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SEMICONDUCTORS From A to Z

BY PHILLIP DAHLEN

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SEMICONDUCTORS
From A to Z

By Phillip Dahlen

In cooperation with the staff of
ELECTRONIC TECHNICIAN/DEALER

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PREFACE

In the summer of 1966 I was asked to prepare a series of four or five articles for publication in Electronic Technician magazine under the same title as this book. By the time the third article was published, reader response indicated that the ET editors had, indeed, accurately measured the pulse of their readers.

It soon became obvious, however, that four or five articles would fall miserably short of the "A to Z" coverage. Moreover, new developments were being announced continually, creating the need for many additional articles. And the letters of praise kept pouring in. Several suggested that the series be published in book form, but TAB BOOKS was way ahead of them, having proposed the idea when the second article appeared. Thus, by the time the 26th and final installment was published, this book was practically ready for printing.

Whether you're interested in catching up or brushing up, you'll find what you want to know within these pages. The first several chapters describe semiconductor characteristics, beginning with the simple diodes we've come to know pretty well. Then you are exposed to transistors, FETs, MOS FETs (just a variation of the main transistor theme), the versatile but quite simple tunnel diode, and of course integrated circuits, which are now replacing entire circuits.

By now, it is rather obvious that electronics is playing a dominant role in the life of every individual. And it is you, and thousands just like you, who will accept the challenge of maintaining the nation's electronic gear in an age when change is becoming increasingly common. It is hoped that this book will help you meet that challenge.

Greatful acknowledgement is hereby given to the editors of ELECTRONIC TECHNICIAN/DEALER, who originally encouraged my development of this material, and to Dr. Kent H. Bracewell, head of Hamline University's physics department, who taught all his students the importance of clarity and accuracy when describing scientific principles.

Phillip Dahlen

Duluth, Minn.
November 1968

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CHAPTER 1

What Is a Semiconductor?

Early experiments indicated that current in electrical circuits flowed from positive to negative. This was proven false when electron tubes were developed. Current was observed to actually flow from the tube's negatively polarized cathode to its positively polarized plate. By this time, however, people were so entrenched in the "conventional" current-flow concept, they continued to speak and think in those terms.

With the development of semiconductor materials, scientists felt it was necessary to speak more exactly regarding the direction in which current flowed in p-n junctions of solid-state circuits. Hence, many technicians have been thoroughly confused in attempting to reconcile "hole" flow, "electron" flow and "current" flow in semiconductor components and circuits. But these matters actually concern only chemists, physicists and other scientists.

It is not necessary to understand the precise atomic principles of semiconductors to successfully troubleshoot and repair solid-state equipment. We intend, therefore, to circumvent this man-made confusion by sticking to the accepted electron theory — as we already know it in electron tube circuitry.

Although we believe no useful purpose can be served by attempting a full, and necessarily artificial, comparison between electron tubes and transistors, we will call attention to an occasional similarity or difference between the two components if an understanding of the similarity or difference seems helpful.

Fig. 1—Forward biased diode electron tube.

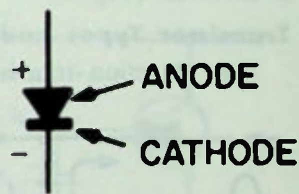

Fig. 2—Forward biased semiconductor diode.

Current Flow

When current flows through an electron tube diode (Fig. 1), the plate must be more positive than the cathode. Similarly, if current is to flow through a semiconductor diode (Fig. 2), its anode must be more positive than its cathode. Note the semiconductor diode schematic shows the anode as an arrow point and the cathode as a bar. The arrow points toward the negative polarity. Before this diode can conduct significant current, it must be forward biased — the cathode must be negatively polarized and the anode positively polarized. Note, also, that current flows through the diode from negative to positive — according to the accepted electron theory — opposite the direction of the arrow (originally pointed to indicate the direction of "convential" current flow). If an ac voltage is applied to this diode it will conduct current only when the anode is positive — on the positive half-cycles of a sinewave, for example. But it will conduct on the negative half-cycles if its position in the circuit is reversed (Fig. 3).

Transistor Types and Functions

Two junction-transistor types, NPN and PNP, are in general use today (Fig. 4). An NPN transistor's P and N material arrangement requires the collector (Fig. 4A) to be more positive than the emitter before the transistor can conduct current. Note the transistor symbol shows the emitter having an arrow. The arrow points toward the negative polarity, the same as the arrow in a diode symbol.

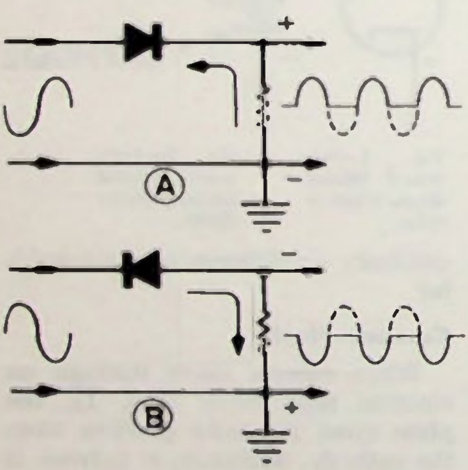

Fig. 3 (A)—Positive peaks passing through diode. (B)—Negative peaks passing through diode.

Because the PNP transistor's P and N material is arranged in a different manner, it is biased differently. Its collector is biased

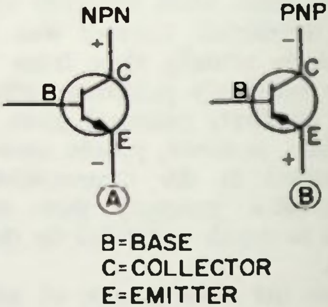

Fig. 4 (A)—The NPN transistor collector is the most positive element. (B)—The PNP transistor collector is the most negative element.

more negative than its emitter (Fig. 4B). The arrow shown on the emitter symbol points away from the positive polarity.

The emitter of an NPN transistor should be negative while the emitter of a PNP transistor should be positive.

When the cathode and plate of an electron tube (Fig. 1) are properly biased, it will conduct. By making the tube's grid more negative than its cathode (Fig. 5) current flow can be decreased and even stopped. Electron tube current flow is controlled by varying the negative bias between the grid and cathode.

Since transistors are made of semiconductor material, their chemical composition does not allow them to be good electrical conductors under normal conditions. Transistors must be "encouraged" to conduct electricity. Biasing the transistor base to encourage current flow is called forward biasing. As previously stated, the collector of an NPN transistor should be more

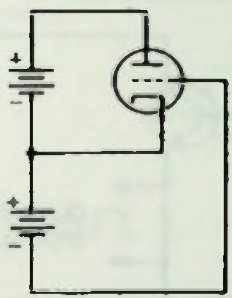

Fig. 5—Biasing circuit for triode electron tube.

positive than the emitter. To encourage current flow, the base should also be more positive than the emitter (which results in forward biasing) as shown in Fig. 6.

The collector of a PNP transistor should be more negative than the emitter. To encourage current flow the base should also be more negative than the emitter (forward biased) as shown in Fig. 7. A forward biased PNP transistor has emitter-to-collector and emitter-to-base polarities, opposite those of forward biased NPN transistors.

If the base lead of a transistor is broken, the transistor will no longer be forward biased and will virtually cease to conduct current. But an electron tube will conduct current when its grid lead is broken.

Transistor Circuits and Signal Characteristics

Three basic transistor circuits are in general use today. The most widely used circuit is the common-emitter. When a positive signal is applied to the base of an NPN transistor in a common-emitter circuit (Fig. 8), the transistor becomes

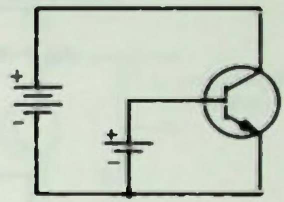

Fig. 6—Biasing circuit for NPN transistor.

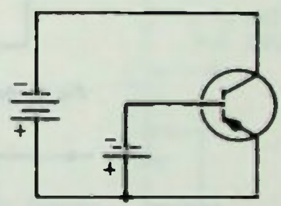

Fig. 7—Biasing circuit for PNP transistor.

more forward biased and conducts more current. As more current flows through the collector and the collector resistor, a larger voltage drop appears across the collector resistor. This results in a voltage drop at the transistor collector and the input signal in this circuit is inverted at the output.

When a positive signal is applied

to the base of an NPN transistor in a grounded-collector circuit (Fig. 9), the transistor becomes more forward biased and more current flows through the emitter and the emitter resistor. A greater voltage drop takes place across the emitter resistor and the emitter becomes more positive. But the input signal in the grounded-collector circuit is not inverted.

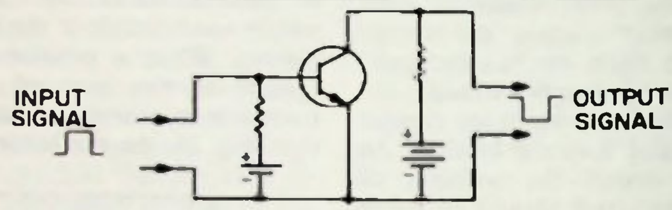

Fig. 8—Basic common emitter circuit.

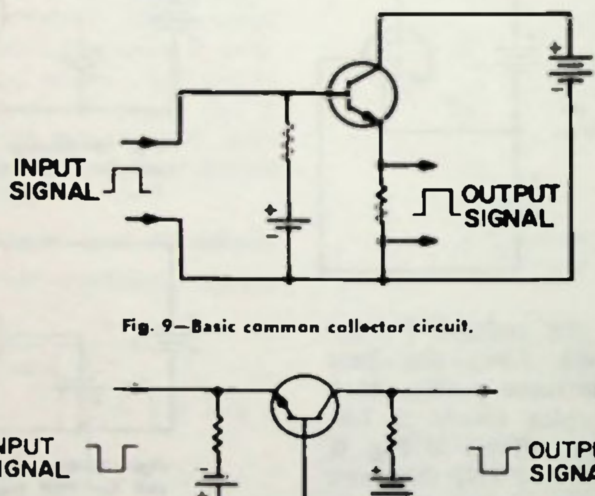

Fig. 9—Basic common collector circuit.

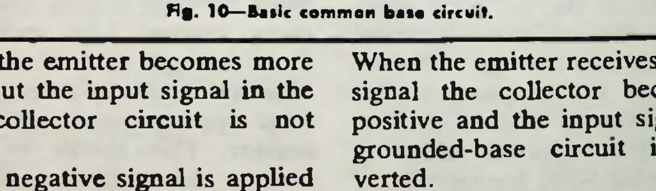

Fig. 10—Basic common base circuit.

When a negative signal is applied to the emitter in a common-base circuit (Fig.10), the base is made more positive than the emitter. The transistor is then forward biased and conducts more current. As more current flows through the collector and the collector resistor, a larger voltage drop appears across the collector resistor, causing the collector to become less positive. When the emitter receives a negative signal the collector becomes less positive and the input signal in the grounded-base circuit is not inverted.

By reversing the polarity of all bias batteries and signals, the descriptions and diagrams for the three basic circuits will apply to PNP transistors.

Compound Circuit

Common-base and common-emitter circuits are frequently combined to form a single circuit with dual function.

We assume the bias supplies for these circuits are from constant voltage sources. An input signal applied directly to a base or emitter resistor will affect the transistor's forward bias in a common-emitter or common-base circuit. An output signal results from the voltage drop across the collector resistor and can be obtained directly from it. The inputs and outputs of these two circuits can be changed (Fig. 11) without affecting their functions.

The revised common-base circuit (Fig. 11A) can be rearranged without making any additional changes in the circuit (Fig. 12). This circuit can be modified (Fig. 13) by regrouping the two bias supplies. The amount of emitter-to-base and emitter-to-collector bias, however, remains unchanged. The base-to-collector voltage must, therefore, also remain the same.

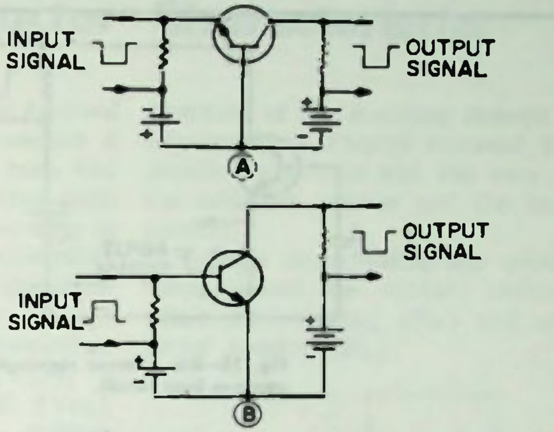

Fig. 11 (A)—Bias sources removed from signals in common base circuit. (B)—Bias sources removed from signals in common emitter circuit.

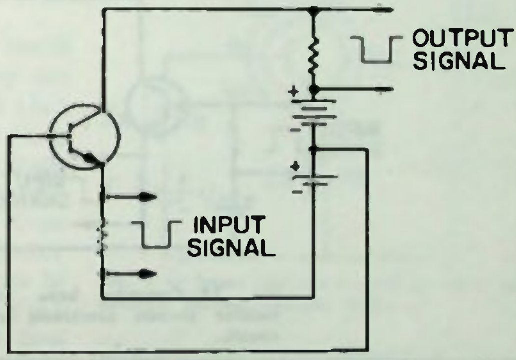

Fig. 12—Another version of common base circuit with bias sources removed from signals.

A negative signal applied to the emitter resistor in the revised common-base circuit will still produce a negative output across the collector resistor.

The biasing network for the revised circuit is similar to that in the revised common-collector circuit (Fig. 11B). In that circuit a positive signal applied to the emitter resistor produces a negative signal across the collector resistor.

A positive signal applied to the base resistor will forward bias the transistor and cause it to conduct more current. A negative signal applied to the emitter resistor will also forward bias the transistor and cause it to conduct more current. In either case, the resulting current will cause a greater voltage drop across the collector resistor, resulting in a negative output signal.

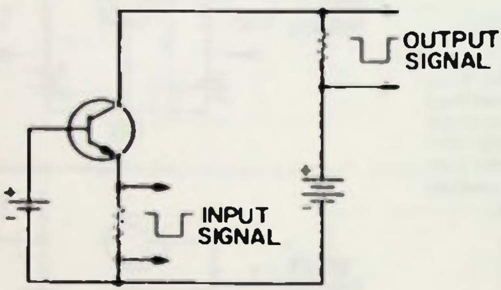

Fig. 13—Bias sources regrouped in common base circuit.

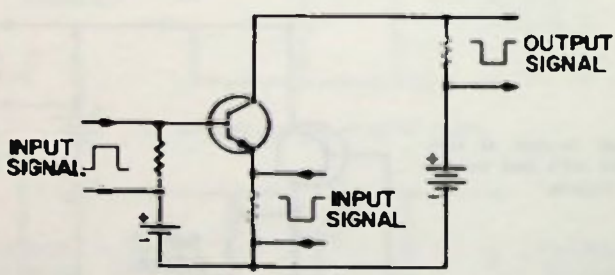

Fig. 14—Common base and common emitter circuits combined as compound circuit.

The two circuits (Fig. 13 and 11B) can be combined (Fig. 14).

Two independent input signals can be applied to this compound circuit where they will be amplified, and mixed at the output.

CHAPTER 2

Semiconductor Characteristics

NPN transistors must be forward biased before they can conduct a significant current — the base and collector being made positive with respect to the emitter. This bias is provided by the collector-to-emitter voltage source (V_{CC}) and the emitter-to-base voltage source (V_{BB}). These sources combine to develop an emitter-to-base voltage (V_{EB}), a collector-to-emitter voltage (V_{CE}) and a collector-to-base voltage (V_{CB}) (Fig. 1).

Two significant currents result from these bias voltages. They are the emitter-to-collector current (I_C) and the emitter-to-base current (I_B). At normal temperatures good transistors will not experience a significant base-to-collector current. Since both the base current and collector current in the transistor originate in the emitter, the emitter current (I_E) must equal the sum of these two currents ($I_C + I_B = I_E$).

PNP transistors must also be forward biased before they can conduct current — the base and collector being made negative with respect to the emitter. By substituting a PNP transistor for the NPN transistor described, the polarity of the biasing voltages is reversed and the direction of the resulting current is also reversed. Though reversed, the emitter current is still the sum of the collector current and the base current.

Voltage drops within the circuit occur across the emitter resistor (R_E), base resistor (R_B) and collector resistor (R_C).

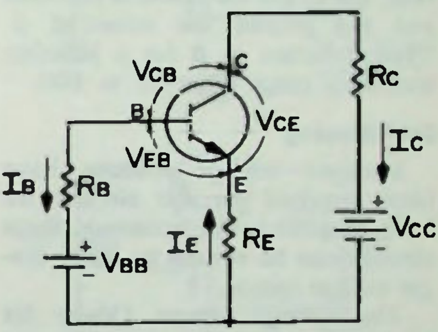

Fig. 1 — Input currents I_B and I_E affect the circuit's output current (I_C).

Relative size of currents

About 92 to 99 percent of the current flowing through the emitter passes through the collector of a junction transistor. The remaining 8 to 1 percent is the base current. The collector current (I_C) is much greater than the base current (I_B). The collector current nearly equals the emitter current (I_E). The ratio of collector current to emitter current is called Alpha ($\alpha = \frac{I_C}{I_E}$). The collector current must always be less than the emitter current since all forward-biased junction transistors normally experience a base current. Therefore, the value of α must always be less than one.

The smaller the base current, the nearer the collector current comes to equaling the emitter current, and the nearer α comes to becoming one. Typical values of α for junction transistors range from 0.92 to 0.98.

The ratio of collector current to base current is called Beta ($\beta = \frac{I_C}{I_B}$). The smaller the base current, the larger the fraction becomes and the greater the value of β. Typical values of β for a junction transistor range from 25 to 100.

Self Biasing

Multiple voltage sources have been required for the circuits we have described. For economy, these circuits can be revised to use a single voltage source.

The voltage source (V_{BB}) for biasing the base of the common-emitter amplifier (Fig. 2A) is able to provide the desired emitter-to-base voltage (V_{EB}) with the necessary base current (I_B) passing through the base resistor (R_{B1}). The same emitter-to-base voltage (V_{EB}) and base current (I_B) can be obtained from the voltage source (V_{CC}) used to provide the collector-to-emitter bias (Fig. 2B). This can be done with a base resistor (R_{B2}) of an appropriate higher value. The increase in resistance is equal to the additional voltage drop ($V_{BB} - V_{CC}$) divided by the base current (I_B). ($R_{B2} = R_{B1} + \frac{V_{CC} - V_{BB}}{I_B}$).

The common-collector amplifier (Fig. 3A) can be revised in the same manner as was the common-

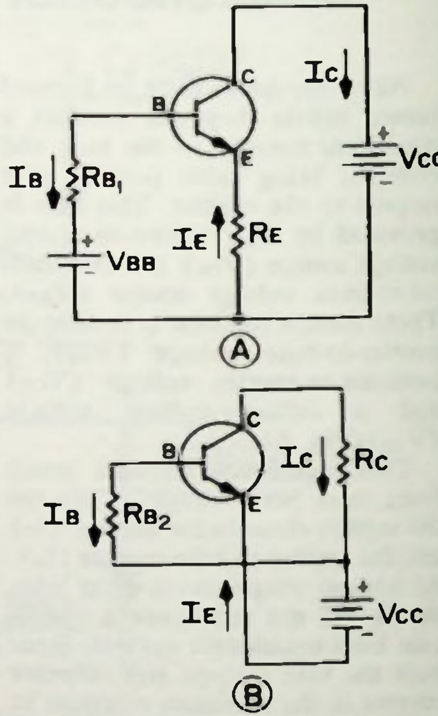

Fig. 2 (A and B)—Common-emitter voltage sources are combined by increasing R_B.

emitter amplifier (Fig. 3B). By increasing the base resistor (R_B), the same emitter-to-base voltage (V_{EB}) and base current (I_B) can be obtained from the higher collector-to-emitter voltage source (V_{CC}).

The common-base diagram (Fig. 4A) can be revised (Fig. 4B) without changing the circuit. The same emitter-to-base voltage (V_{EB}) and base current (I_B) can be obtained by inserting a base resistor (R_B). The value of the base resistor can be determined in the same manner as it was for the other two circuits ($R_B = \frac{V_{CC} - V_{BB}}{I_B}$). The base of the transistor is no longer grounded by

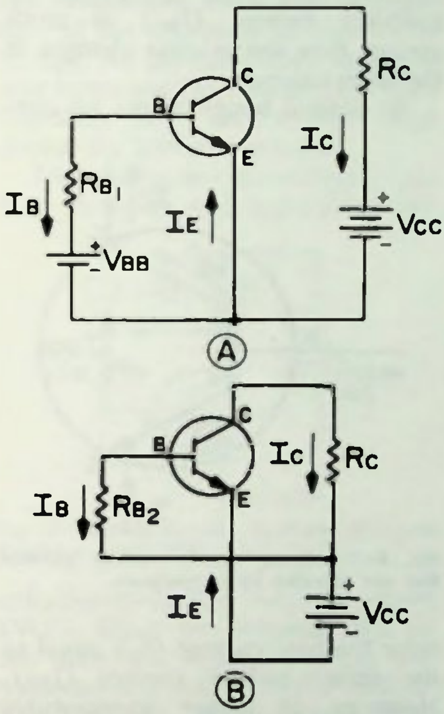

Fig. 3 (A and B)—Common-collector voltage sources are combined by increasing R_B.

its biasing voltage source (V_{EE} or V_{CC}). The effect of an ac ground must, therefore, be obtained by using a capacitor (C_B).

The compound circuit (Fig. 1) can also be modified (Fig. 5) by increasing the base resistor (R_B) to the proper value (R_{B2}). The compound circuit is a combination of the common-base and common-emitter circuits. The common-base circuit (Fig. 4) required a capacitor (C_B) to function without signal loss. The two inputs shown in the com-

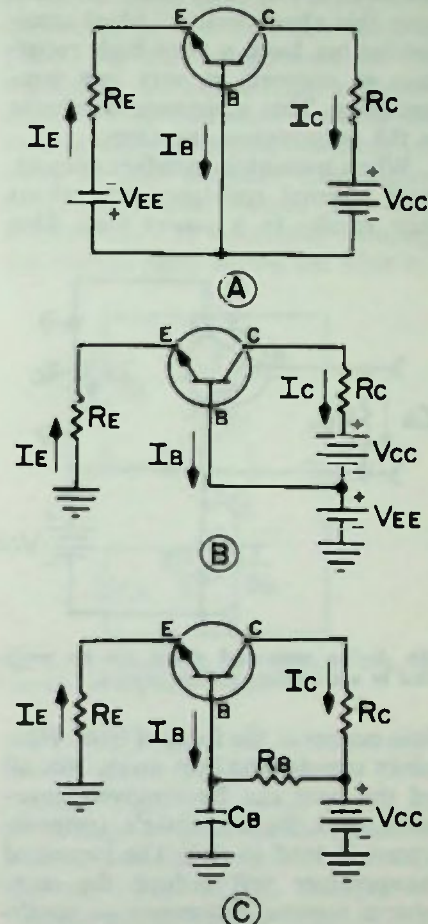

Fig. 4 (A, B and C)—Common-base circuit can be modified for single voltage source.

pound circuit (Fig. 5) should each be tuned to its own separate frequency to eliminate a corresponding signal loss of the other signal at each input.

Effect of Temperature

Most conductors have a low resistance to currents at very low temperatures and higher resistances to currents at higher temperatures. Semiconductor materials do not have this characteristic. Most semiconductors have a very high resistance to currents at very low temperatures. This resistance decreases as the temperature increases.

When transistors conduct current, their internal resistance to current flow results in a power loss. This loss occurs in the form of heat. Heat sinks transfer the heat away. Not all of the heat can be removed, however, and the transistor's temperature will tend to rise. The increased temperature will reduce the transistors internal resistance — resulting in an emitter-to-collector (I_{EC}) and emitter-to-base (I_{EB}) current increase (Fig. 6).

We indicated previously that good transistors will not experience a significant base-to-collector current when operating at normal temperatures. But when the resistances in a transistor are reduced by higher temperatures, the base-to-collector current becomes significant. In germanium transistors this current doubles for about every 9°C rise above room temperature and in silicon transistors this current doubles for about every 11°C rise above room temperature. Having been very low, the relative changes in the base-to-collector current (I_{BC}) is much greater than the relative changes in the other currents.

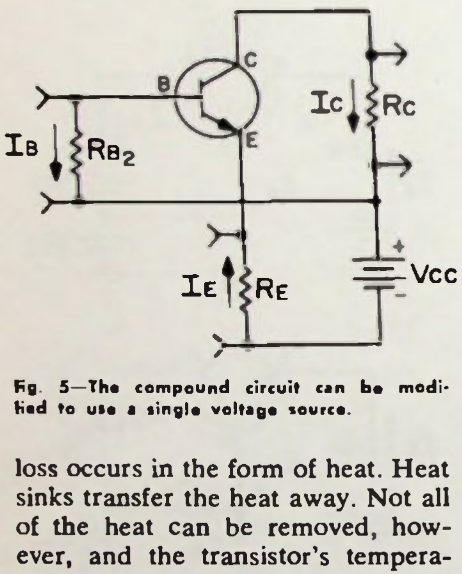

Fig. 5—The compound circuit can be modified to use a single voltage source.

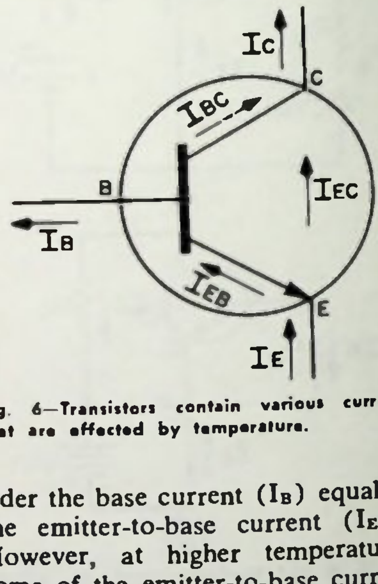

Fig. 6—Transistors contain various currents that are effected by temperature.

At normal temperatures we consider the base current (I_B) equal to the emitter-to-base current (I_{EB}). However, at higher temperatures some of the emitter-to-base current continues through the transistor as the base-to-collector current (I_{BC}), reducing the remaining base current ($I_B = I_{EB} - I_{BC}$).

The collector current (I_C) then contains both the emitter-to-collector current (I_{EC}) and the base-to-collector (I_{BC}). ($I_C = I_{EC} + I_{BC}$). The emitter current (I_E) still contains only the emitter-to-base current (I_{EB}) and the emitter-to-collector current (I_{EC}). ($I_E = I_{EB} + I_{EC}$) Note that the emitter current (I_E) is still equal to the base current (I_B) plus the collector current (I_C). [$I_E = I_B + I_C = (I_{EB} - I_{BC}) + (I_{EC} + I_{BC}) = (I_{EB} + I_{EC})$]

The biasing circuits shown are not satisfactory when the temperature of the transistor is allowed to change. As the transistor's temperature increases, the base current (I_B) decreases and the voltage drop across R_B becomes smaller.

The emitter-to-base voltage (V_{EB}) then increases and approaches the

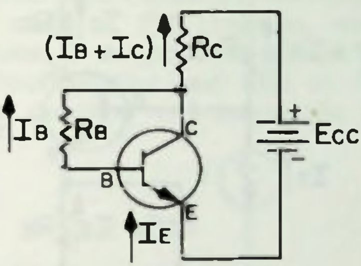

Fig. 7—Collector-to-base feedback improves transistor stabilization.

collector-to-emitter voltage source (V_{CC}). Since the forward bias of the base has been increased, the transistor will conduct more current. The increased current will result in the generation of more heat and result in the further reduction of the transistor's base current (I_B). The temperature and base current will continue to change until either the circuit is overloaded or the transistor burns out. This is called thermal "run-away" or avalanche effect.

Bias Stablization

Circuits can be designed to compensate for the thermo effect in transistors. Collector-to-base feedback (Fig. 7) is one way of improving bias stabilization. Instead of connecting the base resistor (R_B) directly to the voltage source (V_{CC}), it is connected to the collector resistor (R_C). As the temperature of the transistor increases, the collector current (I_C) increases, causing the voltage drop across the collector resistor to also increase. The higher temperature also causes the base current (I_B) to decrease, reducing the voltage drop across the base re-

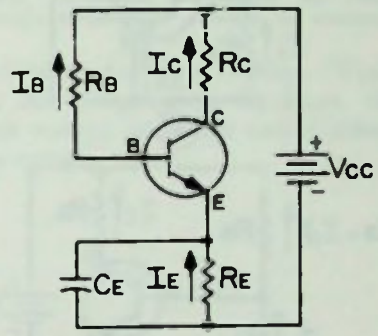

Fig. 8—Transistors can be stabilized with an "emitter swamping" circuit.

sistor (R_B). Since the collector current is much larger than the base current, a greater change occurs in the voltage drop across the collector resistor than across the base resistor. The net effect of the changes in voltages across the two resistors is a reduction in the forward bias of the transistor's base (V_{EB}). This reduction in the base bias reduces the collector current and stabilizes the circuit.

"Emitter swamping" is another way to stabilize a transistor circuit (Fig. 8). As the temperature of the transistor increases, there is an increase in the emitter current (I_E) and a decrease in the base current (I_B). This results in a greater voltage drop across the emitter resistor (R_E) and a smaller voltage drop across the base resistor (R_B).

Since the emitter current is much larger than the base current, the

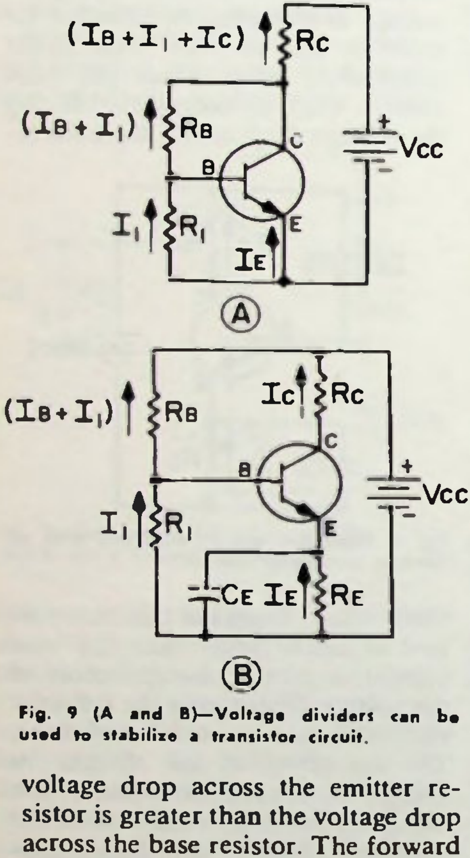

Fig. 9 (A and B)—Voltage dividers can be used to stabilize a transistor circuit.

voltage drop across the emitter resistor is greater than the voltage drop across the base resistor. The forward bias of the base (V_{EB}) is equal to the voltage source (V_{CC}) minus the voltage drop across the base resistor (I_BR_B) and the voltage drop across the emitter resistor (I_ER_E). ($V_{EB} = V_{CC} - I_BR_B - I_ER_E$).

As the temperature of the transistor increases the current change reduces the forward base bias of the transistor and stabilizes the circuit. By connecting a capacitor (C_E) across the emitter resistor (R_E), only the dc portion of the voltage drop affects the circuit bias. The ac portion is able to pass through, reducing the signal loss in the resistor.

The collector-to-base feedback and emitter swamping circuits described provide some bias stabilization. In both circuits the reduced voltage drop across the base resistor (R_B) reduces the stabilizing effect

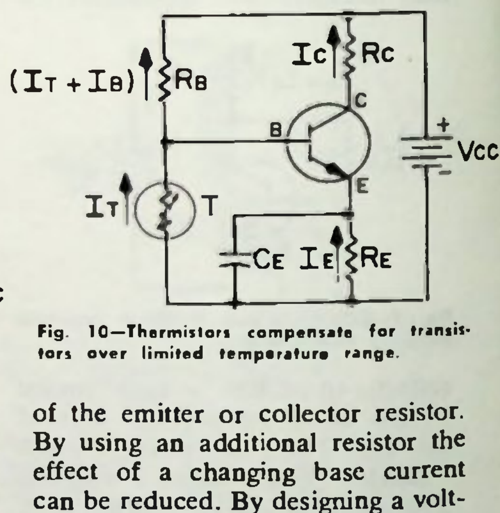

Fig. 10—Thermistors compensate for transistors over limited temperature range.

of the emitter or collector resistor. By using an additional resistor the effect of a changing base current can be reduced. By designing a voltage divider (Fig. 9) so that the current (I_1) passing through the resistors (R_B & R_1) is much greater than the base current (I_B), the voltage drop across the base resistor (R_B) is more greatly influenced by the voltage divider current (I_1) than it is by the base current (I_B). The resulting change in the voltage drop across the base resistor is reduced, increasing circuit stabilization.

The voltage divider can be improved (Fig. 10) by substituting a thermistor (R_T) for the second resistor (R_1). The resistance of the thermistor decreases as the temperature increases. As its resistance decreases, it conducts more current (I_T) and causes a greater voltage drop to occur across the base resistor (R_B). This reduces the forward bias of the base (V_{EB}) and stabilizes the circuit. The thermo characteristics of the transistor and any one thermistor are similar over only a limited range of temperatures. The selection of a thermistor must depend on the temperature range.

Since both transistors and diodes are made of PN junctions, their reaction to temperatures should be relatively similar and shift in the same direction and magnitude. A diode (D) has a slight resistance to a forward biasing current and can be substituted for the thermistor (R_T) in the circuit (Fig. 11) to

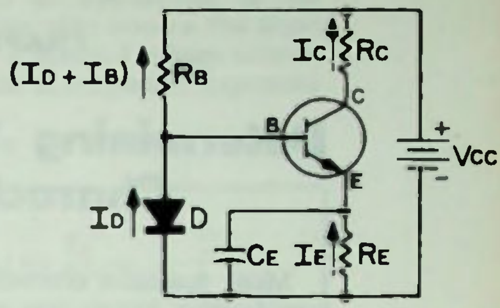

Fig. 11—Diodes compensate for relatively large temperature changes in transistors.

determine the bias voltage (V_{EB}). As the temperature increases, the bias voltage decreases and stabilizes the circuit.

CHAPTER 3

Determining Semiconductor Characteristics

Many transistor characteristics must be taken into consideration before you can understand the circuits well enough to troubleshoot them efficiently.

Emitter Current Gain

A previous chapter indicated that the ratio of collector current (I_C) to emitter current (I_E) is called alpha (α). ($\alpha = \frac{I_C}{I_E}$) This is a ratio of dc currents and is frequently expressed as α_{dc}, H_{FB} or h_{FB}. It was indicated that the emitter current is equal to the sum of the base and collector currents ($I_E = I_B + I_C$). The collector current must, therefore, be smaller than the emitter current and the value of α ($\frac{I_C}{I_E}$) must always be less than one. The value of α in typical junction transistors lies between 0.9 and 1.0.

Resistors reduce the currents flowing through a transistor. But transistor specifications would be too complicated if circuit resistors were taken into consideration. Hence, when specifications refer to fluctuations in the emitter and collector currents, we assume the transistor is connected directly to a constant voltage source — with emitter or collector resistors shorted out of the circuit. The ratio of these fluctuations ($\frac{\Delta i_c}{\Delta i_e}$) is frequently expressed as α or h_{fb}.

Base Current Gain

A previous chapter also indicated that the ratio of the collector current (I_C) to the base current (I_B) is called beta (β). ($\beta = \frac{I_C}{I_B}$). The dc beta is frequently expressed as H_{FE} or h_{FE}.

When specifications refer to fluctuations in the collector and base currents, we also assume the transistor is connected directly to a constant voltage source. The ratio of these fluctuations is frequently expressed as β or h_{fe}.

With algebra (Fig. 1) we can see that a definite

$$\beta = \frac{I_C}{I_B}. \quad \frac{1}{\beta} = \frac{I_B}{I_C}. \quad \alpha = \frac{I_C}{I_E} = \frac{I_C}{I_C + I_B}$$

$$\frac{1}{\alpha} = \frac{I_C + I_B}{I_C} = 1 + \frac{I_B}{I_C} = 1 + \frac{1}{\beta}.$$

$$\frac{1}{\alpha} = 1 + \frac{1}{\beta}. \quad 1 = \alpha + \frac{\alpha}{\beta}. \quad \alpha = 1 - \frac{\alpha}{\beta}.$$

$$\alpha\beta = \beta - \alpha. \quad \beta - \alpha\beta = \alpha.$$

$$\beta\ (1 - \alpha) = \alpha. \quad \beta = \frac{\alpha}{1 - \alpha}$$

$$\alpha\beta + \alpha = \beta. \quad \alpha\ (\beta + 1) = \beta. \quad \alpha = \frac{\beta}{\beta + 1}.$$

Fig. 1—Calculations show a relationship between alpha and beta.

relationship exists between alpha and beta. ($\alpha = \frac{\beta}{\beta+1}$ $\beta = \frac{\alpha}{1 - \alpha}$.)

Since alpha and beta refer to both dc and small signal current ratios, more specific terms will be used henceforth. Alpha will be referred to as H_{FB} or h_{fb} and beta as H_{FE} or h_{fe}. Those in upper case describe dc current ratios while those in lower case describe ac signal current ratios.

If we wish to know a transistor's small-signal base-current gain (h_{fb}), when the small-signal emitter-current gain (h_{fe}) is 0.98, we must make the following calculations:

$$h_{fb} = \frac{h_{fe}}{1 - h_{fe}} = \frac{0.98}{1 - 0.98} = \frac{0.98}{0.02} = \frac{98}{2} = 49.$$

To find the dc emitter-current gain (H_{FE}) when the dc base-current gain (H_{FB}) is 24, we must make the following calculations:

$$H_{FE} = \frac{H_{FB}}{H_{FB} + 1} = \frac{24}{24 + 1} = \frac{24}{25} = 0.96.$$

Voltage Drops Within A Circuit

Kirchhoff's second law states that the applied voltage within any closed circuit is equal to the sum of the

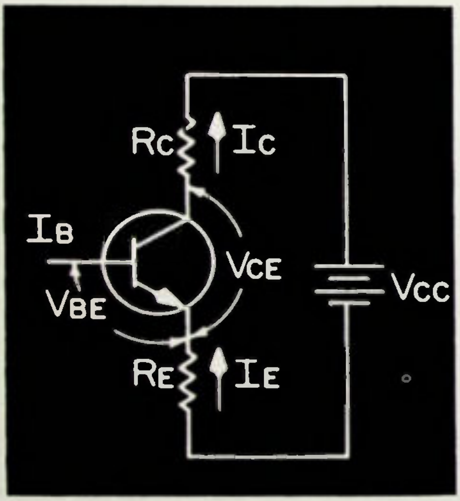

Fig. 2—Voltage drops within a circuit equal the applied voltage.

voltage drops. The voltage drops within the emitter and collector circuits of an NPN transistor are shown in Fig. 2. Here the voltage source (V_{CC}) equals the voltage drop across the collector resistor ($I_C R_C$) plus the collector-to-emitter voltage (V_{CE}) and the voltage drop across the emitter resistor ($I_E R_E$). ($V_{CC} = I_C R_C + V_{CE} + I_E R_E$). Since a transistor's collector current nearly equals its emitter current ($I_C \approx I_E$), the equation can be simplified as follows: ($V_{CC} = I_C R_C + V_{CE} + I_C R_E$).

If the base of a transistor could cut off the collector current completely, no voltage drops would exist across the emitter and collector resistors — the collector-to-emitter voltage would equal the voltage source. (When $I_C = O$, $V_{CE} = V_{CC}$.)

If a transistor could conduct sufficient current to cause the voltage drop across its collector resistor and emitter resistor to equal the voltage source, there would be no collector-to-emitter voltage. (When $I_C R_C + I_C R_E = V_{CC}$, $V_{CE} = O$.) The collector current can be calculated for that condition. (When $V_{CE} = O$, $I_C = \dfrac{V_{CC}}{R_C + R_E}$.)

Graphing A Circuit's Load Line

Transistor circuit design is dependent upon the characteristics of the transistor selected, the desired

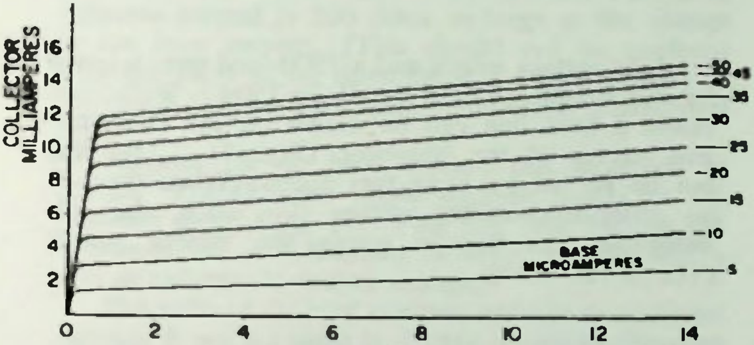

Fig. 3—Curves are used to show transistor dc characteristics.

output load and the voltage source used. The load of the NPN transistor in Fig. 2 consists of its collector resistor and its emitter resistor. ($R_L = R_C + R_E$)

Manufacturers frequently use curves to help describe the characteristics of transistors marketed. Fig. 3 describes the dc characteristics of the NPN transistor selected for the circuit shown in Fig. 2. From this graph we can see that the collector current is more dependent on the base current than it is on the collector-to-emitter voltage.

There is no one collector-to-emitter voltage or collector current at which a transistor must function. The values selected, however, must not exceed the power rating of the transistor.

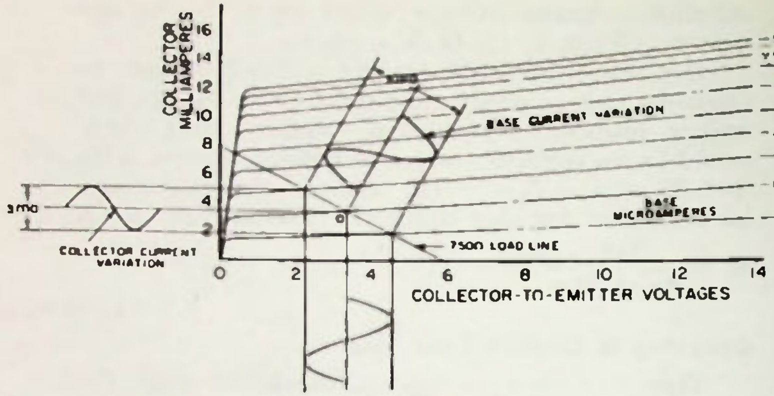

Fig. 4—A load line can be drawn to show transistor voltages and currents in a circuit.

A 6v voltage source and a 750Ω load were selected for this circuit. (V_{EE} = 6v, R_L = 750Ω.) With these values a load line can be drawn on the characteristic curves of the transistor (Fig. 4). This line can be plotted by using two points. When there is no collector-to-emitter voltage (V_{CE} = O), the collector-to-emitter voltage equals the voltage source ($V_{CE} = V_{CC}$ = 6v).

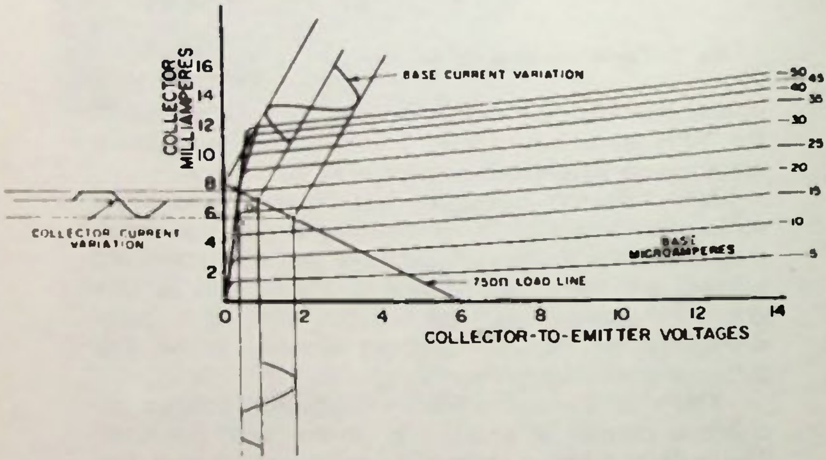

Fig. 5—Too high a Q point may result in a clipped output signal.

Determining A Transistor's Base Current

At any point along this load line we can find the combination of base current, collector current and collector-to-emitter voltage that the transistor will experience under these load-line conditions.

A $10\mu a$ dc base current has been selected as the operating current for this circuit. This operating point on the load line is called the "Q point." Variations in the base current about this Q point cause corresponding variations in the collector current and collector-to-emitter voltage. At this Q point the transistor has a 5ma dc collector current and a 3.4v dc collector-to-emitter voltage. When a $10\mu a$ signal is applied to the transistor's base at this Q point, a 3ma collector current signal and a 2.4v collector-to-emitter voltage signal results. Under these conditions the change in collector current is 300 times as large as the change in the base current. (This should not be confused with the value of h_{fe} since there is a load resistor in the circuit.) This current gain in the circuit is expressed as A_I.

Suppose the dc base current was increased to $22.5\mu a$. The Q point would then be higher on the load-line slope (Fig. 5). With this base current the transistor has a 6.8ma dc collector current and a 0.9v dc collector-to-emitter voltage.

The ratio of dc base currents and the dc collector currents is not the same at the two Q points. Because of the transistor's characteristics, the higher base currents are shown closer together.

There is a point where the load line passes through all of the remaining base-current curves. At that point on the graph there is no base current capable of increasing the collector current above 7.5ma.

The $10\mu a$ base-current signal applied to the higher Q point is the same as that used in Fig. 4. It appears smaller since the characteristic curves are closer together around this Q point. For simplicity this input signal is shown as a sine wave. It should appear more distorted, however, during the positive half cycle since the characteristic curves are much closer together in that area along the load line.

Since only 7.5ma of collector current can result from the base current signal, the output signal is "clipped." The $10\mu a$ base current signal has produced

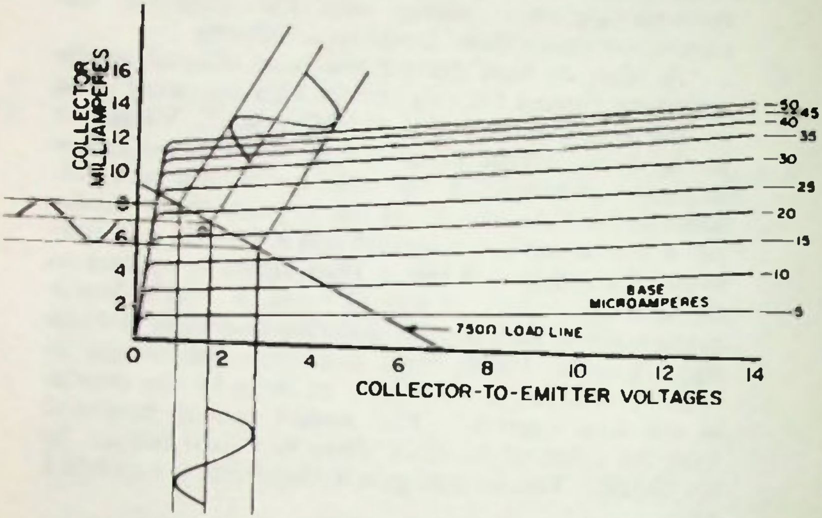

Fig. 6—Signal distortion is reduced by increasing the voltage source.

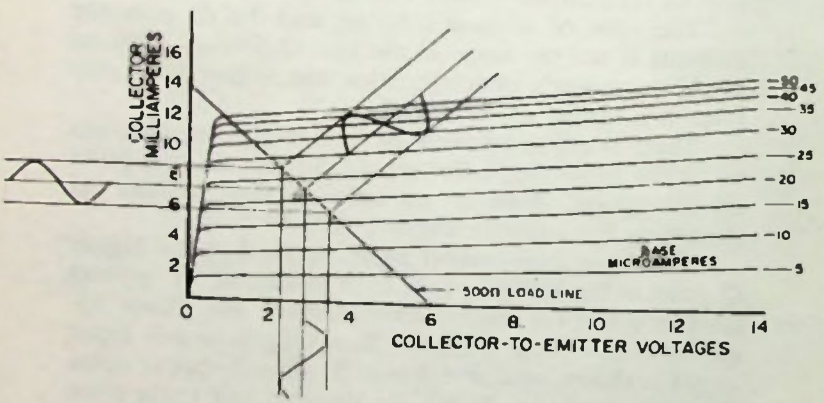

Fig. 7—Reducing the load resistance can improve the output signal.

a 2ma distorted collector current signal and a 1.4v distorted collector-to-emitter voltage signal. Under these conditions the change in collector current is only 200 times as large as the change in base current — this being the current gain (A_I) of the circuit.

The graphs indicate that by increasing the Q point, the transistor's current gain can be decreased. It also indicates that too high a Q point will result in distorting or clipping the output signal.

Changing A Circuit's Load Line

When the voltage source is increased to 7v while the load resistance remains 750Ω, the new load line is parallel and to the right of the old one (Fig. 6). With the new load line the 10μa signal at the same higher Q point will produce a collector current signal and a collector-to-emitter voltage signal that is not clipped.

If the load resistor is reduced to 500Ω the load line will become steeper (Fig. 7). If a 22.5μa dc base current is supplied to provide the same Q point as before, the 10μa base signal will not produce a distorted collector current signal or collector-to-emitter voltage signal.

Although the collector current shown in Fig. 7 has remained nearly the same as that in Fig. 6, the collector-to-emitter voltage has become smaller. This is despite the fact that the dc collector-to-emitter voltage is larger at the Q point in Fig. 7 than in Fig. 6.

Determining A Transistor's Base Voltage

When the voltage source (V_{CC}) in Fig. 2 is varied so that the collector-to-emitter voltage (V_{CE}) remains constant during changes of collector current (I_C), a graph can be made to show the relationship between the base current (I_B) and the base-to-emitter voltage (V_{BE}). The graph in Fig. 8 shows this relationship in the NPN transistor when the collector-to-emitter voltage is held at 2v. We can see from the graph that this relationship changes with temperature. We will deal with the transistor's characteristics at room temperature (25°C).

Since we know the base-to-emitter voltage and the corresponding base current, we can calculate the transistor's effective base-to-emitter resistance (r_b) for each point along the curve.

$$(r_b = \frac{V_{BE}}{I_B})$$

This resistance ranges from 230K at 0.23v and 75K at 0.3 to 11 ⅔K at 0.7v.

From the graph we can see that the transistor's effective base-to-emitter resistance (r_b) changes with the base-to-emitter voltage as well as with temperature.

When the base is conducting 22.5μa this transistor has an effective base-to-emitter resistance (r_b) of about

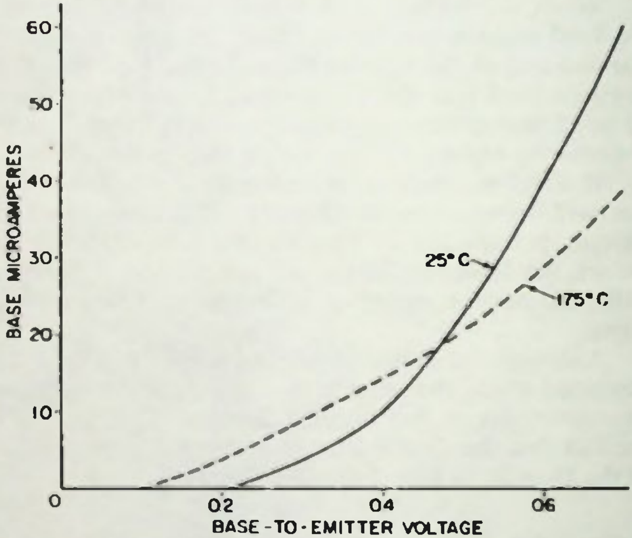

Fig. 8—There is a varying relationship between base current and base bias.

22K with a drop of about 0.5v. We can see from the curve that at 17.5μa the drop is 0.46v while at 27.5μa it is 0.53v. When the dc base current is 22.5μa, a 10μa base current signal corresponds to a 0.07v base-to-emitter voltage input signal. When the transistor has the same load line as shown in Fig. 6, this signal varies the collector-to-emitter output voltage from 1.0v to 2.8v. With this information we can calculate the voltage gain (A_v) of the circuit.

$$(A_v = \frac{V_{CE}}{V_{BE}} = \frac{1.8v}{0.07v} = 25.9)$$

Calculating Bias Circuits

The base resistor (R_B) shown in Fig. 9 is used to bias the transistor. The circuit requires a resistor (R_B) that will permit a 22.5μa base current (I_B) to pass through it from a 6v voltage source (V_{CC}). Since the base is to have a 0.5v forward bias, a 5.5v drop must be developed across the base resistor.

$$(I_B R_B = V_{CC} - V_{BE}.$$

$$R_B = \frac{6.0v - 0.5}{22.5\mu a} = 267K.)$$

It was previously indicated that the transistor's base could be biased more stabily with a voltage divider (Fig. 10). An infinite combination of values can be used for resistors R_B and R_1 that make up the voltage

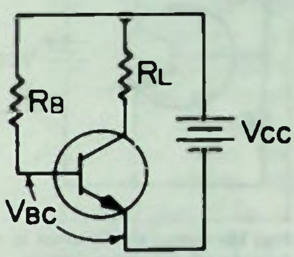

Fig. 9—The transistor is biased by a base resistor.

divider. If R_1 is too large however, the current flowing through R_B will be primarily dependent on the base current (I_B) and the circuit will be less stable, but if R_1 is too small it will short the incoming signal and reduce the gain.

If we let R_1 = 35K, it must conduct 14.3μa to develop a 0.5v base-to-emitter voltage.

$$(I_1 = \frac{V_{BC}}{R_1} = \frac{0.5v}{35K} = 14.3\mu a)$$

The current flowing through the base resistor (R_B) consists of the base current (I_B) plus the current (I_1) that flows through the other resistor in the voltage divider (R_1). Since there is to be a 0.5v base-to-emitter voltage, a 5.5v drop must be developed across the base resistor.

$[(I_B + I_1) R_B = V_{CC} - V_{BE}$.

$$R_B = \frac{6.0v - 0.5v}{22.5\mu a + 14.3\mu a} = \frac{5.5v}{36.8\mu a} = 150K.]$$

Transistor Standardization

At our present level of technology few transistors with the same code number have exactly the same characteristics. Manufacturers cannot afford to design each circuit manufactured to match the transistor used. Instead they sacrifice gain for the economy of being able to exchange transistors. This is done by using negative feedback circuits. A typical negative feedback

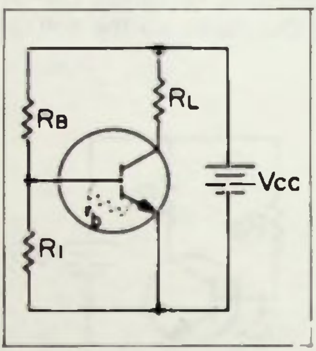

Fig. 10—Better stabilization is obtained with a voltage divider.

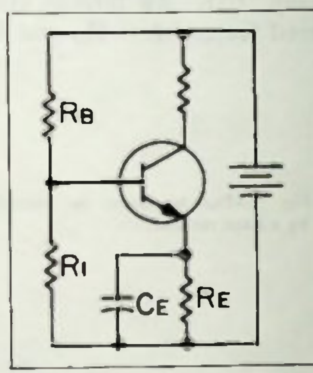

Fig. 11—A negative feedback circuit is used for additional stabilization.

circuit is shown in Fig. 11. If a transistor conducts a higher than average dc emitter current for a given base bias, the additional voltage drop across the emitter resistor (R_E) has the effect of reducing the base's forward bias which in turn reduces the emitter current. The ac signal in the emitter is not affected by this resistor but is shorted to ground by a capacitor (C_E).

Other circuits use the outputs of later stages as a negative feedback signal source for transistor stabilization. If a transistor has more gain than expected, it develops more negative feedback and its efficiency is lowered. If a transistor has less gain than expected it develops less negative feedback and operates at a higher efficiency. In this manner transistors with various characteristics can be used interchangeably in the same circuit.

CHAPTER 4

Frequency Limitations

Measurements show that a transistor's ability to amplify decreases as the signal frequency increases. The frequency at which the gain drops to 0.707 ($\frac{1}{2}\sqrt{2}$) its value at 1kHz is called the cutoff frequency. This is equivalent to a signal drop of 3db.

As previously indicated, the signal gain in a common-base circuit (h_{fb}) is the ratio of the emitter and collector current signals $\frac{(\Delta I_C)}{\Delta I_E}$. The cutoff frequency for this circuit is expressed as f_α or f_{hfb}.

The signal gain in a common emitter circuit (h_{fe}) is the ratio of the base and collector current signals $\frac{(\Delta I_C)}{\Delta I_B}$. In this circuit the cutoff frequency at which the gain has dropped 3db is expressed as f_{beta} or f_{hfe}.

The effect a signal's frequency has on a transistor's gain is shown in Fig. 1. The common-base cutoff frequency (f_{hfb}) is higher than the common-emitter cutoff frequency (f_{hfe}) and is more often used to describe the characteristics of a transistor.

The gain-bandwidth product is the most commonly used measure of a transistor's frequency response. It is the frequency (f_T) at which the common-emitter signal gain (h_{fe}) equals one. This frequency is usually determined by measuring the current gain near f_{hfe} and extrapolating at —6db per octave. When logarithmic scales are used in a graph as shown in Fig. 1 to represent gain and frequency, a —6db per-octave slope is represented by a 45deg. angle. A line was drawn from the value of h_{fo} at f_{hfe} at a 45deg. angle to extrapolate the value of f_T.

Today's transistors have an f_{hfe}, f_T and f_{hfb} frequency response ranging from 0.12, 12 and 18MHz, respectively, for germanium alloy transistors to 50MHz, through 1.7 and 2.5GHz for silicon diffused transistors.

Two Factors Limit High Frequency Response

The base in an NPN transistor circuit (Fig. 2) is more positive than the emitter and, therefore, the base-to-emitter junction is forward biased. In this circuit the base is less positive than the collector and, therefore, the base-to-collector junction is reverse biased. (The same bias conditions occur in a PNP transistor circuit containing voltages of opposite polarity.) Reverse biased semiconductor junctions develop a "diffusion capacity" that is directly proportional to the area of the junction and inversely proportional to the square root or cube root of the voltage applied.

Since only the base-to-collector junction is reverse biased, it is this junction that contains a noticeable capacitance. This capacitance is represented by C_{bc} and generally ranges from 1 to 50pf.

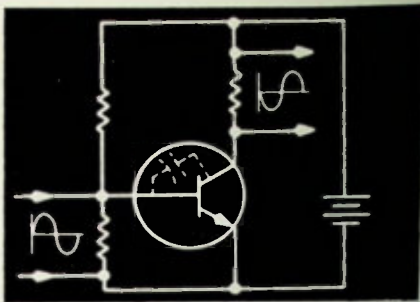

Fig. 3—Transistor gain is reduced by the transistor's internal capacitance.

The base-to-collector capacitance has an adverse effect on the frequency response of the transistor circuit shown in Fig. 3. The input in a common emitter circuit is normally inverted at the output. The capacitance (C_{be}) couples this output with the input. Since the output signal is the inverse of the input signal, it reduces the input signal (negative feedback), reducing the circuit's gain.

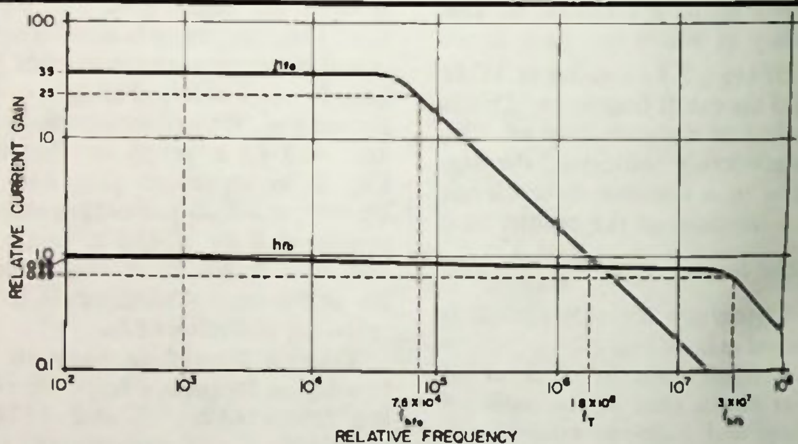

Fig. 1—Transistor gain is reduced at higher frequencies.

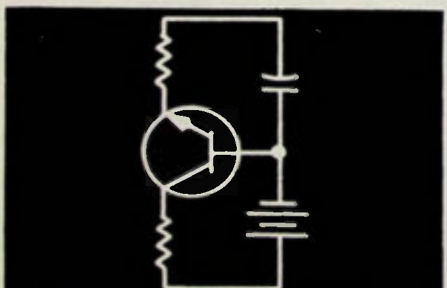

Fig. 2—Base-to-emitter junctions are forward biased while base-to-collector junctions are reverse biased.

Capacitor (C_{bc}) provides a signal opposing the input signal and more base current (I_B) is required to pro-

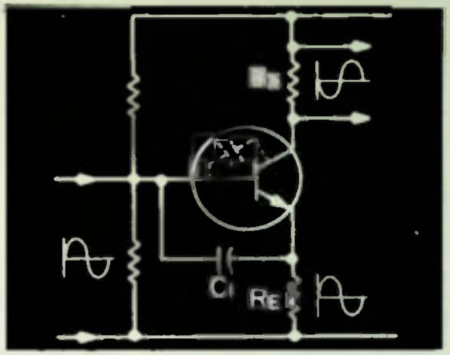

Fig. 4—Transistor internal capacitance can be neutralized by the circuit.

to "neutralize" the effect of the transistor's internal capacitance (C_{bc}) by providing an equal amount of positive feedback. The output signal present in the emitter resistor (R_E) is in phase with the input signal and can be a source of positive feedback. This signal is fed to the base by capacitor C_1.

It has been indicated that the base-to-collector capacitance affects the transistor's frequency response, and that the amount of this capaci-

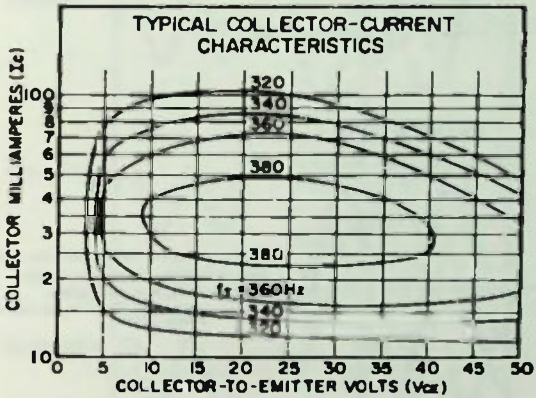

Fig. 5—Voltage and current combinations can affect a transistor's frequency response.

vide an output signal of the same amplitude. This condition has the effect of decreasing the transistor's base-to-emitter impedance.

The feedback resulting from the transistor's internal capacitance (C_{bc}) is called the "Miller Effect." This effect can be reduced by lowering the circuit's input impedance to match the lower impedance of the transistor's base.

The circuit in Fig. 4 is designed

tance (C_{bc}) is dependent on the reverse bias of the base-to-collector junction. For this reason, the transistor's gain bandwidth varies with the collector-to-emitter voltage (V_{CE}) and collector current (I_C). The curves shown in Fig. 5 indicate how voltage and current combinations increase or decrease a transistor's frequency response.

Transistors require a finite time for the current to pass across the base from the emitter to the collector. This "transit time" is dependent on the thickness of the base region. A typical germanium alloy transistor has a 0.001 in. base, a 0.002sec transit time and a 3MHz f_{hfb} cutoff frequency. A silicon diffused transistor may have a 0.0001in. base thickness and a f_{hfb} cutoff frequency of 150MHz.

The delay that a signal experiences when passing through the base results in a phase shift. The higher the signal's frequency, the greater the phase shift resulting from this delay. Slight variations in the distance that signals must travel through the base result in varying phase shifts as shown in Fig. 6. A single base signal can produce out-of-phase collector signals. At lower frequencies a smaller dispersion of phase shifts occur and the resulting total collector signal is at maximum amplitude. Higher frequencies develop a greater dispersion of phase shifts, resulting in a smaller total collector signal.

The transit time effect in transistors does not limit their frequency response as severely as the feedback caused by internal capacitance.

Transistors May Oscillate At High Frequencies

The input signal in the common-emitter circuit shown in Fig. 3 was inverted at the output. As the frequency increases the resulting output signal, because of transit time, experiences a phase shift. This phase shift is illustrated in Fig. 7. At low frequencies (Frequency 1) the output signal has no phase shift and is merely an inverted form of the input signal. At a higher frequency (Frequency 2) some phase shift has occurred, while a still higher frequency (Frequency 3) results in a phase shift so large that the output signal is in phase with the input signal.

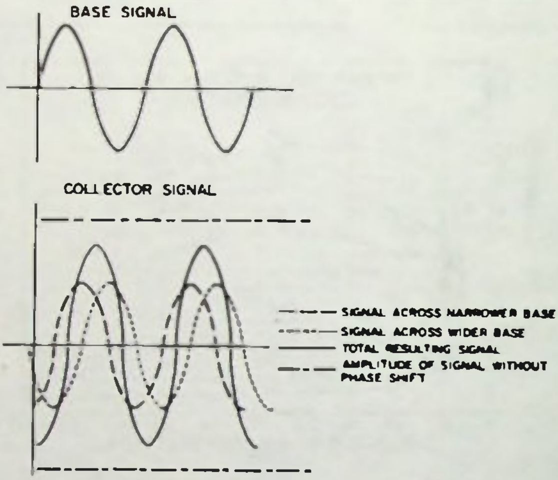

Fig. 6—Varying phase shifts can reduce the collector signal.

The circuit in Fig. 8 illustrates the effect of phase shift. We see that a lower output frequency (f_1), results in an inverted form of the input signal. The transistor's internal capacitance (C_{bc}) then produces negative feeedback and the output signal's amplitude is reduced. At a higher frequency (f_3), the output signal is in phase with the input signal. Under these conditions the transistor's internal capacitance (C_{bc}) produces positive feedback and the transistor can go into oscillation.

The positive feedback that develops at higher frequencies supplies some internal base-to-emitter current (I_{BE}). Less external base current (I_B) is required to produce an output signal of the same amplitude. Hence, the effective base-to-emitter impedance has increased.

Even if spurious oscillations do

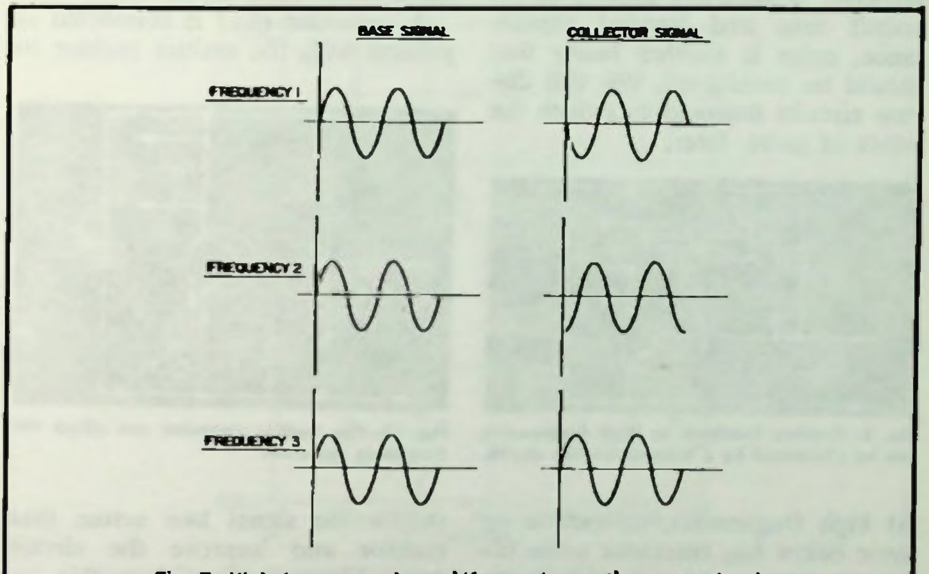

Fig. 7—High frequency phase shifts can invert the output signal.

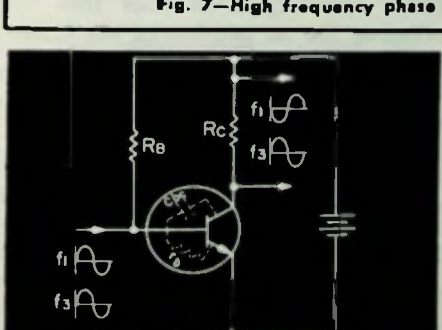

Fig. 8—Transistors tend to oscillate at high frequencies.

not occur, positive feedback can couple the input and output tuned circuits—making them interdependent. A variation in the output circuit will affect the input circuit and vice versa. A negative-feedback process called "unilateralisation" is used to eliminate this postive-feedback problem. (This should not be confused with neutralization circuits designed to remove the effect of negative feedback.) A transformer

(T) can be used (Fig. 9) to invert the output signal. This inverted signal can then pass through a capacitor (C_1) to provide negative feedback equal in amplitude to the transistor's positive feedback. Since the negative feedback cancels the positive feedback, the input and output circuits can be tuned independently.

Although a transistor's frequency response is generally limited by its transit time and internal capacitance, noise is another factor that should be considered. We will discuss circuits designed to reduce the effect of noise later.

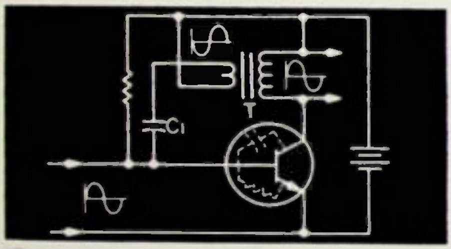

Fig. 9—Positive feedback at high frequencies can be eliminated by a unilateralisation circuit.

At high frequencies, an octave or more below f_{hfb}, transistor noise increases sharply at a rate of 6db per octave. The strongest concentrations of noise are observed around $\frac{1}{2} f_{hfb}$.

Capacitors and Resistors Limit Low Frequencies

Transistors are generally the components that limit a circuit's high frequency response, while resistors and capacitors generally limit a circuit's low frequency response.

Earlier it was indicated that an emitter resistor (R_E) is frequently included in the transistor circuit (Fig. 10) for transistor stabilization. As the transistor warms and conducts more emitter current (I_E), the dc voltage drop ($I_E R_E$) across the emitter resistor increases and the emitter becomes more positive, resulting in a smaller emitter-to-base voltage (V_{EB}). Since a reduction occurs in the forward bias of the transistor's base, the transistor conducts less current and the circuit is stabilized.

A capacitor (C_1) is connected in parallel with the emitter resistor to

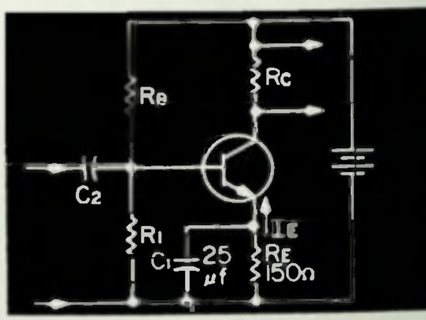

Fig. 10—The emitter capacitor can affect the frequency response.

reduce the signal loss across that resistor and improve the circuit gain. The capacitor does this by shorting the ac portion of the emitter current around the resistor.

Ohm's law can be applied to capacitors as well as resistors. The amount of current (ac or dc) flowing through a resistor is equal to the voltage across the resistor divided by the resistor's value ($I = \frac{E}{R}$). The amount of ac current flowing through a capacitor is equal to the voltage across the capacitor divided by the capacitor's reactance

$(I = \frac{E}{XC})$. The value of a capacitor's reactance is dependent on the value of the capacitor (C) and the frequency (f) of the current flowing through it. This relationship is shown in the following equation:

$$X_C = \frac{1}{2\pi fC} \approx \frac{0.159}{fC}.$$

From the equation we can see that the larger the capacitor's value (C) and the higher the signal's frequency

When the frequency drops to 424Hz the capacitor's impedance equals the value of the emitter resistor. At lower frequencies the emitter resistor is smaller in value than the capacitor's reactance, and more signal current passes through it than through the capacitor.

Since the signal loss increases as the capacitor's reactance increases, the gain in the circuit decreases at lower frequencies.

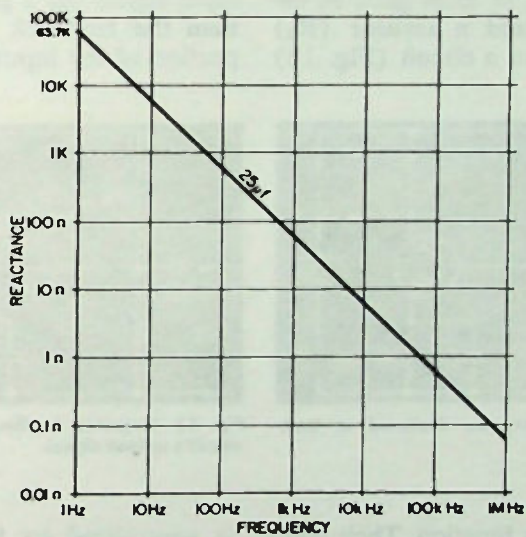

Fig. 11—Capacitor reactance varies with frequency.

(f), the smaller the reactance. The relationship of frequency to reactance is shown in Fig. 11.

The amount of signal current that the 25μf capacitor (C_1) shunts around the 150Ω emitter resistor (R_E) in the circuit (Fig. 10) is dependent on the frequency of the signal. From the graph (Fig. 11) we can see that at 100kHz the capacitor's impedance is 0.637Ω while at 1kHz it has increased to 63.7Ω.

The value of the coupling capacitor (C_2) also has an effect on the circuit's frequency response. As the signal frequency decreases, a point is reached where the circuit's gain starts to decrease because of the coupling capacitor's increasing impedance. At higher impedances less signal current passes through the capacitor. At the frequency where the reactance of the coupling capacitor equals the total resistance in series with it, the circuit's gain will drop —3db.

The lower frequency response of a transistor circuit can be improved by increasing the value of the capacitor to increase its reactance, and with transistors that permit using resistors having values larger than the capacitor's reactance at low frequencies.

Modifications can increase the low frequency response of a circuit at the expense of total gain. A capacitor (C_3) and a resistor (R_2) can be added to a circuit (Fig. 12) used to improve a circuit's low frequency response. One method for obtaining such feedback is shown in Fig. 13. Here a portion of the output signal passes through a capacitor (C_4) and resistor (R_3) and is returned to the input. Since a signal is normally inverted when it is amplified in a common emitter circuit, the returned output signal reduces the amplitude of the input signal. The circuit is designed so the input signal has a greater amplitude than the feedback signal, and the portion of the input signal fed back

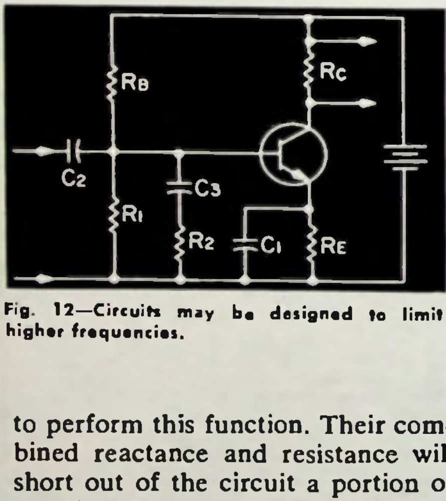

Fig. 12—Circuits may be designed to limit higher frequencies.

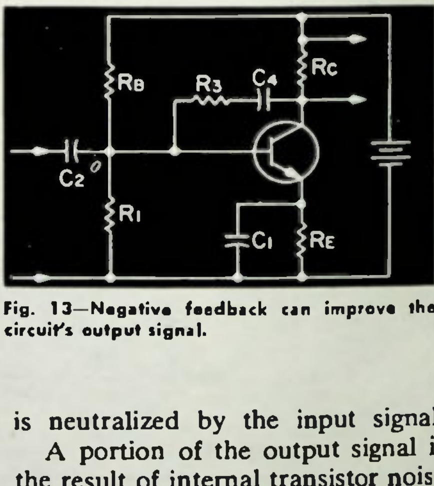

Fig. 13—Negative feedback can improve the circuit's output signal.

to perform this function. Their combined reactance and resistance will short out of the circuit a portion of the high frequency signal. When the remaining high frequency signal is amplified it will more nearly equal the amplified low frequency signal that had been reduced by the other circuit capacitors (C_1 and C_2).

Negative Feedback Circuits

We have seen how negative feedback can be used to improve the high frequency response of a transistor circuit. It can also be is neutralized by the input signal.

A portion of the output signal is the result of internal transistor noise and distortion. This portion of the signal fed back is not the inverse of the input signal and is not neutralized by it. This distortion and noise remains to be fed into the base of the transistor and amplified. When amplified, the noise and distortion signals are inverted and serve to partly neutralize the original noise and distortion in the circuit's output.

Negative feedback can also be produced in the circuit shown in Fig.

10 by removing capacitor C_1. The signal voltage ($\Delta I_E R_E$) developed across the emitter resistor (R_E) is the inverse of the base input signal voltage. It provides negative feedback by reducing the total emitter-to-base voltage (V_{EB}) developed by the input signal.

