SECTION 2
GENERAL INFORMATION ON ELECTRON TUBE CIRCUITS

2.1 DEFINITIONS OF LETTER SYMBOLS USED.

Throughout this technical manual a number of letter symbols are used to indicate components, sources of voltage and current, and to differentiate between points not at the same power and voltage levels. Since the technician must be able to read, speak, and understand the jargon of the trade, he should also learn to recognize the letter symbols used as a form of shorthand notation in technical discussions and on engineering drawings and schematic diagrams. To avoid any conflict or confusion, standard letter symbols are used where available. Since definitions and usage change from time to time, a complete alphabetical listing of all symbols used throughout this manual is included for reference (they need not be memorized).

While the basic symbol such as the capital letters E and I always indicate voltage and current, the lower case and subscript letters (or numbers) are assigned as described in paragraph 2.1.1. Therefore, it is recommended that the user of this manual refer to the definitions listed below to determine the correct meanings of the letter symbols used in the circuits in this technical manual.

2.1.1 **Construction of Symbols.** The letter symbols are made up of single letters with subscripts or superscripts in accordance with the following conventions:

a. Maximum, average, and root-mean-square values are represented by capital (upper case) letters; for example: I, E, P.

b. Where needed to distinguish between values in item a above, the maximum value may be represented by the subscript "m"; for example: E_m, I_m, P_m.

c. Average values may be represented by the subscript "av"; for example: E_{av}, I_{av}, P_{av}.

Note

When items b and c above are used, then item a indicates r-m-s, or effective, values.

d. Instantaneous values of current, voltage, and power which vary with time are represented by the lower case (small) letter of the proper symbol; for example: i, e, p.

e. External resistance, impedance, etc, in the circuit external to a vacuum-tube electrode may be represented by the upper case symbol with the proper electrode subscripts; for example: R_g, R_{sc}, Z_g, Z_{sc}.

f. Values of resistance, impedance, etc, inherent within the electron tube are represented by the lower case symbol with the proper electrode subscripts; for example: r_g, z_g, r_p, z_p, c_{gp}.

g. The symbols "g" and "p" are used as subscripts to identify instantaneous (ac) values of electrode currents and voltages; for example: e_g, e_p, i_g, i_p.

h. The total instantaneous values of electrode currents and voltages (d-c plus a-c components) are indicated by the lower case symbol and the subscripts "b" for plate and "c" for grid; for example: i_b, e_c, i_c, e_b.

i. No-signal or static currents and voltages are indicated by upper case symbol and lower case subscripts "b" for plate and "c" for grid; for example: E_c, I_b, E_b, I_c.

j. R-M-S and maximum values of a varying component are indicated by the upper case letter and the subscripts "g" and "p"; for example: E_g, I_p, E_p, I_g.

k. Average values of current and voltage for the with-signal condition are indicated by adding the subscript "s" to the proper symbol and subscript; for example: I_{bs}, E_{bs}.

l. Supply voltages are indicated by the upper case symbol and double subscript "bb" for plate, "cc" for grid, "ff" for filament; for example: E_{ff}, E_{cc}, E_{bb}.

2.1.2 **List of Symbols.** Since the rules above are somewhat complex, an alphabetical list of the symbols used in this technical manual for electron tube circuitry is presented below.

Symbols	**Definitions**
C	Capacitor, capacitance
C_c	Coupling capacitor
C_d	Distributed circuit capacitance
C_g	Grid-leak or grid capacitor
C_{gk}	Grid-cathode capacitance
C_{gp}	Grid-plate capacitance
C_k	Cathode capacitor
C_n	Neutralizing capacitor
C_p	Plate capacitor
C_{pk}	Plate-cathode capacitance
C_{sc}	Screen capacitor
C_{sup}	Suppressor capacitor
E	Voltage
E_a	Applied voltage
E_{av}	Average voltage
E_b	Plate voltage, static d-c value
E_{bb}	Plate voltage source (supply voltage)
E_C	Capacitor voltage
E_c	Grid voltage, static d-c (bias) value
E_{car}	D-C voltage applied to produce carrier
E_{cc}	Grid bias supply voltage
E_{cc1}	Control grid supply
E_{cc2}	Screen grid supply
E_{cc3}	Suppressor grid supply
E_{co}	Negative tube cutoff voltage
E_f	Filament voltage
E_{ff}	Filament supply voltage
E_g	Root-mean-square value of grid voltage
E_{gm}	Maximum value of varying grid voltage component
E_{in}	Input voltage
E_L	Inductive voltage drop
E_m, E_{max}	Maximum voltage
E_{min}	Minimum voltage
E_o	Output voltage, peak
E_p	Plate voltage r-m-s value
E_R	Resistive voltage drop
E_{sc}	Screen voltage, d-c value
e	Instantaneous voltage
e_b	Instantaneous plate voltage
e_c	Instantaneous grid voltage
e_{car}	Instantaneous carrier voltage

Symbols	Definitions
e_g	Instantaneous a-c component of grid voltage
e_1	Instantaneous voltage on element 1
e_m	Maximum instantaneous voltage
e_o	A-C component of output voltage
e_p	Instantaneous a-c component of plate voltage
e_r	Instantaneous resistive voltage drop
e_{sc}	Instantaneous screen voltage
e_{sig}	Instantaneous signal voltage
e_{su}	Instantaneous suppressor voltage
f	Frequency
f_o, f_r	Resonant frequency
g_m	Transconductance
I	Current
I_{av}	Average current
I_b, I_o	Static d-c plate current
I_C	Capacitive current
I_c	Static d-c grid current
I_f	Filament current
I_g	Grid current, r-m-s value
I_L	Inductive current
I_m, I_{max}	Maximum current
I_{min}	Minimum current
I_p	D-C plate current, r-m-s value
I_R	Current in resistor
I_{rk}	Cathode current
I_T	Total current
I_t	Current for time interval (usually used with subscripts, as t_1, t_2, etc)
i_b	Instantaneous plate current
i_c	Instantaneous grid current
i_g	Instantaneous a-c component of grid current
i_L	Instantaneous inductive current
i_p	Instantaneous a-c component of plate current
i_t	Current per instant of time
i_Z	Instantaneous current through impedance
L	Inductance, inductor
P	Power
P_g	Grid dissipation power
P_i	Input power
P_o	Output power
P_p	Plate dissipation power
P_q	Reactive power
P_s	Apparent power
Q	Figure of merit
R	Resistance, resistor
R_G	Generator internal resistance
R_g	Grid resistance
R_L	Load resistance
R_k	Cathode resistance
R_p	Plate resistance, dc
R_S	Series resistance
R_{sc}	Screen resistance
r_L	A-C load resistance
r_p	Plate resistance, ac
SWR	Standing-wave ratio
t	Time
t_d	Deionization time
t_f	Pulse fall time
t_k	Cathode heating time
t_p	Pulse duration time
t_r	Pulse rise time
T	Transformer
TC	Time constant
V, v	Voltage, volts
W	Watts
X	Reactance
X_C	Capacitive reactance
X_L	Inductive reactance
Y	Admittance
Z	Impedance
Z_{in}	Input impedance
Z_L	Load impedance
Z_o	Characteristic impedance, surge impedance
Z_{out}	Output impedance

2.2 BIASING METHODS.

An electron tube is normally biased through the application of a negative d-c potential (grid bias) between the grid and cathode elements. Grid bias is either obtained from a separate voltage supply source set for a specific fixed voltage, or developed by the flow of cathode or control grid current in the tube; bias developed by tube current is referred to as "self-bias". The grid bias determines the static (quiescent) operating current for the applied plate voltage and thus sets the operating point. When an input signal is applied to the grid, it adds to or subtracts from the initial bias in accordance with instantaneous signal variations and correspondingly varies the plate current and voltage to produce the desired output signal.

