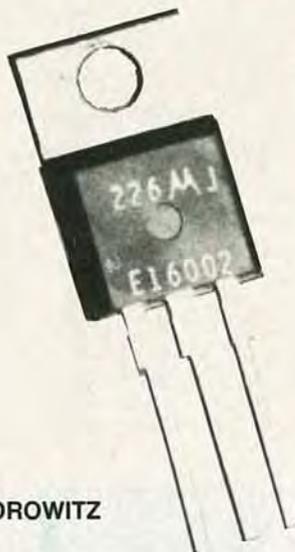


# How to Design Analog Circuits —Multi-Stage Amplifiers



MANNIE HOROWITZ

*For most applications, you'll need more gain than you can get from a one-stage amplifier. This month, we'll show you how to combine one or more amplifier stages to get a more useful circuit.*

A SINGLE STAGE OF AMPLIFICATION MAY provide enough gain for some applications, but rarely, if ever, is the gain from a single-stage amplifier sufficient for audio or RF circuits. For those applications, the output from one amplifier stage must be fed to one or more subsequent stages to obtain enough gain. If that is done, the voltage gain of the first stage is multiplied by the voltage gain of the succeeding stages to determine the overall gain of the entire circuit.

Let's look at some of the methods that are often used to connect several amplifier stages together. Transformer coupling was, at one time, used quite often with audio amplifiers—especially in the power-output stages. Although it is still used for that, transformer coupling is now used mainly in IF and RF circuits.

Another method of coupling is through an R-C network. In that type of circuit, the output of one stage is connected to the input of another through a capacitor. In the previous article in this series (see the November 1982 issue of **Radio-Electronics**), we indicated that a capaci-

tor is used to couple a signal from its source to the input of an amplifier to isolate the source from the bias circuit. If the signal source and the bias circuit were not isolated in that fashion, the bias on the transistor would be altered by the signal source. That will also occur in a two-stage (or more) coupled circuit, but with an additional factor—in a circuit of that type, the collector voltage at the first transistor would affect the base current of the second. In turn, the current demanded by the base circuit of the second would affect the collector voltage of the first.

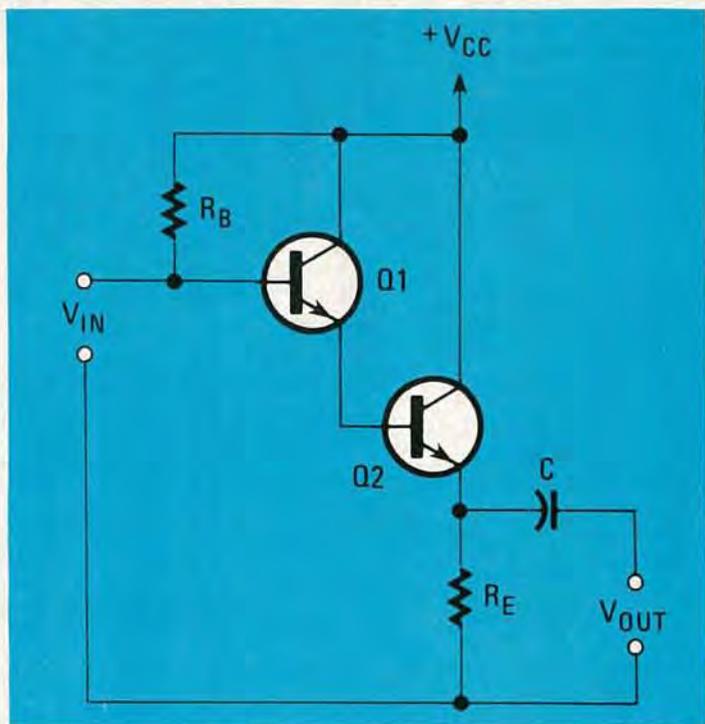
But that interaction can be accounted for and the circuit can be designed so that the use of a capacitor between the two stages is not required. That type of circuit is referred to as a direct-coupled amplifier and there are a number of advantages to using that arrangement, including better low-frequency response.

The circuits we will describe in this article all use bipolar transistors. Many of the problems encountered are not as severe in FET circuits. Thus, once you have mastered the design procedures for

bipolar-transistor circuits, applying what you've learned to FET circuits is relatively easy.

