar rectifier tube.

The pick-up voltage of a charging bulb is that voltage which must be impressed between the plate and filament before ionization or rectification will take place. This will be between 11 volts and 13 volts, depending on the size of the bulb. The arc voltage is the voltage drop across the tube while it is conducting current, and is normally 7 or 8 volts. The maximum peak inverse voltage which determines how much voltage may be applied to the tube without causing arc over is usually fairly low compared to other types of rectifier tubes; it is rarely over 400 volts, and usually considerably less.

Notes

(These extra pages are provided for your use in taking special notes)

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UNIT NO. 3

INTRODUCTION TO ALGEBRA

LESSON NO. 15

YOU NEVER KNOW

.....until you try.

Charlie Chaplin, the internationally famous movie star has earned a fabulous amount of money, and today is a wealtny man. Yet Chaplin came very close to ignoring 'his' opportunity when he came face to face with it years ago.

It seems that the late Michael Selwyn saw Chaplin on the stage, and was greatly impressed by the actor's ability at pantomine. Investigation revealed that he was earning about sixty dollars a week. Selwyn went backstage to talk to Chaplin and offered him an opportunity to enter the movies at $150.00 a week. But Charlie scorned the idea. He knew nothing about acting before movie cameras and no one would pay him that much money.

If Chaplin had not listened to reason, and then visualized the possible opportunities in the new field of entertainment, he probably would never have become a wealthy and famous man. Perhaps he shudders occasionally when he thinks about the time when he almost passed up fame and fortune.

Many young men who today are vainly seeking jobs, or doing work that brings them but little in the way of cash rewards should place the blame for their failure directly upon their own shoulders. They have ignored 'their' opportunity, and are just drifting and coasting along. But.... *you are not one of these men.*

You are proving that you mean business....that you are willing to work, to train for success. Continue your forward march with determination. Stick to each lesson until you have mastered it completely. If you will do this, and we are confident that you will, success will quickly come within your reach.

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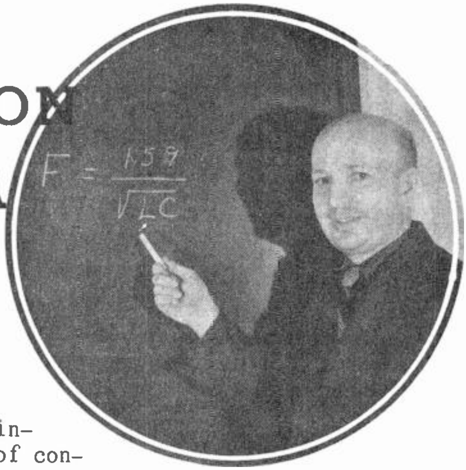
Lesson Fifteen

INTRODUCTION TO ALGEBRA



"Anyone who deals in radio realizes the importance of mathematics. Without the knowledge of multiplication and division we could not put Ohm's law to any practical use. In calculating the inductance of coils, the capacity of condensers, and the resonant frequency of a circuit, etc., we use formulas that are not always as simple as Ohm's law but if we know our "math" these formulas can be solved easily.

"In this lesson we shall study simple equations and their application; exponents and powers; and algebra as applied to radio."

1. INTRODUCTION. *(a) Definition.* The object of algebra is to investigate the properties of numbers and to deal with their combinations and relations. Many of the numbers used in practical work are obtained by counting a group of objects or from the measurement of something. The measure of an object, for instance, the resistance of a conductor, is the number of times it contains a unit of measure. If one Ohm is the unit of measure used, the measure of the resistance of a conductor of 10 ohms is the number 10. It is customary when measuring some object, to give the measure and also the unit of measure. For instance, the resistance of a conductor is 10 ohms, the capacity of a condenser is 8 mfd., or the inductance of a coil is 20 hy.

(b) Definite Numbers. The numbers 5, 8, 10, etc., have definite meanings. For instance, 8 represents a certain idea, that we call eight. It may be 8 ohms, 8 mfd., 8 kc., or 8 of many other units; but in all cases it is a definite number and therefore has a fixed value.

(c) General Numbers. A general number is a number that might change under different conditions. An example is the letter E in Ohm's law which represents voltage. The voltage is not always the same in different applications of Ohm's law, therefore, this letter represents in a general way a certain amount of voltage. Its value might be 100 volts, 10 volts, or any number used in the measurement of voltage. Likewise, I represents the current,

but when used in the formula R = E÷I, we do not think of a particular value for I. Such an idea as represented by E or by I cannot be represented by numerals. The idea is a general-number idea and is usually represented by a letter of the alphabet.

In a certain discussion, the letters used stand for the same value throughout the discussion. For example, when considering a particular application of Ohm's law, the letter E represents a definite value, such as 100 volts.

2. ALGEBRAIC EXPRESSIONS. An algebraic expression is an expression that represents a number my means of the symbols and signs of algebra. Therefore, a numerical algebraic expression is one made up wholly of numerals, while a literal algebraic expression is one that contains letters. Thus, 15-10 + (9-4) is a numerical algebraic expression, while xy - ab is a literal algebraic expression. The number it represents is the value of an algebraic expression.

