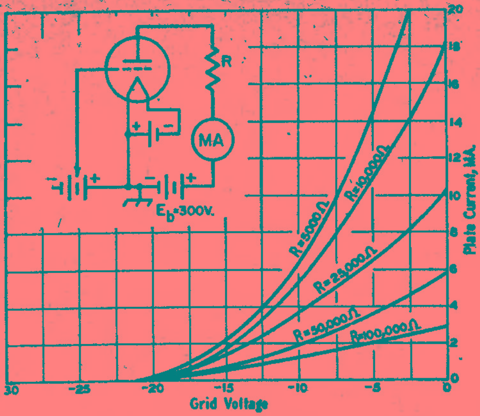


Fig. 3-7 – Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.

triode tube. The many uses of the electronic tube nearly all are based upon this amplifying feature. The amplified output is not obtained from the tube itself, but from the voltage source connected between its plate and cathode. The tube simply controls the power from this source, changing it to the desired form.

To utilize the controlled power, a load must be connected in the plate or "output" circuit, just as in the diode case. The load may be either a resistance or an impedance. The term "impedance" is frequently used even when the load is purely resistive.

Tube Characteristics

The physical construction of a triode determines the relative effectiveness of the grid and plate in controlling the plate current. The control of the grid is increased by moving it closer to the cathode or by making the grid mesh finer.

The plate resistance of a vacuum tube is the ac resistance of the path from cathode to plate. For a given grid voltage, it is the quotient of a small change in plate voltage divided by the resultant change in plate current. Thus if a 1-volt change in plate voltage caused a plate-current change of .01 mA (.00001 ampere), the plate resistance would be 100,000 ohms.

The **amplification factor** (usually designated by the Greek letter μ) of a vacuum tube is defined as the ratio of the change in plate voltage to the change in grid voltage to effect equal changes in plate current. If, for example, an increase of 10 plate volts raised the plate current 1.0 mA, and an increase in (negative) grid voltage of 0.1 volt were required to return the plate current to its original value, the amplification factors of triode tubes would be 100. The amplification factors of triode tubes range from 3 to 100 or so. A **high-μ** tube is one with an amplification of perhaps 30 or more, **medium-μ** tubes have amplification factors in the approximate range 8 to 30 and low-μ tubes in the range below 7 or 8. The μ of a triode is useful in computing stage gains.

The best all-around indication of the effectiveness of a tube as an amplifier is its grid-plate **transconductance** – also called **mutual conductance** or g_m. It is the change in plate current divided by the change in grid voltage that caused the change; it can be found by dividing the amplification factor by the plate resistance. Since current divided by voltage is conductance, transconductance is measured in the unit of conductance, the mho.

Practical values of transconductance are very small, so the micromho (one millionth of a mho) is the commonly used unit. Different types of tubes have transconductances ranging from a few hundred to several thousand. The higher the transconductance the greater the posible amplification.

AMPLIFICATION

The way in which a tube amplifies is best known by a type of graph called the **dynamic characteristic**. Such a graph, together with the circuit used for obtaining it, is shown in Fig. 3-7. The curves are taken with the plate-supply voltage fixed at the desired operating value. The difference between this circuit and the one shown in Fig. 3-6 is that in Fig. 3-7 a load resistance is connected in series with the plate of the tube. Fig. 3-7 thus shows how the plate current will vary, with different grid voltages, when the plate current is made to flow through a load and thus do useful work.

The several curves in Fig. 3-7 are for various values of load resistance. When the resistance is small (as in the case of the 5000-ohm load) the plate current changes rather rapidly with a given change in grid voltage. If the load resistance is high (as in the 100,000-ohm curve), the change in plate current for the same grid-voltage change is relatively small; also, the curve tends to be straighter.

Fig. 3-8 is the same type of curve, but with the circuit arranged so that a source of alternating voltage (**signal**) is inserted between the grid and the grid battery (**"C" battery**). The voltage of the grid battery is fixed at –5 volts, and from the curve it is seen that the plate current at this grid voltage is 2 milliamperes. This current flows when the load resistance is 50,000 ohms, as indicated in the circuit diagram. If there is no ac signal in the grid circuit, the voltage drop in the load resistor is 50,000 × .002 = 100 volts, leaving 200 volts between the plate and cathode.

When a sine-wave signal having a peak value of 2 volts is applied in series with the bias voltage in the grid circuit, the instantaneous voltage at the grid will swing to –3 volts at the instant the signal reaches its positive peak, and to –7 volts at the instant the signal reaches its negative peak. The maximum plate current will occur at the instant the grid voltage is –3 volts. As shown by the graph, it will have a value of 2.65 milliamperes. The minimum plate current occurs at the instant the grid voltage is –7 volts, and has a value of 1.35 mA. At intermediate values of grid voltage, intermdiate plate-current values will occur.

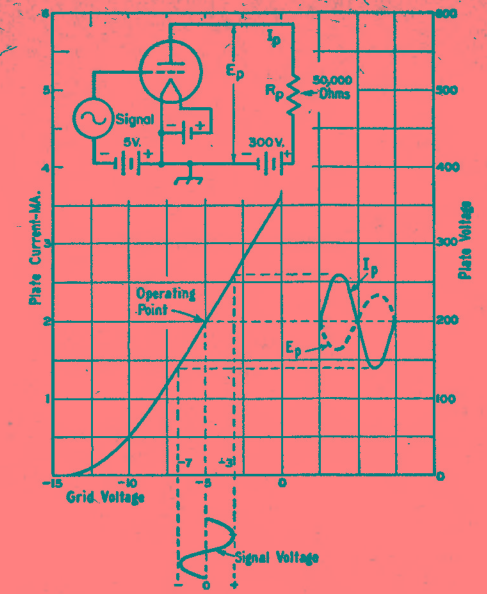


Fig. 3-8 — Amplifier operation. Whan the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve. E_p. I_p is the plate current.

The instantaneous voltage between the plate and cathode of the tube also is shown on the graph. When the plate current is maximum, the instantaneous voltage drop in R_p is 50,000 × .00265 = 132.5 volts; when the plate current is minimum the instantaneous voltage drop in R_p is 50,000 × .00135 = 67.5 volts. The actual voltage between plate and cathode is the difference between the plate-supply potential, 300 volts, and the voltage drop in the load resistance. The plate-to-cathode voltage is therefore 167.5 volts at maximum plate current and 232.5 volts at minimum plate current.

This varying plate voltage is an ac voltage superimposed on the steady plate-cathode potential of 200 volts (as previously determined for no-signal conditions). The peak value of this ac output voltage is the difference between either the maximum or minimum plate-cathode voltage and the no-signal value of 200 volts. In the illustration this difference is 232.5 – 200 or 200 – 167.5; that is, 32.5 volts in either case. Since the grid signal voltage has a peak value of 2 volts, the voltage-amplification ratio of the amplifier is 32.5/2 or 16.25. That is, approximately 16 times as much voltage is obtained from the plate circuit as is applied to the grid circuit.

As shown by the drawings in Fig. 3-8, the alternating component of the plate voltage swings in the *negative* direction (with reference to the no-signal value of plate-cathode voltage) when the grid voltage swings in the *positive* direction, and vice versa. This means that the alternating component of plate voltage (that is, the amplified signal) is 180 degrees out of phase with the signal voltage on the grid.

Bias

The fixed negative grid voltage (called grid bias) in Fig. 3-8 serves a very useful purpose. One object of the type of amplification shown in this drawing is to obtain, from the plate circuit, an alternating voltage that has the same wave shape as the signal voltage applied to the grid. To do so, an operating point on the straight part of the curve must be selected. The curve must be straight in both directions from the operating point at least far enough to accommodate the maximum value of the signal applied to the grid. If the grid signal swings the plate current back and forth over a part of the curve that is not straight, as in Fig. 3-9, the shape of the ac wave in the plate circuit will not be the same as the shape of the grid-signal wave. In such a case the output wave shape will be distorted.

A second reason for using negative grid bias is that any signal whose peak positive voltage does not exceed the fixed negative voltage on the grid cannot cause grid current to flow. With no current flow there is no power consumption, so the tube will amplify without taking any power from the signal source. (However, if the positive peak of the signal does exceed the negative bias, current will flow in the grid circuit during the time the grid is positive.)

Distortion of the output wave shape that results from working over a part of the curve that is not straight (that is, a nonlinear part of the curve) has the effect of transforming a sine-wave grid signal into a more complex waveform. As explained in an earlier chapter, a complex wave can be resolved into a fundamental and a series of harmonics. In other words, distortion from nonlinearity causes the generation of harmonic frequencies — frequencies that are not present in the signal applied to the

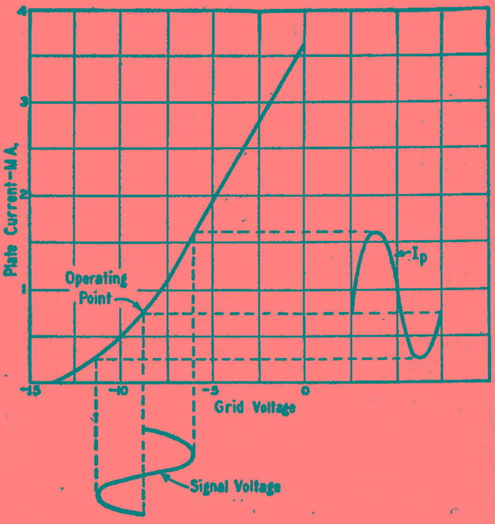
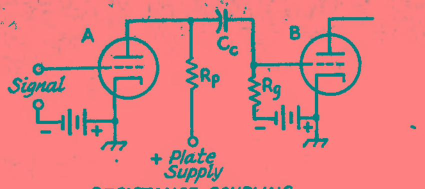
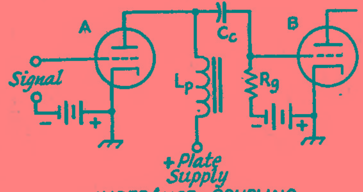


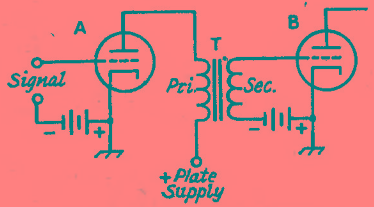
Fig. 3-9 — Harmonic distortion resulting from choice of an operating point on the curved part of tha tube characteristic. The lower half-cycla of plete current does not have the same shape as the upper half-cycle.



RESISTANCE COUPLING



IMPEDANCE COUPLING



TRANSFORMER COUPLING

Fig. 3-10 — Three types of coupling are in common use at audio frequencies. These are resistance coupling, impedance coupling, and transformer coupling. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits can be used to couple to other devices than tubes.

grid. Harmonic distortion is undesirable in most amplifiers, although there are occasions when harmonics are deliberately generated and used.

Audio Amplifier Output Circuits

The useful output of a vacuum-tube amplifier is the *alternating* component of plate current or plate voltage. The dc voltage on the plate of the tube is essential for the tube's operation, but it almost invariably would cause difficulties if it were applied, along with the ac output voltage, to the load. The output circuits of vacuum tubes are therefore arranged so that the ac is transferred to the load but the dc is not.

Three types of coupling are in common use at audio-frequencies. These are **resistance coupling, impedance coupling,** and **transformer coupling.** They are shown in Fig. 3-10. In all three cases the output is shown coupled to the grid circuit of a subsequent amplifier tube, but the same types of circuits can be used to couple to other devices than tubes.

In the resistance-coupled circuit, the ac voltage developed across the **plate resistor** R_p (that is, the ac voltage between the plate and cathode of the tube) is applied to a second resistor, R_g, through a **coupling capacitor,** C_c. The capacitor "blocks off" the dc voltage on the plate of the first tube and prevents it from being applied to the grid of tube *B*. The latter tube has negative grid bias supplied by the battery shown. No current flows on the grid circuit of tube *B* and there is therefore no dc voltage drop in R_g; in other words, the full voltage of the bias battery is applied to the grid of tube *B*.

The grid resistor R_g, usually has a rather high value (0.5 to 2 megohms). The reactance of the coupling capacitor, C_c, must be low enough compared with the resistance of R_g so that the ac voltage drop in C_c is negligible at the lowest frequency to be amplified. If R_g is at least 0.5 megohm, a 0.1-μF capacitor will be amply large for the usual range of audio frequencies.

So far as the alternating component of plate voltage is concerned, it will be realized that if the voltage drop in C_c is negligible then R_p and R_g are effectively in parallel (although they are quite separate so far as dc is concerned). The resultant parallel resistance of the two is therefore the actual load resistance for the tube. That is why R_g is made as high in resistance as possible; then it will have the least effect on the load represented by R_p.

The impedance-coupled circuit differs from that using resistance coupling only in the substitution of a high inductance (as high as several hundred henrys) for the plate resistor. The advantage of using an inductor rather than a resistor at this point is that the impedance of the inductor is high for audio frequencies, but its resistance is relatively low. Thus it provides a higher value of load impedance for ac without an excessive dc voltage drop, and consequently the power-supply voltage does not have to be high for effective operation.

The transformer-coupled amplifier uses a transformer with its primary connected in the plate circuit of the tube and its secondary connected to the load (in the circuit shown, a following amplifier). There is no direct connection between the two windings, so the plate voltage on tube *A* is isolated from the grid of tube *B*. The transformer-coupled amplifier has the same advantage as the impedance-coupled circuit with respect to loss of dc voltage from the plate supply. Also, if the secondary has more turns than the primary, the output voltage will be "stepped up" in proportion to the turns ratio.

Resistance coupling is simple, inexpensive, and will give the same amount of amplification — or **voltage gain** — over a wide range of frequencies; it will give substantially the same amplification at any frequency in the audio range, for example. Impedance coupling will give somewhat more gain, with the same tube and same plate-supply voltage, than resistance coupling. However, it is not quite so good over a wide frequency range; it tends to "peak," or give maximum gain, over a comparatively narrow band of frequencies. With a good transformer the gain of a transformer-coupled amplifier can be kept fairly constant over the audio-frequency range. On the other hand, transformer coupling in voltage amplifiers (see below) is best suited to triodes having amplification factors of about 20 or less, for the reason that the primary inductance of a practicable transform-

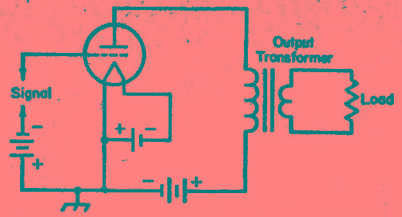


Fig. 3-11 – An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

er cannot be made large enough to work well with a tube having high plate resistance.

Class A Amplifiers

An amplifier in which voltage gain is the primary consideration is called a voltage amplifier. Maximum voltage gain is secured when the load resistance or impedance is made as high as possible in comparison with the plate resistance of the tube. In such a case, the major portion of the voltage generated will appear across the load.

Voltage amplifiers belong to a group called **Class A amplifiers.** A Class A amplifier is one operated so that the wave shape of the output voltage is the same as that of the signal voltage applied to the grid. If a Class A amplifier is biased so that the grid is always negative, even with the largest signal to be handled by the grid, it is called a **Class A$_1$ amplifier.** Voltage amplifiers are always Class A$_1$ amplifiers, and their primary use is in driving a following Class A$_1$ amplifier.

Power Amplifiers

The end result of any amplification is that the amplified signal does some work. For example, an audio-frequency amplifier usually drives a loudspeaker that in turn produces sound waves. The greater the amount of af power supplied to the speaker the louder the sound it will produce.

Fig. 3-11 shows an elementary **power-amplifier** circuit. It is simply a transformer-coupled amplifier with the load connected to the secondary. Although the load is shown as a resistor, it actually would be some device, such as a loudspeaker, that employs the power usefully. Every power tube requires a specific value of load resistance from plate to cathode, usually some thousands of ohms, for optimum operation. The resistance of the actual load is rarely the right value for "matching" this optimum load resistance, so the transformer turns ratio is chosen to reflect the power value of resistance into the primary. The turns ratio may be either step-up or step-down, depending on whether the actual load resistance is higher or lower than the load the tube wants.

The **power-amplification ratio** of an amplifier is the ratio of the power output obtained from the plate circuit to the power required from the ac signal in the grid circuit. There is no power lost in the grid circuit of a Class A$_1$ amplifier, so such an amplifier has an infinitely large power-amplification ratio. However, it is quite possible to operate a Class A amplifier in such a way that current flows in its grid circuit during at least part of the cycle. In such a case power is used up in the grid circuit and the power amplification ratio is not infinite. A tube operated in this fashion is known as a **Class A$_2$ amplifier.** It is necessary to use a power amplifier to drive a Class A$_2$ amplifier, because a voltage amplifier cannot deliver power without serious distortion of the wave shape.

Another term used in connection with power amplifiers is **power sensitivity.** In the case of a Class A$_1$ amplifier, it means the ratio of power output to the grid signal voltage that causes it. If grid current flows, the term usually means the ratio of plate power output to grid power input.

The ac power that is delivered to a load by an amplifier tube has to be paid for in power taken from the source of plate voltage and current. In fact, there is always more power going into the plate circuit of the tube than is coming out as useful output. The difference between the input and output power is used up in heating the plate of the tube, as explained previously. The ratio of useful power output to dc plate input is called the **plate efficiency.** The higher the plate efficiency, the greater the amount of power that can be taken from a tube having a given plate-dissipation rating.

Parallel and Push-Pull

When it is necessary to obtain more power output than one tube is capable of giving, two or more similar tubes may be connected in **parallel.** In this case the similar elements in all tubes are connected together. This method is shown in Fig. 3-12 for a transformer-coupled amplifier. The power output is in proportion to the number of tubes used; the grid signal or exciting voltage required, however, is the same as for one tube.

If the amplifier operates in such a way as to consume power in the grid circuit, the grid power required is in proportion to the number of tubes used.