CHAPTER 5

Field-Effect Transistors

Although junction field effect transistors are also made of P- and N-type material, their design permits an entirely different set of characteristics.

Controlling The Drain's Current

The FET shown in Fig. 1 contains a rod of P-type material capable of conducting current in either direction. The chemistry in portions of the rod has been changed and made into N-type material. The resulting junctions of P and N materials are like the junction in a diode. Since current cannot flow through these junctions when the N-type material is more positive than the P-type material, the current traveling through the rod must pass around them.

The larger the positive voltage applied to the N-type material in junction FETs, the larger the effective size of the junctions, and the smaller the rod's remaining cross-sectional area capable of current flow. As the area permitting current flow diminishes, the resistance of the rod increases and less current is able to flow through the rod.

The portion of the junction FET that controls current flow is called the gate. Voltage is applied to the gate to "pinch-off" the current passing through the rod. Since the gate of a junction FET is biased to reduce current flow through the junction and therefore through the rod, we say that it is reverse biased. This is quite unlike the regular transistor, which has a forward biased base to encourage current flow through the junction and transistor.

Current flows from one end of the P-type rod to the other. The source is where the current enters the rod while the drain is where the current leaves the rod. Since the lead connected to the source is attached in the same manner as the lead connected to the drain, no difference exists between the two ends. If the source and drain leads of a junction FET are interchanged in a circuit, current will flow through the rod in the opposite direction — the drain has become the source while the source has become the drain. Hence, the source and drain of a junction FET are interchangeable. This characteristic also differs from a regular transistor, which has an emitter and collector not normally interchangeable.

Biasing The Gate

A simplified circuit for a P-channel FET is shown in Fig. 2. Since the gate is made of N-type material, a positive potential must be applied to it to regulate the semiconductor device. Because currents flow from negatively to positively charged potentials, the little gate current that does exist flows from the voltage source (V_{GG}) into the gate. The direction of the arrow used to represent the gate points in a direction opposite that of current flow. (The base-to-emitter junction to encourage current flow.

The effective size of the FET's P-N junctions can be partially reduced when the gate is forward biased, and even more current can then flow from source to drain. Silicon junction FETs have a high gate resistance even when the gate is forward biased up to about 0.5v.

It is practical to use a junction FET with this characteristic to operate small-signal amplifiers with a

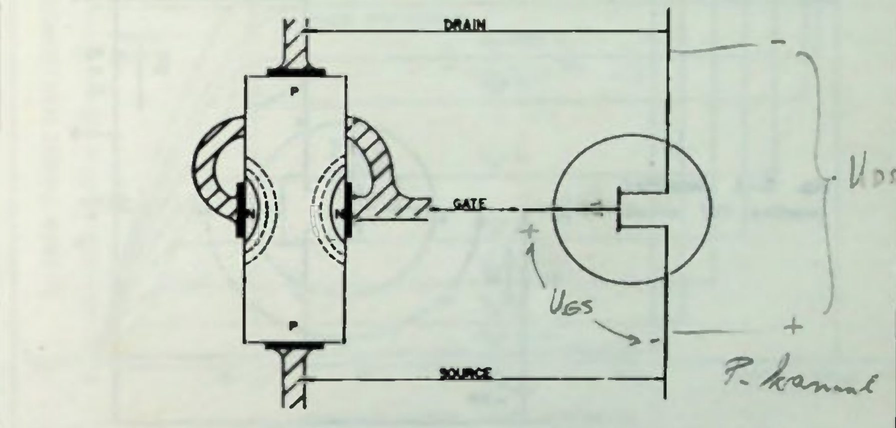

Fig. 1—The structure of a junction field effect transistor.

arrow points in the direction of "conventional" current flow.)

The P-N junctions bordering the gate of a FET are reverse biased to reduce current flow through the junctions and the rod. Since the junctions are reverse biased, their resistance is high and very little current flows through the gate lead. In silicon junction FETs the gate resistance can be 10M or higher.

This FET characteristic differs sharply from that of a regular transistor, which has a forward biased zero-biased gate, if the peak input signal levels off at about 0.5v. Above this forward bias voltage the P-N junction current increases exponentially. Gate signal peaks are clipped with the sudden drop in gate resistance at higher forward bias voltages. To eliminate this effect, the signal's forward bias must either be reduced by lowering the circuit's input impedance, or shifted by modifying the circuit so that it contains a dc reverse bias to counteract the signal's forward bias.

"Pinch-Off" Voltages

Measurements show that when the gate-to-source (V_G) is held constant and an increasing drain-to-source voltage (V_{DS}) is applied to a FET (Fig. 3), there will be a corresponding increase in drain current (I_D) until a certain voltage is reached. Further increases in drain-to-source voltage will not result in any significant increase in drain current until the semiconductor's breakdown voltage (BV_{DS}) is breakdown voltage (BV_{DS}) is called the pinch-off region. (Pinch-off region is where $-V_P < V_{DS} < BV_{DS}$.)

The drain on-current ($I_{D(ON)}$) is defined as the drain current (I_D) that occurs when there is zero gate-to-source voltage ($V_G = O$) and the drain-to-source voltage (V_{DS}) is in the pinch-off region. ($I_{D(ON)} = I_D$ when $V_G = O$, and $-V_P < V_{DS} < BV_{DS}$.)

As the gate-to-source voltage (V_G) in a junction FET increases,

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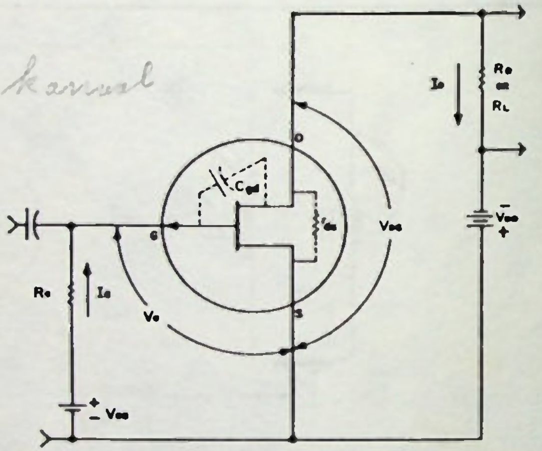


Fig. 2—A simplified junction FET circuit.

reached. The drain-to-source voltage, at which there is no corresponding increase in drain current, is dependent on the gate-to-source voltage. The pinch-off voltage (V_P) has been defined as approximately the negative of the drain-to-source voltage (V_{DS}) for maximum current, when there is no gate-to-source voltage ($V_G = O$). ($V_P \approx -V_{DS}$ when $V_G = O$, and I_D first becomes independent of V_{DS}.)

The range of drain-to-source voltages (V_{DS}) between the negative of the pinch-off voltage ($-V_P$) and the the drain current (I_D) decreases (Fig. 3). The pinch-off voltage (V_P) has also been defined as approximately the gate-to-source voltage (V_G) reverse bias for minimum drain current ($I_{D(OFF)}$). At this gate-to-source voltage the reduced drain current is nearly independent of the drain-to-source voltage. ($V_P \approx V_G$ when $I_D = I_{D(off)}$, and I_D is nearly independent of V_{DS}.)

The characteristic drain-to-source pinch-off voltage at which nearly maximum drain current occurs is nearly equal in magnitude to the characteristic gate-to-source pinch-off voltage at which nearly minimum drain current occurs. Almost the only difference between the two voltages is their polarity.

Junction FET Amplification

One of the specifications for the gain in a regular transistor refers to the ratio of collector and base current signals when the transistor is connected directly to a constant voltage source. This ratio of signal currents, frequently expressed as β or h_{fe}, is not an appropriate measure for describing junction FETs. Since a junction FET has a very high gate resistance, the input signal's current at the gate is so small that a ratio of gate signal current to drain signal current would not be a practical unit of measure.

Junction FET specifications compare the drain current signal (Δi_D), when the source and drain are directly connected to a constant voltage source, with the gate-to-source signal voltage (Δv_G). This ratio ($\frac{\Delta i_D}{\Delta v_G}$) is called transconductance or mutual-conductance, represented as g_m, and has a value generally ranging from 500 to 10,000μmhos. (The unit "mho," as we know, is the inverse of resistance, and is ohm spelled backward.)

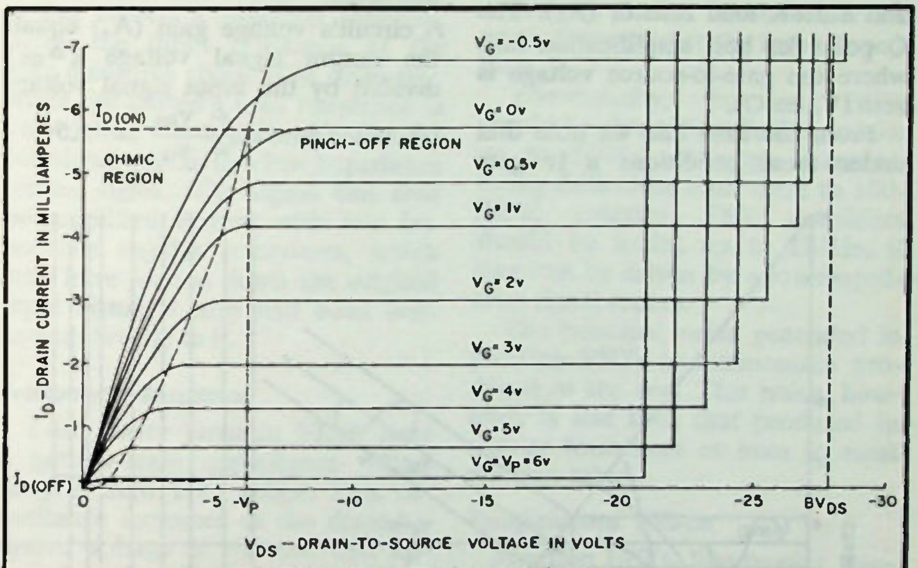


Fig. 3—Junction FET characteristic curves.

If the drain-to-source resistance (r_{ds}) of the FET, shown in Fig. 2, is much greater than the load resistance (R_L), the approximate value of the circuit's voltage gain (A_v) can be easily calculated ($A_v \approx g_m R_L$).

Small signal junction FET amplifiers have a maximum voltage gain (A_v) with a zero gate-to-source bias voltage (V_G) and a drain-to-source voltage (V_{DS}) slightly above the pinch-off voltage (V_P). (A_v = maximum when $V_G = O$, and $V_{DS} > -V_P$.)

Determining A Load Line

A load line can be drawn (Fig. 4) on the FET's characteristic curve (Fig. 3) so it crosses the "$V_G = O$" line near the lower end of the pinch-off region. In this case, we have selected a 20v source (V_{DD}) and a 1.8K load resistor (R_L). The Q point for best amplification falls where the gate-to-source voltage is zero ($V_G = O$).

From the load line we note that under these conditions a 1v gate signal (Δv_G) develops a 0.8ma drain current signal (Δi_D) and a 1.5v drain-to-source voltage signal (Δ_{DS}).

With this information we can calculate the junction FET's transconductance (g_m). ($g_m = \frac{\Delta\ i_D}{\Delta\ v_G} = \frac{0.8 \times 10^{-3}a}{1v} = \frac{1v}{800 \times 10^{-6}a} =$ 800μmho.

A circuit's voltage gain (A_v) equals the output signal voltage (Δ_{DS}) divided by the input signal voltage (Δ v_G). ($A_v = \frac{\Delta\ v_{DS}}{\Delta\ v_G} =$ 1.5v

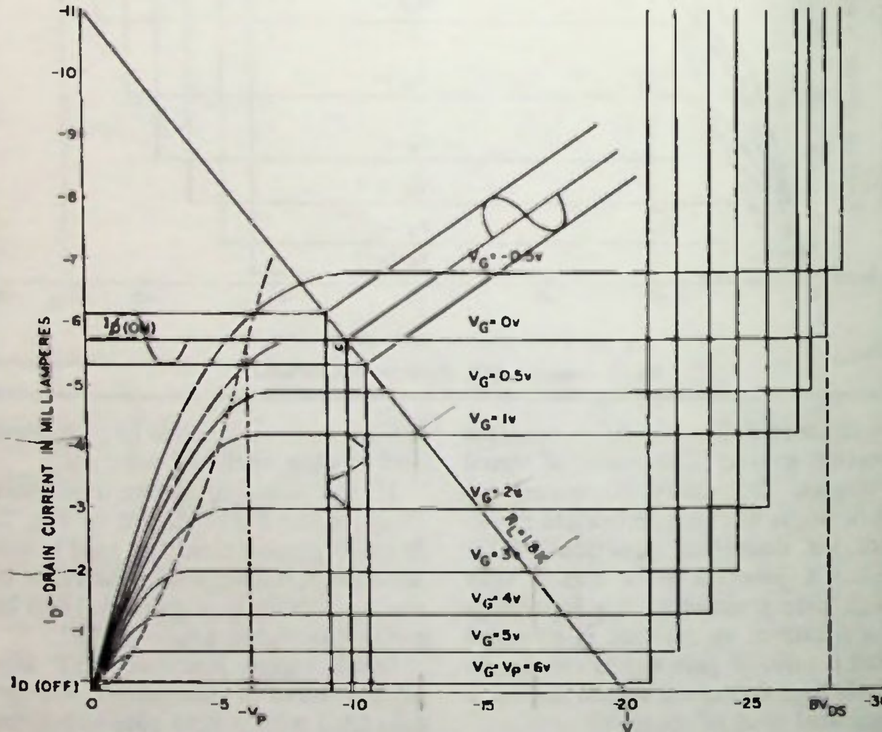


Fig. 4—The load line of a junction FET circuit. (8-kamal)

We can use the other voltage gain equation to verify our calculations. $A_v = 1.5 \approx g_m R_L = 800\mu\text{mho} \times 1.8K = 800 \times 10^{-6}\ \text{mho} \times 1.8 \times 10^3\Omega = 1.44$. The slight discrepancy in this case results from a small error in reading the load line and because the load resistance is not exactly 1.8K.

Although the circuit has developed very little voltage gain (A_v), it has served a practical function. Since the input resistance at the gate measures about 10M or higher, while the circuit's load resistance is only 1.8K, the high impedance input signal results in a low impedance output signal. The signal can now be amplified further with low impedance regular transistors, which may have loaded down the original input signal if they had been connected directly to it.

Frequency Response

Low power junction FETs have a gate-to-drain capacitance (C_{gd}) ranging from 1 to 100pf. This capacitance increases as the drain-to-source voltage is reduced and approaches the pinch-off voltage. As a result of this capacitance (C_{gd}), the "Miller effect" occurs in junction FETs.

A regular transistor's low impedance base input tends to shunt a large portion of the transistor's high impedance internal feedback to the emitter. The Miller effect in regular transistors is, therefore, not as great as in FETs where the input signal at the gate, like the internal feedback signal, is usually also of high impedance.

The Miller effect in FETs can be reduced by increasing the drain-to-source voltage, reducing the circuit's input impedance or with a feedback circuit opposing the semiconductor's internal feedback.

When the voltage gain of a FET circuit is limited to unity, the semiconductor's internal capacitance is large enough to limit the circuit's input impedance to between 1 and 50M. With higher gain the Miller effect still further limits either the upper frequency or the input impedance.

Theoretically, junction FETs should be able to handle frequencies up to 1GHz. Although manufacturing difficulties limit them to 100-MHz, junction FET amplifiers should be useful up to 1MHz, if they can be driven by a low impedance signal source.

The principal noise generated in junction FETs is thermonoise produced in the rod. This noise, however, is less than that produced in regular transistors or even in most electron tubes.

Temperature Effects

Junction FETs, like regular transistors, are normally affected by changes in temperature. Measurements indicate that their pinch-off voltage (V_P) usually increases at about 0.2 percent per 8°C.

Although FET N-P junctions are affected by temperatures, the gate current stability of junction, silicon FETs approaches the base current stability of regular, silicon transistors when the circuit's gate-to-source resistance (R_G) lies between 1 and 10M.

Two opposing effects of temperature change in junction FETs result in a linear decrease in drain on-current ($I_{D(ON)}$) with increasing temperature, causing the current to decrease about 0.6 percent per °C. This drift can be reduced by operating the FET at very low Q-point currents.

The use of junction FETs with lower pinch-off voltages also reduces drain current drift. Measurements show that FETs having a 0.63v pinch-off voltage experience no temperature drift when operating at their normal drain on-current ($I_{D(ON)}$) and their normal transconductance (g_m).

Junction FET Standardization

It was indicated previously that few of the regular transistors with the same code number have exactly the same characteristics, and that by using negative feedback circuits the circuit's gain was sacrificed for the economy of being able to exchange transistors. The same problem exists with junction FETs for many FETs with the same code number had drain on-currents ($I_{D(ON)}$) and transconductances (g_m) that vary by as much as 2-to-1.

CHAPTER 6

MOS FETs

We have seen that many characteristics of the junction FET are superior to those of regular transistors. Many of the characteristics of still another type of semiconductor are even superior to these. This new component is called a depletion-type, insulated-gate, field-effect transistor, and is composed of metal-oxide, semiconductor material. For simplicity we will refer to it as a depletion-type MOS.

Composition of a Depletion-type MOS

The depletion-type MOS, like other semiconductors, contains P- and N-type material. The arrangement of this material is shown in Fig. 1. A slice of P-type silicon material, called substrate, contains two pockets of N-type material located about 5 x 10^{-4}in. (0.0005in.) apart. They are joined by a thin channel, which is also of N-type material. Leads are secured to each pocket, and current can flow from one pocket to the other, through the channel, in the same manner as it flows through the junction FET's rod of P-type material. The pocket in which the current enters is called the *source* while the pocket that the current leaves is called the *drain*.

Unless the N-type material is negatively charged with respect to the P-type substrate, the current traveling from source to drain is restricted to the N-type material by the P-N junctions along its path. Since the current is traveling through only the N-type material, the depletion-type MOS' source and drain leads can be interchanged in a circuit and current will flow through the N-type material in the opposite direction — the drain has become the source while the source has become the drain.

A 4 x 10^{-6}in. (0.000004in.) layer of silicon dioxide forms nearly a perfect insulation between the channel and the gate lead. The electric field, resulting from a charge applied to the gate lead, has the effect of altering the channel's thickness. When a negative charge is applied to the gate lead, the channel becomes thicker and is able to conduct more current. When a positive charge is applied to the gate lead, the channel becomes thinner and is able to conduct less current.

If the depletion-type MOS contained an N-type substrate with pockets and channel of P-type material, positive charges applied to the gate lead would enable the semiconductor to conduct more current while negative charges would restrict current flow.

The arrow shown in Fig. 1 represents the substrate lead. The direction it is pointing indicates that the substrate is of P-type material. If the arrow pointed in the opposite direction, it would indicate that the substrate was of N-type material. For low frequency use the substrate lead and source lead are normally connected together.

Gate Resistances

Unlike regular transistors and junction FETs, the depletion-type MOS has a gate current independent of forward bias. The gate resistance may exceed $10^{15}\Omega$ while its capacitance may be less than 1pf. With these characteristics, the gate is capable of holding a charge for several hours when out of a circuit. It has been claimed that some have even been able to hold their gate charge for days.

Although the gate resistance is very high, the oxide coating, insulating the gate from the channel, is so thin that it can be easily punctured by the application of excessive external voltages as low as 20 to 30v. For this reason it is very important that MOS components be handled with extreme care. A few feet of insulated wire, charged to twice the gate's breakdown voltage, could possibly destroy the gate insulation. Static charges generated by a technician may be sufficient to puncture the insulation and ruin the MOS. For this reason it is advisable to work with a grounded soldering iron, and guard against static

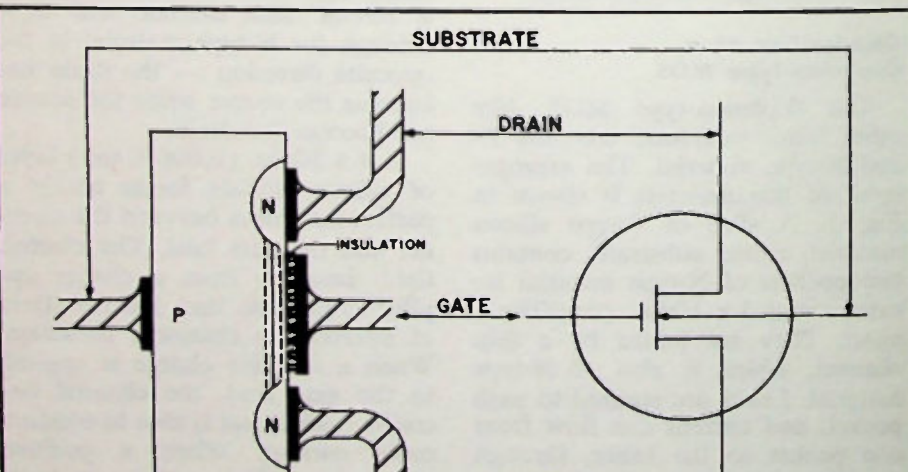


Fig. 1—The structure of a depletion-type MOS.

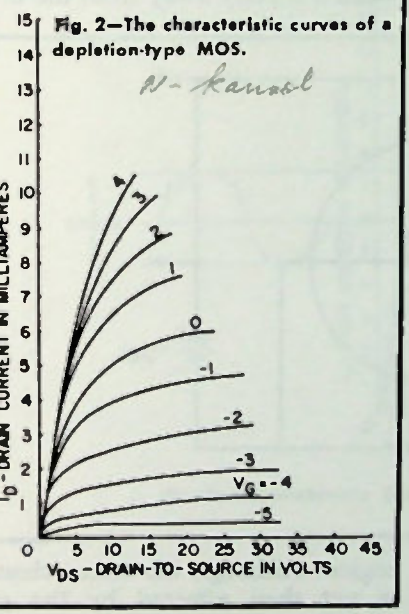
N- [illegible]

charges from test instruments when working with MOS circuits.

Depletion-type MOS Characteristic Curves

The characteristic curves of a depletion-type MOS is shown in Fig. 2. We can see, for this particular semiconductor, a gate charged more negatively than the source results in a reduction of the drain current while a gate charged more positively than the source increases the drain current. The gate voltage curves in Fig. 2 are further apart in the area where the gate voltage is zero and closer together at larger positive or negative potentials. This indicates that when a signal is applied to a gate with zero bias, the gain will be greater than when a signal is applied to a gate with a positive or negative bias.

Fig. 2—The characteristic curves of a depletion-type MOS.



A Depletion-type MOS Common-source Circuit

A common-source circuit (Fig. 3) can be designed for a depletion-type MOS with the characteristics shown in Fig. 2. In this circuit, the gate resistor (R_G) merely serves to prevent static charges from breaking through the MOS' insulated gate signal to the desired amplitude. This resistor does not bias the gate of the semiconductor.

The voltage source (V_{DD}) and load resistor (R_L) for this circuit can be determined by drawing a load line (Fig. 4) similar to those shown earlier.

This line falls across the curves in such a manner that a gate voltage signal (Δv_G) produces a maximum drain-to-source voltage signal (Δv_{DS}) without clipping the signal or overloading the semiconductor. This load line represents a load resistance

(R_L) of 2K $\left(\frac{30v}{15ma}\right)$

and a voltage source of (V_{DD}) of 30v.

A 2v input signal (Δv_G) is shown (Fig. 4) to result in a 5.5v output signal voltage (Δv_{DS}) and 2.75ma output signal current (Δi_D). From these figures we can calculate the circuit's voltage gain

$$\left(A_v = \frac{\Delta v_{DS}}{\Delta v_G} = \frac{5.5v}{2v} = 2.75\right).$$

Since this semiconductor has an insulated gate, the gate current is insignificant. The very small input

signal current $\left(\frac{\Delta v_G}{R_G}\right)$ results in a significant output signal current (Δi_D). This current gain (A_I) is too large to be a practical unit of measure.

Composition of an Enhancement-type MOS

The enhancement-type MOS is still another kind of insulated gate semiconductor. Its arrangement of P- and N-type material (Fig. 5) is nearly the same as that of the depletion-type MOS (Fig. 1). The depletion-type MOS' substrate of P-type material contains two pockets of N-type material joined by a thin channel of N-type material. When a negative charge is applied to the gate lead, the channel becomes thicker and is able to conduct more current. The enhancement-type MOS, however, contains no channel of N-type materials. The two pockets are separated by the substrate of P-type material. When a positive charge is applied to the gate lead, a portion of the substrate changes from P-type to N-type material and a channel of N-type material is formed.

The entire region between the source and the drain of the enhancement-type MOS must be covered by the gate electrode and insulation. The channel would not otherwise reach from source to drain pockets. Since the depletion-type MOS contains a permanent channel, the gate electrode and insulation can be located slightly away from the drain region. Although the entire channel is not then affected by the gate electrode, the semiconductor benefits from a substantial reduction in feedback capacitance between the drain pocket and the gate electrode. The drain-to-gate capacitance of the enhancement-type MOS is, therefore, larger than that of the depletion-type MOS.

The drain and source leads of the enhancement-type MOS can also be interchanged, and its gate resistance is of similar value.

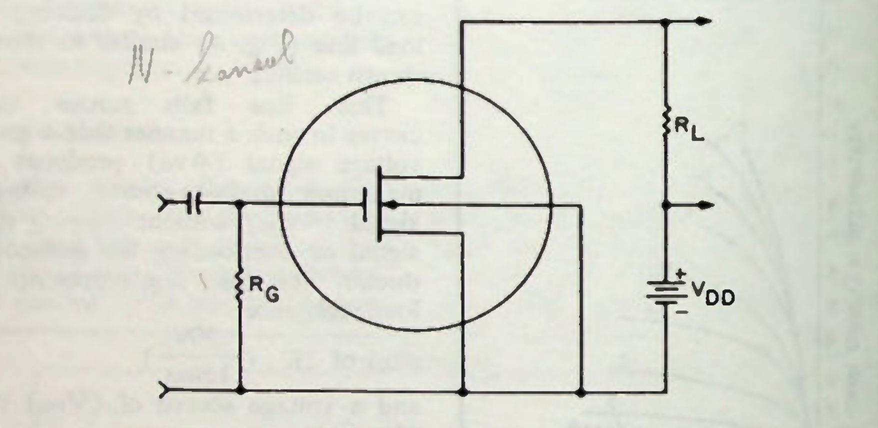


Fig. 3—A depletion-type MOS common-source circuit.

The arrow shown in Fig. 5 represents the substrate lead. The direction it is pointing indicates that the substrate is of P-type material. If the arrow pointed in the opposite direction, it would indicate that the substrate was of N-type material. For low frequency use the substrate to encourage current flow. The basic difference in the biasing of these two semiconductors results from the difference in the input current that they require. The greater base current required by a regular transistor limits its biasing resistances to values given in terms

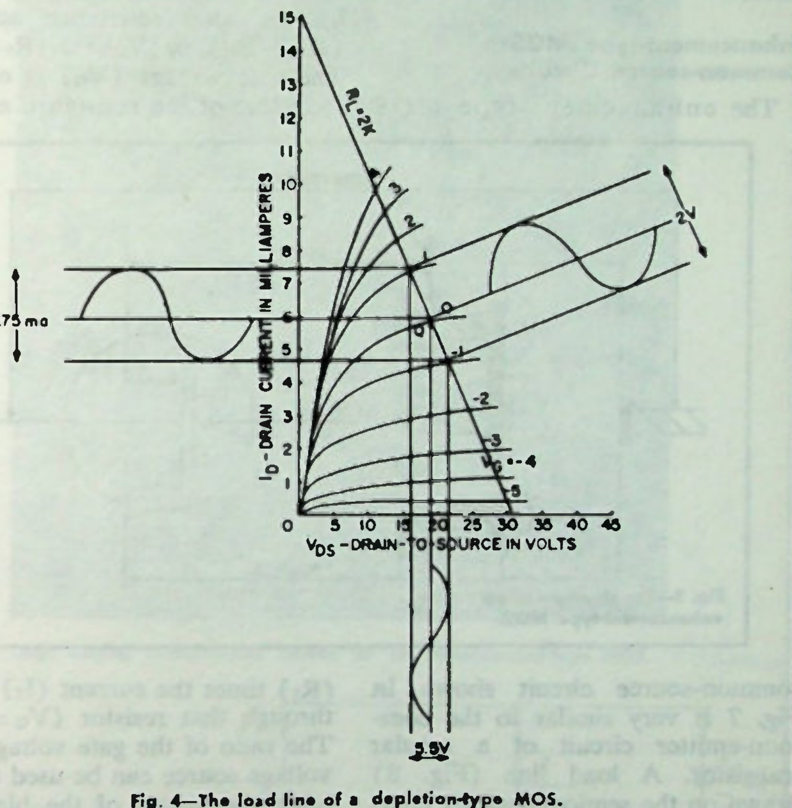


Fig. 4—The load line of a depletion-type MOS.

lead and source lead are normally connected together.

Enhancement-type MOS Characteristic Curves

The gate of an enhancement-type MOS, like the base of a regular transistor, must be forward biased of thousands of ohms (K), while the negligible gate current required by an enhancement-type MOS permits biasing resistances with values given in terms of millions of ohms (M).

The characteristic curves of an

enhancement-type MOS are shown in Fig. 6. The curves for higher bias voltages tend to be slightly further apart than those for lower bias voltages. This is unlike the curves of a regular transistor which tend to be closer together at higher biasing currents than at lower biasing currents.

Enhancement-type MOS Common-source Circuits

The enhancement-type MOS common-source circuit shown in Fig. 7 is very similar to the common-emitter circuit of a regular transistor. A load line (Fig. 8) drawn on the semiconductor's characteristic curves can be used to determine the values for the components in the circuit. For this particular load line the voltage source (V_{DD}) is 24v, the load resistance is 1.3K and the Q-point for the gate voltage (V_G) is 6v.

Since the gate current of an enhancement-type MOS is negligible compared to the base current of a regular transistor, this current need not be considered when determining values for R_1 and R_G, and we can assume that the same current (I_1) flows through both resistors. This current (I_1) is equal to the voltage source (V_{DD}) divided by the total resistance across it ($R_1 + R_G$), or [$V_{DD}=I_1(R_1+R_G)$]. The gate voltage (V_G) is equal to the value of the resistance across it (R_1) times the current (I_1) flowing through that resistor ($V_G = I_1R_1$). The ratio of the gate voltage to the voltage source can be used to determine the values of the biasing resistors.

$$\frac{V_G}{V_{DD}} = \frac{I_1R_1}{I_1(R_1 + R_G)} = \frac{R_1}{R_1 + R_G}.$$

The value of R_1 must be low enough to prevent static charges from destroying the internal gate resistance of the semiconductor

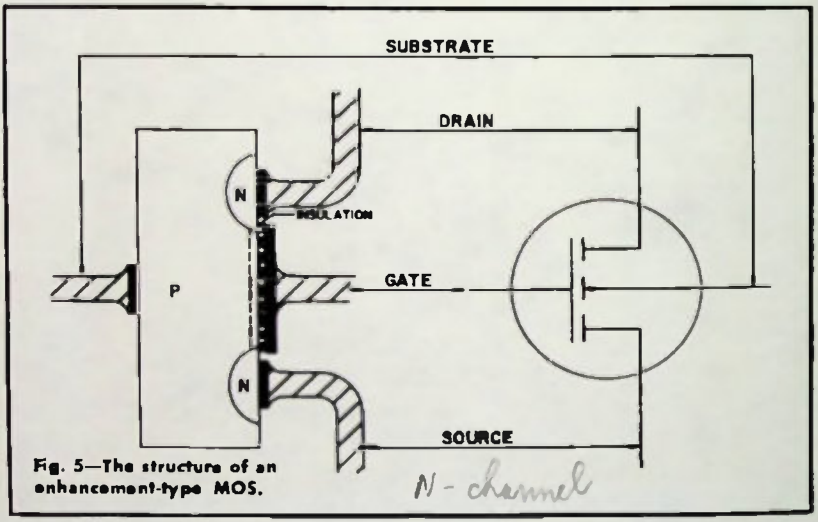


Fig. 5—The structure of an enhancement-type MOS.

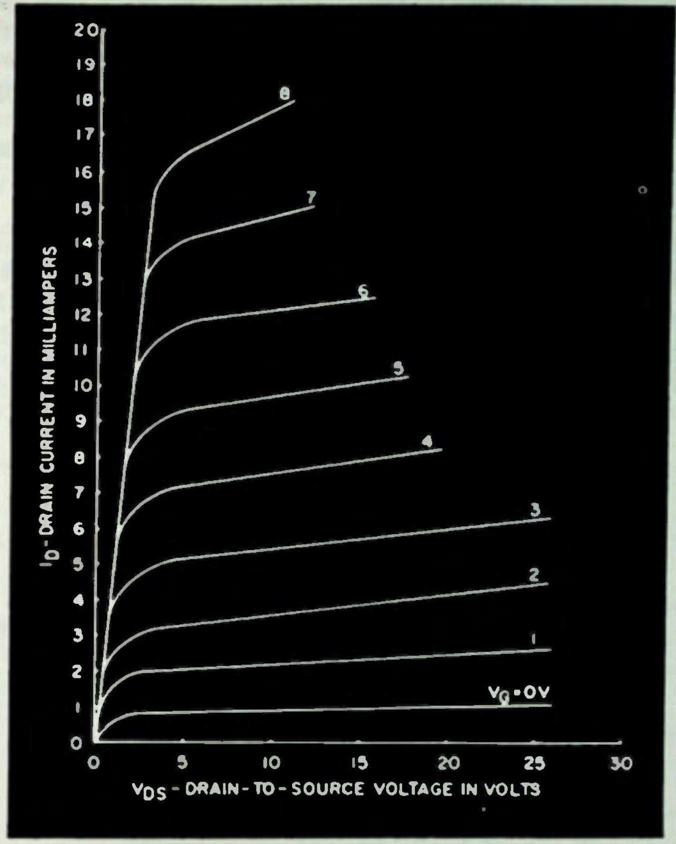
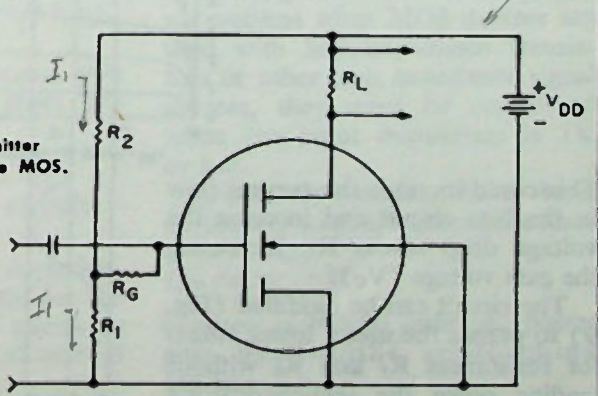


Fig. 6—The characteristic curves of an enhancement-type MOS. N-channel

Fig. 7—A simplified common-emitter circuit for an enhancement-type MOS.



and, if necessary, to load the input signal to the desired amplitude. For this circuit we will arbitrarily assign this resistor the value of 20M (Let $R_1 = 20M$). The corresponding value of R_G can then be calculated.

$$\frac{6v}{24v} = \frac{20M}{20M + R_G}.$$

$$480M = 120M + 6R_G.$$

$$6R_G = 360M. \quad R_G = 60M.$$

A 60M resistance is so large that moisture or other contamination of the circuit wiring can alter the effective value of R_G by shunting it. This could increase the current flow in the bias circuit and increase the voltage drop across R_1, increasing the gate voltage (V_G).

The circuit can be modified (Fig. 9) to permit the use of lower values for resistances R_1 and R_2 without loading down the semiconductor's gate impedance. Since the gate current is very small, the gate voltage (V_G) is essentially the voltage drop across resistor R_1. If we let the total value of the shunting resistors equal 1M (Let $R_1 + R_2 = 1M$), we can calculate the value for each resistor.

$$\frac{V_G}{V_{DD}} = \frac{R_1}{R_1 + R_2}. \quad \frac{6v}{24v} = \frac{R_1}{1M}.$$

$$24R_1 = 6M. \quad R_1 = \tfrac{1}{4}M = 250K.$$

$$R_2 = 1M - 250K = 750K.$$

The load line in Fig. 8 shows us how the circuit in Fig. 9 functions.

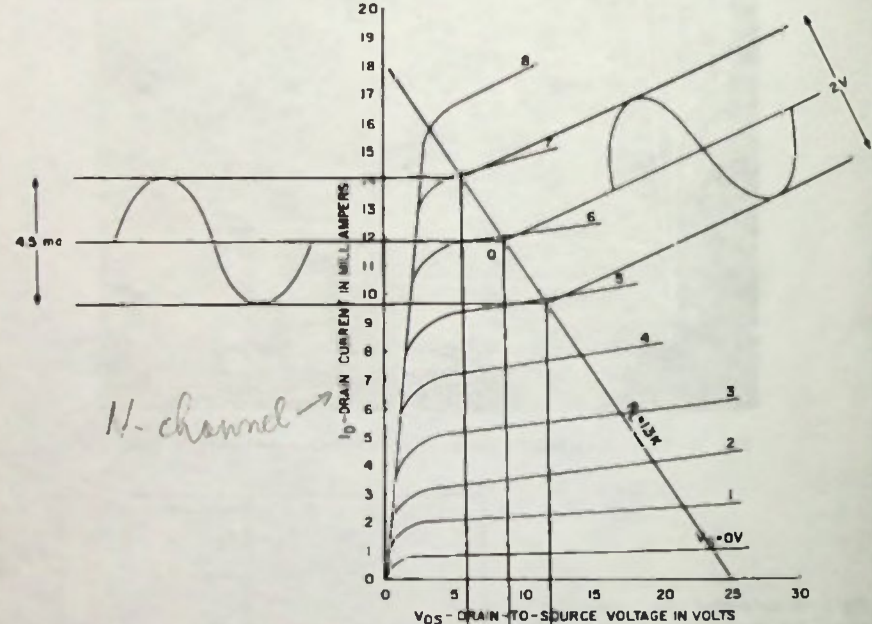


Fig. 8—The load line of an enhancement type MOS.

A 2v input signal (Δv_G) results in a 6v output signal voltage (Δv_{DS}) and a 4.5ma output signal current (Δi_D). From these figures we can calculate the circuit's voltage gain

$$(A_v = \frac{6v}{2v} = 3).$$

Since this semiconductor has an insulated gate, the gate current is insignificant. A very small current resulting from the input signal flows across resistor R_G and through the bias circuit. The value of this current can be calculated as:

General MOS Characteristics

The unique characteristics of MOS devices seem to make them the semiconductor of the future. Except for VHF, almost any electron-tube circuit can be easily adapted for the MOS.

Enhancement-type MOS components are relatively independent of temperature when operated at low drain currents, with some rated for operation with junction temperatures as high as 200°C. MOS components, in general, seem to be less sensitive to temperature than other semiconductors and electron tubes.

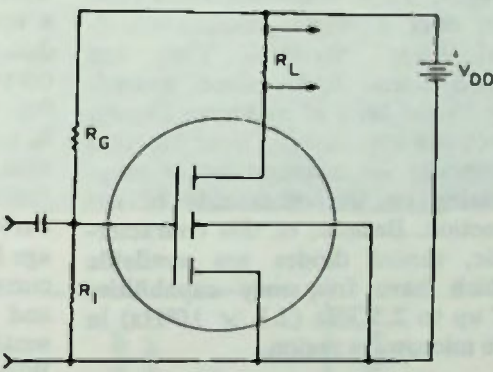


Fig. 9—An improved common-source circuit for an enhancement-type MOS.

N-channel

$$(\Delta i_{in} \approx \frac{\Delta v_G}{R_G} = \frac{2v}{20M} = 0.1\mu a).$$

The circuit's resulting current gain (A_I) can be calculated from this information.

$$A_I = \frac{\Delta i_D}{\Delta i_{in}} = \frac{4.5ma}{0.1\mu a} = 45{,}000.$$

As was the case with the depletion-type MOS, the current gain for the enhancement-type MOS circuit is too large to be a practical unit of measure.

Although noise levels are usually no problem when MOS devices are used with high-impedance transistors or other high impedance signal sources, they must be considered when the input impedance is 1K or less.

Laboratory models have been developed with a transconductance

$$(g_m = \frac{\Delta i_D}{\Delta v_G})$$

in excess of 10,000 μmho and power gain bandwidths approaching 1GHz.

CHAPTER 7

The Tunnel Diode

The structure of tunnel diodes differs considerably from those of regular transistors, FETS and MOS devices. These semiconductors contain only a single junction of P- and N-type material. They are called tunnel diodes since, according to the laws of quantum theory, electrons approaching their junction disappear — instantaneously reappearing on the other side of the junction. Because of this characteristic, tunnel diodes are available which have frequency capabilities of up to 2.5GHz (2.5×10^9Hz) in the microwave region.

Tunnel Diode Characteristic Curves

The voltage-current characteristics of a tunnel diode differ considerably from those of a regular diode (Fig. 1). The forward current of a tunnel diode, unlike that of a regular diode, is reduced as the forward bias voltage increases beyond a certain point. This reduction in current continues until a higher voltage is reached beyond which the current increases in the same manner as that in a conventional diode.

The characteristic curve of a tunnel diode is shown in greater detail in Fig. 2. The "peak point" represents the point on the curve where the diode's current reaches a maximum value before being reduced by a larger bias voltage. The corresponding current and voltage for this point are represented by I_p and V_p, respectively. As the bias voltage continues to increase, a "valley point" is reached on the curve where the increasing bias voltage has resulted in a minimum diode current. The corresponding current and voltage at this point are represented by I_v and V_v, respectively. With still larger forward bias voltages, the tunnel diode's current increases until it reaches the maximum current reached previously at the "peak point." The diode current and forward bias voltage corresponding to this third characteristic point on the graph, the "forward point," are represented by I_p and V_{fp}, respectively. At higher bias voltages, the diode's current continues to increase.

In the portion of the characteristic curve between the peak point and valley point, the diode's current (I_D) decreases with increased volt-

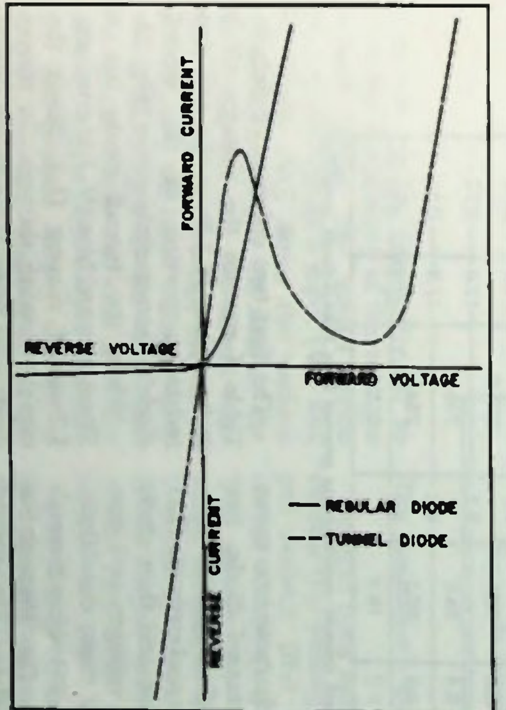


Fig. 1—Comparing a regular diode with a tunnel diode.

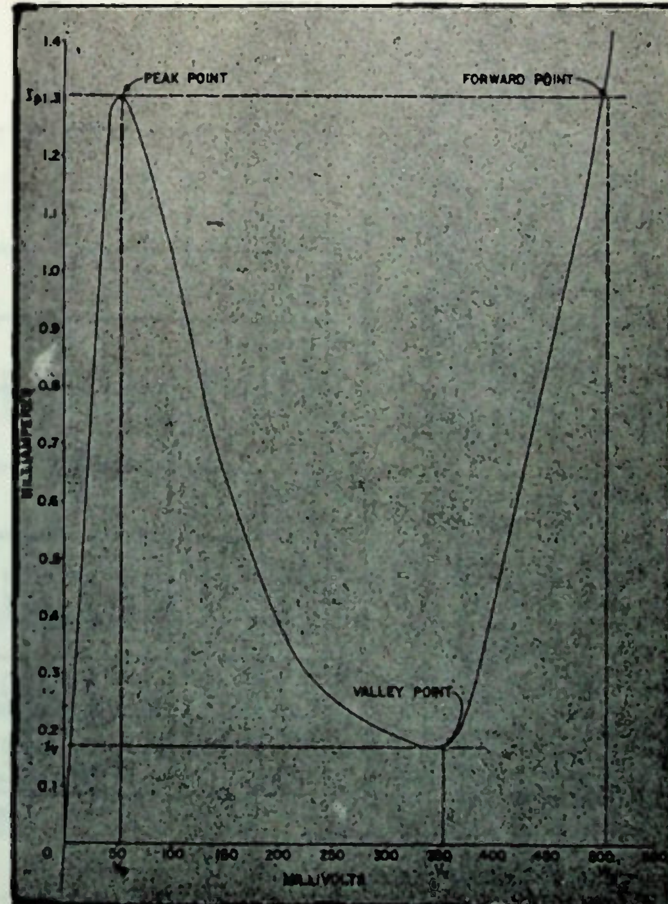


Fig. 2—A tunnel diode characteristic curve.

age (V_D). This is unlike a regular resistor which permits a greater current flow as the voltage increases. Since the reverse condition occurs in the diode over this range of voltages, it is said to have a negative resistance.

A Simplified Amplifier Circuit

A simplified circuit (Fig. 3) was voltage ($V_2 = I_D R_2$). The voltage (V_1) across the input resistor (R_1) is equal to the sum of the other voltage drops within the circuit ($V_1 = V_D + V_2$). With this information the input voltages necessary for providing certain voltages across the tunnel diode can be calculated ($V_1 = V_D + V_2 = V_D + I_D R_2$). Since $R_2 = 92\Omega$, $V_2 = I_D \times 92\Omega$ and $V_1 = V_D + I_D \times 92\Omega$.

TABLE I

V_D (mv)	I_D (ma)	$V_2 = I_D R_2$ (mv)	ΔV_2 (mv)	$V_1 = V_D + V_2$ (mv)	ΔV_1 (mv)	$A_v = \frac{\Delta V_2}{\Delta V_1}$
50	1.3	119.6		169.6		
			9.2		15.8	0.59
75	1.2	110.4		185.4		
			9.2		5.8	1.59
90	1.1	101.2		191.2		
			9.2		5.8	1.59
105	1.0	92.0		197.0		
			9.2		2.8	3.28
117	0.9	82.8		199.8		
			9.2		2.8	3.28
129	0.8	73.6		202.6		
			9.2		3.8	3.22
142	0.7	64.4		206.4		
			9.2		5.8	1.59
157	0.6	55.2		212.2		
			9.2		11.8	0.78
178	0.5	46.0		224.0		
			9.2		12.8	0.72
200	0.4	36.8		236.8		
			9.2		17.8	0.52
227	0.3	27.6		254.6		
			9.2		58.8	0.31
295	0.2	18.4		313.4		

Voltage and current combinations present in a tunnel-diode amplifier.

designed for the characteristic curve (Fig. 2) of the tunnel diode. For simplicity we will first consider various voltages (V_D) and currents (I_D) at the tunnel diode and then study the other circuit voltages and currents required for these conditions.

The current (I_D) flowing through the diode must also flow through the output resistor (R_2). This current will determine the circuit's output

The first two columns of data in Table I contain the voltages (V_D) and currents (I_D) from the negative resistance portion of the tunnel diode's characteristic curve (Fig. 2).

When the tunnel diode has a 50mv forward bias (V_D), it conducts 1.3ma of current (I_D). Since the current flowing through the tunnel diode also flows through the output resistor, the output voltage (V_2) can be calculated ($V_2 = I_DR_2 = 1.3ma \times 92\Omega = 119.6mv$). To develop 50mv across the tunnel diode and 119.6mv across the output resistor, the input voltage must equal the total of these two voltages ($V_1 = V_D + V_2 = 50mv + 119.6mv = 169.6mv$).

The same type of calculations can be made for a 75mv forward bias across the tunnel diode. Under these conditions the diode conducts 1.2-ma, develops a voltage drop of 110.4mv across the load resistor and the input voltage is 185.4mv.

With the calculated data in Table I, curves can be drawn (Fig. 4) to show the relationship of the voltages present across the tunnel diode (V_D) at the circuit's input (V_1) and its output (V_2). Since the various output voltages are the sum of the diode voltages and the output voltages ($V_1 = V_D + V_2$), the curve representing the input voltages must equal the height of the other two curves combined. As seen from the curves, with a 92Ω output resistor, the decrease in output voltage is smaller than the corresponding increase in diode voltage. As a result, the input voltage always increases as the diode voltage increases. Or, with the 92Ω output resistor an increase in input voltage results in a corresponding increase in diode voltage. Under these conditions, the input voltage determines the diode voltage and the output voltage.

From the curves we can see that when the input voltage (V_1) is 200mv, very small input voltage fluctuations (Δv_1) result in large output voltage fluctuations (Δv_2). As the input voltage increases, the output voltage decreases, and any signal present at the input is inverted and amplified at the output.

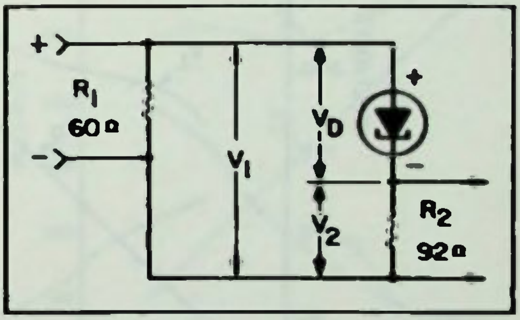


Fig. 3—Voltage drops in a simplified amplifier circuit.

The Amplifier's Output Resistance

As had been indicated, the diode current (I_D) passing through the output resistor (R_2) resulted in the output voltage (V_2), ($V_2 = I_DR_2$). The smaller the output resistor's value, the smaller the output voltage (V_2) and the resulting output signal. If the output resistor's value is too small, very little amplification results.

Another problem is encountered if the value of the output resistor (R_2) is too large. Suppose its value is increased from 92Ω to 150Ω. The resulting output voltage curve (fig. 5) is not only higher, but steeper than before (Fig. 4). With the larger output resistor, the decrease in output voltage can be greater than the corresponding increase in diode voltage. As a result, the input voltage does not always increase as the diode voltage increases. We see from the resulting curves that when the diode voltage is between 95 and 150mv, the decrease in output voltage is greater than the increase in diode voltage — resulting in a decreasing input voltage. At other diode voltages, increases in voltages across the output are smaller than those across the diode, and an increase in diode voltage results in an increase in input voltage. From these curves we see that when there is a 250mv input voltage there can be a diode voltage of 60mv, 125mv or 175mv, and an output voltage of 190mv, 122mv or 77mv. With the larger output resistor the input voltage cannot determine the output voltage and the circuit is unstable. The output resistor's value affects the stability and voltage gain of the circuit.

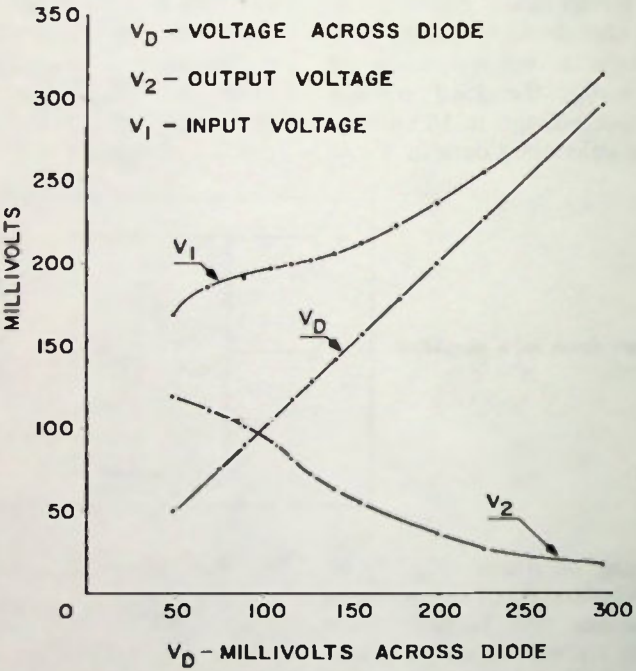


Fig. 4—Comparing voltage drops across components in a stable amplifier circuit.

The Amplifier's Voltage Gain

When comparing the various sets of calculations in Table I, using the smaller 92Ω output resistor, we note that although the increase in the input voltage (ΔV_1) was 15.8mv (185.4mv — 169.6mv), the corresponding increase in output voltage (ΔV_2) was only 9.2mv (119.6mv — 110.4mv).

The circuit's voltage gain (A_v) is equal to the change in output voltage (ΔV_2) divided by the change in input voltage (ΔV_1). ($A_v = \frac{\Delta V_2}{\Delta V_1}$).

This voltage gain can be determined from the change in voltages that have been calculated. ($A_v = \frac{\Delta V_2}{\Delta V_1} = \frac{9.2mv}{15.8mv} = 0.59$).

Table I shows the voltages in the circuit that result when the input voltage is increased to provide various voltages across the tunnel diode. From the last column of the table we can see that over a certain range of input voltages, larger voltage gains are experienced in the tunnel diode circuit.

The circuit's voltage gains (A_v) can be more clearly visualized when plotted against some other variable. This has been done in Fig. 6 by plotting the voltage gain (A_v) against the bias voltage (V_D) across the diode. [*Note, the curve does not follow exactly the points obtained from Table I. This discrepancy results from the fact that the tunnel diode's characteristic curve (Fig. 2) can not be read to as many significant figures as are required for a smoother curve.*]

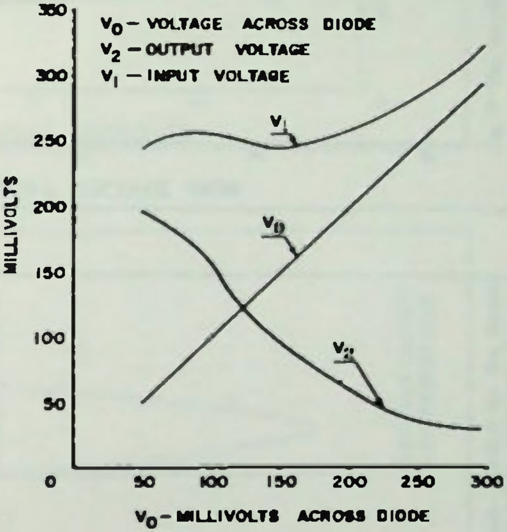


Fig. 5—Comparing voltage drops across components in an unstable amplifier circuit.

This curve (Fig. 6) should not be considered another characteristic curve of the tunnel diode. It is instead a characteristic curve of an amplifier containing a tunnel diode of known characteristics.

Amplification occurs in the tunnel diode circuit when the change in

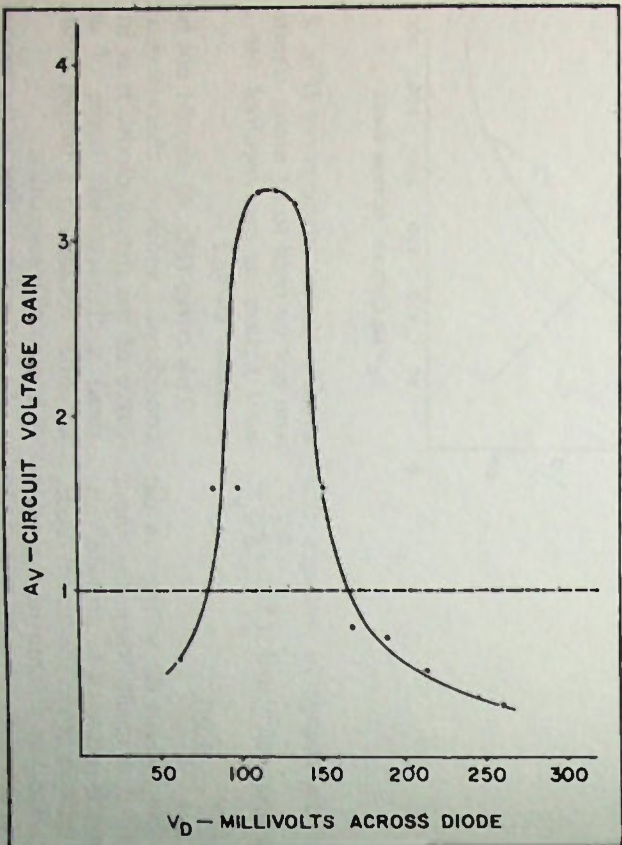


Fig. 6—Voltage gains in a tunnel diode amplifier.

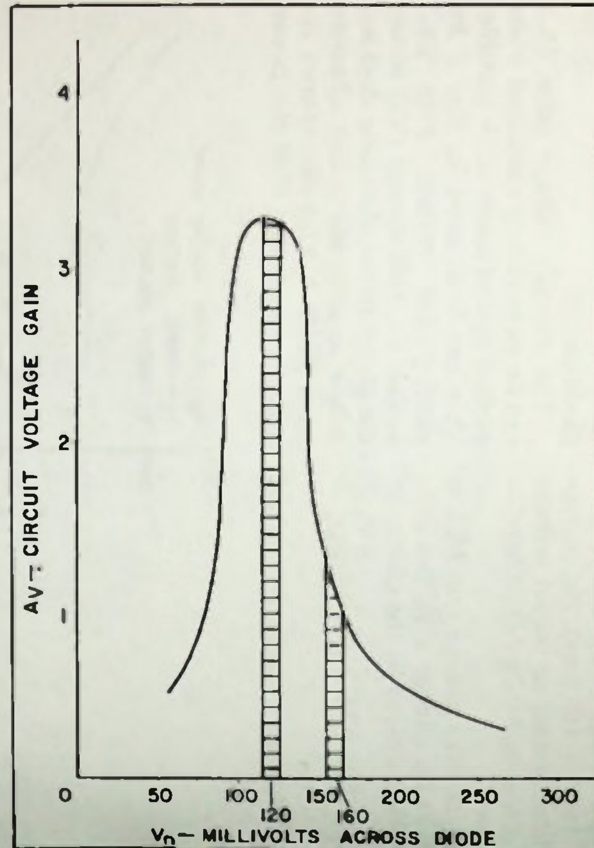


Fig. 7—The effect of diode biasing on a signal's gain.

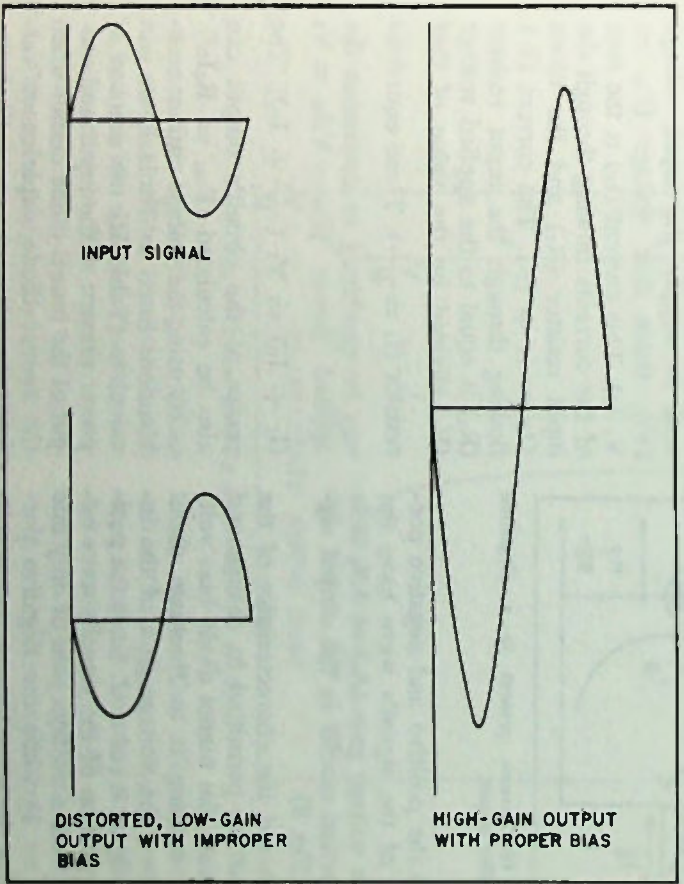


Fig. 8—Comparing input and output signals effected by the tunnel diode bias.

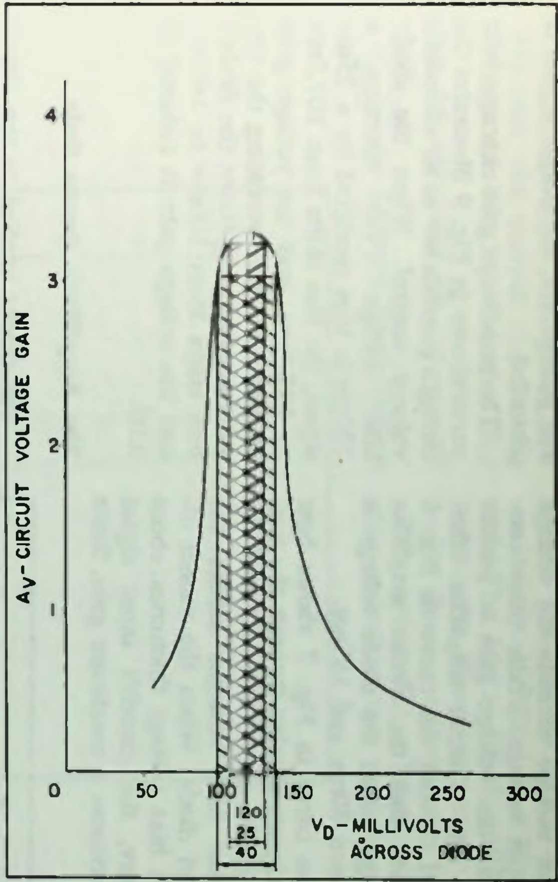


Fig. 9—The non-linear curve limits the gain of input signals.

output voltage exceeds the change in input voltage. Under these conditions the voltage gain is greater than one. (Circuit amplifies when $A_v > 1$.) From the curve in Fig. 6 we see that the circuit amplifies voltages when the diode voltage is between 80mv and 165mv.

The curve in Fig. 7 shows how the signals passing through the tunnel diode circuit are affected by various bias voltages across the tunnel diode. When the tunnel diode's bias voltage fluctuates about 120mv, the circuit's input signal experiences a maximum gain. Since

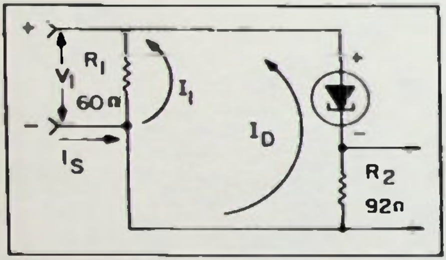


Fig. 10—Currents present in a simplified amplifier circuit.

both the positive and negative portion of the signal's wave have the same voltage gain (A_v) of 3.3, little distortion results in the output signal (Fig. 8).

Should the characteristics of the circuit be permitted to change and increase the tunnel diode bias voltage, causing it to fluctuate about 160mv, the voltage gain of the input signal is reduced. Since the positive portion of the signal's wave experiences a voltage gain of only one ($A_v = 1$) while the negative portion of the wave experiences a voltage gain of 1.3, the output signal is distorted.

The non-linear gain characteristic curve shown in Fig. 9 illustrates the circuit's possible use as an automatic volume control. When the diode bias voltage (V_D) contains a 120mvdc bias modified by a 25mv signal, the bias shifts from 107.5mv to 132.5mv, and the voltage gain (A_v) is 3.25. By increasing the signal's amplitude to 40mv, the diode's bias shifts from 100mv to 140mv, and the voltage gain is reduced to 3.05.

The Amplifier's Power Gain

The power applied to the circuit (Fig. 10) is equal to the current (I_S) flowing from the voltage source, used to supply the input voltage (V_1), times that voltage ($P_{in} = V_1 I_S$). This current (I_S) is the sum of the currents flowing through the input resistor (R_1) and the diode ($I_S = I_1 + I_D$). The current (I_1) flowing through the input resistor (R_1) is equal to the applied voltage (V_1) divided by the value of that resistor ($I_1 = \frac{V_1}{R_1}$). These equations can be combined to determine the applied power [$P_{in} = V_1 I_S = V_1 (I_1 + I_D) = V_1 (\frac{V_1}{R_1} + I_D)$]. The power at the circuit's output can also be calculated ($P_{out} = R_2 I_D^2$).

By using the voltage-current combinations listed in Table I, we can calculate (Table II) the amount of power present at the input and output of the tunnel diode circuit when the tunnel diode experiences various bias voltages.

When the input voltage (V_1) is 169.6mv, the current (I_1) flowing through the 60Ω input resistor (R_1) can be calculated ($I_1 = \frac{V_1}{R_1} = \frac{169.6mv}{60\Omega} \approx 2.82ma$). From Table I we see that under these conditions the tunnel diode current (I_D) is 1.3ma. With this information the source current (I_S) can be determined ($I_S = I_1 + I_D = 2.82ma + 1.30ma = 4.12ma$). The power applied ($P_{in}$) can then be calculated ($P_{in} = V_1 I_S = 169.6mv \times 4.12ma = 169.6 \times 10^{-3}v \times 4.12 \times 10^{-3}a = 698.752 \times 10^{-6} \approx 0.698 \times 10^{-3}w = 0.698mw$). Since we know that under these conditions the tunnel diode current (I_D) is 1.3ma, we can also calculate the output power (P_{out}) across the 92Ω output resistor (R_2). [$P_{out} = R_2 I_D^2 = 92\Omega \times (1.3ma)^2 = 92\Omega \times 1.3 \times 10^{-3}a \times 1.3 \times 10^{-3}a = 155.48 \times 10^{-6}w \approx 0.155 \times 10^{-3}w = 0.155mw$]

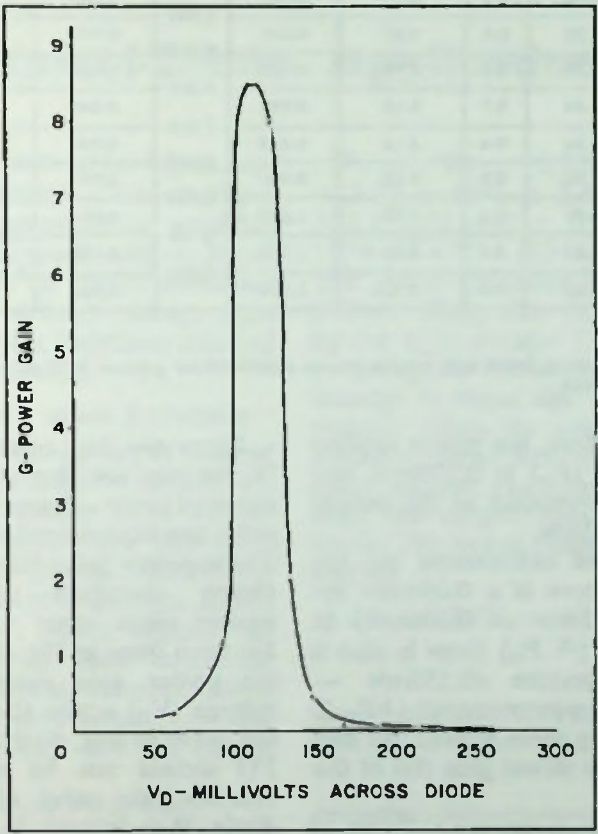


Fig. 11—Power gains in a tunnel diode amplifier.

In a similar manner we can calculate that when the input voltage

TABLE II

V_1 (mv)	$I_1 = \frac{V_1}{R_1}$ (ma)	I_D (ma)	$I_B = I_1 + I_D$ (ma)	$P_{in} = V_1 I_B$ (mw)	ΔP_{in} (mw)	$P_{out} = I_D^2 R_2$ (mw)	ΔP_{out} (mw)	$G = \frac{\Delta P_{out}}{\Delta P_{in}}$
169.6	2.82	1.3	4.12	0.699		0.155		
					0.096		0.023	0.24
185.4	3.09	1.2	4.29	0.795		0.132		
					0.025		0.021	0.84
191.2	3.19	1.1	4.29	0.820		0.111		
					0.023		0.019	0.83
197.0	3.28	1.0	4.28	0.843		0.092		
					0.002		0.017	8.50
199.8	3.33	0.9	4.23	0.845		0.075		
					0.002		0.016	8.00
202.6	3.38	0.8	4.18	0.847		0.059		
					0.007		0.014	2.00
206.4	3.44	0.7	4.14	0.854		0.045		
					0.025		0.012	0.48
212.2	3.54	0.6	4.14	0.879		0.033		
					0.068		0.010	0.11
224.0	3.73	0.5	4.23	0.947		0.023		
					0.083		0.008	0.10
236.8	3.95	0.4	4.35	1.030		0.015		
					0.126		0.007	0.06
254.6	4.24	0.3	4.54	1.156		0.008		
					0.543		0.004	0.01
313.4	5.22	0.2	5.42	1.699		0.004		

Comparing input and output power combinations present in a tunnel-diode amplifier.

(V_1) is 185.4mv, the power applied at the input (P_{in}) is 0.795mw and the power developed at the output (P_{out}) is 0.132mw.

From these calculations we see that when there is a 0.096mw increase (0.795mw — 0.699mw) in power input (ΔP_{in}) there is also a 0.023mw decrease (0.155mw — 0.132mw) in power output (ΔP_{out}). By comparing these figures, we can determine the power gain (G) of the tunnel diode circuit ($G = \frac{\Delta P_{out}}{\Delta P_{in}} = \frac{0.023\text{mw}}{0.096\text{mw}} = 0.24$).

From the last column of Table II, we can see that over a certain range of input voltages larger power gains are experienced in the circuit. These power gains (G) can be more clearly visualized when plotted against some other variable. This has been done in Fig. 11 by plotting the power gain against the bias voltage (V_D) across the diode. Like the curve in Fig. 6, this curve (Fig. 11) should not be considered a characteristic curve of the tunnel diode. It is instead a characteristic curve of an amplifier containing a tunnel diode of known characteristics.

CHAPTER 8

Tunnel-Diode Applications

The importance of having an output or load resistor (R_2) of proper value in a tunnel-diode amplifier circuit was discussed, and we found that too small an output resistance reduced the circuit's voltage gain while too large a resistance resulted in voltage instability.

The Amplifier's Input Resistance

The value of the input resistor (R_1) is also important for the stability of the tunnel diode circuit. This can be seen more effectively if we study the power consumed by taking a different look at the circuit (Fig. 1). The power supplied to the circuit (P_{in}) must equal the power consumed by the tunnel diode (P_D), plus that consumed by the input resistor (P_1) and the output resistor (P_{out}).

$$(P_{in} = P_D + P_1 + P_{out})$$

The power consumed by the tunnel diode can be calculated by multiplying its voltage by its current ($P_D = V_D I_D = 50\text{mv} \times 1.3\text{ma} = 50 \times 10^{-3}\text{v} \times 1.3 \times 10^{-3}\text{a} = 65 \times 10^{-6}\text{w} = 0.065 \times 10^{-3}\text{w} = 0.065\text{mw}$).
The power consumed by the input resistor (R_1) equals the square of the current passing through it times the value of the resistor

[$P_1 = I_1^2 R_1 = (2.82\text{ma})^2 \times 60\Omega = 2.82 \times 10^{-3}\text{a} \times 2.82 \times 10^{-3}\text{a} \times 60\Omega = 477.1440 \times 10^{-6}\text{w} \approx 477 \times 10^{-6}\text{w} = 0.477 \times 10^{-3}\text{w} = 0.477\text{mw}$].
The power consumed by the output resistor (R_2), like that consumed by the input resistor (R_1), is equal to the square of the current passing through it times the value of the resistor. Since the resistor is connected in series with the diode, the same current (I_D) passes through both the output resistor and the diode. The power consumed by this resistor

$$(P_{out} = I_D^2 R_2 = 0.155\text{mw}).$$

We can add this data to determine the total power consumed by the circuit:

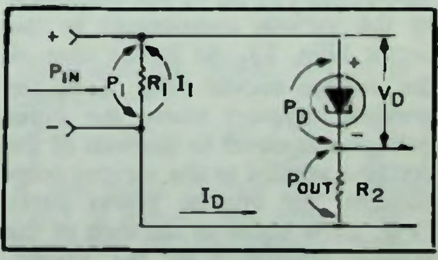


Fig. 1—Power losses in a tunnel diode amplifier.

$$(P_{in} = P_{out} + P_D + P_1 =$$

$$0.065\text{mw} + 0.477\text{mw} + 0.155\text{mw}$$

$$= 0.697\text{mw}).$$

With the use of curves (Fig. 2), we can more effectively see the relationships of the power consumed by the various components in the circuit (Fig. 1). As in the case of the voltage curves (Fig. 4 of the previous chapter) where the input voltage was equal to the sum of the voltages applied to the various components, the applied power curve (Fig. 2) is equal to the sum of the powers consumed by the components.

From the curves we note that as the voltage across the tunnel diode (V_D) increases, the power consumed at the output (P_{out}) continuously decreases; and above approximately 111mv (V_D), the power consumed by the tunnel diode (P_D) also decreases. The total power consumed by these two components (the height of the P_D curve added to the height of the P_{out} curve) decreases with an increase in diode voltage (V_D).

Unlike these curves, the curve representing the power consumed by the input resistor (P_1) increases as the diode voltage (V_D) increases.

As had been indicated earlier in this chapter, the power applied to the circuit is equal to the sum of

V_D (mv)	I_D (ma)	$P_D = V_D I_D$ (mw)	I_1 (ma)	$P_1 = I_1^2 R_1$ (mw)	P_{out} (mw)	$P_{in} = P_{out} + P_D + P_1$ (mw)
50	1.3	0.065	2.82	0.477	0.155	0.697
75	1.2	0.090	3.09	0.573	0.132	0.795
90	1.1	0.099	3.19	0.611	0.111	0.821
105	1.0	0.105	3.28	0.646	0.092	0.843
117	0.9	0.105	3.33	0.665	0.075	0.845
129	0.8	0.103	3.38	0.686	0.059	0.848
142	0.7	0.099	3.44	0.710	0.045	0.854
157	0.6	0.094	3.54	0.752	0.033	0.879
178	0.5	0.089	3.73	0.835	0.023	0.947
200	0.4	0.080	3.95	0.936	0.015	1.031
227	0.3	0.068	4.24	1.079	0.008	1.155
295	0.2	0.059	5.22	1.635	0.004	1.698

Table I: The power consumed by the various components in a tunnel-diode amplifier.

the powers consumed by the components.

$(P_{in} = P_{out} + P_D + P_1)$.

Since the total power consumed by the output resistor (P_{out}) and the diode (P_D) is decreasing as the power consumed by the input resistor (P_1) increases, there is a very small resulting increase in the total power consumed by the circuit (P_{in}), as the diode voltage increases from 105 mv to 129 mv. Over this range of values, a 0.005mw increase in the power applied to the circuit results in a 0.033mw decrease in the power consumed by the output. The power gain curve (Fig. 11 of the previous chapter) shows that the circuit's power gain is greatest over this range of diode voltages. When the input resistor is of its present value, the power applied to the circuit is able to control the amount of power present in the output.

By increasing the value of the input resistor, we can decrease the amount of power consumed at the input resistor when a given voltage is applied to that resistor. This can be shown by means of some basic electrical equations with which the reader is probably familiar. The power consumed by a resistor is equal to the voltage applied to the resistor times the current passing through it ($P = VI$). The current passing through the resistor is equal to the voltage applied to the resistor divided by the value of the resistor ($I = \frac{V}{R}$).
These two equations can be combined to show the relationship between the value of the input resistor

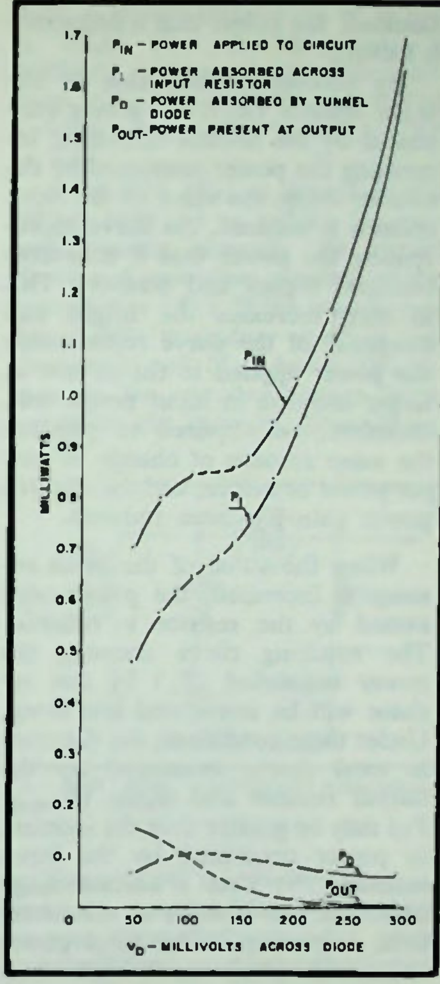


Fig. 2—Comparing the power consumption of components in a stable amplifier circuit.

(R_1) and the power (P_1) that it consumes

$$\left[P_1 = V_1 I_1 = V_1 \left(\frac{V_1}{R_1}\right) = \frac{V_1^2}{R_1} \right]$$

If the value of the input resistor (R_1) is halved, the power that it consumes (P_1) is doubled; while if the value of the input resistor is doubled, the power that it consumes is halved.

By decreasing the value of the input resistor (R_1), the power consumed by the resistor increases, increasing the power consumed by the circuit. When the value of the input resistor is reduced, the curve representing the power that it consumes becomes higher and steeper. This in turn increases the height and steepness of the curve representing the power applied to the circuit. A larger increase in input power will, therefore, be required to produce the same amount of change in output power as before, and the circuit's power gain has been reduced.