In an expression such as 5axy*; 5, a, x, and y are factors of the expression. Any of these factors or the product of any two or more of them is called the COEFFICIENT of the remaining part. Thus, 5ax can be considered the coefficient of y, or 5a the coefficient of xy, BUT USUALLY THE COEFFICIENT IS THE NUMERAL PART ONLY. It may then be called the numerical coefficient and if no numerical part is expressed, such as in axy, then the numerical coefficient is understood to be 1.

3. EXPONENTS. If all the factors in a product are equal (a·a·a·a), the product of the factors is called a power of one of them. The form a·a·a·a is written a^4, the small number 4 indicating how many times (a) is taken as a factor. In the above case, (a) is called the base and 4 the exponent. THE EXPONENT OF A POWER IS THE NUMBER WRITTEN TO THE RIGHT AND A LITTLE ABOVE THE BASE. When the exponent is a positive whole number, it shows how many times the base is taken as a factor. For example, x^2 is read x square or x to the second power, and x^n is read x to the n^{th} power. When no exponent appears, it is understood that the exponent is 1 and is read x.

A term in an algebraic expression is a part of an expression that is not separated by a plus (+) or minus (-) sign. Thus, in the algebraic expression ax-4b+8y, ax, 4b, and 8y are terms.

A MONOMIAL is an algebraic expression consisting of one term (3ay). A BINOMIAL consists of two terms (3a+4bc); and a TRINOMIAL consists of three terms ($3x^2$-2x+2). Any algebraic expression consisting of two or more terms is known as a POLYNOMIAL OR A MULTINOMIAL.

* Note that the times sign (×) is omitted. If 5, a, x, and y denote separate factors, it is customary when using algebraic symbols to omit the sign of multiplication and to write them as a group, such as, 5axy. Even when numerals are to be multiplied, the times sign is omitted, since it is really unnecessary and might be confused with the letter x. Thus, to indicate the product of 5 and 10, the following method is used: 5·10 or (5)(10). The dot placed between the numbers and written above the line indicates that the numbers are to be multiplied. Likewise enclosing each factor in a pair of parentheses has the same significance. For example, 5·a·x·y, (5)(a)(x)(y) and 5axy all mean that 5 is to be multiplied by a, then this product by x, and the result of these two multiplications is then to be multiplied by y. If (a) has a value of 3, (x) of 4, and (y) of 2, then: 5axy = 5·3·4·2 = 120.

Terms that are the same or only differ in their numerical coefficients are called LIKE TERMS. For example, $5a^2x^4$ and $8a^2x^4$ are like terms while $4xa^3$ and $6x^2a^4$ are UNLIKE TERMS.

A FORMULA may be defined as a rule written in algebraic language or may be defined as a rule stated in letters and other symbols.

A *numerical* algebraic expression has a definite value which can be found by going through the necessary operations required. Thus $10-5+(5-1)-6+14 = 10-5+4-6+14 = 17$.

A *literal* algebraic expression has a definite value depending upon the values given the letters in the expression. Thus xyz has a definite value if $x = 3$, $y = 5$, and $z = 7$. Replacing these values in the expression we have $3 \cdot 5 \cdot 7 = 105$. If any other values were assigned to x, y, and z, another definite product would be obtained.

4. ADDITION AND SUBTRACTION OF LITERAL ALGEBRAIC EXPRESSIONS. When we add 5 ohms, 9 ohms, and 14 ohms, we get 28 ohms. That is 5 ohms+9 ohms+14 ohms = 28 ohms. Similarly, $5x+6x+8x = 19x$ and $3ab+4ab+7ab = 14ab$. Literal algebraic expressions are subtracted in the same manner as numerical algebraic expressions. Thus, $17xy-7xy = 10xy$, and $10a^2b-2a^2b-5a^2b = 3a^2b$.

We know that we cannot add and subtract unlike things. That is, we would not think of trying to add 8 ohms and 6 cycles. Likewise we cannot add 8x and 6y. To add 8x and 6y we indicate the addition; thus, $8x+6y$. From the above we have the following rule: MONOMIALS WHICH ARE ALIKE, OR SIMILAR CAN BE ADDED AND SUBTRACTED BY ADDING OR SUBTRACTING THE COEFFICIENTS. IF THE MONOMIALS ARE UNLIKE, THE OPERATIONS OF ADDITION AND SUBTRACTION CAN ONLY BE INDICATED.