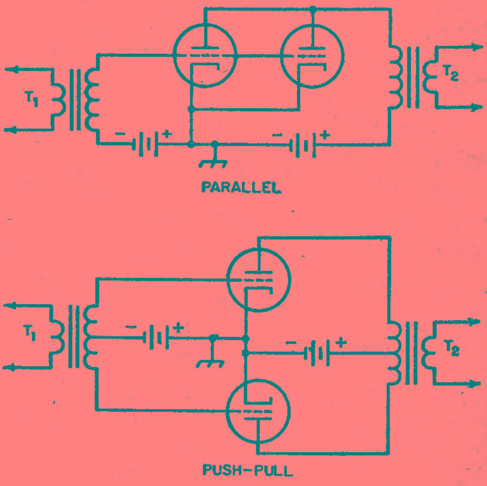


Fig. 3-12 – Parallel and push-pull af amplifier circuits.

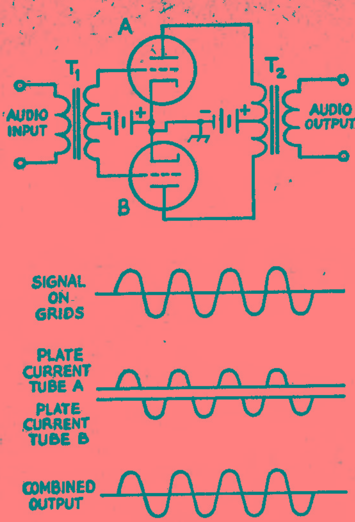


Fig. 3-13 – Class B amplifier operation.

An increase in power output also can be secured by connecting two tubes in **push-pull.** In this case the grids and plates of the two tubes are connected to opposite ends of a balanced circuit as shown in Fig. 3-12. At any instant the ends of the secondary winding of the input transformer, T_1, will be at opposite polarity with respect to the cathode connection, so the grid of one tube is swung positive at the same instant that the grid of the other is swung negative. Hence, in any push-pull-connected amplifier the voltages and currents of one tube are out of phase with those of the other tube.

In push-pull operation the even-harmonic (second, fourth, etc.) distortion is balanced out in the plate circuit. This means that for the same power output the distortion will be less than with parallel operation.

The exciting voltage measured between the two grids must be twice that required for one tube. If the grids consume power, the driving power for the push-pull amplifier is twice that taken by either tube alone.

Cascade Amplifiers

It is readily possible to take the output of one amplifier and apply it as a signal on the grid of a second amplifer, then take the second amplifier's output and apply it to a third, and so on. Each amplifier is called a **stage,** and stages used successively are said to be in cascade.

Class B Amplifiers

Fig. 3-13 shows two tubes connected in a push-pull circuit. If the grid bias is set at the point where (when no signal is applied) the plate current is just cut off, then a signal can cause plate current to flow in either tube only when the signal voltage applied to that particular tube is positive with respect to the cathode. Since in the balanced grid circuit the signal voltages on the grids of the two tubes always have opposite polarities, plate current flows only in one tube at a time.

The graphs show the operation of such an amplifier. The plate current of tube *B* is drawn inverted to show that it flows in the opposite direction, through the primary of the output transformer, to the plate current of tube *A*. Thus each half of the output-transformer primary works alternately to induce a half-cycle of voltage in the secondary. In the secondary of T2, the original waveform is restored. This type of operation is called **Class B amplification.**

The Class B amplifier has considerably higher plate efficiency than the Class A amplifier. Furthermore, the dc plate current of a Class B amplifier is proportional to the signal voltage on the grids, so the power input is small with small signals. The dc plate power input to a Class A amplifier is the same whether the signal is large, small, or absent altogether; therefore the maximum dc plate input that can be applied to a Class A amplifier is equal to the rated plate dissipation of the tube or tubes. Two tubes in a Class B amplifier can deliver approximately twelve times as much audio power as the same two tubes in a Class A amplifier.

A Class B amplifier usually is operated in such a way as to secure the maximum possible power output. This requires rather large values of plate current, and to obtain them the signal voltage must completely overcome the grid bias during at least part of the cycle, so grid current flows and the grid circuit consumes power. While the power requirements are fairly low (as compared with the power output), the fact that the grids are positive during only part of the cycle means that the load on the preceding amplifier or driver stage varies in magnitude during the cycle; the effective load resistance is high when the grids are not drawing current and relatively low when they do take current. This must be allowed for when designing the driver.

Certain types of tubes have been designed specifically for Class B service and can be operated without fixed or other form of grid bias (**zero-bias tubes**). The amplification factor is so high that the plate current is small without signal. Because there is no fixed bias, the grids start drawing current immediately whenever a signal is applied, so the grid-current flow is continuous throughout the cycle. This makes the load on the driver much more constant than is the case with tubes of lower μ biased to plate-current cut off.

Class B amplifiers used at radio frequencies are known as **linear amplifiers** because they are adjusted to operate in such a way that the power output is proportional to the square of the rf exciting voltage. This permits amplification of a modulated rf signal without distortion. Push-pull is not required in this type of operation; a single tube can be used equally well.

Class AB Amplifiers

A **Class AB audio amplifier** is a push-pull amplifier with higher bias than would be normal for pure Class A operation, but less than the cut-off bias required for Class B. At low signal levels the tubes operate as Class A amplifiers, and the plate current is the same with or without signal. At higher signal levels, the plate current of one tube is cut off during part of the negative cycle of the signal applied to its grid, and the plate current of the other tube rises with the signal. The total plate current for the amplifier also rises above the no-signal level with a large signal is applied.

In a properly designed Class AB amplifier the distortion is as low as with a Class A stage, but the efficiency and power output are considerably higher than with pure Class A operation. A Class AB amplifier can be operated either with or without driving the grids into the positive region. A Class AB_1 amplifier is one in which the grids are never positive with respect to the cathode; therefore no driving power is required – only voltage. A Class AB_2 amplifier is one that has grid-current flow during part of the cycle if the applied signal is large; it takes a small amount of driving power. The Class AB_2 amplifier will deliver somewhat more power (using the same tubes) but the Class AB_1 amplifier avoids the problem of designing a driver that will deliver power, without distortion, into a load of highly variable resistance.

Operating Angle

Inspection of Fig. 3-13 shows that either of the two vacuum tubes is working for only half the ac cycle and idling during the other half. It is convenient to describe the amount of time during which plate current flows in terms of electrical degrees. In Fig. 3-13 each tube has "180-degree" excitation, a half-cycle being equal to 180 degrees. The number of degrees during which plate current flows is called the operating angle of the amplifier. From the descriptions given above, it should be clear that a Class A amplifier has 360-degree excitation, because plate current flows during the whole cycle. In a Class AB amplifier the operating angle is between 180 and 360 degrees (in each tube) depending on the particular operating conditions chosen. The greater the amount of negative grid bias, the smaller the operating angle becomes.

An operating angle of less than 180 degrees leads to a considerable amount of distortion, because there is no way for the tube to reproduce even a half-cycle of the signal on its grid. Using two tubes in push-pull, as in Fig. 3-13, would merely put together two distorted half-cycles. An operating angle of less than 180 degrees therefore cannot be used if distortionless output is wanted.

Class C Amplifiers

In power amplifiers operating at radio frequencies distortion of the rf wave form is relatively unimportant. For reasons described later in this chapter, an rf amplifier must be operated with tuned circuits, and the selectivity of such circuits "filters out" the rf harmonics resulting from distortion.

A radio-frequency power amplifier therefore can be used with an operating angle of less than 180 degrees. This is called Class C operation. The advantage is that the plate efficiency is increased, because the loss in the plate is proportional, among other things, to the amount of time during which the plate current flows, and this time is reduced by decreasing the operating angle.

Depending on the type of tube, the optimum load resistance for a Class C amplifier ranges from about 1500 to 5000 ohms. It is usually secured by using tuned-circuit arrangements, of the type described in the chapter on circuit fundamentals, to transform the resistance of the actual load to the value required by the tube. The grid is driven well into the positive region, so that grid current flows and power is consumed in the grid circuit. The smaller the operating angle, the greater the driving voltage and the larger the grid driving power required to develop full output in the load resistance. The best compromise between driving power, plate efficiency, and power output usually results when the minimum plate voltage (at the peak of the driving cycle, when the plate current reaches its highest value) is just equal to the peak positive grid voltage. Under these conditions the operating angle is usually between 120 and 150 degrees and the plate efficiency lies in the range of 60 to 80 percent. While higher plate efficiencies are possible, attaining them requires excessive driving power and grid bias, together with higher plate voltage than is "normal" for the particular tube type.

With proper design and adjustment, a Class C amplifier can be made to operate in such a way that the power input and output are proportional to the square of the applied plate voltage. This is an important consideration when the amplifier is to be plate-modulated for radiotelephony, as described in the chapter on amplitude modulation.

FEEDBACK

It is possible to take a part of the amplified energy in the plate circuit of an amplifier and insert it into the grid circuit. When this is done the amplifier is said to have **feedback.**

If the voltage that is inserted in the grid circuit is 180 degrees out of phase with the signal voltage acting on the grid, the feedback is called **negative,** or **degenerative.** On the other hand, if the voltage is fed back in phase with the grid signal, the feedback is called **positive,** or **regenerative.**

Negative Feedback

With negative feedback the voltage that is fed back opposes the signal voltage. This decreases the amplitude of the voltage acting between the grid and cathode and thus has the effect of reducing the voltage amplification. That is, a larger exciting voltage is required for obtaining the same output voltage from the plate circuit.

The greater the amount of negative feedback (when properly applied) the more independent the amplification becomes of tube characteristics and circuit conditions. This tends to make the frequency-response characteristic of the amplifier flat – that is, the amplification tends to be the same at all frequencies within the range for which the amplifier is designed. Also, any distortion generated in the plate circuit of the tube tends to

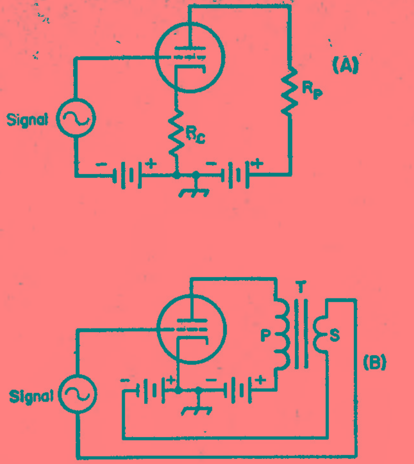


Fig. 3-14 – Simple circuits for producing feedback.

"buck itself out." Amplifiers with negative feedback are therefore comparatively free from harmonic distortion. These advantages are worth while if the amplifier otherwise has enough voltage gain for its intended use.

In the circuit shown at A in Fig. 3-14 resistor R_c is in series with the regular plate resistor, R_p and thus is a part of the load for the tube. Therefore, part of the output voltage will appear across R_c. However, R_c also is connected in series with the grid circuit, and so the output voltage that appears across R_c is in series with the signal voltage. The output voltage across R_c opposes the signal voltage, so the actual ac voltage between the grid and cathode is equal to the *difference* between the two voltages.

The circuit shown At B in Fig. 3-14 can be used to give either negative or positive feedback. The secondary of a transformer is connected back into the grid circuit to insert a desired amount of feedback voltage. Reversing the terminals of either transformer winding (but not both simultaneously) will reverse the phase.

Positive Feedback

Positive feedback increases the amplification because the feedback voltage adds to the original signal voltage and the resulting larger voltage on the grid causes a larger output voltage. The amplification tends to be greatest at one frequency (which depends upon the particular circuit arrangement) and harmonic distortion is increased. If enough energy is fed back, a self-sustaining oscillation – in which energy at essentially one frequency is generated by the tube itself – will be set up. In such case all the signal voltage on the grid can be supplied from the plate circuit; no external signal is needed because any small irregularity in the plate current – and there are always some irregularities – will be amplified and thus give the oscillation an opportunity to build up. Positive feedback finds a major application in such "oscillators," and in addition is used for selective amplification at both audio and radio frequencies, the feedback being kept below the value that causes self-oscillation.

INTERELECTRODE CAPACITANCES

Each pair of elements in a tube forms a small capacitor "plate." There are three such capacitances in a triode – that between the grid and cathode, that between the grid and plate, and that between the plate and cathode. The capacitances are very small – only a few picofarads at most – but they frequently have a very pronounced effect on the operation of an amplifier circuit.

Input Capacitance

It was explained previously that the ac grid voltage and ac plate voltage of an amplifier having a resistive load are 180 degrees out of phase, using the cathode of the tube as a reference point. However, these two voltages are *in* phase going around the circuit from plate to grid as shown in Fig. 3-15. This means that their sum is acting between the grid and plate; that is, across the grid-plate capacitance of the tube.

As a result, a capacitive current flows around the circuit, its amplitude being directly proportional to the sum of the ac grid and plate voltages and to the grid-plate capacitance. The source of the grid signal must furnish this amount of current, in addition to the capacitive current that flows in the grid-cathode capacitance. Hence the signal source "sees" an effective capacitance that is larger than the grid-cathode capacitance. This is known as the Miller Effect.

The greater the voltage amplification the greater the effective input capacitance. The input capacitance of a resistance-coupled amplifier is given by the formula

$$C_{input} = C_{gk} + C_{gp} (A + 1)$$

where C_{gk} is the grid-to-cathode capacitance, C_{gp} is the grid-to-plate capacitance, and A is the voltage amplification. The input capacitance may be as much as several hundred picofarads when the voltage amplification is large, even though the interelectrode capacitances are quite small.

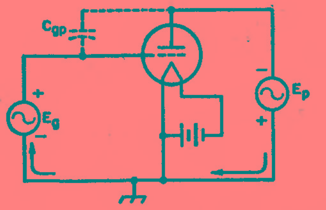


Fig. 3-15 – The ac voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

Output Capacitance

The principal component of the output capacitance of an amplifier is the actual plate-to-cathode capacitance of the tube. The output capacitance usually need not be considered in audio amplifiers, but becomes of importance at radio frequencies.

Tube Capacitance at RF

At radio frequencies the reactances of even very small interelectrode capacitances drop to very low values. A resistance-coupled amplifier gives very little amplification at rf, for example, because the reactances of the interelectrode "capacitors" are so low that they practically short-circuit the input and output circuits and thus the tube is unable to amplify. This is overcome at radio frequencies by using tuned circuits for the grid and plate, making the tube capacitances part of the tuning capacitances. In this way the circuits can have the high resistive impedances necessary for satisfactory amplification.

The grid-plate capacitance is important at radio frequencies because its reactance, relatively low at rf, offers a path over which energy can be fed back from the plate to the grid. In practically every case the feedback is in the right phase and of sufficient amplitude to cause self-oscillation, so the circuit becomes useless as an amplifier.

Special "neutralizing" circuits can be used to prevent feedback but they are, in general, not too satisfactory when used in radio receivers. They are, however, used in transmitters.

SCREEN-GRID TUBES

The grid-plate capacitance can be reduced to a negligible value by inserting a second grid between the control grid and the plate, as indicated in Fig. 3-16. The second grid, called the screen grid, acts as an electrostatic shield to prevent capacitive coupling between the control grid and plate. It is made in the form of a grid or coarse screen so that electrons can pass through it.

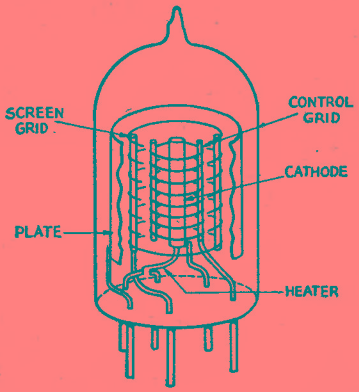


Fig. 3-16 — Representative arrangement of elements in a screen-grid tatrode, with part of plate and screen cut away. This is "single-ended" construction with a button base, typical of miniature receiving tubes. To reduce capacitance between control grid and plata the leads from these elements are brought out at opposite sides; actual tubes probably would have additional shielding between these leads.

Because of the shielding action of the screen grid, the positively charged plate cannot attract electrons from the cathode as it does in a triode. In order to get electrons to the plate, it is necessary to apply a positive voltage (with respect to the cathode) to the screen. The screen then attracts electrons much as does the plate in a triode tube. In traveling toward the screen the electrons acquire such velocity that most of them shoot between the screen wires and then are attracted to the plate. A certain proportion do strike the screen, however, with the result that some current also flows in the screen-grid circuit.

To be a good shield, the screen grid must be connected to the cathode through a circuit that has low impedance at the frequency being amplified. A bypass capacitor from screen grid to cathode, having a reactance of not more than a few hundred ohms, is generally used.

A tube having a cathode, control grid, screen grid and plate (four elements) is called a **tetrode**.

Pentodes

When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons which "splash" from the plate into the interelement space. This is called **secondary emission**. In a triode the negative grid repels the secondary electrons back into the plate and they cause no disturbance. In the screen-grid tube, however, the positively charged screen attracts the secondary electrons, causing a reverse current to flow between screen and plate.

To overcome the effects of secondary emission, a third grid, called the **suppressor grid**, may be inserted between the screen and plate. This grid acts as a shield between the screen grid and plate so the secondary electrons cannot be attracted by the screen grid. They are hence attracted back to the plate without appreciably obstructing the regular plate-current flow. A five-element tube of this type is called a **pentode**.

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, the control grid still can control the plate current in essentially the same way that it does in a triode. Consequently, the grid-plate transconductance (or mutual conductance) of a tetrode or pentode will be of the same order of value as in a triode of corresponding structure. On the other hand, since a change in plate voltage has very little effect on the plate-current flow, both the amplification factor and plate resistance of a pentode or tetrode are very high. In small receiving pentodes the amplification factor is of the order of 1000 or higher, while the plate resistance may be from 0.5 to 1 or more megohms. Because of the high plate resistance, the actual voltage amplification possible with a pentode is very much less than the large amplification factor might indicate. A voltage gain in the vicinity of 50 to 200 is typical of a pentode stage.

In practical screen-grid tubes the grid-plate capacitance is only a small fraction of a picofarad. This capacitance is too small to cause an appreciable increase in input capacitance as described in the preceding section, so the input capacitance of a screen-grid tube is equal to the capacitance between the plate and screen.

