Bipolar Transistor Cookbook — Part 1

The bipolar transistor is the most important “active” circuit element used in modern electronics, and it forms the basis of most linear and digital ICs and op-amps, etc. In its discrete form, it can function as either a digital switch or as a linear amplifier, and is available in many low, medium, and high power forms. This opening episode concentrates on basic transistor theory, characteristics, and circuit configurations. The remaining seven parts of the series will present a wide range of practical bipolar transistor application circuits.

BIPOLAR TRANSISTOR BASICS

A bipolar transistor (first invented in 1948) is a three-terminal (base, emitter, and collector), current-amplifying device in which a small input current can control the magnitude of a much larger output current. The term “bipolar” means that the device is made from semiconductor materials in which conduction relies on both positive and negative (majority and minority) charge carriers.

A normal transistor is made from a three-layer sandwich of n-type and p-type semiconductor material, with the base or “control” terminal connected to the central layer, and the collector and emitter terminals connected to the outer layers. If it uses an n-p-n construction sandwich, as in **Figure 1(a)**, it is known as an npn transistor and uses the standard symbol in **Figure 1(b)**.

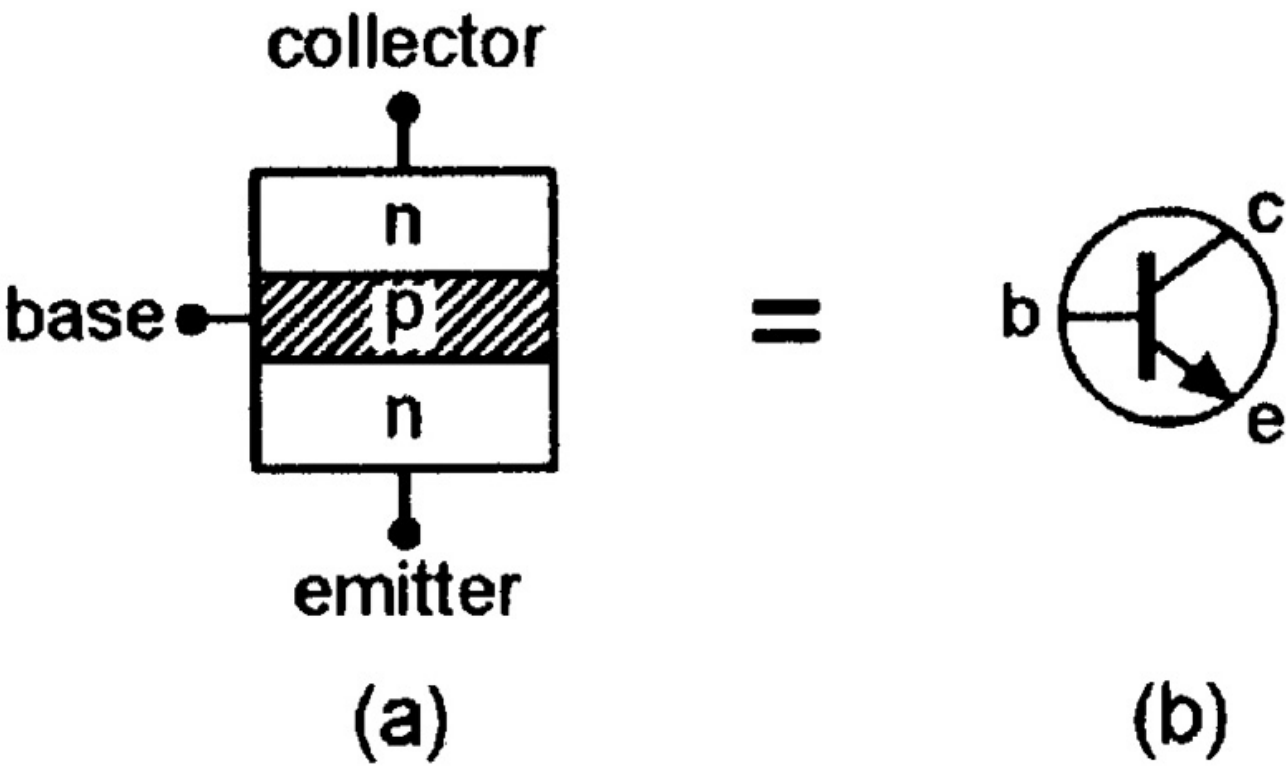


FIGURE 1. Basic construction (a) and symbol (b) of npn transistor.

If it uses a p-n-p structure, as in **Figure 2(a)**, it is known as a pnp transistor and uses the symbol in **Figure 2(b)**.

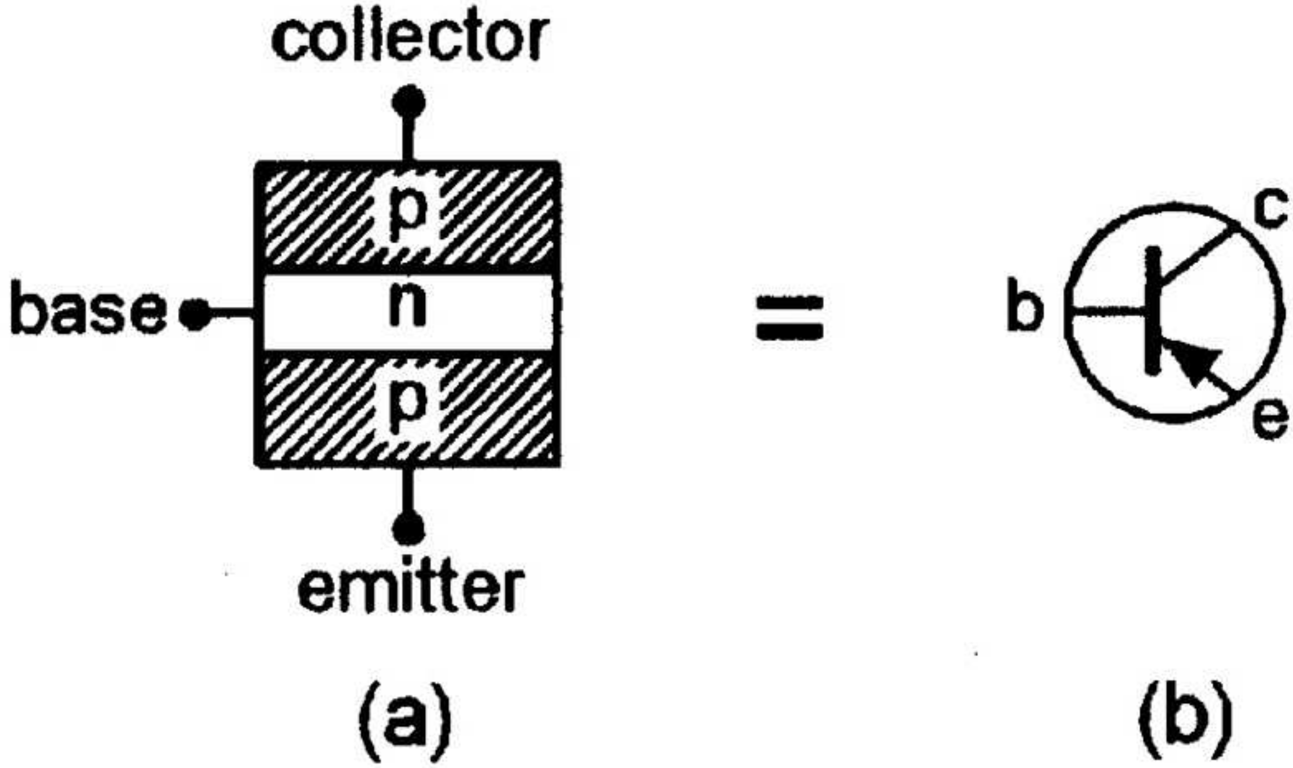


FIGURE 2. Basic construction (a) and symbol (b) of pnp transistor.

In use, npn and pnp transistors each need a power supply of the appropriate polarity, as shown in **Figure 3**.

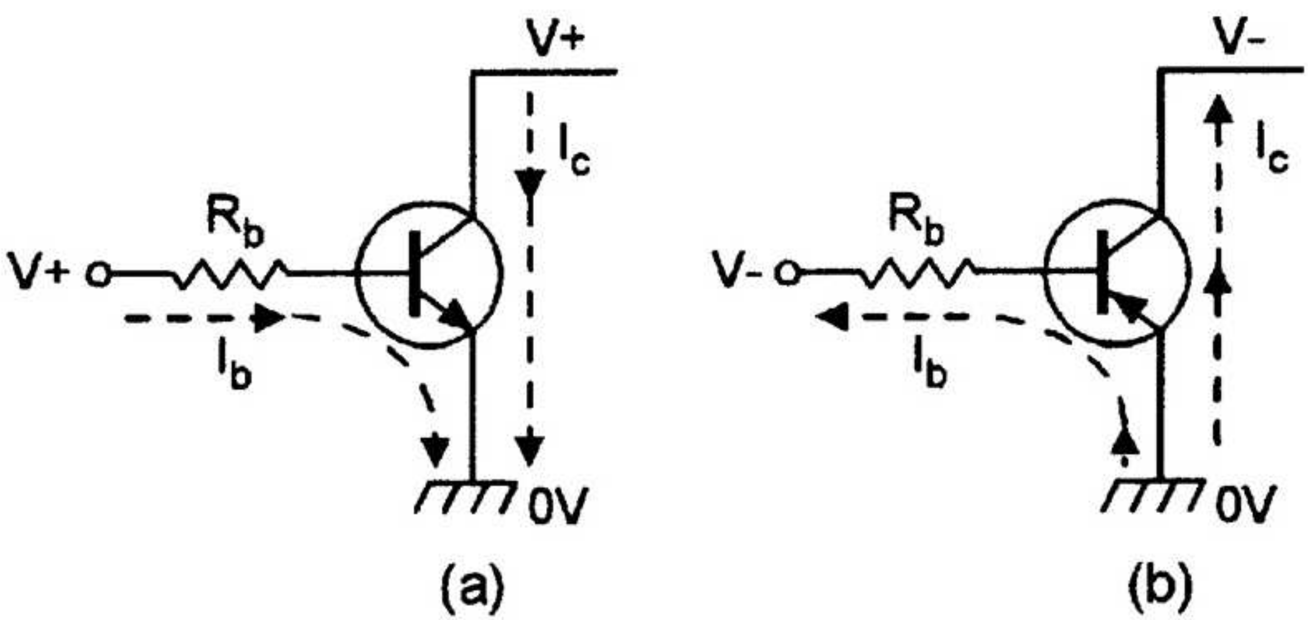


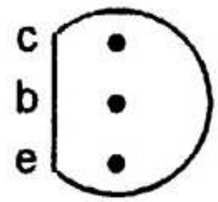
FIGURE 3. Polarity connections to (a) npn and (b) pnp transistors.

An npn device needs a supply that makes the collector positive to the emitter — its output or main-terminal signal current (I_c) flows from collector to emitter, and its amplitude is controlled by an input “control” current (I_b) that flows from base to emitter via an external current-limiting resistor (R_b) and a positive bias voltage. A pnp transistor needs a negative supply — its main terminal current flows from emitter to collector, and is controlled by an emitter-to-base input current that flows to a negative bias voltage.

In the early years of bipolar transistor usage, most transistors were made from germanium semiconductor materials. Such devices had many practical disadvantages: they were fragile, excessively temperature-sensitive, electronically noisy, and had very poor power-handling capacities. Germanium transistors are now obsolete. Virtually all modern bipolar transistors are made from silicon semiconductor materials. Such devices are robust, have good power-handling capacities, are not excessively temperature sensitive, and generate negligible electronic noise.

Today, a very wide variety of excellent silicon bipolar transistor types are readily available. **Figure 4** lists the basic characteristics of two typical general-purpose, low-power types — the 2N3904 (npn) and the 2N3906 (pnp) — which are each housed in a TO-92 plastic case and have the under-side pin connections shown in the diagram.

Parameter	2N3904	2N3906
Transistor type	npn	pnp
I_C (max)	200mA	-200mA
V_{CEO} (max)	40V	-40V
V_{CBO} (max)	60V	-40V
P_T (max)	310mW	310mW
h_{fe} (= a.c. beta)	100-300	100-300
f_T (typ) = gain/bandwidth product	300MHz	250MHz



TO-92 case

FIGURE 4. General characteristics and outlines of the 2N3904 and 2N3906 low-power silicon transistors.

Note, when reading the **Figure 4** list, that $V_{CEO(max)}$ is the maximum voltage that may be applied between the collector and emitter when the base is open-circuit, and $V_{CBO(max)}$ is the maximum voltage that may be applied between the collector and base when the emitter is open-circuit. $I_{C(max)}$ is the maximum mean current that can be allowed to flow through the collector terminal of the device, and $P_{T(max)}$ is the maximum mean power that the device can dissipate, without the use of an external heatsink, at normal room temperature.

One of the most important parameters of the transistor is its forward current transfer ratio, or h_{fe} — this is the current-gain or output/input current ratio of the device (typically 100 to 300 in the two devices listed). Finally, the f_T figure indicates the available gain/bandwidth product frequency of the device, i.e., if the transistor is used in a voltage feedback configuration that provides a voltage gain of x100, the bandwidth is 1/100 of the f_T figure, but if the voltage gain is reduced to x10, the bandwidth increases to f_T/10, etc.

TRANSISTOR CHARACTERISTICS

To get the maximum value from a transistor, the user must understand both its static (DC) and dynamic (AC) characteristics. **Figure 5** shows the static equivalent circuits of npn and pnp transistors.

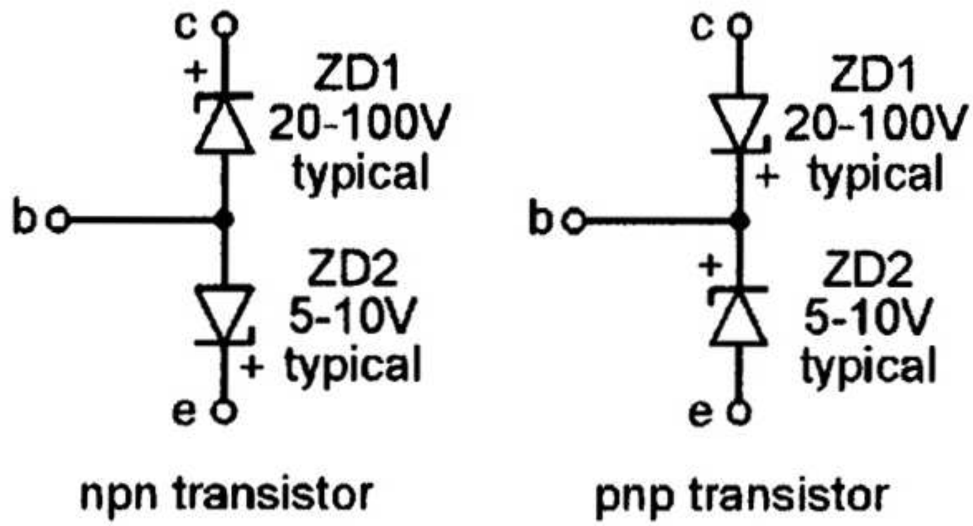


FIGURE 5. Static equivalent circuits of npn and pnp transistors.

A zener diode is inevitably formed by each of the transistor's n-p or p-n junctions, and the transistor is thus (in static terms) equal to a pair of reverse-connected zener diodes wired between the collector and emitter terminals, with the base terminal wired to their "common" point. In most low-power, general-purpose transistors, the base-to-emitter junction has a typical zener value in the range 5V to 10V — the base-to-collector junction's typical zener value is in the range 20V to 100V.

Thus, the transistor's base-emitter junction acts like an ordinary diode when forward-biased and as a zener when reverse-biased. In silicon transistors, a forward-biased junction passes little current until the bias voltage rises to about 600mV, but beyond this value, the current increases rapidly. When forward-biased by a fixed current, the junction's forward voltage has a thermal coefficient of about -2mV/^{0}C. When the transistor is used with the emitter open-circuit, the base-to-collector junction acts like that just described, but has a greater zener value. If the transistor is used with its base open-circuit, the collector-to-emitter path acts like a zener diode wired in series with an ordinary diode.

The transistor's dynamic characteristics can be understood with the aid of **Figure 6**, which shows the typical forward transfer characteristics of a low-power npn silicon transistor with a nominal h_{fe} (current gain) value of 100.

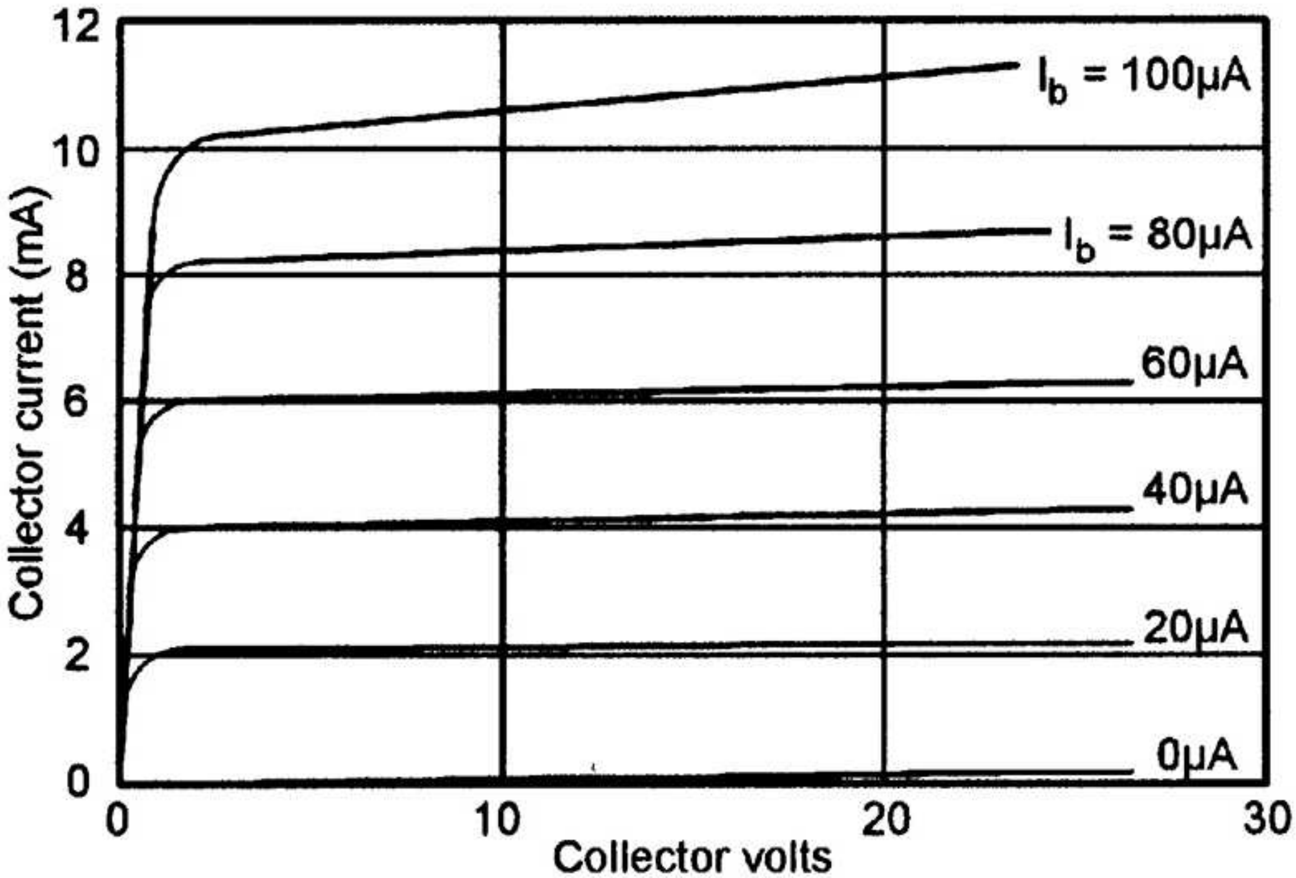


FIGURE 6. Typical transfer characteristics of low-power npn transistors with h_{fe} value of 100 nominal.

Thus, when the base current (I_b) is zero, the transistor passes only a slight leakage current. When the collector voltage is greater than a few hundred millivolts, the collector current is almost directly proportional to the base currents, and is little influenced by the collector voltage value. The device can thus be used as a constant-current generator by feeding a fixed bias current into the base, or can be used as a linear amplifier by superimposing the input signal on a nominal input current.

PRACTICAL APPLICATIONS

A transistor can be used in a variety of different basic circuit configurations, and the remainder of this opening episode presents a brief summary of the most important of these. Note that although all circuits are shown using npn transistor types, they can be used with pnp types by simply changing circuit polarities, etc.

DIODE AND SWITCHING CIRCUITS

The base-emitter or base-collector junction of a silicon transistor can be used as a simple diode or rectifier, or as a zener diode by using it in the appropriate polarity. **Figure 7** shows two alternative ways of using an npn transistor as a simple diode clamp that converts an AC-coupled rectangular input waveform into a rectangular output that swings between zero and a positive voltage value, i.e., which “clamps” the output signal to the zero-volts reference point via either the transistor’s internal base-emitter or base-collector “diode” junction.

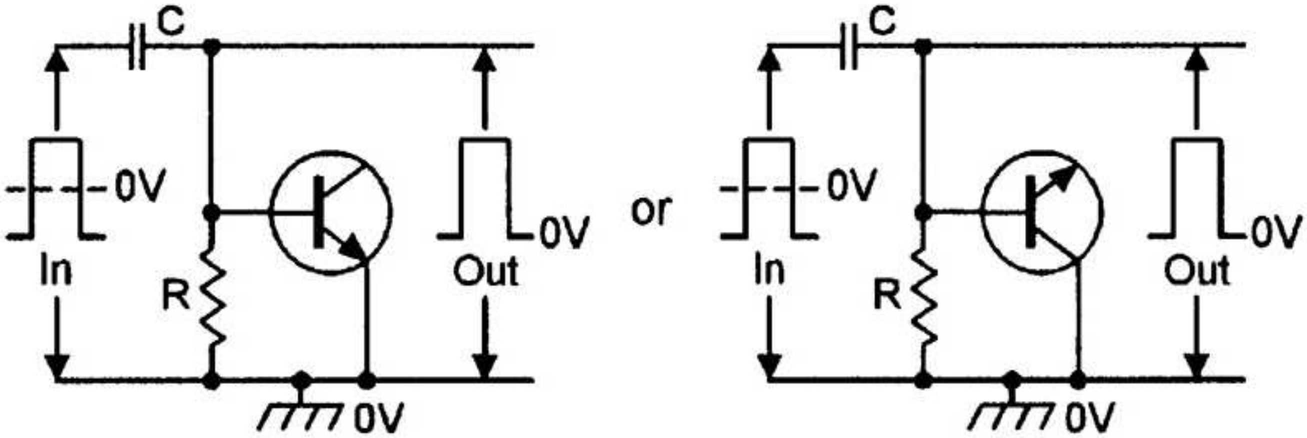


FIGURE 7. Clamping diode circuit, using an npn transistor as a diode.

Figure 8 shows an npn transistor used as a zener diode that converts an unregulated supply voltage into a fixed-value regulated output with a typical value in the range 5V to 10V, depending on the individual transistor. Only the reverse-biased base-emitter junction of the transistor is suitable for use in this application. If the reverse-biased base-collector junction is used, the zener value typically rises into the 30V-100V range, and the transistor may self-destruct (due to over-heating) at fairly low zener current levels.

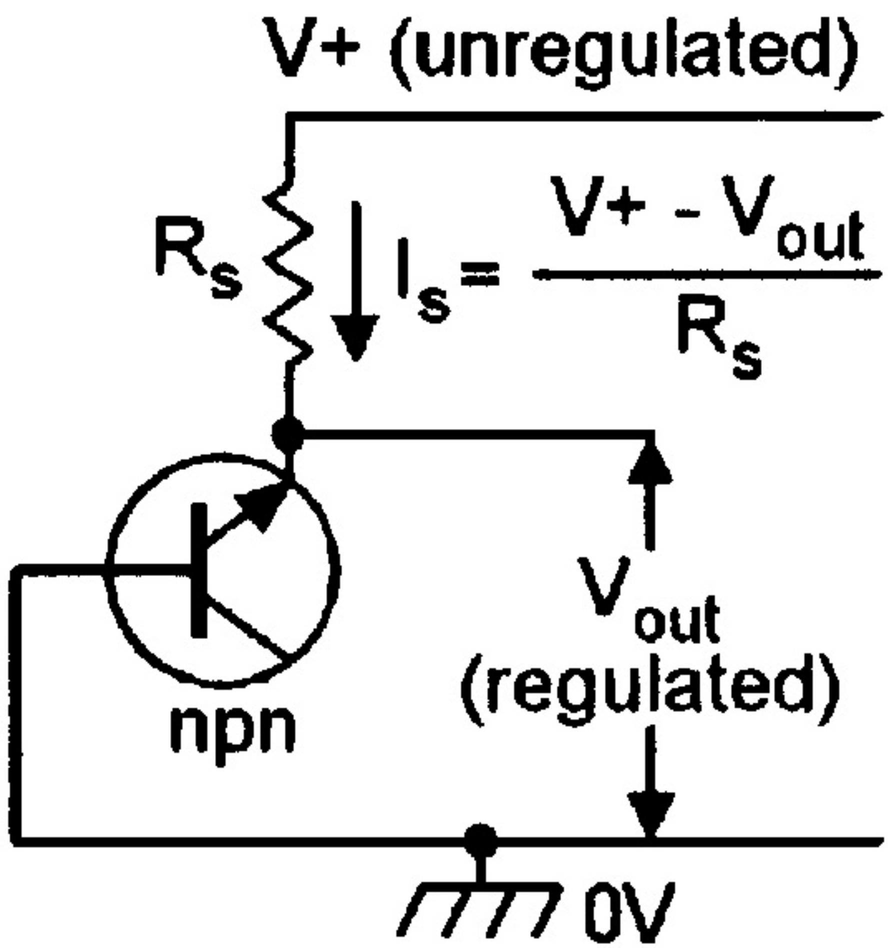


FIGURE 8. A transistor used as a zener diode.

Figure 9 shows a transistor used as a simple electronic switch or digital inverter. Its base is driven (via R_b) by a digital input that is at either zero volts or at a positive value, and load R_L is connected between the collector and the positive supply rail. When the input voltage is zero, the transistor is cut off and zero current flows through the load, so the full supply voltage appears between the collector and emitter. When the input is high, the transistor switch is driven fully on (saturated) and maximum current flows in the load, and only a few hundred millivolts are developed between the collector and emitter. The output voltage is thus an inverted form of the input signal.

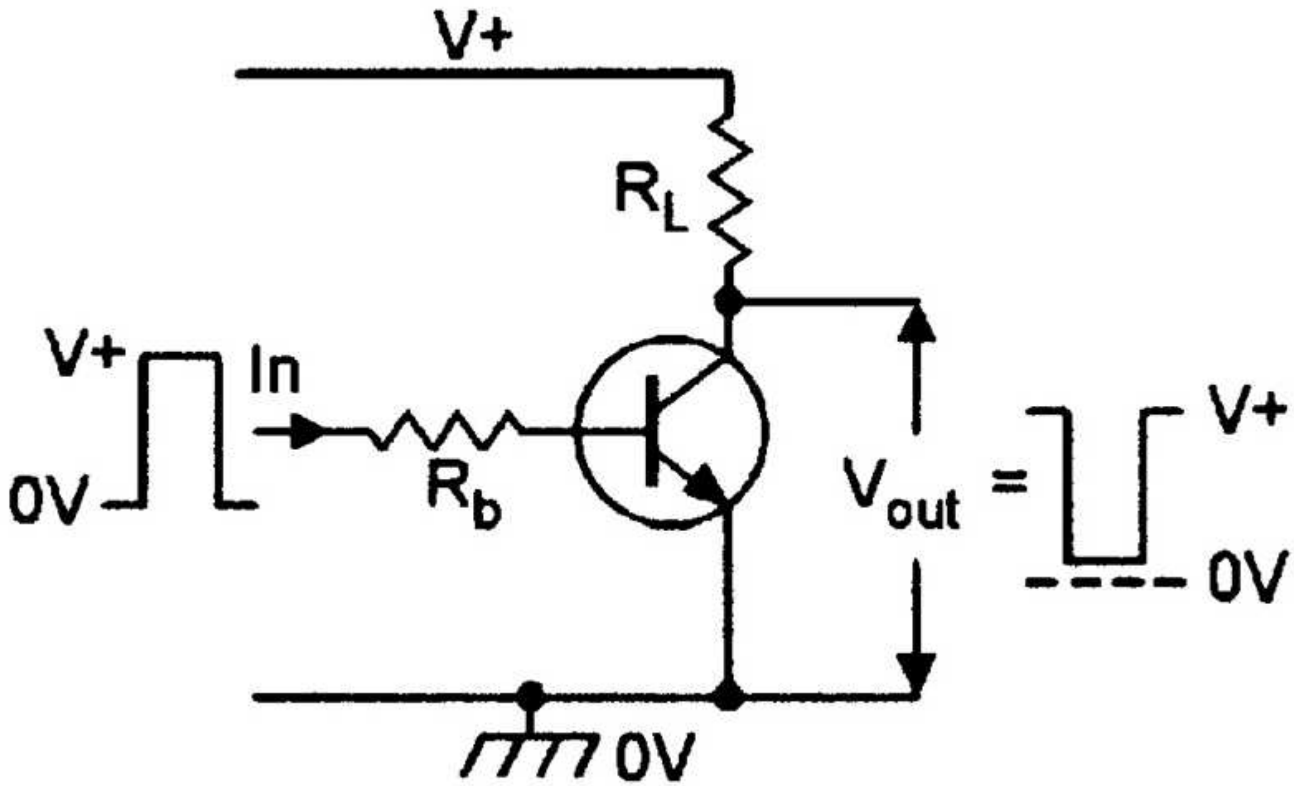


FIGURE 9. Transistor switch or digital inverter.

The basic **Figure 9** circuit is intended for use as a simple digital switch or inverter, driving a purely resistive load. It can be used as an electronic switch that drives a relay coil or other highly inductive load (such as a DC motor) by connecting it as shown in **Figure 10**, in which diodes D1 and D2 protect the transistor from high-value switch-off-induced back EMFs from the inductive load at the moment of power switch-off.

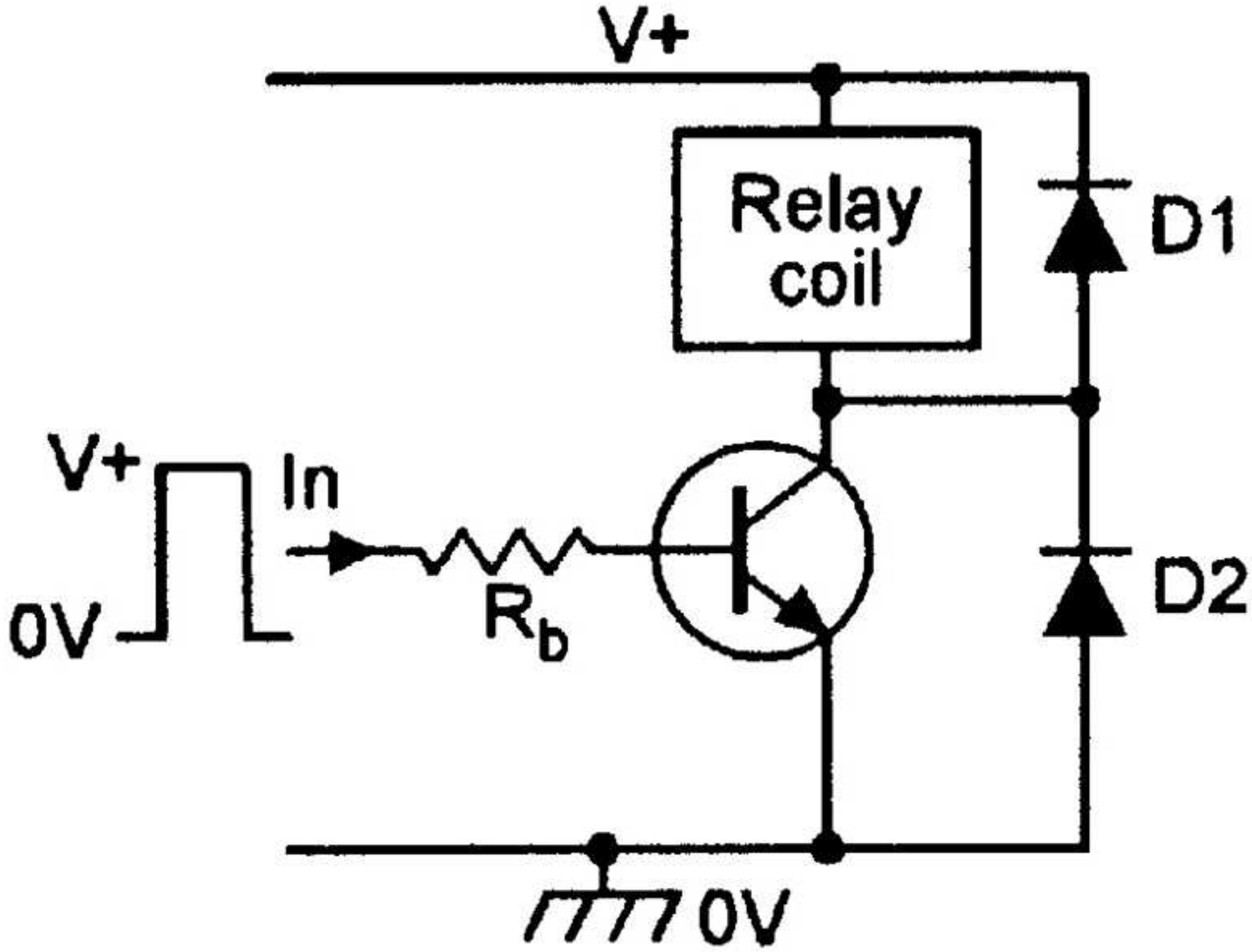


FIGURE 10. Transistor switch (digital inverter) driving a relay coil (or other inductive load).

LINEAR AMPLIFIER CIRCUITS

A transistor can be used as a linear current or voltage amplifier by feeding a suitable bias current into its base and then applying the input signal between an appropriate pair of terminals. The transistor can, in this case, be used in any one of three basic operating modes, each of which provides a unique set of characteristics. These three modes are known as "common-emitter" (**Figure 11**), "common-base" (**Figure 12**), and "common-collector" (**Figures 13 and 14**).

In the common-emitter circuit (which is shown in very basic form in **Figure 11**), resistive load R_L is wired between the transistor's collector and the positive supply line, and a bias current is fed into the base via resistor R_b, whose value is chosen to set the collector at a quiescent half-supply voltage value (to provide maximum undistorted output signal swings). The input signal is applied between the transistor's base and emitter via capacitor C, and the output signal (which is phase-inverted relative to the input) is taken between the collector and emitter. This circuit gives a medium-value input impedance and a fairly high overall voltage gain.

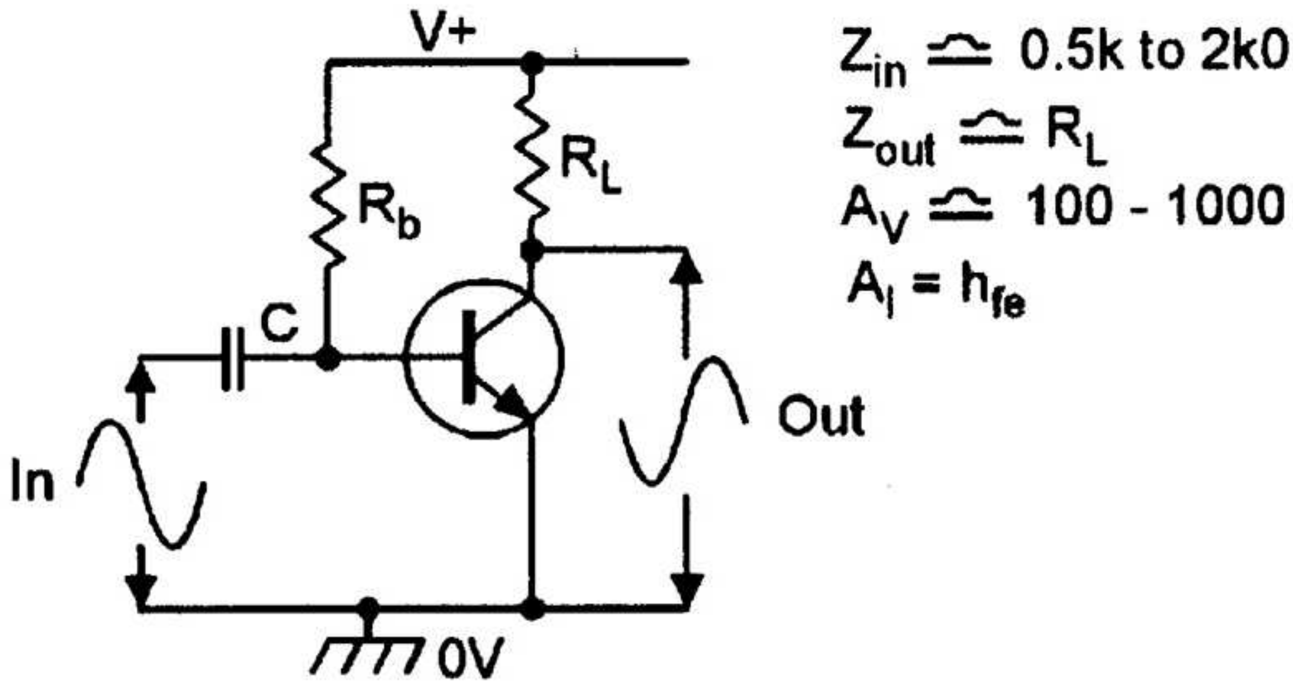


FIGURE 11. Common-emitter linear amplifier.

In the common-base circuit in **Figure 12**, the base is biased via R_b and is AC-decoupled (or AC-grounded) via capacitor C_b. The input signal is effectively applied between the emitter and base via C1, and the amplified but non-inverted output signal is effectively taken from between the collector and base. This circuit features good voltage gain, near-unity current gain, and a very low input impedance.

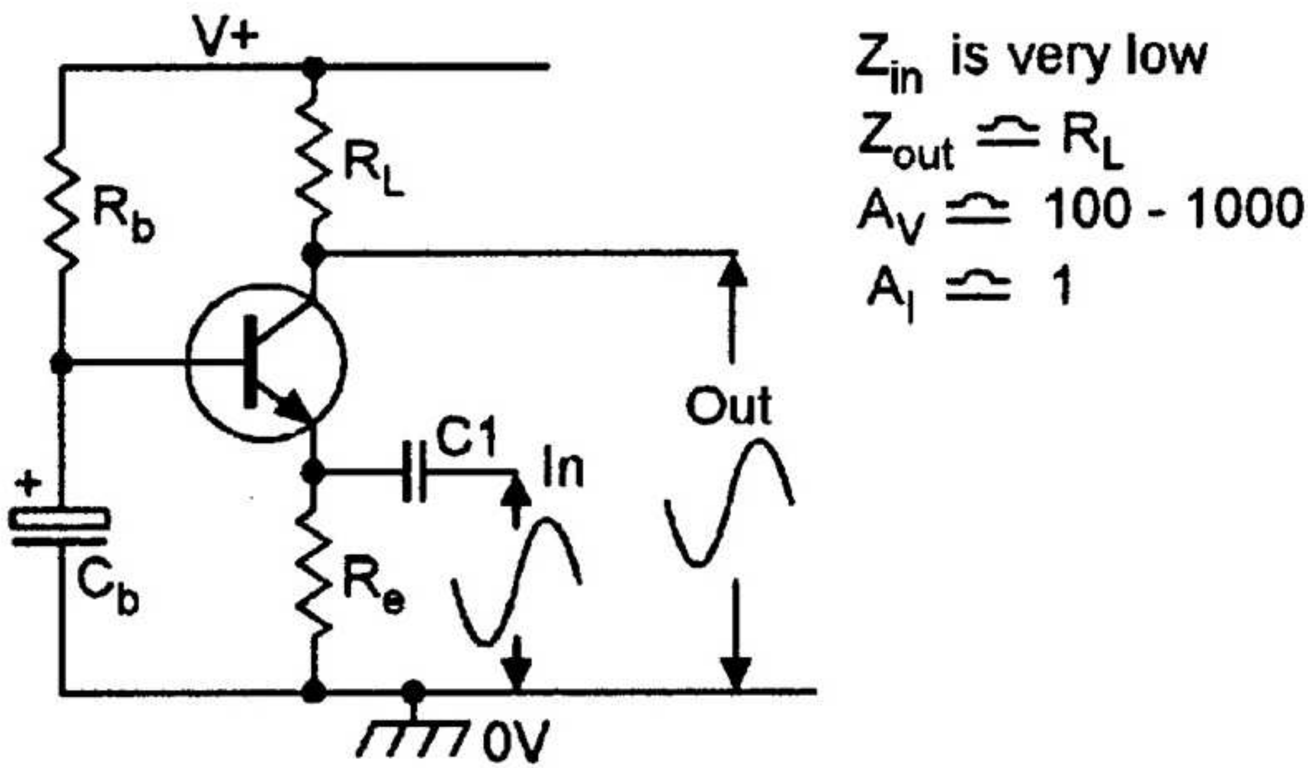


FIGURE 12. Common-base linear amplifier.

In the DC common-collector circuit in **Figure 13**, the collector is shorted to the low-impedance positive supply rail and is thus effectively at “virtual ground” impedance level. The input signal is applied between base and ground (virtual collector), and the non-inverted output is taken from between emitter and ground (virtual collector). This circuit gives near-unity overall voltage gain, and its output “follows” the input signal. It is thus known as a DC-voltage follower (or emitter follower) and it has a very high-input impedance (equal to the product of the R_L and h_{fe} values).

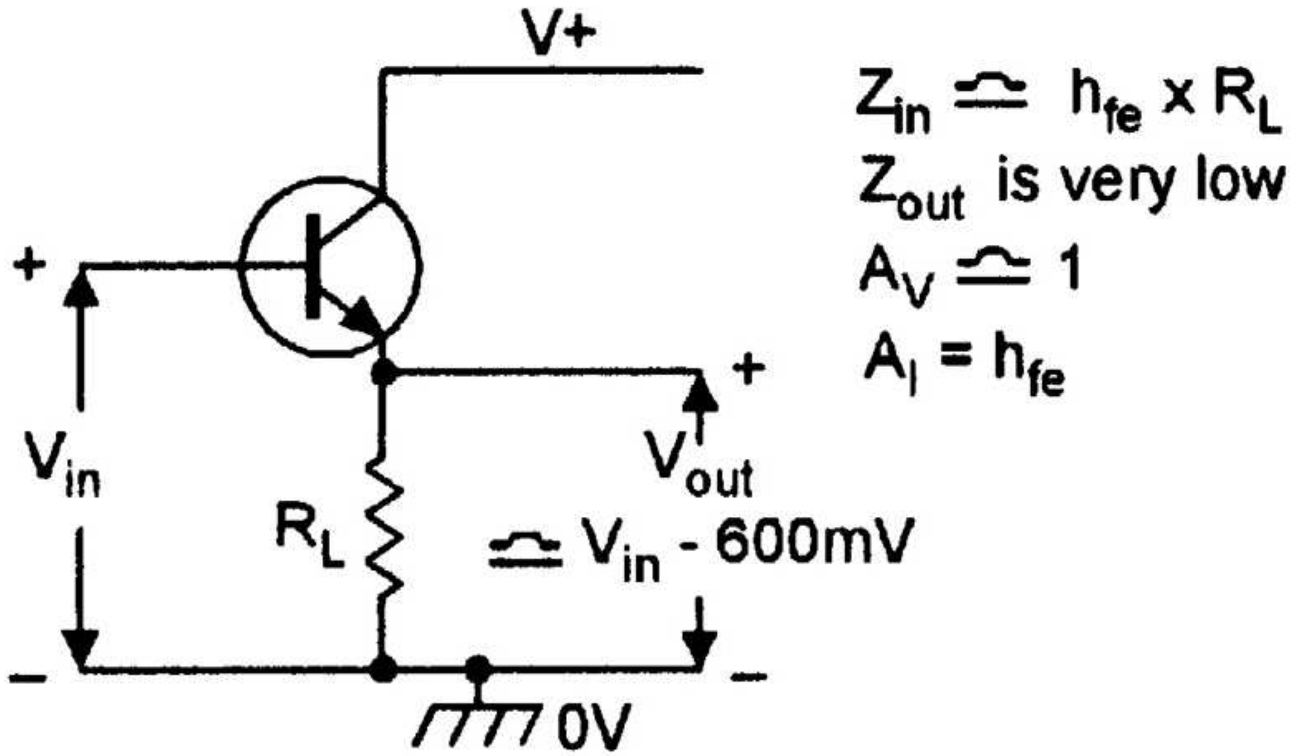


FIGURE 13. DC common-collector linear amplifier or voltage follower.

Note that the above circuit can be modified for AC use by simply biasing the transistor to half-supply volts and AC-coupling the input signal to the base, as shown in the basic circuit in **Figure 14**, in which potential divider R1-R2 provides the half-supply-voltage biasing.

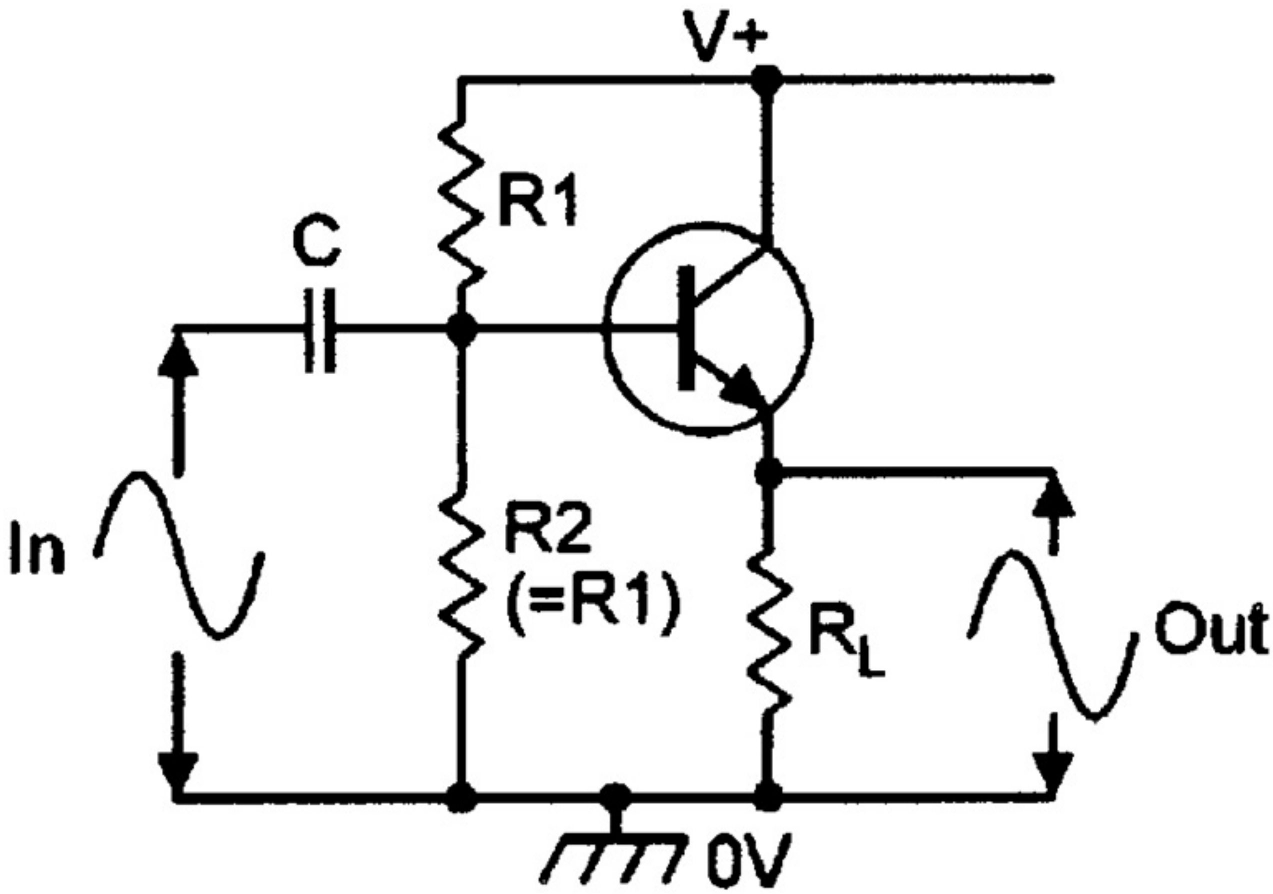


FIGURE 14. AC common-collector amplifier or voltage follower.

The chart in **Figure 15** summarizes the performances of the three basic amplifier configurations. Thus, the common-collector amplifier gives near-unity overall voltage gain and a high input impedance, while the common-emitter and common-base amplifiers both give high values of voltage gain, but have medium to low values of input impedance.

Parameter	Common collector	Common emitter	Common base
Z_{in}	High ($\simeq h_{fe} \times R_L$)	Medium ($\simeq$ 1k0)	Low ($\simeq$ 40R)
Z_{out}	Very low	$\simeq R_L$	$\simeq R_L$
A_V	$\simeq 1$	High	High
A_I	$\simeq h_{fe}$	$\simeq h_{fe}$	$\simeq 1$
Cut-off frequency	Medium	Low	High
Voltage phase shift	Zero	180°	Zero

FIGURE 15. Comparative performances of the three basic circuit configurations.

THE DIFFERENTIAL AMPLIFIER

Figure 16 shows — in basic form — how a pair of amplifiers of the basic **Figure 11** type can be coupled together to make a "differential" amplifier or "long-tailed pair" that produces an output signal that is proportional to the difference between the two input signals. In this case, Q1 and Q2 share a common emitter resistor (the "tail"), and the circuit is biased (via R1-R2 and R3-R4) so that the two transistors pass identical collector currents (thus giving zero difference between the two collector voltages) under quiescent zero-input conditions.

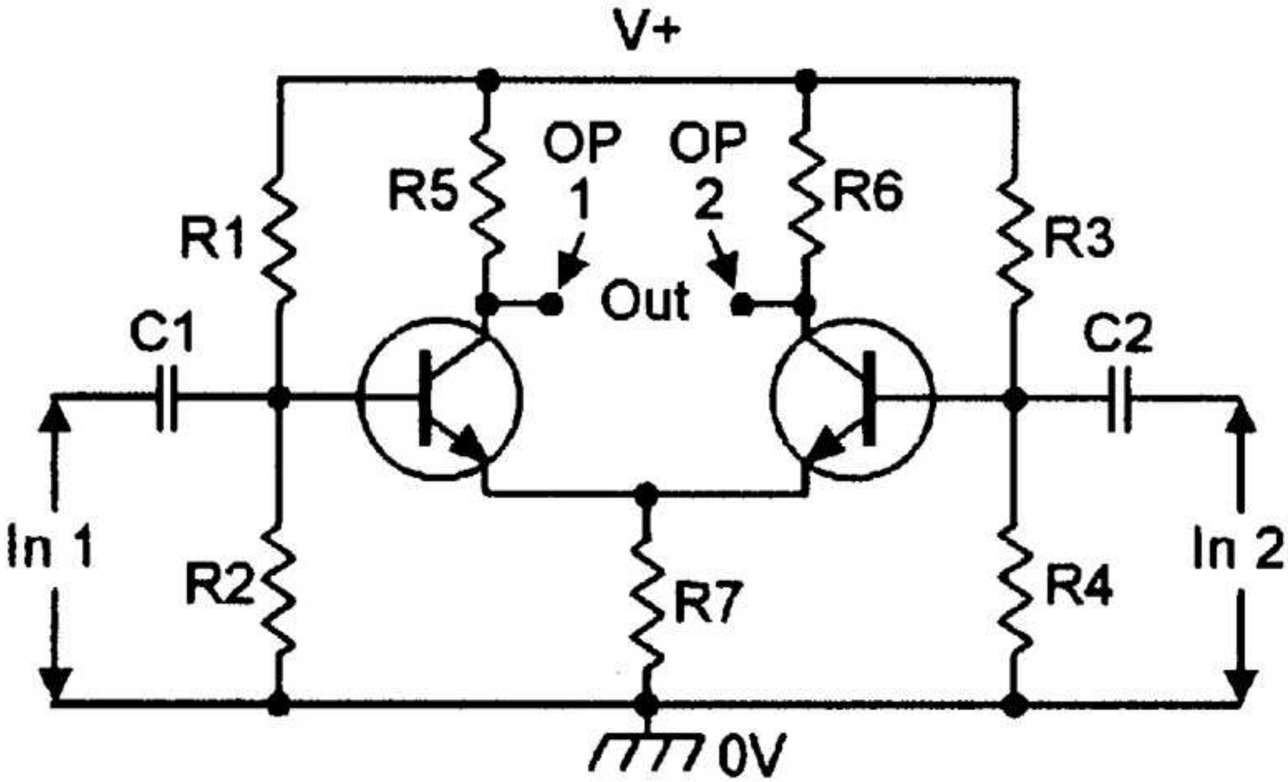


FIGURE 16. Differential amplifier or long-tailed pair.

If, in the above circuit, a rising input voltage is applied to the input of one transistor only, it makes the output voltage of that transistor fall and (as a result of emitter-coupling action) makes the output voltage of the other transistor rise by a similar amount, thus giving a large differential output voltage between the two collectors. If identical signals are applied to the inputs of both transistors, on the other hand, both collectors will move by identical amounts, and the circuit will thus produce a zero differential output signal. The circuit therefore produces an output signal that is proportional to the difference between the two input signals.

THE DARLINGTON CONNECTION

The input impedance of the **Figure 13** emitter follower circuit equals the product of R_L and the transistor's h_{fe} values — if an ultra-high input impedance is wanted, it can be obtained by replacing the single transistor with a pair of transistors connected in the "Darlington" or Super-Alpha configuration, as shown in **Figure 17**. Here, the emitter current of the input transistor feeds directly into the base of the output transistor, and the pair act like a single transistor with an overall h_{fe} value equal to the product of the two individual hfe values, i.e., if each transistor has an h_{fe} value of 100, the pair act like a single transistor with an h_{fe} of 10,000, and the overall circuit presents an input impedance of 10,000 x R_L.

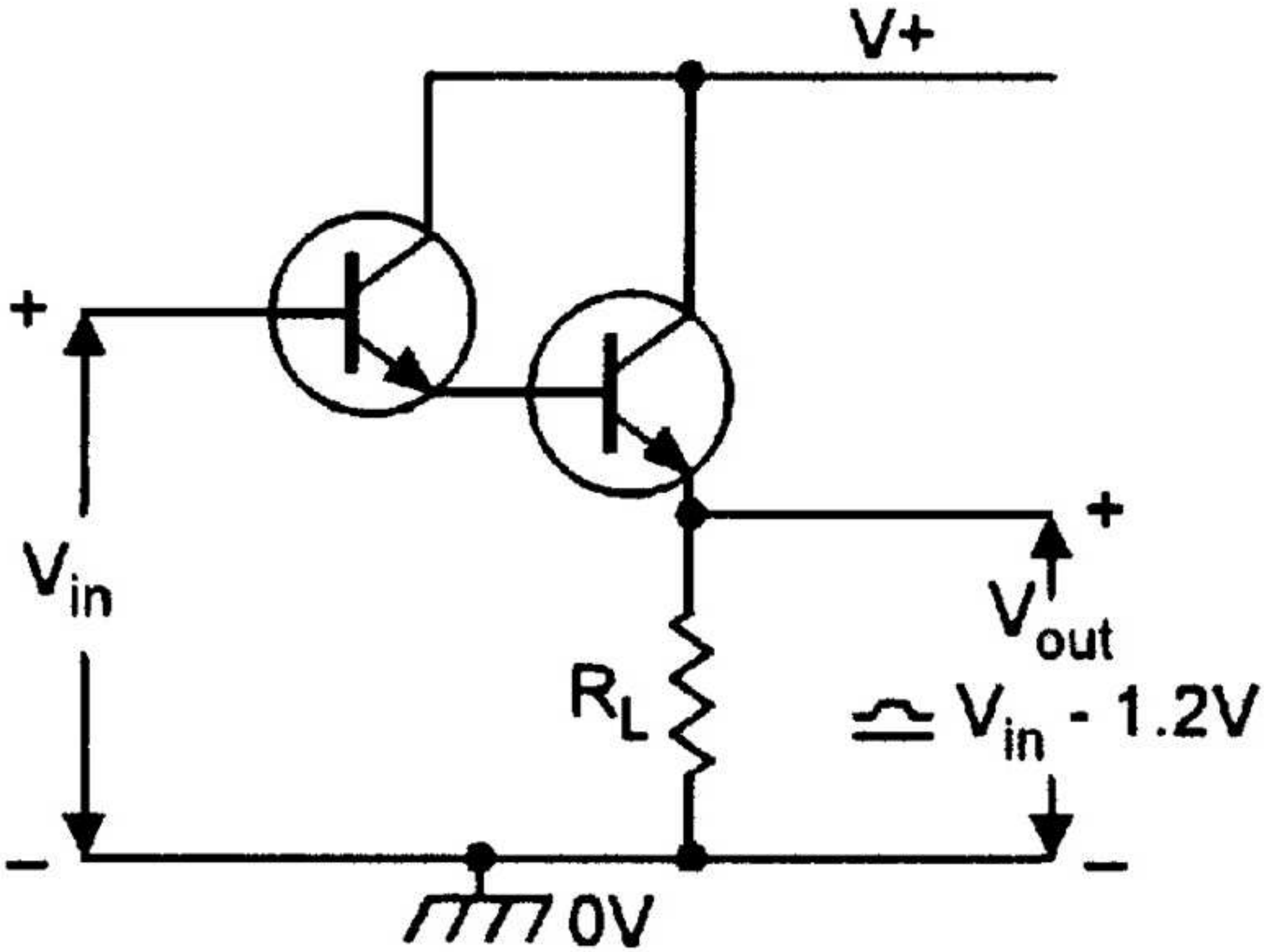


FIGURE 17. Darlington or Super-Alpha DC emitter follower.

MULTIVIBRATOR CIRCUITS

A multivibrator is, in essence, a two-state digital circuit that can be switched from the output-high to the output-low state, or vice versa, via a trigger signal that may be derived from an external source or via an automatic or triggered timing mechanism. Transistors can be used in four basic types of multivibrator circuits, as shown in **Figures 18 to 21**.

The **Figure 18** circuit is a simple, manually-triggered, cross-coupled bistable multivibrator, in which the base bias of each transistor is derived from the collector of the other, so that one transistor automatically turns off when the other turns on, and vice versa.

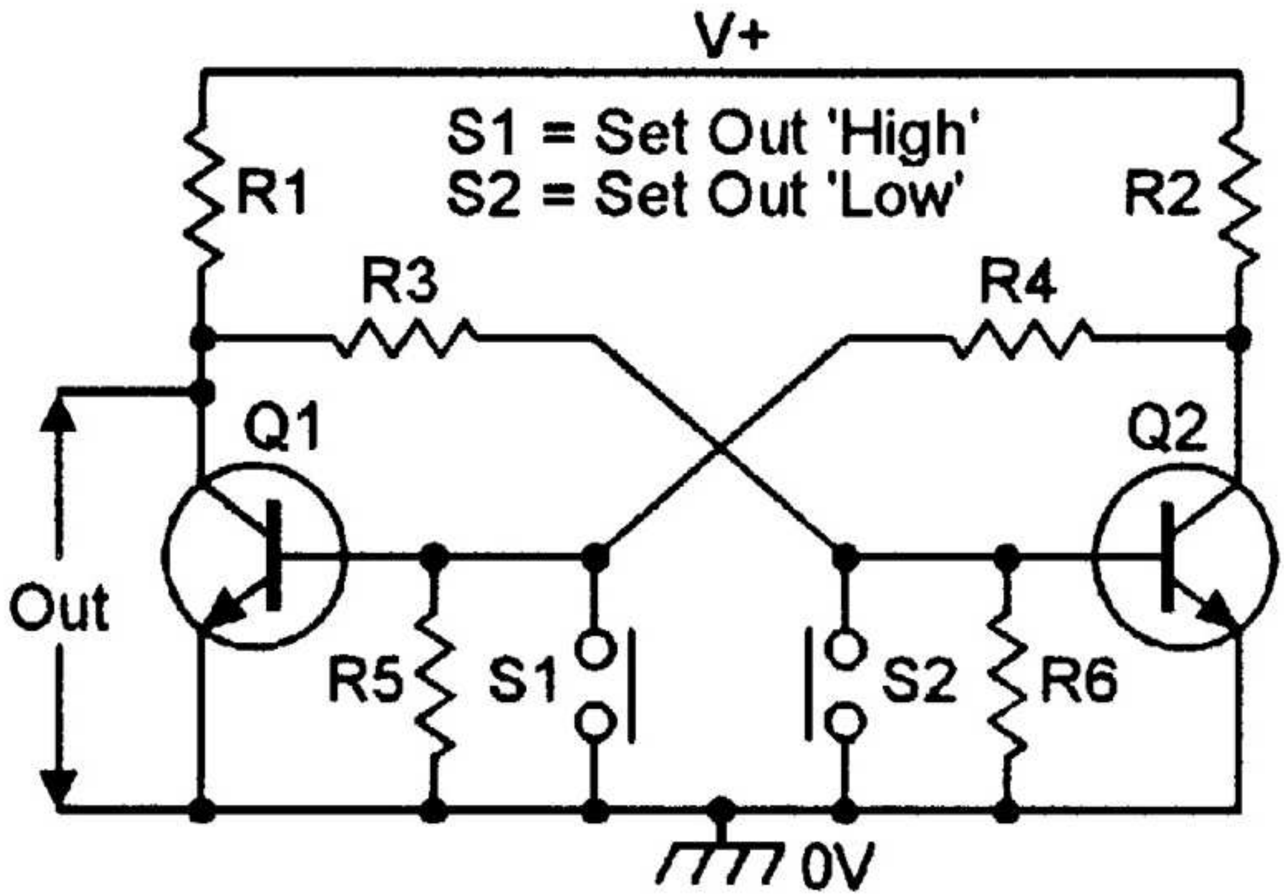


FIGURE 18. Manually-triggered bistable multivibrator.

Thus, the output can be driven low by briefly turning Q2 off via S2, thus shorting Q2's base-emitter path. As Q2 turns off R2-R4 feed base drive to Q1 base, the circuit automatically locks into this state until Q1 is similarly turned off via S1, at which point the output locks into the high state again, and so on ad infinitum.

Figure 19 shows — in basic form — a monostable multivibrator or one-shot pulse generator circuit. Its output (from Q1 collector) is normally low, since Q1 is normally biased on via R5, but switches high for a preset period (determined by the C1-R5 component values) if Q1 is briefly turned off by momentarily closing push-button "Start" switch S1.

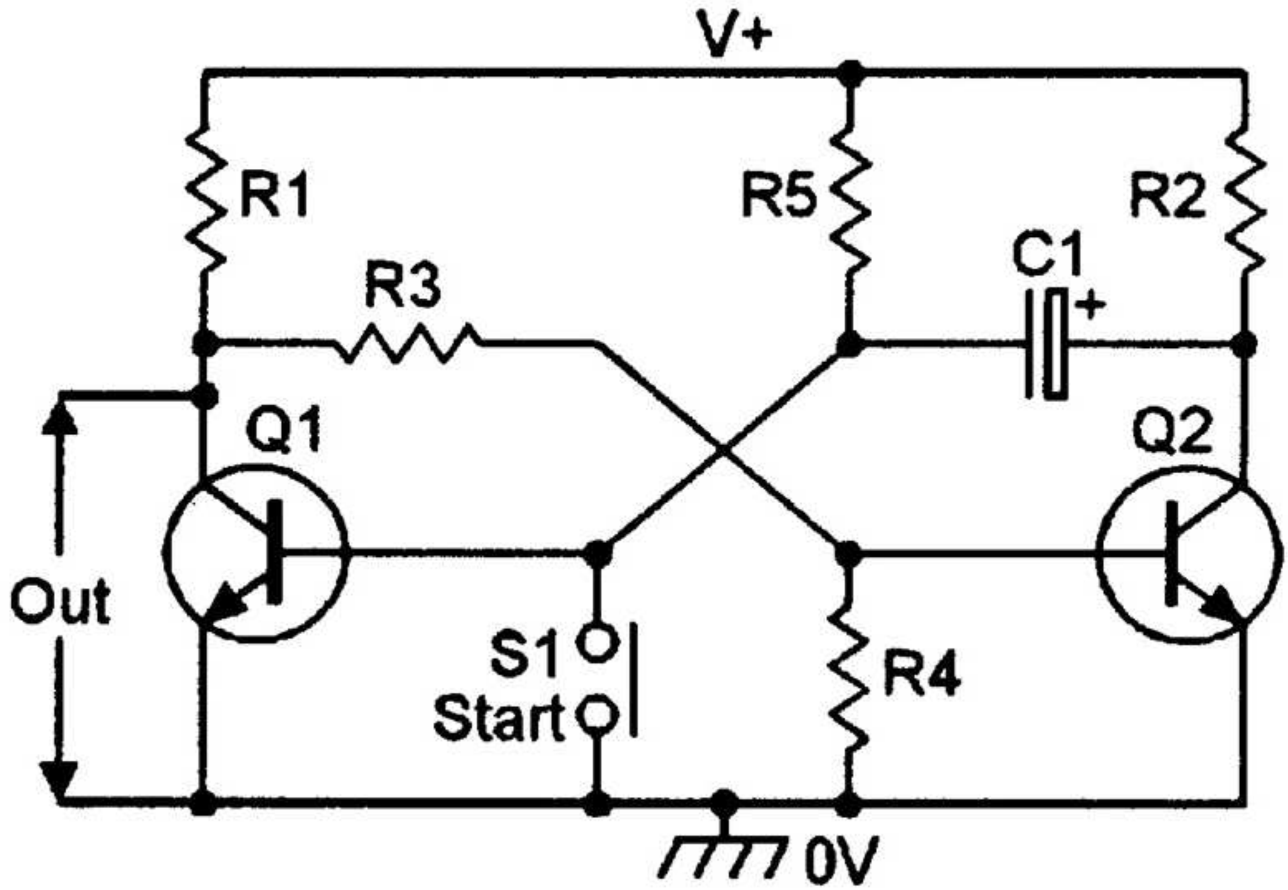


FIGURE 19. Manually-triggered monostable multivibrator.

The actual monostable timing period starts as the push-button “Start” switch is released, and has a period (P) of approximately 0.7 x C1 x R5, where P is in µS, C is in µF, and R is in kilohms.

Figure 20 shows an astable multivibrator, or free-running squarewave generator, in which the on and off periods of the squarewave are determined by the C1-R4 and C2-R3 component values. Basically, this circuit acts like a pair of cross-coupled monostable circuits, which automatically trigger each other sequentially. If the C1-R4 and C2-R3 timing periods are identical, the circuit generates a free-running squarewave output waveform. If the two timing periods are not identical, the circuit generates an asymmetrical output waveform.

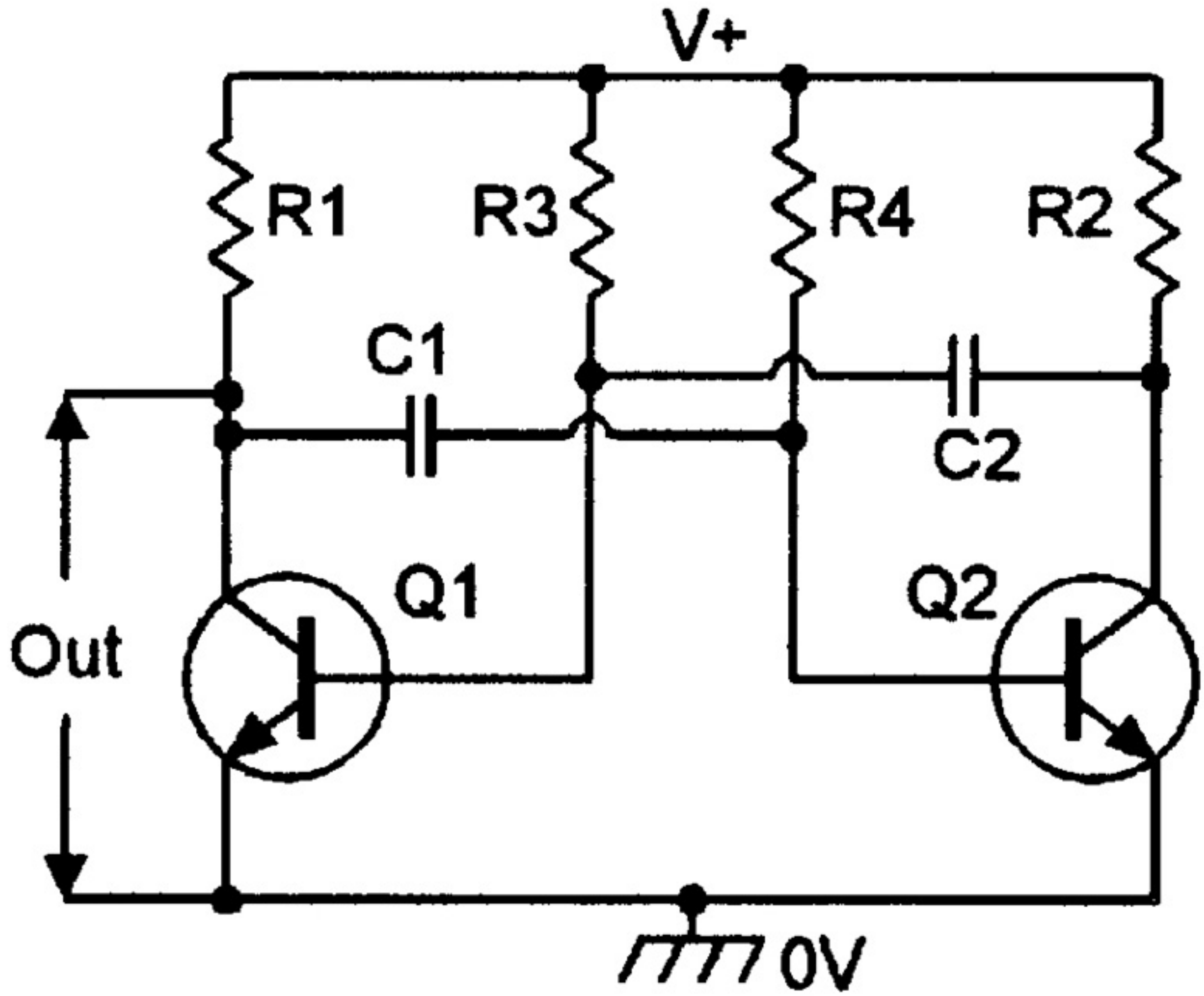


FIGURE 20. Astable multivibrator or free-running squarewave generator.

Finally, **Figure 21** shows a basic Schmitt trigger or sine-to-square waveform converter circuit. The circuit action here is such that Q2 switches abruptly from the “on” state to the “off” state, or vice versa, as Q1 base goes above or below pre-determined “trigger” voltage levels.

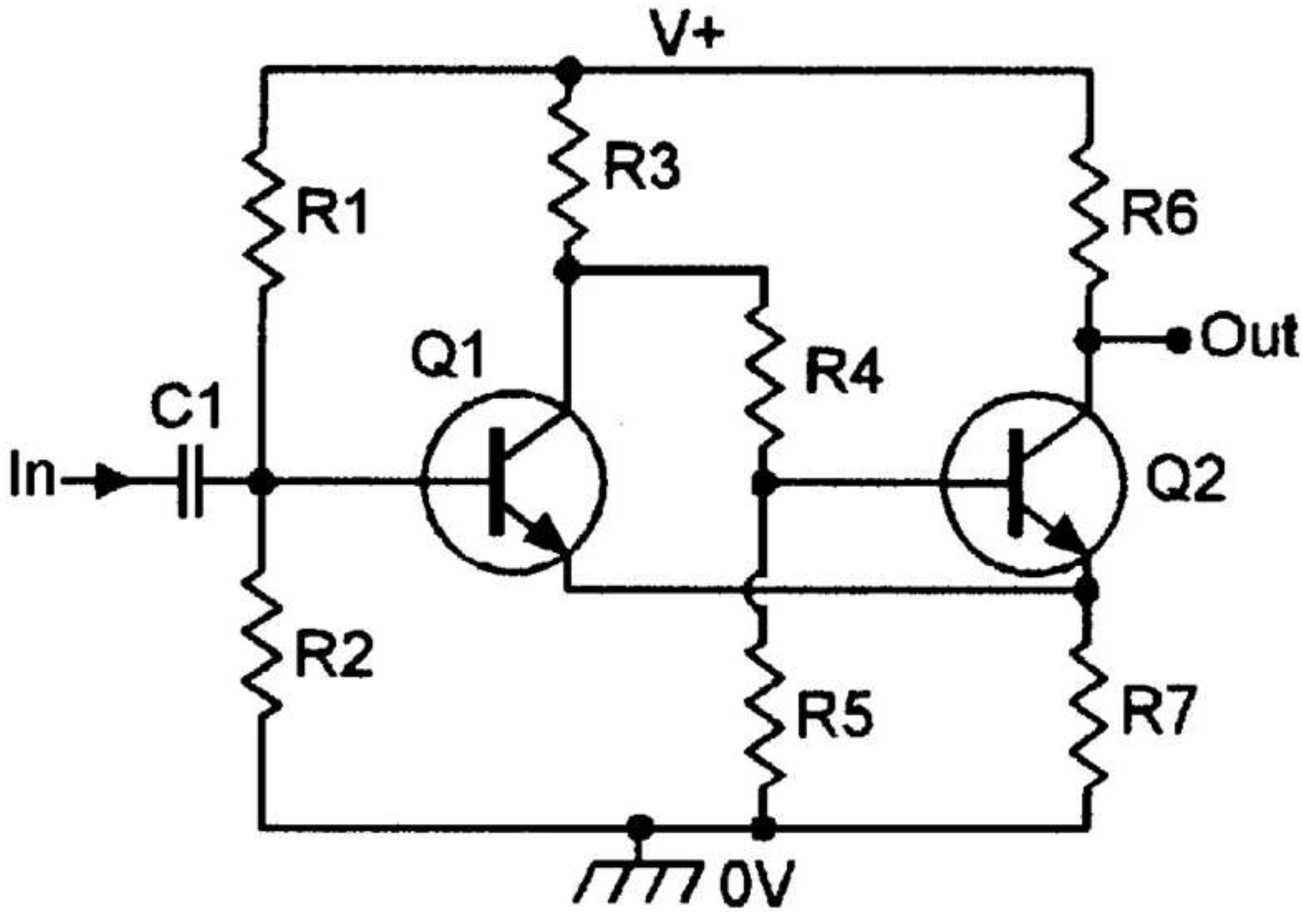


FIGURE 21. Schmitt trigger or sine-to-square waveform converter.

If the circuit's input is fed with a reasonable-amplitude sinewave input, the circuit thus generates a sympathetic squarewave output waveform. **NV**

Bipolar Transistor Cookbook — Part 2

Our first article gave an introductory outline of bipolar transistor principles, characteristics, and basic circuit configurations. This time, we'll concentrate on practical ways of using bipolar transistors in useful common-collector (voltage follower) circuit applications.

COMMON-COLLECTOR AMPLIFIER CIRCUIT

The common-collector amplifier (also known as the grounded-collector amplifier, emitter follower, or voltage follower) can be used in a wide variety of digital and analog amplifier and constant-current generator applications. This month, we start off by looking at practical "digital" amplifier circuits.

DIGITAL AMPLIFIERS

Figure 1 shows a simple NPN common-collector digital amplifier in which the input is either low (at zero volts) or high (at a V_{peak} value not greater than the supply rail value). When the input is low, Q1 is cut off and the output is at zero volts. When the input is high, Q1 is driven on and current I_L flows in R_L, thus generating an output voltage across RL — intrinsic negative feedback makes this output voltage take up a value one base-emitter junction volt-drop (about 600 mV) below the input V_{peak} value. Thus, the output voltage "follows" (but is 600 mV less than) the input voltage.

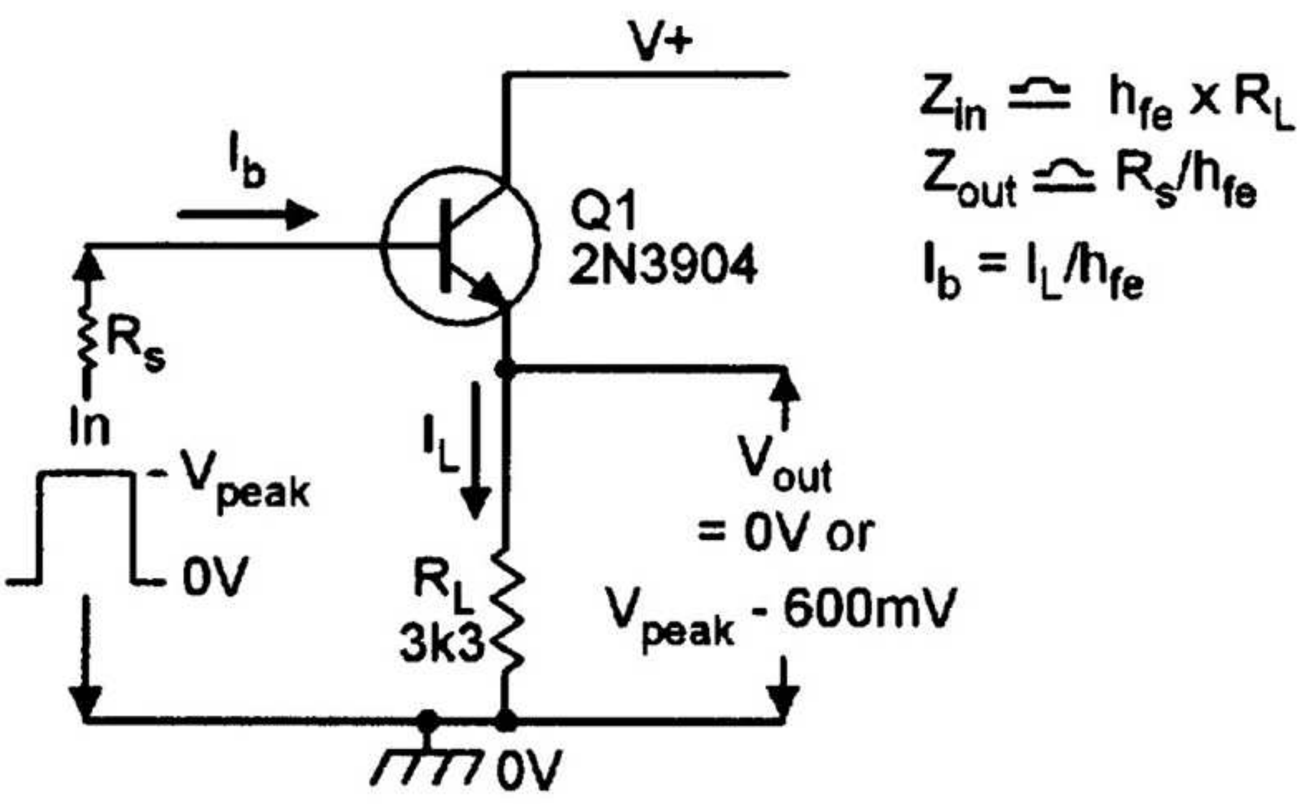


FIGURE 1. Common-collector digital amplifier basic details.

