12.1 Tubes

In 1883, Edison discovered that electrons flowed in an evacuated lamp bulb from a heated filament to a separate electrode (the Edison effect). Fleming, making use of this principle, invented the *Fleming Valve* in 1905, but when DeForest, in 1907, inserted the grid, he opened the door to electronic amplification with the *Audion*. The millions of vacuum tubes are an outgrowth of the principles set forth by these men.[1]

It was thought that with the invention of the transistor and integrated circuits, that the tube would disappear from audio circuits. This has hardly been the case. Recently tubes have had a revival because some *golden ears* like the smoothness and nature of the tube sound. The 1946 vintage 12AX7 is not dead and is still used today as are miniature tubes in condenser microphones and 6L6s in power amplifiers. It is interesting that many feel that a 50 W tube amplifier sounds better than a 250 W solid state amplifier. For this reason, like the phonograph, tubes are still discussed in this handbook.

12.1.1 Tube Elements

Vacuum tubes consist of various elements or electrodes, Table 12-1. The symbols for these elements are shown in Fig. 12-1.

Table 12-1. Vacuum Tube Elements and Their Designation

filament	The cathode in a directly heated tube that heats and emits electrons. A filament can also be a separate coiled element used to heat the cathode in an indirectly heated tube.
cathode	The sleeve surrounding the heater that emits electrons. The surface of the cathode is coated with barium oxide or thoriated tungsten to increase the emission of electrons.
plate	The positive element in a tube and the element from which the output signal is usually taken. It is also called an *anode*.
control grid	The spiral wire element placed between the plate and cathode to which the input signal is generally applied. This element controls the flow of electrons or current between the cathode and the plate.
screen grid	The element in a tetrode (four element) or pentode (five element) vacuum tube that is situated between the control grid and the plate. The screen grid is maintained at a positive potential to reduce the capacitance existing between the plate and the control grid. It acts as an electrostatic shield and prevents self-oscillation and feedback within the tube.
suppressor grid	The grid like element situated between the plate and screen in a tube to prevent secondary electrons emitted by the plate from striking the screen grid. The suppressor is generally connected to the ground or to the cathode circuit.

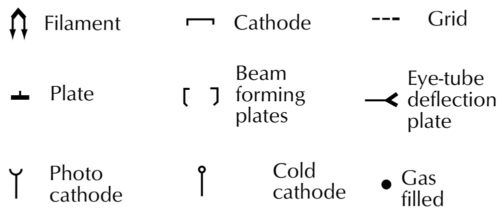


Figure 12-1. Tube elements and their designation.

12.1.2 Tube Types

There are many types of tubes, each used for a particular purpose. All tubes require a type of heater to permit the electrons to flow. Table12-2 defines the various types of tubes.

Table 12-2. The Eight Types of Vacuum Tubes

diode	A two-element tube consisting of a plate and a cathode. Diodes are used for rectifying or controlling the polarity of a signal as current can flow in one direction only.
triode	A three-element tube consisting of a cathode, a control grid, and a plate. This is the simplest type of tubes used to amplify a signal.
tetrode	A four-element tube containing a cathode, a control grid, a screen grid, and a plate. It is frequently referred to as a screen-grid tube
pentode	A five-element tube containing a cathode, a control grid, a screen grid, a suppressor grid, and a plate
hexode	A six-element tube consisting of a cathode, a control grid, a suppressor grid, a screen grid, an injector grid, and a plate.
heptode	A seven-element tube consisting of a cathode, a control grid, four other grids, and a plate.
pentagrid	A seven-element tube consisting of a cathode, five grids, and a plate.
beam-power tube	A power-output tube having the advantage of both the tetrode and pentode tubes. Beam-power tubes are capable of handling relatively high levels of output power for application in the output stage of an audio amplifier. The power-handling capabilities stem from the concentration of the plate-current electrons into beams of moving electrons. In the conventional tube the electrons flow from the cathode to the plate, but they are not confined to a beam. In a beam-power tube the internal elements consist of a cathode, a control grid, a screen grid, and two beam-forming elements that are tied internally to the cathode element. The cathode is indirectly heated as in the conventional tube.

12.1.3 Symbols and Base Diagrams

Table 12-3 gives the basic symbols used for tube circuits. The basing diagrams for various types of vacuum tubes are shown in Fig. 12-2.

Table 12-3. Tube Nomenclature

C	Coupling capacitor between stages
C_{g2}	Screen grid bypass capacitor
C_k	Cathode bypass capacitor
E_{bb}	Supply voltage
E_{ff}	Plate efficiency
E_p	Actual voltage at plate
E_{sg}	Actual voltage at screen grid
E_o	Output voltage
E_{sig}	Signal voltage at input
E_g	Voltage at control grid
E_f	Filament or heater voltage
I_f	Filament or heater current
I_p	Plate current
I_k	Cathode current
I_{sg}	Screen-grid current
I_{pa}	Average plate current
I_{pac}	Average ac plate current
I_{ka}	Average cathode current
I_{sga}	Average screen grid current
g_m	Transconductance (mutual conductance)
mu	Amplification factor (µ)
P_{sg}	Power at screen grid
P_p	Power at plate
P-P	Plate-to-plate or push-pull amplifier
R_g	Grid resistor
R_k	Cathode resistor
R_l	Plate-load impedance or resistance
R_p	Plate-load resistor
R_{sg}	Screen-dropping resistor
R_d	Decoupling resistor
r_p	Internal plate resistance
V_g	Voltage gain

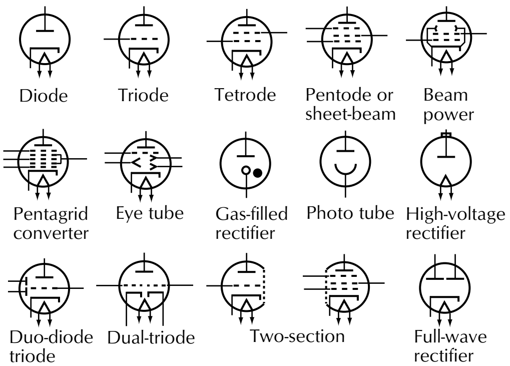


Figure 12-2. Basing diagrams for popular tubes.

12.1.4 Transconductance

Transconductance (g_m) is the change in the value of plate current expressed in microamperes (µA) divided by the signal voltage at the control grid of a tube, and is expressed by conductance. *Conductance* is the opposite of resistance, and the name *mho* (ohm spelled backward) was adopted for this unit of measurement. *Siemens* (S) have been adopted as the SI standard for conductance and are currently replacing mhos in measurement.

The basic mho or siemen are too large for practical usage; therefore, the term micromho (µmho) and microsiemens (µS) is used. One micromho is equal to one-millionth of a mho.

The transconductance (g_m) of a tube in µmhos may be found with the equation

$$g_m = \frac{\Delta I_p}{\Delta E_{sig}} \quad E_{bb} \text{ held constant} \tag{12-1}$$

where,
ΔI_p is the change of plate current,
ΔE_{sig} is the change of control-grid signal voltage,
E_{bb} is the plate supply voltage.

For example, a change of 1 mA of plate current for a change of 1 V at the control grid is equal to a transconductance of 1000 µmho. A tube having a change of 2 mA plate current for a change of 1 V at the control grid would have a transconductance of 2000 µmho.

$$g_m = I_{pac} \times 1{,}000 \tag{12-2}$$

where,
g_m is the transconductance in micromho or microsiemens,
I_{pac} is the ac plate current.

12.1.5 Amplification Factor

Amplification factor (µ) or *voltage gain* (V_g) is the ratio of the incremental plate voltage change to the control-electrode voltage change at a fixed plate current and constant voltage on all other electrodes. This normally is the amount the signal at the control grid is increased in amplitude after passing through the tube.

Tube voltage gain may be computed using the equation

$$V_g = \frac{\Delta E_p}{\Delta E_g} \tag{12-3}$$

where,
V_g is the voltage gain,

ΔE_p is the change in signal plate voltage,
ΔE_g is the change in the signal grid voltage.

If the amplifier consists of several stages, the amount of amplification is multiplied by each stage. The gain of an amplifier stage varies with the type tube and the interstage coupling used. The general equation for voltage gain is

$$V_{gt} = V_{g1}V_{g2}\ldots V_{gn} \qquad (12\text{-}4)$$

where,
V_{gt} is the total gain of the amplifier,
V_{g1}, V_{g2}, and V_{gn} are the voltage gain of the individual stages.

Triode tubes are classified by their amplification factor. A low-μ tube has an amplification factor less than 10. Medium-μ tubes have an amplification factor from 10–50, with a plate resistance of 5 Ω–15,000 Ω. High-μ tubes have an amplification factor of 50–100 with a plate resistance of 50 kΩ–100 kΩ.

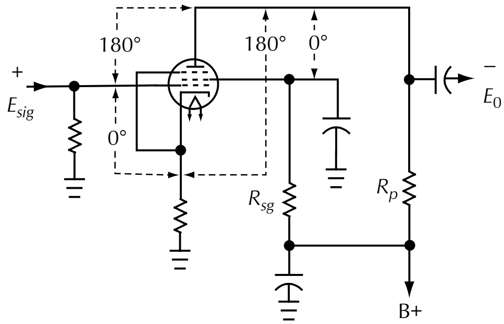
12.1.6 Polarity

Polarity reversals take place in a tube. The polarity reversal in electrical degrees between the elements of a self-biased pentode for a given signal at the control grid is shown in Fig. 12-3A. The reversals are the same for a triode. Note that, for an instantaneous positive voltage at the control grid, the voltage polarity between the grid and plate is 180° and will remain so for all normal operating conditions. The control grid and cathode are in polarity. The plate and screen-grid elements are in polarity with each other. The cathode is 180° out of polarity with the plate and screen-grid elements.

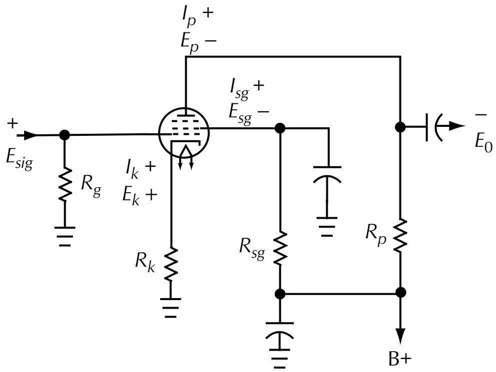
The polarity reversal of the instantaneous voltage and current for each element is shown in Fig. 12-3B. For an instantaneous positive sine wave at the control grid, the voltages at the plate and screen grid are negative, and the currents are positive. The voltage and current are both positive in the cathode resistor and are in polarity with the voltage at the control grid. The reversals are the same in a triode for a given element.

12.1.7 Internal Capacitance

The *internal capacitance* of a vacuum tube is created by the close proximity of the internal elements, Fig. 12-4. Unless otherwise stated by the manufacturer, the internal capacitance of a glass tube is measured using a close-fitting metal tube shield around the glass envelope connected to the cathode terminal. Generally, the capacitance is measured with the heater or filament cold and with no voltage applied to any of the other elements.



A. Polarity reversal of the signal between the elements of a pentode vacuum tube.



B. Polarity reversal of the current and voltage in a pentode vacuum tube.

Figure 12-3. Polarity characteristics of a vacuum tube.

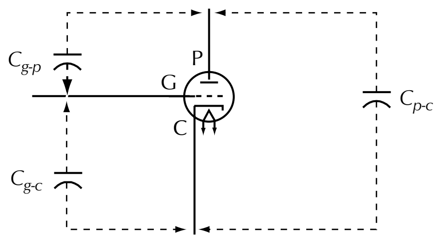


Figure 12-4. Interelectrode capacitance of a triode.

In measuring the capacitance, all metal parts, except the input and output elements, are connected to the cathode. These metal parts include internal and external shields, base sleeves, and unused pins. In testing a midsection tube, elements not common to the section being measured are connected to ground.

Input capacitance is measured from the control grid to all other elements, except the plate, which is connected to ground.

Output capacitance is measured from the plate to all other elements, except the control grid, which is connected to ground.

Grid-to-plate capacitance is measured from the control grid to the plate with all other elements connected to ground.

12.1.8 Plate Resistance

The *plate resistance* (r_p) of a vacuum tube is a constant and denotes the internal resistance of the tube or the opposition offered to the passage of electrons from the cathode to the plate. Plate resistance may be expressed in two ways: the dc resistance and the ac resistance. Dc resistance is the internal opposition to the current flow when steady values of voltage are applied to the tube elements and may be determined simply by using Ohm's law

$$r_{p_{dc}} = \frac{E_p}{I_p} \qquad (12\text{-}5)$$

where,
E_p is the dc plate voltage,
I_p is the steady value of plate current.

The ac resistance requires a family of plate-current curves from which the information may be extracted. As a rule, this information is included with the tube characteristics and is used when calculating or selecting components for an amplifier. The equation for calculating ac plate resistance is

$$r_{p_{ac}} = \frac{\Delta E_p}{\Delta I_p} \quad E_{sig} \text{ held constant} \qquad (12\text{-}6)$$

where,
ΔE_p is the change in voltage at the plate,
ΔI_p is the change in plate current,
E_{sig} is the control grid signal voltage.

The values of E_p and I_p are those taken from the family of curves supplied by the manufacturer for the particular tube under consideration.

12.1.9 Grid Bias

Increasing the plate voltage or decreasing the grid-bias voltage decreases the plate resistance. The six methods most commonly used to bias a tube are illustrated in Fig. 12-5. In Fig. 12-5A bias cell (battery) is connected in series with the control grid. In Fig. 12-5B the tube is self-biased by the use of a resistor connected in the cathode circuit. In Fig. 12-5C the circuit is also a form of self-bias; however, the bias voltage is obtained by the use of a grid capacitor and grid-leak resistor connected between the control grid and ground. In Fig. 12-5D the bias voltage is developed by a grid-leak resistor and capacitor in parallel, connected in series with the control grid. The method illustrated in Fig. 12-5E is called combination bias and consists of self-bias and battery bias. The resultant bias voltage is the negative voltage of the battery, and the bias created by the self-bias resistor in the cathode circuit. Another combination bias circuit is shown in Fig. 12-5F. The bias battery is connected in series with the grid-leak resistor. The bias voltage at the control grid is that developed by the battery and the self-bias created by the combination of the grid resistor and capacitor.

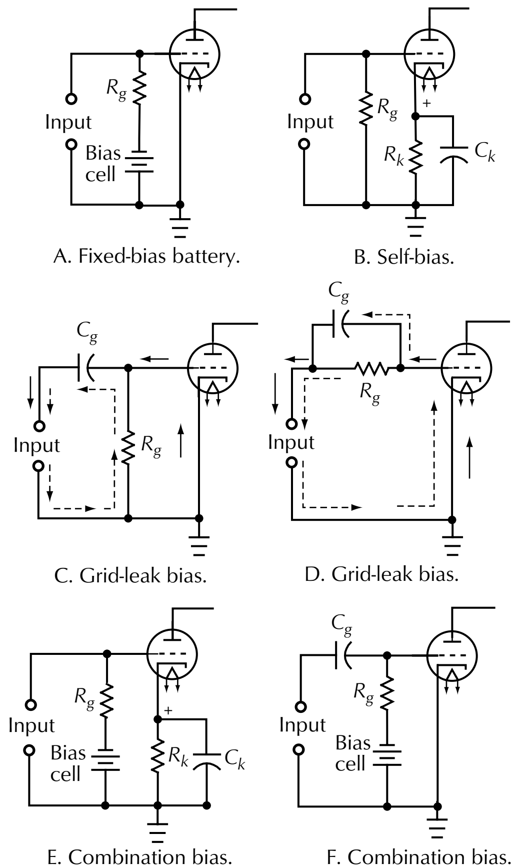


Figure 12-5. Various methods of obtaining grid bias.

If the control grid becomes positive with respect to the cathode, it results in a flow of current between the control grid and the cathode through the external circuits. This condition is unavoidable because the wires of the control grid, having a positive charge, attract electrons passing from the cathode to the plate. It is important that the control-grid voltage is kept negative, reducing grid current and distortion.

Grid-current flow in a vacuum tube is generally thought of as being caused by driving the control grid into the positive region and causing the flow of grid current.

The grid voltage, plate-current characteristics are found through a series of curves supplied by the tube manufacturer, as shown in Fig. 12-6.

The curves indicate that for a given plate voltage the plate current and grid bias may be determined. For example, the manufacturer states that for a plate voltage of 250 V and a negative grid bias of –8 V, the plate current will be 9 mA, which is indicated at point A on the 250 V curve. If it is desired to operate this tube with a plate voltage of 150 V and still maintain a plate current of 9 mA, the grid bias will have to be changed to a –3 V.

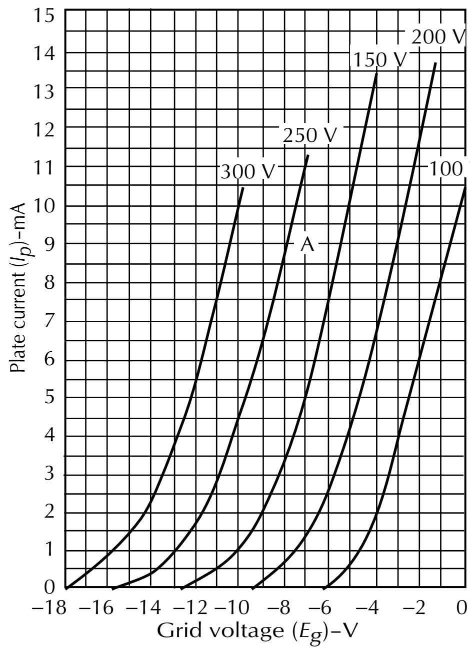


Figure 12-6. Grid voltage, plate-current curves for a triode tube.

12.1.10 Plate Efficiency

The *plate efficiency* (E_{ff}) is calculated by the equation:

$$E_{ff} = \frac{watts}{E_{pa}I_{pa}} \times 100 \tag{12-7}$$

where,

watts is the power output,

E_{pa} is the average plate voltage,

I_{pa} is the average plate current.

The measurement is made with a load resistance in the plate circuit equal in value to the plate resistance stated by the manufacturer.

12.1.11 Power Sensitivity

Power sensitivity is the ratio of the power output to the square of the input voltage, expressed in mhos or siemens and is determined by the equation

$$\text{Power sensitivity} = \frac{P_o}{E_{in}^2} \tag{12-8}$$

where,

P_o is the power output of the tube in watts,

E_{sig} is the rms signal voltage at the input.

12.1.12 Screen Grid

The *screen grid* series-dropping resistance is calculated by referring to the data sheet of the manufacturer and finding the maximum voltage that may be applied and the maximum power that may be dissipated by the screen grid. These limitations are generally shown graphically as in Fig. 12-7. The value of the resistor may be calculated using the equation

$$R_{sg} = \frac{E_{sg} \times (E_{bb} - E_{sg})}{P_{sg}} \tag{12-9}$$

where,

R_{sg} is the minimum value for the screen-grid voltage-dropping resistor in ohms,

E_{sg} is the selected value of screen-grid voltage,

E_{bb} is the screen-grid supply voltage,

P_{sg} is the screen-grid input in watts corresponding to the selected value of E_{sg}.

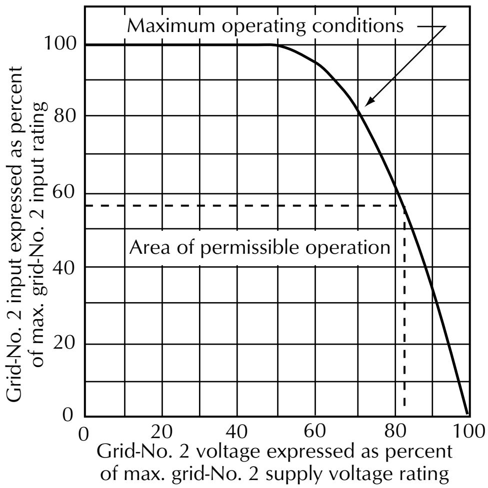


Figure 12-7. Typical graph for determining the maximum power dissipated by the screen grid.

12.1.13 Plate Dissipation

Plate dissipation is the maximum power that can be dissipated by the plate element before damage and is found with the equation

$$\text{Watts dissipation} = E_p I_p \tag{12-10}$$

where,

E_p is the voltage at the plate,

I_p is the plate current.

12.1.14 Changing Parameters

If a tube is to operate at a different plate voltage than published, the new values of bias, screen voltage, and plate resistance can be calculated by the use of conversion factors F_1, F_2, F_3, F_4, and F_5. Assume the following conditions are specified for a single beam-power tube:

Plate voltage	250.0 V
Screen voltage	250.0 V
Grid voltage	–12.5 V
Plate current	45.0 mA
Screen current	4.5 mA
Plate resistance	52,000.0 Ω
Plate load	5,000.0 Ω
Transconductance	4,100.0 µS
Power output	4.5 W

F_1 is used to find the new plate voltage

$$F_1 = \frac{E_{p_{new}}}{E_{p_{old}}} \tag{12-11}$$

For example, the new plate voltage is to be 180 V. The conversion factor F_1 for this voltage is obtained by dividing the new plate voltage by the published plate voltage Eq. 12-11:

$$F_1 = \frac{180}{250} = 0.72$$

The screen and grid voltage will be proportional to the plate voltage

$$E_g = F_1 \times old\ grid\ voltage \tag{12-12}$$

$$E_{sg} = F_1 \times old\ screen\ voltage \tag{12-13}$$

In the example

$$E_g = 0.72 \times (-12.5) = -9V$$

$$E_{sg} = 0.72 \times 250 = 180\text{ V}.$$

F_2 is used to calculate the plate and screen currents

$$F_2 = F_1\sqrt{F_1} \tag{12-14}$$

$$I_P = F_2 \times old\ plate\ current \tag{12-15}$$

$$I_s = F_2 \times old\ screen\ current\,. \tag{12-16}$$

In the example

$$F_2 = 0.72 \times 0.848 = 0.61$$

$$I_P = 0.61 \times 45\text{ mA} = 27.4\text{ mA}$$

$$I_{sg} = 0.61 \times 4.5\text{ mA} = 2.74\text{ mA}.$$

The plate load and plate resistance may be calculated by use of factor F_3:

$$F_3 = \frac{F_1}{F_2}. \quad (12\text{-}17)$$

$$r_p = F_3 \times old\ internal\ plate\ resistance \quad (12\text{-}18)$$

$$R_L = F_3 \times old\ \ plateload\ resistance \quad (12\text{-}19)$$

In the example

$$F_3 = \frac{0.720}{0.610} = 1.18$$

$$r_p = 1.18 \times 52{,}000 = 61{,}360\Omega$$

$$R_L = 1.18 \times 5{,}000 = 5{,}900\Omega.$$

F_4 is used to find the power output

$$F_4 = F_1 F_2 \quad (12\text{-}20)$$

$$Power\ output = F_4 \times old\ power\ output \quad (12\text{-}21)$$

In the example:

$$F_4 = 0.72 \times 0.610 = 0.439$$

$$Power\ output = 0.439 \times 4.5 = 1.97\ \text{W}$$

F_5 is used to find the transconductance where

$$F_5 = \frac{1}{F_3} \quad (12\text{-}22)$$

$$Transconductance = F_5 \times old\ Transconductance \quad (12\text{-}23)$$

In the example:

$$F_5 = \frac{1}{1.18} = 0.847$$

$$\text{transconductance} = 0.847 \times 4{,}100 = 3{,}472\ \mu\text{mho or } \mu\text{S}$$

The foregoing method of converting for voltages other than those originally specified may be used for triodes, tetrodes, pentodes, and beam-power tubes, provided the plate and grid 1 and grid 2 voltages are changed simultaneously by the same factor. This will apply to any class of tube operation, such as class A, AB_1, AB_2, B, or C. Although this method of conversion is quite satisfactory in most instances, the error will be increased as the conversion factor departs from unity. The most satisfactory region of operation will be between 0.7 and 2.0. When the factor falls outside this region, the accuracy of operation is reduced.

12.1.15 Tube Heater

The data sheets of tube manufacturers generally contain a warning that the heater voltage should be maintained within ±10% of the rated voltage. As a rule, this warning is taken lightly, and little attention is paid to heater voltage variations, which have a pronounced effect on the tube characteristics. Internal noise is the greatest offender. Because of heater-voltage variation, emission life is shortened, electrical leakage between elements is increased, heater-to-cathode leakage is increased, and grid current is caused to flow. Thus, the life of the tube is decreased with an increase of internal noise.

12.2 Discrete Solid-State Devices

12.2.1 Semiconductors

Conduction in solids was first observed by Munck and Henry in 1835, and later in 1874 by Braum. In 1905, Col. Dunwoody invented the crystal detector used in the detection of electromagnetic waves. It consisted of a bar of silicon carbide or carborundum held between two contacts. However, in 1903, Pickard filed a patent application for a crystal detector in which a fine wire was placed in contact with the silicon. This was the first mention of a silicon rectifier and was the forerunner of the present-day silicon rectifier. Later, other minerals such as galena (lead sulfide) were employed as detectors. During World War II, intensive research was conducted to improve crystal detectors used for microwave radar equipment. As a result of this research, the original point-contact transistor was invented at the Bell Telephone Laboratories in 1948.

A s*emiconductor* is an electronic device whose main functioning part is made from materials, such as germanium and silicon, whose conductivity ranges between that of a conductor and an insulator.

Germanium is a rare metal discovered by Winkler in Saxony, Germany, in the year 1896. Germanium is a by-product of zinc mining. Germanium crystals are grown from germanium dioxide powder. Germanium in its purest state behaves much like an insulator because it has very few electrical charge carriers. The conductivity of germanium may be increased by the addition of small amounts of an impurity.

Silicon is a nonmetallic element used in the manufacture of diode rectifiers and transistors. Its resistivity is considerably higher than that of germanium.

The relative position of pure germanium and silicon is given in Fig. 12-8. The scale indicates the resistance of conductors, semiconductors, and insulators per cubic centimeter. Pure germanium has a resistance of approximately 60 Ω/cm^3. Germanium has a higher conductivity or less resistance to current flow than silicon and is used in low and medium power diodes and transistors.

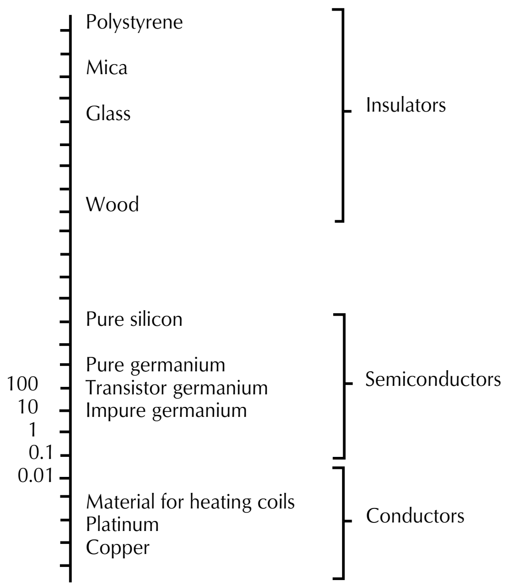


Figure 12-8. Resistance of various materials per cubic centimeter.