Usually bias may be obtained by any of a number of methods; the basic circuits are discussed in the following paragraphs. For ease of explanation, triode type electron tubes are used, but biasing methods are applicable to other types of electron tubes if the currents and voltages taken by the additional electrodes are properly considered.

2.2.1 **Cathode Bias.** When the cathode of an electron tube is biased positively with respect to the grid, the electron tube operates exactly as though an equivalent negative bias is applied to the grid. Since current flow within an electron tube is from cathode to plate, cathode resistor R_k can be inserted between cathode and B- to produce a voltage drop (cathode bias) as long as the plate current flows continuously; see figure 2-1A. Since cathode current always flows in the same direction, the voltage drop remains more positive at the cathode. Thus plate current flow within the electron tube itself produces a positive cathode bias. The amount of bias obtained depends upon the total tube current and the size of the cathode resistor $I_{rk} \cdot R_k$.

This type of bias is restricted in use to amplifiers in which plate current flows for more than half of each cycle of operation, because plate current flow cannot be cut off, or the developed bias will be lost. But it can be used in combination with a fixed minimum bias to limit maximum plate excursions or as a protective bias which limits plate current to a safe value when grid drive is removed. An inherent disadvantage of this circuit is that the developed bias voltage is subtracted from the plate voltage applied to the electron tube. Thus if 30 volts cathode bias is needed with a 250-volt plate supply, the power supply must be able to furnish 280 volts. See figure 2-1B. Its greatest advantage is that it is simple to use and economical of parts. When used in audio, video, or radio-frequency circuits, the cathode resistor must always be adequately bypassed to prevent a constant change of bias with signal. If the cathode is not bypassed, any change in plate current will produce a corresponding change in cathode bias voltage. The change in cathode voltage will be in such a direction as to oppose the effects of the input signal, and therefore will have the same effect as degenerative feedback. While a controlled amount of degenerative feedback can be beneficial in extending the over-all frequency response and in reducing distortion, a large amount of degenerative feedback will result in a serious loss of amplification. When the cathode bias resistor is adequately bypassed, the fluctuating a-c plate current caused by the input signal is effectively shunted around the cathode bias resistor, through a much lower reactance path offered by the bypass capacitor. Thus only the d-c component of plate current flows through the bias resistor, and the total cathode current remains constant at the initial static (d-c) value of no-signal current, and is unchanged by

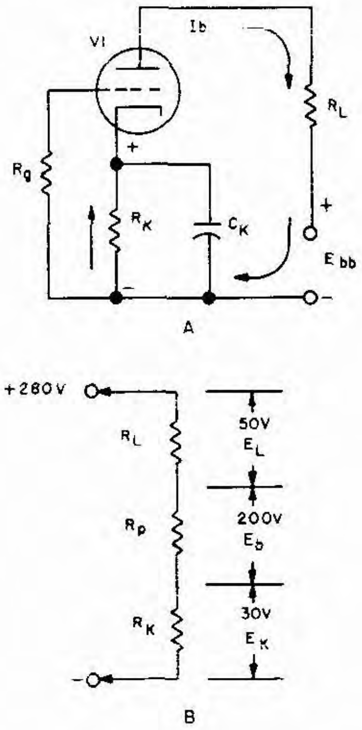


Figure 2-1
Cathode Bias Circuit

the varying input signal. For satisfactory bypassing, the bypass reactance should be about 10% of the d-c bias resistance at the lowest frequencies used. Cathode bias is usually used in audio-frequency or video-frequency amplifiers and in low power r-f amplifiers and test equipment. In some applications, such as high fidelity audio amplifiers or video i-f amplifiers, a relatively small value of unbypassed cathode resistance may be used. The degenerative action of this resistance increases the fidelity (frequency response) of the stage and reduces the possibility of overloading from strong signals.

2.2.2 **Grid-Leak Bias.** Grid-leak bias, sometimes called **signal bias**, is obtained by allowing grid current flow, produced by the a-c input signal, to charge an R-C network in the grid-cathode circuit. Two basic circuits are used to develop this form of bias—the shunt type and the series type. The methods of developing grid voltage in these circuits are similar, but the physical connections of the network components are different. See figures 2-2 and 2-3.

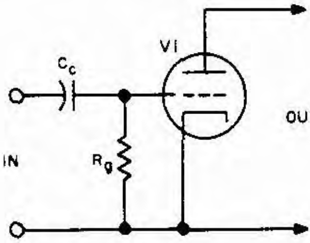


Figure 2-2.
Shunt Grid-leak Bias

The grid-cathode circuit is used as a diode rectifier to develop a d-c voltage that is proportional to the positive peak input (driving) signal amplitude. Grid-leak capacitor *Cc operates as a coupling capacitor* to apply the input (driving) signal to the grid. On the positive input signal excursions the grid is driven *positive causing grid current* to flow between grid and cathode and through grid-leak resistor R_g. The result is to produce a d-c voltage across R_g which is polarized negatively at the grid. The grid-leak capacitor, C_g, is effectively connected in series with the applied input signal, as shown by the equivalent

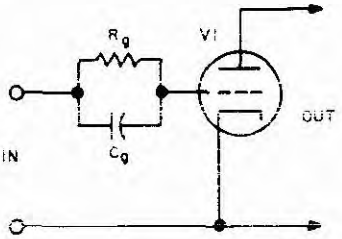
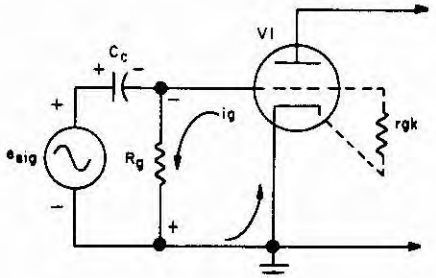
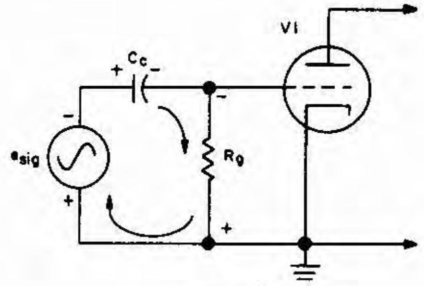


Figure 2-3.
Series Grid-leak Bias

charging circuit of figure 2-4A. As long as the input signal remains positive the coupling capacitor charges. When the input signal becomes negative, grid capacitor C_c begins discharging through the grid-leak (figure 2-4B); during the discharge period the grid is held negative by the charge remaining in the capacitor plus the negative input signal, and no grid current flows. The grid-leak time constant is made long with respect to the signal frequency (usually seven times the period for one input cycle) so that the discharge is relatively slow. The charge time is much faster than the discharge time because the grid leak is effectively shunted by the flow of grid current and offers a low resistance of about 500 ohms (r_{gk}, figure 2-4A) in comparison with the large resistance of R_g. When the signal begins its next positive excursion, C_c again starts to charge while still retaining a certain amount of residual charge. The cycle resumes and the action is repeated. After only a few more cycles, the grid bias stabilizes, since only a small portion of each charge leaks off before the application of the next positive half of the input signal. The residual charge on the coupling capacitor will not permit the next positive portion of the input signal to drive the grid as far positive as the positive portion did in the first half cycle. Thus the additional charge placed on C_c is not as great as it was for the first half cycle, but when added to the residual charge that remained, it causes the total bias to increase. This process of rapid charging and slow discharging of C_c leads to the steady state condition of operation, in which the amount of charge deposited on each positive half cycle exactly equals the amount of charge lost through discharge on each negative half cycle.