The addition or subtraction of polynomials is similar to that of monomials. To add or subtract polynomials, place like terms in the same column in the same manner as you would monomials. Below are examples of the addition and subtraction of polymonials.

$$\text{Addition:}\quad \begin{array}{r} 3xy+7z-14ab-c \\ \underline{-2xy+4z+6ab-2c} \\ xy+11z-8ab-3c \end{array}$$

$$\text{Subtraction:}\quad \begin{array}{r} 17xy-14z+2ac-5b \\ \underline{10xy+4z-7ac-9b} \\ 7xy-18z+9ac+4b \end{array}$$

The results of the above problems and like problems can be tested as to there correctness by substituting definite values for the general numbers. Thus, if x, y, z, a, b, c, were all given a value of 1, we would obtain the following results:

$$\text{Addition:}\quad \begin{array}{r} 3+7-14-1 = -5 \\ \underline{-2+4+6-2 = +6} \\ 1+11-8-3 = 1 \end{array}$$

Of course, we could have substituted any values for the letters but to make the operation as simple as possible, we chose 1.

It often happens that we might want to add or subtract where the coefficients that are to be united are not all numerical. For example, the following problem may be added by uniting the coefficients of x:

$$a^2x+y^4x+b^3x = x(a^2+y^4+b^3)$$

Similarly the sum of 8y, 5y, 4y, may be united and written as (8+5+4)y and here the coefficients can also be united and expressed as one symbol, thusly 17y.

EXERCISES

Perform the following operations:

1. Add 3ax-6bx-2by-7ay to -2ay+4bx-3by (3ax-2bx-5by-9ay Ans.)
2. Add 6xz-4az to 3yz+2xz (3yz+8xz-4az Ans.)
3. Subtract 2wz-3xz from 8xz-4wz (11xz-6wz Ans.)
4. Subtract -7xy+3by from xy-4by (8xy-7by Ans.)
5. Subtract 3ay-2by from 4by+2ab (6by+2ab-3ay Ans.)

5. SIGNS OF GROUPING. When a sign of grouping, such as (y-x), is preceded by a (+) or (-) sign, it indicates that the expression (y-x) is to be added or subtracted from what precedes it. It is possible to remove these signs of grouping without changing the value of the expression. For example, when a plus sign precedes a sign of grouping, the sign of grouping may be removed without changing any of the signs. Thus, a+(x-y)-b = a+x-y-b. However, when the sign of grouping is preceded by a minus sign, the signs within the grouping are changed when the sign of grouping is removed. Thus, a-(x-y)-b = a-x+y-b. Similarly, a-(-b+c-x) = a+b-c+x. When several signs of grouping are used; (that is, one pair within another), they may be removed in the same manner by removing the innermost pair first. For example, $8y^3+5x^2+[6a^2-4z^3-(-b^3+2x)]$ may be simplified in the following manner:

$$8y^3+5x^2+[6a^2-4z^3-(-b^3+2x)]$$

$$= 8y^3+5x^2+[6a^2-4z^3+b^3-2x]$$

$$= 8y^3+5x^2+6a^2-4z^3+b^3-2x \quad \text{Ans.}$$

When it is necessary to put polynomials into signs of groupings, the same rules that applied above are used. If the sign is plus, any terms of a polynomial may be enclosed in a sign of grouping without changing any of the signs. However, if the sign preceding the grouping is minus, then all the signs of the terms included in the grouping must be changed; that is from (-) to (+) or from (+) to (-). For example, the expression ax+ay-xy+2zx-bc-az may be grouped in the following manner:

$$ax+ay-xy+(2zy-bc-az)$$

And if the last four terms are grouped, then the signs are changed thusly:

$$ax+ay-(xy-2zx+bc+az)$$

6. EQUATIONS. An equation may be defined as a statement that two expressions are equal in value. Thus, $F = 159 \div \sqrt{LC}$ is an equation; so are $W = E \times I$ and $X_L = 6.28fL$.

The part to the left of the equality sign is called the FIRST MEMBER of the equation, while the part to the right of the equality sign is known as the SECOND MEMBER.

There is a slight difference between a formula and an equation which should be made clear at this time. If the factor on the left hand side of the equality sign is unknown and all the factors on the right hand side are known, then the expression is called a formula. If both the right and left hand factors contain unknowns, then it is called an equation.

If the inductive reactance of a coil is 3,140 Ohms and the frequency is 5,000 cycles, we have the statement:

$$3{,}140 = (6.28)(5000)L$$

This equation is true when, and only when, L = .1 henry. Such an equation where the unknown has a certain value, is called a CONDITIONAL EQUATION. That is, the equation is true only on the condition that L = .1 henry, and for no other value of L. Not all equations are conditional, and where the unknown may have any value, the equation is called an IDENTICAL EQUATION or an IDENTITY. For instance, in the equation $(x^2-9) \div (x+3) = x-3$, x may have any value and yet the equation is true. Thus, if x = 3, we would have $(9-9) \div 3+3 = 3-3$ or $0 \div 6 = 0$ or $0 = 0$. Similarly, if x = 4, the equation becomes:

$$\frac{(4)^2-9}{4+3} = 4-3$$

$$= \frac{16-9}{4+3} = 4-3$$

$$= \frac{7}{7} = 1$$

$$\text{or } 1 = 1$$

7. SOLUTION OF EQUATIONS. To solve an equation is to find the value or values of the unknown that will make the equation true.