In addition to their applications as radio-frequency amplifiers, pentodes or tetrodes also are used for audio-frequency power amplification. In tubes designed for this purpose the chief function of the screen is to serve as an accelerator of the electrons, so that large values of plate current can be drawn at relatively low plate voltages. Such tubes have quite high power sensitivity compared with triodes of the same power output, although harmonic distortion is somewhat greater.

Beam Tubes

A beam tetrode is a four-element screen-grid tube constructed in such a way that the electrons are formed into concentrated beams on their way to the plate. Additional design features overcome the effects of secondary emission so that a suppressor grid is not needed. The "beam" construction makes it possible to draw large plate currents at relatively low plate voltages, and increases the power sensitivity.

For power amplification at both audio and radio frequencies beam tetrodes have largely supplanted the non beam types because large power outputs can be secured with very small amounts of grid driving power.

Variable-μ Tubes

The mutual conductance of a vacuum tube decreases when its grid bias is made more negative, assuming that the other electrode voltages are held constant. Since the mutual conductance controls the amount of amplification, it is possible to adjust the gain of the amplifier by adjusting the grid bias. This method of gain control is universally used in radio-frequency amplifiers designed for receivers.

The ordinary type of tube has what is known as a sharp-cutoff characteristic. The mutual conductance decreases at a uniform rate as the negative bias is increased. The amount of signal voltage that such a tube can handle without causing distortion is not sufficient to take care of very strong signals. To overcome this, some tubes are made with a variable-μ characteristic – that is, the amplification factor decreases with increasing grid bias. The variable-μ tube can handle a much larger signal than the sharp-cutoff type before the signal swings either beyond the zero grid-bias point or the plate-current cutoff point.

INPUT AND OUTPUT IMPEDANCES

The input impedance of a vacuum-tube amplifier is the impedance "seen" by the signal source when connected to the input terminals of the amplifier. In the types of amplifiers previously discussed, the input impedance is the impedance measured between the grid and cathode of the tube with operating voltages applied. At audio frequencies the input impedance of a Class A_1 amplifier is for all practical purposes the input impedance of the stage. If the tube is driven into the grid-current region there is in addition a resistance component in the input impedance, the resistance having an average value equal to E^2/P, where E is the rms driving voltage and P is the power in watts consumed in the grid. The resistance usually will vary during the ac cycle because grid current may flow only during part of the cycle; also, the grid-voltage/grid-current characteristic is seldom linear.

The output impedance of amplifiers of this type consists of the plate resistance of the tube shunted by the output capacitance.

At radio frequencies, when tuned circuits are employed, the input and output impedances are usually pure resistances; any reactive components are "tuned out" in the process of adjusting the circuits to resonance at the operating frequency.

OTHER TYPES OF AMPLIFIERS

In the amplifier circuits so far discussed, the signal has been applied between the grid and cathode and the amplified output has been taken from the plate-to-cathode circuit. That is, the cathode has been the meeting point for the input and output circuits. However, it is possible to use any one of the three principal elements as the common point. This leads to two additional kinds of amplifiers, commonly called the **grounded-grid amplifier** (or **grid-separation** circuit) and the **cathode follower**.

These two circuits are shown in simplified form in Fig. 3-17. In both circuits the resistor R represents the load into which the amplifier works; the actual load may be resistance-capacitance-coupled, transformer-coupled, may be a tuned circuit if the amplifier operates at radio frequencies, and so on. Also, in both circuits the batteries that supply grid bias and plate power are assumed to have such negligible impedance that they do not enter into the operation of the circuits.

Grounded-Grid Amplifier

In the grounded-grid amplifier the input signal is applied between the cathode and grid, and the

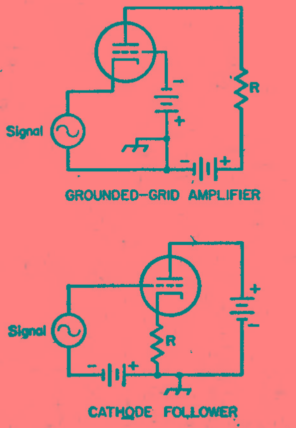


Fig. 3-17 – In the upper circuit, the grid is the junction point between the input and output circuits in the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, R, and may be coupled to a following amplifier by the usual methods.

output is taken between the plate and grid. The grid is thus the common element. The ac component of the plate current has to flow through the signal source to reach the cathode. The source of signal is in series with the load through the plate-to-cathode resistance of the tube, so some of the power in the load is supplied by the signal source. In transmitting applications this fed-through power is of the order of 10 percent of the total power output, using tubes suitable for grounded-grid service.

The input impedance of the grounded-grid amplifier consists of a capacitance in parallel with an equivalent resistance representing the power furnished by the driving source of the grid and to the load. This resistance is of the order of a few hundred ohms. The output impedance, neglecting the interelectrode capacitances, is equal to the plate resistance of the tube. This is the same as in the case of the grounded-cathode amplifier.

The grounded-grid amplifier is widely used at vhf and uhf, where the more conventional amplifier circuit fails to work properly. With a triode tube designed for this type of operation, an rf amplifier can be built that is free from the type of feedback that causes oscillation. This requires that the grid act as a shield between the cathode and plate, reducing the plate-cathode capacitance to a very low value.

Cathode Follower

The cathode follower uses the plate of the tube as the common element. The input signal is applied between the grid and plate (assuming negligible impedance in the batteries) and the output is taken between cathode and plate. This circuit is degenerative; in fact, all of the output voltage is fed back into the input circuit out of phase with the grid signal. The input signal therefore has to be larger than the output voltage; that is, the cathode follower gives a loss in voltage, although it gives the same power gain as other circuits under equivalent operating conditions.

An important feature of the cathode follower is its low output impedance, which is given by the formula (neglecting interelectrode capacitances)

$$Z_{out} = \frac{r_p}{1 + \mu}$$

where r_p is the tube plate resistance and μ is the amplification factor. Low output impedance is a valuable characteristic in an amplifier designed to cover a wide band of frequencies. In addition, the input capacitance is only a fraction of the grid-to-cathode capacitance of the tube, a feature of further benefit in a wide-band amplifier. The cathode follower is useful as a step-down impedance transformer, since the input impedance is high and the output impedance is low.

CATHODE CIRCUITS AND GRID BIAS

Most of the equipment used by amateurs is powered by the ac line. This includes the filaments or heaters of vacuum tubes. Although supplies for the plate (and sometimes the grid) are usually rectified and filtered to give pure dc – that is, direct current that is constant and without a superimposed ac component – the relatively large currents required by filaments and heaters usually make a rectifier-type dc supply impracticable.

Filament Hum

Alternating current is just as good as direct current from the heating standpoint, but some of the ac voltage is likely to get on the grid and cause a low-pitched "ac hum" to be superimposed on the output.

Hum troubles are worst with directly-heated cathodes or filaments, because with such cathodes there has to be a direct connection between the source of heating power and the rest of the circuit. The hum can be minimized by either of the connections shown in Fig. 3-18. In both cases the grid- and plate-return circuits are connected to the electrical midpoint (center tap) of the filament supply. Thus, so far as the grid and plate are concerned, the voltage and current on one side of the filament are balanced by an equal and opposite voltage and current on the other side. The balance is never quite perfect, however, so filament-type tubes are never completely hum-free. For this reason directly-heated filaments are employed for the most part in power tubes, where the hum introduced is extremely small in comparison with the power-output level.

With indirectly heated cathodes the chief problem is the magnetic field set up by the heater. Occasionally, also, there is leakage between the heater and cathode, allowing a small ac voltage to get to the grid. If hum appears, grounding one side of the heater supply usually will help to reduce it, although sometimes better results are obtained if the heater supply is center-tapped and the center-tap grounded, as in Fig. 3-18.

Cathode Bias

In the simplified amplifier circuits discussed in this chapter, grid bias has been supplied by a battery. However, in equipment that operates from the power line, cathode bias is almost universally used for tubes that are operated in Class A (constant dc input).

The cathode-bias method uses a resistor (cathode resistor) connected in series with the cathode,

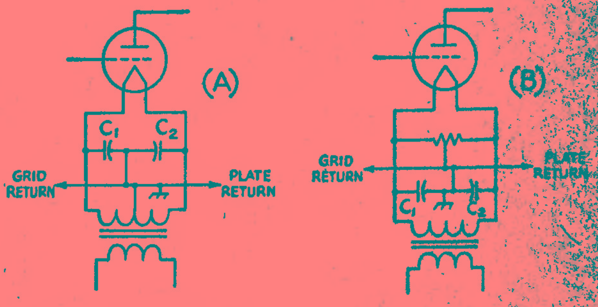


Fig. 3-18 – Filament center-tapping methods for use with directly heated tubes.

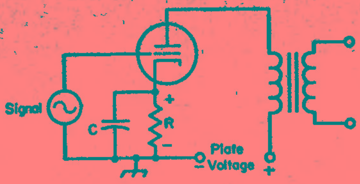


Fig. 3-19 — Cathode biasing. *R* is the cathode resistor and *C* is the cathode bypass capacitor.

as shown at *R* in Fig. 3-19. The direction of plate-current flow is such that the end of the resistor nearest the cathode is positive. The voltage drop across *R* therefore places a *negative* voltage on the grid. This negative bias is obtained from the steady dc plate current.

If the alternating component of plate current flows through *R* when the tube is amplifying, the voltage drop caused by the ac will be degenerative (note the similarity between this circuit and that of Fig. 3-14A). To prevent this the resistor is bypassed by a capacitor, *C*, that has very low reactance compared with the resistance of *R*. Depending on the type of tube and the particular kind of operation, *R* may be between about 100 and 3000 ohms. For good bypassing at the low audio frequencies, *C* should be 10 to 50 microfarads (electrolytic capacitors are used for this purpose). At radio frequencies, capacitances of about 100 pF to 0.1 μF are used; the small values are sufficient at very high frequencies and the largest at low and medium frequencies. In the range 3 to 30 megahertz a capacitance of .01 μF is satisfactory.

The value of cathode resistor for an amplifier having negligible dc resistance in its plate circuit (transformer or impedance coupled) can easily be calculated from the known operating conditions of the tube. The proper grid bias and plate current always are specified by the manufacturer. Knowing these, the required resistance can be found by applying Ohm's Law.

Example: It is found from tube tables that the tube to be used should have a negative grid bias of S volts and that at this bias the plate current will be 12 milliamperes (0.012 amp). The required cathode resistance is then

$$R = \frac{E}{I} = \frac{8}{012} = 667 \text{ ohms}$$

The nearest standard value, 680 ohms, would be close enough. The power used in the resistor is

$$P = EI = 8 \times .012 = 0.096 \text{ watt}$$

A 1/4-watt or 1/2-watt resistor would have ample rating.

The current that flows through *R* is the total cathode current. In an ordinary triode amplifier this is the same as the plate current, but in a screen-grid tube the cathode current is the sum of the plate and screen currents. Hence these two currents must be added when calculating the value of cathode resistor required for a screen-grid tube.

Example: A receiving pentode requires 3 volts negative bias. At this bias and the recommended plate and screen voltages, its plate current is 9 mA and its screen current is 2 mA. The cathode current is therefore 11 mA (0.011 amp). The required resistance is

$$R = \frac{E}{I} = \frac{3}{.011} = 272 \text{ ohms}$$

A 270-ohm resistor would be satisfactory. The power in the resistor is

$$P = EI = 3 \times 0.011 = .033 \text{ watt}$$

The cathode-resistor method of biasing is self-regulating, because if the tube characteristics vary slightly from the published values (as they do in practice) the bias will increase if the plate current is slightly high, or decrease if it is slightly low. This tends to hold the plate current at the proper value.

Calculation of the cathode resistor for a resistance-coupled amplifier is ordinarily not practicable by the method described above, because the plate current in such an amplifier is usually much smaller than the rated value given in the tube tables. However, representative data for the tubes commonly used as resistance-coupled amplifiers are given in the chapter on audio amplifiers, including cathode-resistor values.

"Contact Potential" Bias

In the absence of any negative bias voltage on the grid of a tube, some of the electrons in the space charge will have enough velocity to reach the grid. This causes a small current (of the order of microamperes) to flow in the external circuit between the grid and cathode. If the current is made to flow through a high resistance – a megohm or so – the resulting voltage drop in the resistor will give the grid a negative bias of the order of one volt. The bias so obtained is called contact-potential bias.

Contact-potential bias can be used to advantage in circuits operating at low signal levels (less than one volt peak) since it eliminates the cathode-bias resistor and bypass capacitor. It is principally used in low-level resistance-coupled audio amplifiers. The bias resistor is connected directly between grid and cathode, and must be isolated from the signal source by a blocking capacitor.

Screen Supply

In practical circuits using tetrodes and pentodes the voltage for the screen frequently is taken from the plate supply through a resistor. A typical circuit for an rf amplifier is shown in Fig. 3-20. Resistor *R* is the screen dropping resistor, and *C* is the screen bypass capacitor. In flowing through *R*, the screen current causes a voltage drop in *R* that reduces the plate-supply voltage to the proper value for the screen. When the plate-supply voltage and the screen current are known, the value of *R* can be caluclated from Ohm's Law.

Example: An rf receiving pentode has a rated screen current of 2 milliamperes (0.002 amp) at normal operating conditions. The rated screen voltage is 100 volts, and the plate supply gives 250 volts. To put 100 volts on the screen, the drop across *R* must be equal to the difference between the plate-supply voltage and the screen voltage; that is, 250 − 100 = 150 volts. Then

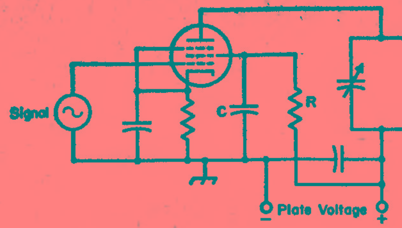


Fig. 3-20 — Screen-voltage supply for a pentode tube through a dropping resistor, *R*. The screen bypass capacitor, *C*, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

$$R = \frac{E}{I} = \frac{150}{.002} = 75{,}000 \text{ ohms}$$

The power to be dissipated in the resistor is

$$P = EI = 150 \times .002 = 0.3 \text{ watt}$$

A 1/2- or 1-watt resistor would be satisfactory.

The reactance of the screen bypass capacitor, *C*, should be low compared with the screen-to-cathode impedance. For radio-frequency applications a capacitance in the vicinity of .01 μF is amply large.

In some vacuum-tube circuits the screen voltage is obtained from a voltage divider connected across the plate supply. The design of voltage dividers is discussed at length elsewhere in this book.

OSCILLATORS

It was mentioned earlier that if there is enough positive feedback in an amplifier circuit, self-sustaining oscillations will be set up. When an amplifier is arranged so that this condition exists it is called an oscillator.

Oscillations normally take place at only one frequency, and a desired frequency of oscillation can be obtained by using a resonant circuit tuned to that frequency. For example, in Fig. 3-21A the circuit *LC* is tuned to the desired frequency of oscillation. The cathode of the tube is connected to a tap on coil *L* and the grid and plate are connected to opposite ends of the tuned circuit. When an rf current flows in the tuned circuit there is a voltage drop across *L* that increases progressively along the turns. Thus the point at which the tap is connected will be at an intermediate potential with respect to the two ends of the coil. The amplified current in the plate circuit, which flows through the bottom section of *L*, is in phase with the current already flowing in the circuit and thus in the proper relationship for positive feedback.

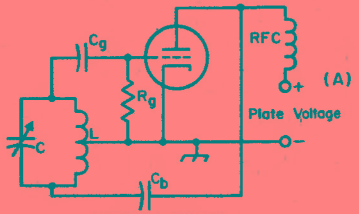
The amount of feedback depends on the position of the tap. If the tap is too near the grid end the voltage drop between grid and cathode is too small to give enough feedback to sustain oscillation, and if it is too near, the plate end of the impedance between the cathode and plate is too small to permit good amplification. Maximum feedback usually is obtained when the tap is somewhere near the center of the coil.

The circuit of Fig. 3-21A is parallel-fed, C_b being the blocking capacitor. The value of C_b is not critical so long as its reactance is low (not more than a few hundred ohms) at the operating frequency.

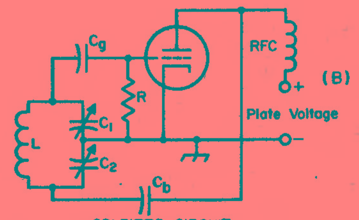
Capacitor C_g is the grid capacitor. It and R_g (the grid leak) are used for the purpose of obtaining grid bias for the tube. In most oscillator circuits the tube generates its own bias. During the part of the cycle when the grid is positive with respect to the cathode, it attracts electrons. These electrons cannot flow through *L* back to the cathode because C_g "blocks" direct current. They therefore have to flow or "leak" through R_g to cathode, and in doing so cause a voltage drop in R_g that places a negative bias on the grid. The amount of bias so developed is equal to the grid current multiplied by the resistance of R_g (Ohm's Law). The value of grid-leak resistance required depends upon the kind of tube used and the purpose for which the oscillator is intended. Values range all the way from a few thousand to several hundred thousand ohms. The capacitance of C_g should be large enough to have low reactance (a few hundred ohms) at the operating frequency.

The circuit shown at B in Fig. 3-21 uses the voltage drops across two capacitors in series in the tuned circuit to supply the feedback. Other than this, the operation is the same as just described. The feedback can be varied by varying the ratio of the reactance of C1 and C2 (that is, by varying the ratio of their capacitances).

Another type of oscillator, called the tuned-plate tuned-grid circuit, is shown in Fig. 3-22. Resonant circuits tuned approximately to the same frequency are connected between grid and cathode and between plate and cathode. The two coils, L1



HARTLEY CIRCUIT



COLPITTS CIRCUIT

Fig. 3-21 — Basic oscillator circuits. Feedback voltage is obteined by tapping the grid and cathode across a portion of the tuned circuit. In the Hartley circuit the tep is on the coil, but in the Colpitts circuit the voltage is obtained from the drop across a capacitor.