## Transformer-coupled circuits

The transformer has characteristics that make it useful as a coupling device. As you know, a simple transformer is made up of two coils of wire wound around a common magnetic-core. For the rest of this discussion, we will assume that one coil, called the primary, consists of  $N_1$  turns of wire, while the other coil, called the secondary, consists of  $N_2$  turns of wire. If a DC current were fed to the primary, a steady magnetic field would be induced in the primary and coupled through the core to the secondary, but no current would be induced in the secondary. If an AC current were fed to the primary, however, an AC current would be induced in the secondary. The signal in the secondary would have the same frequency and vary identically to the signal fed to the primary (assuming an ideal transformer, of course). If there were perfect magnetic coupling between the two



coils, the current appearing in the secondary ( $I_2$ ) due to the input current ( $I_1$ ) would be inversely proportional to the ratio of the turns in the coils, or:

$$\frac{I_2}{I_1} = \frac{N_1}{N_2}$$

The ratio of the voltages across the two coils is directly proportional to the turns ratio, or:

$$\frac{V_1}{V_2} = \frac{N_1}{N_2}$$

Since impedance,  $Z$ , is equal to the product of  $V$  and  $I$ , by multiplying equation 1 by equation 2, we come up with the fact that the impedance ratio is proportional to the turns ratio squared, or:

$$\frac{Z_1}{Z_2} = \left(\frac{N_1}{N_2}\right)^2 = \left(\frac{N_1}{N_2}\right)^2$$

That equation states that an impedance in the secondary,  $Z_2$ , will appear reflected into the primary as an impedance,  $Z_1'$ , equal to  $Z_2$  multiplied by  $(N_1/N_2)^2$ , the square of the turns ratio.

Now, let's turn to a practical design example. Let us say that we have a ceramic phonograph cartridge that must see a resistance of at least 22,000 ohms. The average output voltage from the cartridge is 0.25 volt. After amplification, however,  $\frac{1}{2}$ -watt must be delivered to an 8-ohm loudspeaker at the output. Transformer coupling is to be used between the amplifier stages and between the output stage and the loudspeaker. To do that, we'll use the circuit shown in Fig. 1. Assume that the beta of both Q1 and Q2 is 100.

If the transformer is 60% efficient, at least  $0.5/0.6 = 0.833$  watt must be available from Q2. However, there are other factors such as transistor leakage, losses due to the emitter resistor, and so on. Considering those potential losses, we should be safe if we design the circuit to deliver 1 watt.

To start our design, we must first draw the load lines for output transistor Q2. When we draw the load lines, we must take into account the effect of the transformer in the collector circuit of Q2. Since the primary of the transformer is an inductor, it will present a DC resistance, equal to the resistance of the wire in the coil, but a different AC impedance. Because of that, two load lines, one DC and one AC, must be plotted on the same graph. The supply voltage is specified to be 9 volts.

Let's assume that the resistance of the primary winding of T2 is very low and just about equal to zero. We will also assume that  $R_{E2}$  is very small and let it equal zero. Making those assumptions, the vertical line in Fig. 2 is the DC load

line. (To make the graph clearer, the transistor's characteristic curves have been omitted here.) That is, of course, determined by the equation for the load line,  $V_{CC} = I_C(R_P + R_{E2}) + V_{CE}$ . Since  $R_P$ , the resistance of the primary winding of the transformer, and  $R_{E2}$  were both assumed to be zero,  $V_{CC} = V_{CE}$  for all values of collector current  $I_C$ . That, of course, is the equation of a straight line.

The AC impedance is  $R_L$  reflected into the primary of T2 as a resistance  $R_L'$ . Before we go any further, let's take another look at Q2's collector circuit because something interesting happens here. First, as we said before, the primary of the transformer is an inductor. Just as a capacitor opposes sudden changes in voltage, an inductor opposes sudden changes in current. We will show how the quiescent collector current,  $I_{CQ}$ , is chosen shortly, but for now let's just say that it will be chosen to allow for equal swings between 0 and  $I_{Cmax}$ . With no input-signal applied to the transistor's base,  $I_{CQ}$  (the quiescent collector current) flows through Q2's collector circuit and through the primary of T2. When an AC

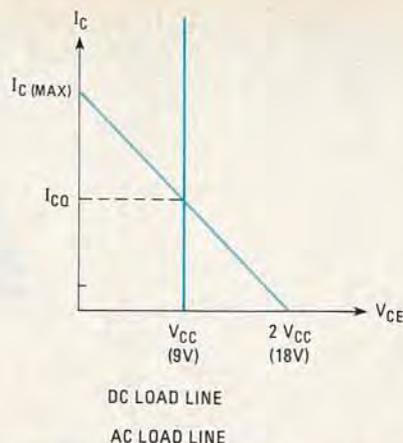


FIG. 2—THE LOAD LINES for the transistors of the circuit shown in Fig. 1 are shown here.

swing between 0 and 18 volts. Plot 18 volts on the  $V_{CE}$  axis; that determines  $V_{Cmax}$  on the load line.