The base elements used to make semiconductor devices are not usable as semiconductors in their pure state. They must be subjected to a complex chemical, metallurgical, and photo lithographical process wherein the base element is highly refined and then modified with the addition of specific impurities. This precisely controlled process of diffusing impurities into the pure base element is called *doping* and converts the pure base material into a semiconductor material. The semiconductor mechanism is achieved by the application of a voltage across the device with the proper polarity so as to have the device act either as an extremely low resistance (the forward biased or conducting mode) or as an extremely high resistance (reversed bias or nonconducting mode). Because the device is acting as both a good conductor of electricity and also, with the proper reversal of voltage, as a good electrical nonconductor or insulator, it is called a semiconductor.

Some semiconductor materials are called *p* or positive type because they are processed to have an excess of positively charged ions. Others are called *n* or negative type because they are processed to have an excess of negatively charged electrons. When a p-type of material is brought into contact with an n-type of material, a *pn junction* is formed. With the application of the proper external voltage, a low resistance path is produced between the n and p material. By reversing the previously applied voltage, an extremely high resistance called the *depletion layer* between the p and n types results. A diode is an example because its conduction depends upon the polarity of the externally applied voltage. Combining several of these pn junctions together in a single device produces semiconductors with extremely useful electrical properties.

The theory of operation of a semiconductor device is approached from its atomic structure. The outer orbit of a germanium atom contains four electrons. The atomic structure for a pure germanium crystal is shown in Fig. 12-9A. Each atom containing four electrons forms covalent bonds with adjacent atoms, therefore there are no "free" electrons. Germanium in its pure state is a poor conductor of electricity. If a piece of "pure" germanium (the size used in a transistor) has a voltage applied to it, only a few microamperes of current caused by electrons that have been broken away from their bonds by thermal agitation will flow in the circuit. This current will increase at an exponential rate with an increase of temperature.

When an atom with five electrons, such as antimony or arsenic, is introduced into the germanium crystal, the atomic structure is changed to that of Fig. 12-9B. The extra electrons (called free electrons) will move toward the positive terminal of the external voltage source.

When an electron flows from the germanium crystal to the positive terminal of the external voltage source, another electron enters the crystal from the negative terminal of the voltage source. Thus, a continuous stream of electrons will flow as long as the external potential is maintained.

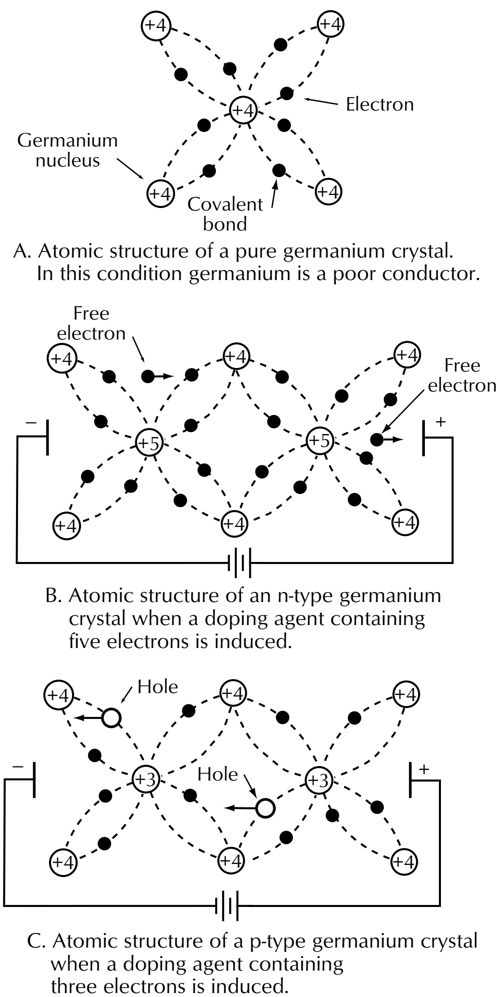


Figure 12-9. Atomic structure of germanium.

The atom containing the five electrons is the *doping agent* or *donor*. Such germanium crystals are classified as *n-type germanium*.

Using a doping agent of indium, gallium, or aluminum, each of which contains only three electrons in its outer orbit, causes the germanium crystal to take the atomic structure of Fig. 12-9C. In this structure, there is a *hole* or *acceptor*. The term *hole* is used to denote a mobile particle that has a positive charge and that simulates the properties of an electron having a positive charge.

When a germanium crystal containing holes is subjected to an electrical field, electrons jump into the holes, and the holes appear to move toward the negative terminal of the external voltage source.

When a hole arrives at the negative terminal, an electron is emitted by the terminal, and the hole is canceled. Simultaneously, an electron from one of the covalent bonds flows into the positive terminal of the voltage source. This new hole moves toward the negative terminal causing a continuous flow of holes in the crystal.

Germanium crystals having a deficiency of electrons are classified *p-type germanium.* Insofar as the external electrical circuits are concerned, there is no difference between electron and hole current flow. However, the method of connection to the two types of transistors differs.

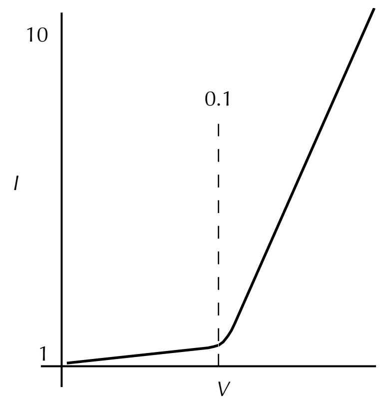
When a germanium crystal is doped so that it abruptly changes from an n-type to a p-type, and a positive potential is applied to the p-region, and a negative potential is applied to the n-region, the holes move through the junction to the right and the electrons move to the left, resulting in the voltage-current characteristic shown in Fig. 12-10A. If the potential is reversed, both electrons and holes move away from the junction until the electrical field produced by their displacement counteracts the applied electrical field. Under these conditions, zero current flows in the external circuit. Any minute amount of current that might flow is caused by thermal-generated hole pairs. Fig. 12-10B is a plot of the voltage versus current for the reversed condition. The leakage current is essentially independent of the applied potential up to the point where the junction breaks down.

12.2.2 Diodes

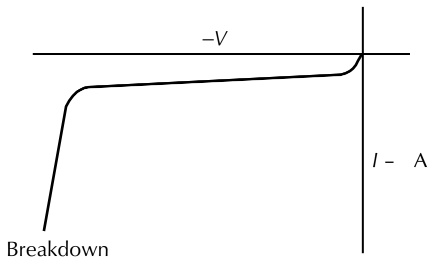
The *diode* is a device that exhibits a low resistance to current flow in one direction and a high resistance in the other. Ideally, when reverse biasing the diode (connecting the negative of the supply to the diode anode), no current should flow regardless of the value of voltage impressed across the diode. A forward-biased diode presents a very low resistance to current flow.

Fig. 12-11 shows the actual diode characteristics. Starting with the diode reverse biased, a small reverse current does flow. The size of this reverse-leakage current has been exaggerated for clarity and typically is in the order of nanoamperes. The forward resistance is not constant therefore does not yield a straight line forward-conduction curve. Instead, it begins high and drops rapidly at relatively low applied voltage. Above a 0.5–1 V drop it approaches a steep straight line slope (i.e., low resistance).

In the reverse-biased region of Fig. 12-11, when the applied voltage (*–V*) becomes large enough, the leakage current suddenly begins to increase very rapidly, and the



A. Voltage-versus-current characteristic of the junction.



B. Voltage-versus-current characteristic of the junction transistor with the battery polarities in the reverse condition.

Figure 12-10. Voltage versus current characteristics

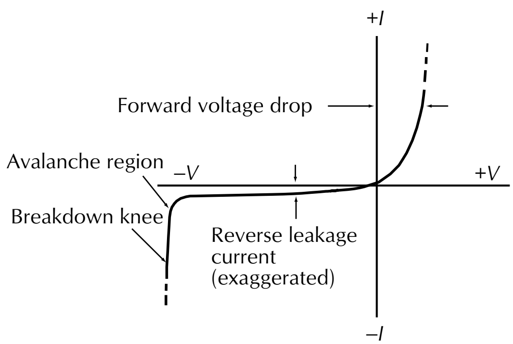


Figure 12-11. Actual diode characteristics.

slope of the characteristic curves becomes very steep. Past the *knee* in the characteristic, even a small increase in reverse voltage causes a large increase in the reverse current. This steep region is called the *breakdown* or *avalanche* region of the diode characteristic.

The application of high reverse voltage causes the diode to break down and stop behaving like a diode. *Peak-reverse-voltage* rating or *prv* is one of the two most important diode parameters. This is also referred to as the *peak-inverse-voltage* rating or *piv.* This rating indicates how high the reverse voltage can be without approaching the knee and risking breakdown. Additional diode parameters are:

Maximum average current	Causes overheating of the device.
Peak repetitive current	Maximum peak value of current on a repetitive basis.
Surge current	Absolute maximum allowed current even if just momentary.

The maximum average current is limited by power dissipation in the junction. This power dissipation is represented by the product of forward voltage drop (V_F) and the forward current (I_F):

$$P = V_F I_F . \tag{12-24}$$

Selenium Rectifiers and Diodes. A *selenium rectifier cell* consists of a nickel-plated aluminum-base plate coated with selenium, over which a low-temperature alloy is sprayed. The aluminum base serves as a negative electrode, and the alloy, as the positive. Current flows from the base plate to the alloy but encounters high resistance in the opposite direction. The efficiency of conversion depends to some extent on the ratio of the resistance in the conducting direction to that of the blocking direction. Conventional rectifiers generally have ratios from 100:1 to 1000:1.

Selenium rectifiers may be operated over temperatures of –55°C to +150°C (–67°F to +302°F). Rectification efficiency is in the order of 90% for three-phase bridge circuits and 70% for single-phase bridge circuits. As a selenium cell ages, the forward and reverse resistance increases for approximately one year, then stabilizes, decreasing the output voltage by approximately 15%. The internal impedance of a selenium rectifier is low and exhibits a nonlinear characteristic with respect to the applied voltage, maintaining a good voltage regulation. They are often used for battery charging.

Selenium rectifiers, because of their construction, have considerable internal capacitance which limits their operating range to audio frequencies. Approximate capacitance ranges are 0.10–0.15 µF/in^2 of rectifying surface.

The minimum voltage required for conduction in the forward direction is termed the *threshold voltage* and is about 1 V therefore selenium rectifiers cannot be used successfully below that voltage.

Silicon Rectifiers and Diodes. The high forward-to-reverse current characteristic of the silicon diode produces an efficiency of about 99%. When properly used, silicon diodes have long life and are not affected by aging, moisture, or temperature when used with the proper heat sink.

As an example, four individual diodes of 400 V *piv* may be connected in series to withstand a *piv* of 1600 V. In a series arrangement, the most important consideration is that the applied voltage be equally distributed between the several units. The voltage drops across each individual unit must be very nearly identical. If the instantaneous voltage is not equally divided, one of the units may be subjected to a voltage exceeding its rated value, causing it to fail. This causes the other rectifiers to absorb the *piv*, often creating destruction of all the rectifiers.

Uniform voltage distribution can be obtained by the connection of capacitors or resistors in parallel with the individual rectifier unit, Fig. 12-12. Shunt resistors are used for steady-state applications, and shunt capacitors are used in applications where transient voltages are expected. If the circuit is exposed to both dc and ac, both shunt capacitors and resistors should be employed.

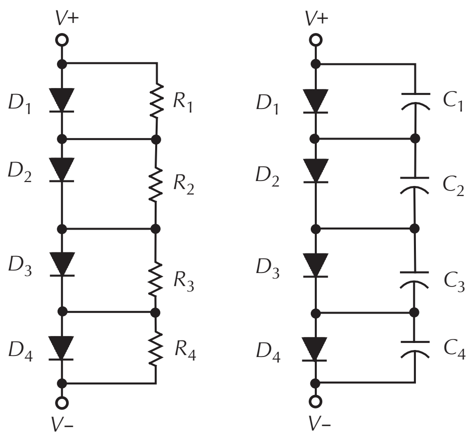


Figure 12-12. Rectifiers connected in series.

When the maximum current of a single diode is exceeded, two or more units may be connected in parallel. To avoid differences in voltage drop across the individual units, a resistor or small inductor is connected in series with each diode, Fig. 12-13. Of the two methods, the inductance is favored because of the lower voltage drop and consumption of power.

Zener and Avalanche Diodes. When the reverse voltage is increased beyond the breakdown knee of the diode characteristics as shown in Fig. 12-11, the diode impedance suddenly drops sharply to a very low value. If the current is limited by an external circuit resistance, operating in the "zener region" is normal for certain

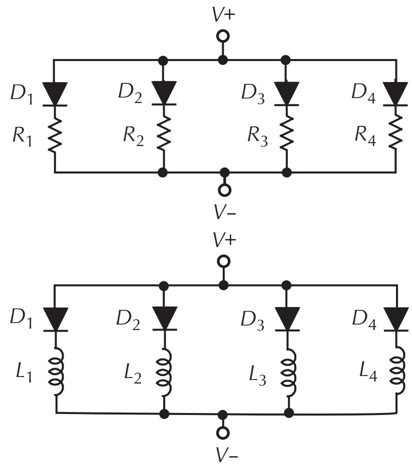


Figure 12-13. Rectifiers connected in parallel.

diodes specifically designed for the purpose. In *zener diodes*, sometimes simply called *zeners*, the breakdown characteristic is deliberately made as vertical as possible in the zener region so that the voltage across the diode is essentially constant over a wide reverse-current range, acting as a voltage regulator. Since its zener-region voltage can be made highly repeatable and very stable with respect to time and temperature, the zener diode can also function as a voltage reference. Zener diodes come in a wide variety of voltages, currents, and powers, ranging from 3.2 V to hundreds of volts, from a few milliamperes to 10 A or more, and from about 250 mW to over 50 W.

Avalanche diodes are diodes in which the shape of the breakdown knee has been controlled, and the leakage current before breakdown has been reduced so that the diode is especially well suited to two applications: high-voltage stacking, and clamping. In other words, they prevent a circuit from exceeding a certain value of voltage by causing breakdown of the diode at or just below that voltage.

Small-Signal Diodes. *Small-signal diodes* or *general-purpose diodes* are low-level devices with the same general characteristics as power diodes. They are smaller, dissipate much less power, and are not designed for high-voltage, high-power operation. Typical rating ranges are:

I_F (forward current):	1–500 mA
V_F (forward voltage drop at I_F):	0.2–1.1 V
piv or *prv:*	6–1000 V
I_R (leakage current at 80% *prv*):	0.1–1.0 μA

Switching Diodes. *Switching diodes* are small-signal diodes used primarily in digital-logic and control applications in which the voltages may change very rapidly so that speed, particularly reverse-recovery time, is of paramount importance. Other parameters of particular importance are low shunt capacitance, low and uniform V_F (forward voltage drop), low I_R (reverse leakage current) and, in control circuits, *prv*.

Noise Diodes. *Noise Diodes* are silicon diodes used in the avalanche mode (reverse biased beyond the breakdown knee) to generate broadband noise signals. All diodes generate some noise; these, however, have special internal geometry and are specially processed so as to generate uniform noise power over very broad bands. They are low-power devices (typically, 0.05–0.25 W) and are available in several different bandwidth classes from as low as 0 kHz–100 kHz to as high as 1000–18,000 MHz.

Varactor Diodes. *Varactor diodes* are made of silicon or gallium arsenide and are used as adjustable capacitors. Certain diodes, when operated in the reverse-biased mode at voltages below the breakdown value, exhibit a shunt capacitance that is inversely proportional to the applied voltage. By varying the applied reverse voltage, the capacitance of the varactor varies. This effect can be used to tune circuits, modulate oscillators, generate harmonics, and to mix signals. Varactors are sometimes referred to as *voltage-tunable trimmer capacitors*.

Tunnel Diodes. The *tunnel diode* takes its name from the tunnel effect, a process where a particle can disappear from one side of a barrier and instantaneously reappear on the other side as though it had tunneled through the barrier element.

Tunnel diodes are made by heavily doping both the p and n materials with impurities, giving them a completely different voltage-current characteristic from regular diodes. This characteristic makes them uniquely useful in many high-frequency amplifiers as well as pulse generators and radio-frequency oscillators, Fig. 12-14.

What makes the tunnel diode work as an active element is the negative-resistance region over the voltage range V_d (a small fraction of a volt). In this region, increasing the voltage decreases the current, the opposite of what happens with a normal resistor. Tunnel diodes conduct heavily in the reverse direction; in fact, there is no breakdown knee or leakage region.

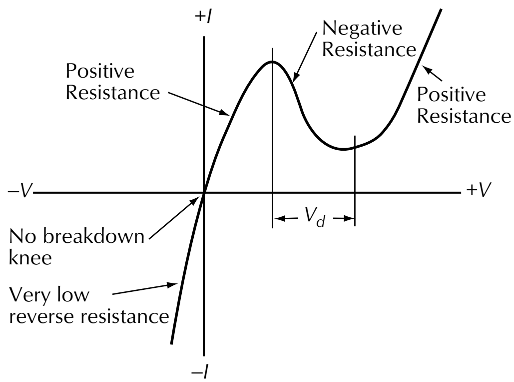


Figure 12-14. Tunnel-diode characteristics showing negative region (tunnel region).

12.2.3 Thyristors

Stack four properly doped semiconductor layers in series, pnpn (or npnp), and the result is a four-layer, or Shockley breakover diode. Adding a terminal (gate) to the second layer creates a gate-controlled, reverse-blocking *thyristor*, or *silicon-controlled rectifier* (*SCR*), as shown in Fig. 12-15A.

The four-layer diode connects (fires) above a specific threshold voltage. In the *SCR*, the gate controls this firing threshold voltage, called the forward blocking voltage.

To understand how four-layer devices work, separate the material of the layers into two three-layer transistor devices. Fig. 12-15B is an equivalent two-transistor representation in a positive-feedback connection. Assuming a_1 and a_2 are the current gains of the two transistor sections with each gain value less than unity, the total base current I_b into the $n_1p_2n_2$ transistor is

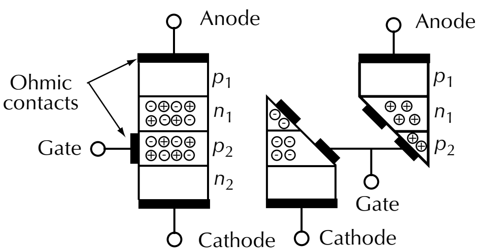
$$I_b = a_1a_2I_b + I_o + I_g \qquad (12\text{-}25)$$

where,

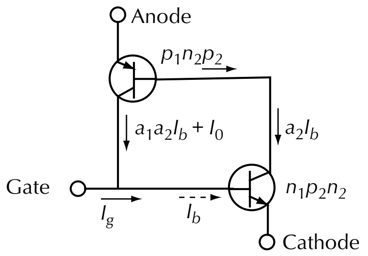
a_1 and a_2 are the transistor current gains,
I_b is the total base current,
I_o is the leakage current into the base of the $n_1p_2n_2$ transistor,
I_g is the current into the gate terminal.

The circuit turns on and becomes self-latching after a certain turn-on time needed to stabilize the feedback action, when the equality of Eq. 12-18 is achieved. This result becomes easier to understand by solving for I_b, which gives

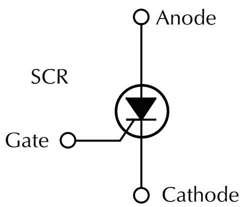
$$I_b = \frac{I_o + I_g}{1 - a_1a_2} \qquad (12\text{-}26)$$



A. Electrical layout of a thyristor.



B. Two-transistor equivalent circuit.



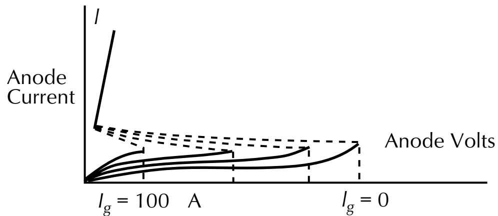
C. SCR layout.

Figure 12-15. Thyristor schematics.

When the product a_1a_2 is close to unity, the denominator approaches zero and I_b approaches a large value. For a given leakage current I_o, the gate current to fire the device can be extremely small. Moreover, as I_b becomes large, I_g can be removed, and the feedback will sustain the on condition since a_1 and a_2 then approach even closer to unity.

As applied anode voltage increases in the breakover diode, where I_g is absent, I_o also increases. When the quality of Eq. 12-18 is established, the diode fires. The thyristor fires when the gate current I_g rises to establish equality in the equation with the anode voltage fixed. For a fixed I_g, the anode voltage can be raised until the thyristor fires, with I_g determining the firing voltage, Fig. 12-16.

Once fired, a thyristor stays on until the anode current falls below a specified minimum holding current for a certain turnoff time. In addition, the gate loses all control once a thyristor fires. Removal or even reverse biasing of the gate signal will not turn off the device although reverse biasing can help speed turnoff. When



Thyristor breakover as function of gate current

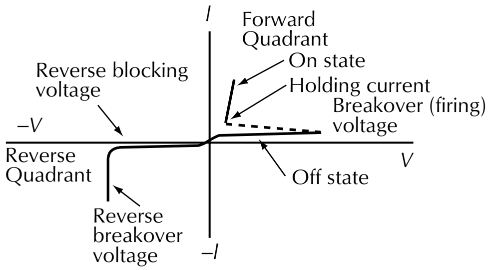


Figure 12-16. Thyristor breakover as a function of gate current and forward voltage.

the device is used with an ac voltage on the anode, the unit automatically turns off on the negative half of the voltage cycle. In dc switching circuits, however, complex means must often be used to remove, reduce, or reverse the anode voltage for turnoff.

Figure 12-17 shows a bilaterally conductive arrangement that behaves very much like two four-layer diodes (*diacs*), or two *SCR*s (*triacs*), parallel and oppositely conductive. When terminal A is positive and above the breakover voltage, a path through $p_1n_1p_2n_2$ can conduct; when terminal B is positive, path $p_2n_1p_1n_3$ can conduct. When terminal A is positive and a third element, terminal G, is sufficiently positive, the $p_1n_1p_2n_2$ path will fire at a much lower voltage than when G is zero. This action is almost identical with that of the *SCR*. When terminal G is made negative and terminal B is made positive, the firing point is lowered in the reverse, or $p_2n_1p_1n_3$, direction.

Because of low impedances in the *on* condition, four-layer devices must be operated with a series resistance in the anode and gate that is large enough to limit the anode-to-cathode or gate current to a safe value.

To understand the low-impedance, high-current capability of the thyristor, the device must be examined as a whole rather than by the two-transistor model. In Fig. 12-17B the $p_1n_1p_2$ transistor has holes injected to fire the unit, and the $n_1p_2n_2$ transistor has electrons injected. Considered separately as two transistors, the space-charge distributions would produce two typical transistor saturation-voltage forward drops, which are quite high when compared with the actual voltage drop of a thyristor.

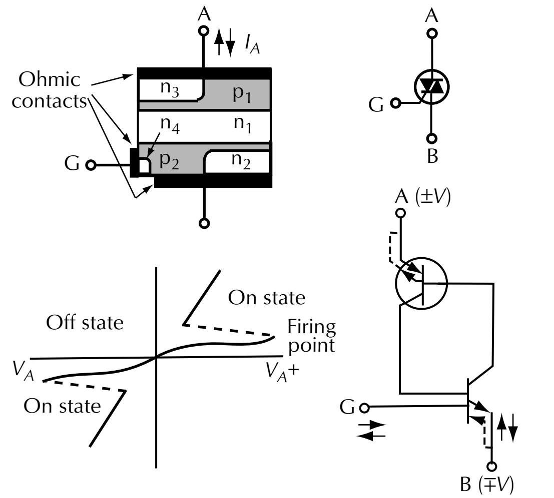


Figure 12-17. Bilateral arrangement to create a triac or ac operating device.

However, when the thyristor shown in Fig. 12-17A is considered, the charges of both polarities exist simultaneously in the same n_1 and p_2 regions. Therefore, at the high injection levels that exist in thyristors, the mobile-carrier concentration of minority carriers far exceeds that from the background-doping density. Accordingly, the space-charge is practically neutralized so that the forward drop becomes almost independent of the current density to high current levels. The major resistance to current comes from the ohmic contacts of the unit and load resistance.

The price paid for this low-impedance capability in a standard thyristor is a long turnoff time relative to turnon time necessary to allow the high level of minority current carriers to dissipate. This long turnoff time limits the speed of a thyristor. Fortunately, this long turnoff time does not add significantly to switching power losses the way that a slow turnon time would.

Turnoff time is the minimum time between the forward anode current ceasing and the device being able to block reapplied forward voltage without turning on again.

Reverse-recovery time, is the minimum time after forward conduction ceases that is needed to block reverse-voltage with ac applied to the anode-cathode circuit.

A third specification, *turnon time*, is the time a thyristor takes from the instant of triggering to when conduction is fully on.

These timing specifications limit the operating frequency of a thyristor. Two additional important specifications, the derivative of voltage with respect to time (dv/dt) and the derivative of current with respect to time (di/dt) limit the rates of change of voltage and current application to thyristor terminals.

A rapidly varying anode voltage can cause a thyristor to turn on even though the voltage level never exceeds the forward breakdown voltage. There is capacitance between the layers which may produce a current large enough to cause firing can be generated in the gated layer. Current through a capacitor is directly proportional to the rate of change of the applied voltage; therefore, the dv/dt of the anode voltage is an important thyristor specification.

Turnon by the dv/dt can be accomplished with as little as a few volts per microsecond in some units, especially in older designs. Newer designs are often rated in tens to hundreds of volts per microsecond.

The other important rate effect is the anode-current di/dt rating. This rating is particularly important in circuits that have low inductance in the anode-cathode path. Adequate inductance would limit the rate of current rise when the device fires.

When a thyristor fires, the region near the gate conducts first; then the current spreads to the rest of the semiconductor material of the gate-controlled layer over a period of time. If the current flow through the device increases too rapidly during this period because the input-current di/dt is too high, the high concentration of current near the gate could damage the device do to localized overheating. Specially designed gate structures can speed up the turnon time of a thyristor, and thus its operational frequency, as well as alleviate this hot-spot problem.

Silicon-Controlled Rectifiers. The *silicon controlled rectifier* (*SCR*) thyristor can be considered a solid-state latching relay if dc is used as the supply voltage for the load. The gate current turns on the SCR which is equivalent to closing the contacts in the load circuit.

If ac is used as the supply voltage, the SCR load current will reduce to zero as the positive ac wave shape crosses through zero and reverses its polarity to a negative voltage. This will shut off the SCR. If the positive gate voltage is also removed it will not turn on during the next positive half cycle of applied ac voltage unless positive gate voltage is applied.

The SCR is suitable for controlling large amounts of rectifier power by means of small gate currents. The ratio of the load current to the control current can be several thousand to one. For example, a 10 A load current might be triggered on by a 5 mA control current.