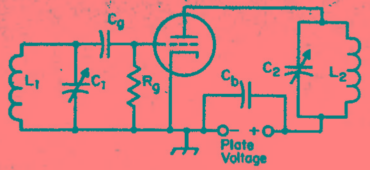


Fig. 3-22 – The tuned-plate tuned-grid oscillator.

and L2, are not magnetically coupled. The feedback is through the grid-plate capacitance of the tube, and will be in the right phase to be positive when the plate circuit, C2-L2, is tuned to a slightly higher frequency than the grid circuit, L1-C1. The amount of feedback can be adjusted by varying the tuning of either circuit. The frequency of oscillation is determined by the tuned circuit that has the higher *Q*. The grid leak and grid capacitor have the same functions as in the other circuits. In this case it is convenient to use series feed for the plate circuit, so C_b is a bypass capacitor to guide the rf current around the plate supply.

There are many oscillator circuits (examples of others will be found in later chapters) but the basic feature of all of them is that there is positive feedback in the proper amplitude and phase to sustain oscillation.

Oscillator Operating Characteristics

When an oscillator is delivering power to a load, the adjustment for proper feedback will depend on how heavily the oscillator is loaded – that is, how much power is being taken from the circuit. If the feedback is not large enough – grid excitation too small – a small increase in load may tend to throw the circuit out of oscillation. On the other hand, too much feedback will make the grid current excessively higher, with the result that the power loss in the grid circuit becomes larger than necessary. Since the oscillator itself supplies this grid power, excessive feedback lowers the over-all efficiency because whatever power is used in the grid circuit is not available as useful output.

One of the most important considerations in oscillator design is frequency stability. The principal factors that cause a change in frequency are (1) temperature, (2) plate voltage, (3) loading, (4) mechanical variations of circuit elements. Temperature changes will cause vacuum-tube elements to expand or contract slightly, thus causing variations in the interelectrode capacitances. Since these are unavoidably part of the tuned circuit, the frequency will change correspondingly. Temperature changes in the coil or the tuning capacitor will alter the inductance or capacitance slightly, again causing a shift in the resonant frequency. These effects are relatively slow in operation, and the frequency change caused by them is called drift.

A change in plate voltage usually will cause the frequency to change a small amount, an effect called dynamic instability. Dynamic instability can be reduced by using a tuned circuit of high effective *Q*. The energy taken from the circuit to supply grid losses, as well as energy supplied to a load, represent an increase in the effective resistance of the tuned circuit and thus lower its *Q*. For highest stability, therefore, the coupling between the tuned circuit and the tube and load must be kept as loose as possible. Preferably, the oscillator should not be required to deliver power to an external circuit, and a high value of grid leak resistance should be used since this helps to raise the tube grid and plate resistances as seen by the tuned circuit. Loose coupling can be effected in a variety of ways – one, for example, is by "tapping down" on the tank for the connections to the grid and plate. This is done in the "series-tuned" Colpitts circuit widely used in variable-frequency oscillators for amateur transmitters and described in a later chapter. Alternatively, the *L/C* ratio may be made as small as possible while sustaining stable oscillations (high C) with the grid and plate connected to the ends of the circuit as shown in Figs. 3-21 and 3-22. Using relatively high plate voltage and low plate current also is desirable.

In general, dynamic stability will be at maximum when the feedback is adjusted to the least value that permits reliable oscillation. The use of a tube having a high value of transconductance is desirable, since the higher the transconductance the looser the permissible coupling to the tuned circuit and the smaller the feedback required.

Load variations act in much the same way as plate-voltage variations. A temperature change in the load may also result in drift.

Mechanical variations, usually caused by vibrations, cause changes in inductance and/or capacitance that in turn cause the frequency to "wobble" in step with the vibration.

Methods of minimizing frequency variations in oscillators are taken up in detail in later chapters.

Ground Point

In the oscillator circuits shown in Figs. 3-21 and 3-22 the cathode is connected to ground. It is not actually essential that the radio-frequency circuit should be grounded at the cathode; in fact, there are many times when an *rf* ground on some other point in the circuit is desirable. The rf ground can be placed at any point so long as proper provisions are made for feeding the supply voltages to the tube elements.

Fig. 3-23 shows the Hartley circuit with the plate end of the circuit grounded. The cathode and control grid are "above ground," so far as the rf is concerned. An advantage of such a circuit is that the frame of the tuning capacitor can be grounded. The Colpitts circuit can also be used with the plate

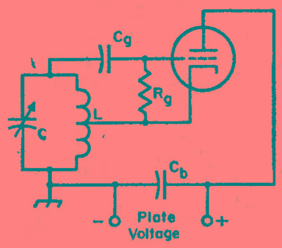


Fig. 3-23 – Showing how the plete may be grounded for rf in a typical oscillator circuit (Hartley).

grounded and the cathode above ground; it is only necessary to feed the dc to the cathode through an rf choke.

A tetrode or pentode tube can be used in any of the popular oscillator circuits. A common variation is to use the screen grid of the tube as the anode for the Hartley or Colpitts oscillator circuit. It is usually used in the grounded anode circuit, and the plate circuit of the tube is tuned to the second harmonic of the oscillator frequency.

VHF AND MICROWAVE TUBES

Until now, it has been assumed that the time it takes for the electrons to travel from the cathode to the plate does not affect the performance of vacuum-tube operation. As the frequency of operation is raised, this time, called the **transit time,** becomes increasingly important. The transit time depends upon the voltage from the cathode to the plate and the spacing between them. The higher the voltage and the smaller the spacing, the shorter the transit time. This is why tubes designed for vhf and uhf work have very small interelectrode spacings. However, the power handling capabilities also get smaller as the spacing decreases so there is a limit above which ordinary triode and pentode tubes cannot be operated efficiently.

Many different tubes have been developed which actually use transit-time effects to an advantage. **Velocity modulation** of the electron stream in a **klystron** is one example. A small voltage applied across the gap in a re-entrant cavity resonator either retards or accelerates an electron stream by means of the resultant electric field. Initially, all the electrons are traveling at the same velocity and the current in the beam is uniform. After the velocity fluctuations are impressed on the beam, the current is still uniform for awhile but then the electrons that were accelerated begin to catch up with the slower ones that passed through when the field was zero. The latter are also catching up with ones that passed through the gap earlier but were retarded. The result is that the current in the beam is no longer uniform but consists of a series of pulses. If the beam now passes through another cavity gap, a current will be induced in the cavity walls and an electric field also will be set up across the gap. If the phase of the electric field is right, the electron pulses or "bunches" pass through the gap and are retarded, thus giving up energy to the electric field. When the electric field reverses, it would normally accelerate the same number of electrons and give back the energy, but fewer electrons now pass through the gap and the energy given up is less. Thus, a net flow of energy is from the beam to the cavity. If the voltage produced across the output cavity is greater than that across the input cavity, amplification results (assuming the two cavity impedances are the same).

The type of klystron that amateurs are most likely to use is the reflex klystron oscillator. Here, the input and output cavity are the same. The electron stream makes one pass through, becomes velocity modulated, and is turned around by the negative charge on an element called the repeller. During the second pass through, the stream is now bunched and delivers some of its energy to the cavity. The dissipated beam is then picked up by the cavity walls and the circuit is completed. This is shown pictorially in Fig. 3-24.

Klystrons either have cavities external to the vacuum part of the tube or built in as an integral part of the tube structure. The 723 reflex klystron is of the latter type, and along with similar types can be purchased surplus. These tubes were used as local oscillators in radar receivers and can be used for the same purpose in amateur applications. They also may be used in low power transmitters.

Along with a heater supply (usually 6.3 volts), two other voltages are necessary for the operation of the reflex klystron. This is shown in Fig. 3-24. Vc is typically 300 volts dc and Vr will vary from 100 to 150 volts dc. The loaded *Q* of the reflex klystron cavity is quite low and oscillations will occur at different frequencies for various values of Vr. This can be used to advantage and either frequency modulation or automatic frequency control (afc) can be applied to the klystron by means of changes in Vr. As the repeller voltage is made more negative, it will be found that oscillations will occur, increase in amplitude, and then drop out. This will be repeated as the voltage is increased and each time the maximum amplitude of the output power will be less. However, the frequency range covered by each different set of oscillation conditions is approximately the same.

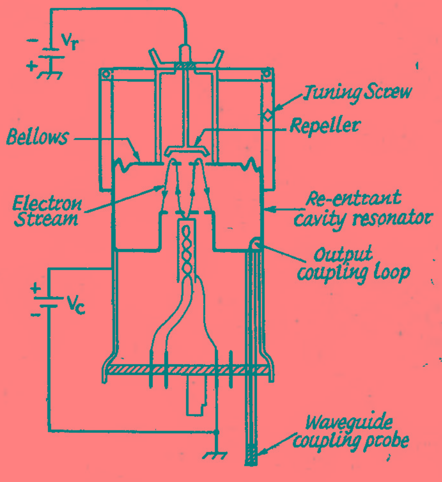


Fig. 3-24 — Cross-sectional view of a typical reflex klystron. The frequency of the cavity resonator is changed by varying the spacing between the grids using a tuning mechanism and a flexible bellows. Modification of this system may be necessary to get certain surplus klystrons into an appropriate amateur bend.

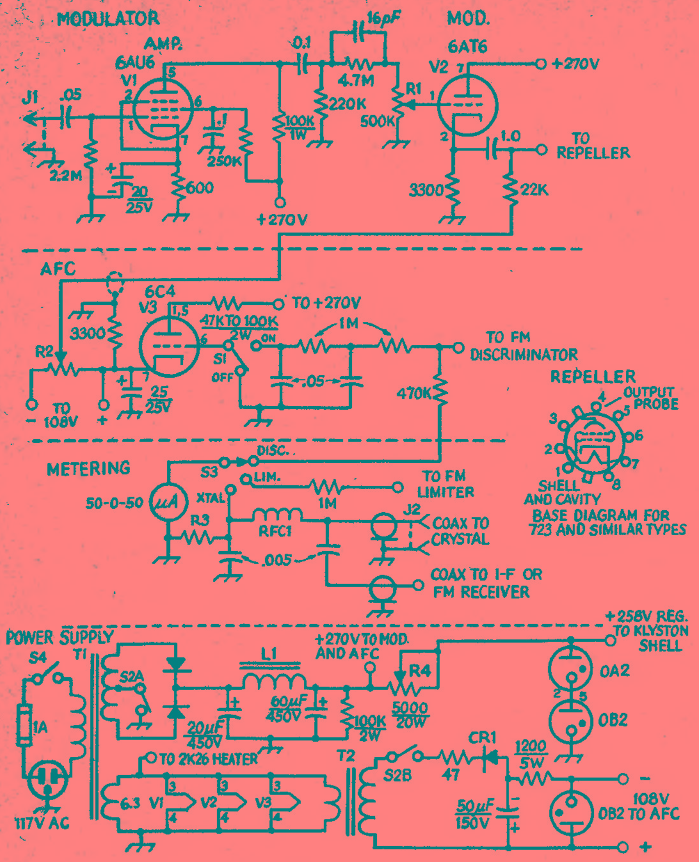


Fig. 3-25 – Schematic diegram and parts information for a power supply and control unit suitable for amateur microwave transceivers. Unless otherwise specified, capacitor values are in μF and resistors are 1/2-watt composition.

CR1 – 1000 PRV, 1-A.
F1 – 1-A fuse and holder.
J1 – Shielded microphone jack.
J2 – Coaxial chassis fitting.
L1 – 10-H 110-mA choke (Stancor C-1001).
R1 – 0.5-megohm potentiometar, audio taper.
R2 – 0.2-meghom potentiometer, carbon, linear taper.
R3 – Meter shunt; value to suit meter used, for 1-mA range.
R4 – 5000 ohms, 20 watts, with slider.
RFC1 – 15 turns No. 24 enamel on 1/2-inch form. (Any rf choke for 30 to 100 MHz is suitable.)
S1 – Toggle switch.
S2 – Toggle switch.
S3 – Single-pole 3-position wafer switch.
S4 – Toggle switch.
T1 – 270-0-270 volts at 70 mA min., 5 volts, 3 A. 6.3 volts, 3.5 A (Stancor PC-8405).
T2 – 6.3 volts, 1.2 A (Stancor P-6134).

Practical metering, afc, modulator, and power supply circuit diagrams are shown in Fig. 3-25 (*QST*, August, 1960) which are suitable for the 723 and 2K26 klystrons. One disadvantage of the system shown is that the shell of the klystron is at 260 volts above ground. An alternate method is to ground the shell and apply –260 volts to the cathode and –(260 + Vr) to the repeller. A 510-volt supply is needed for the repeller, but since the repeller draws negligible current, this should not be difficult.

As is the case with most microwave tubes, coupling power out from the klystron is somewhat more complicated than is the case with low-frequency tubes. A magnetic pickup loop placed in the cavity is connected either to a coaxial fitting or a waveguide probe. The latter (used with the 2K26 and the 723) is inserted into the middle of the waveguide and the coupling to the line is determined by the depth of the probe. Since klystrons (and other microwave tubes) are quite sensitive to variations in loading, some sort of attenuator or an isolator is often necessary to prevent malfunctions.

Other types of microwave tubes that the amateur may encounter are the traveling wave tube (TWT), and the backward wave oscillator (BWO). Here, an electromagnetic wave is slowed down below the speed of light in free space and allowed to interact continuously with an electron stream. While the latter two tubes use magnets for focusing the electron beam, the magnetron and other crossed-field amplifiers also use a magnetic field in conjunction with an electric field in their operation.

Chapter 4

Semiconductor Devices

Materials whose conductivity falls approximately midway between that of good conductors (e.g., copper) and good insulators (e.g., quartz) are called semiconductors. Some of these materials (primarily germanium and silicon) can, by careful processing, be used in solid-state electronic devices that perform many or all of the functions of thermionic tubes. In many applications their small size, long life and low power requirements make them superior to tubes.

The conductivity of a material is proportional to the number of free electrons in the material. Pure germanium and pure silicon crystals have relatively few free electrons. If, however, carefully controlled amounts of "impurities" (materials having a different atomic structure, such as arsenic or antimony) are added the number of free electrons, and consequently the conductivity, is increased. When certain other impurities are introduced (such as aluminum, gallium or indium), an electron deficiency, or hole, is produced. As in the case of free electrons, the presence of holes encourages the flow of electrons in the semiconductor material, and the conductivity is increased. Semiconductor material that conducts by virtue of the free electrons is called n-type material; material that conducts by virtue of an electron deficiency is called p-type.

Electron and Hole Conduction

If a piece of p-type material is joined to a piece of n-type material as at A in Fig. 4-1 and a voltage is applied to the pair as at B, current will flow across the boundary or junction between the two (and also in the external circuit) when the battery has the polarity indicated. Electrons, indicated by the minus symbol, are attracted across the junction from the n material through the p material to the positive terminal of the battery, and holes, indicated by the plus symbol, are attracted in the opposite direction across the junction by the negative potential of the battery. Thus current flows through the circuit by means of electrons moving one way and holes the other.

If the battery polarity is reversed, as at C, the excess electrons in the n material are attracted away from the junction and the holes in the p material are attracted by the negative potential of the battery away from the junction. This leaves the junction region without any current carriers, consequently there is no conduction.

In other words, a junction of p- and n-type materials constitutes a rectifier. It differs from the tube diode rectifier in that there is a measurable, although comparatively very small, reverse current. The reverse current results from the presence of some carriers of the type opposite to those which principally characterize the material.

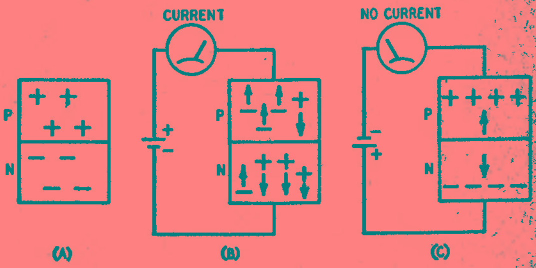


Fig. 4-1 — A p-n junction (A) and its behavior when conducting (B) and nonconducting (C).

With the two plates separated by practically zero spacing, the junction forms a capacitor of relatively high capacitance. This places a limit on the upper frequency at which semiconductor devices of this construction will operate, as compared with vacuum tubes. Also, the number of excess electrons and holes in the material depends upon temperature, and since the conductivity in turn depends on the number of excess holes and electrons, the device is more temperature sensitive than is a vacuum tube.

Capacitance may be reduced by making the contact area very small. This is done by means of a point contact, a tiny p-type region being formed under the contact point during manufacture when n-type material is used for the main body of the device.

SEMICONDUCTOR DIODES

Point-contact and junction-type diodes are used for many of the same purposes for which tube

Typical silicon and germanium diodes of the present era. The larger units are designed to handle high current.

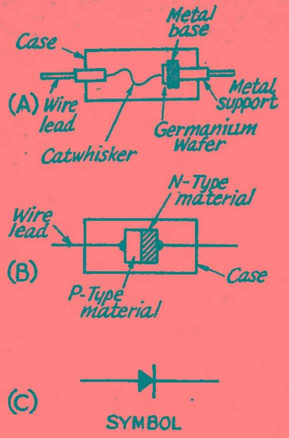
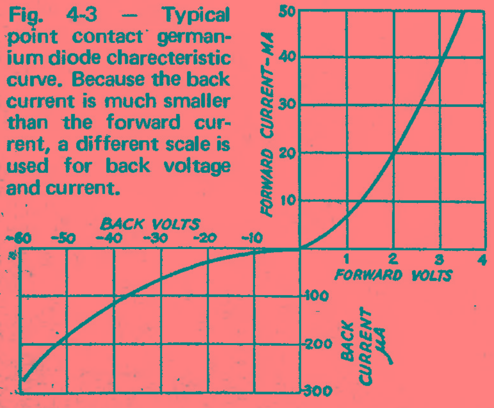


Fig. 4-2 — At A, a germanium point-contact diode. At B, construction of a silicon junction-type diode. The symbol et C is used for both diode types and indicates the direction of minimum resistanca measured by conventional methods. At C, the arrow corresponds to the plate (anode) of a vacuum-tube diode. The bar represents the tube's cathode element.

diodes are used. The construction of such diodes is shown in Fig. 4-2. Germanium and silicon are the most widely used materials; silicon finds much application as a microwave mixer diode. As compared with the tube diode for rf applications, the semiconductor point-contact diode has the advantages of very low interelectrode capacitance (on the order of 1 pF or less) and not requiring any heater or filament power.