This may or may not be a simple matter. We will take up the solving of simple equations first.

EXAMPLE 1. Find the value of x, if x-6 equals 4.

Given equation	$x-6 = 4$
Adding 6 to each member,	$x-6+6 = 4+6$
Collecting terms	$x = 10$ Ans.

EXAMPLE 2. Solve for y, if $y+4 = 9$.

Given equation,	$y+4 = 9$
Subtracting 4 from each member,	$y+4-4 = 9-4$
Collecting terms,	$y = 5$ Ans.

EXAMPLE 3. Solve for x, if $6x-5+4 = 3x-30+5$

Given equation,	$6x-5+4 = 3x-30+5$
Adding 5 to both members,	$6x-5+4+5 = 3x-30+5+5$
Collecting terms	$6x+4 = 3x-20$
Subtracting 4 from both members,	$6x+4-4 = 3x-20-4$
Collecting terms,	$6x = 3x-24$
Subtracting 3x from both members,	$6x-3x = 3x-3x-24$
Collecting terms,	$3x = -24$
Dividing both members by 3,	$x = -8$ Ans.

In the foregoing equations, the primary object is to arrange the equation so that the first member consists only of terms containing the unknown, and the second member, only of terms not containing the unknown. This rearrangement of the terms was accomplished by adding or subtracting various quantities, making certain that both members were treated alike. A much simpler method for rearranging terms is the process known as "transposition". Its underlying principle states that any term of an equation may be moved from one member to the other if its sign be changed. For example, consider the equation: $8x+4 = 3x-1$

It is desirable to place the (+4) in the second member and the (3x) in the first member. By changing the signs of each of these terms as they are moved from one member to the other, we obtain:

$$8x-3x = -4-1$$

This equation may now be solved by collecting terms and then dividing by the coefficient of the unknown. Thus:

$$5x = -5$$

$$x = \frac{-5}{5}$$

$$x = -1$$

It is not necessary that all of the steps indicated be written; however, the three following rules should be observed in solving any simple equation:

(1) Transpose all terms containing the unknown to the first member, and all other terms to the second member. In each case where the term is transposed, be sure to change the sign.
(2) Collect the terms in each member.
(3) Divide each member by the coefficient of the unknown.

EXAMPLE 4. Solve for x in the equation $5x-6+8 = 3x+6$

SOLUTION:

$$5x-3x = 6+6-8$$

$$2x = 4$$

$$x = 2$$

From the foregoing discussion, we find the following rules apply.

(1) If equal numbers are added to equal numbers, the sums are equal.
(2) If equal numbers are subtracted from equal numbers, the remainders are equal.
(3) If equal numbers are multiplied by equal numbers, the products are equal.
(4) If equal numbers are divided by equal numbers, the quotients are equal.
(5) Numbers that are equal to the same number are equal to each other.
(6) Like powers of equal numbers are equal.
(7) Like roots of equal numbers are equal.
(8) The whole of anything equals the sum of all its parts.

8. TESTING THE EQUATION. Equations that have been solved for the unknown may be tested by substituting in the original equation the value of the unknown that was obtained. For instance, in Example 4, the value obtained for x was 2. Substituting 2 in the original equation, we have:

$$5(2)-6+8 = 3(2)+6$$

$$\text{or } 10-6+8 = 6+6$$

$$12 = 12$$

EXERCISES

Solve for the unknown:

1. $8x-2 = 4+2x$ $\qquad x = 1$ Ans.

2. $7-6y-2 = 4y$ $\qquad y = \frac{1}{2}$ Ans.

3. $z+2-3z = 4$ $\qquad z = -1$ Ans.

4. Solve for x: $ax-2+3ax = 4+2ax$ $\qquad x = \frac{3}{a}$ Ans.

5. Solve for y: $by+ay-2 = ay+by+y$ $\qquad y = -2$ Ans.

9. THE EQUATION AS USED FOR SOLVING PROBLEMS. Equations can be used for the solving of problems that are stated in English. In the translation of English to mathematical language, no set rules can be given, but there are a few suggestions that may be helpful in stating a problem in the form of an equation.

(1) Read the problem carefully as it is given in words.
(2) Select the unknown quantity and represent it by some letter of the alphabet. If there are two or more unknowns, try to express the second in terms of the first unknown.
(3) Find two expressions which, according to the problem, represent the same number, and set them equal to each other. Thus the equation to be solved is formed.

Below are listed a few illustrations:

EXAMPLE 1. If twice the value of an unknown resistance is added to six times the unknown resistance and their sum is equal to 96 ohms. What is the value of the resistance?

SOLUTION: Let x be the unknown resistance. Then, if 6 times this value plus 2 times this value is equal to 96, the equation must be:

$$6x+2x = 96$$

$$8x = 96$$

$$x = 12 \text{ Ans.}$$

EXAMPLE 2. The sum of two numbers is 58, and their difference is 12. What are the numbers?