The amount of power delivered from the transistor, 1 watt, determines the  $I_{Cmax}$  point on the load line. Because the signal swings from the transistor's quiescent current point,  $I_{CQ}$ , to a peak current of  $I_{Cmax}$ , and then to a minimum current

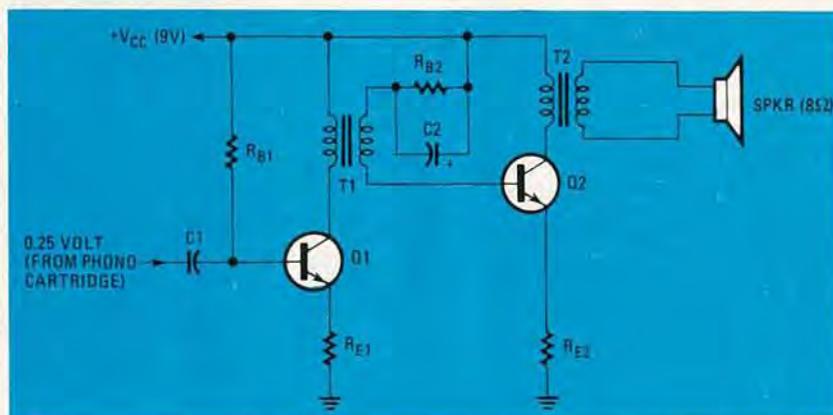


FIG. 1—THIS BASIC TRANSFORMER-COUPLED CIRCUIT uses transformers to couple the two amplifier stages, and to couple the entire circuit to the load, in this case a loudspeaker.

signal is applied to the transistor, however, things begin to change.

Let's first consider what happens when the AC-input swings negative. When that happens, the transistor tries to decrease the collector current by increasing its collector-to-emitter resistance. The inductance of T2's primary, however, opposes the change in  $I_C$  and tries to maintain the flow of current at  $I_{CQ}$ . The result is that  $I_{CQ}$  now flows through a higher collector-to-emitter resistance, and the transistor's collector-to-emitter voltage *increases* to a level higher than the power-supply potential. That same phenomenon occurs when the input signal starts swinging positive. In that case, however, the collector-to-emitter voltage subtracts from the power-supply voltage. As a result, the transistor should be biased so that  $V_{CEQ}$  is higher than  $V_{CC}/2$ . We will design the circuit in Fig. 1 so that the collector-to-emitter voltage of Q2 will

of 0, the RMS current delivered by the transistor is  $I_{Cmax}/2\sqrt{2}$ . Similarly, the RMS voltage due to the 0- to 18-volt swing is  $V_{Cmax}/2\sqrt{2} = 18/2\sqrt{2} = 6.36$  volts. The product of the RMS voltage and the RMS current must be equal to the required 1 watt needed from Q2, so  $6.36(I_{Cmax}/2\sqrt{2}) = 1$ , and  $I_{Cmax} = 0.44$  amps. Plot that point on the  $I_C$  axis and draw the AC load line. Idling current,  $I_{CQ}$ , is half of  $I_{Cmax}$ , or 0.22 amps.

The slope of the AC load line is the negative of the load resistance as seen by Q2, or  $(18 - 0)/(0.44 - 0) = 41$  ohms. Because the load that the loudspeaker presents to the secondary is 8 ohms, the impedance ratio of the primary to secondary is  $41/8 = 5.13$ . The turns ratio of T2 is the square root of the impedance ratio, or 2.3:1.

For stability's sake, we want to make  $R_{E2}$  as large as possible. But we are limited because of gain. Let us assume, for

the moment, that we can accept a gain of 8.2. Then the sum of  $R_{E2}$  and  $r_e$ , the AC emitter resistance of the transistor, can be  $41/8.2 = 5$  ohms. (That is so because the gain of a stage of amplification is equal to the ratio of the output impedance to the resistance in the emitter circuit. Here it is  $41/5 = 8.2$ .) Because the idling current is 0.22 amps, or 220 mA,  $r_e = 26/220 = 0.12$  ohms. Resistor  $R_{E2}$  can be a maximum of  $5 - 0.12 = 4.88$  ohms. Thus, a standard 4.7-ohm resistor can be used for  $R_{E2}$ . Now the actual gain of that circuit is  $41/(4.7 + 0.12) = 8.33$ . That is just a bit above our minimum desired gain of 8.2.