The germanium diode is characterized by relatively large current flow with small applied voltages in the "forward" direction, and small, although finite, current flow in the reverse or "back" direction for much larger applied voltages. A typical characteristic curve is shown in Fig. 4-3. The dynamic resistance in either the forward or back direction is determined by the change in current that occurs, at any given point on the curve, when the applied voltage is changed by a small amount. The forward resistance shows some variation in the region of very small applied voltages, but the curve is for the most part quite straight, indicating fairly constant dynamic resistance. For small applied voltages, the forward resistance is of the order of 200 ohms or less in most such diodes. The back resistance shows considerable variation, depending on the particular voltage chosen for the measurement. It may run from a few thousand ohms to well over a megohm. In applications such as meter rectifiers for rf indicating instruments (rf voltmeters, wavemeter indicators, and so on) where the load resistance may be small and the applied voltage of the order of several volts, the resistances vary with the value of the applied voltage and are considerably lower.

Fig. 4-3 — Typical point contact germanium diode charecteristic curve. Because the back current is much smaller than the forward current, a different scale is used for back voltage and current.



Junction Diodes

Junction-type diodes made of silicon are employed widely as rectifiers. Depending upon the design of the diode, they are capable of rectifying currents up to 40 or 50 amperes, and up to reverse peak voltages of 2500. They can be connected in series or in parallel, with suitable circuitry, to provide higher capabilities than those given above. A big advantage over thermionic rectifiers is their large surge-to-average-current ratio, which makes them suitable for use with capacitor-only filter circuits. This in turn leads to improved no-load-to-full-load voltage characteristics. Some consideration must be given to the operating temperature of silicon diodes, although many carry ratings to 150 degrees C or so. A silicon junction diode requires a forward voltage of from 0.4 to 0.7 volts to overcome the junction potential barrier.

Ratings

Semiconductor diodes are rated primarily in terms of maximum safe inverse voltage (PIV or PRV) and maximum average rectified current. Inverse voltage is a voltage applied in the direction opposite to that which would be read by a dc meter connected in the current path.

It is also customary with some types to specify standards of performance with respect to forward and back current. A minimum value of forward current is usually specified for one volt applied. The voltage at which the maximum tolerable back current is specified varies with the type of diode.

Zener Diodes

The Zener diode is a special type of silicon junction diode that has a characteristic similar to that shown in Fig. 4-4. The sharp break from non-conductance to conductance is called the Zener knee; at applied voltages greater than this breakdown point, the voltage drop across the diode is essentially constant over a wide range of currents. The substantially constant voltage drop over a wide range of currents allows this semiconductor device to be used as a constant voltage reference or control element, in a manner somewhat similar to the gaseous voltage-regulator tube. Voltages for Zener-diode action range from a

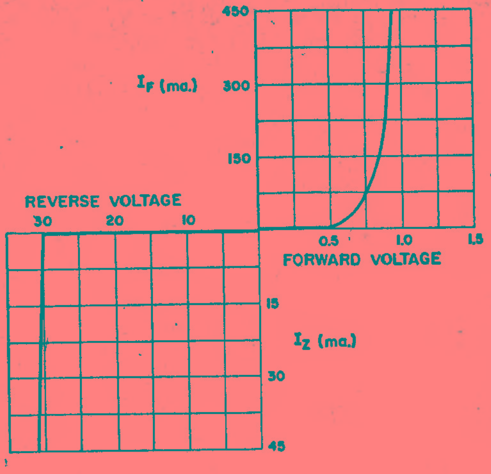


Fig. 4-4 — Typical characteristic of a Zener diode. In this example, the voltage drop is substantially constant at 30 volts in the (normally) reverse direction. Compare with Fig. 4-3. A diode with this characteristic would be called a "30-volt Zener diode."

few volts to several hundred and power ratings run from a fraction of a watt to 50 watts.

Zener diodes can be connected in series to advantage; the temperature coefficient is improved over that of a single diode of equivalent rating and the power-handling capability is increased.

Examples of Zener diode applications are given in Fig. 4-5. The illustrations represent some of the more common uses to which Zeners are put. Many other applications are possible, though not shown here.

Voltage-Variable Capacitor Diodes

Voltage-variable capacitors, Varicaps or varactors, are p-n junction diodes that behave as capacitors of reasonable Q when biased in the reverse direction. They are useful in many applications because the actual capacitance value is dependent upon the dc bias voltage that is applied. In a typical capacitor the capacitance can be varied over a 10-to-1 range with a bias change from 0 to −100 volts. The current demand on the bias supply is on the order of a few microamperes.

Typical applications include remote control of tuned circuits, automatic frequency control of receiver local oscillators, and simple frequency modulators for communications and for sweep-

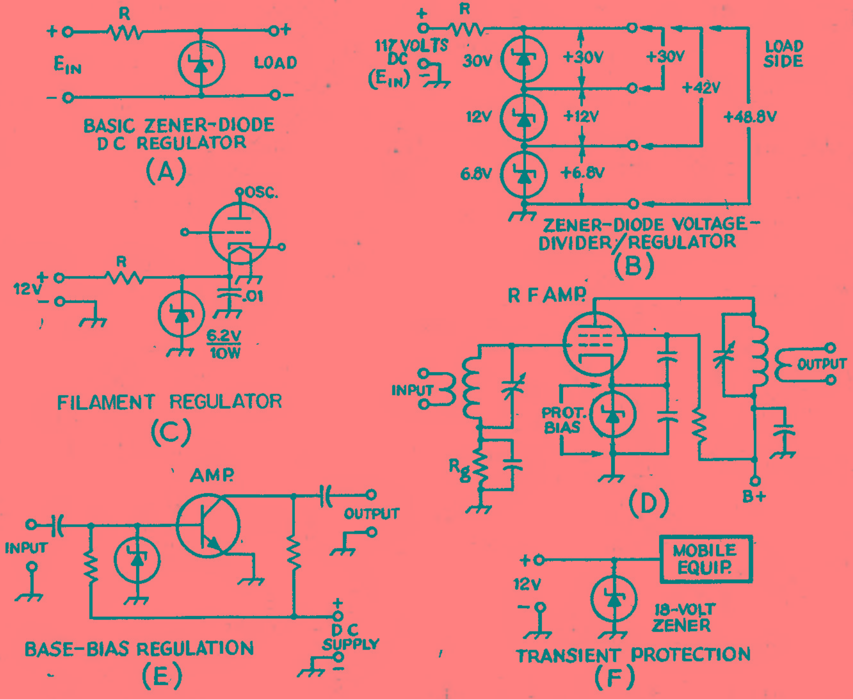


Fig. 4-5 — Zener diodes have many practical uses. Shown at A, is a simple dc voltage regulator which operates in the same menner as a gaseous regulator tube. Several Zener diodes can be connected in series (B) to provide various regulated voltages. At C, the filament line of a tube is supplied with regulated dc to anhance oscillator stability and reduce hum. In the circuit at D a Zener diode sets the bias level of an rf power amplifier. Bias regulation is afforded the bipolar transistor at E by connecting the Zener diode between base and ground. At F, the 18-volt Zener will clip peaks at and above 18 volts to protect 12-volt mobile equipment. (High peaks are frequently caused by transients in the automotive ignition system.)

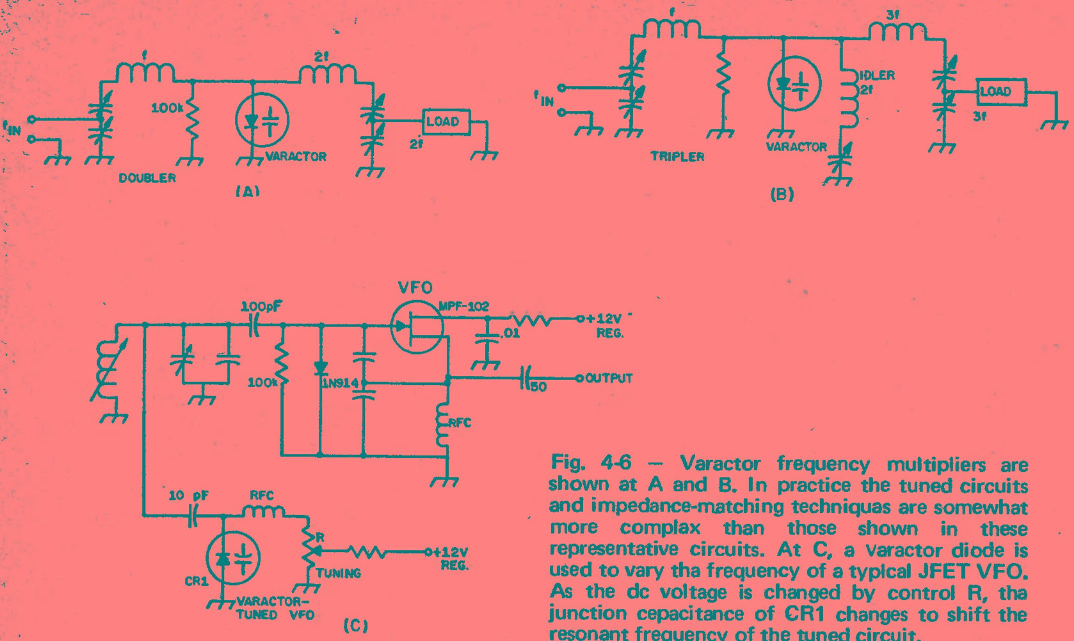


Fig. 4-6 — Varactor frequency multipliers are shown at A and B. In practice the tuned circuits and impedance-matching techniques are somewhat more complex than those shown in these representative circuits. At C, a varactor diode is used to vary the frequency of a typical JFET VFO. As the dc voltage is changed by control R, the junction capacitance of CR1 changes to shift the resonant frequency of the tuned circuit.

tuning applications. Diodes used in these applications are frequently referred to as "Varicap" or "Epicap" diodes.

An important transmitter application of the varactor is as a high-efficiency frequency multiplier. The basic circuits for varactor doublers and triplers are shown in Fig. 4-6, at A and B. In these circuits the fundamental frequency flows around the input loop. Harmonics generated by the varactor are passed to the load through a filter tuned to the desired harmonic. In the case of the tripler circuit at B, an idler circuit, tuned to the second harmonic, is required. Tripling efficiencies of 75 percent are not too difficult to come by, at power levels of 10 to 25 watts.

Fig. 4-6C illustrates how a voltage-variable capacitor diode can be used to tune a VFO. These diodes can be used to tune other rf circuits also, and are particularly useful for remote tuning such as might be encountered in vehicular installations. These diodes, because of their small size, permit tuned-circuit assemblies to be quite compact. Since the *Q* of the diode is a vital consideration in such applications, this factor must be taken into account when designing a circuit. Present-day manufacturing processes have produced units whose *Q*s are in excess of 200 at 50 MHz.

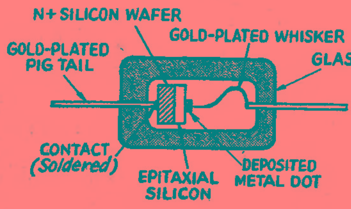


Fig. 4-7 — Cross-sectional view of a hot-carrier diode.

HOT-CARRIER DIODES

The hot-carrier diode is a high-frequency and microwave semiconductor whose characteristics fall somewhere between those of the point-contact diode and the junction diode. The former is comparable to the point-contact diode in high-frequency characteristics, and exceeds it in uniformity and reliability. The hot-carrier diode is useful in high-speed switching circuits and as a mixer, detector, and rectifier well into the microwave spectrum. In essence, the hot-carrier diode is a rectifying metal-semiconductor junction. Typical metals used in combination with silicon of either the n- or p-type are platinum, silver, gold or palladium.

The hot-carrier diode utilizes a true Schottky barrier, whereas the point-contact diode used a metal whisker to make contact with the semiconductor element. In a hot-carrier diode a planar area provides a uniform contact potential and uniform current distribution throughout the junction. This geometry results in lower series resistance, greater power capability, lower noise characteristics, and considerably greater immunity to burnout from transient pulses or spikes. A cross-sectional view of a hot-carrier diode is shown in Fig. 4-7 (courtesy of Hewlett Packard Associates). A comparison in characteristics between a point-contact diode and a hot-carrier diode is given in Fig. 4-8. Detailed information on the characteristics of hot-carrier diodes and their many applications is given in Hewlett Packard *Application Note 907.*

PIN Diodes

Another type of diode is the PIN diode. It might more aptly be described as a variable resistor than as a diode. In its intended application (at vhf and higher) it does not rectify the applied signal, nor does it generate harmonics. Its resistance is controlled by dc or a low-frequency signal, and the high-frequency signal which is being controlled by the diode sees a constant polarity-independent resistance. The dynamic resistance of the PIN diode is often larger than 10,000 ohms, and its junction capacitance is very low.

PIN diodes are used as variable shunt or series resistive elements in microwave transmission lines, and as agc diodes in the signal input lead to vhf and uhf fm receivers. The PIN diode offers many interesting possibilities.

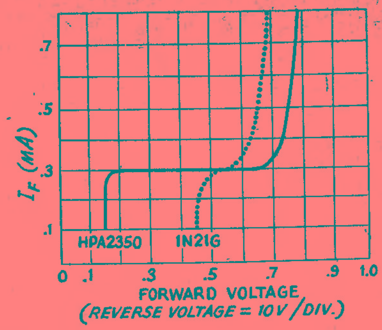


Fig. 4-8 — Curves showing the comparison in characteristics between a 1N21G point-contact diode and a Hewlett-Packard HPA2350 hot-carrier diode.

TRANSISTORS

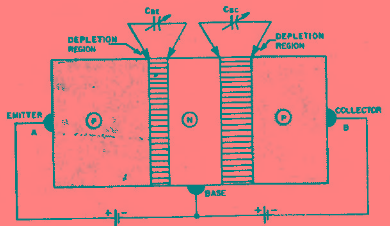


Fig. 4-9 — Illustration of a junction pnp trensistor. Capacitances Cbe and Cbc are discussed in the text, and vary with changes in opereting and signal voltage.

Fig. 4-9 shows a "sandwich" made from two layers of p-type semiconductor material with a thin layer of n-type between. There are in effect two pn junction diodes back to back. If a positive bias is applied to the p-type material at the left, current will flow though the left-hand junction, the holes moving to the right and the electrons from the n-type material moving to the left. Some of the holes moving into the n-type material will combine with the electrons there and be neutralized, but some of them also will travel to the region of the right-hand junction.

If the pn combination at the right is biased negatively, as shown, there would normally be no current flow in this circuit. However, there are now additional holes available at the junction to travel to point *B* and electrons can travel toward point *A*, so a current can flow even though this section of the sandwich is biased to prevent conduction. Most of the current is between *A* and *B* and does not flow out through the common connection to the n-type material in the sandwich.

A semiconductor combination of this type is called a **transistor**, and the three sections are known as the **emitter**, **base** and **collector**, respectively. The amplitude of the collector current depends principally upon the amplitude of the emitter current; that is, the collector current is controlled by the emitter current.

Between each p-n junction exists an area known as the **depletion**, or **transition region**. It is similar in characteristics to a dielectric layer, and its width varies in accordance with the operating voltage. The semiconductor materials either side of the depletion region consitute the plates of a capacitor. The capacitance from base to emitter is shown as C_{be} (Fig. 4-9), and the collector-base capacitance is represented as C_{bc}. Changes in signal and operating voltages cause a nonlinear change in these junction capacitances, which must be taken into account when designing some circuits. A base-emitter resistance, *rb'*, also exists. The junction capacitance, in combination with rb' determines the useful upper frequency limit (f_T or f_a) of a transistor by establishing an *RC* time constant.

Power Amplification

Because the collector is biased in the back direction the collector-to-base resistance is high.

This photo shows various modern-day bipoler and field-effect transistors. Various case styles and power classes are represented here.

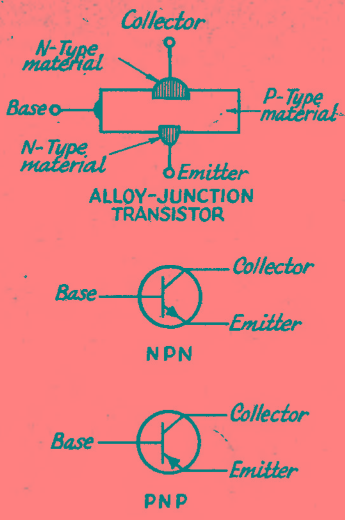


Fig. 4-10 – Schematic and pictorial representations of junction-type transistors. In analogous terms the base can be thought of es a tube's grid, the collector as a plate, and the emitter es a cathode. (See Fig. 4-12.)

On the other hand, the emitter and collector currents are substantially equal, so the power in the collector circuit is larger than the power in the emitter circuit ($P = I^2R$, so the powers are proportional to the respective resistances, if the currents are the same). In practical transistors emitter resistance is of the order of a few hundred ohms while the collector resistance is hundreds or thousands of times higher, so power gains of 20 to 40 dB or even more are possible.

Types

The transistor may be one of the types shown in Fig. 4-10. The assembly of p- and n-types materials may be reversed, so that pnp and npn transistors are both possible.

The first two letters of the npn and pnp designations indicate the respective polarities of the voltages applied to the emitter and collector in normal operation. In a pnp transistor, for example, the emitter is made positive with respect to both the collector and the base, and the collector is made negative with respect to both the emitter and the base.