This circuit's input (base) current equals the I_L value divided by Q1's h_{fe} value (nominally

200 in the 2N3904), and its input impedance equals h_{fe} x R_L, i.e., nominally 660K in the example shown. The circuit's output impedance equals the input signal source impedance (R_S) value divided by h_{fe}. Thus, the circuit has a high input and low output impedance, and acts as a unity-voltage-gain "buffer" circuit.

If this buffer circuit is fed with a fast input pulse, its output may have a deteriorated falling edge, as shown in **Figure 2**. This deterioration is caused by the presence of stray capacitance (C_S) across R_L. When the input pulse switches high, Q1 turns on and rapidly "sources" (feeds) a charge current into C_S, thus giving an output pulse with a sharp leading edge. However, when the input signal switches low again, Q1 switches off and is thus unable to "sink" (absorb) the charge current of C_S, which thus discharges via R_L and makes the output pulse's trailing edge decay exponentially, with a time constant equal to the C_S-R_L product.

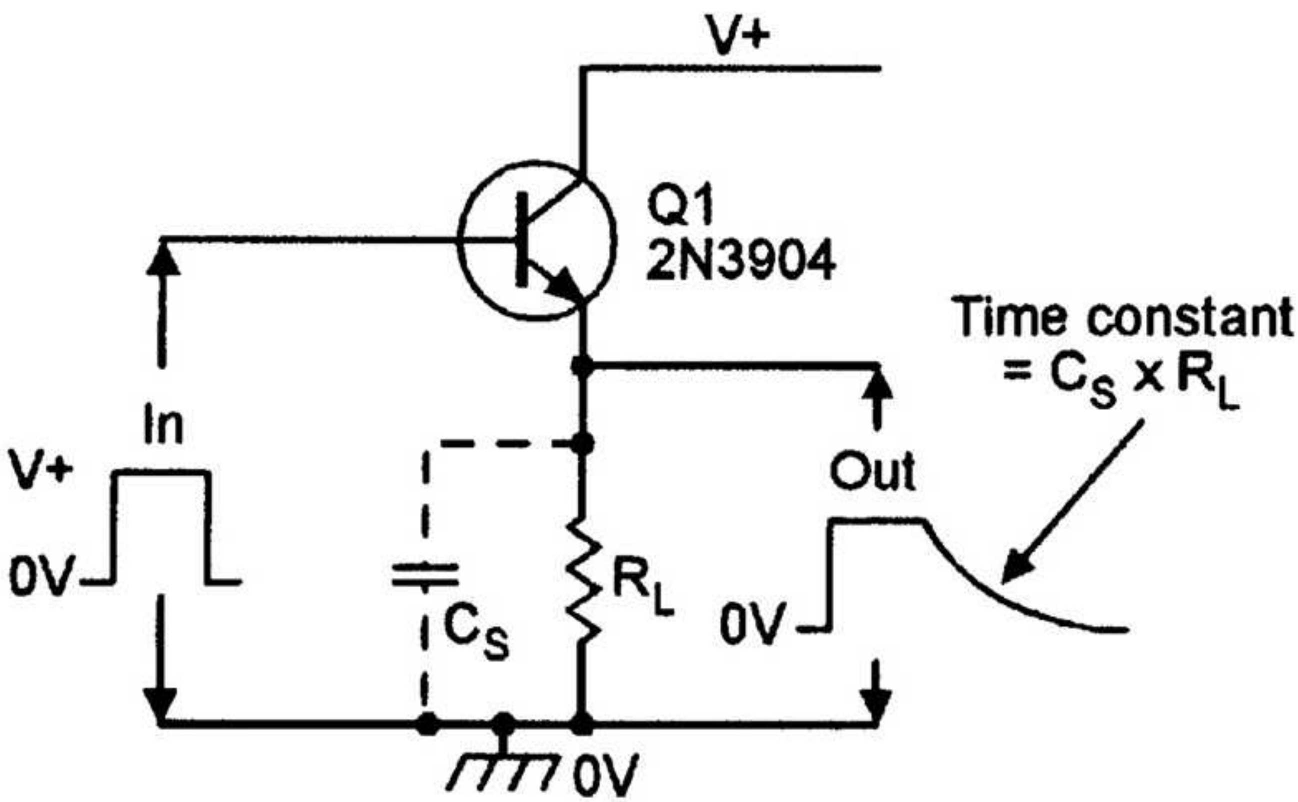


FIGURE 2. Effects of Cs on the circuit's output pulses in Figure 1.

Note from the above description that an NPN emitter follower can efficiently source (but not sink) high currents — a PNP emitter follower gives the opposite action, and can efficiently sink (but not source) high currents.

RELAY DRIVERS

If the basic **Figure 1** switching circuit is used to drive inductive loads such as coils or loudspeakers, etc., it must be fitted with a diode protection network to limit inductive switch-off back-EMFs to safe values. One very useful inductor-driving circuit is the relay driver, and a number of examples of this are shown in **Figures 3** to **7**.

The relay in the NPN driver circuit in **Figure 3** can be activated via a digital input or via switch SW1 — it turns on when the input signal is high or SW1 is closed, and turns off when the input signal is low or SW1 is open.

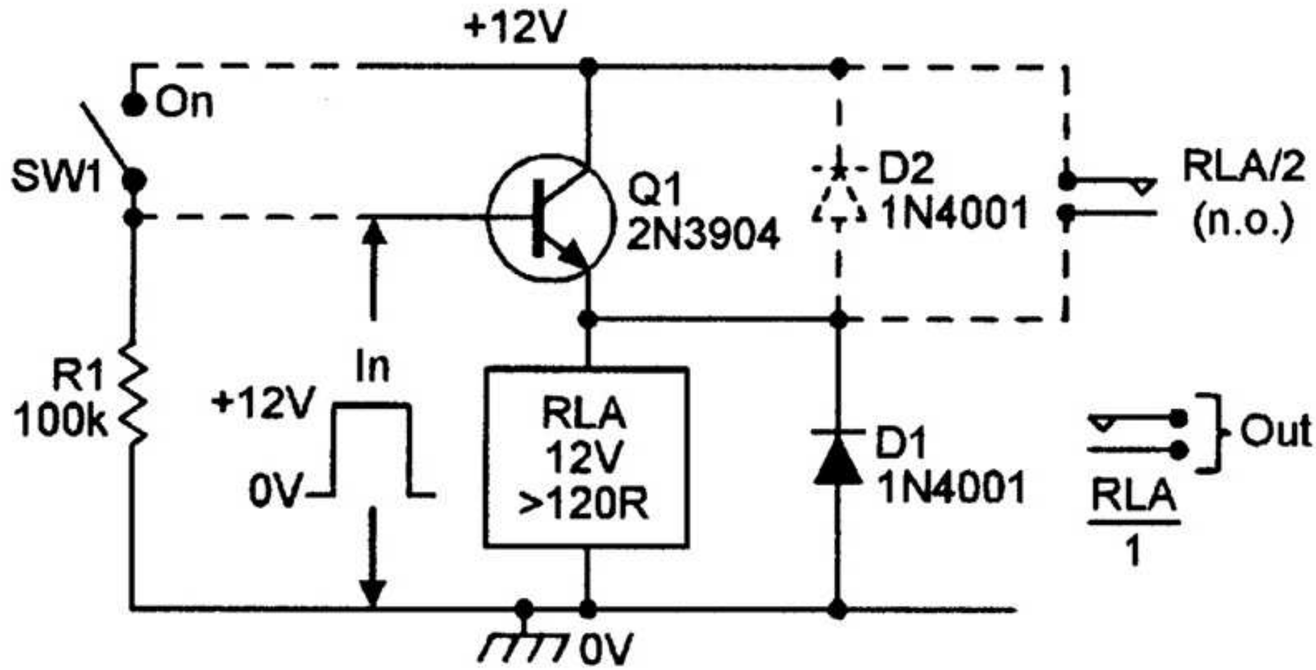


FIGURE 3. Simple emitter-follower relay driver.

Relay contacts RLA/1 are available for external use, and the circuit can be made self-latching by wiring a spare set of normally-open relay contacts (RLA/2) between Q1's collector and emitter, as shown dotted. **Figure 4** is a PNP version of the same circuit; in this case, the relay can be turned on by closing SW1 or by applying a "zero" input signal. *Note in **Figure 3** that D1 damps relay switch-off back-emfs by preventing this voltage from swinging below the zero-volts rail value. Optional diode D2 can be used to stop this voltage swinging above the positive rail.*

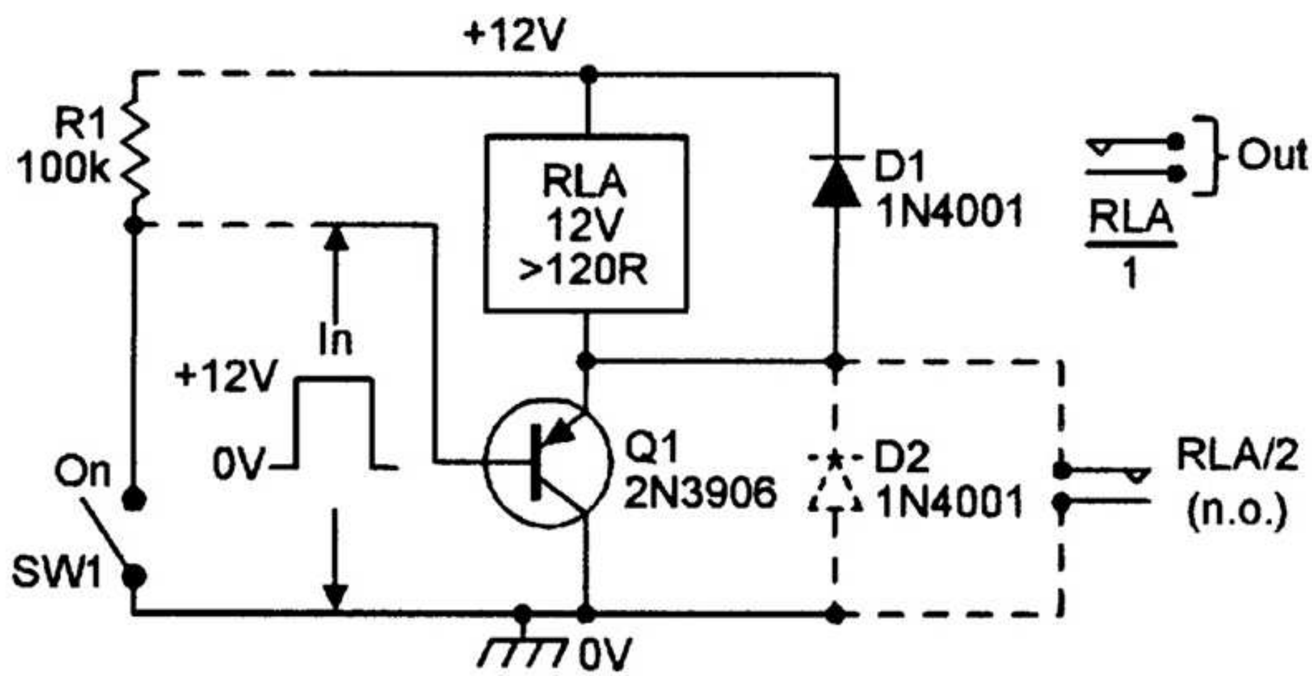


FIGURE 4. PNP version of the relay driver.

The circuits shown in **Figures 3** and **4** effectively increase the relay current sensitivity by a factor of about 200 (the h_{fe} value of Q1), e.g., if the relay has a coil resistance of 120R and needs an activating current of 100 mA, the circuit's input impedance is 24K and the input operating current requirement is 0.5 mA. Sensitivity can be further increased by using a Darlington pair of transistors in place of Q1 (as shown in **Figure 5**), but the emitter "following" voltage of Q2 will be 1.2V (two base-emitter volt drops) below the base input voltage of Q1.

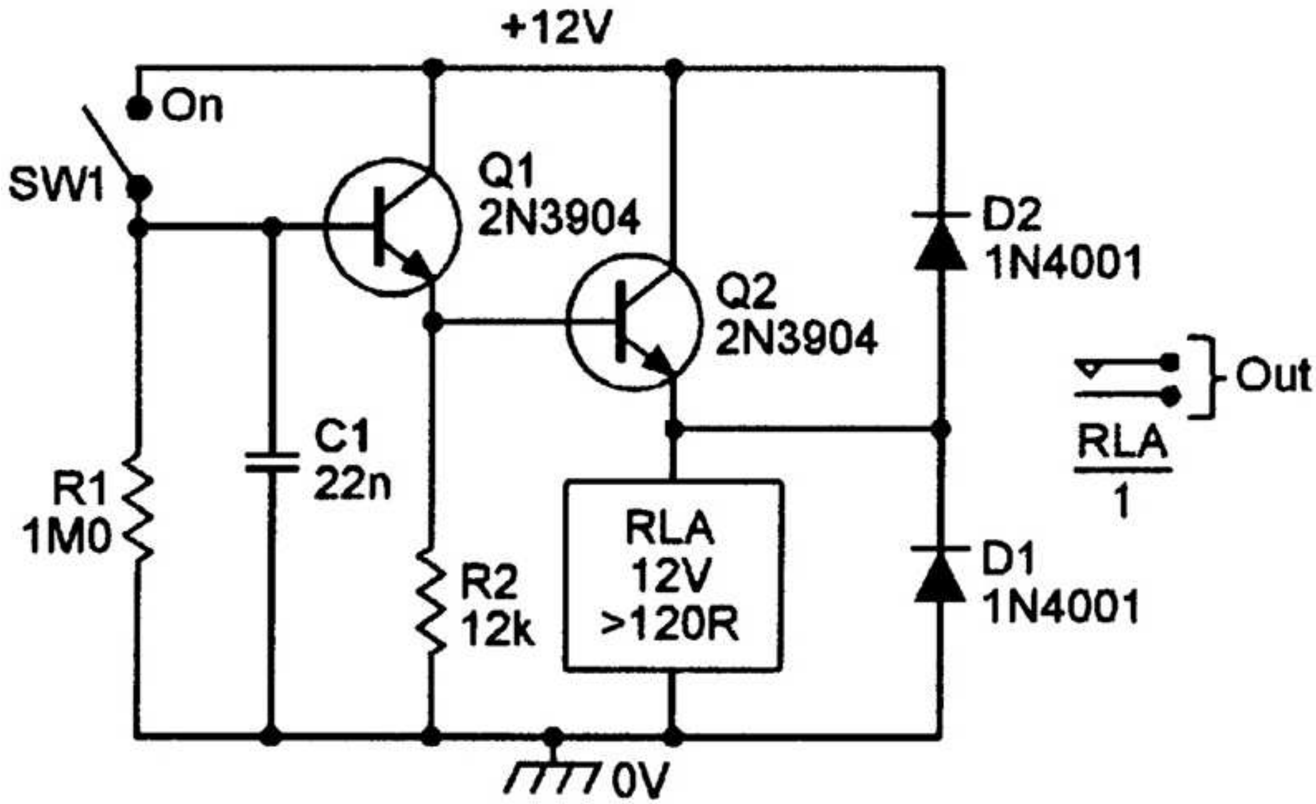


FIGURE 5. Darlington version of the NPN relay driver.

This circuit has an input impedance of 500K and needs an input operating current of 24 µA — C1 protects the circuit against activation via high-impedance transient voltages, such as those induced by lightening flashes, RFI, etc. The Darlington buffer is useful in relay-driving C-R time-delay designs such as those shown in **Figures 6** and **7**, in which C1-R1 generate an exponential waveform that is fed to the relay via Q1-Q2, thus making the relay change state some delayed time after the supply is initially connected. With an R1 value of 120k, the circuits give operating delays of roughly 0.1 seconds per µF of C1 value, i.e., a 10 second delay if C1 = 100 µF, etc. The **Figure 6** circuit makes the relay turn on some delayed time after its power supply is connected. The **Figure 7** circuit makes the relay turn on as soon as the supply is connected, but turn off again after a fixed delay.

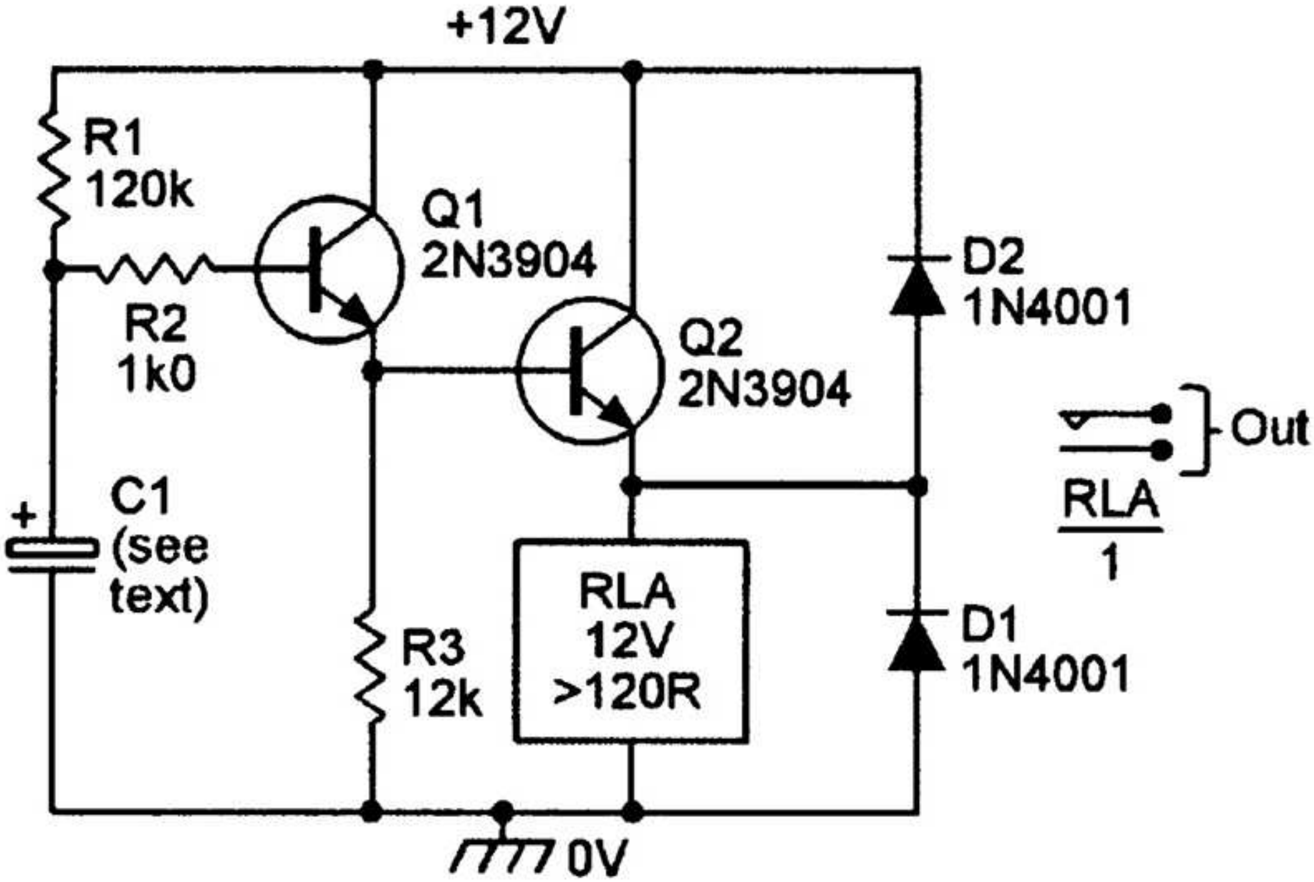


FIGURE 6. Delayed switch-on relay driver.

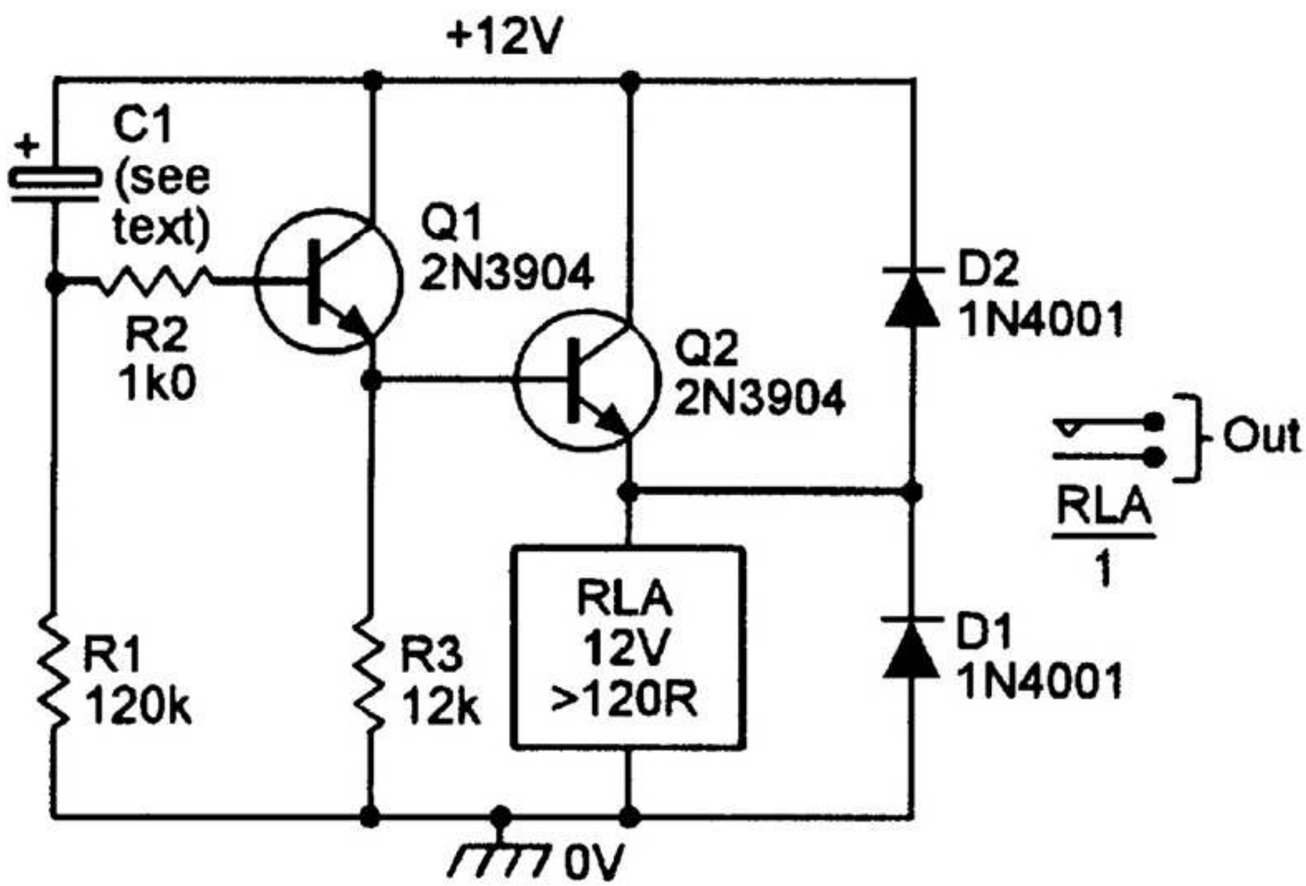


FIGURE 7. Auto-turn-off time-delay circuit.

CONSTANT-CURRENT GENERATORS

A constant-current generator (CCG) is a circuit that generates a constant load current irrespective of wide variations in load resistance. A bipolar transistor can be used as a CCG by using it in the common-collector mode shown in **Figure 8**. Here, R1-ZD1 applies a fixed 5.6V “reference” to Q1 base, making 5V appear across R2, which thus passes 5mA via Q1’s emitter. A transistor’s emitter and collector currents are inherently almost identical, so a 5mA current also flows in any load connected between Q1’s collector and the positive supply rail, provided that its resistance is not so high that Q1 is driven into saturation.

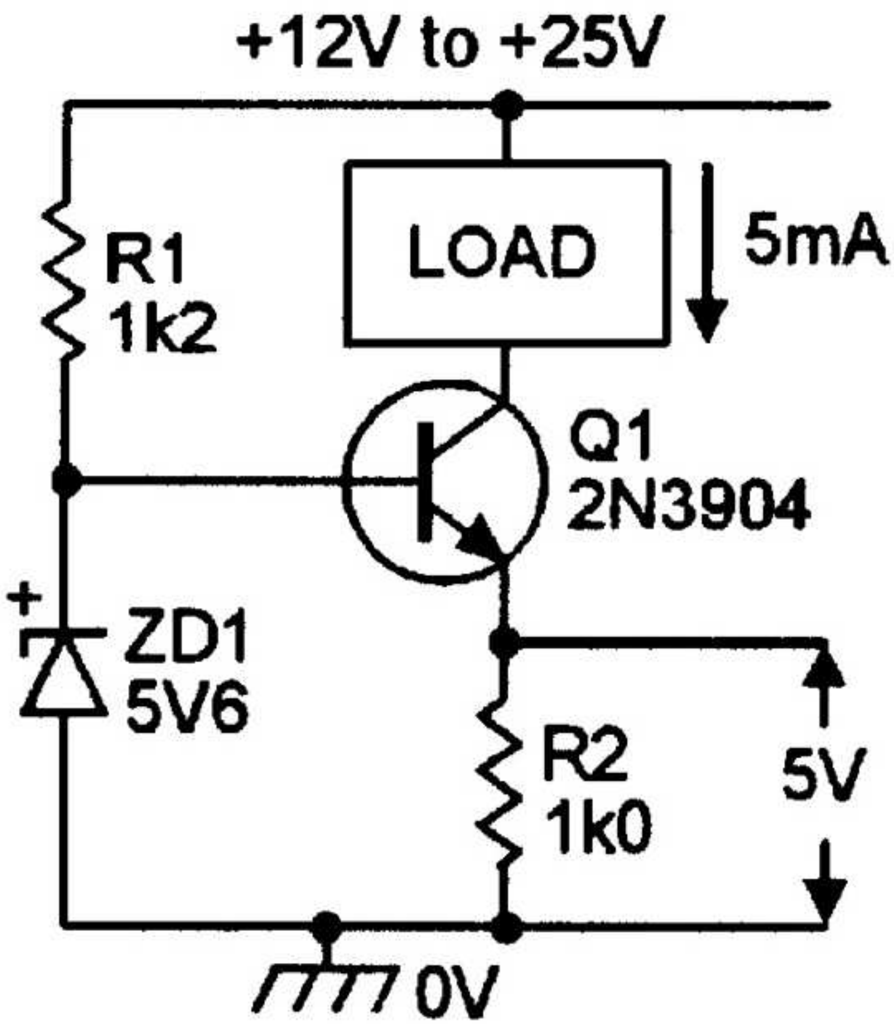


FIGURE 8. Simple 5 mA constant-current generator.

These two points thus act as 5 mA “constant-current” terminals. This circuit’s constant-current value is set by Q1’s base voltage and the R2 value, and can be altered by varying either of these values. **Figure 9** shows how the basic circuit can be “inverted” to give a ground-referenced, constant-current output that can be varied from about 1 mA to 10 mA via RV1.

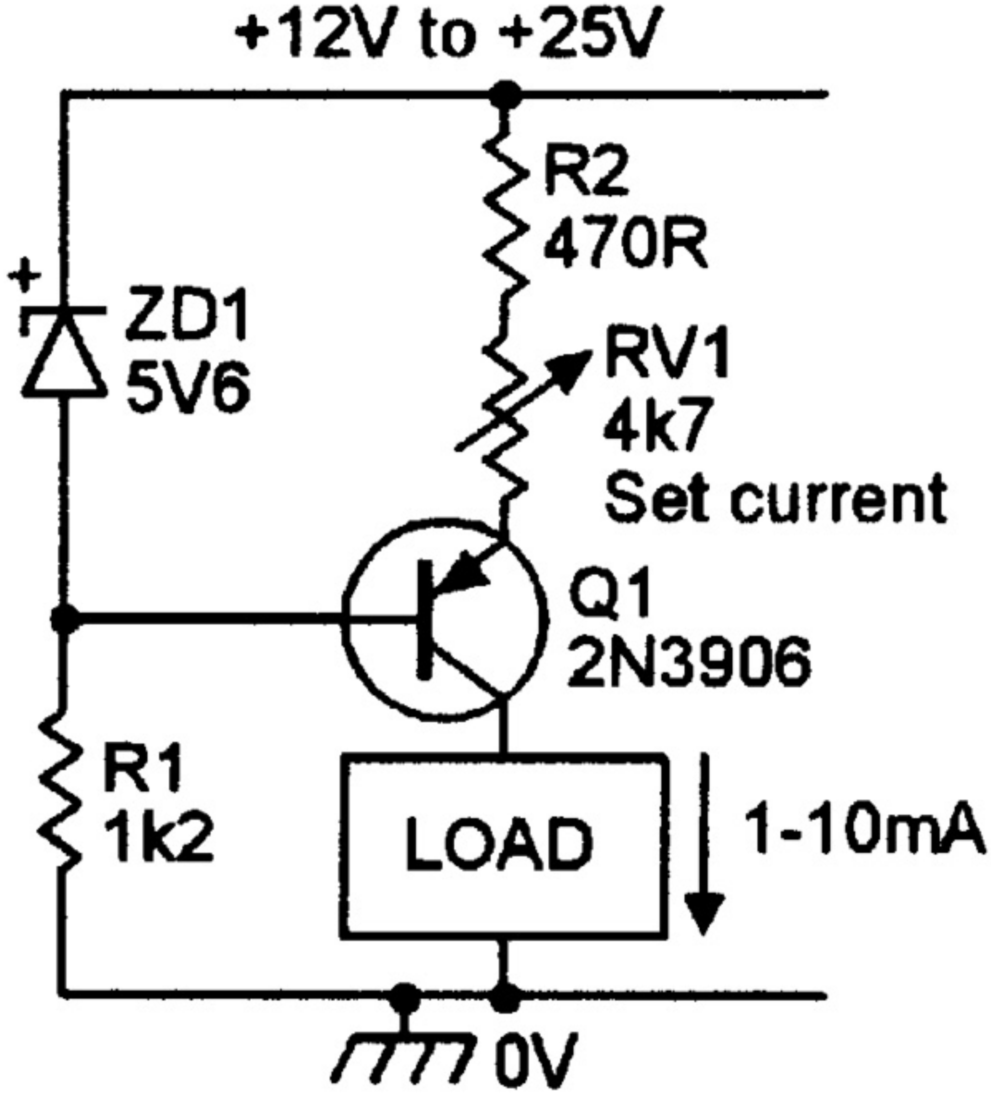


FIGURE 9. Ground-referenced variable (1 mA-10 mA) constant-current generator.

In many practical CCG applications, the circuit's most important feature is its high dynamic output impedance or "current constancy" — the precise current magnitude being of minor importance — in such cases the basic **Figure 8** and **9** circuits can be used. If greater precision is needed, the "reference" voltage accuracy must be improved. One way of doing this is to replace R1 with a 5 mA constant-current generator, as indicated in **Figure 10** by the "double circle" symbol, so that the zener current (and thus voltage) is independent of supply voltage variations.

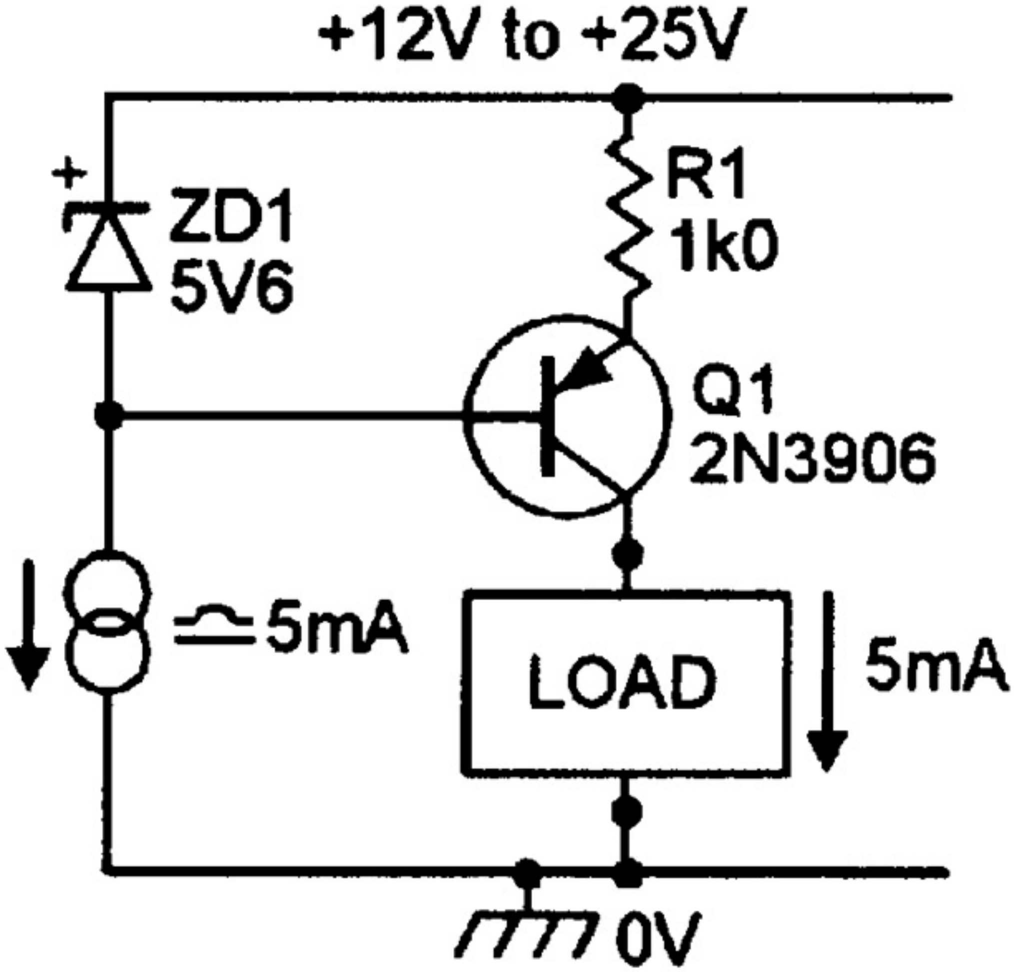


FIGURE 10. Precision constant-current generator.

A red LED acts as an excellent reference voltage generator, and has a very low temperature coefficient, and can be used in place of a zener, as shown in **Figure 11**.

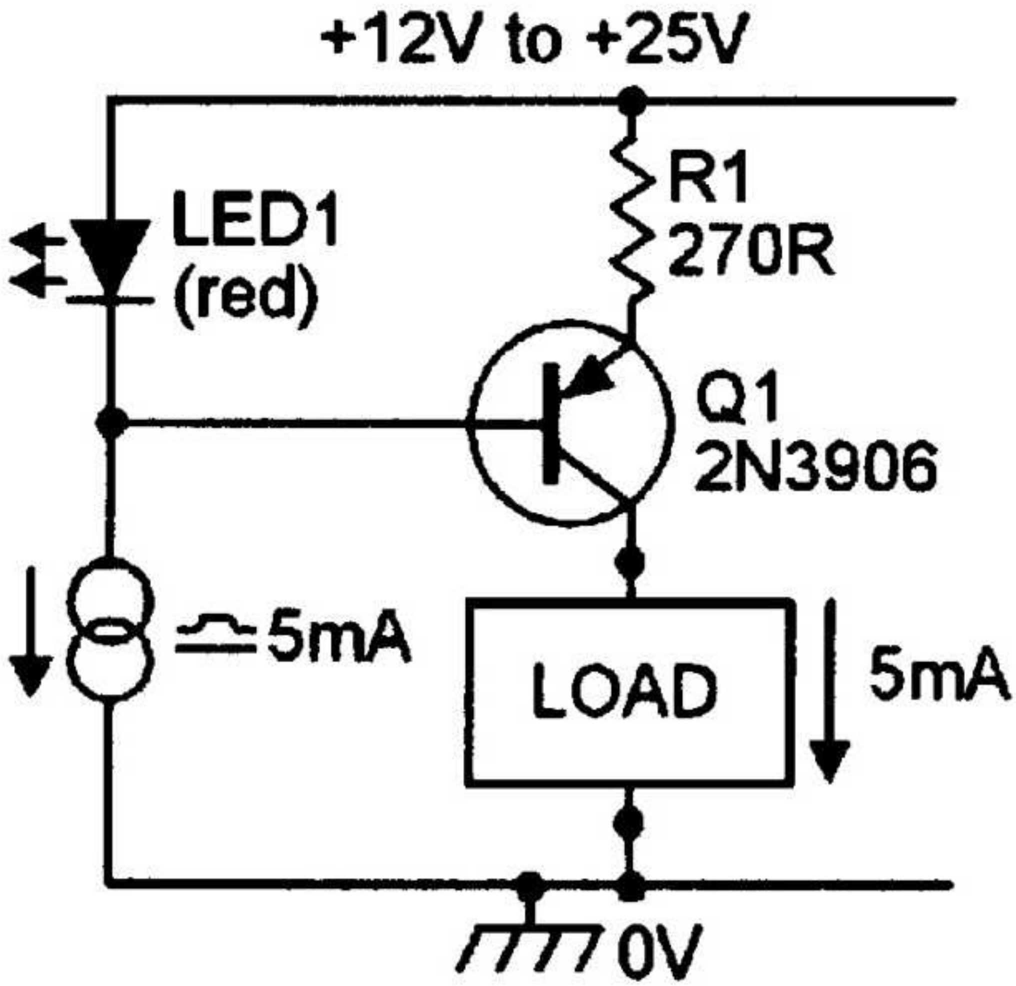


FIGURE 11. Thermally-stabilized constant-current generator, using an LED as a voltage reference.

In this case, the LED generates roughly 2.0V, so only 1.4V appears across R1, which has its value reduced to 270R to give a constant-current output of 5 mA. The CCG (constant-current generator) circuits in **Figures 8** through **11** are all “three-terminal” designs that need both supply and output connections. **Figure 12** shows a two-terminal CCG that consumes a fixed 2 mA when wired in series with an external load.

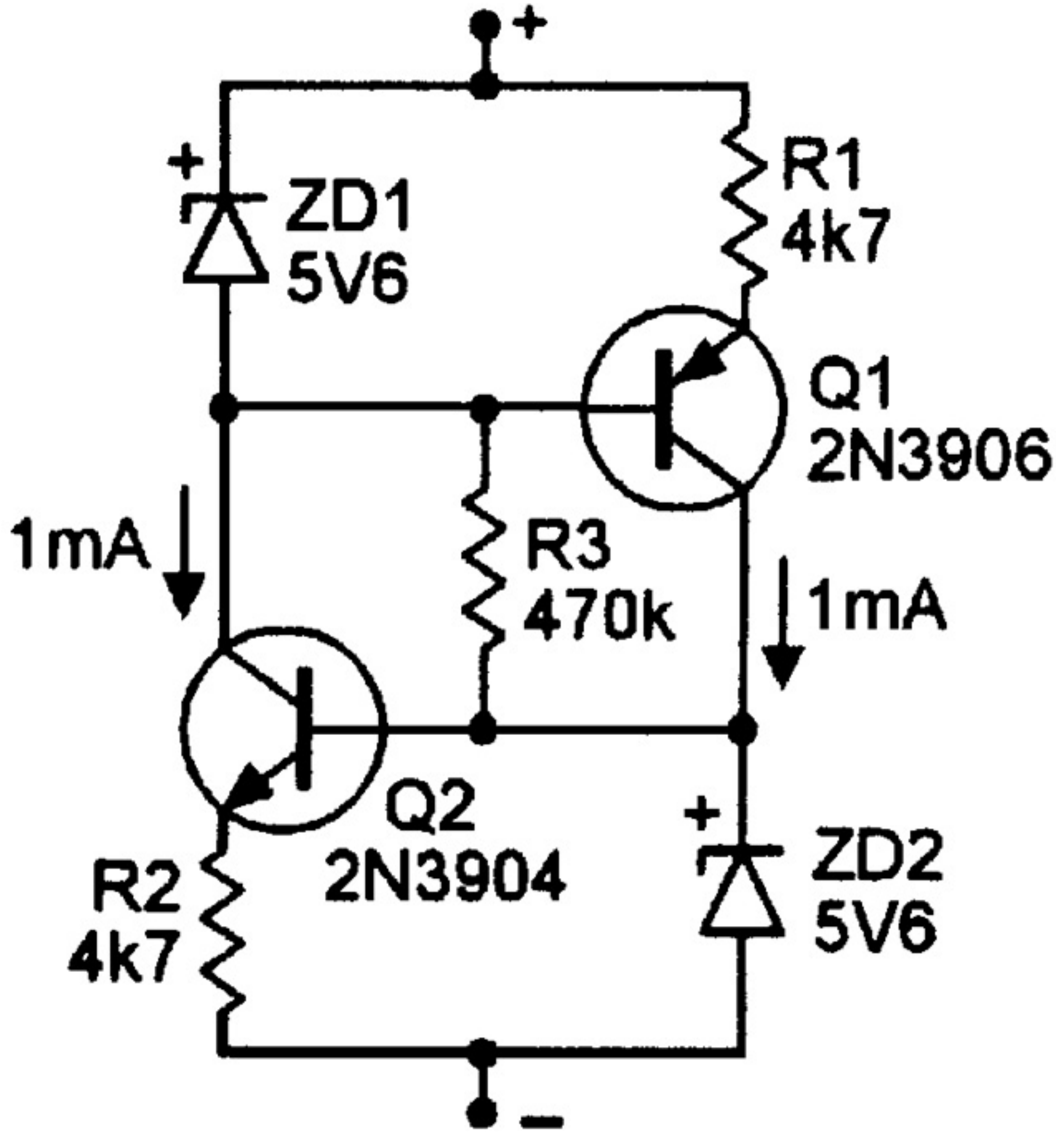


FIGURE 12. Two-terminal 2 mA constant-current generator.

Here, ZD1 applies 5.6V to the base of Q1, which (via R1) generates a constant collector current of 1 mA — this current drives ZD2, which thus develops a very stable 5.6V on the base of Q2 which, in turn, generates a constant collector current of about 1mA, which drives ZD1. The circuit thus acts as a closed-loop current regulator that consumes a total of 2 mA. R3 acts as a start-up resistor that provides the transistor with initial base current. **Figure 13** shows a version of the two-terminal CCG in which the operating current is fully variable from 1 mA to 10 mA via dual ganged variable resistor RV1. *Note that these two circuits each need a minimum operating voltage, between their two main terminals of about 12V, but can operate with maximum ones of 40V.*

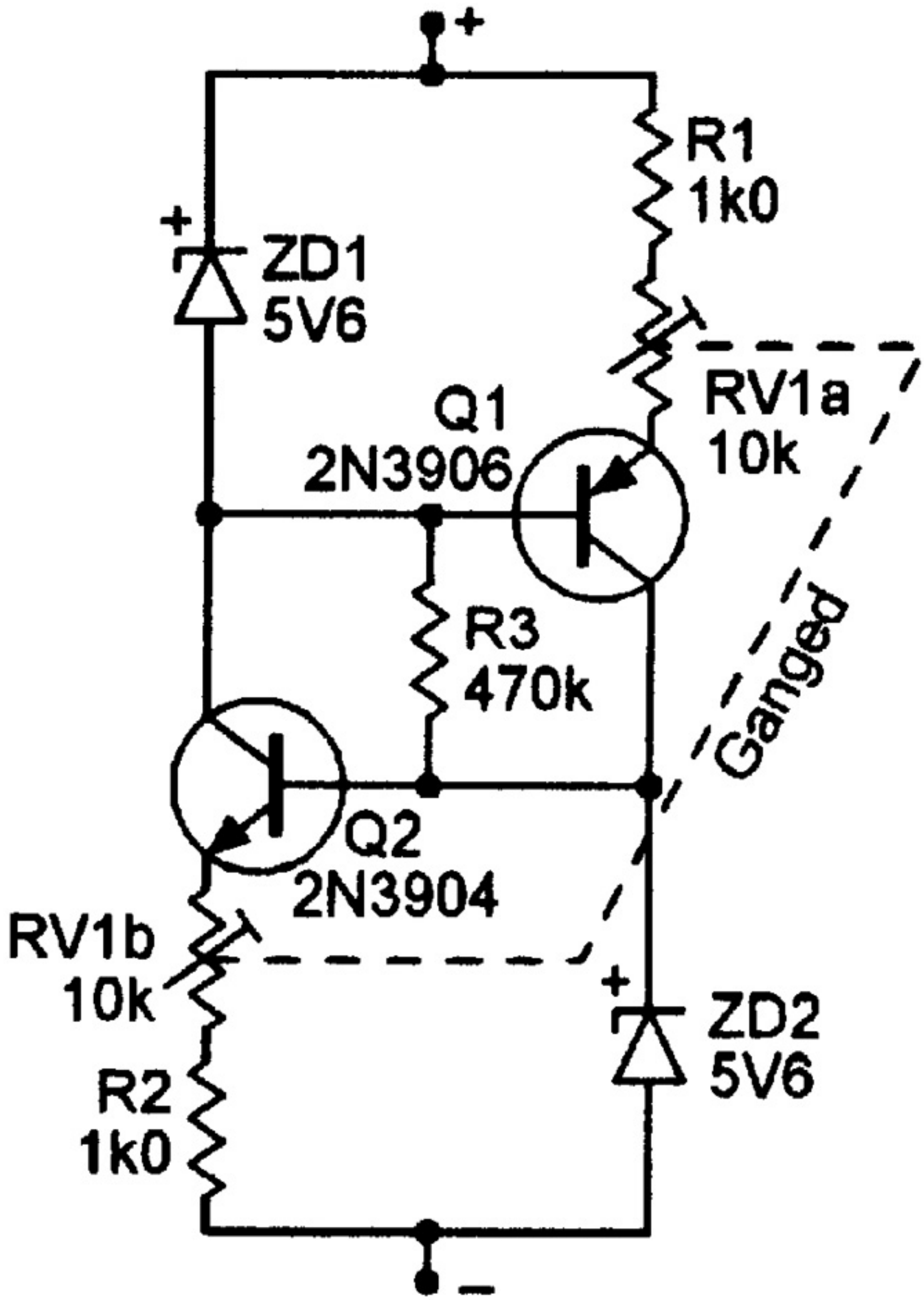


FIGURE 13. Two-terminal variable (1 mA-10 mA) constant-current generator.

LINEAR AMPLIFIERS

A common-collector circuit can be used as an AC-coupled linear amplifier by biasing its base to a quiescent half-supply voltage value (to accommodate maximal signal swings) and AC-coupling the input signal to its base and taking the output signal from its emitter, as shown in the basic circuits in **Figures 14** and **15**. **Figure 14** shows the simplest possible version of the linear emitter follower, with Q1 biased via a single resistor (R1). To achieve half-supply biasing, R1’s value must (ideally) equal Q1’s input resistance — the biasing level is thus dependent on Q1’s h_{fe} value.

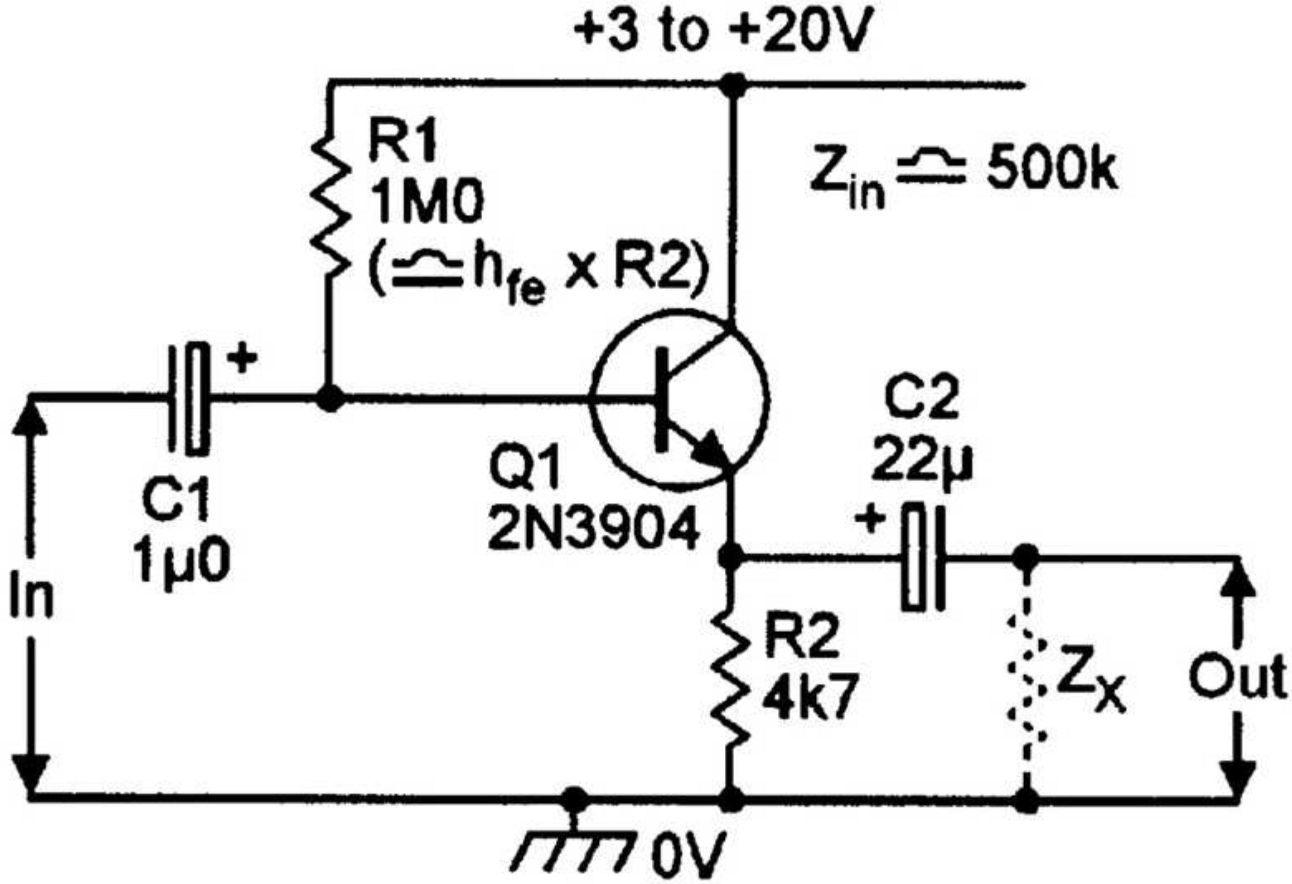


FIGURE 14. Simple emitter follower.

Figure 15 shows an improved version of the basic circuit in which R1-R2 applies a quiescent half-supply voltage to Q1 base, irrespective of variations in Q1’s h_{fe} values. Ideally, R1 should equal the paralleled values or R2 and R_{IN}, but in practice, it is adequate to simply make R1 low relative to RIN, and to make R2 slightly larger than R1.

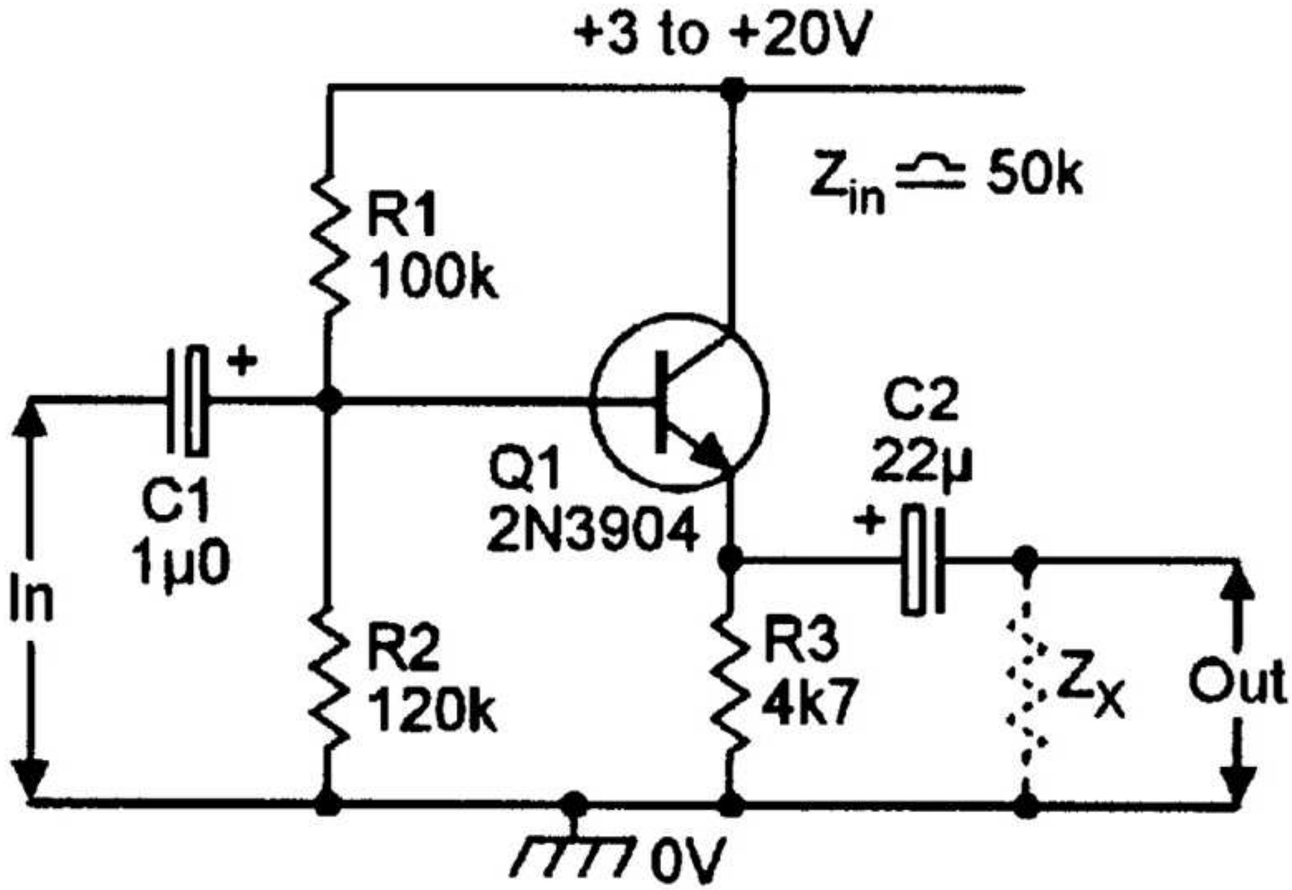


FIGURE 15. High-stability emitter follower.

In these two circuits, the input impedance looking directly into Q1 base equals h_{fe} x Z_{load}, where (in the basic **Figure 14** circuit) Z_{load} is the parallel impedance of R2 and external output load Z_X. Thus, the base impedance value is roughly 1M0 when Z_X is infinite. The input impedance of the complete circuit equals the parallel impedances of the base impedance and the bias network. The circuit in **Figure 14** gives an input impedance of about 500K, and the circuit in **Figure 15** is about 50K. Both circuits give a voltage gain (A_V) that is slightly below unity, the actual gain being given by:

$$A_V = Z_{load}/(Z_b + Z_{load})$$

where $Z_b = 25/I_c$ ohms, and where I_c is the collector current (which is the same as the emitter current) in mA. Thus, at an operating current of 1 mA these circuits give a gain of 0.995 when Z_{load} = 4k7, or 0.975 when Z_{load} = 1k0.

BOOTSTRAPPING

The **Figure 15** circuit's input impedance can easily be boosted by using the basic "bootstrapping" technique shown in **Figure 16**. Here, 47K resistor R3 is wired between the R1-R2 biasing network junction and Q1 base, and the input signal is fed to Q1 base via C1. Note, however, that Q1's output is fed back to the R2-R2 junction via C2, and near-identical signal voltages thus appear at both ends of R3 — very little signal current flows in R3, which appears (to the input signal) to have a far greater impedance than its true resistance value.

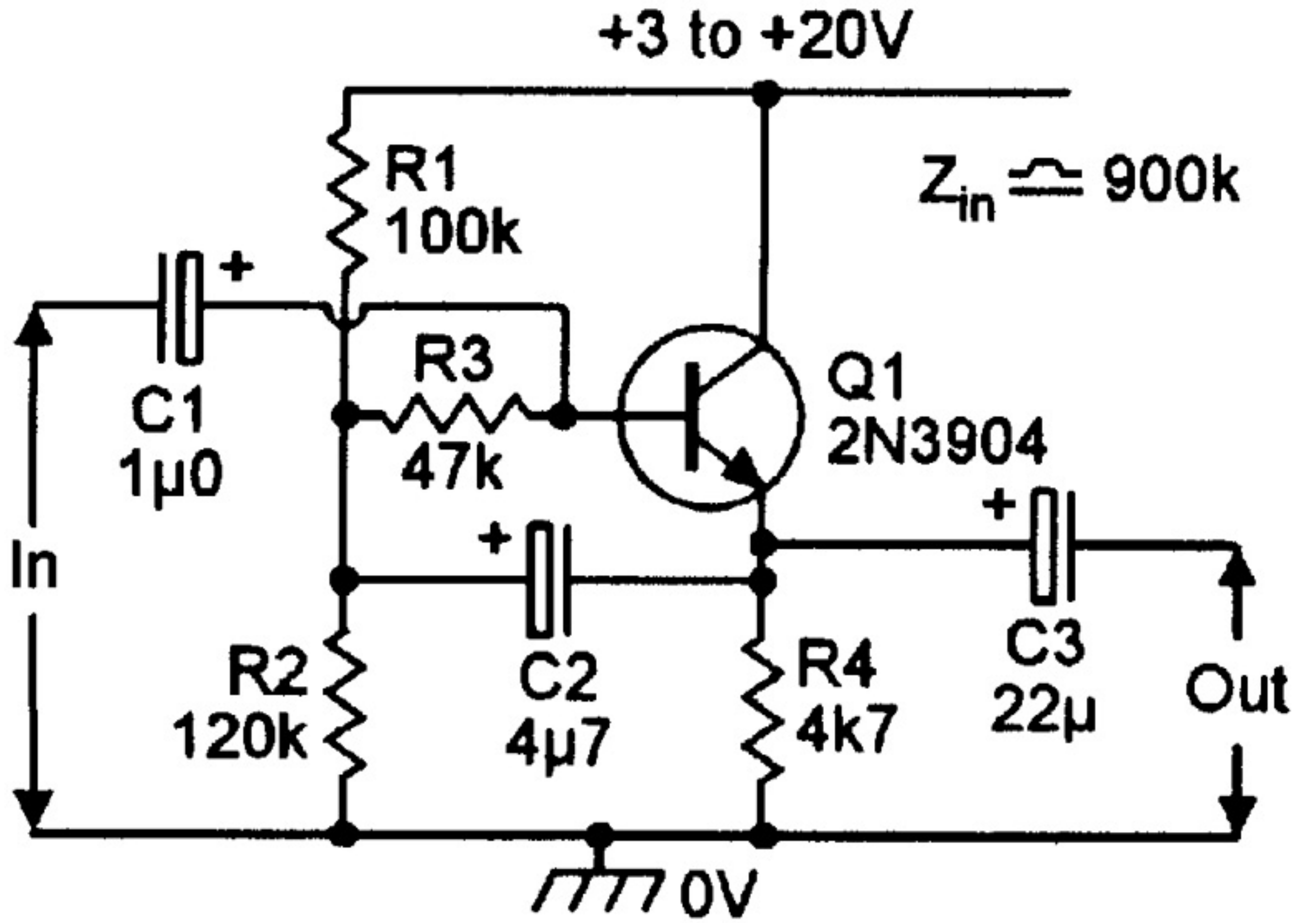


FIGURE 16. Bootstrapped emitter follower.

All practical emitter followers give an A_V of less than unity, and this value determines the resistor "amplification factor," or A_R of the circuit, as follows:

$A_R = 1/(1 - A_V)$

Thus, if the circuit has an A_V or 0.995, A_R equals 200 and the R3 impedance is almost 10M. This impedance is in parallel with R_{IN}, so the **Figure 16** circuit has an input impedance of roughly 900K. The input impedance of the circuit in **Figure 16** can be increased even more by using a pair of Darlington-connected transistors in place of Q1, and increasing the value of R3, as shown in **Figure 17**, which gives a measured input impedance of about 3M3.

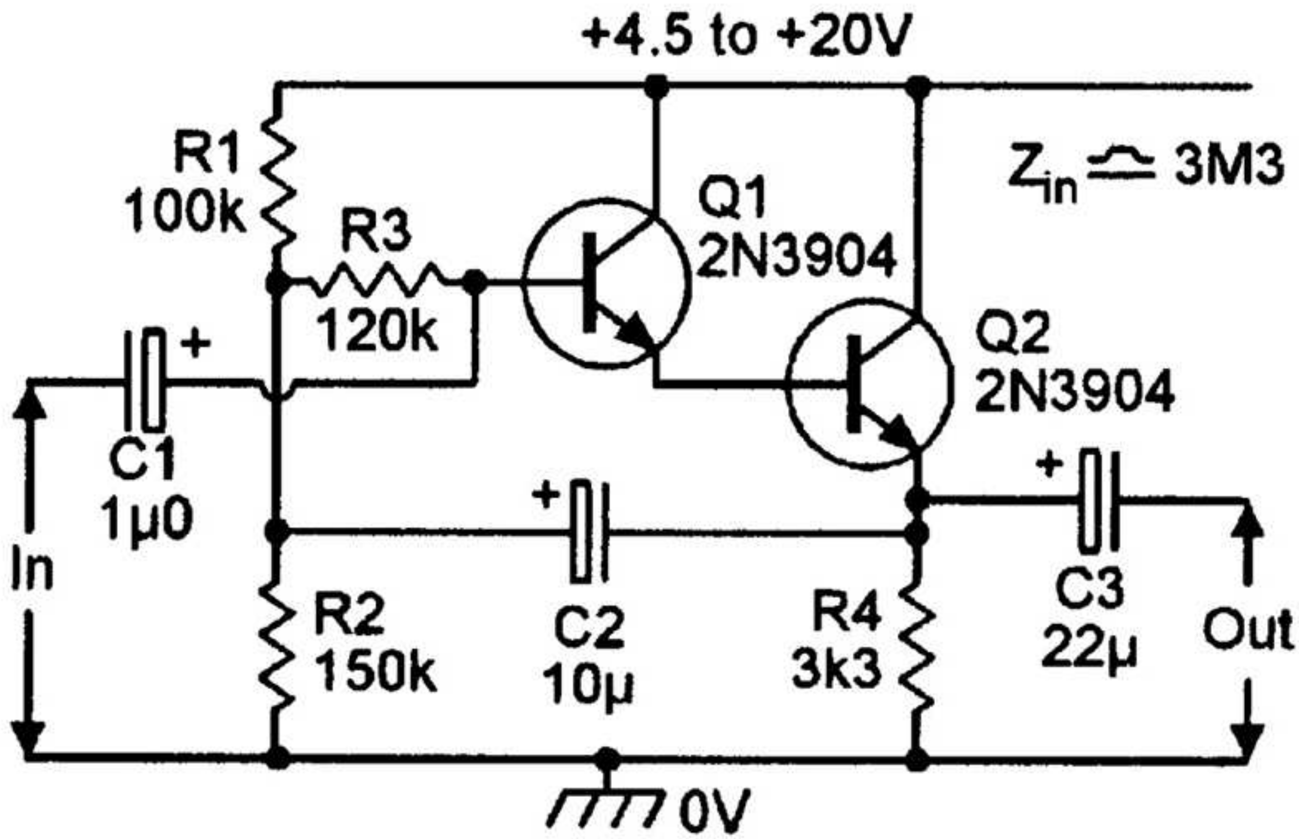


FIGURE 17. Bootstrapped Darlington emitter follower.

An even greater input impedance can be obtained by using the bootstrapped "complementary feedback pair" circuit in **Figure 18**, which gives an input impedance of about 10M. In this case, Q1 and Q2 are, in fact, both wired as common emitter amplifiers, but they operate with virtually 100 percent negative feedback and give an overall voltage gain of almost exactly unity — this "pair" of transistors thus acts like a near-perfect Darlington emitter follower.

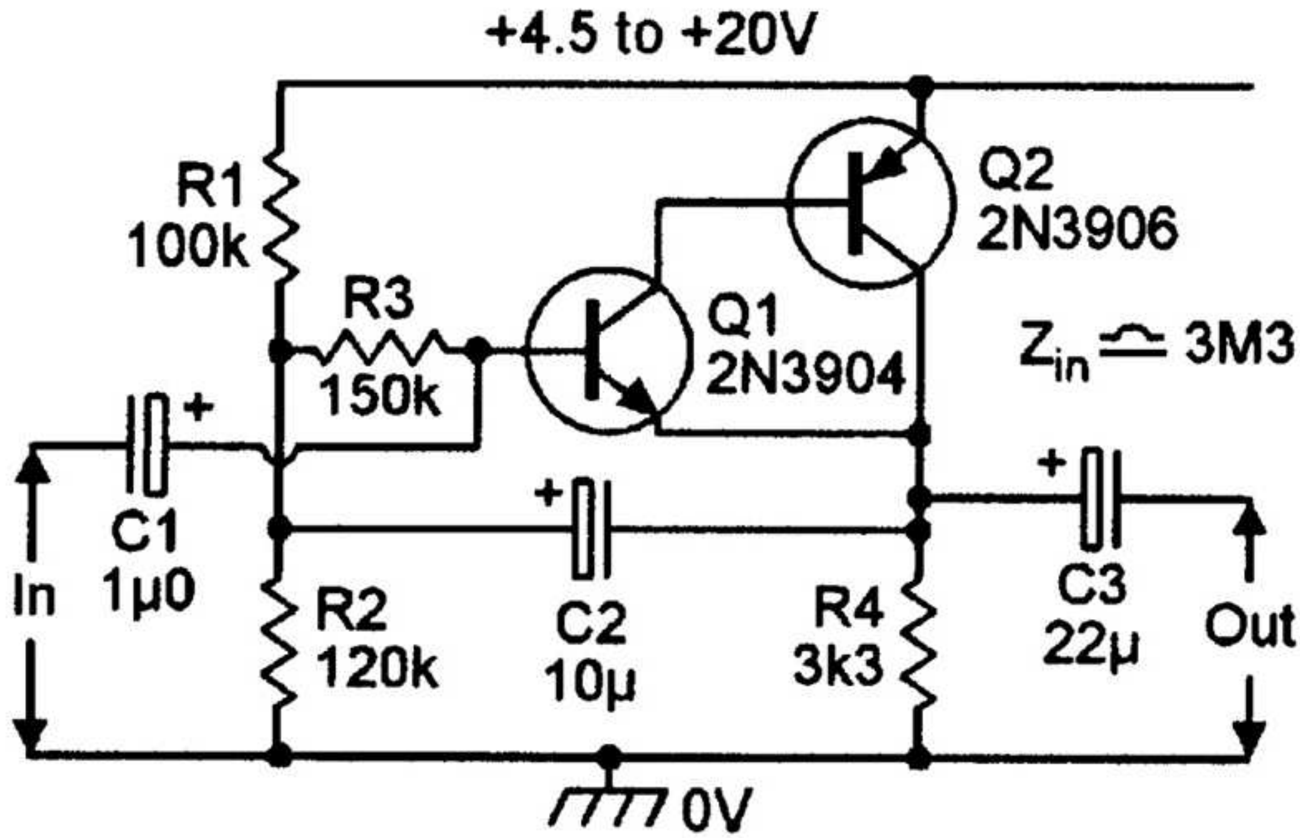


FIGURE 18. Bootstrapped complementary feedback pair.

COMPLEMENTARY EMITTER FOLLOWERS

It was pointed out earlier that an NPN emitter follower can source current, but cannot sink it, and that a PNP emitter follower can sink current, but cannot source it; i.e., these circuits can handle unidirectional output currents only. In many applications, a “bidirectional” emitter follower circuit (that can source and sink currents with equal ease) is required, and this action can be obtained by using a complementary emitter follower configuration in which NPN and PNP emitter followers are effectively wired in series. **Figures 19** to **21** show basic circuits of this type.

The circuit in **Figure 19** uses a dual (split) power supply and has its output direct-coupled to a grounded load (R_L).

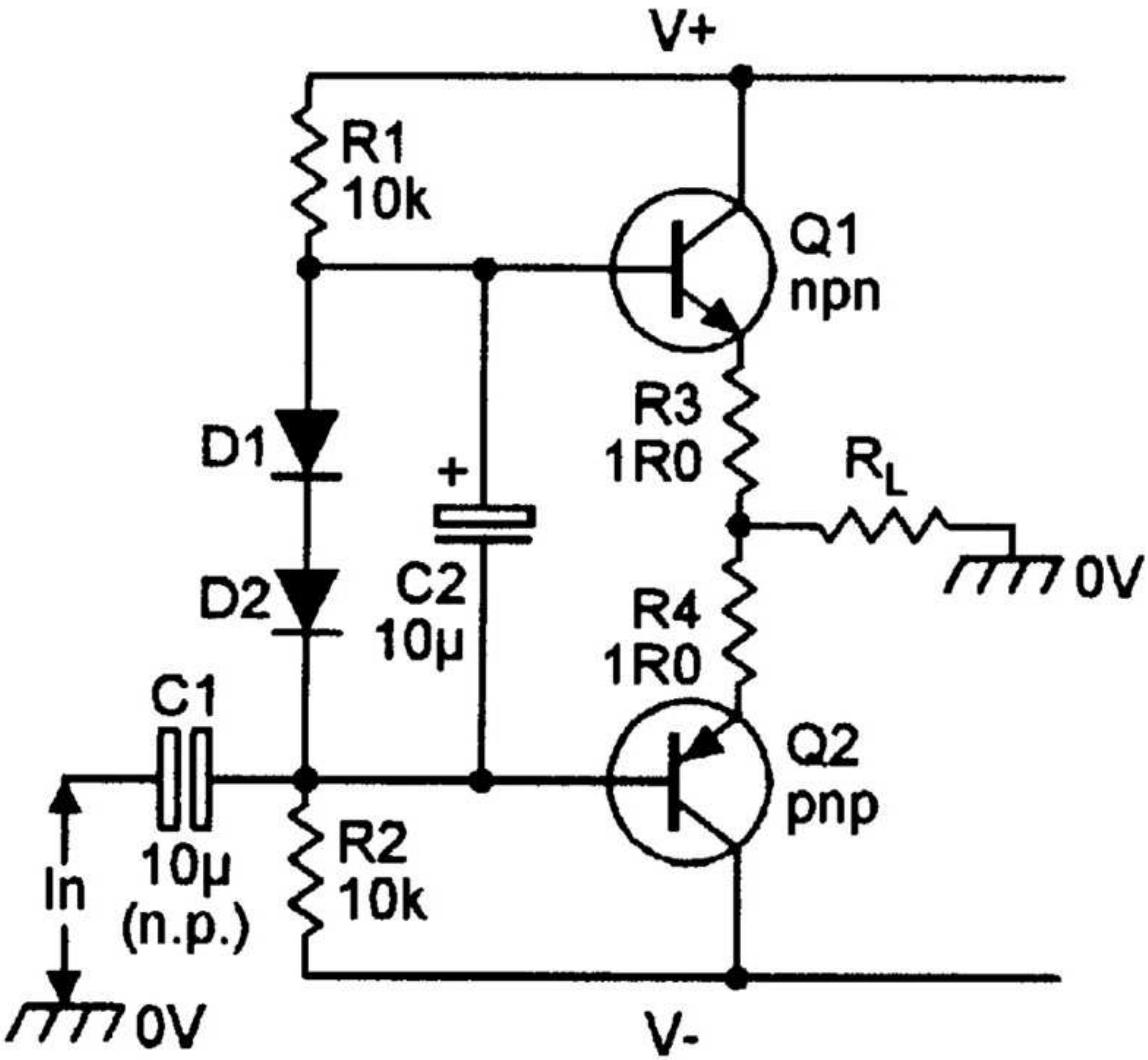


FIGURE 19. Complementary emitter follower using split supply and direct-coupled output load.

The series-connected NPN and PNP transistors are biased at a quiescent "zero volts" value via the R1-D1-D2-R2 potential divider, with each transistor slightly forward-biased via silicon diodes D1 and D2, which have characteristics inherently similar to those of the transistor base-emitter junctions. C2 ensures that identical input signals are applied to the transistor bases, and R3 and R4 protect the transistors against excessive output currents. The circuit's action is such that Q1 sources current into the load when the input goes positive, and Q2 sinks load current when the input goes negative. Note that input capacitor C1 is a non-polarized type. **Figure 20** shows an alternative version of the above circuit, designed for use with a single-ended power supply and an AC-coupled output load — *note in this case, that C1 is a polarized type.*

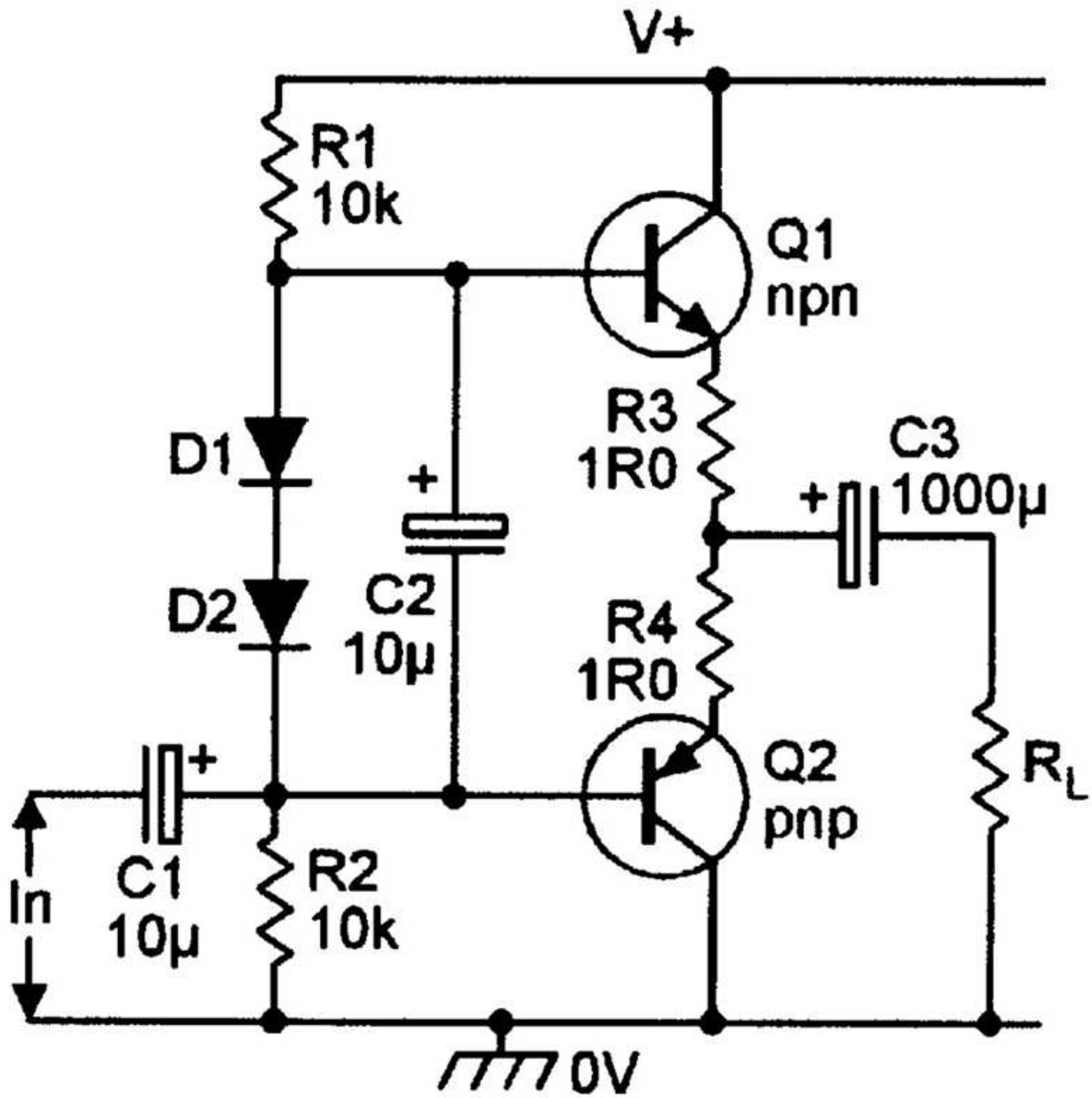


FIGURE 20. Complementary emitter follower using single-ended supply and AC-coupled load.

THE AMPLIFIED DIODE

The Q1 and Q2 circuits in **Figures 19** and **20** are slightly forward-biased (to minimize cross-over distortion problems) via silicon diodes D1 and D2 — in practice, the diode currents (and thus the transistor forward-bias voltages) are usually adjustable over a limited range. If these basic circuits are modified for use with Darlington transistors stages, a total of four biasing diodes are needed — in such cases the diodes are usually replaced by a transistor “amplified diode” stage, as shown by the Q5 circuit in **Figure 21**.

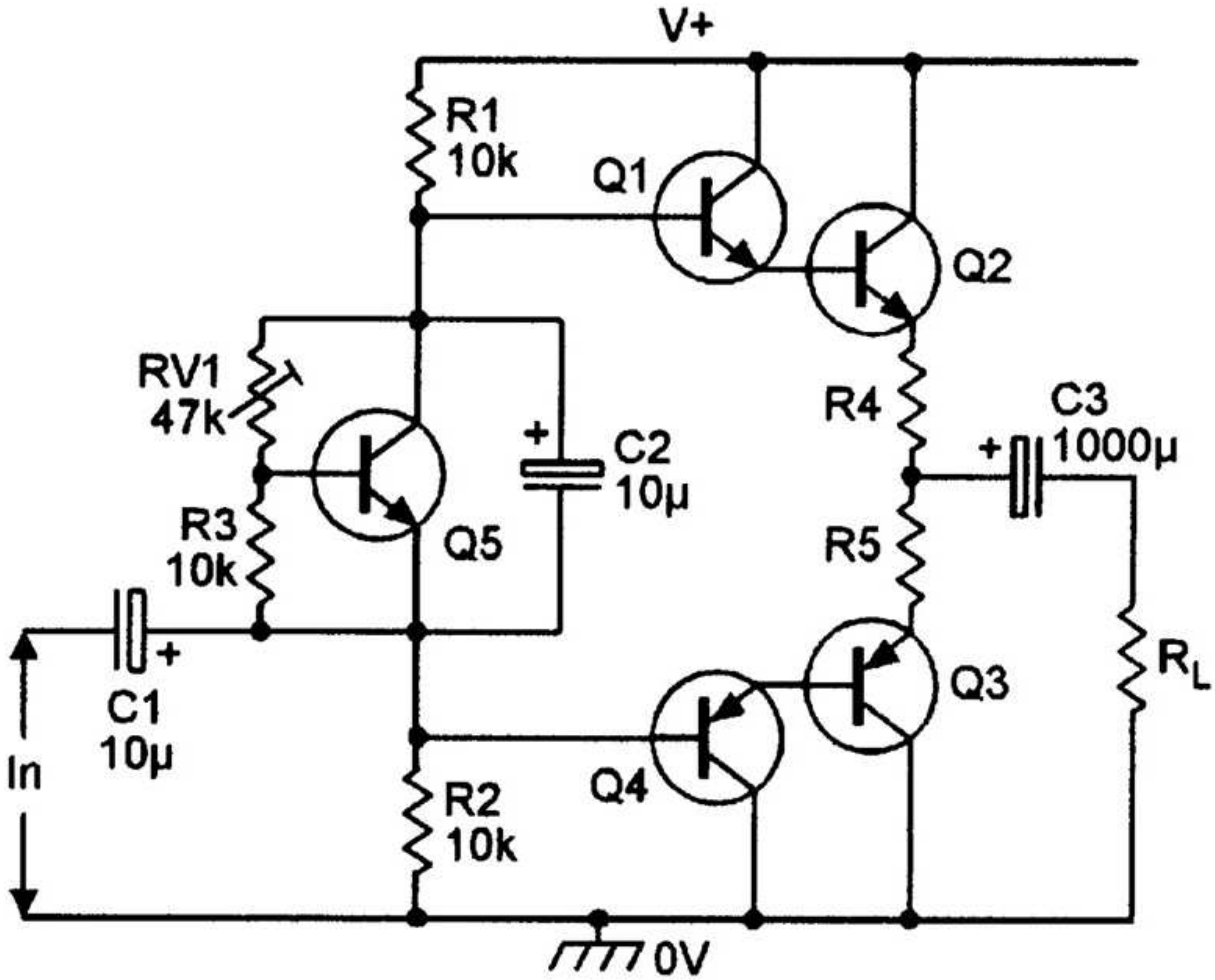


FIGURE 21. Darlington complementary emitter follower, with biasing via an amplified diode (Q5).