A
CHARGE CIRCUIT



B
DISCHARGE CIRCUIT

Figure 2-4.
Simplified Charge and Discharge Circuits

The amount of bias developed in this manner depends upon the amplitude and frequency of the applied input signal and the time constant of the grid network. The greater the amplitude of the input (grid-driving) signal, the greater the amount of bias that will be developed. Also, the longer the time constant of the grid circuit with respect to the input frequency, the greater the amount of bias that will be developed. However, there are practical limitations to the length of time constant which may be employed. Too long a time constant will prevent the bias voltage from changing enough during a cycle, or group of cycles, to bring the stage out of cutoff. Therefore, the stage must wait until the bias has been reduced enough to allow it to come out of cutoff; as a result, the stage will operate intermittently (motorboat). When functioning properly, a grid-leak bias network should produce a d-c voltage that is approximately 90% of the applied positive peak voltage. Under these conditions grid current will flow for only a small portion of the positive peak of the input cycle. Plate current, however, may flow for the entire cycle or for only part of the cycle, depending upon the amplitude of the input signal and the cutoff voltage of the electron tube.

Since the amplitude of the developed grid-leak bias is determined by the driving-signal amplitude, it is evident that Class A, B or C operation may be achieved with grid-leak bias merely by increasing the amount of drive (and by selecting the proper RC value). The limiting factor in this application is the amount of distortion that may be tolerated, because the plate current will be distorted for the portion of the cycle during which grid current flows. It is important to remember that grid-leak bias is developed by the rectification of the input (driving) signal and that loss of the input signal will result in a complete loss of this form of bias. In power amplifier circuits where loss of bias means excessive tube currents and possible damage, a second method of biasing (either fixed or cathode bias) is usually employed as a protective feature.

The shunt type of grid-leak circuit is also used to develop contact bias, which also derives its bias from the flow of grid current, but does not require an input or driving signal to produce it. Since the heated cathode is placed very close to the metal structure of the control grid, a small potential difference called **contact potential** is generated between the cathode and the control grid. Random electron collisions with the grid structure, aided initially by the electrically neutral character of the grid, are the major contributors to contact potential. In diodes a contact potential exists between plate and cathode. The contact potential may be on the order of 0.5 to 1 volt depending upon the structure of the tube, emission characteristics, etc. The 6AT6 and the 6SQ7 are typical examples of the tube types frequently operated with contact bias. The distinguishing feature of contact bias is the extremely high value of grid resistance used. This high value is required because of the minute currents and small potentials involved (with 10 megohms of grid resistance only one microampere of current will produce 1 volt of grid bias). Capacitor C_c (figure 2-2) isolates the grid from the preceding circuit as far as the d-c bias is concerned, but permits the application of an a-c signal to the grid. Contact bias is usually limited to circuits which operate over very small grid swings, say one volt peak to peak, such as an audio preamplifier, or a driven oscillator in which it is desired not to have dc flow through the tuned circuits. Circuits employing contact bias are sensitive to overloading, because any grid current flow due to excessive signal swing will block the tube with a "signal bias" which requires considerable time to leak off through the very long time constant grid network, R_gC_c.

The series grid-leak circuit employs a parallel R_gC_g combination in series between grid and cathode, as shown in figure 2-3. Usually grid resistor R_g is in the range of thousands of ohms (rather than megohms). This type of circuit is almost universally employed in the self-excited type of oscillator because of its self-regulating and self-starting action. Once the operating point is fixed by the value of R_g and C_g employed, the tube continues to oscillate about the grid operating point whether it be Class A, B, or C. If the amplitude of oscillations in the grid tank tends to increase, an increase in positive grid drive will occur causing the grid to draw more current. The increase in grid current will develop an increased bias and a consequent decrease of plate current which tends to reduce the plate current pulses back to their original amplitude. A similar sequence of events occurs if the grid drive tends to decrease, but in this event less bias is developed, so that the plate current pulses are increased, returning them toward their original amplitude. Bias is initally developed by the contact potential method during the first few oscillations, allowing the circuit to be self-starting. Then as the grid swings become greater, the capacitor is charged and discharged at the repetition rate of the tuned circuit, and the grid bias is determined by the time constant of the grid circuit and the value of the grid-voltage swing at that instant, averaged over a number of cycles of oscillation. Thus, if the grid capacitor is increased in value, the grid-leak resistance must be reduced to retain the desired charge and discharge rates. When the grid capacitor is sufficiently large and the grid leak resistance is adequate, quite a large voltage can be produced, and if the discharge rate is slow enough, plate current cutoff can be obtained to produce Class B or C operation. Thus this bias method is ideal for oscillators which are self-excited (produce their own feedback). In separately driven oscillators (buffer amplifiers), the shunt (signal) bias method is used, and the bias voltage required is obtained by supplying sufficient grid current drive from the preceding stage.

2.2.3 **Fixed Bias.** There are a number of methods of obtaining fixed bias. The most obvious of these methods is the use of a separate C-battery or a separate negative power (C-) supply. Two other widely used, but not so obvious, methods are shown in figures 2-5 and 2-6. The advantage of these methods is that they both use the equipment power supply to produce the fixed bias, so that no separate supply is necessary.

The circuit of figure 2-5 utilizes poteniomenter R2 connected in series with resistors R1 and R3 to form a voltage divider between B+ and B-. The cathode of the tube (s) being biased is connected to the moving contact of R2, and the grid is returned through R_g to B-. Resistors R1 and R3 determine the voltage range over which potentiomenter R2 operates. Thus for cutoff bias a positive cathode bias is produced by the voltage divider. For a bias less than cutoff the cathode current passing through R3 and the active portion of R2 determine the operating bias. Bypass capacitor C_k shunts the portion of the resistance affected by cathode current to prevent bias changes with signal current variations, as explained in the discussion of cathode bias in paragraph 2.2.1.

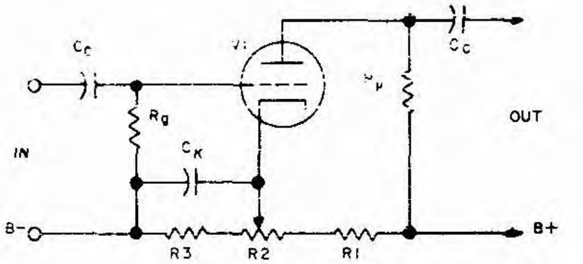
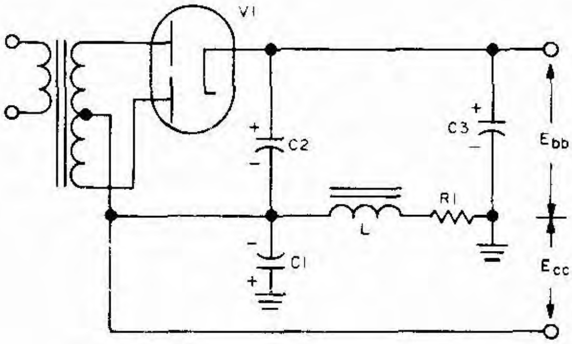
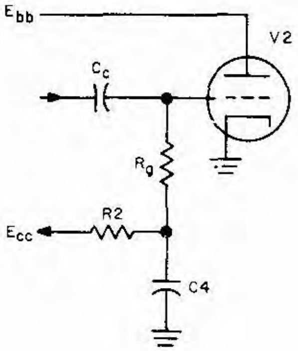