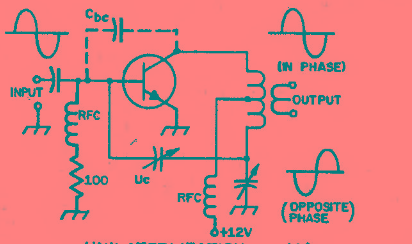
Manufacturers are constantly working to improve the performance of their transistors – greater reliability, higher power and frequency ratings, and improved uniformity of characteristics for any given type number. Recent developments provided the overlay transistor, whose emitter structure is made up of several emitters which are joined together at a common case terminal. This process lowers the base-emitter resistance, *rb'*, and improves the transistor's input time constant, which is determined by *rb'* and the junction capacitance of the device. The overlay transistor is extremely useful in vhf and uhf applications, and is capable of high-power operation well above 1000 MHz. These transistors are quite useful as frequency doublers and triplers, and are able to provide an actual power gain in the process.

Another multi-emitter transistor has been developed for use from hf through uhf, and should be of particular interest to the radio amateur. It is called a balanced-emitter transistor (BET), or "ballasted" transistor. The transistor chip contains several triode semiconductors whose bases and collectors are connected in parallel. The various emitters, however, have built-in emitter resistors (typically about 1 ohm) which provide a current-limiting safety factor during overload periods, or under conditions of significant mismatch. Since the emitters are brought out to a single case terminal the resistances are effectively in parallel, thus reducing the combined emitter resistances to a fraction of an ohm. (If a significant amount of resistance were allowed to exist it would cause degeneration in the stage and would lower the gain of the circuit.)

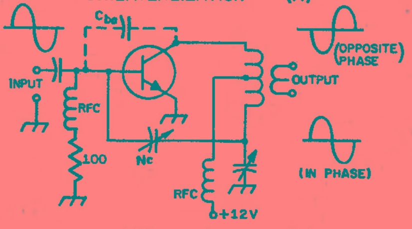
Most modern transistors are of the junction variety. Various names have been given to the several types, some of which are junction alloy, mesa, and planar. Though their characteristics may differ slightly, they are basically of the same family and simply represent different physical properties and manufacturing techniques.

Transistor Characteristics

An important characteristic of a transistor is its beta (β), or current-amplification factor, which is sometimes expressed as h_{FE} (static forward-current



UNILATERLIZATION (A)



NEUTRALIZATION (B)

Fig. 4-11 – Transit-time effects (in combination with base-collector capacitance Cbc) cen cause the positive-feedback condition shown at A. Normally, the phese of the collector signal of an amplifier is the inverse of the base signal. Positive feedback cen be corrected by using unilateralization, feeding an equal amount of opposite-phese signal back to the base through Uc. Neutrelization is shown at B, and deals with negative feedback, as can be seen by the phase ralationships shown.

transfer ratio) or hfe (small-signal forward-current transfer ratio). Both symbols relate to the grounded-emitter configuration. Beta is the ratio of the base current to the collector current. Thus, if a base current of 1 mA causes the collector current to rise to 100 mA the beta is 100. Typical betas for junction transistors range from as low as 10 to as high as several hundred.

A transistor's alpha (α) is the ratio of the emitter and collector currents. Symbols h_{FB} (static forward-current transfer ratio) and hfb (small-signal forward-current transfer ratio), common-base hookup, are frequently used in connection with gain. The smaller the base current, the closer the collector current comes to being equal to that of the emitter, and the closer alpha comes to being 1. Alpha for a junction transistor is usually between 0.92 and 0.98.

Transistors have frequency characteristics which are of importance to circuit designers. Symbol f_T is the gain bandwidth product (common-emitter) of the transistor. This is the frequency at which the gain becomes unity, or 1. The expression "alpha cutoff" is frequently used to express the useful upper-frequency limit of a transistor, and this relates to the common-base hookup. Alpha cutoff is the point at which the gain is 0.707, its value at 1000 Hz.

Another factor which limits the upper frequency capability of a transistor is its transit time. This is the period of time required for the current to flow from emitter to collector, through the semiconductor base material. The thicker the base material, the greater the transit time. Hence, the thicker the base material the more likelihood there will be of phase shift of the signal passing through it. At frequencies near and above f_T or alpha cutoff partial or complete phase shift can occur. This will give rise to positive feedback because the internal capacitance, C_{bc}, (Fig. 4-11) feeds part of the in-phase collector signal back to the base. The positive feedback can cause instability and oscillation, and in most cases will interlock the input and output tuned circuits of an rf amplifier so that it is almost impossible to tune them properly. Positive feedback can be corrected by using a form of neutralization called unilateralization. In this case the feedback conditions are balanced out. These conditions include a resistive as well as a capacitive component, thus changing a network from bilaterial to one which is unilateral. Negative feedback caused by C_{bc}, on the other hand, can be corrected by neutralization. Examples of both techniques are given in Fig. 4-11.

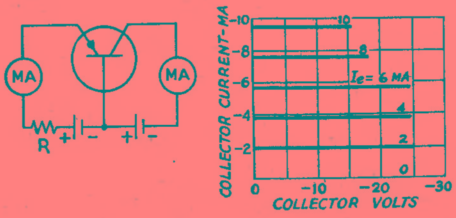


Fig. 4-12 – A typical collector-current vs. collector-voltage characteristic of a junction-type transistor, for various emitter-current values. The circuit shows the setup for taking such measurements. Since the emitter resistance is low, a current-limiting resistor, R, is connected in series with the source of current. The emitter current can be set at a desired value by adjustment of this resistance.

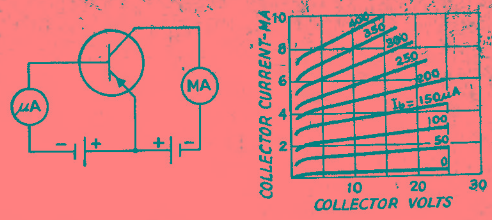


Fig. 4-13 – Collector current vs. collector voltage for various values of base current, for a junction-type transistor. The values are determined by means of the circuit shown.

Characteristic Curves

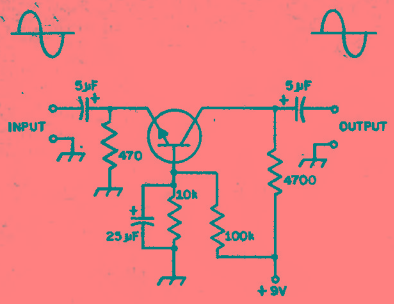
The operating characteristics of transistors can be shown by a series of characteristic curves. One such set of curves is shown in Fig. 4-12. It shows the collector current *vs.* collector voltage for a number of fixed values of emitter current. Practically the collector current depends almost entirely on the emitter current and is independent of the collector voltage. The separation between curves representing equal steps of emitter current is quite uniform, indicating that almost distortionless output can be obtained over the useful operating range of the transistor.

Another type of curve is shown in Fig. 4-13, together with the circuit used for obtaining it. This also shows collector current *vs* collector voltage, but for a number of different values of base current. In this case the emitter element is used as the common point in the circuit. The collector current is not independent of collector voltage with this type of connection, indicating that the output resistance of the device is fairly low. The base current also is quite low, which means that the resistance of the base-emitter circuit is moderately high with this method of connection. This may be contrasted with the high values of emitter current shown in Fig. 4-12.

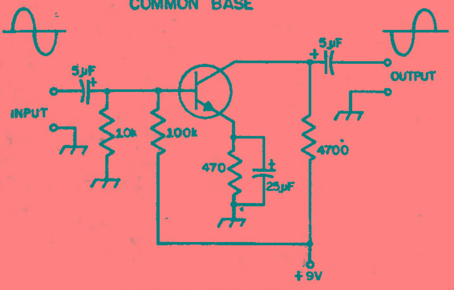
Ratings

The principal maximum ratings for transistors are collector dissipation, collector voltage, collector current, and emitter current. Variations in these basic ratings, such as maximum collector-to-base voltage, are covered in the symbols chart later in this chapter. The designer should study the maximum ratings of a given transistor before selecting it for use in a circuit.

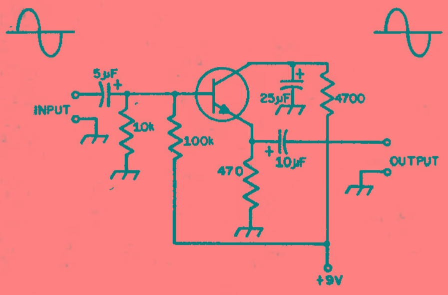
The dissipation rating can be a troublesome matter for an inexperienced designer. Techniques must be employed to reduce the operating



COMMON BASE



COMMON EMITTER



COMMON COLLECTOR

Fig. 4-14 — Basic transistor amplifier circuits. The differences between modes is readily apparent. Typical component values are given for use at audio frequencies. The input and output phase relationships are as shown.

temperature of power transistors, and this usually requires that thermal-conducting materials (heat sinks) be installed on the body of the transistor. The specification sheets list the maximum transistor dissipation in terms of case temperatures up to 25 degrees C. Symbol T_C is used for the case temperature and P_T represents the total dissipation. Silicone grease is often used to assure proper thermal transfer between the transistor and its heat sink. Additional information on the use of heat sinks is given in Chapter 18.

Excessive heat can lead to a condition known as **thermal runaway**. As the transistor gets hotter its internal resistance becomes lower, resulting in an increase of emitter-to-collector and emitter-to-base current. The increased current raises the dissipation and further lowers the internal resistance. The effects are cumulative, and eventually the transistor will be destroyed. It can be seen from this discussion that the use of heat sinks is important, where applicable.

TRANSISTOR AMPLIFIERS

Amplifier circuits used with transistors fall into one of three types, known as the **common-base, common-emitter,** and **common-collector circuits.** These are shown in Fig. 4-14 in elementary form. The three circuits correspond approximately to the grounded-grid, grounded-cathode and cathode-follower circuits, respectively, used with vacuum tubes.

The important transistor **parameters** in these circuits are the **short-circuit current transfer ratio,** the **cut-off frequency,** and the **input** and **output impedances.** The short-circuit current transfer ratio is the ratio of a small change in output current to the change in input current that causes it, the output circuit being short-circuited. The cutoff frequency was discussed earlier in this chapter. The input and output impedances are, respectively, the impedance which a signal source working into the transistor would see, and the internal output impedance of the transistor (corresponding to the plate resistance of a vacuum tube, for example).

Common-Base Circuit

The input circuit of a common-base amplifier must be designed for low impedance, since the emitter-to-base resistance is of the order of $25/I_e$ ohms, where I_e is the emitter current in milliamperes. The optimum output load impedance, R_L, may range from a few thousand ohms to 100,000, depending upon the requirements.

In this circuit the phase of the output (collector) current is the same as that of the input (emitter) current. The parts of these currents that flow through the base resistance are likewise in phase, so the circuit tends to be regenerative and will oscillate if the current amplification factor is greater than 1.

Common-Emitter Circuit

The common-emitter circuit shown in Fig. 4-14 corresponds to the ordinary grounded-cathode vacuum-tube amplifier. As indicated by the curves of Fig. 4-13, the base current is small and the input impedance is therefore fairly high — several thousand ohms in the average case. The collector resistance is some tens of thousands of ohms, depending on the signal source impedance. The common-emitter circuit has a lower cutoff frequency than does the common-base circuit, but it gives the highest power gain of the three configurations.

In this circuit the phase of the output (collector) current is opposite to that of the input (base) current so such feedback as occurs through the small emitter resistance is negative and the amplifier is stable.

Common-Collector Circuit

Like the vacuum-tube cathode follower, the common-collector transistor amplifier has high input impedance and low output impedance. The latter is approximately equal to the impedance of the signal input source multiplied by $(1 - \alpha)$. The input resistance depends on the load resistance, being approximately equal to the load resistance divided by $(1 - \alpha)$. The fact that input resistance is directly related to the load resistance is a disadvantage of this type of amplifier if the load is one whose resistance or impedance varies with frequency.

The current transfer ratio with this circuit is :

$$\frac{1}{1 - a}$$

and the cut-off frequency is the same as in the grounded-emitter circuit. The output and input currents are in phase.

PRACTICAL CIRCUIT DETAILS

The bipolar transistor is no longer restricted to use in low-voltage circuits. Many modern-day transistors have voltage ratings as high as 1400. Such transistors are useful in circuits that operate directly from the 117-volt ac line, following rectification. For this reason, battery power is no longer the primary means by which to operate transistorized equipment. Many low-voltage transistor types are capable of developing a considerable amount of af or rf power, hence draw amperes of current from the power supply. Dry batteries are seldom practical in circuits of this type. The usual approach in powering high-current, high-wattage transistorized equipment is to employ a wet-cell storage battery, or operate the equipment from a 117-volt ac line, stepping the primary voltage down to the desired level by means of a transformer, then rectifying the ac with silicon diodes.

Coupling and Impedance Matching

In contrast to vacuum tubes, bipolar transistors present low input and output impedances when used as amplifiers. Field-effect transistors are the exception, exhibiting terminal impedances similar to those of triode vacuum tubes. Therefore, the designer of bipolar transistor circuits must deal with specific matching techniques that assure efficient power transfer and acceptable stability of operation. Most of the *LC* networks used in tuned transistor amplifiers are of established standard configuration, but in practice call for much higher *C*-to-*L* ratios than are common to circuits using tubes. The low terminal impedances of bipolar transistors result from the fact that current is being amplified rather than voltage. High base or collector current (plus relatively low operating voltages) establishes what may at times seem to be unworkable terminal impedances – ten ohms or less. The greater the power input and output of an amplifier stage the more pronounced the matching problem becomes, requiring the employment of special matching techniques. Low-level amplifying stages are not so seriously affected, and the usual procedure is to use simple *RC*-coupling techniques for audio (and some rf) amplifiers. This being the case, the discussion will relate primarily to common-emitter stages that are called upon to deliver significant amounts of output power.

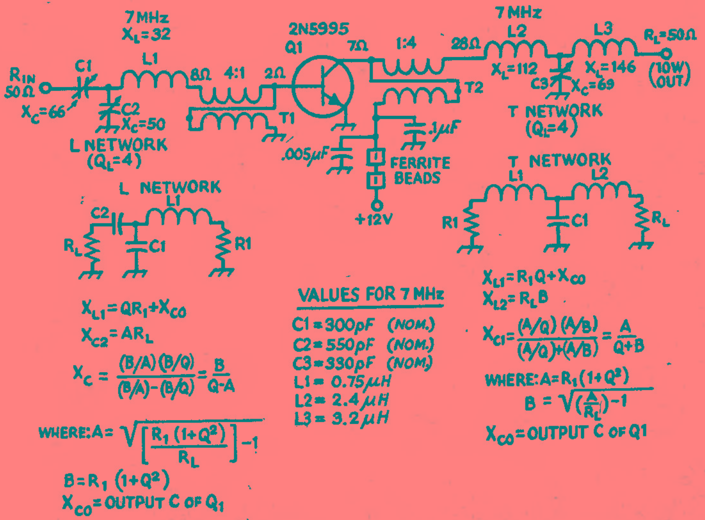
When designing matching network for efficient transfer of power from the collector to a given load impedance, the designer must first establish what the level of power output will be in watts. He must know also what the operating voltage for the collector (collector to emitter) will be. Once these quantities are determined the collector load impedance can be calculated by using the formula:

$$R_L = \frac{V_{cc}^2}{2P_o(\text{watts})}$$

where R_L = Collector load impedance at resonance

V_{cc} = Dc operating voltage, collector to emitter

P_o = Required power output in watts



$X_{L1} = QR_1 + X_{CO}$

$X_{C2} = AR_L$

$$X_C = \frac{(B/A)(B/Q)}{(B/A)-(B/Q)} = \frac{B}{Q-A}$$

WHERE: $A = \sqrt{\left[\frac{R_1(1+Q^2)}{R_L}\right] - 1}$

$B = R_1\,(1+Q^2)$

X_{CO} = OUTPUT C OF Q_1

VALUES FOR 7 MHz

C1 = 300pF (NOM.)
C2 = 550pF (NOM.)
C3 = 330pF (NOM.)
L1 = 0.75μH
L2 = 2.4μH
L3 = 3.2μH

$X_{L1} = R_1Q + X_{CO}$

$X_{L2} = R_LB$

$$X_{C1} = \frac{(A/Q)(A/B)}{(A/Q)+(A/B)} = \frac{A}{Q+B}$$

WHERE: $A = R_1(1+Q^2)$

$B = \sqrt{\left(\frac{A}{R_L}\right) - 1}$

X_{CO} = OUTPUT C OF Q1

Fig. 4-15 – A practical example illustrating the problems encountered when designing networks for use with solid-state power amplifiers. Transformers T1 and T2 can be used to bring the base and collector impedences up to a practical value for matching with *L* and *T* networks. The transformers are broad-band, toroidal types.

Example: An amplifier stage must deliver 10 watts to a known resistive load. The dc voltage from collector to emitter is 13.6. R_L is

$$R_L = \frac{V_{cc}^2}{2P_o} = \frac{184.96}{20} = 9.248 \text{ ohms}$$

It is not difficult to determine from this that an amplifier delivering, say, 25 watts output at a collector supply of 12 volts would have an extremely low collector impedance (2.88 ohms). Few standard *LC* networks are suitable for transforming that value to the typical 50-ohm nonreactive antenna impedance. The situation becomes even more complex when matching a power driver to the base element of a power-amplifier stage. In such a case it would not be uncommon to match an 18-ohm collector impedance to a 3-ohm base impedance, or similar.

Two common networks are illustrated in Fig. 4-15. Additional information is contained in Chapter 6 of this book. An excellent design aid is Motorola's *Matching Network Designs with Computer Solutions*, Application Note AN-267. The bibliography at the end of this chapter lists other recommended texts for amateur and professional designers.