When the value of the input resistor is increased, the power consumed by the resistor is reduced. The resulting curve showing the power consumed (P_1) by this resistor will be lower and less steep. Under these conditions, the decrease in total power consumed by the output resistor and diode ($P_{out} + P_D$) may be greater than the increase in power consumed by the input resistor (P_1) over a certain range of values. As a result of this situation, a portion of the total power consumed (P_{in}) curve will drop over a certain range of diode voltages (V_D). As a result of this situation, we have an unstable power input curve similar to the unstable voltage input curve (Fig. 5 of previous chapter) discussed before. Just as several output voltages could result from a single input voltage, several amounts of power consumed at the output can result from a single amount of input power.

We have seen that when the value of the output resistor (R_2) is too small, the circuit's voltage gain is reduced; and when the value of the resistor is too large, the circuit loses its stability. We have also seen that when the value of the input resistor (R_1) is too small, the circuit's power gain is reduced; and when the value of the resistor is too large, the circuit loses its stability.

Experiments have shown that a tunnel diode circuit is stable when the sum of the positive resistances in the circuit is smaller than the tunnel diode's maximum negative resistance. The total positive resistance calculated for the circuit must include the positive resistance (r_s) present in parts of the tunnel diode that are in series with its negative resistance ($-r_d$), as well as the other positive resistances present in the circuit. (Circuit stable when

$$R_1 + R_2 + r_s < -r_d.)$$

The tunnel diode's negative resistance is frequently expressed in terms of its negative conductance

$$\frac{1}{-r_d} = -g_d). \ (g_d = \frac{\Delta_I}{\Delta_V}.)$$

For the sake of simplicity, we have considered only the positive dc resistances present in the input, output and tunnel diode. At higher frequencies, the impedance of each of these portions of the circuit must be included in their total resistance.

Experiments have also shown that the tunnel diode generally experiences better gain when the value of the output resistor (R_2) is greater than that of the input resistor (R_1). (Normally $R_2 > R_1$.)

More Common Amplifier Circuits

The tunnel diode circuits that

have been shown (Fig. 3 and 10 of the previous chapter and Fig. 1 of this chapter) are identical and were illustrated only for the purpose of discussing different aspects of the circuit. The same circuit can be shown in still a different perspective (Fig. 3). Although this circuit has many practical applications in its present form, occasionally the bias supply must be separated from the input signal. This can be done with the use of coils (Fig. 4).

As many readers know, the amount of current that can pass through a coil is dependent on the resistance of the coil, its impedance

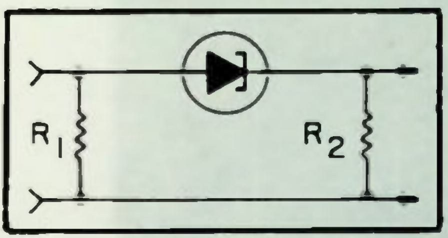


Fig. 3—A more common view of the simplified amplifier circuit.

and the frequency of the voltage across it. When a dc voltage is applied to a coil, the current is dependent solely on the coil's resistance $(I = \frac{V}{R})$.

A coil's ability to conduct current decreases as the frequency of the voltage increases. This reduction in current flow is due to the reactance (X_L) of the coil $(I \sim \frac{V}{X_L})$.

There is a definite relationship between the impedance of the coil (L), the frequency of the voltage (f) and the reactance (X_L). ($X_L = 2\pi fL$.) If the impedance of a coil is $15\mu h$, and the frequency is 100MHz, we can calculate the reactance ($X_L = 2\pi fL \approx 2 \times 3.14159 \times 10^6 Hz \times 15 \times 10^{-6} L \approx 9{,}424.8\Omega \approx 9K$).

A resistor (R_1) is placed in the circuit (Fig. 4) to reduce the current passing through the tunnel diode and secondary winding of the transformer (T_1). If the transformer has a 60Ω secondary winding and the output coil (L_2) has a 92Ω resistance, the resulting voltage drops would be the same as those shown

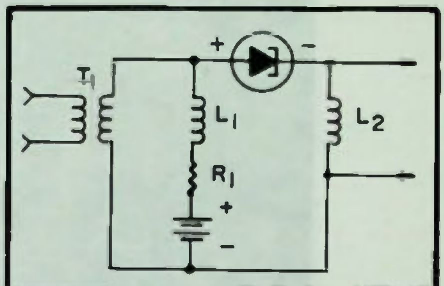


Fig. 4—A more complicated tunnel-diode amplifier circuit.

in Table II of the previous chapter, and a current of 4.23ma would have to pass through the resistor (R_1) before 200mv could be developed across the transformer's secondary winding. When the power is supplied by a 3v battery, the total dc resistance of the coil (L_1) and resistor (R_1) must be sufficient to develop a 2800mv (2.8v) drop in voltage (3000mv − 200mv = 2800mv). We can calculate the total resistance required for these conditions

$$(R = \frac{V}{I} = \frac{2800mv}{4.23ma} \approx 662\Omega).$$

If the coil has a dc resistance of 5Ω, the resistor (R_1) must have a value of 657Ω.

Since the transformer cannot transfer dc currents from one winding to another, the primary winding is free of dc current and the resulting bias. An unbiased input signal can pass through the transformer's primary winding and develop a signal on the biased secondary winding. Due to the high impedance of the coil (L_1) in the battery circuit (Fig. 5). This can be done by substituting a coil (L_3) for the secondary winding of the transformer (T_1). By properly aligning the input coil (L_3) with the output coil (L_2), any signal present across the output coil is induced in an inverted form across the input coil. This signal is then amplified and inverted by the tunnel diode, increasing the amplitude of the signal originally present across the output coil (L_2). The circuit oscillates as

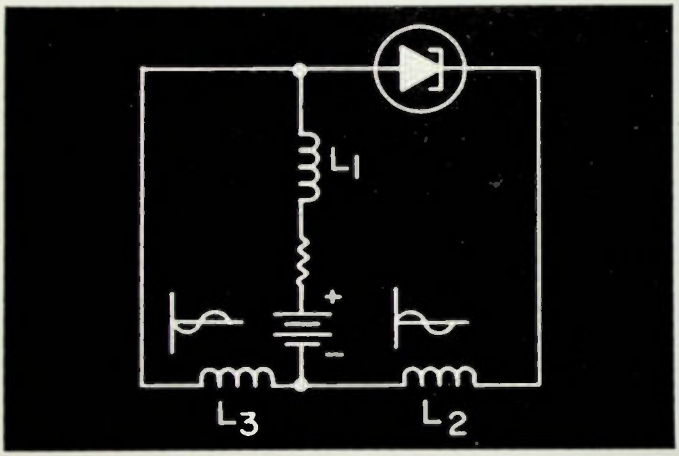


Fig. 5—A tunnel-diode, Hartley-oscillator circuit.

circuit, virtually all of the ac signal current is applied to the tunnel diode portion of the circuit. The signal, imposed on the bias voltage present at the transformer's secondary winding, changes the diode's voltage and the output voltage in the same manner discussed earlier.

Tunnel Diode Oscillators

With a slight modification, the amplifier circuit (Fig. 4) can be changed into a Hartley-oscillator a result of this positive feedback.

The relaxation-oscillator circuit, shown in Fig. 6, is a simpler tunnel-diode oscillator circuit. When a battery is connected to the circuit, the current flowing through the tunnel diode increases until the 1.3ma peak point (I_p) is reached (Fig. 7). When the surge of current passing through the coil (L_1) increases beyond 1.3ma, the voltage drop across the tunnel diode switches from 50mv to 500mv. If the supply voltage is only 225mv, the voltage drop across the tunnel diode is now greater than the voltage source. Since the tunnel diode cannot remain at a higher voltage than its voltage source, the voltage and resulting current passing through the diode must decrease. As the voltage across the diode decreases to 350mv, the valley point (V_p), the current passing through the diode decreases to 0.17ma. The voltage across the tunnel diode is still in excess of the supply voltage and must, therefore, continue to decrease. With any further reduction in current flowing through the tunnel diode, the voltage drop across the diode must switch from 350mv to 5mv. The voltage drop across the tunnel diode is now less than the voltage source, and the coil (L_1) can again conduct current from the voltage source through the tunnel diode. The cycle of voltages and currents then repeats itself as the tunnel diode continues to oscillate.

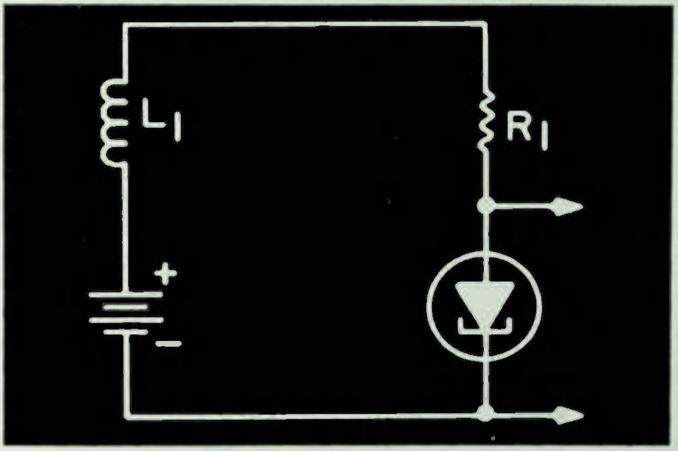


Fig. 6—A tunnel-diode, relaxation-oscillator circuit.

Tunnel Diode Switches

The oscillator and amplifier circuits mentioned function properly only when the total resistance of the circuit is smaller than the negative resistance of the tunnel diode. In a tunnel diode switching circuit (Fig. 8) the situation is different and the load resistor (R_2) is of a value greater than the diode's negative resistance. The load resistor is of a large enough value (10K) that the diode current is normally less than 1.3ma peak current (I_p). Under these conditions (Fig. 9) the voltage drop across the tunnel diode is less than the 50mv peak voltage (V_p). We can consider this maximum voltage (50mv) insignificant, compared to the 8v (8000mv) voltage source, when calculating the current (I_D) passing through the load resistor and tunnel diode

$$\left(I_D \approx \frac{8000\text{mv}}{10{,}000\Omega} = 0.8\text{ma}\right).$$

Under these conditions there is a voltage drop of 25mv across the tunnel diode. When a +0.7ma switching signal is applied through the input resistor (R_1), the tunnel diode conducts a 1.5ma current. 350mv and 500mv. If we assume that the resulting voltage drop across the tunnel diode is about 425mv (midway between those two voltages), we can more easily calculate the voltage drop, across the

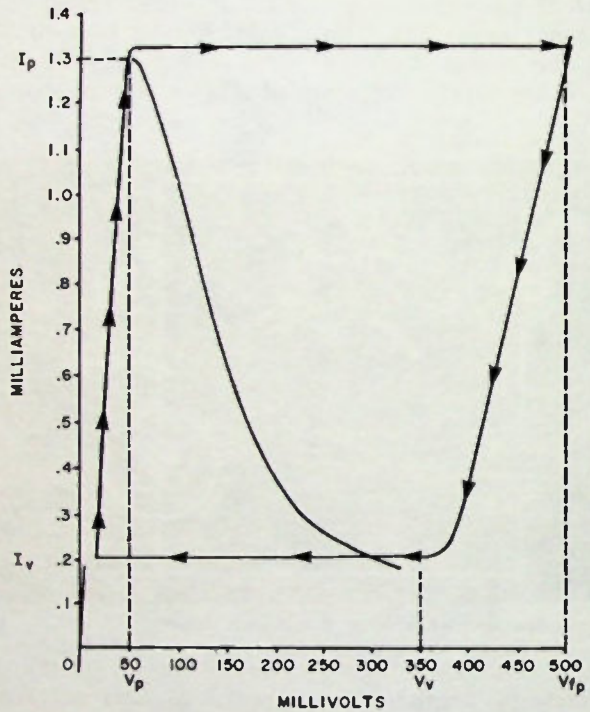


Fig. 7—The tunnel diode's voltage-current combinations in a relaxation oscillator circuit.

This is greater than the peak current (I_p) and the tunnel diode switches to a voltage drop of 520mv. After the +0.7ma switching signal has ended, the tunnel diode remains at a voltage above the valley point (V_v), having a value between load resistor, and the current that the resistor and tunnel diode are conducting while in an ON condition

$$(I_D \approx \frac{8000\text{mv} - 425\text{mv}}{10{,}000\Omega} = 0.7575\text{ma} \approx 0.76\text{ma}).$$

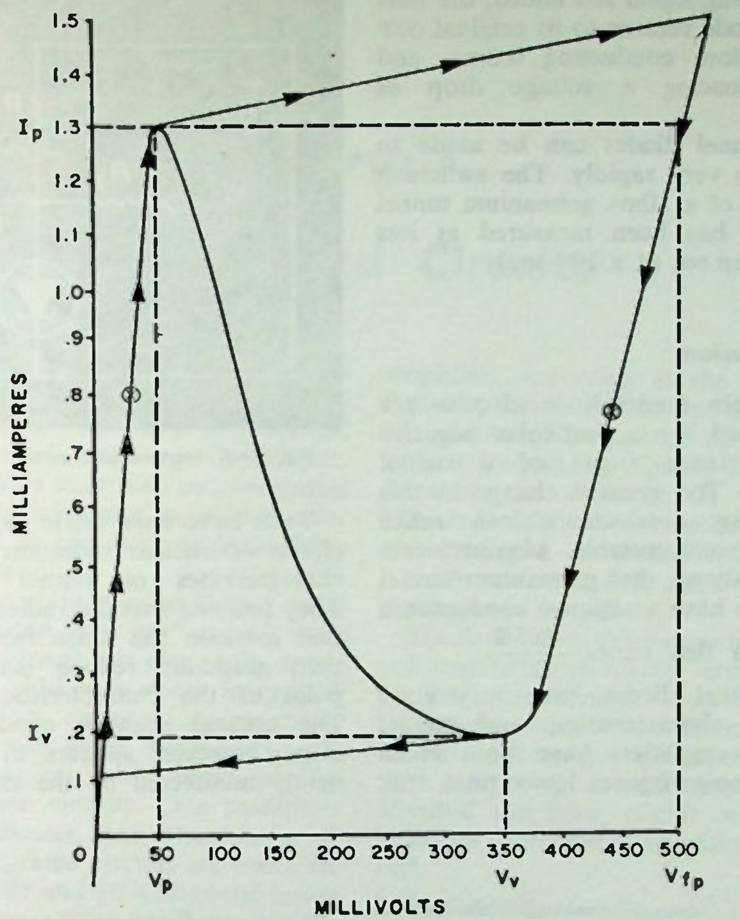


Fig. 9—The tunnel diode's voltage-current combinations in a switching circuit.

The 0.76ma diode current (I_D) corresponds to a diode voltage drop of 438mv. We had originally estimated the voltage drop to be 425mv. The 13mv error in our estimate of diode voltage is so small that the resulting error in our calculated diode current is less than the error resulting from our rounding off the value of the diode current.

When a –0.7ma switching signal is applied through the input resistor (R_1), the tunnel diode conducts a 0.1ma current. This is less than the 0.17ma valley current (I_v), and the voltage drop across the tunnel diode is reduced to 5mv. After the –0.7ma switching signal has ended, the tunnel diode returns to its original OFF condition, conducting 0.8ma, and experiencing a voltage drop of 25mv.

Tunnel diodes can be made to switch very rapidly. The switching speed of a 10ma germanium tunnel diode has been measured at less than 1p sec (1×10^{-9} sec).

Conclusion

Stable tunnel-diode circuits are designed for a particular negative conductance ($-g_d$) of a tunnel diode. Too great a change in this negative conductance can make the circuit unstable. Measurements have shown that germanium tunnel diodes have a negative conductance ($-g_d$) that varies $\frac{-0.5\%}{C^\circ}$.

Tunnel diodes have very good noise characteristics, and tunnel diode amplifiers have been made with noise figures lower than 3db.

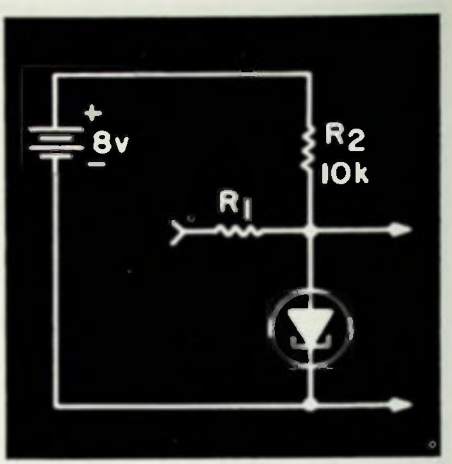


Fig. 8—A tunnel-diode switch circuit.

Tests have been made to see the effects of nuclear radiation on the characteristics of tunnel diodes. They indicate that the radiation will first increase the noise factor and then gradually reduce the valley point of the characteristic curve. The general negative conductance slope, however, appears to remain nearly unaffected by the radiation.

CHAPTER 9

Integrated Circuits

Another device that is becoming increasingly important in semiconductor radios, TV sets and special eqiupment is the integrated circuit. This device must also be understood by the technician if he is to remain efficient in his field as technology advances.

Silicon chips, smaller than a dime (Fig. 1) can now be "grown" with an assortment of electronic components. Although some integrated circuits are currently being made with plug-in bases, they are so reliable that many are wired directly into larger circuits. One manufacturer indicates that when some of their integrated circuits are operated between 25 and 30°C (around room temperature) in an oscillator circuit, they experience a failure rate equal to 0.001 percent per 1000 hours. This is equivalent to running 263,000 integrated circuits for a year without experiencing a single failure. A TV manufacturer indicates that the integrated circuit used in their sets is the least likely component to fail.

One integrated circuit (Fig. 2), sold as part No. CA3005 by an electronics supply company for $2.80, is commonly used as an RF amplifier. According to the manufacturer, this amplifier has an RF, IF, and video frequency capability, with a frequency response ranging from dc to 100MHz. It has a balanced differential amplifier circuit having a controlled constant-current source, and it can be used as a cascode amplifier.

The function of this integrated circuit can be more readily explained by first comparing a variable-current-source amplifier with a constant-current-source amplifier and then comparing an unbalanced differential amplifier circuit with a balanced differential amplifier circuit.

A Variable-Current-Source Amplifier

The variable-current-source amplifier circuit (Fig. 3) is similar to the transistor circuit shown in Fig. 2, chapter 3. The only difference is that the collector resistor (R_C) and emitter resistor (R_E) are combined as a single load resistor (R_L) in the emitter portion of the circuit (Fig. 3). The transistor in this circuit has the same characteristic curve as that shown in Fig. 3 of the forementioned chapter.

Fig. 1—Approximately 135 usable circuits have been formed on a silicon wafer not much larger than a dime. The relative size of a complete integrated circuit unit is shown on the face of a dime. Courtesy of Motorola.

If the transistor could conduct sufficient current to cause the voltage drop across the load resistor to equal the voltage source, there would be no collector-to-emitter voltage. The maximum collector current (I_C) would be the amount of current then passing through the load resistance. (When $C_{CE} = 0$, $I_C = \frac{V_{CC}}{R_L}$.) Since the circuit contains a 1K load resistance and a 16v source, 16ma of current is required for these conditions.

If the base of the transistor could cut off the collector current completely, no voltage drop would exist across the load resistor and the collector-to-emitter voltage would equal the voltage source. (When $I_C = 0$, $V_{CE} = V_{CC} = 16v$.)

Using the points determined in the two preceding paragraphs, the circuit's load line can be drawn on the transistor's characteristic curve (Fig. 4). From this load line we see that when the base current (I_B) fluctuates between 21 and 29μa, the transistor's collector-to-emitter voltage (V_{CE}) fluctuates between 6 and 8v, and the current flowing through the load resistor (R_L) fluctuates between 8 and 10 ma. (*The characteristic curve indicates that this is a fluctuation in collector current rather than a fluctuation in emitter current.* The actual emitter current varies from 8 ma + 21μa to 10ma + 29μa — *the collector current plus the base current* — *or from* 8.021 *to* 10.029ma.) For all practical purposes, there is no need

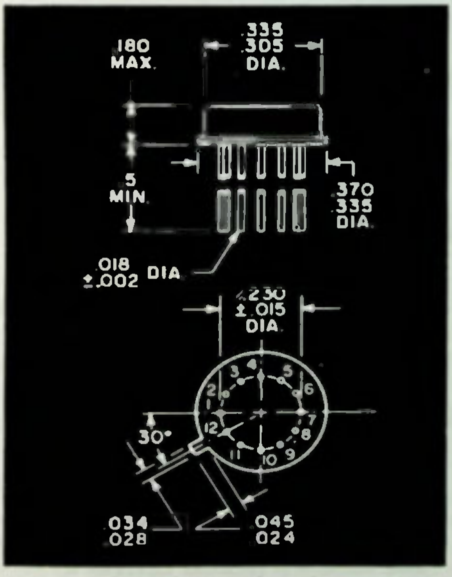


Fig. 2—The dimensions of an integrated circuit sold by an electronics supply company for $2.80. Courtesy of RCA.

for this degree of accuracy, and we can merely refer to the collector current as the current flowing through the load resistor R_L in Fig. 3. We see that a variable amount of current is able to flow from the voltage source (V_{CC}), and that is the reason the circuit has been referred to as a variable-current-source amplifier.

A Constant-Current-Source Amplifier

A second transistor (Q_2) is used to regulate the current in the constant current source amplifier cir-

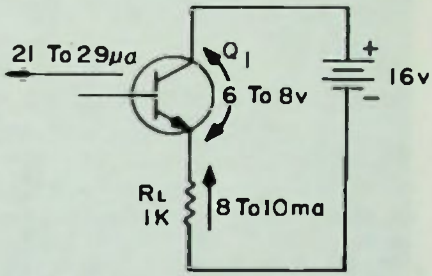


Fig. 3—A variable-current-source amplifier circuit.

cuit (Fig. 5). Resistors R_1 and R_2 are used as a voltage divider to provide this transistor with 25μa of base current (I_B). From the transistor's characteristic curve (Fig. 6), we see that the transistor's collector current (I_C) is almost solely dependent on the transistor's base current (I_B) and nearly independent of variations in its collector-to-emitter voltage (V_{CE}). When this particular transistor experiences a constant 25μa base current, and the collector-to-emitter voltage is varied from 8 to 10v, the collector current varies only from 9.0 to 9.5ma—remaining nearly constant.

The other transistor (Q_1) in this circuit (Fig. 5) is used to vary the collector-to-emitter voltage of the nearly constant current transistor (Q_2). Since the circuit contains a 16v source of voltage (V_{CC}), the collector-to-emitter voltage in transistor Q_1 must vary from 6 to 8v to cause the collector-to-emitter voltage in transistor Q_2 to vary from 10 to 8v (16v $-$6v $=$ 10v, 16v $-$8v $=$ 8v.) When the collector-to-emitter voltage of transistor Q_1 is 6v, the collector-to-emitter voltage of transistor Q_2 is 10v, and transistor Q_2 conducts 9.5ma of current. Since transistor Q_1 is connected in series with transistor Q_2, 9.5ma of current must pass through it as well. When the collector-to-emitter voltage of transistor Q_1 is 8v, the collector-to-emitter voltage of transistor Q_2 is 8v, and transistors Q_1 and Q_2 conduct 9.0ma of current. Lines have been drawn on the characteristic curve (Fig. 7) of transistor Q_1 to indicate that when it's collector-to-emitter voltage (V_{CE}) varies between 6 and 8v, the collector current (I_C) varies between 9.5 and 9.0ma. By drawing a line through the two points representing these voltage-current combinations, the effective load line can be drawn for the circuit (Fig. 5) containing transistor Q_1. From the effective load line we see that the base current (I_B) must vary between 25 and 28.5μa to cause the transistor's emitter voltage to vary between 6 and 8v. In this circuit (Fig. 5), a 3.5μa variation in base current results in a 2v change in collector-to-emitter voltage, while in the circuit discussed earlier (Fig. 3), the same transistor required an 8μa variation in base current to produce the 2v change in collector-to-emitter voltage.

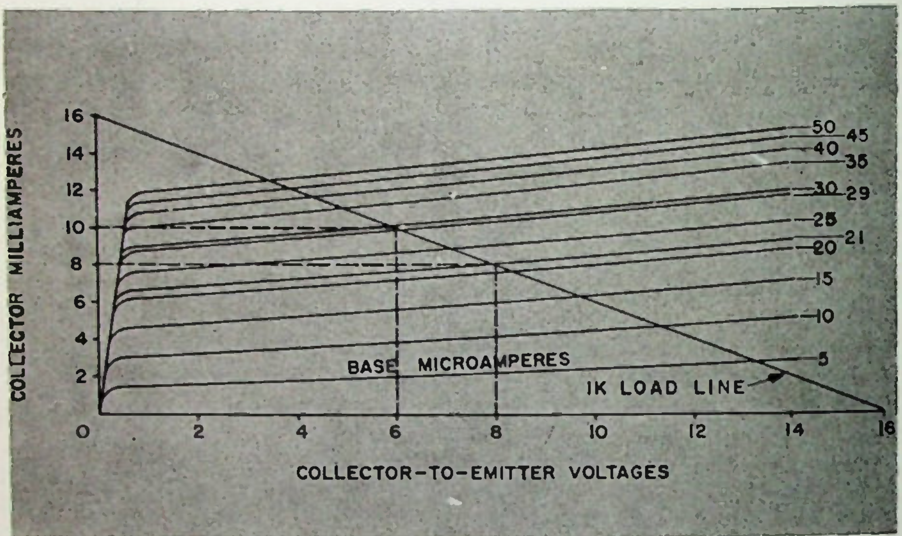


Fig. 4—The load line of the variable-current-source amplifier circuit.

If the effective load line (Fig. 7) for the nearly constant current circuit (Fig. 5) could be extended to both coordinates, we would see that

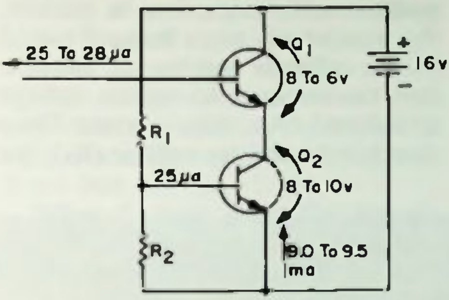


Fig. 5—A constant-current-source amplifier circuit.

it would indicate a maximum current of 11ma and a maximum potential of 44v. The conditions experienced by transistor Q_1 are nearly the same as those it would experience in a circuit (Fig. 8) containing a 44v source and a 4K load resistor. The effective load line, obviously, cannot extend over this range of conditions. In this circuit (Fig. 5), the collector-to-emitter voltage of transistor Q_1 cannot be greater than the actual 16v source. As the collector-to-emitter voltage in transistor Q_1 exceeds 15v, the collector-to-emitter voltage in transistor Q_2 becomes less than 1v, and there is a sharp decrease in the amount of current passing through the transistor and circuit.

Suppose we were able to substitute another transistor, for the current regulating transistor Q_2 in the circuit (Fig. 5), that was able to regulate the current 10 times as effectively as before. When the collector-to-emitter voltage in this transistor changes from 8 to 10v, the collector current changes from 9.00 to 9.05ma. Under these new conditions transistor Q_1 conducts 9.05ma when the collector-to-emitter potential is 6v, and it conducts 9.00ma when the collector-to-emit-

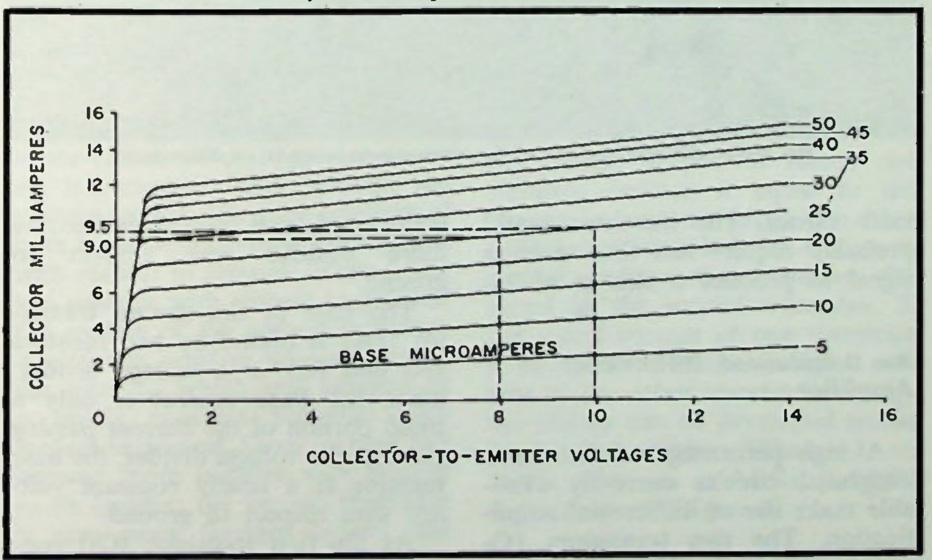


Fig. 6—The characteristic curve of the current regulating transistor.

ter potential is 8v. If these new points were plotted on the transistor's characteristic curve, we would see that the maximum current conducted in the circuit would be 9.2ma and that there would be an effective potential of 368v. These conditions experienced by transistor Q_1 are nearly the same as those it would experience in a circuit (Fig. 8) containing a 368v supply and a 40K load resistor. The transistor would probably require less than a ½μa signal to produce a change of 2v.

An Unbalanced Differential Amplifier

A high percentage of the linear integrated circuits currently available make use of differential amplification. The two transistors (Q_1 and Q_2) in a simplified differential amplifier circuit (Fig 9) have separate collector resistors (R_1 and R_2) but a common emitter resistor (R_3). When the base of the first transistor (Q_1) is made more positive with respect to its emitter, the transistor is more forward biased — the collector current will increase and the collector-to-emitter voltage is reduced. As more current flows through the emitter resistor (R_3), the emitters of both transistors become more positive with respect to ground.

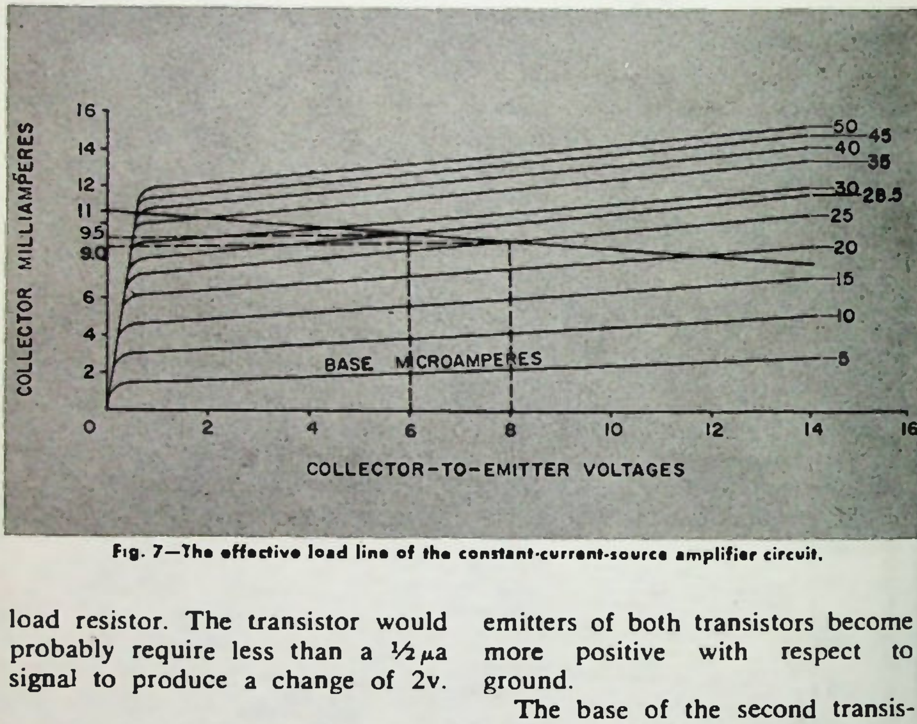


Fig. 7—The effective load line of the constant-current-source amplifier circuit.

The base of the second transistor (Q_2) is biased by two resistors (R_4 and R_5) in a voltage divider. Since the base current is only a small portion of the current passing through the voltage divider, the base remains at a nearly constant voltage with respect to ground.

As the first transistor (Q_1) conducts more current and the emitters of both transistors become more positive with respect to ground, the emitter of the second transistor (Q_2) also becomes more positive with respect to the base — the base remaining at a nearly constant voltage above ground.

Since the base of the second transistor is less positive with respect to the emitter, the transistor (Q_2) is less forward biased and conducts less current.

When the base of the first transistor (Q_1) is made less positive with respect to the emitter, the transistor is less forward biased, and its col-

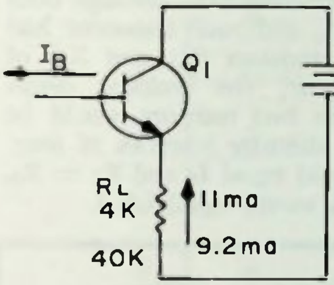


Fig. 8—An equivalent to the constant-current-source amplifier circuit.

lector current decreases as its collector-to-emitter voltage increases. As less current flows through the emitter resistor (R_3), the emitters of both transistors become less positive with respect to ground. While being less positive with respect to ground, the emitter of the second transistor is also less positive with respect to the base — which still remains at a nearly constant potential with respect to ground. Since its base has become more positive with respect to its emitter, the second transistor is now more forward biased and conducts more current.

When the emitter current in the first transistor increases, the emitter current in the second transistor decreases. Since the voltage drop in the common emitter resistor is the sum of the voltage drops caused by the two transistors [$V_{emitter} = (I_1 + I_2)\,R_3$ or $I_1R_3 + I_2R_3$], the amplitude of the emitter signal is reduced by the two opposing changes in current.

As the first transistor's collector current (I_1) increases, the second transistor's collector current (I_2) decreases, and when the first transistor conducts less current, the second transistor conducts more current. As a result of this situation, the voltage drop across one collector resistor (I_1R_1) increases as the voltage drop across the other collector resistor (I_2R_2) decreases and vice-versa. Since the signal present on one collector resistor is present in inverted form on the other collector resistor, the signal applied to the base of the first transistor results in a push-pull signal at the circuit's output.

There is no way of biasing the two transistors in this circuit so that the change in current across one collector resistor is equal to the change in current across the other collector resistor. It is the change in emitter voltage that produces the signal in the second transistor. If the signal current of one transistor is equal to the alternate signal current of the other transistor, no signal voltage can be developed across the emitter resistor (When $\Delta I_1 = -\Delta I_2$, $\Delta V_{emitter} = \Delta I_1R_3 + \Delta I_2R_3 = \Delta I_1R_3 - \Delta I_1R_3 = 0$).

We can more easily compare the

currents flowing through the two transistors (Q_1 and Q_2) by applying values to the differential amplifier circuit (Fig. 10). By changing the forward bias of the first transistor (Q_1), the voltage drop across its collector resistor (R_1) varies from about 1 to 4v, and the voltage drop across the common emitter resistor (R_3) varies from 8 to 10v. The voltage drop across the first transistor (V_1) must, therefore, vary from 7 to 2v [16v — (1v + 8v) = 7v, and 16v — (4v $\times$ 10v) = 2v].

The change in the voltage drop across the common emitter resistor changes the bias of the second transistor (Q_2), and as a result of this change, the voltage drop across its collector resistor (R_2) varies from 4 to 1v. The voltage drop across the second transistor must, therefore, vary from 4 to 5v [16v — (4v + 8v) = 4v, and 16v — (1v + 10v) = 5v].

Since the voltage drop across the 1K common emitter resistor (R_3) varies from 8 to 10v, the current (I_3) flowing through this resistor must vary from 8 to 10 ma

$$\left[\frac{8\text{v}}{1000\Omega} = 0.008\text{a} = 8\text{ma, and } \frac{10\text{v}}{1000\Omega} = 0.010\text{a} = 10\text{ma}\right].$$

Since this resistance (R_3) is in series with both transistors (Q_1 and Q_2), the current passing through the resistor must equal the total of the currents passing through the two transistors ($I_3 = I_1 + I_2$).

If the same amount of current could alternately flow through both transistors, and each transistor had collector resistors (R_1 and R_2) of equal value, the voltage drops across the two resistors would be equal at alternate intervals of time. (If I_1 could equal I_2 and $R_1 = R_2$, then I_1R_1 would equal I_2R_2).

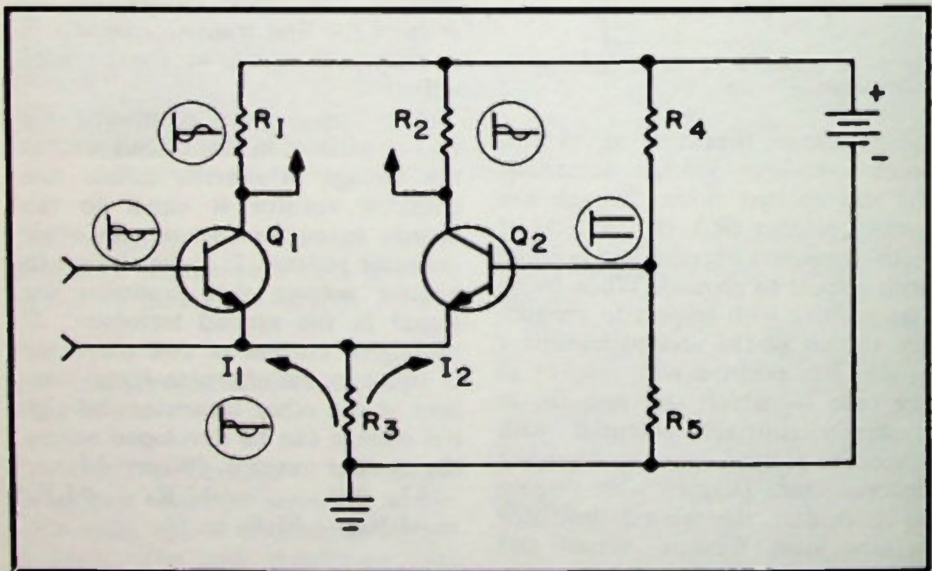


Fig. 9—An unbalanced differential amplifier circuit.

When the forward bias of the first transistor (Q_1) results in a 7v collector-to-emitter potential, the transistor conducts sufficient current to produce an 8v drop across the common emitter resistor (R_3). Under these conditions, 8ma of current passes through this resistor and is the total current passing through the two transistors.

When the two collector resistors (R_1 and R_2) are of equal value, the voltage drop across one is four times that across the other, and 1/5th of the total current passes through one resistor while the remaining 4/5ths of the total current passes through the other resistor. The first transistor must, therefore, be biased to conduct 1.6ma while

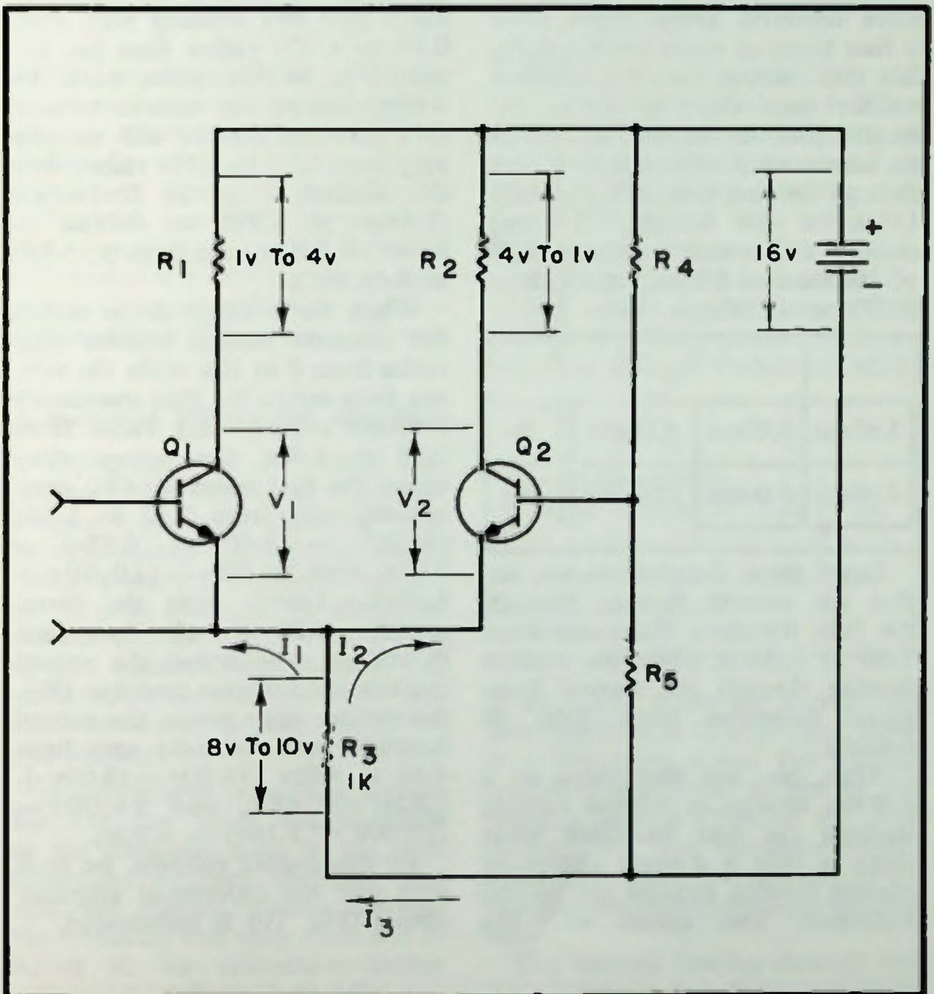


Fig. 10—Voltages present in the unbalanced differential amplifier circuit.

the second transistor must be biased to conduct 6.40ma ($1/5 \times 8.00$ma $= 1.60$ma, and $4/5 \times 8.00$ma $=$ 6.40ma).

When the forward bias of the first transistor results in a 2v collector-to-emitter voltage, the transistor conducts sufficient current to produce a 10v drop across the common emitter resistor. Under these conditions, the common emitter resistor conducts 10ma. Since there is four times as much voltage drop, this time, across the first collector resistor than there is across the second collector resistor, 4/5ths of the current (8.00ma) must now flow through the first transistor and only 1/5th of the current (2.00ma) through the second transistor ($4/5 \times 10.00$ma $= 8.00$ma, and $1/5 \times$ 10.00ma $= 2.00$ma).

1v	4v		
1.60ma	6.40ma	8.00ma	8v
2.00ma	8.00ma	10.00ma	10v

From these calculations we see that the current flowing through the first transistor fluctuates from 1.60 to 8.00ma while the current flowing through the second transistor fluctuates from 2.00 to 6.40ma.

Thus, we see that there is a 6.40ma change in current flowing through the first transistor while there is only a 4.40ma change in current flowing through the second transistor. This agrees with the earlier statement that the signal current through the first transistor must be greater than the signal current through the second transistor. Since this unbalanced condition does exist, the voltage drops across the two resistors are not quite equal at alternate intervals of time. If the value of each collector resistor is 550Ω, the voltage across the first transistor's collector resistor (R_1) will actually vary from 0.88 to 4.40v rather than the desired 1 to 4v fluctuation, while the voltage across the second transistor's collector resistor will actually vary from 3.52 to 1.10v rather than the desired 4 to 1v fluctuation (1.6ma $\times$ 550Ω $=$ 0.016a $\times$ 550Ω $=$ 0.88v, and 8ma $\times$ 550Ω $=$ 4.4v, etc.).

When the voltage drops across the common emitter resistor (R_3) varies from 8 to 10v while the voltage drop across the first transistor's collector resistor (R_1) varies from 0.88 to 4.40v, the voltage drop across the first transistor (V_1) must actually vary from 7.12 to 1.60v [16.00v $-$ (8.00v $+$ 0.88v) $=$ 7.12v, and 16.00v $-$ (10.00v $+$ 4.40v) $=$ 1.60v]. With the corresponding 3.52 to 1.10v fluctuation in voltage drop across the second transistor's collector resistor (R_2), the voltage drop across the second transistor must actually vary from 4.48 to 4.90v [16.00v $-$ (8.00v $+$ 3.52v) $=$ 4.48v, and 16.00v $-$ (10.00v $+$ 1.10v) $=$ 4.90v].

By substituting voltages, we have seen that the differential amplifier circuit (Fig. 10) is unbalanced.

CHAPTER 10

IC Differential Amplifiers

The previous chapter describes a constant-current-source amplifier and an unbalanced differential amplifier. By combining some of the features of these two amplifiers, a differential amplifier can be made that is virtually balanced. Balanced differential amplifiers are frequently used in the integrated linear amplifier circuits found in radios and TV sets.

A Balanced Differential Amplifier

The unbalanced differential amplifier circuit, shown in Fig. 9 of chapter 9, can be revised (Fig. 1) to be more nearly balanced by substituting a current-regulating transistor (Q_3) in place of the common-emitter resistor (R_3). The voltage divider resistors (R_4, R_5, and R_6) serve to provide nearly constant base bias currents to the two transistors (A_2 and Q_3).

If the current regulating transistor has the same characteristics as the transistor used in the constant-current-source amplifier circuit, the current flowing through the transistor will vary from 9.0 to 9.5ma as its collector-to-emitter voltage varies from 8 to 10v.

If the pair of transistors (Q_1 and Q_2), used in the unbalanced differential amplifier circuit (Fig. 10 *of the previous* chapter), are biased as before; 1/5th of the resulting current (1.8ma) will flow (Fig. 2) through the first transistor (Q_1), when there is an 8v drop across the current regulating transistor (Q_3), and the remaining current (7.2ma) will flow through the second transistor (Q_2). When the bias in the first transistor is changed and there is a 10v drop across the current regulating transistor, 4/5ths of the resulting current (7.6ma) will flow through the first transistor (Q_1), while the remaining current (1.9ma) will flow through the second transistor.

1v	4v		
1.8ma	7.2ma	9.0ma	8v
1.9ma	7.6ma	9.5ma	10v

TABLE 1

The current flowing through the first transistor has fluctuated from

1.8 to 7.6ma, resulting in a 0.99 to 4.18v fluctuation across its 550 Ω collector resistor (R_1); and the current flowing through the second transistor has fluctuated from 7.2 to 1.9ma, resulting in a 3.96 to 1.045v fluctuation across its 550 Ω collector resistor (R_2). The voltage drop across the first transistor fluctuates between 7.01 and 1.82v, while the voltage drop across the second transistor fluctuates between 4.04 and 4.955v.

The 0.99 to 4.18v fluctuation ates between 8 and 10v, the current through the first collector resistor (R_1) will vary from 1.80 to 7.24ma, and the current through the second resistor (R_2) will vary from 7.20 to 1.81ma.

The voltage drop across the first resistor will then vary from 0.99 to 3.98v while the voltage drop across the second resistor will vary from 3.96 to 0.9955v. By rounding these voltages off to a practical number of significant figures, we see that the differential amplifier circuit can

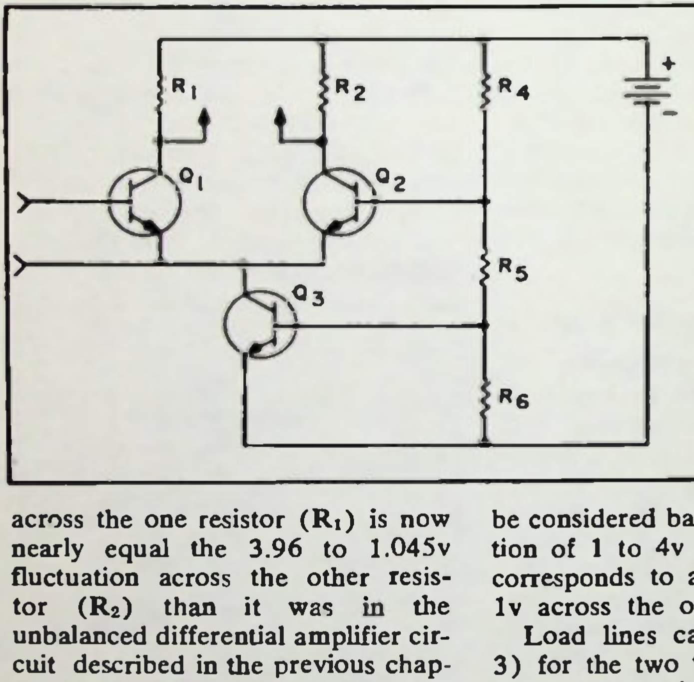


Fig. 1—A balanced differential amplifier circuit.

across the one resistor (R_1) is now nearly equal the 3.96 to 1.045v fluctuation across the other resistor (R_2) than it was in the unbalanced differential amplifier circuit described in the previous chapter. This differential amplifier can be even better balanced if we substitute another transistor with better current stability for the third transistor (Q_3). If the current through this substituted transistor fluctuates only from 9.00 to 9.05ma as the collector-to-emitter voltage fluctu- be considered balanced and a variation of 1 to 4v across one resistor corresponds to a variation of 4 to 1v across the other resistor.

Load lines can be drawn (Fig. 3) for the two transistors (Q_1 and Q_2) using a pair of known voltage-current combinations for each transistor. When the voltage drop across the first transistor's collector resistor (R_1) is 0.99v, the transistor's collector-to-emitter voltage must be 7.01v, and both the resistor and transistor conduct 1.80ma. When

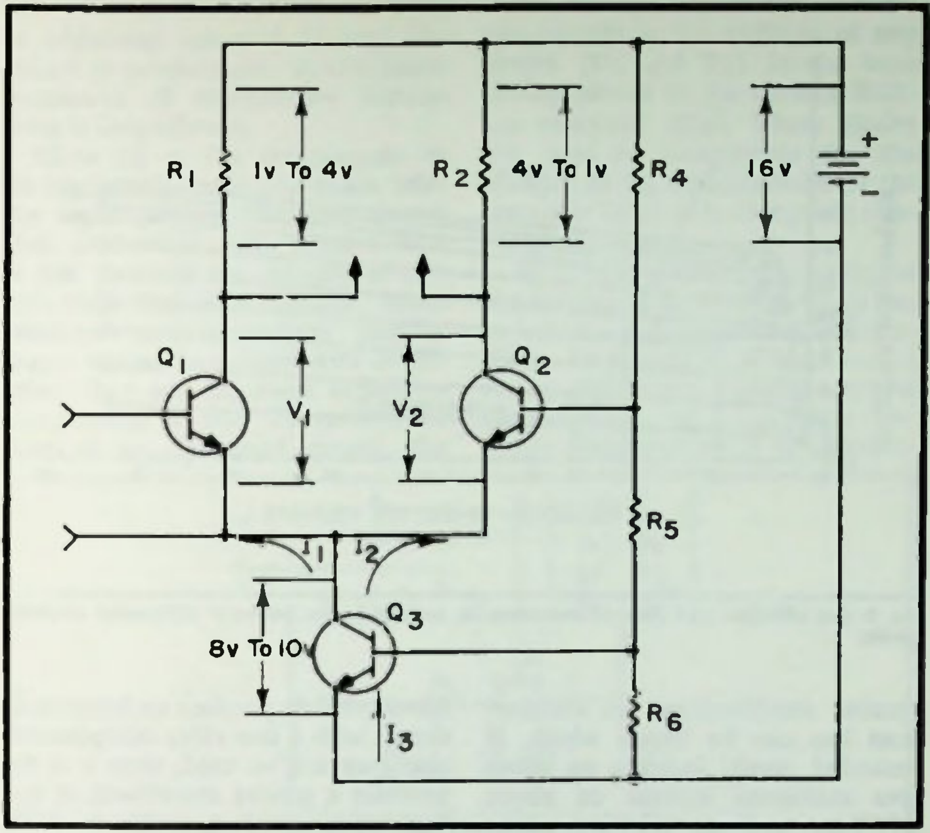


Fig. 2—Voltages present in the balanced differential amplifier circuit.

the voltage drop across the collector resistor is 3.982v, the collector-to-emitter voltage is 2.018v, and the current is 7.24ma. By connecting the two points representing these voltage-current combinations, an effective load line can be drawn. When extended, this load line indicates an effective maximum current of about 9.44ma and an effective maximum potential of about 8.67v. This would correspond to an effective load resistance of about 900 Ω.

When the voltage across the second transistor is 4.04v, it conducts 7.20ma; and when 5.0045v, it conducts 1.81ma. By connecting the points representing these voltage-

1v	4v		
1.80ma	7.20ma	9.00ma	8v
1.81ma	7.24ma	9.05ma	10v

TABLE II

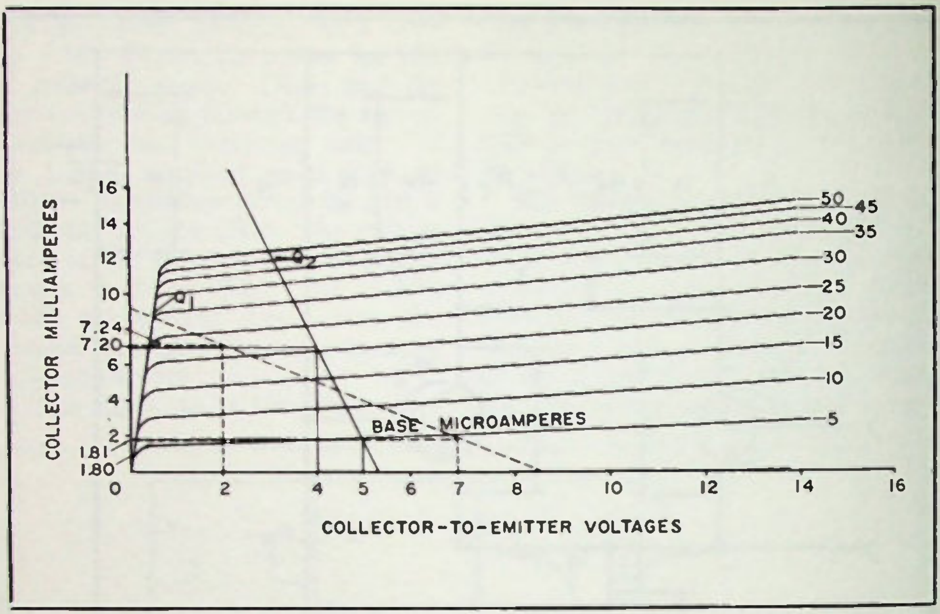


Fig. 3—The effective load lines of transistors Q_1 and Q_2 in the balanced differential amplifier circuit.

current combinations, an effective load line can be drawn which, if extended, would indicate an effective maximum current of about 29.78ma and an effective maximum potential of about 5.25v. This would correspond to an effective load resistance of about 200Ω.

The CA3005 Integrated Circuit

The initial design of an integrated circuit is a major factor in determining the cost of the circuit. Once a circuit has been drawn and the photographic masks are prepared (Fig. 4) for various "growth" processes, the design is ready for mass production. The greater the production run of these circuits, the more economically they can be produced. Because of the nature of the expenses encountered, it is more economical to produce an integrated circuit with a few extra components that may not be used, than it is to produce a greater assortment of integrated circuit designs. The amount

Fig. 4—The enlarged artwork for an integrated circuit is shown on the screen of a 25-ft. long camera. Courtesy of Cutler-Hammer, Inc.

of additional material, if any, required to produce one or two extra transistors of microscopic dimensions is insignificant.

Since all of the components in an integrated circuit are made with the same process, the only factor that determines their relative cost is the area of the integrated circuit chip that they require. Since resistors and capacitors require larger areas than transistors or diodes, they are the most expensive components to use. To reduce the cost of an integrated circuit, the circuit should be designed to include as few capacitors and resistors as possible, even if several more semiconductors are required to accomplish this.

The manufacturer's schematic of integrated circuit CA3005 is shown in Fig. 5. This circuit can be more readily understood if it is shown as a standard transistor circuit (Fig. 6). Note the similarity between the basic portions of this circuit and the balanced differential amplifier circuit discussed earlier (Fig. 1). The major difference between the two circuits is the addition of two diodes (D_1 and D_2) in the base biasing circuit of the current limiting transistor (Q_3). These diodes are used to compensate for the changes in the characteristics of the transistor (Q_3) with changes in temperature.

By attaching external circuitry to the integrated circuit (Fig. 7), it can be used as a balanced differential amplifier. *Because of the arrangement of the integrated circuit components on the substrate, the manufacturer indicates that terminal 9 of the integrated circuit must be at least as positive as any other portion of the integrated circuit. This terminal has, therefore, been connected directly to the collector voltage source* (V_{CC}). *Not all integrated circuits have such a requirement.* This diagram can be simplified without actually changing the circuit (Fig. 8). No actual change has been made in the integrated circuit — the unused leads are no longer shown, and for simplicity series resistors R_2 and R_3 are shown as a single resistor ($R_2 + R_3 = R_6$), and resistors

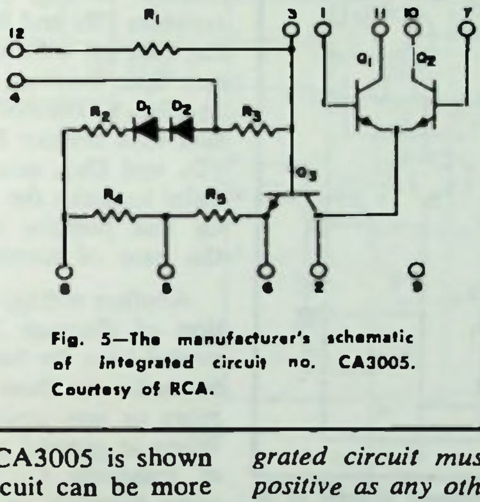


Fig. 5—The manufacturer's schematic of integrated circuit no. CA3005. Courtesy of RCA.

R_4 and R_5 are also shown as a single resistor ($R_4 + R_5 = R_7$).

The 1K collector resistors attached to leads 10 and 11 serve the same function as the 550Ω resistors described with Fig. 1 and 2. A 0.1µf capacitor connected to these two resistors shunts to ground any portion of the ac signal that would otherwise appear across the collector voltage source (V_{CC}).

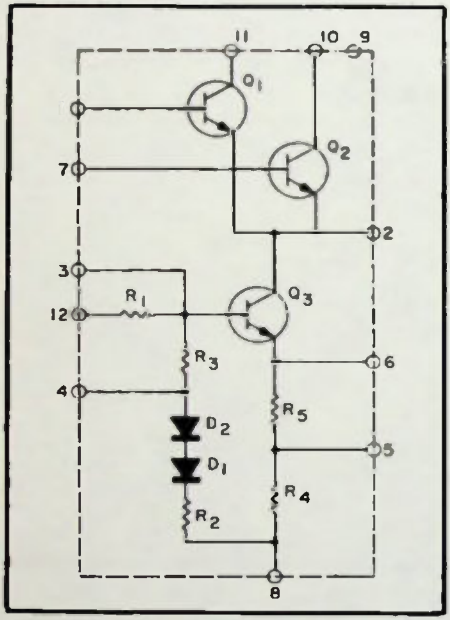


Fig. 6—A more conventional diagram of integrated circuit no. CA3005.

The collector and emitter voltage sources (V_{CC} and V_{EE}) not only supply power to the circuit, but act as a voltage divider for the circuit. Since the common battery terminal and the base terminal of transistor Q_2 are both connected to ground, one voltage source (V_{CC}) serves to make the transistor's collector more positive than the base while the other voltage source (V_{EE}) serves to make the transistor's emitter more negative than the base. The corresponding transistor (Q_2) in Fig. 1 received the same base bias current through a resistor voltage divider circuit (R_4, R_5 and R_6).

Another 0.1µf capacitor connected to the emitter voltage source (V_{EE}) shunts to ground the ac signal that would otherwise appear across the battery.

The base of the current limiting transistor is biased in a manner similar to the one described previously. In Fig. 1, voltage divider resistors (R_5 and R_6) serve to make the base of transistor Q_3 less positive than the base of transistor Q_2. In Fig. 8, resistor R_1, in conjunction with resistor R_6 and the diodes (D_1 and D_2), acts as a voltage divider to make the base of transistor Q_3 less positive than ground and the base of transistor Q_2.

Another voltage divider (left portion of diagram in Fig. 8), connected to a 3v battery, is designed to make the base of transistor Q_1 more or less positive than ground. When at ground potential, the base of transistor Q_1 is biased at the same potential as transistor Q_2. By changing the base bias of transistor Q_1, that transistor (Q_1) conducts more or less current while transistor Q_2 conducts less or more current. Their alternate fluctuations of current occur in the same manner discussed earlier for Fig. 2.

Certain terms are used by the manufacturer to list integrated circuit characteristics. The input offset voltage (V_{IO}) is the difference in the dc voltages that must be applied to the input terminals to obtain a zero output voltage differential at the output terminals. This is the amount of voltage that must be applied between terminal 1 and ground (Fig. 8) to reduce to zero any difference in potential that may occur between terminals 10 and 11 of the integrated circuit. The manufacturer indicates that a typical input offset voltage for an integrated circuit of type CA3005 is 2.6mv, while the maximum input offset voltage is 5mv.

It is not economically feasible to produce every integrated circuit with identical characteristics. The characteristics vary slightly with production batches. Some integrated circuits of the same design have a typical input offset voltage of only 0.8mv and a maximum voltage of 1mv. These circuits are classified as type CA3006. This is the only rated difference listed by the manufacturer for these two types of integrated circuits. The price of type CA3006 integrated circuits was listed by an electronics supply company for $4.00 more than the price of the type CA3005 circuits. (As had been mentioned in the previous chapter, that electronics supply company listed the type CA3005 circuits for $2.80.)

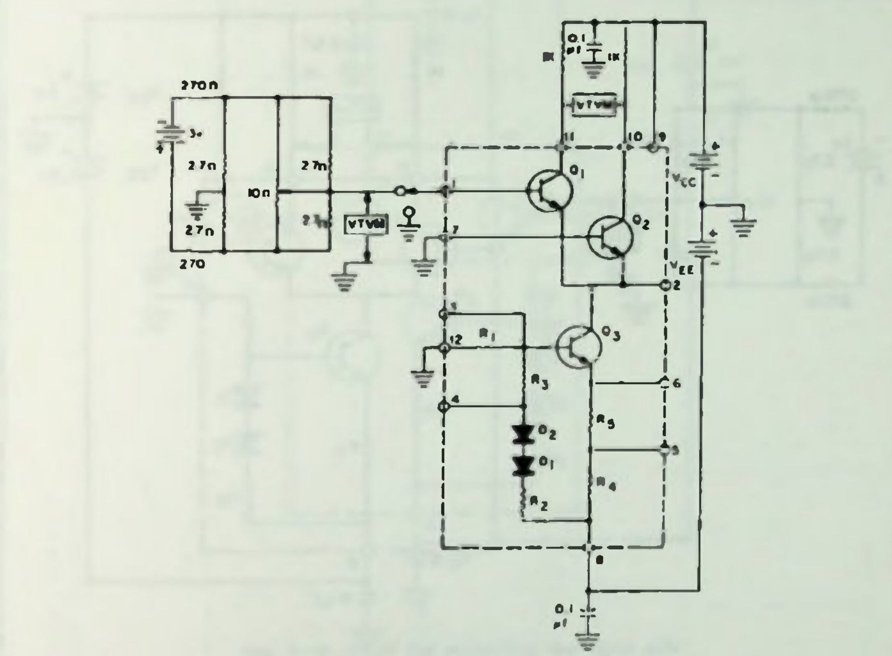


Fig. 7—A differential amplifier using the CA3005 integrated circuit.

The input offset current (I_{IO}) is the difference in the amount of current passing through the base leads of the two transistors (Q_1 and Q_2) when their collectors are connected directly to the voltage source (V_{CC}). This is measured (Fig. 9) by connecting one milliammeter between terminal 1 and ground and another milliammeter between terminal 7 and ground, while terminals 10 and 11 are connected, along with terminal 9, directly to the voltage source. The typical input offset current for these circuits is 1.4μa.

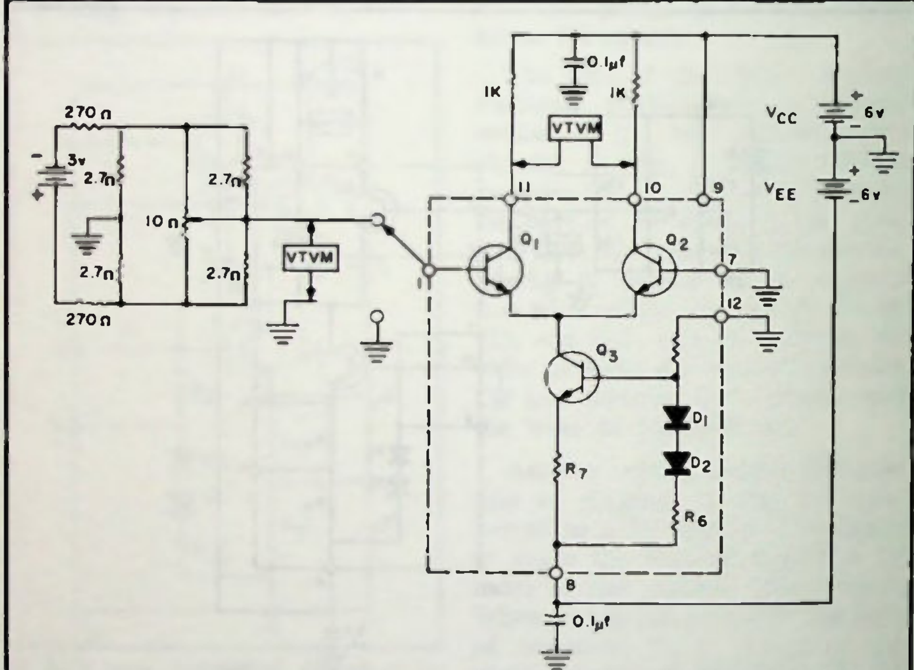


Fig. 8—A simplified diagram of the CA3005 differential amplifier circuit.

The input bias current (I_1) is the average value of the two currents measured when determining the input offset current. That current, as you probably know, would be half the sum of the two currents.

The quiescent operating current is the average dc collector current of either of the two transistors (Q_1 or Q_2). This current is measured (Fig. 10) by connecting the base of both transistors to ground (terminals 1 and 7 to ground) and connecting milliammeters between the transistor's collectors and the voltage source (between terminal 11 and V_{CC}) and between terminal 10 and V_{CC}). A typical value for both of these currents is 1ma.

There will be a slight difference in the two quiescent operating current measurements since the input offset voltage is now zero — the bases of both transistors being grounded. The ratio of these two currents is called the quiescent operating current ratio, and it has a typical value of 1.05.

The characteristic quiescent operating current can be increased by changing the bias of the current regulating transistor (Q_3). If terminal 5 (Fig. 6) is connected to terminal 8 and the emitter voltage source (V_{EE}), the emitter of the transistor will be more negatively biased with respect to its base. [We have reduced the value of resistor R_7 in Fig. 10 (R_7 now equals R_5 instead of $R_4 + R_5$).] The base of transistor Q_3 has been made more positive with respect to its emitter, and the transistor conducts more current to the other two transistors (Q_1 and Q_2). As a result of this change, the circuit's typical quiescent operating current has increased to 2.7ma.

By disconnecting terminal 5 and connecting terminal 4 to terminal 8 (Fig. 6) and the emitter voltage source (V_{EE}), a resistor (R_2) and two diodes (D_1 and D_2) are shorted out of the circuit, making the base of transistor Q_3 less positive with respect to its emitter. Under these conditions, the current limiting transistor conducts less current to the other two transistors (Q_1 and Q_2), and the characteristic quiescent operating current is reduced to 0.45ma.

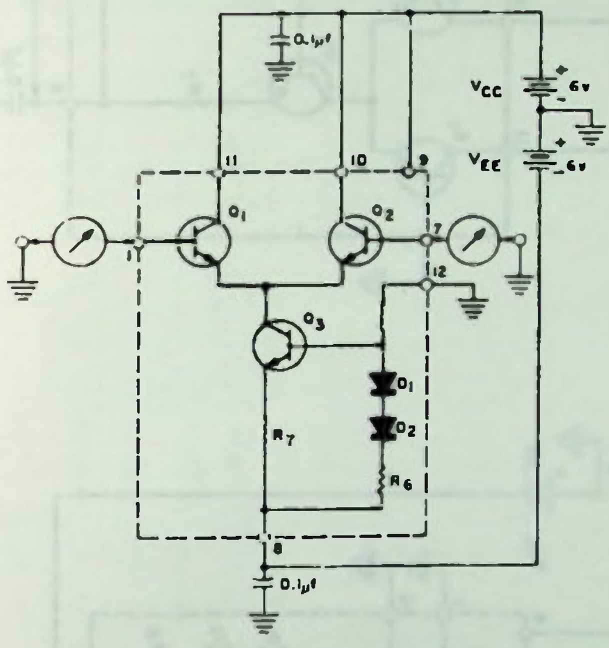


Fig. 9—A circuit for measuring the input offset current and input bais current of the CA3005 differential amplifier circuit.

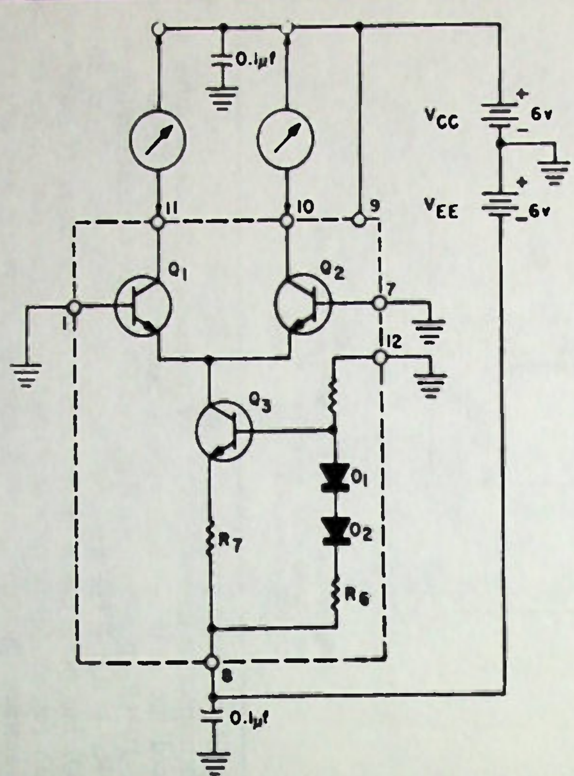


Fig. 10—A circuit for measuring the quiescent operating current of the CA3005 differential amplifier circuit.

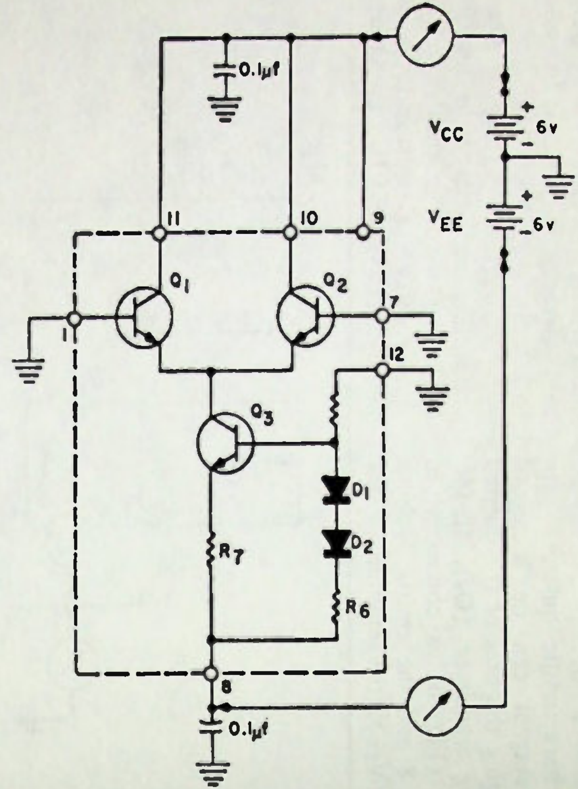


Fig. 11—A circuit for measuring the device power dissipation of the CA3005 differential amplifier circuit.

If both terminals 4 and 5 are connected to terminal 8 and the emitter voltage source, the bias of transistor Q_3 is again changed, and the characteristic quiescent operating current is 1.25ma.

The device dissipation (P_T) is the total power drain of the integrated circuit (Fig. 11) when no signal is applied to it and there are no external load resistances. This power is equal to the product of the total collector current (I_{CC}) times the collector voltage (V_{CC}) plus the product of the total emitter current (I_{EE}) times the emitter voltage (V_{EE}).

$$(P_T = I_{CC}V_{CC} + I_{EE}V_{EE}.)$$

CHAPTER 11

IC Applications in FM Circuits

The two previous chapters describe the principles of unbalanced and balanced differential amplifier circuits and relate them to the CA 3005 integrated circuit. Now, let's consider the application of a similar type.

An Integrated FM IF Strip

A similar integrated circuit, type μA703 (Fig. 1), has been designed for an FM IF strip. The manufacturer's diagram of this circuit is shown in Fig. 2. This diagram can be changed to a form that we are more familiar with (Fig. 3). A fourth transistor (Q_4) is used in the circuit in place of the diodes (D_1 and D_2) used in the other integrated circuit for temperature compensation. The temperature characteristics of transistor Q_3 should be quite similar to those of transistor Q_4. This additional transistor (Q_4) provides the same function as resistor R_6, described with Fig. 1 in chapter 10. The collector-to-emitter voltage drop across transistor Q_4 produces the desired base-to-emitter bias voltage for the current-limiting transistor (Q_3).

Still another transistor Q_5 (Fig. 3) provides the same function as resistor R_5, described with Fig. 1 of the previous chapter. It is used as part of a voltage divider to supply transistor Q_2 with nearly constant base-bias current.

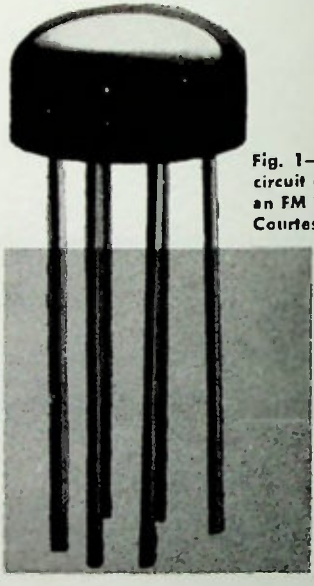


Fig. 1—An integrated circuit designed for an FM IF strip. Courtesy of Scott.

The manufacturer's schematic of the FM IF strip is shown in Fig. 4. A simplified version of the 3rd IF amplifier stage is shown in Fig. 5. Transistors Q_1 and Q_2, in this circuit (Fig. 5), function in nearly the same manner as transistors Q_1 and Q_2 described in chapter 10. When no current is developed across the secondary winding of transformer T_2, nearly all the dc bias current passes through the winding, and the base of transistor Q_1 has the same amount of bias as the base of transistor Q_2. Since there is very little dc resistance in transformer T_3, nearly the same amount of collector current from the two transistors flows through the common collector resistor and terminal 1. Both transistors (Q_1 and Q_2) are maintained at nearly identical dc collector-to-emitter voltages.

When a radio-frequency (RF), positive signal current is induced into the secondary winding of transformer T_2, the base of transistor Q_1 becomes more forward biased, the transistor conducts more current and a reduction occurs in its collector-to-emitter voltage. The RF portion of the collector current from transistor Q_1 is shorted to ground by the 0.02μf capacitor connected to terminal 2 of the integrated circuit. This capacitor prevents an ac signal from developing across the common collector resistor. The RF change in collector-to-emitter voltage across transistor Q_1 can result in a RF change in the potential of the emitter only — the ac potential of the collector being shorted to ground.

The RF positive signal present at the base of transistor Q_1 results in an amplified RF positive signal at the emitter of the same transistor. Since the emitter of transistor Q_2 is connected to the emitter of transistor Q_1, there is also an RF positive potential developed at its emitter. Since the base of transistor Q_2 remains at a constant potential, the transistor is less forward biased as the emitter becomes more positive, and the RF positive potential at the emitter results in an RF decrease in current flowing through transistor Q_2. Because of the impedance of the primary winding of transformer T_3 the winding has a relatively high ac resistance, compared to its dc resistance, and a signal

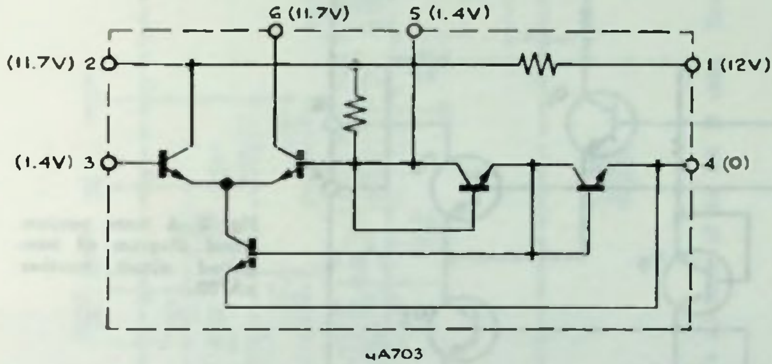


Fig. 2—The manufacturer's schematic of integrated circuit number μA703.

voltage develops across the primary winding of the transformer.

The RF positive signal present at the emitter of transistor Q_2 induces an RF positive signal across the primary winding of transformer T_3. The RF reduction in current flow through transistor Q_2 does not result in an RF change in potential across the common collector resistor. The portion of the RF decrease in current through transistor Q_2 — not neutralized by the RF increase in current through transistor Q_1 — is shorted to ground by the $0.02\mu f$ capacitor connected to terminal 2 of the integrated circuit.