Figure 2-5.
Fixed (Voltage-Divider) Bias

The circuit of figure 2-6 uses the entire current flow of the power supply to develop a fixed "back-bias". The term **back-bias** refers to a combination of self (cathode) bias and fixed bias from another source. In this instance the self-bias is obtained from the cathode current of the tube being biased, and the fixed bias is obtained from the total power supply current flowing back through the chassis and the fixed bias resistor connected in series with the negative supply lead and chassis. There are two variations to this circuit, one using resistor R_1 alone and the other using the d-c resistance of the loudspeaker field coil L (or a separate choke coil) to act as R_1. Use of the latter variation results in an economy of parts since the field coil also provides the bias. The back-bias circuit produces its result by virtue of the connection of the negative B supply lead to the chassis through field or choke coil L and resistor R_1, when used. Since all of the tubes on the chassis have their cathodes grounded, the entire chassis current passes through the choke (and resistor) to B-. The voltage drop across the choke coil (and R_1) is the bias. By utilizing a voltage-divider arrangement at each grid, the other stages may also be back-biased from the same source. If more bias is required, more current is returned through the chassis, or resistor R_1 is added to produce the required voltage drop. It is usually necessary to use a decoupling filter (R2, C4 figure 2-6B) to prevent power supply ripple voltage from appearing on the grid. Where more than one stage uses back bias (E_{cc}) a decoupling filter at the grid of each stage is necessary to prevent signal voltage fluctuations on the grid of one stage from affecting operation of the other stages through common impedance coupling. Where more than one



A
BACK BIAS CIRCUIT



B
DECOUPLING CIRCUIT

Figure 2-6.
Back-Bias Circuit

value of fixed bias is necessary to suit the requirements of the individual stages a voltage divider arrangement may be used. In the case of equipments requiring large negative bias voltages a separate negative power supply is usually used to obtain the bias.

2.3 CLASSES OF AMPLIFIER OPERATION.

Amplifier operation is divided into three general classes: Class A, Class B, and Class C. The classes are determined by the operating bias applied, the amplitude of the input signal, and the amount of time during the operating cycle that plate current is allowed to flow.

These general classes of operation cover a large range of operation; since it is possible to operate an amplifier partially within one class and partially within another, additional designations are added to indicate whether or not grid current flows. When grid current does not flow at all during the entire cycle of operation, the suffix (or subscript) 1 is added to the class letter; when grid current flows, for even a portion of the cycle, the number 2 is added. Thus Class AB1 operation indicates that the circuit is biased to operate somewhere between the limits of Class A and Class B operation and that no grid current flows during the operating cycle. Similarly, Class AB2 operation indicates that the circuit is biased to operate somewhere between the limits of Class A and Class B operation, and that the input signal exceeds the bias sufficiently to produce grid current flow over a portion of the cycle.

Class A operation produces the lowest power conversion efficiency and the least amount of distortion, while Class C produces the greatest possible power efficiency with the most distortion, with Class B being intermediate between the two values. When no grid current is drawn during the operating cycle, there is no power loss in the grid circuit, and a minimum of driving power is required. When grid current is drawn, a power loss occurs in the grid circuit, and the driver stage must be able to supply power at the required maximum drive voltage.

2.3.1 **Class A Operation.** In normal Class A operation, the electron tube is operated on the linear portion of the grid-plate transfer characteristic curve, and plate current flows continuously over the entire input cycle. Because operation is on the linear portion of the curve, distortion is low (on the order of 2 to 3 percent maximum) and voltage gain is high. Because of plate current flow over the entire cycle, plate efficiency is extremely low (on the order of 20 to 30 percent) and power output is low.

Class A operation is usually employed where voltage amplification rather than power output is desired, or where low power output is desired with a minimum of distortion. Thus it is used in the r-f and i-f stages of receivers, in low-power test equipment, in the oscillator and driver stages of low-power transmitters, in video amplifiers, and in audio-amplifier applications where voltage gain with good fidelity is desired.

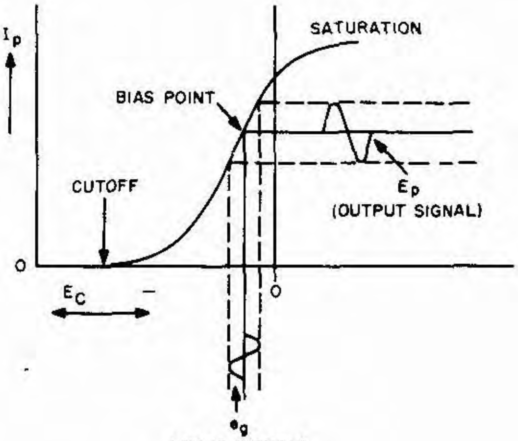


Figure 2-7.
Class A Operation

Normally, no grid current flows during any part of the cycle and pure Class A operation is defined as A_1. However, in common usage the symbol stands alone, the subscript is dropped, and it is understood that unless otherwise *indicated no grid current flows.* When the input signal amplitude exceeds the bias voltage and grid current does flow over a portion of the cycle, the subscript 2 is used, e. g., A_2. When grid current flows, a larger input signal is required to drive the tube to the same output, and distortion increases; but at the same time, plate current flow reduces during the grid-current portion of the cycle and the overall plate efficiency is increased. Figure 2-7 shows a graph of A_1 operation, and figure 2-8 shows A_2 operation.

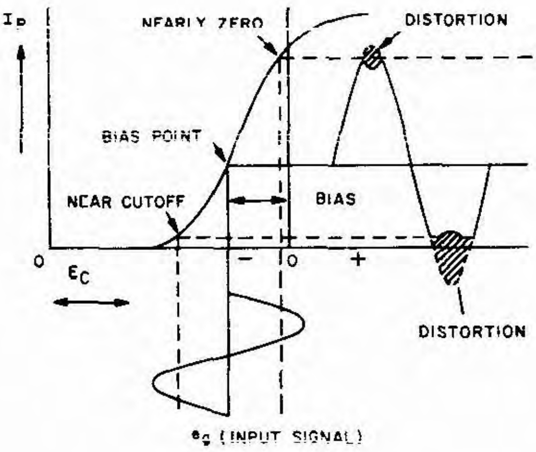


Figure 2-8.
Class A_2 Operation

2.3.2 **Class B Operation.** Biasing of Class B amplifiers is such that, with no grid input signal, the plate current is biased to cutoff; consequently, no plate current flows for approximately half the operating cycle (180°), and grid drive is sufficient to cause grid current flow. Since plate current flows for only half the cycle, it is necessary to use two tubes in a push-pull arrangement to reproduce the full positive and negative input swings, and to minimize distortion. An exception to this rule is in an r-f amplifier where the missing half of the signal is supplied by the tank circuit, and the distortion can be tolerated.

Class B operation is normally characterized by moderate efficiency (40 to 60 %) in the plate region, with low driving power, and somewhat greater output distortion (4 to 6%). Figure 2-9 illustrates typical Class B operation for a single tube.

Class B amplifiers are used extensively for audio output stages where high-power outputs are required; they are also used as the driver and power amplifier stages of transmitters and as audio modulators.

Special tubes have been developed, particularly for Class B operation, which normally rest at plate current cutoff without bias applied. These tubes, together with improved transformer design, make it possible at the present state of the art to design Class B stages which compare favorably in performance with Class A stages, and yet provide the increased power output and efficiency of Class B operation.