In the **Figure 21** circuit, Q5's collector-to-emitter voltage equals the Q5 base-emitter volt drop (about 600 mV) multiplied by (RV1+R3)/R3 — so, if RV1 is set to zero ohms, 600 mV are developed across Q5, which thus acts like a single silicon diode. However, if RV1 is set to 47K, about 3.6V is developed across Q5, which thus acts like six series-connected silicon diodes. RV1 can be used to precisely set the Q5 volt drop and thus adjust the quiescent current values of the Q2-Q3 output stages.

High-power versions of the basic **Figure 21** circuit are widely used as the basis of many modern "Hi-Fi" audio power amplifier circuits. Some simple circuits of this type will be described later in this Bipolar Transistor Cookbook series. **NV**

Part 1 is available **here**.

Bipolar Transistor Cookbook — Part 3

Last time in this series, I described practical ways of using bipolar transistors in useful common-collector (voltage follower) circuit applications, including those of relay drivers, constant-current generators, linear amplifiers, and complementary emitter followers. This article moves on and shows various ways of using bipolar transistors in simple, but useful common-emitter and common-base configurations.

COMMON-EMITTER AMPLIFIER CIRCUITS

The common-emitter amplifier (also known as the common-earth or grounded-emitter circuit) has a medium value of input impedance and provides substantial voltage gain between input and output. The circuit's input is applied to the transistor's base, and the output is taken from its collector — the circuit's basic operating principles were briefly described in the opening installment of this eight-part series. The common-emitter amplifier can be used in a wide variety of digital and analog voltage amplifier applications. This section of the *Cookbook* series starts off by looking at "digital" application circuits.

DIGITAL CIRCUITS

Figure 1 shows a simple npn common-emitter digital amplifier, inverter, or switch, in which the input signal is at either zero volts or a substantial positive value, and is applied to the transistor's base via series resistor R_b, and the output signal is taken from the transistor's collector. When the input is zero, the transistor is cut off and the output is at full positive supply rail value. When the input is high, the transistor is biased on and a collector current flows via R_L, thus pulling the output low. If the input voltage is large enough, Q1 is driven fully on and the output drops to a "saturation" value of a few hundred mV. Thus, the output signal is an amplified and inverted version of the input signal.

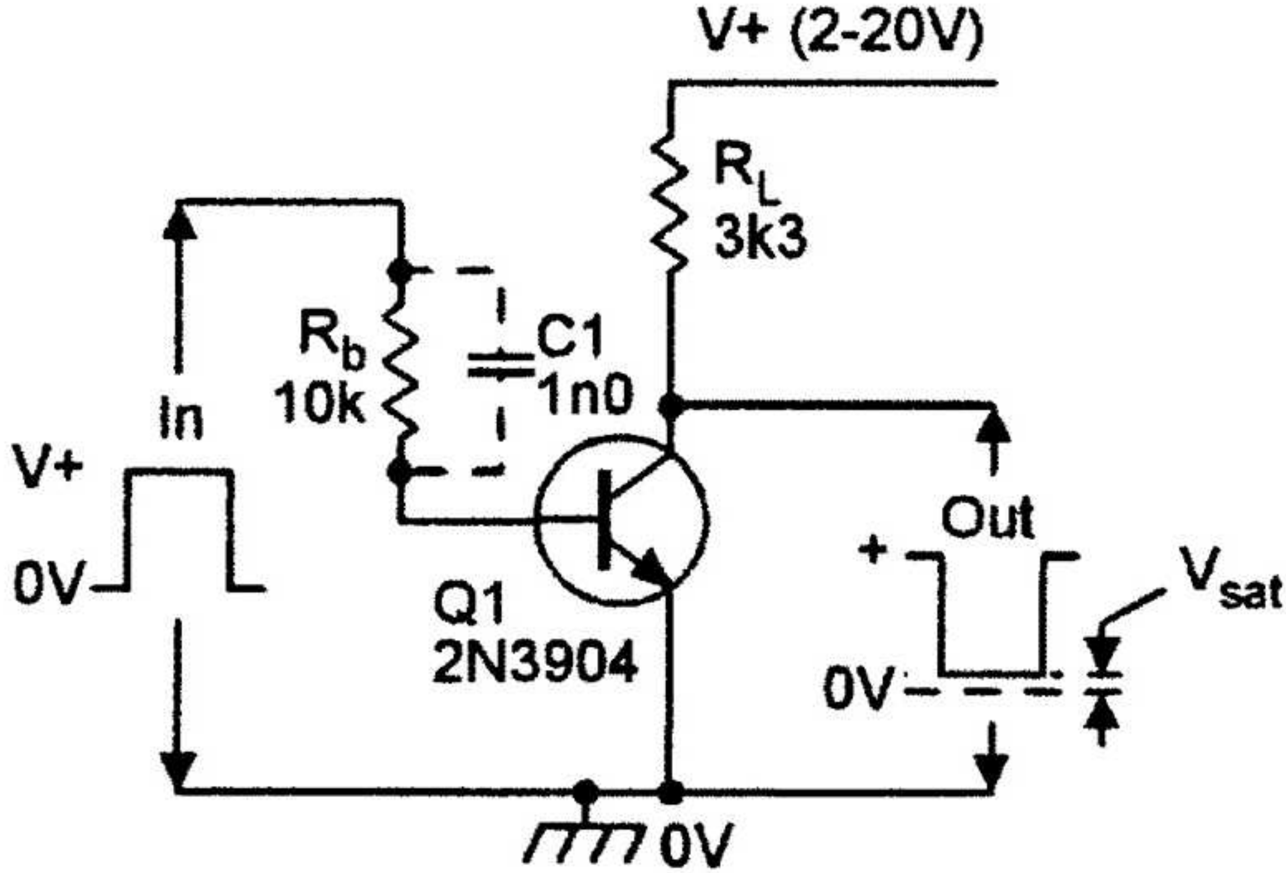


FIGURE 1. Digital inverter/switch (npn)

In **Figure 1**, resistor R_b limits the input base-drive current to a safe value. The circuit's input impedance is slightly greater than the R_b value, which also influences the rise and fall times of the output signal — the greater the R_b value, the worse these become. This snag can be overcome by shunting R_b with a "speed-up" capacitor (typically about 1n0), as shown dotted in the diagram. In practice, R_b should be as small as possible, consistent with safety and input-impedance requirements, and must not exceed R_L x h_{fe}.

Figure 2 shows a pnp version of the digital inverter/switch circuit. Q1 switches fully on, with its output a few hundred mV below the positive supply value, when the input is at zero volts, and turns off (with its output at zero volts) when the input rises to within less than 600mV of the positive supply rail value.

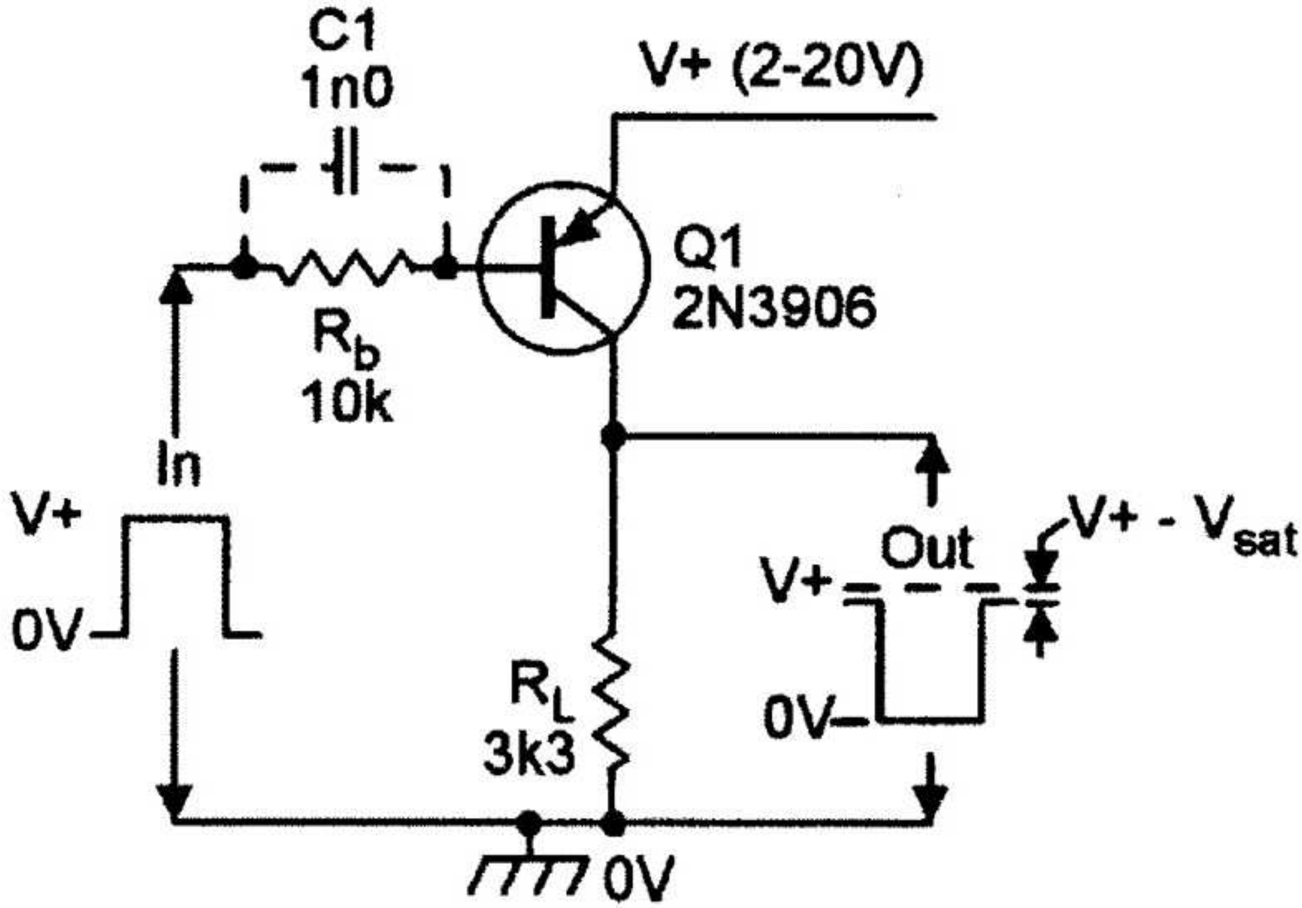


FIGURE 2. Digital inverter/switch (pnp)

The sensitivity of the **Figure 1 and 2** circuits can be increased by replacing Q1 with a pair of Darlington- or Super-Alpha-connected transistors. Alternatively, a very-high-gain non-inverting digital amplifier/switch can be made by using a pair of transistors wired in either of the ways shown in **Figures 3 or 4**.

The **Figure 3** circuit uses two npn transistors. When the input is at zero volts, Q1 is cut off, so Q2 is driven fully on via R2, and the output is low (saturated). When the input is “high,” Q1 is driven to saturation and pulls Q2 base to less than 600mV, so Q2 is cut off and the output is high (at V+).

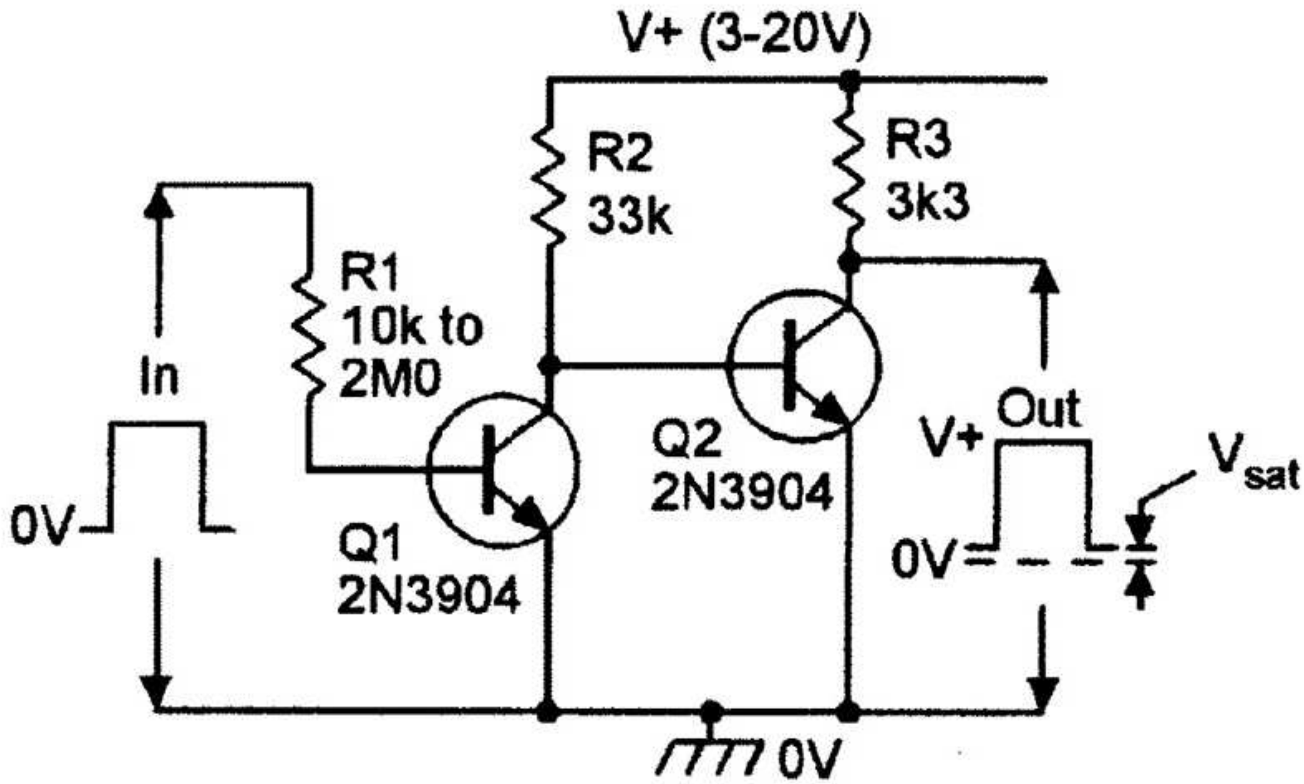


FIGURE 3. Very-high-gain non-inverting digital amplifier/switch using npn transistors

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The **Figure 4** circuit uses one npn and one pnp transistor. When the input is at zero volts, Q1 is cut off, so Q2 is also cut off (via R2-R3) and the output is at zero volts. When the input is "high," Q1 is driven on and pulls Q2 into saturation via R3. Under this condition, the output takes up a value a few hundred mV below the positive supply rail value.

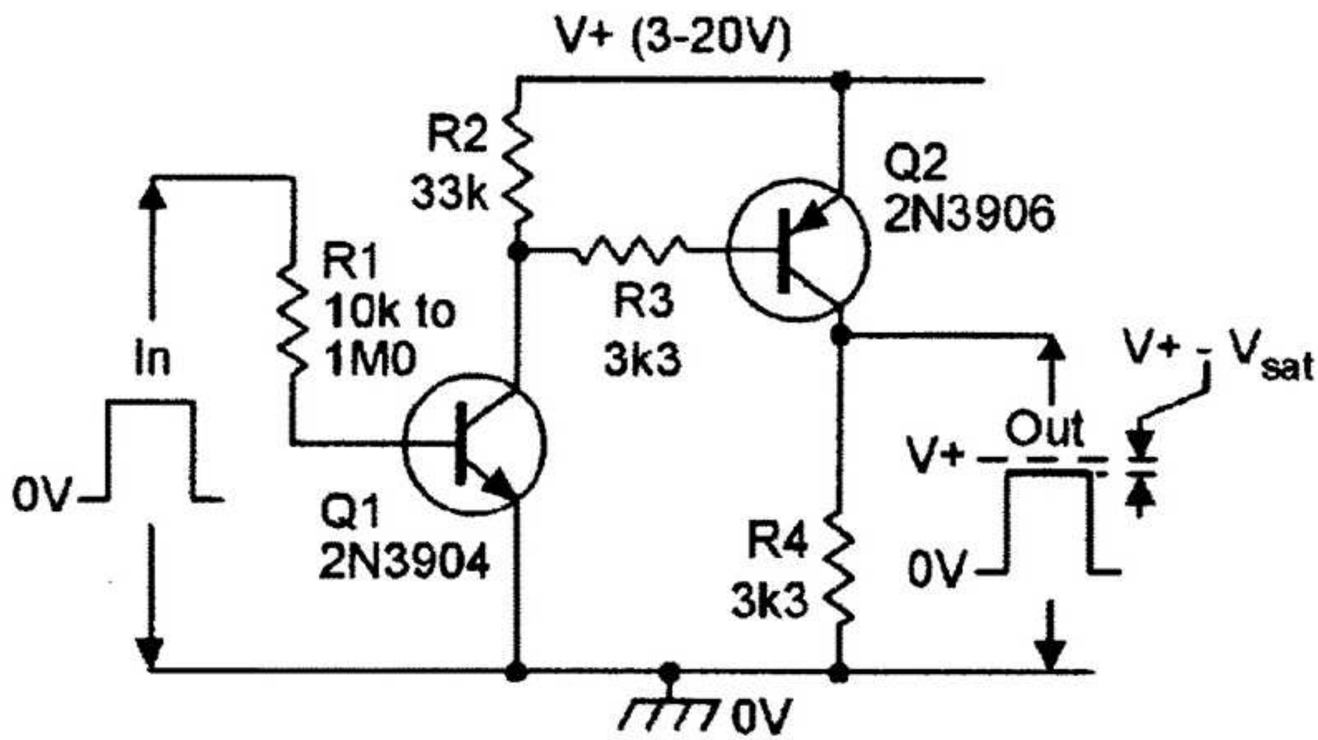


FIGURE 4. Alternative non-inverting digital amplifier/switch using an npn-pnp pair of transistors

Figure 5 shows (in basic form) how a complementary pair of the **Figure 4** circuits can be used to make a DC-motor direction-control network, using a dual power supply. The circuit operates as follows.

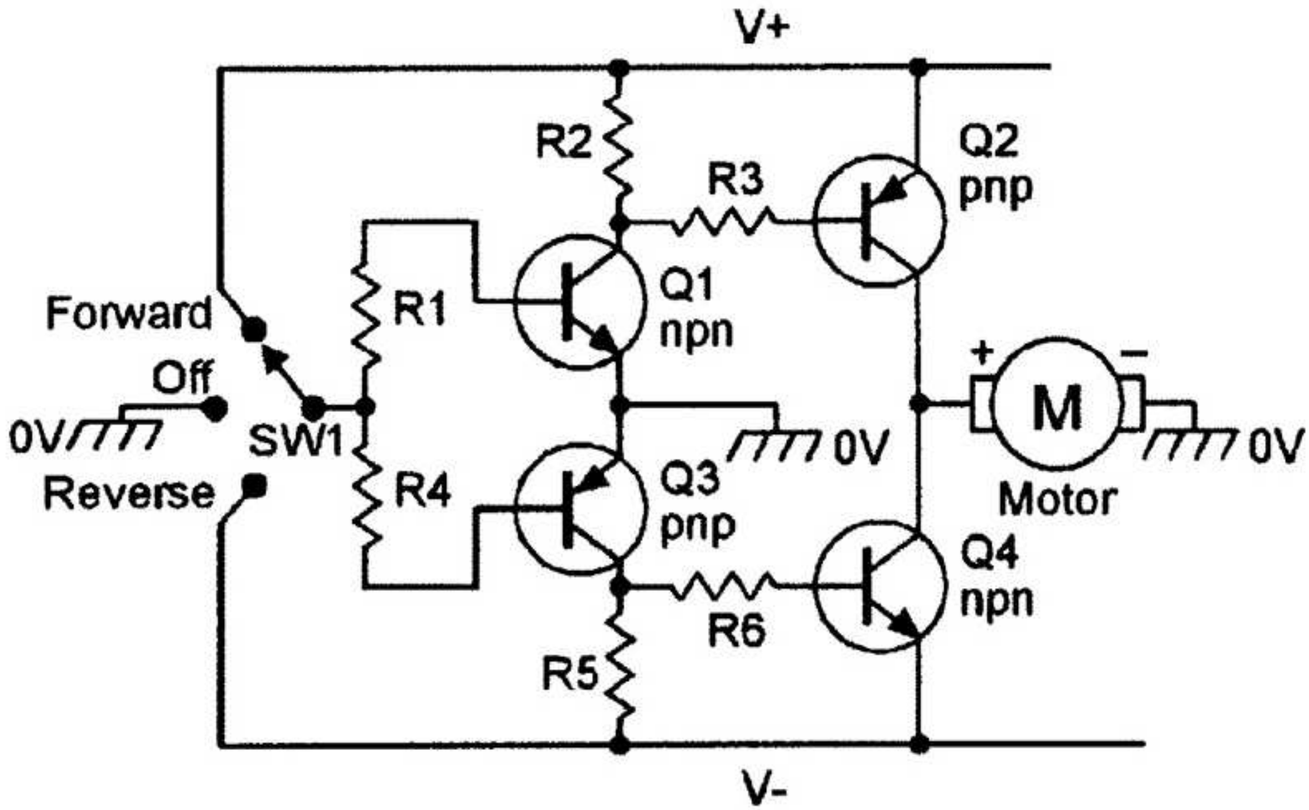


FIGURE 5. DC-motor direction-control circuit

When SW1 is set to “Forward,” Q1 is driven on via R1, and pulls Q2 on via R3, but Q3 and Q4 are cut off. The “live” side of the motor is thus connected (via Q2) to the positive supply rail under this condition, and the motor runs in the forward direction.

When SW1 is set to “Off,” all four transistors are cut off, and the motor is inoperative.

When SW1 is set to “Reverse,” Q3 is biased on via R4, and pulls Q4 on via R6, but Q1 and Q2 are cut off. The “live” side of the motor is thus connected (via Q4) to the negative supply rail under this condition, and the motor runs in the reverse direction.

RELAY DRIVERS

The basic digital circuits of **Figures 1 through 4** can be used as efficient relay drivers if fitted with suitable diode protection networks. **Figures 6 through 8** show examples of such circuits.

The **Figure 6** circuit raises a relay’s current sensitivity by a factor of about 200 (= the current gain of transistor Q1), and greatly increases its voltage sensitivity. R1 gives base drive protection, and can be larger than 1k0, if desired. The relay is turned on by a positive input voltage.

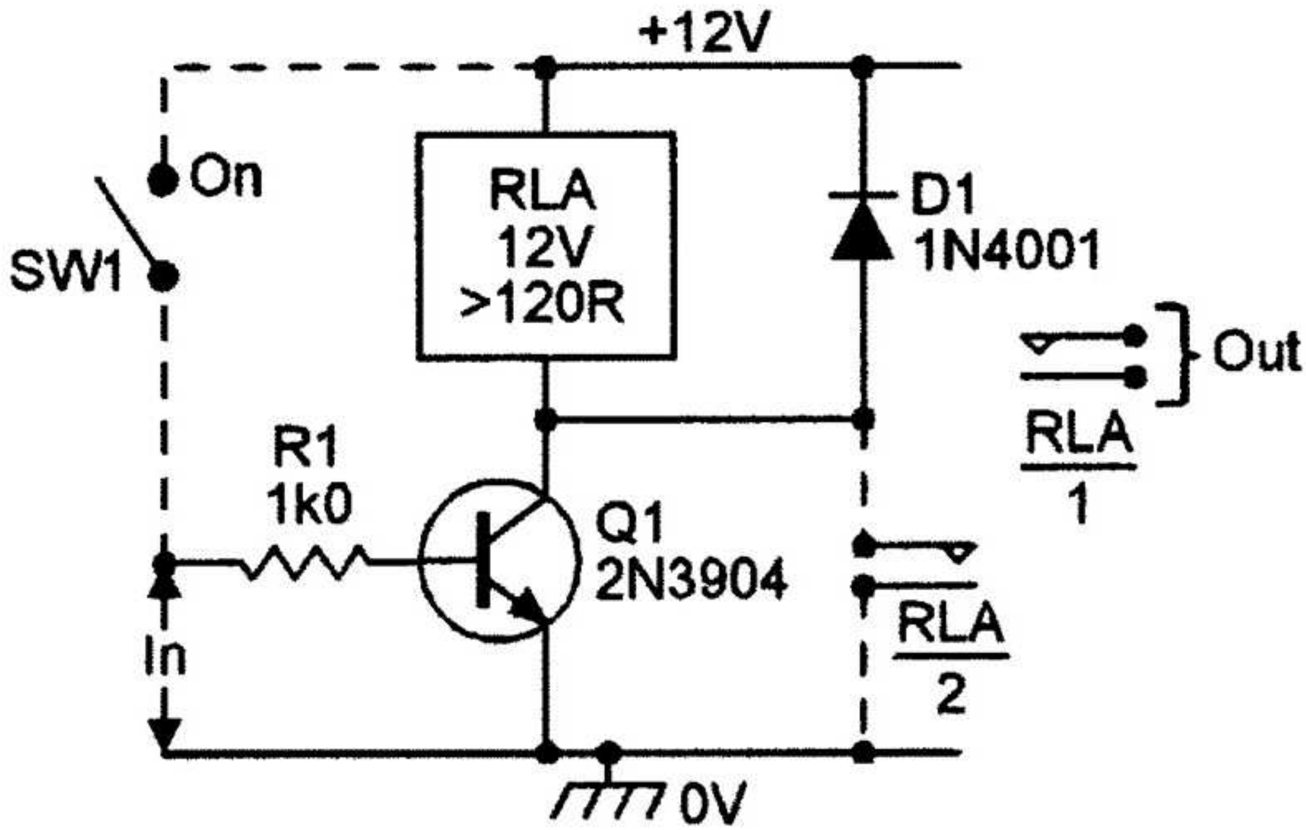


FIGURE 6. Simple relay-driving circuit

The current sensitivity of the relay can be raised by a factor of about 20,000 by replacing Q1 with a Darlington-connected pair of transistors. **Figure 7** shows this technique used to make a circuit that can be activated by placing a resistance of less than 2M0 across a pair of stainless metal probes. Water, steam, and skin contacts have resistances below this value, so this simple little circuit can be used as a water, steam, or touch-activated relay switch.

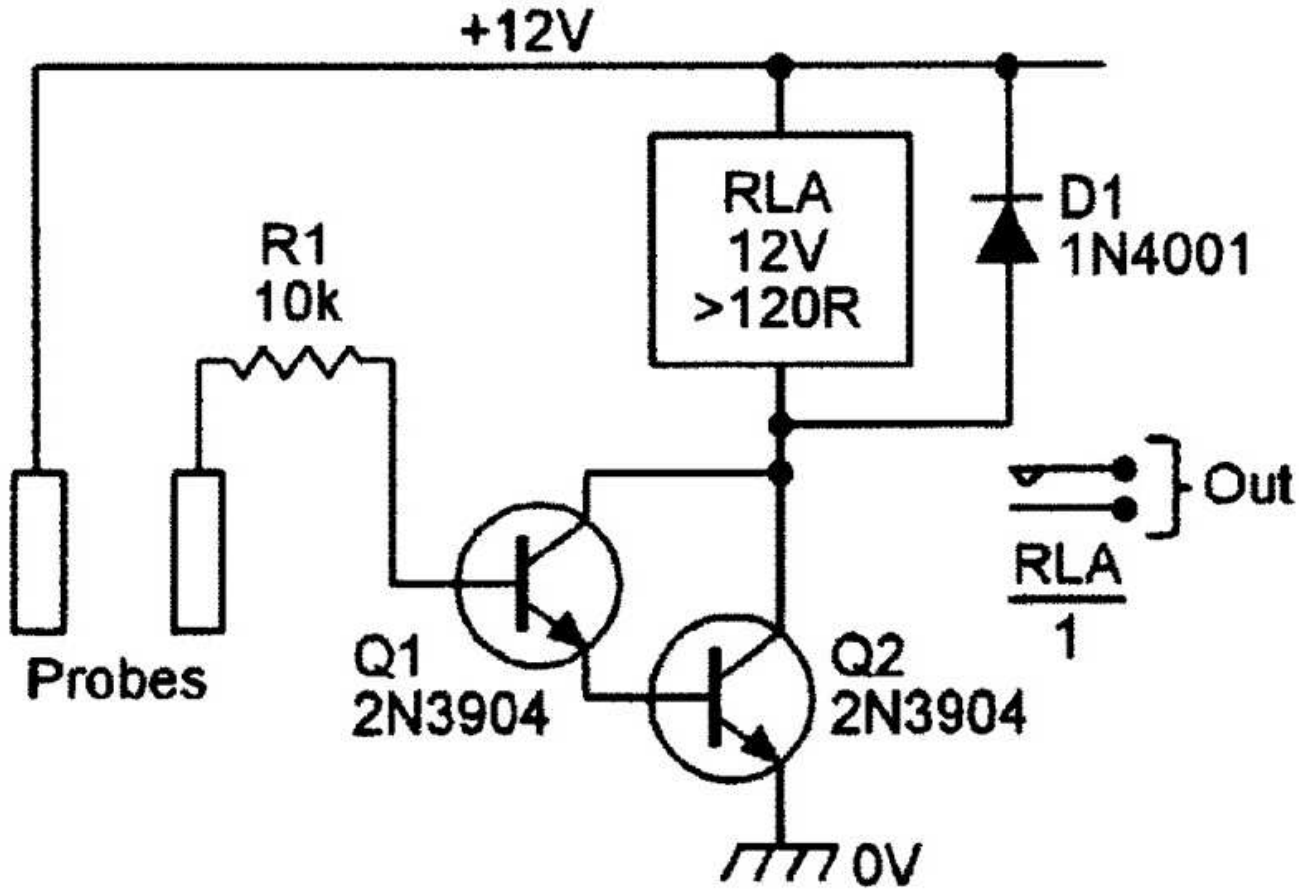


FIGURE 7. Touch, water, or steam-activated relay switch

Figure 8 shows another ultra-sensitive relay driver, based on the **Figure 4** circuit, that needs an input of only 700mV at 40µA to activate the relay. R2 ensures that Q1 and Q2 turn completely off when the input terminals are open circuit.

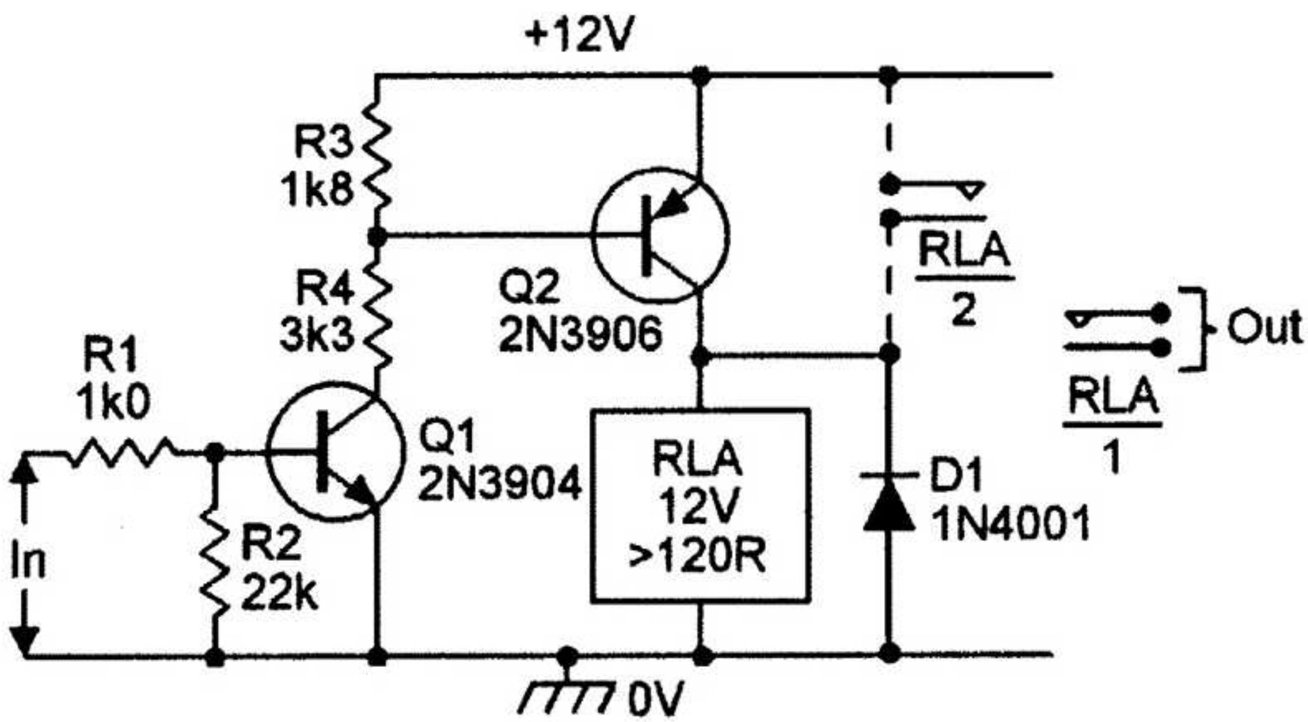


FIGURE 8. Ultra-sensitive relay driver (needs an input of 700mV at 40μA)

LINEAR BIASING CIRCUITS

A common-emitter circuit can be used as a linear AC amplifier by applying a DC bias current to its base so that its collector takes up a quiescent half-supply voltage value (to accommodate maximal undistorted output signal swings), and by then feeding the AC input signal to its base and taking the AC output from its collector (as shown in **Figure 9**).

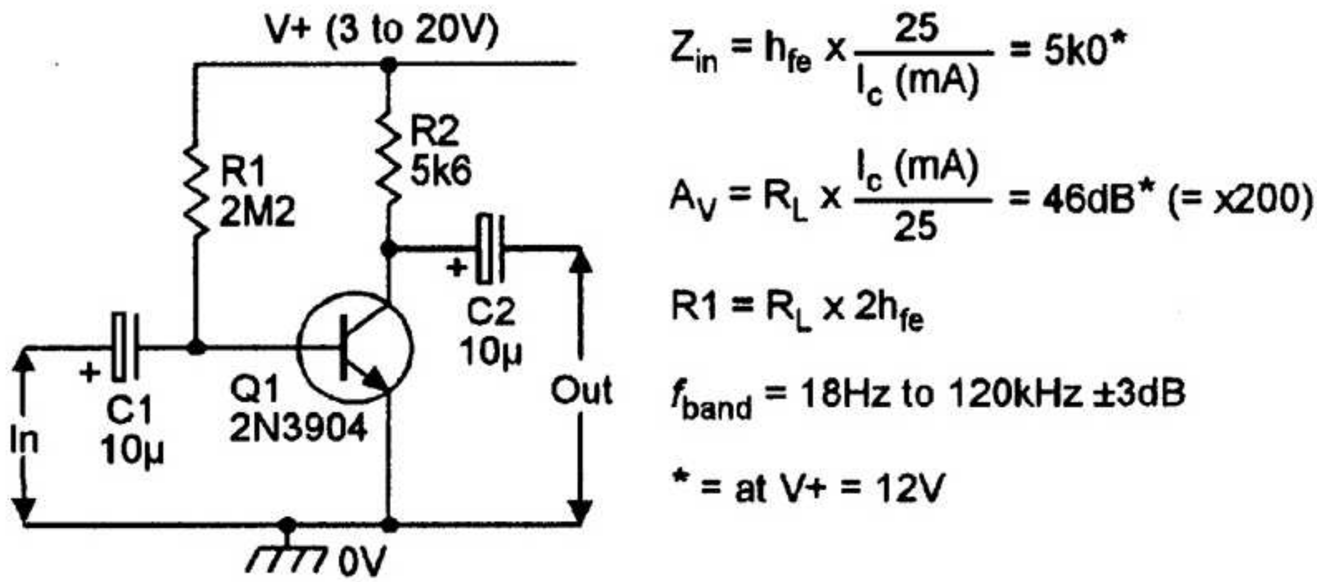


FIGURE 9. Simple npn common-emitter amplifier

The first step in designing a circuit of the basic **Figure 9** type is to select the value of load resistor R2. The lower this is, the higher the amplifier's upper cut-off frequency will be (due to the smaller shunting effects of stray capacitance on the effective impedance of the load), but the higher Q1's quiescent operating current will be. In the diagram, R2 has a compromise value of 5k6, which gives an upper "3dB down" frequency of about 120kHz and a quiescent current consumption of 1mA from a 12V supply.

To bias the **Figure 9** circuit's output to half-supply volts, R1 needs a value of R2 x 2h_{fe}, and (assuming a nominal h_{fe} of 200) this works out at about 2M2 in the example shown. The formula for the circuit's input impedance (looking into Q1 base) and voltage gain are both given in the diagram. In the example shown, the input impedance is roughly 5k0, and is shunted by R1 — the voltage gain works out at about x200, or 46dB.

The quiescent biasing point of the **Figure 9** circuit depends on Q1's h_{fe} value. This weakness can be overcome by modifying the circuit as shown in **Figure 10**, where biasing resistor R1 is wired in a DC feedback mode between Q1's collector and base, and has a value of R2 x h_{fe}. The feedback action is such that any shift in the output level (due to variations in h_{fe}, temperature, or component values) causes a counter-change in the base-current biasing level, thus tending to cancel the original shift.

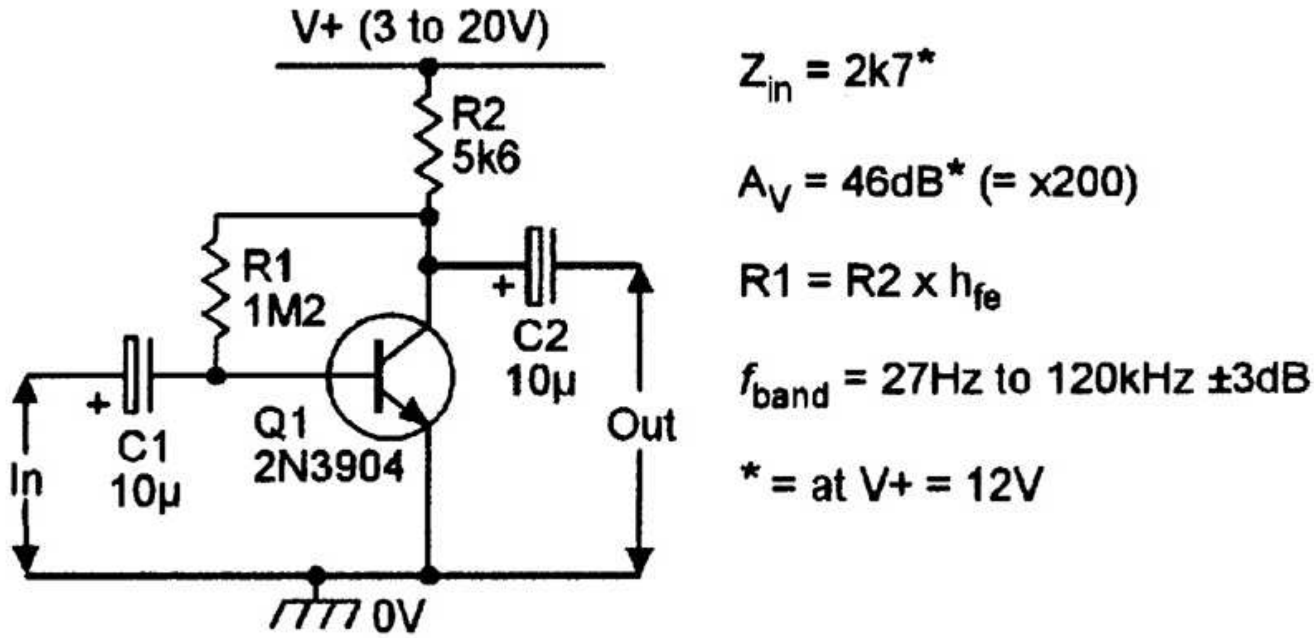


FIGURE 10. Common-emitter amplifier with feedback biasing

The **Figure 10** circuit has the same values of bandwidth and voltage gain as the **Figure 9** design, but has a lower total value of input impedance. This is because the AC feedback action reduces the apparent impedance of R1 (which shunts the 5k0 base impedance of Q1) by a factor of 200 (= A_V), thus giving a total input impedance of 2k7. If desired, the shunting effects of the biasing network can be eliminated by using two feedback resistors and AC-decoupling them as shown in **Figure 11**.

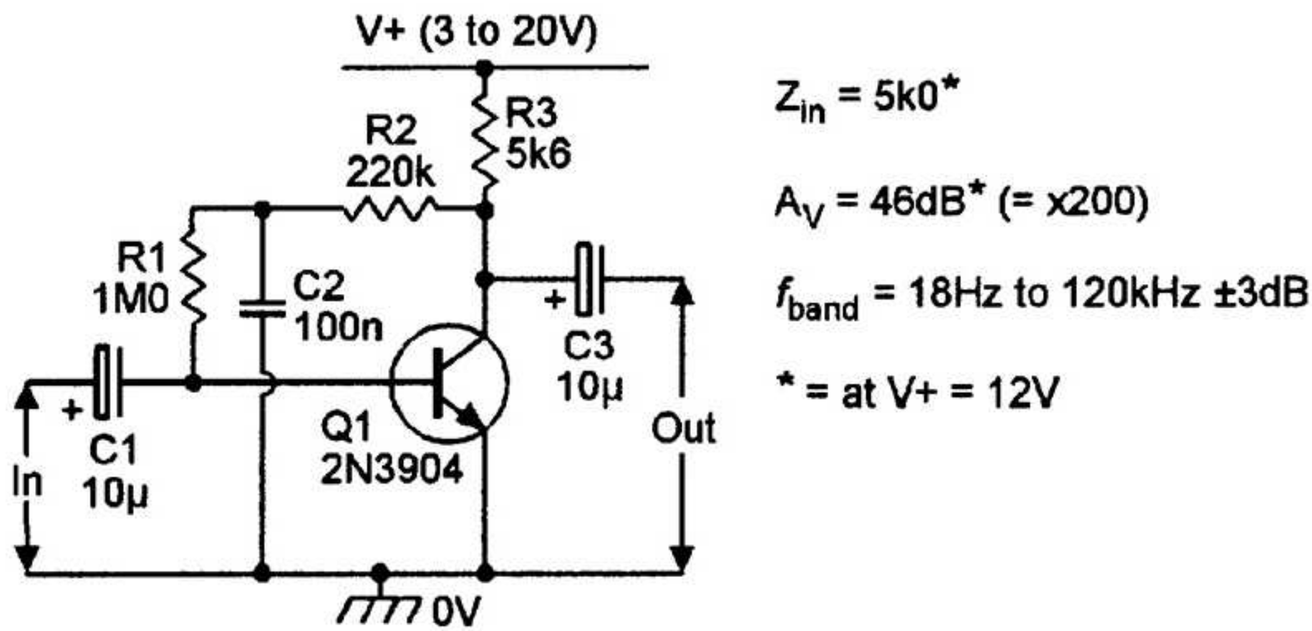


FIGURE 11. Amplifier with AC-decoupled feedback biasing

Finally, the ultimate in biasing stability is given by the “potential-divider biasing” circuit of **Figure 12**. Here, potential divider R1-R2 sets a quiescent voltage slightly greater than V+/3 on Q1 base, and voltage follower action causes 600mV less than this to appear on Q1 emitter. V+/3 is thus developed across 5k6 emitter resistor R3, and (since Q1’s emitter and collector currents are almost identical) a similar voltage is dropped across R4, which also has a value of 5k6, thus setting the collector at a quiescent value of 2V+/3. R3 is AC-decoupled via C2, and the circuit gives an AC voltage gain of 46dB.

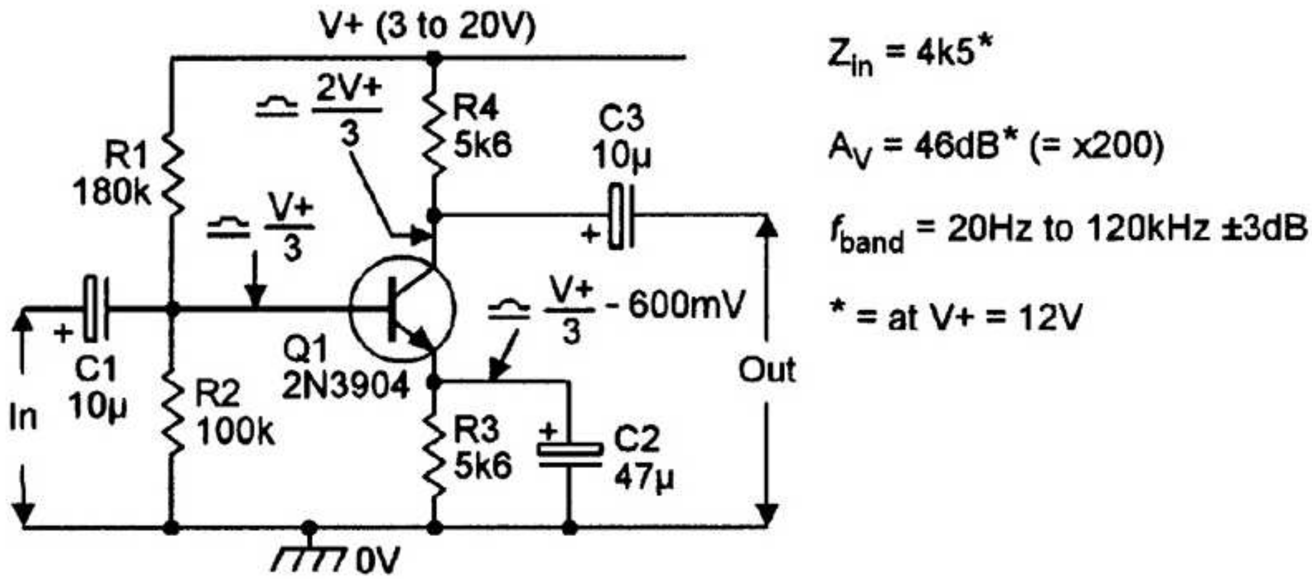


FIGURE 12. Amplifier with voltage-divider biasing

CIRCUIT VARIATIONS

Figures 13 to 16 show some useful common-emitter amplifier variations. **Figure 13** shows the basic Figure 12 design modified to give an AC voltage gain of x10 — the gain actually equals the R4 collector load value divided by the effective “emitter” impedance value, which in this case (since R3 is decoupled by series-connected C2-R5) equals the value of the base-emitter junction impedance in series with the paralleled values of R3 and R5, and works out at roughly 560R, thus giving a voltage gain of x10. Alternative gain values can be obtained by altering the R5 value.

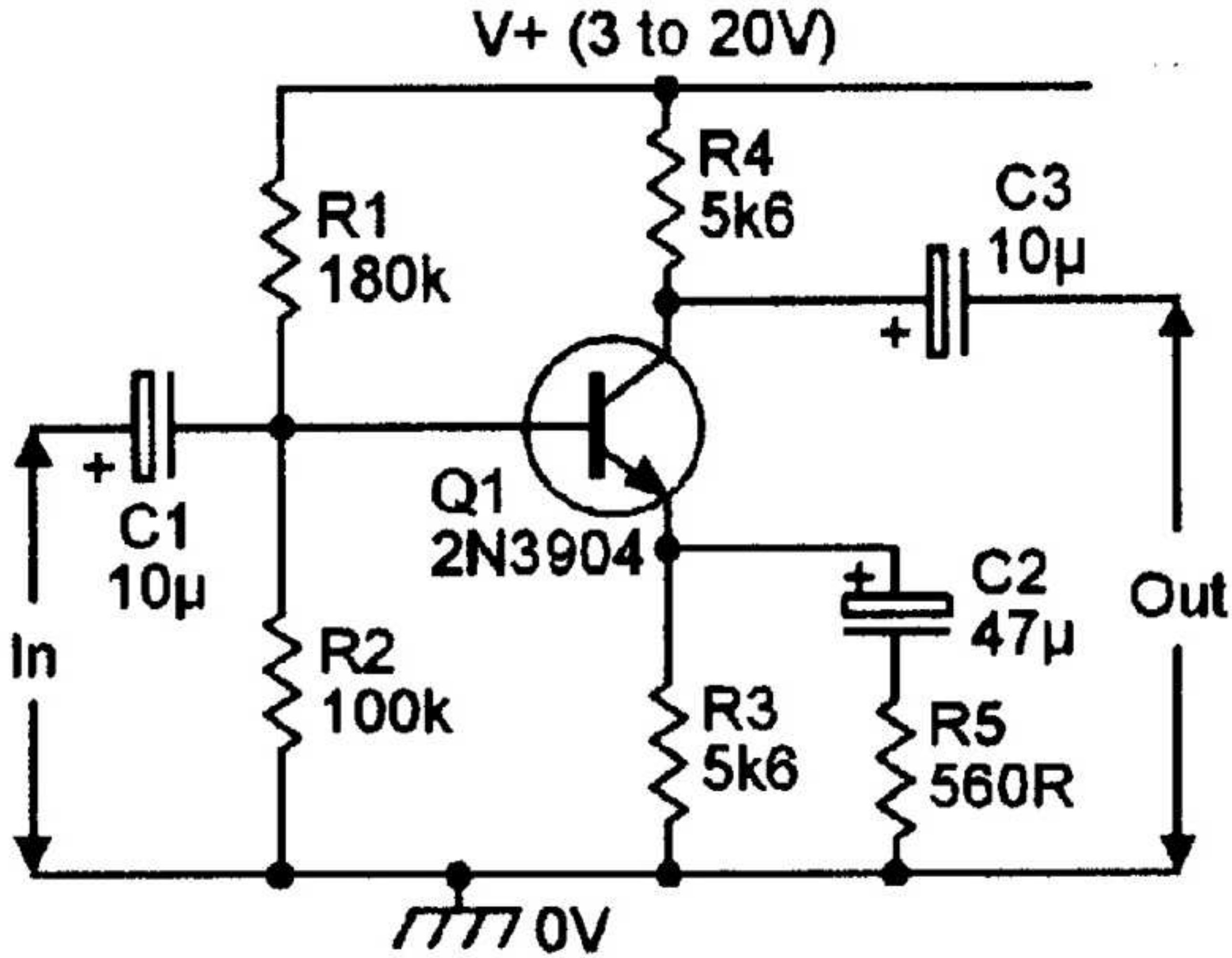


FIGURE 13. Fixed-gain (x10) common-emitter amplifier

Figure 14 shows a useful variation of the above design. In this case, R3 equals R4, and is not decoupled, so the circuit gives unity voltage gain. Note, however, that this circuit gives two unity-gain output signals, with the emitter output in phase with the input and the collector signal in anti-phase. This circuit thus acts as a unity-gain phase splitter.

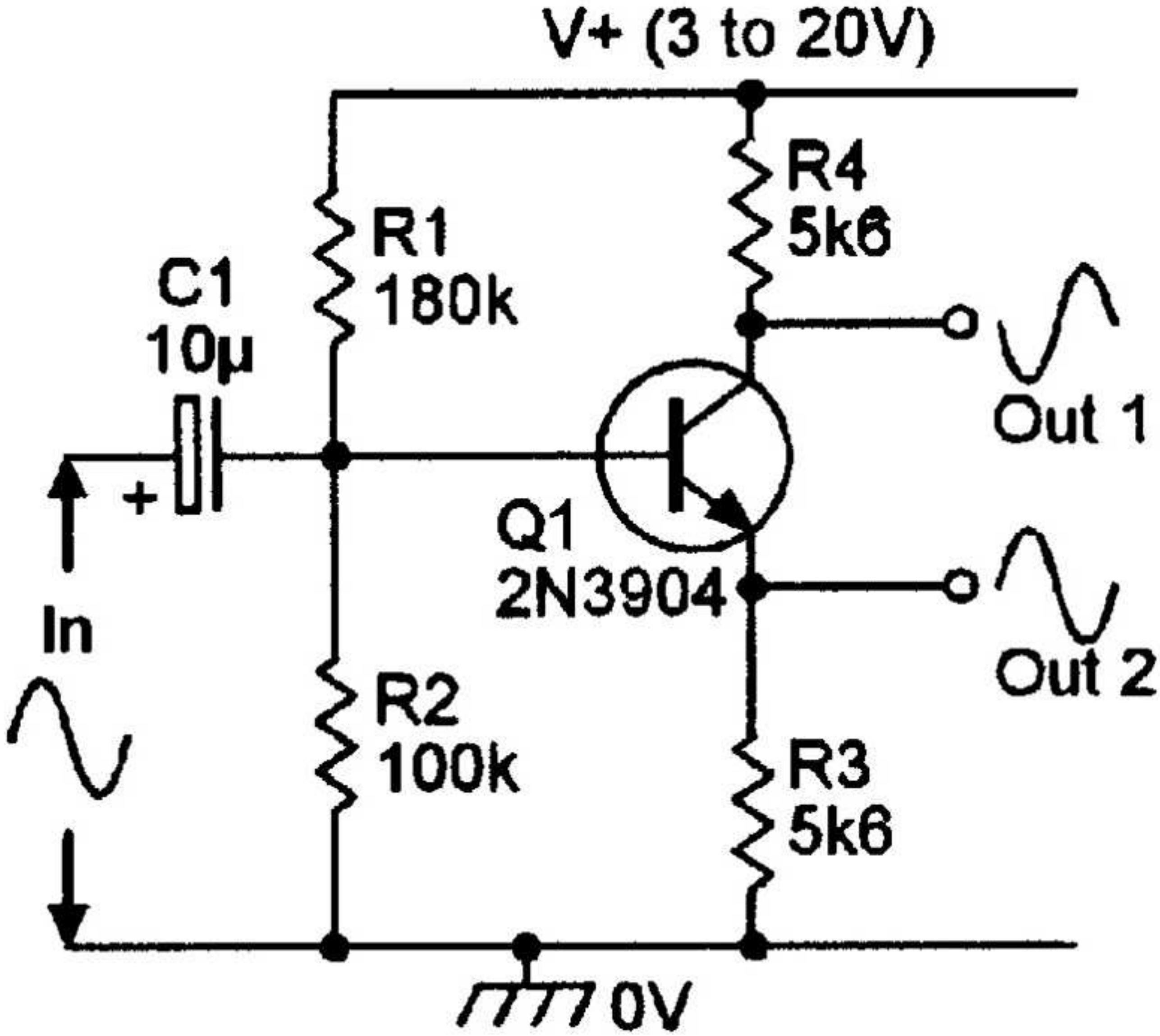


FIGURE 14. Unity-gain phase splitter

Figure 15 shows another way of varying circuit gain. This design gives high voltage gain between Q1 collector and base, but R2 gives AC feedback to the base, and R1 is wired in series between the input signal and Q1 base — the net effect is that the circuit's voltage gain (between input and output) equals R2/R1, and works out at x10 in this particular case.

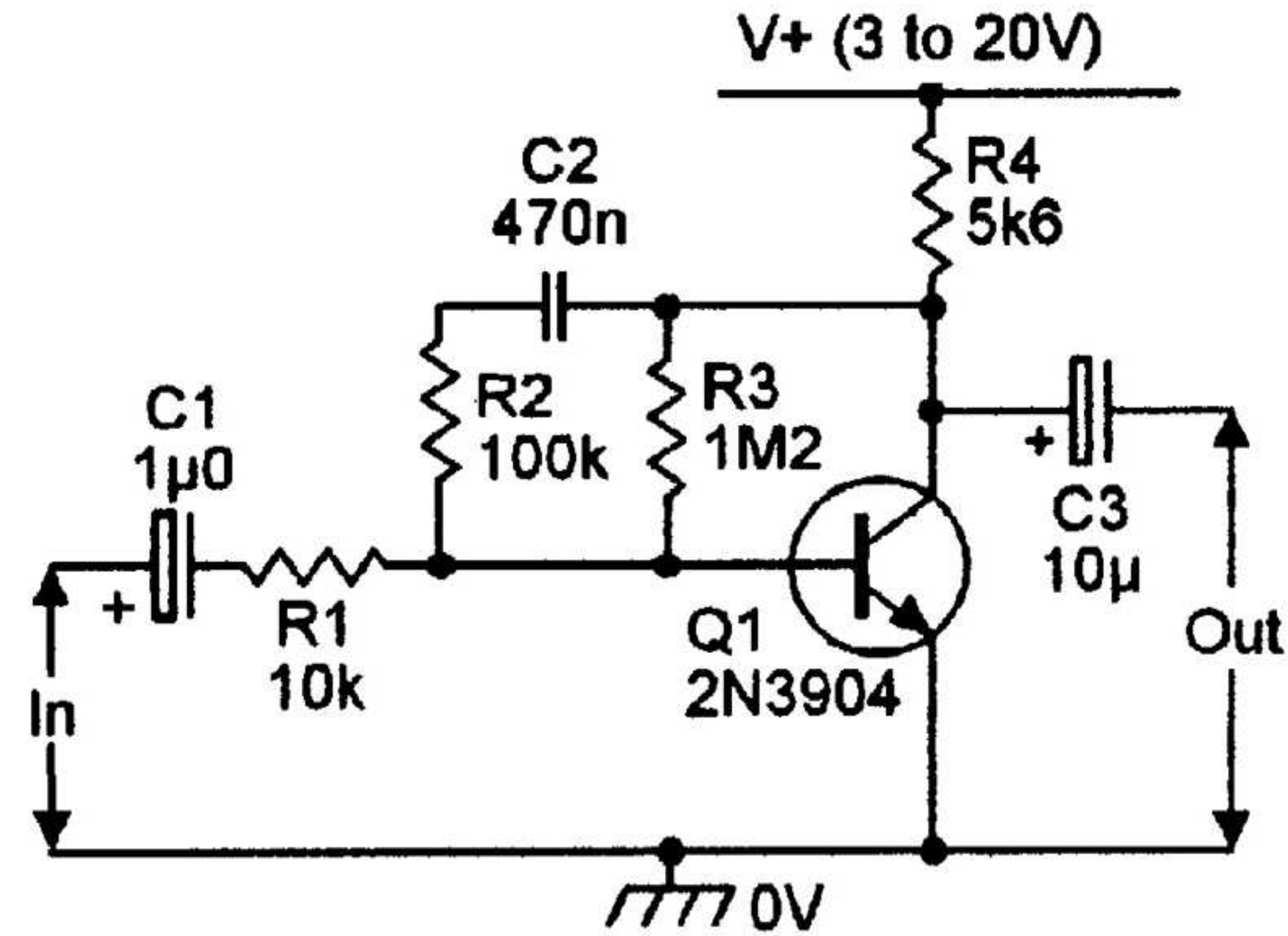


FIGURE 15. Alternative fixed-gain (x10) amplifier

Finally, **Figure 16** shows how the **Figure 10** design can be modified to give a wide-band performance by wiring DC-coupled emitter follower buffer Q2 between Q1 collector and the output terminal, to minimize the shunting effects of stray capacitance on R2, and thus extending the upper bandwidth to several hundred kHz.

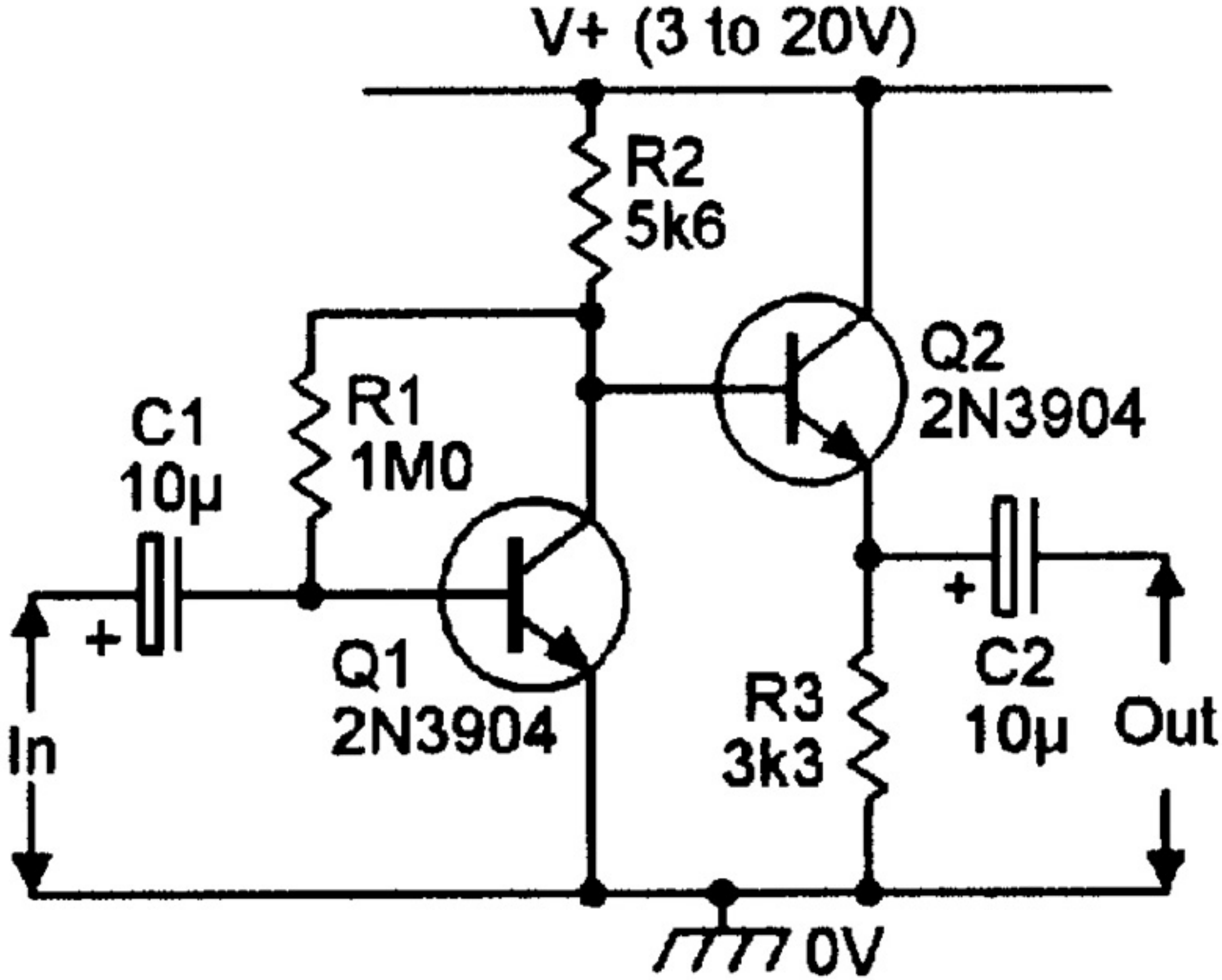


FIGURE 16. Wide-band amplifier

HIGH-GAIN CIRCUITS

A single-stage common-emitter amplifier circuit cannot give a voltage gain much greater that 46dB when using a resistive collector load — a multi-stage circuit must be used if higher gain is needed. **Figures 17 to 19** show three useful high-gain, two-transistor voltage amplifier designs.

The **Figure 17** circuit acts like a direct-coupled pair of common-emitter amplifiers, with Q1's output feeding directly into Q2 base, and gives an overall voltage gain of 76dB (about x6150) and an upper -3dB frequency of 35kHz. Note that feedback biasing resistor R4 is fed from Q2's AC-decoupled emitter (which "follows" the quiescent collector voltage of Q1), rather than directly from Q1 collector, and that the bias circuit is thus effectively AC-decoupled.

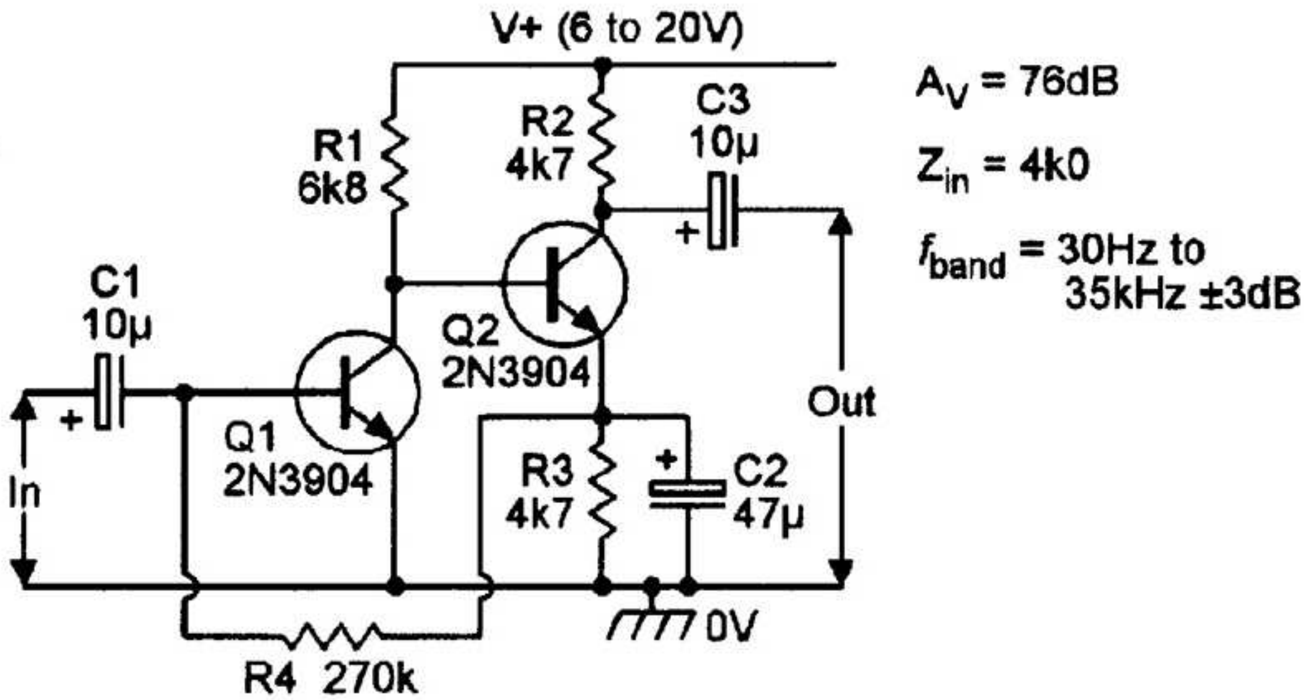


FIGURE 17. High-gain two-stage amplifier

Figure 18 shows an alternative version of the above design, using a pnp output stage — its performance is the same as that of **Figure 17**.

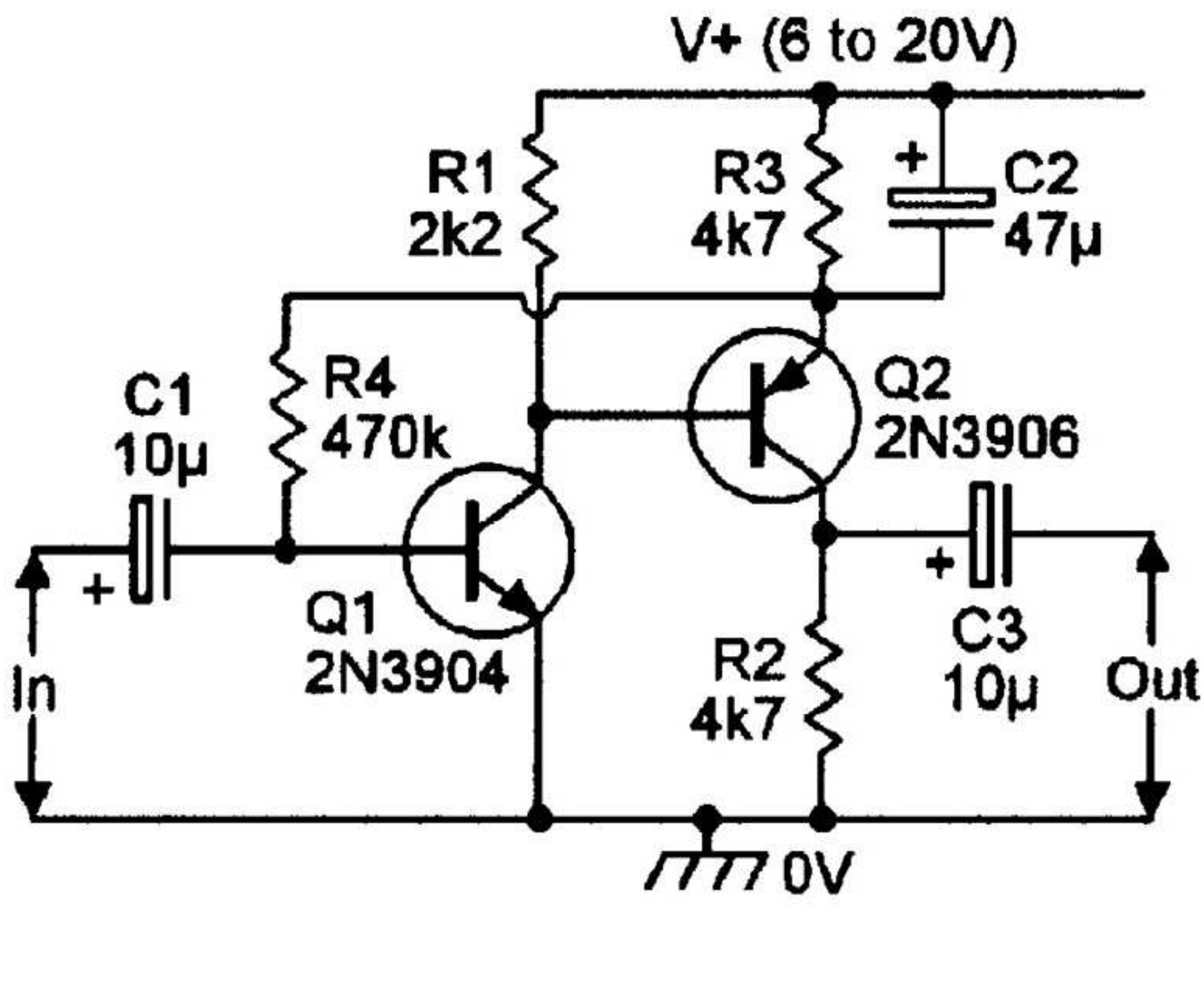


FIGURE 18. Alternative high-gain two-stage amplifier

The **Figure 19** circuit gives a voltage gain of about 66dB. Q1 is a common-emitter amplifier with a split collector load (R2-R3), and Q2 is an emitter follower and feeds its AC output signal back to the R2-R3 junction via C3, thus “bootstrapping” the R3 value (as described in last month’s installment) so that it acts as a high AC impedance. Q1 thus gives a very high voltage gain. This circuit’s bandwidth extends up to about 32kHz, but its input impedance is only 330R.

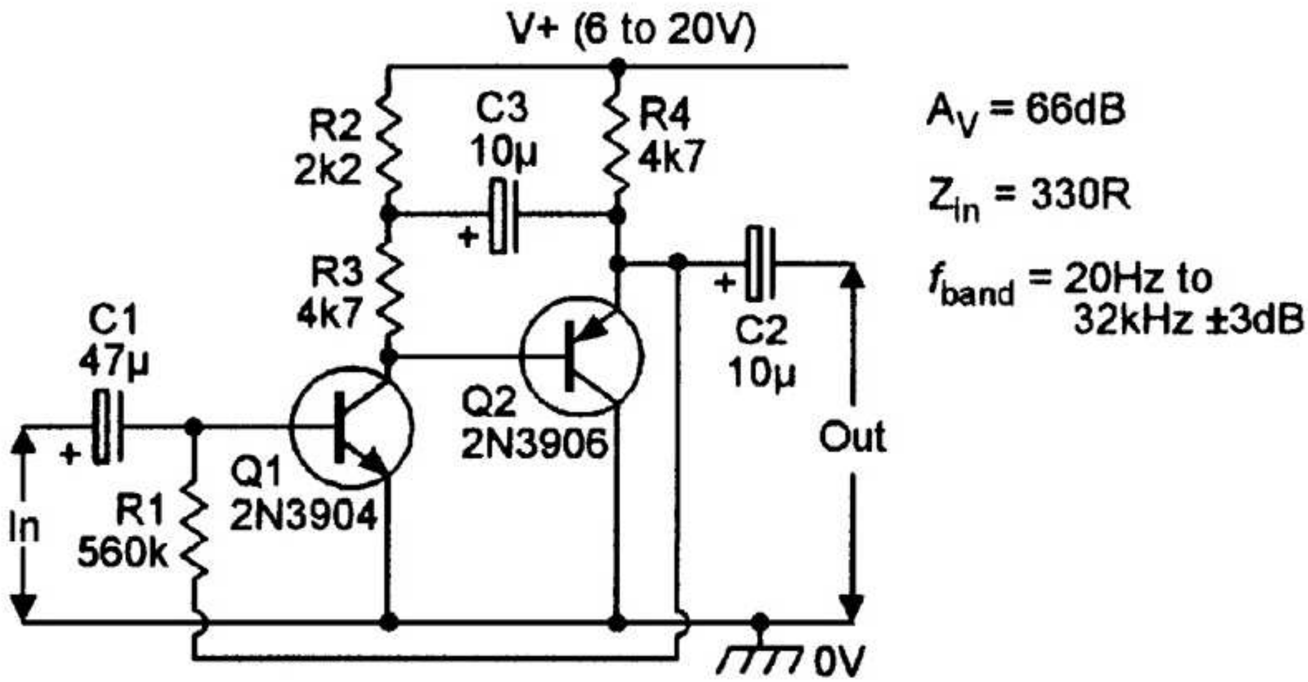


FIGURE 19. Bootstrapped high-gain amplifier

COMMON-BASE AMPLIFIER CIRCUITS

In a so-called “common-base” transistor amplifier, the input signal is applied to the transistor’s emitter, and the output is taken from the transistor’s collector. The common-base amplifier has a very low input impedance, gives near-unity current gain and a high voltage gain, and is used mainly in wide-band or high-frequency voltage amplifier applications. **Figure 20** shows an example of a common-base amplifier that gives a good wide-band response.

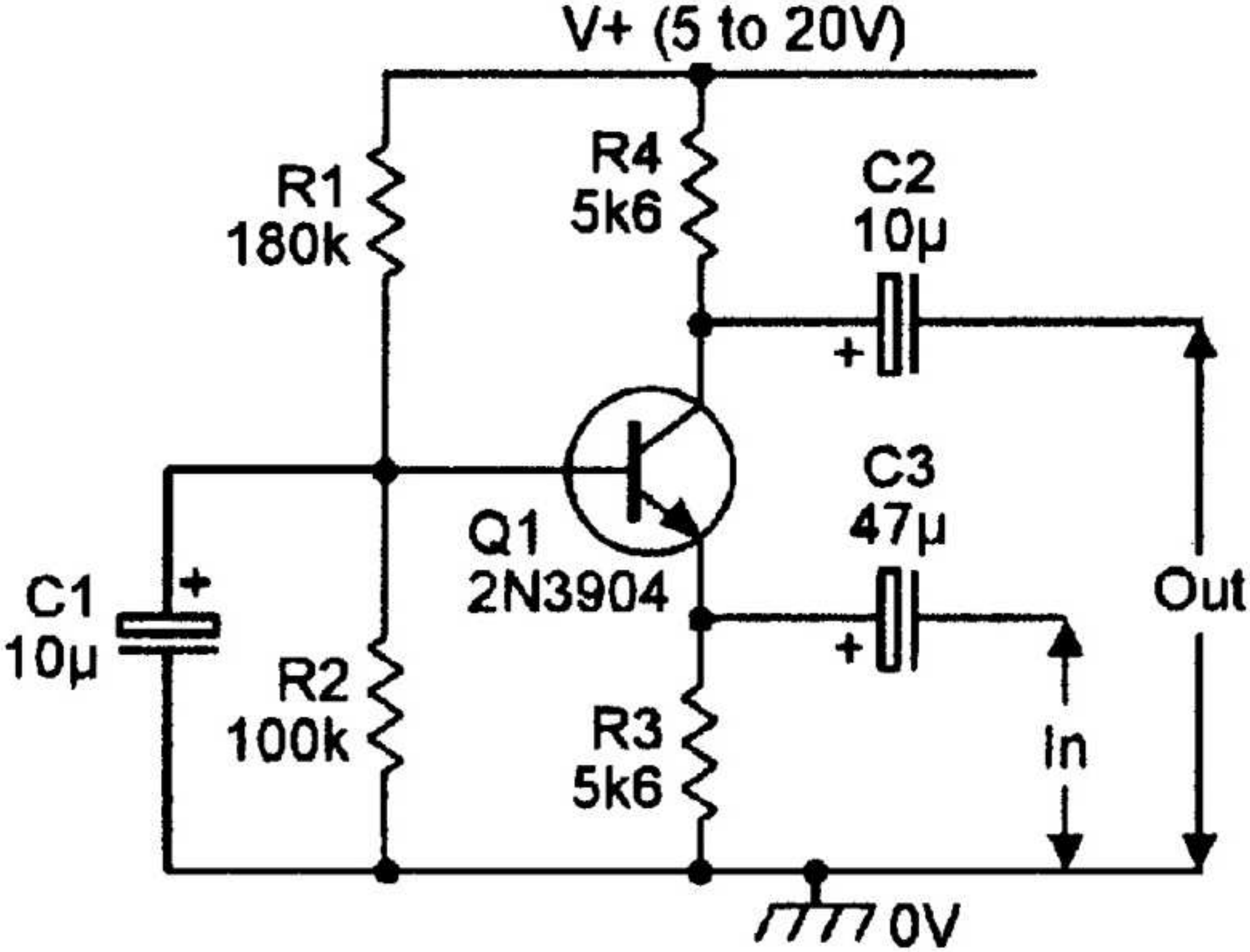


FIGURE 20. Common-base amplifier

The **Figure 20** circuit is biased in the same way as **Figure 12**. Note, however, that the base is AC-decoupled via C1, and the input signal is applied to the emitter via C3. The circuit has a very low input impedance (equal to that of Q1’s forward-biased base-emitter junction), gives the same voltage gain as the common-emitter amplifier (about 46dB), gives zero phase shift between input and output, and has a -3dB bandwidth extending to a few MHz.