Broadband toroidal-wound transformers and baluns are frequently used to match difficult impedances. They can be used in combination with tuned circuits or networks to arrive at practical network values. Resonant networks are employed to provide needed selectivity for assurance of clean output waveforms from amplifiers. A practical upper limit for network Q_L (loaded Q) is 5, though some professional engineers design for values higher than 5. It should be understood that the higher the Q_L the greater the chance for electrical instability. It is recommended that the amateur adhere to the practice of designing his networks for Q_L values between 3 and 5. Values as low as 1 are suitable for some circuits, especially low-pass harmonic filters of the variety used in the 50-ohm output line from many amplifiers.

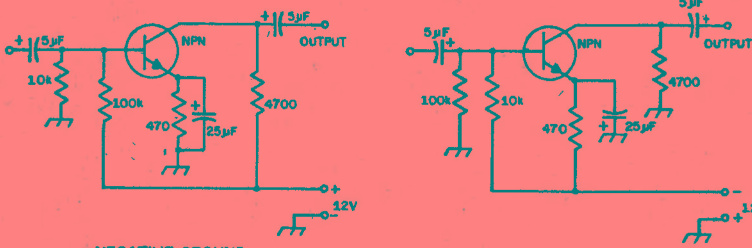
Bias and Bias Stabilization

Transistors must be forward biased in order to conduct significant current. In the npn design case the collector and base must be positive with respect to the emitter. The same is true when working with a pnp device, but the base and collector must be negative with respect to the emitter. The required bias is provided by the collector-to-emitter voltage, and by the emitter-to-base voltage. These bias voltages cause two currents to flow – emitter-to-collector current and emitter-to-base current. Either type of transistor, pnp or npn, can be used with a negative- or positive-ground power source by changing the circuit hookup as shown in Fig. 4-16. Forward bias is still properly applied in each instance. The lower the forward bias, the lower the collector current. As the forward bias is increased the collector current rises and the junction temperature increases. If the bias is continuously increased a point will be reached where the transistor becomes overloaded and burns out. This condition, called *thermal runaway* was discussed earlier in the chapter. To prevent damage to the transistor, some form of bias stabilization should be included in the design. Some practical bias-stabilization techniques are given in Fig. 4-17. At A and B, R1 in series with the emitter, is for the purpose of "swamping out" the resistance of the emitter-base diode; this swamping helps to stabilize the emitter current. The resistance of R1 should be large compared with that of the emitter-base diode, which is approximately equal to 25 divided by the emitter current in mA.

Since the current in R1 flows in such a direction as to bias the emitter negatively with respect to the base (a pnp transistor is assumed), a base-emitter bias slightly greater than the drop in R1 must be supplied. The proper operating point is achieved through adjustment of voltage divider R2R3, which is proportioned to give the desired value of no-signal collector current.

In the transformer-coupled circuit, input signal currents flow through R1 and R2, and there would be a loss of signal power at the base-emitter diode if these resistors were not bypassed by C1 and C2. The capacitors should have low reactance com-

Fig. 4-16 – An example of how the circuit polarity can be changed to accommodate either a positive or negative power-supply ground. Npn transistors are shown here, but the same rules apply to pnp types.



NEGATIVE GROUND

POSITIVE GROUND

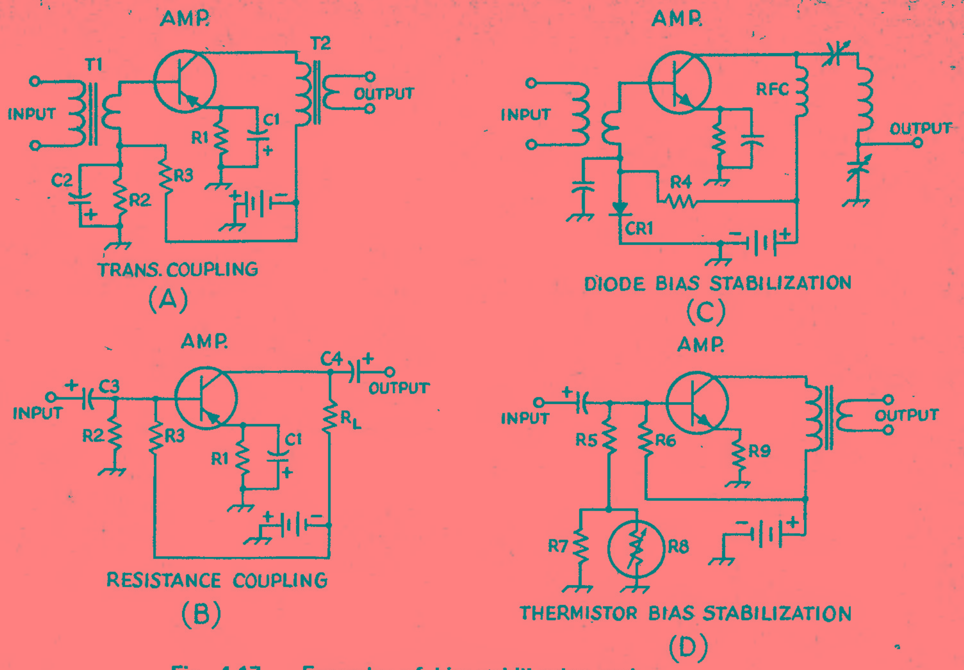


Fig. 4-17 – Examples of bias-stabilizetion techniques. A text discussion is given.

pared with the resistances across which they are connected. In the resistance-coupled circuit R2 serves as part of the bias voltage divider and also as part of the load for the signal-input source. As seen by the signal source, R3 is in parallel with R2 and thus becomes part of the input load resistance. C3 must have low reactance compared with the parallel combination of R2R3 and the base-to-emitter resistance of the transistor. The load impedance will determine the reactance of C4.

The output load resistance in the transformer-coupled case will be the actual load as reflected at the primary of the transformer, and its proper value will be determined by the transistor characteristics and the type of operation (Class A, B). The value of R_L in the resistance-coupled case is usually such as to permit the maximum ac voltage swing in the collector circuit without undue distortion, since Class-A operation is usual with this type of amplifier.

Transistor currents are sensitive to temperature variations, and so the operating point tends to shift as the transistor heats. The shift in operating point is in such a direction as to increase the heating, leading to thermal runaway. The heat developed depends on the amount of power dissipated in the transistor, so it is obviously advantageous in this respect to operate with as little internal dissipation as possible; i.e., the dc input should be kept to the lowest value that will permit the type of operation desired and should never exceed the rated value for the particular transistor used.

A contributing factor to the shift in operating point is the collector-to-base leakage current (usually designated I_{co}) – that is, the current that flows from the collector to base with the emitter connection open. This current, which is highly temperature sensitive, has the effect of increasing the emitter current by an amount much larger than I_{co} itself, thus shifting the operating point in such a way as to increase the collector current. This effect is reduced to the extent that I_{co} can be made to flow out of the base terminal rather than through the base-emitter diode. In the circuits of Fig. 4-17, bias stabilization is improved by making the resistance of R1 as large as possible and both R2 and R3 as small as possible, consistent with gain and power-supply economy.

It is common practice to employ certain devices in the bias networks of transistor stages to enhance bias stability. Thermistors or diodes can be used to advantage in such circuits. Examples of both techniques are given in Fig. 4-17 at C and D. Thermistors (temperature-sensitive resistors) can be used to compensate the rapid increase in collector current which is brought about by an increase in temperature. As the temperature in that part of the circuit increases, the thermistor's resistance decreases, reducing the emitter-to-base voltage (bias). As the bias is reduced in this manner, the collector current tends to remain the same, thus providing bias stabilization.

Resistors R5 and R7 of Fig. 4-17D are selected to give the most effective compensation over a particular temperature range.

A somewhat better bias-stabilization method is shown in Fig. 4-17C. In this instance, a diode is used between the base of the transistor and ground, replacing the resistor that is used in the circuits at A and B. The diode establishes a fixed value of forward bias and sets the no-signal collector current of the transistor. Also, the diode

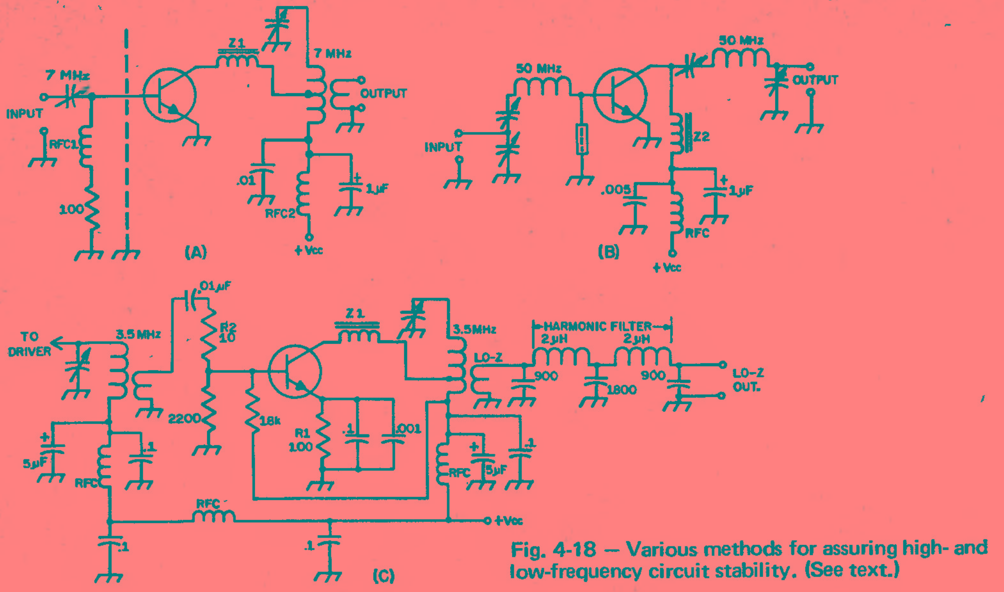


Fig. 4-18 — Various methods for assuring high- and low-frequency circuit stability. (See text.)

bias current varies in direct proportion with the supply voltage, tending to hold the no-signal collector current of the transistor at a steady value. If the diode is installed thermally close to the transistor with which it is used (clamped to the chassis near the transistor heat sink), it will provide protection against bias changes brought about by temperature excursions. As the diode temperature increases so will the diode bias current, thus lowering the bias voltage. Ordinarily, diode bias stabilization is applied to Class B stages. With germanium transistors, diode bias stabilization reduces collector-current variations to approximately one fifth of that obtainable with thermistor bias protection. With silicon transistors, the current variations are reduced to approximately one-fifteenth the thermistor-bias value.

Frequency Stability

Parasitic oscillations are a common source of trouble in transistor circuits. If severe enough in magnitude they can cause thermal runaway and destroy the transistor. Oscillation can take place at any frequency from just above dc to the f_T of the device, and these parasitics can often pass unnoticed if the waveforms are not examined with an oscilloscope. In addition to posing a potential danger to the device itself, the oscillations can cause distortion and unwanted radiation of spurious energy. In an amateur transmitter this condition can lead to violation notices from the FCC, interference to other services, and TVI. In the case of receivers, spurious energy can cause "birdies" and poor noise figures.

A transistor chosen for high-frequency operation (f_T at least five times greater than the proposed operating frequency) can easily oscillate *above* the operating frequency if feedback conditions are correct. Also, the device gain in the spectrum below the operating frequency will be very high, giving rise to low-frequency oscillation. At vhf and uhf phase shifts come into play, and this condition can encourage positive feedback, which leads to instability. At these higher frequencies it is wise to avoid the use of rf chokes and coupling capacitors whenever possible. The capacitors can cause shifts in phase (as can the base semiconductor material in the transistor), and the rf chokes, unless of very low *Q*, can cause a tuned-base tuned-collector condition. Some precautionary measures against instability are shown in Fig. 4-18. At A, RFC1 has its *Q* lowered by the addition of the 100-ohm series resistor. Alternatively, RFC1 could be shunted by a low-value resistor, but at some sacrifice in driving power. One or more ferrite beads can be slipped over the pigtail of an rf choke to lower the *Q* of the inductor. This method may be preferred in instances where the addition of a low-value resistor might establish an undesirable bias condition, as in the base return of a Class C stage. Parasitic choke Z1 consists of three ferrite beads slipped over a short piece of wire. The choke is installed as close to the collector terminal as possible. This low-*Q* choke will help prevent vhf or uhf instability. RFC2 is part of the decoupling network in the collector supply lead. It is bypassed for the operating frequency by means of the .01-µF capacitor, but is also bypassed for low frequencies by the addition of the 1-µF capacitor. In the vhf amplifier at B, Z1 and Z2 are ferrite-bead chokes. They present a high impedance to the base and collector elements, but because they are low-*Q* chokes there is little chance for them to permit a tuned-base tuned-collector oscillation. At C, the stage operates Class A, a typical arrangement in the low-level section of a transmitter, and the emitter is above ground by virtue of bias resistor R1. It must be bypassed to assure maximum stage gain. Here the emitter is bypassed for the operating frequency and for vhf. By not bypassing the emitter for low frequencies the stage is degenerative at lf. This will lessen the chance of low-frequency oscillation. The supply leads, however, are bypassed for the operating frequency and for lf, thus preventing unwanted feedback between stages along the supply leads. Z1 is a ferrite-bead vhf/uhf parasitic choke. The 10-ohm resistor, R2, also helps suppress vhf parasitics. The emitter lead should be kept as short as possible in all three circuits to enhance stability and to prevent degeneration at the operating frequency. It is wise to use rf shields between the input and output halves of the rf amplifier stage to prevent unwanted coupling between the base and collector tuned circuits. At operating frequencies where toroid cores are suitable, the shields can often be omitted if the tuned circuits use toroidal inductors. Toroidal transformers and inductors have self-shielding properties – an asset to the designer.

FIELD-EFFECT TRANSISTORS

Still another semiconductor device, the field-effect transistor, (FET) is superior to bipolar transistors in many applications because it has a high input impedance, its characteristics more nearly approach those of a vacuum tube.

The Junction FET

Field-effect transistors are divided into two main groups: junction FETs, and MOSFETs. The basic JFET is shown in Fig. 4-19.

The reason for the terminal names will become clear later. A dc operating condition is set up by starting a current flow between source and drain. This current flow is made up of free electrons since the semiconductor is n-type in the channel, so a positive voltage is applied at the drain. This positive voltage attracts the negatively charged free electrons and the current flows (Fig. 4-20). The next step is to apply a gate voltage of the polarity shown in Fig. 4-20. Note that this reverse-biases the gates with respect to the source, channel, and drain. This reverse-bias gate voltage causes a depletion layer to be formed which takes up part of the channel, and since the electrons now have less volume in which to move the resistance is greater and the current between source and drain is reduced. If a large gate voltage is applied the depletion regions meet, causing pinch off, and consequently the source-drain current is reduced nearly to zero. Since the large source-drain current changed with a relatively small gate voltage, the device acts as an amplifier. In the operation of the JFET, the gate terminal is never foward biased, because if it were the source-drain current would all be diverted through the forward-biased gate junction diode.

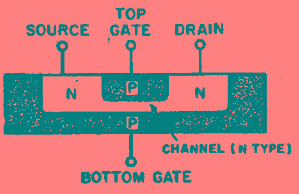


Fig. 4-19 — The junction field-effect transistor.

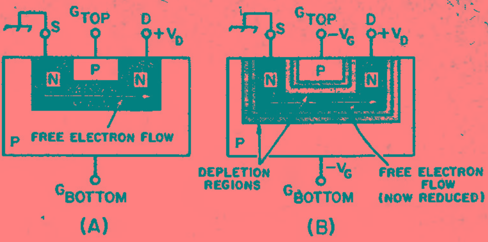


Fig. 4-20 — Operation of the JFET under applied bias. A depletion region (light shading) is formed, compressing the channel and increasing its resistance to current flow.

The resistance between the gate terminal and the rest of the device is very high, since the gate terminal is always reverse biased, so the JFET has a very high input resistance. The source terminal is the *source* of current carriers, and they are *drained* out of the circuit at the drain. The gate *opens* and *closes* the amount of channel current which flows in the pinch-off region. Thus the operation of a FET closely resembles the operation of the vacuum tube with its high grid input impedance. Comparing the JFET to a vacuum tube, the source corresponds to the cathode, the gate to the grid, and the drain to the plate.

MOSFETs

The other large family which makes up field-effect transistors is the insulated-gate FET, or MOSFET, which is pictured schematically in Fig. 4-21. In order to set up a dc operating condition, a positive polarity is applied to the drain terminal. The substrate is connected to the source, and both are at ground potential, so the channel electrons are attracted to the positive drain. In order to regulate this source-drain current, voltage is applied to the gate contact. The gate is insulated from the rest of the device by a layer of very thin dielectric material, so this is not a p-n junction between the gate and the device – thus the name insulated gate. When a negative gate polarity is applied, positive-charged holes from the p-type substrate

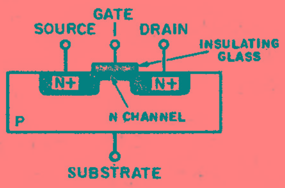


Fig. 4-21 — The insulated-gate field-effect transistor.

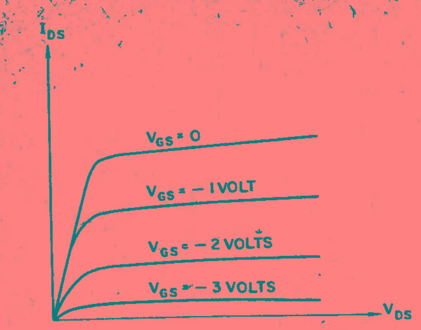


Fig. 4-21A—Typical JFET characteristic curves.