The major time-related specification associated with SCRs is the dv/dt rating. This characteristic reveals how fast a transient spike on the power line can be before it false-triggers the SCR and starts its conducting without gate control current. Apart from this time-related parameter and its gate characteristics, SCR ratings are similar to those for power diodes.

SCRs can be used to control dc by using *commutating circuits* to shut them off. These are not needed on ac since the anode supply voltage reverses every half cycle. SCRs can be used in pairs or sets of pairs to generate ac from dc in inverters. They are also used as protective devices to protect against excessive voltage by acting as a short-circuit switch. These are commonly used in power supply *crowbar* overvoltage protection circuits. SCRs are also used to provide switchedpower-amplification, as in solid-state relays.

Triacs. The triac in Fig. 12-18 is a three-terminal semiconductor that behaves like two SCRs connected back to front in parallel so that they conduct power in both directions under control of a single gate-control circuit. Triacs are widely used to control ac power by phase shifting or delaying the gate-control signal for some fraction of the half cycle during which the power diode could be conducting. Light dimmers found in homes and offices and variable-speed drills are good examples of triac applications.

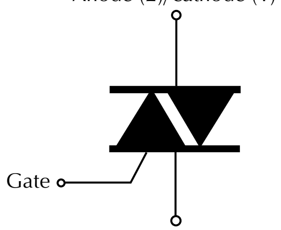


Figure 12-18. Schematic of a triac.

Light-Activated Silicon Controlled Rectifiers. When sufficient light falls on the exposed gate junction, the SCR is turned on just as if the gate-control current were flowing. The gate terminal is also provided for optional use in some circuits. These devices are used in projector controls, positioning controls, photo relays, slave flashes, and security protection systems.

Diacs. The *diac* is shown in Fig. 12-19. It acts as two zener (or avalanche) diodes connected in series, back to back. When the voltage across the diac in either direction gets large enough, one of the zeners breaks down. The action drops the voltage to a lower level, causing a current increase in the associated circuit. This device is used to trigger SCRs or triacs.

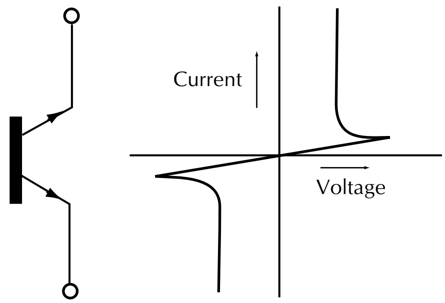


Figure 12-19. Schematic of a diac.

Opto-Coupled Silicon-Controlled Rectifiers. The *opto-coupled SCR* is a combination of a light-emitting diode (LED) and a photo silicon-controlled rectifier (photo-SCR). When sufficient current is forced through the LED, it emits an infrared radiation that triggers the gate of the photo-SCR. A small control current can regulate a large load current, and the device provides insulation and isolation between the control circuit (the LED) and the load circuit (the SCR). Opto-coupled transistors and Darlington transistors that operate on the same principle will be discussed later.

12.2.4 Transistors

There are many different types of transistors,[1] and they are named by the way they are grown, or made. Fig. 12-20A shows the construction of a *grown- junction* transistor. An *alloy-junction* transistor is shown in Fig. 12-20B. During the manufacture of the material for a grown junction, the impurity content of the semiconductor is altered to provide npn or pnp regions. The grown material is cut into small sections, and contacts are attached to the regions. In the alloy-junction type, small dots of n- or p-type impurity elements are attached to either side of a thin wafer of p- or n-type semiconductor material to form regions for the emitter and collector junctions. The base connection is made to the original semiconductor material.

Drift-field transistors, Fig. 12-20C, employ a modified alloy junction in which the impurity concentration in the wafer is diffused or graded. The drift field speeds up the current flow and extends the frequency response of the alloy-junction transistor. A variation of the drift-field transistor is the *microalloy diffused* transistor, as shown in Fig. 12-20D. Very narrow base dimensions

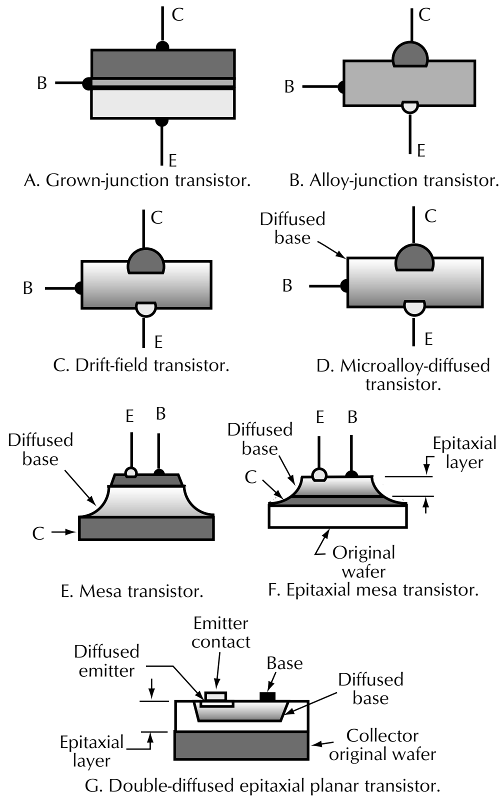


Figure 12-20. Construction of various transistors.

are achieved by etching techniques, resulting in a shortened current path to the collector.

Mesa transistors shown in Fig. 12-20E use the original semiconductor material as the collector, with the base material diffused into the wafer and an emitter dot alloyed into the base region. A flat-topped peak or mesa is etched to reduce the area of the collector at the base junction. Mesa devices have large power-dissipation capabilities and can be operated at very high frequencies. *Double-diffused epitaxial mesa transistors* are grown by the use of vapor deposition to build up a crystal layer on a crystal wafer and will permit the precise control of the physical and electrical dimensions independently of the nature of the original wafer. This technique is shown in Fig. 12-20F.

The *planar transistor* is a highly sophisticated method of constructing transistors. A limited area source is used for both the base diffusion and emitter diffusion, which provides a very small active area, with a large wire contact area. The advantage of the planar construction is its high dissipation, lower leakage current, and lower collector cut-off current, which increases the stability and reliability. Planar construction is also used with several of the previously discussed base designs. A double-diffused epitaxial planar transistor is shown in Fig. 12-20G.

The *field-effect transistor*, or FET as it is commonly known, was developed by the Bell Telephone Laboratories in 1946, but it was not put to any practical use until about 1964. The principal difference between a conventional transistor and the FET is the transistor is a current-controlled device, while the FET is voltage controlled, similar to the vacuum tube. Conventional transistors also have a low input impedance, which may at times complicate the circuit designer's problems. The FET has a high input impedance with a low output impedance, much like vacuum tube.

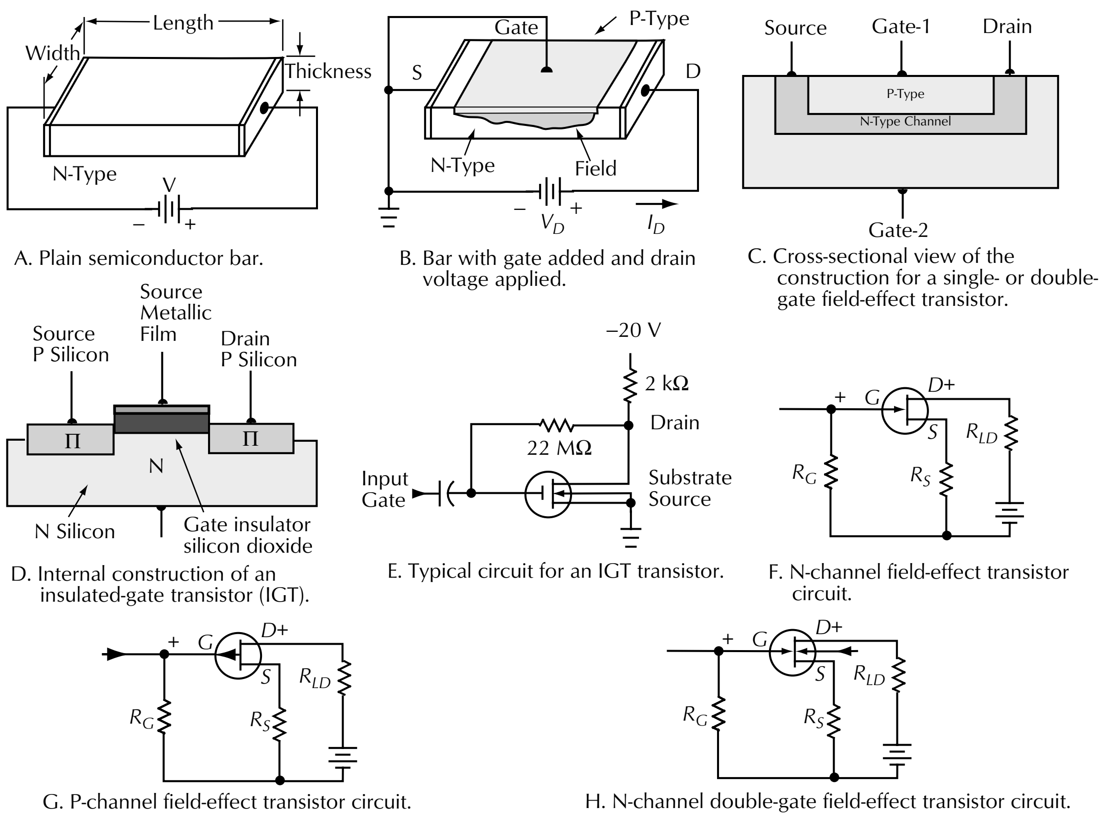
The basic principles of the FET operation can best be explained by the simple mechanism of a pn junction. The control mechanism is the creation and control of a depletion layer, which is common to all reverse-biased junctions. Atoms in the n-region possess excess electrons that are available for conduction, and the atoms in the p-region have excess holes that may also allow current to flow. Reversing the voltage applied to the junction and allowing time for stabilization, very little current flows, but a rearrangement of the electrons and holes will occur. The positively charged holes will be drawn toward the negative terminals of the voltage source, and the electrons, which are negative, will be attracted to the positive terminal of the voltage source. This results in a region being formed near the center of the junction having a majority of the carriers removed and therefore called the depletion regions.

Referring to Fig. 12-21A, a simple bar composed of n-type semiconductor material has a nonrectifying contacts at each end. The resistance between the two end electrodes is

$$R = \frac{PL}{WT} \tag{12-27}$$

where,
P is the function of the material sensitivity,
L is the length of the bar,
W is the width,
T is the thickness.

Varying one or more of the variables of the resistance of the semiconductor, changes the bar. Assume a p-region in the form of a sheet is formed at the top of the bar shown in Fig. 12-21B. A pn junction is formed by diffusion, alloying, or epitaxial growth creating a reverse voltage between the p- and n-material producing two depletion regions. Current in the n-material is



A. Plain semiconductor bar.

B. Bar with gate added and drain voltage applied.

C. Cross-sectional view of the construction for a single- or double-gate field-effect transistor.

D. Internal construction of an insulated-gate transistor (IGT).

E. Typical circuit for an IGT transistor.

F. N-channel field-effect transistor circuit.

G. P-channel field-effect transistor circuit.

H. N-channel double-gate field-effect transistor circuit.

Figure 12-21. Field-effect transistors (FETs).

caused primarily by means of excess electrons. By reducing the concentration of electrons or majority carriers, the resistivity of the material is increased. Removal of the excess electrons by means of the depletion region causes the material to become practically nonconductive.

Disregarding the p-region and applying a voltage to the ends of the bar cause a current and create a potential gradient along the length of the bar material, with the voltage increasing toward the right, with respect to the negative end or ground. Connecting the p-region to ground causes varying amounts of reverse-bias voltage across the pn junction, with the greatest amount developed toward the right end of the p-region. A reverse voltage across the bar will produce the same depletion regions. If the resistivity of the p-type material is made much smaller than that of the n-type material, the depletion region will then extend much farther into the n-material than into the p-material. To simplify the following explanation, the depletion of p-material will be ignored.

The general shape of the depletion is that of a wedge, increasing the size from left to right. Since the resistivity of the bar material within the depletion area is increased, the effective thickness of the conducting portion of the bar becomes less and less, going from the end of the p-region to the right end. The overall resistance of the semiconductor material is greater because the effective thickness is being reduced. Continuing to increase the voltage across the ends of the bar, a point is reached where the depletion region is extended practically all the way through the bar, reducing the effective thickness to zero. Increasing the voltage beyond this point produces little change in current.

The p-region controls the action and is termed a gate. The left end of the bar, being the source of majority carriers, is termed the source. The right end, being where the electrons are drained off, is called the drain. A cross-sectional drawing of a typical FET is shown in Fig. 12-21C, and three basic circuits are shown in Fig. 12-21F–H.

Insulated-gate transistors (*IGT*) are also known as field-effect transistors, metal-oxide silicon or semiconductor field-effect transistors (*MOSFET*), metal-oxide silicon or semiconductor transistors (*MOST*), and insulated-gate field-effect transistors (*IGFET*). All these devices are similar and are simply names applied to them by the different manufacturers.

The outstanding characteristics of the IGT are its extremely high input impedance, running to $10^{15}\ \Omega$. IGTs have three elements but four connections—the gate, the drain, the source, and an n-type substrate, into which two identical p-type silicon regions have been diffused. The source and drain terminals are taken from these two p-regions, which form a capacitance between the n-substrate and the silicon-dioxide insulator and the metallic gate terminals. A cross-sectional view of the internal construction appears in Fig. 12-21D, with a basic circuit shown in Fig. 12-21E. Because of the high input impedance, the IGT can easily be damaged by static charges. Strict adherence to the instructions of the manufacturer must be followed since the device can be damaged even before putting it into use.

IGTs are used in electrometers, logic circuits, and ultra sensitive electronic instruments. They should not be confused with the conventional FET used in audio equipment.

Transistor Equivalent Circuits, Current Flow, and Polarity. Transistors may be considered to be a T configuration active network, as shown in Fig. 12-22.

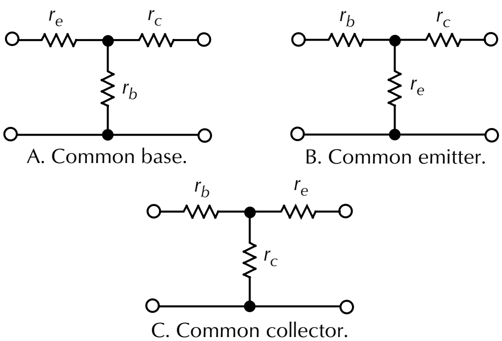


Figure 12-22. Equivalent circuits for transistors.

The current flow, phase, and impedances of the npn and pnp transistors are shown in Fig. 12-23 for the three basic configurations, common emitter, common base and common collector. Note phase reversal only takes place in the common-emitter configuration.

The input resistance for the common-collector and common-base configuration increases with an increase of the load resistance R_L. For the common emitter, the input resistance decreases as the load resistance is increased; therefore, changes of input or output resistance are reflected from one to the other.

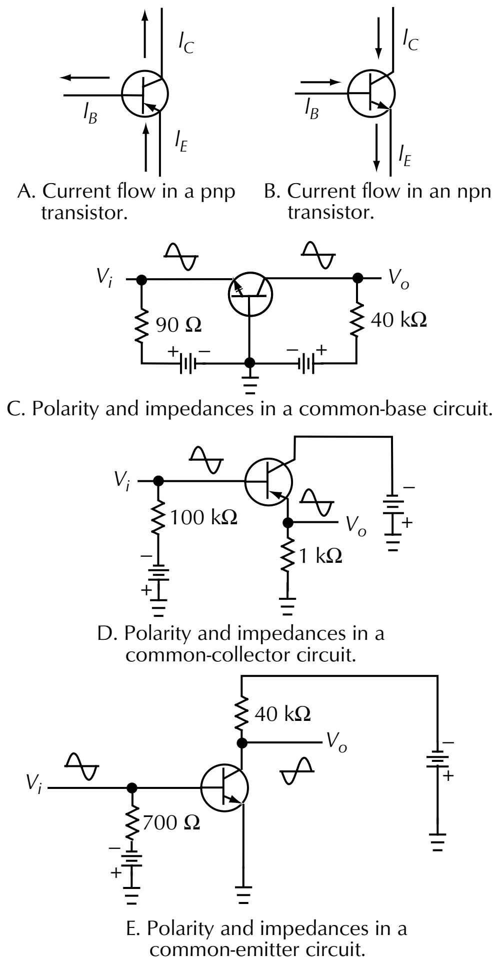


Figure 12-23. Current, polarity. and impedance relationships.

Fig. 12-24 shows the signal-voltage polarities of a p-channel field-effect transistor. Note the similarity to tube characteristics.

Voltage, power, and current gains for a typical transistor using a common-emitter configuration are shown in Fig. 12-25. The current gain decreases as the load resistance is increased, and the voltage gain increases as the load resistance is increased. Maximum power gain occurs when the load resistance is approximately 40,000 Ω, and it may exceed unity.

For the common-collector connection, the current gain decreases as the load resistance is increased and the voltage gain increases as the load resistance is increased, but it never exceeds unity. Curves such as these help the designer to select a set of conditions for a specific result.

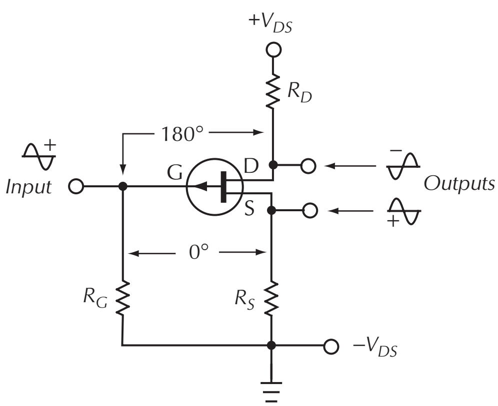


Figure 12-24. Signal-voltage polarities in a p-channel field-effect transistor (FET).

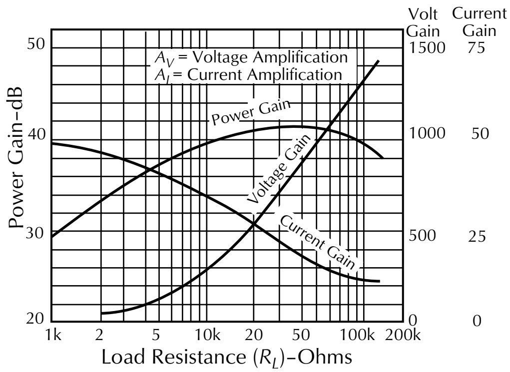


Figure 12-25. Typical voltage, power, and current gains for a conventional transistor using a common-emitter configuration.

The power gain varies as the ratio of the input to output impedance and may be calculated with the equation

$$dB = 10\log\frac{Z_o}{Z_{in}} \tag{12-28}$$

where,
Z_o is the output impedance in ohms,
Z_{in} is the input impedance in ohms.

Forward-Current-Transfer Ratio. An important characteristic of a transistor is its *forward- current-transfer ratio*, or the ratio of the current in the output to the current in the input element. Because of the many different configurations for connecting transistors, the forward transfer ratio is specified for a particular circuit configuration. The forward- current-transfer ratio for the common-base configuration is often referred to as *alpha* (α) and the common-emitter forward- current-transfer ratio as *beta* (ß). In common-base circuitry, the emitter is the input element, and the collector is the output element. Therefore, α_{dc} is the ratio of the dc collector current I_C to the dc emitter current I_E. For the common emitter, the $ß_{dc}$ is then the ratio of the dc collector current I_C to the base current I_B. The ratios are also given in terms of the ratio of signal current, relative to the input and output, or in terms of ratio of change in the output current to the input current, which causes the change.

The term α and ß are also used to denote the frequency cutoff of a transistor and is defined as the frequency at which the value of alpha for a common-base configuration, or ß for a common-emitter circuit, falls to 0.707 times its value at a frequency of 1000 Hz.

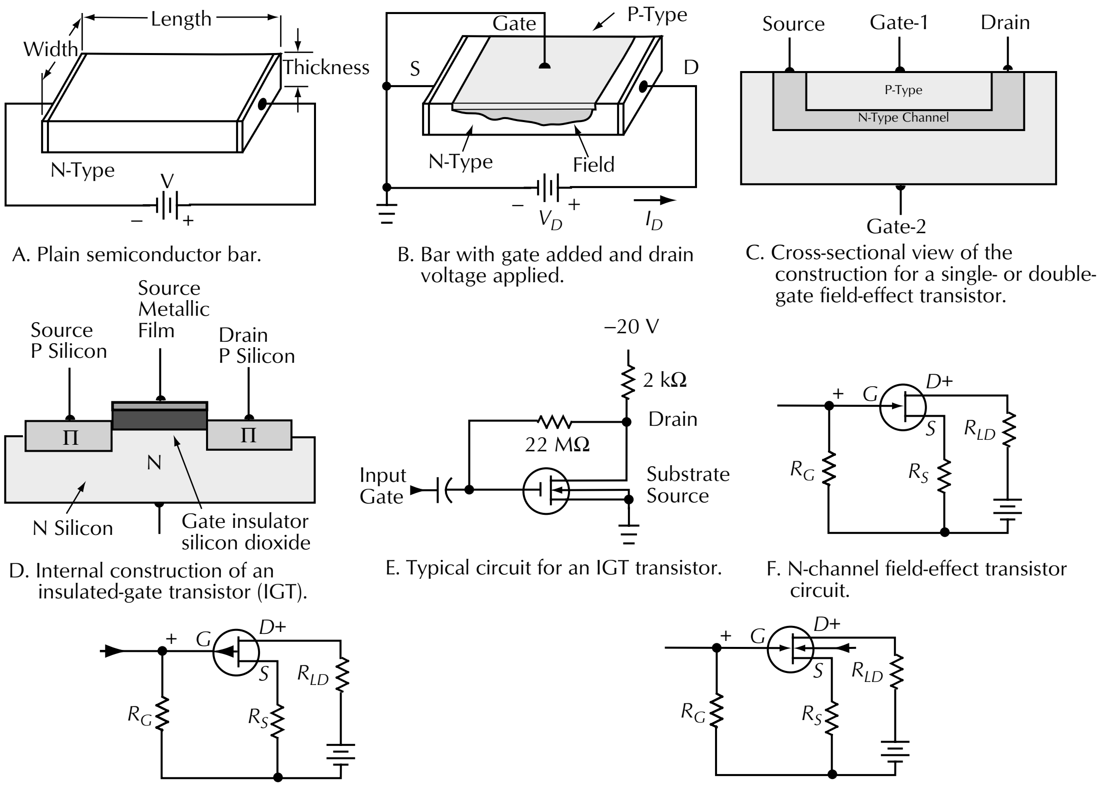
Gain-bandwidth product is the frequency at which the common-emitter forward-current-transfer ratio ß is equal to unity. It indicates the useful frequency range of the device and assists in the determination of the most suitable configuration for a given application.

Bias Circuits. Several different methods of applying bias voltage to transistors are shown in Fig. 12-26, with a master circuit for aiding in the selection of the proper circuit shown in Fig. 12-27. Comparing the circuits shown in Fig. 12-26, their equivalents may be found by making the resistors in Fig. 12-27 equal to zero or infinity for analysis and study. As an example, the circuit of Fig. 12-26D may be duplicated in Fig. 12-27 by shorting out resistors R4 and R5 in Fig. 12-27.

The circuit Fig. 12-26G employs a split voltage divider for R_2. A capacitor connected at the junction of the two resistors shunts any ac feedback current to ground. The stability of circuits A, D, and G in Fig. 12-26 may be poor unless the voltage drop across the load resistor is at least one-third the value of the power supply voltage V_{cc}. The final determining factors will be gain and stability.

Stability may be enhanced by the use of a thermistor to compensate for increases in collector current with increasing temperature. The resistance of the thermistor decreases as the temperature increases, decreasing the bias voltage so the collector voltage tends to remain constant. Diode biasing may also be used for both temperature and voltage variations. The diode is used to establish the bias voltage, which sets the transistor idling current or the current flow in the quiescent state.

When a transistor is biased to a nonconducting state, small reverse dc currents flow, consisting of leakage currents that are related to the surface characteristics of the semiconductor material and saturation currents. Saturation current increases with temperature and is related to the impurity concentration in the material.



A. Plain semiconductor bar.

B. Bar with gate added and drain voltage applied.

C. Cross-sectional view of the construction for a single- or double-gate field-effect transistor.

D. Internal construction of an insulated-gate transistor (IGT).

E. Typical circuit for an IGT transistor.

F. N-channel field-effect transistor circuit.

G. P-channel field-effect transistor circuit.

H. N-channel double-gate field-effect transistor circuit.

Figure 12-26. Basic design circuit for transistor bias circuits.

Collector-cutoff current is a dc current caused when the collector-to-base circuit is reverse biased and the emitter-to-base circuit is open. Emitter-cutoff current flows when the emitter to base is reverse biased and the collector-to-base circuit is open.

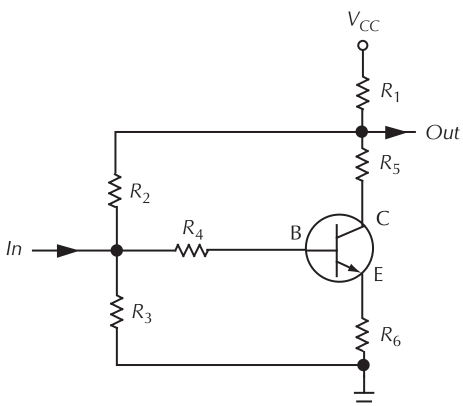


Figure 12-27. Basic bias circuits for transistors

Small and Large-Signal Characteristics. The transistor like the vacuum tube is nonlinear and can be classified as a nonlinear active device. Although the transistor is only slightly nonlinear, these nonlinearities become quite pronounced at very low and very high current and voltage levels. If an ac signal is applied to the base of a transistor without a bias voltage, conduction will take place on only one-half cycle of the applied signal voltage, resulting in a highly distorted output signal. To avoid high distortion, a dc biased voltage is applied to the transistor, and the operating point is shifted to the linear portion of the characteristic curve. This improves the linearity and reduces the distortion to a value suitable for small-signal operation. Even though the transistor is biased to the most linear part of the characteristic curve, it can still add considerable distortion to the signal if driven into the nonlinear portion of the characteristic.

Small-signal swings generally run from less than 1 μV to about 10 mV so it is important that the dc biased voltage be of large enough that the applied ac signal is small compared to the dc bias current and voltage. Transistors are normally biased at current values between 0.1 mA and 10 mA. For large-signal operation, the design procedures become quite involved mathematically and require a considerable amount of approximation and the use of nonlinear circuit analysis.

It is important to provide an impedance match between cascaded stages because of the wide difference of impedance between the input and output circuits of transistors. If the impedances are not matched, an appreciable loss of power will take place.

The maximum power amplification is obtained with a transistor when the source impedance matches the internal input resistance, and the load impedance matches the internal output resistance. The transistor is then said to be image matched.

If the source impedance is changed, it affects the internal output resistance of the transistor, requiring a change in the value of the load impedance. When transistor stages are connected in tandem, except for the grounded-emitter connection, the input impedance is considerably lower than the preceding stage output impedance, therefore an interstage transformer should be used to supply an impedance match in both directions.