Figure 21 shows an excellent wideband amplifier — the “cascode” circuit — which gives the wide bandwidth benefit of the common-base amplifier, together with the medium input impedance of the common-emitter amplifier. This is achieved by wiring Q1 and Q2 in series, with Q1 connected in the common-base mode and Q2 in the common-emitter mode.

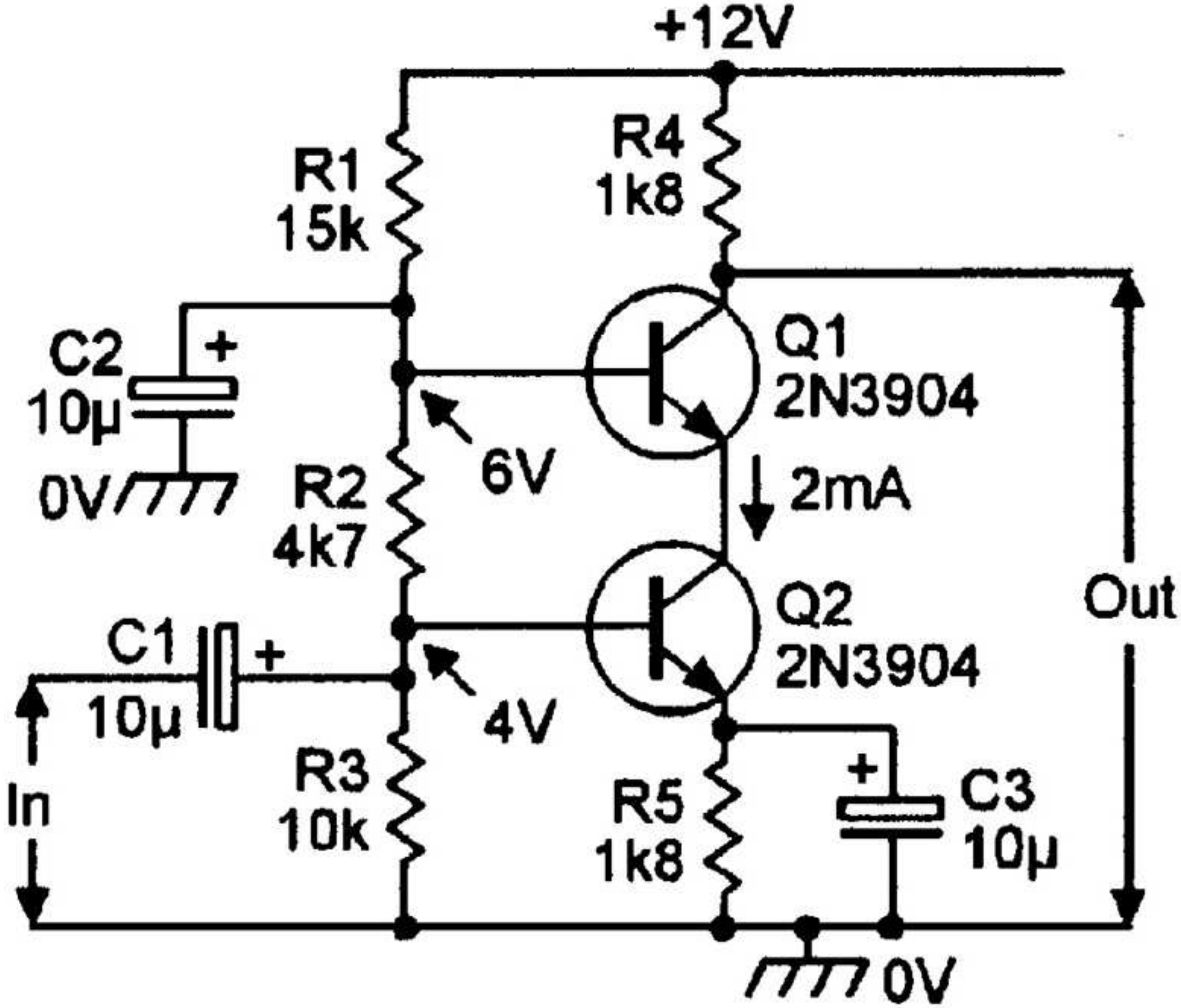


FIGURE 21. Wide-band cascode amplifier

The input signal is applied to the base of Q2, which uses Q1 emitter as its collector load and thus gives unity voltage gain and a very wide bandwidth, and Q1 gives a voltage gain of about 46dB. Thus, the complete circuit has an input impedance of about 1k8, a voltage gain of 46dB, and a -3dB bandwidth that extends to a few MHz.

Figure 22 shows a close relative of the common-base amplifier — the "long-tailed pair" phase splitter — which gives a pair of anti-phase outputs when driven from a single-ended input signal. Q1 and Q2 share a common emitter resistor (the "tail"), and the circuit bias point is set via RV1 so that the two transistors pass near-identical collector currents (giving zero difference between the two collector voltages) under quiescent conditions.

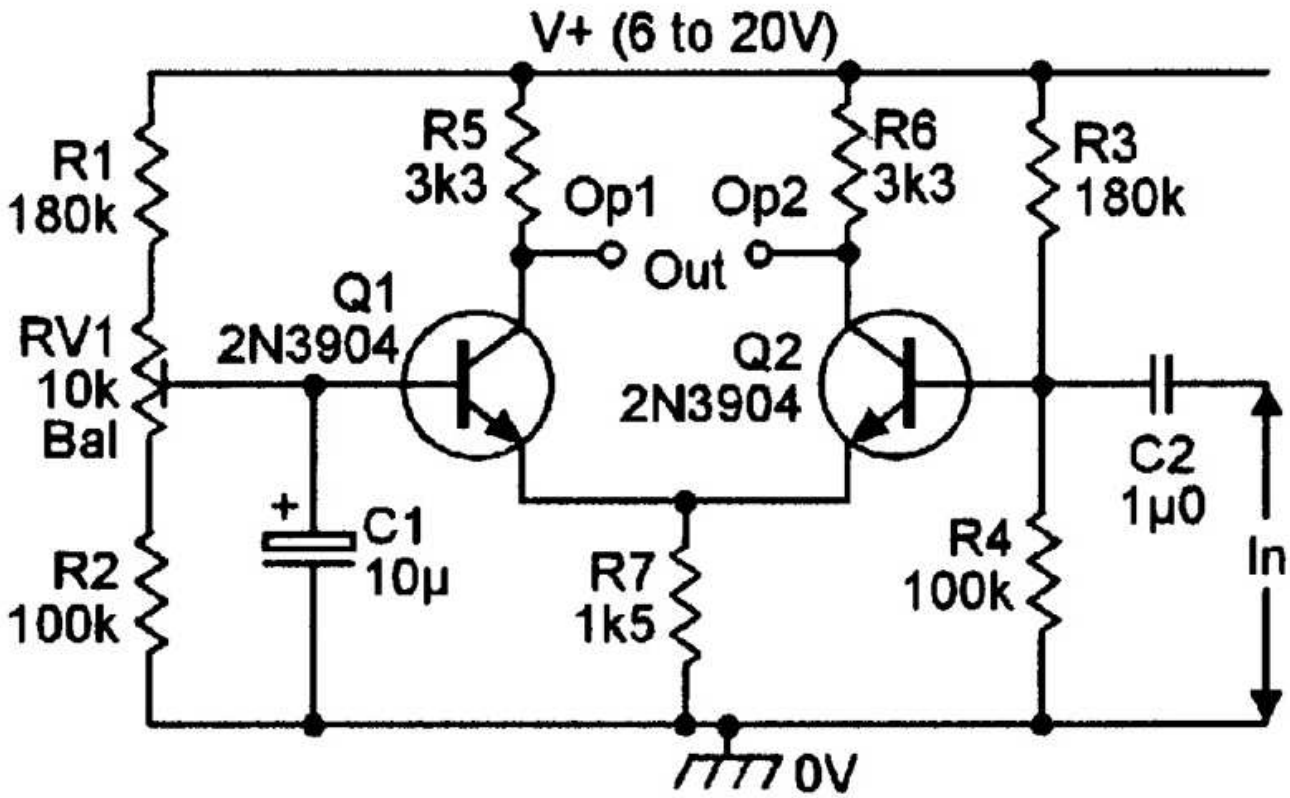


FIGURE 22. "Long-tailed pair" phase splitter

Q1 base is AC-grounded via C1, and AC-input signals are applied to Q2 base via C2. The circuit acts as follows.

Suppose that a sinewave input signal is fed to Q2 base. Q2 acts as an inverting common-emitter amplifier, and when the signal drives its base upward, its collector inevitably swings downward, and vice versa. Simultaneously, Q2's emitter "follows" the input signal, and as its emitter voltage rises, it inevitably reduces the base-emitter bias of Q1, thus making Q1's collector voltage rise, etc.

Q1 thus operates in the common-base mode and gives the same voltage gain as Q2, but gives a non-inverting amplifier action. This "phase-splitter" circuit thus generates a pair of balanced anti-phase output signals from a single-ended input.

Finally, **Figure 23** shows how the above circuit can be made to act as a differential amplifier that gives a pair of anti-phase outputs that are proportional to the difference between the two input signals — if identical signals are applied to both inputs, the circuit will (ideally) give zero output.

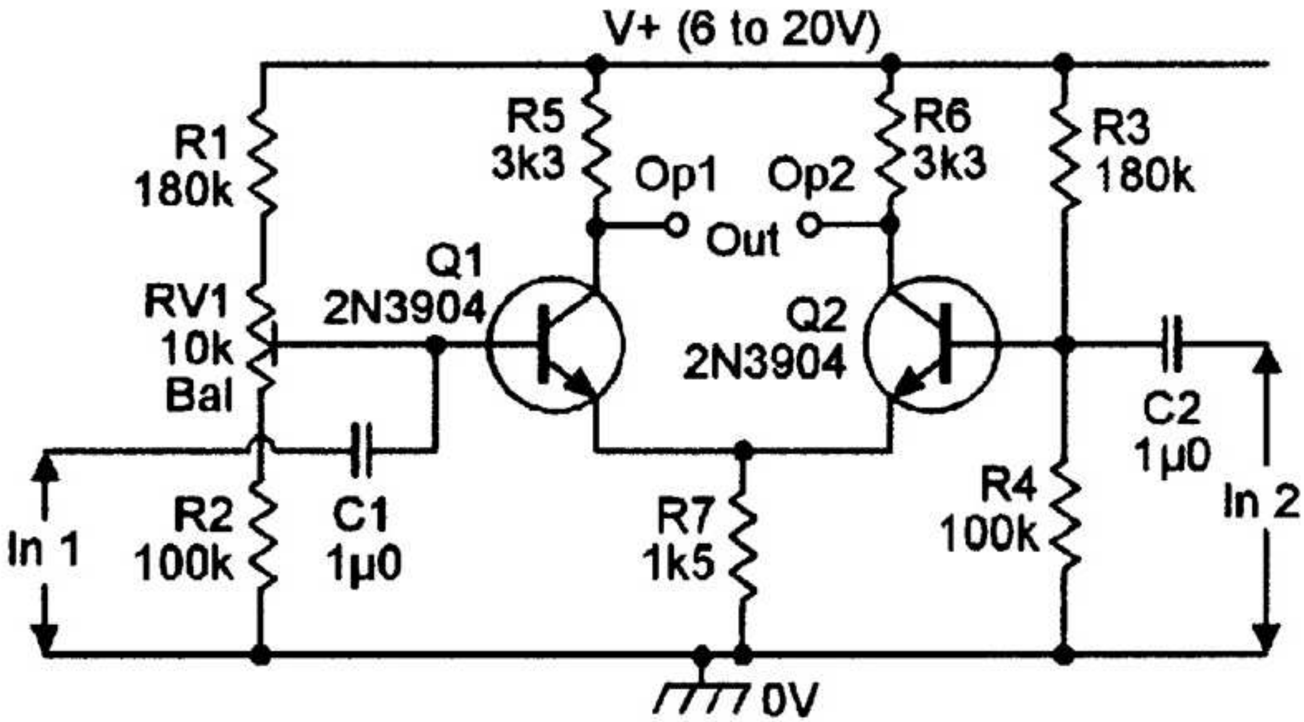


FIGURE 23. Simple differential amplifier or long-tailed pair

The second input signal is fed to Q1 base via C1, and the R7 “tail” provides the coupling between the two transistors. **NV**

Bipolar Transistor Cookbook — Part 4

Our last edition of the Transistor Cookbook series described practical ways of using bipolar transistors in simple, but useful common-emitter and common-base configurations. This time, we'll show various ways of using bipolars in practical small-signal audio amplifier applications.

AUDIO AMPLIFIER BASICS

Transistor amplifiers have many useful applications in mono and stereo audio systems. For most practical purposes, each channel of a stereo system can be broken down into three distinct circuit sections, or blocks, as shown in **Figure 1**. The first section is the selector/pre-amplifier block. It lets the user select the desired type of input signal source and applies an appropriate amount of amplification and frequency correction to the signal so that the resulting output signal is suitable for use by the second circuit block.

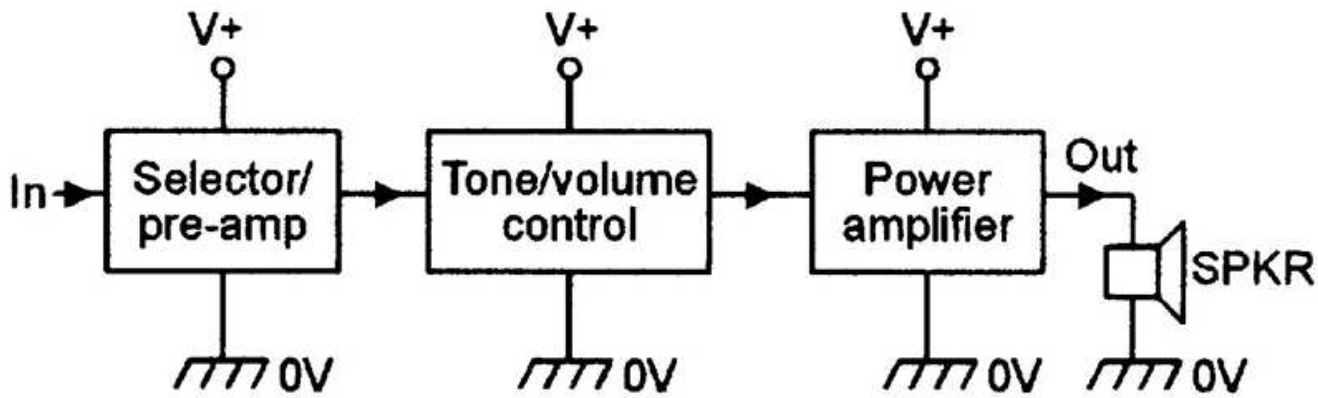


FIGURE 1.

The second section is the tone-/volume-control block, which lets the user adjust the system's frequency characteristics and output signal amplitude to suit personal tastes. This section may contain additional filter circuits and gadgets, such as scratch and rumble filters, and audio mixer circuitry, etc. Its output is fed to the system's final section — the audio power amplifier — which drives the loudspeakers. A variety of practical pre-amplifier, tone-control, and associated circuits are described here. Audio power amplifier circuits will be dealt with in a future episode of the series.

SIMPLE PRE-AMPS

The basic function of an audio pre-amplifier is that of modifying the input signal characteristics so that they give the level frequency response and nominal 100mV mean output amplitude needed to drive the amplifier's tone-control system. If the input comes from a radio tuner, CD player, etc., the signal characteristics are usually such that they can be fed directly to the tone-control sections, by-passing the pre-amplifier circuit. If they are derived from a microphone or an old-style record (disk) pick-up, they usually need modification via a pre-amp stage.

Microphones and pick-ups are usually either magnetic or ceramic/crystal devices. Magnetic types usually have a low output impedance and a low signal sensitivity or mean output amplitude (about 2mV nominal). Their outputs thus need to be fed to high-gain pre-amplifier stages. Ceramic/crystal types usually have a high output impedance and a high sensitivity (about 100mV nominal). Their outputs thus need to be fed to a high-impedance pre-amp stage with near-unity voltage gain.

Most microphones have a flat frequency response and can be used with simple pre-amp stages. **Figure 2** shows a unity-gain pre-amp that can be used with most high-impedance ceramic/crystal microphones. It is an emitter follower circuit with a bootstrapped (via C2-R3) input network, and has an input impedance of about two megohms — its supply is decoupled via C4-R5.

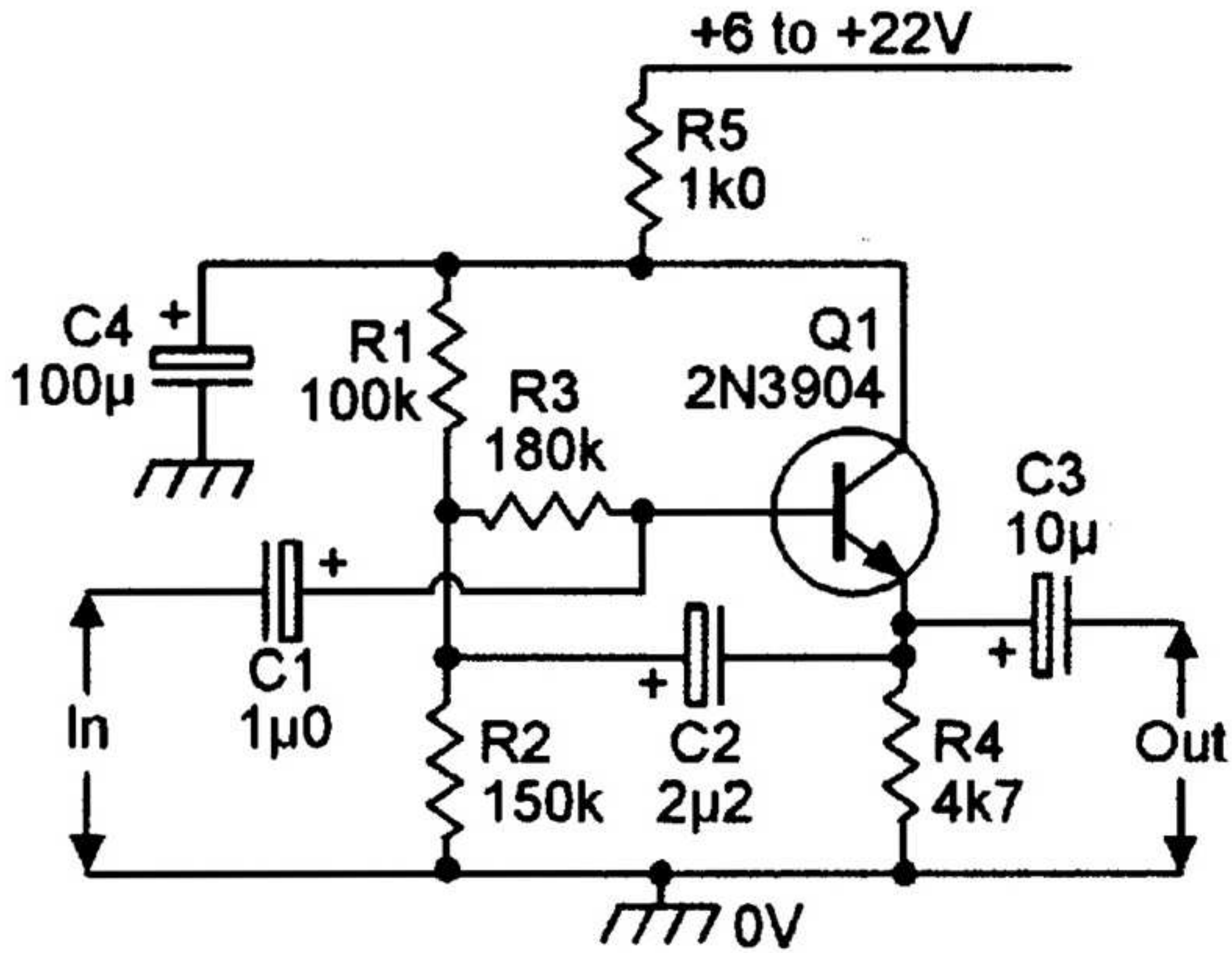


FIGURE 2.

Figures 3 and **4** show pre-amp circuits that can be used with magnetic microphones. The single-stage circuit in **Figure 3** gives 46dB (x200) of voltage gain, and can be used with most magnetic microphones.

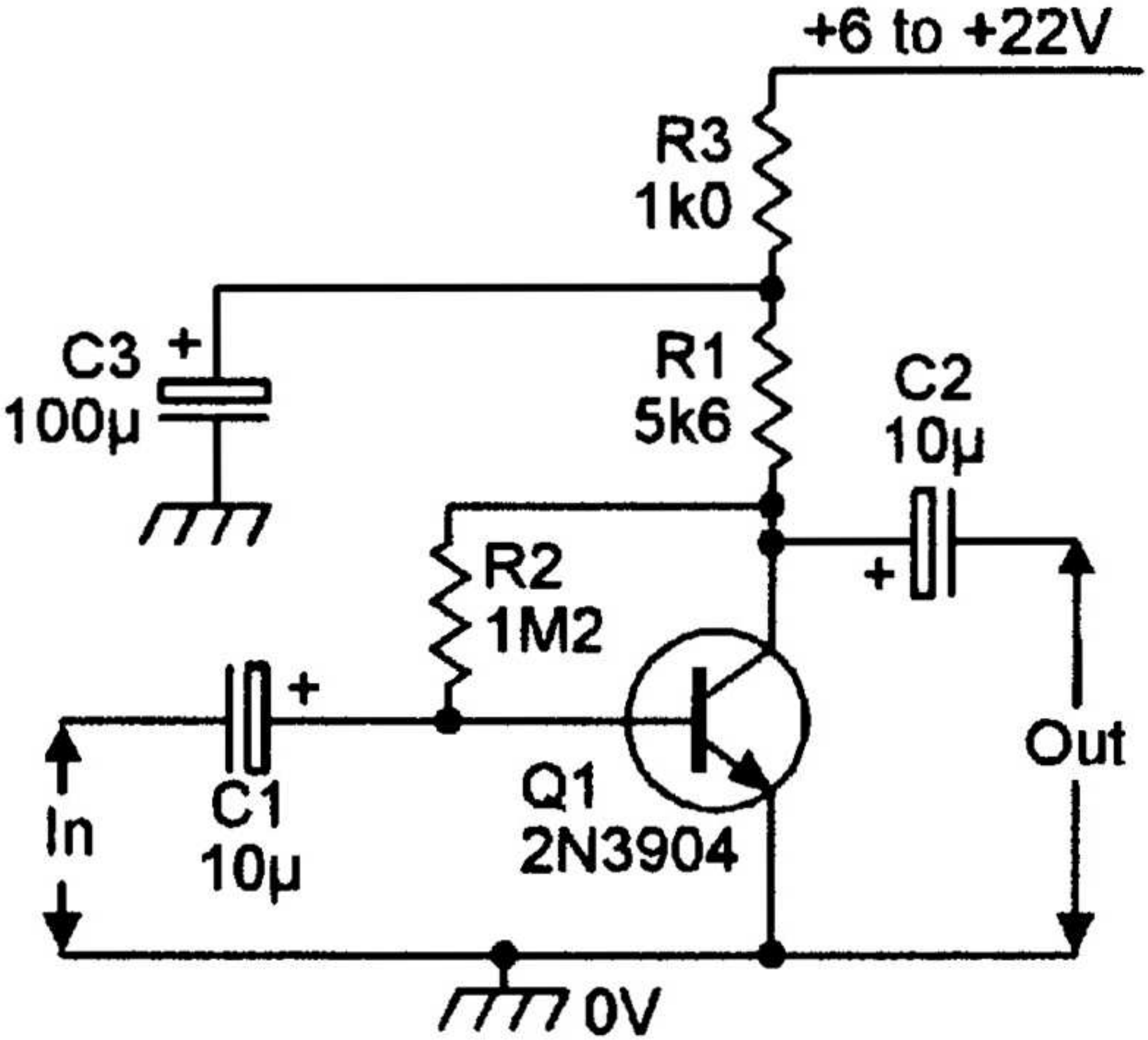


FIGURE 3.

The two-stage circuit in **Figure 4** gives 76dB of voltage gain, and is meant for use with magnetic microphones with very low sensitivity.

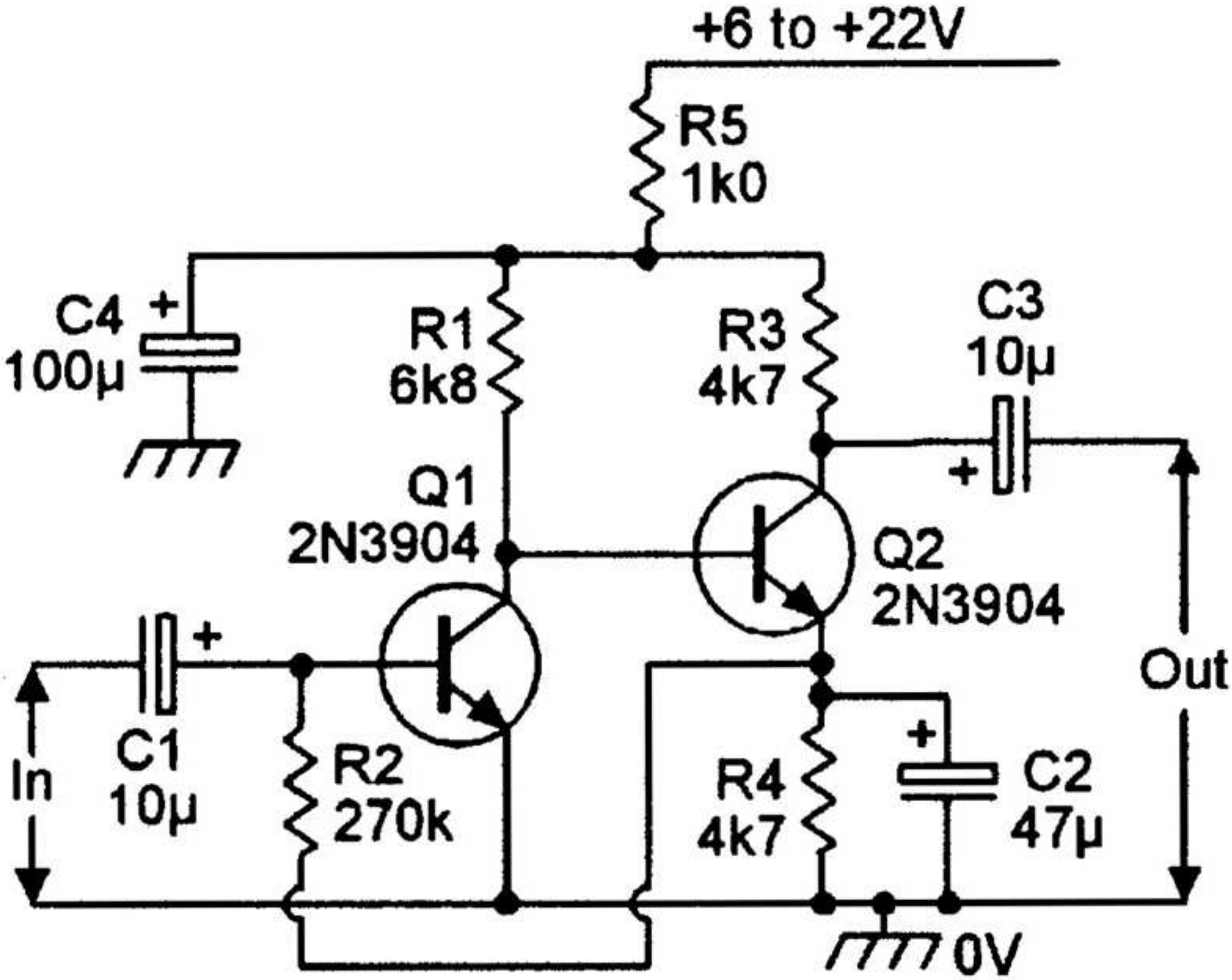


FIGURE 4.

RIAA PRE-AMP CIRCUITS

If a constant-amplitude 20 Hz to 20 KHz variable-frequency signal is recorded on a standard 33.3 RPM phonograph disk (record) using conventional stereo recording equipment, and the record is then replayed, it generates the highly non-linear frequency response curve, shown in **Figure 5** — the dotted line shows the idealized shape of this curve, and the solid line shows its practical form. The idealized response is flat between 500 Hz and 2120 Hz, but rises at a rate of 6dB/octave (20dB/decade) above 2120 Hz, and falls at a 6dB/octave rate between 500Hz and 50 Hz. The response is flat to frequencies below 50 Hz.

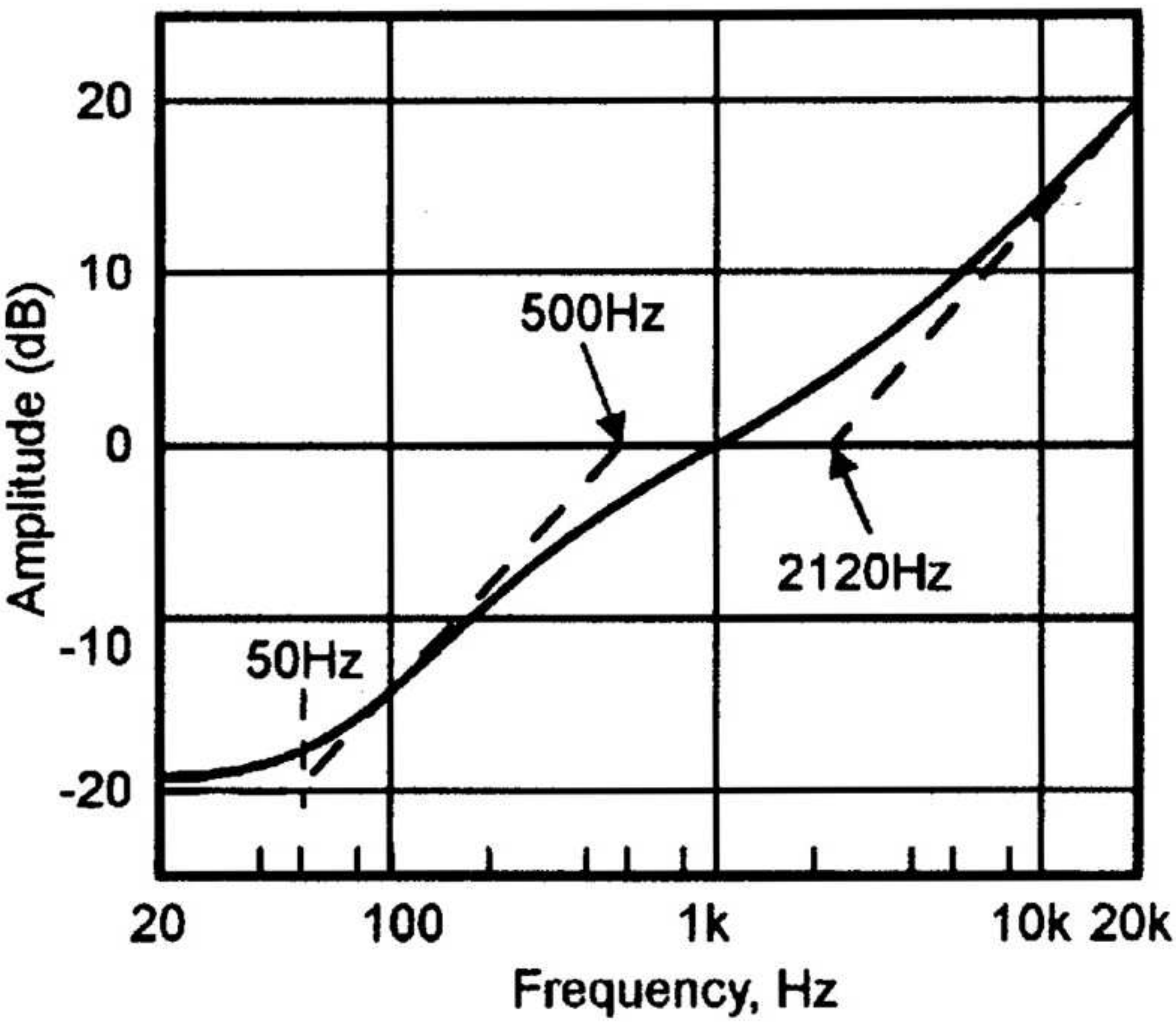


FIGURE 5.

These responses enable disk recordings to be made with good signal-to-noise ratios and wide dynamic ranges, and are used on all normal records. Consequently, when a disk is replayed, its output must be passed to the power amplifier via a pre-amp with an equalization curve that is the exact inverse of that used to make the original disk recording, so that a linear overall record-to-replay response is obtained. **Figure 6** shows the shape of the necessary RIAA.

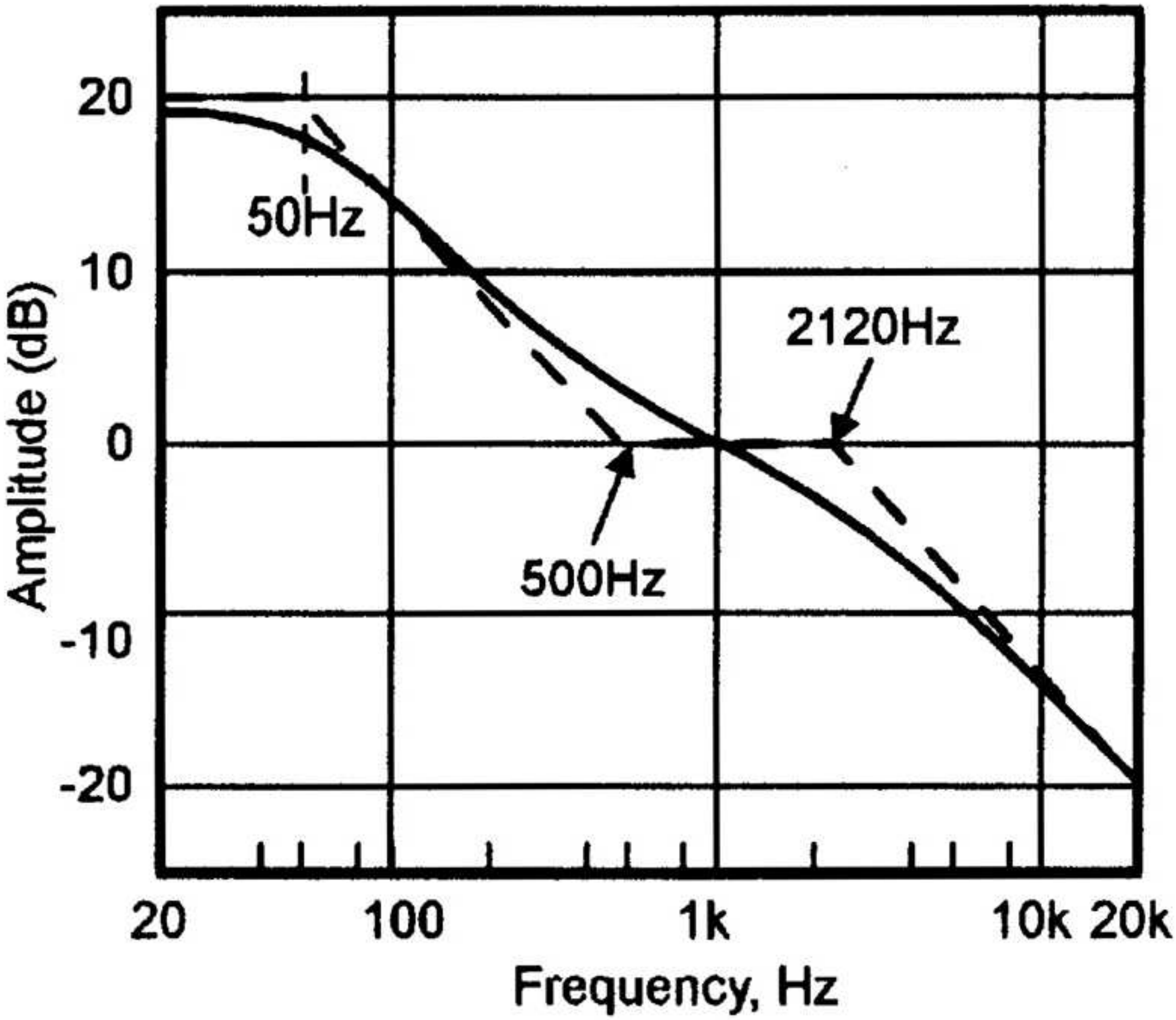


FIGURE 6.

(Record Industry Association of America) equalization curve. A practical RIAA equalization circuit can be made by wiring a pair of C-R feedback networks into a standard pre-amp (so that the gain falls as the frequency rises), with one network controlling the 50 Hz to 500 Hz response, and the other the 2120 Hz to 20 KHz response. **Figure 7** shows such an amplifier.

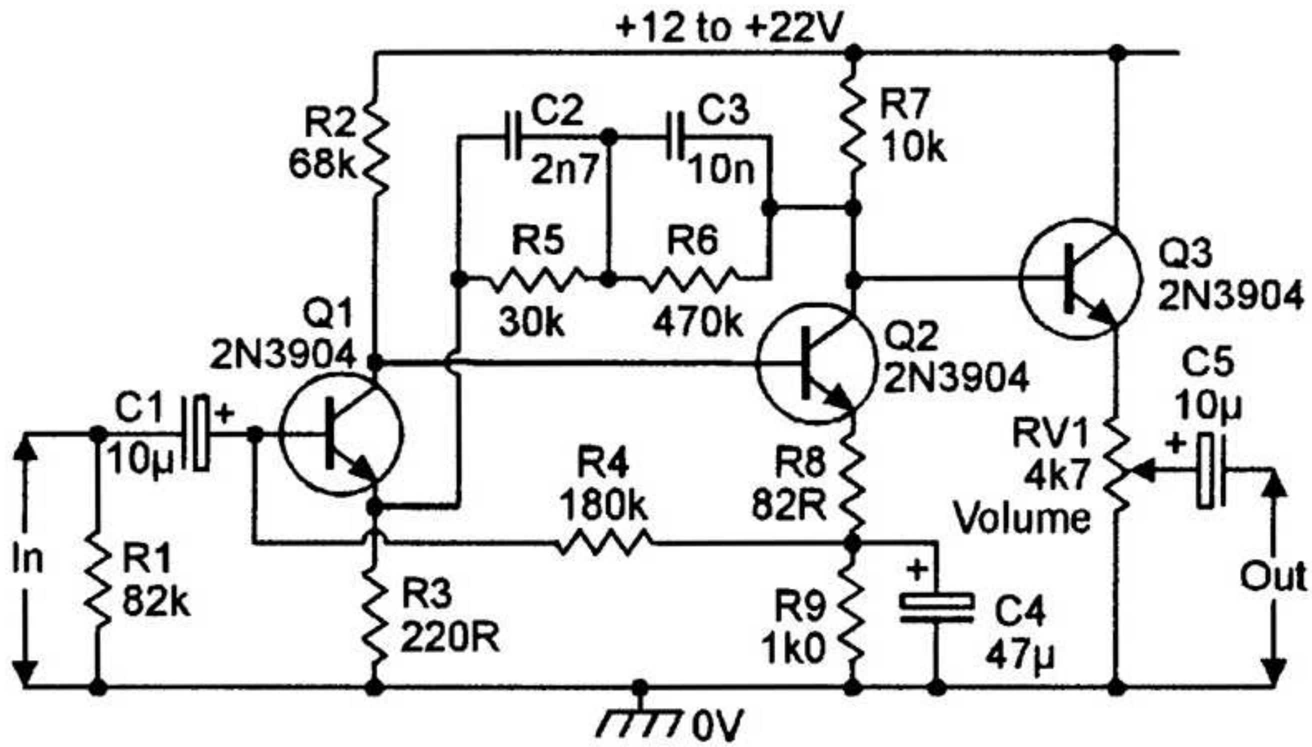


FIGURE 7.

The **Figure 7** circuit can be used with any magnetic pick-up cartridge. It gives a 1V output from a 6mV input at 1 KHz, and provides equalization that is within 1dB of the RIAA standard between 40 Hz and 12 KHz. The actual pre-amp is designed around Q1 and Q2, with C2-R5 and C3-R6 forming the feedback equalization network. Q3 is an emitter follower buffer stage, and drives optional volume control RV1.

Ceramic/crystal pick-ups usually give a poorer reproduction quality than magnetic types, but give output signals of far greater amplitude. They can thus be used with a very simple type of equalization pre-amp, and are consequently found in many popular record player systems. **Figures 8** and **9** show alternative phonograph pre-amplifier circuits that can be used with ceramic or crystal pick-up cartridges. In each case, the pre-amp/equalizer circuit is designed around Q1, and Q2 is an emitter follower output stage that drives optional volume control RV1. The **Figure 8** circuit can be used with any pick-up cartridge that has a capacitance in the 1000pF to 10,000pF range. Two-stage equalization is provided via C1-R2 and C2-R3, and is typically within 1.6dB of the RIAA standard between 40 Hz and 12 KHz.

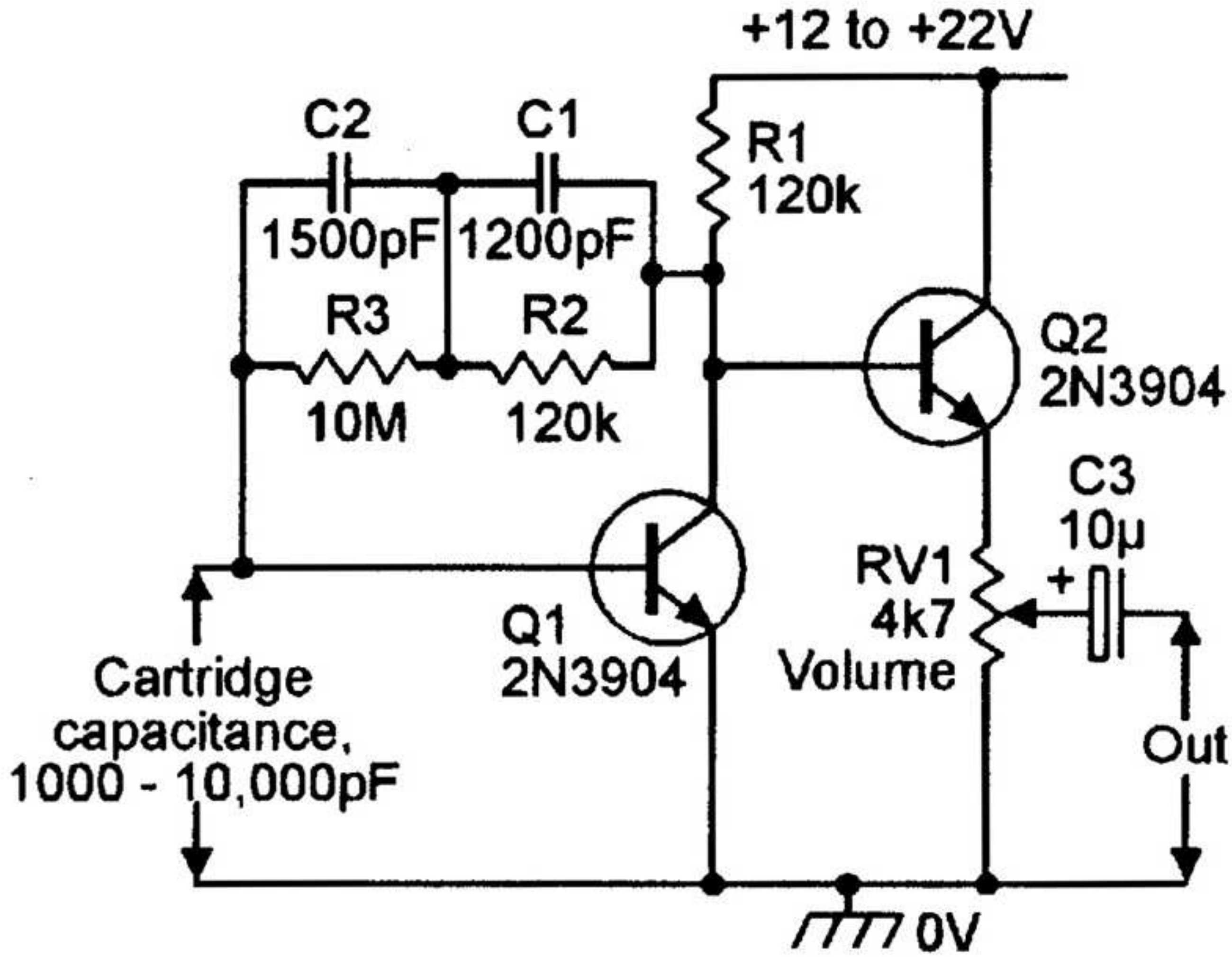


FIGURE 8.

The alternative **Figure 9** circuit can only be used with pick-ups with capacitance values in the range 5000pF to 10,000pF, since this capacitance forms part of the frequency response network The other part is formed by C1-R3. At 50-60 Hz, this circuit has a high input impedance (about 600K), and causes only slight cartridge loading. As the frequency is increased, however, the input impedance decreases sharply, thus increasing the cartridge loading and effectively reducing the circuit gain. The equalization curve approximates the RIAA standard, and the performance is adequate for many practical applications.

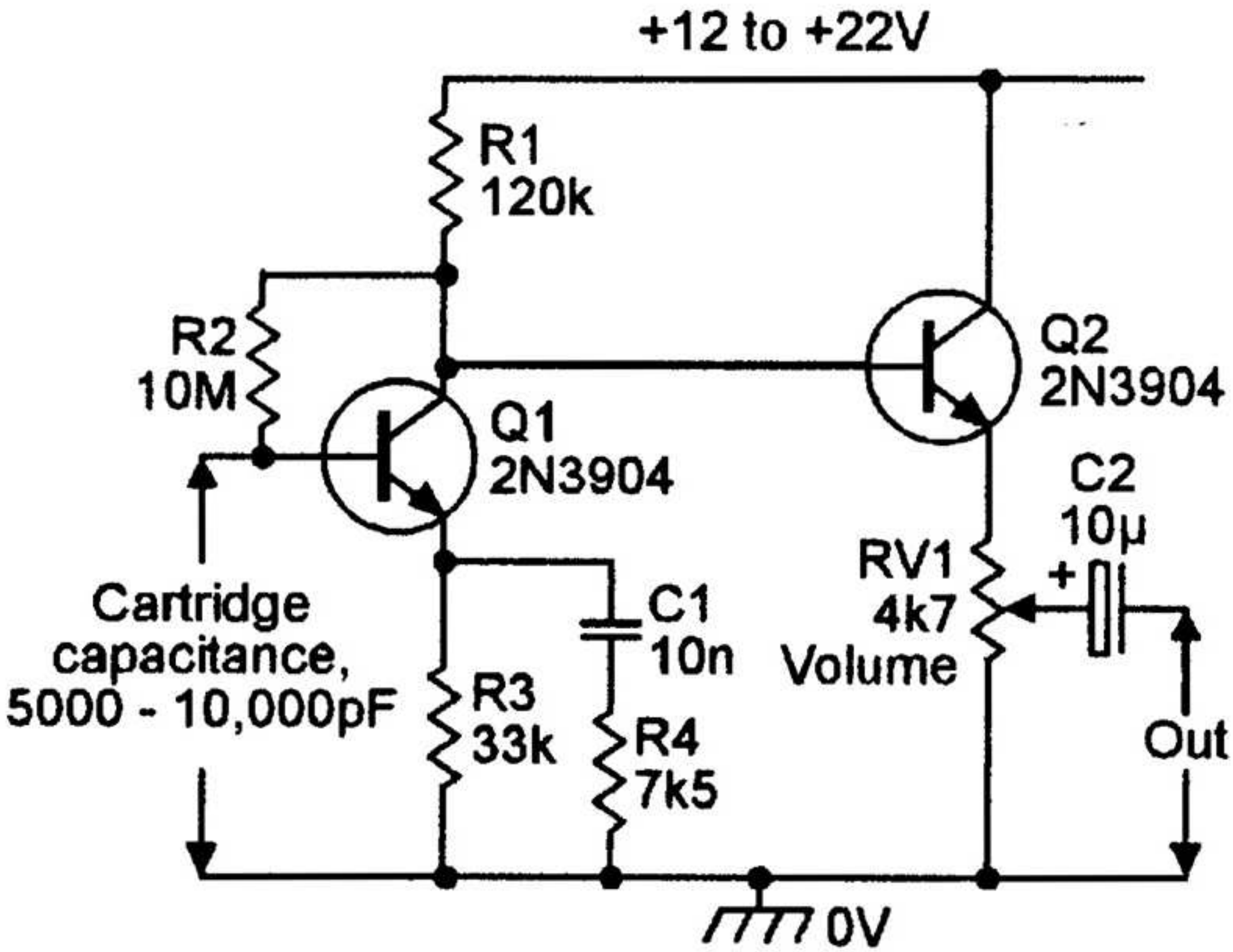


FIGURE 9.

A UNIVERSAL PRE-AMP

Most audio amplifiers use pre-amps with variable characteristics, such as a high-gain linear response for use with magnetic microphones, low-gain linear response for use with a radio tuner, and high-gain RIAA equalization for use with a magnetic pick-up cartridge, etc. To meet this requirement, it is normal to fit the system with a single universal pre-amp circuit of the type shown in **Figure 10**. This is basically a high-gain linear amplifier that can have its characteristics altered by switching alternative types of resistor/filter networks into its feedback loops.

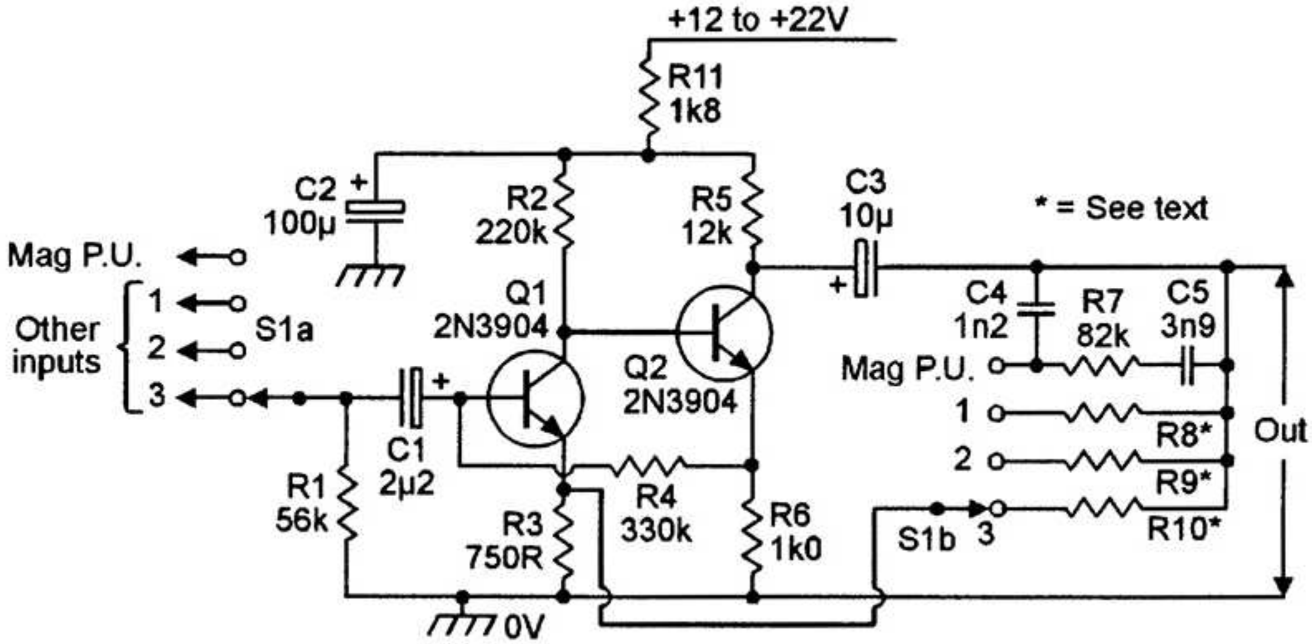


FIGURE 10.

Thus, when the selector switch is set to the “MAG P.U.” position, S1a connects the input to the magnetic pick-up cartridge, and S1b connects the C4-R7-C5 RIAA equalization network into the feedback loop. In the remaining switch positions, alternative input sources are selected via S1a, and appropriate linear-response gain-controlling feedback resistors (R8, R9 and R10) are selected via S1b. The values of these feedback resistors should be selected (between 10K and 10M) to suit individual requirements — the circuit gain is proportional to the feedback resistor value.

VOLUME CONTROL

The volume control circuitry of an audio amplifier system is normally placed between the output of the pre-amp and the input of the tone-control circuitry, and consists of a variable potential divider or pot. This pot can form part of an active circuit, as shown in **Figures 7** through **9**, but a snag here is that rapid variations of the control can briefly apply DC potentials to the next circuit, possibly upsetting its bias and generating severe signal distortion.

Figure 11 shows the ideal form and location of the volume control. It is fully DC-isolated from the pre-amp’s output via C1, and from the input of the tone-control circuitry via C2. Variation of the RV1 slider thus has no effect on the DC bias levels of either circuit. RV1 should be a log-type of pot.

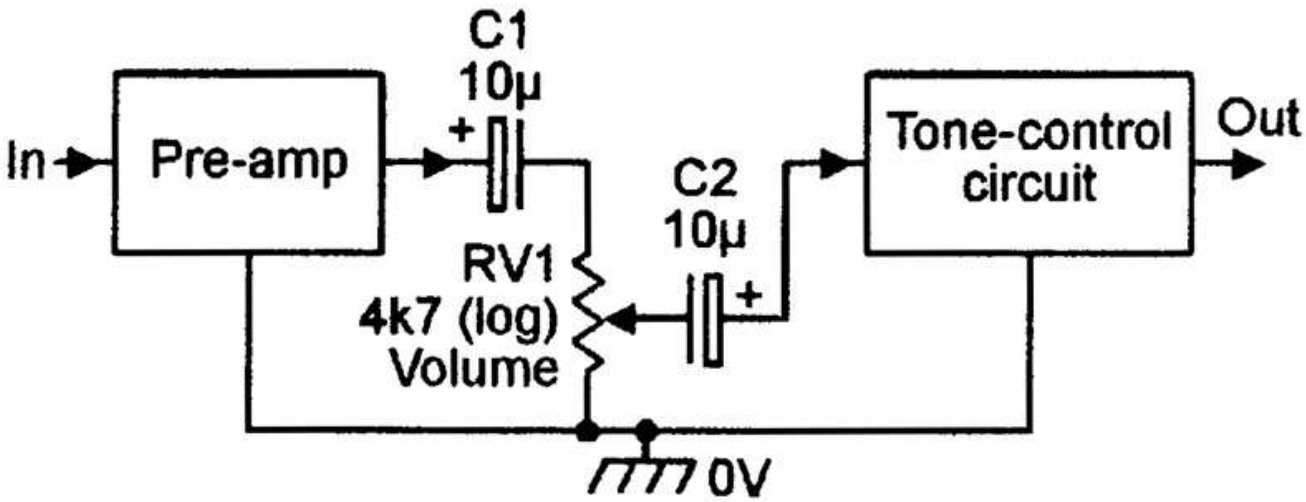


FIGURE 11.

TONE CONTROL CIRCUITS

A tone control network lets the user alter the frequency response of the amplifier system to suit a personal mood or requirement. Simple tone control networks consist of collections of C-R filters, through which the audio signals are passed — these networks are passive, and cause some degree of signal attenuation. **Figure 12** shows the practical circuit of a passive tone control network that gives about 20dB of signal attenuation when the bass and treble controls are in the flat position, and gives maximum bass and treble boost and cut values of about 20dB relative to the flat performance. The input to this circuit can be taken from the circuit's volume control, and the output can be fed to the input of the main power amplifier.

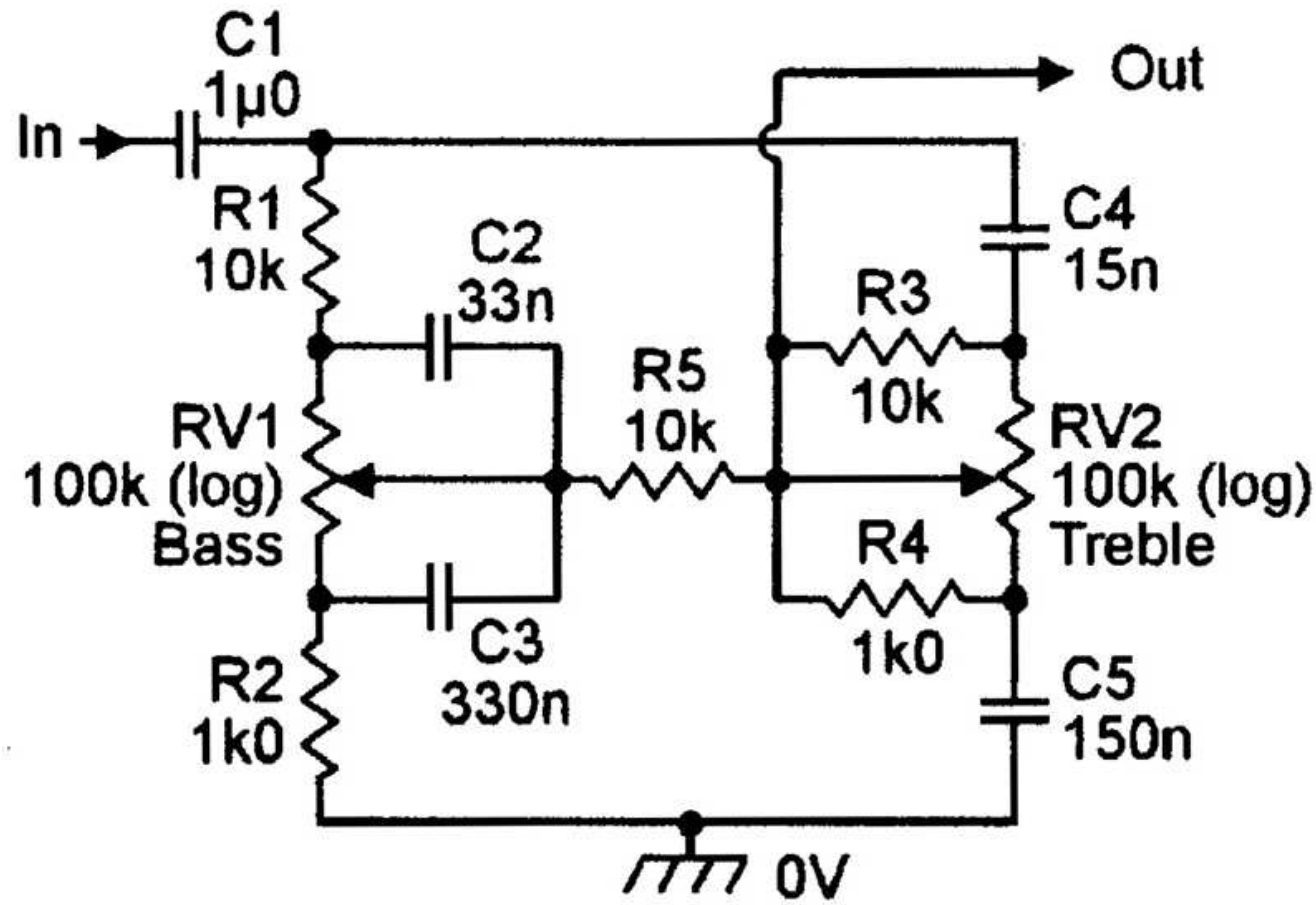


FIGURE 12.

The basic action of the **Figure 12** tone control network can be understood with the help of **Figures 13** and **14**, which show (a) the basic circuit and its equivalents under (b) boost, (c) cut, and (d) flat conditions of the bass and treble tone control networks, respectively. Brief explanations of these two diagrams are as follows. In the **Figure 13** bass control diagram, C1 is shorted out via RV1 when RV1 is in the maximum boost position, to give the equivalent circuit of (b), which gives only slight bass attenuation. When RV1 is in the maximum cut position, it shorts out C2, to give the equivalent circuit of (c), which gives roughly 40dB of bass attenuation. Finally, when RV1 is in the flat position, it gives the equivalent circuit of (d), which gives about 20dB of signal attenuation at all frequencies. Thus, this bass control circuit gives a maximum of about 20dB of bass boost or cut relative to the flat signals.

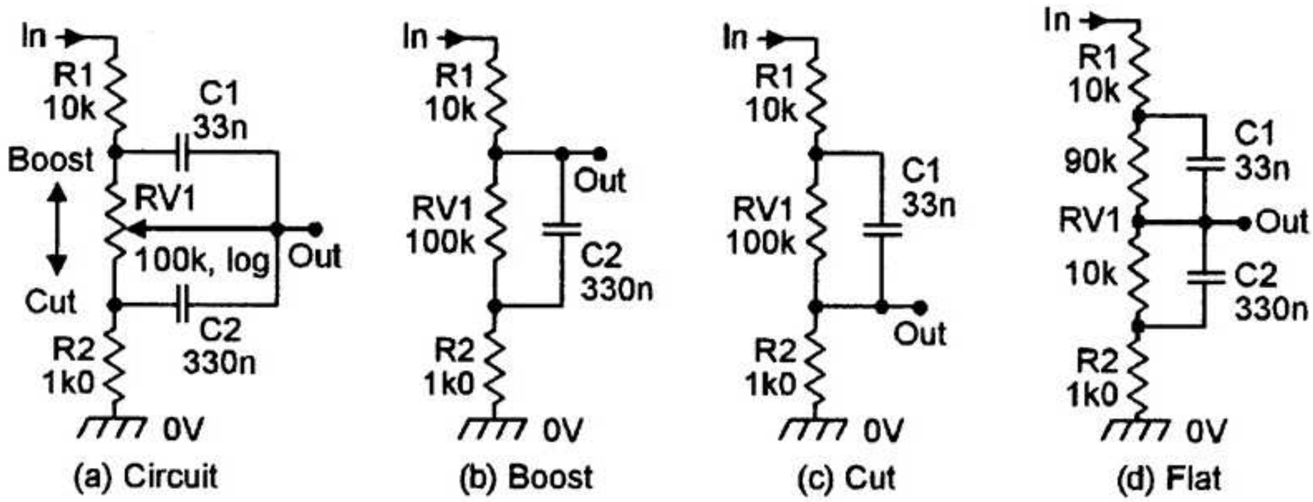


FIGURE 13.

In the **Figure 14** treble control diagram, R1 is shorted out when RV1 is in the maximum boost position, to give the equivalent circuit of (b), and R2 is shorted out when RV1 is in the maximum cut position, to give the equivalent circuit of (c). When RV1 is set to the flat position, the circuit equivalent is that of (d), which gives about 20dB of signal attenuation at all frequencies. The net result is that this treble control circuit gives a maximum of about 20dB of treble boost or cut relative to the flat signals.

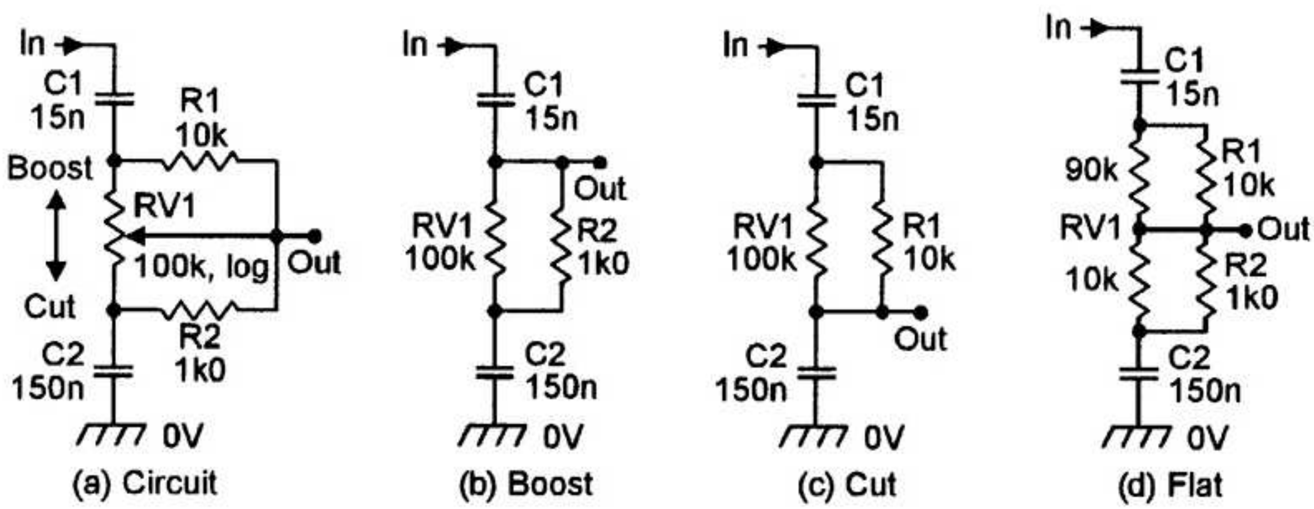


FIGURE 14.

A passive tone control network of the basic type described above can easily be wired into the feedback path of a transistor amplifier so that the system gives an overall signal gain (rather than attenuation) when its controls are in the flat position. **Figure 15** shows a practical example of an active tone control circuit of this type. In this particular example, the design uses a modified version of the basic **Figure 12** tone control circuit, which enables the tone-control circuit to use three (rather than four) tone-control capacitors.

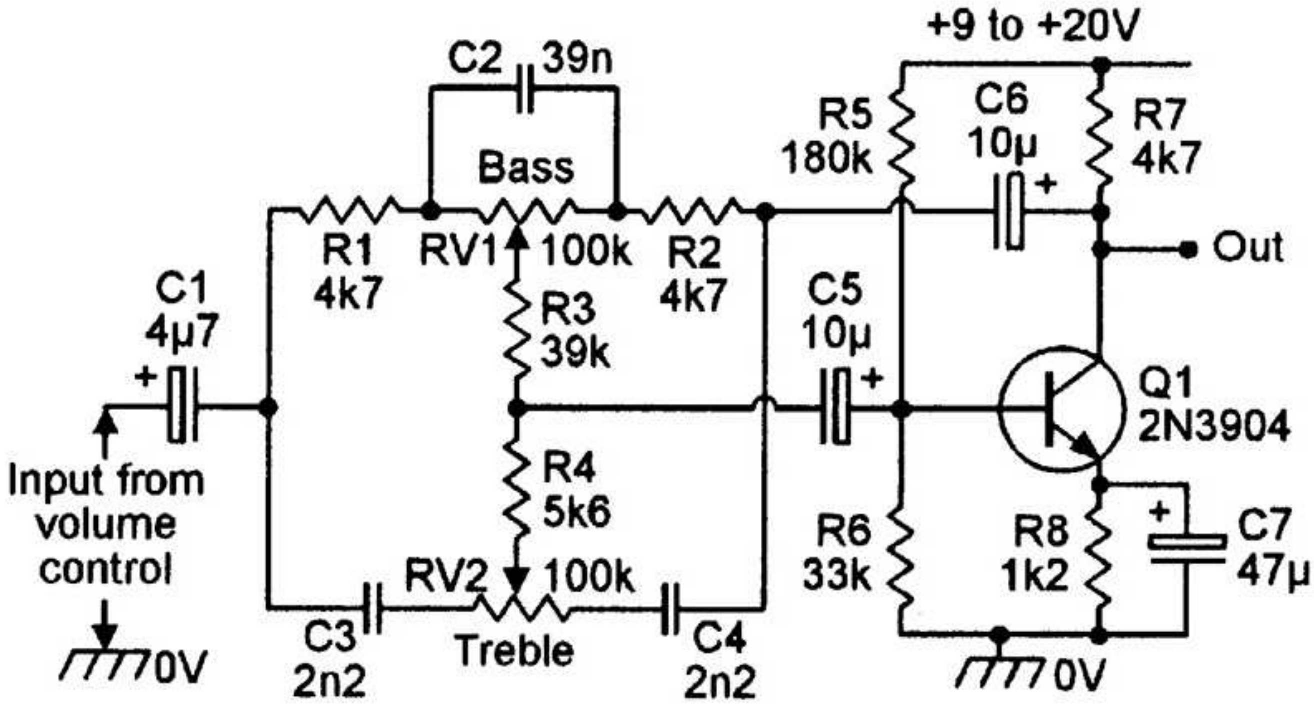


FIGURE 15.

AUDIO MIXER CIRCUITS

One useful gadget that can be fitted in the area of the volume/tone-control section of an audio amplifier is a multi-channel audio mixer, which enables several different audio signals to be mixed together to form a single composite output signal. This can be useful in, for example, enabling the user to hear the emergency sounds of a front-door or baby-room microphone, etc., while listening to normal entertainment sources.

Figure 16 shows an example of a simple three-channel audio mixer that gives unity gain between the output and each input. Each input channel comprises a single 100nF capacitor (C1) and 100K resistor (R1), and presents an input impedance of 100K. The circuit can be given any desired number of input channels by simply adding more C1 and R1 components. In use, the mixer should be placed between the output of the tone-control circuitry and the input of the main power amplifier, with one input taken from the tone-control output and the others taken from the desired signal sources.

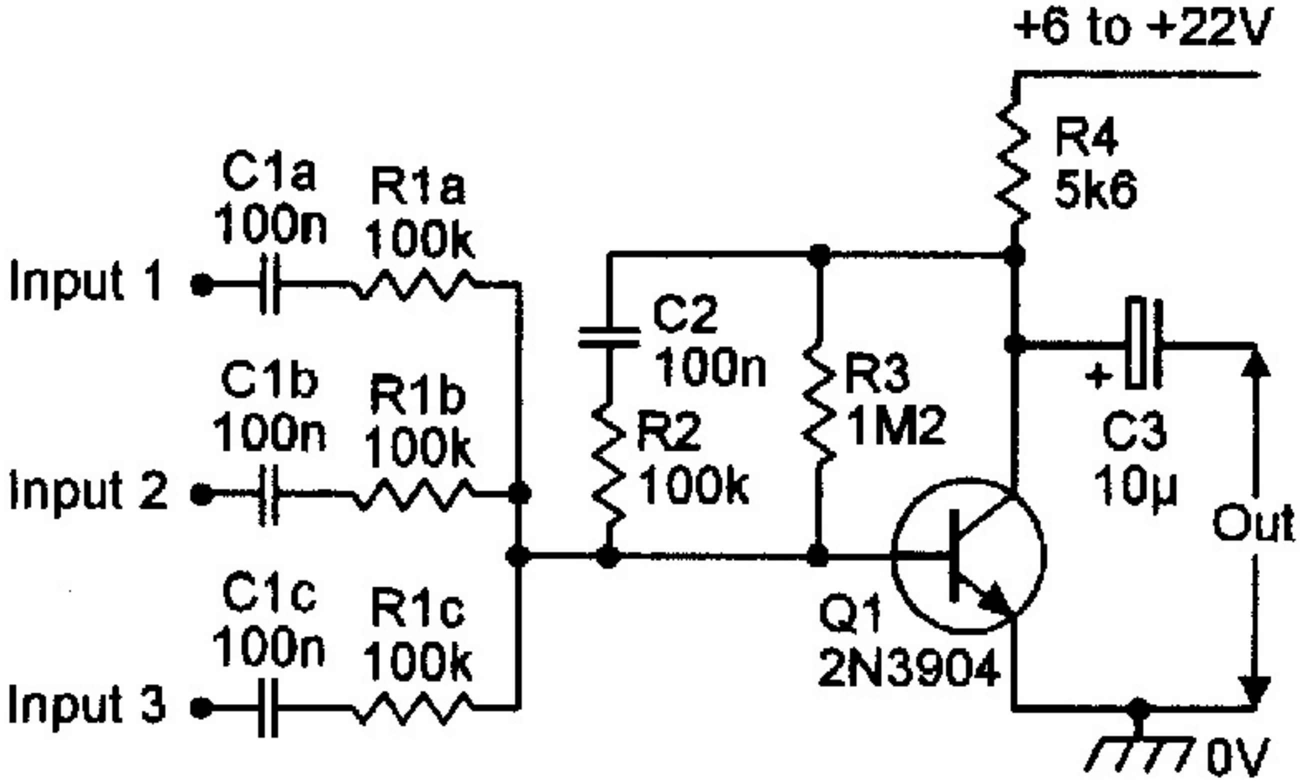


FIGURE 16.

Figure 17 shows a simple way of adding independent volume and on/off control to any desired number of input channels of the basic **Figure 16** audio mixer circuit — RV1 controls the volume, and S1 provides the on/off function.

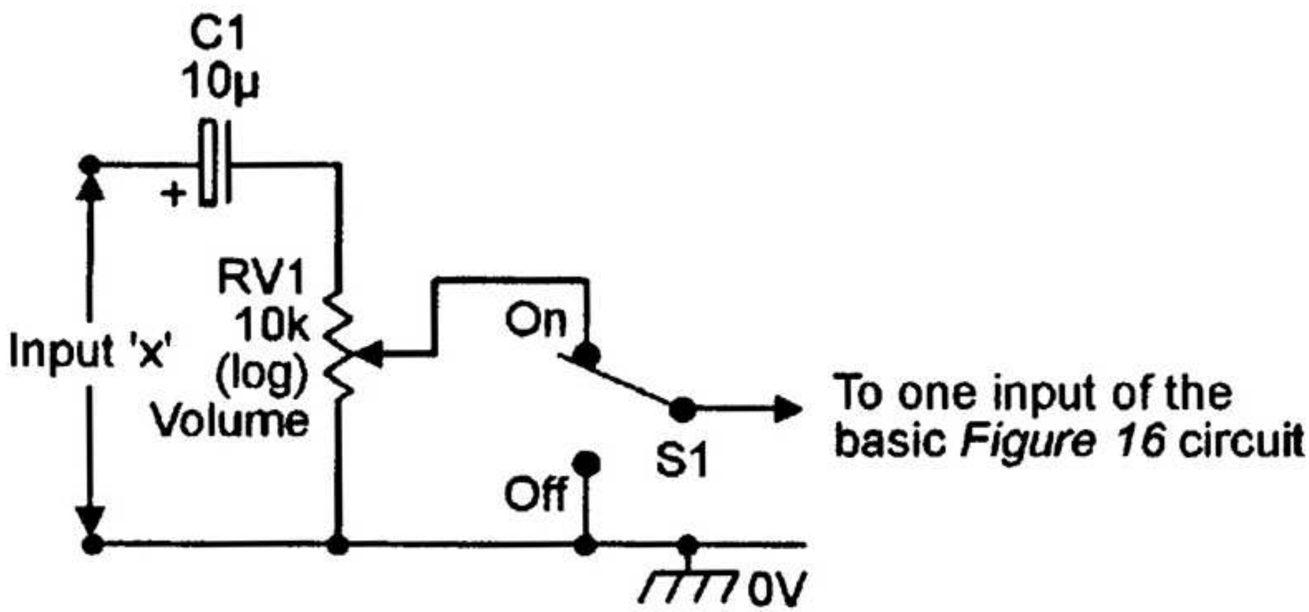


FIGURE 17.

SCRATCH/RUMBLE FILTERS

A common annoyance when playing old records/disks is that of scratch and/or rumble sounds. The scratch noises are mainly high-frequency (greater than 10 KHz) sounds picked up from the disk surface, and the rumbles are low-frequency (less than 50 Hz) sounds that are mostly caused by slow variations in motor-drive speed. Each of these noises can be greatly reduced or eliminated by passing the player's audio signals through a filter that rejects troublesome parts of the audio spectrum. **Figures 18** and **19** show suitable circuits.

The high-pass rumble filter in **Figure 18** gives unity voltage gain to signals above 50 Hz, but gives 12dB per octave rejection to those below this value, i.e., it gives 40dB of attenuation at 5 Hz, etc. Emitter-follower Q1 is biased at half-supply volts from the R1-R2-C3 low-impedance point, but has negative feedback applied via the R3-C2-C1-R4 filter network. The circuit's frequency turnover point can be altered by changing the C1-C2 values (which must be equal). Thus, if the C1-C2 values are halved (to 110nF), the turn-over frequency doubles (to 100 Hz), etc.

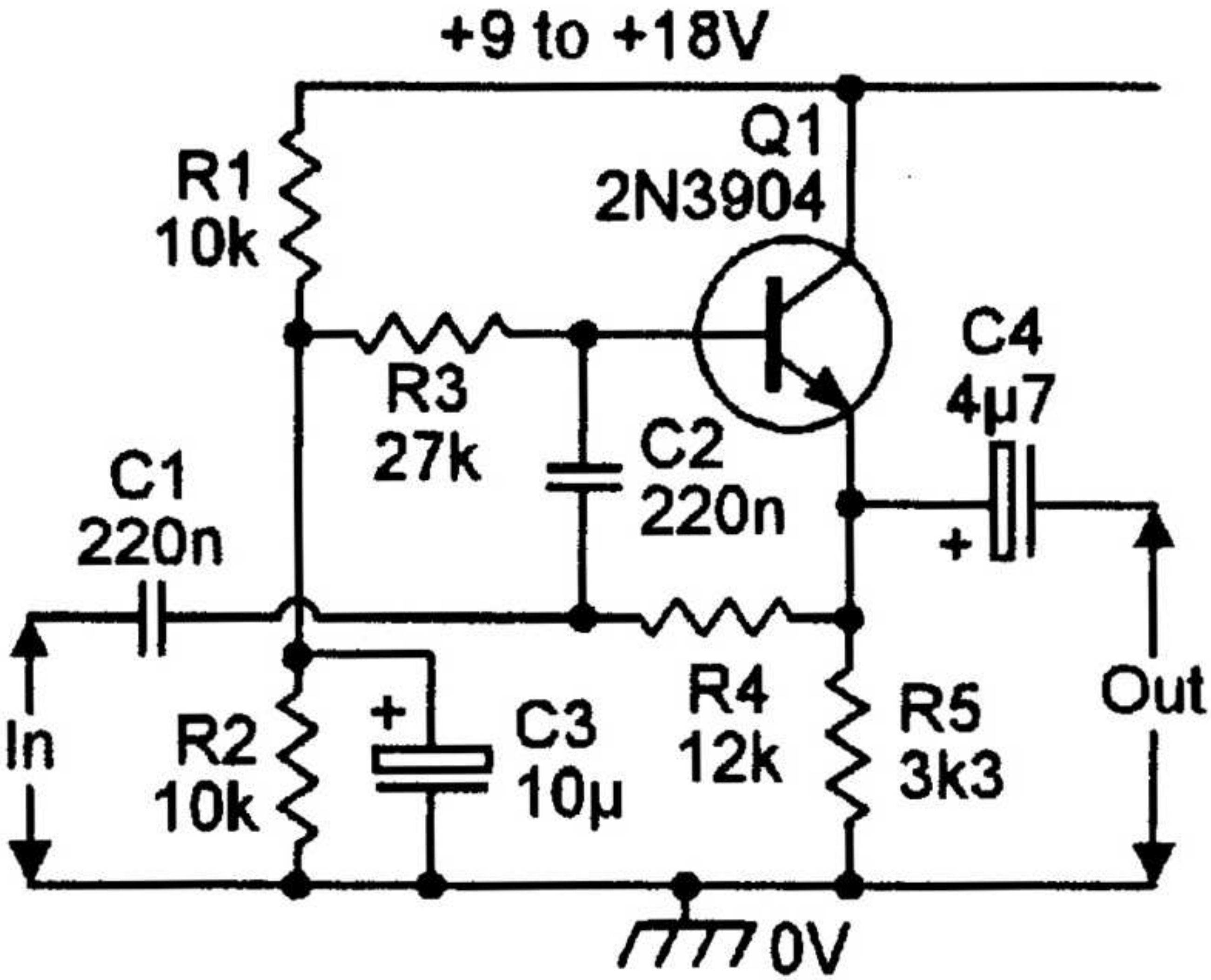


FIGURE 18.