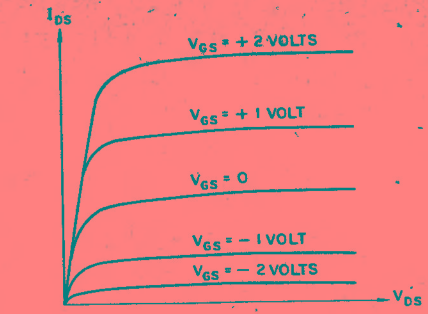


Fig. 4-21B—Typical MOSFET characteristic curves.

are attracted toward the gate and the conducting channel is made more narrow; thus the source-drain current is reduced. When a positive gate voltage is connected, the holes in the substrate are repelled away, the conducting channel is made larger, and the source-drain current is increased. The MOSFET is more flexible since either a positive or negative voltage can be applied to the gate. The resistance between the gate and the rest of the device is extremely high because they are separated by a layer of thin dielectric. Thus the MOSFET has an extremely high input impedance. In fact, since the leakage through the insulating material is generally much smaller than through the reverse-biased p-n gate junction in the JFET, the MOSFET has a much higher input impedance. Typical values of R_{in} for the MOSFET are over a million megohms, while R_{in} for the JFET ranges from megohms to over a thousand megohms. There are both single-gate and dual-gate MOSFETs available. The latter has a signal gate, Gate 1, and a control gate , Gate 2. The gates are effectively in series making it an easy matter to control the dynamic range of the device by varying the bias on Gate 2. Dual-gate MOSFETs are widely used as agc-controlled rf and i-f amplifiers, as mixers and product detectors, and as variable attenuators. The isolation between the gates is relatively high in mixer service. This helps lessen oscillator "pulling" and reduces oscillator radiation. The forward transadmittance (transconductance, or g_m) of modern MOSFETs is as high as 18,000, and they are designed to operate efficiently well into the uhf spectrum.

Characteristic Curves

The characteristic curves for the FETs described above are shown in Figs. 4-21A and 4-21B, where drain-source current is plotted against drain-source voltage for given gate voltages.

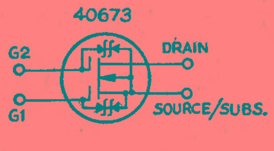


Fig. 4-22 — Schematic presentation of a gate-protected MOSFET. Back-to-back Zener diodes are bridged internally from gates 1 and 2 to the source/substrate element.

Classifications

Field-effect transistors are classed into two main groupings for application in circuits, ENHANCEMENT MODE and DEPLETION MODE. The enhancement-mode devices are those specifically constructed so that they have *no* channel. They become useful only when a gate voltage is applied that causes a channel to be formed. IGFETs can be used as enhancement-mode devices since both polarities can be applied to the gate without the gate becoming forward biased and conducting.

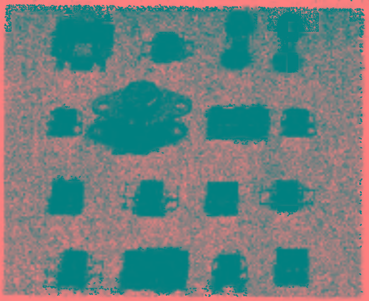
A depletion-mode unit corresponds to Figs. 4-19 and 4-21 shown earlier, where a channel exists with no gate voltage applied. For the JFET we can apply a gate voltage and deplete the channel, causing the current to decrease. With the MOSFET we can apply a gate voltage of either polarity so the device can be depleted (current decreased) or enhanced (current increased).

To sum up, a depletion-mode FET is one which has a channel constructed; thus it has a current flow for zero gate voltage. Enhancement-mode FETs are those which have no channel, so no current flows with zero gate voltage.

Gate-Protected FETs

Most JFETs are capable of withstanding up to 80 volts pk-pk from gate to source before junction damage occurs. Insulated-gate FETs, however, can be damaged by allowing the leads to come in contact with plastic materials, or by the simple act of handling the leads with one's fingers. Static charges account for the foregoing, and the damage takes the form of punctured dielectric between the gate or gates and the remainder of the internal elements. Devices of the MFE3006 and 3N140 series are among those which can be easily damaged.

Gate-protected MOSFETs are currently available, and their gates are able to withstand pk-pk voltages (gate to source) of up to 10. Internal Zener diodes are connected back to back from each gate to the source/substrate element. The 40673 and 3N200 FETs are among the types which have built-in Zener diodes. Dual-gate MOSFETs which are gate-protected can be used as single-gate protected FETs by connecting the two gate leads in parallel. A gate-protected MOSFET is shown schematically in Fig. 4-22.



A collection of modern ICs. Various case styles of metal and epoxy materials are illustrated.

INTEGRATED CIRCUITS

Just as the term implies, integrated circuits (ICs) contain numerous components which are manufactured in such a way as to be suitably interconnected for a particular application, and on one piece of semiconductor base material. The various elements of the IC are comprised of bi-polar transistors, MOSFETs, diodes, resistances, and capacitances. There are often as many as ten or more transistors on a single IC chip, and frequently their respective bias resistors are formed on the chip. Generally speaking, ICs fall into four basic categories – differential amplifiers, operational amplifiers, diode or transistor arrays, and logic ICs.

IC Structures

The basic IC is formed on a uniform chip of n-type or p-type silicon. Impurities are introduced into the chip, their depth into it being determined by the diffusion temperature and time. The geometry of the plane surface of the chip is determined by masking off certain areas, applying photochemical techniques, and applying a coating of insulating oxide. Certain areas of the oxide coating are then opened up to allow the formation of interconnecting leads between sections of the IC. When capacitors are formed on the chip, the oxide serves as the dielectric material. Fig. 4-23 shows a representative three-component IC in both pictorial and schematic form. Most integrated circuits are housed in TO-5 type cases, or in flat-pack epoxy blocks. ICs may have as many as 12 or more leads which connect to the various elements on the chip.

Types of IC Amplifiers

Some ICs are called differential amplifiers and others are known as operational amplifiers. The basic differential-amplifier IC consists of a pair of transistors that have similar input circuits. The inputs can be connected so as to enable the transistors to respond to the difference between two voltages or currents. During this function, the circuit effectively suppresses like voltages or currents. For the sake of simplicity we may think of the differential pair of transistors as a push-pull amplifier stage. Ordinarily, the differential pair of transistors are fed from a controlled, constant-current source (Q3 in Fig. 4-24A. Q1 and Q2 are the differential pair in this instance). Q3 is commonly called a transistor current sink. Excellent balance exists between the input terminals of differential amplifiers because the base-to-emitter voltages and current gains (beta) of

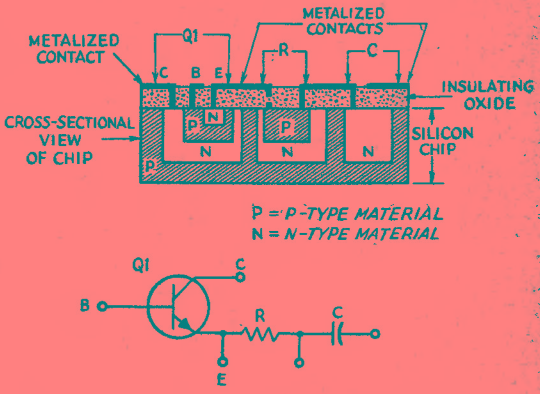


Fig. 4-23 — Pictorial and schematic illustrations of a simple IC device.

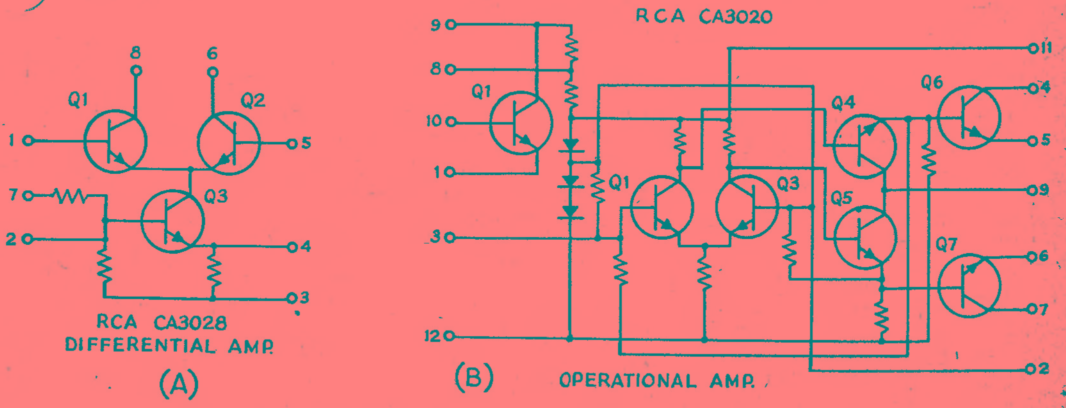
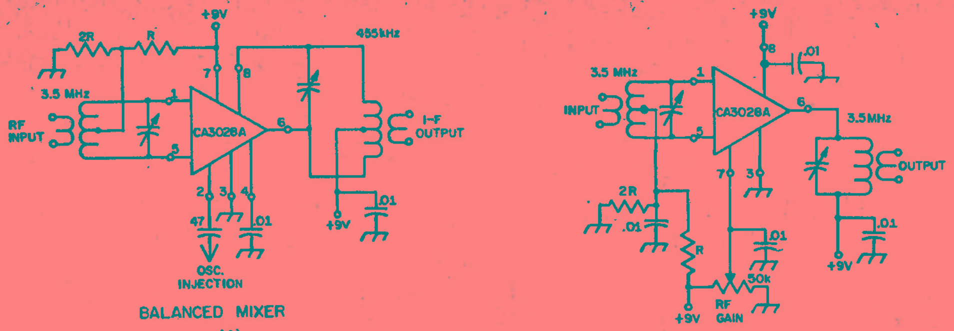


Fig. 4-24 — At A, a representative circuit for a typical differential IC. An Operational Amplifier IC is illustrated at B, also in representative form.



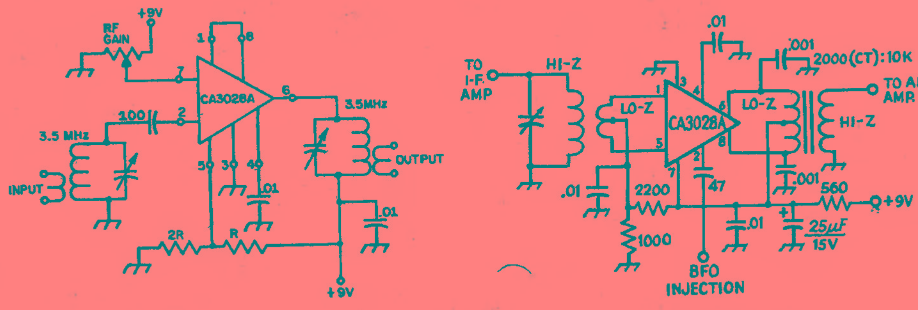
BALANCED MIXER

(A)

BALANCED DIFFERENTIAL AMPLIFIER

(B)

Fig. 4-25 — Some typical circuit applications for a differential amplifier IC. The internal circuit of the CA3028A IC is given in Fig. 4-24 at A.



CASCODE AMP.

(C)

PRODUCT DETECTOR

(D)

the two transistors are closely matched. The match results from the fact that the transistors are formed next to one another on the same silicon chip.

Differential ICs are useful as linear amplifiers from dc to the vhf spectrum, and can be employed in such circuits as limiters, product detectors, frequency multipliers, mixers, amplitude modulators, squelch, rf and i-f amplifiers, and even in signal-generating applications. Although they are designed to be used as differential amplifiers, they can be used in other types of circuits as well, treating the various IC components as discrete units.

Operational-amplifier ICs are basically very-high-gain direct-coupled amplifiers that rely on feedback for control of their response characteristics. They contain cascaded differential amplifiers of the type shown in Fig. 4-24A. A separate output stage, Q6–Q7, Fig. 4-24B, is contained on the chip. Although operational ICs can be successfully operated under open-loop conditions, they are generally controlled by externally applied negative feedback. Operational amplifiers are most often used for audio amplification, as frequency-shaping (peaking, notching, or bandpass) amplifiers, or as integrator, differentiator, or comparator amplifiers.

Diode-ICs are also being manufactured in the same manner as outlined in the foregoing section. Several diodes can be contained on a single silicon wafer to provide a near-perfect match between diode characteristics. The diode arrangement can take the form of a bridge circuit, series-connected groups, or as separate components. Diode ICs of this kind are useful in balanced-modulator circuits, or to any application requiring closely matched diodes.

Fig. 4-25 demonstrates the versatility of just one type of IC, an RCA CA3028A differential amplifier. Its internal workings are shown in Fig. 4-24A, permitting a comparison of the schematic diagram and the block representations of Fig. 4-25. The circuit at B in Fig. 4-25 is characterized by its high input and output impedances (several thousand ohms), its high gain, and its stability. This circuit can be adapted as an audio amplifier by using transformer or RC coupling. In the circuits of B and C terminal 7 is used to manually control the rf gain, but agc can be applied to that terminal instead. In the circuit at D the CA3028A provides low-noise operation and exhibits good conversion gain as a product detector. The CA3028A offers good performance from dc to 100 MHz.

PRACTICAL CONSIDERATIONS

Some modern-day ICs are designed to replace nearly all of the discrete components used in earlier composite equipment. One example can be seen in the RCA CA3089E flat-pack IC which contains nearly the entire circuit for an fm receiver. The IC contains 63 bipolar transistors, 16 diodes, and 32 resistors. The CA3089E is designed for an i-f of 10.7 MHz and requires but one outboard tuned circuit. The chip consists of an i-f

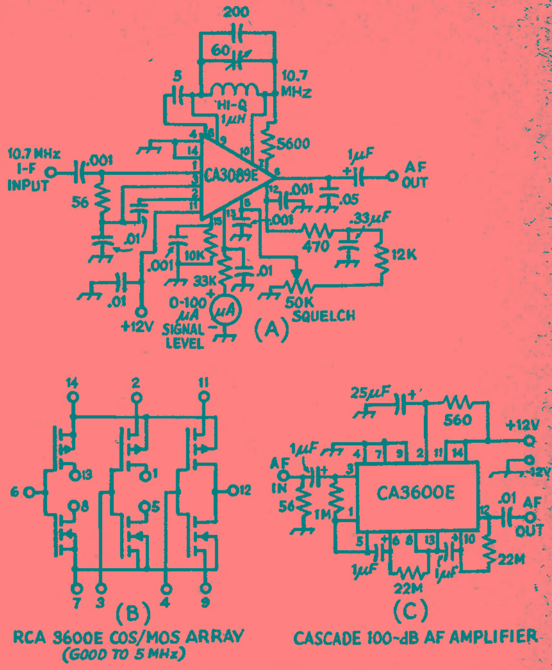


Fig. 4-26 — The circuit at A shows practical component values for use with the CA3089E fm subsystem IC. A COS/MOS array IC is illustrated at B in schematic-diagram form. It consists of three complementary-symmetry MOSFET pairs. The illustration at C shows how the CA3600E can be connected in cascade to provide at least 100 dB of audio amplification.

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (µF) ; OTHERS ARE IN PICOFARADS (pF OR µµF); RESISTANCES ARE IN OHMS ; k = 1000, M = 1000 000.

amplifier, quadrature detector, audio preamplifier, agc, afc, squelch, and a tuning-meter circuit. Limiting of -3 dB takes place at the 12-µV input level. When using an IC of this kind it is necessary only to provide a front-end converter for the desired frequency of reception, an audio amplifier, and a power supply (plus speaker, level controls, and meter).

There are two IC subsystem units designed for a-m receiver use. Each is similar in complexity to the CA3089E illustrated in Fig: 4-26. These components are identified as CA3088E and CA3123E. The latter is described in RCA Data File No. 631. Both ICs are readily adaptable to communications receiver use and should become popular building blocks for amateurs who desire compact, portable receiving equipment.

TRANSISTOR ARRAYS

Amateur designers should not overlook the usefulness of transistor- and diode-array ICs. These devices contain numerous bipolar or MOSFET transistors on a common substrate. In most instances the transistors can be employed as one would treat discrete npn devices. An entire receiver can be made from one transistor-array IC if one wishes to construct a compact assembly. The CA3049 is a dual independent differential rf/i-f amplifier chip with an f_T of 1.3 GHz. It is especially well suited to applications which call for double-balanced mixers, detectors, and modulators. Another device of similar usefulness is the CA3018A. The CA3045 should also be of interest to the amateur. Matched electrical characteristics of the transistors in these ICs offer many advantages not available when using discrete transistors. Fig. 4-27 shows the internal workings of the CA3018A and CA3045 ICs.

COS/MOS (complementary-symmetry metal-oxide silicon) ICs are becoming increasingly popular, and one RCA part, the CA3600E, contains an array of complementary-symmetry MOSFET pairs (three) which can be used individually or in cascade, as shown in Fig. 4-26 at B. Detailed information is given in RCA File No. 619. The CA3600E is a high-input impedance, micropower component which is suitable for use as a preamplifier, differential amplifier, op amp, comparator, timer, mixer, chopper, or oscillator.

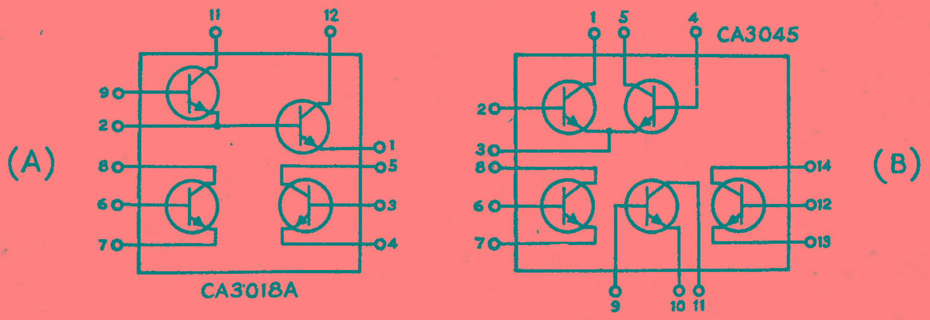


Fig. 4-27 — Transistor arrays offer unlimited application because several circuit combinations are possible. The CA3018A IC at A has a Darlington-connected pair plus two separate transistors. At B, two transistors are internally connected in a differential amplifier fashion. Three separate transistors are available for use in other functions. The arrays shown here are useful into the vhf spectrum.