When working between a grounded base and a grounded-emitter circuit, a step-down transformer is used. Working into a grounded-collector stage, a step-up transformer is used. Grounded-collector stages can also be used as an impedance-matching device between other transistor stages.

When adjusting the supply voltages for a transistor amplifier employing transformers, the *battery* voltage must be increased to compensate for the dc voltage drop across the transformer windings. The data sheets of the manufacturer should be consulted before selecting a transformer to determine the source and load impedances.

Transistor Noise Figure (*nf*). In a low-level amplifier, such as a preamplifier, noise is the most important single factor and is stated as the *SNR* or *nf*. Most amplifiers employ resistors in the input circuit which contribute a certain amount of measurable noise because of thermal activity. This power is generally about –160 dB, re: 1 W for a bandwidth of 10,000 Hz. When the input signal is amplified, the noise is also amplified. If the ratio of the signal power to noise power is the same, the amplifier is noiseless and has a noise figure of unity or more. In a practical amplifier some noise is present, and the degree of impairment is called the noise figure (*nf*) of the amplifier, expressed as the ratio of signal power to noise power at the output:

$$nf = \frac{S_1 \times N_o}{S_o \times N_1} \tag{12-29}$$

where,
S_1 is the signal power,
N_1 is the noise power,
S_o is the signal power at the output,
N_o is the noise at the output.

$$nf_{dB} = 10\log(nf \text{ of the power ratio}) \tag{12-30}$$

For an amplifier with various *nf*, the *SNR* would be:

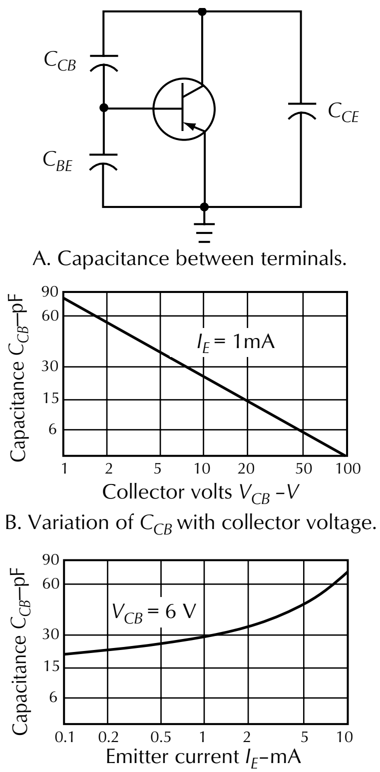
nf	*SNR*
1 dB	1.26
3 dB	2
10 dB	10
20 dB	100

An amplifier with an *nf* below 6 dB is considered excellent.

Low-noise factors can be obtained by the use of an emitter current of less than 1 mA, a collector voltage of less than 2 V, and a signal-source resistance below 2000 Ω.

Internal Capacitance. The paths of internal capacitance in a typical transistor are shown in Fig. 12-28. The width of the pn junction in the transistor varies in accordance with voltage and current, the internal capacitance also varies. Variation of collector-base capacitance *C* with collector voltage and emitter current is shown in Figs. 12-28B and C. The increase in the width of the pn junction between the base and collector, as the reverse bias voltage (V_{CB}) is increased, is reflected in lower capacitance values. This phenomenon is equivalent to increasing the spacing between the plates of a capacitor. An increase in the emitter current, most of which flows through the base-collector junction, increases the collector-base capacitance (C_{CB}). The increased current through the pn junction may be considered as effectively reducing the width of the pn junction. This is equivalent to decreasing the spacing between the plates of a capacitor, therefore, increasing the capacitance.

The average value of collector-base capacitance (C_{CB}) varies from 2–50 pF, depending on the type transistor and the manufacturing techniques. The collector-emitter capacitance is caused by the pn junction. It normally is five to ten times greater than that of the collector-base capacitance and will vary with the emitter current and collector voltage.



A. Capacitance between terminals.

B. Variation of C_{CB} with collector voltage.

C. Variation of C_{CB} with emitter current.

Figure 12-28. Internal capacitance of a transistor.

Punch-Through. *Punch-through* is the widening of the space charge between the collector element and the base of a transistor. As the potential V_{CB} is increased from a low to a high value, the collector-base space-charge is widened. This widening effect of the space-charge narrows the effective width of the base. If the diode space-charge does not avalanche before the space-charge spreads to the emitter section, a phenomenon termed punch-through is encountered, as shown in Fig. 12-29.

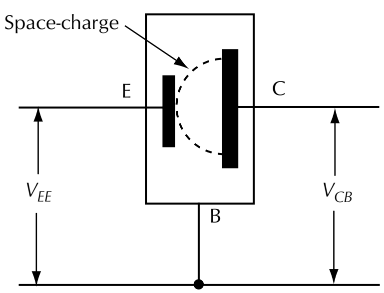


Figure 12-29. Spreading of the space charge between the emitter and the collector, which creates *punch-through*.

The effect is the base disappears as the collector-base space-charge layer contacts the emitter, creating relatively low resistance between the emitter and the collector. This causes a sharp rise in the current. The transistor action then ceases. Because there is no voltage breakdown in the transistor, it will start functioning again if the voltage is lowered to a value below where punch-through occurs.

When a transistor is operated in the punch-through region, its functioning is not normal, and heat is generated internally that can cause permanent damage to the transistor.

Breakdown Voltage. *Breakdown voltage* is that voltage value between two given elements in a transistor at which the crystal structure changes and current begins to increase rapidly. Breakdown voltage may be measured with the third electrode open, shorted, or biased in either the forward or reverse direction. A group of collector characteristics for different values of base bias are shown in Fig. 12-30. The collector- to-emitter breakdown voltage increases as the base-to-emitter bias is decreased from the normal forward values through zero to reverse. As the resistance in the base-to-emitter circuit decreases, the collector characteristics develop two breakdown points. After the initial breakdown, the collector-to-emitter voltage decreases with an increasing collector current, until another breakdown occurs at the lower voltage.

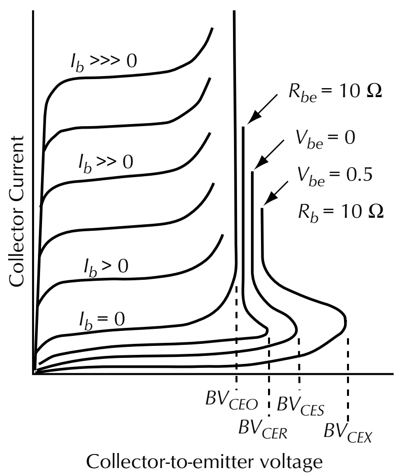


Figure 12-30. Typical collector characteristic curves showing locations of various breakdown voltages.

Breakdown can be very destructive in power transistors. A breakdown mechanism, termed *second breakdown*, is an electrical and thermal process in which current is concentrated in a very small area. The high current, together with the voltage across the transistor, causes intense heating, melting a hole from the collector to the emitter. This causes a short circuit and internal breakdown of the transistor.

The fundamental limitation to the use of transistors is the breakdown voltage (BV_{cer}) The breakdown voltage is not sharp so it is necessary to specify the value of collector current at which breakdown will occur. This data is obtained from the data sheet of the manufacturer.

Transistor Load Lines. Transistor load lines are used to design circuits. An example of circuit design uses a transistor with the following characteristics:

Maximum collector current:	10 mA
Maximum collector voltage:	–22 V
Base current:	0 to 300 µA
Maximum power dissipation:	300 mW

The base-current curves are shown in Fig. 12-31A. The amplifier circuit is to be Class A, using a common-emitter circuit, as shown in Fig. 12-31B. By proper choice of the operating point, with respect to the transistor characteristics and supply voltage, low distortion, Class-A performance is easily obtained within the transistor power ratings.

The first requirement is a set of collector-current, collector-voltage curves for the transistor to be employed. Such curves can generally be obtained from the data sheets of the manufacturer. Assuming that such data is at hand and referring to Fig. 12-31A, a curved line is plotted on the data sheet, representing the maximum power dissipation by the use of the equation

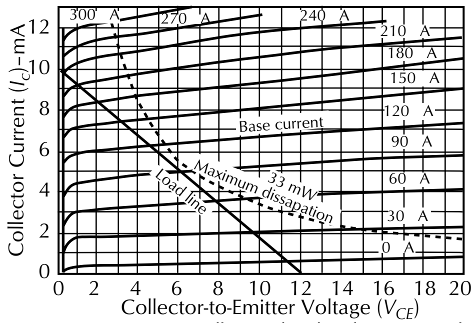
$$I_c = \frac{P_c}{V_c} \tag{12-31}$$

or

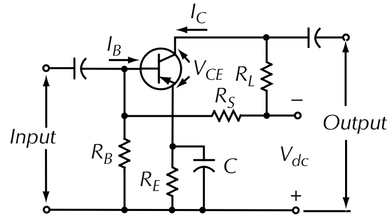
$$V_c = \frac{P_c}{I_c} \tag{12-32}$$

where,
I_c is the collector current,
P_c is the maximum power dissipation of the transistor,
V_c is the collector voltage.

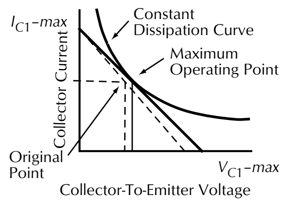
At any point on this line at the intersection of V_cI_c, the product equals 0.033 W or 33 mW. In determining the points for the dissipation curve, voltages are selected along the horizontal axis and the corresponding current is equated using:



A. Common-emitter-collector family of curves, with load line and maximum dissipation power curve.



B. Amplifier circuit used for load-line calculations.



C. Load line moved to right for maximum power output. Dotted lines are the original load line and operating point.

Figure 12-31. Load-line calculation curves.

$$I_C = \frac{P_C}{V_{CE}} \tag{12-33}$$

The current is determined for each of the major collector-voltage points, starting at 16 V and working backward until the upper end of the power curve intersects the 300 µA base-current line. After entering the value on the graph for the power dissipation curve, the area to the left of the curve encompasses all points within the maximum dissipation rating of the transistor. The area to the right of the curve is the overload region and is to be avoided.

The operating point is next determined. A point that results in less than a 33 mW dissipation is selected somewhere near the center of the power curve. For this example, a 5 mA collector current at 6 V, or a dissipation of 30 mW, will be used. The selected point is indicated on the graph and circled for reference. A line is drawn through the dot to the maximum collector current, 10 mA, and downward to intersect the V_{CE} line at the bottom of the graph, which, for this example, is 12 V. This line is termed the *load line*. The load resistance R_L may be computed with

$$R_L = \frac{dV_{CE}}{dI_C} \tag{12-34}$$

where,
R_L is the load resistance,
dV_{CE} is the range of collector-to-emitter voltage,
dI_C is the range of collector current.

In the example

$$\begin{aligned} R_L &= \frac{0-12}{0-0.01} \\ &= \frac{12}{0.01} \\ &= 1200\Omega \end{aligned}$$

Under these conditions, the entire load line dissipates less than the maximum value of 33 mW, with 90 μA of base current and 5 mA of collector current. The required base current of 90 μA may be obtained by means of one of the biasing arrangements shown in Fig. 12-26.

To derive the maximum power output from the transistor, the load line may be moved to the right and the operating point placed in the maximum dissipation curve, as shown in Fig. 12-31C. Under these conditions, an increase in distortion may be expected. As the operating point is now at 6.5 V and 5 mA, the dissipation is 33 mW. Drawing a line through the new operating point and 10 mA (the maximum current), the voltage at the lower end of the load line is 13.0 V; therefore, the load impedance is now 1300 Ω.

12.3 Integrated Circuits

An *integrated circuit* (IC) is a device consisting of hundreds and even thousands of components in one small enclosure, and came into being when manufacturers learned how to grow and package semiconductors and resistors.

The first ICs were small scale and usually too noisy for audio circuits; however, as time passed, the noise was reduced, stability increased, and the operational amplifier (op-amp) IC became an important part of the audio circuit. With the introduction of medium-scale integration (MSI) and large-scale integration (LSI) circuits, power amplifiers were made on a single chip with only capacitors, gain, and frequency compensation components externally connected.

Typical circuit components might use up a space 4 mils × 6 mils (1 mil = 0.001 inch) for a transistor, 3 mils × 4 mils for a diode and 2 mils × 12 mils for a resistor. These components are packed on the surface of the semiconductor wafer and interconnected by a metal pattern that is evaporated into the top surface. Leads are attached to the wafer that is then sealed and packaged in several configurations, depending on their complexity.

ICs can be categorized by their method of fabrication or use. The most common are monolithic or hybrid and linear or digital. Operational amplifiers and most analog circuits are linear while flip-flops and on-off switch circuits are digital.

An IC is considered *monolithic* if it is produced on one single chip and *hybrid* if it consists of more than one monolithic chip tied together and/or includes discrete components such as transistors, resistors, and capacitors.

With only a few external components, ICs can perform math functions, such as trigonometry, squaring, square roots, logarithms and antilogarithms, integration, and differentiation. ICs are well suited to act as voltage comparators, zero-crossing detectors, ac and dc amplifiers, audio and video amplifiers, null detectors, and sine-, square-, or triangular-wave generators, and all at a fraction of the cost of discrete-device circuits.

12.3.1 Monolithic Integrated Circuits

All circuit elements, both active and passive, are formed at the same time on a single wafer. The same circuit can be repeated many times on a single wafer and then cut to form individual 50 mil^2 ICs.

Bipolar transistors are often used in ICs and are fabricated much like the discrete transistor by the planar process. The differences are the contact- to-the-collector region is through the top surface rather than the substrate, requiring electrical isolation between the substrate and the collector. The integrated transistor is isolated from other components by a pn junction that creates capacitance, reducing high-frequency response and increasing leakage current, which in low-power circuits can be significant.

Integrated diodes are produced the same way as transistors and can be regarded as transistors whose terminals have been connected to give the desired characteristics.

Resistors are made at the same time as transistors. The resistance is characterized in terms of its sheet resistance, which is usually 100–200 Ω/square material for diffused resistors and 50–150 Ω/square material for deposited resistors. To increase the value of a resistor, square materials are simply connected in series.

It is very difficult to produce resistors with much closer tolerance than 10%; however, it is very easy to produce two adjacent resistors to be almost identical. When making comparator-type circuits, the circuits are balanced and are made to perform on ratios rather than absolute values. Another advantage is uniformity in temperature. As the temperature of one component varies, so does the temperature of the other components, allowing good tracking between components and circuits so integrated circuits are usually more stable than discrete circuits.

Capacitors are made as thin-film integrated capacitors or junction capacitors. The thin-film integrated capacitor has a deposited metal layer and a n+ layer isolated with a carrier-free region of silicon dioxide. In junction capacitors, both layers are diffused low-resistance semiconductor materials. Each layer has a dopant of opposite polarity; therefore, the carrier-free region is formed by the charge depleted area at the pn junction.

The MOSFET transistor has many advantages over the bipolar transistor for use in ICs as it occupies only $1/25$ the area of the bipolar equivalent due to lack of isolation pads. The MOSFET acts like a variable resistor and can be used as a high-value resistor. For instance, a 100 kΩ resistor might occupy only 1 mil^2 as opposed to 250 mil^2 for a diffused resistor.

The chip must finally be connected to terminals or have some means of connecting to other circuits, and it must also be packaged to protect it from the environment. Early methods included using fine gold wire to connect the chip to contacts. This was later replaced with aluminum wire ultrasonically bonded.

Flip-chip and beam-lead methods eliminate the problems of individually bonding wires. Relatively thick metal is deposited on the contact pads before the ICs are separated from the wafer. The deposited metal is then used to contact a matching metal pattern on the substrate. In the flip-chip method, globules of solder deposited on each contact pad ultrasonically bond the chip to the substrate.

In the beam-lead method, thin metal tabs lead away from the chip at each contact pad. The bonding of the leads to the substrate reduces heat transfer into the chip and eliminates pressure on the chip.

The chip is finally packaged in either hermetically sealed metal headers or is encapsulated in plastic, which is an inexpensive method of producing ICs.

12.3.2 Hybrid Integrated Circuits

Hybrid circuits combine monolithic and thick- and thin-film discrete components for obtaining the best solution to the design.

Active components are usually formed as monolithics; however, sometimes discrete transistors are soldered into the hybrid circuit.

Passive components such as resistors and capacitors are made with thin- and thick-film techniques. Thin films are 0.001–0.1 mil thick, while thick films are normally 60 mils thick.

Resistors can be made with a value from ohms to megohms with a tolerance of 0.05% or better.

High-value capacitors are generally discrete, miniature components that are welded or soldered into the circuit, and low-value capacitors can be made as film capacitors and fabricated directly on the substrate.

Along with being certain that the components will fit into the hybrid package, the temperature must also be taken into account. The temperature rise T_R of the package can be calculated with the following equation:

$$\begin{aligned} T_R &= T_C - T_A \\ &= P_T T \theta_{CA} \end{aligned} \tag{12-35}$$

where,
T_C is the case temperature,
T_A is the ambient temperature,
P_T is the total power dissipation,
θ_{CA} is the case-to-ambient thermal resistance.

The θ_{CA} for a package in free air can be approximated at 35°C/W/in^2 or a device will have a 35°C rise in temperature above ambient if 1 W is dissipated over an area of 1 in^2.

12.3.3 Operational Voltage Amplifiers (op-amp)

One of the most useful ICs for audio is the op-amp. Op-amps can be made with discrete components, but they would be very large and normally unstable to temperature and external noise.

An op-amp normally has one or more of the following features:

- Very high input impedance (> 10^6–10^{12} Ω),
- Very high open-loop (no feedback) gain,
- Low output impedance (< 200 Ω),
- Wide frequency response (> 100 MHz),

- Low input noise,
- High symmetrical slew rate and/or high input dynamic range,
- Low inherent distortion.

By adding external feedback paths, gain, frequency response, and stability, can be controlled.

Op-amps are normally two-input differential devices; one input inverting the signal, and the second input not inverting the signal, hence called *noninverting*. Several typical op-amp circuits are shown in Fig. 12-32.

Because there are two inputs of opposite polarity, the output voltage is the difference between the inputs where

$$E_{O(+)} = A_V E_2 \tag{12-36}$$

$$E_{O(-)} = A_V E_1 \tag{12-37}$$

E_O is calculated with the equation

$$E_O = A_V \times (E_1 - E_2) \tag{12-38}$$

Often one of the inputs is grounded, either through a direct short or a capacitor, therefore, the gain is either

$$E_O = A_V E_1 \tag{12-39}$$

or

$$E_O = A_V E_2 \tag{12-40}$$

To provide both a positive and negative output with respect to ground, a positive and negative power supply is required, as shown in Fig. 12-33. The supply should be regulated and filtered. Often a + and − power supply is not available, such as in an automobile, so the op-amp must operate on a single supply, as shown in Fig. 12-34. In this supply, the output dc voltage is set by adjusting R_1 and R_2 so the voltage at the noninverting input is about one-third the power supply voltage.

The diodes and zener diodes in Fig. 12-35 are used to protect the op-amp from damage caused by transients, reverse voltage, and overdriving. D_6 and D_7 clip the inputs before overdriving, D_1 and D_2 protect against reverse polarity, D_4 and D_5 regulate the supply, and D_3 limits the total voltage across the op-amp.

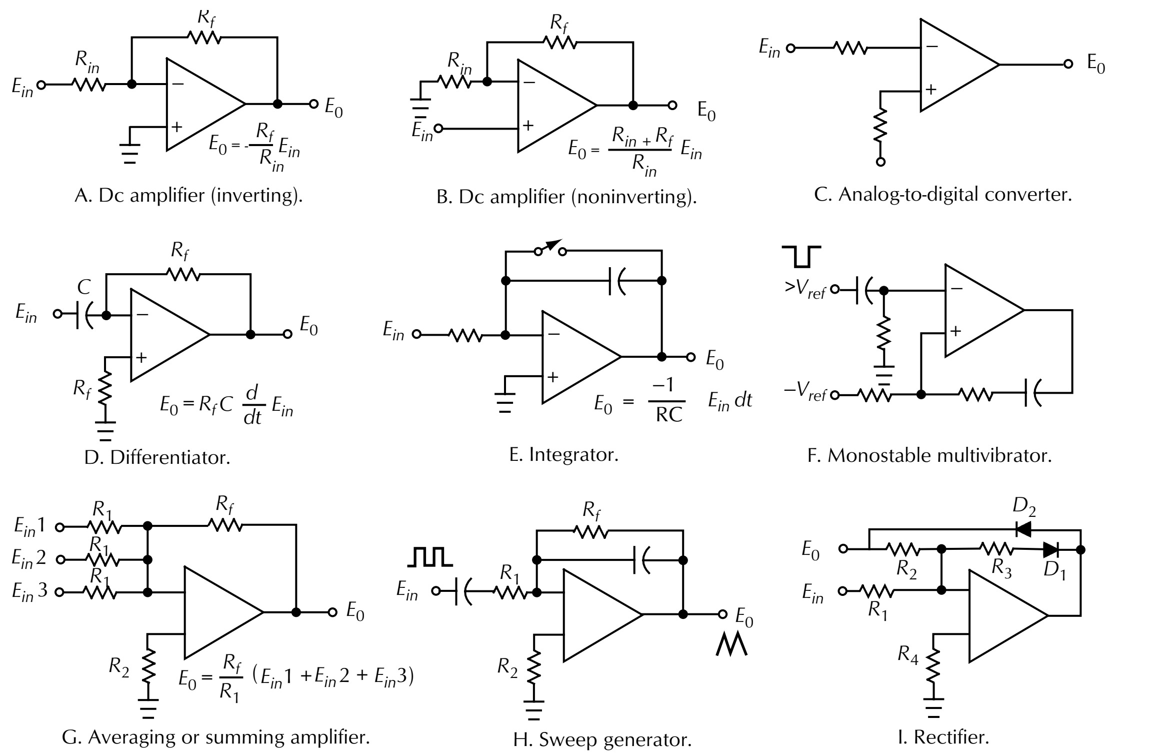


Figure 12-32. Typical op-amp circuits.

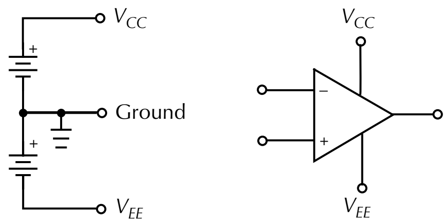


Figure 12-33. Positive and negative type power supply.

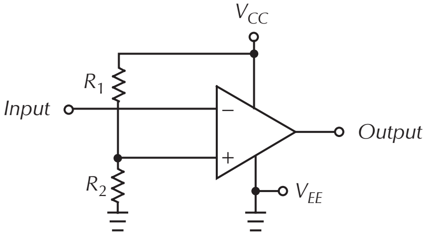


Figure 12-34. Simple circuit for operating on a single ended power supply.

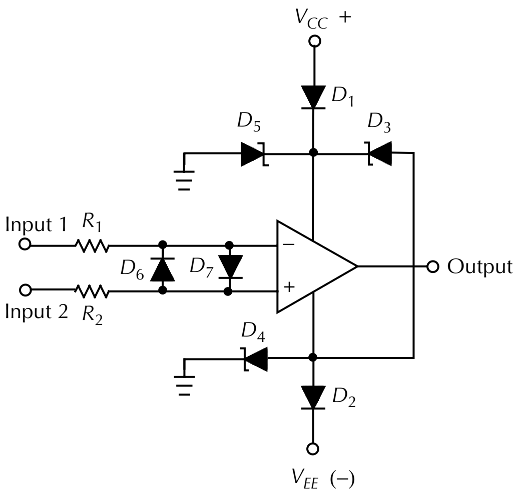


Figure 12-35. Diode protection circuits for op-amps.

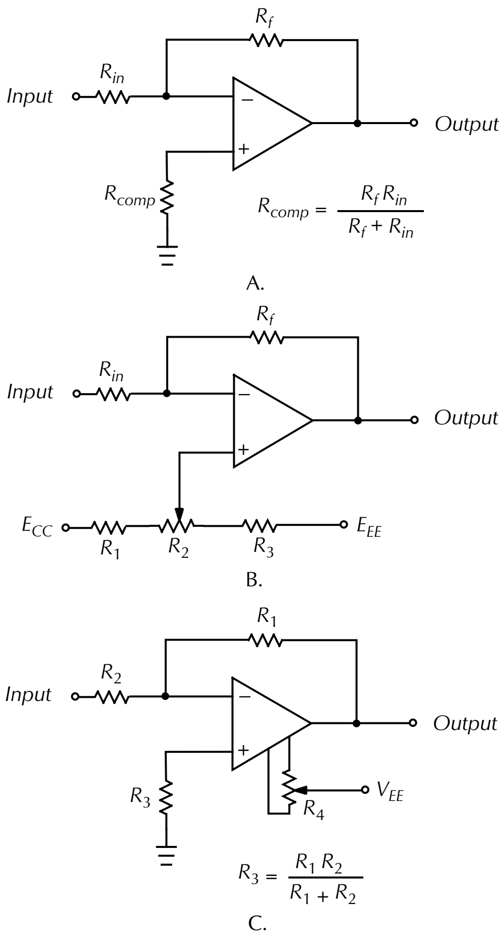


Figure 12-36. Various methods of correcting dc error.

The dc error factors result in an output offset voltage E_{Oo}, which exists between the output and ground when it should be zero. The dc offset error is most easily corrected by supplying a voltage differential between the inverting and noninverting inputs, which can be accomplished by one of several methods, Fig. 12-36. Connecting the feedback resistor R_f usually causes an offset and can be found with the equation

$$E_{Oo} = I_{bias} R_f \tag{12-41}$$

To obtain minimum offset, make the compensating resistor shown in Fig. 12-36A equal to

$$R_{comp} = \frac{R_f R_{in}}{R_f + R_{in}} \tag{12-42}$$

If this method is not satisfactory, the methods of Figs. 12-36B or C might be required.

Many op-amps are internally compensated. Often it is advantageous to compensate a device externally to optimize bandwidth and slew rate, lowering distortion. Internally compensated op-amp ICs come in standard packages—the 8 pin TO-99 metal can, the 8 pin dual-in-line package (MINI DIP), and the 14 pin DIP.

Inverting Amplifiers. In the *inverting amplifier* the + input is grounded and the signal is applied to the minus (–) input, Fig. 12-37. The output of the circuit is determined by the input resistor R_1 and the feedback resistor R_f.

$$E_O = I_{in}\left(\frac{-R_f}{R_1}\right) \tag{12-43}$$

where,
E_{in} is the signal input voltage in volts,
R_f is the feedback resistor in ohms,
R_1 is the input resistor in ohms.

The low frequency rolloff is

$$f_C = \frac{1}{2\pi R_1 C_1} \tag{12-44}$$

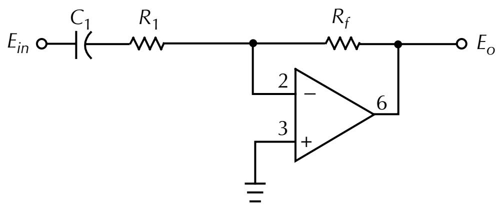


Figure 12-37. A simple inverting amplifier.