2.3.3 **Class AB Operation.** In this type of operation, the bias is set between Class A and Class B so that the amplifier operates Class A for small input signals and Class B (really Class AB_1) for large input signals. Plate

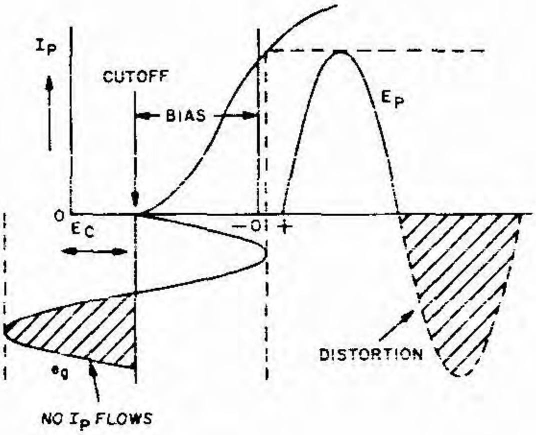


Figure 2-9.
Class B Operation

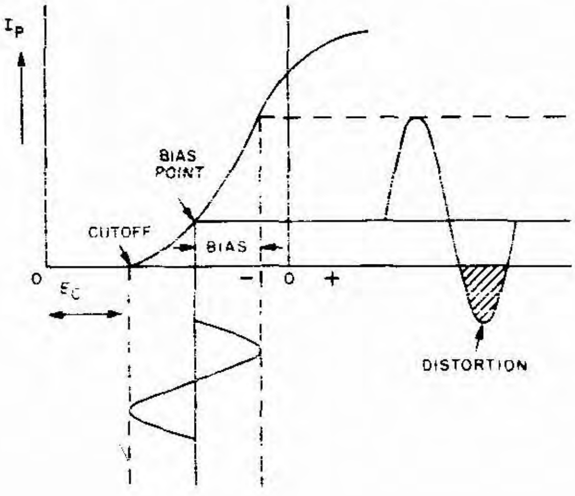


Figure 2-10.
Class AB Operation

current flows for more than half the operating cycle, but not for the entire cycle. In addition, the input signal, although just about equal to the bias, is not great enough to produce grid current flow.

This type of operation is used where good fidelity and increased efficiency are desired but without sufficient drive to produce full Class B operation or excessive distortion.

As can be seen from figure 2-10, the bias for Class AB operation is about halfway between the values for Class A and cutoff. The power output and efficiency are higher than with Class A operation, but at the expense of more distortion. To reduce the effective distortion, Class AB audio amplifiers are operated push-pull, in which case the efficiency is much greater than that of Class A operation with about equal distortion.

Class AB$_2$ amplifiers require low-impedance grid drive sources because of grid-current flow and, therefore, are often transformer-coupled to power drivers. This class of operation provides power output and efficiency ratings more nearly equal to those of Class B operation.

2.3.4 **Class C Operation.** In Class C operation, the bias is adjusted to twice the cutoff value or more (usually does not exceed four times cutoff value). Plate current flows only on the peaks of the driving signal, the grid is always driven positive, and extremely high efficiency is produced (70 to 85%). Because of the high distortion produced, Class C has not been found useful for audio-frequency-amplifier operation. But it has proven excellent for use in r-f amplifiers, where tank circuits provide the missing portion of the signal, and extremely high powers can be conveniently and efficiently handled. The fly-wheel effect of the tuned tank circuit serves to smooth the intermittent pulses of plate current into sine-wave oscillations in the tank circuit.

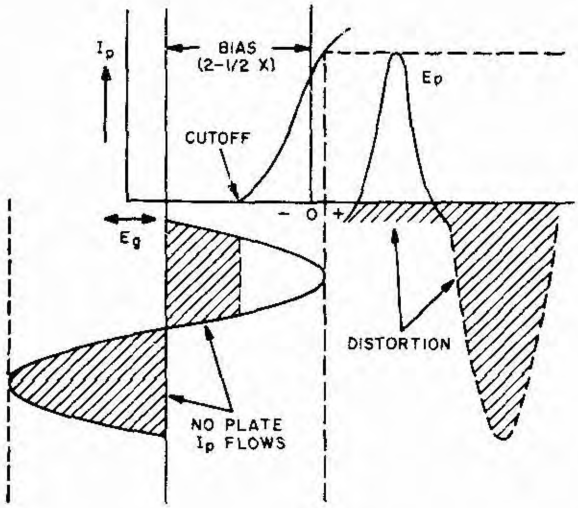


Figure 2-11.
Class C Operation

Since grid current always flows, this type of operation should be designated as Class C$_2$ but, being generally understood, this notation is usually never used. Figure 2-11 illustrates Class C operation.

2.4 COUPLING METHODS.

Usually more than one amplifier, operated in cascade, is needed to increase the amplitude of the feeble input signal to the required output value. Cascaded amplifier stages are connected (coupled) together by resistance-capacitance networks, impedance-capacitance networks, transformers, or direct coupling as described in the following paragraphs. While all coupling networks are frequency-responsive, some coupling methods provide better response than others, and their basic principles of operation remain the same, regardless of whether or not they are employed singly as input or output coupling devices, or in cascade. Discussion in this section will be limited to the actual types of coupling and important considerations involved in the various functional classes on a relative or comparative basis. Circuit discussions in other sections will fully cover the limiting parameters for the specific circuit arrangement.

2.4.1 **R-C Coupling.** Although R-C coupling involves the use of two resistors and a capacitor, as shown in figure 2-12, it is usually referred to as resistance coupling. Because of the rather wide frequency response offered by this form of coupling, plus small size and economy, the resistance coupler finds almost universal use where voltage amplification is required with little or no power output. Basically the resistance coupler is a pi-type, hi-pass network, with plate resistor R$_L$ and grid resistor R$_g$ forming the legs of the pi and capacitor C$_c$ the body. Since the plate and grid resistors are not frequency-responsive, it can be seen that basically over-all frequency response is limited by the capacitive reactance of C$_c$ between the plate and grid circuits, plus the effect of shunt wiring and electrode-to-ground capacitances across the network. At d-c or zero frequency the coupling capacitor separates, or blocks, the plate voltage of the driving stage from the grid bias of the driven stage, so that bias ana plate or element voltages are not affected between stages.

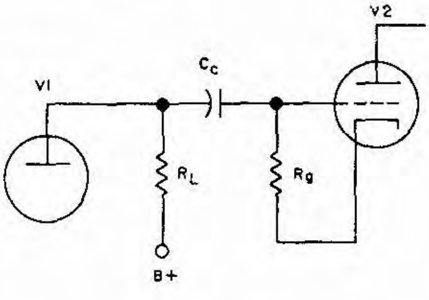
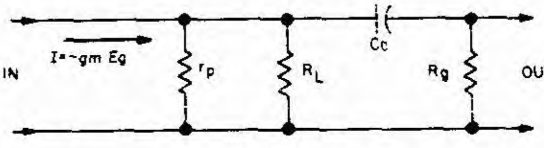


Figure 2-12.
R-C Coupling

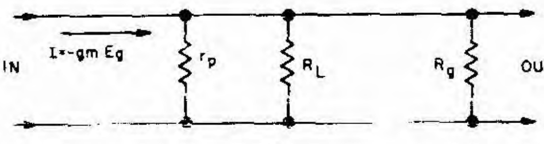
In the conventional R-C amplifier signal voltage variations on the grid produce plate current variations through plate load resistor R_L, and the resulting voltage developed across R_L represents an amplified replica of the input signal but 180 degrees out of phase. The amplified signal is coupled through capacitor C_c, and applied to the grid of the next stage across grid load resistor R_g. The same cycle of operation is repeated for each stage of the cascaded amplifier.

Since load resistor R_L is in parallel with the internal tube plate resistance, r_p, the load resistor of a triode is never made less than twice r_p. Usually the maximum undistorted output for a triode is obtained with a value of R_L that is 5 to 10 times the plate resistance of the tube. With pentodes the extremely high r_p makes *this impractical;* best results are usually obtained with a load resistor of 1/4 to 1/10 the value of r_p, or with a value that is as high as is practical for the plate voltage and current needed without requiring excessive supply voltage.