SOLUTION: Let x be the smaller number. Then, since the smaller is 12 less than the greater, the greater number may be represented by x+12. Next, the problem states that the sum of the numbers is

58 and so the sum of x and (x+12) must be equal to 58. This makes the equation read:

$$x+(x+12) = 58$$

$$x+x+12 = 58$$

$$2x+12 = 58$$

$$2x = 58-12$$

$$2x = 46$$

$$x = 23 \quad \text{(the smaller number)}$$

$$x+12 = 35 \quad \text{(the larger number)}$$

Check: $35+23 = 58$

EXERCISES

Solve the following problems.

(1) One number is three times another and their sum is 12. What are the numbers? Ans. 3,9.

(2) A room is 10 feet longer than it is wide. The total distance around the four sides is 60 feet. What are the dimensions of the room? Ans. 10 ft. x 20 ft.

(3) A radio receiver complete with tubes cost $125.00. The receiver itself costs $110.00 more than the tubes. What is the price of the tubes? Ans. $7.50.

10. MULTIPLICATION. *(A) Unlike Terms.* It is well known that the factors of a product may be written in any order.

For example: $4\cdot 2 = 2\cdot 4$

Similarly $x\cdot y = y\cdot x$

Further $3x^2\cdot 4 = 3\cdot 4\cdot x^2 = 12x^2$

And $2x\cdot 4y = 2\cdot 4\cdot x\cdot y = 8xy$

Also $5x^2\cdot 6y^3 = 5\cdot 6\cdot x^2\cdot y^3 = 30x^2y^3$

(B) Like Terms. From the definition of an exponent, we have learned that $x\cdot x\cdot x\cdot x = x^4$ and $x\cdot x\cdot x\cdot x\cdot x = x^5$.

Therefore: $x^2\cdot x^3 = x\cdot x\cdot x\cdot x\cdot x = x^5 = x^{2+3}$

In like manner: $5xy^2z^3\cdot 4x^3y^4 = 20x^4y^6z^3$

Therefore, we have the following rule: THE EXPONENT OF ANY

LETTER IN THE PRODUCT IS EQUAL TO THE SUM OF THE EXPONENTS OF THAT LETTER IN THE FACTORS. This may be expressed in general terms thusly:

$$n^a \text{ times } n^b \text{ equals } n^{a+b}$$

In multiplying literal terms of positive and negative numbers, the same law of signs applies as with numerical terms. That is, when like signs are multiplied, the sign of the product is positive, and when unlike signs are multiplied, the sign of the product is negative.

For example:	$2x^2 \cdot +4x^3$	=	$+8x^5$
Similarly	$-2x^2 \cdot -4x^3$	=	$+8x^5$
And	$2x^2 \cdot -4x^3$	=	$-8x^5$
Also	$-2x^2 \cdot 4x^3$	=	$-8x^5$

For the multiplication of two monomials we have the following procedure: Observing the rule of signs, first obtain the product of the numerical coefficients followed by all the letters that occur in the factors, each letter having as its exponent the sum of the exponents of that letter in both factors.

For example:	$(4x^2)(2xz^2)$	=	$8x^3z^2$
Similarly	$(2abz^2)(x^2y^3z)(4a^2b^3)$	=	$8a^3b^4x^2y^3z^3$
And	$(-2a)(4ab)(ax)$	=	$-8a^3bx$
Also	$(-2zx)(-3ac)(2acxz)$	=	$12a^2c^2x^2z^2$*

(C) Multiplication of a Polynomial and a Monomial. It is easily seen that $4(3+5)$ is equivalent to $(4\cdot3)+(4\cdot5)$, each expression being equal to 32. Similarly $x(y+z) = xy+xz$. Therefore, the rule is: Multiply each term of the polynomial by the monomial and write in order the resulting terms with the proper signs.

EXAMPLE: Multiply $4xy^3-7ab-6a^2x^2$ by $2a^3x^4y$

Process
$$\begin{array}{r} 4xy^3-7ab-6a^2x^2 \\ \underline{2a^3x^4y\qquad\qquad\quad} \\ 8a^3x^5y^4-14a^4bx^4y-12a^5x^6y \end{array}$$

(D) Multiplication of Polynomials. RULE: Multiply the multiplicand by each term of the multiplier in turn, and add the partial products.

* It will be noted that the product is a positive number. When the two negative numbers were multiplied, the sign became positive and remained positive when multiplied by the remaining positive number.

EXAMPLE 1. Multiply $x-4$ by $x+6$.

Process:
$$\begin{array}{r} x-4 \\ \underline{x+6} \\ x^2-4x \\ \underline{6x-24} \\ x^2+2x-24 \end{array}$$

EXAMPLE 2. Multiply $a^2+3ab-2b^2$ by $2ab-2b^2$.