Noninverting Amplifier. In the *noninverting amplifier,* Fig. 12-38. the signal is applied to the plus input, while the minus input is part of the feedback loop. The output is

$$E_O = I_{in}\left(\frac{1+R_f}{R_1}\right) \tag{12-45}$$

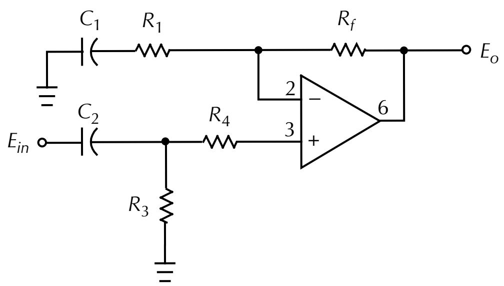


Figure 12-38. A simple noninverting amplifier.

The low frequency rolloff is in two steps.

$$f_{C_1} = \frac{1}{2\pi R_1 C_1} \tag{12-46}$$

$$f_{C_2} = \frac{1}{2\pi R_3 C_2} \tag{12-47}$$

To keep low frequency noise gain at a minimum, keep $f_{C_1} > f_{C_2}$.

Power Supply Compensation. The power supply for wide band op-amp circuits should be bypassed with capacitors, Fig. 12-39A, between the plus and minus pin and common. The leads should be as short as possible and as close to the IC as possible. If this is not possible, bypass capacitors should be on each printed circuit board.

Input Capacitance Compensation. Stray input capacitance can lead to oscillation in feedback op-amps because it represents a potential phase shift at the frequency of

$$f = \frac{1}{2\pi R_f C_s} \tag{12-48}$$

where,
R_f is the feedback resistor,
C_s is the stray capacitance.

One way to reduce this problem is to keep the value of R_f low. The most useful way, however, is to add a compensation capacitor, C_f, across R_f as shown in Fig. 12-39B. This makes C_f/R_f and C_s/R_{in} a frequency compensated divider.

Output Capacitance Compensation. Output capacitance greater than 100 pF can cause problems, requiring a series resistor R_o being installed between the output of the IC and the load and stray capacitance as shown in Fig. 12-39C. The feedback resistor (R_f) is connected after R_o to compensate for the loss in signal caused by R_o. A compensating capacitor (C_f) bypasses R_f to reduce gain at high frequencies.

Gain and Bandwidth. A perfect op-amp would have infinite gain and infinite bandwidth. In real life however, the dc open loop voltage gain is around 100,000 or 100 dB and the bandwidth where gain is 0 is 1 MHz, Fig. 12-40.

To determine the gain possible in an op-amp, for a particular bandwidth, determine the bandwidth, follow vertically up to the open loop gain response curve and horizontally to the voltage gain. This, of course, is with no feedback at the upper frequency. For example, for a frequency bandwidth of 0–10 kHz, the maximum gain of the op-amp in Fig. 12-40 is 100. To have lower distortion, it would be better to have feedback at the required upper frequency limit. To increase this gain beyond 100 would require a better op-amp or two op-amps with lower gain connected in series.

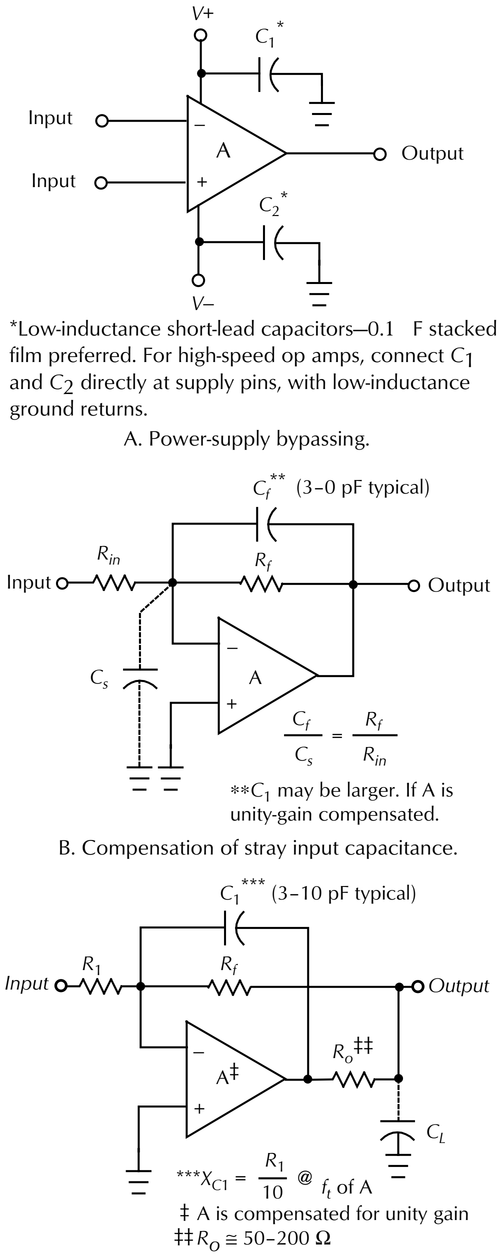


Figure 12-39. Stability enhancement techniques.

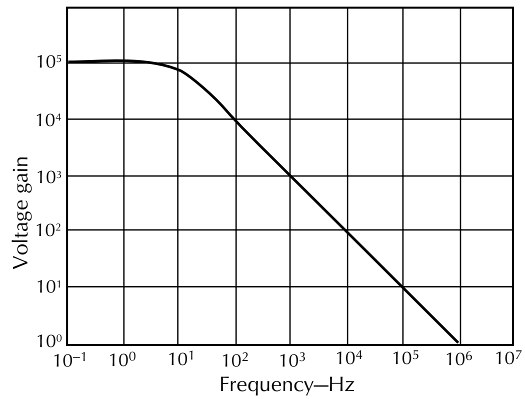


Figure 12-40. Typical open loop gain response.

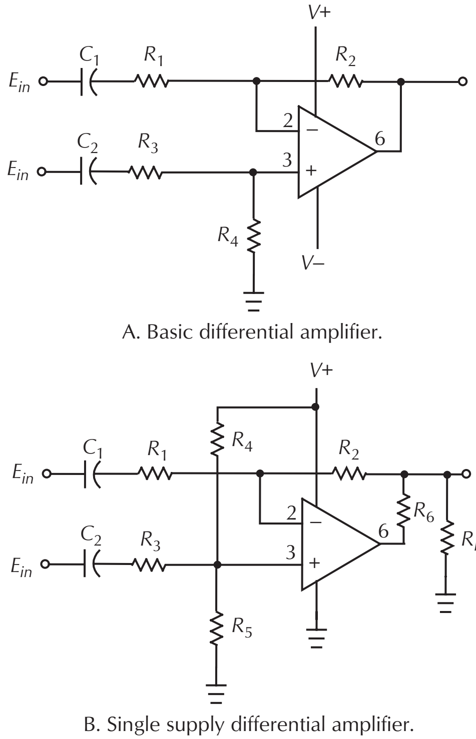


Figure 12-41. Differential amplifiers.

Differential Amplifiers. Two *differential amplifier* circuits are shown in Fig. 12-41. The ability of the differential amplifier to block identical signals is useful to reduce hum and noise that is picked up on input lines such as in low-level microphone circuits. This rejection is called common-mode rejection and sometimes eliminates the need for an input transformer.

In Fig. 12-41A, capacitors C_1 and C_2 block dc from the previous circuit and provide a 6 dB/octave rolloff below

$$f_{C_1} = \frac{1}{2\pi R_1 C_1} \tag{12-49}$$

$$f_{C_2} = \frac{1}{2\pi(R_3 + R_4)C_2} \tag{12-50}$$

The output voltage is

$$E_O = (E_{in_2} - E_{in_1})\frac{R_2}{R_1}. \tag{12-51}$$

To reduce the common-mode Rejection Ratio (CMRR),

$$\frac{R_2}{R_1} \equiv \frac{R_4}{R_3} \tag{12-52}$$

and

$$f_{C_1} = f_{C_2} \tag{12-53}$$

Summing Inverter Amplifiers. In the summing inverter, Fig. 12-32G, the virtual ground characteristic of the amplifier's summing point is used to produce a scaling adder. In this circuit, I_{in} is the algebraic sum of the number of inputs.

$$\begin{aligned} I_{in_1} &= \frac{E_{in_1}}{R_{in_1}} \\ I_{in_2} &= \frac{E_{in_2}}{R_{in_2}} \\ I_{in_n} &= \frac{E_{in_n}}{R_{in_n}} \end{aligned} \tag{12-54}$$

and the total input current is

$$\begin{aligned} I_{in} &= I_{in_1} + I_{in_2} + \ldots I_{in_n} \\ &= I_f \end{aligned} \tag{12-55}$$

and

$$I_f = \frac{-E_O}{R_f} \tag{12-56}$$

therefore

$$I_{in_1} + I_{in_2} + \ldots I_{in_n} = \frac{-E_o}{R_f}. \tag{12-57}$$

The output voltage is found with the equation

$$E_O = \left[R_{in_1}\left(\frac{R_f}{R_{in_1}}\right) + R_{in_2}\left(\frac{R_f}{R_{in_2}}\right) + \ldots R_{in_n}\left(\frac{R_f}{R_{in_n}}\right)\right]. \tag{12-58}$$

It is interesting that even though the inputs mix at one point, all signals are isolated from each other and one signal does not effect the others and one impedance does not effect the rest.

Operational Transconductance Amplifiers. The *operational transconductance amplifier* (OTA) provides transconductance gain and current output rather than voltage gain and output as in an operational amplifier. The output is the product of the input voltage and amplifier transconductance, and it can be considered an infinite impedance current generator.

Varying the bias current on the OTA can completely control the open-loop gain of the device and can also control the total power input.

OTAs are useful as multipliers, automatic gain control (agc) amplifiers, sample and hold circuits, multiplexers, and multivibrators to name a few.

12.3.4 Dedicated Analog Integrated Circuits for Audio Applications

By Les Tyler and Wayne Kirkwood, THAT Corp.

The first ICs used in audio applications were general-purpose op-amps like the famous Fairchild µA741. Early op-amps like the classic 741 generally had drawbacks that limited their use in professional audio, from limited slew rate to poor clipping behavior.

Early on, integrated circuit manufacturers recognized that the relatively high-volume consumer audio market would make good use of dedicated ICs tailored to specific applications such as phono preamplifiers and companders. The National LM381 preamplifier and Signetics NE570 compander addressed the needs of consumer equipment makers producing high-volume products such as phono preamplifiers and cordless telephones. Operational Transconductance Amplifiers, such as the RCA CA3080, were introduced around 1970 to primarily serve the industrial market. It was not long before professional audio equipment manufacturers adapted OTAs for professional audio use as early voltage controlled amplifiers or “VCAs.” However, through the 1970s all these integrated circuits were intended more for use in consumer and industrial applications than professional audio.

In the mid-1970s, semiconductor manufacturers began to recognize that professional audio had significantly different requirements from the needs of consumer audio or industrial products. The Philips TDA1034 was the first op-amp to combine low noise, 600 Ω drive capability and high slew rate—all important characteristics to pro audio designers. Shortly after its introduction, Philips transferred production of the TDA1034 to the newly purchased Signetics division which re-branded it the NE5534. At about the same time, Texas Instruments and National Semiconductor developed general-purpose op-amps using a combination of bipolar and FET technology (the TI TLO70- and TLO80- series, and the National LF351-series, sometimes called "BIFET"). These parts offered high slew rates, low distortion, and modest noise (though not the 600 Ω drive capability of the 5534). While not specifically aimed at pro audio, these characteristics made them attractive to pro audio designers. Along with the NE5534, these op-amps became pro audio industry standards much like the 12AX7 of the vacuum tube era.

Op-amps are fundamentally general-purpose devices. The desire to control gain via a voltage, and the application of such technology to tape noise reduction, in particular, created a market for integrated circuits that were dedicated to a specific function. This paralleled the way that phono preamplifiers spawned ICs designed for preamplification. In many ways, the VCA drove the development of early pro audio ICs.

The design of audio VCAs benefitted from the early work of Barrie Gilbert, inventor of the "Gilbert Cell" multiplier, who in 1968 published "a precise four-quadrant multiplier with subnanosecond response."[1] Gilbert discovered a current-mode analog multiplication cell using current mirrors that was linear with respect to both of its inputs. Although its primary appeal at the time was to communications system designers working at RF frequencies, Gilbert laid the groundwork for many audio VCA designs.

In 1972, David E. Blackmer received US Patent 3,681,618 for an "RMS Circuit with Bipolar Logarithmic Converter" and in the following year patent 3,714,462 for a "Multiplier Circuit" useful as an audio voltage-controlled amplifier. Unlike Gilbert, Blackmer used the logarithmic properties of bipolar transistors to perform the analog computation necessary for gain control and rms level detection. Blackmer's development was targeted at professional audio.[2,3]

Blackmer's timing could not have been better as the number of recording tracks expanded and, due to reduced track width coupled with the effect of summing many tracks together, tape noise increased. The expanded number of recorded tracks also increased mix complexity. Automation became a desirable feature for recording consoles because there just were not enough hands available to operate the faders.

Companies such as dbx Inc. and Dolby Laboratories benefited from this trend with tape noise reduction technologies, and in the case of dbx, VCAs for console automation. Blackmer's discrete transistor-based rms level detectors and VCAs, made by dbx, were soon used in companding multi-track tape noise reduction and console automation systems.

The early Blackmer VCAs used discrete NPN and PNP transistors that required careful selection to match each other. Blackmer's design would benefit greatly from integration into monolithic form. For some time this proved to be very difficult. Nonetheless, Blackmer's discrete audio VCAs and Gilbert's transconductance cell laid the groundwork for dedicated audio integrated circuits. VCAs became a major focus of audio IC development.

Electronic music, not professional recording, primarily drove the early integration of monolithic VCAs and dedicated audio ICs. In 1976, Ron Dow of Solid State Music (SSM) and Dave Rossum of E-mu Systems developed some of the first monolithic ICs for analog synthesizers. SSM's first product was the SSM2000 monolithic VCA.[4] Solid State Music, later to become Solid State Microtechnology, developed an entire line of audio ICs including microphone preamplifiers, VCAs, voltage-controlled filters, oscillators and level detectors. Later, Douglas Frey developed a VCA topology known as the operational voltage-controlled element, "OVCE," that was first used in the SSM2014.[5] Doug Curtis, of Interdesign and later founder of Curtis Electro Music (CEM), also developed a line of monolithic ICs for the synthesizer market that proved to be very popular with manufacturers such as Oberheim, Moog, and ARP.[6] VCAs produced for electronic music relied on NPN transistor gain cells to simplify integration.

In the professional audio market, Paul Buff of Valley People, David Baskind and Harvey Rubens of VCA Associates, and others in addition to Blackmer also advanced discrete VCA technology. Baskind and Rubens eventually produced a VCA integrated circuit that ultimately became the Aphex/VCA Associates "1537."[7]

Blackmer's VCAs and rms detectors used the precise logarithmic characteristics of bipolar transistors to perform mathematical operations suitable for VCAs and rms detection. The SSM, CEM, and Aphex products used variations on the linear multiplier, where a differential pair, or differential quad, is varied to perform

VCA functions and analog voltage-controlled filtering. Close transistor matching and control of temperature-related errors are required for low distortion and control feed-through in all VCA topologies.

The Gilbert multiplier, the CA3080-series of OTAs, and the VCAs produced by SSM, CEM, and Aphex all relied solely on NPN transistors as the gain cell elements. This greatly simplified the integration of the circuits. Blackmer's log-antilog VCAs required, by contrast, precisely matched NPN and PNP transistors. This made Blackmer's VCAs the most difficult to integrate. dbx finally introduced its 2150-series monolithic VCAs in the early 1980s, almost six years after the introduction of the SSM2000.[8]

Many of the earlier developers of VCAs changed ownership or left the market as analog synthesis faded. Analog Devices currently produces many of the SSM products after numerous ownership changes. THAT Corporation assumed the patent portfolio of dbx Inc. Today Analog Devices, THAT Corporation, and Texas Instruments' Burr Brown division are the primary manufacturers making analog integrated circuits specifically for the professional audio market.

12.3.4.1 Voltage-Controlled Amplifiers

Modern integrated circuit voltage controlled amplifiers take advantage of the inherent and precise matching of monolithic transistors which, when combined with on-chip trimming, lowers distortion to very low levels. Two types of IC audio VCAs are commonly used and manufactured today: those based on Douglas Frey's Operational Voltage Controlled Element "OVCE"[9] and those based on David Blackmer's bipolar log-antilog topology.[10]

The Analog Devices SSM2018. The Frey OVCE gain cell was first introduced in the SSM2014 manufactured by Solid State Microtechnology (SSM).[11] SSM was acquired by Precision Monolithics, Inc, which was itself acquired by Analog Devices, who currently offers a Frey OVCE gain cell branded the SSM2018T. Frey's original patents, US 4,471,320 and US 4,560,947, built upon the work of David Baskind and Harvey Rubens (see US patent 4,155,047) by adding corrective feedback around the gain cell core.[12,13,14]. Fig. 12-42 shows a block diagram of the SSM2018T VCA.

The OVCE is unique in that it has two outputs: V_G and $V_{1\text{-}G}$. As the V_G output increases gain with respect to control voltage, the $V_{1\text{-}G}$ output attenuates. The result is that the audio signal "pans" from one output to the other as the control voltage is changed.

The following expression shows how this circuit works mathematically:

$$\begin{aligned} V_{out1} &= V_G \\ &= 2K \times V_{in} \end{aligned} \tag{12-59}$$

and

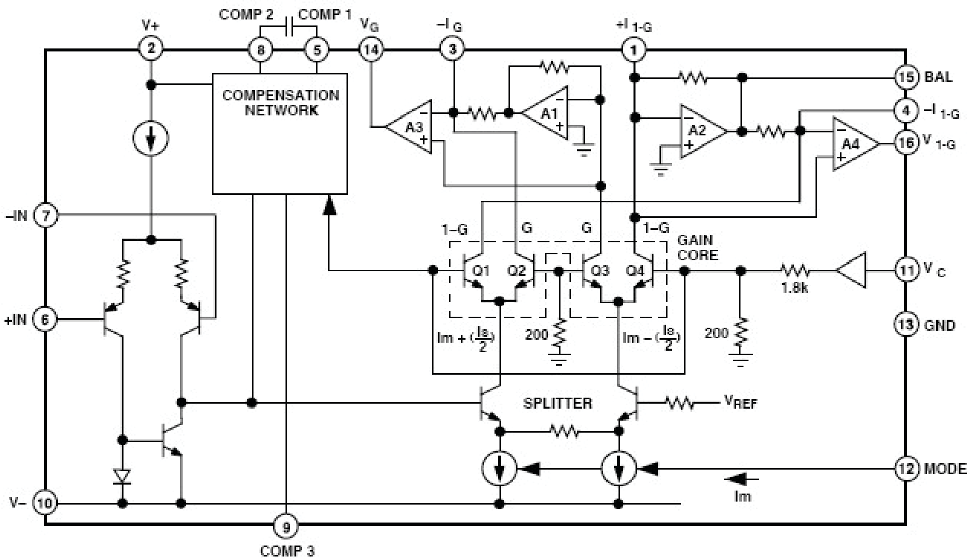


Figure 12-42. A block diagram of the SSM2018T VCA. Courtesy Analog Devices, Inc.

$$\begin{aligned} V_{out2} &= V_{1-G} \\ &= 2(1-K) \times V_{in} \end{aligned} \tag{12-60}$$

where,

K varies between 0 and 1 as the control voltage is changed from full attenuation to full gain.

When the control voltage is 0 V, $K = 0.5$ and both output voltages equal the input voltage. The value K is exponentially proportional to the applied control voltage; in the SSM2018T, the gain control constant in the basic VCA configuration is –30 mV/dB, so the decibel gain is directly proportional to the applied control voltage. This makes the part especially applicable to audio applications.

The SSM2018 has many applications as a VCA, but its use as a voltage-controlled panner (VCP) is perhaps one of the most unique, Fig. 12-43.

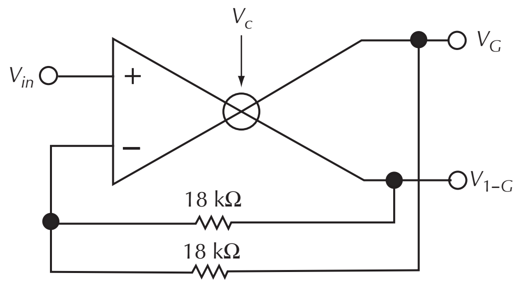


Figure 12-43. SSM2018 as a VCP. Courtesy Analog Devices, Inc.

THAT Corporation's 2180 and 2181 VCAs. The Blackmer VCAs now offered by THAT Corporation (which registered the trademark "Blackmer" for this application) exploit the mathematical property that adding a constant to the logarithm of a number is equivalent, in the linear domain, to multiplying the number by the antilog of the constant.

The equation for determining the output is

$$\begin{aligned} I_{out} &= \text{antilog}[(\log I_{in}) + E_C] \\ &= I_{in} \times [\text{antilog} E_C] \end{aligned} \tag{12-61}$$

I_{in} is multiplied by the antilog of E_C to produce I_{out}. Conveniently, and fortunately for Blackmer, the exponential response of E_C is linear in dB.

Consider the unity-gain case when $E_C = 0$.

$$\begin{aligned} I_{out} &= \text{antilog}[(\log I_{in}) + 0] \\ &= I_{in} \times [\text{antilog} 0] \\ &= I_{in} \times 1 \\ I_{out} &= I_{in} \end{aligned}$$

Blackmer VCAs exploit the logarithmic properties of a bipolar junction transistor, or "BJT." In the basic Blackmer circuit, the input signal I_{in} (the Blackmer VCA works in the current, not the voltage domain) is first converted to its "log-domain" equivalent. A control voltage, E_C, is added to the log of the input signal. Finally, the antilog is taken of the sum to provide an output signal I_{out}. This multiplies I_{in} by a control constant, E_C. When needed, the input signal voltage is converted to a current via an input resistor, and the output signal current is converted back to a voltage via an op-amp and feedback resistor.

Like the Frey OVCE, the Blackmer VCA's control voltage (E_C) is exponentiated in the process. This makes the control law exponential, or linear in dB. Many of the early embodiments of VCAs for electronic music were based on linear multiplication and required exponential converters, either external or internal to the VCA, to obtain this desirable characteristic.[15] Fig. 12-44 shows the relationship between gain and E_C for a Blackmer VCA.

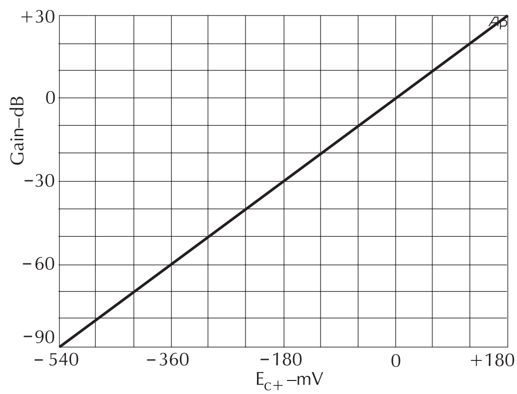


Figure 12-44. THAT 2180 gain versus E_C+. Courtesy THAT Corporation.

Audio signals are of both polarities; that is, the sign of I_{in} in the above equations will be either positive or negative at different times. Mathematically, the log of a negative number is undefined, so the circuit must be designed to handle both polarities. The essence of David Blackmer's invention was to handle each phase—positive and negative—of the signal waveform with different "genders" of transistors—NPN and PNP—and to provide a class A-B bias scheme to deal with the crossover region between the two. This made it possible to generate a sort of "bipolar" log and antilog. A block diagram of a Blackmer VCA is shown in Fig. 12-45.

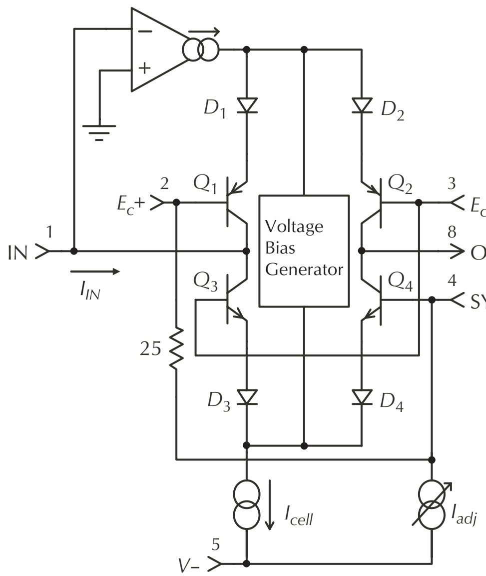


Figure 12-45. THAT 2180 equivalent schematic. Courtesy THAT Corporation.

Briefly, the circuit functions as follows. An ac input signal current *Iin* flows in pin 1, the input pin. An internal operational transconductance amplifier (OTA) maintains pin 1 at virtual ground potential by driving the emitters of Q_1 and (through the Voltage Bias Generator) Q_3. Q_3/D_3 and Q_1/D_1 act to log the input current, producing a voltage (V_3) that represents the "bipolar" logarithm of the input current. (The voltage at the junction of D_1 and D_2 is the same as V_3, but shifted by four forward V_{be} drops.)

Pin 8, the output, is usually connected to a virtual ground. As a result, Q_2/D_2 and Q_4/D_4 take the bipolar antilog of V_3, creating an output current flowing to the virtual ground which is a precise replica of the input current. If pin 2 (E_C+) and pin 3 (E_C–) are held at ground, the output current will equal the input current. For pin 2 positive or pin 3 negative, the output current will be scaled larger than the input current. For pin 2 negative or pin 3 positive, the output current is scaled smaller than the input.

The log portion of the VCA, D_1/Q_1 and D_3/Q_3, and the antilog stages, D_2/Q_2 and D_4/Q_4 in Fig. 12-45, require both the NPN and the PNP transistors to be closely matched to maintain low distortion. As well, all the devices (including the bias network) must be at the same temperature. Integration solves the matching and temperature problems, but conventional "junction-isolated" integration is notorious for offering poor-performing PNP devices. Frey and others avoided this problem by basing their designs exclusively on NPN devices for the critical multiplier stage. Blackmer's design required "good" PNPs as well as NPNs.

One way to obtain precisely matched PNP transistors that provide discrete transistor performance is to use an IC fabrication technology known as "dielectric isolation." THAT Corporation uses dielectric isolation to fabricate integrated PNP transistors that equal or exceed the performance of NPNs. With dielectric isolation, the bottom layers of the devices are available early in the process, so both N and P-type collectors are possible. Furthermore, each transistor is electrically insulated from the substrate and all other devices by an oxide layer, which enables discrete transistor performance with the matching and temperature characteristics only available in monolithic form.