The low-pass scratch filter in **Figure 19** gives unity voltage gain to signals below 10 KHz, but gives 12dB per octave rejection to those above this value. This circuit is similar to that in **Figure 18**, except that the positions of the main filter network components are transposed. The circuit's turn-over frequency can be altered by changing the C2-R4 values; e.g., values of 3.3nF give a frequency of 7.5 KHz.

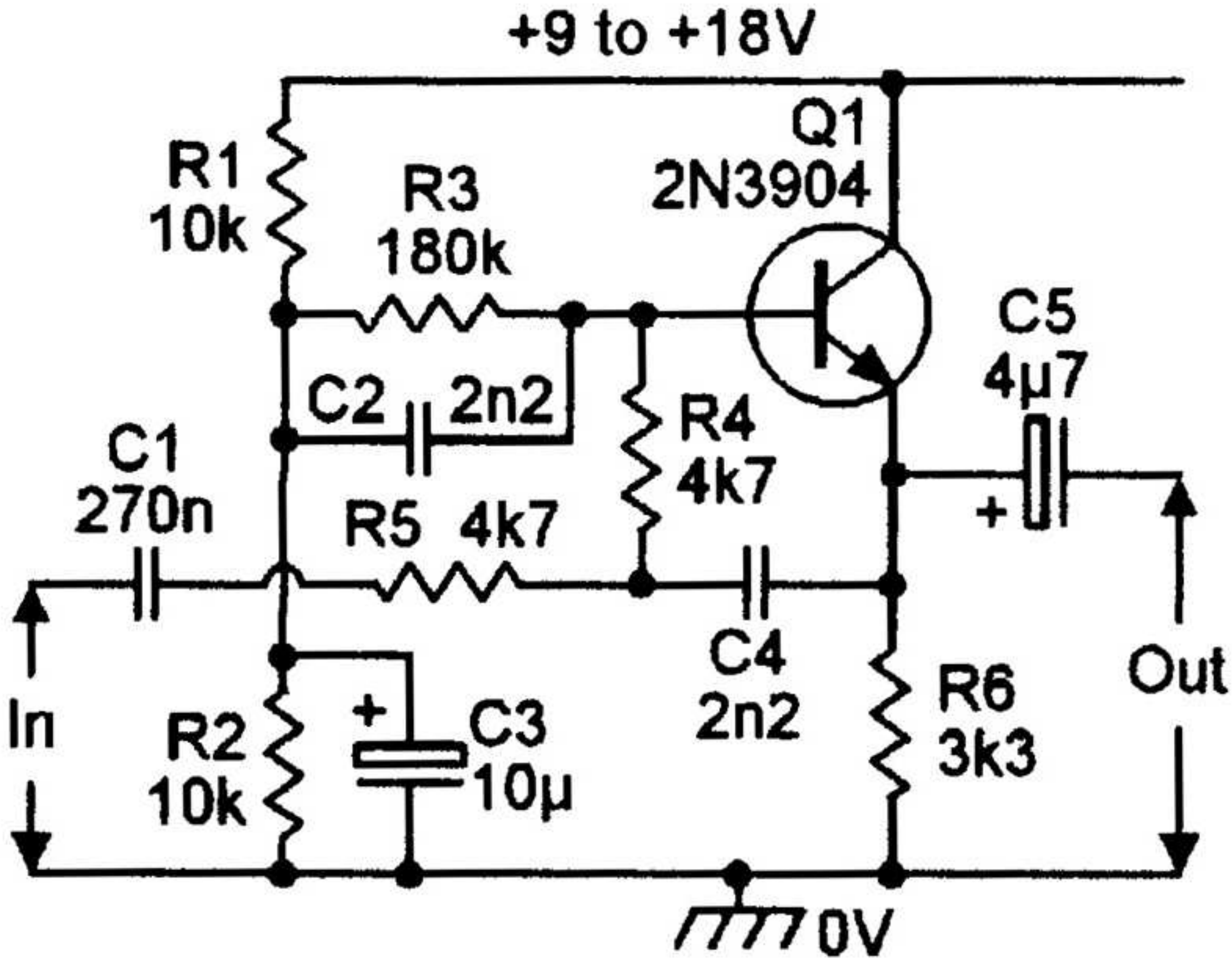


FIGURE 19.

The **Figure 18** and **19** circuits can be combined, to make a composite scratch and rumble filter, by connecting the output of the high-pass filter to the input of the low-pass filter. If desired, the filters can be provided with bypass switches, enabling them to be easily switched in and out of circuit, by using the connections shown in **Figure 20**.

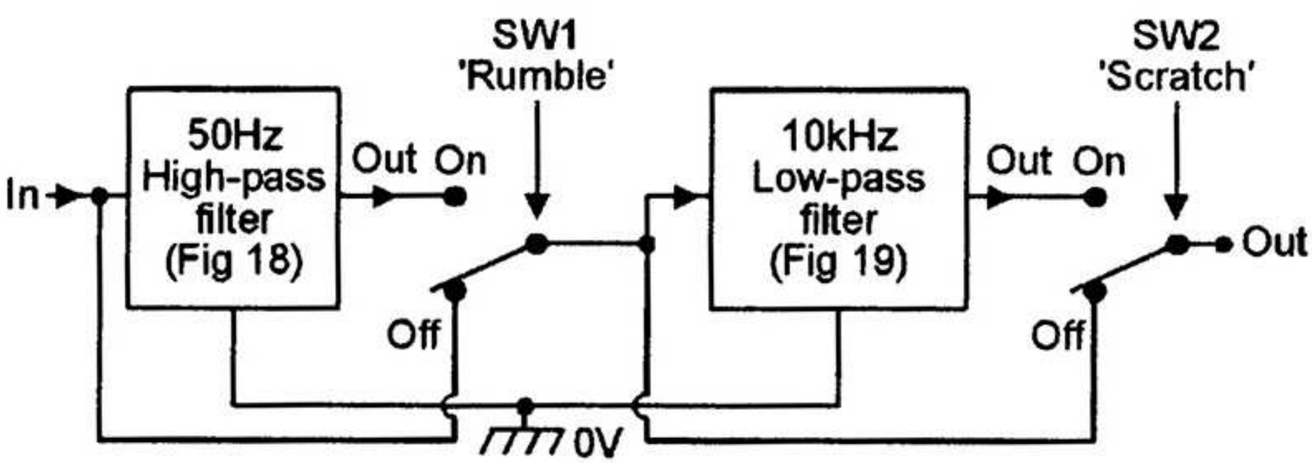


FIGURE 20.

Note that if the **Figure 18** and **19** designs are built as a single unit, a few components can be saved by making the R1-R2-C3 biasing network common to both circuits. **NV**

Bipolar Transistor Cookbook — Part 5

The two most widely used types of transistor waveform generator circuits are the oscillator types that produce sine waves and use transistors as linear amplifying elements, and the multivibrator types that generate square or rectangular waveforms and use transistors as digital switching elements.

This month's installment describes practical ways of using bipolars in the linear mode to make simple, but useful sine wave and white-noise generator circuits. Next month's edition of the series will deal with practical multivibrator types of bipolar waveform generator circuits.

OSCILLATOR BASICS

To generate reasonably pure sine waves, an oscillator has to satisfy two basic design requirements, as shown in **Figure 1**. First, the output of its amplifier (A1) must be fed back to its input via a frequency-selective network (A2) in such a way that the sum of the amplifier and feedback network phase shifts equals zero degrees (or 360°) at the desired oscillation frequency, i.e., so that x° + y° = 0° (or 360°). Thus, if the amplifier generates 180° of phase shift between input and output, an additional 180° of phase shift must be introduced by the frequency-selective network.

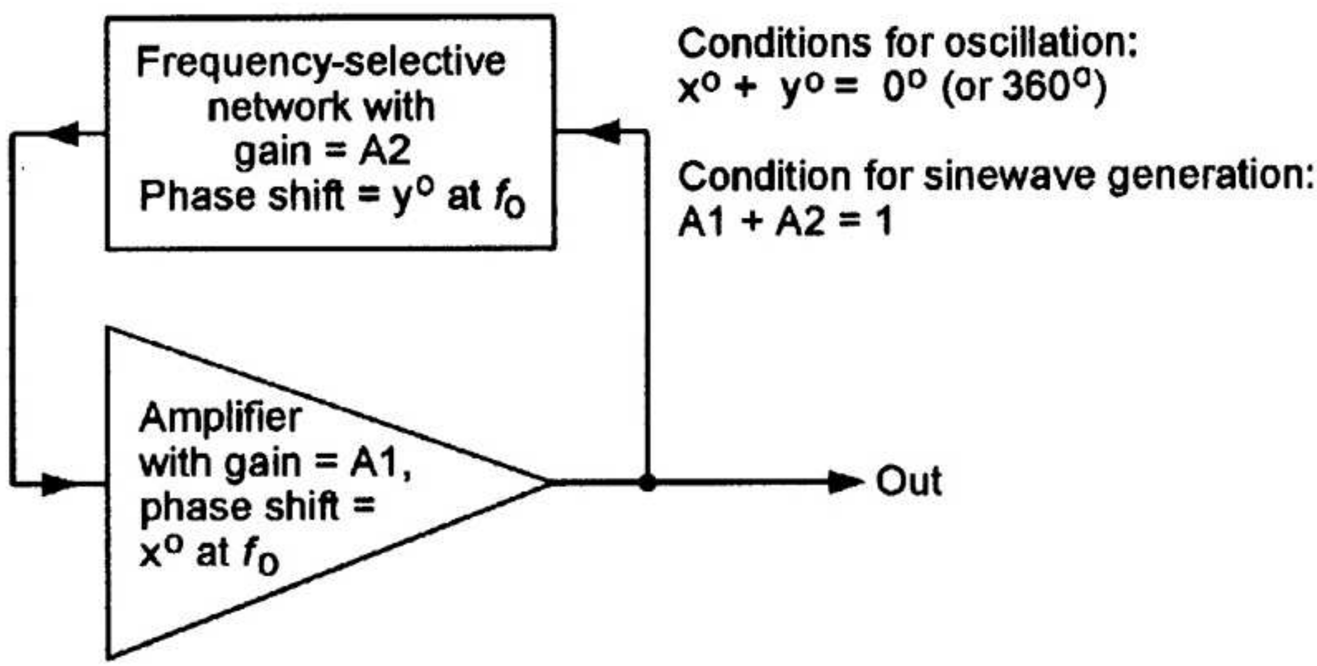


FIGURE 1. Essential circuit and conditions needed for sine wave generation.

The second requirement is that the amplifier's gain must exactly counter the losses of the frequency-selective feedback network at the desired oscillation frequency, to give an overall system gain of unity, e.g., A1 x A2 = 1. If the gain is below unity, the circuit will not oscillate, and if greater than unity, it will be over-driven and will generate distorted waveforms. The frequency-selective feedback network usually consists of either a C-R or L-C or crystal filter; practical oscillator circuits that use C-R frequency-selective filters usually generate output frequencies below 500 kHz; ones that use L-C frequency-selective filters usually generate output frequencies above 500 kHz; ones that use crystal filters generate ultra-precise signal frequencies.

C-R OSCILLATORS

The simplest C-R sine wave oscillator is the phase-shift type, which usually takes the basic form as shown in **Figure 2**. Here, three identical C-R high-pass filters are cascaded to make a third-order filter that is inserted between the output and input of the inverting (180° phase shift) amplifier; the filter gives a total phase shift of 180° at a frequency, fo, of about 1/(14RC), so the complete circuit has a loop shift of 360° under this condition and oscillates at fo if the amplifier has enough gain (about x29) to compensate for the filter's losses and, thus, give a mean loop gain fractionally greater than unity.

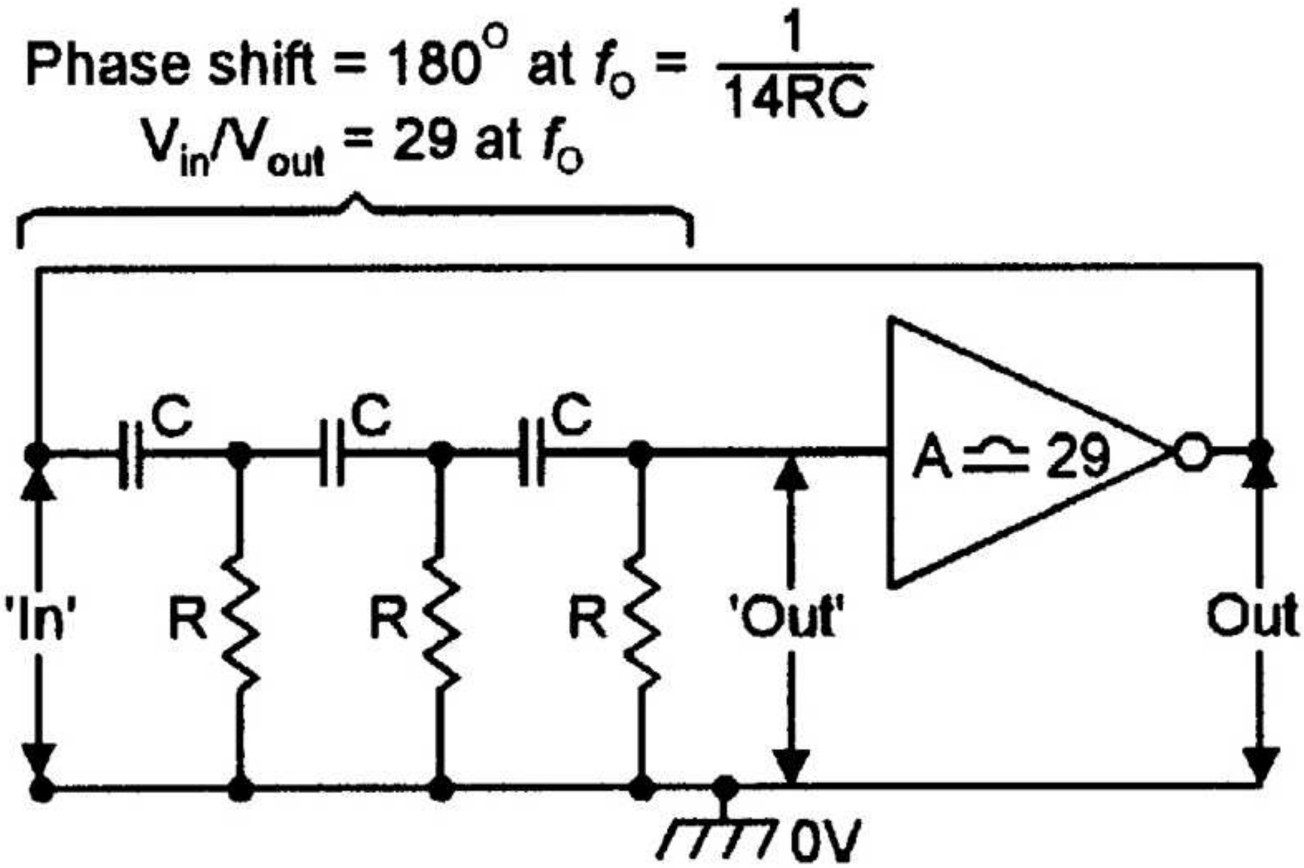


FIGURE 2. Third-order, high-pass filter used as the basis of a phase-shift oscillator.

Note in **Figure 2** that each individual C-R high-pass filter stage tends to pass high-frequency signals, but rejects low-frequency ones. Its output is 3 dB down at a break frequency of 1/(2 RC), and falls at a 6 dB/octave rate as the frequency is decreased below this value. Thus, a basic 1 kHz filter gives 12 dB of rejection to a 250 Hz signal, and 20 dB to a 100 Hz one. The phase angle of the output signal leads that of the input and equals arctan 1/(2fCR), or +45° at fc. Each C-R stage is known as a first-order filter. If a number (n) of such filters are cascaded, the resulting circuit is known as an "nth-order" filter and has a slope, beyond fc of (n x 6 dB)/octave.

Figure 3 shows the circuit of a practical 800 Hz phase-shift oscillator that can operate from any DC supply in the 9V to 18V range. To initially set up the circuit, simply trim RV1 so that the circuit generates a reasonably pure sine wave output as seen on an oscilloscope — the signal's output level is fully variable via RV2.

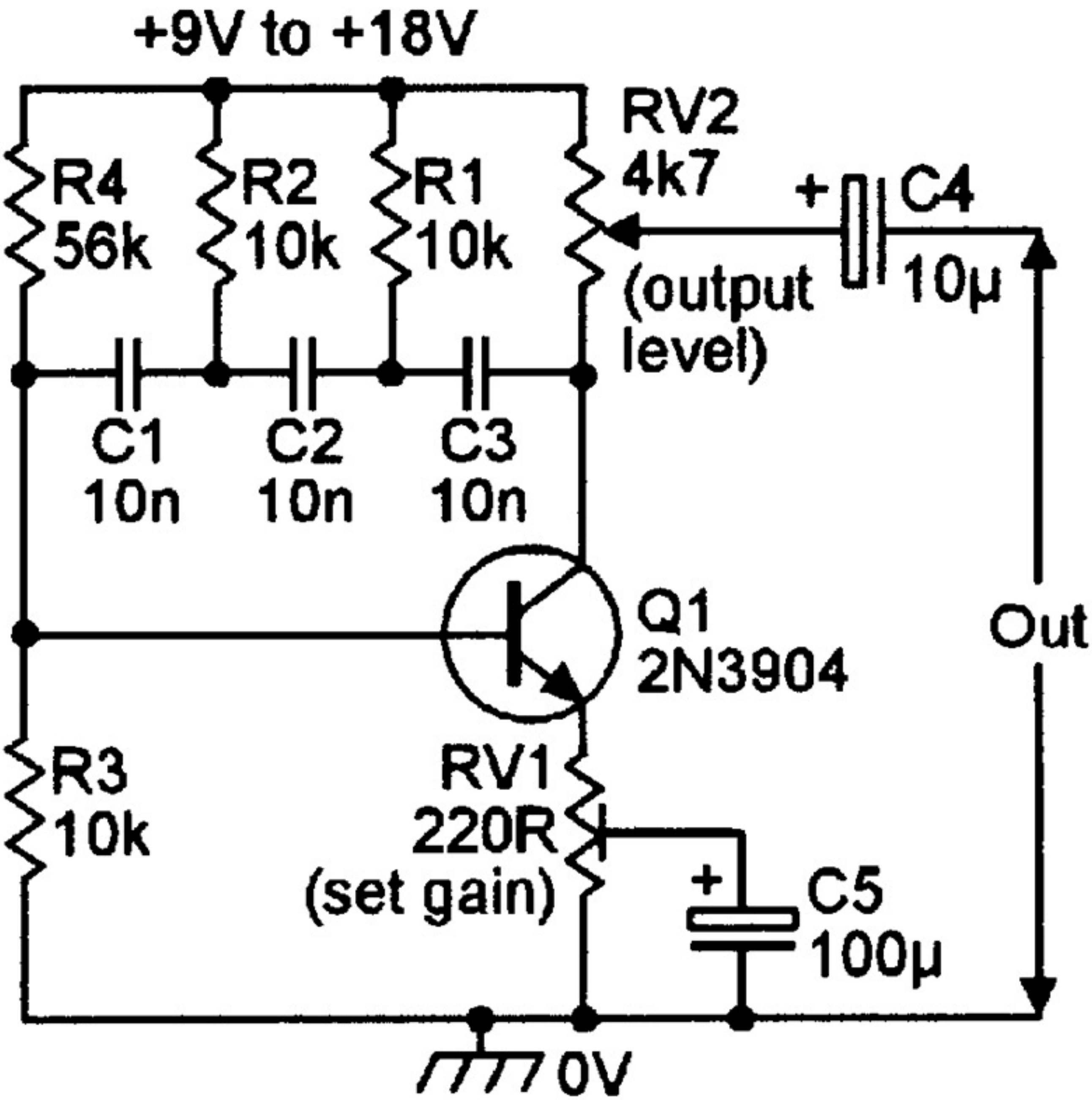


FIGURE 3. 800 Hz phase-shift oscillator.

Major disadvantages of simple phase-shift oscillators of the **Figure 3** type are that they have fairly poor inherent gain stability, and that their operating frequency can not easily be made variable. A far more versatile C-R oscillator can be built using the Wien bridge network.

Figure 4 shows the basic elements of the Wien bridge oscillator. The Wien network consists of R1-C1 and R2-C2, which have their values balanced so that C1=C2=C, and R1=R2=R. This network’s phase shifts are negative at low frequencies, positive at high ones, and zero at a center frequency of 1/(6.28CR), at which the network has an attenuation factor of three. The network can thus be made to oscillate by connecting a non-inverting x3 high input impedance amplifier between its output and input terminals, as shown in the diagram.

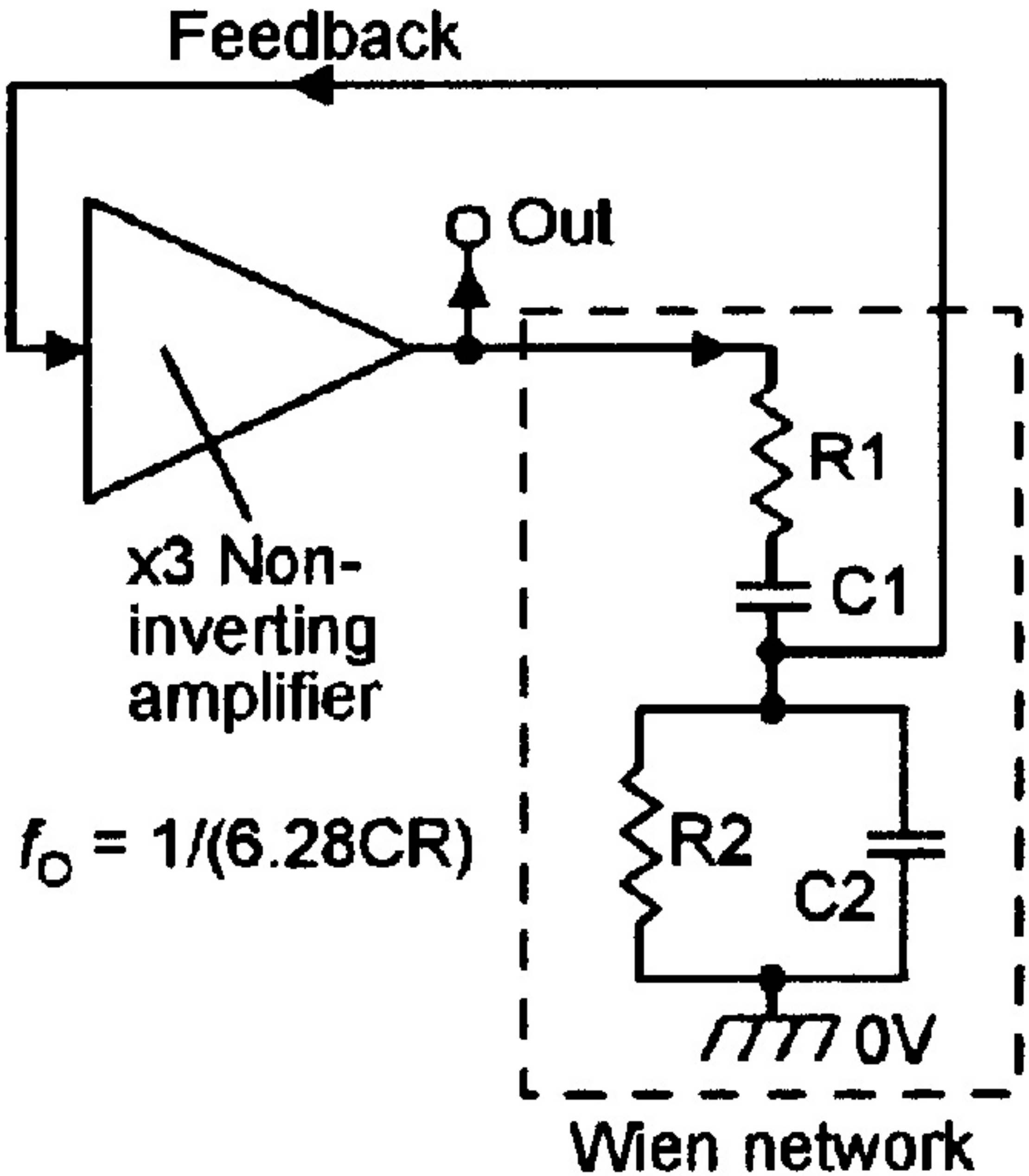


FIGURE 4. Basic Wien oscillator circuit.

Figure 5 shows a simple fixed-frequency Wien oscillator in which Q1 and Q2 are both wired as low-gain common-emitter amplifiers. Q2 gives a voltage gain slightly greater than unity and uses Wien network resistor R1 as its collector load and Q1 presents a high input impedance to the output of the Wien network and has its gain variable via RV1. The component values show that the circuit oscillates at about 1 kHz — in use, RV1 should be adjusted so that a slightly distorted sine wave output is generated.

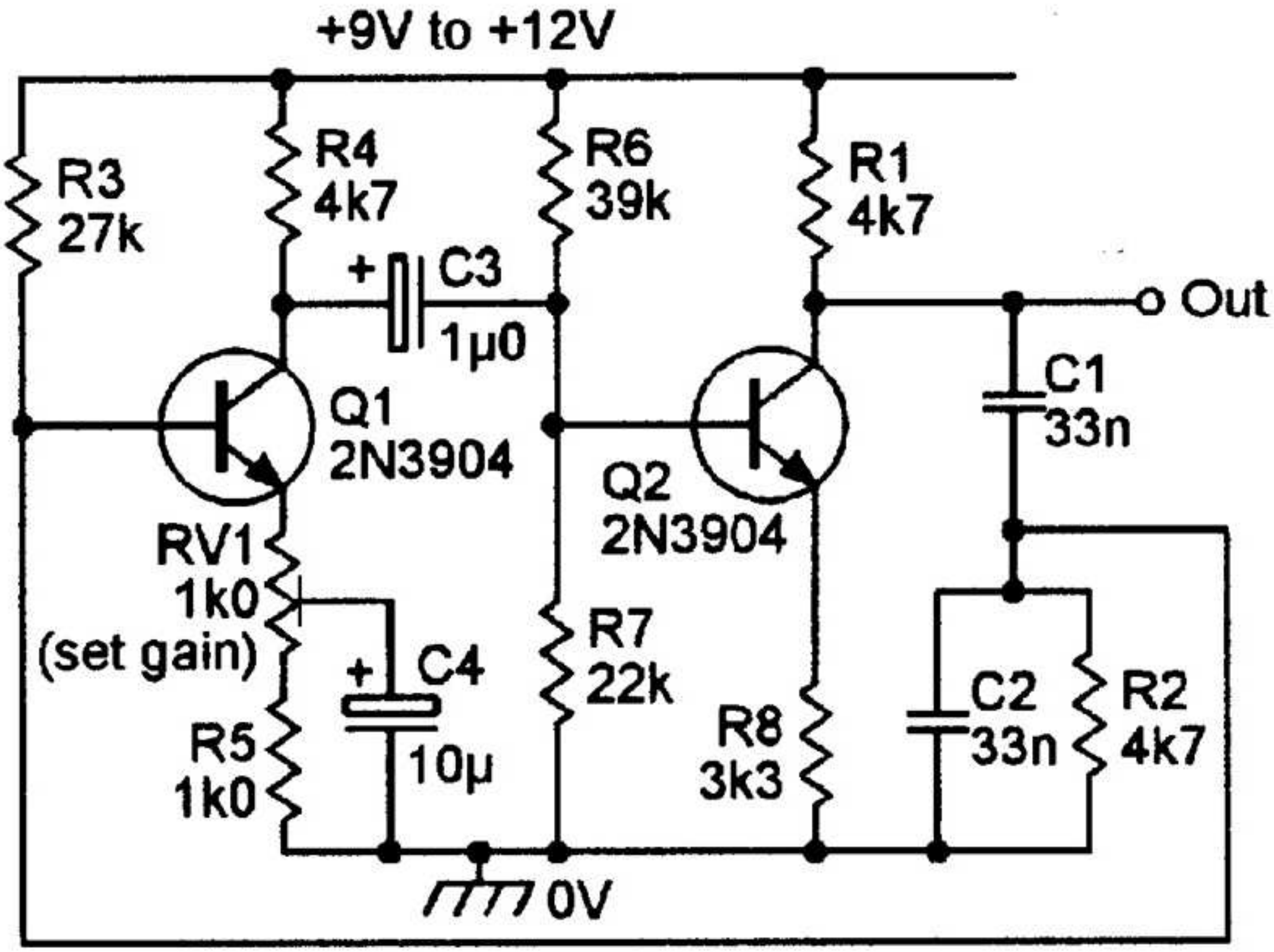


FIGURE 5. Practical 1 kHz Wien oscillator.

Figure 6 shows an improved Wien oscillator design that consumes 1.8 mA from a 9V supply and has an output amplitude that is fully variable up to 6V peak-to-peak via RV2. Q1-Q2 are a direct-coupled complementary common-emitter pair, and give a very high input impedance to Q1 base, a low output impedance from Q2 collector, and non-inverted voltage gains of x5.5 DC and x1 to x5.5 AC (variable via RV1). The red LED generates a low-impedance 1.5V that is fed to Q1 base via R2 and therefore biases Q2's output to a quiescent value of +5V. The R1-C1 and R2-C2 Wien network is connected between Q2's output and Q1's input, and in use RV1 is simply adjusted so that, when the circuit's output is viewed on an oscilloscope, a stable and visually clean waveform is generated. Under this condition, the oscillation amplitude is limited at about 6V peak-to-peak by the onset of positive-peak clipping as the amplifier starts to run into saturation. If RV1 is carefully adjusted, this clipping can be reduced to an almost imperceptible level, enabling good-quality sine waves, with less than 0.5% THD, to be generated.

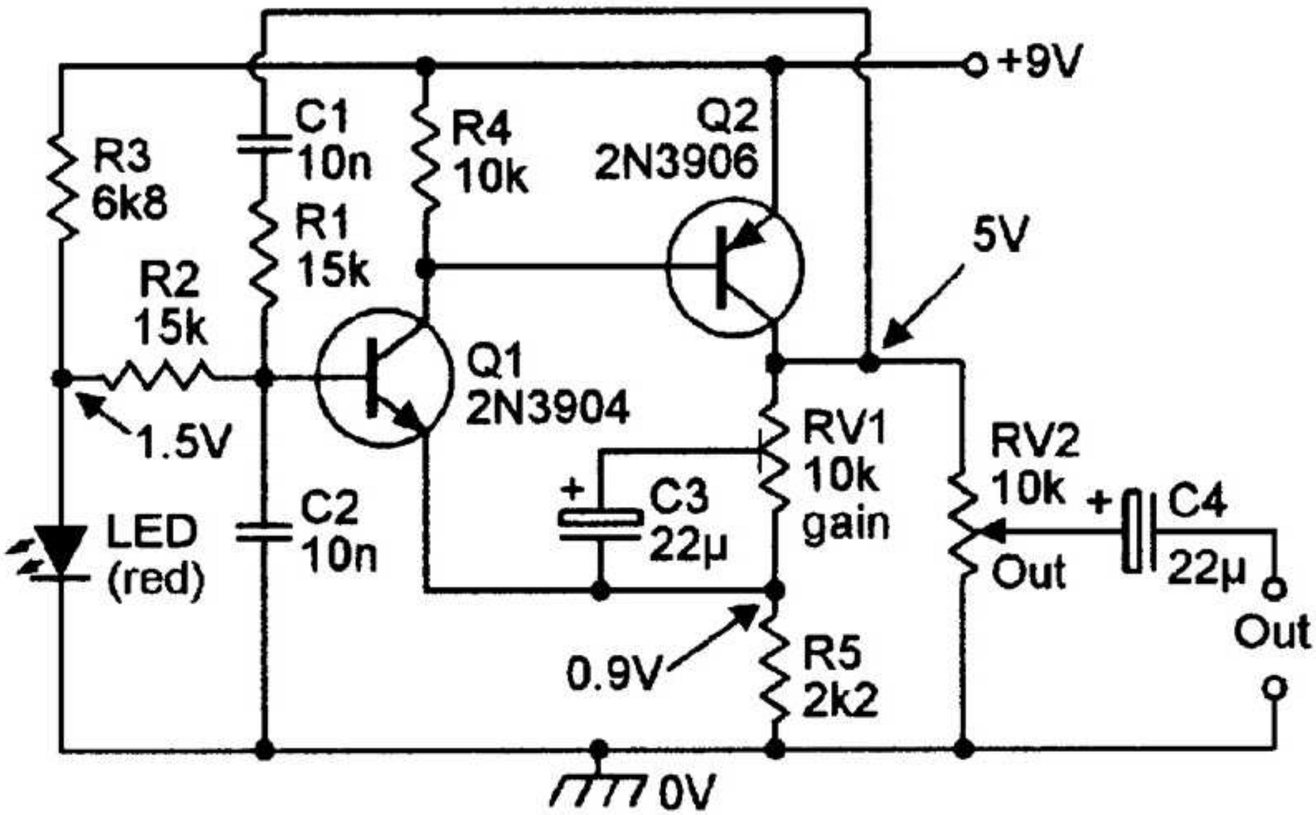


FIGURE 6. 1 kHz Wien bridge sine wave generator with variable-amplitude output.

The **Figure 6** circuit can be modified to give limited-range variable-frequency operation by reducing the R1 and R2 values to 4.7K and wiring them in series with ganged 10K variable resistors. Note, however, that variable-frequency Wien oscillators are best built using op-amps or other linear ICs, in conjunction with automatic-gain-control feedback systems, using various standard circuits of this type that have been published in previous editions of this magazine.

L-C OSCILLATORS

C-R sine wave oscillators usually generate signals in the 5 Hz to 500 kHz range. L-C oscillators usually generate them in the 5 kHz to 500 MHz range, and consist of a frequency-selective L-C network that is connected into an amplifier's feedback loop.

The simplest L-C transistor oscillator is the tuned collector feedback type shown in **Figure 7**. Q1 is wired as a common-emitter amplifier, with base bias provided via R1-R2 and with emitter resistor R3 AC-decoupled via C2. L1-C1 forms the tuned collector circuit, and collector-to-base feedback is provided via L2, which is inductively coupled to L1 and provides a transformer action. By selecting the phase of this feedback signal, the circuit can be made to give zero loop phase shift at the tuned frequency, so that it oscillates if the loop gain (determined by the turns ratio of T1) is greater than unity.

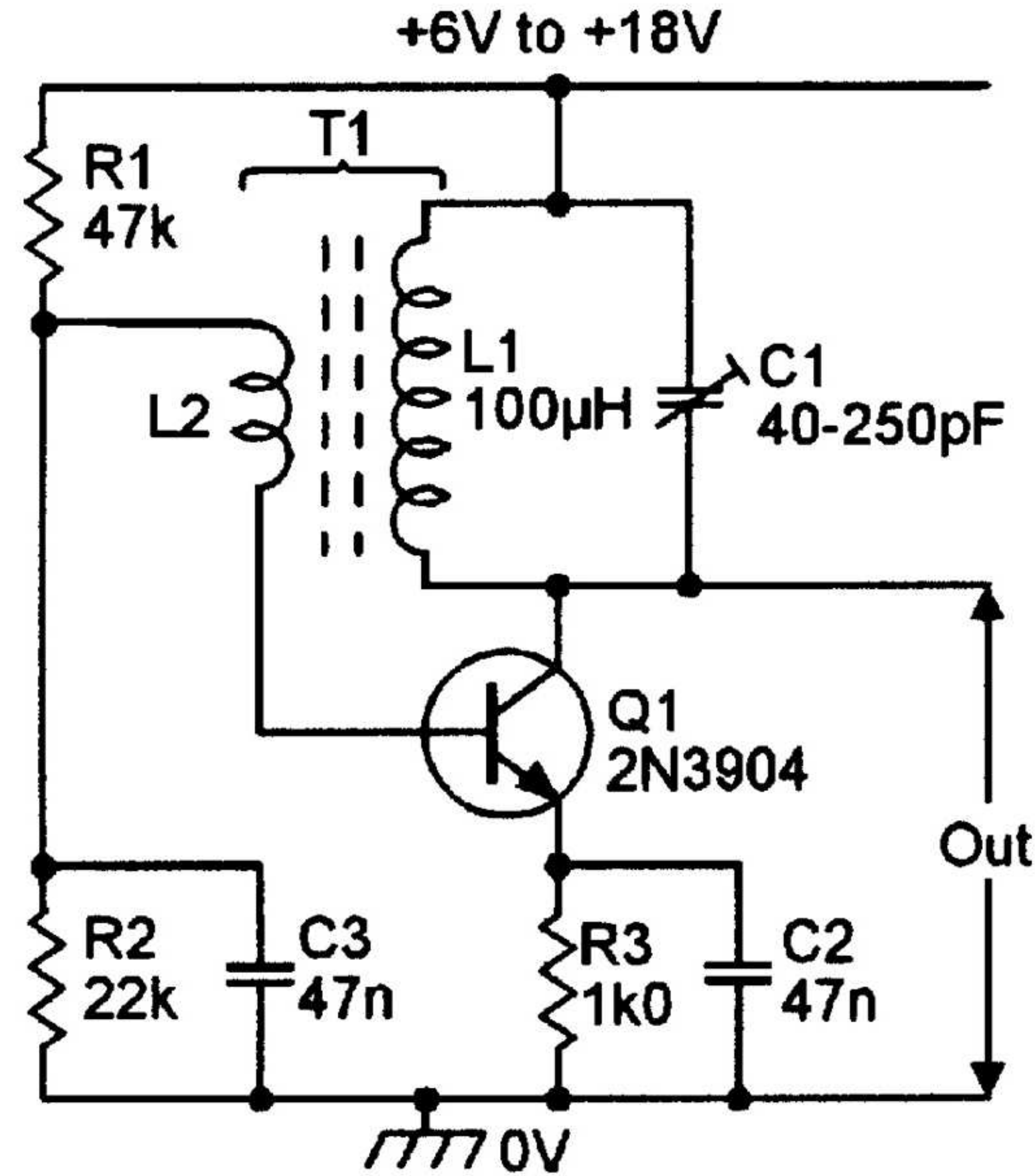


FIGURE 7. Tuned collector feedback oscillator.

A feature of any L-C tuned circuit is that the phase relationship between its energizing current and induced voltage varies from -90° to +90°, and is zero at a center frequency given by f = 1/(2 LC). Thus, the **Figure 7** circuit gives zero overall phase shift, and oscillates at, this center frequency. With the component values shown, the frequency can be varied from 1 MHz to 2 MHz via C1. This basic circuit can be designed to operate at frequencies ranging from a few tens of Hz by using a laminated iron-cored transformer, up to tens or hundreds of MHz using RF techniques.

CIRCUIT VARIATIONS

Figure 8 shows a simple variation of the **Figure 7** design — the Hartley oscillator. Its L1 collector load is tapped about 20% down from its top, and the positive supply rail is connected to this point; L1 thus gives an auto-transformer action, in which the signal voltage at the top of L1 is 180° out of phase with that at its low (Q1 collector) end. The signal from the top of the coil is fed to Q1’s base via C2, and the circuit thus oscillates at a frequency set by the L-C values.

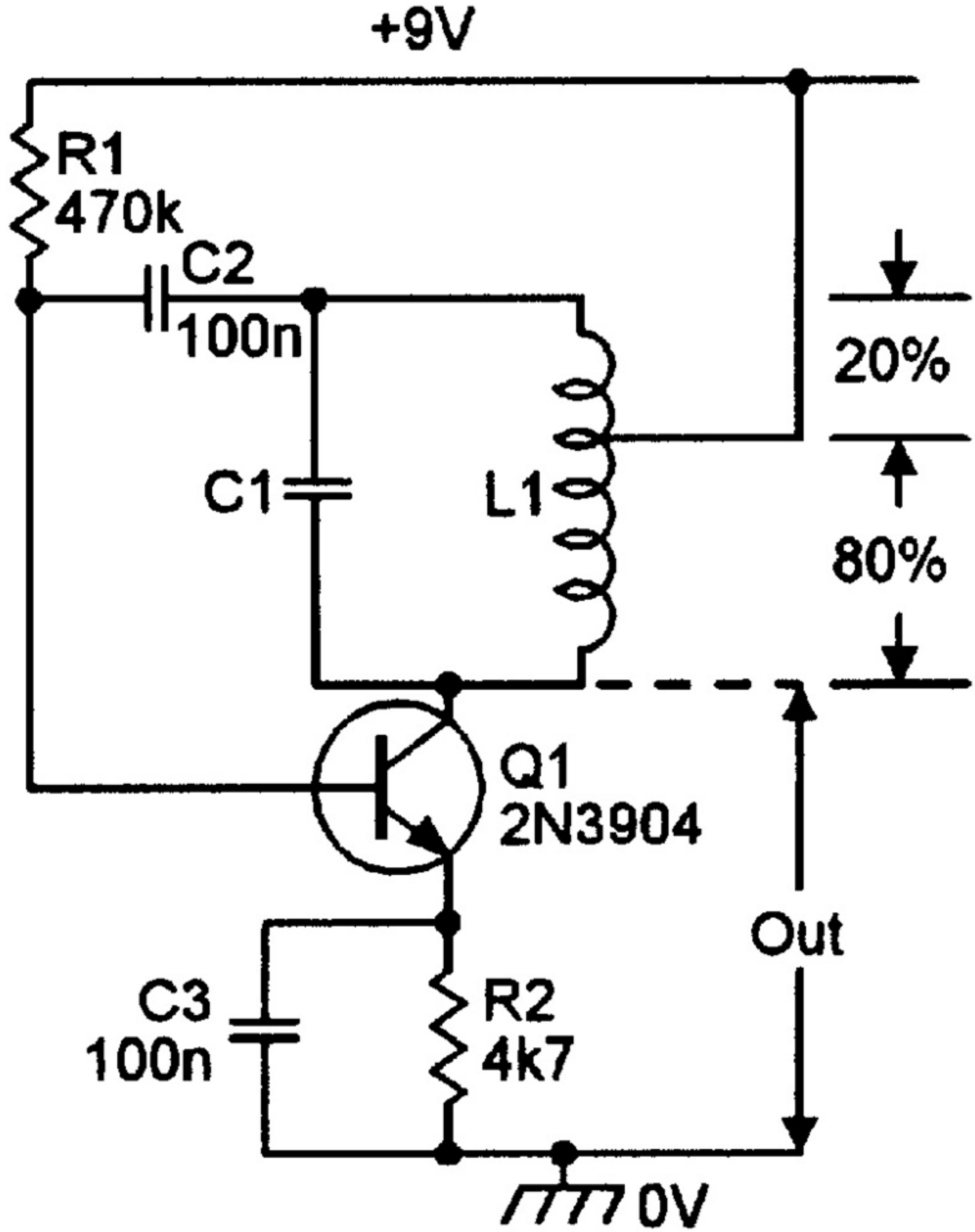


FIGURE 8. Basic Hartley oscillator.

Note from the above description that oscillator action depends on some kind of common signal tapping point being made into the tuned circuit, so that a phase-splitting autotransformer action is obtained. This tapping point does not have to be made into the actual tuning coil, but can be made into the tuning capacitor, as in the Colpitts oscillator circuit shown in **Figure 9**. With the component values shown, this particular circuit oscillates at about 37 kHz.

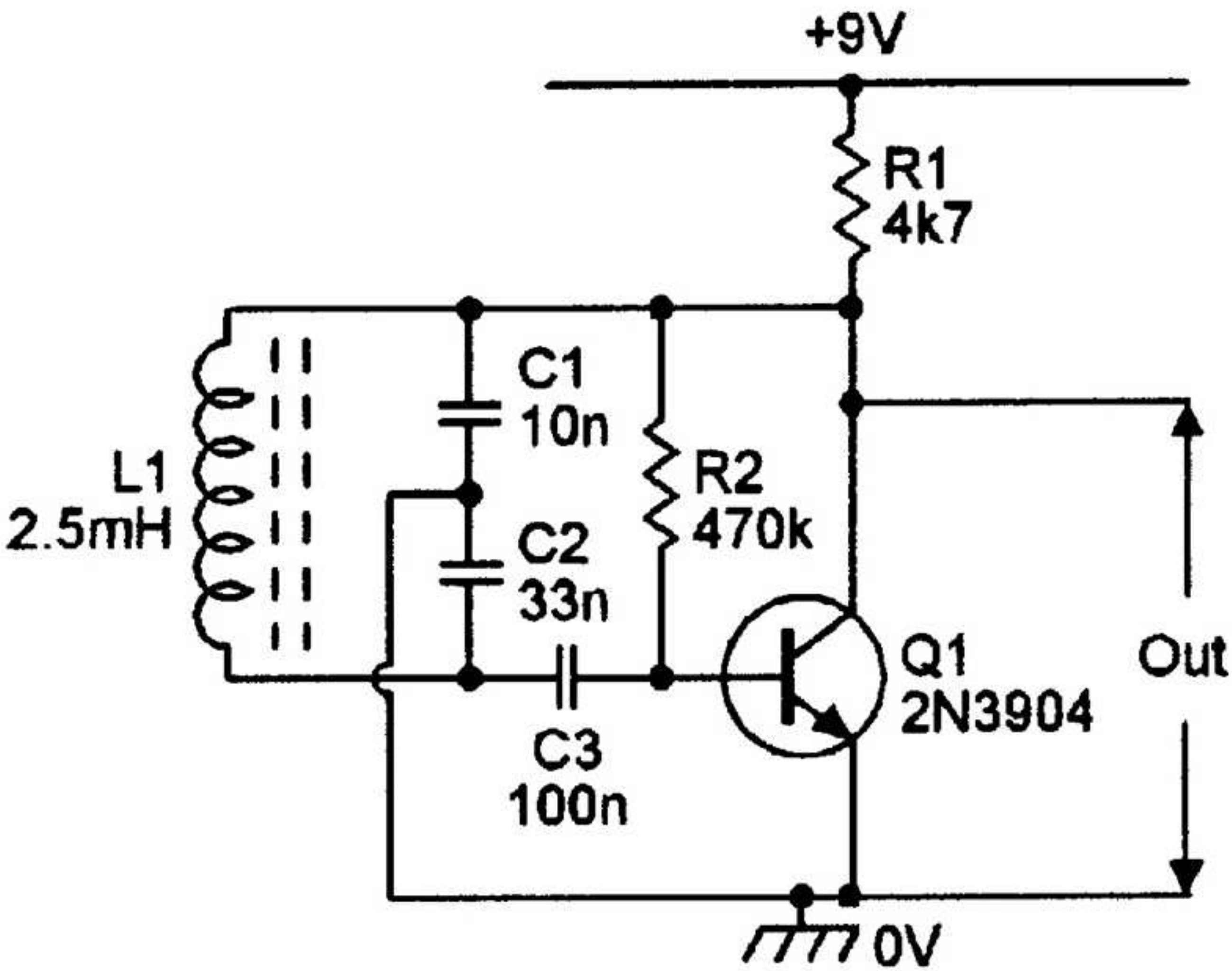


FIGURE 9. 37 kHz Colpitts oscillator.

A modification of the Colpitts design, known as the Clapp or Gouriet oscillator, is shown in **Figure 10**. C3 is wired in series with L1 and has a value that is small relative to C1 and C2. Consequently, the circuit's resonant frequency is set mainly by L1 and C3, and is almost independent of variations in transistor capacitances, etc. The circuit thus gives excellent frequency stability. With the component values shown, it oscillates at about 80 kHz.

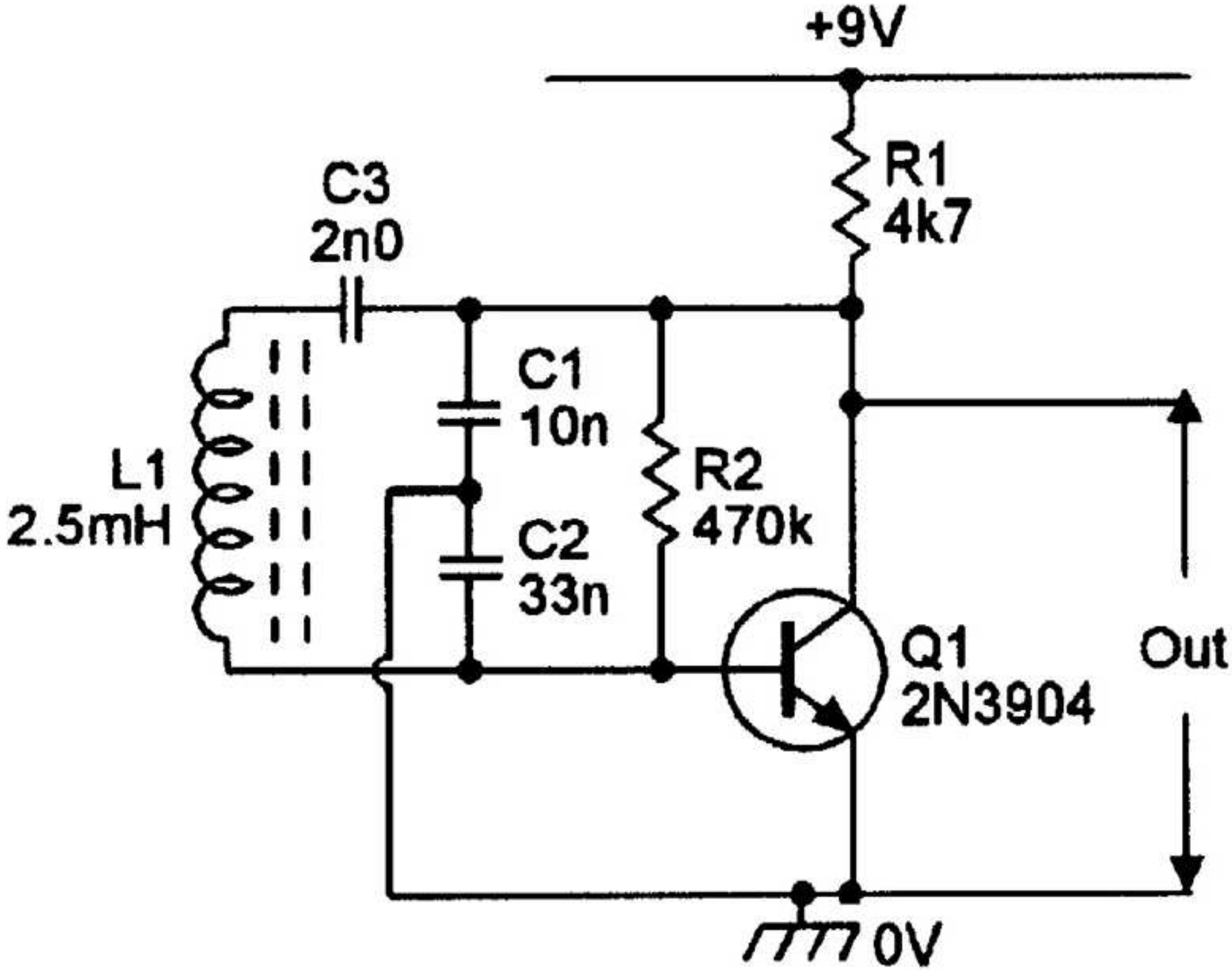


FIGURE 10. 80 kHz Gouriet or Clapp oscillator.

Figure 11 shows a Reinartz oscillator, in which the tuning coil has three inductively-coupled windings. Positive feedback is obtained by coupling the transistor's collector and emitter signals via windings L1 and L2. Both of these inductors are coupled to L3, and the circuit oscillates at a frequency determined by L3-C1. The diagram shows typical coil-turns ratios for a circuit that oscillates at a few hundred kHz.

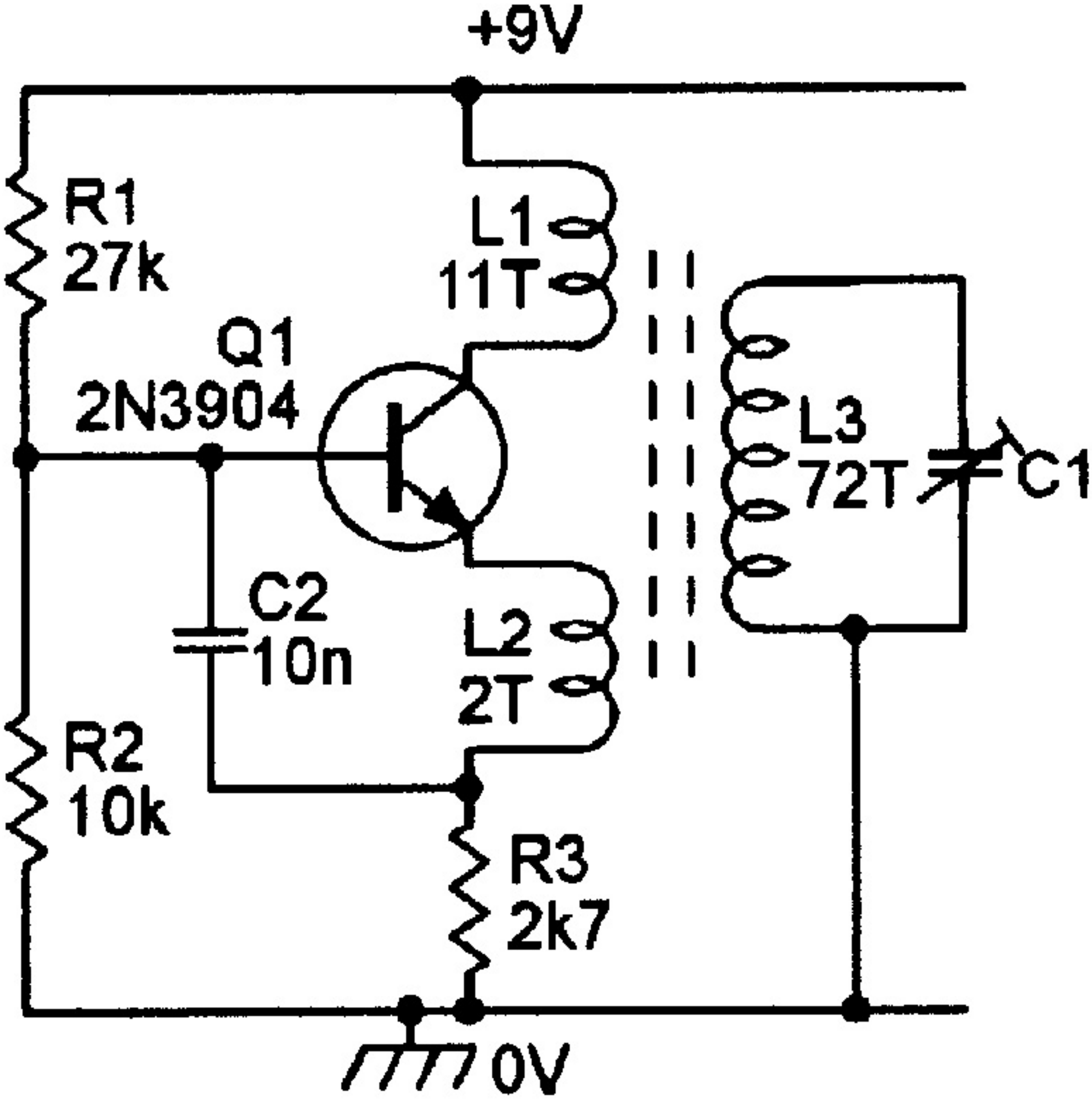


FIGURE 11. Basic Reinartz oscillator.

Finally, **Figures 12** and **13** show emitter follower versions of Hartley and Colpitts oscillators. In these circuits, the transistors and L1-C1 tuned circuits each give zero phase shift at the oscillation frequency, and the tuned circuit gives the voltage gain necessary to ensure oscillation.

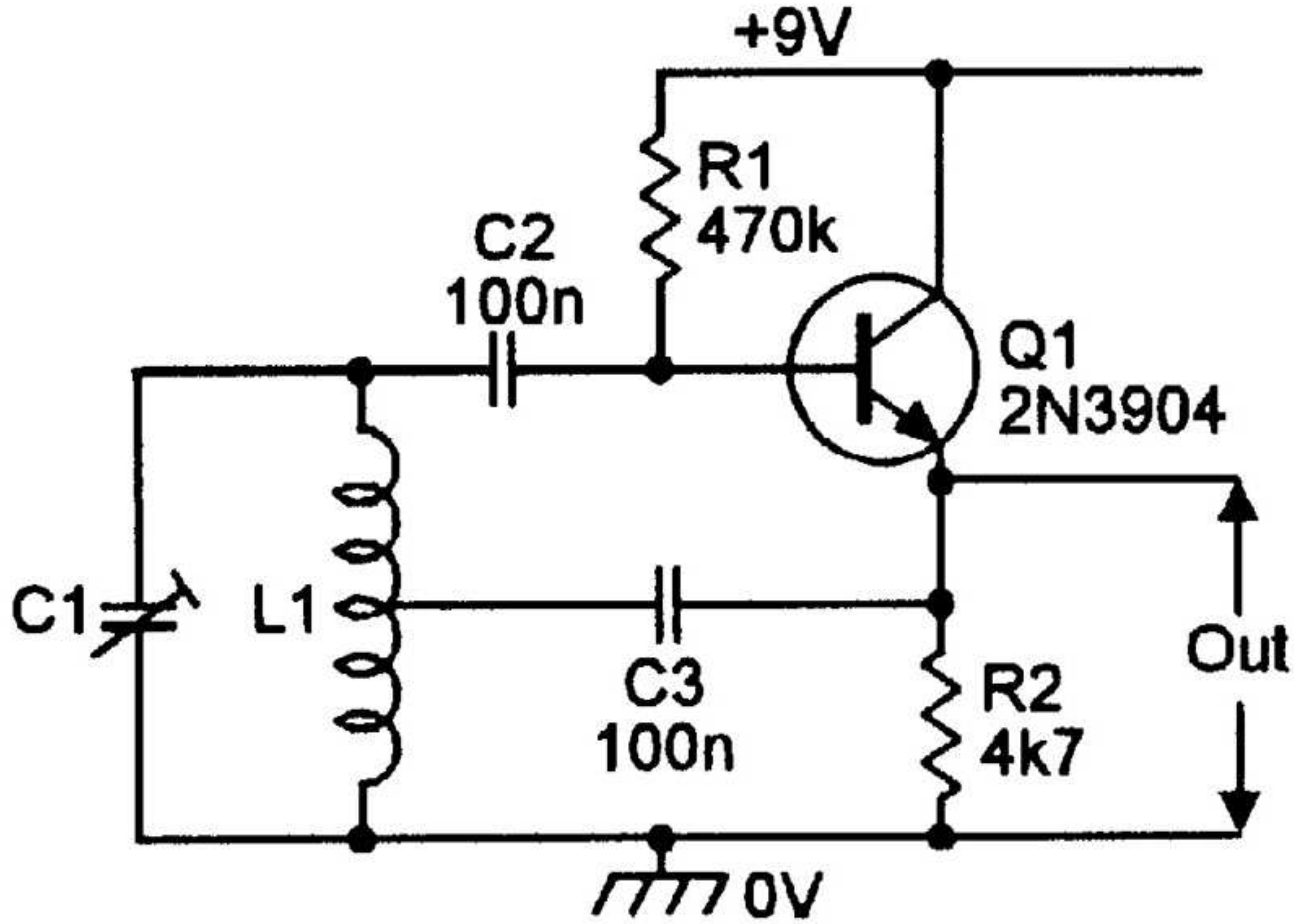


FIGURE 12. Emitter follower version of the Hartley oscillator.

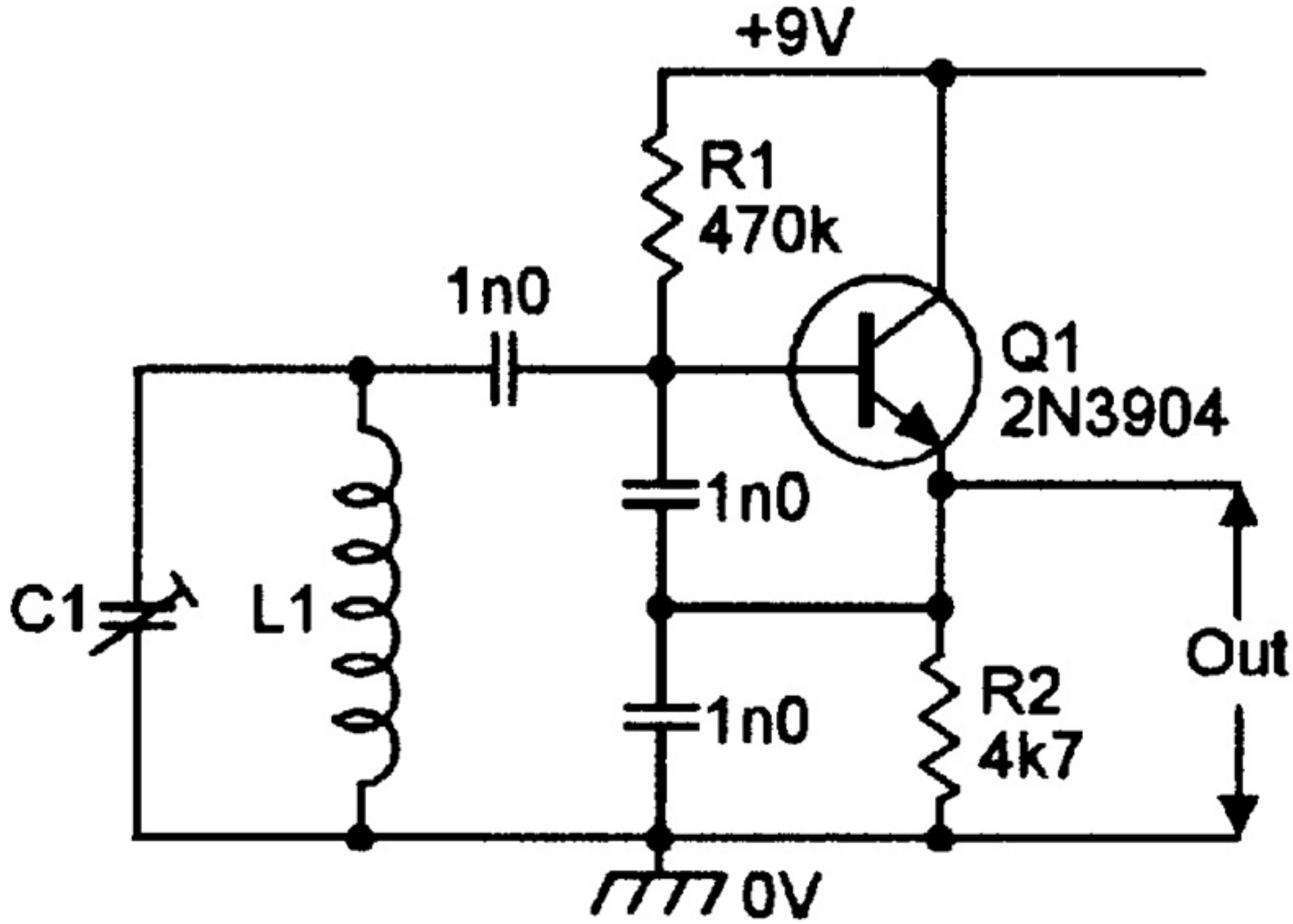


FIGURE 13. Emitter follower version of the Colpitts oscillator.

MODULATION

The L-C oscillator circuits of **Figures 7** to **13** can easily be modified to give modulated (AM or FM) rather than continuous-wave (CW) outputs. **Figure 14**, for example, shows the **Figure 7** circuit modified to act as a 456 kHz beat-frequency oscillator (BFO) with an amplitude-modulation (AM) facility. A standard 465 kHz transistor IF transformer (T1) is used as the L-C tuned circuit, and an external AF signal can be fed to Q1's emitter via C2, thus effectively modulating Q1's supply voltage and thereby amplitude-modulating the 465 kHz carrier signal. The circuit can be used to generate modulation depths up to about 40%. C1 presents a low impedance to the 465 kHz carrier but a high impedance to the AF modulation signal.

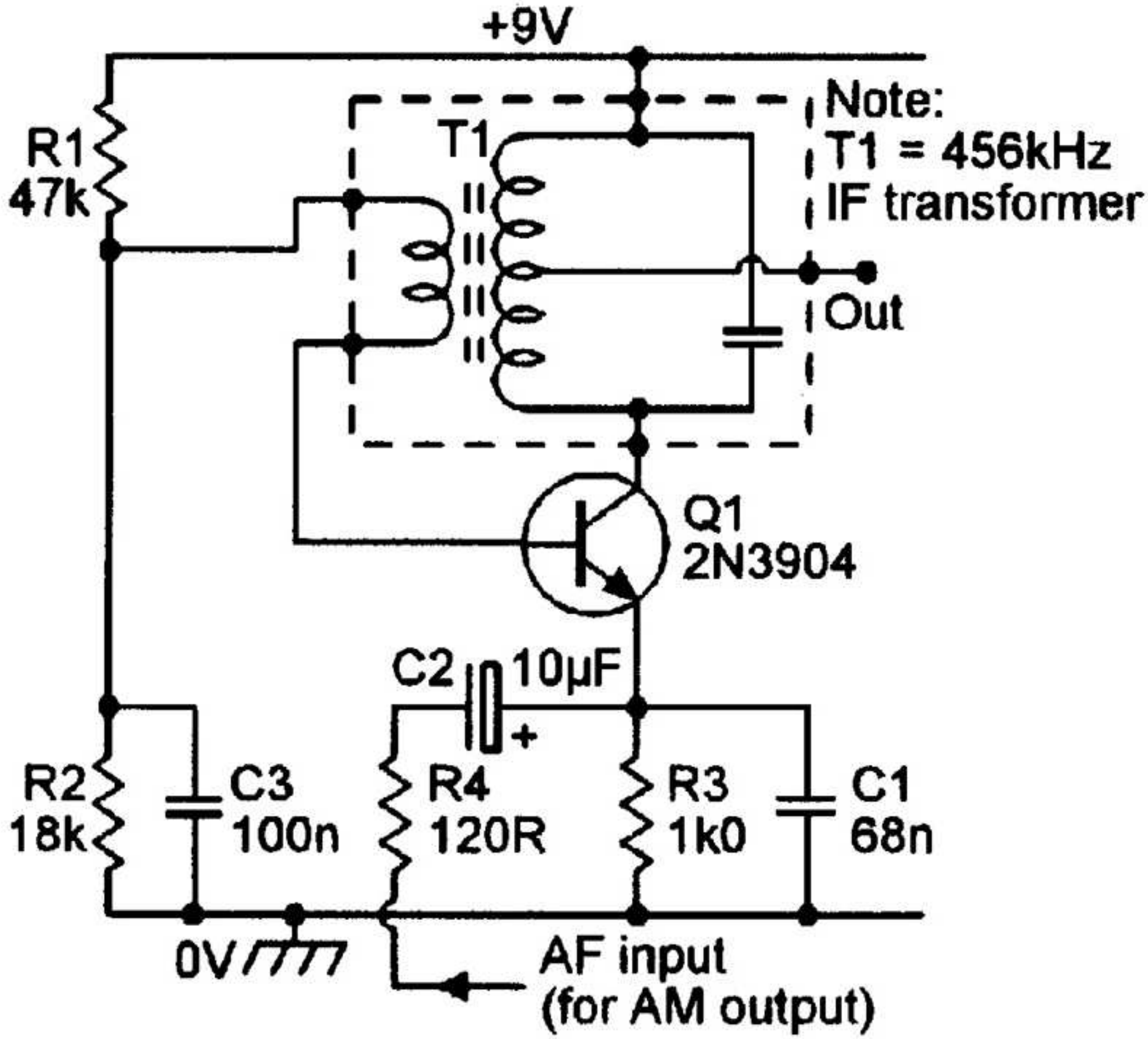


FIGURE 14. 465 kHz BFO with AM facility.

Figure 15 shows the above circuit modified to give a frequency-modulation (FM) facility, together with varactor tuning via RV1. 1N4001 silicon diode D1 is used as an inexpensive varactor diode which, when reverse biased (as an inherent part of its basic silicon diode action) inherently exhibits a capacitance (of a few tens of pF) that decreases with applied reverse voltage. D1 and blocking capacitor C2 are wired in series and effectively connected across the T1 tuned circuit (since the circuit's supply rails are shorted together as far as AC signals are concerned).

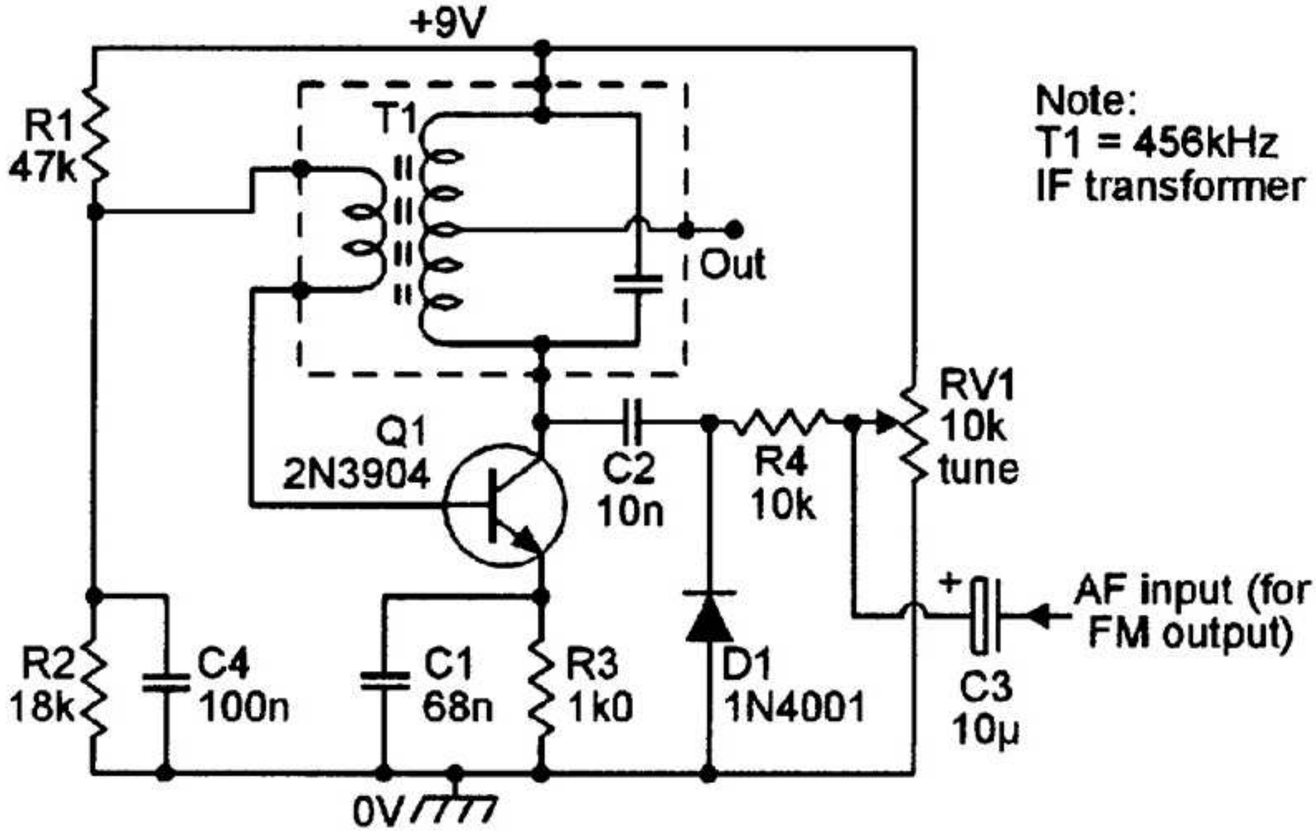


FIGURE 15. 465 kHz BFO with varactor tuning and FM facility.

Consequently, the oscillator's center frequency can be varied by altering D1's capacitance via RV1, and FM signals can be obtained by feeding an AF modulation signal to D1 via C3 and R4.

CRYSTAL OSCILLATORS

Crystal-controlled oscillators give excellent frequency accuracy and stability. Quartz crystals have typical Q values of about 100,000 and provide about 1,000 times greater stability than a conventional L-C tuned circuit. Their operating frequency (which may vary from a few kHz to 100 MHz) is determined by the mechanical dimensions of the crystal, which may be cut to give either series or parallel resonant operation. Series-mode devices show a low impedance at resonance — parallel-mode types show a high impedance at resonance.

Figure 16 shows a wide-range crystal oscillator designed for use with a parallel-mode crystal. This is actually a Pierce oscillator circuit, and can be used with virtually any good 100 kHz to 5 MHz parallel-mode crystal without need for circuit modification.

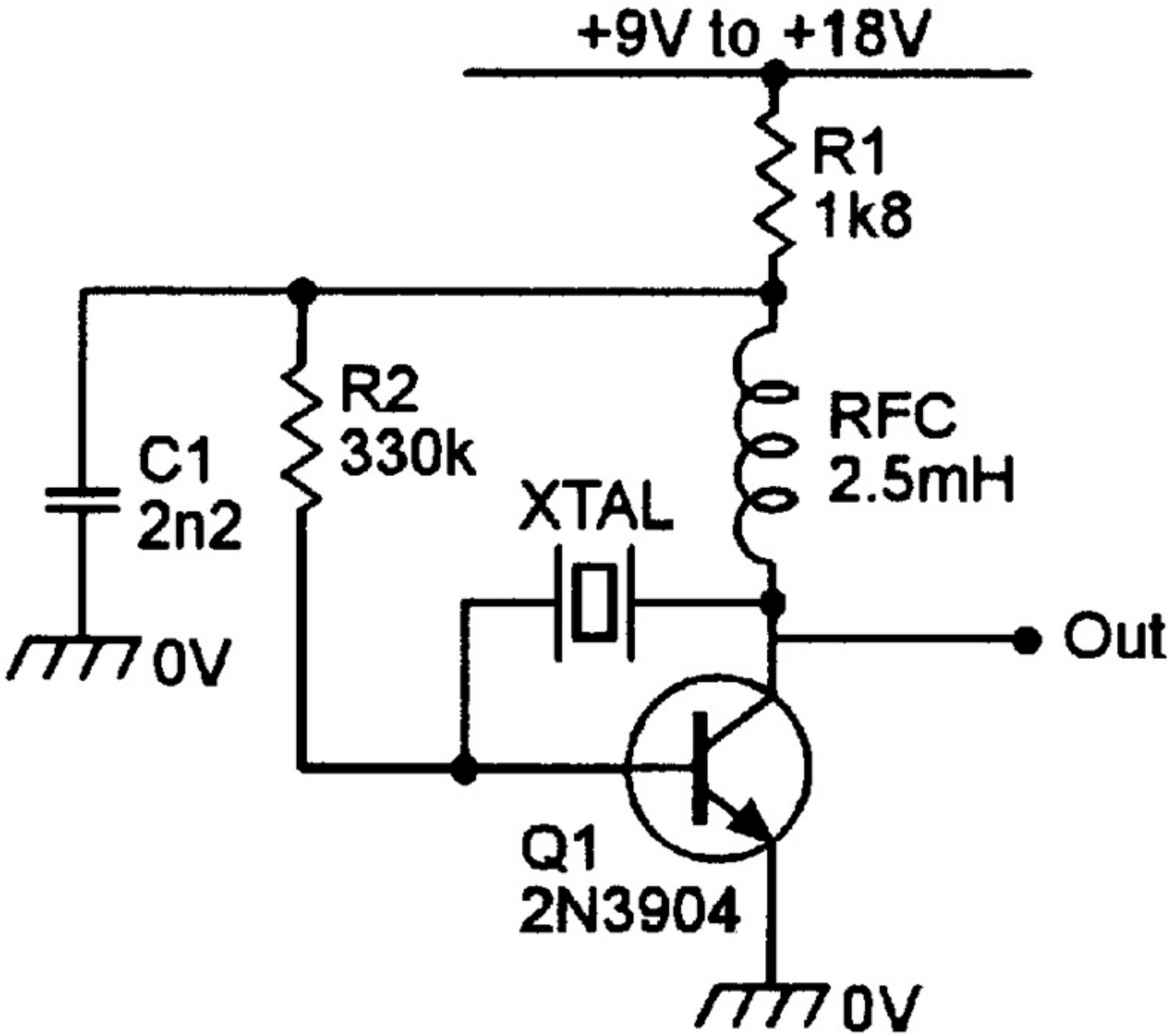


FIGURE 16. Wide-range Pierce oscillator uses parallel-mode crystal.

Alternatively, **Figure 17** shows a 100 kHz Colpitts oscillator designed for use with a series-mode crystal. Note that the L1-C1-C2 tuned circuit is designed to resonate at the same frequency as the crystal, and that its component values must be changed if other crystal frequencies are used.