If the transistor is to idle at 0.22 amperes, the base current is  $I_{CQ}/\beta = 0.22/100 = 0.0022$  amp. That current is determined from the circuit consisting of  $V_{CC}$ ,  $R_{B2}$ , the resistance of the secondary of T1 (assumed to be 0 ohms), the base-emitter voltage of Q2, and the resistance  $R_{E2}$  multiplied by  $\beta$ . Knowing everything but  $R_{B2}$ , we can determine it from the equation relating all these factors,  $0.0022 = (V_{CC} - V_{BE})/(R_{B2} + \beta R_{E2}) = (9 - 0.6)/[R_{B2} + 100(4.7)]$ ; solving for  $R_{B2}$ , we find it is equal to 3348 ohms. A standard 3300-ohm resistor can be used. If that resistor is only to affect the DC bias, no signal voltage must appear across it. For that to happen,  $R_{B2}$  must be bypassed by a capacitor. The reactance of that capacitor,  $X_C$ , should be about .01 of the resistance of  $R_{E2}$ , or 330 ohms at the lowest frequency,  $f_L$ , to be amplified. The value of the required capacitor can be found from  $C = 1/(2\pi f_L X_C) = 1/2072.4f_L$ .

Because the gain of the circuit is 8.33 and the RMS voltage at the collector of Q2 is 6.3 volts, the voltage at the secondary of T1 must be at least  $6.3/8.33$ , or 0.76 volts RMS to drive the output circuit to its desired level. The resistance across the secondary winding of T1 is the 3300 ohms of  $R_{B2}$  in parallel with  $\beta R_{E2}$  ( $r_e$  is negligible), or 3300 ohms in parallel with 470 ohms. That is equal to 411 ohms.

A reasonable load for T1 to present to the collector of Q1 is 2000 ohms. As the secondary of the transformer sees 411 ohms, and as the turns ratio of T1 is equal to the square root of the resistance ratio, that turns ratio is equal to  $\sqrt{2000/411}$ , or 2.2:1. That, of course, is also the voltage ratio. So if 0.76 volt must be across the secondary of T1 for T2 to deliver  $\frac{1}{2}$  watt,  $2.2 \times 0.76 = 1.67$  volts must be across the primary of T1. As the cartridge delivers 0.25 volt, the gain of the Q1 amplifier must be at least  $1.67/0.25 = 6.68$ . Hence  $R_{E1} + r_e$  of Q1 must be equal to or less than  $2000/6.68 = 300$  ohms. We can now plot a load line for Q1, and find its quiescent operating point, as shown in Fig. 2.

When we worked around Q2, we did not decide just what the AC load resistance should be. Now, we know it—it is 2000 ohms plus the 300-ohm resistance of the emitter, or a total of 2300 ohms. Thus,

$I_{Cmax}$  is equal to 18-volts/2300-ohms, or 7.8 mA. The quiescent operating point,  $I_{CQ}$  is one half of that for Q1, or 3.9 mA. That makes  $r_e$  for the Q1 stage equal to  $26/3.9$ , or 6.7 ohms. Resistor  $R_{E1}$  must then be a maximum of 300 ohms less 6.7 ohms, or less than 293.3 ohms. Use a standard 270-ohm resistor for that.

Since the idling collector current is 3.9 mA and  $\beta$  is 100, the base current is  $3.9/100$ , or 0.039 mA. Using that,  $R_B$  can be determined from the base current circuit; i.e.,  $(V_{CC} - V_{BE})/(R_B + \beta R_E) = (9 - 0.6)/[R_B + 100(270)]$ . Solving for  $R_B$ , it is equal to 188,000 ohms and a standard 180,000-ohm resistor can be used.

Transformer coupling is used quite often in radios where one IF stage is coupled to the other in a circuit similar to that shown in Fig. 1. In those circuits, however, a capacitor shunts the coil windings to form a resonant circuit so that only the IF frequency will pass through the amplifier circuit and all other frequencies will be rejected.

### Resistor-capacitor coupling

Transformers are fine when you are working with RF frequencies. But at AF frequencies, transformers can get quite expensive and bulky. That's why R-C coupling is now being used for many applications, although transformer-coupling is still common in RF circuits.