In Fig. 12-45, it can also be seen that the Blackmer VCA has two E_C inputs having opposite control response—E_C+ and E_C–. This unique characteristic allows both control inputs to be used simultaneously. Individually, gain is exponentially proportional to the voltage at pin 2, and exponentially proportional to the negative of the voltage at pin 3. When both are used simultaneously, gain is exponentially proportional to the difference in voltage between pins 2 and 3. Overall, because of the exponential characteristic, the control voltage sets gain *linearly* in *decibels* at 6 mV/dB.

Fig. 12-46 shows a typical VCA application based on a THAT2180 IC. The audio input to the VCA is a current; an input resistor converts the input voltage to a current. The VCA output is also a current. An op-amp and its feedback resistor serve to convert the VCA's current output back to a voltage.

As with the basic topologies from Gilbert, Dow, Curtis, and other transconductance cells, the current input/output Blackmer VCA can be used as a variable conductance to tune oscillators, filters, and the like. An example of a VCA being used to control a first-order state-variable filter is shown in Fig. 12-47 with the response plot in Fig. 12-48.

When combined with audio level detectors, VCAs can be used to form a wide range of dynamics processors, including compressors, limiters, gates, duckers, companding noise reduction systems, and signal-controlled filters.

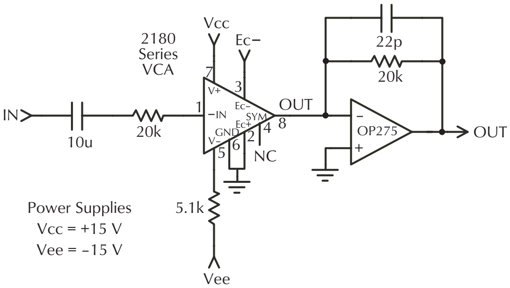


Figure 12-46. Basic THAT 2180 VCA application. Courtesy THAT Corporation.

12.3.4.2 Peak, Average, and RMS Level Detection

It is often desirable to measure audio level for display, dynamics control, noise reduction, instrumentation, etc. Level detectors take different forms: among the most common are those that represent peak level, some form of average level over time, and "root-mean-square" (more simply known as "rms" level).

Peak signal level is usually interpreted to mean the highest instantaneous level within the audio bandwidth. Measuring peak level involves a detector with very fast charge (attack) response and much slower decay. Peak levels are often used for headroom and overload indication and in audio limiters to prevent even brief overload of transmission or storage media. However, peak measurements do not correlate well with perceived loudness, since the ear responds not only to the amplitude, but also to the duration of a sound.

Average-responding level detectors generally average out (or "smooth") full or half-wave rectified signals to provide envelope information. While a pure average response (that of an R-C circuit) has equal rise (attack) and fall (decay) times, in audio applications, level detectors often have faster attack than decay. The familiar VU meter is average responding, with a "response time" and "return time" of the indicator both equal to 300 ms. The "PPM" meter, commonly used in Europe for audio program level measurement, combines a specific quick attack response with an equally specific, slow fall time. PPM metering provides a reliable indication of meaningful peak levels.[16]

Rms level detection is unique in that it provides an ac measurement suitable for the calculation of signal power. Rms measurements of voltage, current, or both indicate "effective" power. Effective power is the heating power of a dc signal equivalent to that offered by an ac signal. True rms measurements are not affected by the signal waveform complexity, while peak and average readings vary greatly depending on the nature of the waveform. For example, a resistor heated by a 12 Vac rms signal produces the same number of watts—and heat—as a resistor connected to 12 Vdc. This is true regardless of whether the ac waveform is a pure sinusoid, a square wave, a triangle wave or music. In instrumentation, rms is often referred to as "true rms" to distinguish it from average-responding instruments which are calibrated to read "rms" only for sinusoidal inputs. Importantly, in audio signal-processing applications, rms response is thought to closely approximate the human perception of loudness.[17]

12.3.4.3 Peak and Average Detection with Integrated Circuits

The fast response of a peak detector is often desirable for overload indication or dynamics control when a

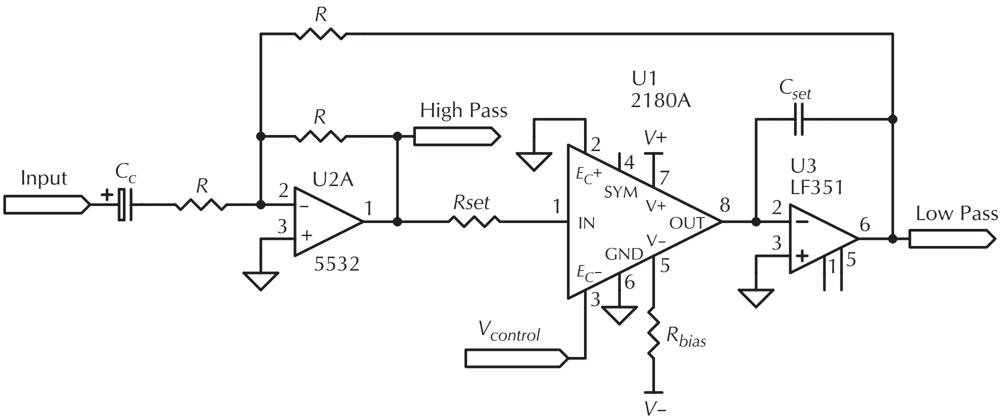


Figure 12-47. VCA state-variable filter. Courtesy THAT Corporation.

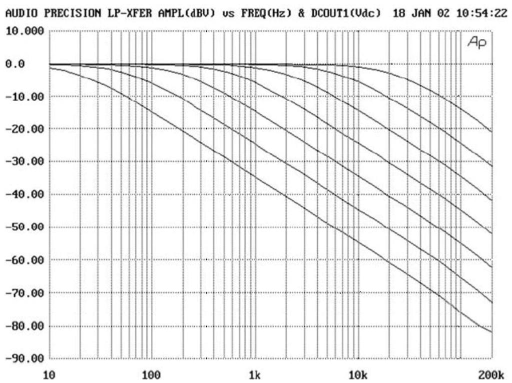


Figure 12-48. State-variable filter response. Courtesy THAT Corporation.

signal needs to be limited to fit the strict level confines of a transmission or storage medium. A number of op-amp-based circuits detect peak levels using full or half-wave rectification. General-purpose op-amps are quite useful for constructing peak detectors and are discussed previously in this chapter. The recently discontinued Analog Devices PKD01 was perhaps the only peak detector IC suited for audio applications.

Average-responding level detection is performed by rectification followed by a smoothing resistor/capacitor (R-C) filter whose time constants are chosen for the application. If the input is averaged over a sufficiently long period, the signal envelope is detected. Again, general-purpose op-amps serve quite well as rectifiers with R-C networks or integrators serving as averaging filters.

Other than meters, most simple electronic audio level detectors use an asymmetrical averaging response that attacks more quickly than it decays. Such circuits usually use diode steering to charge a capacitor quickly through a relatively small-value resistor, but discharge it through a larger resistor. The small resistor yields a fast attack, and the large resistor yields a slower decay.

12.3.4.4 Rms Level Detection Basics

Rms detection has many applications in acoustic and industrial instrumentation. As mentioned previously, rms-level detectors are thought to respond similarly to the human perception of loudness. This makes them particularly useful for audio dynamics control.

Rms is mathematically defined as the square root of the mean of the square of a waveform. Electronically, the mean is equal to the average, which can be approximated by an R-C network or an op-amp-based integrator. However, calculating the square and square root of waveforms is more difficult.

Designers have come up with a number of clever techniques to avoid the complexity of numerical rms calculation. For example, the heat generated by a resistive element may be used to measure power. Power is directly proportional to the square of the voltage across, or current through, a resistor, so the heat given off is proportional to the square of the applied signal level. To measure large amounts of power having very complex waveforms, such as the RF output of a television transmitter, a resistor "dummy load" is used to heat water. The temperature rise is proportional to the transmitter power. Such "caloric" instruments are naturally slow to respond, and impractical for the measurement of sound. Nonetheless, solid-state versions of this concept have been integrated, as for example US patent 4,346,291, invented by Roy Chapel and Macit Gurol.[18] This patent, assigned to instrumentation manufacturer Fluke, describes the use of a differential amplifier to match the power dissipated in a resistive element, thus measuring the true rms component of current or voltage applied to the element. While very useful in instrumentation, this technique has not made it into audio products due to the relatively slow time constants of the heating element.

To provide faster time constants to measure small rms voltages or currents with complex waveforms such as sound, various analog computational methods have been employed. Computing the square of a signal generally requires extreme dynamic range, which limits the usefulness of direct analog methods in computing rms value. As well, the square and square-root operations require complex analog multipliers, which have traditionally been expensive to fabricate.

As with VCAs, the analog computation required for rms level detection is simplified by taking advantage of the logarithmic properties of bipolar junction transistors. The seminal work on computing rms values for audio applications was developed by David E. Blackmer, who received US Patent 3,681,618 for an "RMS Circuit with Bipolar Logarithmic Converter."[17] Blackmer's circuit, discussed later, took advantage of two important log-domain properties to compute the square and square root. In the "log domain," a number is squared by multiplying it by 2; the square root is obtained by dividing it by 2.

For example, to square the signal V_{in} use

$$V_{in}^{\ 2} = \text{antilog}[(\log V_{in}) \times 2]. \qquad (12\text{-}62)$$

To take the square root of Vlog,

$$\sqrt{V\log} = \text{antilog}\left[\frac{\log(V\log)}{2}\right]. \qquad (12\text{-}63)$$

12.3.4.5 Rms Level Detection ICs

Because rms level detectors are more complex than either peak- or average-responding circuits, they benefit greatly from integration. Fortunately, a few ICs are suitable for the professional audio applications. Two ICs currently in production are the Analog Devices AD636 and the THAT Corporation THAT2252.

Analog Devices AD636. The AD636 has enjoyed wide application in audio and instrumentation. Its predecessor, the AD536, was used in the channel dynamics processor of the SSL 4000 series console in conjunction with a dbx VCA. Thousands of these channels are in daily use worldwide.

The AD636 shown in Fig. 12-49 provides both a linear-domain rms output and a dB-scaled logarithmic output. The linear output at pin 8 is ideal for applications where the rms input voltage must be read with a dc meter. Suitably scaled, 1 Vrms input can produce 1 Vdc at the buffer output, pin 6.

In audio applications such as signal processors, it is often most useful to express the signal level in dB. The AD636 also provides a dB-scaled current output at pin 5. The linear dB output is particularly useful for use with exponentially controlled VCAs such as the SSM2018 or THAT2180-series.

Averaging required to calculate the mean of the sum of the squares is performed by a capacitor, C_{AV} connected to pin 4. Fig. 12-50 shows an AD636 used as an audio dB meter for measurement applications.

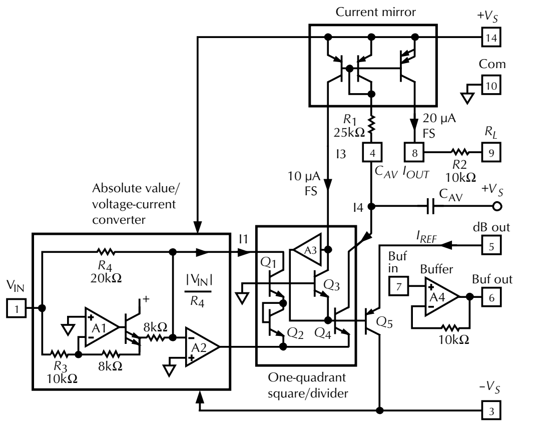


Figure 12-49. The AD636 block diagram. Courtesy Analog Devices, Inc.

THAT Corporation THAT2252. The 2252 IC uses the technique taught by David Blackmer to provide wide dynamic range, logarithmic "linear dB" output, and relatively fast time constants. Blackmer's detector delivers a fast attack with a slow "linear dB" decay characteristic in the "log domain."[17] Because it was specifically developed for audio applications, it has become a standard for use in companding noise reduction systems and VCA-based compressor/limiters.

A simplified schematic of Blackmer's rms detector, used in the THAT2252, is shown in Fig. 12-51.

The audio input is first converted to a current I_{in} by an external resistor (not shown in Fig 12-51). I_{in} is full-wave rectified by a current mirror rectifier formed by OA1 and Q_1-Q_3, such that IC_4 is a full-wave rectified

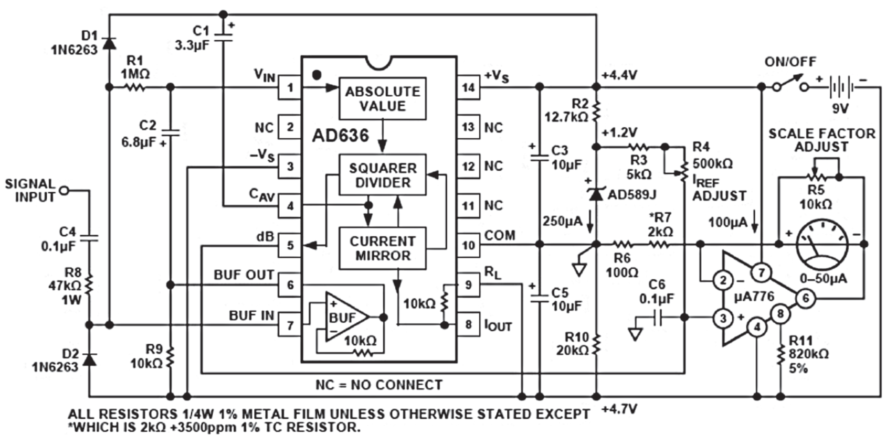


Figure 12-50. AD636 as an audio dB meter. Courtesy Analog Devices, Inc.

version of I_{in}. Positive input currents are forced to flow through Q_1, and mirrored to Q_2 as IC_2; negative input currents flow through Q_3 as IC_3; both IC_2 and IC_3 thus flow through Q_4. (Note that pin 4 is normally connected to ground through an external 20 Ω resistor.)

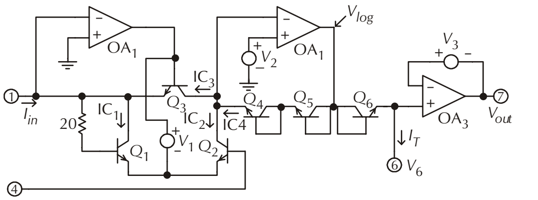


Figure 12-51. Block diagram of a THAT2252 IC. Courtesy THAT Corporation.

Performing the absolute value before logarithmic conversion avoids the problem that, mathematically, the log of a negative number is undefined. This eliminates the requirement for "bipolar" logarithmic conversion and the PNP transistors required for log-domain VCAs.

OA2, together with Q_4 and Q_5, forms a log amplifier. Due to the two diode-connected transistors in the feedback loop of OA2, the voltage at its output is proportional to twice the log of IC_4. This voltage, $V_{\log}$, is therefore proportional to the log of $I_{in}{}^2$ (plus the bias voltage V_2).

To average $V_{\log}$, pin 6 is usually connected to a capacitor C_T and a negative current source R_T, see Fig. 12-52. The current source establishes a quiescent dc bias current, I_T, through diode-connected Q_6. Over time, C_T charges to 1 V_{be} below $V_{\log}$.

Q_6's emitter current is proportional to the antilog of its V_{be}. The potential at the base (and collector) of Q_6 represents the log of $I_{in}{}^2$ while the emitter of Q_6 is held at ac ground via the capacitor. Thus, the current in Q_6 is proportional to the square of the instantaneous change in input current. This "dynamic" antilogging causes the capacitor voltage to represent the log of the mean of the square of the input current. Another way to characterize the operation of Q_6, C_T, and R_T is that of a "log domain" filter.[20]

In the THAT2252, the square root portion of the rms calculation is not computed explicitly but is implied by the constant of proportionality for the output. Since, in the log domain, taking the square root is equivalent to dividing by two, the voltage at the output (pin 7) is proportional to the mean of the square at approximately 3 mV/dB and proportional to the square root of the mean of the square at approximately 6 mV/dB.

The attack and release times of rms detectors are locked in a relationship to each other and separate controls for each are not possible while still maintaining rms response. Varying the value of C_T and R_T in the THAT2252, and C_{AV} in the AD636 allow the time constant to be varied to suit the application. More complex approaches, such as a "non-linear" capacitor, are possible with additional circuitry.[21]

Fig. 12-52 shows a typical application for the THAT2252. The input voltage is converted to a current by R_{in}. C_{in} blocks input dc and internal op-amp bias currents. The network around pin 4 sets the waveform symmetry for positive vs. negative input currents. Internal bias for the THAT2252 is set by R_b and bypassed by a 1 µF capacitor. R_T and C_T set the timing of the log-domain filter. The output signal (pin 7) is 0 V when the input signal current equals a reference current determined by I_{bias} and I_T. It varies in dc level above and below this value to represent the dB input level at the rate of ~6 mV/dB.

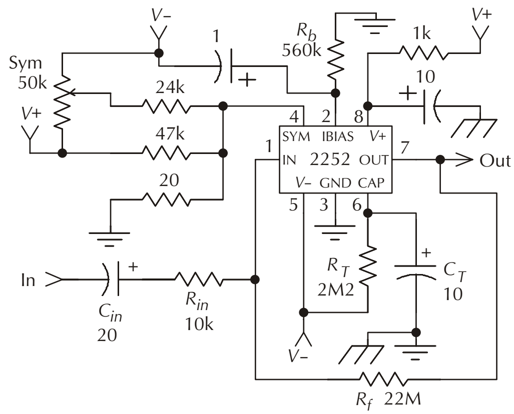


Figure 12-52. Typical application of a THAT2252 IC. Courtesy THAT Corporation.

Fig. 12-53 shows the tone burst response of a THAT2252, while Fig.12-54 is a plot of THAT2252 output level versus input level. The THAT2252 has linear dB response over an almost 100 dB range.

The Analog Devices AD636 and THAT Corporation THAT2252 provide precise, low-cost rms detection due to their integration into monolithic form. On their own, rms detectors are very useful at monitoring signal level, controlling instrumentation, and other applications. When combined with VCAs for gain control, many different signal processing functions can be realized including noise reduction, compression and limiting.

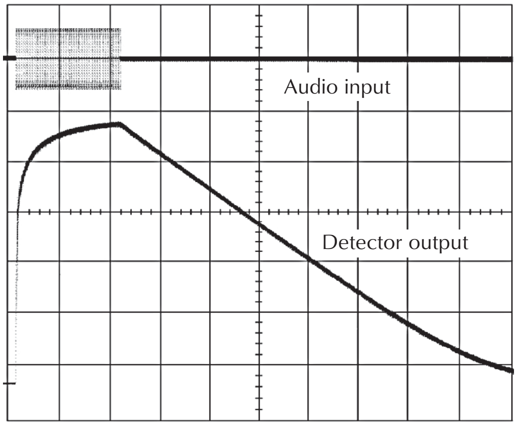


Figure 12-53. THAT2252 tone burst response. Courtesy THAT Corporation

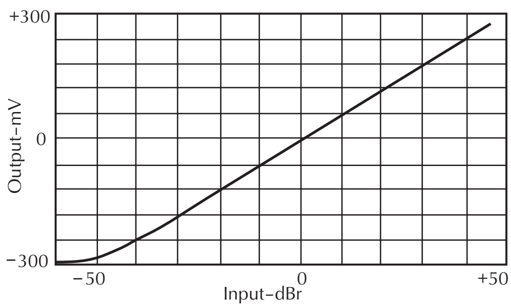


Figure 12-54. THAT2252 input versus output. Courtesy THAT Corporation.

12.3.5 Integrated Circuit Preamplifiers

The primary applications of preamplifiers for professional audio in the post-tape era are for use with microphones. Before the development of monolithic ICs dedicated to the purpose, vacuum tubes, discrete bipolar or field-effect transistors,[22] or general-purpose audio op-amps were used as preamplifiers.[23] Dynamic microphones generally produce very small signal levels and have low output impedance. Ribbon microphones are notorious for low output levels. For many audio applications, significant gain (40–60 dB) is required to bring these mic-level signals up to pro audio levels. Condenser microphones, powered by phantom power, external power supplies, or batteries, often produce higher signal levels requiring less gain.

To avoid adding significant noise to the microphone's output, professional audio preamplifiers must have very low input noise. Transformer-coupled preamps ease the requirement for very low-noise amplification, since they take advantage of the voltage step-up possible within the input transformer. Early transformerless, or "active," designs required performance that eluded integration until the early 1980s. Until semiconductor process and design improvements permitted it and the market developed to generate sufficient demand, most microphone preamplifiers were based on discrete transistors, or discrete transistors augmented with commercially available op-amps.

Virtually all professional microphones use two signal lines to produce a balanced output. This allows a preamplifier to distinguish the desired "differential" audio signal—which appears as a voltage *difference* between the two signal lines—from hum and noise pickup—which appears as a "common-mode" signal with the *same* amplitude and polarity on both signal lines. "common-mode rejection" quantifies the ability of the preamplifier to reject common-mode interference while accepting differential signals.

Therefore, one goal of a pro-audio mic preamp is to amplify differential signals in the presence of common-mode hum. As well, the preamp should ideally add no more noise than the thermal noise of the source impedance—well below the self-noise of the microphone and ambient acoustic noise.

"Phantom" power is required for many microphones, especially professional condenser types. This is usually a +48 Vdc power supply applied to both polarities of the differential input through 6.8 kΩ resistors (one for each input polarity). Dc supply current from the microphone returns through the ground conductor. Phantom power appears in common-mode essentially equal on both inputs. The voltage is used to provide power to the circuitry inside the microphone.

12.3.5.1 Transformer Input Microphone Preamplifiers

Many microphone preamplifiers use transformers at their inputs. Transformers, although costly, provide voltage gain that can ease the requirements for low noise in the subsequent amplifier. The transformer's voltage gain is determined by the turns ratio of the secondary vs. the primary. This ratio also transforms impedance, making it possible to "match" a low-impedance microphone to a high-impedance amplifier without compromising noise performance.

A transformer's voltage gain is related to its impedance ratio by the following equation:

$$Gain = 20\log\sqrt{\frac{Z_s}{Z_p}} \tag{12-64}$$

where,
Gain is the voltage gain in dB of the transformer,
Z_p is the primary transformer impedance in ohms,
Z_s is the secondary transformer impedance in ohms.

A properly designed transformer with a 150 Ω primary and 15 kΩ secondary produces 20 dB of "free" voltage gain without adding noise.

Well-made transformers also provide high common-mode rejection, which helps avoid hum and noise pickup. This is especially important with the low output voltages and long cable runs common with professional microphones. As well, transformers provide galvanic isolation by electrically insulating the primary circuit from the secondary while allowing signal to pass. While usually unnecessary in microphone applications, this provides a true "ground lift," which can eliminate ground loops in certain difficult circumstances.

Transformer isolation is also useful when feeding "phantom power" (a +48 Vdc current-limited voltage to power internal circuitry in the microphone) down the mic cable from the preamp input terminals. Phantom power may be connected through a center tap on the primary to energize the entire primary to +48 Vdc, or supplied through resistors (usually 6.8 kΩ) to each end of the primary of the transformer. (The latter connection avoids dc currents in the coils, which can lead to premature saturation of the core magnetics.) The galvanic isolation of the transformer avoids any possibility of the 48 Vdc signal from reaching the secondary windings.

12.3.5.2 Active Microphone Preamplifiers Eliminate Input Transformers

As is common in electronic design, transformers do have drawbacks. Perhaps the most prominent one is cost: a Jensen Transformer, Inc. JT-115K-E costs approximately $75 US or $3.75 per dB of gain.[24] From the point of view of signal, transformers add distortion due to core saturation. Transformer distortion has a unique sonic signature that is considered an asset or a liability—depending on the transformer and whom you ask. Transformers also limit frequency response at both ends of the audio spectrum. Furthermore, they are susceptible to picking up hum from stray electromagnetic fields.

Well designed active "transformerless" preamplifiers can avoid these problems, lowering cost, reducing distortion and increasing bandwidth. However, transformerless designs require far better noise performance from the active circuitry than transformer-based preamps do. Active mic preamps usually require capacitors (and other protection devices) to block potentially damaging effects of phantom power.[25]

12.3.5.3 The Evolution of Active Microphone Preamplifier ICs

Active balanced-input microphone preamplifier ICs were not developed until the early 1980s. Early IC fabrication processes did not permit high-quality low-noise devices, and semiconductor makers were uncertain of the demand for such products.

"Active transformerless" microphone preamplifiers must have fully differential inputs because they interface to balanced microphones. The amplifiers described here, both discrete and IC, use a current feedback "CFB" topology with feedback returned to one (or both) of the differential input transistor pair's emitters. Among its many attributes, current feedback permits differential gain to be set by a single resistor.

Current feedback amplifiers have a history rooted in instrumentation amplifiers. The challenges of amplifying low-level instrumentation signals are very similar to microphones. The current feedback instrumentation amplifier topology, known at least since Demrow's 1968 paper,[26] was integrated as early as 1982 as the Analog Devices AD524 developed by Scott Wurcer.[27] A simplified diagram of the AD524 is shown in Fig. 12-55. Although the AD524 was not designed as an audio preamp, the topology it used later became a de facto standard for IC microphone preamps. Demrow and Wurcer both used a bias scheme and fully balanced topology in which they wrapped op-amps around each of the two input transistors to provide both ac and dc feedback. Gain is set by a single resistor connected between the emitters (shown as 40 Ω, 404 Ω, and 4.44 kΩ) and feedback is provided by two resistors (R_{56} and R_{57}). The input stage is fully symmetrical and followed by a precision differential amplifier to convert the balanced output to single-ended. Wurcer's AD524 required laser-trimmed thin film resistors with matching to 0.01% for an 80 dB common-mode rejection ratio at unity gain.

Audio manufacturers, using variations on current feedback and the Demrow/Wurcer instrumentation amp, produced microphone preamps based on discrete low-noise transistor front-ends as early as 1978; an example is the Harrison PC1041 module.[28] In December of 1984, Graeme Cohen also published his discrete transistor topology; it was remarkably similar to the work of Demrow, Wurcer and the Harrison preamps.[29]

Solid State Music, or "SSM," which later became Solid State Microtechnology, developed the first active

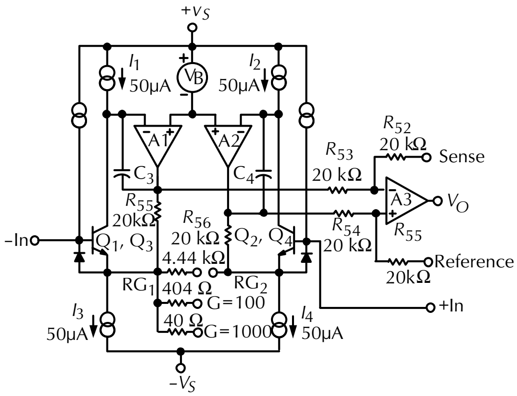


Figure 12-55. AD524 block diagram. Courtesy Analog Devices, Inc.

microphone preamp IC for professional audio around 1982.[30] SSM specialized in producing niche-market semiconductors aimed at the professional audio business. The SSM2011 was almost completely self-contained, requiring only a handful of external resistors and capacitors to provide a complete preamp system. One unique feature of the SSM2011 was an on-chip LED overload and signal presence indicator.