Process:
$$\begin{array}{r} a^2+3ab-2b^2 \qquad\qquad \\ \underline{2ab-2b^2 \qquad\qquad\qquad} \\ 2a^3b+6a^2b^2-4ab^3 \qquad\quad \\ \underline{-2a^2b^2-6ab^3+4b^4} \\ 2a^3b+4a^2b^2-10ab^3+4b^4 \end{array}$$

Problems in multiplication can be tested by substituting numerical values for the letters. It is better to use values larger than 1 since any power of 1 is 1. For example, let a = 2 and b = 2 in the foregoing example.

$$\begin{array}{rcl} a^2+3ab-2b^2 & = 4+12-8 & = 8 \\ \underline{2ab-2b^2} & = \underline{8-8} & = \underline{0} \\ 2a^3b+6a^2b^2-4ab^3 & & \ 0 \\ \underline{-2a^2b^2-6ab^3+4b^4} & & \\ 2a^3b+4a^2b^2-10ab^3+4b^4 & = 32+64-160+64 & = 0 \end{array}$$

The work is correct if the product of the values of the two factors equals the value of the product. In this problem, the value in each case was 0.

EXERCISES

1. Multiply
   $4ax^2-2b^2y-6$ by $6a^2xy$ ($24a^3x^3y^3-12a^2b^2xy^4-36a^2xy^3$ Ans.)

2. Multiply $6xy+2x^2y-3xy^2$ by $2xy$ ($12x^2y^2+4x^3y^2-6x^2y^3$ Ans.)

3. Multiply x^2-4 by $x+3$ ($x^3+3x^2-4x-12$ Ans.)

4. Multiply x^2y-3ax by $axy+3a$ ($ax^3y^2-3a^2x^2y+3ax^2y-9a^2x$ Ans.)

5. Multiply a^2+ab+b^2 by $a-b$ (a^3-b^3 Ans.)

11. DIVISION. *(A) Law of exponents.* Division is the inverse of multiplication and, as in multiplication, there are certain rules that apply. First, the numerical part of the quotient is the quotient of the absolute values. Second, the sign of the quotient is plus when the signs of the dividend and divisor are alike, and minus when the signs are unlike.

In dividing powers of the same base, if the exponent of the dividend is the larger, the exponent of the quotient equals the difference of the exponents, or, the exponent of the dividend minus the exponent of the divisor. This can readily be seen in the following example.

$$\frac{x^5}{x^3} = \frac{x\cdot x\cdot x\cdot x\cdot x}{x\cdot x\cdot x} = x^2 \text{ or } x^5 \div x^3 = x^{5-3} = x^2$$

If the exponents of the dividend and divisor are equal, the quotient is 1. Thus, $x^5 \div x^5 = 1$. This can be shown by the illustration:

$$\frac{2^2}{2^2} = \frac{4}{4} = 1$$

This same rule applies even though the exponent of the dividend is smaller than that of the divisor. For example, let us consider this division:

$$x^3 \div x^5 = \frac{x^3}{x^5} = \frac{x\cdot x\cdot x}{x\cdot x\cdot x\cdot x\cdot x} = \frac{1}{x\cdot x}$$

Now using the rule of subtracting the exponent of the divisor from the exponent of the dividend, we would obtain:

$$x^3 \div x^5 = x^{3-5} = x^{-2}$$

Thus by definition, $x^{-2} = 1 \div x^2$, and in a like manner, $x^{-3} = 1 \div x^3$ etc. This law of exponents can be put in general terms thusly:

$$n^a \div n^b = n^{a-b}$$

(B) Division of one monomial by another. It is easier to perform the work of division in regular steps. First, determine the sign of the quotient; next, the coefficient of the quotient; finally, the letters and exponents.

EXAMPLE: Divide $30x^6y^4$ by $-6x^2y^3$
Process as carried out in steps:

$$30 \div -6 = -5$$

$$x^6 \div x^2 = x^4$$

$$y^4 \div y^3 = y$$

Combining the quotients: $30x^6y^4 \div -6x^2y^3 = -5x^4y$ Ans.

The different steps carried out in the above example are omitted in actual practice and only the last is written down.

The division of monomials can also be performed as a cancellation. For instance, the above example may be written:

$$\frac{30x^6y^4}{-6x^2y^3} = \frac{\overset{-1}{5}\cdot\cancel{6}\cdot x\cdot x\cdot x\cdot x\cdot\cancel{x}\cdot\cancel{x}\cdot\cancel{y}\cdot\cancel{y}\cdot\cancel{y}\cdot y}{-\cancel{6}\cdot\cancel{x}\cdot\cancel{x}\cdot\cancel{y}\cdot\cancel{y}\cdot\cancel{y}} = -5x^4y \text{ Ans.}$$

The work in division can be checked in the same manner as multiplication; that is, by substituting numerical values for the letters. It may also be checked by multiplying the divisor by the quotient; the product of this multiplication will be the dividend.