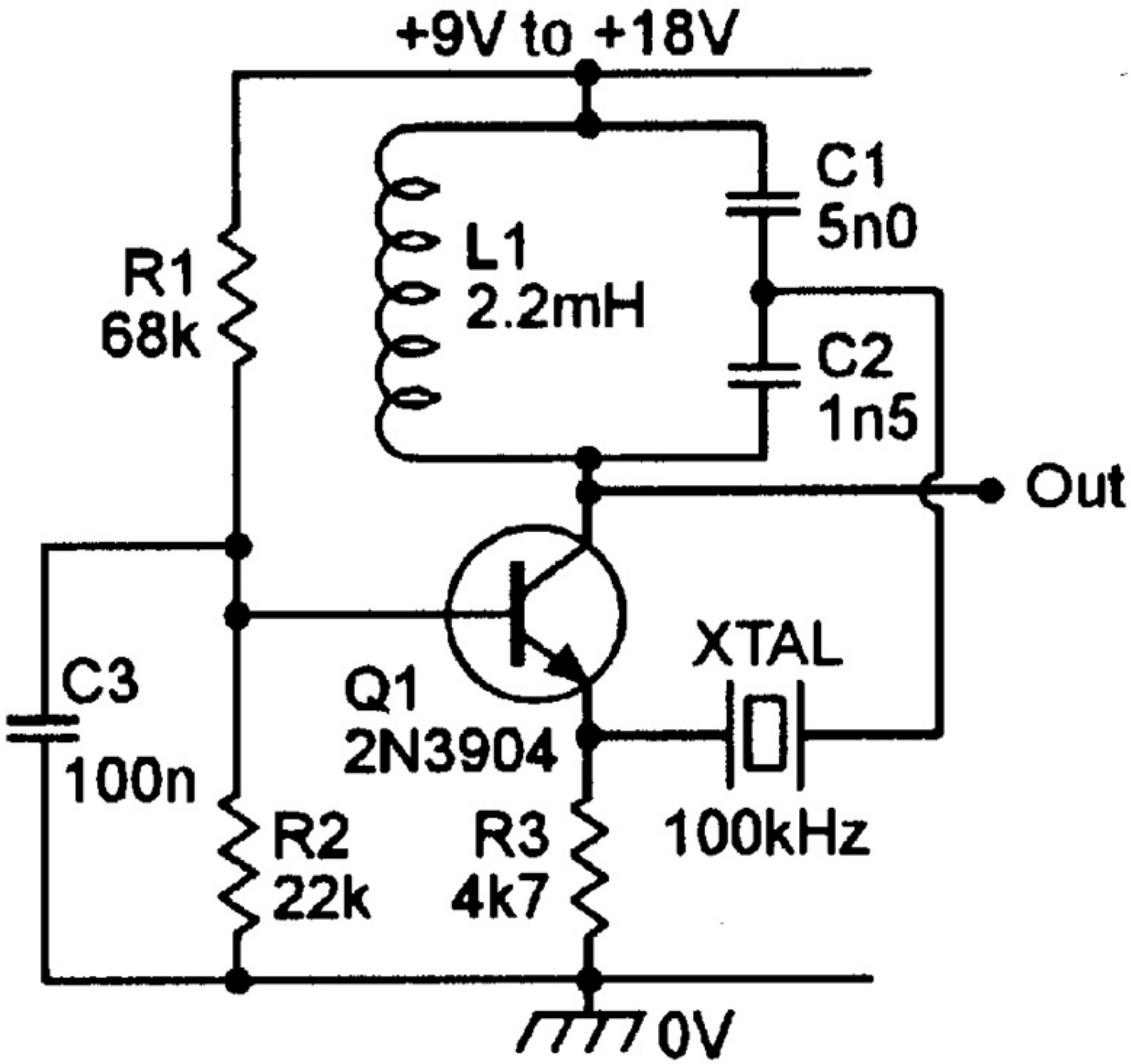


FIGURE 17. 100 kHz Colpitts oscillator uses series-mode crystal.

Finally, **Figure 18** shows an exceptionally useful two-transistor oscillator that can be used with any 50 kHz to 10 MHz series-resonant crystal. Q1 is wired as a common-base amplifier and Q2 as an emitter follower, and the output signal (from Q2 emitter) is fed back to the input (Q1 emitter) via C2 and the series-resonant crystal. This excellent circuit will oscillate with any crystal that shows the slightest sign of life.

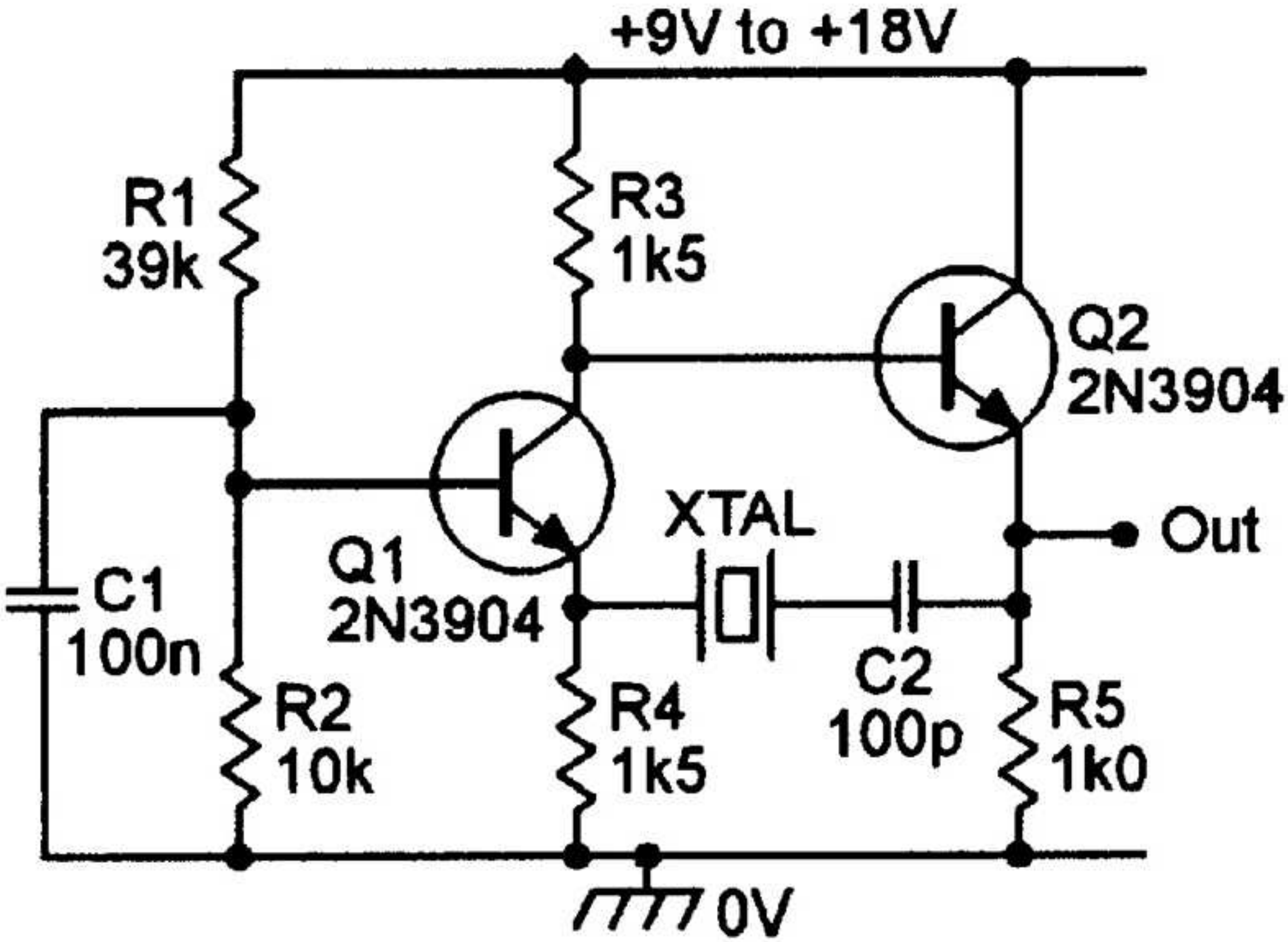


FIGURE 18. Wide-range (50 kHz-10 MHz) oscillator can be used with almost any series-mode crystal.

WHITE NOISE GENERATORS

One useful linear but non-sinusoidal waveform is that known as white noise, which contains a full spectrum of randomly generated frequencies, each with equal mean power when averaged over a unit of time. White noise is of value in testing AF and RF amplifiers, and is widely used in special-effects sound generator systems.

Figure 19 shows a simple white noise generator that relies on the fact that all zener diodes generate substantial white noise when operated at a low current. R2 and ZD1 are wired in a negative-feedback loop between the collector and base of common-emitter amplifier Q1, thus stabilizing the circuit's DC working levels, and the loop is AC-decoupled via C1. ZD1 thus acts as a white noise source that is wired in series with the base of Q1, which amplifies the noise to a useful level of about 1.0 volts, peak-to-peak. Any 5.6V to 12V zener diode can be used in this circuit.

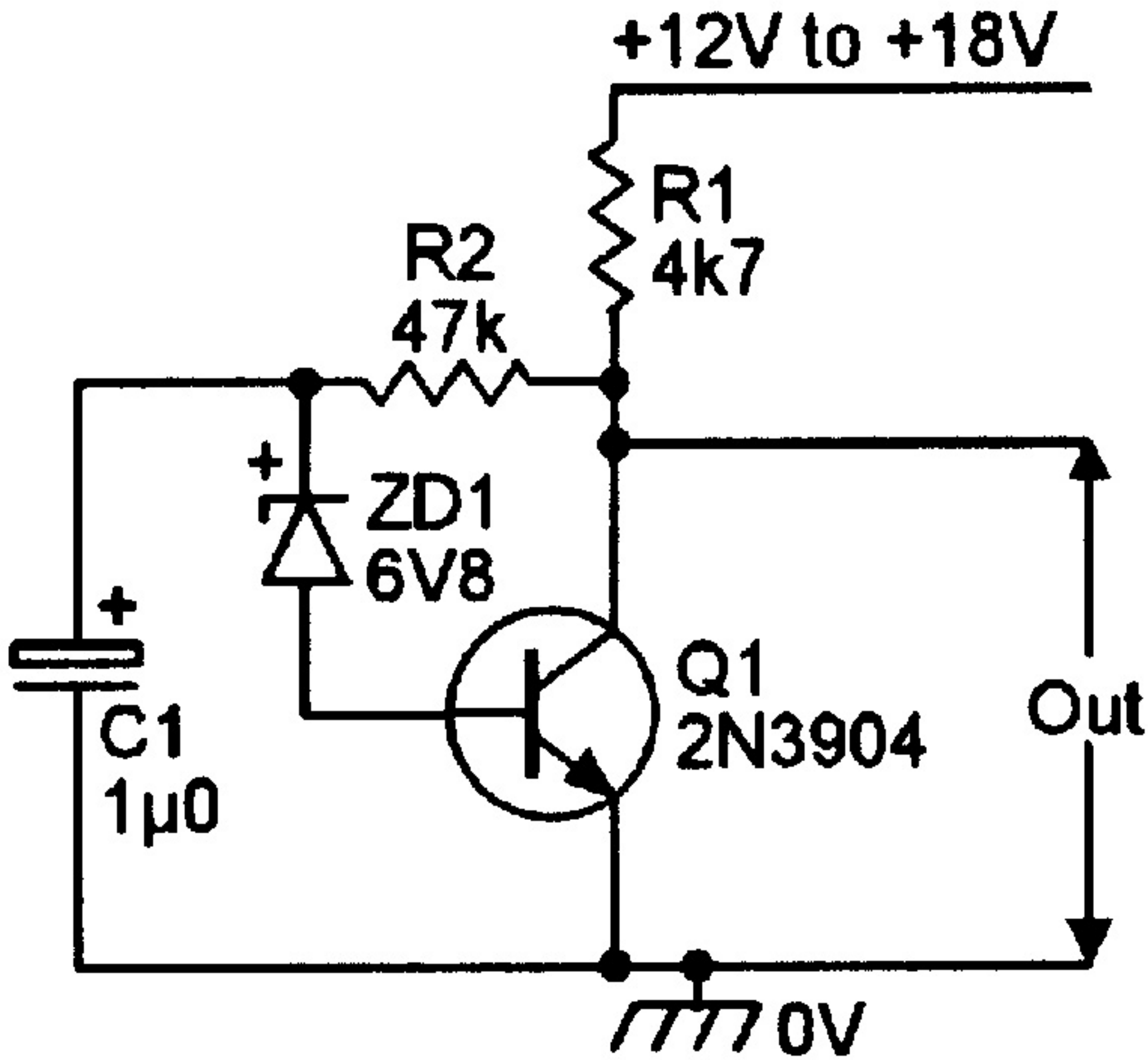


FIGURE 19. Transistor-Zener white noise generator.

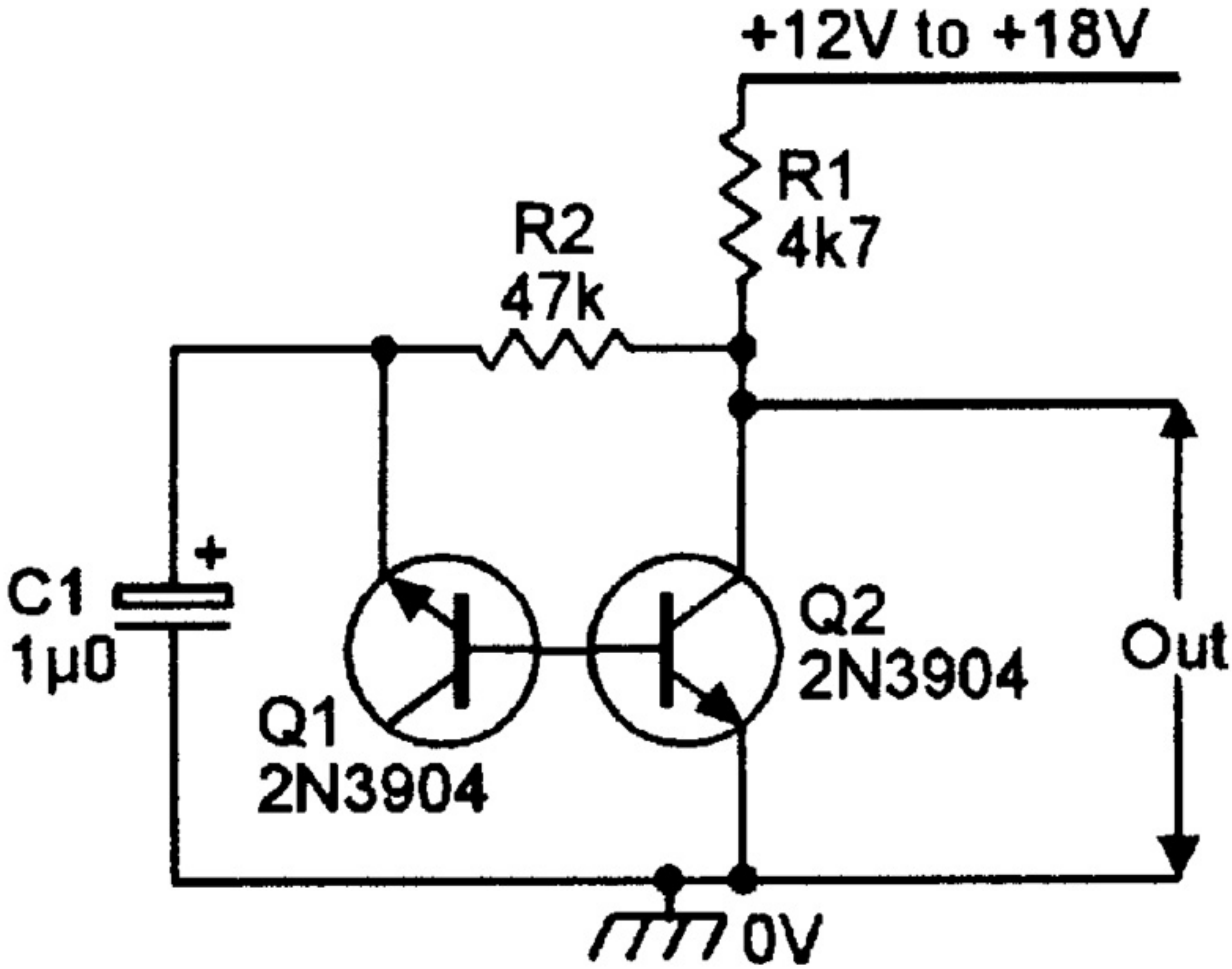


FIGURE 20. Two-transistor white noise generator.

Figure 20 is a simple variation of the above design, with the reverse-biased base-emitter junction of a 2N3904 transistor (which “zeners” at about 6V) used as the noise-generating zener diode. **NV**

Bipolar Transistor Cookbook — Part 6

The two most widely used types of transistor waveform generator circuits are the oscillator types that produce sine waves and use transistors as linear amplifying elements, and the multivibrator types that generate square or rectangular waveforms and use transistors as digital switching elements.

Our last installment covered practical circuits of the oscillator type. This time, we describe ways of using bipolars to make practical multivibrator types of waveform generator circuits.

MULTIVIBRATOR CIRCUIT TYPES

Multivibrators are two-state (output high or output low) circuits that can be switched between one state and the other via a suitable trigger signal, which may be generated either internally or externally. There are four basic types of multivibrator (multi) circuits, and they are all useful in waveform generating applications. Of these four, the astable has two quasi-stable states and is useful as a free-running square wave generator. The monostable has one stable and one quasi-stable state and is useful as a triggered pulse generator. The bistable has two stable states and is useful as a triggered stop/go or high/low waveform generator. Lastly, the Schmitt has two stable input-voltage-sensitive states and is useful as a sine-to-square waveform converter or threshold switch.

ASTABLE MULTIVIBRATOR BASICS

Figure 1 shows the circuit and generated waveforms of a simple 1 kHz astable multivibrator, in which the two transistors are cross-coupled (from collector to base) via timer networks C1-R1 and C2-R2. The basic circuit action is such that, at the moment that power is initially switched to the circuit, inevitable differences in the precise characteristics of Q1 and Q2 make one transistor turn on slightly faster than the other, and the cross-coupling then causes a regenerative switching action to take place in which one transistor switches abruptly on and the other switches abruptly off.

FIGURE 1. Circuit and waveforms of a basic 1 kHz astable multivibrator.

After a delay determined by the C1-R1 or C2-R2 time constant, the off transistor starts to turn on again, and the cross-coupling then causes another regenerative action in which the two transistors abruptly change state again. The whole process then repeats add infinitum. Thus, the basic **Figure 1** circuit acts as a self-oscillating regenerative switch in which the on and off periods are controlled by the C1-R1 and C2-R2 time constants. If these time constants are equal (C1=C2=C, and R1=R2=R), the circuit acts as a square wave generator and operates at a frequency of about 1/(1.4CR). The frequency can be decreased by raising the C or R values, or increased by reducing the C or R values, or can be made variable by using twin-gang variable resistors (in series with 10K limiting resistors) in place of R1 and R2.

Outputs can be taken from either collector, and the two outputs are in anti-phase. The **Figure 1** circuit’s operating frequency is almost independent of supply-rail values in the range 1.5 V to 9.0 V; the upper voltage limit is set by the fact that, as the transistors change state at the end of each half-cycle, the base-emitter junction of the off one is reverse biased by an amount almost equal to the supply voltage and will zener (and upset the timing action) if this voltage exceeds the junction’s reverse breakdown voltage value (which is typically about 10 V).

This problem can be overcome by wiring a silicon diode in series with the input of each transistor, to raise its effective zener value to that of the diode, as shown in **Figure 2**.

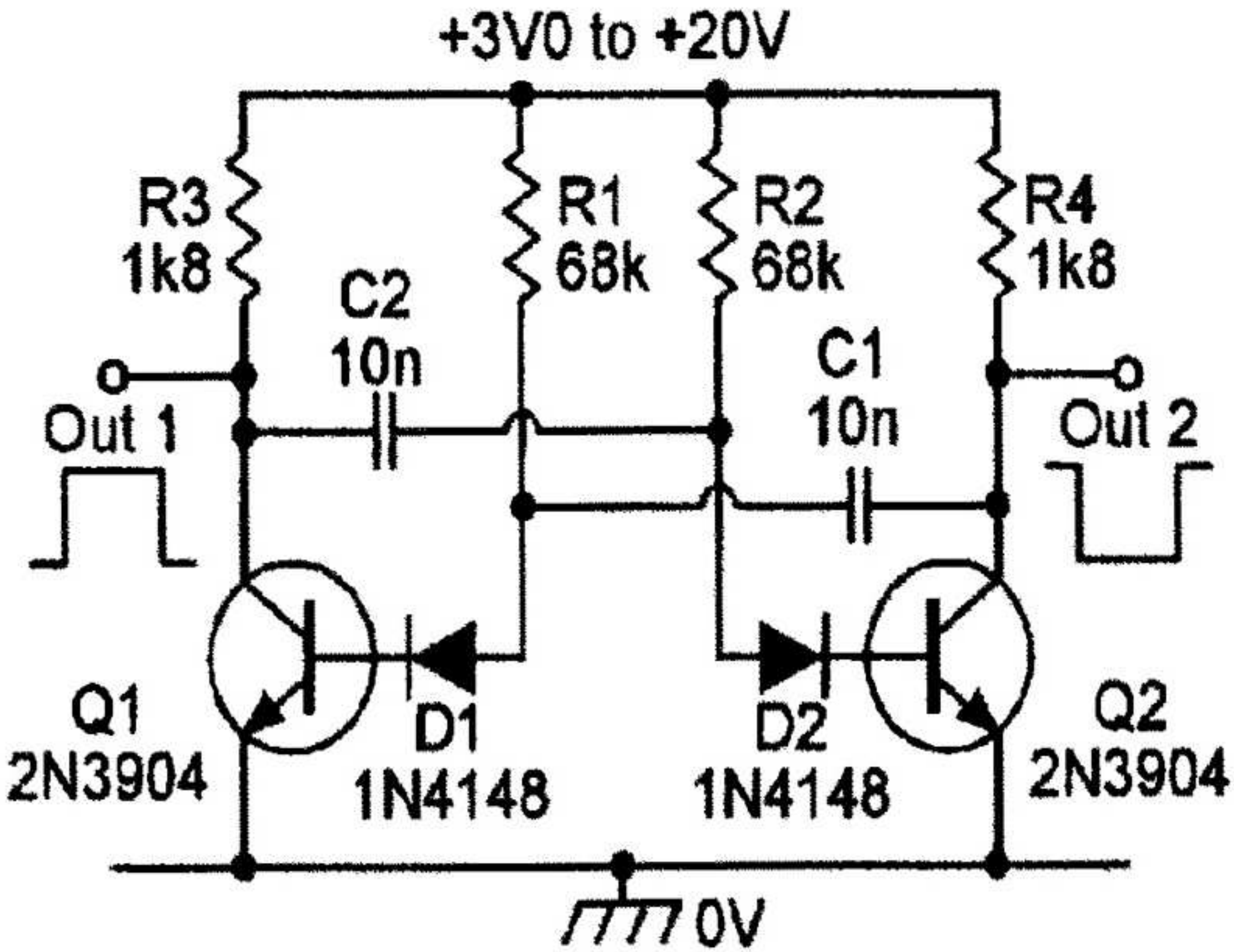


FIGURE 2. Wide-supply-voltage example of a 1 kHz astable multivibrator.

This protected circuit can be used with any supply in the range 3 V to 20 V, and gives a frequency variation of only 2% when the supply is varied from 6 V to 18 V. This variation can be reduced to a mere 0.5% by wiring an additional compensation diode in series with the collector of each transistor, as shown in the circuit of **Figure 3**.

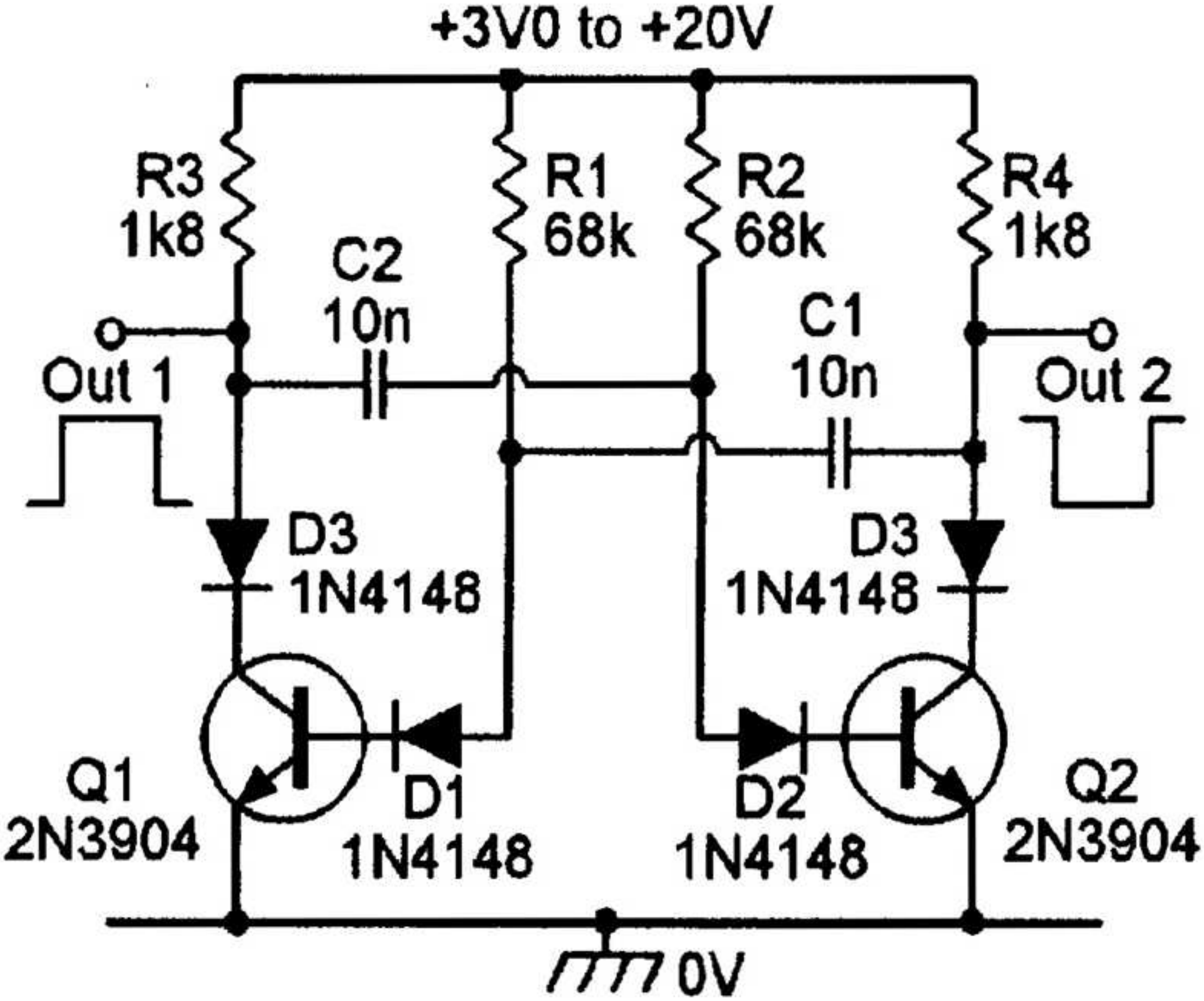


FIGURE 3. High-stability version of the basic Figure 2 1 kHz astable multivibrator circuit.

ASTABLE CIRCUIT VARIATIONS

The basic **Figure 1** astable circuit can be usefully modified in several ways, either to improve its performance or to alter the type of output waveform that it generates. Some of the most popular of these variations are shown in **Figures 4** through **9**.

One weakness of the basic **Figure 1** circuit is that the leading edges of its output waveforms are slightly rounded — the larger the values of timing resistors R1-R2 relative to collector load resistors R3-R4, the squarer the edges become. The maximum usable R1-R2 values are, in fact, limited to h_{fe} x R3 (or R4), and one obvious way of improving the waveforms is to replace Q1 and Q2 with Darlington connected pairs of transistors and then use very large R1 and R2 values, as in the **Figure 4** circuit, in which R1 and R2 can have values up to 12M, and the circuit can use any supply from 3 V to 18 V.

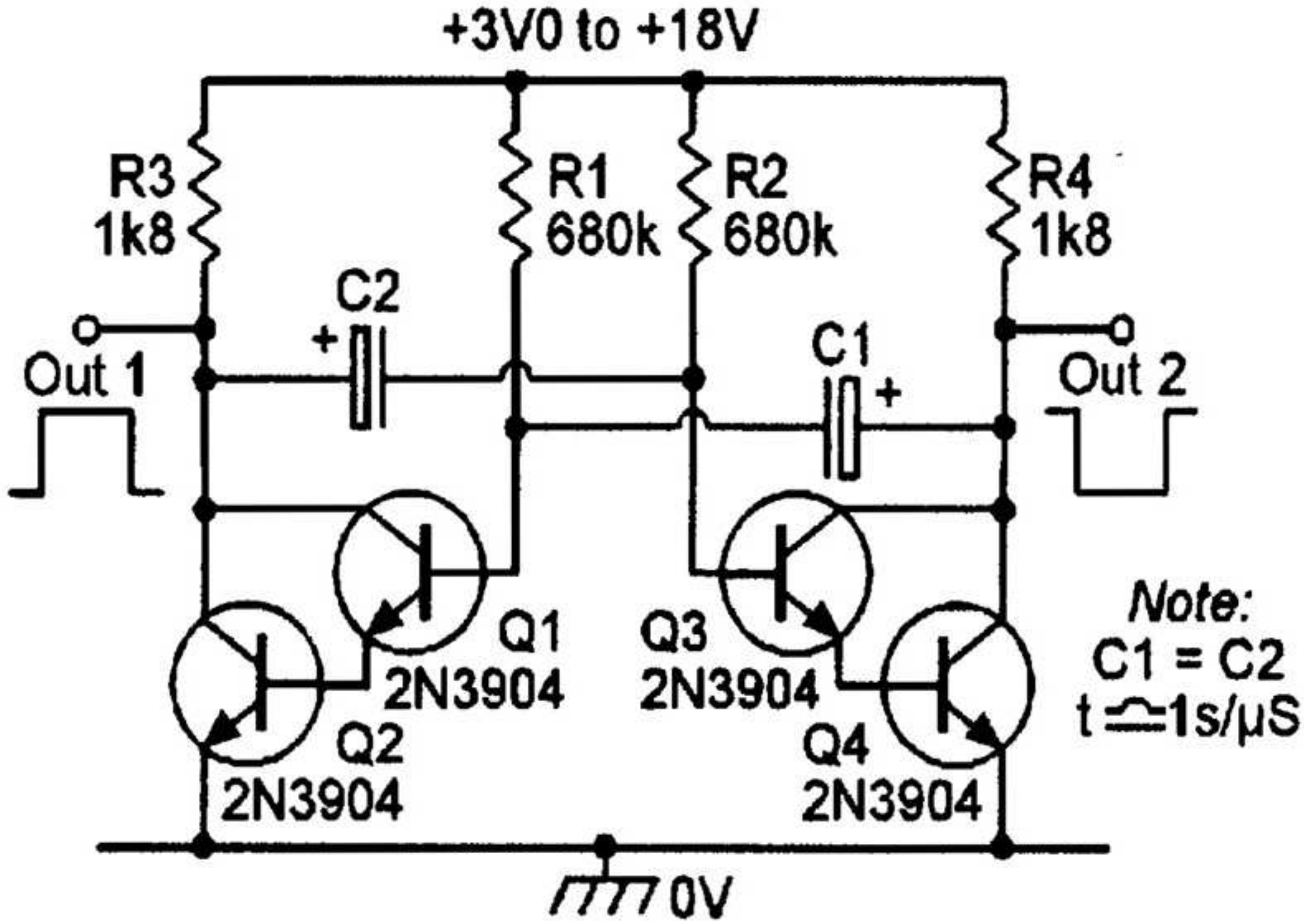


FIGURE 4. Long-period astable multivibrator.

With the R1-R2 values shown, the circuit gives a total period or cycling time of about one second per µF when C1 and C2 have equal values, and gives an excellent square wave output. The leading-edge rounding of the **Figure 1** circuit can be eliminated by using the modifications of **Figure 5**, in which steering or waveform-correction diodes D1 and D2 automatically disconnect their respective timing capacitors from the transistor collectors at the moment of transistor switching. The circuit’s main time constants are set by C1-R1 and C2-R2, but the effective collector loads of Q1 and Q2 are equal to the parallel resistances of R3-R4 or R5-R6.

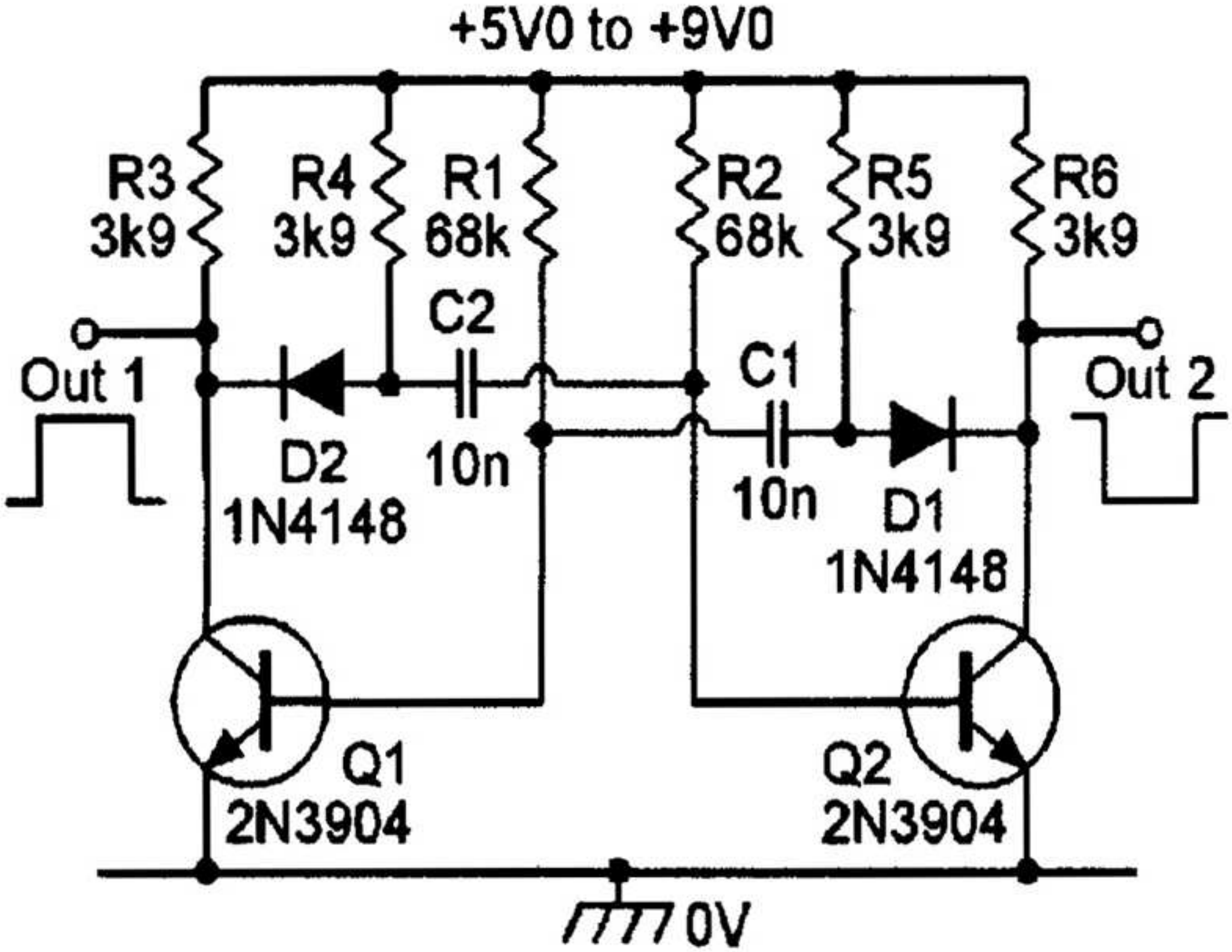


FIGURE 5. 1 kHz astable with waveform correction via steering diodes D1 and D2.

A minor weakness of the basic **Figure 1** circuit is that if its supply is slowly raised from zero to its normal value, both transistors may turn on simultaneously, and the oscillator will not start. This snag can be overcome by using the sure-start circuit of **Figure 6**, in which the timing resistors are connected to the transistor collectors in such a way that only one transistor can be on at a time.

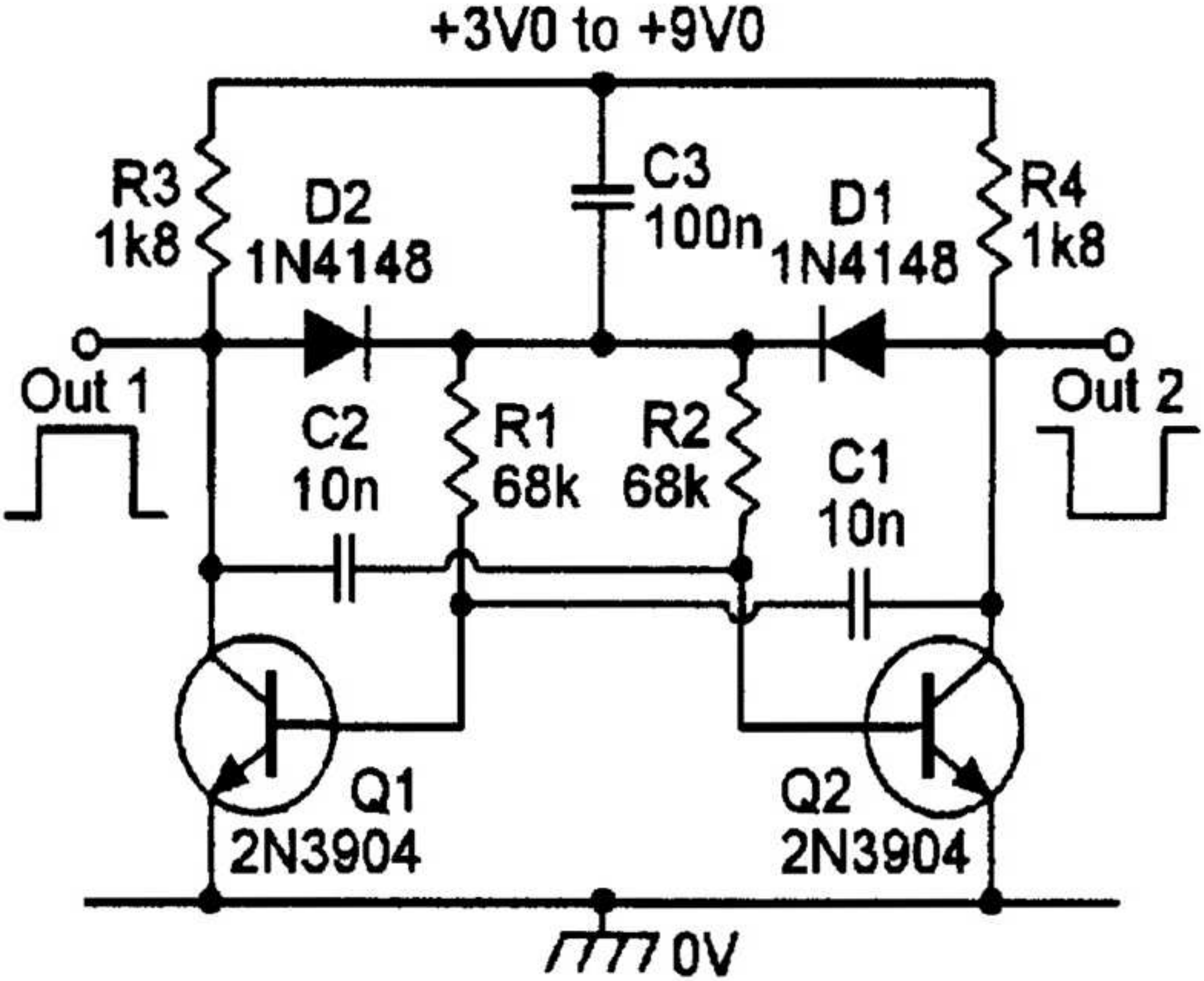


FIGURE 6. 1 kHz astable with sure-start facility.

All astable circuits shown so far give symmetrical output waveforms, with a 1:1 mark/space ratio. A non-symmetrical waveform can be obtained by making one set of astable time constants larger than the other. **Figure 7** shows a fixed-frequency (1,100 Hz) generator in which the mark/space ratio is variable from 1:10 to 10:1 via RV1.

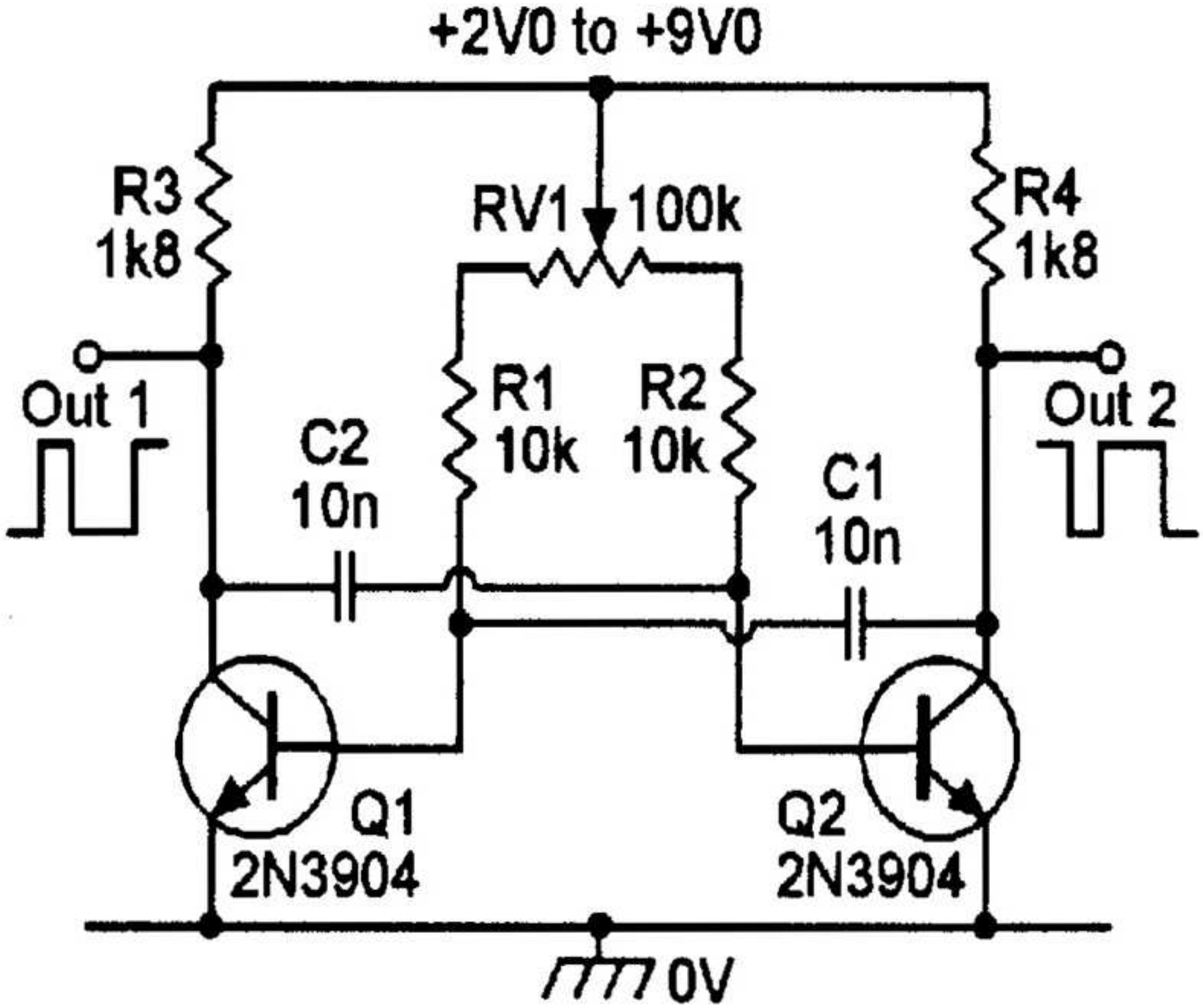


FIGURE 7. Basic 1,100 Hz variable mark/space ratio generator.

The leading edges of the output waveforms of the above circuit may be objectionably rounded when the mark-space control is set to its extreme positions. Also, the circuit may not start if its supply is applied too slowly. Both of these snags are overcome in the circuit of **Figure 8**, which is fitted with both sure-start and waveform-correction diodes.

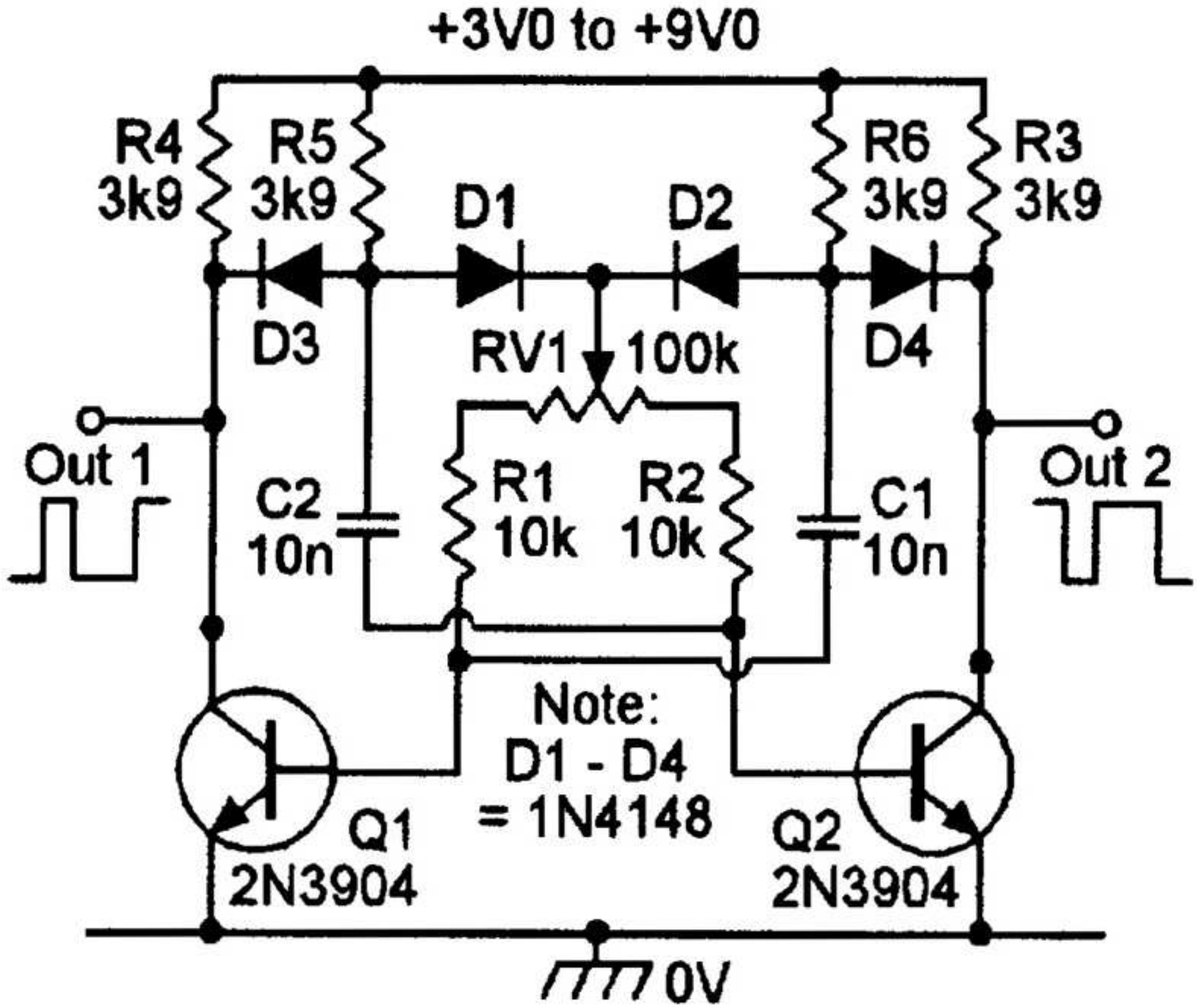


FIGURE 8. 1,100 Hz variable mark/space ratio generator with waveform correction and sure-start facility.

Finally, **Figure 9** shows a basic astable circuit modified so that its frequency is variable over a 2:1 range (from 20 kHz down to 10 kHz) via a single pot, and so that its generated waveform can be frequency modulated via an external low-frequency signal. Timing resistors R3 and R4 have their top ends taken to RV1 pot and the frequency is greatest when the pot is at the positive supply line. Frequency modulation is obtained by feeding the low-frequency signal to the tops of R3-R4 via C4; C3 presents a low impedance to the carrier signal, but a high impedance to the modulating one.

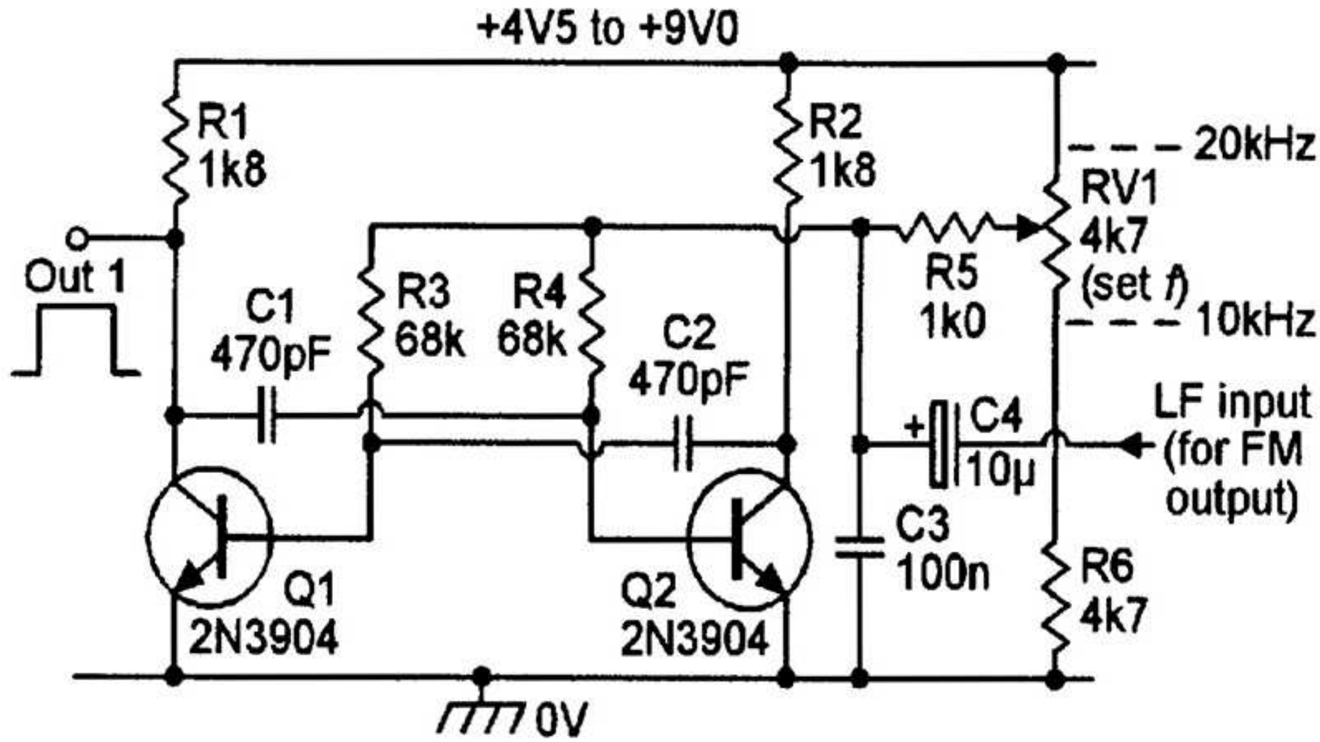
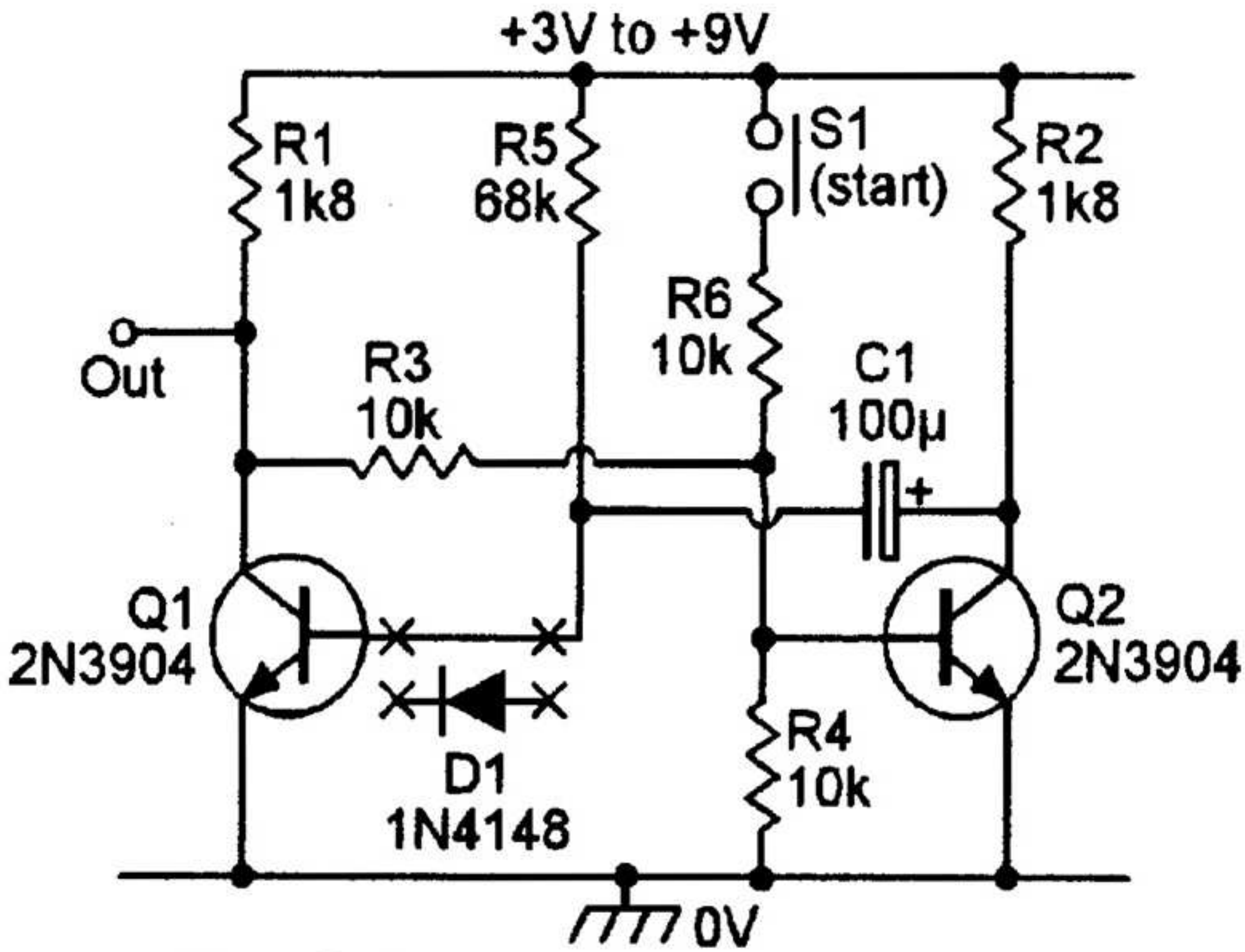


FIGURE 9. Astable with variable-frequency and FM facility.

MONOSTABLE BASICS

Monostable multivibrators are pulse generators, and may be triggered either electronically or manually. **Figure 10** shows a circuit of the latter type, which is triggered by feeding a positive pulse to Q2 base via S1 and R6. This circuit operates as follows. Normally, Q1 is driven to saturation via R5, so the output (Q1 collector) is low. Q2 (which derives its base-bias from Q1 collector via R3) is cut off under this condition, so C1 is fully charged. When a start signal is applied to Q2 base via S1, Q2 is driven on and its collector goes low, reverse biasing Q1 base via C1 and thus initiating a regenerative switching action in which Q1 is turned off (and its output switches high) via C1's negative charge, and Q2 is driven on via R1-R3 after S1 is released. As soon as the switching is complete, C1 starts to discharge via R5, until its charge falls to such a low value that Q1 starts to turn on again, thus initiating another regenerative action in which the transistors revert to their original states and the output pulse terminates, completing the action.



Note: Delay (p) = 50mS/µS
= 5sec with C1 value shown

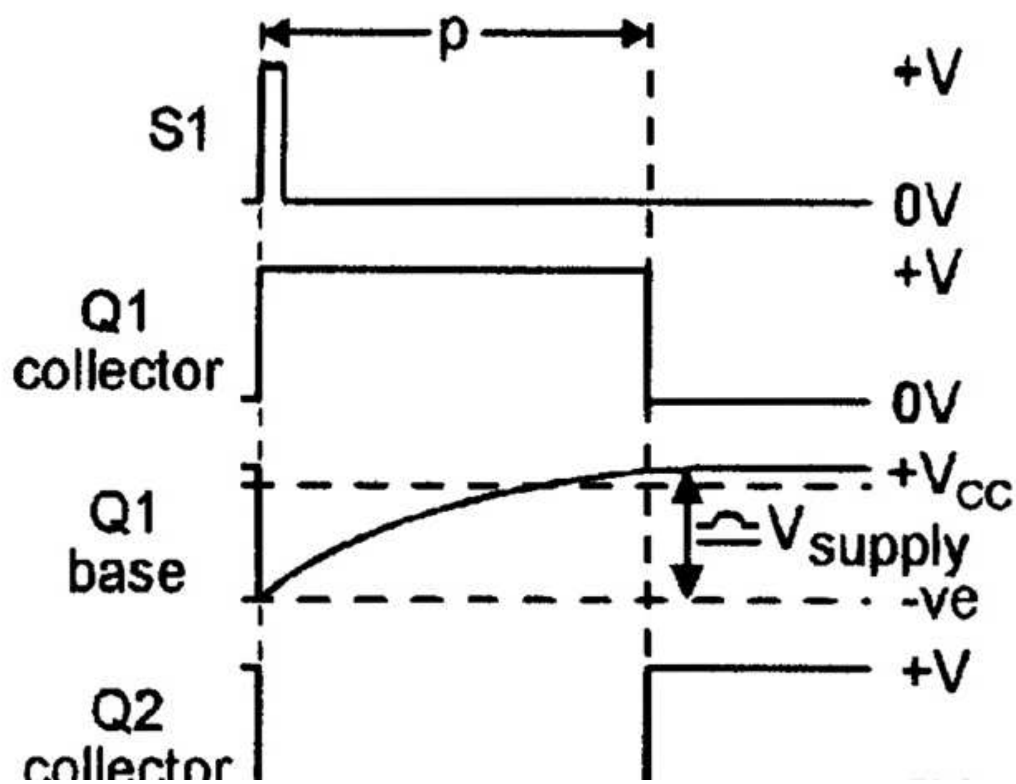


FIGURE 10. Basic manually-triggered monostable pulse generator.

Thus, a positive pulse is developed at the Q1 output each time an input trigger signal is applied via S1. The pulse period (P) is determined by the R5-C1 values, and approximates 0.7 x R5 x C1, where P is in mS, C is in µF, and R is in kilohms, and equals about 50mS/µF in the example shown. In practice, the **Figure 10** circuit can be triggered either by applying a negative pulse to Q1 base or a positive one to Q2 base (as shown). Note that the base-emitter junction of Q1 is reverse biased by a peak amount equal to V_{SUPPLY} during the operating cycle, thus limiting the maximum usable supply voltage to about 9 V. Greater supply voltages can be used by wiring a silicon diode in series with Q1 base, as shown by D1 in the diagram, to give the same frequency correction action as described earlier for the astable circuit.

LONG DELAYS

The value of timing resistor R5 must be large relative to R2, but must be less than the product of R1 and Q1's h_{fe} value. Very long timing periods can be obtained by using a Darlington or Super-Alpha pair of transistors in place of Q1, thus enabling large R5 values to be used, as shown in the **Figure 11** circuit, which gives a pulse period of about 100 seconds with the component values shown.

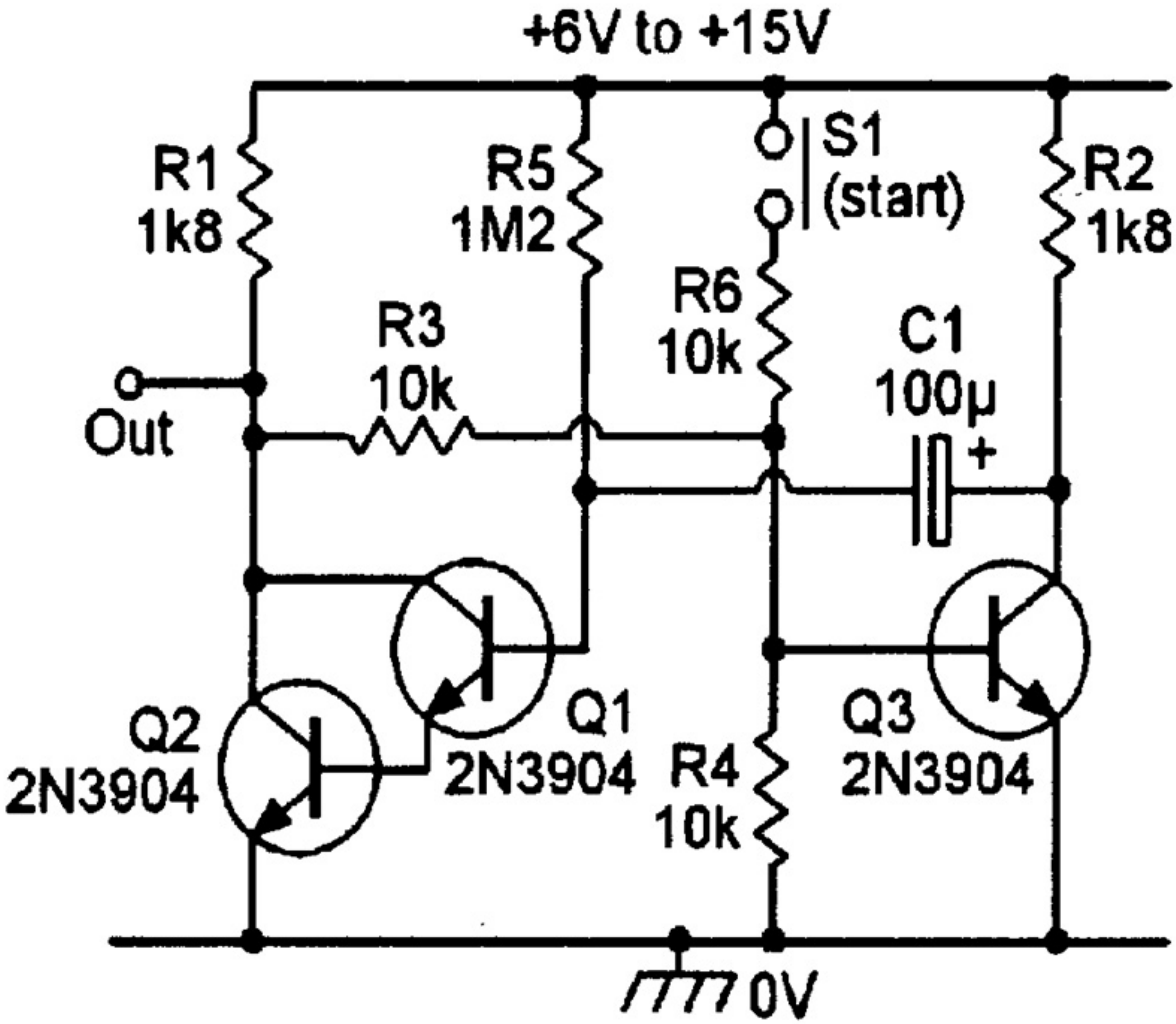


FIGURE 11. Long-period (100 second) monostable circuit.

An important point to note is that the basic **Figure 10** circuit actually triggers at the moment of application (via S1 and R6) of a positive-going pulse to the base of Q2. If this pulse is removed before the monostable completes its natural timing period, the pulse will end regeneratively in the way already described, but if the trigger pulse is not removed (via S1) at this time, the monostable period will end non-regeneratively and will have a longer period and fall-time than normal. This problem can be eliminated by using electronic (rather than manual) triggering, as described in the next section.

ELECTRONIC TRIGGERING

Figures 12 and **13** show alternative ways of applying electronic triggering to the monostable pulse generator. In each case, the circuit is triggered by a square wave input with a short rise time. This waveform is differentiated by C2-R6, to produce a brief trigger pulse. In the Figure 12 circuit, the differentiated input signal is discriminated by D1, to provide a positive trigger pulse on Q2 base each time an external trigger signal is applied. In the **Figure 13** circuit, the differentiated signal is fed to Q3, which enables the trigger signal to be quite independent of Q2. Note in the latter circuit that speed up capacitor C3 is wired across feedback resistor R3 to help improve the shape of the circuit’s output pulse.

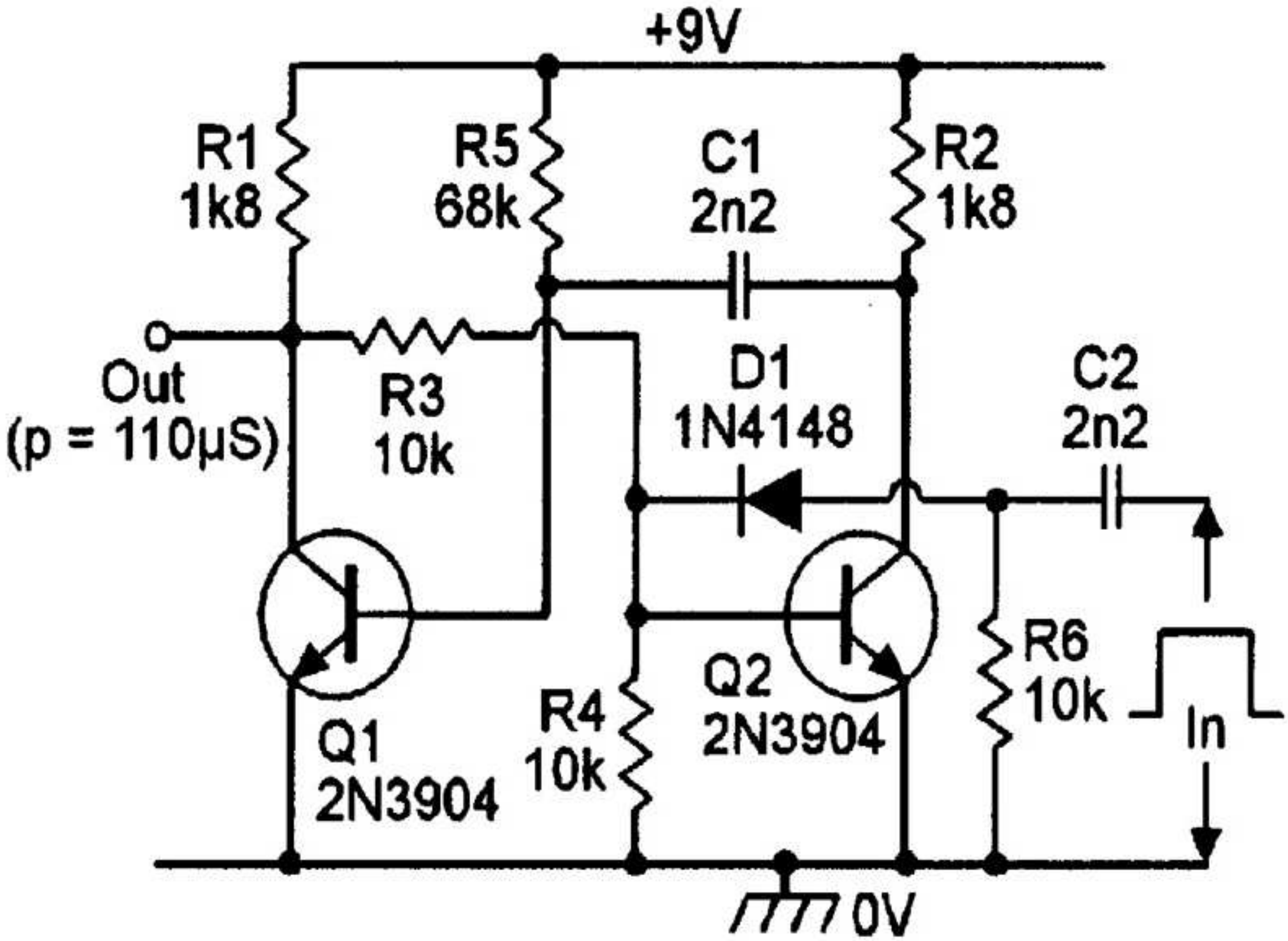


FIGURE 12. Electronically triggered monostable.

The **Figure 12** and **13** circuits each give an output pulse period of about 110 mS with the component values shown. The period can be varied from a fraction of a millisecond to many seconds by choice of the C1-R5 values. The circuits can be triggered by sine or other non-rectangular waveforms by feeding them to the monostable input via a Schmitt trigger or similar sine/square converter circuit (see **Figure 20**).

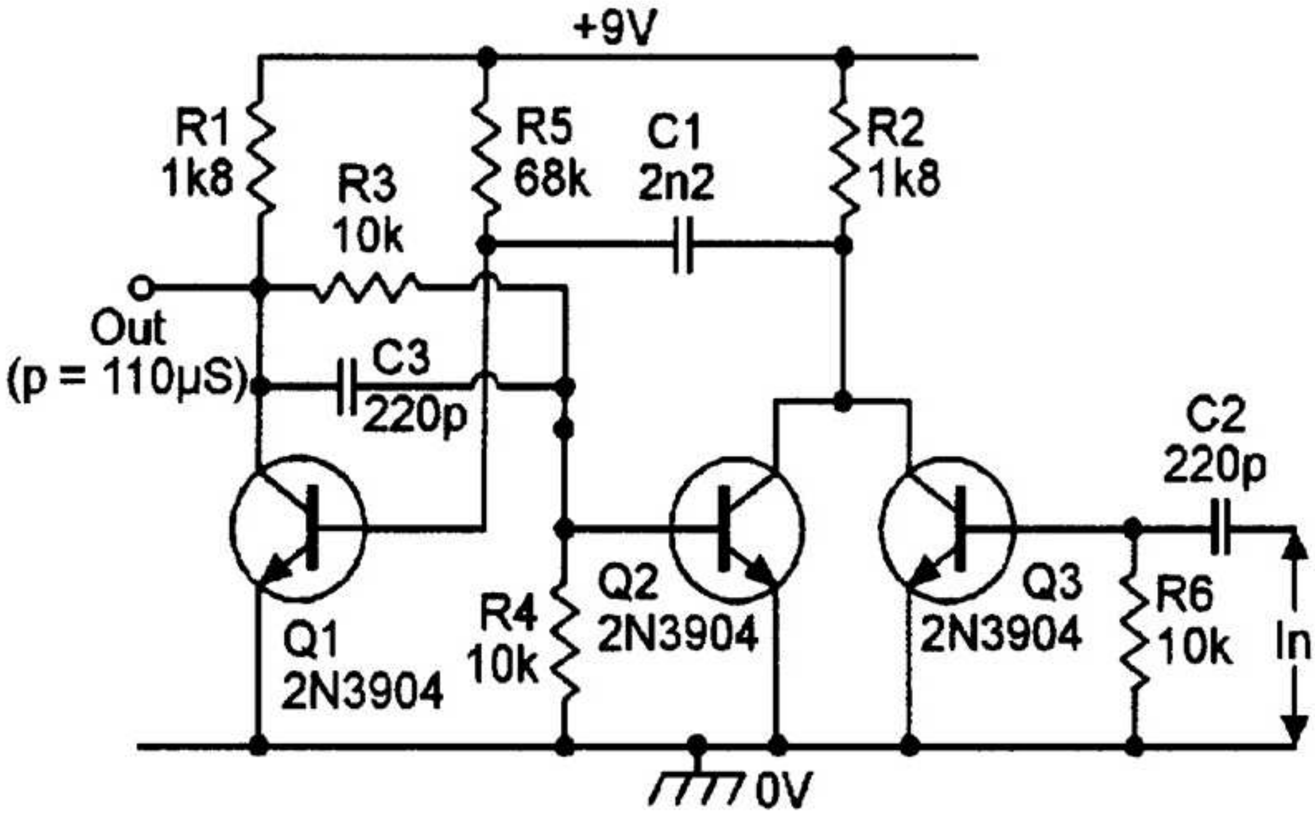


FIGURE 13. Monostable with gate-input triggering.

BISTABLE CIRCUITS

Bistable multivibrators make good stop/go waveform generators, and **Figure 14** shows a basic manually-triggered version of such a circuit, which is also known as an R-S (Reset-Set) flip-flop. Its output can be set to the high state by briefly closing S1 (or by applying a negative pulse to Q1 base via a current-limiting resistor), thus turning Q1 off (and simultaneously turning Q2 on via the R3 cross-coupling), and the circuit then latches into this state until it is reset to the low state by briefly closing S2 (or by applying a negative current-limited pulse to Q2 base), thus turning Q2 off and therefore turning Q1 on via the R4 cross coupling. The circuit then latches into this new state until it is set again via S1, and so on.

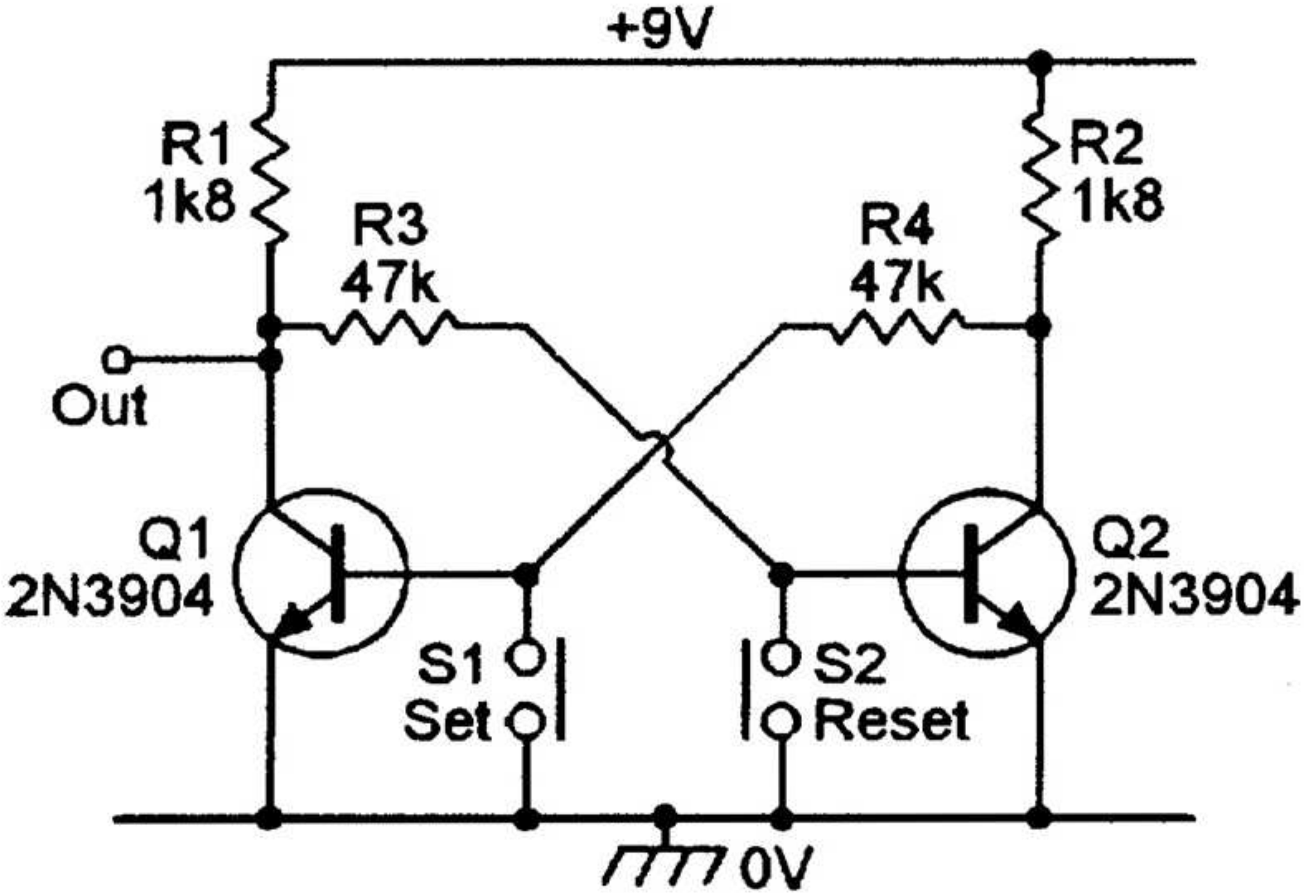


FIGURE 14. Basic manually-triggered R-S bistable multivibrator.

The latching action of the basic **Figure 14** circuit relies on the fact that the saturation voltage (typically 200 mV) of the ON transistor is significantly lower than the base-biasing voltage (typically 600 mV) of the opposing device. In practice, these ideal conditions may not be met if the transistor is not a good-quality silicon type, or if it operates at an excessive temperature or with a low-value collector load. In cases of doubt, the circuit's reliability can be greatly enhanced by using the modifications shown in the improved circuit of **Figure 15**, in which resistors R5 and R6 act as simple potential dividers with R3 and R4, respectively, thus reducing the undesirable effects of high saturation voltages, etc.

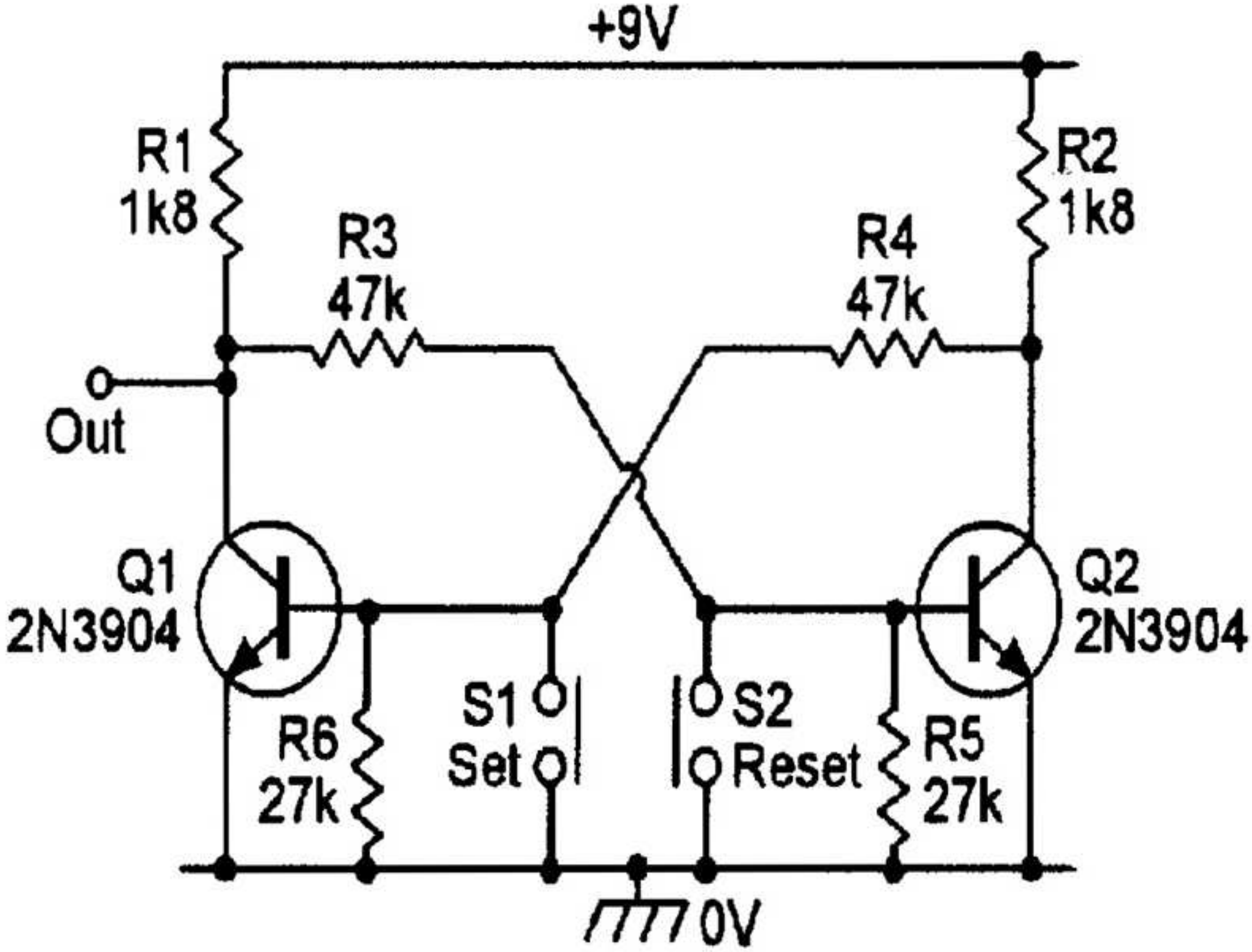


FIGURE 15. Improved manually-triggered R-S bistable multivibrator with switch-low triggering.