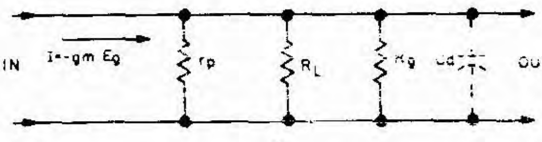
At low frequencies (below 100 cps), the reactance of C_c, plus R_g in series, parallels the load resistor. But the reactance of C_c is in series between the plate and grid, and produces a large voltage drop (or loss of gain), while R_g shunts the grid and has the least voltage developed across it. Therefore, the interstage grid resistance has less over-all effect on the gain than the reactance of C_c, and the distributed circuit and electrode capacitance to ground is not effective and may be neglected. Low-frequency response, therefore, is controlled by the size of the coupling capacitor. A simplified equivalent low-frequency circuit (constant current generator form) is shown in figure 2-13A. Only those circuit elements which are of significance at low frequencies are shown. When the coupling capacitor is excessively large, distortion is caused by excessive phase shift, and motorboating may occur. Motorboating is caused by regenerative feedback at low frequencies through impedance coupling in multiple cascaded R-C stages with a common power supply.



A
LOW FREQUENCY



B
MID FREQUENCY



C
HIGH FREQUENCY

Figure 2-13.
Simplified Equivalent Circuits
(Constant Current Generator Form)

Over the mid-frequency range (100 to 20,000 cps), amplification is relativly constant, and the coupling capacitor reactance is much lower than the resistance of grid resistor R_g. Therefore, the reactance of coupling capacitor C_c may be neglected as in the simplified equivalent mid-frequency circuit of figure 2-13B. *To insure that* the plate resistor is not shunted excessively by the grid resistance of the following stage and to maintain a high input resistance, R_g is made two to four times R_L and never less than R_L. Usually in low-level stages the grid resistor is in the megohm range (from 1 to 5 megohms), with its value for grid leak bias ranging from 5 to 10 megohms; whereas in power amplifier stages the grid resistor is on the *order of* 0.5 megohm and usually never more than 1 megohm. High values of grid resistance tend to cause adverse effects due to the possibility of grid current flow caused by imperfect evacuation, grid emission, and leakage effects in the electron tube.

In the high-frequency range (over 20,000 cps), the coupling capacitor reactance is neglected entirely and shunt capacitive effects become the limiting parameters. The shunt capacitance consists of the plate-to-ground (tube output) capacitance, the grid-to-ground (input) capacitance, plus the distributed wiring and network capacitance to ground. The effects of these individual capacitances are cumulative and may be lumped together, as shown by C_d in figure 2-13C. The over-all effect of this capacitance is to shunt the signal to ground and reduce the response.

In high-fidelity amplifier circuits, increased low-frequency response is usually desired and is achieved by bass compensating networks. Likewise, in video amplifiers both low- and high-frequency response must be increased, and is achieved through the use of shunt and series peaking circuits. See paragraph 2.5 for a discussion of R-C and R-L compensation considerations.

Because of the low amount of gain available with R-C coupling at radio frequencies, it is seldom used in r-f amplifiers, but it may be encountered in special cases, particularly in test equipment.

2.4.2 **Impedance Coupling.** When an impedance is substituted for the plate load resistor in a resistance-coupled circuit, the impedance coupling circuit results. Actually, an original circuit derivation involved a substitution of an impedance for the grid resistor also by making the grid inductor series resonant with the coupling capacitor, bass response was improved. However, the grid impedance offered more of a disadvantage because of high-frequency shunt losses, so the L-C-R circuit as shown in figure 2-14 became the standard impedance coupler. The impedance coupler operates in the same manner as the resistance coupler as far as C_c and R_g are concerned; the basic difference is in effect of the plate impedance. By using an impedance in the plate circuit, less voltage drop occurs for a given voltage supply. Therefore, a lower supply source will provide the same effective plate voltage, there is less I^2R (power) loss, and a better over-all efficiency results. Low-frequency

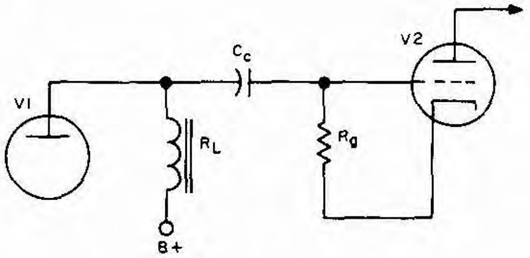


Figure 2-14.
Impedance Coupling

response is dependent upon obtaining a high inductive reactance in the plate circuit, and requires a large number of turns for good low-frequency response. The distributed capacitance associated with a winding of many turns produces a large shunting capacitive reactance with a consequent drop in high-frequency response. Since the impedance of plate reactor varies with frequency, the response is not as uniform as that of the resistance coupler.

The impedance coupling circuit is used where a limited response over a relatively narrow band of frequencies is required. As a result, impedance coupling is mostly employed in tuned or untuned amplifier applications such as i-f or r-f stages, and its use in audio applications has generally been discontinued in favor of the R-C or transformer-coupled circuit, except for special designs.

As in the resistance-coupled amplifier, series and shunt peaking compensating networks may also be employed to extend frequency response. Bass boost circuits in particular can be advantageously used. However, as frequency compensation is increased, the circuitry becomes complex and its use is restricted to a particular application, so that the basic type of coupling circuit is no longer of much significance.

2.4.3 **Transformer Coupling.** When the primary of a transformer is connected as the plate load, and the secondary provides the output signal, either to the next stage or an output device, we have what is known as transformer coupling; see figure 2-15.

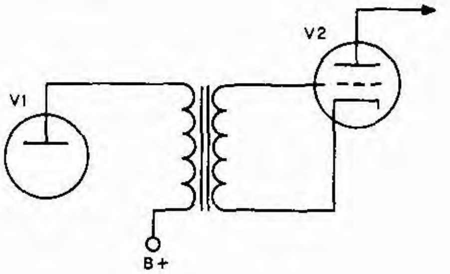


Figure 2-15.
Transformer Coupling

With transformer coupling, response, gain, and output considerations become more difficult to predict, because they depend primarily on the transformer design. Basically, the use of transformer coupling provides additional gain, achieved through the use of a step-up turns ratio of primary to secondary, but this gain usually does not exceed 2 or 3 to 1. Since there is no physical d-c connection between stages, plate and bias voltages are kept separate, and the a-c signal is coupled from the plate of one stage to the grid of the following stage by the mutual inductive coupling between primary and secondary windings.

Low-frequency response is primarily dependent on the inductive reactance of the primary, dropping off approximately 3 db at the frequency where the inductive reactance is equal to the plate resistance plus the primary resistance ($2\pi fL = R_p + R_{pri}$).

Gain at the mid frequency is the highest, and is essentially equal to the amplification factor of the electron tube multiplied by the transformer turns ratio. Response is considered to be relatively flat over a range of frequencies above and below the mid frequency, but such results are usually achieved only in the ideal case. In practice, the response curve is a continually changing curve, dropping off on either side of the mid frequency, with possible humps produced by circuit and transformer resonances.

Since the secondary contains more turns than the primary, a larger shunt capacitance to ground is produced and, together with the primary and secondary leakage reactances, limits the high-frequency response.

Transformer coupling is generally used for interstage applications with electron tubes having plate resistances of 5 to 10 thousand ohms maximum, since higher plate resistances require excessively large transformer primary inductances. For output stages, lower plate resistances are used, and the transformer is carefully designed to handle larger plate currents. Generally speaking, a lower plate resistance improves bass response, while a higher amplification factor provides greater gain, and lower plate current produces less d-c core saturation effects.

Since the impedance transformation in a transformer varies as the square of the turns ratio between primary and secondary, output and input matching is possible and is extensively employed. For interstage applications, matching is not always used, because power output is not required; in this case, more attention is given to the step-up ratio to provide a higher voltage gain.