(C) Division of a Polynomial by a Polynomial. To divide a polynomial by a polynomial, the following procedure should be observed. First, arrange the dividend and divisor in descending powers. This means that the highest power of the unknown should be written first, followed by the next lower power of the unknown, etc., the last term ordinarily being one which does not contain the unknown. For example, $x+x^3-3+4x^2$ would be arranged in this manner: x^3+4x^2+x-3. Next, divide the first term of the dividend by the first term of the divisor to find the first term of the quotient. Now multiply the divisor by the first term of the quotient writing the product under like terms of the dividend, and subtract from the dividend. Then bring down other terms of the dividend. Next, divide the first term of this remainder by the first term of the divisor to find the second term of the quotient. Now, multiply the divisor by the second term of the quotient, writing the product under the remainder, subtract, and continue on for other terms that might be in the quotient. The division will be complete when there is no remainder or when the remainder is of lower power than the highest power of the divisor.

By the above procedure, it can be seen that the process is very similar to long division in arithmetic.

EXAMPLE: Divide $a^2+7a+12$ by $a+3$.

SOLUTION

$$\begin{array}{r|l} & \underline{\quad a+4} \\ a+3 & a^2+7a+12 \\ & \underline{a^2+3a} \\ & \quad 4a+12 \\ & \quad \underline{4a+12} \end{array}$$

EXAMPLE 2. Divide a^4-3a^2-54 by $a-3$

Since neither the third nor first power of (a) is contained in the dividend, it is assumed that their coefficients are zero, and space is left for them in writing the dividend in powers of descending order.

$$\begin{array}{r|l} & \underline{a^3+3a^2+6a+18} \\ a-3 & a^4 \quad -3a^2 \quad -54 \\ & \underline{a^4-3a^3} \\ & +3a^3-3a^2 \\ & \underline{+3a^3-9a^2} \\ & \quad +6a^2 \quad -54 \\ & \quad \underline{+6a^2-18a} \\ & \qquad 18a-54 \\ & \qquad \underline{18a-54} \end{array}$$

EXAMPLE 3. Divide x^4-1 by x^2-2

SOLUTION:

$$\begin{array}{r|l} & x^2 \quad +2 \\ \hline x^2-2 & x^4 \qquad -1 \\ & \underline{x^4-2x^2} \\ & \quad +2x^2-1 \\ & \quad \underline{+2x^2-4} \\ & \qquad\quad +3 \end{array}$$

$$x^2+2+\frac{3}{x^2-2} \quad \text{Ans.}$$

Where there is a remainder in the answer, it is indicated as above. The answer may be checked thusly; dividend = divisor × quotient + remainder.

EXERCISES

1. Divide $7a^5c^4xy^2$ by $-7a^4c^4y$ ($-axy$ Ans.)
2. Divide $24a^5y^3-96a^5y^6$ by $8a^4y^3$ ($3a-12ay^3$ Ans.)
3. Divide $x^2+13x-48$ by $x-3$ ($x+16$ Ans.)
4. Divide x^4-1 by x^2+1 (x^2-1 Ans.)
5. Divide a^3-8 by $a-2$ (a^2+2a+4 Ans.)

12. CLEARING AN EQUATION OF FRACTIONS. When an equation contains a fraction in which the unknown quantity appears, it must be changed so that the equation does not contain any fractions. The process of changing the equation so the unknown does not appear in the fraction is known as the "clearing of fractions". ANY EQUATION CAN BE CLEARED OF FRACTIONS BY MULTIPLYING BOTH MEMBERS BY THE LEAST COMMON DENOMINATOR OF THE FRACTIONS. This can best be explained by the following examples:

EXAMPLE 1. Solve for x in the equation $(2x \div 3)+(x \div 5) = 45$. By inspection it is seen that the least common denominator is 15; that is, $3 \cdot 5$. Thus, multiply each side of the equation by 15:

$$\frac{15 \cdot 2x}{3} + \frac{15 \cdot x}{5} = 15 \cdot 45$$

$$\frac{30x}{3} + \frac{15x}{5} = 675$$

Now by cancellation we obtain:

$$\frac{\overset{10}{\cancel{30}}x}{\cancel{3}} + \frac{\overset{3}{\cancel{15}}x}{\cancel{5}} = 10x + 3x = 675$$

Therefore; $13x = 675$, $x = 51\frac{12}{13}$

EXAMPLE 2. Solve for x in the equation $(3x \div 4)+(2x \div 3) = 5 \div 12$. By inspection it is clear that the least common denominator is 12. Then when each side is multiplied by the common denominator we have:

$$\frac{12(3x)}{4} + \frac{12(2x)}{3} = \frac{12(5)}{12}$$

By cancellation:

$$9x+8x = 5$$

$$17x = 5$$

$$x = \frac{5}{17}$$

13. REARRANGING FORMULAS. Quite often it is necessary to rearrange a formula that is not set up for ready solution of the unknown. That is, the unknown is on the right hand side of the formula with the known factors. It is always better to have the unknown on the left hand side of the equation and the known factors on the right hand side. Therefore, the equation must be rearranged so that it appears in the form that is wanted. For example, in the formula $F = 1 \div 2\pi\sqrt{LC}$, assume that the frequency and capacity are known and the inductance is unknown. It is desirable to have L in the left member and the F and C in the right member with the rest of the knowns.