Let's look at the circuit shown in Fig. 3. This time, let us say that the input

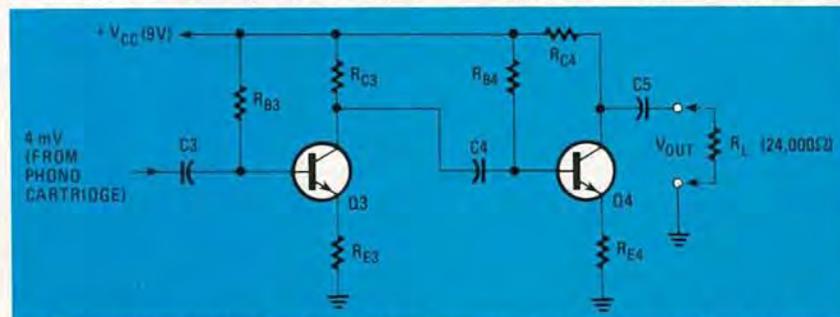


FIG. 3—AMONG THE ADVANTAGES of using R-C coupled circuits such as this one are the elimination of bulky and expensive transformers, and increased voltage gain.

signal comes from a low-impedance magnetic-type phonograph cartridge with an output of 4 millivolts. Assume we want to build that voltage up to a sufficient level to drive the power amplifier of Fig. 1. Thus the circuit in Fig. 1 would be driven by an amplifier circuit rather than by a ceramic cartridge. In this case, 0.25 volt is required at the output of the R-C circuit of Fig. 3 and that output must drive a load equal to the input impedance of the transformer amplifier, or 180,000 ohms in parallel with  $\beta$  multiplied by 276.7 ohms (the emitter resistance of Q1), or 24,000 ohms. The overall voltage gain of the amplifier must then be  $0.25\text{-volt}/4\text{-millivolt} = 62.5$ . Theoretically, the voltage gains of the two stages can be made identical. If that were done, the gain of each stage would be  $\sqrt{62.5}$ , or 7.9.

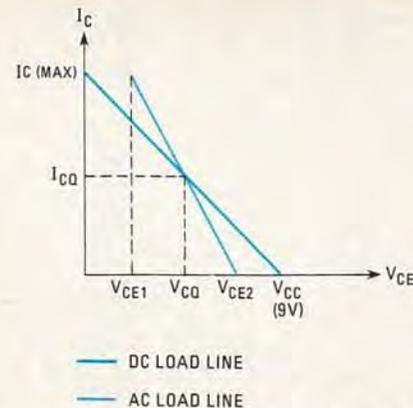


FIG. 4—THE LOAD LINES for the transistors in the circuit shown in Fig. 3.

Usually, however, the voltage gain of the second stage is made lower than that of the first stage because the output load of the second stage generally has a lower impedance.

Let's start the design by considering the second stage. Resistor  $R_{C4}$  should be chosen so that its value is between 10% to 20% of  $R_L$ ; let's make it 3300 ohms. Now the DC load on Q4 is 3300 ohms and the AC load is 3300 ohms in parallel with the 24,000 ohm load, or 2933 ohms. To have a gain of somewhat more than 5, make  $R_{E4}$  equal to about 20% of 2933 ohms, or about 470 ohms.

The DC load line (see Fig. 4) extends from  $V_{CC}$  to a point on the  $I_C$  axis equal to  $I_{Cmax}$ . From the load-line equation,  $V_{CC} = I_{Cmax}(R_{C4} + R_{E4}) + V_{CE4}$ , we find

that  $I_{Cmax} = 2.4$  mA when  $V_{CE4} = 0$  and  $V_{CC} = 9$ . And, as  $I_C$  must swing between 2.4 mA and 0,  $I_{CQ} = 1.2$  mA and  $V_{CQ} = 4.5$  volts.

The AC load line must pass through that point. We can determine  $V_{CE1}$  from the negative of the equation for the slope of the AC load line, or  $(V_{CE2} - V_{CQ})/(I_{CQ} - 0) = (2933 + 470) = 3403$  ohms. Substituting  $V_{CQ} = 4.5$  volts and  $I_{CQ} = 1.2$  mA into the equation,  $V_{CE2} = 8.6$  volts. As  $V_{CE1}$  is the same distance below  $V_{CQ}$  as  $V_{CE2}$  is above it, and because  $(V_{CE2} - V_{CQ}) = (8.6 - 4.5) = 4.1$  volts, then  $V_{CE1} = (4.5 - 4.1) = 0.4$  volts when  $I_C$  is equal to 2.4 mA.