An RF negative pulse present at the base of transistor Q_1 causes an RF decrease in current flow through transistor Q_1 and an RF increase in current flow through transistor Q_2. This results in an RF negative signal across the primary winding of transformer T_3. Again, any portion of the RF increase in current through one transistor, not neutralized by the RF decrease in current through the other transistor, is shorted to ground through the $0.02\mu f$ capacitor, and no signal is developed across the common collector resistor.

Transistor Q_1 functions in this circuit as a common-collector transistor, while transistor Q_2 functions as a common-base transistor.

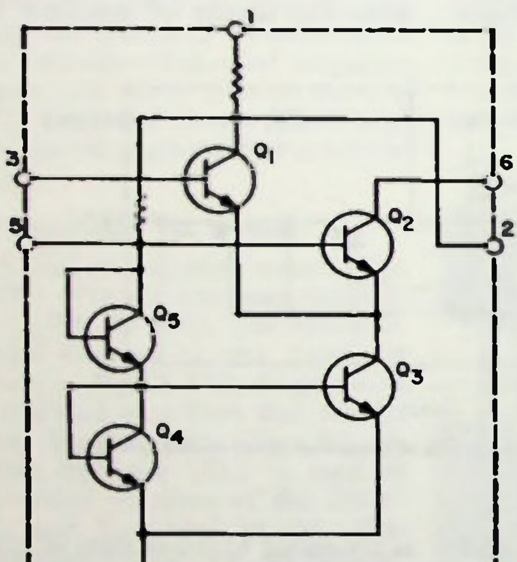


Fig. 3—A more conventional diagram of integrated circuit number μA703.

Because of the constant-current-source characteristics and the method of temperature compensation — with transistors said to operate up to 900MHz — the integrated circuit reportedly has a stable bandwidth greater than 150MHz. The manufacturer indicates that each five-transistor integrated circuit

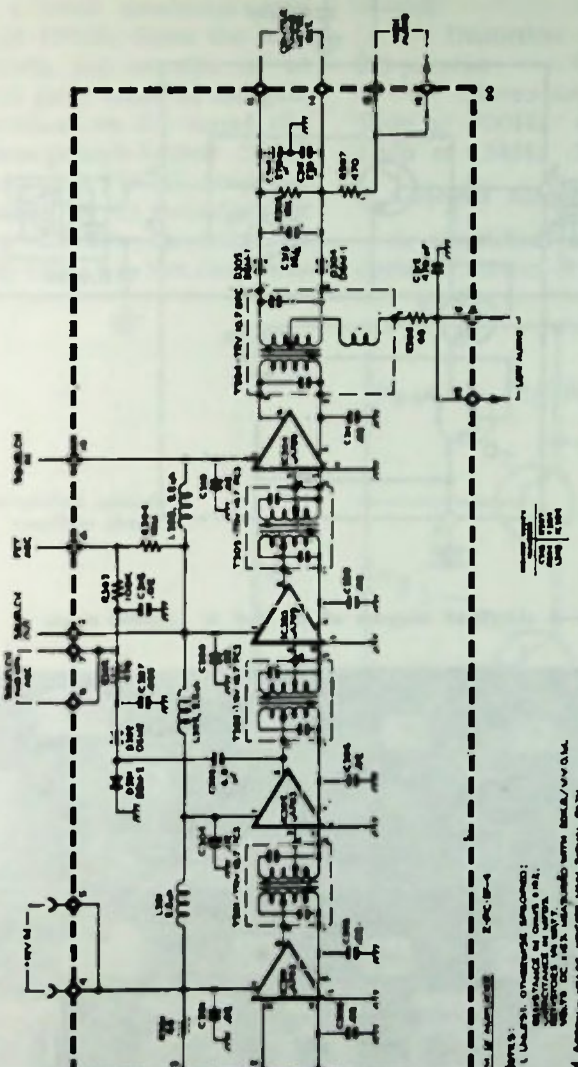


Fig. 4—The manufacturer's schematic of the FM IF strip. Courtesy of Scott.

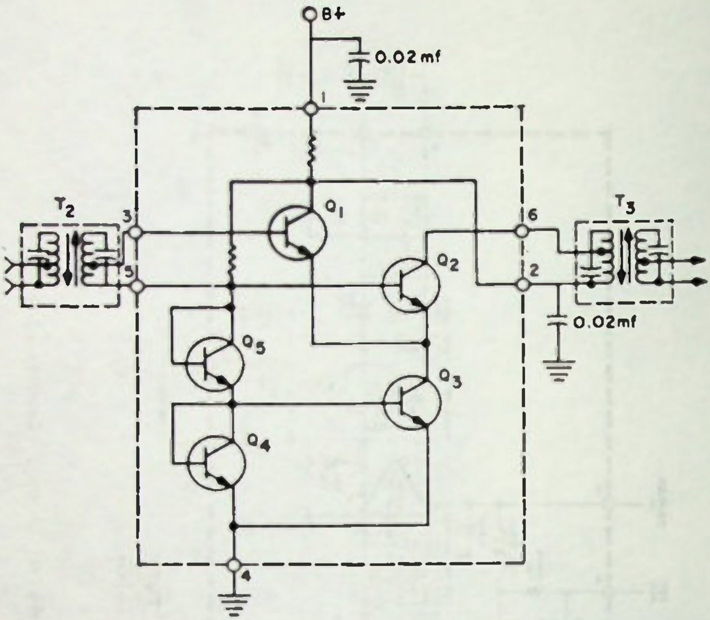


Fig. 5—A simplified diagram of the 3rd IF amplifier stage.

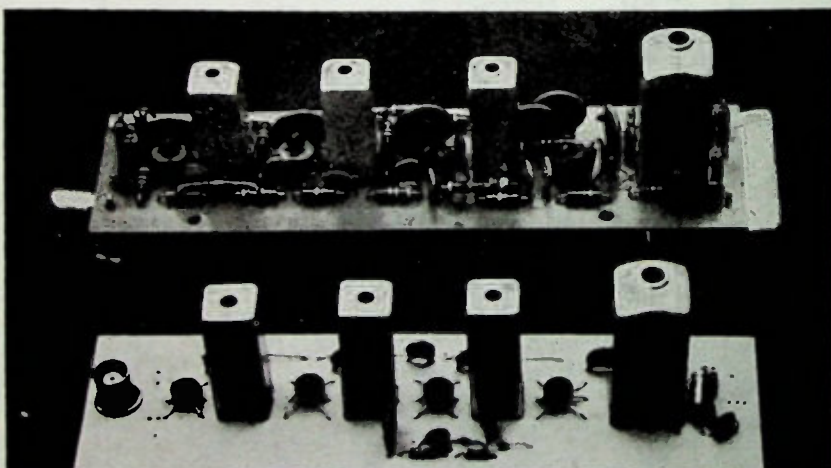


Fig. 6 — A comparative photograph showing the FM IF strip with transistors (above) and with integrated circuits (below). Courtesy of Fairchild.

has a minimum gain of 26db at the 10.7MHz IF. With four of these integrated circuits in series, the IF strip has a total maximum gain capability of 104db. Since the integrated circuits are capable of so large a total gain, some of the gain can be sacrificed in the tuned circuits to insure high-loaded "Q's."

The integrated FM IF strip has been compared by the manufacturer with its predecessor, a transistor FM IF strip (Fig. 6). Not only does the integrated FM IF strip require fewer components than the other strip, it is reported to have better specifications. The manufacturer's comparative specifications are as follows:

Transistor IF	Integrated Circuit IF
Sensitivity	
1.7μv	1.5μv
Capture Ratio	
3db	1.75db
Selectivity	
45db	46db
AM rejection	
—46db	—52db
Distortion at 400Hz	
0.3 percent	0.3 percent
Stereo separation	
35db at 400Hz	41db at 400Hz
19db at 15kHz	30db at 15kHz

A Cascade Amplifier

A simplified cascade amplifier circuit is shown in Fig. 7. Here, the base of transistor Q_1 is forward biased with resistors R_1 and R_2, while the base of transistor Q_2 is forward biased with resistors R_3 and R_4. Since transistors Q_1 and Q_2 are connected in series, and a collector-to-emitter voltage drop will occur in transistor Q_1, the emitter of transistor Q_2 must be more positive than the emitter of transistor Q_1. Hence, if the same amount of forward bias is to be applied to both transistors, the base of tran-

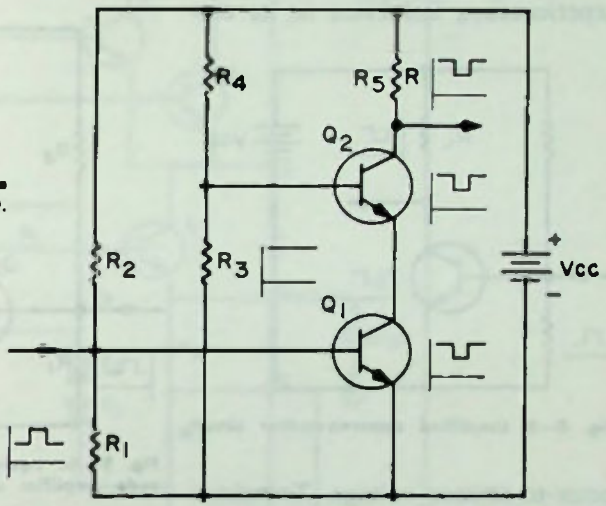


Fig. 7—A simplified cascade amplifier circuit.

sistor Q_2 must be biased more positive than the base of transistor Q_1.

When a positive signal is applied to the base of transistor Q_1, the transistor conducts more current, and a reduction in its collector-to-emitter voltage occurs. Because of the reduction in the voltage drop across transistor Q_1, the emitter of transistor Q_2 becomes less positive. The base of transistor Q_2 is then more positive with respect to its emitter, and since the transistor is now more forward biased, it also experiences a reduction in its collector-to-emitter voltage. Transistor Q_2 functions as though it were in a common-base circuit.

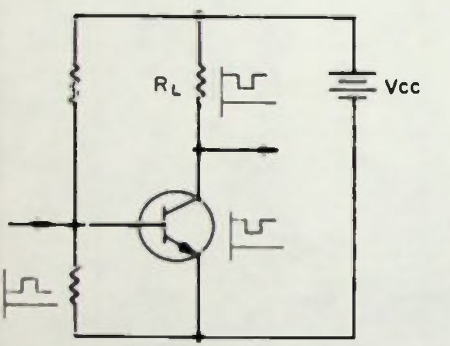


Fig. 8—A simplified common-emitter circuit.

In a simple common-emitter circuit (Fig. 8), a reduction in the transistor's collector-to-emitter voltage results in a corresponding increase in the voltage drop across the load resistor

$$(\Delta V_{CE} = -\Delta I_C R_L).$$

This relationship is not true for transistor Q_1 in the cascade amplifier circuit (Fig. 7). There, the increase in the voltage drop across the load resistor (R_5) is greater than the decrease in the collector-to-emitter voltage drop of transistor Q_1:

$$(\Delta_{CE} < -\Delta I_C R_5)$$

Instead, it is equal to the sum of the collector-to-emitter voltage changes in the two transistors (Q_1 and Q_2).

Transistor Q_2 functions (Fig. 9) as a negative resistor (r_e) in series with a load resistor (R_L). When a positive signal is applied to the base of transistor Q_1 (Fig. 9), the transistor conducts more current. The transistor's increased collector current results in a greater voltage

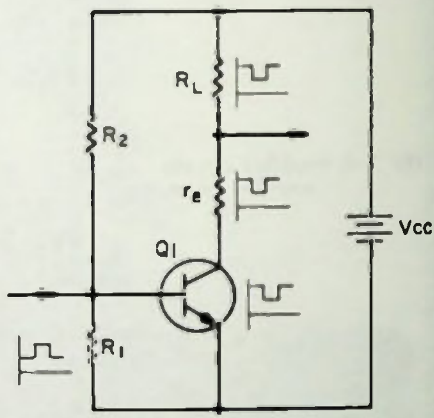


Fig. 9—An equivalent to the simplified cascode amplifier circuit.

drop across the load resistor (R_L) and a reduced voltage drop across the effective negative resistance, r_e. (Although the voltage drop across resistor R_L increases as the voltage drop across resistor r_e decreases, the squarewave pulses developed across both are shown as negative pulses. A greater voltage drop across resistor R_L will make the output less positive with respect to the positive supply-voltage lead, while the reduced voltage drop across the effective negative resistor r_e will make the output less positive with respect to the negative lead.) Transistor Q_1 in Fig. 9 requires a smaller change in its collector-to-emitter voltage, than the transistor in Fig. 6 in chapter 10). By attaching external circuitry to the integrated circuit (Fig. 10), it can be used as a cascode amplifier. This diagram can be simplified without actually

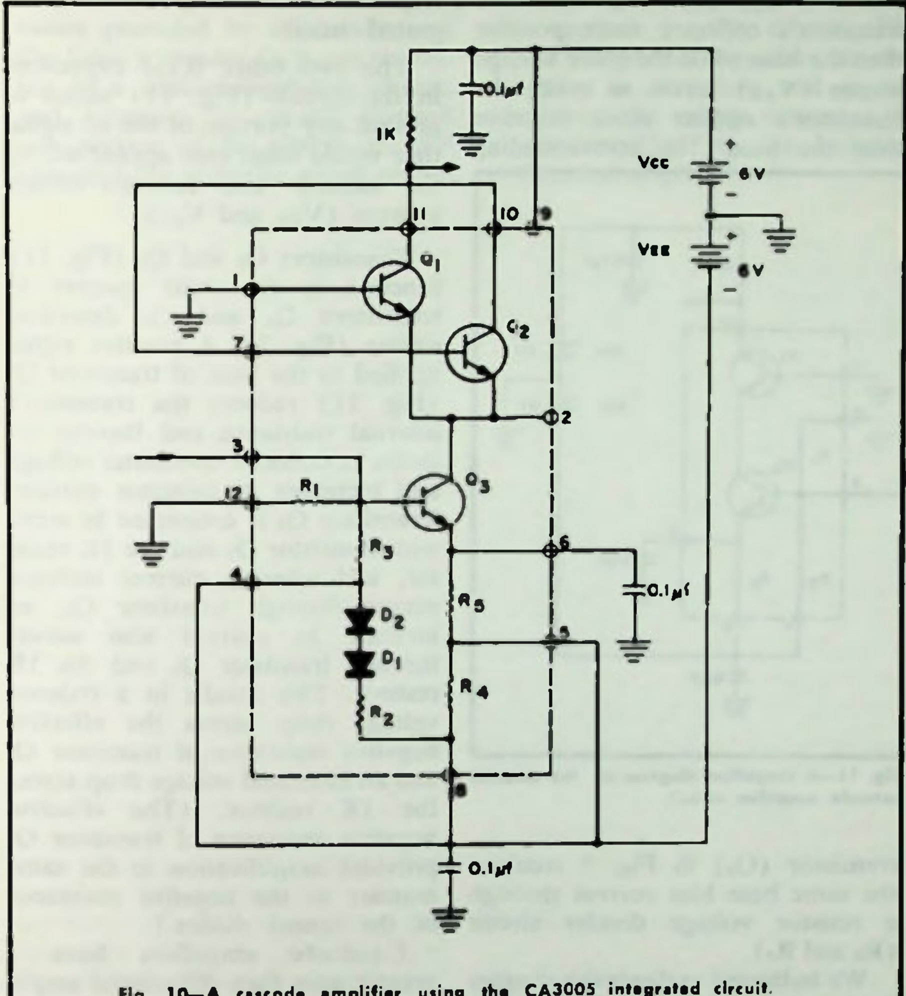


Fig. 10—A cascode amplifier using the CA3005 integrated circuit.

8, to produce the same output signal across an equal load resistor (R_L).

The two previous chapters showed integrated circuit CA3005 as a standard transistor circuit (Fig. changing the circuit (Fig. 11). The unused leads and components are no longer shown.

The collector and emitter voltage sources (V_{CC} and V_{EE}) supply power to the circuit and act as a voltage divider. Since the common battery terminal and the base terminal of transistor Q_1 are both connected to ground, one voltage source (V_{CC}) serves to make the transistor's collector more positive than the base while the other voltage source (V_{EE}) serves to make the transistor's emitter more negative than the base. The corresponding

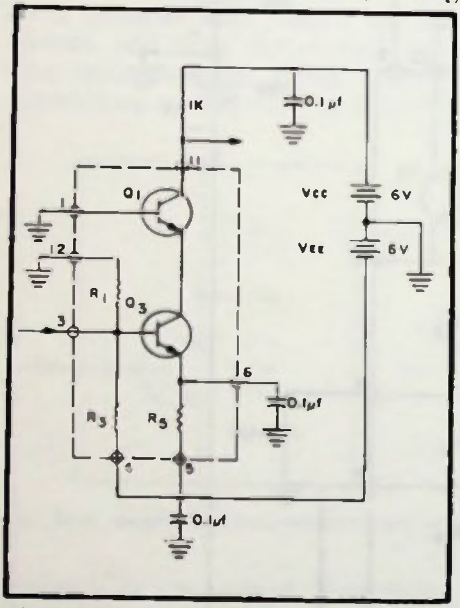


Fig. 11—A simplified diagram of the CA3005 cascode amplifier circuit.

transistor (Q_2) in Fig. 7 received the same base bias current through a resistor voltage divider circuit (R_3 and R_4).

We indicated earlier in this chapter that the base of Transistor Q_1 (Fig. 7) is biased at a more negative potential than the base of transistor Q_2. The corresponding transistor in Fig. 11 (transistor Q_3) is biased at a more negative potential than transistor Q_1 and ground with a resistor voltage divider (R_1 and R_3).

Resistor R_5 and the 0.1μf capacitor connected to terminal 6 serve as an "emitter swamping" circuit to improve the stability of the integrated circuit.

The two other 0.1μf capacitors in the circuit (Fig. 11) shunt to ground any portion of the ac signal that would otherwise appear across the emitter- and collector-voltage sources (V_{EE} and V_{CC}).

Transistors Q_3 and Q_1 (Fig. 11) function in the same manner as transistors Q_1 and Q_2 described earlier (Fig. 7). A positive signal applied to the base of transistor Q_3 (Fig. 11) reduces the transistor's internal resistance and thereby reduces its collector-to-emitter voltage and increases its collector current. Transistor Q_3 is connected in series with transistor Q_1 and the 1K resistor, and when a current increase occurs through transistor Q_3, an increase in current also occurs through transistor Q_1 and the 1K resistor. This results in a reduced voltage drop across the effective negative resistance of transistor Q_1 and an increased voltage drop across the 1K resistor. (The effective negative resistance of transistor Q_1 provides amplification in the same manner as the negative resistance of the tunnel diodes.)

Cascade amplifiers have a greater gain than differential amplifiers. When connected in a circuit designed to operate at 100MHz, the CA3005 integrated circuit reportedly has a 20db power gain (G_p) when operating as a cascade amplifier and a 16db power gain when operating as a balanced differential amplifier.

When operating as a cascade amplifier, the integrated circuit no longer has the temperature compensation provided by diodes D_1 and D_2. Since transistor Q_3 is not operating in a current-regulating circuit, and increases in current through one portion of the circuit are not balanced by virtually equal reductions in current through another portion of the circuit — as in the balanced differential amplifier circuit described earlier — the current drawn by the cascode amplifier is not entirely independent of the signal applied to the circuit. The signal amplified by the cascode amplifier is, therefore, not as isolated from the voltage source as the signal amplified by the balanced differential amplifier.

CHAPTER 12

IC Applications in TV

A line of TV sets currently on the market uses an integrated circuit, IC201, in the sound IF and audio circuits. The silicon chip upon which this circuit is grown measures less than 0.05in. and, when in its case, the IC201 is not much larger than the end of a pencil (Fig. 1).

The manufacturer's schematic of the integrated circuit is shown in Fig. 2, while a schematic of how the integrated circuit is incorporated in a TV set is shown in Fig. 3.

These two schematics are combined and converted to a more conventional form in Fig. 4. The general layout of the integrated circuit's sub components in this diagram is nearly the same as that shown in Fig. 2. The only modification, for the sake of simplicity, is in the change in layout of the sub components connected to terminals 6 and 7.

The dc voltages indicated at the terminals of the integrated circuit (Fig. 4) are within ±15 percent of the VTVM voltages that can be read when the TV set is functioning properly. If the voltage at terminal 10 is not within the indicated range, the fault probably lies in the B+ supply. An improper bias voltage at terminal 1 might be caused by an open secondary winding of the input transformer (T201), while an abnormal reading at terminal 5, 6 or 7 may be caused by a defective discriminator transformer (T203).

When checking a TV set that has no audio, the integrated circuit is the component that should be least suspected. To service a TV set containing an integrated circuit, however, technicians should understand how the circuit functions.

Differential Amplifiers

The bases of the two transistors (Q1 and Q2), in the first two stages (Fig. 5) of the integrated circuit, have been forward biased to produce emitter currents in both transistors. This bias circuitry is not shown since it will be described later. (A similar unbalanced differential amplifier circuit was described and shown in Fig. 9 of chapter 9.)

A positive signal applied to the base of transistor Q1 (Fig. 5) will cause it to conduct more collector-to-emitter current. This will result in a greater voltage drop across the common emitter resistor

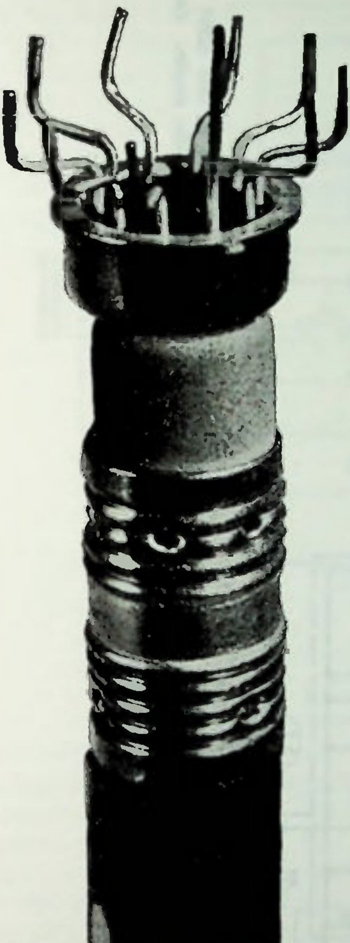


Fig. 1—The integrated circuit currently used in many TV sets is so small it can be placed on the end of a pencil.

(R1). The emitters of both transistors will, therefore, become more positive with respect to ground.

The base of transistor Q2 is biased at a relatively constant potential above ground. As the transistor's emitter becomes more positive with respect to ground, its base becomes less positive with respect to the emitter. The effective reduction in the second transistor's base bias reduces its collector-to-emitter current. This, in turn, reduces the voltage drop across resistor R2, and the collector of transistor Q2 becomes more positive.

In chapter 9 it was also explained that in a differential amplifier circuit, such as this, the collector-to-emitter signal current in the second transistor (Q2) must be smaller than the collector-to-emitter signal current in the first transistor (Q1). The circuit still experiences a gain, however, since the entire collector-to-emitter signal current in the first transistor (Q1) alters the voltage drop only across resistor R1 while the collector-to-emitter signal current in transistor Q2 alters the voltage drop across both resistor R1 and resistor R2. A voltage gain, rather than a current gain, is experienced in the circuit.

The first two amplifier stages (Fig. 5) are shown as a part of the circuit in Fig. 6. A positive signal applied to the base of transistor Q1 results in a voltage-amplified, positive signal across resistor R2, increasing the base forward bias of transistor Q3.

As the emitter current of transistor Q3 increases, a greater voltage drop is developed across resistor R3, and the signal is amplified further.

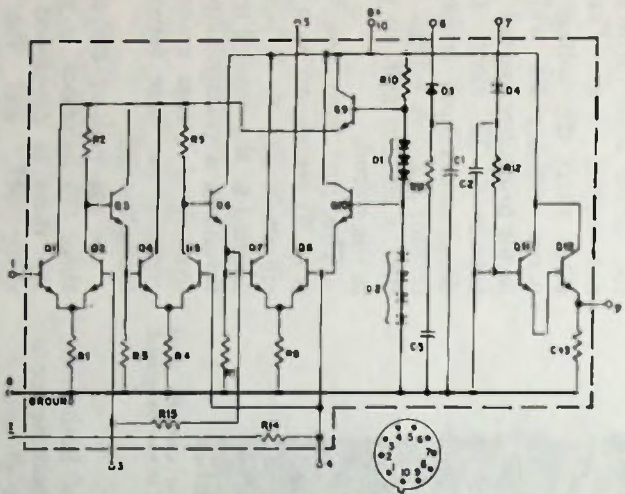


Fig. 2—The manufacturer's schematic of IC201.
Courtesy of RCA.

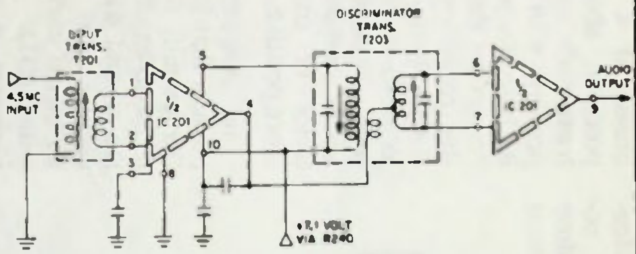


Fig. 3—The manufacturer's schematic of how IC201 is incorporated in a TV receiver's sound IF and audio circuits. Courtesy of RCA.

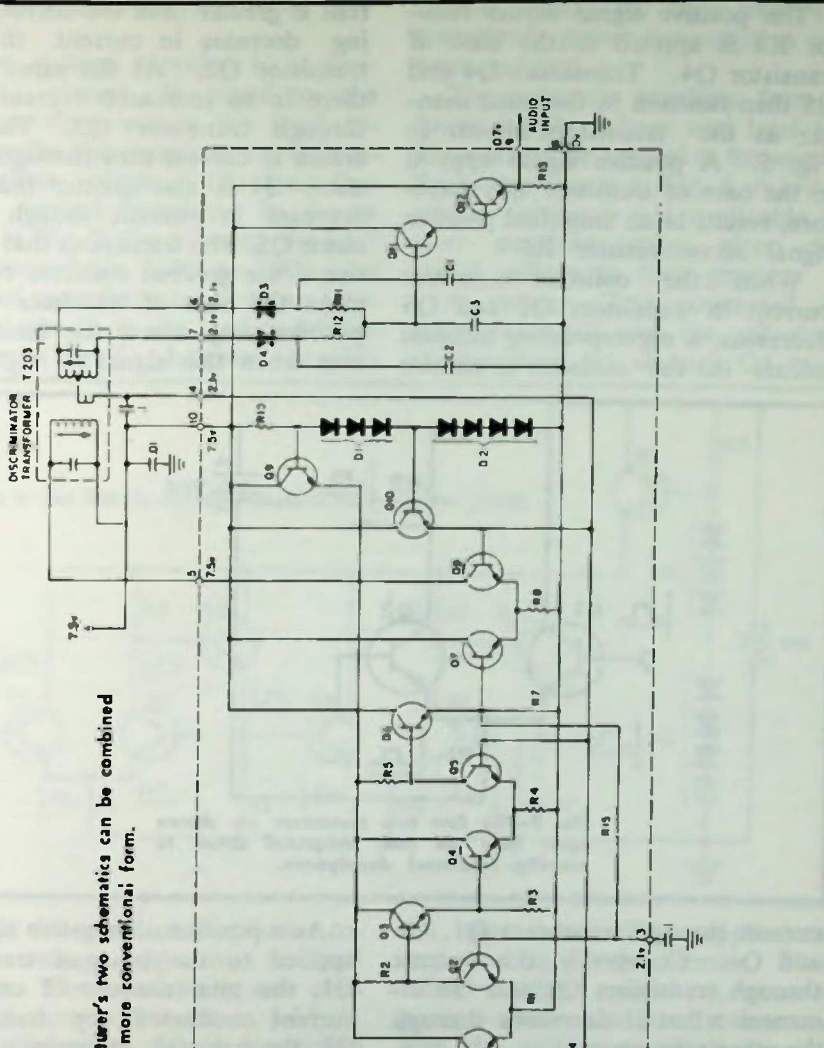


Fig. 4—The manufacturer's two schematics can be combined and converted to a more conventional form.

The positive signal across resistor R3 is applied to the base of transistor Q4. Transistors Q4 and Q5 then function in the same manner as the transistors shown in Fig. 5. A positive signal applied to the base of transistor Q1, therefore, results in an amplified positive signal across resistor R5.

When the collector-to-emitter current in transistors Q2 and Q5 decreases, a corresponding increase occurs in the collector-to-emitter rent is greater than the corresponding decrease in current through transistor Q2. At the same time, there is an increased current flow through transistor Q3. The increase in current flow through transistor Q4 is also greater than the decrease in current though transistor Q5. The transistors that experience the greatest collector current when the base of transistor Q1 is positive, experience the least current when this signal is negative.

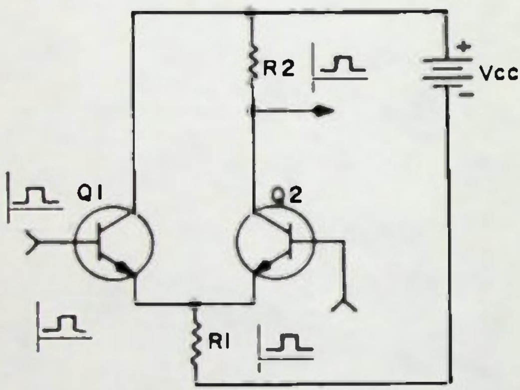


Fig. 5—The first two transistors are shown apart from the total integrated circuit to simplify functional descriptions.

current through transistors Q1, Q3 and Q4. Conversely, this current through transistors Q2 and Q5 increases when it decreases through the other transistors.

As previously indicated, the collector-to-emitter signal current through transistor Q1 is greater than that through transistor Q2. This situation is also true for transistors Q4 and Q5.

When a positive signal is applied to the base of transistor Q1, its increase in collector-to-emitter current through transistors Q1, Q3

As a positive or negative signal is applied to the base of transistor Q1, the total amount of collector current conducted by transistors Q1 through Q5 increases or decreases as though these transistors were but a single component.

IC resistors cost more to produce than transistors since they require a larger area on the chip. The cost of producing resistors R1 through R4 can be reduced if their values can be kept to a minimum.

Voltage-Regulating Transistor

The voltage supplied to a transistor circuit is equal to the voltage drop across the load resistor plus the transistor's collector-to-emitter voltage. The load line of a typical transistor is shown here in Fig. 7. By reducing the load resistance from 750Ω to 500Ω, the cost of manufacturing it in an integrated circuit is also reduced. This change (Fig. 8), however, has increased the transistor's collector-to-emitter voltage drop across the load resistor. The transistor is now consuming more power ($P = I_C V_{CE}$), and producing more destructive heat, to provide nearly the same amount of amplification as before.

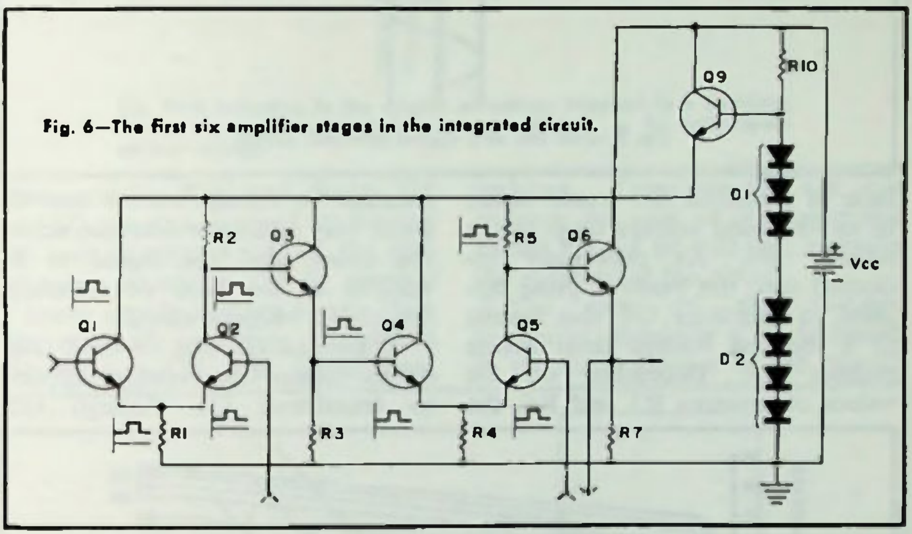


Fig. 6—The first six amplifier stages in the integrated circuit.

A reduction in the amount of voltage supplied to the transistor circuit (Fig. 9) reduces only slightly the voltage across the load resistor and appears almost entirely as a reduction in collector-to-emitter voltage. The signal voltage remains virtually the same while the transistor's power consumption has been reduced.

The economy resulting from a lower supply voltage and load resistance can be applied to the amplifier stages shown in Fig. 6. A lower supply voltage is not applicable to later amplifier stages since the signal voltage is then greater.

A resistor (R6) connected in series (Fig. 10) with the collectors of transistors Q1 through Q5 cannot reduce the voltage applied to that portion of the circuit. The value of that resistor is now a portion of the total load resistance, with the voltage supplied remaining the same. As indicated previously, when a signal is applied to the base of transistor Q1, the total amount of collector current conducted by transistors Q1 through Q5 increases or decreases as though these transistors were but a single transistor.

A positive pulse applied to the

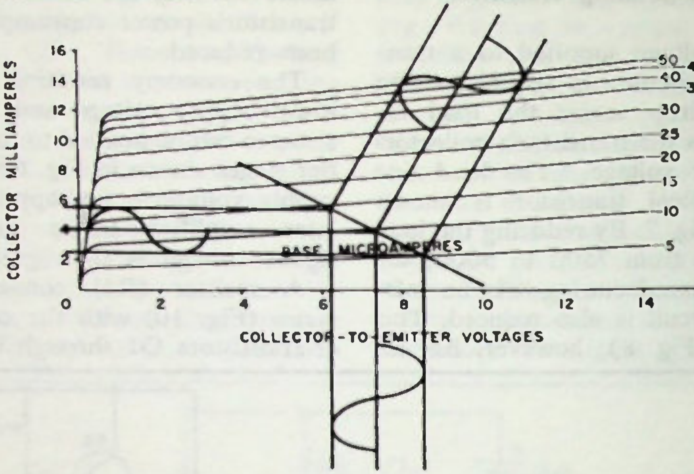


Fig. 7—Load line of a typical transistor circuit.

base of transistor Q1 would result in an increased voltage drop across resistor R6. As previously indicated too, the positive pulse applied to transistor Q1 also results in a reduced voltage drop across resistor R5. Depending on the values of resistors R5 and R6, the increase in voltage across one resistor may equal the decrease across the other, and the signal to be applied to the base of transistor Q6 could be neutralized.

Transistor Q9 can be used (Fig. 4) to reduce the voltage applied to transistors Q1 through Q5.

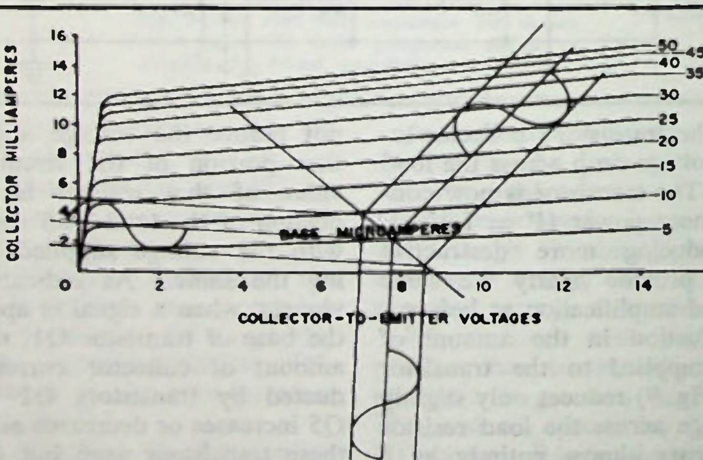


Fig. 8—The transistor's collector-to-emitter voltage increases as the load resistance decreases.

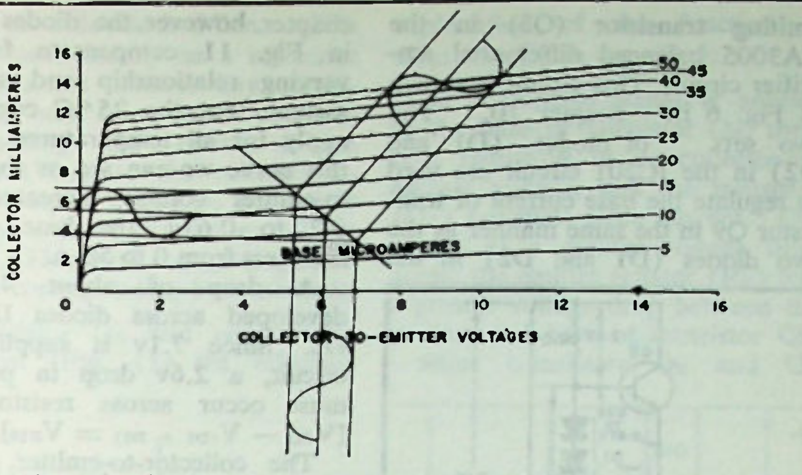


Fig. 9—A reduction in the amount of voltage supplied to a transistor circuit appears almost entirely as a reduction in its collector-to-emitter voltage.

Since resistor R10 and diodes D1 and D2 are also required for biasing the base of transistor Q10, the cost of using transistor Q9 to provide a lower supply voltage is less than the cost of providing transistors Q1 through Q5 with larger load resistors. Transistor Q9 not only reduces the cost of the integrated circuit but, as we will see, increases the stability of the circuit.

The base of the voltage-regulating transistor (Q9) is biased in the same manner as the current

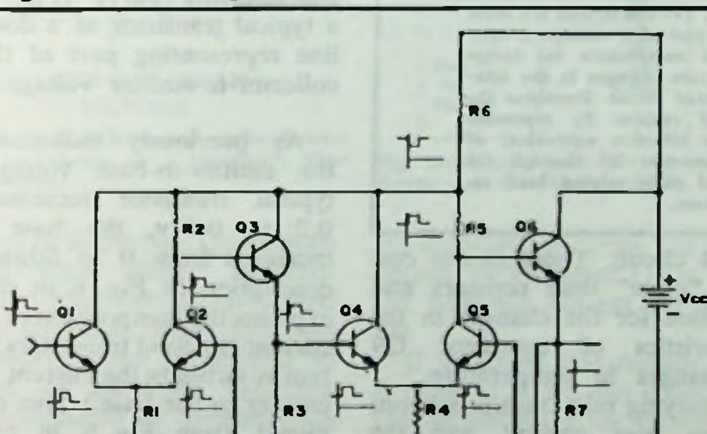


Fig. 10—A resistor connected in series with the first five amplifier stages will increase their total load resistance rather than reduce the supply voltage.

limiting transistor (Q3) in the CA3005 balanced differential amplifier circuit. This circuit is shown in Fig. 6 in chapter 10. The two sets of diodes (D1 and D2) in the IC201 circuit are used to regulate the base current of transistor Q9 in the same manner as the two diodes (D1 and D2) in the

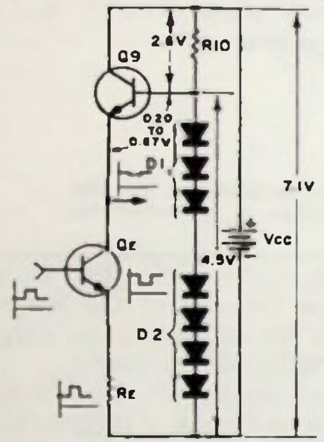


Fig. 11—The diodes are used as part of a voltage divider and compensate for temperature changes in the integrated circuit. Transistor Q_E and resistor R_E represent the effective equivalent of transistors Q1 through Q5 and their related load resistors.

CA3004 circuit. These diodes cost less to "grow" than resistors and compensate for the changes in the characteristics of transistor Q9 with changes in temperature.

The varying relationship between the base bias current and the base-to-emitter voltage of a typical transistor was shown in Fig. 8 in chapter 3.

As indicated in the present chapter, however, the diodes shown in Fig. 11 compensate for this varying relationship and we can assume that the 25 °C curve will apply for all temperatures. From this curve we can see as the base-to-emitter voltage increases from 0.2 to 0.67v, the base current increases from 0 to 50μa.

A drop of about 4.5v is developed across diodes D1 and D2. Since 7.1v is supplied the circuit, a 2.6v drop in potential must occur across resistor R10 [$V_{CC} - V_{(D1 + D2)} = V_{R10}$].

The collector-to-emitter voltage (V_{CE}) of transistor Q9 is the sum of its collector-to-base voltage (V_{CB}) and its emitter-to-base voltage (V_{EB}). ($V_{CE} = V_{CB} + V_{EB}$.) The voltage drop across resistor R10 determines the collector-to-base voltage of transistor Q9 ($V_{CE} = 2.6$v). This collector-to-base voltage is shown on the characteristic curve (Fig. 12) of a typical transistor as a dot-dashed line representing part of the total collector-to-emitter voltage.

As previously indicated, when the emitter-to-base voltage of a typical transistor increases from 0.2 to 0.67v, the base current increases from 0 to 50μa. (The description of Fig. 6 in chapter 2 explains the composition of the base current. In good transistors this current is virtually the current from the emitter to the base.) This data, obtained from Fig. 8 in chapter 3, can be applied to the typical transistor characteristic curve (Fig. 12). There the collector-to-base voltage (V_{CB}) is shown as a dot-dashed line. (This line is used merely as a reference and does not indicate any corresponding base or collector current.) By adding the emitter-to-base voltage (V_{EB}), to the collector-to-base voltage (V_{CB}), we can show the total collector-to-emitter voltage (V_{CE}). This total voltage is plotted against corresponding base currents.

Transistor Q_E and resistor R_E represent (Fig. 11) the effective equivalent of transistors Q1 through Q5 and their related resistors. Just as the total current conducted by these amplifier stages increases as a positive pulse is applied to the base of transistor Q1, the current conducted by transistor Q_E increases as a positive signal is applied to its base. We have substituted a single transistor and resistor for these stages of amplification merely to simplify the explanation of the function of transistor Q9 in the circuit.

As a positive signal is applied to the base of transistor Q_E, there is a reduction in its collector-to-emitter voltage, and the transistor (Q_E) conducts more current. This results in a greater voltage drop across the emitter resistor (R_E) and a greater voltage drop between the emitter and base of transistor Q9.

Since transistors Q_E and Q9 are connected in series, the collector current of transistor Q_E is virtually the collector current of transistor Q9.

If we knew that resistor R_E had a value of 375Ω, the collector current of transistor Q_E varied from 2 to 8ma, and Fig. 12 represented the characteristic curves of transistor Q9, we could calculate the range of voltage in this portion of the circuit.

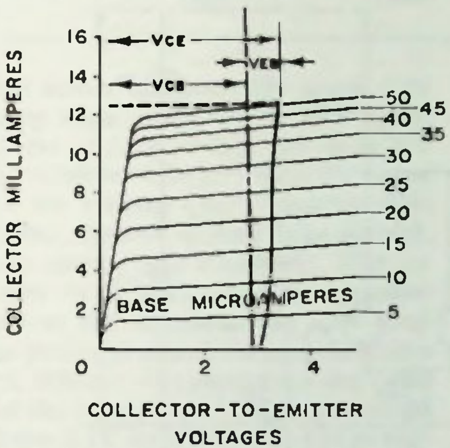


Fig. 12—Collector-to-base and emitter-to-base voltages are shown on a transistor's characteristic curve.

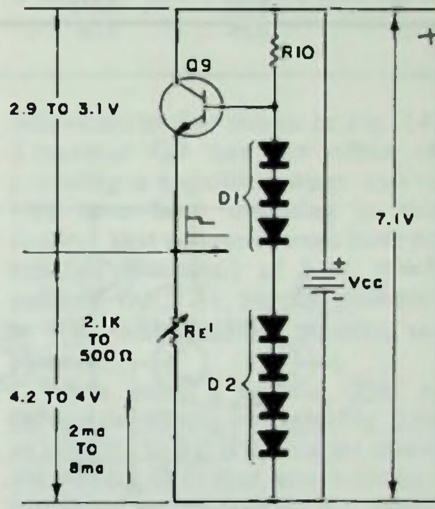


Fig. 13—A variable resistor demonstrates the function of a voltage-regulating transistor.

From Fig. 12 we see that the collector-to-emitter voltage of transistor Q9 would vary from 2.9 to 3.1v under these conditions. The voltage drop across resistor R_E would vary from ¾ to 3v ($V = IR$, $2ma \times 375\Omega = 2 \times 10^{-3}a \times 375\Omega = 0.750v$, $8ma \times 375\Omega = 3.00v$). The voltage drop across transistor Q_E would then vary from 3.45 to 1.00v [Since the voltage source equals the sum of the voltage drops in the circuit, $V_{CC} - (V_{Q9} + V_{RE}) = V_{QE}$, $7.1v - (2.9v + 0.75v) = 3.45v$, $7.1v - (3.1v + 3.00v) = 1.00v$].

The 2.9 to 3.1v change in collector-to-emitter voltage across transistor Q9 occurred as the collector current varied from 2 to 8ma. This change would correspond to a 33⅓Ω effective resistance ($R = \frac{\Delta v}{\Delta I} = \frac{3.1v - 2.9v}{8ma - 2ma} = \frac{0.2v}{6ma} = 33⅓\Omega$). If the transistor used as transistor Q9 has a greater beta or steeper base-to-emitter voltage vs base current curve, the effective resistance would be reduced and there would be a smaller fluctuation in voltages as the current varied.

Although we see a relatively large variation in voltages across transistor Q_E and resistor R_E, the voltage drop across transistor Q9 remains nearly unchanged, and as the voltage across transistor Q_E and resistor R_E changes, their total voltage remains nearly the same.

The function of the voltage regulating transistor may be more clearly seen in the circuit shown in Fig. 13. There a variable resistor $R_{E'}$ is substituted for transistor Q_E and resistor R_E in Fig. 11. When the value of the resistor ($R_{E'}$) is 2.1K, 2ma of current flows through the resistor and transistor (Q9). This results in a 2.9v drop in poten-

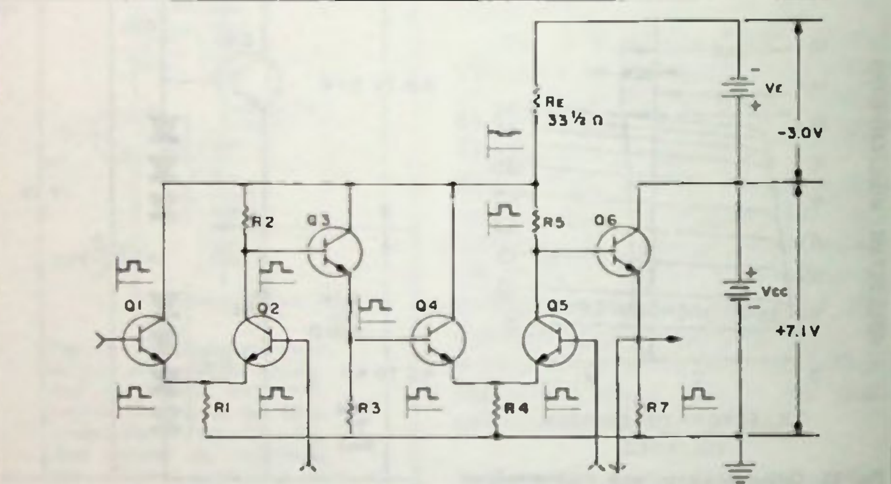


Fig. 14—The equivalent to a voltage-regulating transistor is substituted for transistor Q9.

Factory Sales of Integrated Circuits
During First Two Months of 1966 and 1967

Circuit Type	First Two Months 1966 Units	Dollars	Average Value	First Two Months 1967 Units	Dollars	Average Value
Digital	2,847,000	14,914,000	5.24	6,297,000	22,059,000	3.50
Analog	179,000	3,615,000	20.20	857,000	7,782,000	9.08
Total	3,025,000	18,529,000	6.13	7,154,000	29,841,000	4.17

Percent of Increase or Decrease
First Two Months 1966 and 1967

Circuit Type	Units	Dollars	Average Value
Digital	121.2	47.9	33.2
Analog	378.8	115.3	55.0
Total	136.5	61.1	32.0

tial across transistor Q9 and a 4.2v drop in potential across the resistor ($2ma \times 2.1K = 2 \times 10^{-3}a \times 2.1 \times 10^{3}\Omega = 4.2v$). When the value of the resistor ($R_{E'}$) is reduced to 500Ω, 8ma of current flows through the resistor and transistor. This results in a 3.1v drop in potential across the transistor and a 4v drop in potential across the resistor ($8ma \times 500\Omega = 4v$). Although the value of the load resistor ($R_{E'}$) has varied from 2.1K to 500Ω, the voltage regulating transistor has permitted the voltage drop across the resistor to vary only from 4.2v to 4v.

If the characteristics of transistor Q9 are the same as those shown in Fig. 12, and its base is biased as shown in those curves, then the circuit shown in Fig. 6 is equivalent to that shown in Fig. 14. Transistor Q9 has the effect of providing a negative voltage source (We have been assuming in this chapter that voltage sources have no resistive elements.) of 3.0v, which reduces the 7.1v supply potential to 4.1v, with a 33⅓Ω effective resistance.

When using a resistor (R6) to reduce the supply voltage (Fig. 10), an inverted signal is developed across the resistor (R6) that may neutralize the signal across resistor R5, which is to be amplified. The inverted signal developed (Fig. 14) across the effective resistance (R_E) of transistor Q9 is much smaller than the signal developed across resistor R5, and the reduction in the ladder signal is insignificant.

CHAPTER 13

IC TV Sound Circuits

The previous chapter (chapter 12) explained how a temperature-compensated, voltage-regulating transistor (Q9) is used in integrated circuit, IC201, to provide a nearly constant supply voltage to the first five amplifier stages. These stages contain still another circuit to improve their stability further.

A Negative Feedback Circuit

A resistor (R15) couples the base of transistor Q2 to the emitter of transistor Q6 (Fig. 1). Since the portion of the integrated circuit shown in this figure is used to amplify FM IF signals, the circuit handles frequencies of around 4.5MHz. The 0.1μf capacitor connected to the base of transistor Q2 is, therefore, able to short these ac signals to ground. The ac portion of the transistor (Q6) emitter current develops an ac voltage across resistors R7 and R15, while virtually none of this ac voltage appears at the base of transistor Q2. If the value of resistor R15 is sufficiently large, it does not significantly reduce the ac voltage across resistor R7.

If a temperature change or slight production variance results in an increased dc current, a larger dc voltage drop will result across resistor R7. This increases the forward bias of transistor Q2, increasing its collector current and the current flow through resistor R2. With a greater voltage drop across that resistor, the base of transistor Q3 becomes less positive and a reduction occurs in its dc collector current. This, in turn, results in reduced current through transistor Q4, increased current through transistor Q5 and a reduced dc current through transistor Q6. The dc voltage drop across resistor R7 is, therefore, reduced and negative feedback has taken place.

Transistors Q7 and Q8 function (Fig. 2) in the same manner as transistors Q1 and Q2 (described in Fig. 5, chapter 12. A positive pulse applied to the base of transistor Q_7 causes that transistor to conduct more current and transistor Q8 to conduct less current. This results in a reduced voltage drop across the primary winding of transformer T203, and the winding becomes more positive.

This differential amplifier circuit (Fig. 2) need not be balanced and the increased current through transistor Q7 may not necessarily equal the decreased current through transistor Q8. A $0.01\mu f$ capacitor shunts any of the resulting 4.5MHz signal, which may otherwise be developed across the voltage source (Vcc), to ground.

Transistor Q10 (Fig. 3) functions in the same manner as transistor Q9 (described in Fig. 13 in chapter 12). Just as the equivalent to the circuit containing transistor Q9 (Fig. 4, chapter 12) is shown as Fig. 14 in that chapter, the equivalent to the circuit containing transistor Q10 (Fig. 3) is shown in Fig. 4. Transistor Q10 supplies a temperature-compensated, virtually constant voltage source to the base of transistor Q5, the base of transistor Q8, a secondary winding of transformer T203 and resistor R14, which is connected to a secondary winding of transformer T201.

When no signal is induced into the secondary winding of transformer T201, the base of transistor Q1 has a 2.1v bias — there having been an insignificant dc voltage drop across the transformer's secondary winding and resistor R14. The 4.5MHz signal induced in the transformer's secondary winding increases and decreases the transistor's base bias at that frequency.

Specifications indicate that the 4.5MHz signal applied to terminal 1 of the integrated circuit (Fig. 4 in chapter 12) experiences a gain of about 60 db with a bandwidth of 70 to 80 kHz before it leaves the circuit at terminal 5.

The frequency of the FM IF signal, amplified in the first portion of the integrated circuit, varies with the audio signal it contains. This varying IF signal is applied to the primary winding of discriminator transformer T203 (Fig. 3).

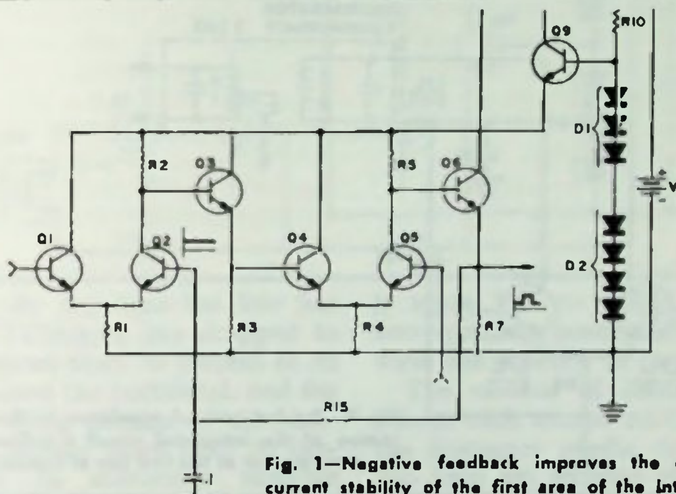


Fig. 1—Negative feedback improves the dc current stability of the first area of the integrated circuit.

The Discriminator Transformer

When a transformer has a single secondary winding, the voltages in the two leads at the ends of the winding (Fig. 5) are 180deg out of phase — when one lead is positive, the other is negative. If another secondary winding is connected to the first secondary winding (Fig. 6), the voltages in the two leads connected to the first secondary winding are still 180deg out of phase. The voltage induced in the third lead, however, need not have a fixed phase relationship with the other voltages. The phase of the voltage at the third lead depends on how the second coil is wound and the frequency of the signal induced.

Some less experienced technicians may not be familiar with the term "phase angle." This term is used either to indicate portions of a sinewave or to indicate the degree of time lag between two sinewaves of the same frequency.

Positive and negative changes of voltage generally appear on the screen of an oscilloscope as a sinewave — whether the signal observed is from an ac power line or some unmodulated RF source.

When a line is horizontal (zero deg), it has no height (Fig. 7), and the corresponding portion of the sinewave (Fig. 8) is at zero voltage. After rotating 45deg, the line has risen slightly more than 70 percent of its full height, and the corresponding portion of the sinewave is slightly more than 70 percent of its maximum positive voltage. After rotating 90deg, the line is at its full height, and the corresponding portion of the sinewave is at its peak positive voltage. At 135deg, the line is again only slightly more than 70 percent of its full height, and the corresponding portion of the sinewave is slightly more than 70 percent of its maximum positive voltage. At 180deg, the line is again horizontal, and the corresponding portion of the sinewave is at a zero

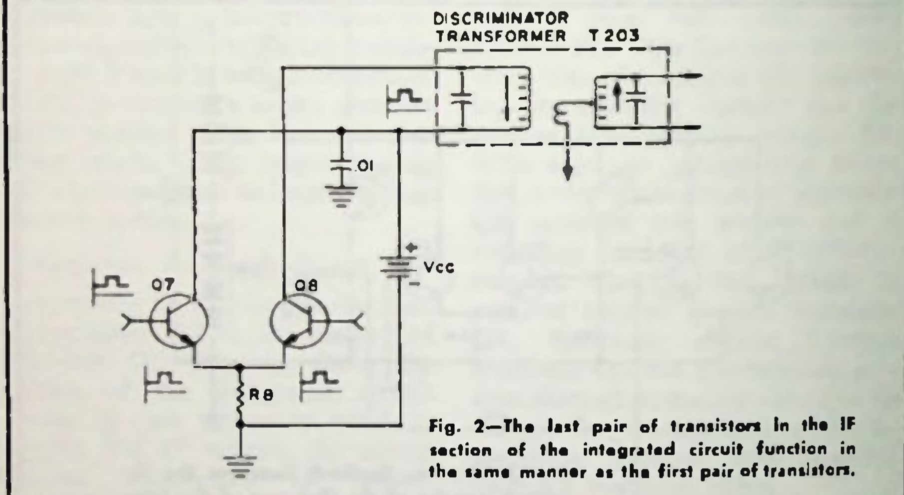


Fig. 2—The last pair of transistors in the IF section of the integrated circuit function in the same manner as the first pair of transistors.

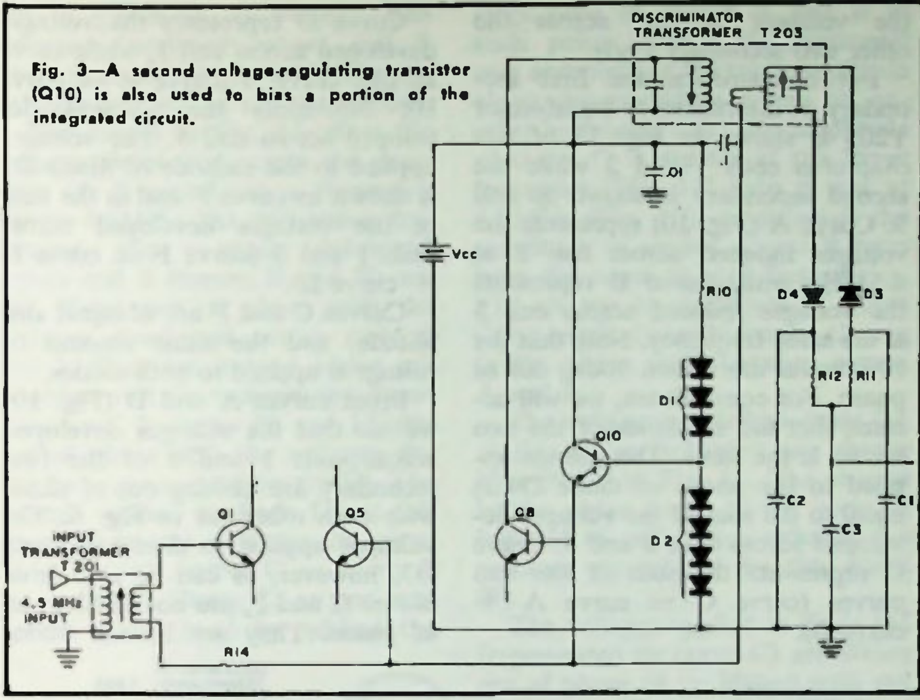


Fig. 3—A second voltage-regulating transistor (Q10) is also used to bias a portion of the integrated circuit.

voltage. By the time the line has rotated 225deg, it has dropped to slightly more than 70 percent of its length below the horizontal, and the corresponding portion of the sinewave is slightly more than 70 percent of its maximum negative voltage. By the time the line has rotated 270deg, it has dropped its full length below the horizontal, and the corresponding portion of the sinewave is at its peak negative voltage. After rotating 315deg, the line is again slightly more than 70 percent of its full length below the horizontal, and the voltage at the corresponding portion of the sinewave has been reduced to about 70 percent of the maximum negative voltage. When the line has rotated 360deg and has returned to its initial horizontal position (zero deg), it again has no height, and the corresponding portion of the sinewave has returned to zero volts.

The number of times the line rotates each second corresponds to the frequency of the sinewave. If two lines are rotating at the same frequency, the angle between the two rotating lines corresponds to the angle between their two corresponding sinewaves.

The Discriminator Circuit

The ac signal in the third lead in the secondary of the discriminator transformer (Fig. 9) is shorted to ground by a 0.1μf and a 0.01μf capacitor. The discriminator transformer is designed in such a manner that at 4.5MHz the voltages induced across the grounded extra secondary lead are 90deg out of phase with the voltages induced across the other two secondary leads.

For convenience, the first secondary of discriminator transformer T203 is shown in Fig. 13 of this chapter as coils 1 and 2 while the second secondary is shown as coil 3. Curve A (Fig. 10) represents the voltages induced across coil 2 at 4.5MHz while curve B represents the voltages induced across coil 3 at the same frequency. Note that the two curves are shown 90deg out of phase. For convenience, we will assume that the amplitude of the two curves is the same. The voltage applied to the anode of diode D4 is equal to the sum of the voltages developed across coils 2 and 3. Curve C represents the sum of the two curves (curve C = curve A + curve B).

Curve D represents the voltages developed across coil 1, while curve E, like curve B (curve E = curve B), represents the voltages developed across coil 3. The voltages applied to the cathode of diode D3 is shown as curve F and is the sum of the voltages developed across coils 1 and 3 (curve F = curve D + curve E).

Curves C and F are of equal amplitude, and the same amount of voltage is applied to both diodes.

From curves A and D (Fig. 10) we see that the voltages developed across coils 1 and 2 of the first secondary are 180deg out of phase with each other, as in Fig. 6. The voltages applied to diodes D4 and D3, however, as can be seen from curves C and F, are not 180deg out of phase. They are instead 90deg

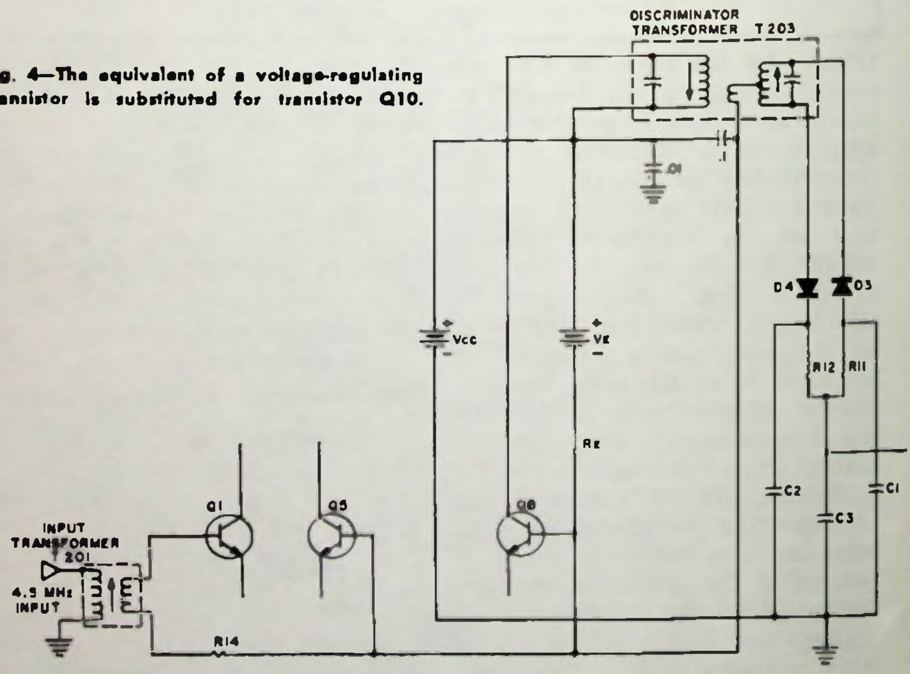


Fig. 4—The equivalent of a voltage-regulating transistor is substituted for transistor Q10.

out of phase ($\theta = 90$deg). The voltages developed across coil 3 have resulted in a phase shift in the voltages applied to the two diodes.

The curves in Fig. 11 show the voltages developed across the same coils (1, 2 and 3) at a frequency below 4.5MHz. At this lower frequency, the voltages developed across coil 3 (curves B and E) are not 90deg out of phase with the voltages developed across coils 1 and 2. They are instead 18deg out of phase with the voltages developed across coil 2 (curve A) and 162deg out of phase with the voltages developed across coil 1 (curve D). The voltage applied to diode D4 at this lower frequency is shown by curve C in Fig. 11 (curve C = curve A + curve B). Curve C, as you may know, was formed by adding the amplitudes of curves A and B at each point along their horizontal axis and plotting the total amplitude along corresponding points on another horizontal axis. The amplitude of curve C is greater at the lower frequency (Fig. 11) than it was at 4.5MHz (Fig. 10) — although the amplitude of curves A and B have been the same at both frequencies. The change in amplitude of curve C has resulted merely from a change in the phase angle between curves A and B.

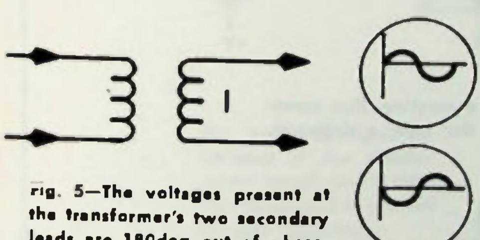


Fig. 5—The voltages present at the transformer's two secondary leads are 180deg out of phase.

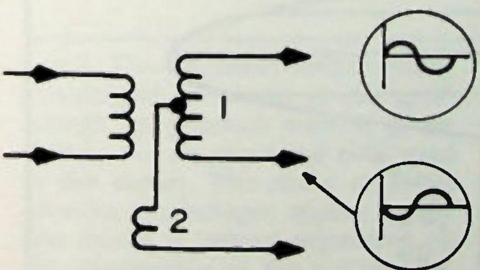


Fig. 6—The voltages present at two of the transformer's secondary leads remain 180deg out of phase, while the phase relationship of the voltages present at the third lead varies with the signal's frequency.

The voltages applied to diode D3 at this lower frequency are shown by curve F in Fig. 11. The change in the phase angle of the voltages induced in coil 3 has reduced the voltages represented by this curve (F).

The voltages applied to diode D4 (represented by curve C) are 90deg out of phase ($\theta = 90$deg) with the voltages applied to diode D3 (represented by curve F).

When the frequency of the IF signal is greater than 4.5MHz, the voltages induced across coil 3 have still another phase relationship with the voltages induced across coils 1 and 2 (Fig. 12). In this third example, the voltages induced across coil 3 (curve B) are 45deg out of phase with the voltages induced across coil 2 (curve A) and 135deg out of phase with the voltages induced across coil 1 (curve D). As a result of this phase shift, larger voltages are applied to diode D3 than to diode D4 (amplitude of curve E > amplitude of curve C).

Again, the voltages applied to diode D4 are 90deg out of phase ($\theta = 90$ deg) with the voltages ap-

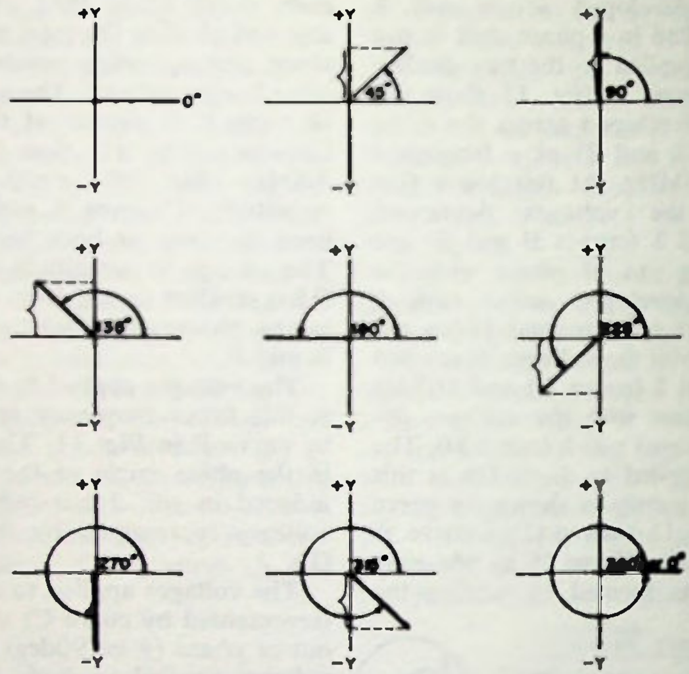
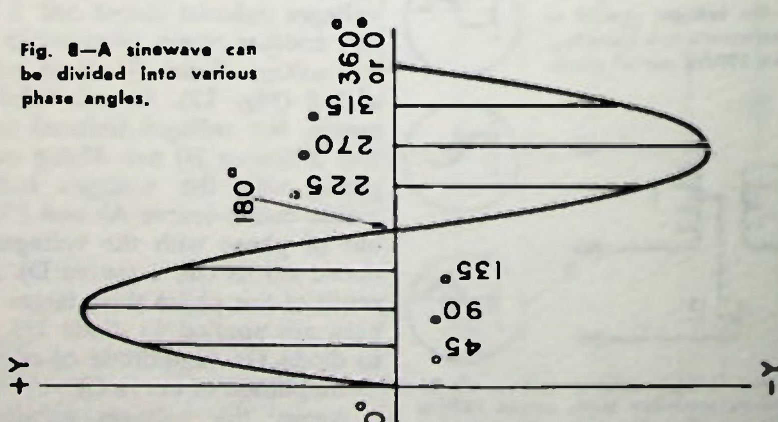


Fig. 7—The height of a rotating line corresponds to the voltages that form a sinewave.

Fig. 8—A sinewave can be divided into various phase angles.



plied to diode D3. This phase relationship is the same as those that occurred at 4.5MHz and the other lower frequency. The phase relationship of the voltages applied to the two diodes (θ) will not change with the frequency of the IF signal — unless the frequency is one at which the phase of the voltages induced across coil 3 is the same as the phase of those induced across coil 1 or 2. The frequency shift normally experienced, however, is not great enough to produce such a phase shift in the discriminator coils used in this circuit. The phase angle (θ) between the voltages applied to the two diodes is instead dependent on the relationship of the voltages induced across coil 3 to the voltages induced across coils 1 and 2. Curves C and F are 90deg out of phase (θ = 90deg) only when the amplitudes of curves B and E are the same height as curves A and D. (Curve B = curve E, and the amplitude of curve A = amplitude of curve D. θ = 90deg when amplitude of curve B = amplitude of curve A = amplitude of curve D.) In the next chapter we will see that, with the capacitors in the detector circuit (Fig. 13), we need not be concerned with the phase angles of the voltages applied to the diodes (D4 and D5).

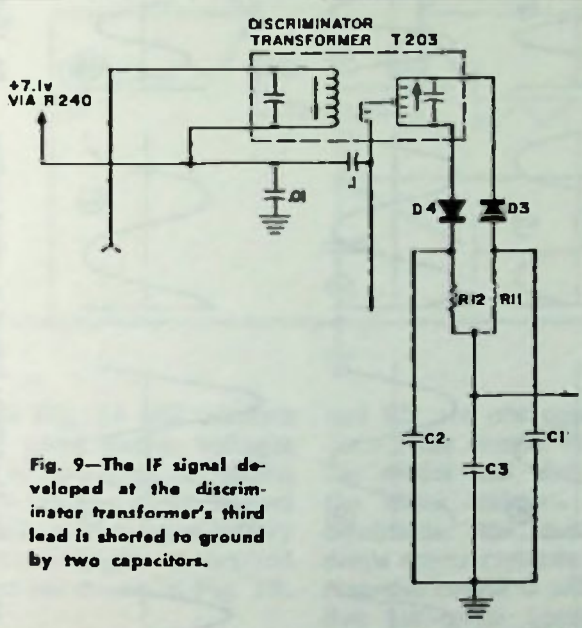


Fig. 9—The IF signal developed at the discriminator transformer's third lead is shorted to ground by two capacitors.

Curves C and F, shown in Fig. 13, represent the voltages applied to the two diodes (D4 and D5) at 4.5MHz (these curves are respectively the same as curves C and F in Fig. 10). In the first chapter we observed that electrical currents flow from negative to positive. These currents can pass only from the anode to the cathode of a diode.

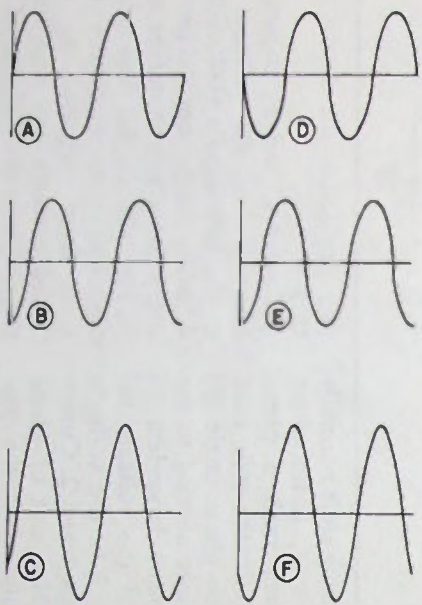


Fig. 10—Voltages induced in various portions of the discriminator transformer at 4.5MHz.

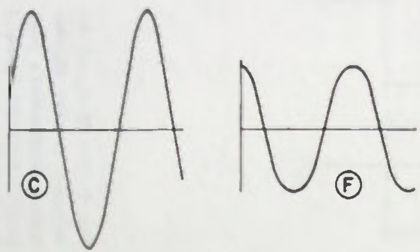


Fig. 11—Voltages induced in various portions of the discriminator transformer at a frequency below 4.5MHz.

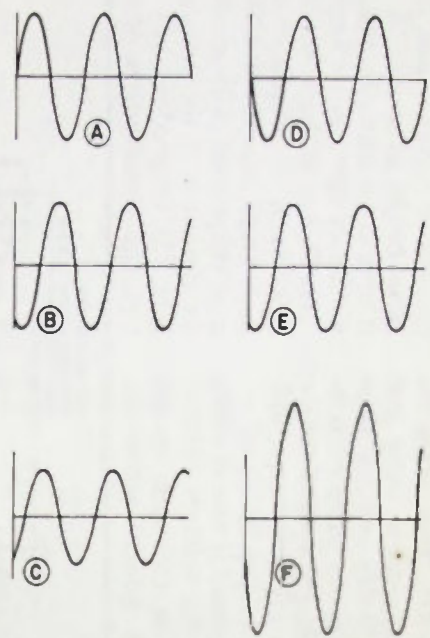


Fig. 12—Voltages induced in various portions of the discriminator transformer at a frequency above 4.5MHz.

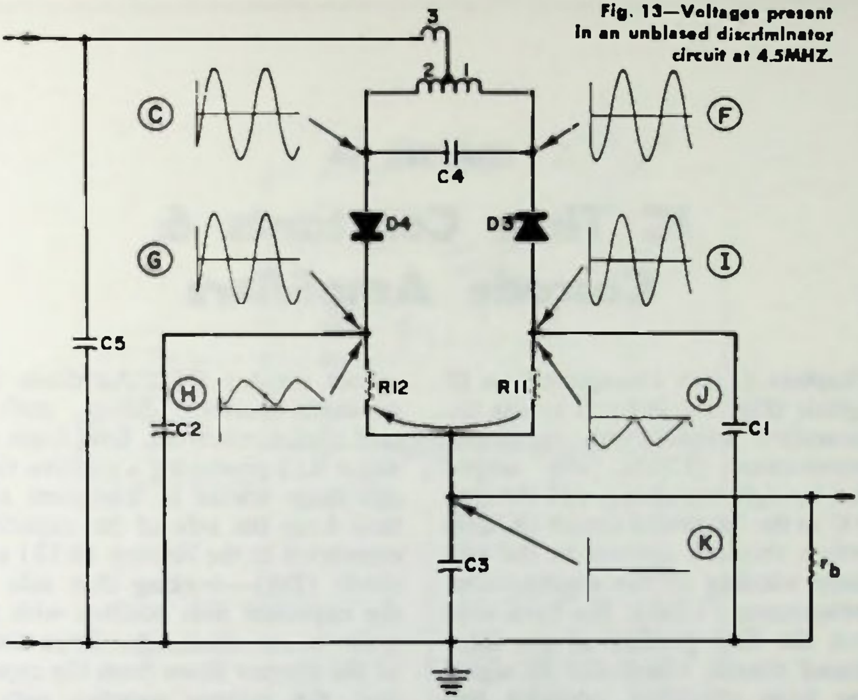


Fig. 13—Voltages present in an unbiased discriminator circuit at 4.5MHZ.

Diode D4 in Fig. 13 will conduct current only when postive voltages are applied to its anode (as shown in Fig. 3A in the first chapter), and diode D3 will conduct current only when negative voltages are applied to its cathode (as shown in Fig. 3B, chapter 1).

Diode D4 (Fig. 13) is connected in series with resistor R12, and diode D3 is connected in series with resistor R11. When capacitors C1 and C2 are not connected in the circuit, the current flowing through the diodes will also flow through the series resistors. Under these conditions, the resulting voltage drops across resistors R12 and R11 resemble curves G and I. Only positive half-cycles appear across resistor R12, while only negative half-cycles appear across resistor R11. The resulting waveforms are the same as those shown in Fig. 3 of chapter 1.

CHAPTER 14

IC Time Constants & Cascade Amplifiers

Chapters 12 and 13 explain how IF signals (Fig. 1), induced across the secondary winding of the input transformer (T201), are amplified by eight transistors (Q1 through Q8) in the integrated circuit (IC201) before they are applied to the primary winding of the discriminator transformer (T203). We have seen that the first portion of the integrated circuit, where the IF signal has been amplified, contains two voltage regulating circuits (that use transistors Q9 and Q10) and a negative feedback circuit (that uses resistor R15). As the frequency of the IF signals shift above and below 4.5MHz, in response to the audio signal it contains, the amplitude of the voltages applied to diodes D4 and D3 also shifts—increasing across one diode while decreasing across the other, and vice versa. The resulting voltage drops across resistors R12 and R11 are altered by capacitors in the discriminator circuit (Fig. 2).