The circuits of **Figures 14** and **15** both give a switch-low triggering action, in which the circuit changes state when an ON transistor is turned OFF by pulling its base low via a switch or by applying a negative pulse to its base.

Figure 16 shows an alternative version of the basic manually-triggered bistable, in which the circuit gives a switch-high action in which the circuit changes state when an OFF transistor is turned ON by pulling its base high via a switch or by applying a positive current-limited pulse to its base.

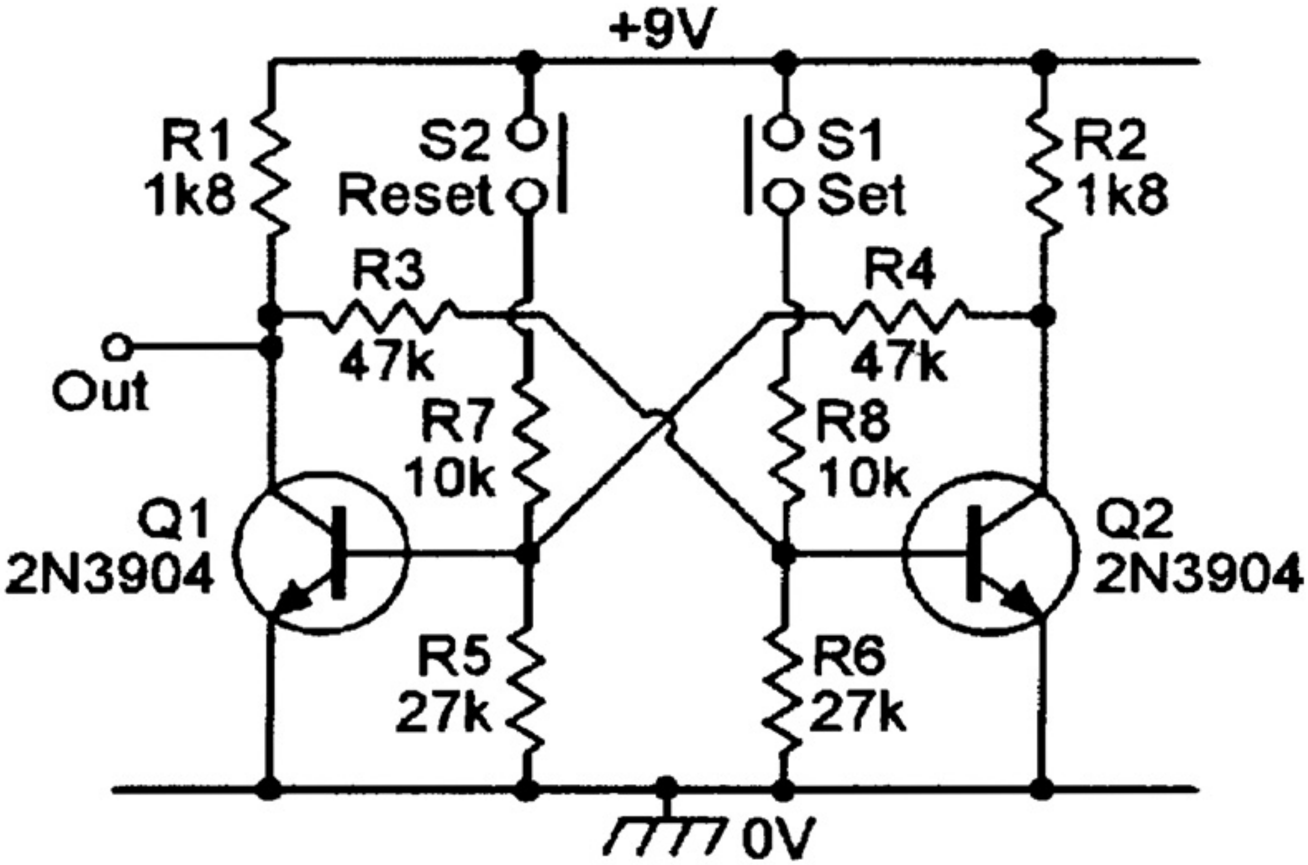


FIGURE 16. Manually-triggered R-S bistable with switch-high triggering.

Note that when power is initially applied to the basic **Figure 14** to **16** circuits, the output initially settles into a randomly-determined state that depends on the relative characteristics of the two transistors and their associated passive components.

If desired, the basic circuit can be made to automatically switch into a desired initial power up state by automatically feeding a suitable switch-on trigger pulse to the base of one or the other of the two transistors, as shown in **Figure 17**, which shows the basic **Figure 15** circuit modified (via R7-C1 and current-limiting resistor R8) so that the circuit automatically switches into the set (Q1 output high) state at power-up.

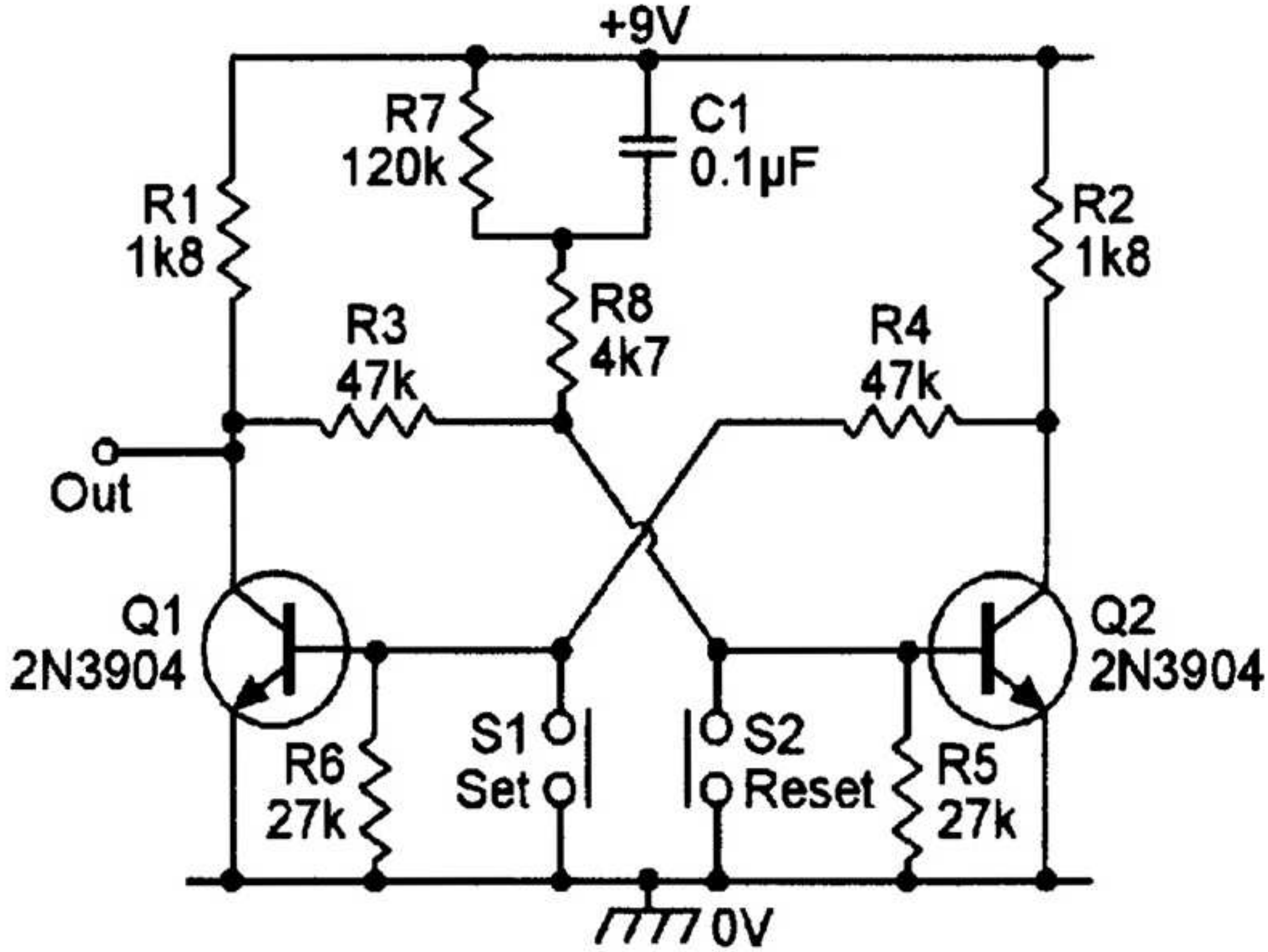


FIGURE 17. Basic Figure 15 circuit modified to give SET action at initial power-up.

One of the most useful applications of the basic bistable multivibrator is as a push-button-controlled timer circuit, in which the output automatically goes high at power-up or on the closure of a push-button start switch, but goes low again automatically after a pre-set delay. **Figure 18** shows the basic **Figure 17** circuit modified to give such action. Here, the Q1 output automatically goes high (via R7-C1 and R8) at the moment of initial power-up, thereby activating (via emitter follower Q3) an adjustable delayed-pulse generator, which automatically feeds a reset pulse to Q1 base via D1-R9 at the end of the desired delay period, thereby completing the circuit’s operating cycle.

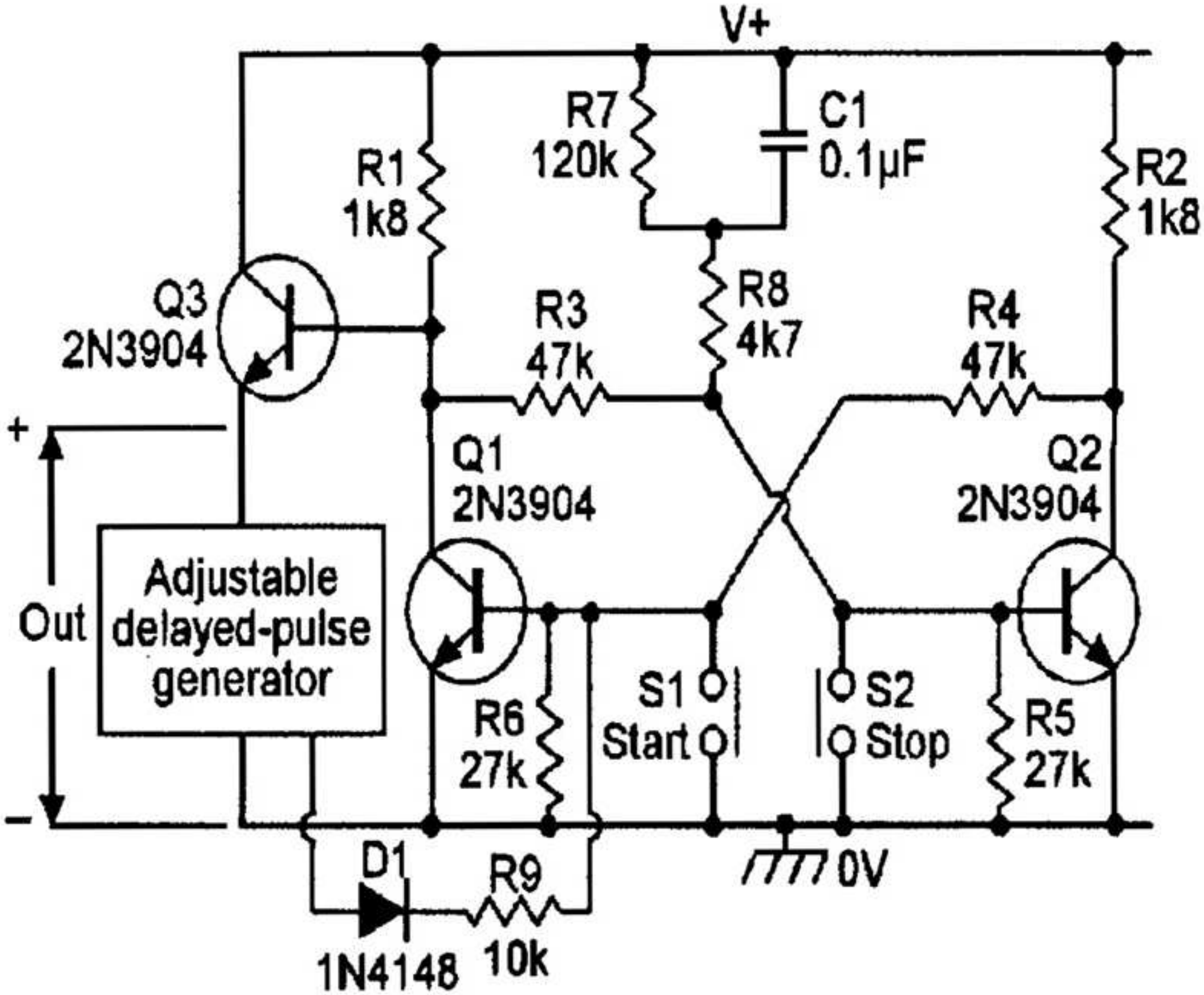


FIGURE 18. Basic circuit of a manually-triggered bistable multivibrator with timed auto-reset action.

Finally, before leaving the basic bistable multivibrator circuit, note that it can, by connecting two steering diodes and associated components as shown in **Figure 19**, be modified to give a divide-by-two or counting action in which it changes state each time a negative-going trigger pulse is applied. The circuit generates a pair of anti-phase outputs, known as Q and not-Q (denoted by a bar over the Q sign in the diagram). In practice, greatly improved versions of this counting type of circuit are readily available in CMOS or TTL digital IC form.

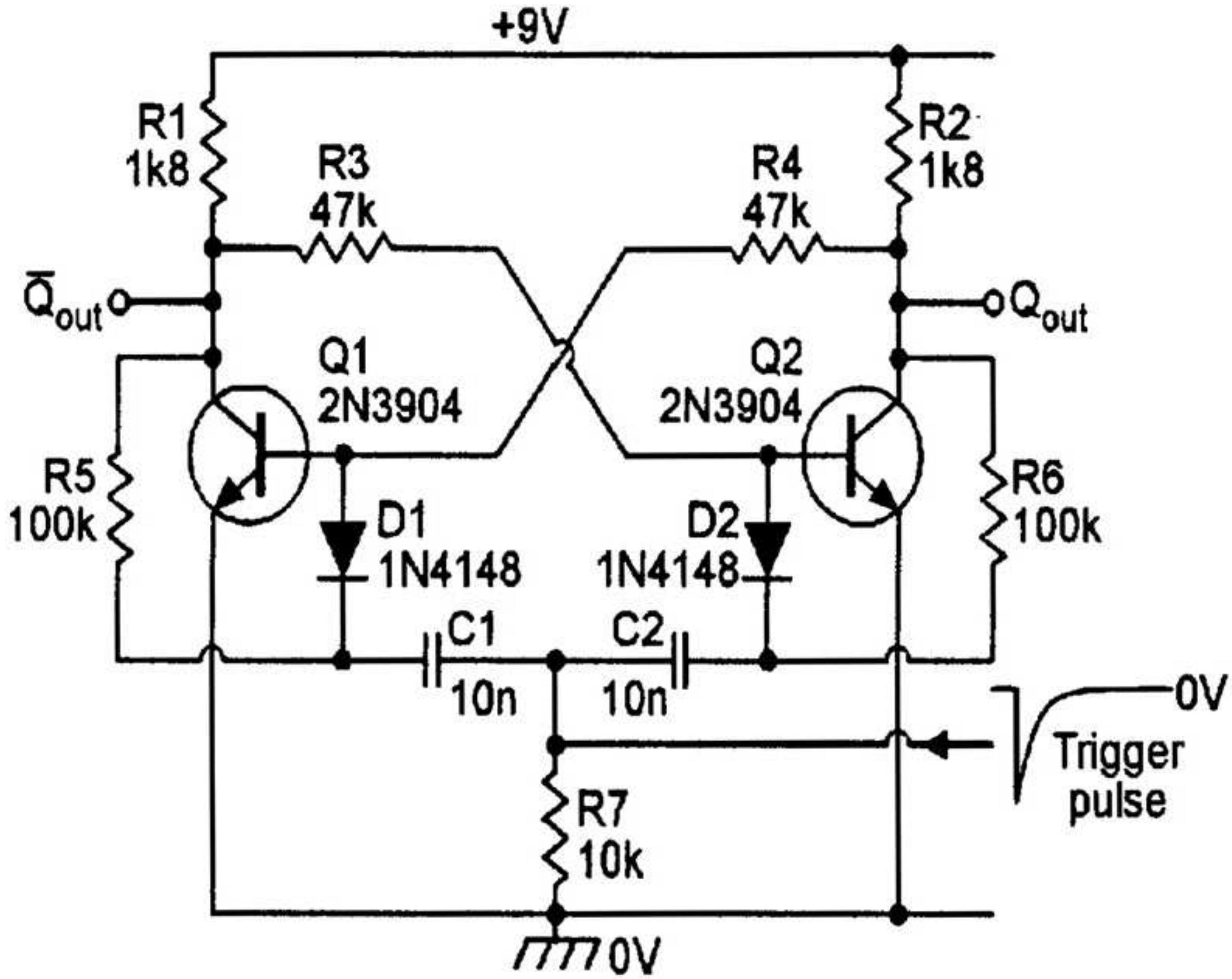


FIGURE 19. Divide-by-two bistable circuit.

THE SCHMITT TRIGGER

The final member of the multivibrator family is the Schmitt trigger. This is a voltage-sensitive bistable switching circuit that changes its output state when the input goes above or below pre-set upper and lower threshold levels; to complete this month's discussion, **Figure 20** shows a simple Schmitt trigger circuit used as a sine-to-square waveform converter that gives a good performance up to a few hundred kHz and needs a sine wave input signal amplitude of at least 0.5V RMS.

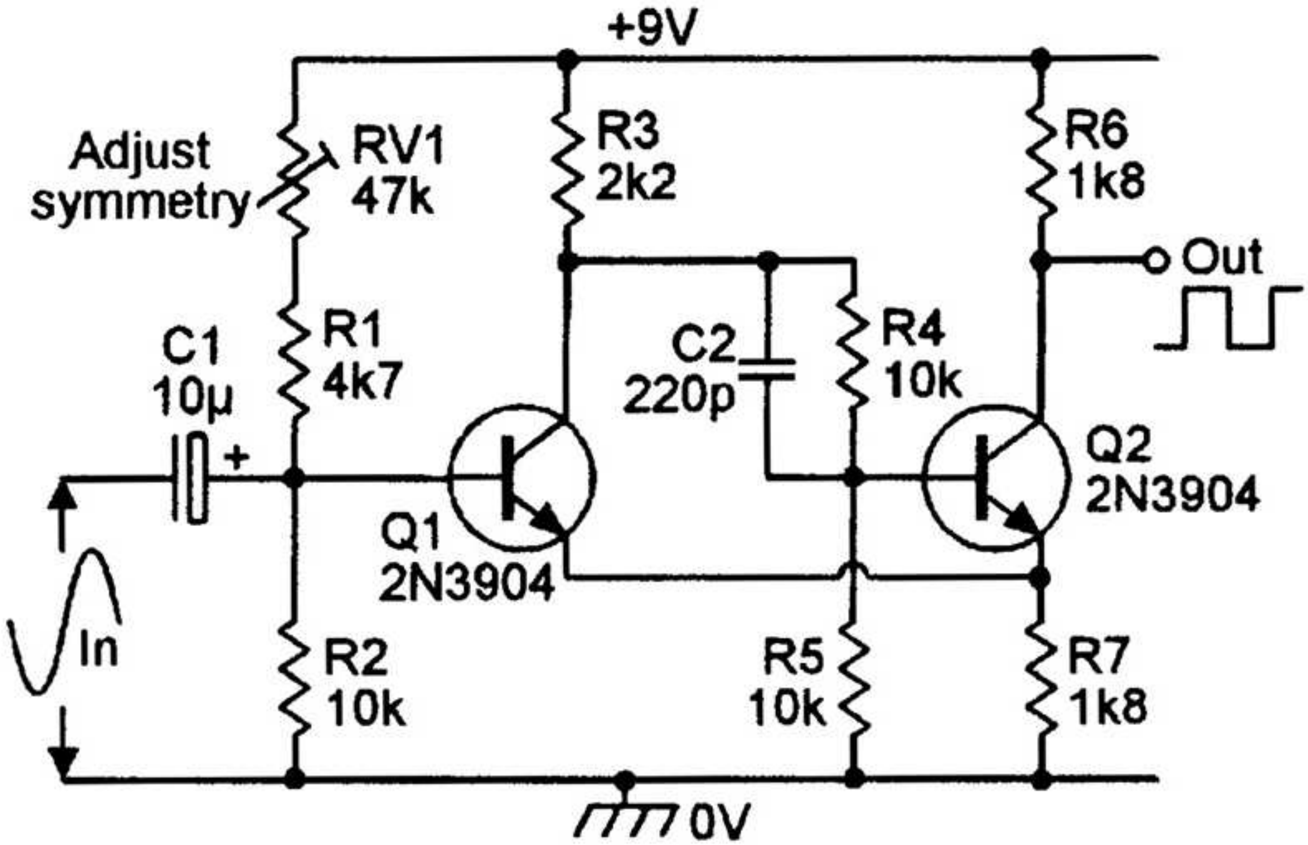


FIGURE 20. Schmitt sine/square converter.

The output signal symmetry varies with input signal amplitude; RV1 should be adjusted to give best results. Next month, we'll describe a variety of audio power amplifier circuits and associated gadgets. **NV**

Bipolar Transistor Cookbook — Part 7

One of the most popular applications of transistors is in audio power amplifiers. This month we describe the operating principles of various circuits of this type and present a selection of practical audio power amplifier circuit designs. The installment concludes by presenting a practical 'scratch and rumble' filter circuit, which can be used to eliminate these unwanted sounds when playing old-fashioned records/discs through any type of audio power amplifier system.

POWER AMPLIFIER BASICS

A transistor power amplifier's job is that of converting a medium-level medium-impedance AC input signal into a high-level low-impedance state suitable for driving a low-impedance external load. This action can be achieved by operating the transistor(s) in either of two basic modes, known as 'class-A' or 'class-B.'

Figure 1(a) shows a basic class-A audio amplifier circuit; Q1 is a common-emitter amplifier with a loudspeaker collector load, and is so biased that its collector current has a quiescent value halfway between the desired maximum and minimum swings of output current, as shown in **Figure 1(b)**, so that maximal low-distortion output signal swings can be obtained. The circuit consumes a high quiescent current, and is relatively inefficient; 'efficiency' is the ratio of AC power feeding into the load, compared with the DC power consumed by the circuit, and at maximum output power is typically about 40%, falling to 4% at one tenth of maximum output, etc.

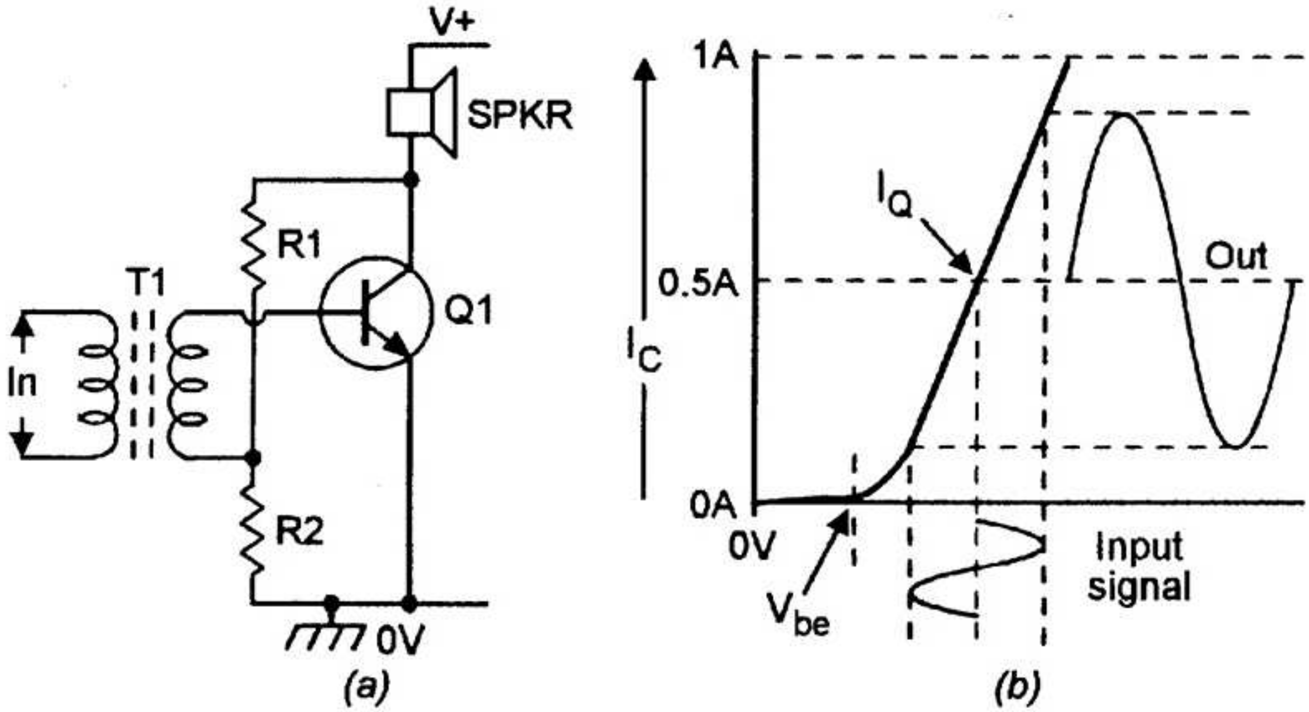


FIGURE 1. Basic circuit (a) and transfer characteristics (b) of a class-A amplifier.

Figure 2 shows an example of a low-power (up to a few dozen milliwatts) high-gain general-purpose class-A amplifier that draws a quiescent current of about 20 mA and is suitable for driving a medium impedance (greater than 65Ω) loudspeaker or headset. Q1 and Q2 are wired as direct-coupled common-emitter amplifiers, and give an overall voltage gain of about 80 dB. Q1's base bias is derived (via R2) from Q2's emitter, which is decoupled via C3 and thus 'follows' the mean collector voltage of Q1. The bias is thus stabilized by DC negative feedback. Input pot RV1 acts as the circuit's volume control.

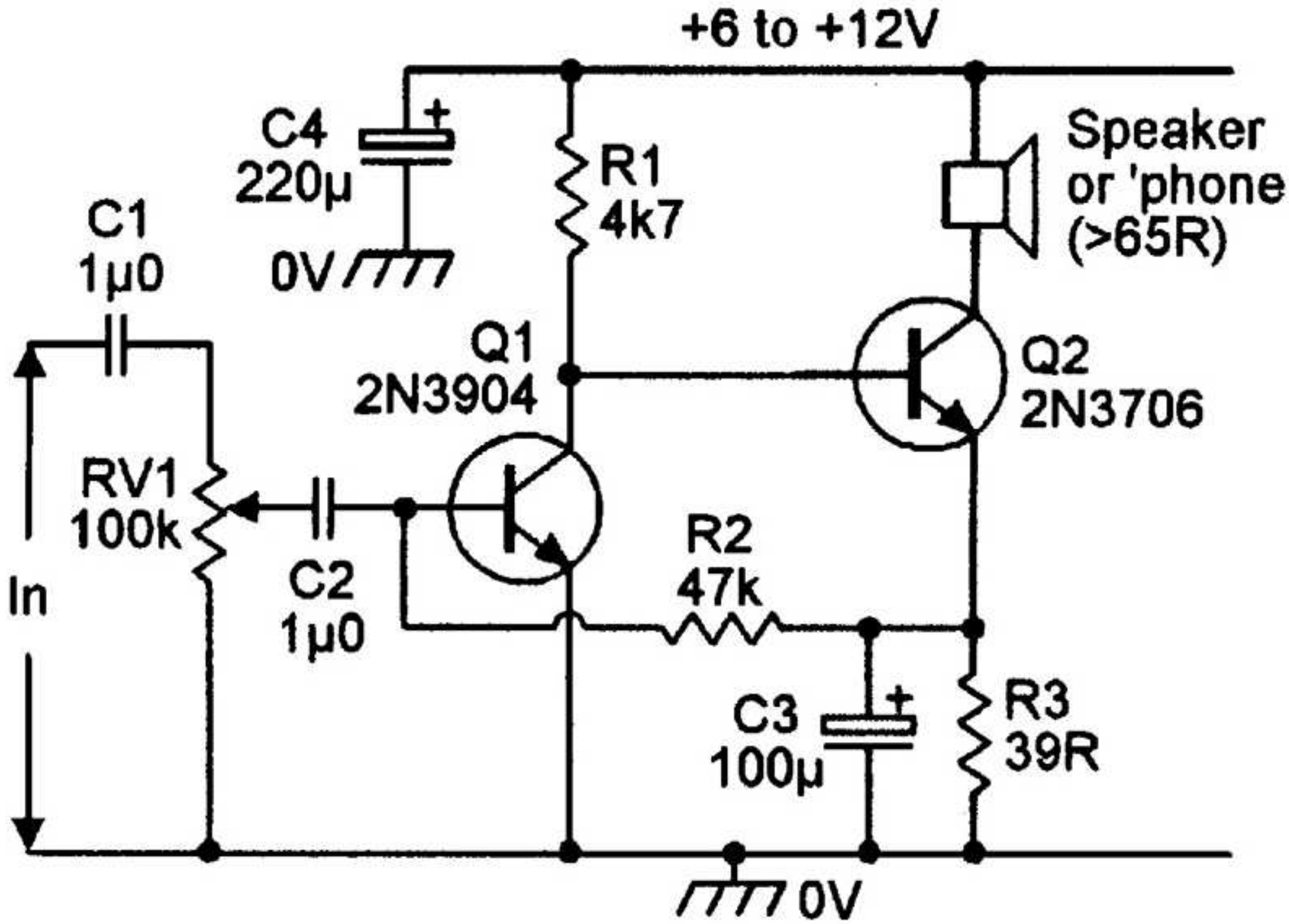
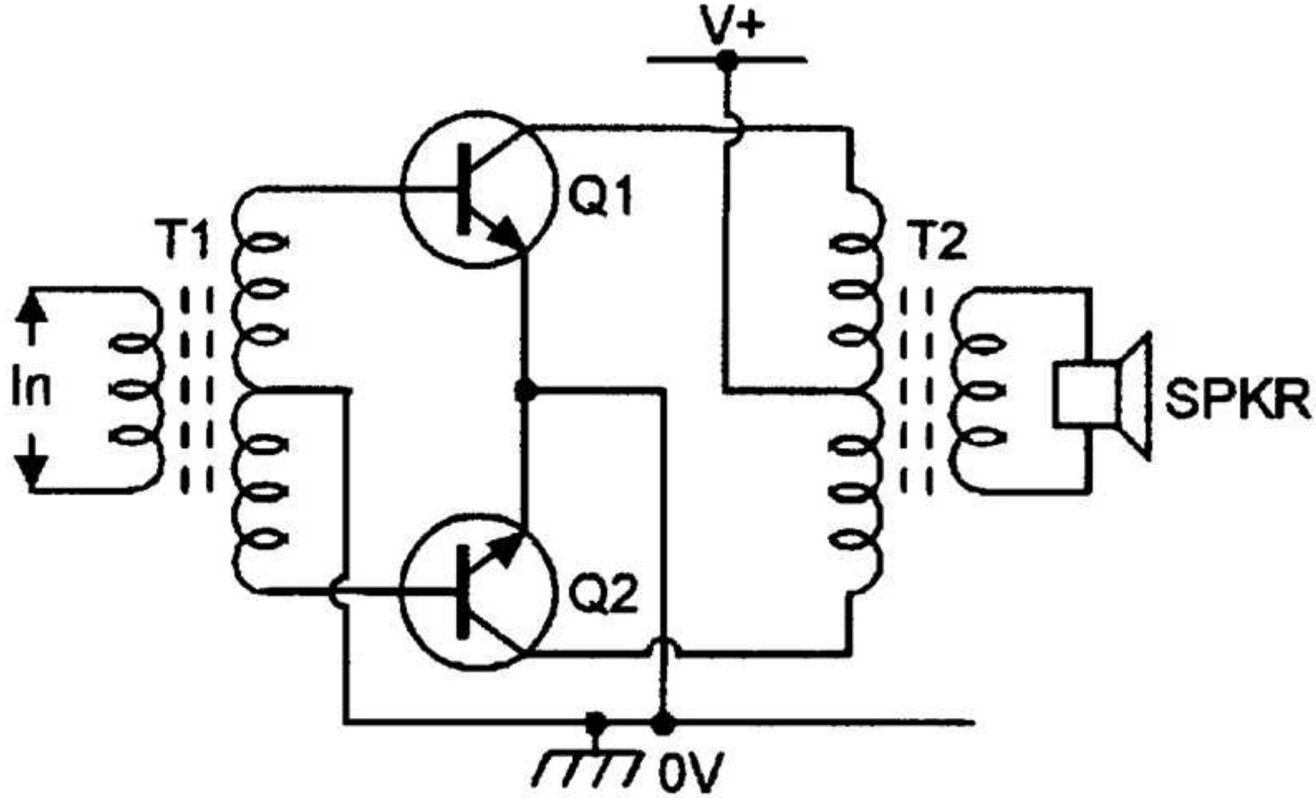


FIGURE 2. General-purpose high-gain low-power audio amplifier.

A basic class-B amplifier consists of a pair of transistors, driven in anti-phase but driving a common output load, as shown in **Figure 3(a)**. In this particular design, Q1 and Q2 are wired in the common-emitter mode and drive the loudspeaker via push-pull transformer T2, and the anti-phase input drive is obtained via phase-splitting transformer T1. The essential features of this type of amplifier are that both transistors are cut off under quiescent conditions, that neither transistor conducts until its input drive signal exceeds its base-emitter 'knee' voltage, and that one transistor is driven on when the other is driven off, and vice versa. The circuit consumes near-zero quiescent current, and has high efficiency (up to 78.5%) under all operating conditions, but it generates severe cross-over distortion in the amplifier's output signal, as shown in **Figure 3(b)**. The basic class-B circuit must thus be modified if it is to be used as a practical audio power amplifier; the modified circuit is known as a 'class-AB' amplifier.


V+
Q1
T1
T2
In
SPKR
Q2
0V
(a)

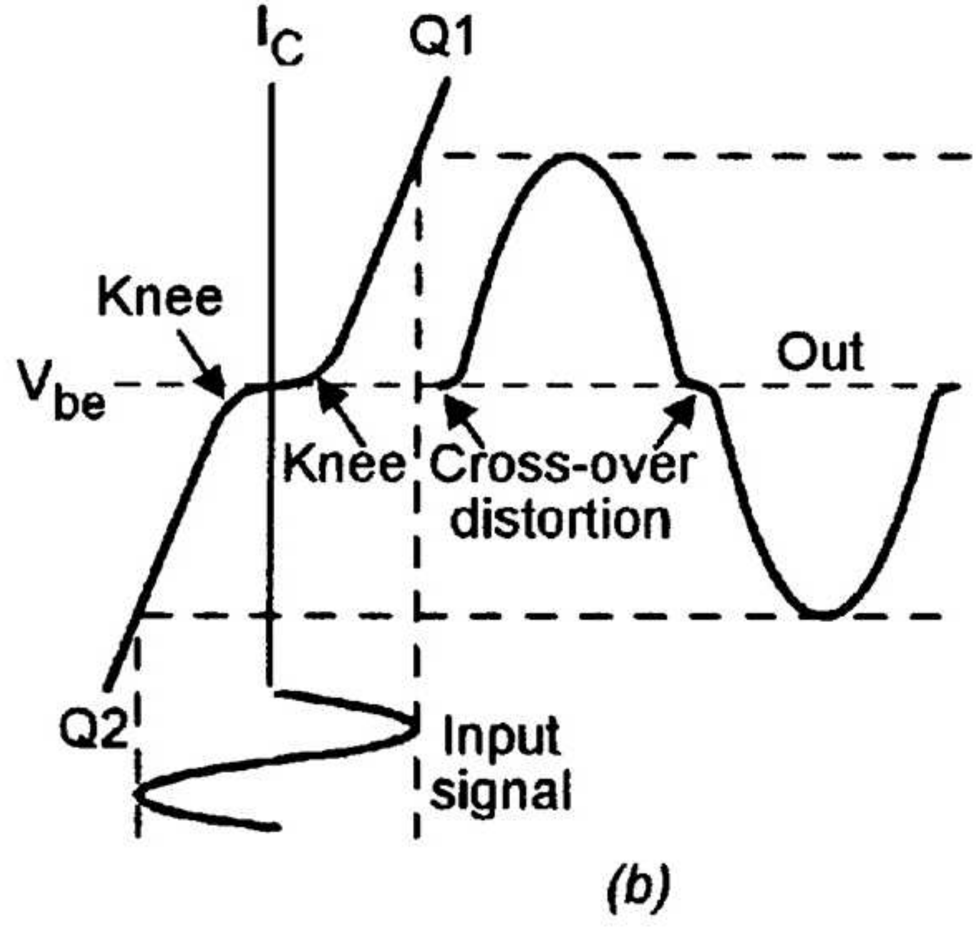

I_C
Q1
Knee
V_{be}
Out
Knee
Cross-over
distortion
Q2
Input
signal
(b)

FIGURE 3. Basic circuit (a) and transfer characteristics (b) of a class-B amplifier.

CLASS-AB BASICS

The cross-over distortion of the class-B amplifier can be eliminated by applying slight forward bias to the base of each transistor, as shown in **Figure 4**, so that each transistor passes a modest quiescent current. Such a circuit is known as a class-AB amplifier. Circuits of this type were widely used in early transistor power amplifier systems but are now virtually obsolete, since they require the use of transformers for input phase-splitting and output loudspeaker driving, and must have closely matched transistor characteristics if a good low-distortion performance is to be obtained.

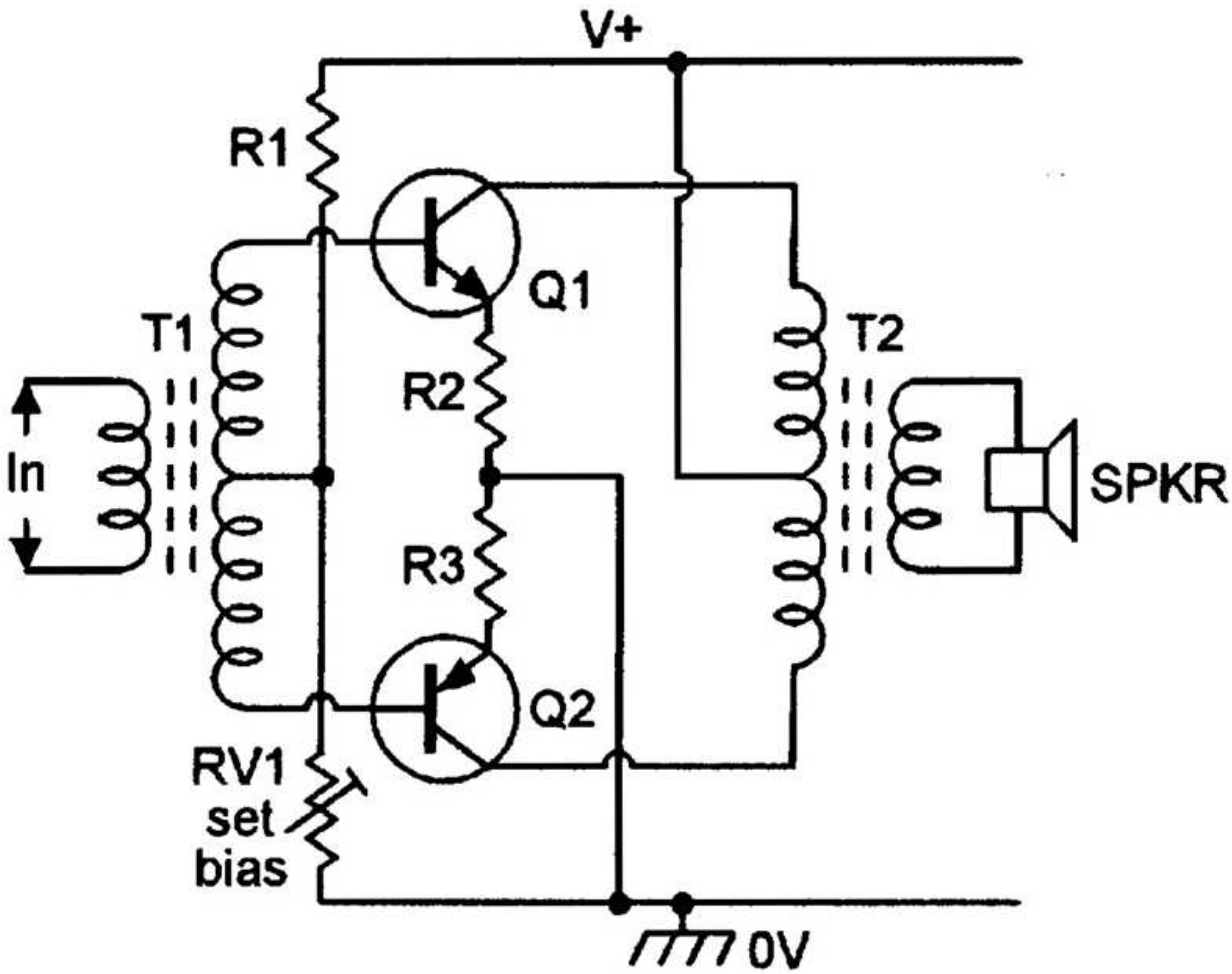


FIGURE 4. Basic circuit of a class-AB amplifier.

Figure 5 shows the basic circuit of a class-AB amplifier that suffers from none of the snags mentioned above. It is a complementary emitter follower, and is shown using a split (dual) power supply. Q1 and Q2 are biased (via R1-RV1-R2) so that their outputs are at zero volts and zero current flows in the loudspeaker load under quiescent conditions, but have slight forward bias applied (via RV1), so that they pass modest quiescent currents and thus do not suffer from cross-over distortion problems. Identical input signals are applied (via C1 and C2) to the bases of both emitter followers. This circuit's operation was described in Part 2 of this 'Cookbook' series.

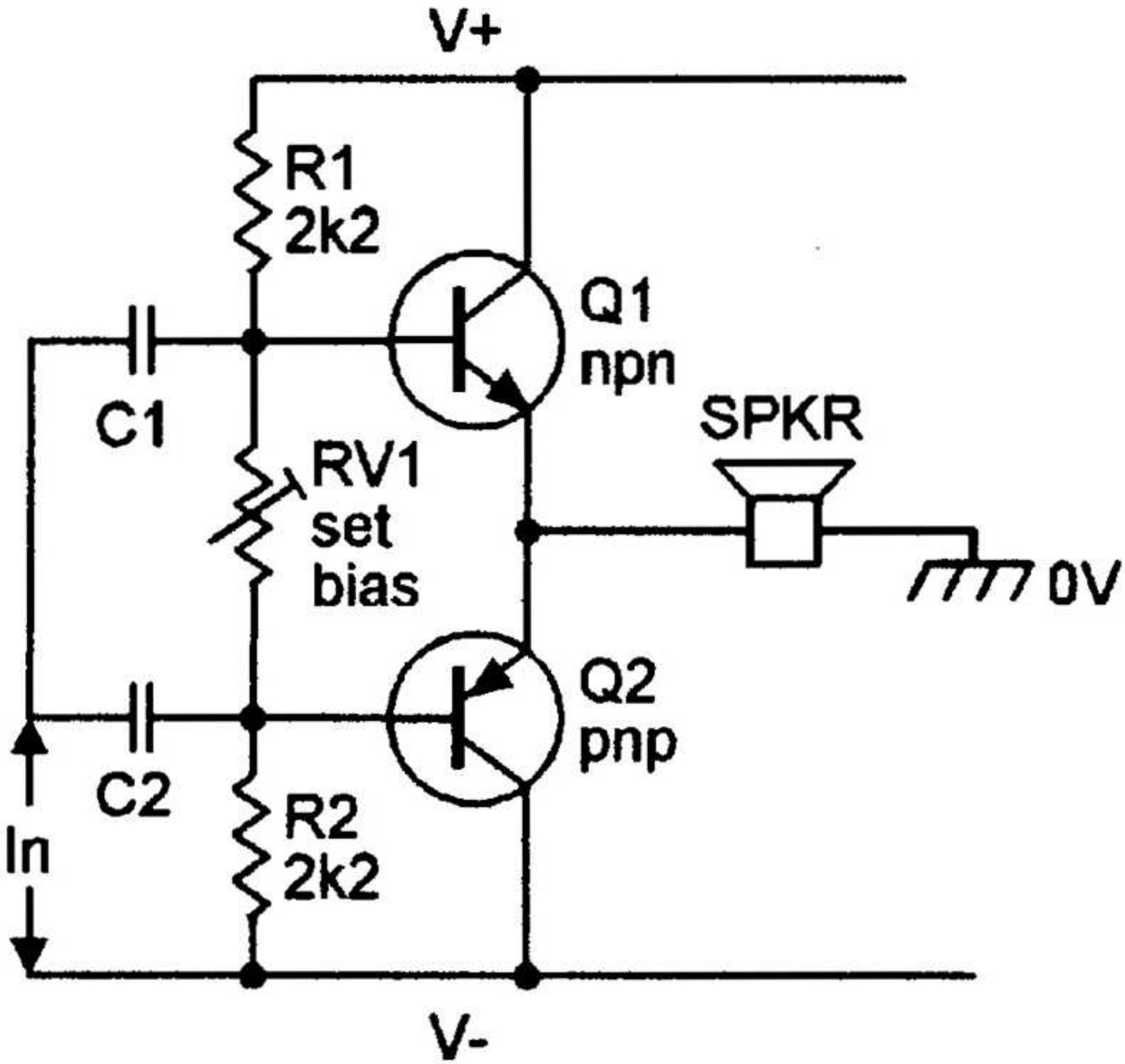
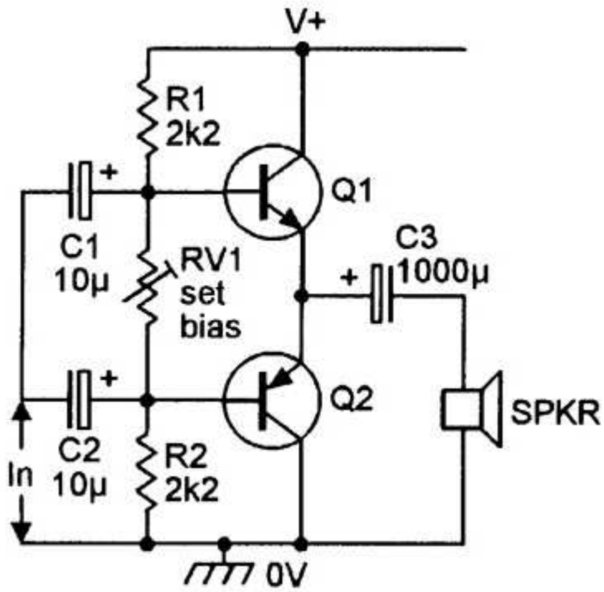


FIGURE 5. Basic class-AB amplifier with complementary emitter follower output and dual power supply.

The basic **Figure 5** circuit does not require the use of transistors with closely matched electrical characteristics, and gives direct drive to the speaker. It can be modified for use with a single-ended power supply by simply connecting one end of the speaker to either the zero or the positive supply rail, and connecting the other end to the amplifier output via a high-value blocking capacitor, as shown in **Figure 6**.



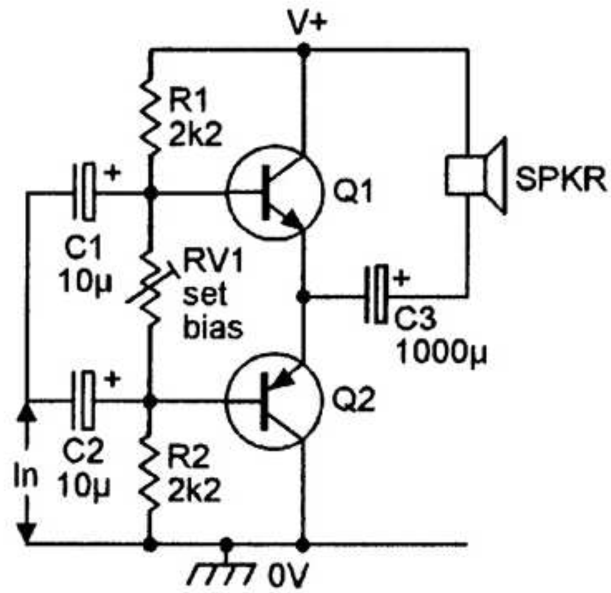


FIGURE 6. Alternative versions of the class-AB amplifier with a single-ended power supply.

The basic **Figure 5** and **6** circuits form the basis of virtually all modern audio power amplifier designs, including those in IC form. Many modifications and variations can be made to the basic circuit.

CIRCUIT VARIATIONS

The **Figure 5** circuit gives unity overall voltage gain, so an obvious circuit modification is to provide it with a voltage-amplifying driver stage, as in **Figure 7**. Here, common emitter amplifier Q1 drives the Q2-Q3 complementary emitter followers via collector load resistor R1 and auto-biasing silicon diodes D1 and D2 (the function of these diodes was explained in Part 2 of this series). Q1's base bias is derived from the circuit's output via R2-R3, thus providing DC feedback to stabilize the circuit's operating points, and AC feedback to minimize signal distortion. In practice, a pre-set pot is usually wired in series with D1-D2, to enable the Q2-Q3 bias to be trimmed; low-value resistors R4 and R5 are wired in series with Q2 and Q3 emitters to prevent thermal runaway, etc.

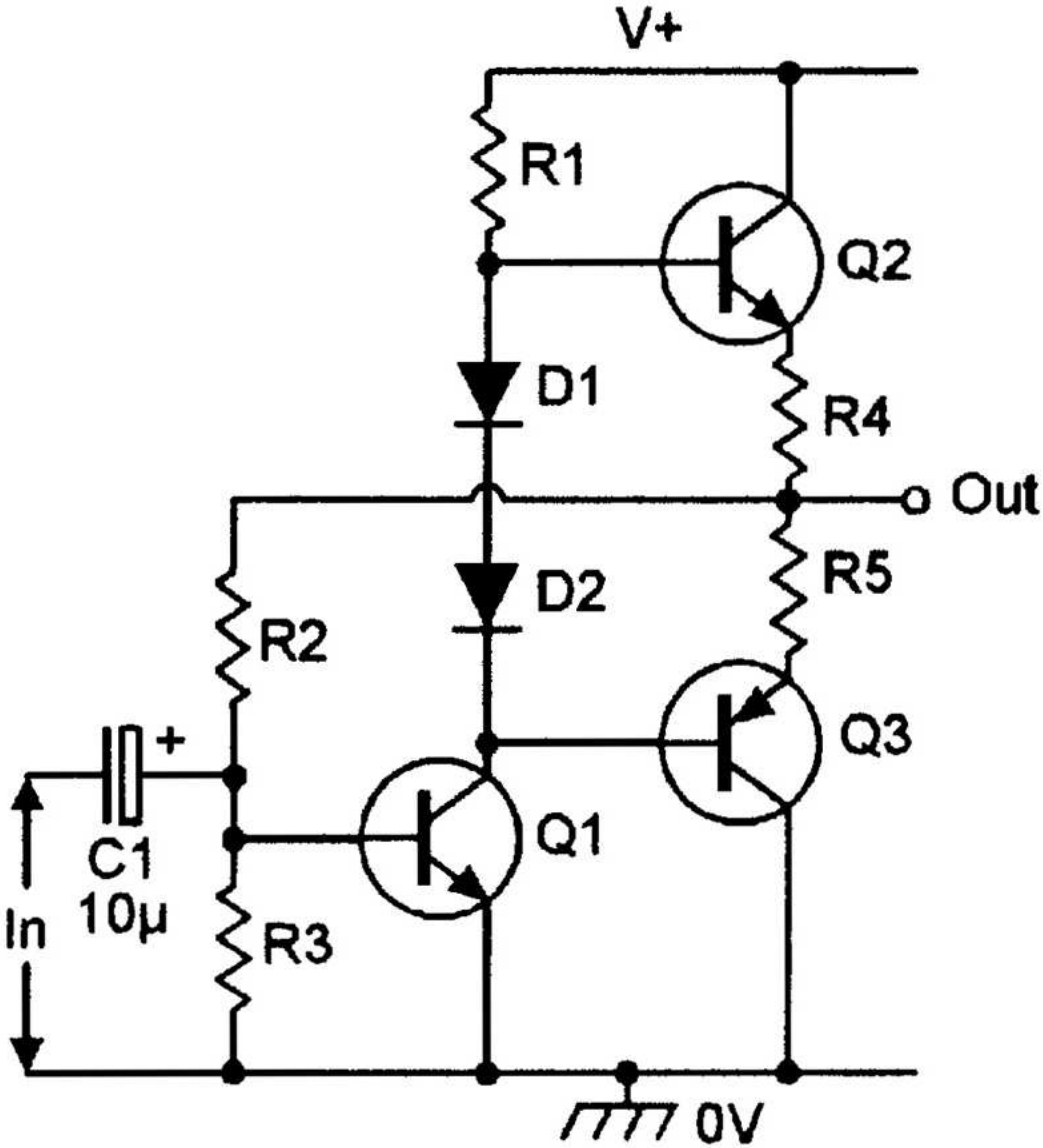


FIGURE 7. Complementary amplifier with driver and auto-bias.

The input impedance of the basic **Figure 5** circuit equals the product of the loudspeaker load impedance and the hfe of Q1 or Q2. An obvious circuit improvement is to replace the individual Q1 and Q2 transistors with high-gain pairs of transistors, to increase the circuit's input impedance, and enable it to be used with a driver with a high-value collector load. **Figures 8** to **10** show three alternative ways of modifying the **Figure 7** circuit in this way.

In **Figure 8**, Q2-Q3 are wired as a Darlington NPN pair, and Q3-Q4 as a Darlington PNP pair; note that four base-emitter junctions exist between Q2 base and Q4 base, so this output circuit must be biased via a chain of four silicon diodes.

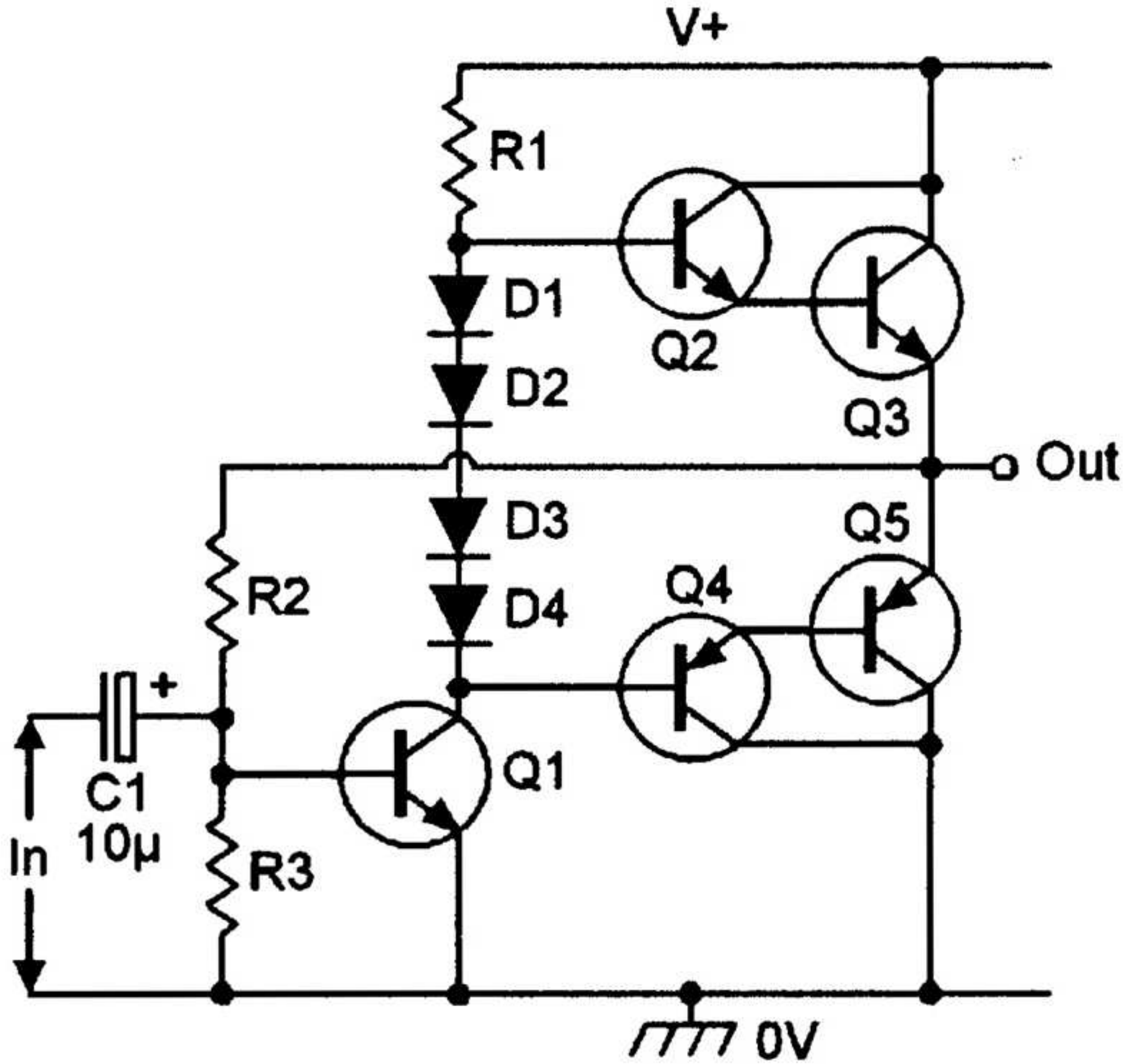


FIGURE 8. Amplifier with Darlington output stages.

In **Figure 9**, Q2-Q3 are wired as a Darlington NPN pair, but Q3-Q4 are wired as a complementary pair of common-emitter amplifiers that operate with 100% negative feedback and provide unity voltage gain and a very high input impedance. This design is known as a 'quasi-complementary' output stage, and is probably the most popular of all class-AB amplifier configurations; it calls for the use of three biasing diodes.

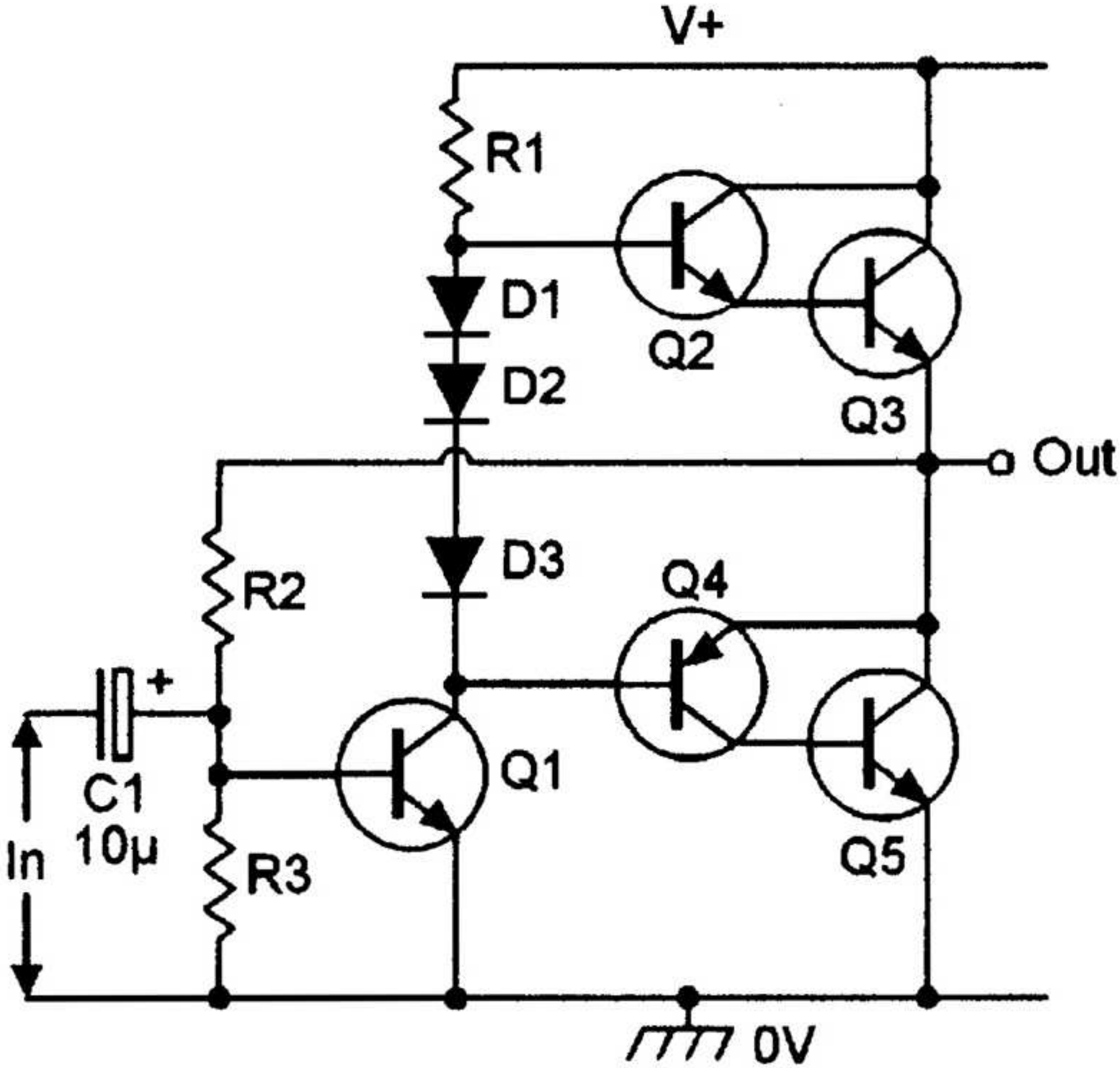


FIGURE 9. Amplifier with quasi-complementary output stages.

In **Figure 10**, both Q2-Q3 and Q4-Q5 are wired as complementary pairs of unity-gain common-emitter amplifiers with 100% negative feedback; they are mirror images of each other, and form a complementary output stage that needs only two biasing diodes.

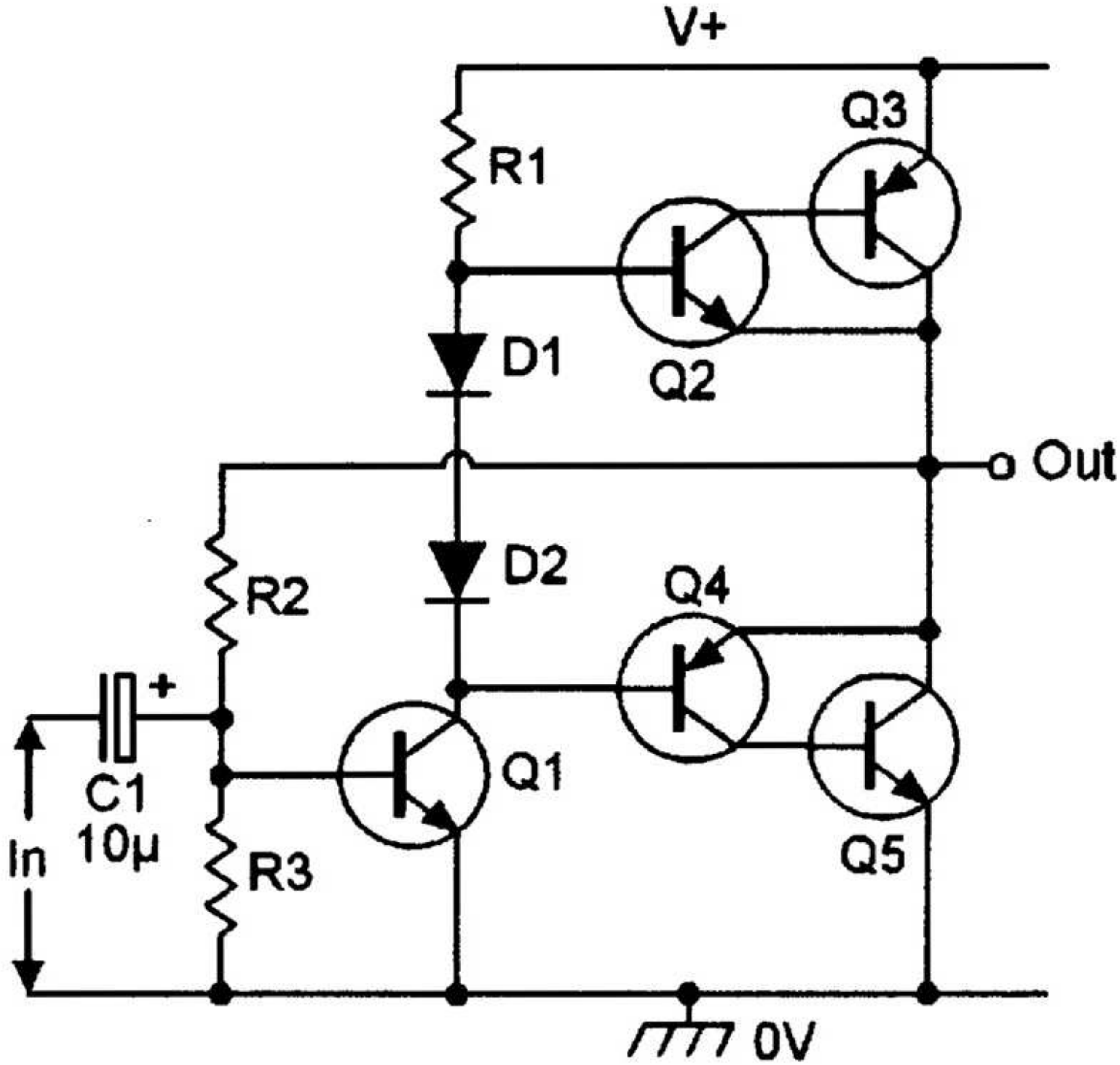


FIGURE 10. Amplifier with complementary output stages.

The circuits of **Figures 7** to **10** all call for the use of a chain of silicon biasing diodes. If desired, each of these chains can be replaced by a single transistor and two resistors, wired in the 'amplified diode' configuration described in Part 2 of this series and repeated here, in very basic form, in **Figure 11**.

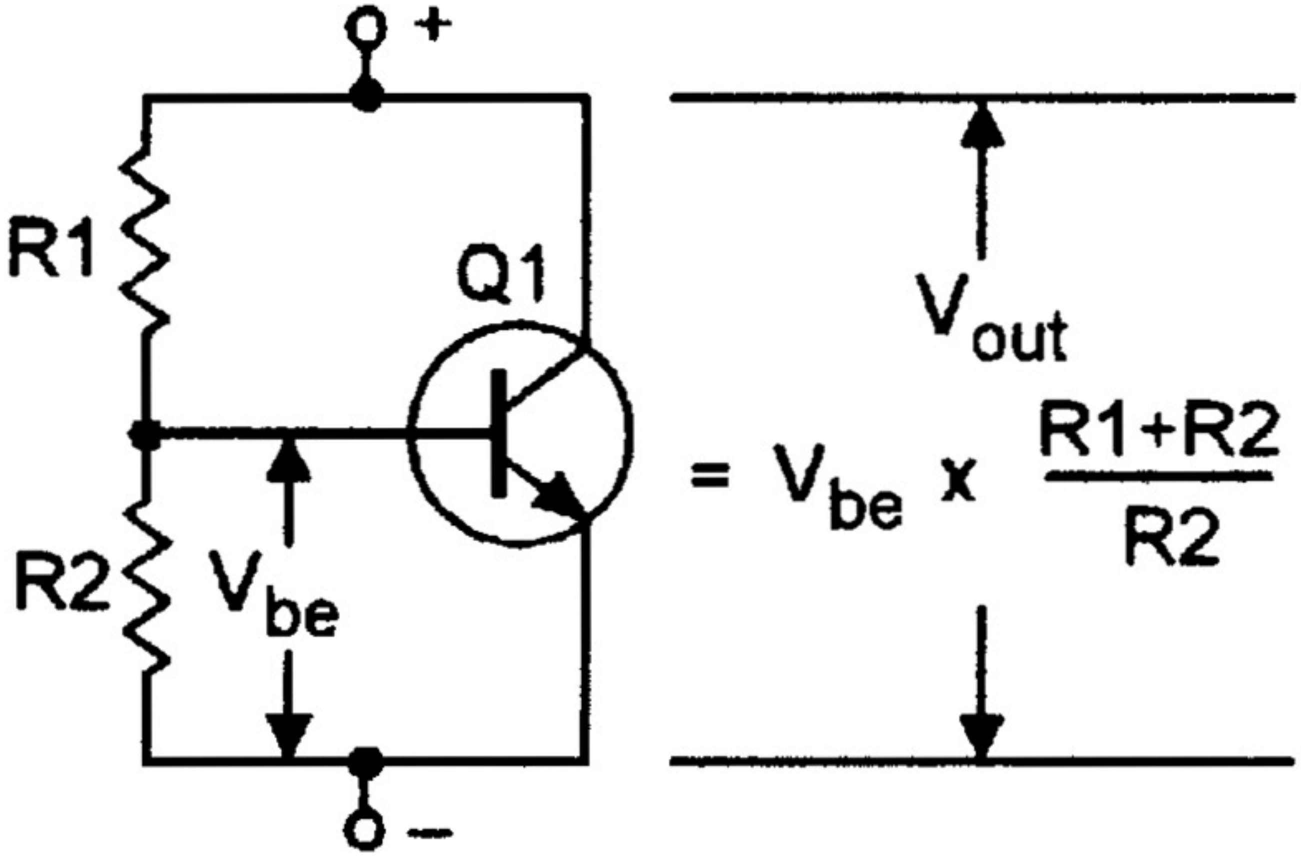


FIGURE 11. Fixed-gain amplified diode circuit.

Thus, if R1 is shorted out, the circuit acts like a single base-emitter junction diode, and if R1 is not shorted out, it acts like (R1+R2)/R2 series-wired diodes. **Figure 12** shows the circuit modified so that it acts as a fully adjustable amplified silicon diode, with an output variable from 1 to 5.7 base-emitter junction voltages.

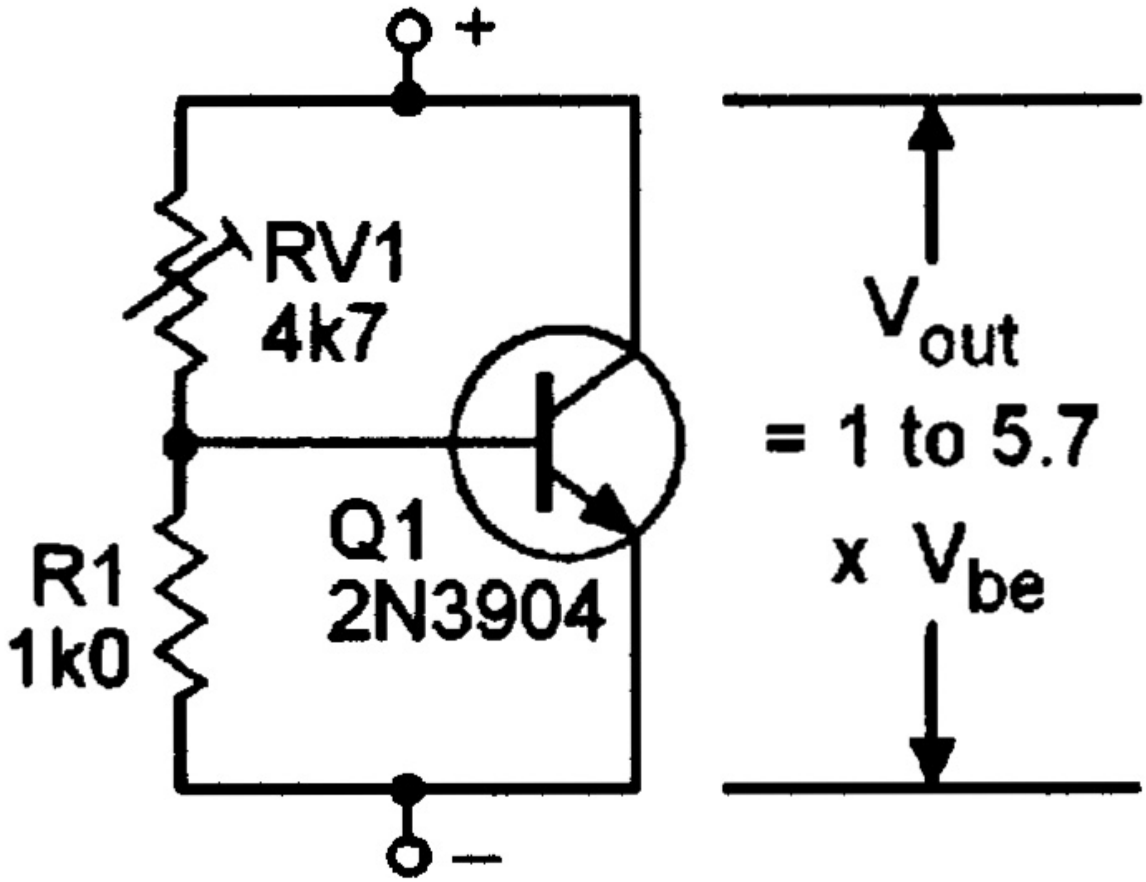


FIGURE 12. Adjustable amplified diode circuit.

In **Figure 13**, the Q1 collector load comprises R1 and R2 in series, and the circuit's output signal (which also appears across SPKR), is fed back to the R1-R2 junction via C2, thus bootstrapping R2's value so that its AC impedance is boosted by (typically) a factor of about 20, and the circuit's voltage gain is boosted by a similar amount.

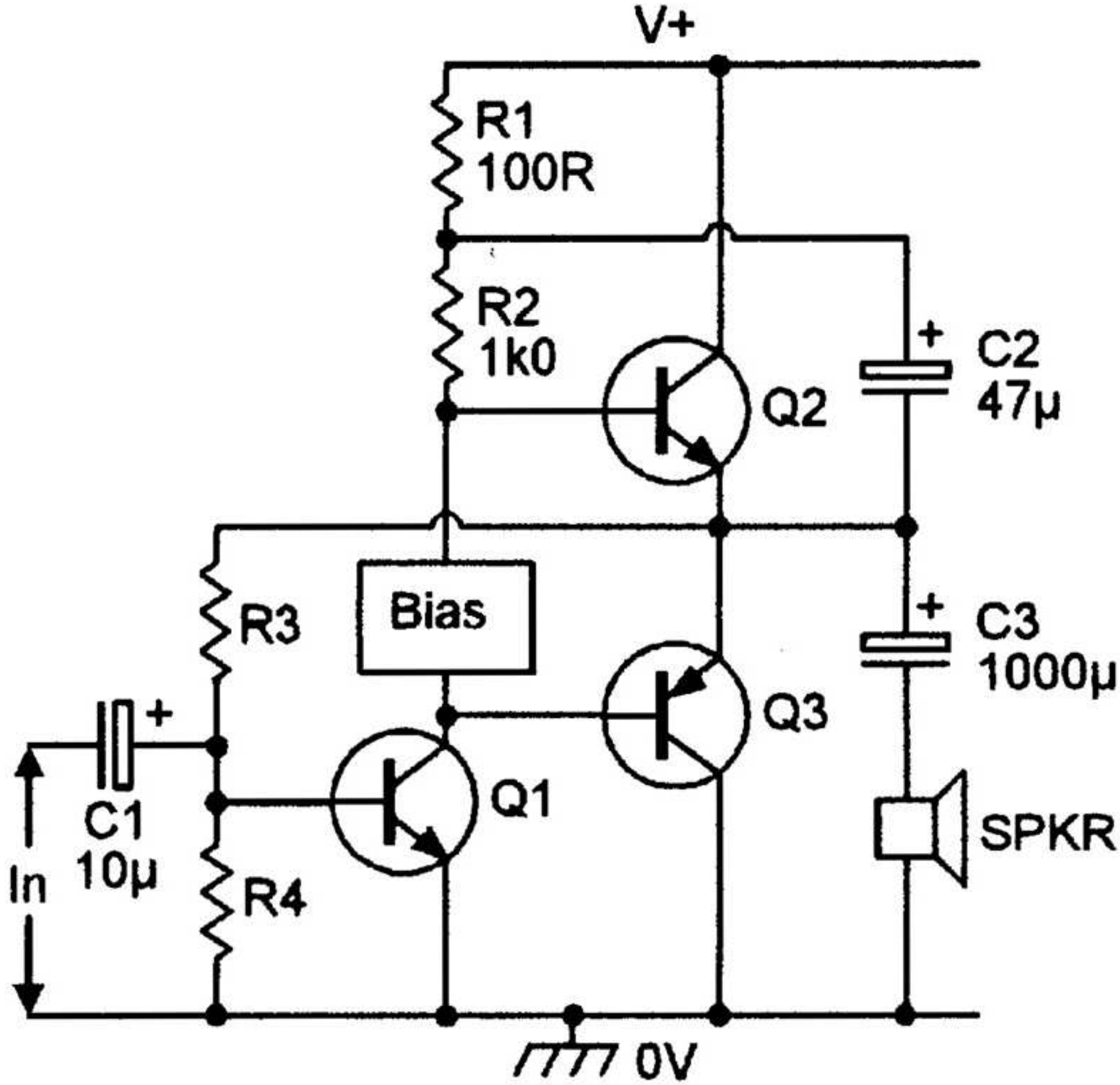


FIGURE 13. Amplifier with bootstrapped driver stage.

Another useful modification that can be made to the basic **Figure 7** circuit is to add bootstrapping to its R1 collector load, to boost its effective impedance and thus raise the circuit's overall voltage gain (the 'bootstrapping' technique was also described in Part 2 of this series). **Figures 13** and **14** show examples of bootstrapped class-AB power amplifier circuits.

Figure 14 shows a version of the circuit that saves two components; in this case, the SPKR forms part of Q1's collector load, and directly bootstraps R1.