The output must be able to swing the full 0.25 volt RMS, or  $2/\sqrt{2}(0.25) = 0.707$  volt peak-to-peak in order to be able to drive the amplifier in Fig. 1 suf-

ficiently. The maximum peak-to-peak swing of the transistor is  $(V_{CE2} - V_{CE1}) = (8.6 - 0.4) = 8.2$  volts. Thus a 0.707-volt swing is well within the capabilities of this amplifier stage.

The actual gain of this stage can be determined after we find just what  $r_{e4}$  is in the Q4 circuit. It is equal to  $26/1.2 = 21.7$  ohms. Thus, the total resistance in the emitter circuit is  $470 + 21.7 = 491.7$  ohms, and the gain is just about equal to  $2933/491.7 = 6$ . Because the total gain must be at least equal to 62.5, the gain of the first stage, Q3, must be greater than 62.5/6, or 10.42.

We will now determine  $R_{B4}$  in a manner similar to that used for determining  $R_{B2}$  in the circuit of Fig. 1. Use the equation  $V_{CC} = I_B R_{B4} + V_{BE} + \beta I_B R_{E4}$ . Note that  $\beta = 100$  and  $I_B = I_{CQ}/\beta = 1.2/100 = 1.2 \times 10^{-5}$  amps. After substituting into the equation, we can calculate that  $R_{B4}$  is equal to 653,000 ohms and a standard 620,000-ohm resistor can be used.

The load on the output of the Q3 circuit is 620,000 ohms in parallel with  $\beta(R_{E4} + r_e)$ , or 45,291 ohms. As  $R_{C3}$  should be less than 10% of that, use a 4300-ohm resistor for that. Now the DC load on the transistor is 4300 ohms and the AC load is that resistor in parallel with 45,291 ohms, or 3927 ohms.

Following the procedure used to find  $R_{E4}$ , we can determine that  $R_{E3}$  should be 330 ohms if we are to get a gain of more than the required minimum of 10.42. To check the gain, we must first determine  $I_{Cmax}$ . It is  $(9 - 0.6)/(3927 + 330) = 1.97$  mA, and  $I_{CQ}$  is one-half of  $I_{Cmax}$ , or about 1 mA. That makes  $r_e$  equal to 26/1, or 26 ohms. The total emitter resistance is  $330 + 26$ , or 356 ohms. Consequently, the actual gain of the Q3 circuit is about  $3927/356$ , or 11. As that is more than the required minimum of 10.42, the circuit fulfills our requirements.

The value of  $R_{B3}$  can be found from the equation  $V_{CC} = I_B R_{B3} + V_{BE} + \beta I_B R_{E3}$ . When we note that  $I_B = \beta I_{CQ}$ , we can determine that  $R_{B3}$  is equal to about 897,000 ohms, and you can use a standard 910,000-ohm resistor.

The resistance the magnetic cartridge sees is 35,600 ohms (the resistor in the emitter circuit of Q3, multiplied by the beta of that transistor) in parallel with 910,000 ohms, or just about 35,000 ohms. That is pretty close to the resistance most cartridges must see. If the resistance the cartridge must see is specified to be less than 35,000 ohms by the manufacturer, an additional shunting resistor must be wired across the cartridge. The shunting resistor can be calculated by using the equation for resistors in parallel and substituting 35,000 ohms and the resistance that the cartridge must see into the equation. Place that calculated resistor directly across the cartridge, not on the circuit side of C3. The reason it must be placed that way is to keep it from upset-

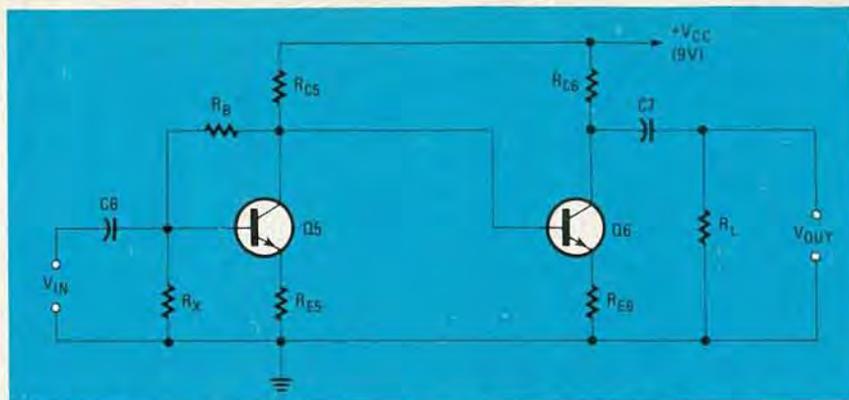


FIG. 5—DIRECT-COUPLED CIRCUIT. Those circuits have been developed to overcome the drawbacks of using capacitors as coupling devices.

ting the bias voltage established by the other components in the circuit. The function of the isolating capacitor has been detailed in the previous article on audio-amplifier circuits.