SSM later produced the SSM2015 and the SSM2016 designed by Derek Bowers.[31] The SSM2016, and the SSM2011 and 2015 which preceded it, did not use a fully balanced topology like Wurcer's AD524 and the Harrison PC1041. The SSM parts used an internal op-amp to convert the differential stage output to single-ended. This allowed external feedback resistors to be used, eliminating the performance penalty of on-chip diffused resistors. The SSM2016 was highly successful but required external precision resistors and up to three external trims. SSM was later acquired by Precision Monolithics and eventually by Analog Devices ("ADI"). The SSM2016 was extremely successful and, after its discontinuance in the mid-1990s, became highly sought after.

Analog Devices introduced the SSM2017 "self contained" preamp, also designed by Bowers, as a replacement for the SSM2016. The SSM2017 used internal laser-trimmed thin-film resistors that permitted the fully balanced topology of the AD524 and discrete preamps to be realized as an IC. Analog Devices manufactured the SSM2017 until about 2000 when it was discontinued. A year or two later, ADI released the 2019 which is available today.

The Burr Brown division of Texas Instruments offered the INA163, which had similar performance to the SSM2017, but was not pin compatible with it. After the 2017 was discontinued, TI introduced its INA217 in the SSM2017 pinout. Today, TI produces a number of INA-family instrumentation amplifiers suitable for microphone preamps including the INA103, INA163, INA166, INA217, and the first digitally gain-controlled preamp: the PGA2500.

In 2005, THAT Corporation introduced a series of microphone preamplifiers in pinouts to match the familiar SSM2019/INA217 as well as the INA163. The THAT1510 and the performance-enhanced THAT1512 use dielectric isolation to provide higher bandwidth than the junction-isolated INA and SSM-series products. (Dielectric isolation is explained in the section on audio VCAs.)

While all offer relatively high performance, the three different families of parts have different strengths and weaknesses. Differences exist in gain-bandwidth, noise floor, distortion, gain structure, and supply consumption. The optimum part for any given application will depend on the exact requirements of the designer. A designer considering any one of these parts should compare their specs carefully before finalizing a new design.

12.3.5.4 Integrated Circuit Mic Preamplifier Application Circuits

The THAT1510-series block diagram is shown in Fig. 12-57. Its topology is similar to those of the TI and ADI parts. A typical application circuit is shown in Fig. 3#. The balanced mic-level signal is applied to the input pins, In+ and In−. A single resistor (R_G), connected between pins R_{G1} and R_{G2}, sets the gain in conjunction with the internal resistors R_A and R_B. The input stage consists of two independent low-noise amplifiers in a balanced differential amplifier configuration with both ac and dc feedback returned to the emitters of the differential pair. This topology is essentially identical to the AD524 current feedback amplifier as described by Wurcer *et al.*

The output stage is a single op-amp differential amplifier that converts the balanced output of the gain stage into single-ended form. The THAT1500 series offers a choice of gains in this stage: 0 dB for the 1510, and −6 dB for the 1512. Gain is controlled by the input-side resistor values: 5 kΩ for the 1510 and 10 kΩ for the 1512.

The gain equations for the THAT1510 are identical to that of the SSM2017/2019, and the INA217. The INA163 and THAT1512 have unique gain equations.

For the THAT1510, SSM 2019, & INA217 the equation is

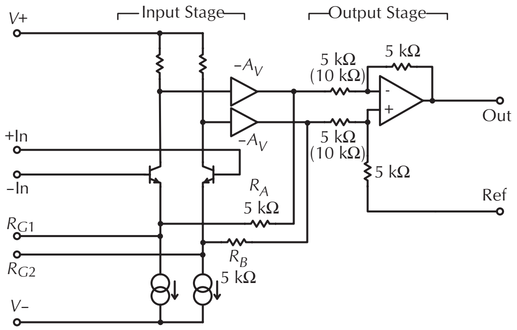


Figure 12-56. THAT1510/1512 block diagram. Courtesy THAT Corporation.

$$Av = 1 + \frac{10\ \text{k}\Omega}{R_G}, \qquad (12\text{-}65)$$

for the INA163 it is

$$Av = 1 + \frac{6\ \text{k}\Omega}{R_G}, \qquad (12\text{-}66)$$

and for the THAT1512 it is

$$Av = 0.5 + \frac{5\ \text{k}\Omega}{R_G}, \qquad (12\text{-}67)$$

where,
Av is the voltage gain of the circuit.

All these parts can reach unity gain but the value of R_G required varies considerably. For the 1510, 2017, 2019, 163, and 217, gain is 0 dB ($Av = 1$) when R_G is open: this is the minimum gain of all these ICs. For the 1512, gain is –6 dB (Av = 0.5) with R_G open. To go from 60 dB to 0 dB gain, R_G must span a large range: 10 Ω to 10 kΩ for the 1510 and its equivalents.

R_G is typically a reverse log potentiometer (or set of switched resistors) to provide smooth rotational control of gain. In many applications, and, as shown in Fig. 12-57, a large value capacitor is placed in series with R_G to limit the dc gain of the device, thus preventing shifts in output DC offset with gain changes. For 60 dB of gain with the THAT1512, R_G = 5 Ω (6 Ω in the case of the INA163). Because of this, C_G must be quite large, typically ranging from 1000 uF to 6800 uF to preserve low frequency response. Fortunately, C_G does not have to support large voltages: 6.3 V is acceptable.

Parts from all manufacturers exhibit excellent voltage noise performance of ~1 nV/$\sqrt{\text{Hz}}$ at high gains. Differences in noise performance begin to show up at lower gains, with the THAT 1512 offering the best performance ~34 nV/$\sqrt{\text{Hz}}$ at 0 dB gain) of the group.

These parts are all generally optimized for the relatively low source impedances of dynamic microphones with typically a few hundred ohm output impedance.

Fig. 12-57 provides an application example for direct connection to a dynamic microphone. Capacitors C_1–C_3 filter out radio frequencies that might cause interference (forming an "RFI" filter). R_1 and R_2 provide a bias current path for the inputs and terminate the microphone output. R_G sets the gain as defined in the previous equation. C_G blocks dc in the input stage feedback loop, limiting the dc gain of this stage to unity and avoiding output offset change with gain. C_6 and C_9 provide power supply bypass.

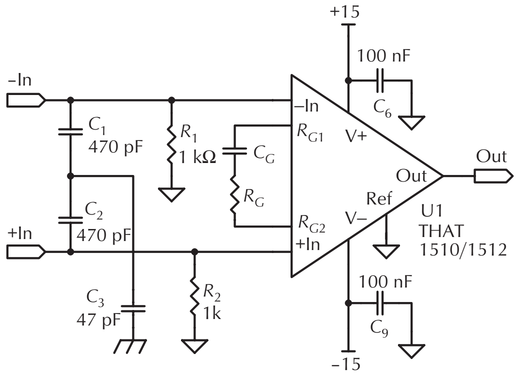


Figure 12-57. THAT1510/1512 Basic Application. Courtesy THAT Corporation.

Fig. 12-58 shows the THAT1512 used as a preamp capable of being used with phantom power. C_1–C_3 provide RFI protection. R_5 and R_6 feed phantom power to the microphone. R_9 terminates the microphone. C_4 and C_5 block 48 Vdc phantom potential from the THAT1512. R_3, R_4 and D_1–D_4 provide current limiting and overvoltage protection from phantom power faults. R_1 and R_2 are made larger than previously shown to reduce the loading on C_4 and C_5.

Many variations are possible on these basic circuits, including digital control of gain, dc servos to reduce or eliminate some of the ac-coupling needed, and exotic power supply arrangements that can produce response down to dc. For more information on possible configurations, see application notes published by Analog Devices, Texas Instruments, and THAT Corporation. (All available at their respective web sites: www.analog.com, www.ti.com, www.thatcorp.com.)

Modern integrated-circuit microphone preamplifiers provide a simple "building block" with performance equaling discrete solutions without a costly input transformer.

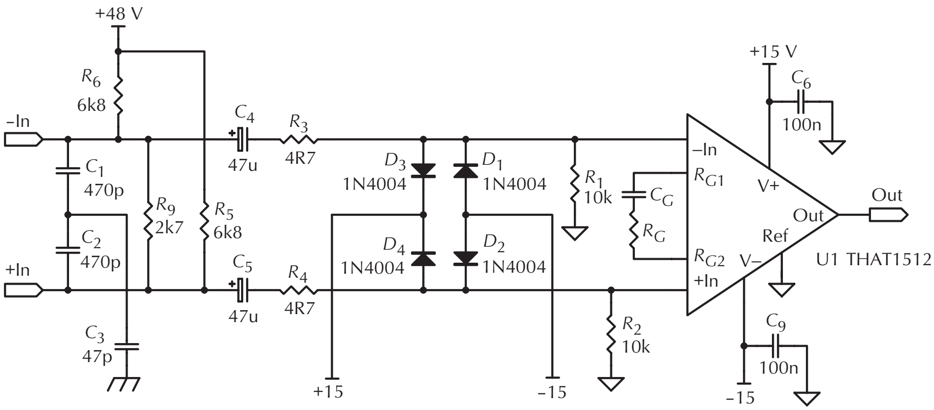


Figure 12-58. THAT preamp circuit with phantom power. Courtesy THAT Corporation.

12.3.6 Balanced Line Interfaces

In professional audio, interconnections between devices frequently use "balanced lines." This is especially important when analog audio signals are sent over long distances, where the ground references for the send and receive ends are different, or where noise and interference may be picked up in the interconnection cables.

Differences in signal ground potentials arise as a result of current flowing into power-line safety grounds. These currents, flowing through finite ground impedances between equipment, can produce up to several volts potential difference between the ground references within a single building. These currents, usually at the power line frequency and its harmonics, produce the all-too-familiar hum and buzz known to every sound engineer.

Two other forms of interference, electrostatic and magnetic, also create difficulty. Cable shielding reduces electrostatic interference from fields, typically using braided copper, foil wrap, or both. Magnetic interference from fields is much harder to prevent via shielding. The impact of magnetic fields in signal cables is reduced by balanced cable construction using twisted pair cable. Balanced circuits benefit from the pair's twist by ensuring that magnetic fields cut each conductor equally. This in turn ensures that the currents produced by these fields appear in "common-mode," wherein the voltages produced appear equally in both inputs.

The balanced line approach comes out of telephony, in which voice communications are transmitted over many miles of unshielded twisted pair cables with reasonable fidelity and freedom from hum and interference pickup. Two principles allow balanced lines to work. First, interference—whether magnetic or electrostatic—is induced equally in both wires in the twisted paired-conductor cable, and second, the circuits formed by the source and receiver, plus the two wires connecting them form a "balanced bridge,"[32] Fig. 12-59. Interfering signals appear identically (in *common-mode*) at the two (+ and –) inputs, while the desired audio signal appears as a difference (the *differential* signal) between the two inputs.

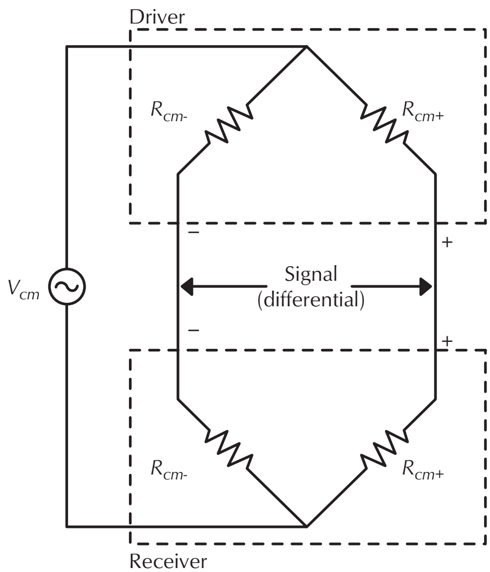


Figure 12-59. Balanced Bridge. Courtesy THAT Corporation.

A common misconception in the design of balanced interfaces is that the audio signals must be transmitted as equal and opposite polarity on both lines. While this is desirable to maximize headroom in many situations, it is unnecessary to preserve fidelity and avoid noise pickup. It is enough if the bridge formed by the combination of the circuit's two common-mode source *impedances* (not the *signals*) working against the two common-mode load *impedances* remains balanced in all circumstances.

In telephony, and in early professional audio systems, transformers were used at both the inputs and outputs of audio gear to maintain bridge balance. Well made output transformers have closely matched common-mode source impedances and very high common-mode impedance. (Common-mode impedance is the equivalent impedance from one or both conductors to ground.) The floating connections of most transformers—whether used for inputs or outputs—naturally offer very large common-mode impedance. Both of these factors, matched source impedances for output transformers, and high common-mode impedance (to ground) for both input and output transformers, work together to maintain the balance of the source/load impedance bridge across a wide range of circumstances. As well, transformers offer "galvanic isolation" which is sometimes helpful when faced with particularly difficult grounding situations.

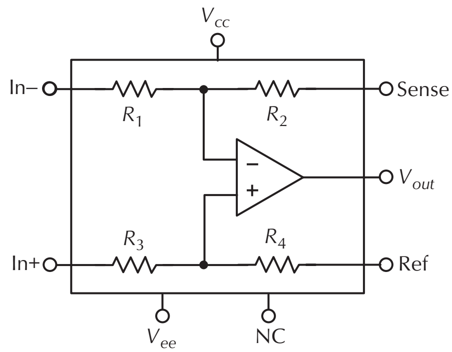
On the other hand, as noted previously in the section on preamplifiers, transformers have drawbacks of high cost, limited bandwidth, distortion at high signal levels, and magnetic pickup.

12.3.6.1 Balanced Line Inputs

Transformers were used in early balanced line input stages, particularly in the days before inexpensive op-amps made it attractive to replace them. The advent of inexpensive op-amps, especially compared to the cost of transformers, motivated the development of active "transformerless" inputs. As the state of the art in op-amps improved, transformer-coupled inputs were replaced by less expensive, high-performance active stages based on general-purpose parts like the Texas Instruments TL070- and TL080-series, the National Semiconductor LF351-series, and the Signetics NE5534.

As with microphone preamplifiers, common-mode rejection is an important specification for line receiver inputs. The most common configuration for active balanced line input stages used in professional audio are the simple circuit shown in Fig. 12-60. To maintain high common-mode rejection ("CMR"), the four resistors used must match very closely. To maintain a 90 dB CMR, for example, the resistor ratio R_1/R_2 must match that of R_3/R_4 within 0.005%. The requirement for precision-matched resistors to provide high CMR drove the development of specialized line receiver ICs.

To maintain the high CMR potential of precision balanced line receivers, the interconnections between stages must be made through low-resistance connections, and the impedances in both lines of the circuit must be very nearly identical. A few ohms of contact resistance external to the line driver and receiver (due, for example, to oxidation or poor contact) or any imbalance in the driving circuit, can significantly reduce CMR by unbalancing the bridge circuit. The imbalance can be at the source, in the middle at a cable junction, or near the input of the receiving equipment. Although many balanced line receivers provide excellent CMR under ideal conditions, few provide the performance of a transformer under less-than-ideal real world circumstances.



Part no.	Gain	R_1, R_3	R_2, R_4
THAT1240	0 dB	9 kΩ	9 kΩ
THAT1243	−3 dB	10.5 kΩ	7.5 kΩ
THAT1246	−6 dB	12 kΩ	6 kΩ

Figure 12-60. 1240 basic circuit. Courtesy THAT Corporation.

12.3.6.2 Balanced Line Outputs

Transformers were also used in early balanced output stages, for the same reasons as they are used in inputs. However, to drive 600 Ω loads, an output transformer must have more current capacity than an input transformer that supports the same voltage levels. This increased the cost of output transformers, requiring more copper and steel than input-side transformers, and putting pressure on designers to find alternative outputs. Early active stages were either discrete or used discrete output transistors to boost the current available from op-amps. The NE5534, with its capability to directly drive a 600 Ω load, made it possible to use op-amps without additional buffering as output stages.

One desirable property of transformer-coupled output stages was that the output voltage was the same regardless of whether the output was connected differentially or in single-ended fashion. While “professional” audio gear has traditionally used balanced input stages, sound engineers commonly must interface to “consumer” and “semi-pro” gear that use single-ended input connections referenced to ground. Transformers behave just as well when one terminal of their output winding is shorted to the ground of a subsequent single-ended input stage. On the other hand, an active-balanced output stage which provides “equal and opposite” drive to the positive and negative outputs will likely have trouble if one output is shorted to ground.

This led to the development of a “cross-coupled” topology by Thomas Hay of MCI that allowed an active balanced output stage to mimic this property of transformers.[33] When loaded equally by reasonable impedances (e.g., 600 Ω or more) Hay’s circuit delivers substantially equal—and opposite-polarity voltage signals at either output. However, because feedback is taken differentially, when one leg is shorted to ground, the feedback loop automatically produces twice the voltage at the opposing output terminal. This mimics the behavior of a transformer in the same situation.

While very clever, this circuit has at least two drawbacks. First, its resistors must be matched very precisely. A tolerance of 0.1% (or better) is often needed to ensure stability, minimize sensitivity to output loading, and maintain close matching of the voltages at either output. (Though, as noted earlier, this last requirement is unnecessary for good performance.) The second drawback is that the power supply voltage available to the two amplifiers limits the voltage swing at each output. When loaded differentially, the output stage can provide twice the voltage swing than it can when driving a single-ended load. But this means that headroom is reduced 6 dB with single-ended loads.

One way to ensure the precise matching required by Hay’s circuit is to use laser-trimmed thin-film resistors in an integrated circuit. SSM was the first to do just that when they introduced the SSM2142, a balanced line output driver with a cross-coupled topology.

12.3.6.3 Integrated Circuits for Balanced Line Interfaces

Instrumentation amplifier inputs have similar requirements to those of an audio line receiver. The INA105, originally produced by Burr Brown and now Texas Instruments, was an early instrumentation amplifier that featured laser-trimmed resistors to provide 86 dB common-mode rejection. Although its application in professional audio was limited due to the performance of its internal op-amps, the INA105 served as the basis for the modern audio balanced line receiver.

In 1989, the SSM Audio Products Division of Precision Monolithics introduced the SSM2141 balanced line receiver and companion SSM2142 line driver. The SSM2141 was offered in the same pinout as the INA105 but provided low noise and a slew rate of almost 10 V/µs. With a typical CMR of 90 dB, the pro-audio industry finally had a low-cost, high-performance replacement for the line input transformer. The SSM2142 line driver, with its cross-coupled outputs, became a low-cost replacement for the output transformer. Both parts have been quite successful.

Today, Analog Devices (who acquired Precision Monolithics) makes the SSM2141 line receiver and the SSM2142 line driver. The SSM2143 line receiver, designed for 6 dB attenuation, was introduced later to offer increased input headroom. It also provides overall unity gain operation when used with an SSM2142 line driver, which has 6 dB of gain.

The Burr Brown division of Texas Instruments now produces a similar family of balanced line drivers and receivers, including dual units. The INA134 audio differential line receiver is a second-source to the SSM2141. The INA137 is similar to the SSM2143 and also permits gains of ±6 dB. Both devices owe their pinouts to the original INA105. Dual versions of both parts are available as the INA2134 and 2137. TI also makes cross-coupled line drivers known as the DRV134 and DRV135.

THAT Corporation also makes balanced line drivers and receivers. THAT’s 1240-series single and 1280-series dual balanced line receivers use laser-trimmed resistors to provide high common rejection in the familiar SSM2141 (single) and INA2134 (dual) pinouts. For lower cost applications, THAT offers the 1250- and 1290-series single and dual line receivers. These parts eliminate laser trimming, which sacrifices CMR to reduce cost. Notably, THAT offers both dual and single line receivers in the unique configuration of ±3 dB gain, which can optimize dynamic range for many common applications.

THAT Corporation also offers an unique line receiver, the THAT1200-series, based on technology licensed from William E. Whitlock of Jensen transformers (US patent 5,568,561).[34] This design, dubbed InGenius (a trademark of THAT Corporation), bootstraps the common-mode input impedance to raise it into the megohm range of transformers. This overcomes the loss of common-mode rejection when the impedances feeding the line receiver are slightly unbalanced and permits transformer-like operation. The InGenius circuit will be discussed in a following section.

THAT also offers the THAT1646 balanced line driver, which has identical pinout to the SSM2142 and DRV134/135. THAT's 1606 balanced line driver is unique among these parts in that it provides not only a differential output, but also a differential input—enabling a more direct connection to digital to analog converters.

The THAT1646 and 1606 use a unique output topology unlike conventional cross-coupled outputs which THAT calls "OutSmarts," (another trademark). OutSmarts is based on US patent 4,979,218 issued to Chris Strahm, then of Audio Teknology Incorporated.[35] Conventional cross-coupled outputs lose common-mode feedback when one output is shorted to ground to accommodate a single-ended load. This allows large signal currents to flow into ground increasing crosstalk and distortion. Strahm's circuit avoids this by using an additional feedback loop to provide current feedback. Application circuits for the THAT1646 will be described in the section "Balanced Line Outputs."

12.3.6.4 Balanced Line Input Application Circuits

Conventional balanced line receivers from Analog Devices, Texas Instruments and THAT Corporation are substantially equivalent to the THAT1240 circuit shown in Fig. 12-61. Some variations exist in the values of R_1–R_4 from one manufacturer to the other that will influence input impedance and noise. The ratio of R_1/R_3 to R_2/R_4 establishes the gain with $R_1 = R_3$ and $R_2 = R_4$. V_{out} is normally connected to the sense input resistor with the Reference pin grounded.

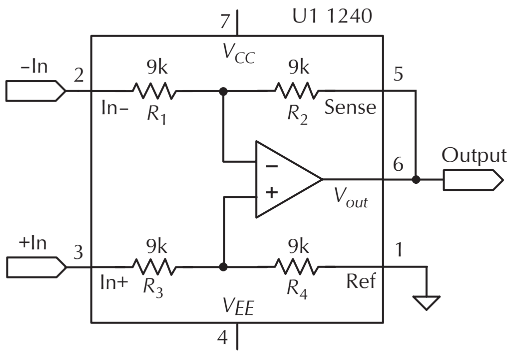


Figure 12-61. THAT 1240 with 0 dB gain. Courtesy THAT Corporation.

Line receivers usually operate at either unity-gain (SSM2141, INA134, THAT1240, or THAT1250) or in attenuation (SSM2143, INA137, THAT1243, or THAT1246, etc.). When a perfectly balanced signal (with each input line swinging ½ the differential voltage) is converted from differential to single-ended by a unity gain receiver, the output must swing twice the voltage of either input line for a net voltage "gain" of +6 dB. With only +21 dBu output voltage available from a line receiver powered by bipolar 15 V supplies, additional attenuation is often needed to provide headroom to accommodate pro audio signal levels of +24 dBu or more. The ratios R_1/R_2 and R_3/R_4 are 2:1 in the SSM2143, INA137, and THAT1246 to provide 6 dB attenuation. These parts accommodate up to +27 dBu inputs without clipping their outputs when running from bipolar 15 V supplies. The THAT1243, and THAT's other "±3 dB" parts (the 1253, 1283, and 1293) are unique with their 0.707 attenuation. This permits a line receiver that accommodates +24 dBu inputs, but avoids additional attenuation that increases noise. A –3 dB line receiver is shown in Fig. 12-62.

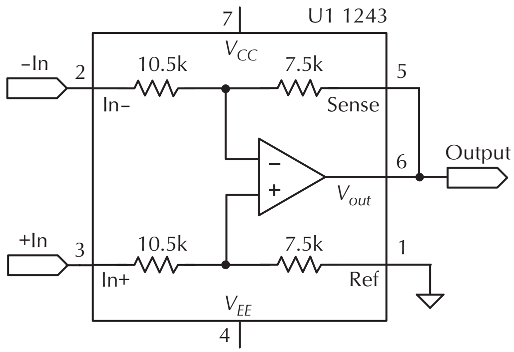


Figure 12-62. THAT 1243 with 3 dB attenuation. Courtesy THAT Corporation.

The ±6 dB parts from all three manufacturers (and the ±3 dB parts from THAT) may be configured for gain instead of attenuation. To accomplish this, the reference and sense pins are be used as inputs with the In- pin connected to V_{out} and the In+ pin connected to ground. A line receiver configured for 6 dB gain is shown in Fig. 12-63.

Balanced line receivers may also be used to provide sum-difference networks for "mid-side" (M/S or M-S) encoding/decoding as well as general-purpose applications requiring precise difference amplifiers. Such applications take advantage of the precise matching of resistor ratios possible via monolithic, laser-trimmed resistors. In fact, while these parts are usually promoted

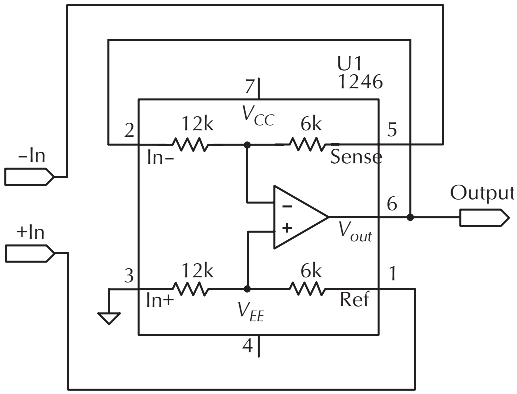


Figure 12-63. THAT 1246 with 6 dB gain. Courtesy THAT Corporation.

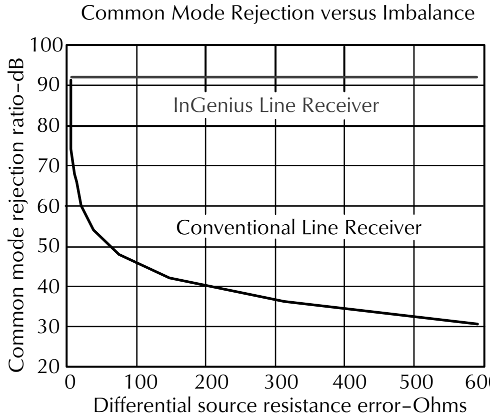


Figure 12-65. CMR imbalance versus source. Courtesy THAT Corporation.

as input stages, they have applications to many circuits where precise resistor ratios are required. The typical 90 dB common-mode rejection advertised by many of these manufacturers requires ratio matching to within 0.005%.

Any resistance external to the line receiver input appears in series with the highly-matched internal resistors. A basic line receiver connected to an imbalanced circuit is shown in Fig. 12-64. Even a slight imbalance, one as low as 10 Ω from connector oxidation or poor contact, can degrade common-mode rejection. Fig. 12-65 compares the reduction in CMR for low common-mode impedance line receivers vs. the THAT1200-series or a transformer.

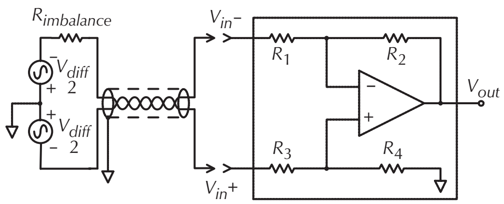


Figure 12-64. Balanced circuit with imbalance. Courtesy THAT Corporation.