The limitations of frequency response generally restrict the use of transformer coupling to audio circuits which do not require an exceptionally wide bandpass or frequency response, but do require voltage or power outputs. Wide use is found for transformer coupling in the radio and intermediate frequency ranges where selective, high-Q bandpass filters and tuned transformers are universally used. See Tuned Interstage IF Amplifiers Circuit in Section 6, for a discussion of tuned transformer coupling.

2.4.4 **Direct Coupling.** Direct coupling is characterized by the direct connection of tube elements; that is, the plate of the driver stage is physically connected to the grid of the driven stage, and the coupling network is eliminated. A basic two-stage d-c amplifier is shown in figure 2-16.

Since the plate and bias circuits are not isolated by a transformer or coupling capacitor the direct-coupling circuitry

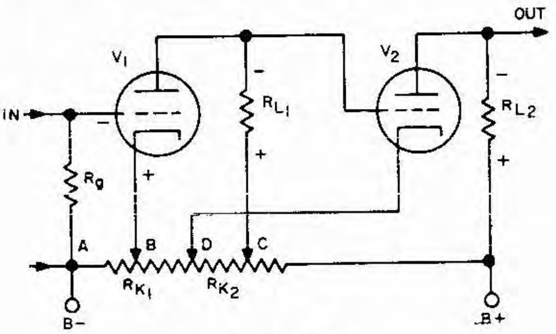


Figure 2-16.
Direct-Coupled Amplifier Circuit

is slightly complicated by the arrangement necessary to produce an effective negative bias. Usually a voltage-divider bias arrangement similar to that shown in figure 2-16 is used.

Examining figure 2-16, it is evident that the cathode of V1 is biased by R_{k1} between points A and B on the divider, while the plate is connected through R_{L1} to point C, which is at a potential equal to approximately half the supply voltage. Plate current through load resistor R_{L1} (which is also the grid resistor for V2) provides a negative voltage drop and bias on V2. The cathode of V2 is tapped back at point D on the divider, and produces a voltage in opposition to the drop across R_{L1}. When R_{k2} is correctly adjusted, the voltage on the cathode of V2 is always more positive than the plate of V1 (and the grid of V2), producing an effective negative bias on V2 of the desired value. Voltages must be chosen so that the grid of V2 is not driven positive by the input signal. Since the plate of V2 is at the highest positive point, it is more positive than the cathode and conduction occurs. Thus the plates are maintained positive and the grids negative, to provide the conditions required for operation.

Because there is no coupling network inserted between the output of one tube and the input of the following tube, there is no phase distortion, time delay, or loss of frequency response. Since the plate and grid of the tubes are directly connected, the low-frequency response is extended down to dc (zero frequency). The high-frequency response is limited only by the tube interelectrode-to-ground capacitance, plus the circuit distributed wiring capacitance. By appropriate matching (or mismatching) of tubes, high values of amplification and power output may be obtained.

Since the use of more than two stages requires plate voltages two or more times the normal value for one tube, plate supply considerations limit d-c coupling to a few stages. Any change in the supply voltage affects the bias of all the tubes and is cumulative; therefore, special voltage supply regulation circuits are necessary. Noise and thermal effects in tubes produce circuit instability and drift that limit the use of this type of coupling in audio or r-f amplifiers.

Because of its ability to amplify direct current or zero frequency, d-c coupled circuitry is often used in computer circuits, and in the output circuits of video amplifiers. Because response is practically instantaneous and no time delay occurs, it is especially valuable for pulse circuits. D-C amplifiers are discussed in greater detail in Section 6, see Direct Coupled (D-C) Amplifier circuit.

2.5 TIME CONSTANTS.

In the previous paragraphs on coupling circuits, it has been shown that lumped inductance, capacitance, and resistance are used to couple together various electronic circuits. It has also been shown that these basic components are affected in various ways when responding to signals of different frequencies. Properly connected they can be used to shape, control, or distort signal waveforms. Since R-C and L-R combinations contain lumped capacitance or inductance, they always require a finite time to charge or discharge, and thus provide a simple arithmetical figure (time constant) which is useful in the discussion of the operation and performance of these circuits.

2.5.1 **R-C Circuits.** The time constant (TC) of an R-C circuit is defined as the time in seconds that is required to charge (or discharge) the R-C network to an arithmetical value of 63.2 percent. As normally used, it is defined as the time it takes the charge to reach 63.2 percent of the maximum charge, or to discharge 63.2 percent and to reach 36.8 percent of the initial (maximum) value. To obtain the time constant, the value of the resistance in ohms is multiplied by the value of the capacitor in farads (or megohms and microfarads); that is, TC = R x C.

Consider now the action of a series R-C circuit when a voltage is applied. The capacitor charges heavily in an exponential manner, following the curve of figure 2-17. At the end of a period of time equal to one time constant the voltage across the capacitor reaches 63.2% of the maximum value of voltage applied. If the charge is continued, at the end of the second time constant interval it rises 63.2% of the value remaining at the end of the first time constant (36.8%), to reach a new value of 86.4% maximum voltage (23.2% remaining). Thus at the end of five time constants the voltage has reached a value of 99.3% of maximum, or almost full charge. Theoretically, the condition of complete charge (or discharge) will not be obtained but the difference is so infinitesimally small after five time constants that the remaining value can be neglected.

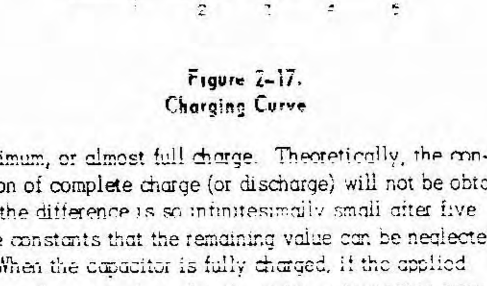


Figure 2-17.
Charging Curve

When the capacitor is fully charged, if the applied voltage is removed and the capacitor is connected across

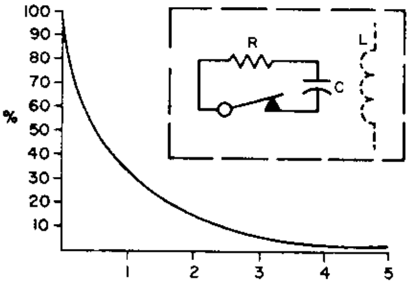


Figure 2-18.
Discharge Curve

the resistor, it begins to discharge through the resistor at an exponential rate. Discharge occurs as shown by the curve, figure 2-18. At the end of five time constant intervals of discharge the voltage across the capacitor is approximately zero.

Comparing the illustrations showing the charge and discharge, it is evident that the curves are identical in shape, but that one is the reverse of the other. That is, while it takes one time constant interval to reach 63.2% of the applied voltage (or charge) with 36.8% to go, conversely, it also takes one time constant interval to reach 36.8% of the initial capacitor voltage (63.2% discharge). The vertical portion of each illustration represents the amount of voltage, current, or capacitor charge in percent.

A time constant interval may be considered short when the RC product is equal to, or less than, 1/10 of the period of the applied voltage. Likewise, a time constant may be considered long when the RC product is equal to, or greater than, ten times the period of the applied (pulse) voltage.

Short time constant circuits are employed in differentiating circuits which change a square wave into positive and negative pips. Conversely, long time constant circuits are use to reproduce a square wave with fidelity, or to integrate (sum up) a series of pulses and change them into a single sawtooth, rectangular, or square wave. The figure below briefly illustrates the signal-shaping effects of long and short time constant circuits. For a complete discussion of the waveshaping effect of the R-C circuit, refer to Section 17 of this technical manual.