### Direct-coupled amplifiers

Using capacitors to couple transistor amplifiers does have some important drawbacks. For one thing, capacitors do not pass low frequencies very readily. Besides, capacitor-resistor circuits introduce phase shifts, are not linear, and add distortion. As a result, circuits have been developed that do not use coupling capacitors. An example of one such circuit is shown in Fig. 5. While that circuit does not appear to be much different than the one you see in Fig. 3, the differences that do exist are significant. Note primarily that the coupling capacitor between the collector of Q5 and the base of Q6 has been omitted. The overall gain is still the product of the gains of the two transistor circuits.

The circuit around Q6 can be designed in a fashion similar to that used for the design of the circuit around Q4 in Fig. 3. But here, we must also consider that the base voltage of Q6 is the same as the collector voltage of Q5. Voltage at the collector of Q5 must not be upset if Q6's base voltage is low, as it tends to be. To compensate for that, a parallel R-C network may be wired in series with  $R_{E6}$ . Considering the DC current that flows through the series resistor circuit, the DC voltage developed across the combination is much greater than it would have been had only  $R_{E6}$  been in the circuit. Voltage at the base of Q6 is the voltage across the series combination of resistors added to the  $V_{BE}$  of the transistor. The capacitor across the added series resistor is there to bypass the AC signal so that the new resistor will not affect the AC gain of Q6. The reactance of that bypass capacitor should be less than 10% of the value of the resistor it is bypassing at the lowest frequency to be amplified.

Note the  $R_B$  and  $R_X$  resistors in the base circuit of Q5. If you look back at the bias circuits described in a previous article, you will find that the circuit involving

those resistors is essential for stabilizing transistor collector-current to withstand variations of temperature. It is used here because the effect of any increase in leakage current in Q5 is multiplied in magnitude due to the DC amplification of the direct-coupled circuit. So if the leakage current in Q5 is multiplied in that transistor by a factor of 10 due to the increase in temperature, it can be multiplied further by the beta of Q6 before it appears in its collector circuit. Excellent stability is an important consideration when designing direct-coupled amplifier circuits.

More than just the collector current of Q5 flows through  $R_{C5}$ . When determining the voltage drop across that resistor, be sure to add other currents that flow through that resistor to the  $I_C$  normally expected. Bias current for Q5 flows through  $R_{C5}$  (and  $R_B$ ) along with the base bias-current for Q6. The sum of those currents may, at times, be a substantial factor in determining the quiescent conditions around the circuit, along with the expected  $I_C$ .

### Special direct-coupled circuits

Last time, we described the emitter-follower or common-collector circuit. Its characteristics were a high input impedance and a low output impedance; its power gain was equal to beta. The Darlington circuit goes one step farther and improves even on those excellent characteristics by a factor of beta.

Darlington amplifiers consist of two cascaded emitter followers, as shown in Fig. 6. If the beta of Q1 is  $\beta_1$ , and the beta of Q2 is  $\beta_2$ , the current and power gains of the overall circuit are  $\beta_1 \times \beta_2$ . But the voltage gain still remains slightly less than 1, as it did for the emitter follower.

The input impedance is  $R_E$  multiplied by the product of the betas, in parallel with  $R_B$ . The value of  $R_B$  can be quite large in that circuit as there is only a small amount of base current in Q1. That Q1 base current must equal  $(V_{CC} - 2V_{BE})/(\beta_1\beta_2R_E + R_B)$  for a substantial amount of emitter current to flow in Q2.

The output impedance of that circuit is  $R_B/\beta_1\beta_2$  in parallel with  $R_E$ . The values

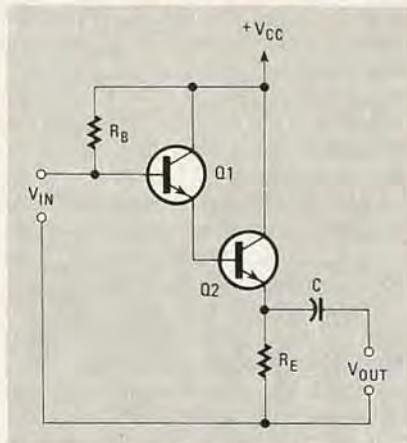
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## ANALOG CIRCUIT

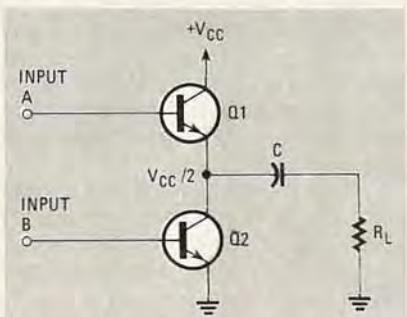
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of  $r_e$  and  $r_b$  are negligible when compared to  $R_E$  and  $R_B$  and can therefore be ignored. The load across  $R_E$  can be very small, but should be no less than ten times the calculated output impedance.