The degradation of common-mode rejection from impedance imbalance comes from the relatively low-impedance load of simple line receivers interacting with external impedance imbalances. Since unwanted hum and noise appears in common-mode (as the same signal in both inputs), common-mode loading by common-mode input impedance is often a significant source of error. (The differential input impedance is the load seen by differential signals; the common-mode input impedances is the load seen by common-mode signals.) To reduce the effect of impedance imbalance, the common-mode input impedance, but not the differential impedance, must be made very high.

12.3.6.5 Balanced Line Receivers with the common-mode Performance of a Transformer

The transformer input stage has one major advantage over most active input stages: its common-mode input impedance is extremely high regardless of its differential input impedance. This is because transformers offer floating connections without any connection to ground. Active stages, especially those made with the simple SSM2141-type IC have common-mode input impedances of approximately the same value as their differential input impedance. (Note that for simple differential stages such as these, the common-mode and differential input impedances are not always the same.) Op-amp input bias current considerations generally make it difficult to use very high impedances for these simple stages. A bigger problem is that the noise of these stages increases with the square root of the impedances chosen, so large input impedances inevitably cause higher noise.

Noise and op-amp requirements led designers to choose relatively low impedances (10 k~25 kΩ). Unfortunately, this means these stages have relatively low common-mode input impedance as well (20 k~50 kΩ). This interacts with the common-mode output impedance (also relative to ground) of the driving stage, and added cable or connector resistance. If the driver, cable, or connectors provide an unequal, non-zero common-mode output impedance, the input stage loading will upset the natural balance of any common-mode signal, converting it from common-mode to differential. No amount of precision in the input stage's resistors will reject this common-mode-turned-to-differential signal. This can completely spoil the apparently fine performance available from the precisely matched resistors in simple input stages.

An instrumentation amplifier, Fig. 12-66, may be used to increase common-mode input impedance. Input resistors R_{i1} and R_{i2} must be present to supply a bias current return path for buffer amplifiers OA1 and OA2. R_{i1} and R_{i2} can be made large—in the MΩ range—to minimize the effect of impedance imbalance. While it is possible to use this technique to make line receivers with very high common-mode input impedances, doing so requires specialized op-amps with bias-current compensation or FET input stages. In addition, this requires two more op-amps in addition to the basic differential stage (OA3).

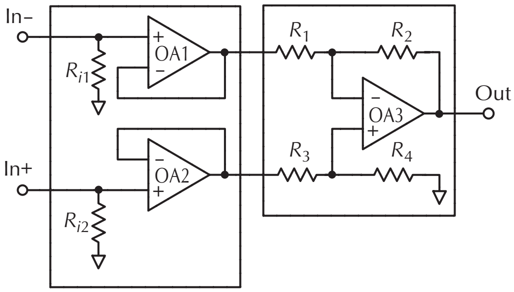


Figure 12-66. Instrumentation amplifier. Courtesy THAT Corporation.

With additional circuitry, even higher performance can be obtained by modifying the basic instrumentation amplifier circuit. Bill Whitlock of Jensen Transformers developed and patented (US patent 5,568,561) a method of applying "bootstrapping" to the instrumentation amplifier in order to further raise common-mode input impedance.[34] THAT Corporation incorporated this technology in its "InGenius" series of input stage ICs.

12.3.6.6 InGenius High Common-Mode Rejection Line Receiver ICs

Fig. 12-67 shows the general principle behind ac bootstrapping in a single-ended connection. By feeding the ac component of the input into the junction of R_a and R_b, the effective value of R_a (at ac) can be made to appear quite large. The dc value of the input impedance (neglecting R_s being in parallel) is $R_a + R_b$. Because of bootstrapping, R_a and R_b can be made relatively small values to provide op-amp bias current, but the ac load on R_s (Z_{in}) can be made to appear to be extremely large

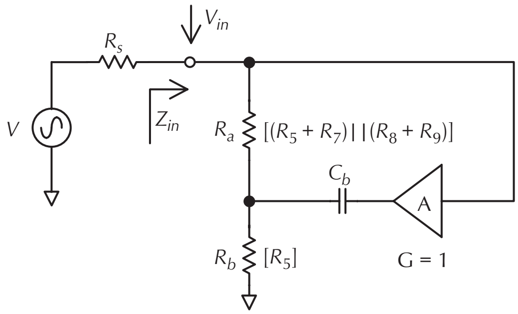


Figure 12-67. Single ended bootstrap. Courtesy THAT Corporation.

A circuit diagram of an InGenius balanced line receiver using the THAT1200 is shown in Fig. 12-68. (All the op-amps and resistors are internal to the IC.) R_5–R_9 provides dc bias to internal op-amps OA1 and OA2. Op-amp OA4, along with R_{10} and R_{11} extract the common-mode component at the input and feed the ac common-mode component back through Cb to the junction of R_7 and R_8. Because of this positive feedback, the effective value of R_7 and R_8—at ac—are multiplied into the MΩ range. In its data sheet for the 1200-series ICs, THAT cautions that C_b should be at least 10 µf to maintain common-mode input impedance (Z_{inCM}) of at least 1 MΩ at 50 Hz. Larger capacitors can increase Z_{inCM} at low power-line frequencies up to the IC's practical limit of ~10 MΩ. This limitation is due to the precision of the gain of the internal amplifiers.

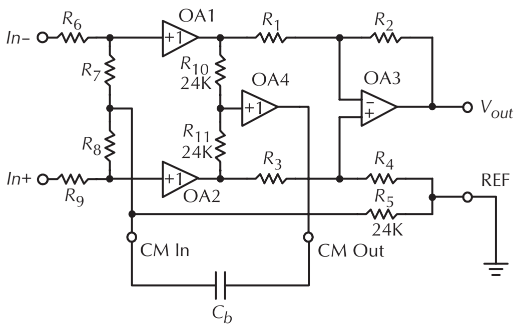


Figure 12-68. Balanced line receiver. Courtesy THAT Corporation.

The outputs of OA1 and OA2 contain replicas of the positive and negative input signals. These are converted to single-ended form by a precision differential amplifier OA3 and laser-trimmed resistors R_1–R_4. Because OA1 and OA2 isolate the differential amplifier, and the positive common-mode feedback ensures very high common-mode input impedance, a 1200-series input stage provides 90 dB CMR even with high levels of imbalance.

It took Bill Whitlock and Jensen Transformers to provide an active input as good as a transformer operating under conditions likely to be found in the real world.

A basic application circuit using the THAT1200-series parts is shown in Fig. 12-69.

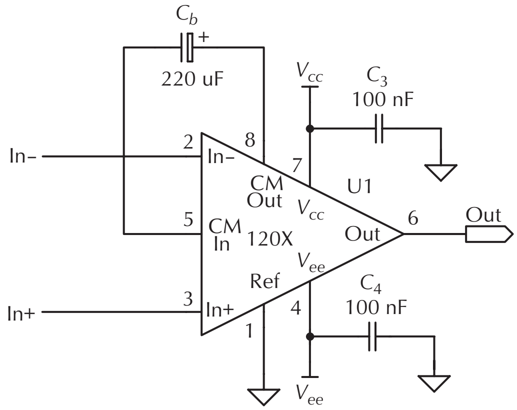


Figure 12-69. InGenius basic application. Courtesy THAT Corporation.

12.3.6.7 Balanced Line Drivers

The Analog Devices SSM2142 and Texas Instruments DRV-series balanced line drivers use a cross-coupled method to emulate a transformer's floating connection and provide constant level with both single-ended (grounded) terminations and fully-balanced loads. A block diagram of a cross-coupled line driver is shown in Fig. 12-70. The force and sense lines are normally connected to each output either directly or through small electrolytic coupling capacitors. A typical application of the SSM2142 driving an SSM2141 (or SSM2143) line receiver is provided in Fig. 12-71.

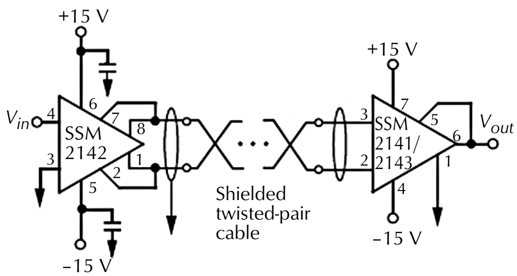


Figure 12-71. SSM2142 driving a SSM2141 line receiver. Courtesy Analog Devices, Inc.

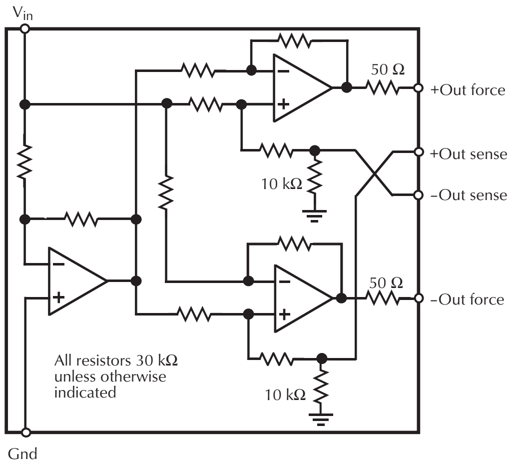


Figure 12-70. SSM2142 cross coupled output. Courtesy Analog Devices, Inc.

If one output of the cross-coupled line driver outputs is shorted to ground in order to provide a single-ended termination, the full short-circuit current of the device will flow into ground. Although this is not harmful to the device, and is in fact a recommended practice, large, clipped signal currents will flow into ground which can produce crosstalk within the product using the stage, as well as in the output signal line itself.

THAT Corporation licensed a patented technology developed by Chris Strahm of Audio Teknology Incorporated. US patent 4,979,218, issued in December 1990, describes a balanced line driver that emulates a floating transformer output by providing a current-feedback system where the current from each output is equal and out of phase to the opposing output.[35] THAT trademarked this technology as "OutSmarts" and introduced its THAT1646 line driver having identical pinout and functionality to the SSM2142. THAT also offers a version of the 1646 with differential inputs known as the THAT1606. Fig. 12-72 is a simplified block diagram of the THAT1646.

The THAT1646 OutSmarts internal circuitry differs from other manufacturer's offerings. Outputs D_{out-} and D_{out+} supply current through 25 Ω build-out resistors. Feedback from both sides of these resistors is returned into *two* internal common-mode feedback paths. The driven side of the build-out resistors are fed back into the common-mode C_{in-} input while the load side of the build out resistors, through the sense- and sense+ pins, provide feedback into the C_{in+} input. A current feedback bridge circuit allows the 1646 to drive one output shorted to ground to allow a single-ended load to be connected. The output short increases gain by 6 dB, similarly to

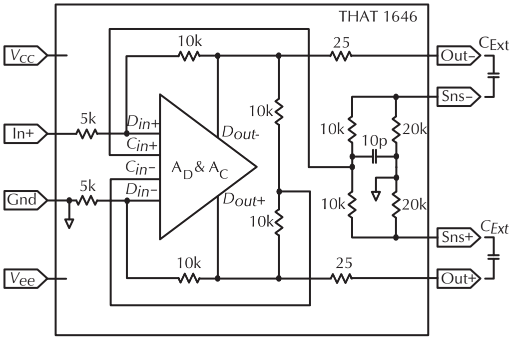


Figure 12-72. THAT 1646 block diagram. Courtesy THAT Corporation.

conventional cross-coupled topologies. However, it does so without loss of the common-mode feedback loop. The resulting current feedback prevents large, clipped signal currents flowing into ground. This reduces the crosstalk and distortion produced by these currents.

A typical application circuit for the THAT1646 is shown in Fig. 12-73.

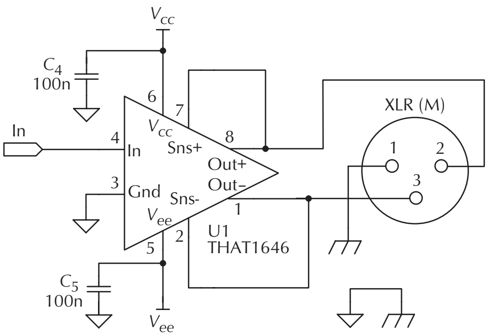


Figure 12-73. THAT 1646 application. Courtesy THAT Corporation.

To reduce the amount of common-mode dc offset, the circuit in Fig. 12-74 is recommended. Capacitors C_1 and C_2, outside the primary signal path, minimize common-mode dc gain, which reduces common-mode output offset voltage and the effect of OutSmarts at low frequencies. Similar capacitors are used in the ADI and TI parts to the same effect, although OutSmarts current feedback does not apply.

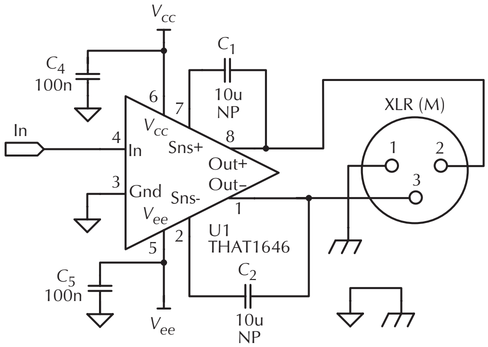


Figure 12-74. THAT 1646 CMR offset reduction circuit. Courtesy THAT Corporation.

THAT's 1606 version of OutSmarts provides a differential input for easier connection to a digital-to-analog converter's output. A typical application of the THAT1606 is shown in Fig. 12-75. Another advantage to the 1606 is that it requires only single low-value capacitor (typically a film type) versus the two larger capacitors required by the THAT1646, SSM2142 or DRV134.

Active balanced line drivers and receivers offer numerous advantages over transformers providing lower cost, weight, and distortion, along with greater bandwidth and freedom from magnetic pickup. Whenused properly, active devices perform as well, and in many ways better, than the transformers they replace. With careful selection of modern integrated circuit building blocks from several IC makers, excellent performance is easy to achieve.

12.3.7 Digital Integrated Circuits

Digital ICs produce an output of either 0 or 1. With digital circuits, when the input reaches a preset level, the output switches polarity. This makes digital circuitry relatively immune to noise.

Bipolar technology is characterized by very fast propagation time and high power consumption, while MOS technology has relatively slow propagation times, low power consumption, and high circuit density. Fig. 12-76 shows typical circuits and characteristics of the major bipolar logic families.

Table 12-4 gives some of the terminology common to digital circuitry and digital ICs.

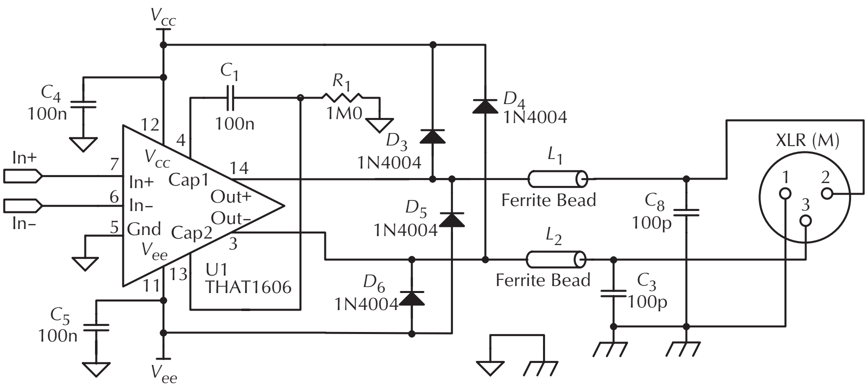


Figure 12-75. THAT 1606 application. Courtesy THAT Corporation.

Table 12-4. Digital Circuit Terminology

Adder	Switching circuits that generate sum and carry bits.
Address	A code that designates the location of information and instructions.
AND	A Boolean logic operation that performs multiplication. All inputs must be true for the output to be true.
Asynchronous	A free-running switching network that triggers successive instructions.
Bit	Abbreviation for binary digit; a unit of binary information.
Buffer	A noninverting circuit used to handle fan-out or convert input and output levels.
Byte	A fixed-length binary-bit pattern (word).
Clear	To restore a device to its standard state.
Clock	A pulse generator used to control timing of switching and memory circuits.
Clock rate	The frequency (speed) at which the clock operates. This is normally the major speed of the computer.
Counter	A device capable of changing states in a specified sequence or number of input signals.
Counter, binary	A single input flip-flop. Whenever a pulse appears at the input, the flip-flop changes state (called a *T flip-flop*).
Counter, ring	A loop or circuit of interconnected flip-flops connected so that only one is on at any given time. As input signals are received, the position of the on state moves in sequence from one flip-flop to another around the loop.
Fan-in	The number of inputs available on a gate.
Fan-out	The number of gates that a given gate can drive. The term is applicable only within a given logic family.
Flip-flop	A circuit having two stable states and the ability to change from one state to the other on application of a signal in a specified manner.
Flip-flop D	D stands for delay. A flip-flop whose output is a function of the input that appeared one pulse earlier; that is, if a one appears at its input, the output will be a one a pulse later.
Flip-flop JK	A flip-flop having two inputs designated *J* and *K*. At the application of a clock pulse, a one on the J input will set the flip-flop to the one or on state; a one on the K input will reset it to the zero or off state; and ones simultaneously on both inputs will cause it to change state regardless of the state it had been in.
Flip-flop RS	A flip-flop having two inputs designated *R* and *S*. At the application of a clock pulse, a one on the S input will set the flip-flop to the one or on state, and a one on the *R* input will reset it to the zero or off state. It is assumed that ones will never appear simultaneously at both inputs.
Flip-flop R, S, T	A flip-flop having three inputs, *R*, *S*, and *T*. The R and S inputs produce states as described for the RS flip-flop above; the T input causes the flip-flop to change states.
Flip-flop T	A flip-flop having only one input. A pulse appearing on the input will cause the flip-flop to change states.
Gate	A circuit having two or more inputs and one output, the output depending on the combination of logic signals at the inputs. There are four gates: AND, OR, NAND, NOR. The definitions below assume positive logic is used.
Gate, AND	All inputs must have one-state signals to produce a zero-state output.
Gate, NAND	All inputs must have one-state signals to produce a one-state output.
Gate, NOR	Any one or more inputs having a one-state signal will yield a zero-state output.
Gate, OR	Any one or more inputs having a one-state signal is sufficient to produce a one-state output.

Table 12-4. Digital Circuit Terminology (Continued)

Inverter	The output is always in the opposite logic state as the input. Also called a NOT circuit.
Memory	A storage device into which information can be inserted and held for use at a later time.
NAND gate (D = ABC for positive inputs)	The simultaneous presence of all inputs in the positive state generates an inverted output.
Negative logic	The more negative voltage (or current) level represents the one-state; the less negative level represents the zero-state.
NOR gate (D = A + B + C for positive inputs)	The presence of one or more positive inputs generates an inverted output.
NOT	A Boolean logic operator indicating negation. A variable designated NOT will be the opposite of its AND or OR function. A switching function for only one variable.
OR	A Boolean operator analogous to addition (except that two truths will only add up to one truth). Of two variables, only one need be true for the output to be true.
Parallel operator	Pertaining to the manipulation of information within computer circuits in which the digits of a word are transmitted simultaneously on separate lines. It is faster than serial operation but requires more equipment.
Positive logic	The more positive voltage (or current) level represents the one-state; the less positive level represents the zero-state.
Propagation delay	A measure of the time required for a change in logic level to spread through a chain of circuit elements.
Pulse	A change of voltage or current of some finite duration and magnitude. The duration is called the *pulse width* or *pulse length*; the magnitude of the change is called the *pulse amplitude* or *pulse height*.
Register	A device used to store a certain number of digits in the computer circuits, often one word. Certain registers may also include provisions for shifting, circulating, or other operations.
Rise time	A measure of the time required for a circuit to change its output from a low level (zero) to a high level (one).
Serial operation	The handling of information within computer circuits in which the digits of a word are transmitted one at a time along a single line. Though slower than parallel operation, its circuits are much less complex.
Shift register	An element in the digital family that uses flip-flops to perform a displacement or movement of a set of digits one or more places to the right or left. If the digits are those of a numerical expression, a shift may be the equivalent of multiplying the number by a power of the base.
Skew	Time delay or offset between any two signals.
Synchronous timing	Operation of a switching network by a clock pulse generator. Slower and more critical than asynchronous timing but requires fewer and simpler circuits.
Word	An assemblage of bits considered as an entity in a computer.

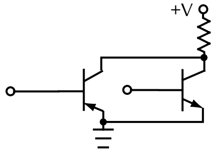
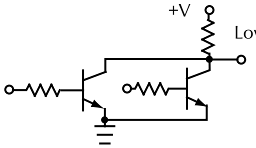
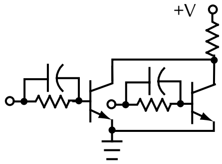
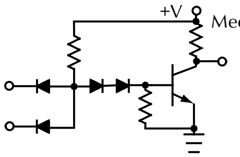
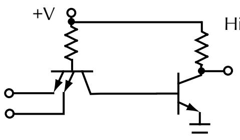
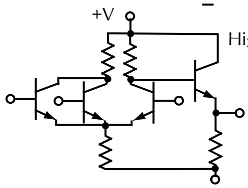
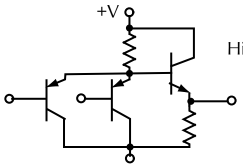
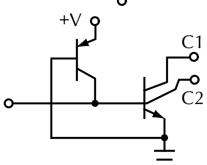
Symbol	Circuit diagram	Speed*	Power*	Fan-out*	Noise immunity*	Trade name	Remarks
DCTL		Medium	Medium	Low	Low	Series 53	Variations in input characteistics result in base current "hogging" problem. Proper operation not always guaranteed. More susceptibile to noise because of low operating and signal voltages.
RTL		Low	Low	Low	Low	RTL	Very similar to DCTL. Resistors resolve current "hogging" problem and reduce power dissipation. However,operating speed is reduced.
RCTI		Low	Low	Low	Low	Series 51	Though capacitors can increase speed capability, noise immunity is affected by capacitive coupling of noise signals.
DTL		Medium	Medium	Medium	Medium to high	930 DTL	Use of pull-up resistor and charge-control technique improves speed capabilities. Many variations of this circuit exist, each having specific advantages.
TTL		High	Medium	Medium	Medium to high	SUHL Series 54/74	Very similaar to DTL. Has lower parasitic capacity at inputs. With the many existing variations, this has become very popular.
CML (ECL)		High	High	High	Medium to high	MECL ECCSL	Similar to a differential amplifier, the reference voltage sets the threshold voltage. High speed, high fan-out operation is posible with associated high power dissipation. Also known as emitter-coupled logic (ECL)
CTL		High	High	Medium	Medium	CTML	More difficult manufacturing process results in compromises of active device characteristcs and higher cost.
I^2L		High	Low	High	Medium	I^2L	Provides smallest and most dense bipolar gate. Simple manufacturing process and higher component packing density than the MOS process. Also known as merged-transistor logic (MTL)
*Low =		<5 MHz	<5 mW	<5	<300 mV		
Medium +		5 –15 MHz	5–15 mW	5–10	300–500 mV		
High =		>15 MHz	>15mW	>10	>500 mV		

Figure 12-76. Typical digital circuits and their characteristics for the major logic families. (Adapted from Reference 4.)

References

1. Gilbert, "A precise Four Quadrant Multiplier with Subnanosecond Response," IEEE *J. of Solid State Circuits*, Vol. SC-3, No. 4, December 1968.
2. Blackmer, "RMS Circuits with Bipolar Logarithmic Converter," United States Patent 3,681,618, August 1, 1972.
3. Blackmer, "Multiplier Circuits," United States Patent 3,714,462, January 30, 1973.
4. "Solid State Music Dual Linear-Antilog Voltage Controlled Amplifier," SSM2000 Data Sheet, September, 1976.
5. Frey, "An Integrated Generalized Voltage Controlled Building Block", *Journal of the Audio Engineering Society*, Preprint No. 2403, November 1986.
6. Curtis Electro Music Website, http://curtiselectromusic.com/Doug_Curtis.html.
7. Baskind, Rubens et al, "The Design and Integration of A High Performance Voltage Controlled Attenuator," *Journal of the Audio Engineering Society*, Preprint No. 1555, November 1979.
8. Welland, "Compensation for VCA op-amp errors," United States Patent 4,434,380, February 28,1984
9. Analog Devices, "SSM2018T Trimless Voltage Controlled Amplifiers," SSM2018T Datasheet, July 2002.
10. Blackmer, "Multiplier Circuits," United States Patent 3,714,462, January 30, 1973.
11. Frey, "An Integrated Generalized Voltage Controlled Building Block," *Journal of the Audio Engineering Society*, Preprint No. 2403, November 1986.
12. Frey, "Voltage-controlled element," United States Patent 4,471,320, September 11, 1984.
13. Frey, "Monolithic voltage-controlled element," United States Patent 4,560,947, December 24, 1985.
14. Rubens, Baskind, "Voltage controlled attenuator," United States Patent 4,155,047, May 15, 1979.
15. "Solid State Music-Dual Linear-Antilog Voltage Controlled Amplifier," SSM2000 Data Sheet, September 1976.
16. IEEE Std 152-1991-IEEE standard for audio program level measurement, June 22, 1992.
17. Blackmer, "RMS Circuits with Bipolar Logarithmic Converter," United States patent 3,681,618, August 1, 1972.
18. Chapel and Gurol, "Thermally Isolated Monolithic Semiconductor Die," United States Patent 4,346,291, August 24, 1982.
19. Analog Devices, AD636 Datasheet.
20. Adams, "Networks for the Log Domain," United States Patent 4,430,626, February 7, 1984.
21. THAT Corporation, "Adaptive Attack and Release Rates Using THAT Corporation RMS Detectors," Design Note DN-114, 2000.
22. Jensen, "JE-990 Discrete Operational Amplifier," *Journal of the Audio Engineering Society*, January/February 1980.
23. Jung, *IC Op-Amp Cookbook*, No. 20969, Howard W. Sams, 1974.
24. www.jensentransformers.com, May 2008.
25. Hebert and Thomas, "The 48 Volt Phantom Menace," Audio Engineering Society Convention Paper 5335, May 2001.
26. Demrow, "Evolution From Operational Amplifier To Data Amplifier," Analog Devices, 1968.
27. Wurcer, "A Programmable Instrumentation Amplifier for 12b Resolution Systems," International Solid State Circuits Conference, February 1982. See also Analog Devices, Analog Dialog, Vol. 30 No. 2, 1996.
28. Harrison Systems, PC1041 Microphone Preamp Schematic, 1978.
29. Cohen, "Double Balanced Microphone Amplifier," Audio Engineering Society Preprint 2106, September 1984.
30. Solid State Music, SSM2011 Datasheet, 1982.
31. Bowers, "An Ultra-Low-Noise Monolithic Microphone Preamplifier," Audio Engineering Society Preprint 2495, 1987.
32. Whitlock, "A New Balanced Audio Input Circuit For Maximum Common-Mode Rejection In Real-World Environments," www.jensen-transformers.com. 1996.
33. Hay, "Differential Technology In Recording Consoles And The Impact Of Transformerless Circuitry On Grounding Technique," Audio Engineering Society Preprint 1723, October 1980.
34. Whitlock, "Differential Line Receiver With Common-Mode Bootstrapping," US patent 5,568,561, October 22, 1996.
35. Strahm, "Balanced Line Output Circuit," United States Patent 4,979,218, December, 18, 1990.