Capacitor Time Constants

As you may know, capacitor C2 in Fig. 2 has the function of partly shorting to ground the ac portion of the voltage drops developed across resistor R12. As diode D4 conducts current during positive half cycles, electrons flow from resistor R12 producing a positive voltage drop across it. Electrons also flow from the side of the capacitor connected to the resistor (R12) and diode (D4)—making that side of the capacitor also positive with respect to its other side. Since some of the current flows from the capacitor, the resistor supplies only a portion of the current flowing through the diode. The resistor, therefore, does not develop as great a positive voltage drop as would have occurred without the capacitor. For this reason, the positive pulse shown in curve H (Fig. 2) is not as large as the corresponding one shown in curve G. (Curve G represents the voltages developed across resistor R12 when the capacitor is not connected in the discriminator circuit; while curve H represents the voltages developed across the resistor when capacitor C2 is in the circuit.)

The side of the capacitor that has lost the electrons will remain positive until electrons are returned to replace those that have left it.

The number of electrons that a

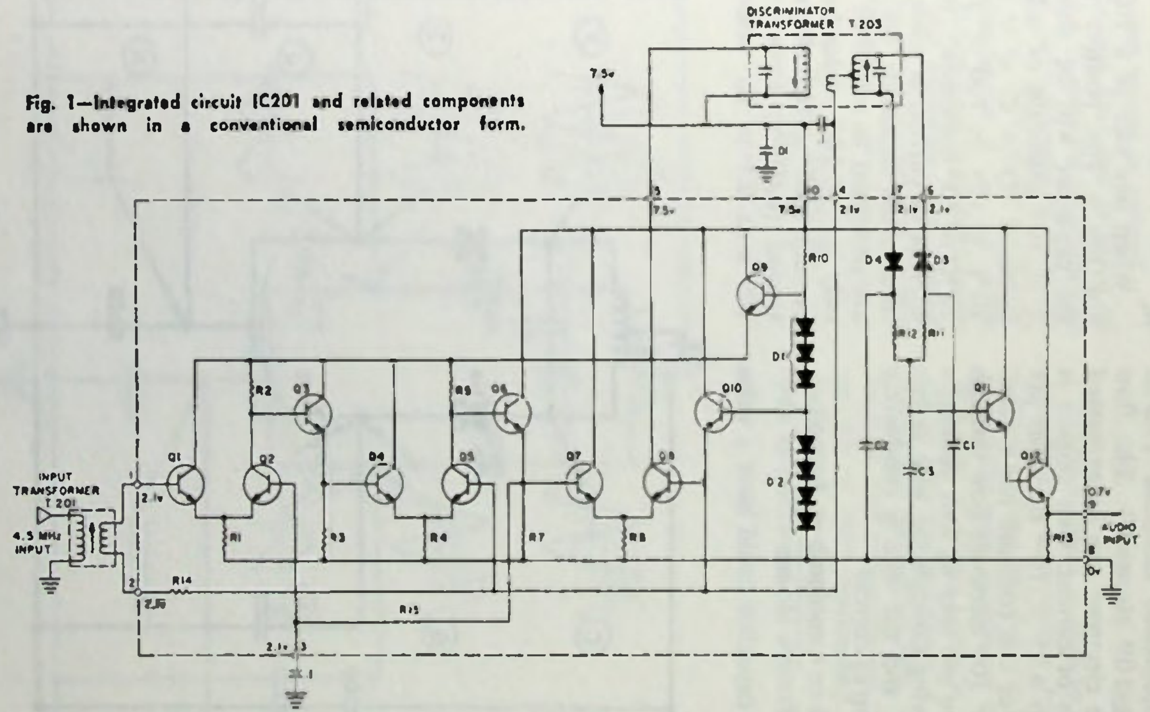


Fig. 1—Integrated circuit IC201 and related components are shown in a conventional semiconductor form.

capacitor may lose or gain is expressed in terms of coulombs (Q). One coulomb is equal to 6,250,000,000,000,000,000 electrons ($1Q = 6.25 \times 10^{18}$ electrons). The flow of these electrons can be expressed in terms of current. One ampere is equal to 6.25×10^{18} electrons per second or one coulomb per second. If 25×10^{18} electrons flow through a wire in one second, four coulombs are flowing through the wire in one second, and the wire is conducting four amp of current

$$(I = \frac{Q}{T} = \frac{4 \text{ coulomb}}{1 \text{ sec}} = 4 \text{ amp}).$$

If a capacitor could have a value of one Farad, it could gain or loose 6.25×10^{18} electrons or one coulomb for every volt applied across it.

When one side of a $10\mu f$ capacitor has a 10v positive potential, the capacitor will be able to lose 1×10^{-4} coulombs or 6.25×10^{14} electrons ($Q = C \times V = 10 \times 10^{-6}f \times 10v = 100 \times 10^{-6}\,Q = 10^{-4}Q$ or 10^{-4} coulombs). When only 0.01ma of current is allowed to return to the capacitor, the electrons are permitted to return at a rate of only 0.1×10^{-4} coulombs per sec. or only 6.25×10^{13} electrons each second ($0.01 \times 10^{-3}a = 0.1 \times 10^{-4}\,a = 0.1 \times 10^{-4}$ coulombs per

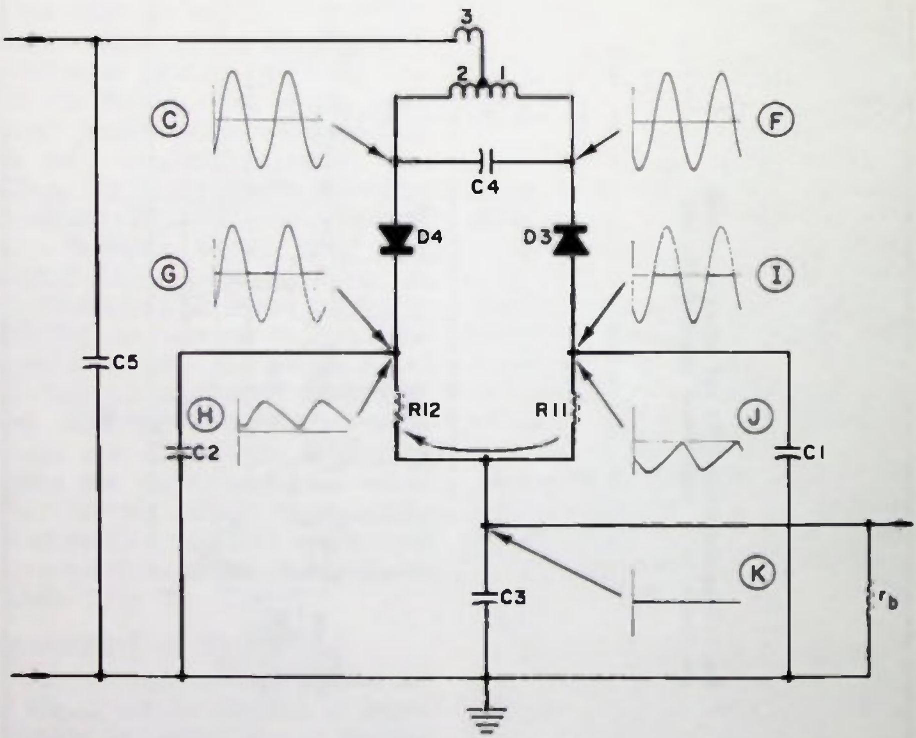


Fig. 2—The voltages present in an unbiased discriminator circuit at 4.5MHz.

second). At that rate, it will require 10sec for the electrons to return the 6.25×10^{14} electrons that had been lost before the voltage across the capacitor can return to normal.

A capacitor that has a value of $0.1\mu f$ and a 2v positive potential will loose 2×10^{-7} coulombs of electrons ($Q = 0.1 \times 10^{-6}f \times 2v = 0.2 \times 10^{-6}$ coulombs or 2×10^{-7} coulombs). When 5ma is allowed to return to the capacitor, the voltage across the capacitor returns to normal in 4msec

$$(T = \frac{2 \times 10^{-7} \text{ coulombs}}{5\text{ma}} = \frac{2 \times 10^{-7} \text{ coulombs}}{5 \times 10^{-3}\text{a}} = \frac{2 \times 10^{-7} \text{ coulombs}}{5 \times 10^{-3} \text{ coulombs/sec}} = 0.4 \times 10^{-4} \text{ sec} = 4\text{msec}).$$

When a diode is used to conduct 10ma from a capacitor for 1 sec then a resistor will require 10sec to return the same number of electrons at a rate of 1ma ($10\text{ma} \times 1\text{sec} = 10 \times 10^{-3}$ coulombs/sec $\times$ 1sec $= 10 \times 10^{-3}$ coulombs, 1ma $\times$ 10sec $= 1 \times 10^{-3}$ coulombs/sec $\times$ 10sec $= 10 \times 10^{-3}$ coulombs). When, during positive pulses, a diode is able to remove more electrons from

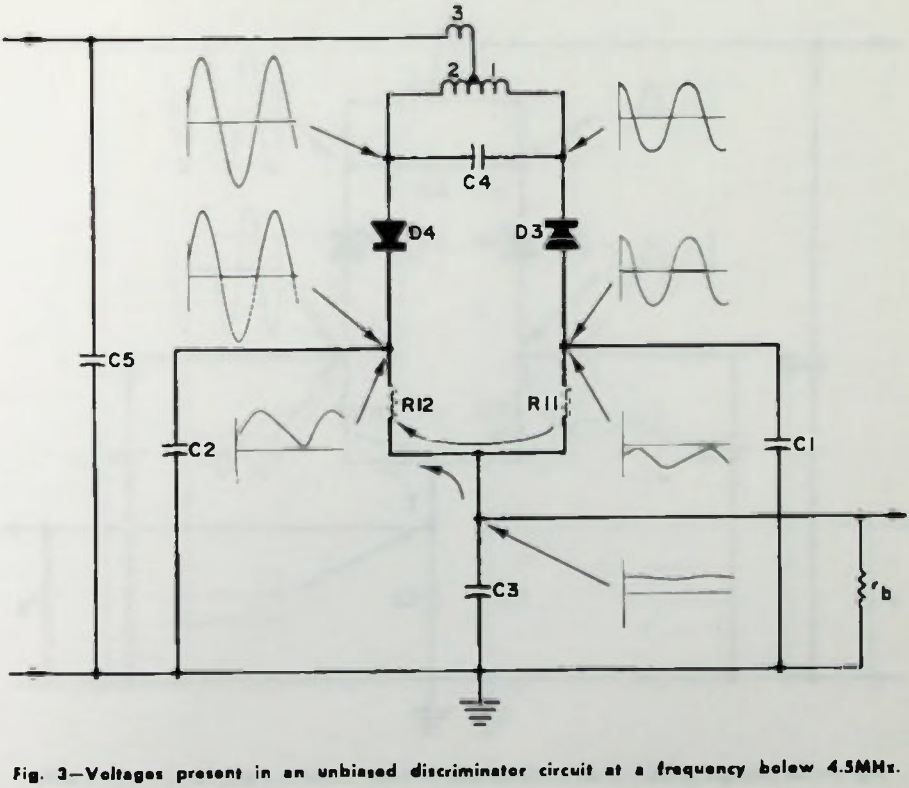


Fig. 3—Voltages present in an unbiased discriminator circuit at a frequency below 4.5MHz.

a capacitor than a resistor has time to return with a smaller current, the voltage remaining across the capacitor maintains a voltage drop across the resistor.

Part of the current flowing through the diode (D4) during positive half cycles (Fig. 2) comes from the capacitor (C2) as well as the resistor (R12). The positive voltage drop across the resistor is, therefore, less than the drop developed when all the current comes from the resistor. During the half cycle that the diode is not conducting current, the resistor (R12) does not allow a sufficient current to flow through it to the capacitor (C2) for the capacitor to loose the entire positive voltage drop developed across it. Curve G shows the voltage drops developed across the resistor (R12) when the capacitor (C2) is not in the circuit, while curve H shows the voltage drops developed across the resistor when the capacitor is in the circuit.

Part of the current flowing through diode D3 during negative half cycles (Fig. 2) goes into the capacitor (C1) as well as the resistor (R11). This reduces the negative voltage drop developed across the resistor (R11). During the positive half cycles that the diode (D3) is not conducting current, the resistor (R11) impedes the outward

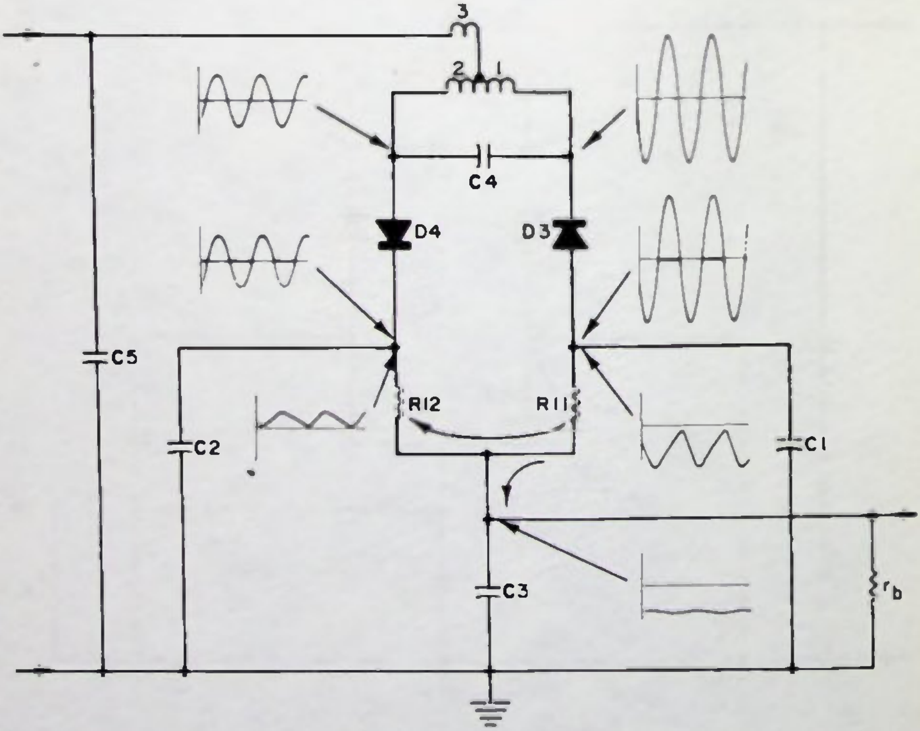


Fig. 4—The voltages present in an unbiased discriminator circuit at a frequency above 4.5MHz.

flow of electrons that had entered the capacitor (C1) and prevents the capacitor from losing a portion of the negative voltage developed across it.

Curve I shows the voltage drops developed across the resistor (R11) when the capacitor (C1) is not in the circuit, while curve J shows the voltage drops developed across the resistor when the capacitor is in the circuit.

The capacitors (C2 and C1) have reduced the maximum and minimum voltage drops developed across the resistors (R12 and R11) and have had the effect of shorting to ground some of the IF signal that would have otherwise been developed across the resistors.

Voltages induced across the secondary windings of the discriminator transformer at 4.5MHz (Fig. 2) result in a positive voltage drop across resistor R12 and a negative voltage drop across resistor R11. The major portion of the IF signal, that would otherwise appear across these resistors, is shorted to ground by capacitors C2 and C1, and the remaining voltages across the two resistors are nearly equal—though of opposite polarity. The positive voltage drop across resistor R12 cancels the negative voltage drop across resistor R11—current flows

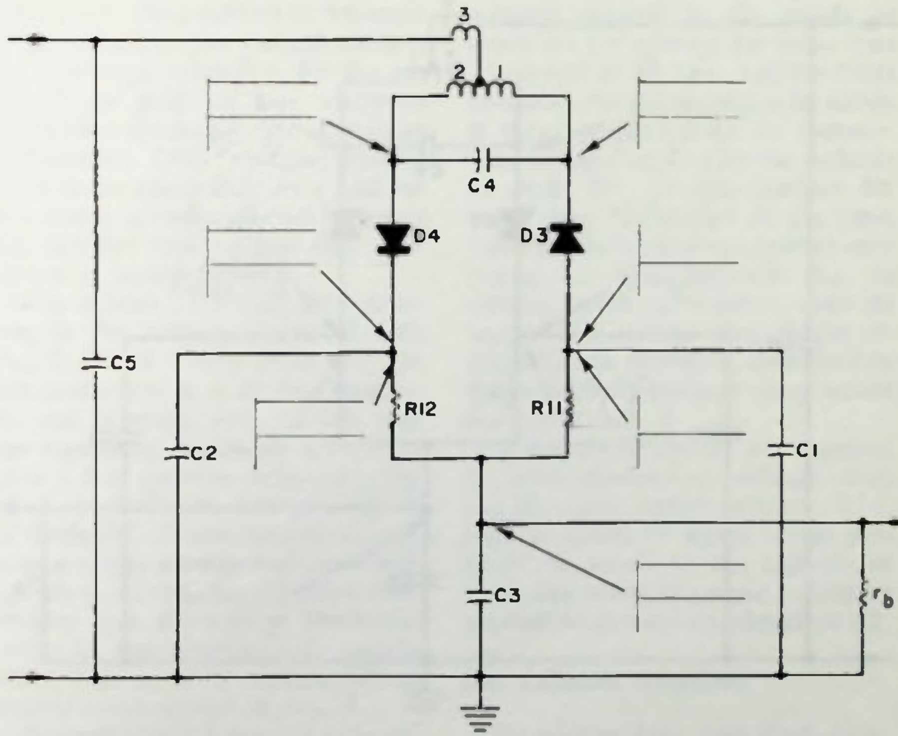


Fig. 5—Voltages present in a biased discrimnator circuit without an IF signal.

from resistor R11 to resistor R12, and no current resulting from the IF signal remains to flow through the effective base resistance (r_b) of the transistor in the next portion of the circuit. At 4.5MHz, the IF signal does not produce a voltage drop across the effective base resistor (r_b) or the capacitor (C3) connected in parallel with it.

Voltages induced across the secondary windings of the discriminator transformer at a frequency below 4.5MHz (Fig. 3) result in a positive voltage drop across resistor R12 that is larger than the negative voltage drop across resistor R11. As a result of the smaller voltage drop across resistor R11, less current flows through that resistor than flows through resistor R12. The remaining current through resistor R12 flows from the effective base resistor (r_b). The IF signal below 4.5MHz results in a positive voltage drop across the effective base resistor (r_b) and the capacitor (C3) connected in series with it.

Voltages induced across the secondary windings of the discriminator transformer at a frequency above 4.5MHz (Fig. 4) result in a positive voltage drop across resistor R12 that is smaller than the negative voltage drop across resistor R11. As a result of the larger voltage drop

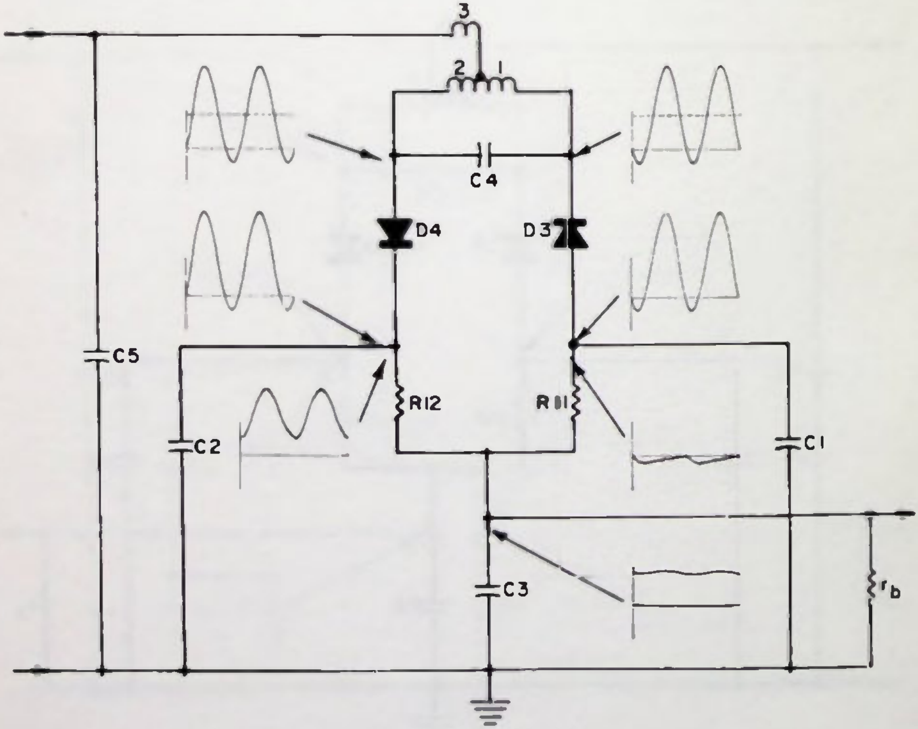


Fig. 6—The voltages present in a biased discriminator circuit at 4.5MHz.

across resistor R11, more current flows through that resistor than flows through resistor R12. The extra current from resistor R12 flows through the effective base resistor (r_b). The IF signal above 4.5MHz results in a negative voltage drop across the effective base resistor (r_b) and the capacitor (C3) connected in series with it.

As the IF signal fluctuates above and below 4.5MHz, the resulting voltage drop across the effective base resistor (r_b) and capacitor C3 fluctuates between positive and negative. The value of capacitor C3 is larger than the value of capacitors C1 and C2, and it (C3) gains and loses more electrons than the other capacitors, when subject to the same voltage change. Not enough current flows through resistor r_b for the capacitor to gain or lose electrons faster than the audio rate of voltage fluctuations. Only voltage fluctuations of the audio frequency and below occur across capacitor C3 and the effective base resistor (r_b) connected in parallel with it.

The diodes (D3 and D4) function in the manner described with Fig. 2, 3 and 4 only when they are not connected to a dc bias supply. We see, however, (Fig. 1) that voltage-regulating transistor Q10 supplies a 2.1v positive potential to the anode of diode D4 and the cathode of diode D3. Unless another voltage source is also used to develop a voltage drop across the effective base resistor (r_b), the voltage drops that occur in the discriminator circuit, when there is no IF signal, will resemble those shown in Fig. 5.

Since diodes D3 and D4 will conduct curent only from their cathodes to their anodes, and electrical currents flow only from negative to positive, only diode D4 can conduct current. A current flowing through diode D4 will develop a positive voltage drop across resistor R12, while no voltage drop will occur across resistor R11; since diode D3 is not conducting current. All of the current flowing through resistor R12 will, therefore, have to flow through the effective base resistor (r_b), which will also develop a positive voltage drop.

When a 4.5MHz signal is induced in the secondary of the transformer, in a circuit containing the potentials shown in Fig. 5, the resulting signals resemble those shown in Fig. 6. The voltages applied to the anode of diode D4 are positive for more than 75 percent of the time, and the diode conducts current as long as its anode is more positive than its cathode. The voltages applied to the cathode of diode D3 are also positive for more than 75 percent of the time, and this diode conducts current only during the short intervals that its cathode is more negative than its anode. The voltage drop across resistor R12 is, therefore, considerably larger than the voltage drop across resistor R11.

If weaker IF signals were applied to both diodes, no voltage drop would occur across resistor R11, and the entire IF signal would pass from the anode to the cathode of diode D4 where the signal would be shorted to ground by capacitor C2.

The Cascade Amplifier

Transistors Q11 and Q12 (Fig. 7) form a cascade amplifier (the signals amplified by the first transistor are then amplified by the second transistor) in integrated circuit, IC201.

The emitter current of transistor Q12 passes through resistor R13. The voltages shown in Fig. 1 indicate that when no signal is applied to the integrated circuit (and the base of transistor Q11) there is a 0.7v potential drop across resistor R13. The emitter of transistor Q13 is, therefore, normally at that potential above ground.

The base of transistor Q12 is only a fraction of a volt of dc potential is present across diodes D3 and D4 (Fig. 2, 3 and 4). The manufacturer indicates that normally 4 to 6 P-P volts of 4.5MHz IF signal is induced in the secondary winding of the discriminator transformer (Fig. 2). The slight dc bias across the diodes (D4 and D3) is, therefore, nearly insignificant when compared to the signal voltages.

The effective base resistor (r_b) described with Fig. 2, 3 and 4 is the effective resistance between the base of transistor Q11 (Fig. 7) and

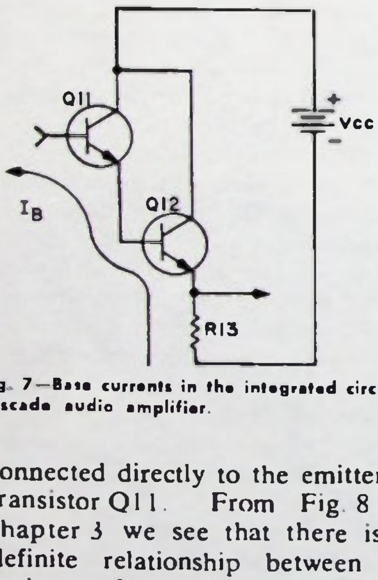


Fig. 7—Base currents in the integrated circuit's cascade audio amplifier.

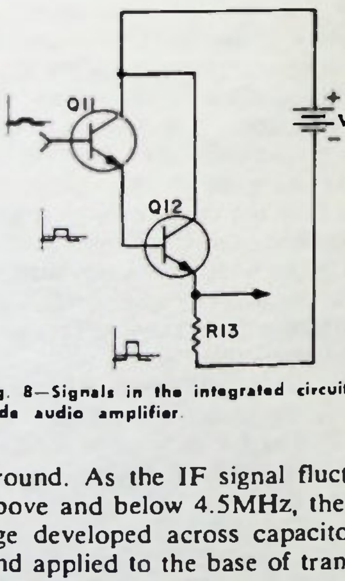


Fig. 8—Signals in the integrated circuit's cascade audio amplifier.

connected directly to the emitter of transistor Q11. From Fig. 8 in chapter 3 we see that there is a definite relationship between the emitter-to-base voltage (V_{EB}) and the base current (I_B) of transistor Q12—which in this case is also the emitter current (I_E) of transistor Q11. The emitter of transistor Q11 is, therefore, more positive than the emitter of transistor Q12, which is more positive than ground.

The base of transistor Q11 is at least as positive as its emitter and ground. As the IF signal fluctuates above and below 4.5MHz, the voltage developed across capacitor C3 and applied to the base of transistor Q11 (Fig. 8) increases and decreases, providing the audio signal that is transmitted by the sound intermediate frequency.

As the base of transistor Q11 becomes more positive, there is a reduction in the transistor's collector-to-emitter voltage and the transistor (Q11) conducts more current. The emitter current conducted by transistor Q11 is also the base current of transistor Q12. As the emitter of transistor Q11 becomes more positive, because of the reduced voltage drop across the transistor, the base of transistor Q12 also becomes more positive and its emitter current also increases. This results in an increased voltage drop across its emitter resistor (R13).

The audio signal developed in the discriminator circuit and applied to the base of transistor Q11 is amplified by transistors Q11 and Q12 and appears across resistor R13—the output of the integrated circuit. From there, the audio signal passes through a volume control and to additional stages of audio amplification outside the integrated circuit before the audio is applied to the speaker.

CHAPTER 15

The Varicap

Voltage-variable capacitance diodes, varicaps, have been used in the past for microwave tuning (as varactors) and in some FM and TV AFC circuits. With recent improvements, these diodes can now substitute for mechanical variable capacitors to tune AM and FM radios.

General Capacitor Characteristics

The basic principles of nonvariable capacitors also apply to varicaps. These principles may be merely a review for experienced technicians.

The study of capacitor time constants in the previous chapter indicated that electrons enter or leave a capacitor as the potential drop across it changes. The number of electrons that enter or leave the two portions of the capacitor is dependent on the ac potential across the capacitor and the value of the capacitor.

The material that separates the two portions of some capacitors virtually prohibits the flow of electrons between the two portions of the capacitor. Although electrons are not flowing from one capacitor lead to the other, the concentration of electrons in one portion of the capacitor increases when a negative potential is applied to it, and the concentration of electrons in the other portion of the capacitor decreases when a positive potential is applied to it. When an ac voltage is applied to the capacitor, the potential across each side of the capacitor alternates between negative and positive, and electrons enter and leave each side of the capacitor. This results in an alternating current flow in the two capacitor leads.

Since, in some capacitors, virtually no electrons pass from one portion of the capacitor to the other, virtually no current can pass from one capacitor lead to the other. After the capacitor has been charged or discharged, virtually no current will flow through its leads. In the process of charging and discharging, however, electrons enter and leave the two portions of the capacitor. The resulting current flow through the capacitor leads is not from one capacitor lead to the other, but between the capacitor leads and the portions of the capacitor gaining and losing electrons.

The ac potential applied across a capacitor results in ac currents through the capacitor leads. This ac current only appears to flow between the two leads, and no dc current flows between the leads of a capacitor that contains a perfect insulator between its two portions.

The apparent ability of a capacitor to conduct an ac current increases with the frequency of the ac voltage applied across it and is expressed in terms of the capacitor's reactance

$$(I = \frac{X}{X_C}).$$

This relationship is illustrated in Fig. 11, chapter 4, and is shown in the equation:

$$X_C = \frac{1}{2\pi fC}.$$

The material separating the two portions of a capacitor is usually not nearly a perfect insulator, and some electrons may pass through it. These capacitors experience a "leakage" of electrons and some current may "leak" through capacitors at frequencies from dc up. The amount of current resulting from the capacitor's leakage, unlike the current resulting from the capacitor's reactance, is not dependent on the frequency of the ac voltage applied to the capacitor. It depends instead on the composition and quality of the capacitor.

If a manufacturer increases the resistance (R_p) of the material that is used to separate the two portions of a capacitor, he reduces the leakage current. This resistance (R_p) has the same effect as a resistor connected in parallel with the capacitor. By reducing the leakage curent, the manufacturer produces a higher quality (Q_p) capacitor, since the capacitor's reactance is then a greater factor in determining the total current conducted. This quality factor (Q_p) can be calculated if the leakage resistance (R_p) and capacitor reactance (X_C) is known

$$(Q_p = \frac{R_p}{X_C}).$$

The two portions of the capacitor that gain and lose electrons, the leads connected to them and the connections all resist the flow of electrons. They have the same effect as a resistor (R_s) connected in series with the capacitor. This resistance (R_s) reduces the ac current that results from the capacitor's reactance (X_C), and by reducing the resistance the manufacturer produces a higher quality (Q_s) capacitor. This quality factor (Q_s) can be calculated if the effective series resistance (R_s) and the capacitor reactance (X_C) is known

$$(Q_s = \frac{X_C}{R_s}).$$

The over-all quality (Q) of a capacitor is reduced by both its leakage resistance (R_p) being too low and its effective series resistance (R_s) being too large. This over-all quality factor (Q) must, therefore, be less than the two quality factors (Q_p and Q_s) already calculated

$$(\frac{1}{Q} = \frac{1}{Q_p} + \frac{1}{Q_s} = \frac{X_C}{R_p} + \frac{R_s}{X_C}),$$

but equally dependent on both.

The over-all quality factor (Q) of a capacitor depends on the relationship of the capacitor's leakage resistance (R_p) and effective series resistance (R_s) with its reactance (X_C). This factor (Q) changes as the capacitor's reactance (X_C) changes with the frequency of the applied ac voltage

$$(X_C = \frac{1}{2\pi f C}).$$

Most circuit applications require capacitors with loaded Q's exceeding 20 and often 100 or more. The low unloaded Q's of the previously available voltage-variable capacitors limited the application of these components.

Most capacitors basically resemble the one shown as the top illustration in Fig. 1. The lower illustration in that figure is, of course, a conventional diagram of a capacitor. The ground side (right side) is the same for both.

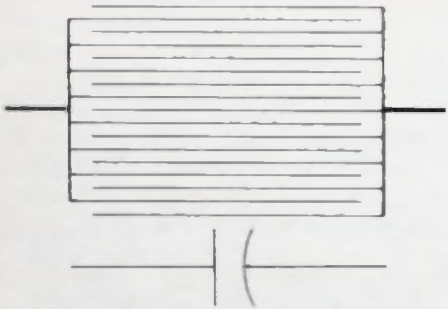


Fig. 1 — The basic structure of most capacitors (above) and a conventional diagram of a capacitor (below).

Capacitors are constructed so that the maximum area of the conductive surfaces attached to their two leads are exposed to one another. The capacitor functions by gaining and losing electrons in these surfaces. The greater the number of surfaces and the greater the area of these surfaces, the more electrons they can gain or lose and the greater the value of the capacitor.

The capacitor shown in Fig. 1 contains 32 surfaces exposed to one another. Only the bottom surface of the material represented by the top line is exposed to another surface and only the top surface of the material represented by the bottom line is exposed to another surface, while both surfaces of the material represented by the other lines are exposed to other surfaces.

The right side of the capacitor shown in Fig. 1 is the ground side since the outer material connected to the right side shields the material connected to the left side.

As you may know, unlike electrical charges attract each other just as unlike magnetic poles attract each other. As one set of surfaces loses electrons, when a voltage is applied across a capacitor, its charge becomes more positive — the charge of the remaining protons contained in the atoms of material that the conductive surfaces are made of. The excess electrons in the other set of surfaces have a negative charge and are attracted by the surfaces containing the positive charge. The greater the attraction of electrons for the positive surface, the more electrons the negative surface can contain.

Each proton in an atom has a positive charge virtually equal to the negative charge of an electron. If the proton's positive charge is attracting excess electrons on the negative surface, the proton has less energy to attract electrons on its own positive surface. The greater the proton's attraction for the negative surface, the more electrons the positive surface can lose.

When the negative surfaces are close to the positive surfaces, their unlike charges have a greater attraction toward one another than when the negative and positive surfaces are further apart. The spacing of the conductive surfaces, like their surface areas, is a factor that determines their capacitance.

The capacitance of some capacitors, such as mica trimmer capacitors, is changed by increasing or reducing the spacing between the two sets of surfaces. When this spacing is reduced (Fig. 2), the component's capacitance increases.

In some other capacitors, like the mechanical ones generally used to tune AM or FM radios, the spacing of the two sets of surfaces remains the same while the area of each set, adjacent to the other set, is increased or decreased. When the area of adjacent surfaces is reduced (Fig. 3), the capacitance is also reduced.

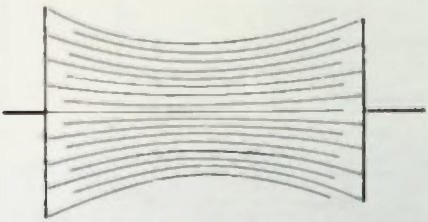


Fig. 2 — Reducing the spacing between the surfaces in a capacitor increases the capacitance.

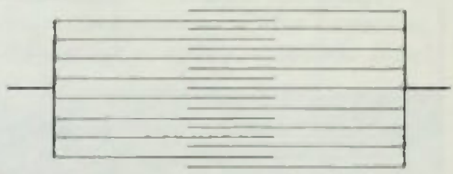


Fig. 3 — Reducing the area of each set of surfaces, adjacent to another set, reduces the capacitance.

As has been indicated, the resistance of the material separating the charged surfaces in a capacitor determines the amount of "leakage" in the capacitor and its quality (Q_P and Q). The separating material also effects its capacitance.

Just as a magnet's ability to attract objects varies with the material its field must travel through, the attractive force of unlike electrical charges varies with the material this force must travel through. The material that seperates the two sets of charged surfaces in a capacitor is called the dielectric. The dielectric's ability to transmit the attractive force of unlike charges is called the dielectric constant. The dielectric constant of air or a vacuum is 1.0, paper is 4.0, pyrex glass is 4.2, clear indian mica is 7.5, aluminum oxide is 10.0 and tantalum oxide is 11.0.

The value of a capacitor can be calculated using the factors that have been described in the equation

$$C = 2.235\frac{KA}{d}(N - 1)\,10^{-11},$$

where:

C = capacitance in Farads

K = dielectric constant

A = average area of effective surfaces (described with Fig. 3)

d = distance between plates

N = number of surfaces exposed to one another described with Fig. 1.

Diode Capacitors

The P-type material in a diode's anode (Fig. 4A) forms a junction with the N-type material in its cathode. This junction of P- and N-type material normally resists the flow of electrons (Fig. 4B). As first described with Fig. 3 in the first chapter and then described in more detail with Fig. 13 in chapter 13, a diode will conduct current (Fig. 4C) when a negative voltage is applied to its cathode and a positive voltage is applied to its anode. When a positive voltage is applied to its cathode and a negative voltage is applied to its anode, the junction of P- and N-type material has a relatively high resistance and functions as a barrier to current flow (Fig. 4D). The greater the amount of this reverse bias, the greater the barrier developed (Fig. 4E).

The P- and N-type material forms the two portions of the voltage-variable capacitance diode that gain or lose electrons. When the diode is reverse biased (its anode more negative than its cathode) a barrier separates the two portions of the diode. Like the capacitors described earlier in this chapter, the

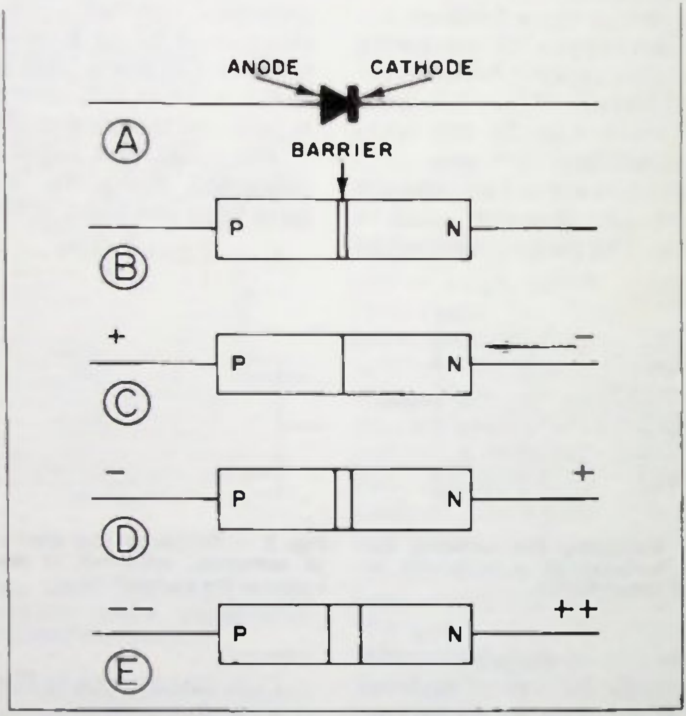


Fig. 4(A) — A diode's anode contains P-type material while its cathode contains N-type material. (B) — The junction of P- and N-type material forms a barrier blocking the flow of electrons. (C) — Current will flow through a diode when its cathode is more negative than its anode. (D) — The junction resists the flow of current when the anode is more negative than the cathode. (E) — The barrier becomes greater as the reverse bias increases.

ac current then flowing through the diode's leads is basically the result of electrons gained or lost in the P- and N-type portions of the diode, rather than the result of an ac current "leaking" between the two portions of the diode.

As the reverse bias is increased, the barrier between the two portions of the diode becomes greater and the diode's capacitance is reduced. As the reverse bias is reduced, the bias barrier becomes smaller and the diode's capacitance increases. This relationship between diode capacitance and reverse bias voltage is shown in Fig. 5 for a voltage-variable capacitance diode recently developed for tuning AM radios. Two of these diodes are shown (Fig. 6) next to a two-stage mechanical tuning capacitor they replace.

Voltage-Current Phases

A review of voltage-current phase relationships may help some technicians understand the application of voltage-variable capacitance diodes in resonant circuits for tuning receivers.

A relationship between electron flow (current) and the voltage across a capacitor is described in the previous chapter. The circuit shown in Fig. 7 is used to explain this relationship in greater detail. When the switch is closed in the circuit, electrons flow from one portion of the capacitor to the battery and from the battery to the other portion of the capacitor. The largest current (electron flow)

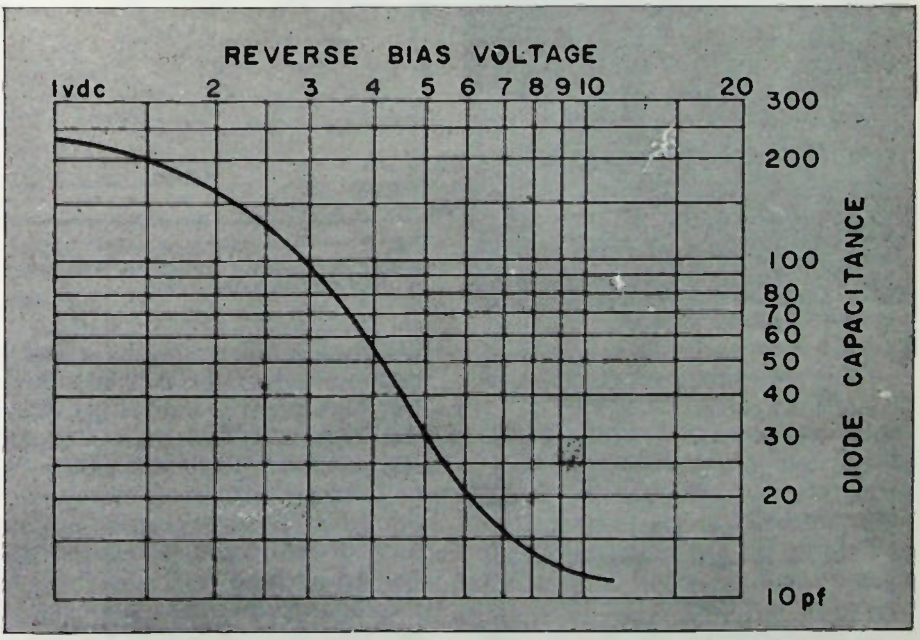


Fig. 5 — The relationship between reverse bias voltage and capacitance in a voltage-variable capacitance diode recently developed for tuning AM radios. Courtesy of Motorola.

occurs just after the switch is closed. before the voltage first appears across the capacitor. The smallest current occurs as the voltage across the capacitor becomes as large as the voltage across the battery. This voltage-current relationship is shown in Fig. 8.

Sinewaves were described with Figs. 7 and 8 in chapter 13. There, the various portions of a sinewave are compared with the angles of a rotating line. The height of a rotating line increases and decreases more rapidly when it is nearly horizontal than when it is nearly vertical up or down. Sinewave voltages change more rapidly at around 0 and 270deg than they do around 90 and 270deg.

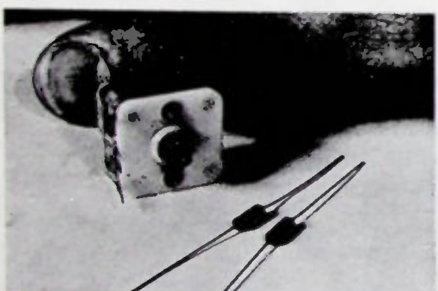


Fig. 6 — Tuning diodes are compared to a miniature mechanical tuning capacitor they can replace.

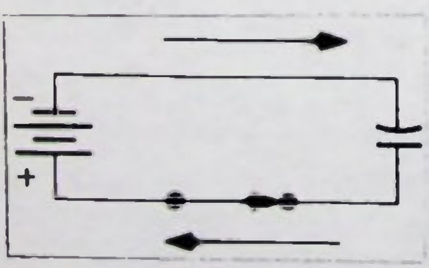


Fig. 7 — The greatest dc current flows through a capacitor when the switch is first closed.

When an ac voltage, resembling a sinewave, is applied across a capacitor, the voltage alternates between positive and negative. The slowest change in voltage occurs when portions of the capacitor are at nearly the maximum positive or negative voltage (90 or 270deg) and the fastest change occurs at the lower voltages or when portions of the capacitor shift between positive and negative polarities (0 and 180deg).

In Fig. 9 we see that as the voltage in one portion of a capacitor

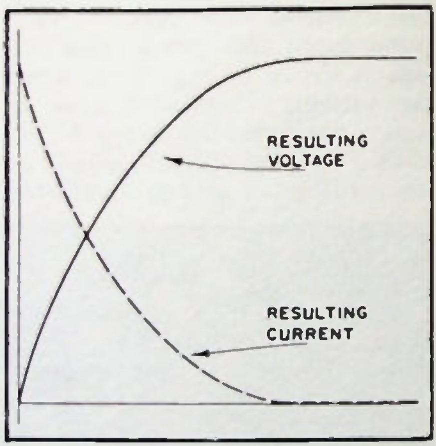


Fig. 8 — The voltage-current relationship that occurs when a dc voltage is applied across a capacitor.

becomes positive (0 to 90deg), that portion of the capacitor loses electrons and a current flows out of the connecting lead. As it becomes less positive and then negative (90 to 270deg), electrons return through the lead. Electrons again reverse their direction and flow out of the connecting lead as this portion of the capacitor becomes less negative and then positive (270 to 360deg).

Since the greatest change in voltage occurs when the voltage shifts between positive and negative, the maximum flow of electrons (current) out of the capacitor lead (the same lead described in the preceding paragraph) occurs at 0 and 360-deg, while the maximum flow of electrons into the lead occurs at 180deg.

Coil Impedances

Resonant circuits are equally dependent on the impedance of capacitors and the impedance of coils. Unless the impedance of coils and their related voltage-current phase relationships are clearly understood, the function of voltage-variable capacitance diodes, as well as mechanical diodes, are beyond the scope of the average technician. Although coils are not semiconductors, a review of their function is important for a clear understanding of semiconductor circuits.

The function of coils in electronic circuits is dependent on the magnetic fields they produce. Similar magnetic fields can be produced by a permanent magnet.

When a permanent magnet approaches the end of a coil (Fig. 10A), it induces a current through the coil's windings, and one end of the coil becomes more negative than the other. While the magnet remains stationary near the coil (Fig. 10B), no current is induced in the coil. As the magnet moves away from the coil (Fig. 10C), a current is again induced through the coil's winding. This current, however, flows in a direction opposite that of the current induced by the approaching magnet, and the coil's other lead is now more negative.

When the south pole of a magnet moves away from a coil (Fig. 11), the induced current flows through the coil in the same direction as a current induced when the north pole of a magnet approaches the coil.

The circuits shown in Fig. 12 help explain the function of a coil. When the switch is first closed (Fig. 12A), an electrical current flows through both the coil and the resistor connected in parallel with it. As the current begins to flow through the coil, it produces a magnetic field in the coil corresponding to a permanent magnet approaching the coil. Just as an approaching perma-

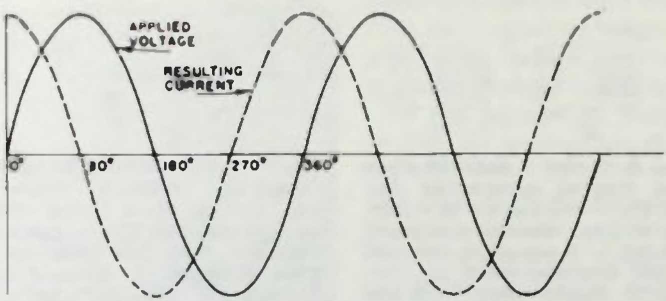


Fig. 9 — The voltage-current relationship that occurs when an ac voltage is applied across a capacitor.

nent magnet induces a current in the coil, the increasing magnetic field induces a current opposing the current from the battery.

Once the magnetic field has attained its maximum strength and ceases to increase (Fig. 12B), it, like a permanent magnet remaining stationary near a coil, no longer induces a current. The maximum amount of current can then flow from the battery through the coil.

When the switch is opened (Fig. 12C), no voltage from the battery is applied across the coil and the magnetic field decreases. Just as a receding permanent magnet induces a current through a coil, the depleting field also induces a current through the coil, which flows

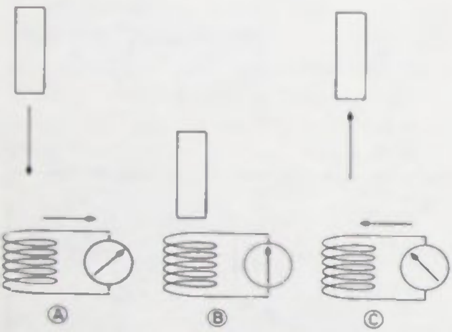


Fig. 10 (A) — A current is induced when a permanent magnet approaches one end of a coil (B) — No current is induced when the magnet remains stationary near the coil (C) — A current is induced in the opposite direction when the permanent magnet moves away from the coil.

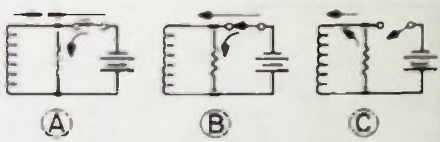


Fig. 12 (A) — When the switch is first closed, an induced current impedes any current flow from the battery through the coil. (B) — Once the coil's magnetic field has reached its maximum strength, no current is induced through the coil to impede any current from the battery through the coil. (C) — When the switch is opened, an induced current flows through the coil in the same direction as the battery current.

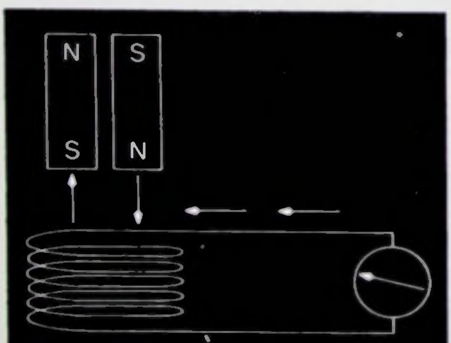


Fig. 11 — The same current is induced when the north pole of a magnet approaches a coil as when the south pole of a magnet is moved away from that end of the coil.

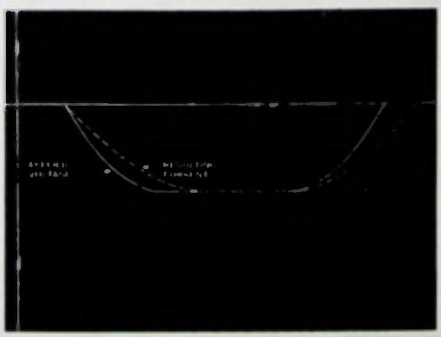


Fig. 13 — The negative current lags behind the increase and decrease in negative voltage.

through the resistor. The current now induced flows in a direction opposite that of the current previously induced. The induced current flows in the same direction as the current that had been flowing from the battery.

The increasing negative current (Fig. 13) in the circuit described lags behind the increasing negative voltage, and the decreasing negative current lags behind the decreasing negative voltage. If the polarity of the battery is reversed in the circuit (Fig. 12), the increasing positive current (Fig. 14) lags behind the increasing positive voltage and the decreasing positive current lags behind the decreasing positive voltage.

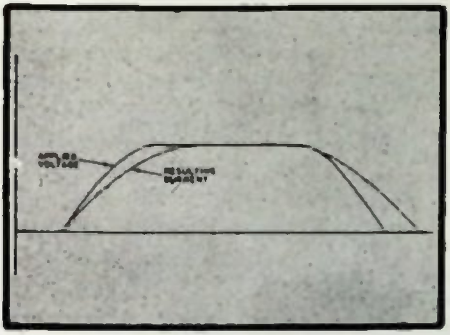


Fig. 14 — The positive current lags behind the increase and decrease in positive voltage.

By combining the curves shown in Fig. 13 and 14 we see the current that results when a positive voltage is applied to the circuit, until the maximum positive current flows through the coil (Fig. 14); the voltage drops to zero, until the current ceases to flow through the coil; a negative voltage is applied, until the maximum negative current flows through the coil; etc. The positive and negative changes in current always lag behind the positive and negative changes in voltage.

If the permanent magnet shown in Fig. 10 approaches the coil and departs from it at a more rapid rate, the induced current becomes greater. As indicated earlier in this chapter, sinewave voltages do not change at a uniform rate. With an ac sinewave voltage applied across a coil, there is a more rapid change in the magnetic field when the applied voltage changes polarity than

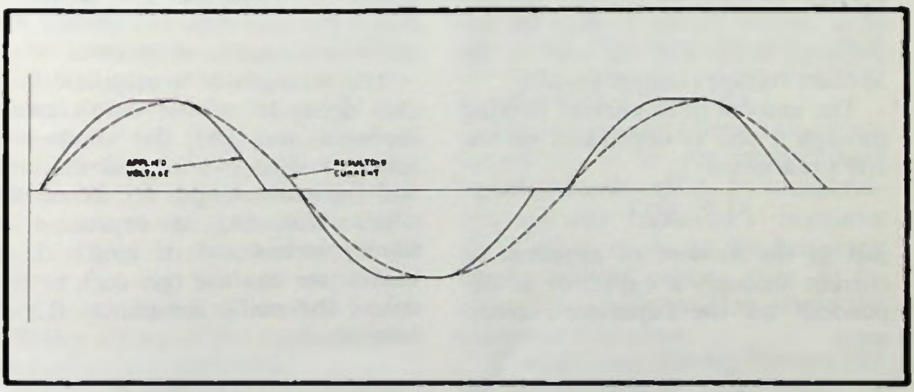


Fig. 15 — As a dc voltage is switched between positive and negative, the resulting current lags behind the voltage.

when the voltage reaches its maximum positive or negative value. The coil's induced current is therefore the greatest, further impeding or enhancing the current from the power source (Fig. 16), when the applied voltage changes polarity. Measurements indicate that when an ac voltage is applied across a coil, the current always lags 90deg behind the voltage.

When comparing the voltage-current phase relationship for a capacitor (Fig. 9) with the voltage-current phase relationship for a coil (Fig. 16), we see that in a capacitor the current leads the voltage 90deg while in a coil the voltage leads the current 90deg. The greatest amount of ac current passes through the capacitor's leads and the coil as the applied voltage changes polarity.

The amount of ac current flowing through a coil is dependent on the coil's reactance

$$(I \approx \frac{V}{X_L}),$$

just as the amount of apparent ac current through a capacitor is dependent on the capacitor's reactance

$$(I \approx \frac{V}{X_C}).$$

Although a capacitor appears to have less resistance to ac currents at higher frequencies, a coil's resistance to ac currents increases as the applied voltages and magnetic field fluctuate at higher frequencies. The coil's reactance can be calculated with the equation $X_L = 2\pi fL$, where L is the inductance of the coil.

The coil's inductance (L) is dependent on the magnetic field it produces. The greater the field's concentration, the greater the coil's inductance and the smaller the ac current it will conduct.

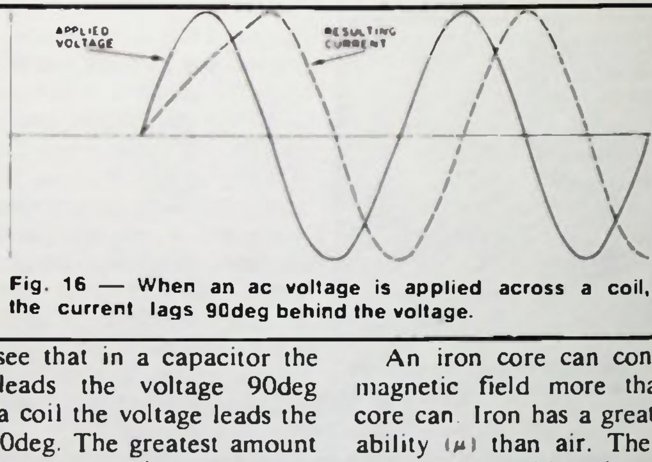


Fig. 16 — When an ac voltage is applied across a coil, the current lags 90deg behind the voltage.

An iron core can concentrate a magnetic field more than an air core can. Iron has a greater permeability (μ) than air. The greater a core's permeability, the greater the coil's resistance to ac currents.

The strength of a magnetic field also depends on the core's cross-sectional area (A), the number of turns of wire (N) around the core and the coil's length (l). When the core's area (A) is expressed in square inches and its length (l) in inches, we can use this data to calculate the coil's inductance (L) in henries,

$$L = \frac{\mu N^2 A 10^{-8}}{l}.$$

CHAPTER 16

Varicap Applications

For several years varicaps have been used in FM tuners to compensate for frequency drift. Until fairly recently the diode's restricted tuning range has limited its general consumer applications to these automatic frequency control (AFC) circuits, although it has had important microwave applications.

In the past some technicians have been able to repair FM receivers without understanding the function of this diode. Their luck, however, is now beginning to run out, and an understanding of this semiconductor is essential if they are to service new electronic circuits effectively.

A thorough understanding of capacitors and coils is required before varicap functions can be comprehended, and to help the reader review or develop these fundamentals the preceding chapter shows the voltage-current relationships that occur in these components. As indicated in Fig. 1, the phase angle of current flowing through capacitor leads (represented by dashed lines) is 90deg ahead of the applied voltage (shown as solid lines).

When a capacitor is connected to a regulated-voltage signal generator (Fig. 2), the current (I) flowing through the capacitor leads is dependent on the capacitor's reactance (X_C),

$$(I = \frac{E}{X_C}),$$

which in turn is dependent on the frequency (f) of the applied voltage

$$(X_C = \frac{1}{2\pi fC}).$$

In Fig. 1 we see that as the frequency (f) of the applied voltage increases, the capacitor's reactance decreases and more current flows through its leads. (The upper pair of curves (A) in Fig. 1, 4, 5 and 8 represent a lower frequency than the center pair of curves (B), while the lower pair of curves (C) represent a higher frequency.)

The voltage (E) across a capacitor connected to a regulated-current signal generator (Fig. 3) is dependent on the capacitor's reactance (X_C), ($E = IX_C$). As the frequency of the applied current (Fig. 4) increases, the voltage drop across the capacitor decreases.

The previous chapter showed that the voltage-current relationship in a capacitor differed from the relationship in a coil. As indicated in Fig. 5, the phase angle of current flowing through a coil (represented by dashed lines) is 90deg behind the applied voltage (shown as solid lines).

When a coil is connected to a regulated-voltage signal generator

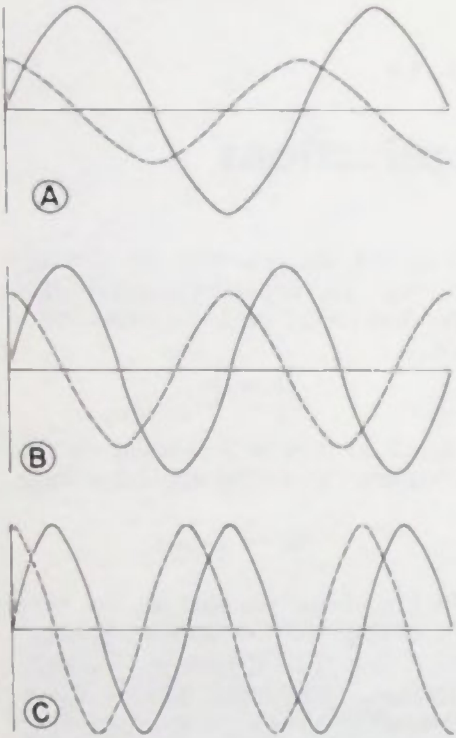


Fig. 1 — The phase angle of current flowing through a capacitor's leads (dashed curves) is 90deg ahead of the applied voltage (solid curves). The current increases as the frequency of the applied voltage increases from a lower (A) to an intermediate (B) to a higher (C) frequency.

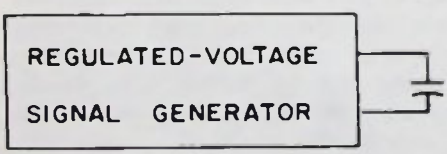


Fig. 2 — A capacitor is connected to the output of a regulated-voltage signal generator.

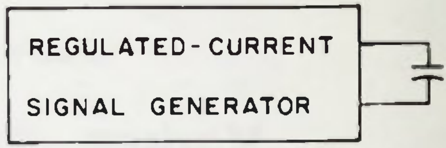


Fig. 3 — A capacitor is connected to the output of a regulated-current signal generator.

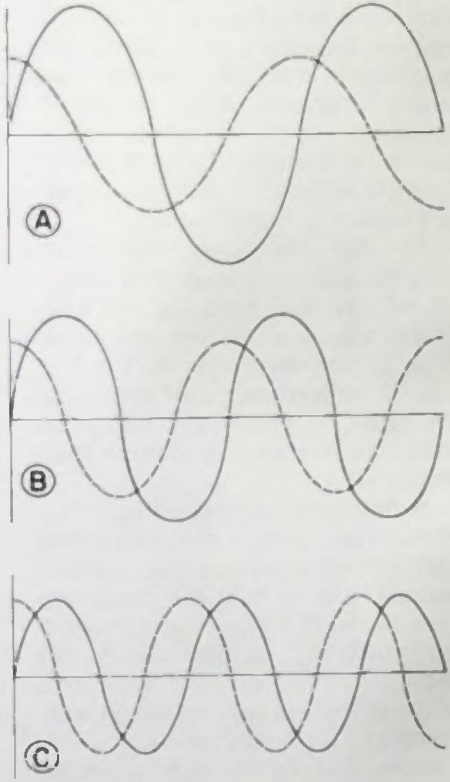


Fig. 4 — The voltage drop across a capacitor (solid curves) decreases as the frequency of the applied current (dashed curves) increases from a lower (A) to an intermediate (B) to a higher (C) frequency.

(Fig. 6), the current (I) flowing through the coil is dependent on the coil's reactance (X_L),

$$(I = \frac{E}{X_L}),$$

which in turn is dependent on the frequency (f) of the applied voltage ($X_C = 2\pi f L$). In Fig. 5 we see that as the frequency (f) of the applied voltage increases, the coil's reactance also increases and less current flows through the coil.

The voltage (E) across a coil connected to a regulated-current signal generator (Fig. 7) is dependent on the coil's reactance (X_L), ($E = IX_L$). As the frequency of the applied current (Fig. 8) increases, the voltage drop across the coil also increases.

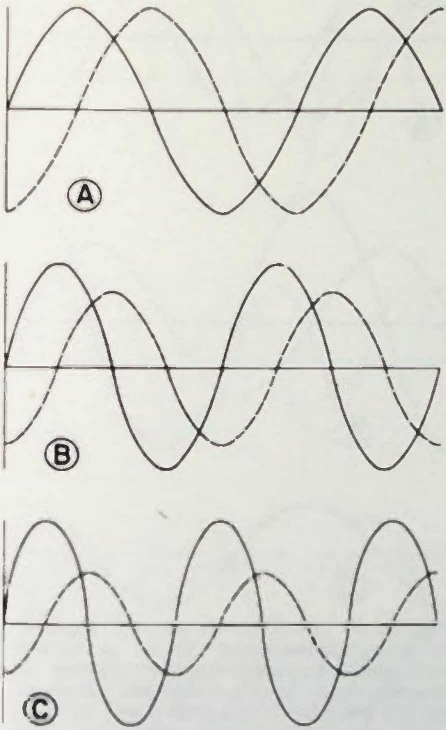


Fig. 5 — The phase angle of current flowing through the coil (dashed curves) is 90deg behind the applied voltage (solid curves). The current decreases as the frequency of the applied voltage increases from a lower (A) to an intermediate (B) to a higher (C) frequency.

Both series and parallel capacitor-coil tuned circuits depend on the capacitor and coil voltage-current relationships.

Series-Resonant Circuits

The capacitor and coil in Fig. 9

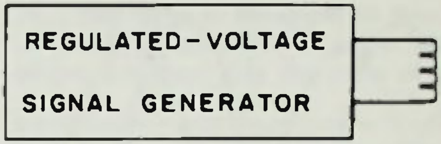


Fig. 6 — A coil is connected to the output of a regulated-voltage signal generator.

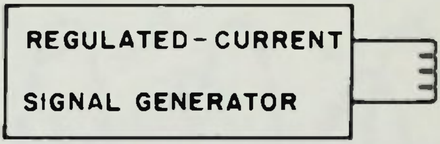


Fig. 7 — A coil is connected to the output of a regulated-current signal generator.

are connected in series to a regulated-current signal generator, and the same amount of current must flow through both the capacitor leads and coil. When the frequency (f) of the current from the signal generator is below the circuit's resonant frequency (f_r), the coil's reactance (X_L) is smaller than the capacitor's reactance (X_C). (When $f < f_r$, $X_L < X_C$.) Under these conditions (Fig. 10), a smaller voltage drop occurs across the coil (solid curve A) than across the capacitor (solid curve B).

Since the coil and capacitor are connected in series and the same current must flow through both, the phase angle of the current in both components (dashed curves A and B) must be the same. The phase angle of the voltage across the coil (solid curve A) leads the coil current (dashed curve A) 90deg, while the phase angle of the voltage across the capacitor (solid curve B) lags behind the capacitor lead current (dashed curve B) 90deg. The voltages across the two components are therefore 180deg out of phase with each other, or of opposite polarity. When the voltage across the coil

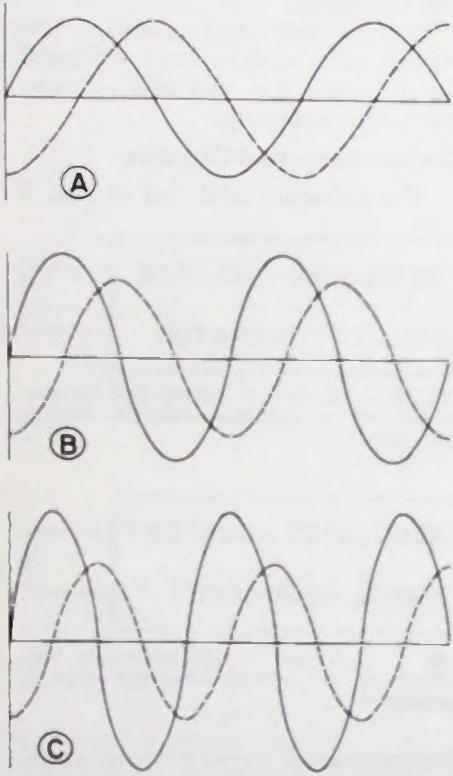


Fig. 8 — The voltage drop across a coil (solid curves) increases as the frequency of the applied current (dashed curves) increases from a lower (A) to an intermediate (B) to a higher (C) frequency.

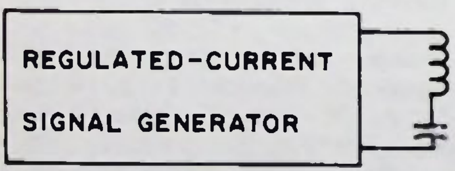


Fig. 9 — A capacitor and coil are connected in series to a regulated-current signal generator.

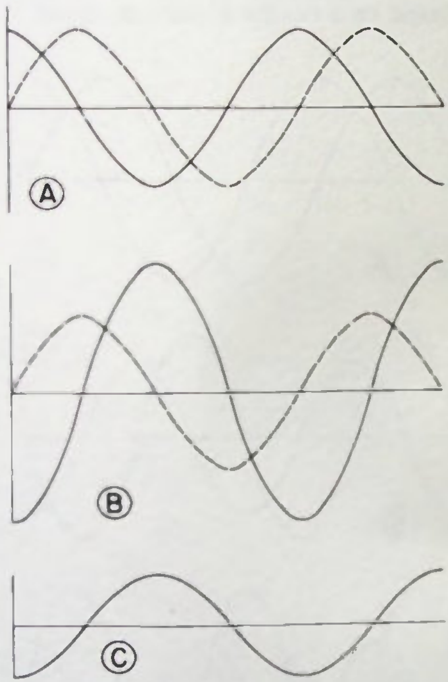


Fig. 10 — When the ac current applied to a series-resonant circuit is below resonant frequency (dashed curves A and B), there is a greater voltage drop across the capacitor (solid curve B) than across the coil (solid curve A). The curve that results (curve C) when adding the amplitudes of these two curves (solid curves A and B) represents the voltage at the signal generator. The signal generator voltage is in phase with the capacitor voltage.

(solid curve A) is positive, the voltage across the capacitor (solid curve B) is negative.

The voltage at the regulated-current signal generator is equal to the total voltage across the coil and capacitor. If the voltage across the coil is +8v and the voltage across the capacitor is −16v, the total voltage at the signal generator is −8v.

From these calculations it is apparent that at this frequency, with the components used, an ac voltmeter would indicate a voltage across the coil greater than the voltage at the signal generator and a voltage across the capacitor equal to the voltage at the signal generator. Since the same amount of current flows from the regulated-current signal generator through the coil and capacitor, the apparent power supplied by the signal generator ($P_S = IE_S$) is less than the apparent power at the coil ($P_L = IE_L$) and equal to the apparent power at the capacitor ($P_C = IE_C$). We know, however, that the parts of a

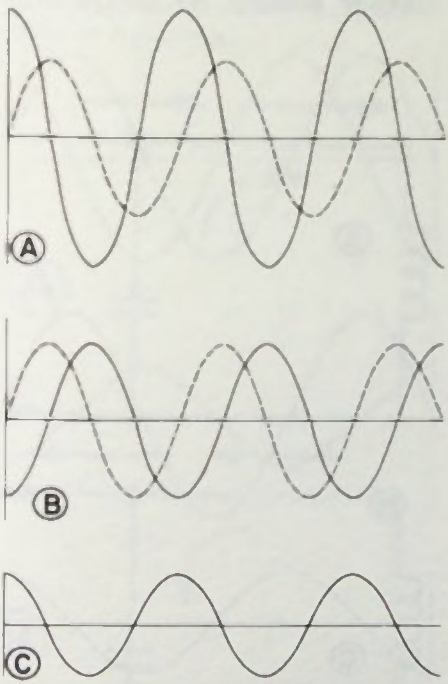


Fig. 11 — When the ac current applied to a series-resonant circuit is above resonant frequency (dashed curves A and B), there is a greater voltage drop across the coil (solid curve A) than across the capacitor (solid curve B). The curve that results (curve C) when adding the amplitudes of these two curves (solid curves A and B) represents the voltage at the signal generator. The signal generator voltage is in phase with the coil voltage.

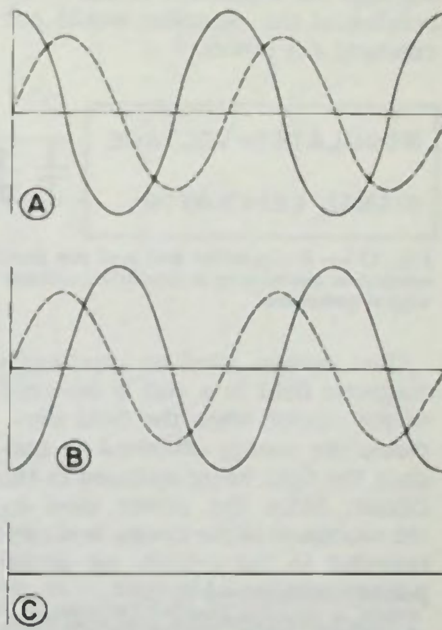


Fig. 12 — When the ac current applied to a series-resonant circuit is at resonant frequency (dashed curves A and B), the voltage drop across the capacitor (solid curve B) is as large as that across the coil (solid curve A). No curve results (line C) when adding the amplitudes of these two curves (solid curves A and B), and virtually no voltage is present at the signal generator.

circuit cannot absorb more power than is supplied the circuit and, therefore, the apparent power at the coil and capacitor cannot be the actual power consumed.

Measurements with a wattmeter indicate that the apparent power ($P = IE$) is the true power only in ac or dc resistance circuits. If the circuit contained a capacitor that could have a perfect insulator separating its two segments, no electrons could pass between the two segments, and current entering one capacitor lead would return out the same lead without ever passing through the capacitor. Under these conditions the capacitor would not consume any power.

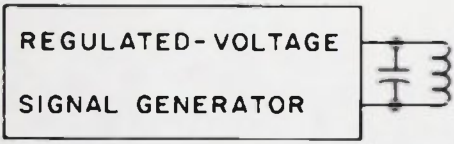


Fig. 13 — A capacitor and coil are connected in parallel to a regulated-voltage signal generator.

The current used to produce a magnetic field in a coil is returned to the circuit when the field dissipates, the energy absorbed to produce the field being returned to the circuit. Since the power used by the reactance in the circuit is always returned to the circuit, no actual power is dissipated by it.

When the capacitor and coil (Fig. 9) are connected in series to the signal generator and the frequency (f) of the resulting current is above the circuit's resonant frequency (f_r), the coil's reactance (X_L) is greater than the capacitor's reactance (X_C). (When $f > f_r$, $X_L > X_C$.) Under these conditions (Fig. 11), a larger voltage drop occurs across the coil (solid curve A) than across the capacitor (solid curve B). If the voltage across the coil is +16v and the voltage across the capacitor is −8v, the total voltage at the signal generator is +8v.

When the frequency (f) of the current from the signal generator is equal to the voltage drop across the capacitor. Since the two equal voltages are 180deg out of phase, of opposite polarity, no voltage drop

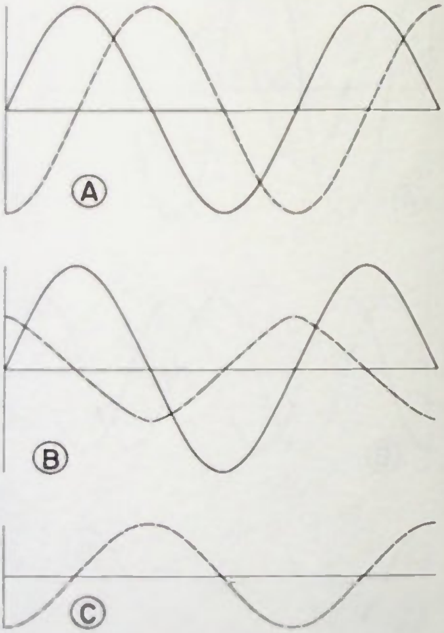


Fig. 14 — When the ac voltage applied to a parallel-resonant circuit is below resonant frequency (solid curves A and B), more current flows through the coil (dashed curve A) than through the capacitor leads (dashed curve B). The curve that results (curve C) when adding the amplitudes of these two curves (dashed curves A and B) represents the current from the signal generator. The signal generator current is in phase with the coil current.

the same as the circuit's resonant frequency (f_r), the coil's reactance (X_L) is equal to the capacitor's reactance (X_C) (When $f = f_r$, $X_L = X_C$.) The voltage drop across the coil (Fig. 12) is then occurs across the signal generator ($+12v - 12v = 0v$).

When the frequency of the signal generator current (Fig. 9) is either above or below the resonant frequency, the capacitor and coil impedances result in a signal generator voltage that does not occur at the resonant frequency.

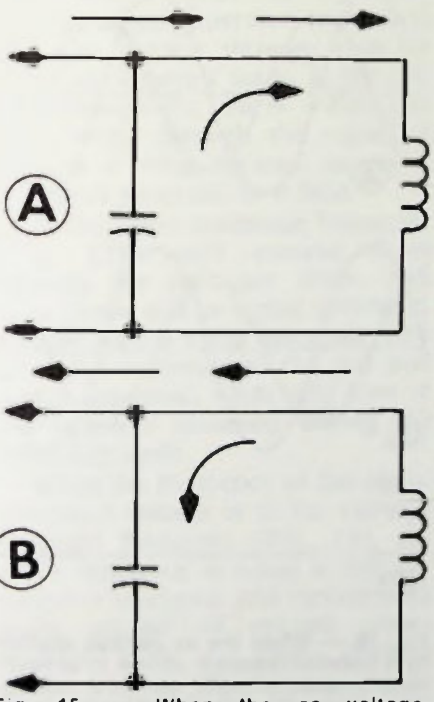


Fig. 15 — When the ac voltage is below resonant frequency, electrons flow from the signal generator and capacitor through the coil during half a cycle (A) and flow through the coil, entering the capacitor and signal generator, during the other half cycle (B).

Parallel-Resonant Circuits

The capacitor and coil in Fig. 13 are connected in parallel to a regulated-voltage signal generator, and the same amount of voltage is present across both the capacitor and coil. When the frequency of the signal generator voltage is below the circuit's resonant frequency (Fig. 14), the coil's reactance is smaller than the capacitor's reactance and more current flows through the coil

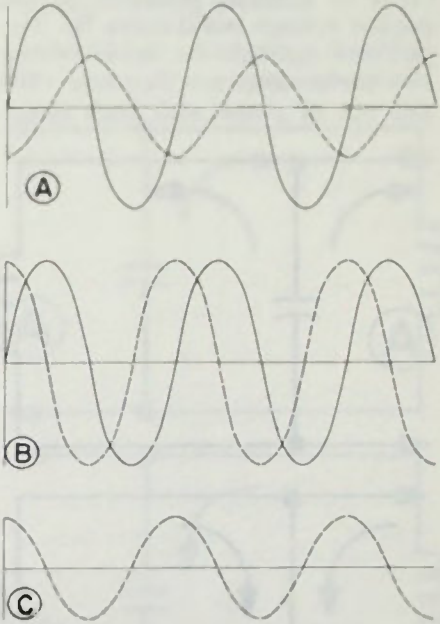


Fig. 16 — When the ac voltage applied to a parallel-resonant circuit is above resonant frequency (solid curves A and B), more current flows through the capacitor leads (dashed curve B) than through the coil (dashed curve A). The curve that results (curve C) when adding the amplitudes of these two curves (dashed curves A and B) represents the current from the signal generator. The signal generator current is in phase with the capacitor lead current.

(dashed curve A) than through the capacitor leads (dashed curve B).