SOLUTION: Squaring both sides of the equation, we have:

$$F^2 = \frac{1}{4\pi^2 LC}$$

Multiplying both sides of the equation by L and cancelling in the right hand member at the same time, we have:

$$LF^2 = \frac{1}{4\pi^2 C}$$

Dividing both sides of the equation by F^2 and cancelling in the left hand member at the same time, we obtain:

$$L = \frac{1}{4\pi^2 CF^2}$$

Thus we have the desired formula.

14. COMBINING FORMULAS. Sometimes it is desirable to combine formulas. For example, consider the formula $Z = \sqrt{R^2+(X_l-X_c)^2}$, which is the formula for finding the impedance of a resistance, inductive reactance, and capacitive reactance in series. If the inductive reactance or capacitive reactance is not known, but the line frequency, inductance and capacity are known, then by using the two

basic formulas for inductive reactance and capacitive reactance, we may combine them in the above formula to produce a formula that has the known terms in it.

$X_l = 2\pi FL$, (The formula for inductive reactance.)

$X_c = \frac{1}{2\pi FC}$, (The formula for capacitive reactance.)

Substituting the formulas for inductive reactance and capacitive reactance into the formula for finding the impedance of a series circuit containing resistance, inductive reactance and capacity reactance, the following formula is obtained:

$$Z = \sqrt{R^2 + \left(2\pi FL - \frac{1}{2\pi FC}\right)^2}$$

Thus the original formula becomes an important formula in AC circuit theory.

15. THE USE OF EXPONENTS FOR WRITING NUMBERS. The writing of very large or small numbers in the ordinary manner requires the use of a great many zeros. For instance, if 105 megohms is to be converted into ohms, the ordinary method of writing the number would be 105,000,000 ohms. It is, however, easy to see that 105,000,000 is equal to $105 \times 1,000,000$ and also, 1,000,000 is the same as 10^6. For this reason, the common method of writing such large numbers is to write the first two or three significant figures followed by a times sign and the number 10 to the correct power. In this case, 105,000,000 ohms would be written 105×10^6 ohms. The following table shows other examples of this method of notation:

324,000,000,000 is equal to 324×10^9
1,370,000 is equal to 137×10^4
3.000,000,000,000 is equal to 3×10^{12}

If two such large numbers are to be added, the exponents of each must be the same. For example, to add 45×10^6 to 37×10^5, we must change one or the other of the exponents so that each will be either 5 or 6. Let us, in this case make both exponents 5. Considering 45×10^6, we see that if the 10^6 is divided by 10, it will become 10^5. However, in order to keep the number at the same value, we must multiply the 45 by 10. The number then becomes: 450×10^5 and the sum is:

$$\begin{array}{r} 450 \times 10^5 \\ \underline{37 \times 10^5} \\ 487 \times 10^5 \end{array}$$

It would, of course, have been possible to make the exponents of both numbers 6. Considering 37×10^5, we see that if 10^5 is multiplied by 10, it will become 10^6, and so that the number will not be changed, the 37 must be divided by 10, making it 3.7. Thus the number becomes 3.7×10^6, and when added to 45×10^6, the sum is 48.7×10^6, which is the same result as before, but in slightly different form.

In a like manner, the writing of very small numbers also necessitates the use of a large number of zeros. For example, 5 mfd. is equal to .000005 farad. This number is the same as $5 \times (1 \div 1,000,000)$. The fraction, however, which is one-millionth may also be written as 10^{-6}. Thus, the number may be expressed as 5×10^{-6}, which is the same as 5 millionths of a farad. This notation is easily remembered by learning the following table:

$$
\begin{aligned}
.1 &= 10^{-1} \\
.01 &= 10^{-2} \\
.001 &= 10^{-3} \\
.0001 &= 10^{-4} \\
&\text{etc.}
\end{aligned}
$$

Note that the value of the exponent is the same as the number of decimal places in the number. Other examples are:

$$
\begin{aligned}
.000325 &= 325 \times 10^{-6} \\
.0000007 &= 7 \times 10^{-7} \\
.0000000076 &= 76 \times 10^{-10} \\
.0005 &= 5 \times 10^{-4}
\end{aligned}
$$

Again, when two small numbers written in this notation are to be added, the exponents of each must be the same. To add 4×10^{-5} to 32×10^{-6}, we must change one or the other of the numbers so that the exponents are the same. Considering 4×10^{-5}, we see that the exponent may be changed to −6 if the 10^{-5} is divided by 10; however, in order not to change the value of the number, the 4 must be multiplied by 10. When this is done, the number is expressed as; 40×10^{-6}, and when added to 32×10^{-6}, the sum is 72×10^{-6}.

When two numbers, written in this notation are to be multiplied, it is not necessary that the exponents of each be equal. For example, to multiply 3×10^{-5} by 4×10^{-2}, the 3 and 4 are multiplied together obtaining 12 and the exponent of the product will be the sum of the exponents of the two factors. This makes the product: 12×10^{-7}.

Notes

(These extra pages are provided for your use in taking special notes)

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