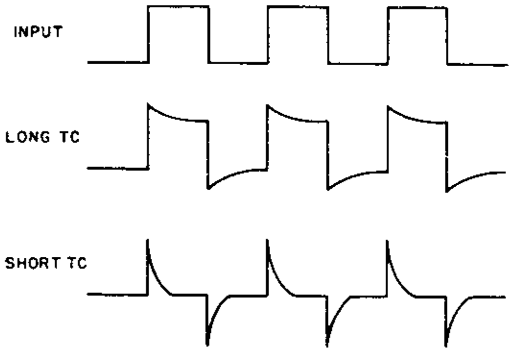


Figure 2-19.
Waveshaping Effects of RC Circuits

Time constants are sometimes used in audio coupling networks to define the low-frequency response. For example, the frequency at which the response of a single resistance-coupled stage falls 3 db (theoretical cutoff frequency) is:

$$f_o = \frac{1}{2\pi RC}$$

Therefore, if the time constant of a network is 3000 microseconds, the low-frequency response limit of the network is found as follows:

$$f_o = \frac{1}{2\pi 3000 \times 10^{-6}}$$

Time constant is also related to frequency by the formula

$$F = \frac{1}{TC}$$

where F is in cycles per second, and TC is in seconds. Thus a time constant of .01 second corresponds to a frequency of 100 cycles per second. Since TC equals R times C, this time constant could be produced by a 0.01-μf capacitor and a 1-megohm resistor (or 0.1 μf and 100K).

R-C circuits are also used as decoupling networks, feedback networks, and high-and low-pass filters. Figure 2-20 shows a few basic R-C circuit arrangements which illustrate possible combinations.

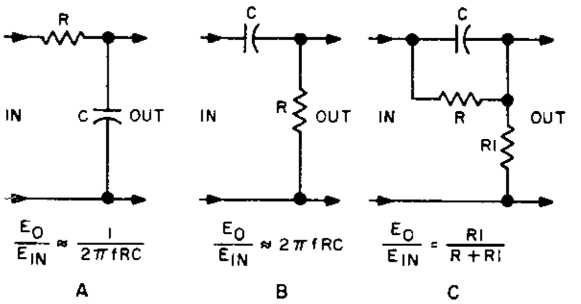


Figure 2-20.
R-C Circuits

Part A of figure 2-20 illustrates a typical low-pass filter arrangement, where the low frequencies are passed without attenuation, but the higher frequencies are attenuated. Input and output relationships are shown by the formula, and for a specific time constant vary directly as the capacitive reactance. Part B shows a typical high-pass circuit which passes the higher frequencies and attenuates the low frequencies. Here the attenuation increases with a decrease of frequency, with greater output for the higher frequencies. Part C shows a parallel R-C network which is inserted in series with the coupled stages; it functions to bypass the higher frequencies across the resistor, with R and R1 acting as a voltage divider at the lower frequencies.

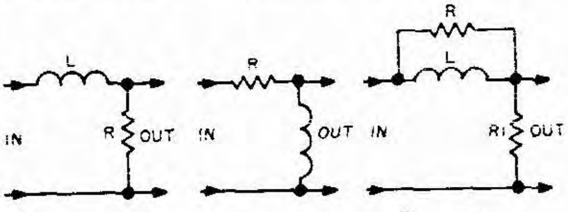
Since most electron tube elements are biased through a series resistor bypassed by a capacitor, it is evident that quite a number of time constant circuits exist in any circuit configuration. The effect that the time constant has with regard to frequency response and time delay becomes an important factor in the design of pulse circuits. See Sections 17 and 23 for a complete discussion of these factors.

2.5.2 **R-L Circuits.** The time constant (in seconds) of a circuit containing an inductor and a resistor connected in series can be found by dividing L (henrys) by R (ohms), $TC = \frac{L}{R}$

If a d-c voltage is applied to a series L-R circuit, the output current does not rise instantaneously to a maximum value at the instant the voltage is applied. The rise of current is slowed down because the inductance produces a self-induced counter voltage which opposes the applied voltage, causing the inductance to oppose any change in current flow. Thus, current flow begins at a zero rate and increases exponentially as the self-induced counter emf decreases. When a voltage is applied to an R-L circuit, the voltage across the inductor varies as shown in figure 2-18. At the end of five L/R time intervals, the counter voltage across L drops to approximately zero volts. During the same time interval the voltage drop across R, which results from the current flow in the circuit, begins at zero volts and increases at an exponential rate until, at the end of five time constant intervals, the voltage across R nearly equals the applied voltage (similar to figure 2-17).

When the applied voltage is removed, the inductor again opposes the change in current flow by attempting to keep the current flowing. The counter emf now produced causes the current to decrease at an exponential rate. The voltage across R (and across the inductance) during the decay period, shown in figure 2-18, drops to approximately zero volts at the end of five time constant intervals.

Figure 2-21 shows the circuits for basic L-R networks similar to the R-C networks of figure 2-20. Operation is identical for the comparable types of networks.



$$\frac{E_O}{E_{IN}} \approx \frac{1}{2\pi f L/R} \qquad \frac{E_O}{E_{IN}} \approx 2\pi f L/R \qquad \frac{E_O}{E_{IN}} = \frac{R1}{R+R1}$$

Figure 2-21.
L-R Circuits

Part A of figure 2-21 illustrates a typical low-pass filter arrangement. Since the reactance of L increases directly with frequency, the higher frequencies are attenuated and the lower frequencies are passed. Part B shows a typical high-pass circuit, which develops a larger voltage across the inductive reactance at the higher frequencies than at the lower frequencies. In part C the inductor acts as a variable shunt for R at the lower frequencies and as a voltage divider with R1 at the higher frequencies.

Since the inductor has distributed (turns) capacitance across it, undesired resonant responses may occur in L-R circuits containing large values of inductance; therefore, the use of these networks is usually limited to high-frequency applications. A practical application in resistance-coupled circuits is the insertion of a small inductance in series with the plate resistor or in series with the grid coupling capacitor to improve video response. These circuits are called shunt and series peaking circuits, respectively, and are illustrated in figure 2-22.

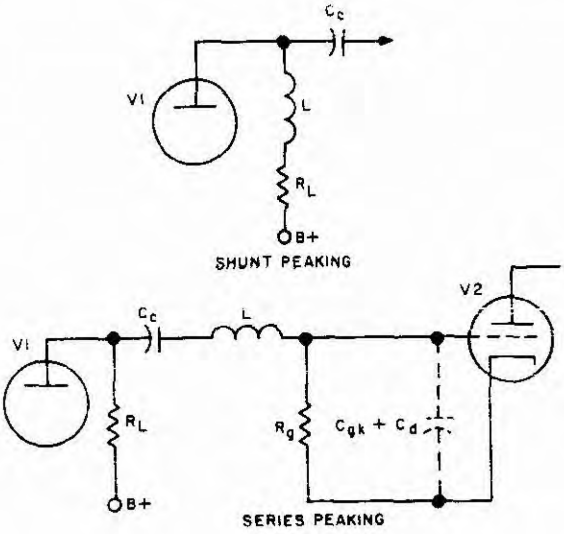


Figure 2-22.
Shunt and Series Peaking Circuits

In the shunt peaking circuit, the increased inductive reactance at the higher video frequencies produces an effectively higher load impedance and larger output signal for coupling to the next stage. When the peaking inductance value is selected to resonate with the shunt capacitance to ground of the electron tube, a parallel resonant circuit is obtained which also boosts amplification at and around the resonant frequency. The series peaking inductance is usually selected to produce a series resonant circuit with the grid-to-ground tube and circuit capacitance. By combining both shunt and series peaking in a stage, greater high-frequency response is obtained. Excessive compensation, however, may cause unwanted transient responses. See section 6, Video Amplifiers Circuit, for a more complete discussion of peaking circuits.