Since the coil and capacitor are connected in parallel and the same voltage is applied to both, the phase angle of the voltage drop across both components (solid curves A and B) must be the same. The phase angle of the current through the coil (dashed curve A) lags behind the coil voltage (solid curve A) 90deg, while the phase angle of the current through the capacitor leads (dashed curve B) is 90deg ahead of the capacitor voltage (solid curve B). The currents through the leads of the two components are therefore 180-deg out of phase with each other, or of opposite polarity. When the current through the coil (dashed curve A) is negative, the current through the capacitor leads (dashed curve B) is positive.

The current from the regulated-voltage signal generator is equal to the total current through both the coil and capacitor leads. If the current through the coil is -16ma and the current through the capacitor leads is $+8$ma, the total current at the signal generator is -8ma.

At the below resonance frequency (Fig. 15), more current flows

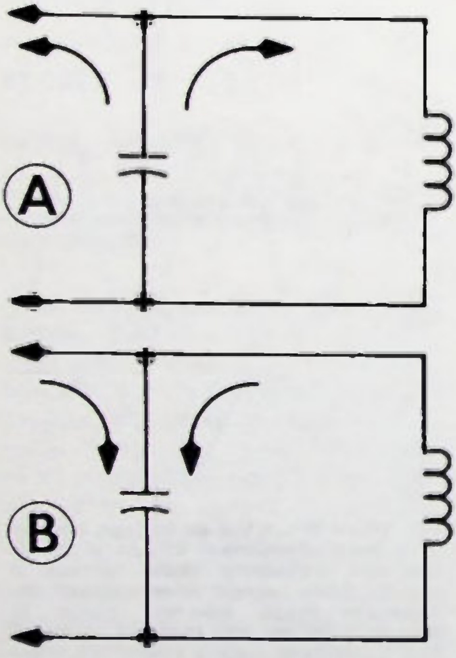


Fig. 17 — When the applied ac voltage is above resonant frequency, electrons flow from the capacitor to the coil and signal generator during half a cycle (A) and flow in the opposite direction during the other half cycle (B).

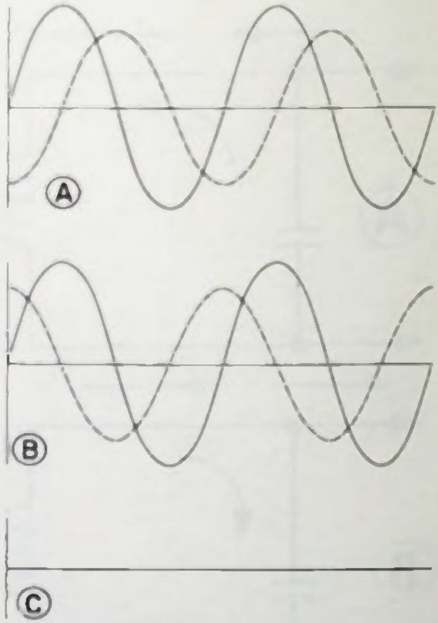


Fig. 18 — When the ac voltage applied to a parallel-resonant circuit is at resonant frequency (solid curves A and B), the current through the capacitor leads (dashed curve B) is as great as the current through the coil (dashed curve A). No curve results (line C) when adding the amplitudes of these two curves (dashed curves A and B), and virtually no current flows from the signal generator.

through the coil than through the capacitor leads or signal generator. During half a cycle electrons flow from the signal generator and capacitor through the coil, while during the other half cycle electrons flowing through the coil enter the capacitor and signal generator.

When the frequency of the signal generator voltage is above the circuit's resonant frequency (Fig. 16), the coil's reactance is larger than the capacitor's reactance and less current flows through the coil (dashed curve A) than through the capacitor leads (dashed curve B). The current from the regulated-voltage signal generator is equal to the total current through both the coil and capacitor leads. If the current through the coil is −8ma and the current through the capacitor leads is +16ma, the total current at the signal generator is +8ma.

At the above resonance frequency (Fig. 17), more current flows through the capacitor leads than through the coil or signal generator. During half a cycle electrons flow from the capacitor to the coil and signal generator, while they flow in the opposite direction during the other half cycle.

When the frequency of the signal generator voltage is at the circuit's resonant frequency (Fig. 18), the coil's reactance is equal to the capacitor's reactance and virtually the same amount of current passes through both of them (dashed curves A and B). If the current through the coil is −12ma and the current through the capacitor leads is +12ma, no current remains to flow from the signal generator and theoretically the electrons flow (Fig. 19) only between the capacitor and coil.

From the series-resonance curves (Fig. 12) we see that the signal generator voltage is a minimum at the resonant frequency, while from the parallel-resonance curves (Fig. 18) we see that the signal generator current is a minimum at the resonant frequency. The voltage or current became greater at frequencies above and below the resonant frequency.

Varicap Tuning Circuits

Succeeding chapters describe the use of series resonant

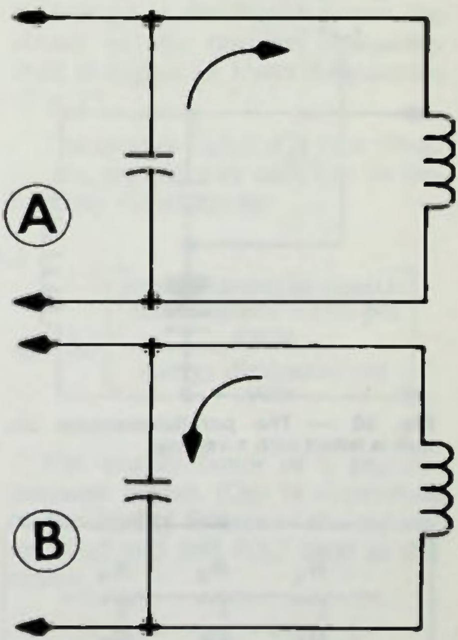


Fig. 19 — When the applied ac voltage is at resonant frequency, virtually all of the electrons flow from the capacitor through the coil during half a cycle (A) and return during the other half cycle (B) without flowing through the signal generator.

circuits for filtering out undesired signals in semiconductor circuits, we are now more concerned with the parallel-resonant circuits used to tune receivers. The parallel-resonant circuit shown in Fig. 20 is tuned with a varicap. The negative voltage applied to the anode of this diode (D1) is isolated from the balance of the tuned circuit by a capacitor (C1), and this capacitor (C1) and diode (D1) are connected in series, parallel to the coil (L1).

Just as series resistances can be added (Fig. 21) to determine the total resistance in a circuit ($R_T = R_1 + R_2 + R_3$), series capacitor reactances can be added to determine the total capacitive reactance in a circuit ($X_{CT} = X_{C1} + X_{C2} + X_{C3}$). By substituting the equation for determining capacitive reactance

$$(X_C = \frac{1}{2\pi f C}),$$

we can calculate the total capacitance in the series:

$$\frac{1}{2\pi f C_T} = \frac{1}{2\pi f C_1} + \frac{1}{2\pi f C_2} + \frac{1}{2\pi f C_3}.$$

Both sides of the equation can be multiplied by the same quantity:

$$\frac{2\pi f}{2\pi f C_T} = \frac{2\pi f}{2\pi f C_1} + \frac{2\pi f}{2\pi f C_2} + \frac{2\pi f}{2\pi f C_3}.$$

This can be simplified to the following:

$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}.$$

If the varicap in Fig. 20 has the characteristics illustrated in Fig. 9 of the preceding chapter, the diode will have a 50 pf capacitance when its reverse bias is 4.25vdc. When connected in series with a 200pf capacitor, the total capacitance in parallel with the coil is 40pf.

$$(\frac{1}{C_T} = \frac{1}{D1} + \frac{1}{C1} = \frac{1}{50\text{pf}} + \frac{1}{200\text{pf}} = 0.02/\text{pf} + 0.005/\text{pf} = 0.025/\text{pf}.\ C_T = \frac{1}{0.025/\text{pf}} = 40\text{pf.})$$

By reducing the reverse bias, the total capacitance is increased, reducing the capacitive reactance and changing the resonant frequency.

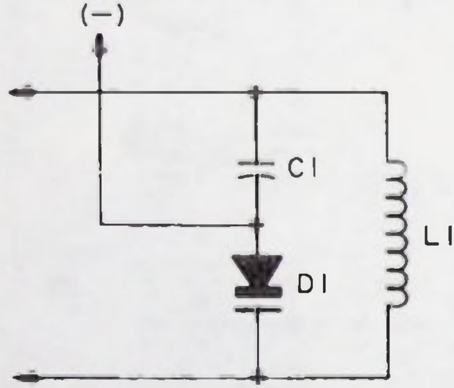


Fig. 20 — The parallel-resonant circuit is tuned with a varicap.

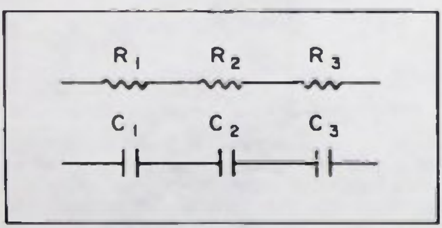


Fig. 21 — These resistors and capacitors are connected in series.

CHAPTER 17

Review of Capacitor & Coil Impedances

Practical tuned circuits do not have perfect characteristics. Their quality is limited. A portion of the ac current applied to a tuned circuit is unaffected by reactance. Their resonant-frequency current is not reduced to zero by the resulting impedance (Fig. 1) as theory would dictate for a perfect resonant circuit (Fig. 18 in chapter 16). Neither does the voltage drop developed across a parallel-resonant tuned circuit decrease to zero at frequencies above and below the resonant frequency (Fig. 2).

Resonant Circuit Q Factor

The amount of signal current flowing through a receiver's parallel-resonant circuit varies with the frequency of the signal. From the curves shown in Fig. 14, 16 and 18 of the preceding chapter, we see that a minimum amount of applied signal current flows through this circuit at a resonant frequency, while greater signal currents occur at higher or lower frequencies (Fig. 1) — the tuned circuit presenting maximum impedance to the flow of current at the resonant frequency. Because of the circuit's greater impedance at the resonant frequency, when signal currents of different frequencies but the same strength pass through a parallel-resonant circuit, a greater voltage is developed across the circuit at the resonant frequency than at higher or lower frequencies (Fig. 2).

The quality factor (Q) of a tuned circuit, capacitor or coil, can be defined by the equation:

$$Q = 2\pi \left[\frac{\text{energy stored in electric and magnetic fields per cycle}}{\text{energy dissipated per cycle}} \right]$$

The quality factor of a parallel-resonant circuit (Q_P) is dependent on the quality factors of the capacitor (Q_C) and coil (Q_L) used in the circuit

$$\left(\frac{1}{Q_P} = \frac{1}{Q_C} + \frac{1}{Q_L}\right).$$

In chapter 15 we saw that a capacitor's quality factor was:

$$\frac{1}{Q_C} = \frac{X_C}{R_{C_P}} + \frac{R_{C_S}}{X_C}.$$

Where: Q_C = Capacitor quality factor.

X_C = Capacitor reactance.

R_{Cp} = Capacitor's effective parallel resistance.

R_{Cs} = Capacitor's effective series resistance.

A coil's quality factor is determined in a similar manner:

$$\frac{1}{Q_L} = \frac{X_L}{R_{Lp}} + \frac{R_{Ls}}{X_L}.$$

In most of the older parallel-resonant circuits the values of R_{Cp} and R_{Lp} were so large while the value of R_{Cs} was so small, with respect to X_C or X_L, that with little error the equation could be simplified to:

$$\frac{1}{Q_P} \approx \frac{1}{Q_L} \approx \frac{R_{Ls}}{X_L} \text{ or}$$

$$Q_P \approx \frac{X_L}{R_{Ls}}.$$

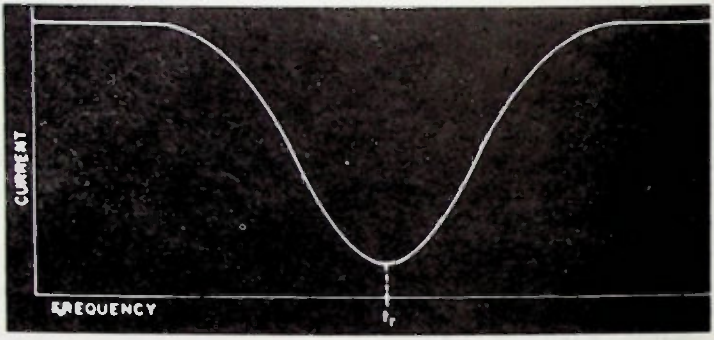


Fig. 1 — The ac current from a constant-voltage signal generator is smallest at the resonant frequency of a parallel-resonant circuit.

Where: Q_L = Coil quality factor.

X_L = Coil reactance.

R_{Lp} = Coil's effective parallel resistance.

R_{Ls} = Coil's effective series resistance.

These equations can be combined to the following:

$$\frac{1}{Q_P} = \frac{1}{Q_C} + \frac{1}{Q_L} =$$

$$\frac{1}{Q_C} + \frac{X_L}{R_{Lp}} + \frac{R_{Ls}}{X_L} \text{, or}$$

$$\frac{X_C}{R_{Cp}} + \frac{R_{Cs}}{X_C} + \frac{X_L}{R_{Lp}} + \frac{R_{Ls}}{X_L}.$$

In the new receivers that use varicap-tuning diodes, however, the capacitor's quality factor (Q_C) is small enough to have a significant effect on the quality factor (Q_P) of the parallel-resonant tuning circuits.

The amount of voltage developed across a parallel-resonant circuit differs with the Q of the circuit (Fig. 3). The greater the Q, the greater the voltage developed and the sharper the reduction in voltage as the signal moves above or below the resonant frequency (f_r).

A tuned circuit's bandwidth (BW)

is defined as the range of signal frequencies that develop a voltage across a tuned circuit that is at least 0.707 times as large as the resonant voltage. The effect of Q on tuned-circuit bandwidth can be more clearly seen with the aid of a selectivity (S) curve (Fig. 4). The amplitude of this curve is determined by the equation:

$$S = \frac{\text{nonresonant voltage}}{\text{resonant voltage}}$$

Since the peak voltage across a parallel-resonant circuit is always the resonant voltage, all curves drawn for these tuned circuits have a maximum value of one, whatever the circuit's Q. The bandwidth for these circuits can be calculated with the equation:

$$BW = \frac{f_r}{Q}$$

Tuning Receivers

When a signal in a receiver is at the resonant frequency of the receiver's parallel-resonant circuit, the circuit's capacitive impedance equals its inductive (coil type) impedance (when $f = f_r$, $X_C = X_L$). In chapter 15 it was indicated that:

$$X_C = \frac{1}{2\pi fC}$$

and $X_L = 2\pi fL$. These equations can be combined, and at the resonant frequency:

$$\frac{1}{2\pi f_r C} = 2\pi f_r L.$$

With algebra, this equation can be converted to a more convenient form. $1 = 4\pi^2 f_r^2 LC$.

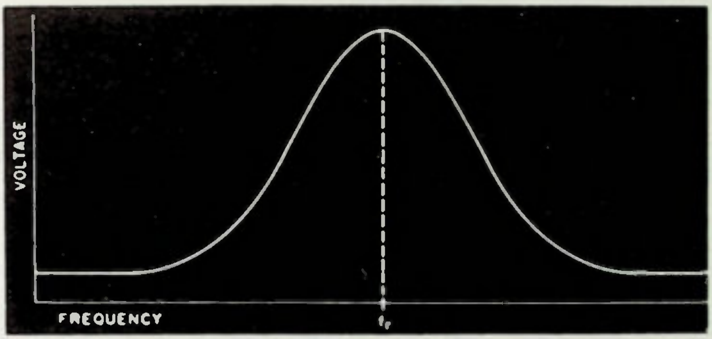


Fig. 2 — The ac voltage from a constant-current signal generator is largest at the resonant frequency of a parallel-resonant circuit.

$$f_r^2 = \frac{1}{4\pi^2 LC}. \quad f_r = \frac{1}{2\pi \sqrt{LC}}$$

From the last equation we see that the larger the parallel-resonant circuit's inductance (L) or capacitance (C), the lower the resonant frequency; and the smaller the circuit's inductance or capacitance, the higher the resonant (tuned) frequency. Receivers can be tuned by changing either the inductance or capacitance in their parallel-resonant tuning circuits.

Some of the older receivers were mechanically tuned by changing the parallel-resonant tuned circuit's inductance by moving a metal core in-and-out of a coil, changing the effective number of windings in a coil or by varying the spacing between coil turns. Most receivers were mechanically tuned, however, by changing the parallel-resonant tuned circuit's capacitance by changing the spacing between capacitor plates or effective capacitor plate surface areas. Modern receivers can now be electronically tuned by varying the bias of varicap tuning diodes.

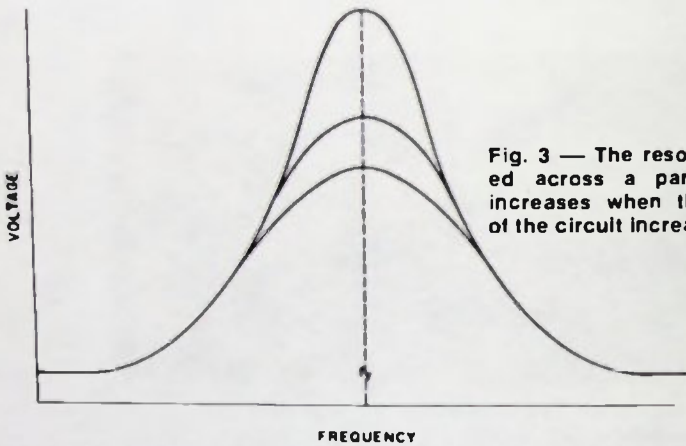


Fig. 3 — The resonant voltage developed across a parallel-resonant circuit increases when the quality factor (Q) of the circuit increases.

Superheterodyne Circuits

Most receivers currently on the market have superheterodyne circuits (Fig. 5) containing a variable-tuned radio-frequency (RF) circuit, a variable-tuned oscillator circuit and a mixer circuit. The RF and oscillator circuits are tuned simultaneously, the oscillator circuit always being tuned to a frequency below that of the RF circuit.

Measurements indicate that when signals of two different frequencies are mixed, the combination produces a waveform having frequencies equal to both the sum and the difference between the two applied frequencies. When an FM receiver's RF circuit is tuned to a 96.0-MHz station and the oscillator is tuned to 85.3MHz, the resulting signals can be combined to produce 181.3 and 10.7MHz signals. The intermediate frequency (IF) amplifier circuits are tuned to handle only those frequencies around 10.7MHz.

Superheterodyne receivers have the advantage of containing additional amplifiers tuned only to the IF frequency, rather than additional amplifiers that must all be tuned to the frequency of the station being received.

Superheterodyne FM receivers at one time had a common problem with their oscillator circuits. As the oscillator's temperature changed, its frequency would shift. When cold, the RF circuit might be tuned to a 96.0MHz station while the oscillator frequency has shifted and is producing a 84.7MHz signal. Since the IF amplifiers are tuned to 10.7MHz, the signals passing through them must originate from a 95.4MHz station (84.7MHz osc. + 10.7MHz IF = 95.4MHz RF). Although the RF circuit is not tuned to that station, programs may still be received if that station's signal is strong. As the oscillator circuit warms up, the frequency it produces may gradually shift to 85.3MHz and the signals from the 96.0MHz station are heard. An automatic frequency control (AFC) circuit was required to prevent this apparent drifting of stations. It was in this circuit that technicians first encountered varicap tuning diodes.

FM AFC Circuits

FM receivers produce an audio input very similar to the audio input produced by integrated circuit IC201 (Fig. 1 in chapter 14). As the FM receiver's IF signal shifts in response to the shifting frequency of the FM station, the output voltage of the discriminator circuit (Figs. 2, 3, and 4 in chapter 14) also shifts. The audio signal consists of these rapid changes in output voltage.

By reversing the anode and cathode leads of diodes D3 and D4 in Fig. 4, chapter 14, the induced current flow is reversed and a positive voltage is developed across capacitor C3 when the IF signal shifts above its mean

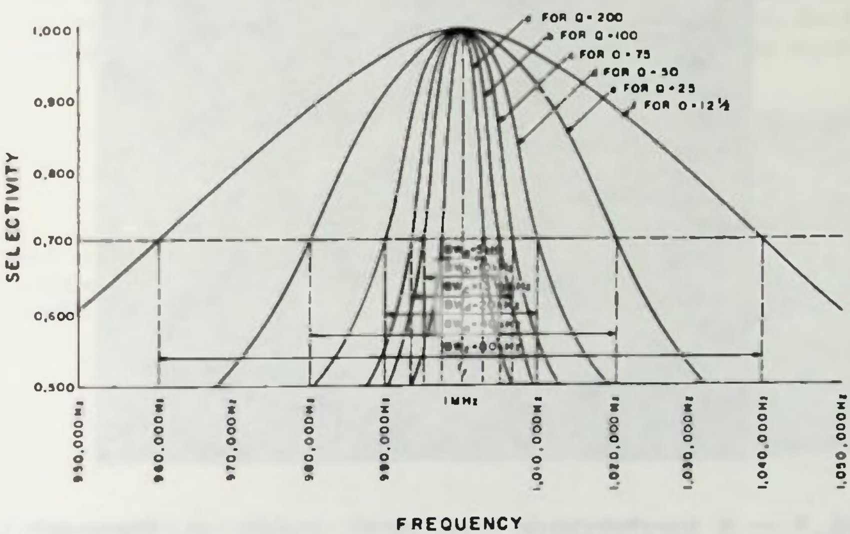


Fig. 4 — The greater a parallel-resonant circuit's quality factor (Q), the smaller the bandwidth (BW) of the resonant frequency.

frequency (4.5MHz for the TV integrated circuit, 10.7MHz for an FM receiver discriminator). It is this output voltage that regulates the AFC oscillator.

A typical FM AFC oscillator circuit is shown in Fig. 6. Any change in the transistor's emitter voltage results in an amplified change in its collector voltage — like the signals present in the common base circuit shown in Fig. 10 in the first chapter. A portion of the amplified signal is returned to the emitter by a 2.7pf capacitor, causing positive feedback. The maximum amount of collector signal voltage is developed across the tuned circuit at the circuit's resonant frequency. At that frequency the positive feedback is large enough to cause the tranistor to oscillate.

The tuned circuit in Fig. 6 is basically the same as the tuned circuit in Fig. 20, chapter 16, except that here two capacitors, rather than one, are connected in series with the varicap. The audio signal normally present with the discriminator circuit's output voltage is filtered from the oscillator circuit with a 0.01μf capacitor and 1M resistor.

Assuming that under some bias condition the varicap has a 2pf capacitance and ignoring the negligible effect of the 3.3K resistor, we can use the series capacitance equation derived in chapter 16 to determine the total capacitance connected in parallel with the

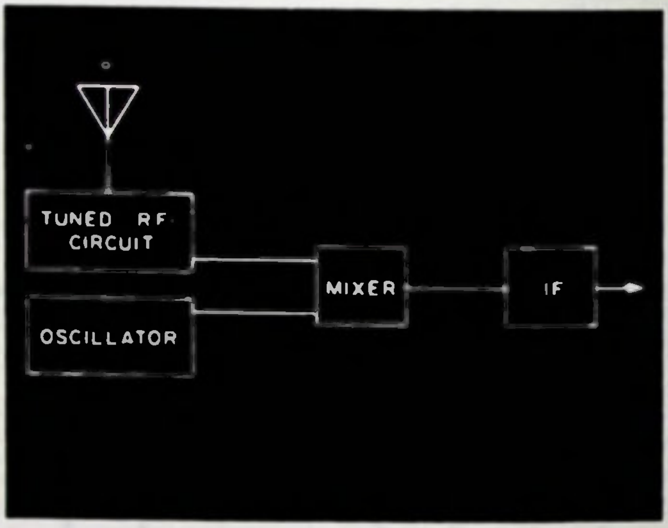


Fig. 5 — A superheterodyne circuit contains a variable-tuned radio-frequency (RF) circuit, a variable-tuned oscillator circuit and a mixer circuit, which produce an intermediate-frequency (IF) signal for additional amplification.

2 to 7.5pf tuning and trimmer capacitors.

$$\frac{1}{C_T} \approx \frac{1}{6.8pf} + \frac{1}{1000pf} + \frac{1}{2pf} =$$

0.147/pf + 0.001/pf + 0.500/pf = 0.648/pf.

$$C_T \approx \frac{1}{0.648/pf} \approx 1.5pf.$$

From these calculations we see that when capacitors are connected in series, their total capacitance is always less than that of the smallest capacitor.

$$I_1 = \frac{V}{R_1}, I_2 = \frac{V}{R_2}, I_3 = \frac{V}{R_3},$$

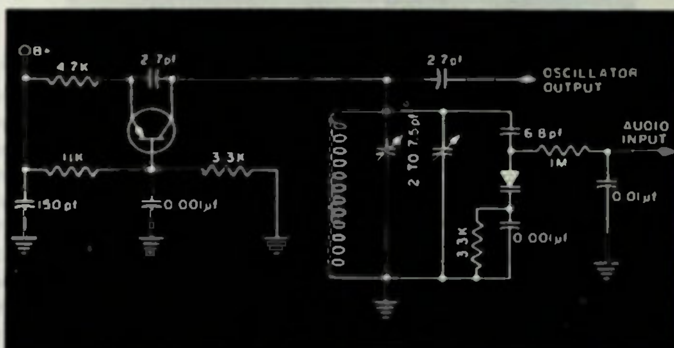
$$I_T = \frac{V}{R_T}.$$

By substituting the second set of equations for parts of the first, we can calculate the circuit's total parallel resistance.

$$\frac{V}{R_T} = \frac{V}{R_1} + \frac{V}{R_2} + \frac{V}{R_3}.$$

If both sides of the equation are divided by V, then we have the well-known parallel resistance equation:

Fig. 6 — A typical AFC circuit for an FM receiver.



An equation can also be derived to determine the total value of capacitors connected in parallel with the coil in the tuned circuit (Fig. 6).

When resistors are connected in parallel (Fig. 7), the total current passing through the circuit is equal to the sum of the currents passing through each resistor. $I_T = I_1 + I_2 + I_3$.

As you know, current is equal to voltage divided by resistance, and since the resistors are connected in parallel, the same voltage is applied across all of them.

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}.$$

The total reactance of parallel capacitors (Fig. 7) can be determined with the parallel resistance equation:

$$\frac{1}{X_{CT}} = \frac{1}{X_{C1}} + \frac{1}{X_{C2}} + \frac{1}{X_{C3}}.$$

By substituting

$$\frac{1}{2\pi fC}$$

for X_C in that equation, we get the following:

$$\frac{1}{\frac{1}{2\pi f C_T}} = \frac{1}{\frac{1}{2\pi f C_1}} + \frac{1}{\frac{1}{2\pi f C_2}} + \frac{1}{\frac{1}{2\pi f C_3}}.$$

This can be simplified to: $2\pi f C_T = 2\pi f C_1 + 2\pi f C_2 + 2\pi f C_3$. After dividing both sides of the equation by $2\pi f$, we get: $C_T = C_1 + C_2 + C_3$. The total capacitance of a parallel capacitor circuit is equal to the sum of the individual capacitances.

When the FM oscillator tuning and trimmer capacitors (Fig. 6) have a 4.0pf total capacitance, and the varicap and the two capacitors connected in parallel with it have a 1.5-pf total capacitance, the tuned circuit contains 5.5pf (4.0pf + 1.5pf) of capacitance in parallel with the coil.

As the bias voltage across the varicap changes, the total capacitance of the tuned circuit also changes. When the IF signal drifts to a frequency slightly higher than normal, the discriminator's output voltage becomes slightly larger than normal and additional reverse bias develops across the varicap. This, in turn, reduces the diode's capacitance, lowering the oscillator's resonant frequency and returning the IF signal to normal.

TV AFTC Circuits

Automatic fine tuning controls in the newer TV receivers operate basically on the same principle as FM AFC circuits. A discriminator circuit operating on the same principle as the one discussed earlier (Fig. 4, chapter 14), is connected to the receiver's IF circuit. When the frequency of the IF signal increases and shifts into the discriminator's negative voltage area, a smaller positive potential is applied across the varicap diode in the TV receiver's UHF tuner (Fig. 8). The tuner's oscillator section is shown in a more familiar form in Fig. 9.

In this circuit coil L7 serves merely to supply the transistor's collector with negative dc current,

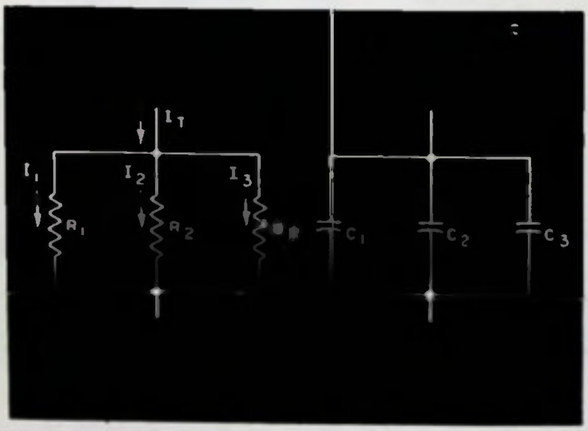


Fig. 7 — In both parallel resistor and parallel capacitor circuits, the total current passing through the circuit is equal to the sum of the currents passing through each component.

while the coil's impedance prevents UHF signals from passing through it to ground. This coil is required since no dc current is able to pass through coil L6 because of a variable capacitor connected in series with it. Its impedance is so much greater than that of coil L6 ($X_{L7} > X_{L6}$) that it (L7) has virtually no effect on the tuning of the oscillator circuit.

From the curves in Fig. 11 chapter 15, we saw that circuit (Fig. 9) is varied by the capacitor connected in series with it, tuning the oscillator's parallel-resonant circuit. This type of circuit is required for UHF tuners since even very small coils have relatively large impedances at these frequencies.

From Figs. 7 and 8 in chap- chapter 4 we see that at ultrahigh frequencies (UHF) the phase shift and capacitance in a transistor results in positive feedback.

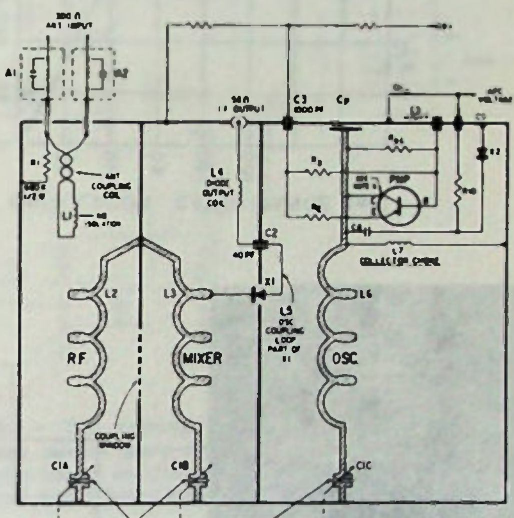


Fig. 8 — A schematic of a UHF tuner used in a TV set. Courtesy of Zenith.

when a coil and a capacitor are connected in series and the coil's impedance is greater than the capacitor's impedance ($X_L > X_C$), the resulting ac voltage drop across the pair of components is like that across just a coil — their phase angles are the same. The capacitor has served the function of reducing the coil's effective impedance.

The effective impedance of coil L6 in the TV receiver's oscillator

At the resonant frequency of the parallel-resonant circuit (the effective impedance of coil L6 in conjunction with the capacitors connected in parallel with it), there is sufficient collector signal voltage for the transistor, with positive internal feedback, to oscillate.

The varicap diode (X2), connected in parallel with the oscillator tuning coil (L6), also has a function in tuning the oscillator circuit.

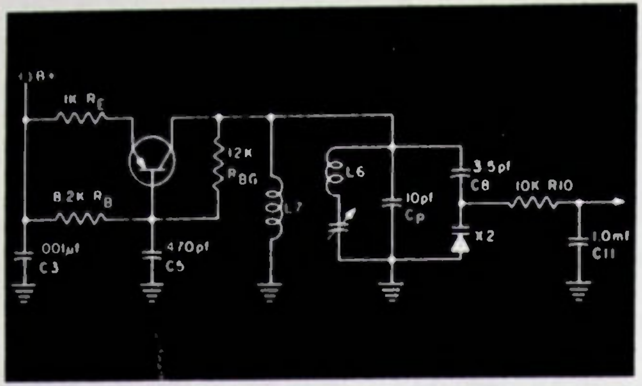


Fig. 9 — The UHF tuner's oscillator circuit is shown in a more familiar form.

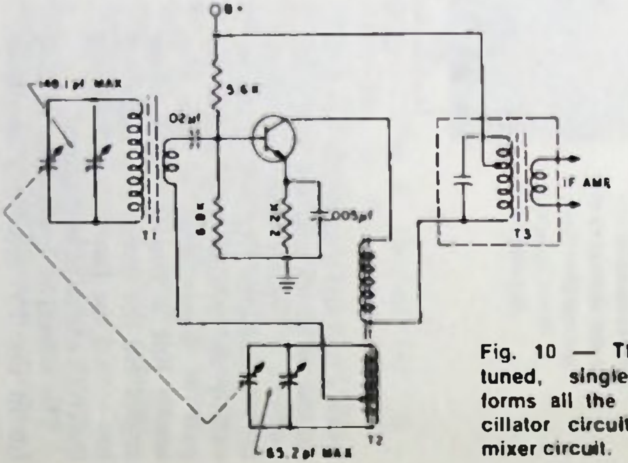


Fig. 10 — This mechanical-capacitor tuned, single-transistor circuit performs all the functions of a tuned oscillator circuit, tuned RF circuit and mixer circuit.

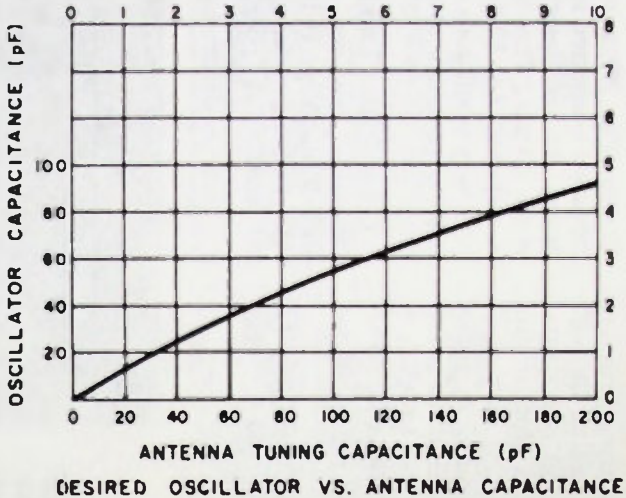


Fig. 11 — The desired combination of oscillator and antenna tuned circuit capacitances are shown for a transistor AM receiver. Courtesy of Motorola.

As the applied dc potential from the IF discriminator circuit decreases with a higher IF frequency, the varicap's capacitance increases, and the oscillator circuit oscillates at a lower frequency. This circuit, in effect, adjusts the receiver's tuner oscillator for the best TV signal reception.

of one transformer (T2) and its parallel capacitors (65.2pf maximum total parallel value components) form a parallel resonant circuit, and at their resonant frequency maximum collector signals are induced across the transformer's secondary winding (T2). These signals return through the ferrous an-

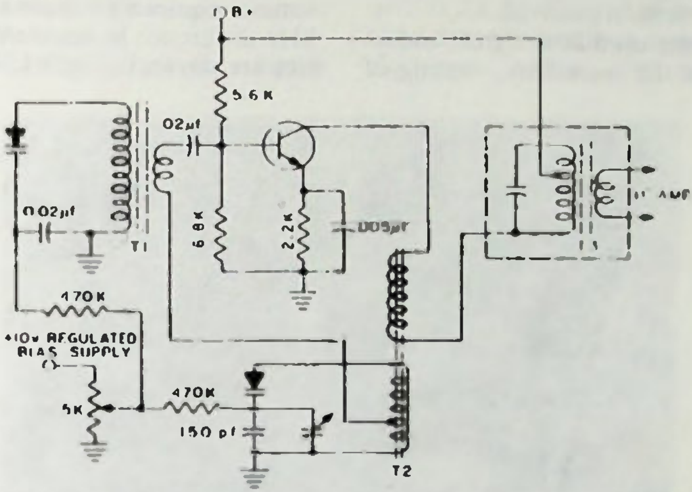


Fig. 12 — The varicap tuned, single-transistor circuit performs all the functions of a tuned oscillator circuit, tuned RF circuit and mixer circuit.

AM Varicap Tuning

The single transistor circuit shown in Fig. 10 is typical of the circuits used in many transistor radios, and it performs all the functions of a tuned oscillator circuit, tuned RF circuit and mixer circuit.

In this circuit transistor collector-current signals pass through the primary windings of transformers T2 and T3. The secondary winding

tenna secondary (T1) and 0.02μf capacitor to the base of the transistor. This positive feedback circuit results in an oscillation at the tuned resonant frequency of the transformer's secondary (T2).

The parallel-resonant circuit, formed by the primary winding of the ferrous antenna (T1) and the capacitors connected in parallel with it (148.1pf maximum total parallel value components), is tuned to the radio frequency (RF) of the station being received. The greater the voltage induced across the antenna's primary winding, the greater the voltage also induced across its secondary winding. The largest signal voltage is induced across these windings when the tuned parallel-resonant frequency of the primary winding is the frequency of the station received.

Both the oscillator signal, induced across the secondary winding of transformer T2, and the RF signal, induced across the secondary winding of the ferrous antenna (T1), pass through the 0.02μf capacitor and are applied to the base of the transistor. The pair of signals are amplified by the transistor and the resulting collector-current signal passes through the primary winding of the IF transformer (T3). The primary winding of this transformer is tuned, with a capacitor connected in parallel with it, to a 455kHz resonant frequency. This frequency is the difference between the RF and oscillator frequencies.

The oscillator and RF resonant frequencies are tuned in this receiver by changing capacitor values in the oscillator and RF parallel-resonant circuits. The corresponding oscillator and RF tuning capacitor values required to maintain a 455-kHz difference in resonant frequencies are shown in Fig. 11.

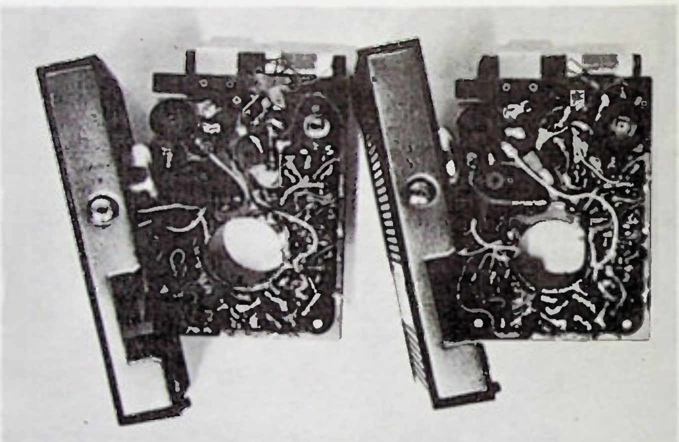


Fig. 13 — The underside of two AM transistor receiver circuit boards is shown. The left one is tuned with a mechanical-tuning capacitor and the one on the right is tuned with a varicap.

The circuit shown in Fig. 10 can be modified (Fig. 12) to eliminate the mechanical tuning capacitors. The varicap diode functions in the RF tuned circuit (Fig. 12) in the same manner as it does in the parallel-resonant circuit shown in Fig. 20, chapter 16. Variations in the dc potential applied across the varicap change the component's capacitance and the resonant frequency of the parallel-resonant circuit. The 5K potentiometer (Fig. 12), connected between a positive voltage source and ground, varies the varicap potential and the resonant frequency of both the RF and oscillator circuits. The 470K resistors between the potentiometer and varicaps serve to isolate IF and oscillator signals, while the varicap bias current is so small that any bias voltage drop across the resistors is insignificant.

The varicap capacitance range is more than adequate for the RF tuning circuit and, therefore, a trimmer capacitor was not required for that circuit. A trimmer capacitor, connected either in parallel with the 150pf capacitor or in parallel with both the 150pf capacitor and the varicap, is used to adjust the oscillator circuit capacitance so that a 455kHz difference exists between IF and oscillator frequencies.

The varicap manufacturer substituted its diode in place of the tuning capacitor in a portable AM radio. Using tuning circuits similar to those shown in Fig. 12, the capacitances in the antenna and oscillator circuits vary with the applied voltage as indicated by the curves in Fig. 15.

Tests were made in accordance with IEEE Standard 186 before and after this receiver was converted to electronic tuning. The results of these tests are shown in Table I.

The curve in Fig. 15 indicates small capacitive changes as the bias voltage is varied between 7 and 10v, and we noted that the last station at the higher-frequency end of the dial covered a much larger portion of the tuning dial than the other stations. Since the potentiometer rotates a greater number of degrees than the mechanical tuning capacitor, the stations were more widely

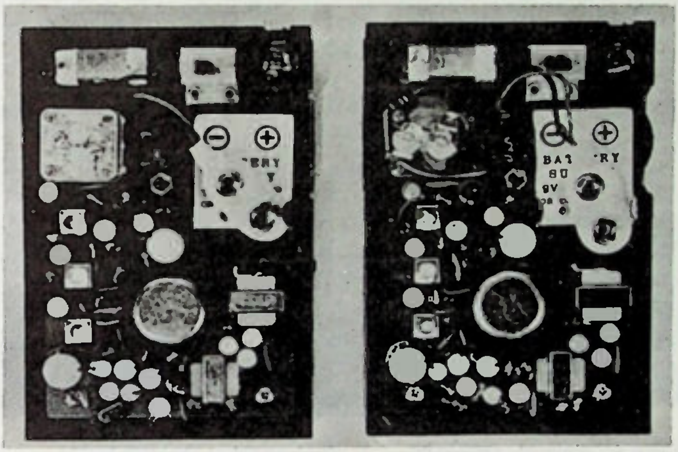


Fig. 14 — The top side of two AM transistor receiver circuit boards is shown. The left one is tuned with a mechanical-tuning capacitor and the one on the right is tuned with a varicap.

spaced across the dial of the converted radio.

We also converted one of a pair of AM portable radios (Fig. 13 and 14) for comparative studies. After the conversion was made, both radios could tune in all local stations, which in this city are spread over nearly the entire broadcast band. The reception sounded as clear on one receiver as on the other.

We found that the capacitors connected in series with the varicaps had to be ceramic types since slight capacitor leakage detuned the radio, and sometimes a pair of stations would be alternately tuned "in" and "out" without moving the dial. It was also necessary to use a separate battery for the tuning bias supply since varying audio signals could load down a common battery and provide distortion by changing the tuning. A voltage-regulator circuit could have been included to eliminate this problem and permit single battery operation.

The converted tuning circuit did require a little additional space, and even more space would have been required if a voltage regulation circuit was included. There was no saving in component cost when constructing the varicap circuit. Then why a varicap tuning circuit?

Varicap tuning circuits permit the construction of all electronic automatic-search-tuning circuits, eliminating the need for complicated mechanical automatic-search-tuning systems that require relays, motors or solenoids. An oscillator step-counter circuit is used in place of the mechanical components. In this circuit, an oscillator produces a signal that slowly changes the varicap bias in steps so small they seem nearly continuous. These steps in

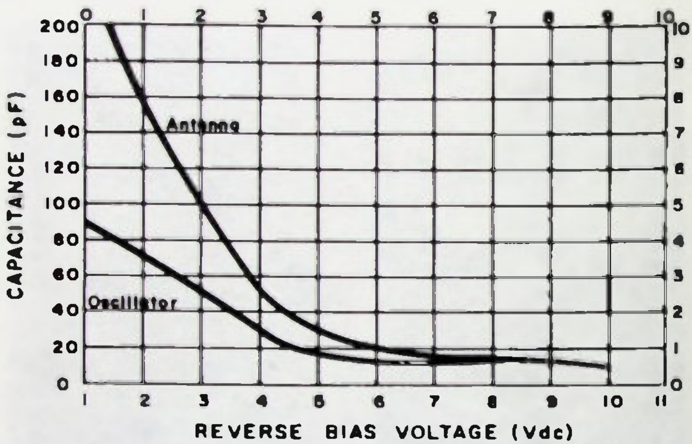


Fig. 15 — The capacitance in the antenna and oscillator circuits vary with the varicap reverse-bias voltage. Courtesy of Motorola.

TABLE 1
Comparative Characteristics of a Mechanical Capacitor-Tuned Radio and a Diode-Tuned Radio Courtesy of Motorola

Measurement Frequency 600kHz	1000kHz	1400kHz	
Signal level (microvolts per meter) required for 6db signal-to-noise ratio			
65μv/M	60μv/M	125μv/M	Mechanical Capacitor
78μv/M	106μv/M	105μv/M	Diode Capacitor
Signal level (microvolts per meter) required for 20db signal-to-noise ratio			
250μv/M	250μv/M	300μv/M	Mechanical Capacitor
240μv/M	250μv/M	205μv/M	Diode Capacitor
Signal level (microvolts per meter) required to obtain 50mw output			
575μv/M	500μv/M	1000μv/M	Mechanical Capacitor
700μv/M	1250μv/M	700μv/M	Diode Capacitor
Image Rejection			
41db	40db	39db	Mechanical Capacitor
40db	29db	34.5db	Diode Capacitor
IF Rejection = 36db			Mechanical Capacitor
IF Rejection = 33db			Diode Capacitor
Tuning Range			
Minimum Frequency	Maximum Frequency		
518.2kHz	1619.3kHz		Mechanical Capacitor
461.4kHz	1614.4kHz		Diode Capacitor
Output level of 5mv/M, 1MHz, RF signal producing 10% distortion			
		0.45mw	Mechanical Capacitor
		0.36mw	Diode Capacitor

bias potential continue to change the varicap capacitance until a station is received and its audio signal blocks the oscillator circuit. The step-count circuit then maintains a constant potential across the varicap, keeping the receiver tuned to the station received. When reactivated, the varicap bias is again changed by the oscillator step-count circuit until another station is received.

Remote tuning is another advantage of varicap tuned circuits. A varicap tuned receiver can be constructed with a single potentiometer to control several RF tuned circuits, eliminating the need for superheterodyne circuits with IF amplifiers. Antenna lead signal loss can be reduced by locating the receiver next to the antenna, on the roof of a building or in the trunk of a car and tuning it remotely with a potentiometer.

CHAPTER 18

Photo-Sensitive and Photo-Emissive Devices

The varicap tuned receiver described in the previous chapter could be designed to tune remotely with a light coupled component designed to isolate the tuned circuit and reduce interference. One integrated circuit currently on the market is isolated with a solid-state optical coupling. The principles of photoemissive and photosensitive semiconductors must be thoroughly understood if their circuits are to be effectively serviced in the future.

Photovoltaic Cells

The photocells that have probably gained the greatest fame are the photovoltaic, or solar cells, that have been used to power small transmitters and other gear in space capsules. Large panels of these cells provide several watts of electrical power from the sun's light. Smaller batteries of these cells are currently being sold to power 9v portable radios with light. Some small relay stations use these cells to recharge their batteries. A photovoltaic cell (Fig. 1), like a diode, contains a junction of two types of material.

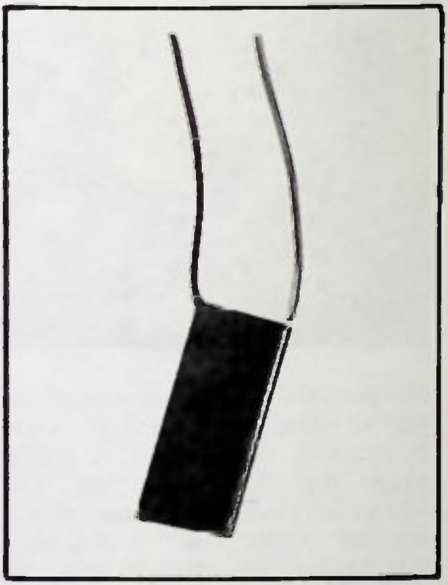


Fig. 1 — A series of these photovoltaic cells can be used to supply the electrical power for a transistor radio.

The photovoltaic cell in Fig. 2 consists of a thin layer of cuprous oxide on a copper plate. When light strikes the cuprous oxide, it drives off electrons which travel to the copper plate. A strip of copper secured to the front surface of the photovoltaic cell acts as an electrode and allows electrons to return to the cuprous-oxide surface. Electrons can be more efficiently returned to the cuprous-oxide surface if an opaque copper screen is substituted for the single copper electrode. The current produced by the light flows in a direction opposite that of the current produced if the component is used as a rectifier in a power supply.

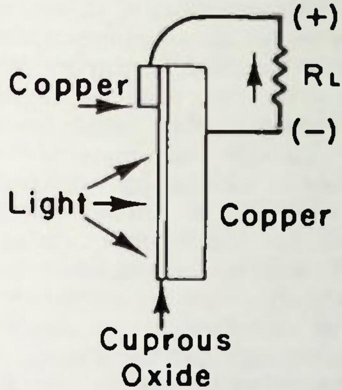


Fig. 2 — The substrate of a back-wall photovoltaic cell collects a negative charge when exposed to light.

Copper can be deposited so thinly on the cuprous oxide surface that it is transparent (Fig. 3) though still a conductor of electricity. More of the electrons driven off the cuprous oxide by light flow through the transparent layer of copper (since the incident light at this junction is more intense) than through the copper backing (where the light intensity has been reduced by the cuprous-oxide material). The transparent copper surface therefore becomes more negative than the copper backing, and the current flows from the front terminal lead to the backing — the opposite direction that it flowed when the copper screen was used.

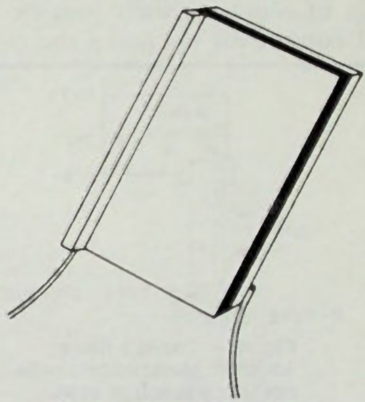


Fig. 3 — The substrate of a front-wall photovoltaic cell collects a positive charge when exposed to light.

A similar photovoltaic cell contains a coating of selenium in place of the cuprous oxide, and a transparent layer of gold or platinum in place of the transparent layer of copper. When light strikes this cell, a current is produced like the one produced by the copper cell.

The most efficient type of photovoltaic cell currently on the market (Fig. 4) contains a junction of P- and N-type silicon. When light strikes this photovoltaic cell, electrons are driven from the P-type material and flow through the junction to the N-type material. Electrons flow through this cell as they do through the ones described in Fig. 2.

The curve in Fig. 5 indicates that the current produced by these photovoltaic cells corresponds more linearly to the amount of incident light when the load resistance is kept to a minimum. Low-impedance circuits are used with these photovoltaic cells in some cameras and light meters. These cells have the advantage of supplying their own electrical current for operating the meter

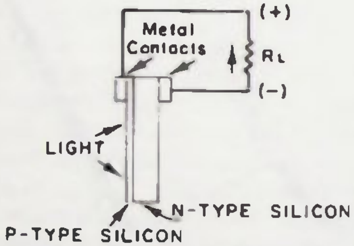


Fig. 4 — Today's more efficient photovoltaic cells contain a junction of P- and N-type silicon.

or controlling the size of the camera iris. Other photosensitive semiconductors must be used, however, when a high sensitivity to light is required.

Photovoltaic cells have a nearly instantaneous response to changes in light intensity, but their output decreases as the frequency of the light fluctuations increase. Their output drops about 10 percent below their peak current at 5kHz and more than 60 percent below their peak at frequencies above 10kHz. Their nonlinear frequency response characteristics make these components undesirable for audio or high-frequency signal applications.

Photoconductive Cells

Photoconductive cells, sometimes called light dependent resistors, generally contain a thick film of photosensitive semiconductor material deposited on an insulating substrate with metallic leads attached to each end. This photosensitive material generally consists of cadmium selenide, cadmium sulfide or lead sulfide, with impurities of silvers, antimony or indium sometimes added to increase their light sensitivity. When light strikes this material, electrons are freed from their bonds and can flow through the material. The amount of incident light determines the number of broken electron bonds and the resulting resistance of the material — the fewer bonds broken, the larger the resistance.

The manufacturer can determine the resistance and the wattage of a photoconductive cell by controlling the dimensions of the photosensitive material. In some photoconductors the electrodes are separated by only a narrow strip of photosensitive material (Fig. 6). For a large varying resistance, the electrodes in some photoconductors are separated by a relatively long strip of photosensitive material — twisted in a serpentine design to conserve space (Fig. 7). In still other photoconductors (Fig. 8), the photosensitive material is twisted between the two electrodes to form the largest possible surface contact between the electrode surfaces and thereby increase the component's current-carrying capacity.

The dark resistance of some cadmium-sulfide cells are 100,000 times greater than their resistance when illuminated. Other photoconductive cells are capable of carrying up to ½a of current. Because of these characteristics, photoconductive cells have the advantage of being able to control relays directly as light intensities change, without the use of amplifiers.

Photoconductive cells are probably the only semiconductor component a technician will encounter that does not contain a junction of P- and N-type material. Because of these junctions, most types of semiconductors must be connected only to dc bias supplies. This component, however, is not subject to such restrictions and will generally function with a lower noise level when connected to an ac bias supply.

Unfortunately, the resistance of photoconductive cells does not decrease or increase as rapidly as the changes in incident light intensity. This limits the frequency response of most photoconductive cells to between 750Hz and 1kHz, even lower than the frequency response of photovoltaic cells.

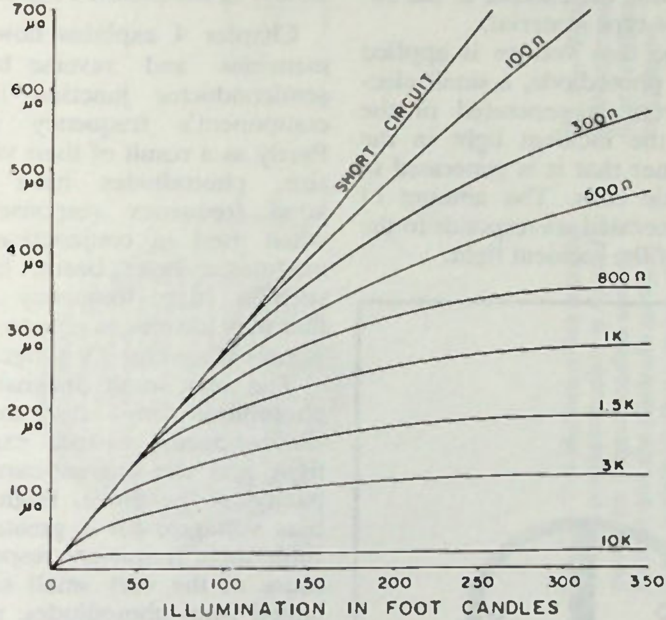


Fig. 5 — There is a more linear relationship between the electrical current from a photovoltaic cell and the incident light intensity when the load resistance is low.

Photodiodes

Photodiodes, like photoconductive cells, contain a junction of P-and N-type material, but they are constructed on a much smaller scale — some being no larger than the head of a pin. The photodiode in Fig. 9 contains a thin deposit of silicon on a metal plate used as an anode. The silicon in contact with the point-contact cathode is chemically treated so that it is of P-type material while the balance of the silicon is of N-type material.

When no bias voltage is applied across the photodiode, a small electrical current is generated in the diode by the incident light in the same manner that it is generated in photovoltaic cells. The amount of current generated corresponds to the intensity of the incident light.

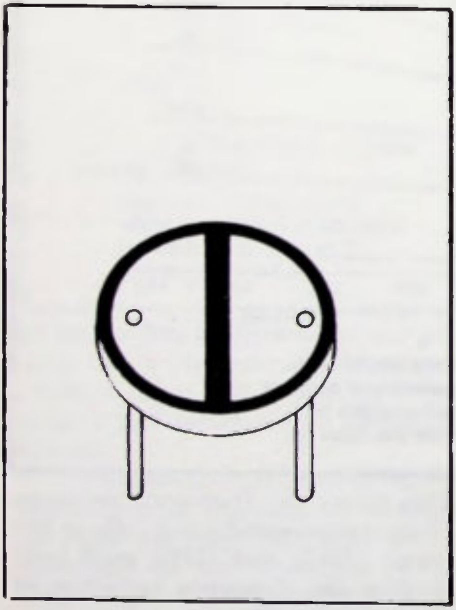


Fig. 6 — The electrons in some photoconductors are separated by only a narrow strip of photosensitive material. (Fig. 3 on page 61 of the July 1967 issue identifies a colored photograph of this component.)

When the diode is reverse biased, an electron barrier is formed across its junction and virtually no current flows through the diode. When light strikes this junction, the electrons in the P-type material pass through the junction into the N-type material and the diode conducts a current that corresponds to the intensity of the incident light.

Chapter 4 explains how the dimensions and reverse bias of a semiconductor junction limit the component's frequency response. Partly as a result of their very small size, photodiodes have a very good frequency response. Some, when used in conjunction with a modulated laser beam, have had such a high frequency response that they have been able to carry the signals of several TV programs.

The very small dimensions of a photodiode limit the amount of semiconductor surface exposed to light and the current-carrying capacity of the diode. High reverse-bias voltages, for a greater signal, limit their frequency response. Because of the very small signal obtained from photodiodes, noise can become a problem.

Phototransistors

Phototransistors amplify the current that results directly from the incident light and therefore have a stronger output signal. With this greater output signal, noise is not as great a factor in these components as in the photodiodes.

The phototransistor in Fig. 10 contains a thin layer of silicon on a metal plate connected to the base lead. The silicon in contact with the

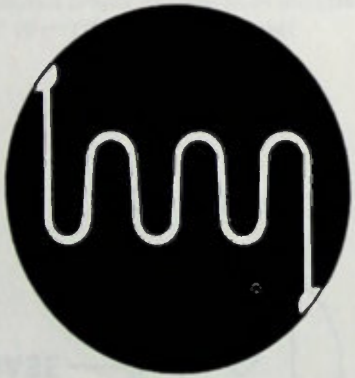


Fig. 7 — The electrodes in some photoconductors are separated by a relatively long strip of photosensitive material. (Fig. 8 on page 61 of the July 1967 issue identifies a colored photograph of this component.)

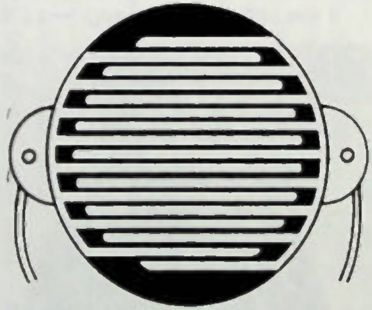


Fig. 8 — The photosensitive material in some photoconductors is shaped to permit the largest possible surface contact with the electrode surfaces. (From a distance this component looks like the one in the colored photograph identified by Fig. 3 on page 61 of the July 1967 issue.)

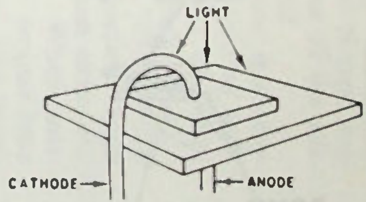


Fig. 9 — When no bias voltage is applied across this diode, incident light develops a positive potential at its cathode.

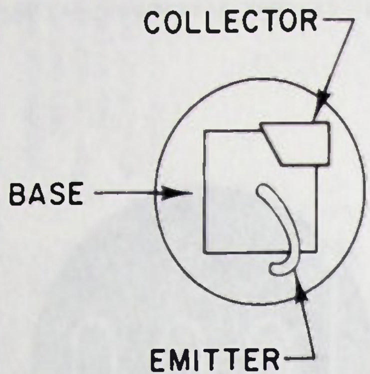


Fig. 10 — Light striking this NPN transistor makes its base more positive than its emitter. (Fig. 9 on page 61 of the July 1967 issue identifies a colored photograph of this component.)

Fig. 11 — This phototransistor has a sufficient signal output to drive a relay without any additional amplification.

Fig. 12 — Light striking this PNP transistor makes its base more negative than its emitter.

point-contact emitter and beneath the collector lead is chemically treated so that it is of N-type material while the balance of the silicon is of P-type material.

As an NPN transistor, this component is normally biased as indicated in Fig. 6 in the first chapter. The transistor's collector lead (Fig. 10) shields the collector-base junction from the light, while light can strike the emitter-base junction. When incident light strikes the emitter-base junction, electrons flow (like in the photodiode) from the P-type material to the N-type material, making the P-type base more positive and increasing the base current. Fluctuations in light intensity change the phototransistor's base current as though a signal had been applied to the base of the transistor for further amplification by the component.

The PNP transistor in Fig. 11 is assembled in a slightly different manner. The central portion of a small cylindrical piece of silicon is secured to a metal ring (Fig. 12), which is connected to the base lead. The top and bottom portions of the silicon cylinder have been changed to P-type material, while the portion within the disc is N-type silicon. The lower end of the cylinder is secured to a metal plate, which is connected to the collector lead, while a point-contact emitter is connected to the cylinder's other end.

As a PNP transistor, this component is normally biased as indicated in Fig. 7 in the first chapter. As you may recall, the polarity of the PNP bias voltages are opposite those for NPN transistors.

When incident light strikes the end of the silicon cylinder, it penetrates the P-type material and electrons flow from the P-type material to the N-type material, making the N-type base more negative and increasing the base current. The resulting fluctuations in base current are amplified by the transistor.

The NPN transistor (Fig. 10) was designed for optimum frequency response, while the PNP transistor (Fig. 12) was designed for optimum power.

CHAPTER 19

FET Light-Sensitive Devices

The previous chapter described the operation of front-wall and back-wall photovoltaic cells (solar cells), photoconductors (photoresistors) and phototransistors. There are still two other photosensitive semiconductors that electronic technicians must prepare to encounter — the photofet and photomos.

Photofet Transistors

The basic structure of a junction field-effect transistor (FET) was illustrated in Fig. 1 of the fifth chapter. Virtually no electrons can flow from the channel (P-type material) to the gate (N-type material) in this transistor when the N-type material is at a more positive potential than the P-type material. Under these conditions, the junction of P- and N-type material acts like the junction of a reverse-biased diode and resists current flow. The greater the reverse bias, the greater the junction's effective size and the smaller the remaining area in the P-type material for source-to-drain channel current. The reverse-biased junction has reduced the source-to-drain current.

The photofet (Fig. 1) operates on a similar principle. A thin slice of P-type silicon material (the substrate that will act as the gate) is secured to a metal surface. A film of N-type material is deposited on the substrate. This film functions as the photofet channel, and source and drain leads are secured to it.

When the substrate of P-type material is made more negative than the film of N-type material, the junction separating these two types of material is reverse biased and resists the flow of electrons. The greater the applied reverse bias, the greater the effective size of the junction, and the thinner the conductive portion of the film between source and drain. The reverse biased junction has reduced the source-to-drain current.

When exposed to light, the photofet's junction of P- and N-type material functions like the photodiode junction described with Fig. 9 in the previous chapter. When light strikes the photofet junction, electrons in the P-type material pass through the reverse biased junction into the N-type material. Light has the effect of reducing the reverse bias junction resistance, increasing the source-to-drain current.

A basic photofet circuit is shown in Fig. 2. The gate biasing voltage (V_{GG}) is adjusted so that the drain-to-source current (I_D) is reduced to nearly zero when the photofet is not exposed to light.

When exposed to light, the photofet's gate current increases in proportion to the intensity of the incident light. This additional current is labeled λi_G. The resulting change in the voltage drop across the gate resistor (R_G) can be determined by the equation $\lambda v_G = \lambda i_G R_G$. Since the potential of the voltage source (V_{GG}) remains virtually constant, the change in the gate-to-source voltage is equal to the change in the voltage drop across the gate resistor.

The greater the intensity of the incident light, the greater the resulting voltage drop across the gate resistor, the smaller the gate-to-source voltage (gate reverse bias) and the greater the resulting drain-to-source current. The smaller the intensity of the incident light, the greater the resulting drain-to-source current (I_D). Changes in the intensity of the incident light result in corresponding changes in the gate-to-source current, the drain-to-source current and the resulting voltage drop across the load resistor (R_L).

Neglecting limitations resulting from noise factors, the greater the value of a photofet's gate resistors (R_G), the greater the photofet's sensitivity. In chapter 5 it was explained that FETs are high-impedance devices because of their very small drain current. Measurements indicate that for one photofet the drain current (I_D) varied from 20na (20×10^{-9}a) per foot candle when this resistance (R_G) is zero to 20ma (20×10^{-3}a) per foot candle when this resistance (R_G) is 500M. Increasing the gate resistance 500M increased the sensitivity 10^6 times.

Photofets have a greater light sensitivity than any

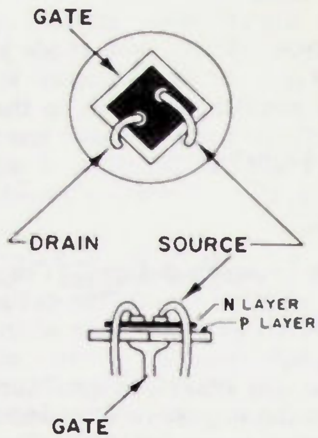


Fig. 1 — Light striking the photofet's epitaxial junction reduces the semiconductor's source-to-drain resistance.

of the photosensitive semiconductors described in the last chapter. Their frequency response, although high, is not quite as high as some photodiodes. Their noise level, however, is generally better than that of a diode and transistor or diode and FET combination. They reportedly are at least ten times as sensitive as a phototransistor with four times the gain bandwidth product.

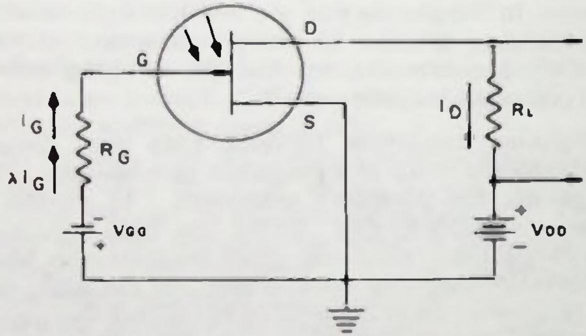


Fig. 2 — The photofet gate is reverse biased for minimum drain-to-source current when not exposed to light.

Photomos Transistors

Photomos transistors (photosensitive MOS transistors) are still in the experimental stage. At last report, experimental models failed to have a desirable low-noise factor or sufficient stability. Future photomos transistors will probably resemble the MOS transistors illustrated in Figs. 1 and 5, chapter 6. The insulation and gate electrode could be made transparent to permit the impact of incident light on the semiconductor material beneath their surface. This would change the effective channel thickness and resulting resistance between source and drain.

Light-Emitting Diodes

There are two basic methods of generating visible light: incandescence and luminescence. Incandescent light is generally produced by electrically heating tungsten lamp filaments or other material to a high temperature. The higher the temperature, the greater the light output.

Luminescence is light radiation that results from exciting atoms with an external source of energy. When atoms in a gas are excited by an electrical discharge, the gas appears to glow (luminesce). The fluorescent coatings on the face of a CRT glow as a result of electrons striking it. Atoms in fluorescent coatings, used in fluorescent lamps, are excited when absorbing ultraviolet light produced by an electrical discharge through mercury vapor. Most of the energy the excited coating releases when returning to normal is in the form of visible light.

At normal temperatures all atoms contain electrons. When energy is applied to an atom from an outside source (such as electrons bombarding the fluorescent surface of a CRT), some of these electrons absorb the energy and are excited into a "higher energy level." As the electrons return to normal, they release the energy that they have absorbed. This energy may be released in the form of ultraviolet light, visible light, heat (infrared), microwaves, lower-frequency forms of radiation or a combination of these forms. The only requirement is that the energy released must equal the energy absorbed. The higher the frequency emitted, the greater the energy released. The stronger the signal emitted, the greater the number of atoms releasing energy.

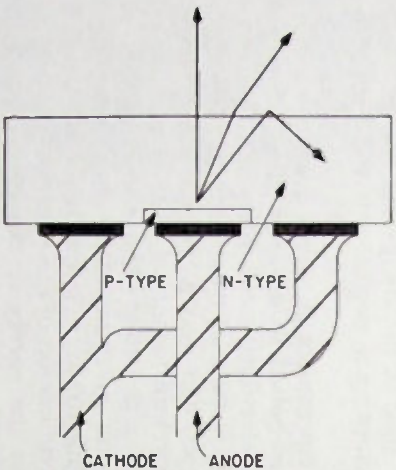


Fig. 3 — By forming both P- and N-type material at the bottom surface, electrodes can be attached without obscuring the light produced. Light striking the top semiconductor surface at an angle that is greater than 16deg from the normal is internally reflected and lost.

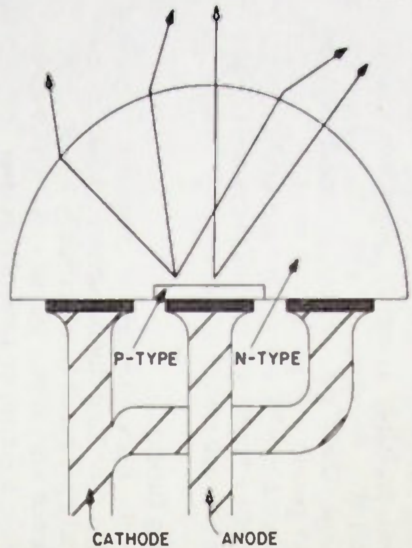


Fig. 4 — The light-emitting diode becomes 10 times as efficient when the top surface is shaped into a hemisphere to reduce internal reflections at the top surface.

The form of energy released by an excited atom (ultraviolet light, visible light, etc.) depends on the type of atom excited (boron, gallium, indium), its temperature and the way electrons in one atom are bonded to those in another atom. Tests have shown that the electron stresses in some diode junctions are such that the diode will emit visible or infrared light when their atoms are exposed to an electrical current. (In this manner some diodes also emit microwaves.)

In some of the more efficient diodes the radiant energy produced is greater than the applied electrical energy. The atoms absorb both thermal and electrical energy as they are excited, emitting radiation when returning to their more stable energy level. Bismuth-telluride (BiTe) semiconductors are currently on the market for refrigeration purposes. At 80°F the junctions in one battery of 32 thermocouple diodes are reportedly cooled to 25°F by 12.5a of current at 3.4v applied potential. Specifications for that cooling system indicate that it can remove 35Btu of heat each hour.

Most light-emitting diodes now on the market are made of gallium-arsenide (GaAs) semiconductors. The semiconductor chip in one such diode measures 50mil by 50mil by 3mil (1 mil = 10^{-3}in.) and consists of N-type material, chemically treated so that the bottom surface is changed to P-type material. A metal electrode is connected to the bottom P-type material and strands of wire are connected to the top surface. When the diode is forward biased, light must pass around the strands of wire at the top surface.

A more convenient light-emitting diode design is shown in Fig. 3. There only the central portion of the bottom surface is changed to P-type material. Both P- and N-type material are present in the bottom surface where the electrodes can be attached without obscuring the light produced.

A large portion of the light generated at the diode junction is lost by internal reflection (Fig. 3). Gallium arsenide has an index of refraction 3.6 times that of air. This means that when light strikes the top surface of the semiconductor material at an angle that is greater than 16deg from the normal (the normal is an imaginary line perpendicular to the crystal's surface), the light is internally reflected. The reflected light is absorbed by the crystal and lost.

By shaping the top surface into a hemisphere, the amount of reflected light is reduced (Fig. 4) and the diode becomes 10 times as efficient as before with the flat top surface (Fig. 3). The standard diameter for these hemispherical crystals ranges between 36 and 72-mil with junctions between 10 and 20mil. Although the cathode is shown as two connected leads, the semiconductor's hemispherical shape permits the use of a metal

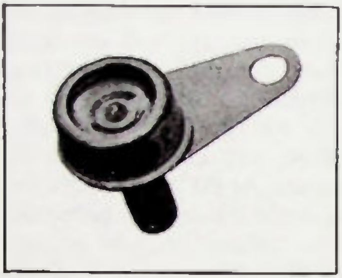


Fig. 5 — This light-emitting diode contains a hemispherically shaped gallium-arsenide semiconductor. Courtesy of Texas Instruments.

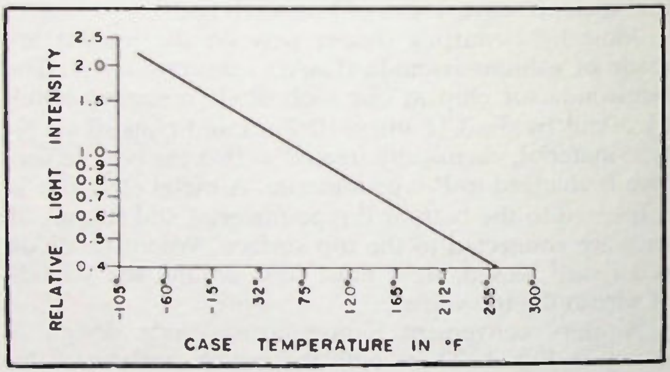


Fig. 6 — A typical relationship between the case temperature and relative light intensity in a light-emitting diode with constant-current forward bias.

ring around the hemisphere's perimeter as an electrode connected to a single cathode lead. A light-emitting diode made with a hemispherical semiconductor is shown in Fig. 5.

Despite the fact that some diodes consume heat when radiating energy, the presence of heat can temporarily or even permanently reduce the number of atoms having the required electron stresses for producing light. Heat can reduce the efficiency of a light-emitting diode. The graph in Fig. 6 shows the relationship between the case temperature of a typical light-emitting diode (like the one shown in Fig. 5) and the relative intensity of the light it produces, when the bias current is kept constant. This graph indicates that when the diode is operating at 120°F, the light it produces is only 80 percent as intense as the light it produces at 75°F. If it were practical to cool the diode to about −70°F, the light produced would be twice as intense as that produced at 75°F.

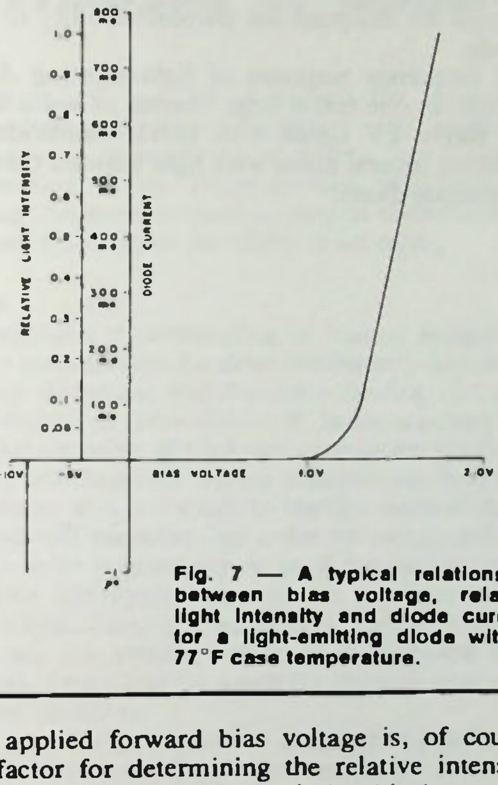


Fig. 7 — A typical relationship between bias voltage, relative light intensity and diode current for a light-emitting diode with a 77°F case temperature.

The applied forward bias voltage is, of course, a major factor for determining the relative intensity of the light produced. A typical relationship between bias voltage, relative light intensity and diode current at 77°F is indicated by the graph in Fig. 7. This graph indicates that when the forward bias is 1.7v, the light-emitting diode conducts 760ma. When the forward bias is 1.5v, the diode conducts 400ma and the light produced is half as intense as that produced with a 1.7v bias. When the forward bias is 1.4v, the diode conducts

220ma and the light produced has only a fourth of the intensity of that produced with a 1.7v bias.

Readings from this curve (Fig. 7) indicate a relatively large current change for a small voltage change. In many cases a 0.25v increase in bias potential, above some low-intensity light-producing voltage, will result in a current increase exceeding the diode's maximum current rating. A slight change in diode junction resistance, as a result of temperature changes, may also result in a damaging current even if the bias voltage did not increase. For this reason, light-emitting diode circuits must be designed for current stability to protect the diode.

The frequency response of light-emitting diodes is very high. In one test a large number of voice channels and a single TV signal with 1MHz bandwidth, were transmitted several miles with light emitted from a gallium arsenide diode.

CHAPTER 20

Principles of Optics

Chapters 18 and 19 described the operation of photo voltaic cells, photoconductive cells, photodiodes, photosensitive transistors, photofets, photomos and light-emitting diodes. These devices, plus those in the following chapters, require a general understanding of lenses and fiber optics for effective servicing.

Lenses

An in-depth understanding of lenses, including the complex mathematics for determining lens curvature for minimum distortion and the compounding of lenses to reduce shape and color distortion, is not required for the successful servicing of photosemiconductor applications normally encountered. Unless a technician is repairing TV cameras, it is fairly safe to say that none of the lenses that he will encounter are color corrected (achromatic) since there is generally no need for having the photosensitive semiconductors respond to more than one color of light. Generally, an understanding of lens focal lengths and the images formed by lens optical systems is all that is required for servicing them in new optical-electronic products.

The image (Fig. 1) used to demonstrate the function of lenses on an optical bench is merely a black silhouette of a leaf on a clear cellulose film, mounted in a supporting rectangle. When light from some relatively distant source passes around this silhouette, a fuzzy shadow of the leaf (Fig. 2) is formed on the screen without the aid of a lens.