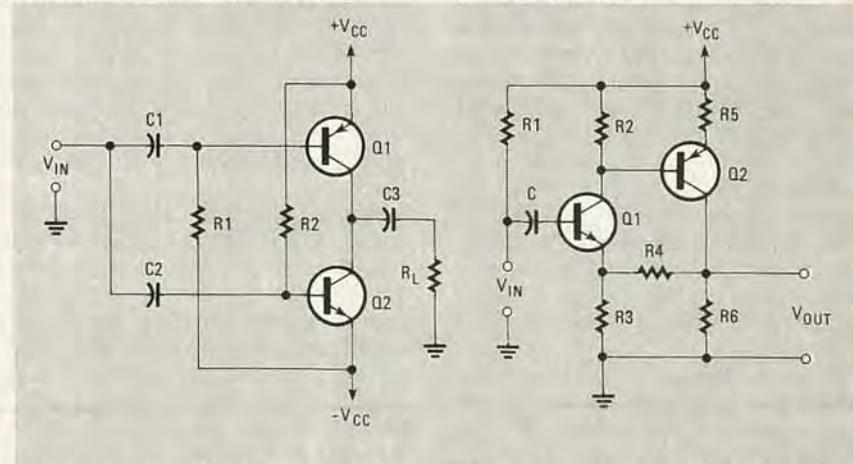
In addition to the direct-coupled Darlington arrangement, transistors can be connected in series or in parallel. When



**FIG. 6—THE DARLINGTON AMPLIFIER** consists of two cascaded emitter-follower circuits.



**FIG. 7—WHEN TWO TRANSISTORS ARE** connected in series, the supply voltage ideally will be split evenly between the two devices.



**FIG. 8—COMPLEMENTARY TRANSISTOR CIRCUITS.** The one shown in *a* is a push-pull audio amplifier. The one shown in *b* can be modified to meet a variety of circuit requirements.

connected in parallel, a two-transistor combination can theoretically deliver twice the current each individual device is capable of supplying. For that to be true,

the current they must pass must divide evenly between the two devices. That of course, never happens. Actually, the current rating of each transistor should be reduced by at least 20% to keep the devices in the circuit from being destroyed. One method of equalizing the currents in the transistors is to place a small resistor in series with the emitters of each one. That, however, reduces the amount of power the circuit can deliver for a given power-supply voltage. As for the series circuit, ideally the supply voltage divides evenly between the two devices, as shown in Fig. 7.

Complementary pairs are used quite frequently in direct-coupled circuits. One common use for that type of arrangement is in a push-pull audio-amplifier circuit, such as the one shown in Fig. 8-*a*. In that amplifier, the positive half of the input signal's cycle is fed to one transistor, in this case Q2, while the negative half is fed to the other, in this case Q1. The two halves of the signal combine across  $R_L$  to reproduce the full cycle. In a practical application,  $R_L$  could be a loudspeaker.

The complementary-pair circuit shown in Fig. 8-*b* can be modified depending on the requirements of the circuit. In the circuit as shown, the current gain is the product of the betas of the two transistors, the voltage gain is  $1 + (R_4/R_3)$ , and the power gain is the product of the voltage and current gains. The input impedance is approximately equal to  $R_1$  while the output impedance is  $R_4R_6/(R_4 + \beta_2R_6)$ , where  $\beta_2$  is the beta of Q2.

Two variations of the circuit shown in Fig. 8-*b* are frequently used. In one,  $R_4$  is omitted while  $R_3$  and  $R_5$  are shorted. Now, voltage gain drops to 1 and the output impedance becomes equal to  $R_6$ .

In the second variation,  $R_2$  and  $R_3$  are omitted while  $R_4$  and  $R_5$  are shorted. Now the characteristics are the same as for the first variation except that the output impedance drops to near zero.

Another form of direct-coupled circuit is the differential amplifier. That extremely useful circuit will be the subject of the next part of this article. **R-E**