An even less distinct image of the leaf is formed (Fig. 3) when only a lens is placed near the screen. The pattern then seen is a white ring surrounding a white

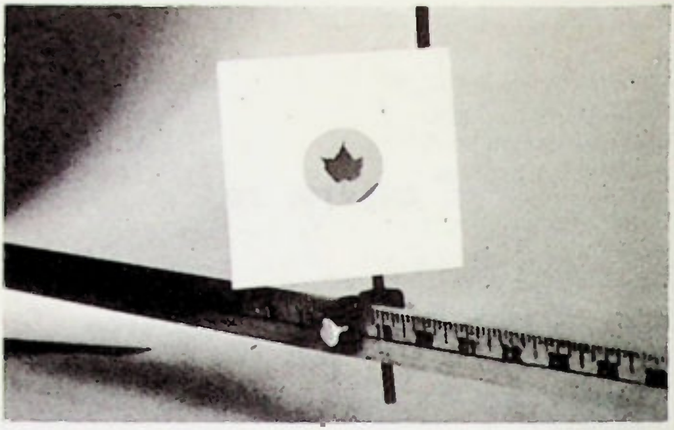


Fig. 1 — A black silhouette of a leaf on a clear cellulose film is the image used to demonstrate the function of lenses on an optical bench.

disc. The outer edge of the ring is the same as the outer edge of the projected disc shown in Fig. 2, while the inner edge of the ring is the shadow cast by the lens. (The shadow of the lens holder can also be seen on the screen.) The leaf image formed in the central disc is not in sharp enough focus to be seen in Fig. 3. This demonstrates a problem common to many lenses. There is no sharp focal point for the entire lens. The outer edge of the lens tends to focus the image differently than the inner portion of the lens, and as a result, when light passes through the entire lens, it fails to focus sharply enough to be seen in the picture.

Fig. 2 — A distant light can be used to cast the shadow of the leaf on a screen without using a lens.

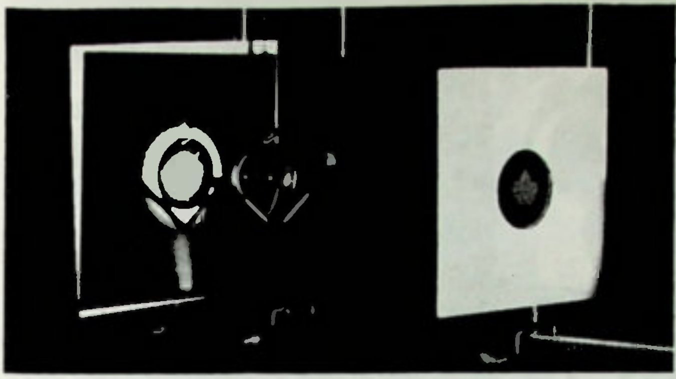


Fig. 3 — An unshielded lens usually fails to project a satisfactory image.

Without moving the lens, screen or mounted leaf pattern, the problems shown in Fig. 3 can be corrected (Fig. 4) by a mask which permits light to pass only through the central portion of the lens before striking the screen. The image of the leaf now appears in sharp focus on the screen.

Since the lense shown in Fig. 4 is nearer the screen than the cellulose film, the image on the screen is smaller than the design on the film. If a photosensitive semiconductor is substituted for the screen, the entire image of the leaf could be projected on the semiconductor despite the mounted leaf's design being considerably larger. This semiconductor is not capable of identifying the shape of the leaf's stationary image since it contains

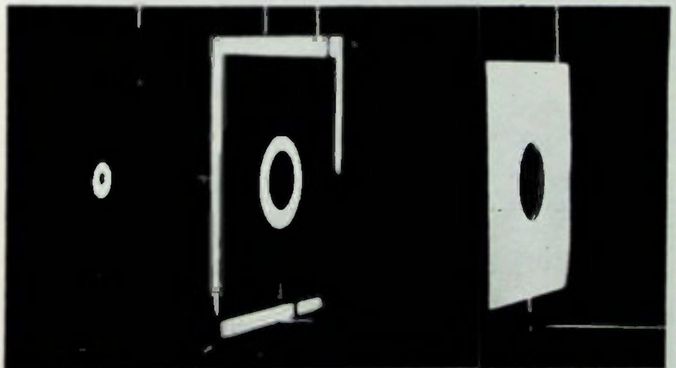


Fig. 4 — When the lens is nearer the focused image than the film, the image on the screen is smaller than the design on the film.

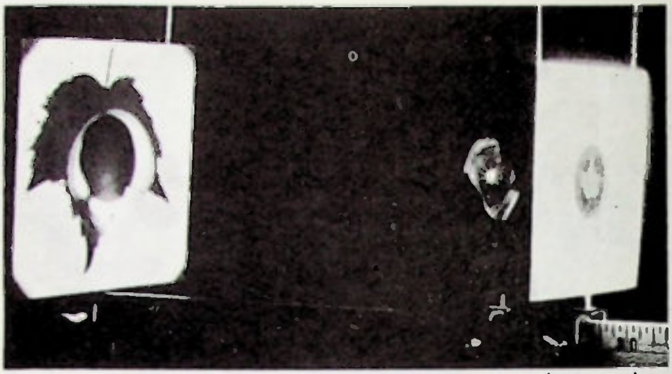


Fig. 5 — A lens usually fails to form a satisfactory image when there is no shield to block light passing around the sides of the lens or through its outer circumference.

merely a single photosensitive surface which indicates only the average amount of light exposed to it. It can, however, be used to determine whether or not the entire image of the leaf design is projected on its photosensitive surface, since this image is black and reduces the total amount of light exposed to the semiconductor.

By placing the lens nearer the cellulose film than the screen (Fig. 5), the projected image becomes larger than before. The image shown (Fig. 5) is not satisfactory, however, since we again (as in Fig. 3) failed to use a mask. The white circle in the central portion of the projected image is formed by light passing around the edge of the lens. The shadow of the lens can be seen within that circle.

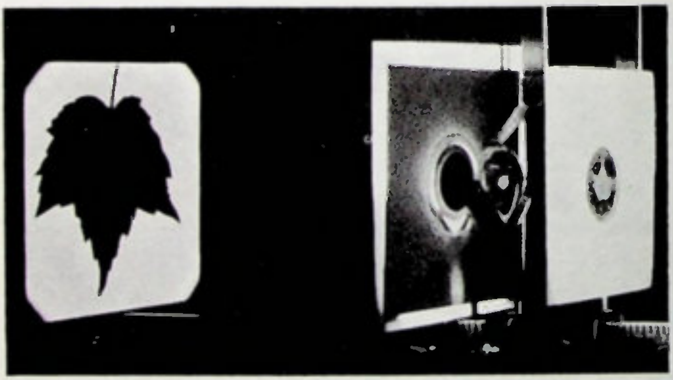


Fig. 6 — When the lens is nearer the film than the focused image, the image on the screen is larger than the design on the film.

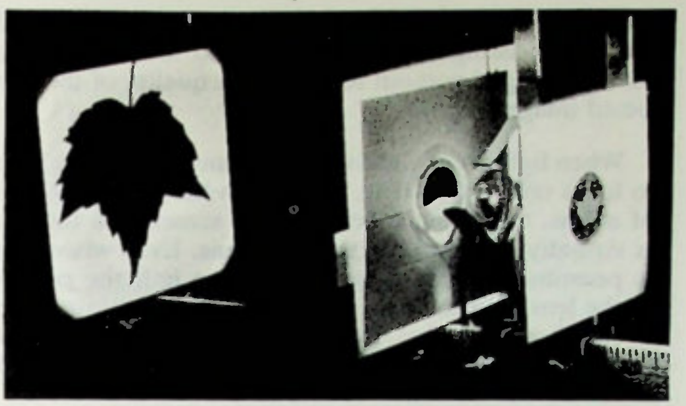


Fig. 7 — Nearly half the light normally passing through a lens can be blocked without affecting the quality of the image formed.

This problem can be corrected by again using a mask (Fig. 6) which blocks the light passing around the lens or through its outer circumference. Only light passing through the central portion of the lens is used to form the image seen. The projected image is now complete (Fig. 6) and in better focus than before (Fig. 5).

Since the lens in Fig. 6 is nearer the cellulose film than the screen, the image on the screen is larger than the design on the film. In some rare instance, a photosensitive semiconductor might be secured to the upper portion of the screen and used in an electronic circuit for counting the stems of passing leaves — the image has been magnified so that the photocell can detect it.

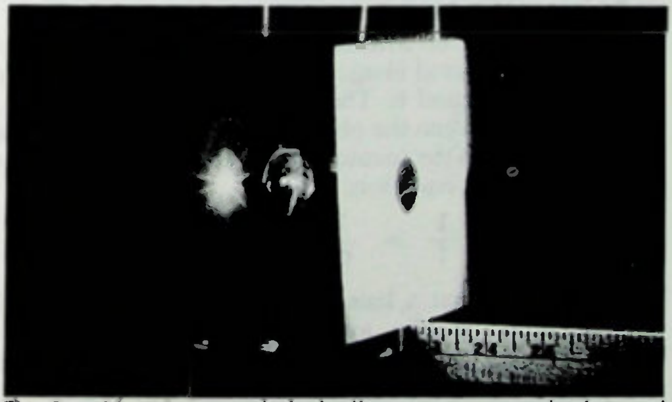


Fig. 8 — Light from a relatively distant source can be focused on a screen at virtually the focal length of the lens.

It is interesting to note that we can block nearly half the light passing through the lens to form the image shown in Fig. 6 without reducing the quality of the projected image (Fig. 7).

When light from a relatively distant source is brought to focus on a screen (Fig. 8) through the central portion of a lens, the distance between the screen and the lens is virtually the focal length of the lens. Even when light is permitted to pass through less than half the portion of the lens used to focus it in Fig. 8, the remaining light is still focused as before — the image formed (Fig. 9) is merely a little less intense. This fact is also important.

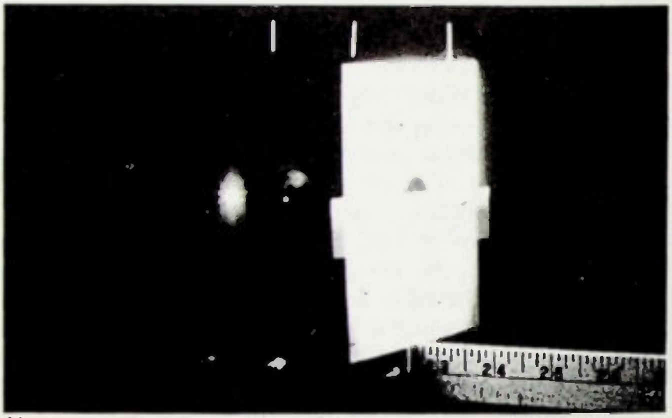


Fig. 9 — The shape of an image formed at the lens' focal length is not affected by reducing to less than half the portion of the lens used to focus it.

The lens must be a certain distance between the object and its projected image (Fig. 10) to form the images shown in Fig. 4 and 6. The focal length of the lens (f), the distance between the object and the lens (d_1) and the distance between the focused image and the lens (d_2) can be applied to the equation:

$$\frac{1}{f} = \frac{1}{d_1} + \frac{1}{d_2}.$$

If we know that a lens has a 5-in. focal length and that the object is 20in. away, we can calculate the distance from the lens where the image can be focused on the screen.

$$\frac{1}{5\text{in.}} = \frac{1}{20\text{in.}} + \frac{1}{d_2}.$$

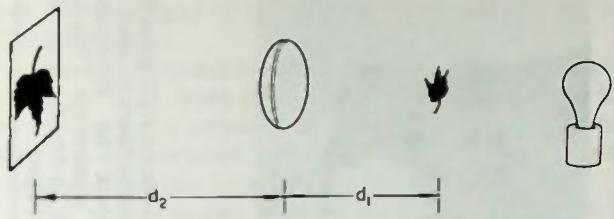


Fig. 10 — There is a definite relationship of distances in a single lens system between the object and the lens (d_1), and between the lens and the projected image (d_2).

$$\frac{1}{d_2} = \frac{1}{5\text{in.}} - \frac{1}{20\text{in.}} = 0.20/\text{in.} - 0.05/\text{in.}$$

$$= 0.15/\text{in.} \qquad d_2 = \frac{1}{0.15/\text{in.}} = 6\tfrac{2}{3}\text{in.}$$

These calculations indicate that when an object is 20in. from the 5-in. focal length lens, its image is formed 6⅔ in. from the other side of the lens.

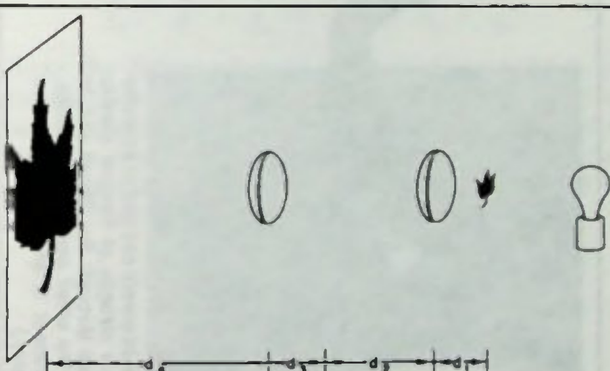


Fig. 11 — A definite relationship exists between distances in a double lens system, between the object and the first lens (d_1) and between the second lens and the projected image (d_4).

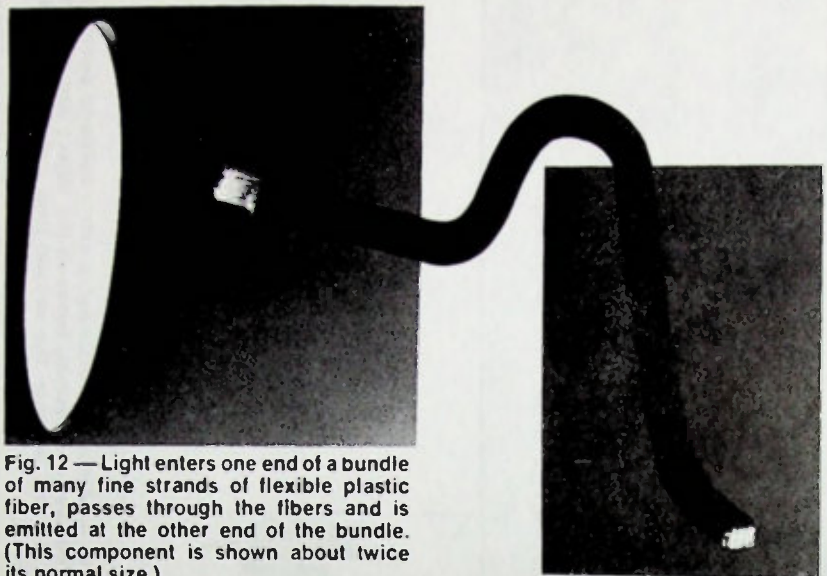


Fig. 12 — Light enters one end of a bundle of many fine strands of flexible plastic fiber, passes through the fibers and is emitted at the other end of the bundle. (This component is shown about twice its normal size.)

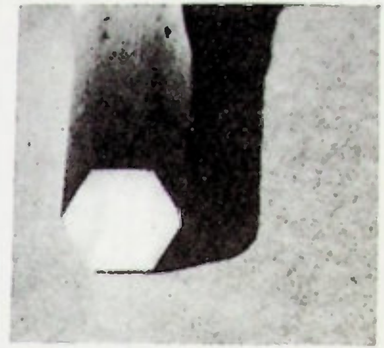


Fig. 13 — Some rigid glass fiber optics contain fibers so small that they cannot be distinguished with the unaided eye. (This component is shown about six times its normal size.)

We know from Fig. 4 that since the image in our example is nearer the lens than the object is, the image is smaller than the object. If the image and object distances are interchanged (Fig. 6), the image would be larger than the object.

The same equation can be used to show that when the object is 10in. from the 5-in. focal length lens, its image is formed on a screen 10in. from the other side of the lens. In this instance the image and object are the same distance from the lens and are the same size.

The equation can also be used to show that a 5-in. focal length lens will focus the light from a source 1000 in. away onto a screen 5.025in. from the lens. As indicated for Fig. 8, the lens is focusing light at virtually its focal point. If the light had been from an even more distant source, its image would have been focused even nearer the lens' focal length.

There are occasions when a greater magnification or reduction in image size requires more than a single lens. When two lenses are used (Fig. 11), d_1 represents the distance of the object from the first lens, d_2 represents the distance of the imaginary image (the image that the first lens could have formed had there been a screen) from the first lens, d_3 represents the distance of that imaginary image from the second lens and d_4 represents the distance of the final focused image from the second lens. These distances can be applied to the equations:

$$\frac{1}{f_1} = \frac{1}{d_1} + \frac{1}{d_2} \text{ and } \frac{1}{f_2} = \frac{1}{d_3} + \frac{1}{d_4}.$$

Since the distance between the two lenses (d_L) in this optical system must equal the distance of the imaginary image from the first lens (d_2) plus the distance of the imaginary image from the second lens (d_3), ($d_L = d_2 + d_3$), we can combine the two preceding equations and arrange them in a more convenient form.

$$\frac{1}{f_1} + \frac{1}{f_2} = \frac{1}{d_1} + \frac{1}{d_2} + \frac{1}{d_3} + \frac{1}{d_4}$$

$$= \frac{1}{d_1} + \frac{1}{d_L} + \frac{1}{d_4}.$$

$$\frac{1}{d_4} = \frac{1}{f_1} + \frac{1}{f_2} - \frac{1}{d_L} - \frac{1}{d_1}.$$

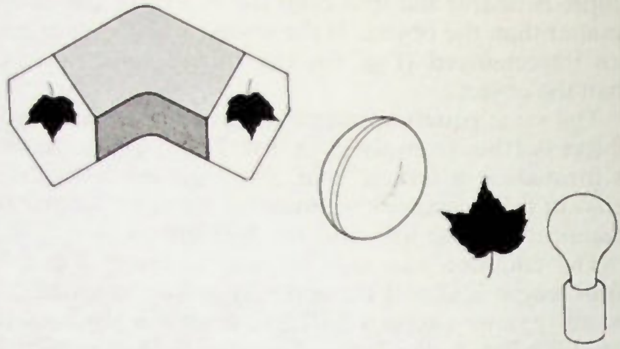


Fig. 14 — Each segment of an image focused on one end of this fiber rod is carried independently by one of thousands of fine glass fibers.

In a double lens system the focal lengths of the lenses used (f_1 and f_2) remain constant while the object and focused image distances (d_1 and d_4) can vary and are mutually dependent — like the object and focused image distances (d_1 and d_2) in a single lens system. Since a change in the distance between the two lenses (d_1) can vary the relationship between the object and focused

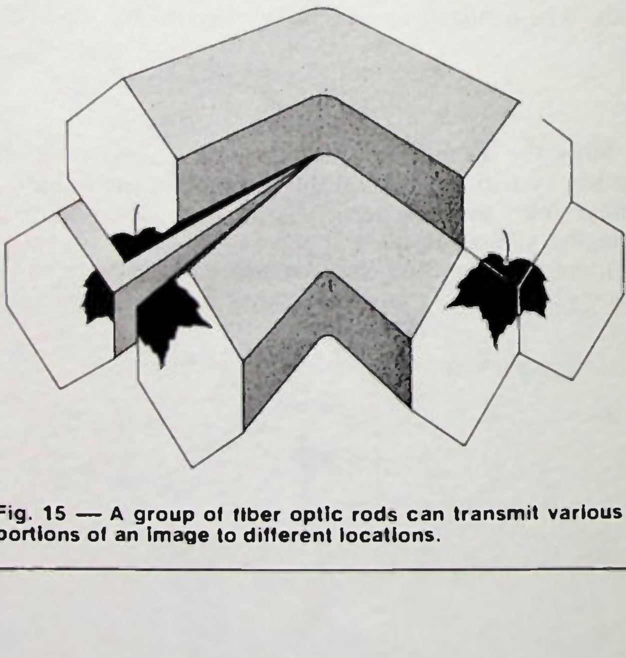


Fig. 15 — A group of fiber optic rods can transmit various portions of an image to different locations.

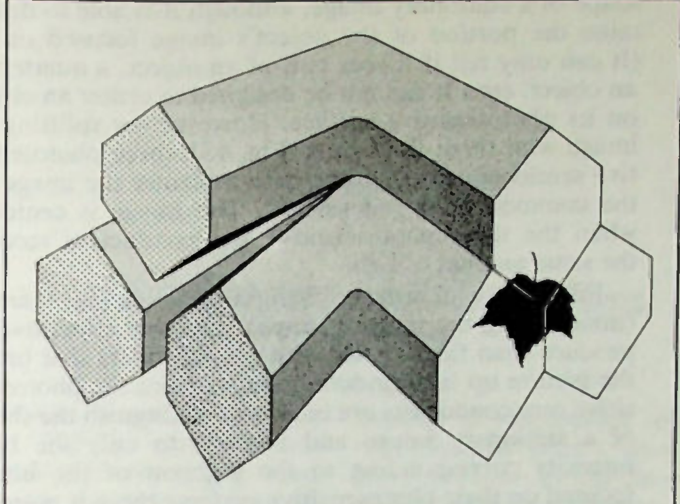


Fig. 16 — Less expensive fiber optics do not transmit the shape of an image but merely the light that forms the image.

image distances (d_1 and d_4), changes in this distance (d_L) have the effect of changing the total effective focal length (f_T) of the double lens system

$$\left(\frac{1}{f_T} = \frac{1}{f_1} + \frac{1}{f_2} - \frac{1}{d_L}\right).$$

Fiber Optics

Light can also be "piped" to perform many desired functions. When light is applied to one end of a bundle of glass or plastic fibers (Fig. 12), it travels through the fibers and appears at the other end of the bundle.

There are many types of fiber optics on the market, ranging from the flexible. relatively coarse plastic fibers shown in Fig. 12 to rigid glass fibers that are so fine that one fiber can not be distinguished from another with the unaided eye (Fig. 13).

When an image is focused on one end of a fiber optic rod, like the one shown in Fig. 13, each fiber independently transmits a minute segment of the image from one end of the rod to the other (Fig. 14). In some rods the grain is so fine that an unaided eye can not detect it in the transmitted image.

Earlier in this chapter it was indicated that a single photosensitive semiconductor is unable to identify the shape of a stationary image, although it is able to determine the portion of the object's image focused on it. (It can only tell if it sees half of an object, a quarter of an object, etc.) It can not be designed to center an object on its photosensitive surface. However, by splitting an image with three fiber rods (Fig. 15), three photosensitive semiconductors can be used to center the image on the common fiber rod surface. The image is centered when the three photosensitive semiconductors receive the same amount of light.

Fiber optics that transmit light in a coherent manner (maintaining the image's shape) are more expensive to produce than fiber optics with twisted fibers that break the picture up in a random manner. Since the photosensitive semiconductors are unable to distinguish the shape of a stationary image and respond to only the light intensity corresponding to the segment of the image focused on their photosensitive surface, there is no need to use more expensive coherent fiber optics in the products that we plan to describe.

Three photosensitive semiconductors connected to the fiber rods shown in Fig. 16 can be used to center the image on the common fiber rod surface as effectively as three photosensitive semiconductors connected to the fiber rods shown in Fig. 15.

CHAPTER 21

Optic-Electronic Couplers

The devices described in the last three chapters can be combined to perform interesting functions in new electronic products that are already on the market or soon will be on the market.

Incandescent Couplers

Incandescent lamps and photoconductive cells have been combined into a single optical-electronic insulator component (Fig. 1). These couplers prevent the transmission of virtually all signals from their output to their input, eliminating undesirable signal radiation or interference at the input. Since there is no electrical connection between input and output, there is generally no need for concern with differences in potential between the two.

A voltage applied across the incandescent lamp (Fig. 2) causes the lamp filament to glow, and the light emitted reduces the resistance of the photoconductive cell. Within the limits of the lamp's rated capacity, the greater the applied voltage, the smaller the output resistance.

An incandescent lamp cannot be heated to its full brightness the instant the maximum voltage is applied, nor can its brilliance diminish as rapidly as the applied voltage. The incandescent lamp's slow response time restricts its use as a variable light source to the lower audio frequencies (generally below 200Hz). This is lower than the frequency response of the photoconductive cells, and therefore generally no reason exists for producing incandescent couplers with photosensitive semiconductors designed for a higher frequency response.

Studies indicate that incandescent lamps can produce noise at higher frequencies than their frequency response. When exposing a photofet to an incandescent lamp used in a flashlight, we found that by tapping on the flashlight a microphonic audio signal was produced at the lamp filament's mechanical resonant frequency. This problem can be eliminated by diffusing the light (breaking up the image of the filament) before it strikes the photocell.

The incandescent coupler can be used in place of a potentiometer (Fig. 3) in such applications as a remote receiver's volume and tuning controls. (Volume controls are shown in almost all radio receiver schematics, while a radio tuning potentiometer is shown in Fig. 12, chapter 17. Potentiometers (left portion of Fig. 3) generally function as a voltage divider and their total internal resistance remains relatively constant. By moving the potentiometer tap across the resistive material, the total resistance is divided into two parts, and it is the ratio of the tap-

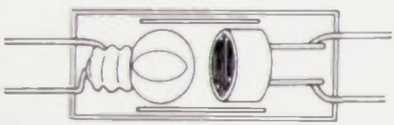


Fig. 1—Some of the optical-electronic insulators now on the market contain an incandescent lamp and photoconductive cell combined as a single component.

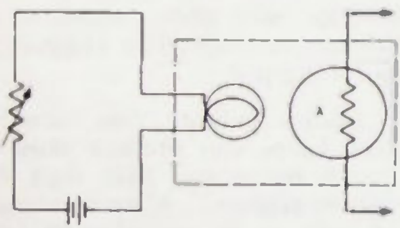


Fig. 2—Reducing the resistance between the battery and incandescent lamp increases the brightness of the lamp and reduces the resistance of the photoconductive cell.

ped resistance to the total resistance that determines the portion of the applied voltage present at the tap. This is quite similar to the function of a voltage divider containing an incandescent coupler and a resistor (right portion of Fig. 3). Although the total resistance in this circuit ($R_1 + R_2$) does not remain constant (the value of R_2 changes with light intensity), it is the ratio of one of these resistances to the total resistance (like the ratio of resistances in a potentiometer) that determines the portion of the applied voltage present at the common lead. A remote potentiometer can alter the voltage applied to the incandescent lamp, which in turn alters the voltage drop within the voltage divider circuit.

A stereo balance indicator (Fig. 4) can be designed by combining two

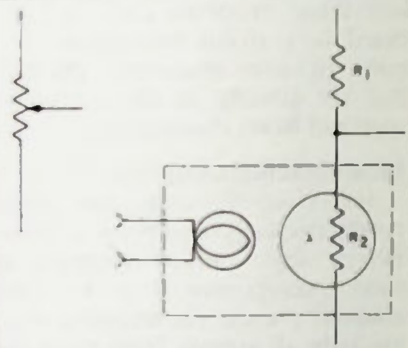


Fig. 3—An incandescent coupler and resistor can be made to function somewhat like a potentiometer voltage divider.

incandescent couplers in a Wheatstone bridge circuit. Each incandescent lamp in the indicator is connected with a capacitor to the output of one of the amplifiers in the stereo system. The amplifier's audio output signal lights the corresponding incandescent lamp, and the stronger the signal, the brighter the lamp and the smaller the photoconductive resistance.

When the same signal is applied to both inputs of the stereo amplifier (a monophonic record, tape or radio program), the output of both channels and the resulting photoconductive resistances will be equal when the stereo amplifier is balanced. Under these conditions the Wheatstone bridge circuit is balanced and there is no deflection of the balance indicator meter.

When the stereo system amplifies the same signals and the amplifiers are not balanced, the output of both channels and the resulting photoconductive resistances will not be equal. Under these conditions the Wheatstone bridge is out of balance and there is a corresponding deflection of the balance indicator meter, showing which amplifier has the greater gain. The meter deflection will be reduced to zero as the amplifier gains are adjusted so that their outputs are equal and both incandescent lamps are of equal brightness.

Luminescent Coupler

The frequency response of optical-electronic insulators can be improved by using a light-emitting diode instead of an incandescent lamp. One luminescent coupler component currently on the market reportedly has a frequency response exceeding 60kHz and as much as 5kv input-to-output insulation.

A small (gallium-arsenide light-emitting diode/silicon light-sensitive diode) optical-electronic insulator has been designed to function as part of an integrated circuit (Fig. 5) and transmit ac or dc signals while providing ±100v input-to-output insulation. When functioning as a broadband pulse transformer, the optically coupled integrated circuit reportedly has a frequency response extending from dc to 100kHz.

Luminescent couplers can eliminate a problem frequently encountered with present electronic scope switches. The two scope switch inputs do not have separate floating grounds; and ground leads basically must be connected to the same portion of the circuit being tested. With such scope switches it is not yet possible to compare signals across two resistors biased at different potentials above ground. By using luminescent couplers however, an electronic scope switch can now be designed having separate floating grounds. The input leads are connected to ungrounded conventional voltage-divider circuits and the signal is amplified by a battery-powered field-effect transistor (FET) before passing through a luminescent coupler to the conventional, common-ground, scope switch amplifiers. (Batteries or a power supply driven by an RF transformer is required for this application since using a conventional ac power supply would tend to reduce the desired high impedance between the two floating grounds by capacitively coupling them to the common ground.)

Contactless Meter Relays

Optical meter relays have been on the market for only a relatively short time and may still be merely a curiosity for some technicians. Any meter reading on an optical meter relay can be used to switch circuits off or on. Their possible applications are numerous: If used in a VOM or VTVM, it can automatically switch off the circuit whenever the meter needle is deflected beyond

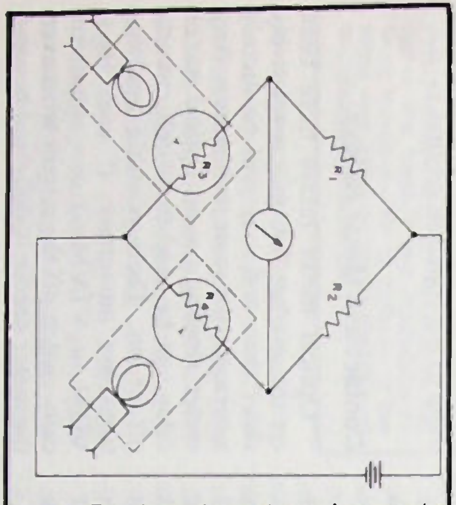


Fig. 4—Two incandescent couplers can be used in a Wheatstone bridge circuit to operate a stereo balance indicator.

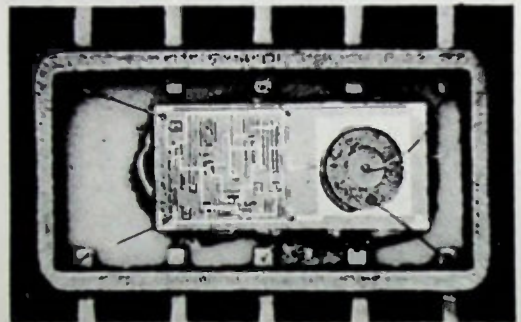


Fig. 5—A luminescent coupler (left) can be made small enough to be considered part of an integrated circuit (right). Courtesy of Texas Instruments Inc.

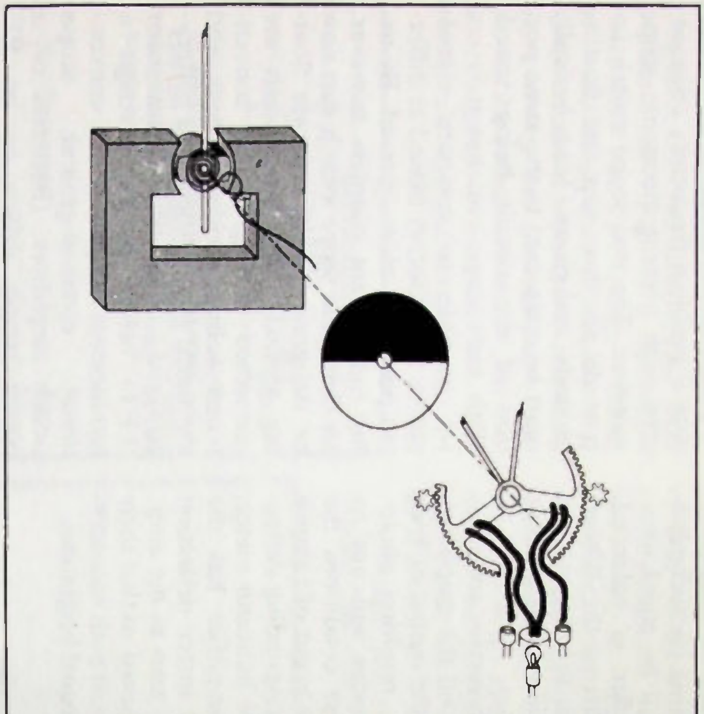


Fig. 6—Contactless meter relays have been designed to switch circuits on and off without making electrical contact with the meter needle.

a predetermined point. When connected to a power supply, the meter needle may be used to disconnect circuits when the needle falls below a certain point and reconnect those circuits when the needle rises above a certain point.

The basic meter movement in a contactless meter relay resembles case behind this disk. A pair of flexible fiber optics are secured to each of two supports, which are rotated by pinions secured to knob shafts. By turning the knobs, the pinions move the fiber-optics supports, changing their relative position behind the black-and-white disk.

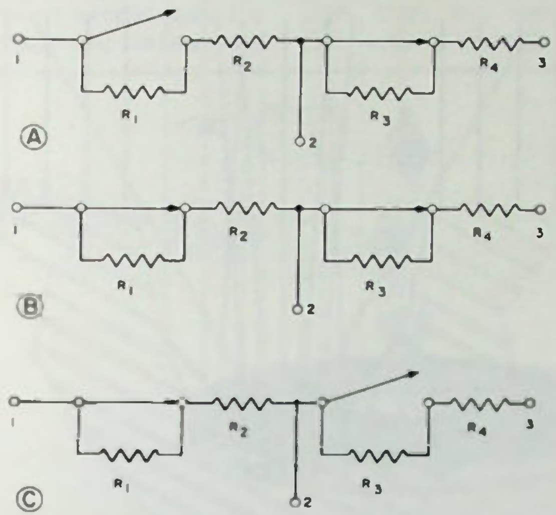


Fig. 7—Equivalent terminal resistances in a contactless meter (A) when the meter reading is below a minimum value, (B) when the meter reading is between minimum and maximum values and (C) when the meter reading is above a maximum value.

that of a conventional meter (left portion of Fig. 6), with the exception of a lightweight disk (central portion of Fig. 6) that is secured to rotate with the meter needle. Half of the disk is black to absorb light, while the other half is white to reflect incident light.

The optical system (right portion of Fig. 6) is secured to the meter When the meter needle, black-and-white disk and fiber-optics supports are in their present position, light, passing through the central pair of fiber optics from the light bulb, is reflected on the white surface of the nearby black-and-white disk. The reflected light passes through the outer pair of fiber optics and illuminates a pair of photoconductive cells, reducing the electrical resistance of these cells.

When the meter needle moves left of center, the attached black-and white disk must also move. Under these conditions, the left pair of fiber optics is then exposed to a black surface and not enough light is reflected to reduce the resistance of the left photoconductive cell, while the right pair of fiber optics is exposed to the white surface and enough light is reflected to reduce the resistance of the right photoconductive cell.

When the meter needle moves right of center, the attached black-and-white disk must also move again. Under these conditions, the left pair of fiber optics is exposed to the white surface and enough light is reflected to reduce the resistance of the left photoconductive cell, while the right pair of fiber optics is exposed to the black surface and not enough light is reflected to reduce the resistance of the right photoconductive cell.

When the meter needle is right of center, the left photoconductive cell is not exposed to light and its resistance is at a maximum (R_1 + R_2), while the right photoconductive cell is exposed to light and its resistance is reduced to a minimum (R_4). (The electrical equivalent of this situation is shown in Fig. 7A.) When the meter needle is centered (Fig. 7B), both photoconductive cells are exposed to light and their

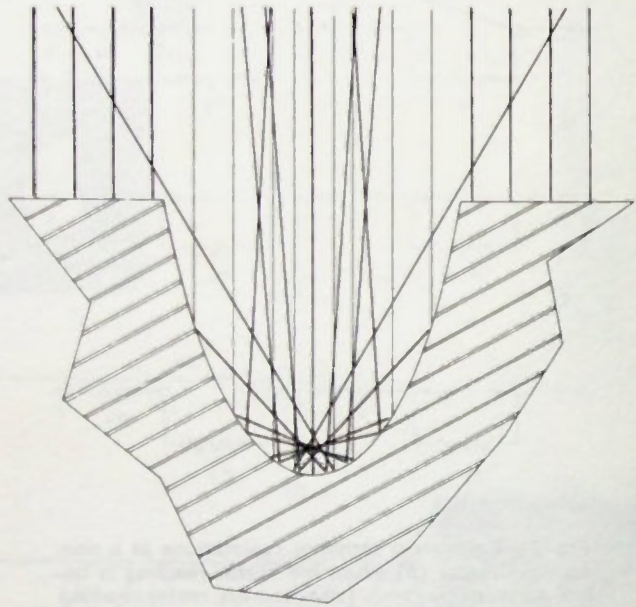


Fig. 8—Reflections that occur when light is focused vertically onto the surface of a phonograph record.

resistance is reduced to a minimum (R_2 and R_4). When the meter needle is left of center (Fig. 7C), the left photoconductive cell is exposed to light and its resistance is reduced to a minimum (R_2), while the right photoconductive cell is not exposed to light and its resistance is at maximum ($R_3 + R_4$).

Pointers are secured to the two fiber-optics supports, and they indicate on the meter needle scale the readings at which the meter will switch electrical circuits on or off. The position of these pointers and the fiber optics is changed by the knob-connected pinions.

Needleless Phonograph

An optical system has been developed that shines light vertically onto the surface of stereo and monophonic records (like those in current use) and produces an image of the record groove that can be converted into an audio signal by photosensitive semiconductors.

When light is focused vertically onto the surface of a phonograph record (Fig. 8), some of the light is reflected vertically — reflected from the horizontal surface adjoining the groove, from the horizontal center of the groove and from multiple reflections on the sides of the groove that total a vertical reflection. Only the vertically reflected light returns through the optical system. The light reflected vertically as a result of multiple reflections along the sides of the record groove becomes polarized as it is reflected and can be removed with a polarized filter. Only light reflected from horizontal surfaces remains.

An optical system designed to perform the function described is shown in Fig. 9. There a pair of lenses (top of figure) is used to focus the lamp's light into a nearly

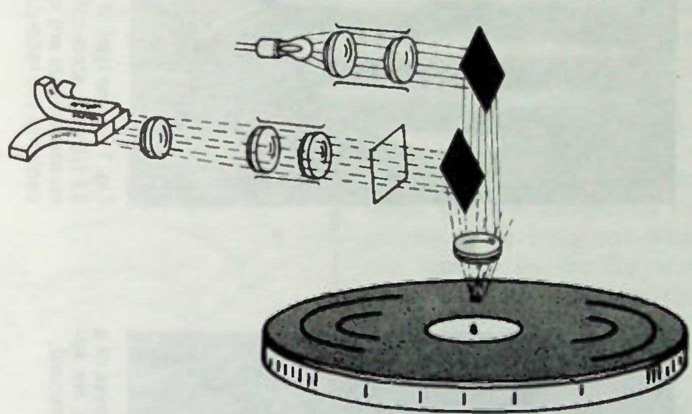


Fig. 9—Basic arrangement of components in an optical system designed for playing phonograph records. (Protected by French Patent No. 1,502,418. United States, Canadian, British, West German and Japanese patents pending.)

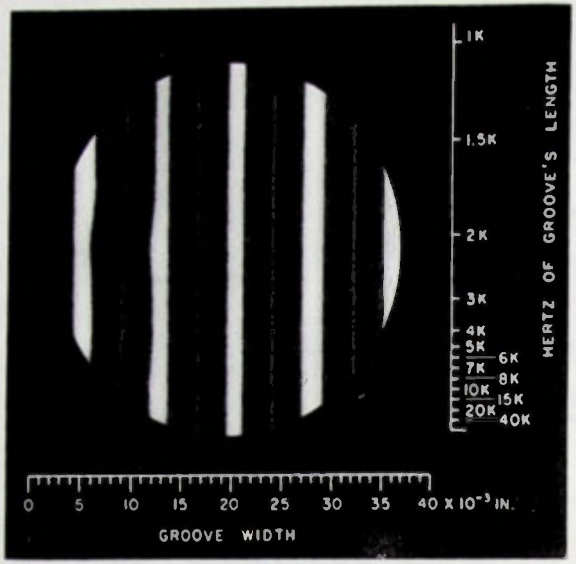


Fig. 10—The type of image produced by the grooves of a 78rpm phonograph record. Hertz measurements are appropriate for grooves 9in. from the center of the record.

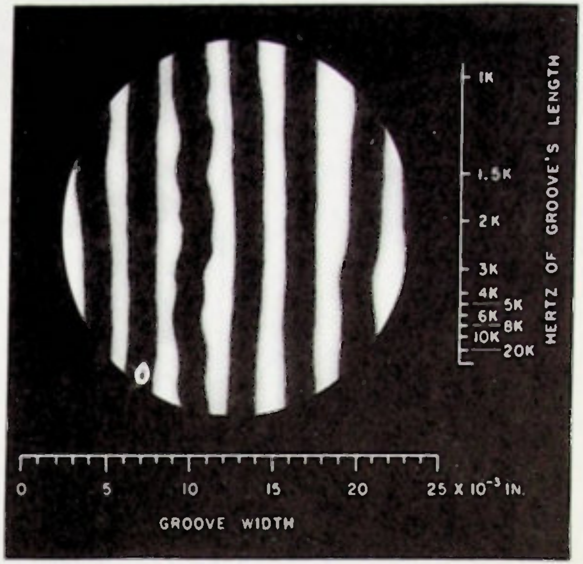


Fig. 11—The type of image produced by the grooves of a 33 1/3rpm stereophonic phonograph record. Hertz measurements are appropriate for grooves 9in. from the center of the record.

parallel beam (a beam of nearly infinite focal length). This light is reflected downward and through an objective lens by a mirror, or prism. A second mirror, or prism, casts its shadow over a portion of the objective lens and light from the lamp passes through only slightly less than half of the lens.

Light reflected vertically from the record passes through both halves of the objective lens, but the second mirror, or prism, is positioned so

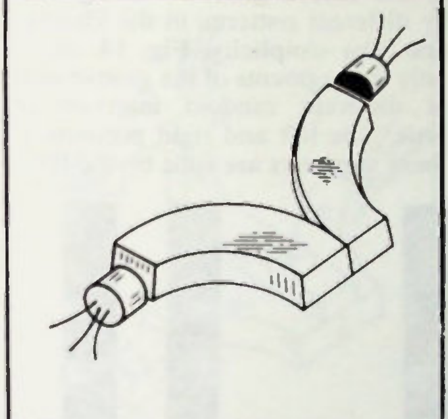


Fig. 12—A pair of fiber optics is used to split the groove image and transmit a portion of the image to each photocell.

that it reflects only the light that passes through the shadowed portion of the lens. Light passing through that portion of the lens does not contain reflected images of the light source but merely the less intense image of the record.

From Figs. 7 and 9 of the previous chapter we see that the shape of an image formed at the lens' focal length (now the light focused on the record) is not affected by reducing to less than half the portion of the lens used to form it, and nearly half the light normally passing through a lens (now the light returning from the record) can be blocked without affecting the quality of the image it forms. By using a single lens, rather than two lenses, to perform this dual function, we are able to project and receive virtually a vertical beam of light.

The light reflected by the record and second mirror, or prism, passes through a polarized filter to remove light reflected by the sides of the record groove. The remaining light then undergoes further magnification, and the image of the record is focused on a set of fiber optics.

The central portion of Fig. 10 shows the type of image obtained from a 78rpm record. Each broad black band is formed by the record groove. The irregular white bands are formed by the flat surface adjoining the grooves, and the narrow white lines are formed by the bottoms of the grooves. The scale superimposed beneath the groove's image indicates the groove's actual physical dimensions, while the vertical scale indicates the Hertz (cycle per second frequency) corresponding to the length of recorded sound waves at a distance of 9in. from the center of the record.

A somewhat similar image is obtained from a 33 1/3 rpm stereophonic record (Fig. 11). Standard stereophonic records are cut in such a manner that one recorded audio channel can be seen on one side of the groove, while the other recorded audio channel can be seen on the other side. This is most clearly demonstrated by the third groove from the left. Because 33 1/3 rpm grooves are cut much finer than

78rpm grooves, the central portions of these smaller grooves are not normally visible unless the optics are critically adjusted.

The two scales in Fig. 11 are basically the same as those shown in Fig. 10. The horizontal scales differ mainly because of a different degree of image magnification, while the vertical scales differ also because of different record speeds.

We observed that black-and-white images were produced by these vertical reflections even when red, yellow or transparent records were substituted for the regular black ones. Since as indicated in the previous chapter, there is no need for having the semiconductors respond to more than one color of light, a color filter can be used to eliminate the need for achromatic lenses — the current model uses these lenses.

When the image of one of the 33 1/3 rpm grooves shown in Fig. 11 is focused on a pair of fiber optics (Fig. 12), the resulting image is split in two and each photosensitive semiconductor sees only a portion of the groove.

As the record rotates (Fig. 13), each photocell sees only its portion of the record groove— designated by different patterns in the illustration. For simplicity, Fig. 14 shows only the segments of the groove seen at different random intervals of time. The left and right portions of these segments are split by the fiber

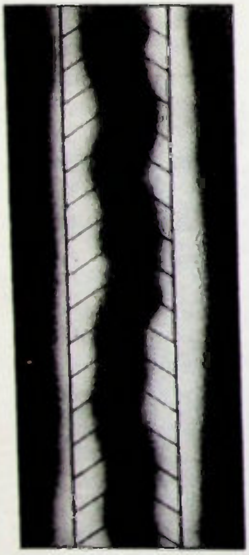


Fig. 13—Different patterns are used to illustrate the portions of a 33 1/3 rpm record groove seen by the two photocells—a different view for each photocell.

Fig. 14—Only the segments of the 33 1/3 rpm record groove seen at various random intervals of time are shown.

Fig. 15—The amount of light exposed to the left and right photocells changes at various intervals of time according to the waveform of the 33 1/3 rpm recorded sound.

optics (Fig. 15), one portion being seen by one photocell and the other portion being seen by the other photocell.

From Fig. 15 we see that the amount of light exposed to each photocell changes at various intervals of time according to the waveform of the recorded sound. The resulting changes in photocell output corresponds to the audio signals recorded on the record. Once amplified, these signals are like those produced by a conventional phonograph.

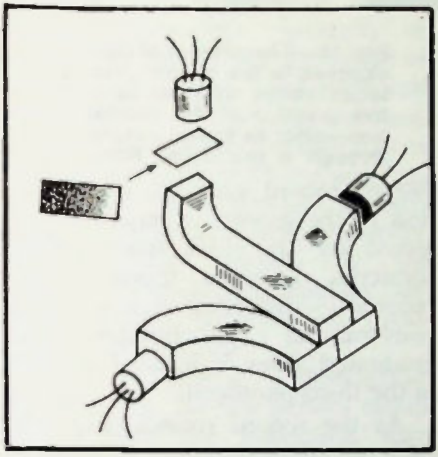


Fig. 16 — Some needleless phonographs may use a third set of fiber optics to view the white line formed at the very bottom of 78rpm record grooves.

When the photocells are properly tracking a record groove, the average amount of light exposed to each photocell remains equal. By connecting the photocells to a dc amplifier (direct-coupled analog integrated circuits are becoming relatively inexpensive and should lend themselves well to this application), the average dc output from each amplifier is equal when the photocells are properly tracking the record groove.

The Wheatstone bridge circuit shown in Fig. 4 can be used in conjunction with the photocells for tracking the record groove. The lamps in this circuit are connected directly to the amplifiers, their average brilliance depending on the amplifiers' dc output, and a motor with speed-reducing gears is substituted for the meter. If the groove's image is right of center on the fiber optics, one photocell sees more light than the other, one amplifier has a greater dc output than the other, the lamps are no longer of equal brilliance, the Wheatstone bridge becomes unbalanced and the motor receives power to center the

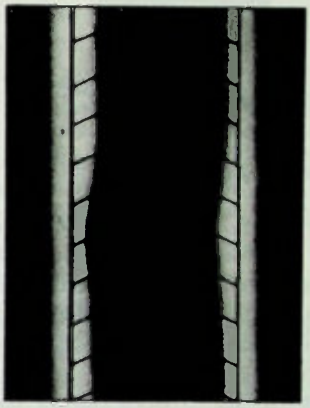


Fig. 17 — Different patterns are used to illustrate the portions of a 78rpm record groove seen by the three photocells. A slightly different view of the record groove is transmitted to each of the three photocells.

Fig. 18 — Only the segments of the 78rpm record groove seen at various random intervals of time are shown.

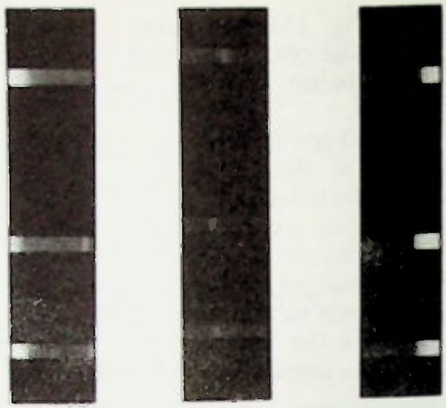


Fig. 19—The amount of light exposed to the central photocell varies with the relative position of the central line—after its image passes through a graduated filter.

phonograph tone arm over the record groove. This function of the Wheatstone bridge circuit in the phonograph is similar to its function in the stereo balance indicator described earlier.

A third set of fiber optics may be used in some needleless phonographs (Fig. 16) to view the white line formed at the very bottom of 78rpm record grooves. (This portion of the groove is generally protected by its sides from surface scratches normally found on old records — distorting their sound on conventional phonographs.) A graduated filter is located in front of the third photocell.

As the record rotates (Fig. 17), each photocell sees only its portion

Fig. 20 — One experimental model of a needleless phonograph with all the optics enclosed at the end of the tone arm.

of the record groove — designated by different patterns in the illustration. For simplicity, Fig. 18 shows only the segments of the groove seen at different random intervals of time. As in Fig. 14, the first two photocells see the left and right portions of the groove and are used in tracking the record. The third photocell sees only the central portion of the groove (Fig. 19).

Because of time delays in the tracking system, the third set of fiber optics is positioned at the average left-to-right location of the groove's center. As the image of the line seen by the third photocell appears to move left or right of center, the image is seen through less or more dense portions of the graduated filter and its brightness appears to vary as its relative position shifts. This apparent motion of the groove's central line changes the output of the third photocell and produces an audio signal.

The needleless phonograph shown in Fig. 20 uses prisms rather than mirrors for reflecting light within the optical system. Additional prisms were used to contain the optical system in an enclosure at the end of the tone arm.

By inserting fiber optics between the lamp and objective lens and extending the length of the fiber optics shown in Fig. 9, the lamp and photocells can be contained beneath the phonograph and only a relatively small tone arm — without moving parts or electronic circuits — can be positioned above the record. The tone arm can be connected to the light source and photocells with a relatively inexpensive cable of plastic fiber optics like the one shown in Figs. 12 and 13 of the previous chapter. This will permit the construction of needleless tone arms no larger than tone arms in current use.

Records played with a needleless phonograph will experience virtually no wear, and the frequency response of recorded music need not be limited by the mass of a vibrating needle and related components.

CHAPTER 22

Semiconductors for Power Supplies

All of the semiconductor circuits described in the previous chapters have required dc bias supplies. Most of these circuits can function in a fairly satisfactory manner with batteries, but greater stability, and frequently greater convenience, results when they receive their bias from ac-driven power supplies.

Many of the newer power supplies and power regulating circuits require semiconductors that have not yet been described in this book. These semiconductors must be understood if new power circuits are to be effectively serviced.

Unijunction Transistors

Although the basic structure of a unijunction transistor (Fig. 1) may appear somewhat similar to that of a field-effect transistor, this semiconductor operates on an entirely different principle, somewhat resembling that of a tunnel diode.

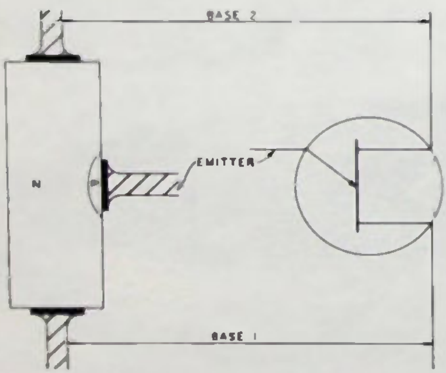


Fig. 1 — The structure of a unijunction transistor.

When the two bases (B_1 and B_2)

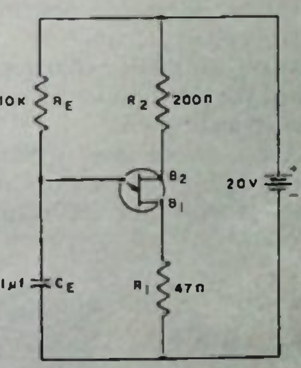


Fig. 2 — A simple unijunction transistor switching circuit.

of a unijunction transistor are connected in a circuit (Fig. 2), current flows through the transistor from base-one to base-two and the resulting voltage drop between the two bases (V_{B2B1}) is dependent on the amount of current through the channel of N-type material (I_{B2}) and the total resistance of that channel (r_{BB}). Just as the voltage drop at a potentiometer tap is less than the total voltage drop across the potentiometer, the voltage drop at the P-N junction (the voltage drop in the N-type material at the junction, not the voltage drop across the junction) within the semiconductor channel is less than the total voltage drop across the channel. The junction is, therefore, more positive than base-one, while less positive than base-two.

When a capacitor (C_E) connected to the transistor's emitter (Fig. 2) is discharged (both plates at the same potential), the emitter (constructed of P-type material) is less positive than the conductive channel (constructed of N-type material), and virtually no current flows between the conductive channel and the emitter (the diode formed by the P-N junction is reverse biased, and virtually no current flows through the diode).

As a small current flows through resistor R_E, the voltage across the capacitor increases, and once the charge across the capacitor becomes so great that the P-type material is more positive than the N-type material, the diode formed is forward biased and current will flow from base-one to the emitter. (The emitter is still less positive than base-two, so the current must flow from the more negative base-one to the emitter.)

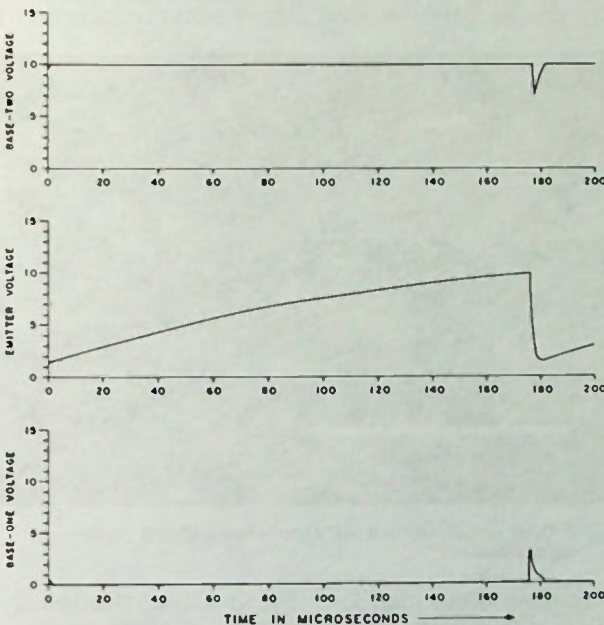


Fig. 3 — Emitter, base-one and base-two voltages present in the switching circuit.

Once a current flows between base-one and the emitter, the semiconductor material between base-one and the emitter behaves like the material in a tunnel diode. The flow of current between base-one and the emitter reduces the resistance of this semiconductor material. The reduction in resistance reduces the voltage drop between base-one and the emitter, which in turn further reduces the material's resistance. (This condition is indicated by the negative resistance slope of a tunnel diode characteristic curve shown in Figs. 1 and 2, chapter 7.) The current flowing between base-one and the emitter reduces the potential drop across the capacitor (C_E) to zero. When the emitter is no longer more positive than base-one, the current ceases to flow through the emitter, and the capacitor again becomes charged by resistor R_E.

While current was flowing from base-one to the emitter, the resistance of N-type material near base-one was reduced. As a result of this reduced resistance, more current flowed not only to the emitter but to base-two as well. The base-one to emitter current increased the base-one to base-two current.

The entire channel of N-type material conducts more current when the emitter voltage exceeds a peak-point voltage (V_p). The emitter, base-one and base-two voltages that result are shown in Fig. 3.

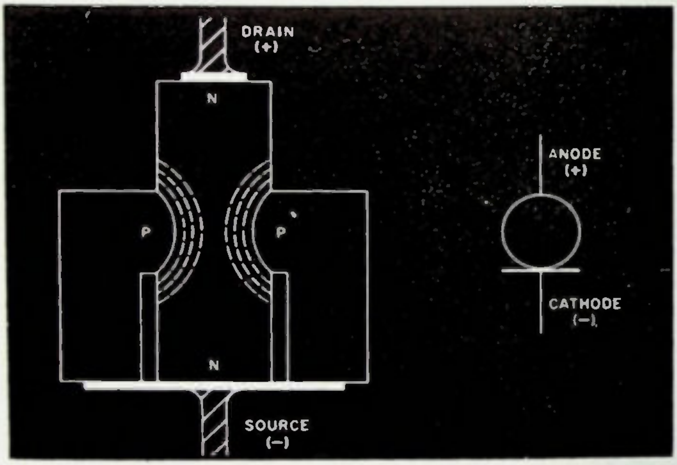


Fig. 4 — Structure of a reverse-biased field-effect diode.

Field-Effect Diodes

Field-effect, or current-regulating diodes (Fig. 4), are very similar to the field-effect transistors described with Fig. 1 in chapter 5. The basic difference is that the diodes are only double-lead components and the gate material in field-effect diodes is connected to the same lead as the end of the conductive channel designated the source. (Another difference between the two particular examples shown is that in the transistor the gates are shown as being made of N-type material, while in the diode they are shown as being made of P-type material—the type of material for the conductive channel having also been changed.

If the diode gates were not connected to the source lead and the P-type material was made more negative than the N-type conductive channel, the P-N junctions would be reverse biased. The greater the reverse bias, the greater the effective junction size and the smaller the remaining N-type channel for conducting current. The opposite condition would occur if we were to reverse the polarity of the applied bias voltage.

By reversing the voltage across the field-effect diode shown in Fig. 4, we are in effect changing the diode so that the gates are connected to the drain rather than the source (Fig. 5). The P-type gate then becomes positive with respect to the source and N-type channel. The P-N junctions are then forward biased, reducing the effective size of these junctions and increasing the effective size of the N-type channel—permitting it to conduct more current.

The curve in Fig. 6 indicates the characteristics of the field-effect diode just described. When the anode is more positive than the cathode, moderate changes in anode-to-cathode voltage results in nearly insignificant changes in diode current. When the polarity of the applied voltage is reversed, much greater changes in diode current result from changes in diode voltage.

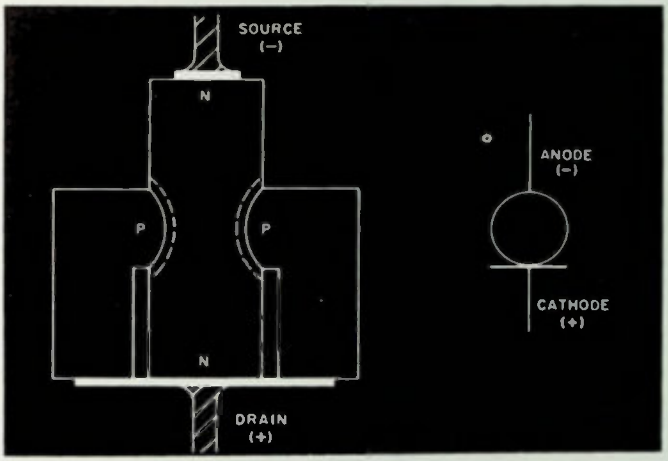


Fig. 5 — Structure of a forward-biased field-effect diode.

Zener Diodes

The physical structure of a zener diode (Fig. 7) is basically the same as the structure of a conventional diode. In fact, when forward biased, both types of diodes conduct current in the same manner — both then having the same general forward-bias characteristic curve.

Within certain limits, zener diodes and conventional diodes also conduct current in the same manner when reverse biased—both again having the same general characteristic curve.

As indicated in earlier chapters, when a P-N junction is reverse biased, a barrier is formed between the P- and N-type material and only a nearly insignificant amount of leakage current is permitted through the barrier and the component. If this reverse bias becomes too great, however, the barrier breaks down and the current becomes considerably greater in both zener and conventional diodes. The basic difference between these two types of diodes is the stability of this barrier breakdown voltage.

The resistance of P- and N-type material in regular diodes is not as great as that of the material used in zener diodes (there is a slight chemical difference), and the resulting junction barriers formed are not as thin in regular diodes as they are in zener diodes. The electric field developed across a thin junction (zener junction) is greater than that developed across a wider junction (regular diode junction). It is the intensity of this electric field in zener diodes that is the primary cause of its junction breakdown, while in the wider junction of less resistive semiconductor material (regular diodes) the electric field is not as great and the temperature of the semiconductor material has a direct effect on the junction breakdown. (The strength of an electrical field developed across a P-N junction is the strength of the mutual attraction of positive and negative charges across the diode barrier.

Zener diodes are used (instead of regular diodes) for voltage regula-

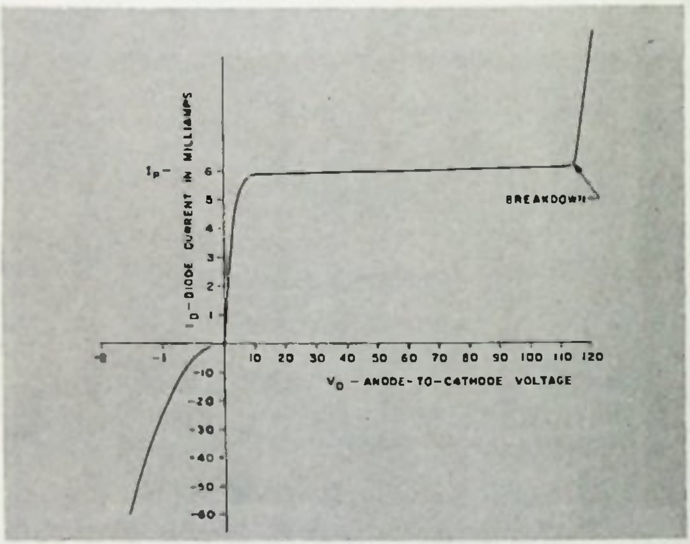


Fig. 6 — The characteristic curve of a field-effect diode.

tion since their breakdown voltage remains nearly constant—heat having very little effect on the breakdown voltage.

Many zener diodes generate noise when their reverse bias current nearly equals the current encountered during junction-barrier breakdown—the "knee" region of the zener diode characteristic curve (Fig. 8). This is true because the breakdown current does not increase suddenly, but increases in increments as the breakdown voltage is reached. To reduce zener diode noise, the resistance in a zener diode circuit should be low enough to permit a zener diode current greater than the current encountered at breakdown.

Complex Diodes

Just as there is no limit to the possible assortment of designs for integrated circuits, there is no limit to the assortment of stacked P-N junctions that can be fabricated to produce different types of complex diodes. Rather than attempt to describe every possible variation, the balance of this chapter and the following chapter is restricted to six complex diodes most frequently encountered in power-supply and power-regulating circuits.

The characteristics of these complex diodes can be more readily understood by studying their more simple diode and/or transistor equivalent circuits. It would, therefore, be merely repetitious to also discuss the arrangement of junctions within these complex diodes.

TC Diodes and Thyrectors

Unfortunately, both temperature-compensating diodes and thyrectors are occasionally identified by the same symbol (Fig. 9). Unless the letters "TC" are shown to the right of the symbol (Fig. 9) to indicate that a temperature-compensating diode is being used, technicians must

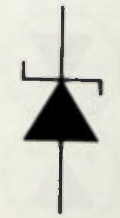


Fig. 7 — Symbol for a zener diode.

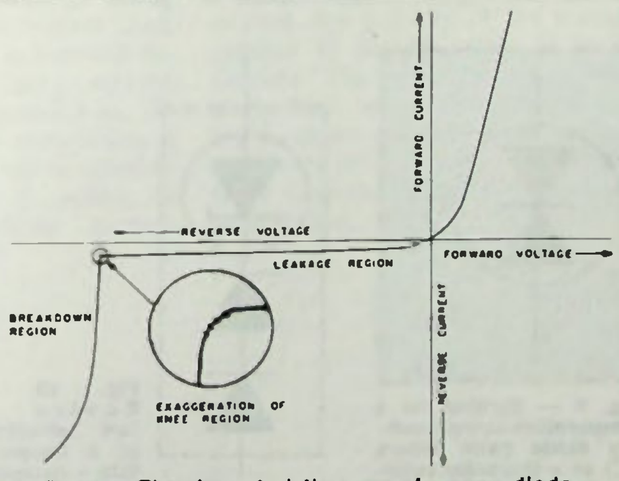


Fig. 8 — The characteristic curve of a zener diode.

identify the component by checking the function of the circuit or a parts list.

Temperature-compensating diodes, sometimes called reference diodes, are an improved and more complex version of zener diodes. Earlier in this chapter it had been indicated that, compared to regular diodes, heat had very little effect on zener diode breakdown voltages. Measurements, however, do indicate that the breakdown voltage of a reverse-biased zener diode actually increases slightly as the diode's temperature increases. Similar measurements also indicate that the small voltage drop across a forward-biased regular diode decreases slightly as the temperature increases. These two opposing characteristics can be combined, and a temperature-compensating diode is formed by connecting one or more forward-biased regular diodes in series with a reverse-biased zender diode (Fig. 10). As the temperature increases, a reduction in the voltage drop across the regular diodes equals a corresponding increase in the voltage drop across the zener diode, and the total voltage drop remains nearly unchanged.

When the voltage across the temperature-compensating diode forward biases the regular diode (it then being capable of conducting current) and reverse biases the zener diode, both the regular diode and zener diode will conduct current once the zener breakdown voltage has been reached. Under these conditions the temperature compensating diode will function as though it was merely a zener diode—except for greater temperature stability (the left portion of the characteristic curve shown in Fig. 11).

When the polarity of the voltage across the temperature compensating diode is reversed, the zener diode will be forward biased (it then being capable of conducting current as indicated by the right portion of the characteristic curve shown in Fig. 8) and the regular diode will be reverse biased (it then being incapable of conducting any significant

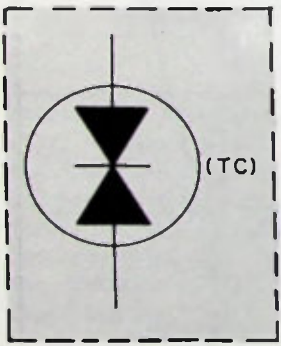


Fig. 9 — Symbol for a temperature-compensating diode (with letters TC) or a thyrector (without those letters).

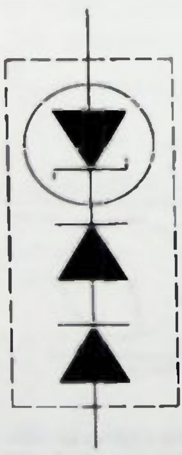


Fig. 10 — Equivalent structure of a temperature-compensating diode.

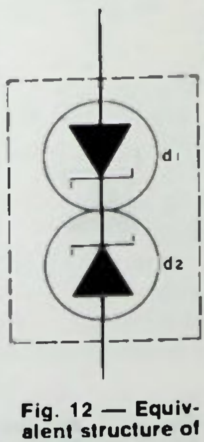


Fig. 12 — Equivalent structure of a thyrector.

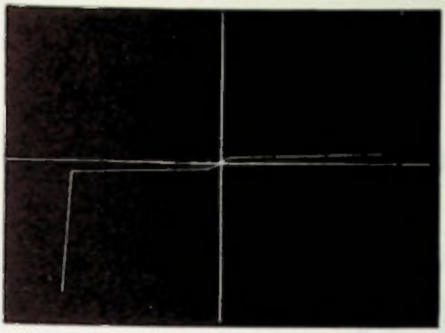


Fig. 11 — The characteristic curve of a temperature-compensating diode.

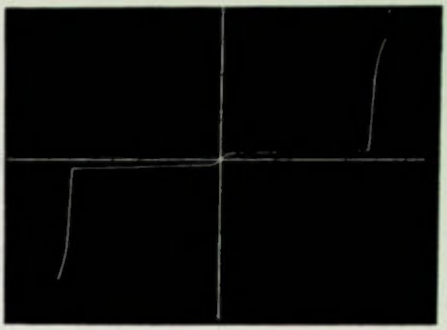


Fig. 13 — The characteristic curve of a thyrector.

current). Under these conditions, only a very small current flows through the temperature-compensating diode (the right portion of the characteristic curve shown in Fig. 11).

Thyrectors, sometimes called surge protectors or trigger diodes, contain the equivalent of two zener diodes (Fig. 12) connected so that one is reverse biased when the other is forward biased.

When the top lead (Fig. 12) is more positive than the bottom lead, one zener diode (d_1) is forward biased while the other (d_2) is reverse biased. Under these conditions, diode d_1 is capable of conducting a significant current at any reasonable voltage and diode d_2 is capable of conducting a significant current when the applied voltage exceeds its breakdown voltage.

When the polarity of the applied voltage is reversed (the bottom lead now more positive than the top lead), the conditions in the two zener diodes are reversed. Diode d_2 is now forward biased and capable of conducting a significant current at any reasonable voltage, while diode d_1 is now reverse biased and capable of conducting a significant current only when the applied voltage exceeds its breakdown voltage.

Because of the characteristics described, the polarity of the voltage applied to the thyrector is insignificant. The thyrector has a relatively high internal resistance that breaks down only when the applied ac voltage or dc voltage of either polarity exceeds its zener breakdown voltage (Fig. 13).

CHAPTER 23

More Semiconductors for Power Supplies

In addition to the semiconductors described in the previous chapter, there are others now being used in modern solid-state power supplies and power regulating circuits. These new diodes must also be understood if the electronic circuits in which they are used are to be effectively serviced.

Four-Layer Diodes

Four-layer diodes (sometimes called Shockley diodes in honor of William Shockley who in 1949 first predicted the possibility of a junction transistor) contain the equivalent of two transistors connected in series (Fig. 1).

When a nominal voltage is applied across this diode, only a small current leaks through the unbiased transistors. As the applied voltage increases, the amount of leakage current through each transistor also increases until each transistor supplies the other with a sufficient base current to develop a large enough base-to-emitter voltage to forward bias the other transistor. Under these conditions the current flowing through one transistor is sufficiently large to switch the other on—each transistor maintaining the other in an *on* condition.

The right portion of the characteristic curve shown in Fig. 2 corresponds to the conditions just described. Once the voltage drop across the four-layer diode exceeds the breakover voltage (V_{BR}), there is enough leakage current to switch its equivalent transistors to an *on* condition, reducing the diode resistance and the voltage drop across the diode. Even though the voltage drop

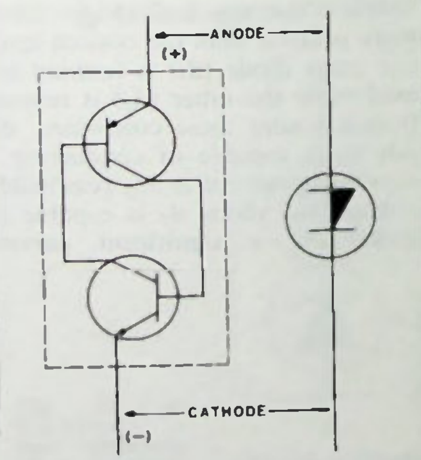


Fig. 1 — Equivalent structure of a four-layer diode.

across the diode is reduced, there is still enough current flowing through each transistor to keep the other in an *on* condition.

When the polarity of the applied voltage is reversed, the four-layer diode's equivalent transistors are reverse biased and block the flow of current like reverse-biased diodes (the left portion of the characteristic curve shown in Fig. 2).

A low-impedance ac voltage supply can be designed so that its output voltage is virtually unaffected by the load resistance. By connecting such a power supply directly to a four-layer diode (Fig. 3), we can see how the diode functions in response to the applied voltage (Fig. 4). From the resulting curve we see that the diode is switched on and conducts current once the forward-bias voltage exceeds the diode's breakover voltage (V_{BR}). The diode then continues to conduct current until the forward-bias voltage returns to zero. The diode will resist the flow of current as the polarity of the applied voltage changes and the diode is reverse biased, even if the reverse-bias voltage is also greater than the forward-bias breakover voltage.

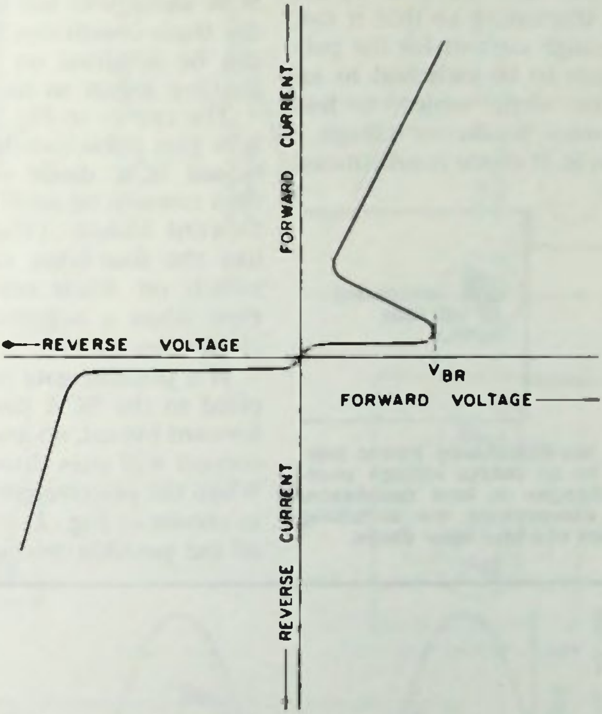


Fig. 2 — The characteristic curve of a four-layer diode.

SCR Diodes

The structure of an SCR (semiconductor controlled rectifier) diode is nearly the same as that of a four-layer diode (Fig. 1) except for the addition of a gate lead (Fig. 5). When the gate lead is not used, these two diodes function in the same manner. The gate lead serves merely to forward bias one of the equivalent transistors so that it can conduct enough current for the pair of transistors to be switched to an *on* condition while subject to less than the normal breakover voltage.

When an SCR diode is substituted for a four-layer diode in a low-impedance ac circuit (Fig. 3), the SCR diode will be switched on like the four-layer diode (Fig. 4) when its forward-bias breakover voltage is exceeded.

If the output of the low-impedance ac voltage supply is adjusted so that the forward-bias voltage never exceeds the breakover voltage, the SCR diode will not switch on. Under these conditions the SCR diode can be switched on by applying a positive signal to its gate (Fig. 6).

The curves in Fig. 7 show a positive gate pulse switching a forward-biased SCR diode on. The diode then remains on until it is no longer forward biased. (The SCR diode, like the four-layer diode, will not switch on when reverse biased—even when a negative pulse is applied at its gate.)

If a positive gate pulse is not applied to the SCR diode while it is forward biased, no anode-to-cathode current will pass through the diode. When the positive gate pulse occurs, as shown in Fig. 7, less than a third of the possible average cathode-to-

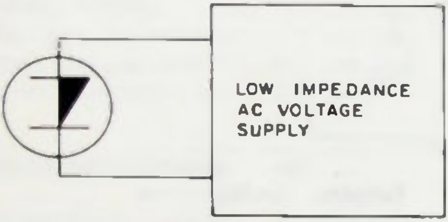


Fig. 3 — A low-impedance power supply, having an ac output voltage unaffected by changes in load resistance, is used to demonstrate the switching characteristics of a four-layer diode.

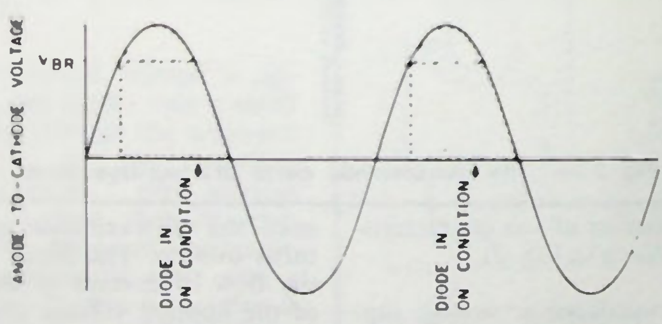


Fig. 4 — The four-layer diode is switched on when the forward-bias voltage exceeds its breakover voltage (V_{BR}).

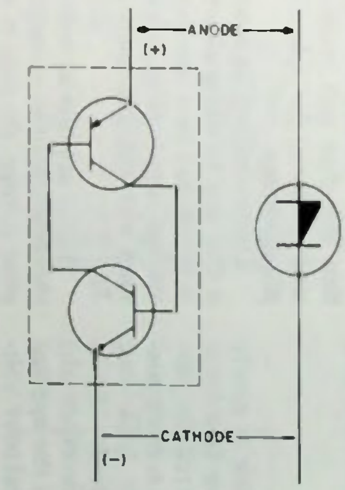


Fig. 5 — Equivalent structure of an SCR diode.

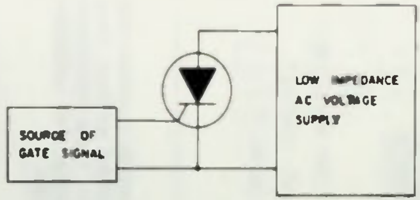


Fig. 6 — A gate pulse circuit is used to switch on the SCR diode when the ac output voltage from the low-impedance power supply is less than the breakover voltage.

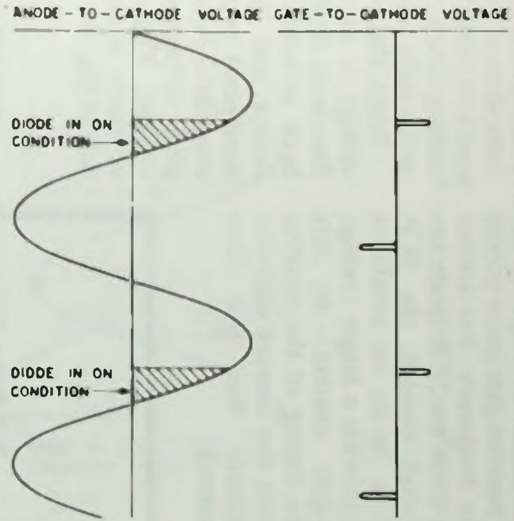


Fig. 7 — A positive cathode-to-gate voltage switches the SCR diode on when it is forward biased.

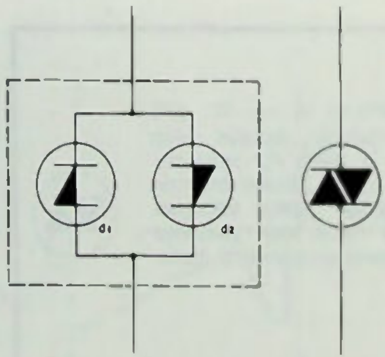


Fig. 8 — Equivalent structure of a diac.

gate current passes through the diode. If the positive gate pulse occurs at the same frequency, but sooner (nearer the times the SCR diode becomes forward biased), the SCR diode will be in an *on* condition longer and provide a larger average cathode-to-gate current. In this manner, the timing of the positive pulse controls the average cathode-to-gate current that will flow through a circuit.

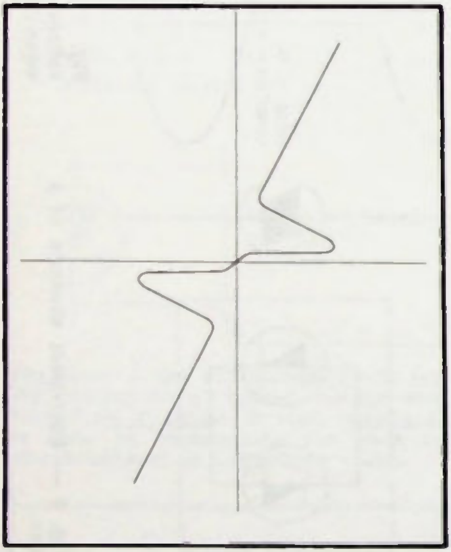


Fig. 9 — The characteristic curve of a diac.

Diacs

The structure of a diac is nearly the same as that of two four-layer diodes (Fig. 8) connected in parallel. When the top lead is more positive than the bottom lead, one equivalent four-layer diode (d_2) is forward biased while the other (d_1) is reverse biased. When the applied voltage exceeds the breakover voltage of diode d_2, that diode is switched to an *on* condition while the reverse-biased diode, d_1, does not conduct significant current. The resulting characteristic curve is shown as the right portion of Fig. 9.

When the polarity of the applied voltage is reversed, the equivalent four-layer diode that had previously been switched on (d_2) is reverse biased and fails to conduct any significant current, while the other diode (d_1) is switched on as the applied voltage exceeds its breakover voltage. The resulting characteristic curve is shown as the left portion of Fig. 9.

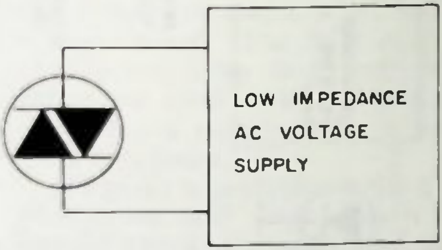


Fig. 10 — A low-impedance power supply, having an ac output voltage unaffected by changes in load resistance, is used to demonstrate the switching characteristics of a diac.

From these conditions we see that the diac can function equally well whatever the polarity of the applied voltage.

The switching characteristics of a diac can be more clearly seen if it is connected to a low-impedance ac voltage supply (Fig. 10) like the one used with the four-layer diode (Fig. 3).

As the applied ac voltage exceeds the diac's positive or negative breakover voltage, the diac is switched on and conducts current until the por-

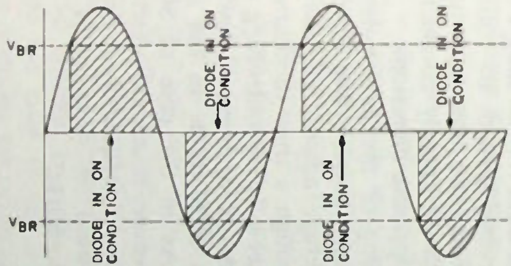


Fig. 11 — The diac is switched on when the forward- or reverse-bias voltage exceeds its breakover voltage (V_{BR}).

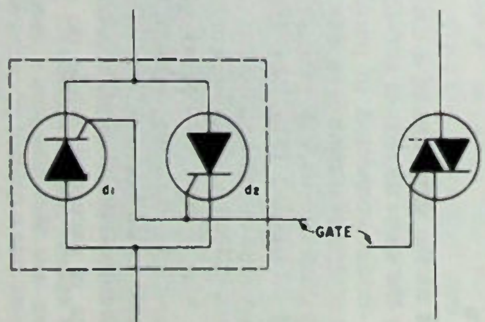


Fig. 12 — Equivalent structure of a triac.

Fig. 13 — A gate pulse circuit is used to switch on the triac when the ac output voltage from the low-impedance power supply is less than the breakover voltage.

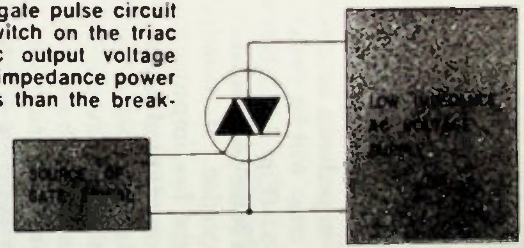
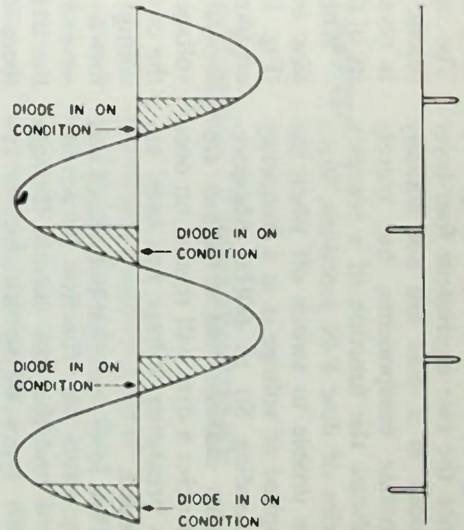


Fig. 14 — A positive gate voltage switches the triac on when it is forward biased, while a negative gate voltage switches the triac on when it is reverse biased.



tion of the cycle that the applied voltage returns to zero (Fig. 11). Since the two equivalent four-layer diodes in a diac are not actually separate components, but merely indicate the function of a complex series of diac P-N junctions, diacs are unable to switch off when the applied ac voltage is at a frequency exceeding 50 to 60Hz. Measurements indicate that at higher frequencies a diac will remain on once its breakover voltage has been exceeded.

We have seen that both diacs and thyrectors can function in ac circuits and that the current through both types of semiconductors increases abruptly when the applied voltage exceeds a certain amount. This, however, is where their similarity ends. The current through a thyrector returns to normal when the applied voltage returns to normal, while a diac remains in an *on* condition until the applied voltage returns to zero.

Triacs

Just as the structure of a diac is nearly the same as that of two four-layer diodes connected in parallel, the structure of a triac is nearly the same as that of two SCR diodes connected in parallel (Fig. 12). When the top lead is more positive than the bottom lead, one equivalent SCR diode (d_2) is forward biased while the other (d_1) is reverse biased. Under these conditions, SCR diode d_2 will be turned on when its breakover voltage is exceeded, while no significant current will flow through the reverse-biased SCR diode (d_1). The opposite condition will occur (diode d_1 turned on while diode d_2 is reverse biased) when the polarity of the applied voltage is reversed.

When a triac is substituted for a diac in a low-impedance ac circuit (Fig. 10), the triac will be switched on like the diac (Fig. 11) when its forward- or reverse-bias breakover voltage is exceeded. If, however, the output of the low-impedance ac voltage supply is adjusted so that the forward- or reverse-bias voltage never exceeds the breakover voltage, the triac will not switch on. Under these conditions the triac can be switched on by applying a signal to its gate (Fig. 13).

The curves in Fig. 14 show a positive gate pulse switching the triac on when it is forward biased and a negative gate pulse switching it on when it is reverse biased. The triac remains on and continues to conduct current until it is no longer forward or reverse biased. Its equivalent SCR diodes are alternately turned on, each alternately conducting current in opposite directions.

The timing of the gate pulse is just as effective a control in regulating the average amount of current flowing through a triac in alternate directions as it is in controlling the average amount of current flowing through an SCR diode in but one direction. As in the case of SCR diodes, however, triacs are effective only at lower frequencies, being primarily used at frequencies of around 50 to 60Hz.

CHAPTER 24

Constant-Voltage Transformer

Electronic products serviced by electronic technicians contain a power supply of some kind, whether powered by batteries or an external ac voltage source. And whatever their application, these power supplies contain basic circuits designed to perform one or more of the following functions: ac voltage regulation, ac current regulation, ac-to-dc power conversion, dc-to-ac power conversion, dc voltage regulation, dc current regulation or filtering.

AC Voltage Regulation

A component most frequently used for obtaining desired ac voltages is the transformer. It contains one or more coils that function on the same basic principles as the coils shown in Figs. 10, 11, and 12 in chapter 15. When current flows through a coil, a magnetic field is produced (Fig. 1A). This field can be concentrated (Fig. 1B) with a soft iron core.

Each cubic inch of soft iron contains millions of microscopic permanent magnets (called domains) formed by iron molecules. When the core is not magnetized, these domains are orientated in random directions, each canceling out the other's magnetic field (Fig. 2A). An external magnetic field magnetizes the soft iron by lining up these domains (Fig. 2B). [*With appropriate equipment, one can hear the flopping of these domains as they line up.*] These domains will remain aligned (the metal will remain magnetized) until they are disarranged by thermomolecular motion (heated molecules bouncing around and knocking the domains out of alignment), physical blows (such as a hammer tapping the iron) or an external magnetic field of alternating polarity that is drawn away from the magnetized iron.

When the domains are aligned (Fig. 2B), the north magnetic pole

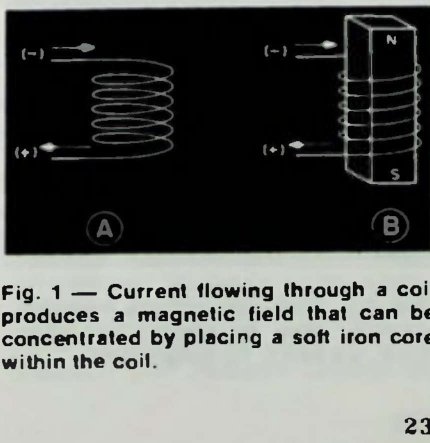


Fig. 1 — Current flowing through a coil produces a magnetic field that can be concentrated by placing a soft iron core within the coil.

of each domain faces the south magnetic pole of an adjacent domain. When two magnetized rods are bent into a shape somewhat resembling a horseshoe (Fig. 3), the alignment of north and south magnetic poles will still remain the same — north magnetic poles facing south magnetic poles. The two vertical portions of the transformer core are shown split merely to demonstrate that although the domains remain aligned, the direction of alignment in the vertical left portion of the core is opposite that in the vertical right portion.

Two iron-core coils, like the one shown in Fig. 1B, are used in the transformer shown in Fig. 4. (Terminal 1 on the left coil corresponds to terminal 3 on the right coil, and terminal 2 corresponds to terminal 4 — one coil being an upsidedown version of the other.) When a negative potential is connected to terminal 1 and a positive potential is connected to terminal 2, a current flows through the coil producing a north magnetic pole at the upper portion of its core and the lower portion of the right core.

As indicated in chapter 15, when the applied current increases the strength of a magnetic field, the magnetic field induces a current that flows through the coil in a direction (from terminal 2 to terminal 1) opposing that of the applied current (from terminal 1 to terminal 2). The increasing magnetic field also induces a current in the same direction (from terminal 4 to terminal 3) through the secondary coil.

When there is a reduction in the

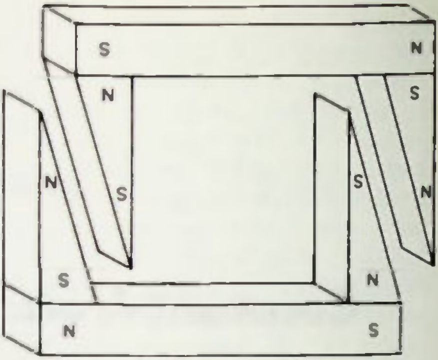


Fig. 3 — Magnetic fields in the primary and secondary portion of a transformer core are aligned in opposite directions.

Fig. 2 — Soft iron contains magnetic domains that are aligned only when the object is magnetized.

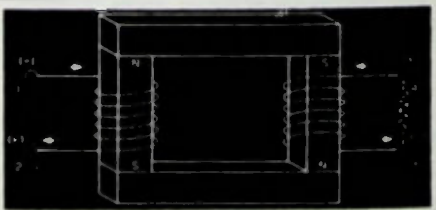


Fig. 4 — Changes in current through the primary winding of a transformer induce current through the secondary winding.

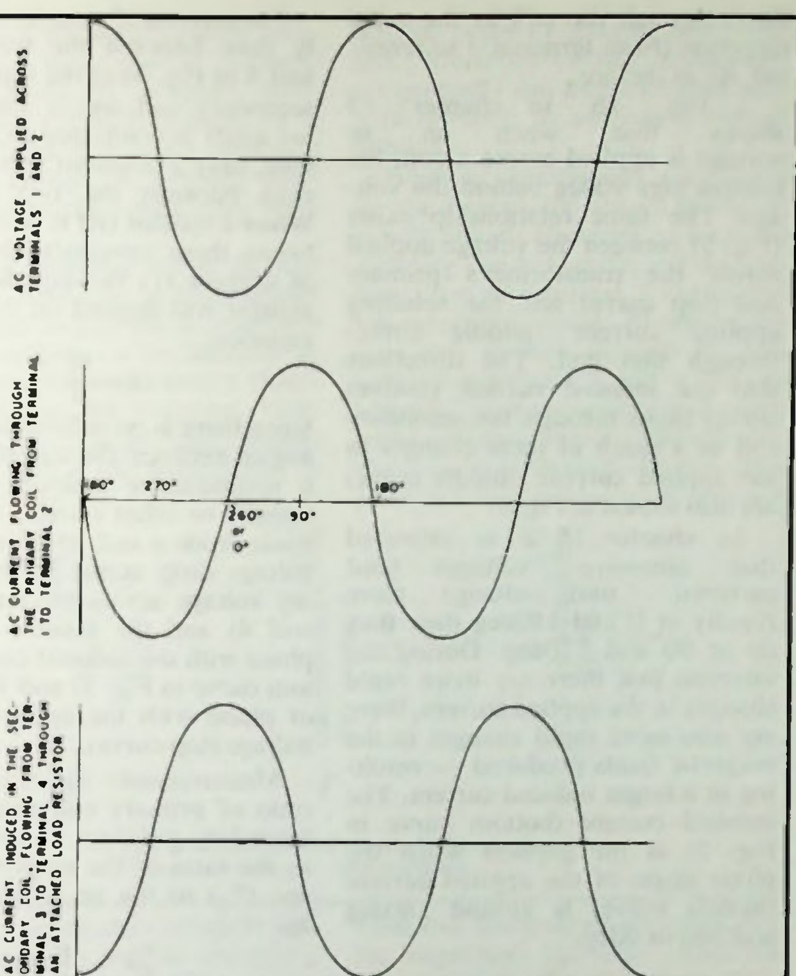


Fig. 5 — There is a relationship between the phase angles of an ac voltage applied to a transformer's primary winding, the ac current through that winding and the current induced in the secondary winding.

amount of applied current, the strength of the magnetic field decreases and a current is induced in the opposite direction (from terminal 3 to terminal 4 in Fig. 4) through the secondary coil.

By reversing the direction of the applied current (from terminal 2 to terminal 1 in Fig. 4), the polarity of the magnetic field produced is also reversed. However, as the applied current increases the strength of the reversed magnetic field also increases. The induced current still flows through the coil in the same direction (from terminal 3 to terminal 4) as before.

Fig. 16 in chapter 15 shows that when an ac voltage is applied across a coil, the current lags 90deg behind the voltage. The same relationship exists (Fig. 5) between the voltage applied across the transformer's primary coil (top curve) and the resulting applied current (middle curve) through that coil. The directions that the induced current (bottom curve) flows through the secondary coil as a result of these changes in the applied current (middle curve) are also shown in Fig. 5.

In chapter 15 it is indicated that sinewave voltages (and currents, too) change more rapidly at 0 and 180deg than they do at 90 and 270deg. During the intervals that there are more rapid changes in the applied current, there are also more rapid changes in the magnetic fields produced — resulting in a larger induced current. The induced current (bottom curve in Fig. 5) is the greatest when the phase angle of the applied current (middle curve) is around 180deg and 360 or 0deg.

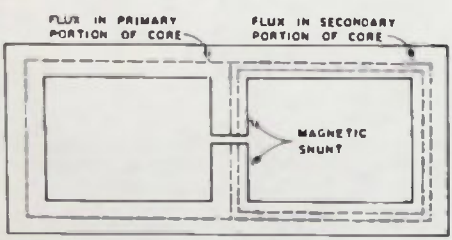


Fig. 6 — "Constant-voltage" transformers contain a magnetic shunt designed to restrict magnetic saturation to their secondary core.

No current, of course, will actually flow between the terminals (3 and 4 in Fig. 4) of the transformer's secondary coil unless some circuit (or load) is connected to it. Otherwise, only a potential voltage would exist between the two terminals. When a resistor (R) is connected between these terminals, the amount of current (I) flowing through the resistor will depend on the familiar equation:

$$I = \frac{V}{R}.$$

Since there is no difference in phase angles between the current through a noninductive resistor (a carbon resistor or other resistor that is not made from a coil of wire) and the voltage drop across it, the secondary voltage across the terminals (3 and 4) and the resistor (R) is in phase with the induced current (bottom curve in Fig. 5) and 180deg out of phase with the applied primary voltage (top curve).

Measurements indicate that the ratio of primary coil turns (N_p) to secondary coil turns (N_s) is the same as the ratio of the ac primary voltage (E_p) to the ac secondary voltage (E_s).

$$\frac{N_p}{N_s} = \frac{E_p}{E_s}.$$

If there are twice as many turns in the secondary coil as there are in the

Fig. 7 — Primary and secondary coils in a "constant-voltage" transformer are basically the same as those in a regular transformer.

primary coil, the ac secondary voltage will be twice the ac primary voltage. (Since dc primary voltages do not produce changes in a magnetic field that could induce a secondary current, dc primary voltages do not produce secondary voltages.)

When a relatively stable ac line voltage is available, a transformer can be used to provide a relatively stable ac voltage — the desired ac voltage being obtained with a transformer having the required ratio of turns in its primary and secondary coils. And if the transformer is rated at an adequate wattage, moderate changes in the ac secondary current will have little effect on the ac secondary voltage delivered by the transformer.

There are, however, occasions when the ac line voltage is not sufficiently stable. "Constant-voltage" transformers can then be used effectively to reduce these changes to less than 1 percent. These transformers are particularly useful in solid-state power supplies since semiconductors need not then absorb excess power whenever the line voltage becomes greater than normal. "Constant-voltage" transformers generally use a capacitor to help improve their ac secondary voltage stability.

When a capacitor is substituted for the resistor (R) shown in Fig. 4, it will absorb electrons at terminal 3 during one-half cycle and then return them to terminal 3 during the following half cycle. Electrons are thus able to flow through the coil between terminals 3 and 4 without actually being conducted through some external circuit. Virtually no electrons pass through the capacitor

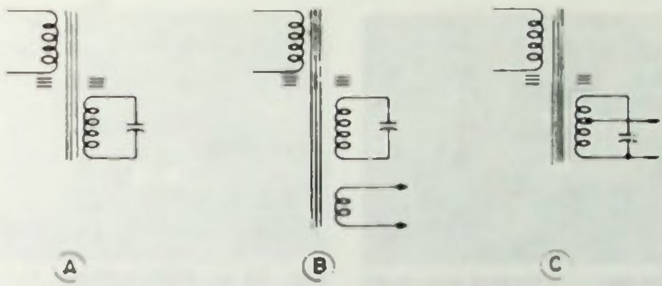


Fig. 8 — Schematic equivalents of three "constant-voltage" transformers.

from one terminal to the other, and the capacitor, therefore, does not consume a significant amount of the induced energy. The induced energy is used only to increase the strength of the magnetic field of the coil's core — aligning more domains.

When the transformer core's domains are all aligned in one direction and the applied ac primary voltage alternates, causing the domains to all become aligned in the opposite direction, the electrical energy induced is absorbed by the capacitor (C). When the capacitor discharges this energy through the secondary coil, a magnetic field is produced that is strong enough to align most of the domains again in the first direction — the capacitor and coil consuming only a small portion of the induced energy, the balance of the energy remaining to realign the domains.

As the applied ac primary voltage again alternates and provides a magnetic field to align all the domains in the first direction, the capacitor across the secondary coil also supplies energy for producing a magnetic field capable of aligning most of the domains in the first direction.

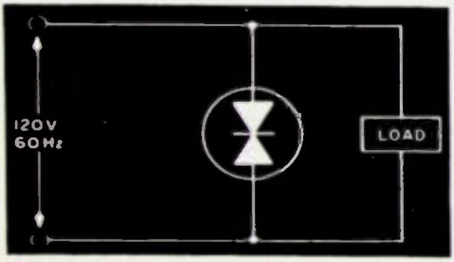


Fig. 9 — Excess ac voltages are shunted from the load circuit by the thyrector.

The total magnetic field produced by the primary and secondary coils is stronger than the field required for aligning all of the domains. It is this magnetic "saturation" of the core that is used for controlling "constant-voltage" transformers.

Although we wish to saturate the secondary portion of a "constant-voltage" transformer, we do not wish to saturate the core in the primary portion of the transformer since that would in turn reduce the impedance of the transformer's primary coil. For this reason, the core of a "constant-voltage" transformer (Fig. 6) differs slightly from that of a regulator transformer (Fig. 3). It contains two rods of soft iron separated by a small gap located between the primary and secondary portions of the transformer core. The magnetic attraction between these two rods, in effect, shorts part of the magnetic field produced in the primary portion of the core from the secondary portion of the core and part of the magnetic field produced in the secondary portion of the core from the primary portion of the core. With this shunt, the magnetic field in the secondary portion of the core can be saturated

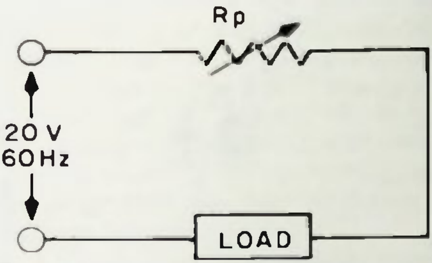


Fig. 10 — High-wattage variable resistors can be used for controling ac currents applied to a load circuit.

without saturating the magnetic field in the primary portion of the core.

The primary and secondary coils in the "constant-voltage" transformer (Fig. 7) are basically the same as those in the regular transformer (Fig. 4). With each half cycle of applied ac primary voltage, the alternating magnetic field that is produced realigns the iron domains in the opposite direction — the change in their alignment inducing nearly enough energy across the capacitor to align all of the domains in the opposite direction on the following half cycle. The magnetic field produced by the applied ac primary voltage is greater than that required to align the balance of the domains, even when this applied ac voltage is less than the normal line voltage. When the ac line voltage is normal, or becomes even greater than normal, the resulting primary magnetic field produced is even greater than what is required on the following half cycle to align all of the domains in the secondary portion of the soft iron core.

Most of the magnetic field produced by the transformer's primary coil travels through the soft iron core to the secondary coil — only a relatively small portion traveling through the air. Since increases in the ac voltage across the primary winding will only increase the small portion of the magnetic field traveling through the air (the secondary portion of the core already being saturated), the voltage developed at the secondary winding is nearly independent of any changes in the ac voltage applied at the primary.

A schematic of the "constant-voltage" transformer is shown in Fig. 8A. If a second coil is wound in the secondary portion of the transformer coil (Fig. 8B), the voltage induced across it will also be affected by the saturation of the magnetic field in the secondary portion of the core. The output voltage across this coil will also remain relatively stable as the ac line voltage at the primary winding changes.

A taped secondary coil (Fig. 8C) will function in the same basic manner as the two secondary coils (Fig. 8B). Measurements indicate that the energy absorbed by a capacitor across this coil is also able to saturate the secondary portion of the "constant-voltage" transformer core, even when a moderate load is applied across the taped portion of the secondary coil.

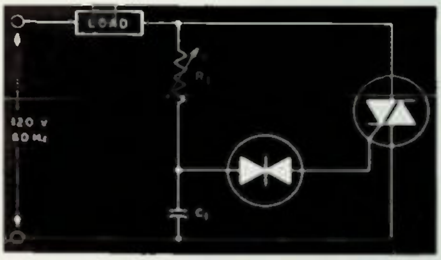


Fig. 11 — A solid-state circuit for controling the ac current applied to a load.

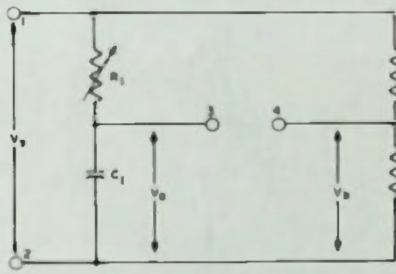


Fig. 12 — The basic capacitor-time-constant circuit in an ac current regulator.

Another ac voltage regulating component used in solid-state power supplies is the thyrector (Fig. 9). Whenever an applied ac voltage excedes its zener breakdown voltage, the semiconductor's internal resistance drops and it conducts sufficient current to load down the applied ac voltage, reducing to normal the voltage across any circuit connected in parallel with the thyrector. The zener breakdown voltage varies according to the particular thyrector type selected.

AC Current Regulation

Early circuits, designed to control the amount of ac current flowing through a circuit to regulate the amount of applied power, merely contained a large variable resistor (Fig. 10). Some of the electrical power was absorbed by the resistor and converted to heat, thereby reducing the amount of power applied to the load.

The average ac current flowing through this circuit is controlled through the use of capacitor time constants. The application of these time constants can be more readily understood with the use of a simplified diagram (Fig. 12) showing a portion of the power regulating circuit.

When terminal 1 is more positive than terminal 2 (Fig. 12), electrons flow from one plate of the capacitor (C_1) through the variable resistor (R_1) and to the positive terminal (1).

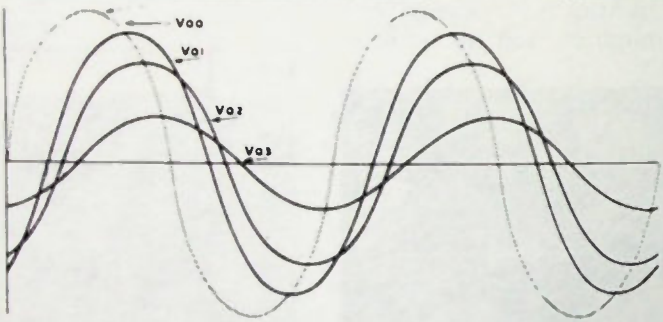


Fig. 13 — The relationship between the supply voltage (V_s) and capacitor voltage (V_a) varies with changes in the series resistance.

There were two major disadvantages of such a circuit. Useful energy was lost in the power-regulating resistor, and the resulting heat had to be removed before it damaged neighboring components.

Solid-state ac current regulators have eliminated these problems. They control the amount of power applied to an ac circuit (motors, lamps, heaters, etc.) with the use of switching diodes, which regulate the average ac current applied. One such ac current regulating circuit is shown in Fig. 11.

The rate of electron flow (amount of current) through the resistor (R_1) determines the rate that a voltage is developed across the capacitor. Since the resistor (R_1) restricts the amount of current into or out of the capacitor (C_1), it (R_1) can be varied to change the rate at which a voltage will be developed across the capacitor. After a sufficient period of time, the voltage across the capacitor (Va) will equal the applied ac voltage (Vs).

If the resistance (R_1) in series with the capacitor (C_1) is reduced to zero, the voltage developed across the capacitor (V_{a0}) is the same as the applied ac voltage (V_s). (The V_{a0} curve would then be the same as the V_s curve in Fig. 13.)

When there is some resistance between the applied ac voltage and the capacitor (C_1), the resistor (R_1) restricts the current to the capacitor (Fig. 12), and the voltage across the capacitor (V_{a1} in Fig. 13) does not increase as rapidly as the applied ac voltage (V_s). The applied ac voltage has already passed its peak by the time the capacitor voltage (V_{a1}) has reached its peak — the instant the two voltages are equal.

By increasing the resistance (R_1), there is a further delay in the voltage developed across the capacitor (V_{a2}), and the applied voltage (V_s) is even further past its peak when the two voltages become equal — resulting in a smaller capacitor voltage than before.

A further increase in the resistance (R_1) results in even a greater time lag in the voltage developed across the capacitor (V_{a3}), and this voltage is even less than before.

From the equivalent triac circuit in chapter 23 (Fig. 12) we can see that any voltage at the gate resulting from leakage current, is not as great as the voltage applied across a triac. Two equal resistors can be used as a voltage divider to supply a voltage corresponding to the voltage present at the gate of some triacs. To simplify the explanation of how a thyrector functions in the ac current regulating circuit (Fig. 11), voltage divider resistors (R_2 and R_3) were substituted for the triac (Fig. 12).

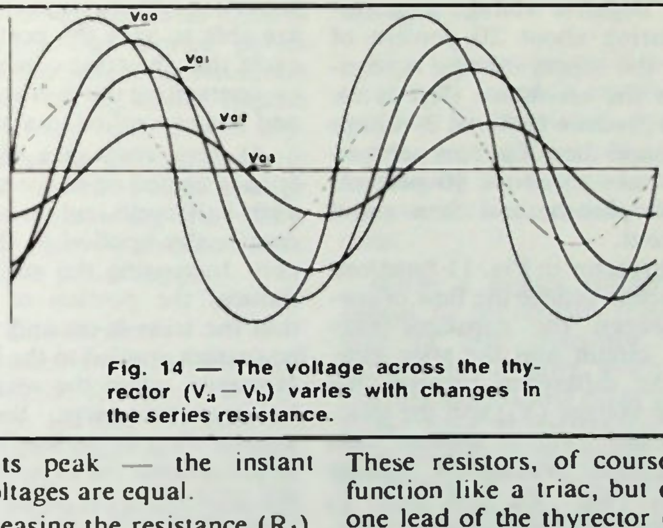


Fig. 14 — The voltage across the thyrector ($V_a - V_b$) varies with changes in the series resistance.

These resistors, of course, do not function like a triac, but do supply one lead of the thyrector (shown in Fig. 12 as just two open terminals) with a voltage (V_b) equivalent to the triac's unloaded gate voltage.

Since resistors R_2 and R_3 are equal, the voltage at terminal 4 (V_b) is half the applied voltage (V_s). This voltage at terminal 4 (V_b) is compared in Fig. 14 with voltages that may be present at terminal 3 (V_a) when there are various resistances (R_1) in series with the capacitor (C_1). (The various curves for V_a are the same in both Fig. 13 and 14. In Fig. 14, V_b has been substituted for V_s, shown in Fig. 13. $V_b = \frac{1}{2} V_s$.)

From Fig. 14 we see that with the larger value of resistor R_1 required for producing voltage V_{a3}, the positive voltage at terminal 3 (V_{a3}) is greater than the positive voltage at terminal 4 (V_b) during about 20 percent of the time that the terminal 4 voltage (V_b) and supply voltage (V_s) are positive. (The negative voltage at terminal 3 is also greater than the negative voltage at terminal 4 during about 20 percent of the time the supply voltage is negative.) As the resistance (R_1) is reduced to produce terminal 3 voltage V_{a2}, V_{a1} and then V_{a0}, this percentage increases to about 40 percent, about 60 percent and then about 100 percent.

The thyrector in Fig. 11 functions as a switch to reduce the flow of current between the capacitor-time-constant circuit and the triac gate. When the difference between the capacitor voltage (V_a) and the triac-gate voltage (V_b) is greater than the thyrector breakdown voltage ($V_{a0} - V_b$, $V_{a1} - V_b$, and $V_{a2} - V_b$ are greater than the breakdown voltage while $V_{a3} - V_b$ in Fig. 14 is not), the thyrector is switched to an ON condition. Current then flows through the thyrector and switches the triac on for the remaining portion of the positive or negative half cycle of the applied voltage (V_s).

The curves in Fig. 13 and 14 show the capacitor voltage (V_a) when no thyrector is included in the circuit. When the thyrector is in the circuit, the current it conducts reduces the capacitor voltage to zero each time the breakdown voltage is exceeded.

By changing the resistance (R_1) in series with the capacitor (C_1), we are able to vary the portion of each cycle that the triac conducts current — controlling the average ac current and power applied to a load circuit.

At zero resistance ($R_1 = 0$), the triac is turned on at the beginning of each half cycle and an ac current is continually applied to the load circuit. Increasing the resistance (R_1) reduces the portion of each cycle that the triac is on and the average ac current applied to the load circuit. However, when the resistance (R_1) becomes too large, the capacitor voltage (V_a) is no longer sufficient to breakdown the thyrector and turn the triac on. This limits the control range of this circuit (Fig. 13).

CHAPTER 25

Power Conversion

Only some of the battery-powered semiconductor circuits can be operated without additional circuits for converting the applied power to the required potential. Receivers plugged into an ac outlet require dc voltages from their power supplies. Batteries for driving portable TV sets do not have high enough voltages for operating the CRT. This voltage must be obtained from an appropriate power supply.

DC-to-AC Conversion

Transformers are generally used in power supplies to increase the applied voltage to that required for operating an electronic circuit. As you know, an ac voltage applied to the primary winding of a transformer will induce an ac voltage across the transformer's secondary winding. A dc voltage applied to this primary winding, however, will not induce either an ac or dc voltage across the transformer's secondary winding. The dc voltage must first be changed to an ac voltage before a transformer can be used in a circuit to increase or decrease a voltage or current.

Oscillators or switching circuits can, of course, perform the function of converting dc voltages to ac voltages.

Fig. 1 shows an oscillator circuit commonly found in inverter power supplies. In this circuit, resistors R_1 and R_2 are used as a voltage divider for biasing the base of both transistors in their common-emitter circuits.

When a supply voltage is connected to the circuit (Fig. 1), the base of one transistor will, by chance, be a little more positive than the base of the other transistor. If the base of the transistor Q_1 is more positive than that of transistor Q_2, transistor Q_1, will conduct more current than the other transistor. Current will then flow through the transformer's primary winding (L_1) between the transistor (Q_1) and the positive supply voltage. This current through winding L_1 will induce a voltage across the other winding (L_2), making the base of transistor Q_2 more positive than the base of transistor Q_1. Transistor Q_2 will then conduct more current through the winding (L_1). Since current is now flowing in the reverse direction

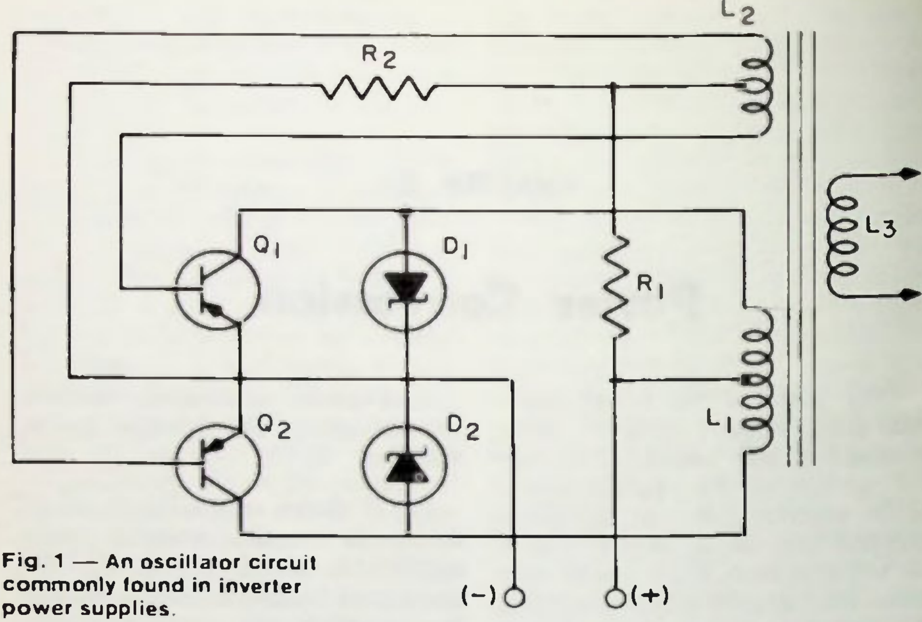


Fig. 1 — An oscillator circuit commonly found in inverter power supplies.

through the transformer's primary winding (L_1), the voltage now induced across the other winding (L_2) is of reverse polarity and the base of transistor Q_1 is now again more positive than the base of transistor Q_2.

If no voltage was induced across winding L_2 then the base voltage of both transistors would be the same, and both transistors would be in a nearly on condition — due to the base voltage from the voltage divider resistors (R_1 and R_2). When a current flows through the primary winding (L_1) from either of the two transistor collector circuits, the resulting voltage induced across the other winding (L_2) makes the base of the non-conducting transistor more positive (turning it on) and the base of the conducting transistor less positive (turning it off).

The conducting transistor turns itself off while turning the other transistor on. The second transistor then conducts current that turns itself off and the first transistor on. In this manner, the circuit oscillates, each transistor alternately conducting current at the transformer's resonant frequency.

As the circuit oscillates, an ac voltage is also induced across another secondary winding (L_3) of the transformer. The voltage across this winding (L_3) is generally used to drive 112vac motors or electronic equipment, using energy from a 12vdc car battery.

Zener diodes are also included in this circuit to extend the life of the transistors by protecting them from surge voltages that occur as one transistor turns off while the other is turning on. During this moment the greatest change occurs in the transformer's magnetic field and

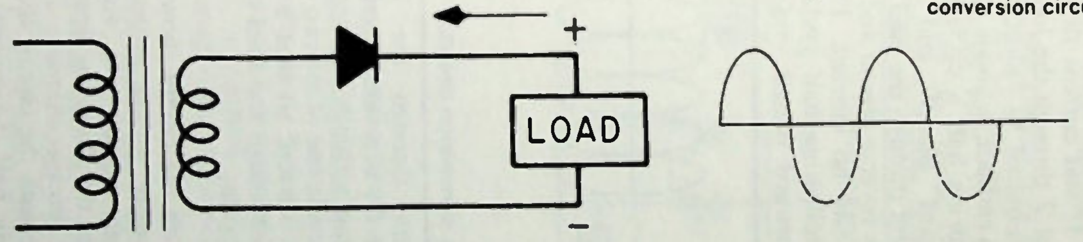


Fig. 2 — A single diode ac-to-dc conversion circuit.

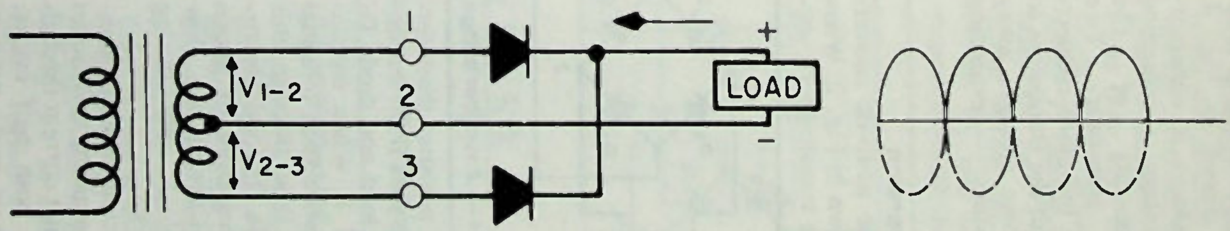


Fig. 3 — A double diode ac-to-dc conversion circuit.

current induced in the primary winding (L_1) by the changing magnetic field (as shown in Fig. 4, chapter 23) produces a collector-to-emitter voltage exceeding the rated capacity of the non-conducting transistor. Two zener diodes are therefore connected in parallel with the transistors to shunt these excess voltages away from the transistors — extending their life.

AC-to-DC Conversion

A single diode ac-to-dc conversion circuit is shown in Fig. 2. When terminal 1 of the transformer's secondary is more positive than terminal 2, electrons pass from terminal 2 through the load and diode D_1 back to terminal 1. When terminal 1 is more negative than terminal 2, no electrons flow through this circuit since diode D_1 conducts virtually no electrons from its anode to its cathode. Current can flow through terminal 1 only during positive half cycles.

A slightly different ac-to-dc conversion circuit containing two diodes (Fig. 3) can conduct current through the load during both half cycles. When terminal 1 is more positive than terminals 2 and 3, electrons flow through terminal 2, the load, diode D_1 and terminal 1 as they did before in the single diode circuit. Then, when the polarity of the applied ac voltage reverses, terminal 3 is more positive than terminals 2 and 1; and current flows from terminal 2 through the load, diode D_2 and terminal 3 — electrons traveling from cathode to anode of diode D_2

During half a cycle diode D_1 is forward biased and conducting current through the load while diode D_2 is reverse biased and conducting virtually no current. Then, during the following half cycle, diode D_2 is forward biased and conducting current through the load while diode D_1 is reverse biased and conducting virtually no current. Each half cycle one of the two diodes conducts current so that a dc current flows through the load during both half cycles.

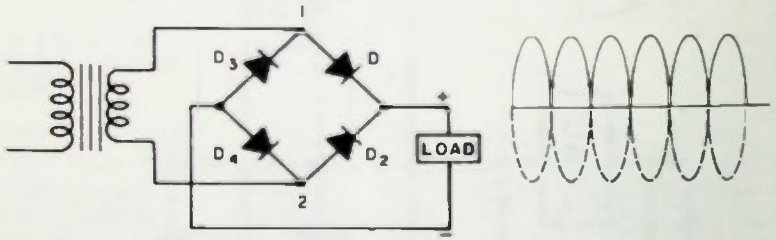


Fig. 4 — A rectifier bridge ac-to-dc conversion circuit.

With the circuit shown in Fig. 2 the entire voltage across the transformer, minus the voltage drop across the diode, is applied to the load during alternate half cycles; while with the circuit shown in Fig. 3 only half the voltage across the transformer ($V_{1-2} = V_{2-3} = \frac{1}{2} V_{1-3}$), minus the voltage drop across the diodes, is applied to the load during every half cycle.

Still another circuit (Fig. 4), containing four diodes to form a rectifier bridge, can be used to apply most of the voltage across the transformer to the load during every half cycle.

Diodes D_1 and D_2 function in this circuit (Fig. 4) as they did in the previous circuit (Fig. 3). During each half cycle one diode or the other conducts electrons from its

D_1 and terminal 1 — the electrons having flowed from cathode to anode through both diodes. Then, when the polarity of the applied ac voltage reverses, terminal 2 is more positive than terminal 1, and electrons flow from terminal 1 through diode D_3, the load, diode D_2 and terminal 2. During one half cycle diodes D_1 and D_4 are conducting while during the other half cycle diodes D_2 and D_3 are conducting current through the load.

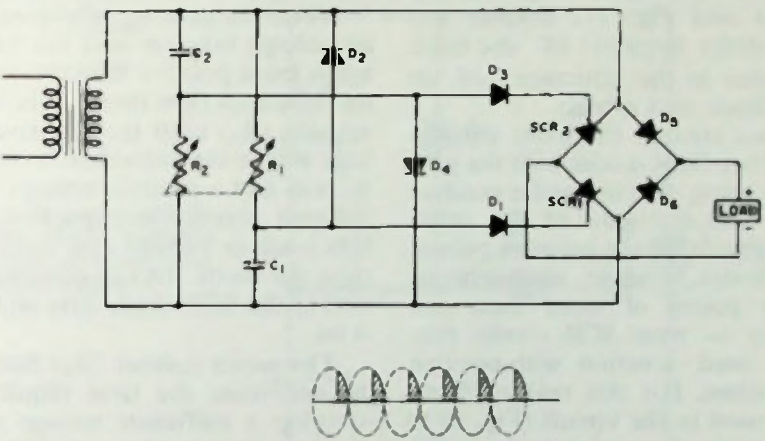


Fig. 5 — A variable output, rectifier bridge ac-to-dc conversion circuit.

cathode to its anode and the transformer — diode d_1 conducting when terminal 1 is positive and diode D_2 conducting when terminal 2 is positive.

Diodes D_3 and D_4 are wired in the circuit so that their anodes, rather than their cathodes, are connected to the load. When terminal 1 is more positive than terminal 2, electrons flow from terminal 2 through diode D_4, the load, diode

AC to Varied DC

Silicon controlled rectifiers, SCR_1 and SCR_2 (Fig. 5), can be substituted for two of the diodes, D_3 and D_4, in the rectifier bridge circuit (Fig. 4) to vary the average dc voltage applied across a load. This voltage is controlled by trigger voltages applied to the SCR diode gates.

Fig. 11 in chapter 24 explains how capacitor time constants (also used in the circuit shown in Fig. 5 of this chapter) can be used for producing a voltage to trigger the gate of a triac. Changing the resistance (R_1) in series with the capacitor (C_1) varies the portion of each cycle that the triac conducts current. During each half cycle the voltage developed across the capacitor (C_1) is discharged through the thyrector and triac.

A slightly different circuit (Fig.6) can vary the portion of each cycle a single SCR diode conducts current. These two circuits (Fig. 6, this chapter and Fig. 11, chapter 24) must differ because of the basic difference in the structure of an SCR diode and a triac.

Triacs contain the basic equivalent of two SCR diodes with the gate of one being able to handle positive pulses and the gate of the other being able to handle negative pulses. SCR diodes, however, are unable to handle pulses of more than one polarity — most SCR diodes presently used function with positive gate pulses. For this reason, diode D_1 is used in the circuit (Fig. 6) to prevent the flow of any damaging negative current through the SCR diode gate.

During the half cycle when the load (Fig. 6) is more positive than the capacitor (C_1), electrons flow from the capacitor through the variable resistor (R_1) to the load. As the capacitor's positive charge becomes greater than the small positive voltage at the SCR diode gate, current flows through diode D_1. This gate current turns the SCR diode on, causing the SCR to also conduct current for the remainder of the positive half cycle.

When the positive capacitor voltage becomes greater than the SCR gate voltage, diode D_1 conducts current because its anode is more positive than its cathode, while diode D_2 does not conduct current since its cathode is still more positive than its anode.

During the following half cycle the load is more negative than the capacitor (C_1). The cathode of diode D_2 is then more negative than its anode and the diode conducts current, the flow of electrons resulting in a negative capacitor voltage.

When the polarity of the applied ac voltage reverses and the load is again more positive than the capacitor, electrons flow through the series resistor (R_1) until the negative voltage across the capacitor is reduced to zero and a positive voltage is developed across the capacitor. It is this positive voltage that again permits the diode (D_1) to conduct current to the SCR diode gate and turn it on.

The series resistor (R_1) functions to determine the time required to develop a sufficient voltage across the capacitor to trigger the thyrector and turn the triac on (Fig. 11 in chapter 24). It (R_1) also functions to determine the time required to reduce the negative voltage across the capacitor (C_1) to zero and then to increase the positive capacitor voltage to that required to turn the SCR diode on (Fig. 6). Because the thyrector requires a higher triggering voltage from the capacitor and diode D_2 provides a negative capacitor voltage that has to be removed before the capacitor can develop a positive voltage, a larger than normal flow of electrons is required to trigger the triac or SCR diode; and the series resistor (R_1) has greater range and stability for adjusting the capacitor time constant. With these circuits the triac and SCR diode can be turned on during almost any portion of the half cycle.

The trigger voltages for the two silicon controlled rectifiers, SCR_1 and SCR_2, in the rectifier bridge circuit (Fig. 5) can be obtained in the same manner as the trigger voltages obtained for the single SCR diode (Fig. 6). One diode (SCR_1) is controlled as before by resistor R_1, capacitor C_1, and diodes D_1 and D_2; while the other diode (SCR_2) is controlled by a similar complement of components — resistor R_2, capacitor C_2, and diodes D_3 and D_4. During one half cycle diode D_1 supplies a positive triggering current to the gate of one diode (SCR_1) and during the following half cycle diode D_3 supplies a positive triggering current to the gate of the other diode (SCR_2).

Just as diodes D_3 and D_4 conduct current in their rectifier bridge circuit (Fig. 4) during alternate half cycles, diodes SCR_1 and SCR_2 also conduct current in their rectifier bridge circuit (Fig. 5) during alternate half cycles. However, diodes D_3 and D_4 must conduct current through their load during each complete half cycle, while a single control connected to a pair of ganged resistors (R_1 and R_2) can be used to vary the portion of each cycle that the two SCR diodes conduct current through their load. Since the SCR diodes are either conducting current or not conducting, like a switch turned on or off, very little power is consumed in reducing the load voltage.

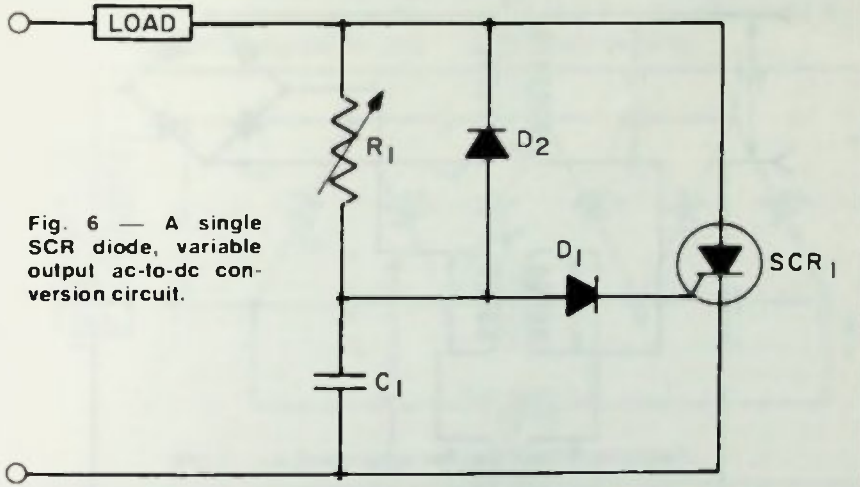


Fig. 6 — A single SCR diode, variable output ac-to-dc conversion circuit.

AC to Regulated DC

Transformers (T_1, T_2 and T_3 in Fig. 7) can be used to regulate a rectifier bridge circuit — like the one described in Fig. 4 — so that when the load draws more current, the voltage across the load increases to compensate for the additional current drain.

In this circuit (Fig. 7), transformers T_1 and T_2 are matched so that they produce nearly identical secondary ac voltages — but 180deg out of phase (when v_1 is positive, v_2 is negative and vice versa). The secondary voltages therefore tend to cancel out ($v_T = v_1 - V_2 = O$), adding no ac voltage to the load or rectifier bridge circuit.

When there is no dc current through the secondary windings, the ac current through both primary windings induces alternating magnetic fields that align almost all the magnetic domains first in one direction and then in the opposite direction during each cycle.

Any dc current through both secondary windings will induce magnetic fields that will work to align the magnetic domains in but one direction. During half a cycle the

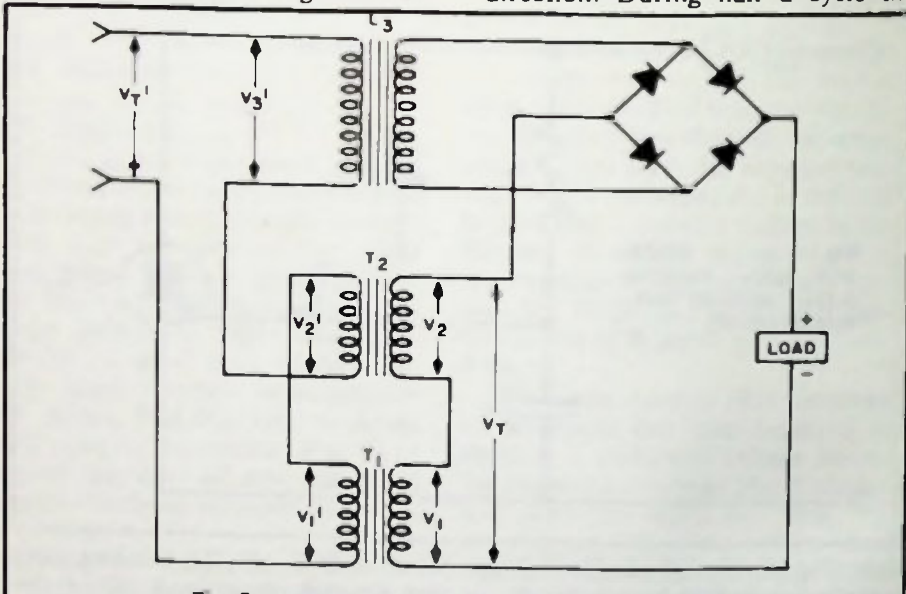


Fig. 7 — An ac to regulated dc conversion circuit.

The ac primary current through the transformers (T_1 and T_2) produces alternating magnetic fields in their cores, while the magnetic fields produced there by the dc secondary current do not alternate. These primary and secondary magnetic fields affect the alignment of magnetic domains in the soft iron core of each transformer (T_1 and T_2).

dc induced magnetic fields will work with the ac induced magnetic fields to align domains in one direction, and during the following half cycle the dc induced magnetic fields will oppose the ac induced magnetic fields' work to realign domains in the opposite direction.

When the dc secondary current induces a magnetic field to align some of the domains, the magnetic field induced by the ac primary current is greater than that required to align the remaining domains in that direction. The transformer cores contain only a limited number of domains to be aligned. During the following half cycle, when the flow of ac primary current is reversed, the number of domains aligned in the reverse direction is reduced by the dc current, which is still producing a magnetic field to align the domains in the initial direction.

With larger dc secondary currents, there is virtually no increase in the number of domains aligned in one direction — there are no more to align — and the number of domains aligned in the opposite direction the following half cycle is reduced — the larger magnetic field induced by the dc secondary current is opposing that induced by the ac primary current. In this manner, the strength of the dc secondary current affects the number of magnetic domains that alternate directions as a result of the ac primary current.

Earlier chapters indicated that a soft iron core can be used to concentrate the magnetic field produced in a coil. The alignment of the soft iron core's domains determines the strength of the core's magnetic field. As the alignment of the core's magnetic domains alternates as a result of the ac primary current, the core's magnetic field also alternates. The strength of the core's alternating magnetic field is greater when there is a small secondary dc current and smaller when there is a larger secondary dc current, since the secondary current can reduce the number of domains that can be alternated by the ac primary current.

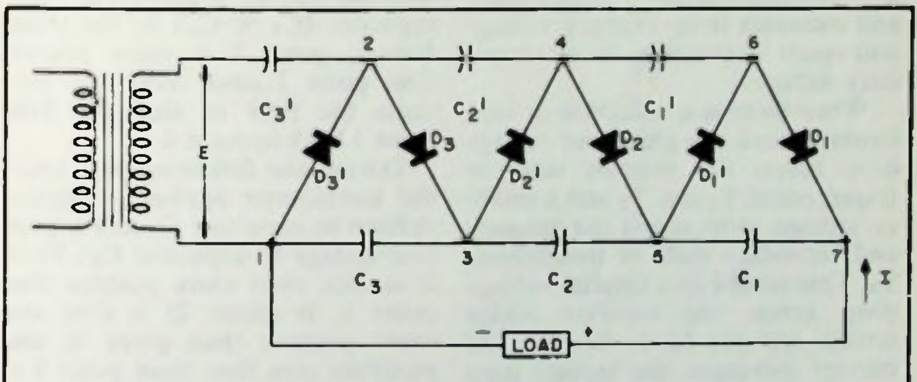


Fig. 8 — A three-stage voltage multiplier circuit.

In transformers T_1 and T_2 the primary ac current induced by the cores' alternating magnetic fields opposes the applied ac primary current. The greater the cores' alternating magnetic field, the greater this resistance to the applied ac primary current — the greater the impedance of the primary coils. In this manner, the dc secondary current can control the ac voltage drop across the primary windings of transformers T_1 and T_2. The greater the dc secondary current, the smaller the cores' alternating magnetic fields and the smaller the coil impedances and the ac voltage drop across the primary coils. The smaller the dc secondary current, the larger the cores' alternating magnetic fields and the larger the coil impedances and the ac voltage drop across the primary coils.

Since the applied ac voltage remains unchanged, a reduction in the voltage drop across the primary coils of transformers T_1 and T_2 will result in an increase in the voltage drop across transformer T_3 [$v_3 = Vt - (v_1 + v_2)$]. Transformer T_3 is not a "constant-voltage" transformer and increases in its primary voltage will result in increases in its secondary voltage.

When there is a reduction in load current, there is a greater ac voltage drop across the primary coils of transformers T_1 and T_2 and a smaller voltage drop across the primary and secondary coils of transformer T_3. This results in a smaller voltage drop across the rectifier bridge circuit and the load. As the load current increases, the voltage drop across the primary coils of transformers T_1 and T_2 decreases and the voltage drop across the primary and secondary coils of transformer T_3 increases. This results in a greater voltage across the rectifier bridge circuit and the load. More current can be conducted through the load as the voltage across it increases.

Voltage Multipliers

The ac-to-dc conversion circuits that have been described in this chapter produce dc voltages smaller than the ac voltages applied to the circuits. Voltage multipliers, however, are able to supply dc voltages greater than the ac voltage applied to the circuit — with a corresponding reduction in current.

During the half cycle when the transformer (Fig. 8) applies a positive voltage to capacitor C_3 and a negative voltage to capacitor C_3', electrons flow through diode D_3' — electrons flow from negative to positive, from the diode's cathode to its anode. No electrons are then flowing through diode D_3 since it is reverse biased.

During the portion of the ac cycle when no voltage is applied to either capacitor (C_3 or C_3') by the transformer, point 2 is more positive than point 1 since diode D_3' prevents the flow of electrons from point 1 back to point 2.

During the following half cycle, the transformer applies a negative voltage to capacitor C_3 and a positive voltage to capacitor C_3'. Point 2 is then even more positive than point 1. It (point 2) is then also more positive than point 3, and electrons now flow from point 3 to point 2 through diode D_3.

During the next portion of the ac cycle when no voltage is applied to either capacitor (C_3 or C_3'), point 3 is more positive than point 2, which is still more positive than point 1.

This mechanism not only develops a dc voltage across capacitors C_3 and C_3', but in the same manner also develops dc voltages across capacitors C_2', C_2, C_1' and C_1. In this circuit, the voltage across the load is equal to the total voltage across capacitors C_1, C_2 and C_3 ($v_T = v_{C1} + v_{C2} + v_{C3}$).

With the components used in this circuit (Fig. 8), 50vac from the transformer results in 270vdc across the load, when 5ma of current flows through the load. Fewer stages of the voltage multiplier would result in a smaller dc voltage, while more stages would result in a larger dc voltage. The amount of dc voltage produced can be approximated with the equation:

$$V = 2NE - \left(\frac{2}{3}N^3\right)\left(\frac{I}{fC}\right);$$

when the value of the capacitors in each stage increases so that $C_N = NC_1$, and N equals the number of stages. There are three stages in the circuit shown in Fig. 8; and $C_2 = 2C_1 = 2 \times 50\mu f = 100\mu f$, $C_3 = 3C_1 = 3 \times 50\mu f = 150\mu f$; while $V = 2 \times 3 \times$

$$50 - \left(\frac{2}{3} \times 3^3\right)\left(\frac{5 \times 10^{-3}}{60 \times 50 \times 10^{-6}}\right) = 300 - 30 = 270v.$$

From the equation we can see that if the load current is increased, the dc voltage is reduced, while if the frequency of the applied ac voltage increases or the values of the circuit capacitors increases, the voltage increases.

CHAPTER 26

Filters & Regulators

Single or multiple diode rectifier bridge circuits alone are not capable of providing dc voltages and currents suitable for use in most solid-state electronic circuits. The curves showing the dc voltages produced by a single diode rectifier circuit (Fig. 2 in chapter 25) and multiple diode rectifier circuits (Figs 3, 4, and 5 in chapter 25) indicate that these dc voltages are not constant. The dc voltage fluctuations (called ripple) occur at either the frequency of the applied ac voltage in a single diode rectifier circuit or at twice that frequency in most multiple diode rectifier circuits. Ripple can produce hum in amplifier circuits, cause errors in counting and switching circuits, and even damage some semiconductors.

Filters

A single capacitor (Fig. 1B) can be used to reduce the ripple present at the output of an ac-to-dc converter (Fig. 1A). The greater its capacitance, the greater the number of electrons absorbed and released by the capacitor to reduce the ripple.

Many diodes used in rectifier bridge circuits have a small negative resistance and oscillate at high frequencies. This has occasionally caused picture interference in solid-state TV sets. Unfortunately some large value capacitors, despite theory, are unable to filter these high frequencies. It is then necessary to shunt these large-value capacitors with small-value capacitors to filter out the high-frequency oscillations.

When a resistor (R_1) and capacitor (C_2) are added to the filter circuit (Fig. 1C), the voltage developed across capacitor C_2 is smaller than that developed

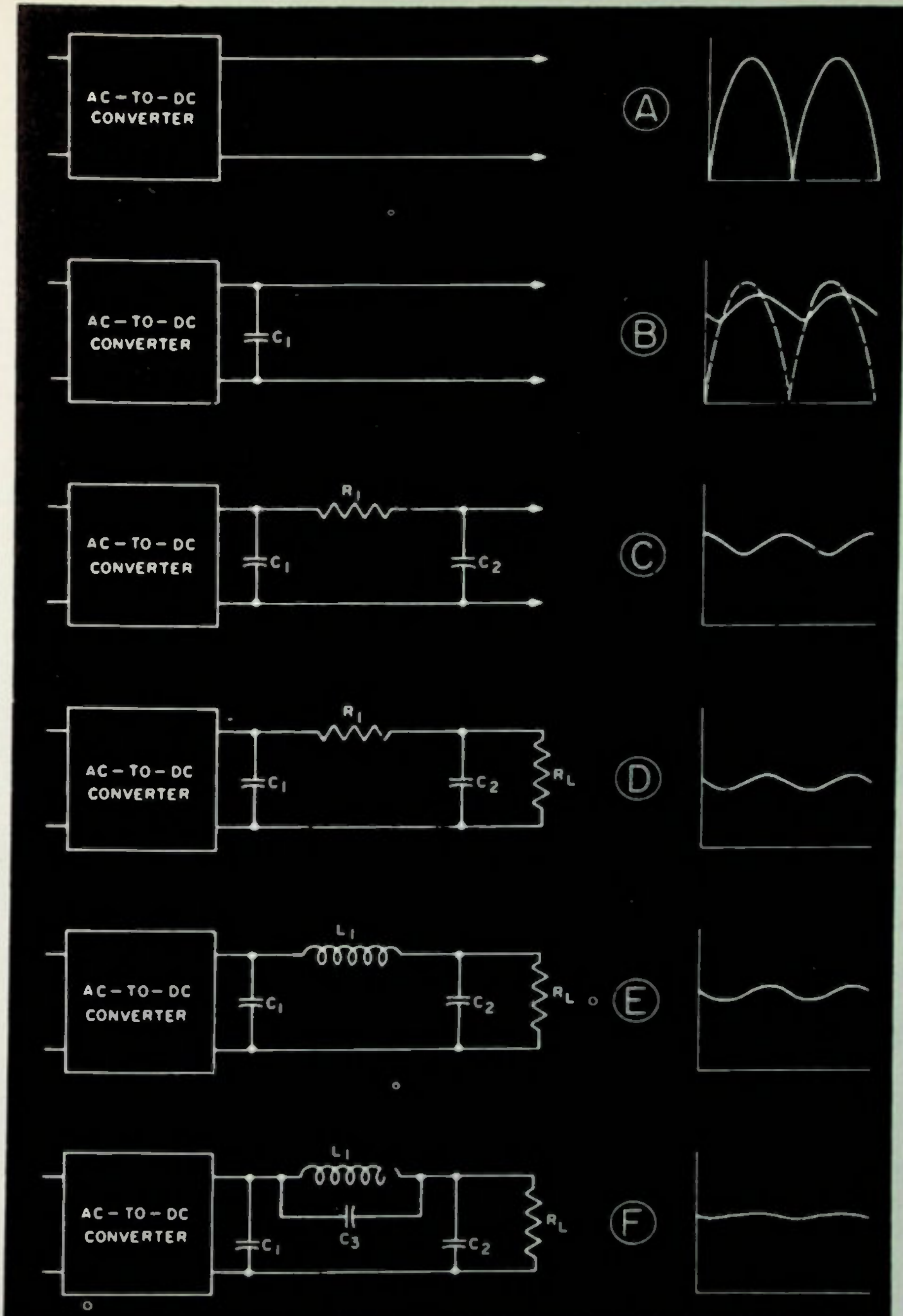


Fig. 1(A)—The unfiltered dc output voltage from an ac-to-dc converter; (B) this voltage after it has been filtered by a single capacitor; (C) as it appears across a second capacitor; (D) across the load of an R-C filter circuit; (E) across the load of a brute-force filter circuit; and (F) across the load of a resonant-filter circuit.

across capacitor C_1, and these voltages are out of phase. The additional resistor (R_1) and capacitor (C_2) in this R-C filter circuit (Fig. 1C) function like resistor R_1 and capacitor C_1 in the circuit shown in Fig. 12 in chapter 24 — the change in phase angles and AC voltages being like that shown in Fig. 13 of chapter 24. The load resistor (R_L) draws current (Fig. 1D) from both capacitors (C_1 and C_2); and the ripple present in the voltage drop across the load resistor is less than the ripple voltage previously present across each capacitor, since capacitor ripple currents are out of phase and tend to cancel each other.

In chapter 15 we described how coil impedances reduce ac currents. Only the resistance of the wire coil windings reduces the dc current. A choke (L_1), which is a coil with a high ac resistance and a low dc resistance, can be used (Fig. 1E) to produce a greater reduction in ripple current without also causing a greater reduction in dc current. As a result of this brute force filter circuit, the dc voltage drop across the load resistor (R_L) is greater with less ripple.

The impedance of the choke (L_1) used in this circuit (Fig. 1E) must not equal the impedance of the second capacitor (C_2) or they will form a series resonant circuit, and the ripple voltage across the capacitor (C_2) and load (R_L) will be even greater than that across the first capacitor (C_1).

A choke (L_1) and capacitor (C_1) can be selected (Fig. 1F) that have equal impedances at the ripple frequency. Together they will act as a parallel resonant circuit, and practically no ripple current is conducted. This resonant-filter circuit causes a greater reduction in ripple that the R-C filter circuit of the brute force filter.

The common-base amplifier circuits described in chapter 1, Fig. 10, and then converted to a more conventional form in Fig. 4, chapter 3, were used to amplify signals. A small signal applied between the transistor's base and emitter resulted in a larger signal between its base and collector. Conversely, a signal applied between the base and collector in a common base circuit (Fig. 2) will result

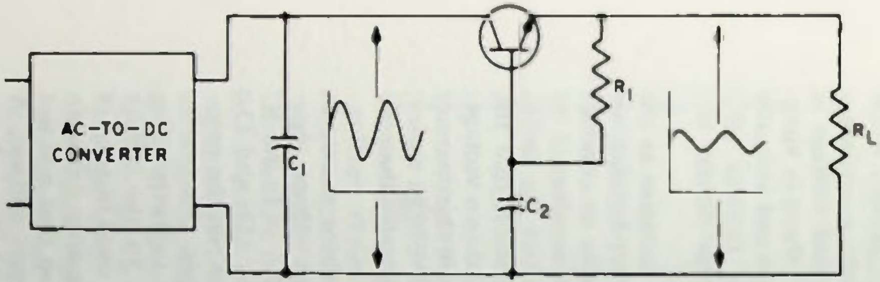


Fig. 2—An electronic filter circuit.

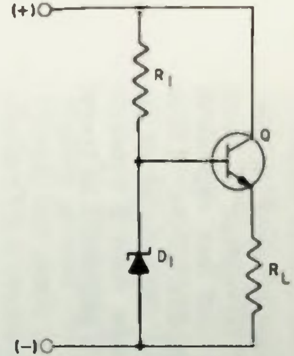


Fig. 4—A self-compensating transistor dc-voltage-regulating circuit.

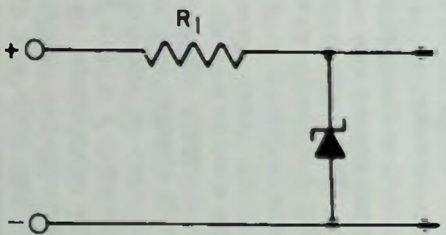


Fig. 3—A zener-diode dc-voltage-regulating circuit.

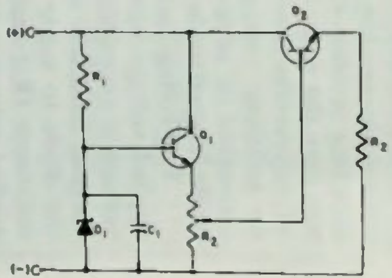


Fig. 5—A variable dc-voltage-regulating circuit.

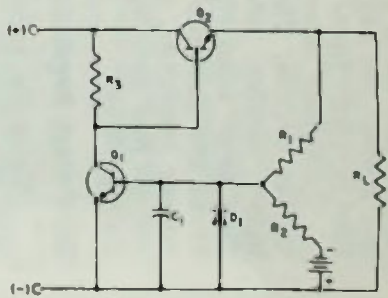


Fig. 6—A dc voltage matching circuit.

in a smaller signal between the base and emitter. In this manner, the circuit can function as an electronic filter to reduce ripple.

Electronic filters are frequently used in low-voltage circuits. They have the advantage of being lighter, smaller, more efficient and less expensive than large capacitors and inductive filters. These filters are generally designed in semiconductor dc voltage regulators, which maintain such a constant voltage that virtually no ripple is permitted to travel through them.

DC Voltage Regulation

About the simplest solid-state voltage regulator (Fig. 3) consists merely of a zener diode (described in Fig. 7, chapter 22) and a voltage-dropping resistor (R_1). When the applied voltage is less than the zener-breakdown voltage, there is virtually no current flowing through the diode and virtually no resulting voltage drop across the resistor (R_1) connected in series with it—the voltage across the diode is equal to the applied voltage.

The voltage across the diode will increase as the applied voltage increases, until the zener-breakdown voltage is exceeded and the diode begins to conduct current. Sufficient current will then be conducted by the diode to cause a voltage drop across the series resistance (R_1) that is equal to the amount that the applied voltage exceeds the zener-breakdown voltage. The voltage across the zener diode can in this manner remain virtually constant as the applied voltage varies, as long as the applied voltage remains greater than the zener-breakdown voltage.

The single transistor voltage regulator shown in Fig. 4 is similar to the integrated circuit voltage regulator described in Fig. 13, chapter 12. In the IC circuit two groups of diodes (D_1 and D_2) are used to provide a voltage drop between the negative supply voltage and transistor base that varies with changes in temperature, to offset corresponding changes in transistor characteristics. In the circuit shown in Fig. 4, a zener diode (D_1) is used instead of the other diodes to maintain a voltage drop between the negative supply and transistor base that will not change with any changes in the supply voltage. A temperature-compensating diode could have been used in place of the zener diode to supply a nearly constant voltage drop that would change slightly with temperature to match transistor characteristics.

When a constant dc voltage is applied to the circuit (Fig. 4), it will function like the IC voltage regulator, and with even moderate changes in load resistance (R_L), the voltage drop across the load remains virtually unchanged.

An increase in supply voltage will result in a slightly greater voltage drop across the load resistor (V_{RL}), but since the voltage drop across the zener diode (V_{D1}) remains unchanged, the base-to-emitter voltage (V_{BE}) is reduced ($V_{BE} = V_{D1} - V_{RL}$). The transistor base is, therefore, less forward biased, and with the resulting reduction in collector current, the

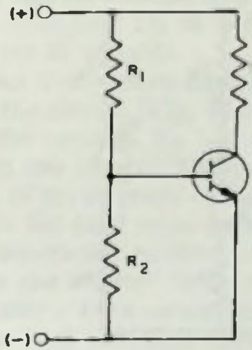


Fig. 7 — A simple dc-current-regulating circuit.

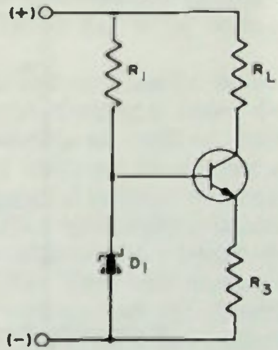


Fig. 8 — A self-compensating dc-current-regulating circuit.

voltage drop across the load resistor is reduced to approximately what it was with the lower supply voltage. In this manner, moderate changes in supply voltage results in only very minor changes in the voltage drop across the load resistance — the circuit has kept it virtually constant.

A two-transistor voltage regulator is shown in Fig. 5. There transistor Q_1 functions as it did in the circuit shown in Fig. 4, providing a virtually constant voltage drop across resistor R_2. With a capacitor (C_1) connected in parallel with the zener diode (D_1), the transistor (Q_1) also acts as an electronic filter, functioning as a reversed common-base amplifier like in Fig. 2.

Since the voltage drop across resistor R_2 remains virtually constant, any tapped voltage from that resistor is also regulated and can be used for biasing the base of transistor Q_2. That transistor (Q_2) then functions in the same manner described for transistor Q_1 in Fig. 4. Transistor Q_2 supplies a regulated voltage across the load resistor (R_L)—that voltage being determined by the setting of the resistor (R_2) tap.

Many power supplies on the market, particularly the solid-state high-voltage ones, regulate the output voltage by comparing it with a reference voltage. For simplicity the circuit shown in Fig. 6 uses a battery to supply the reference voltage, while many circuits operating on this principle use a separate zener-diode regulated power supply to provide this voltage.

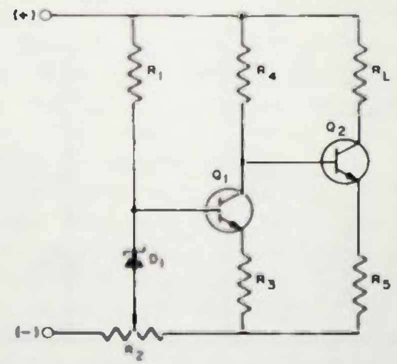


Fig. 9 — A variable dc-current-regulating circuit.

Currents from the positive and negative voltage sources pass through resistors R_1 and R_2; and when the voltages are equal, they cancel out and no bias voltage is present at the base of transistor Q_1.

When the voltage from the power supply is greater than the battery reference voltage, the positive current is greater than the negative current and the base of transistor Q_1 is forward biased. It (Q_1) then conducts current, causing a voltage drop across its collector resistor (R_3). This voltage drop reduces the forward bias at the base of transistor Q_2; and it conducts less current, reducing the voltage at the output of the power supply. In this manner, transistors Q_1 and Q_2 function to keep the power supply voltage from becoming significantly greater than the battery reference voltage.

Should there be a failure resulting in the absence of voltage from the power supply but not the battery, diode D_1 would short-circuit the battery, protecting transistor Q_1 from any reverse bias voltage that would damage it.

Capacitor C_1 is used in this voltage regulating circuit, like the others, to remove ripple.

DC Current Regulation

The single transistor current-regulating circuit shown in Fig. 7 operates on the same principle as the integrated circuit current regulator described in Fig. 5, chapter 9 (resistors R_1 and R_2 are the same in both circuits; load resistor R_L is substituted for transistor Q_1 in the IC circuit; and for convenience of polarity, PNP transistor Q_1 is substituted for NPN transistor Q_2 in the IC circuit).

An even more constant current can be obtained from the circuit (Fig. 8) by substituting a zener diode (D_1) for resistor R_2 and inserting an emitter resistor (R_3) in the circuit. Then any increase in current as a result of an increase in the applied voltage or a reduction in the load resistance (R_L) will slightly increase the transistor's emitter current and the voltage drop across the emitter resistor (R_3). This will reduce the transistor's base-to-emitter voltage and the resulting emitter and collector currents, keeping the collector current nearly constant.

The operation of transistor Q_1 in the circuit shown in Fig. 9 is basically the same as in the previous circuit (Fig. 8). A tapped resistor (R_2) has been added, however, so that the base bias voltage can be adjusted to select the constant collector current desired. Since transistor Q_1 has a nearly constant collector current, the voltage drop across resistor R_4 is also nearly constant, supplying a regulated voltage to the base of transistor Q_2.

Transistor Q_2 also operates like transistor Q_1 in Fig. 8, providing a nearly constant regulated current to its load resistor (R_L).

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