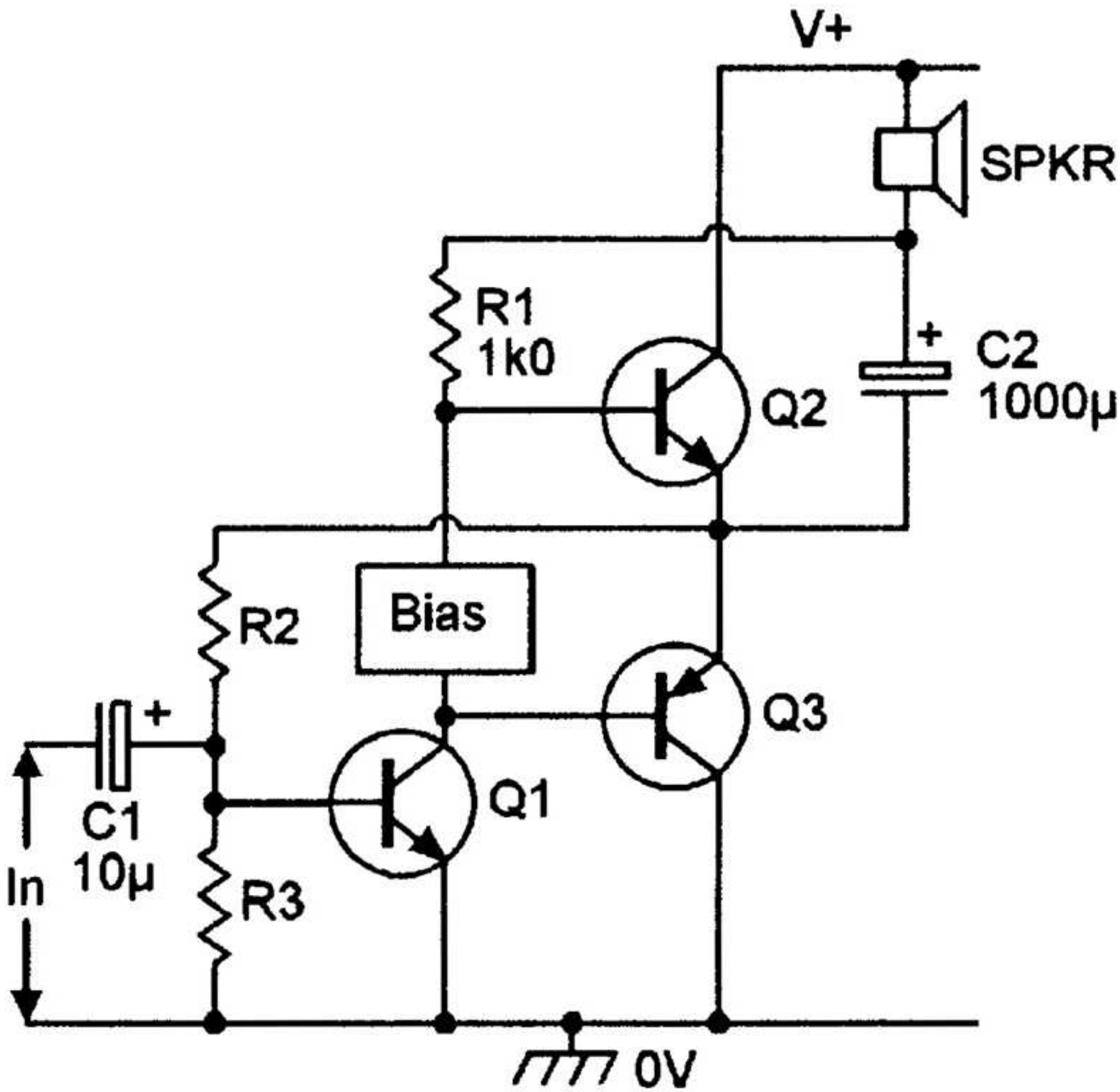


FIGURE 14. Alternative amplifier with bootstrapped driver stage.

PRACTICAL CLASS AB AMPLIFIERS

The easiest way to build a class-AB audio amplifier is to do so using one of the many readily-available audio ICs of this type. In some cases, however, particularly when making 'one off' projects, it may be cheaper or more convenient to use a discrete transistor design, such as one of those shown in **Figures 15** or **16**.

Figure 15 shows a simple class-AB amplifier that can typically drive 1W into a 3Ω speaker. Here, common-emitter amplifier Q1 uses collector load LS1-R1-D1-RV2, and drives the Q2-Q3 complementary emitter follower stage. The amplifier's output is fed (via C2) to the LS1-R1 junction, thus providing a low impedance drive to the loudspeaker and simultaneously bootstrapping the R1 value so that the circuit gives high voltage gain. The output is also fed back to Q1 base via R4, thus providing base bias via a negative feedback loop. In use, RV1 should be trimmed to give minimal audible cross-over distortion consistent with low quiescent current consumption (typically in the range 10 mA to 15 mA).

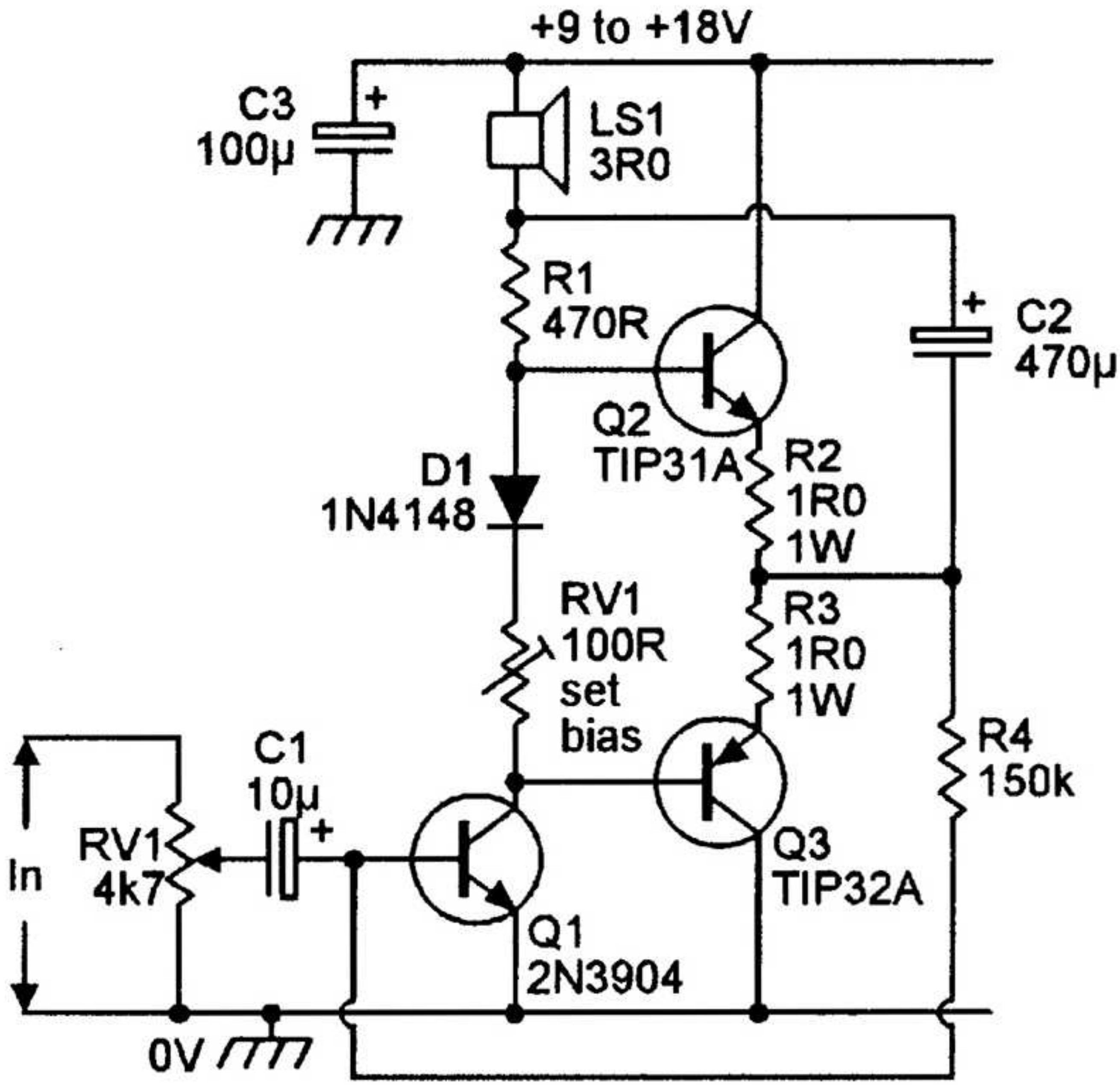


FIGURE 15. Simple 1 watt amplifier.

Figure 16 shows a rather more complex audio power amplifier that can deliver about 10W into an 8Ω load when powered from a 30 V supply.

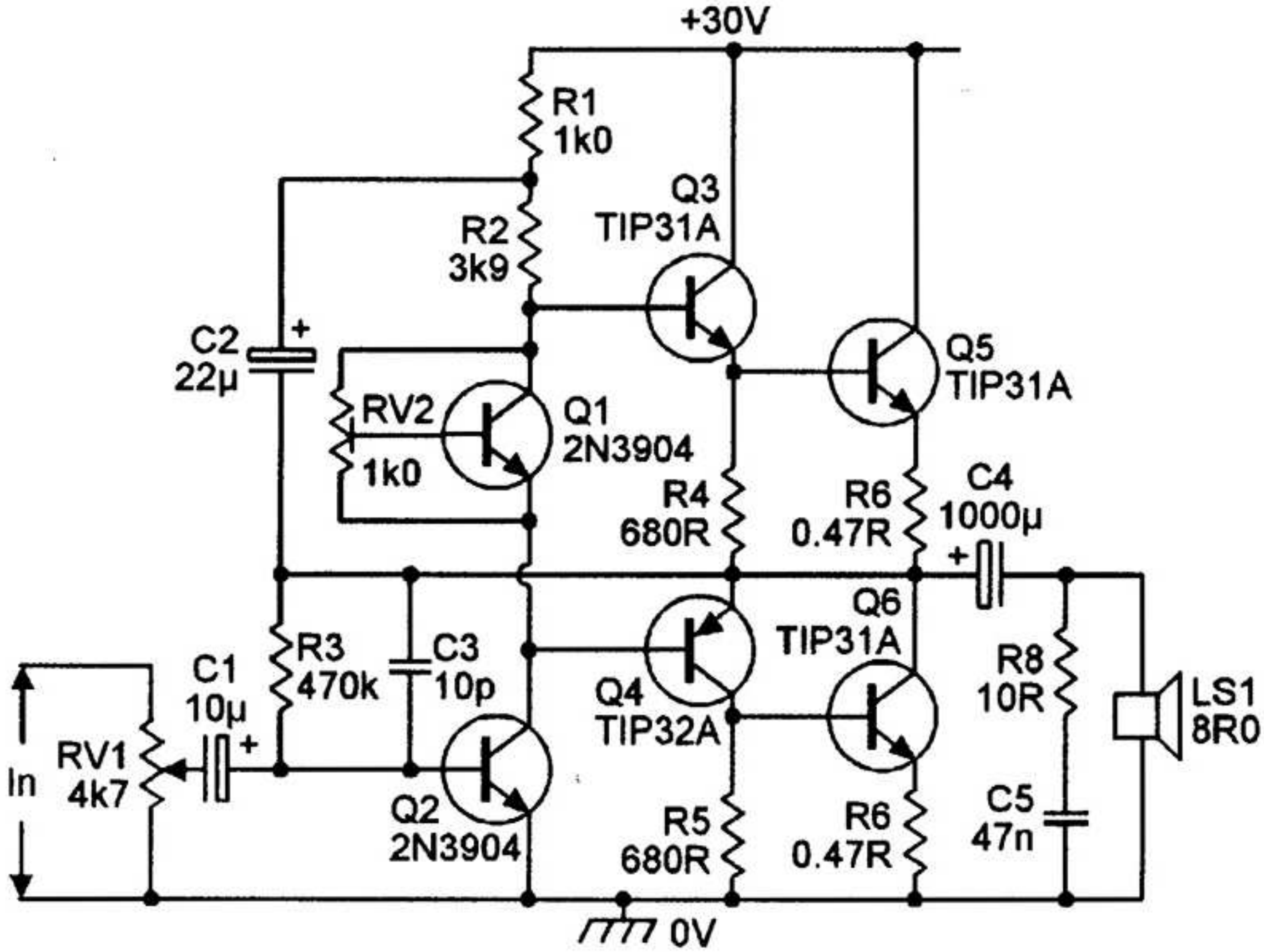


FIGURE 16. 10 watt audio amplifier.

This circuit uses high-gain quasi-complementary output stages (Q3 to Q6) and uses an adjustable amplifier diode (Q1) as an output biasing device. The Q2 common emitter amplifier stage has its main load resistor (R2) bootstrapped via C2, and is DC biased via R3, which should set the quiescent output voltage at about half-supply value (if not, alter the R3 value). The upper frequency response of the amplifier is restricted via C3, to enhance circuit stability, and C5-R8 are wired as a Zobel network across the output of the amplifier to further enhance the stability. In use, the amplifier should be initially set up in the way already described for the **Figure 15** circuit.

ALTERNATIVE DRIVERS

In the basic **Figure 7** circuit, the Q1 driver stage uses parallel DC and AC voltage feedback via potential divider network R2-R3. This circuit is simple and stable, but suffers from fairly low gain and very low input resistance, and can be used over only a very limited range of power supply voltages. A simple variation of this circuit is shown in **Figure 17**. It uses current feedback via R1-R2, thus enabling the circuit to be used over a wide range of supply voltages.

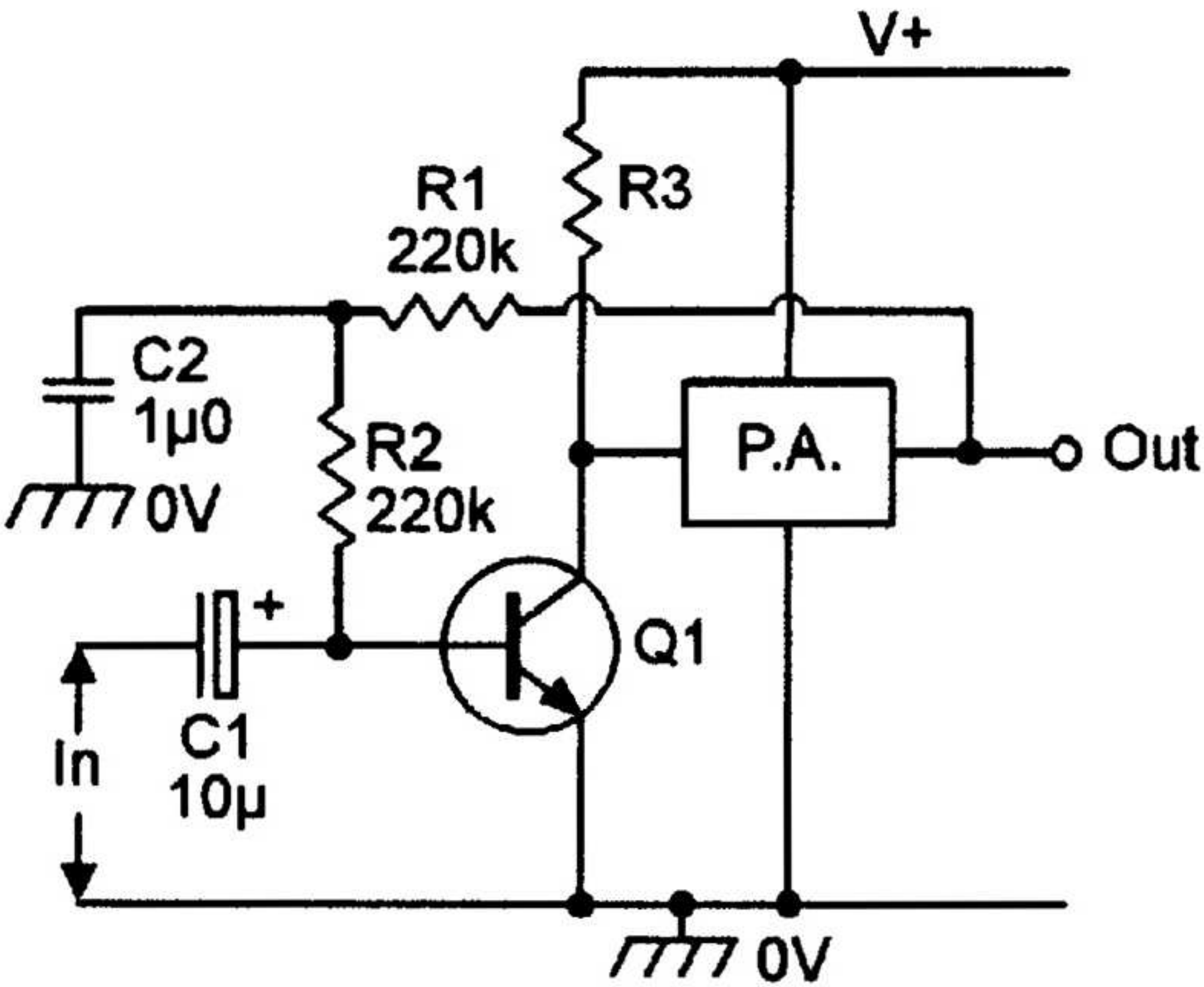


FIGURE 17. Driver stage with decoupled parallel DC feedback.

The feedback resistors can be AC-decoupled (as shown) via C2 to give increased gain and input impedance, at the expense of increased distortion. Q1 can be a Darlington type, if a very high input impedance is required.

Figure 18 shows an alternative configuration of driver stage. This design uses series DC and AC feedback, and gives greater gain and input impedance than the basic **Figure 7** circuit, but uses two transistors of opposite polarities.

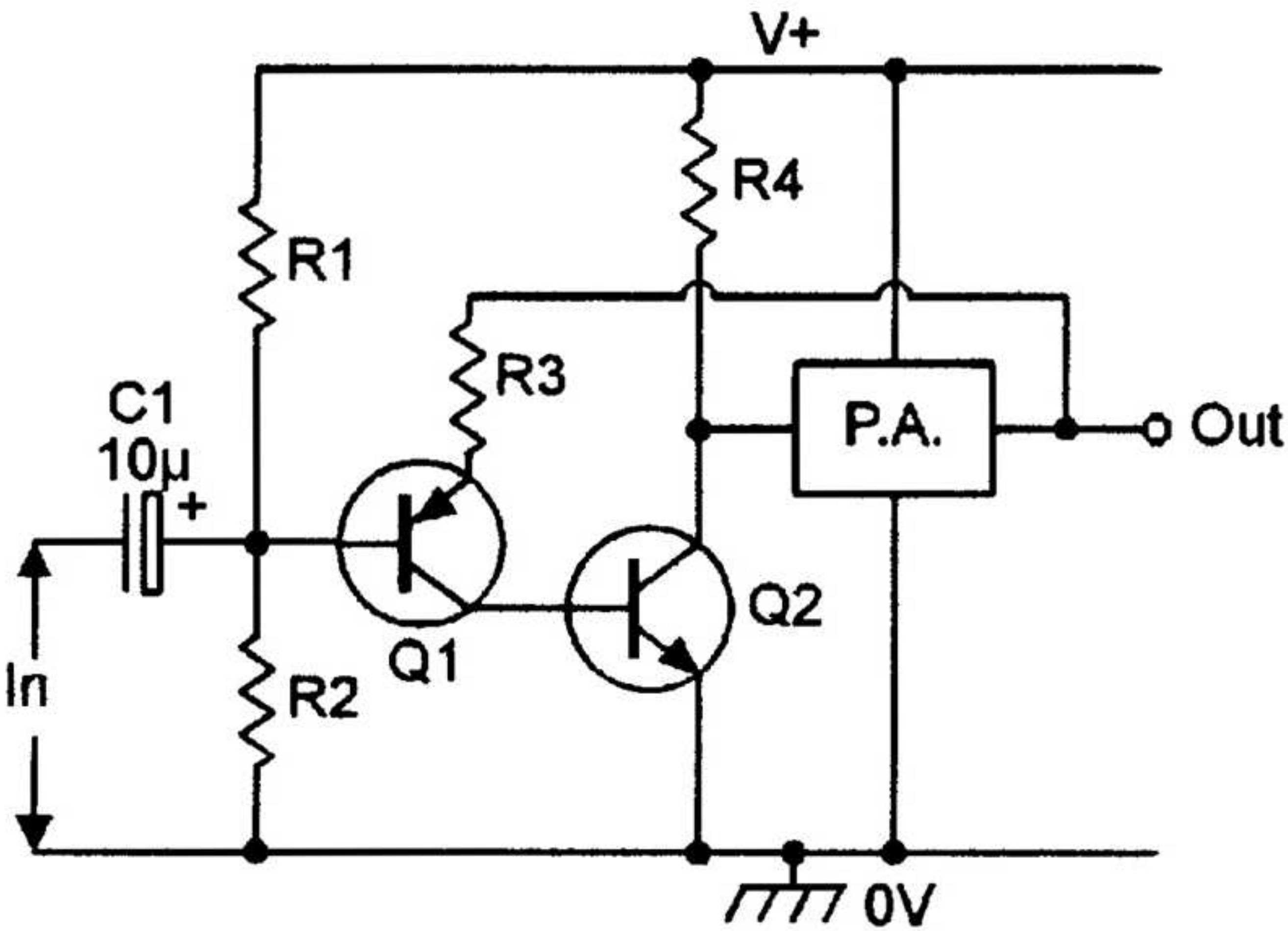


FIGURE 18. Driver stage with series DC feedback.

Finally, to complete this look at audio power amplifiers, **Figure 19** shows a circuit that has direct-coupled ground-referenced inputs and outputs, and uses split power supplies. It has a long-tailed pair input stage, and the input and output both center on zero volts if R1 and R4 have equal values. The circuit can be used with a single ended power supply by grounding one supply line and using AC coupling of the input and output signals. This basic circuit forms the basis of many IC power amplifier designs.

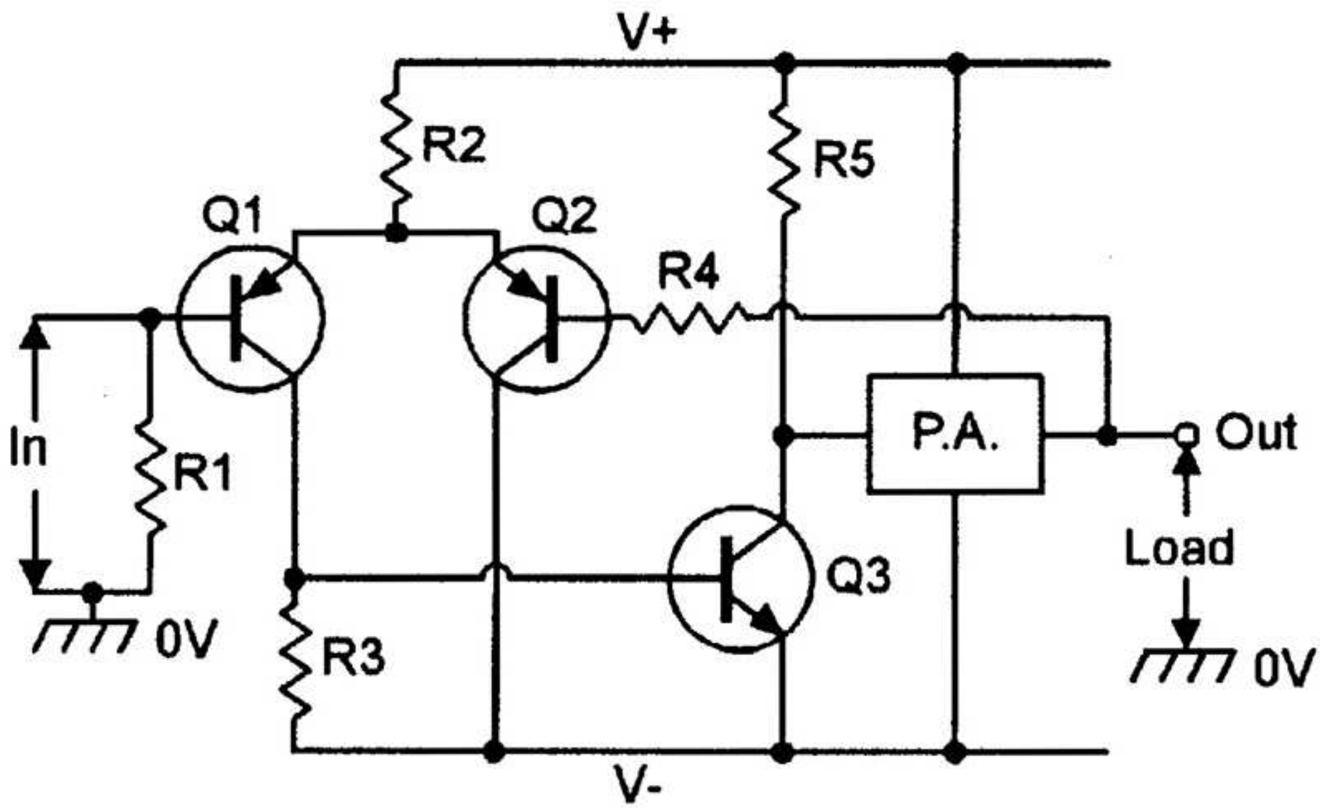


FIGURE 19. Driver stage with long-tailed pair input.

SCRATCH/RUMBLE FILTERS

A common annoyance when playing old records (discs) through audio power amplifiers is that of scratch and/or rumble sounds. The scratch noises are mainly high-frequency (greater than 10 kHz) sounds picked up from the disc surface, and the rumbles are low-frequency (less than 50 Hz) sounds that are mostly caused by slow variations in motor-drive speed.

Each of these noises can be greatly reduced or eliminated by coupling the player's audio signals into the audio power amplifier input via a filter that rejects the troublesome parts of the audio spectrum. **Figures 20** and **21** show suitable circuits.

The high-pass rumble filter of **Figure 20** gives unity voltage gain to signals above 50 Hz, but gives 12dB per octave rejection to those below this value, i.e., it gives 40dB of attenuation at 5 Hz, etc. Emitter-follower Q1 is biased at half-supply volts from the R1-R2-C3 low-impedance point, but has negative feedback applied via the R3-C2-C1-R4 filter network. The circuit's frequency turn-over point can be altered by changing the C1-C2 values (which must be equal); thus, if the C1-C2 values are halved (to 110nF), the turn-over frequency doubles (to 100 Hz), etc.

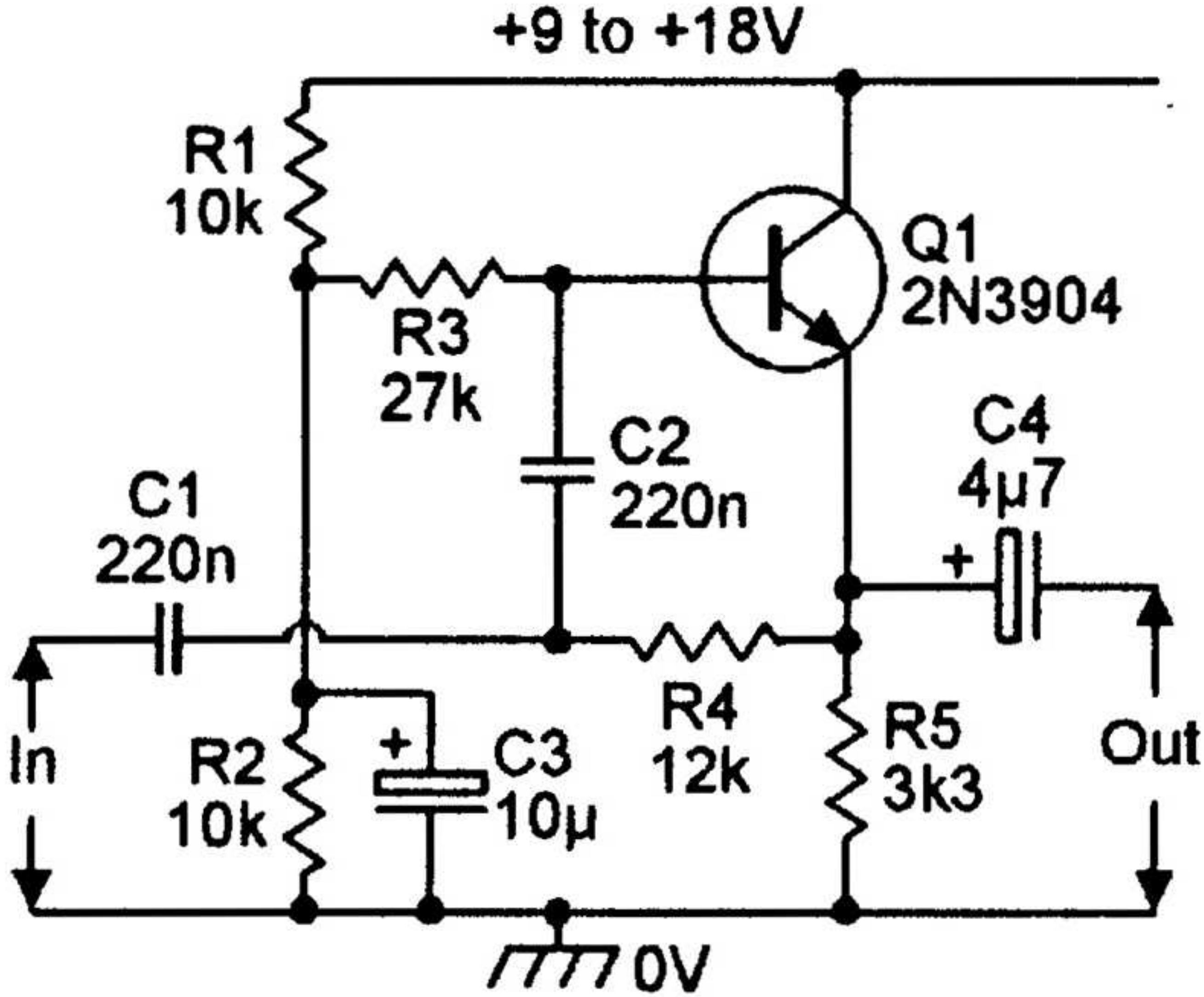


FIGURE 20. 50 Hz rumble or high-pass filter.

The low-pass scratch filter of **Figure 21** gives unity voltage gain to signals below 10 kHz, but gives 12dB per octave rejection to those above this value.

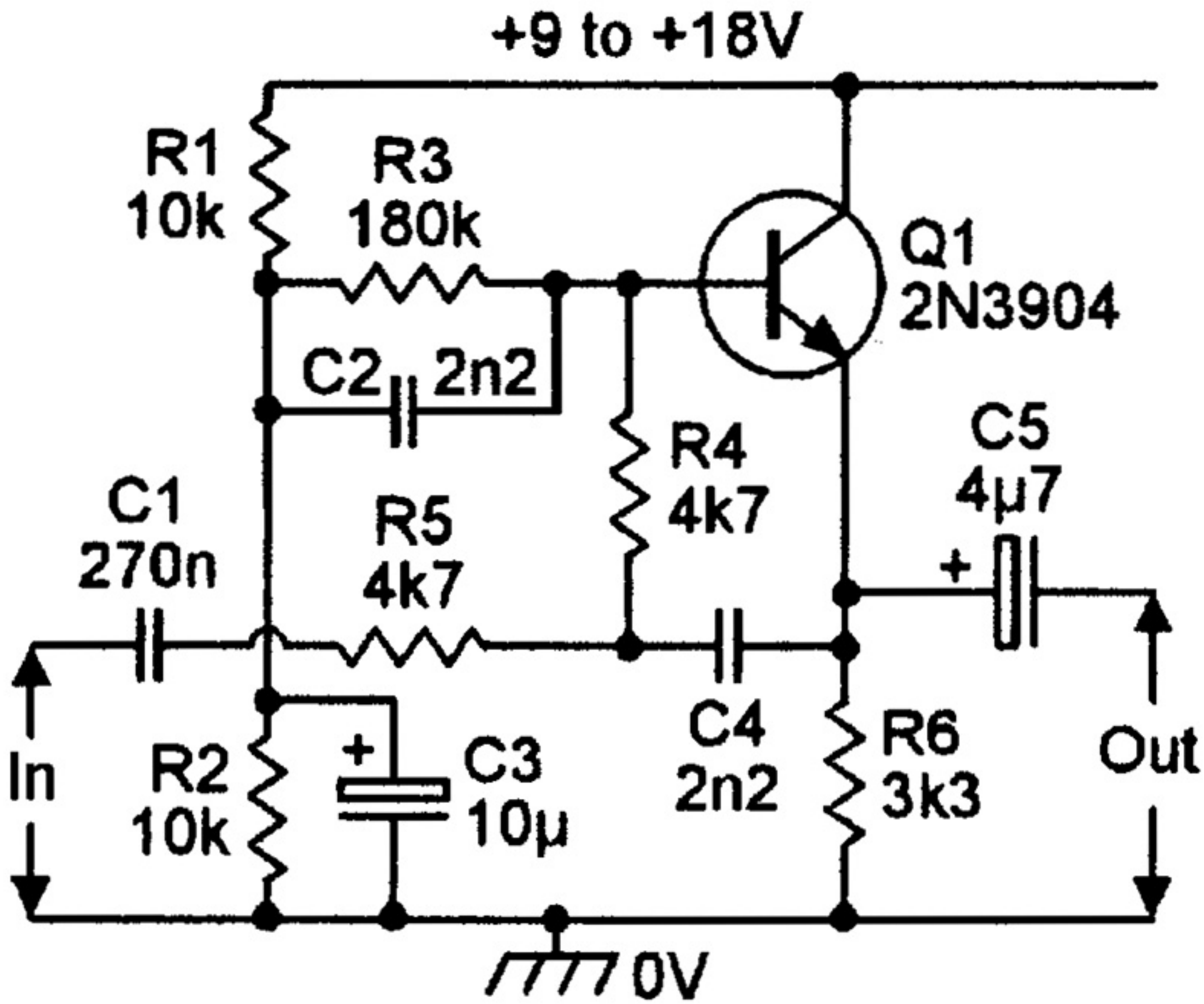


FIGURE 21. 10 kHz scratch or low-pass filter.

This circuit is similar to that of **Figure 20**, except that the positions of the resistors and capacitors are transposed in the C2-R4-C4-R5 filter network. The circuit's turn-over frequency can be altered by changing the C2-C4 values, e.g., values of 3.3nF give a frequency of 7.5 kHz.

The **Figure 20** and **21** circuits can be combined to make a composite scratch and rumble filter, by connecting the output of the high-pass filter to the input of the low-pass filter; if desired, the filters can be provided with bypass switches, enabling them to be easily switched in and out of circuit, by using the connections of **Figure 22**.

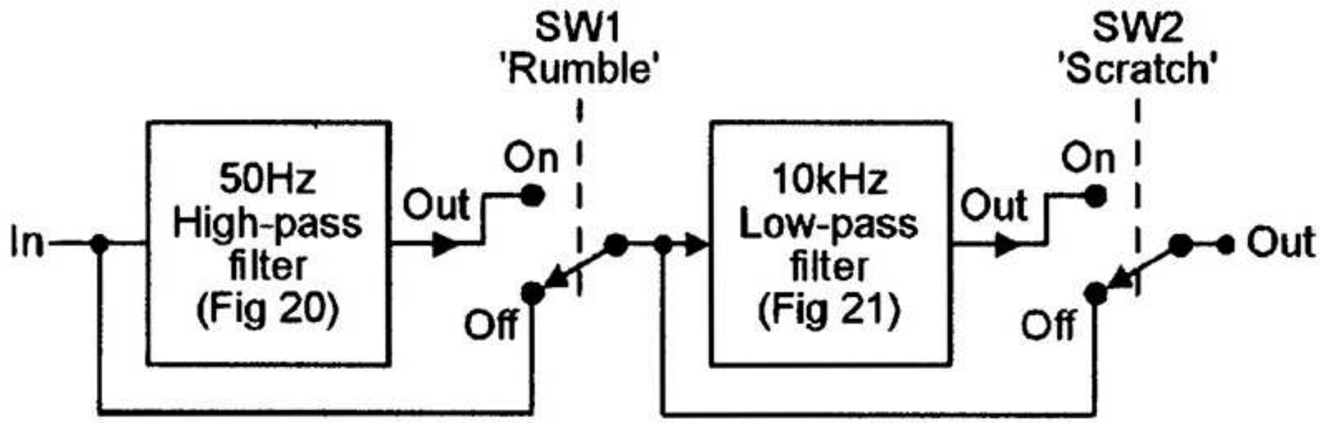


FIGURE 22. Complete scratch/rumble filter, with switching.

Note that if the **Figure 20** and **21** designs are to be built as a single unit, a few components can be saved by making the R1-R2-C3 biasing network common to both circuits.

The next, (and final) installment will describe a miscellaneous collection of useful transistor circuits and gadgets. **NV**

Bipolar Transistor Cookbook — Part 8

The opening piece of this eight-part series described basic transistor principles and configurations (**Part 1**); subsequent articles went on to describe a wide variety of practical transistor circuits ranging from common-collector amplifiers (**Part 2**), common-emitter and common-base amplifiers (**Part 3**), and small-signal audio amplifiers (**Part 4**), to various practical oscillator (**Part 5**), multivibrator waveform generator (**Part 6**), and audio power amplifier (**Part 7**) circuits. This final episode rounds off the "Transistor Cookbook" subject by presenting a miscellaneous collection of practical and useful transistor circuits and gadgets.

A NOISE LIMITER CIRCUIT

Unwanted electronic "noise" can be a great nuisance; when listening to very weak broadcast signals, for example, peaks of background noise often completely swamp the broadcast signal, making it unintelligible. This problem can often be overcome by using the noise limiter circuit in **Figure 1**. Here, the signal-plus-noise waveform is fed to amplifier Q1 via RV1. Q1 amplifies both waveforms equally, but D1 and D2 automatically limit the peak-to-peak output swing of Q1 to about 1.2 V. Thus, if RV1 is adjusted so that the signal output is amplified to this peak level, the noise peaks will not be able to greatly exceed the signal output, and intelligibility is greatly improved.

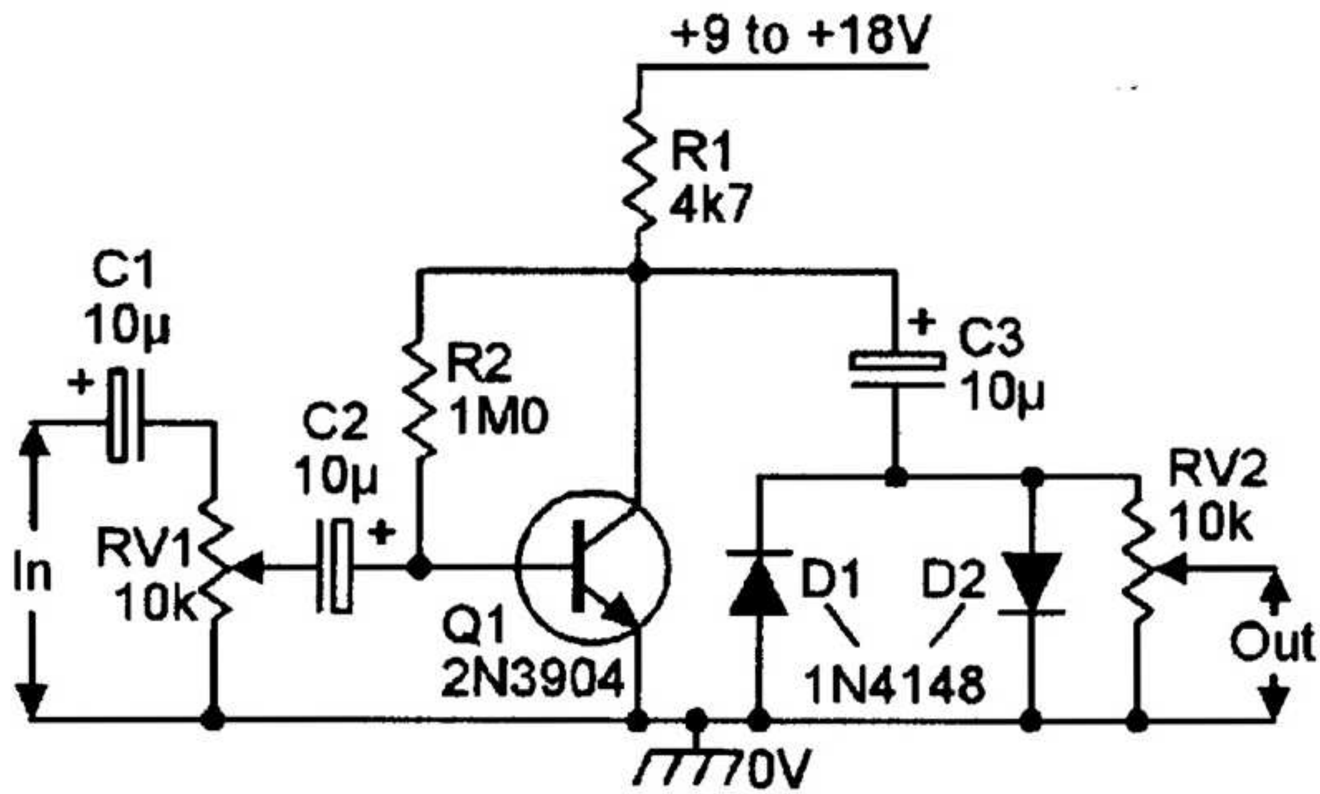


FIGURE 1. Noise limiter.

ASTABLE MULTIVIBRATOR CIRCUITS

The astable multivibrator circuit has many practical uses. It can be used to generate a non-symmetrical 800 Hz waveform that produces a monotone audio signal in the loudspeaker when S1 is closed (**Figure 2**). The circuit can be used as a Morse code practice oscillator by using a Morse key as S1; the tone frequency can be changed by altering the C1 and/or C2 values.

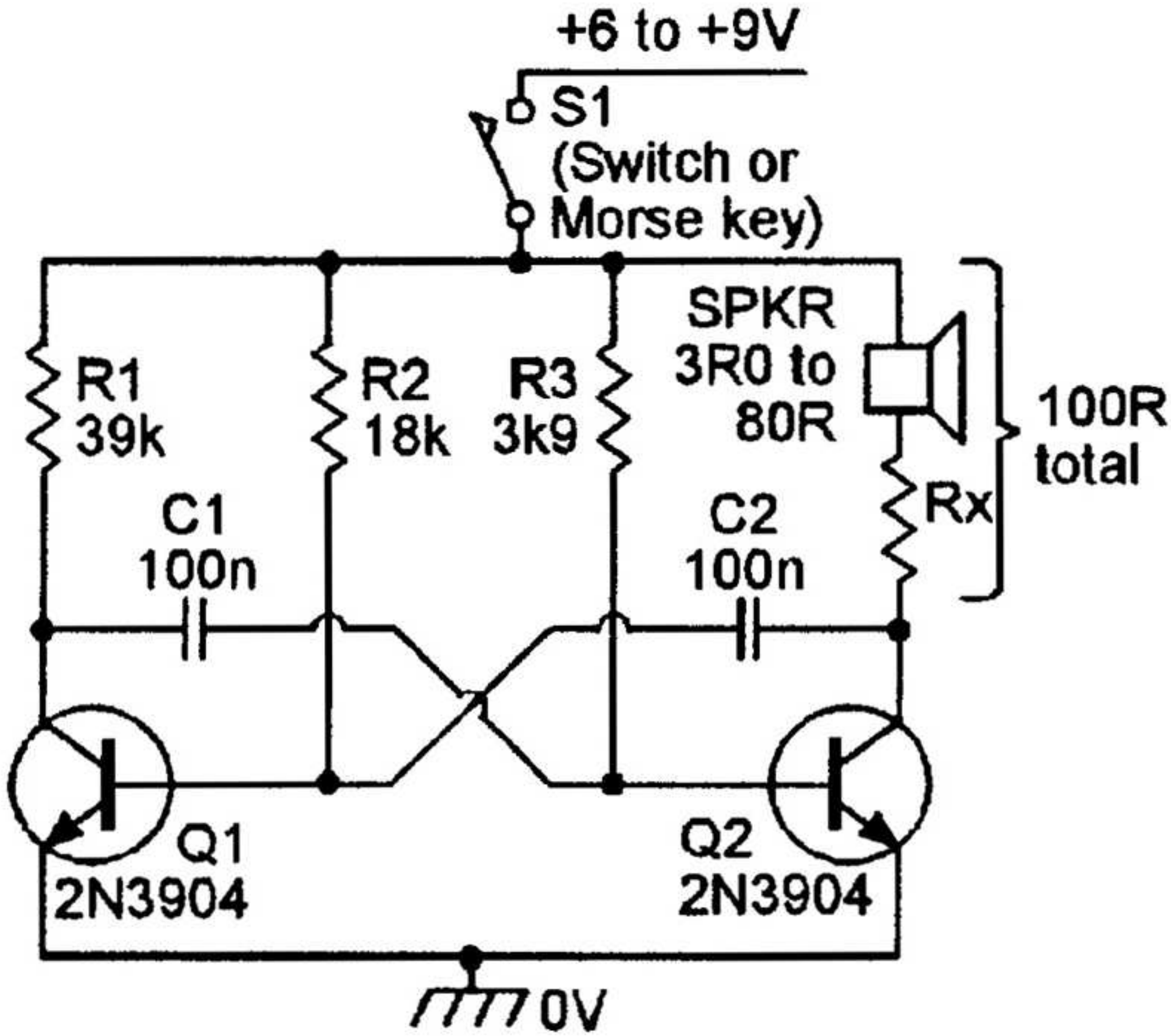


FIGURE 2. Morse code practice oscillator.

Figure 3 shows an astable multivibrator used as the basis of a "signal injector-tracer" item of test gear. When SW1 is in INJECT position 1, Q1 and Q2 are configured as a 1 kHz astable, and feed a good square wave into the probe terminal via R1-C1. This waveform is rich in harmonics, so if it is injected into any AF or RF stage of an AM radio, it produces an audible output via the radio's loudspeaker, unless one of the radio's stages is faulty. By choosing a suitable injection point, the injector can be used to trouble-shoot a defective radio.

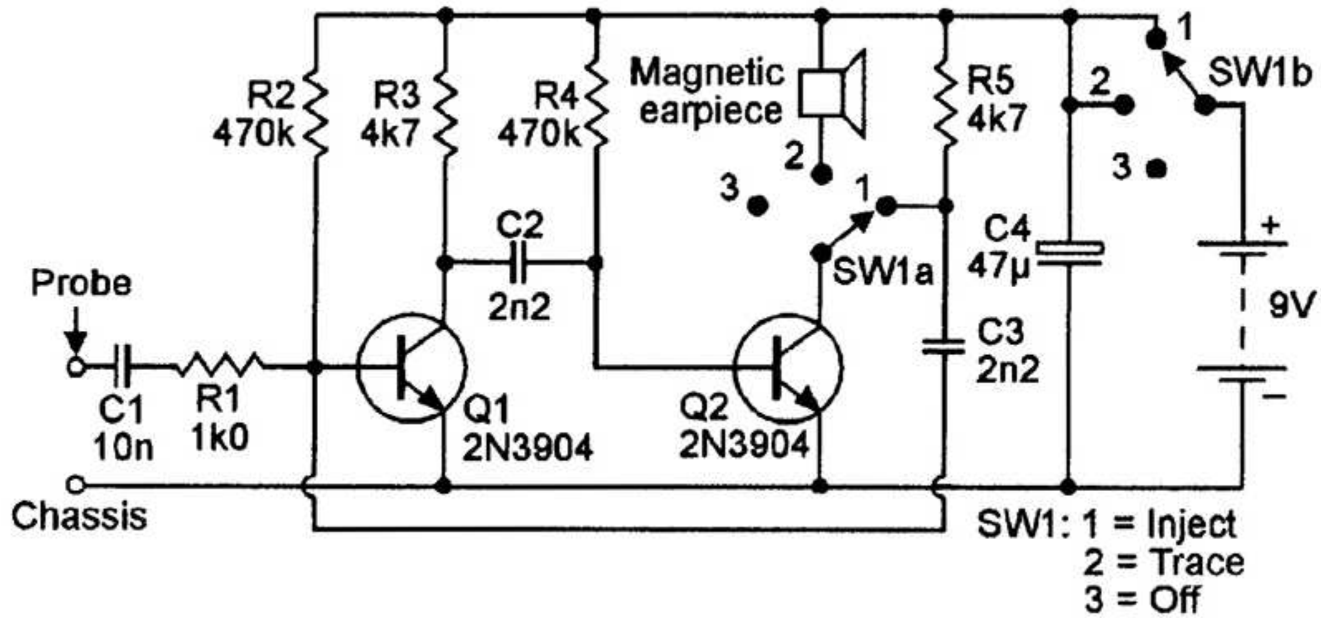


FIGURE 3. Signal injector-tracer.

When SW1 is switched to TRACE position 2, the **Figure 3** circuit is configured as a cascaded pair of common-emitter amplifiers, with the probe input feeding to Q1 base, and Q2 output feeding into an earpiece or head-set. Any weak audio signals fed to the probe are directly amplified and heard in the earpiece, and any amplitude-modulated RF signals fed to the probe are demodulated by the non-linear action of Q1. The resulting audio signals are then amplified and heard in the earpiece. By connecting the probe to suitable points in a radio, the tracer can thus be used to trouble-shoot a faulty radio, etc.

LIE DETECTOR

The lie detector of **Figure 4** is an experimenter's circuit, in which the victim is connected (via a pair of metal probes) into a Wheatstone bridge, formed by R1-RV1-Q1 and R3-R4; the 1 mA center-zero meter is used as a bridge-balance detector. In use, the victim makes firm contact with the probes and, once he/she has attained a relaxed state (in which the skin resistance reaches a stable value), RV1 is adjusted to set a null on the meter. The victim is then cross-questioned and, according to theory, the victim's skin resistance will then change, causing the bridge to go out of balance if he/she lies or shows any sign of emotional upset (embarrassment, etc.) when being questioned.

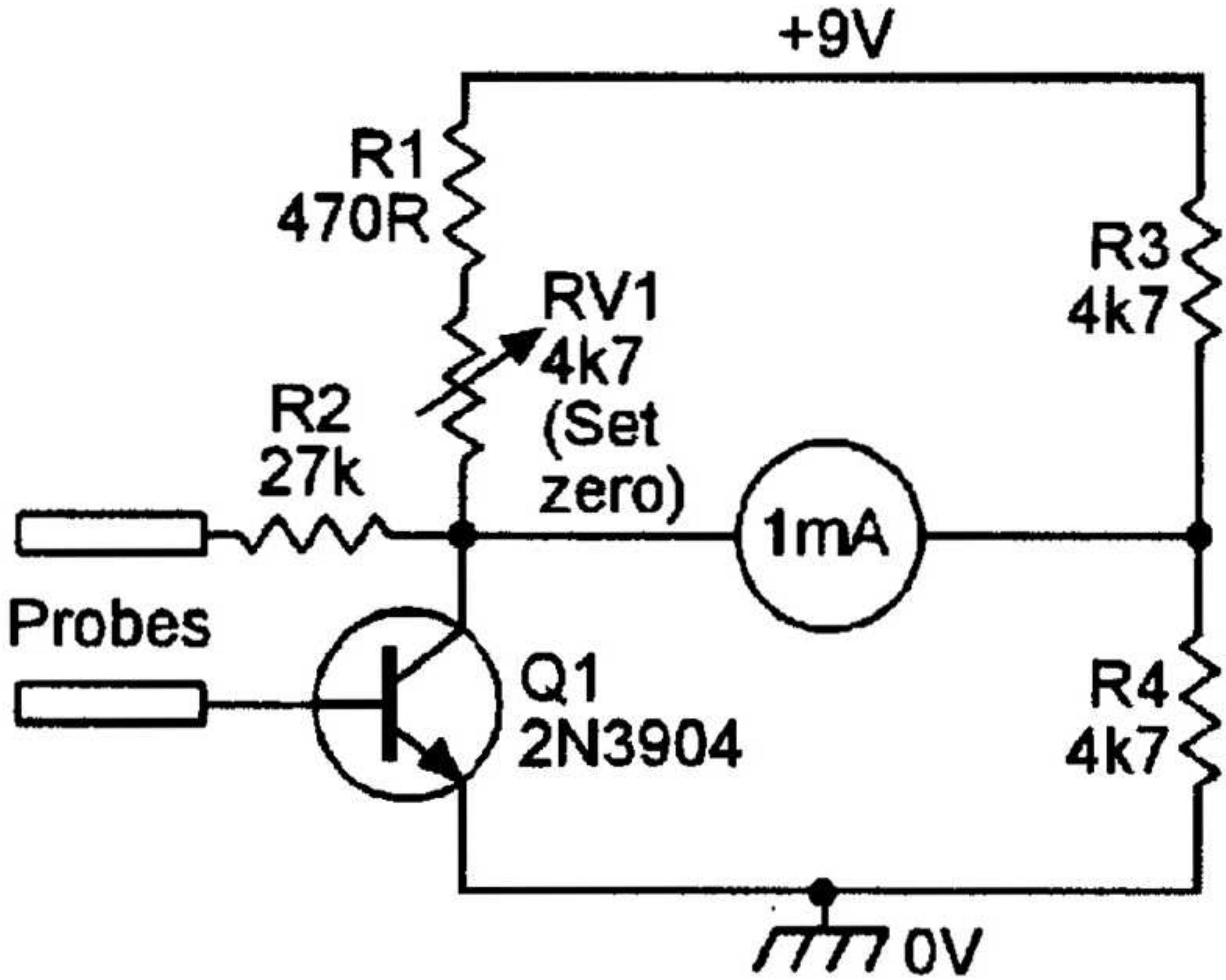


FIGURE 4. Simple lie detector.

CURRENT MIRRORS

A current mirror is a constant-current generator in which the output current magnitude is virtually identical to that of an independent input control current. This type of circuit is widely used in modern linear IC design. **Figure 5** shows a simple current mirror using ordinary npn transistors; Q1 and Q2 are a matched pair and share a common thermal environment.

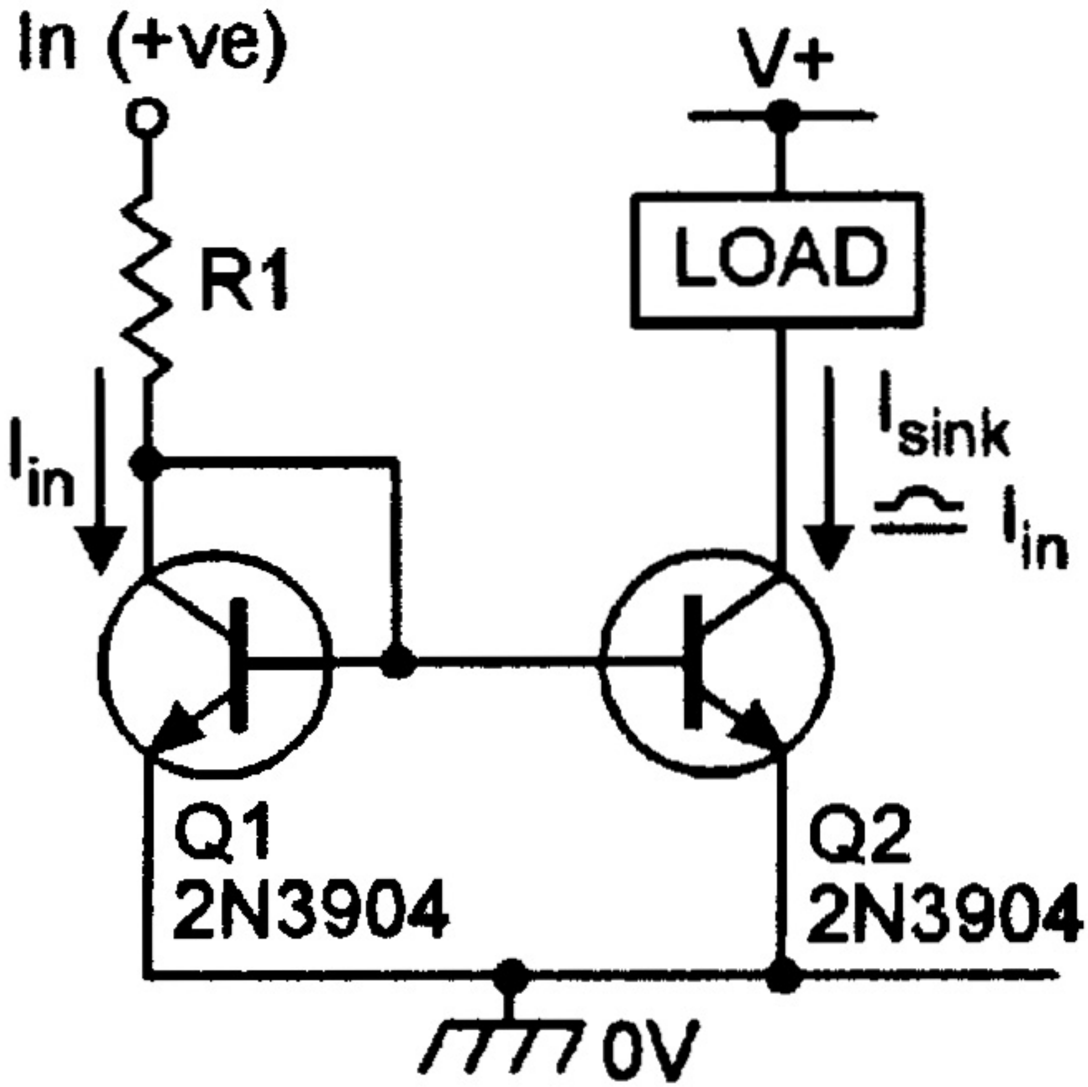


FIGURE 5. An npn current mirror.

When input current Iin is fed into diode-connected Q1, it generates a proportionate forward base-emitter voltage, which is applied directly to the base-emitter junction of matched transistor Q2, causing it to sink an almost identical (mirror) value of collector current, I_{sink}. Q2 thus acts as a constant current sink that is controlled by I_{in}, even at collector voltages as low as a few hundred millivolts.

Figure 6 shows a pnp version of the simple current mirror circuit. This works in the same basic way as already described, except that Q2's collector acts as a constant current source that has its amplitude controlled by Iin. Note that both of these circuits still work quite well as current-controlled, constant-current sinks or sources, even if Q1 and Q2 have badly matched characteristics, but in this case may not act as true current mirrors, since their I_{sink} and I_{in} values may be very different.

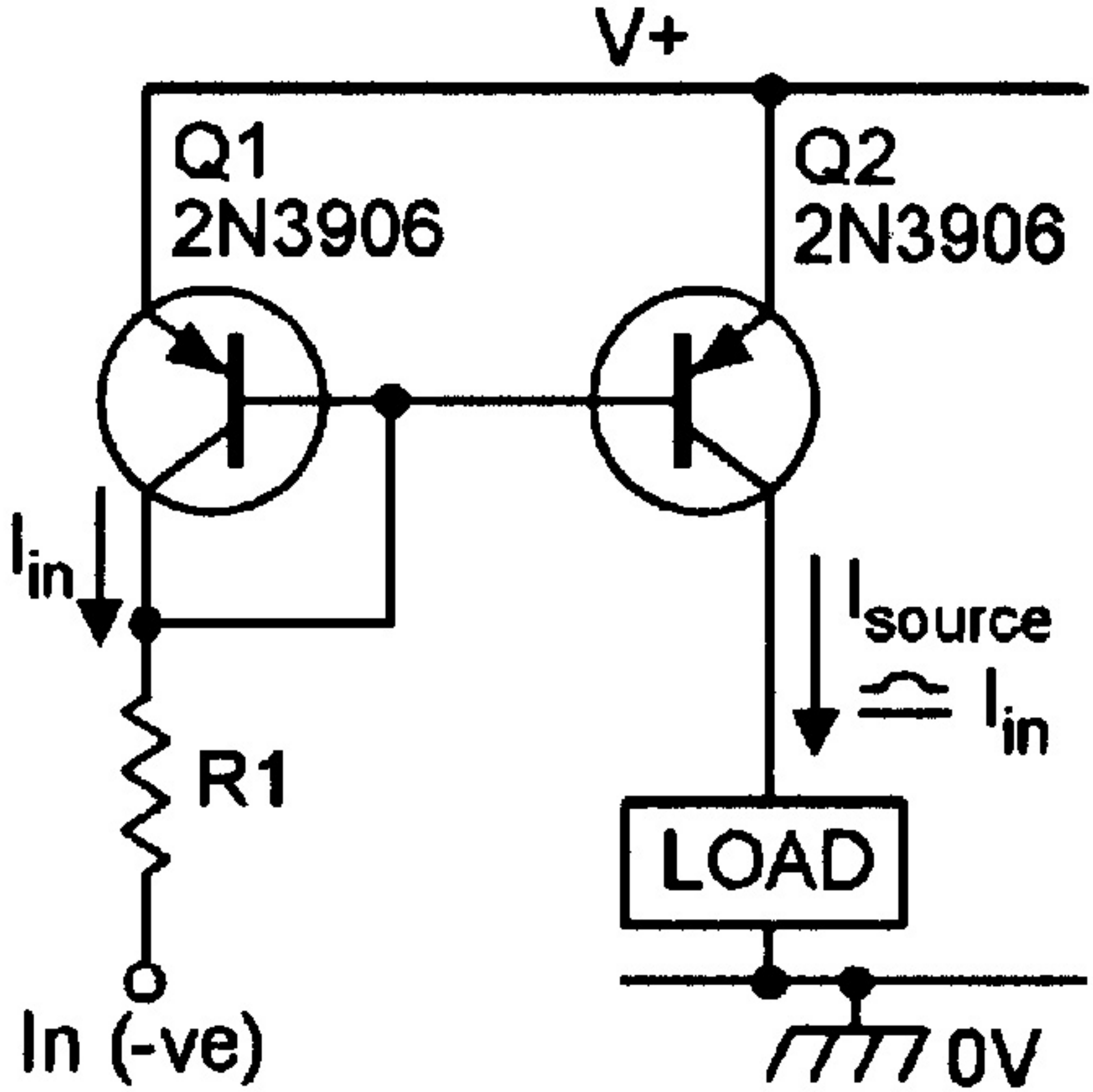


FIGURE 6. A pnp current mirror.

AN ADJUSTABLE ZENER

Figure 7 shows the circuit of an adjustable zener that can have its output voltage pre-set over the range 6.8 V to 21 V via RV1. The circuit action is such that a fixed reference voltage (equal to the sum of the zener and V_{be} values) is generated between Q1's base and ground (because of the value of zener voltage used) and has a near-zero temperature coefficient. The circuit's output voltage is equal to V_{ref} multiplied by (RV1 + R1)/R1, and is thus pre-settable via RV1. This circuit is used like an ordinary zener diode, with the R_S value chosen to set its operating current at a nominal value in the range 5 to 20 mA.

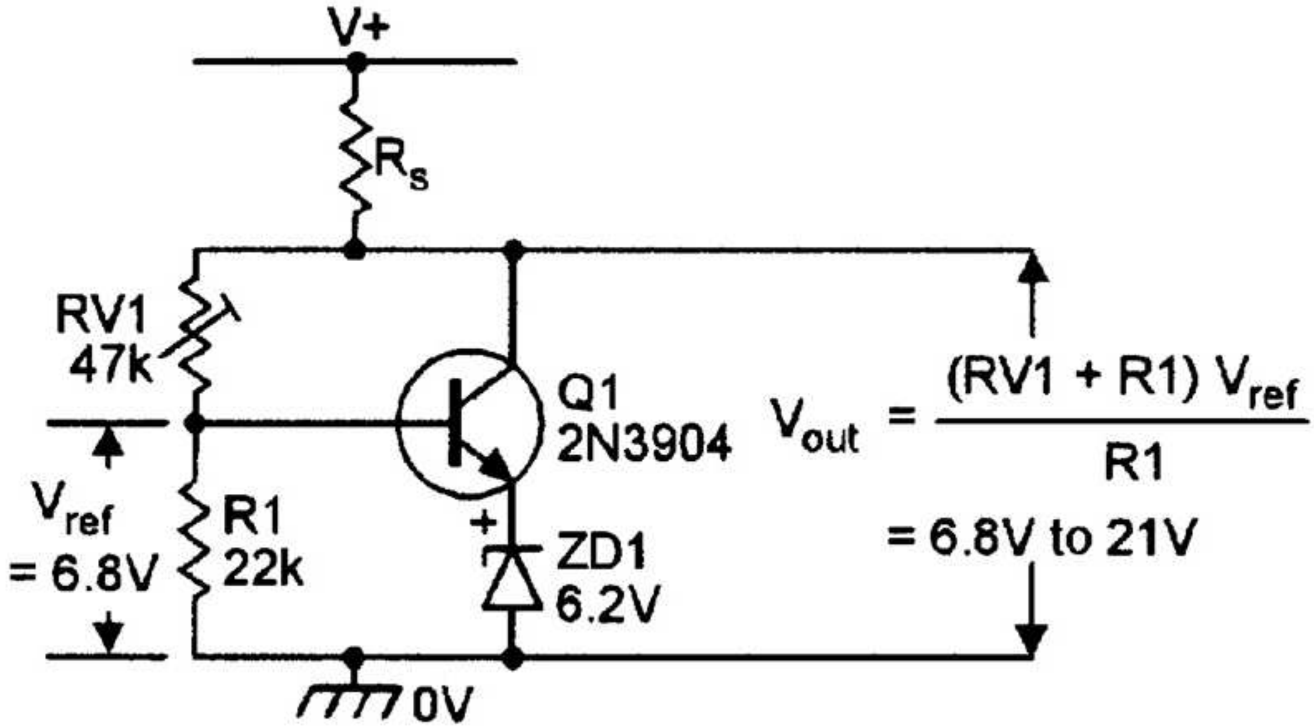


FIGURE 7. Adjustable zener.

L-C OSCILLATORS

L-C oscillators have many applications in test gear and gadgets, etc. **Figure 8** shows an L-C medium-wave (MW) signal generator or beat-frequency oscillator (BFO), with Q1 wired as a Hartley oscillator that uses a modified 465 kHz IF transformer as its collector load. The IF transformer's internal tuning capacitor is removed, and variable oscillator tuning is available via VC1, which enables the output frequency (on either fundamentals or harmonics) to be varied from well below 465 kHz to well above 1.7 MHz. Any MW radio will detect the oscillation frequency if placed near the circuit; if the unit is tuned to the radio's IF value, a beat note will be heard, enabling CW and SSB transmissions to be clearly heard.

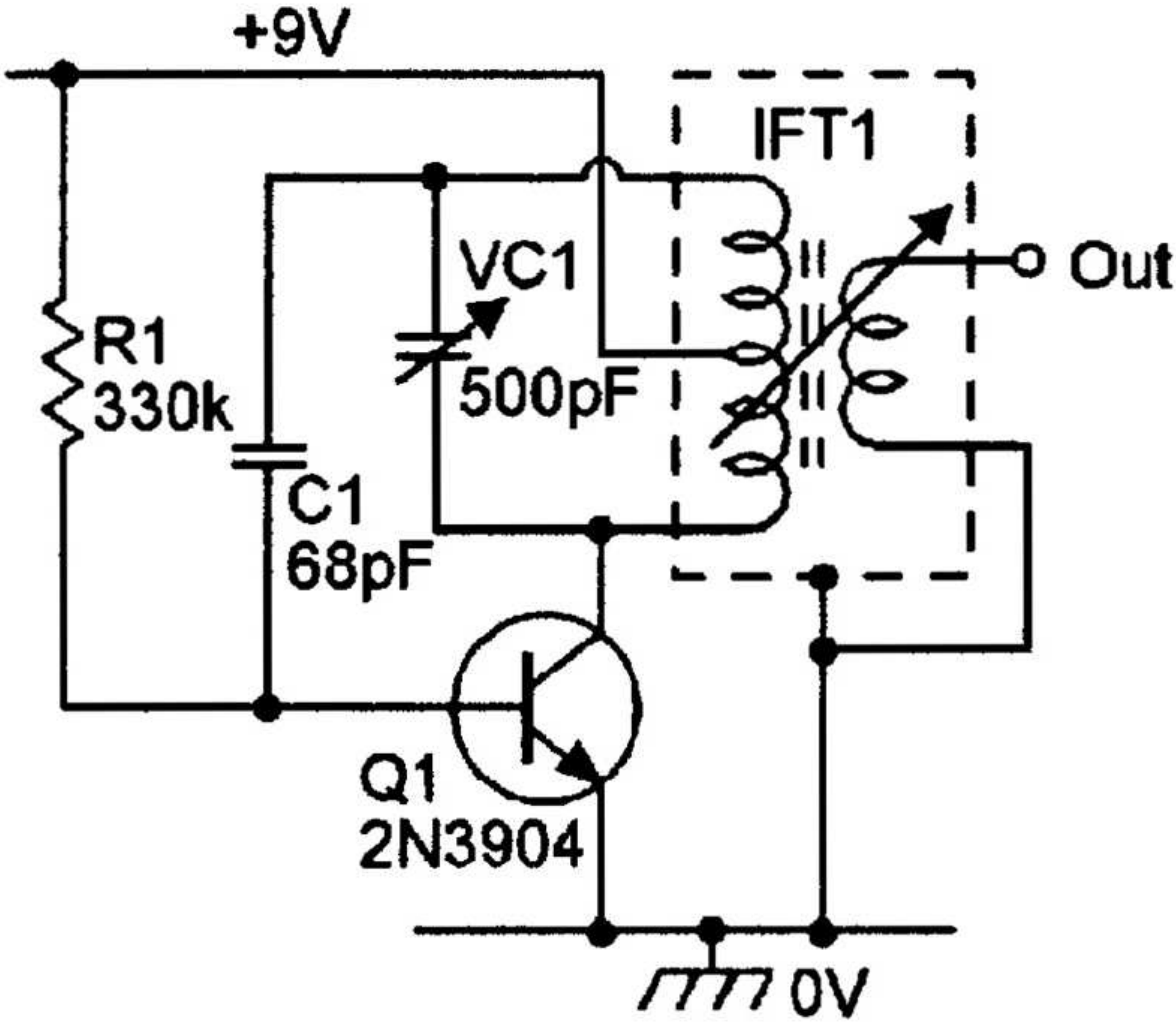


FIGURE 8. MW signal generator/BFO.

Figure 9 shows the above oscillator modified so that, when used in conjunction with a MW radio, it functions as a simple metal/pipe locator. Oscillator coil L1 is hand-wound and comprises 30 center-tapped turns of wire, firmly wound over about a 25 mm (one-inch) length of a 75 to 100 mm (three- to four-inch) diameter non-metallic former or search head and connected to the main circuit via a three-core cable. The search head can be fixed to the end of a long non-metallic handle if the circuit is to be used in the classic metal detector mode, or can be hand-held if used to locate metal pipes or wiring hidden behind plasterwork, etc. Circuit operation relies on the fact that L1's electromagnetic field is disturbed by the presence of metal, causing the inductance of L1 and the frequency of the oscillator to alter. This frequency shift can be detected on a portable MW radio placed near L1 by tuning the radio to a local station and then adjusting VC1 so that a low frequency beat or whistle note is heard from the radio. This beat note changes if L1 (the search head) is placed near metal.

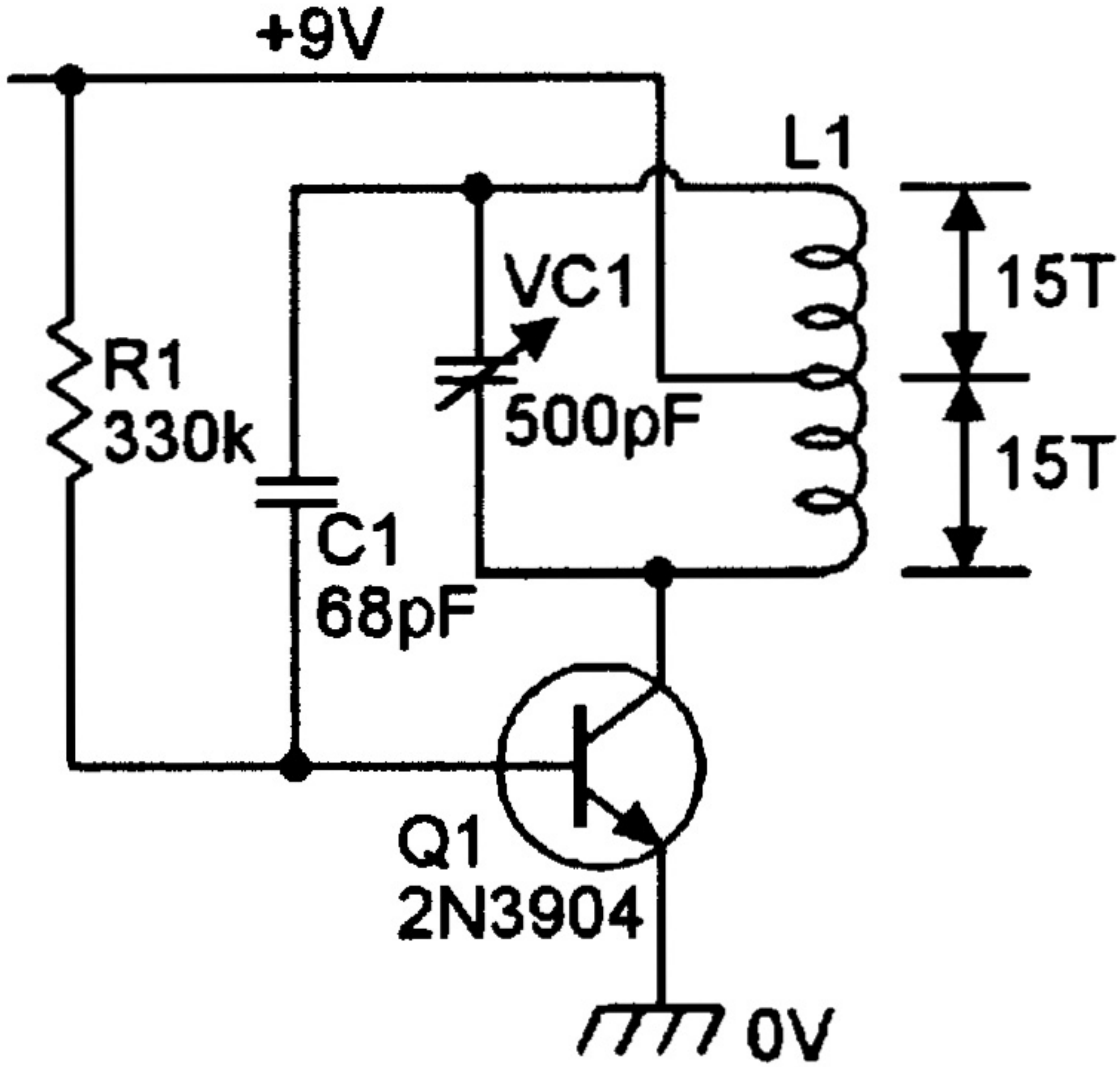


FIGURE 9. Metal/pipe locator.

Figure 10 shows another application of the Hartley oscillator. In this case, the circuit functions as a DC-to-DC converter, which converts a 9 V battery supply into a 300-V DC output. T1 is a 9V-0-9V to 250 V transformer, with its primary forming the L part of the oscillator. The supply voltage is stepped up to about 350 V peak at T1 secondary, and is half-wave rectified by D1 and used to charge C3. With no permanent load on C3, the capacitor can deliver a powerful but non-lethal belt. With a permanent load on the output, the output falls to about 300 V at a load current of a few mA.

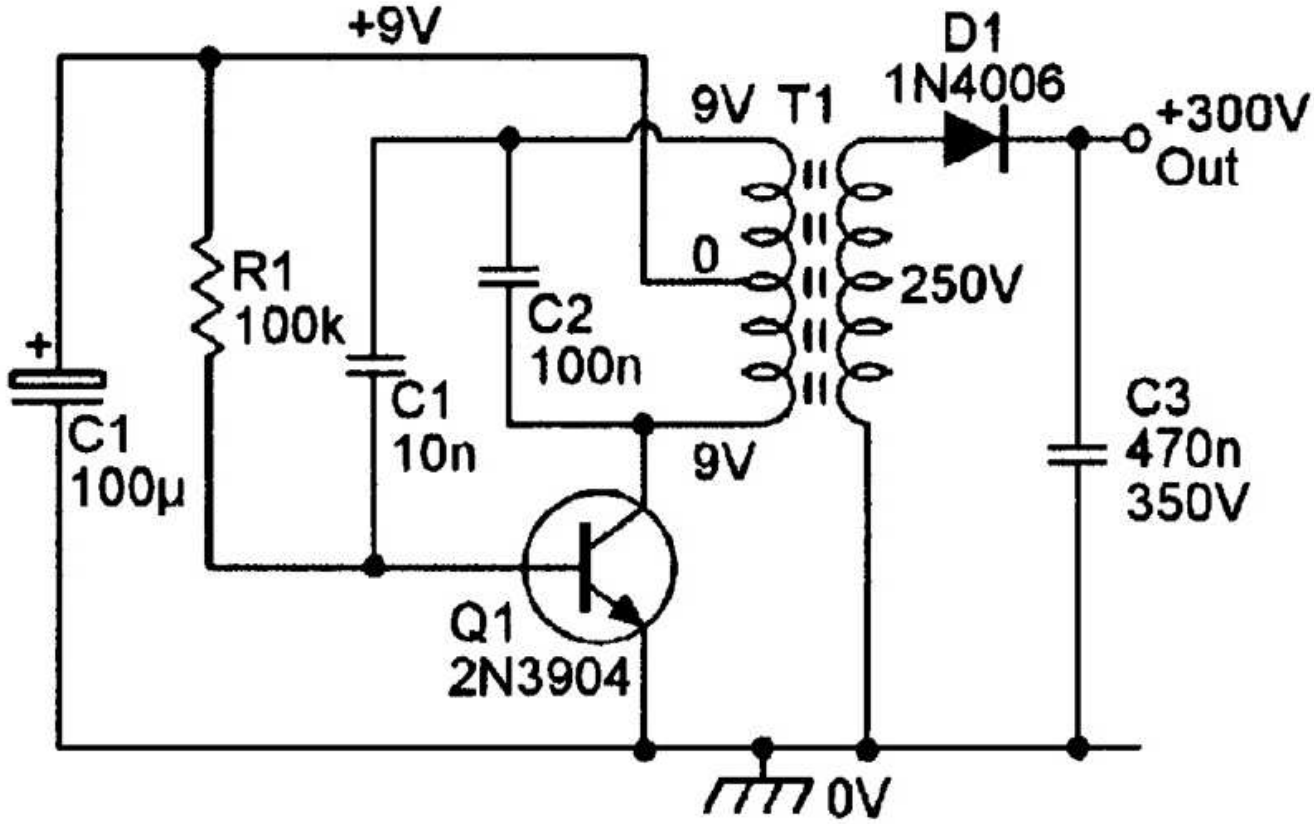


FIGURE 10. 9 V to 300 V DC-to-DC converter.

FM TRANSMITTERS

Figures 11 and **12** show a pair of low-power FM transmitters that generate signals that can be picked up at a respectable range on any 88 to 108 MHz FM-band receiver. The **Figure 11** circuit uses IC1 as a 1 kHz squarewave generator that modulates the Q1 VHF oscillator, and produces a harsh 1 kHz tone signal in the receiver; this circuit thus acts as a simple alarm-signal transmitter.

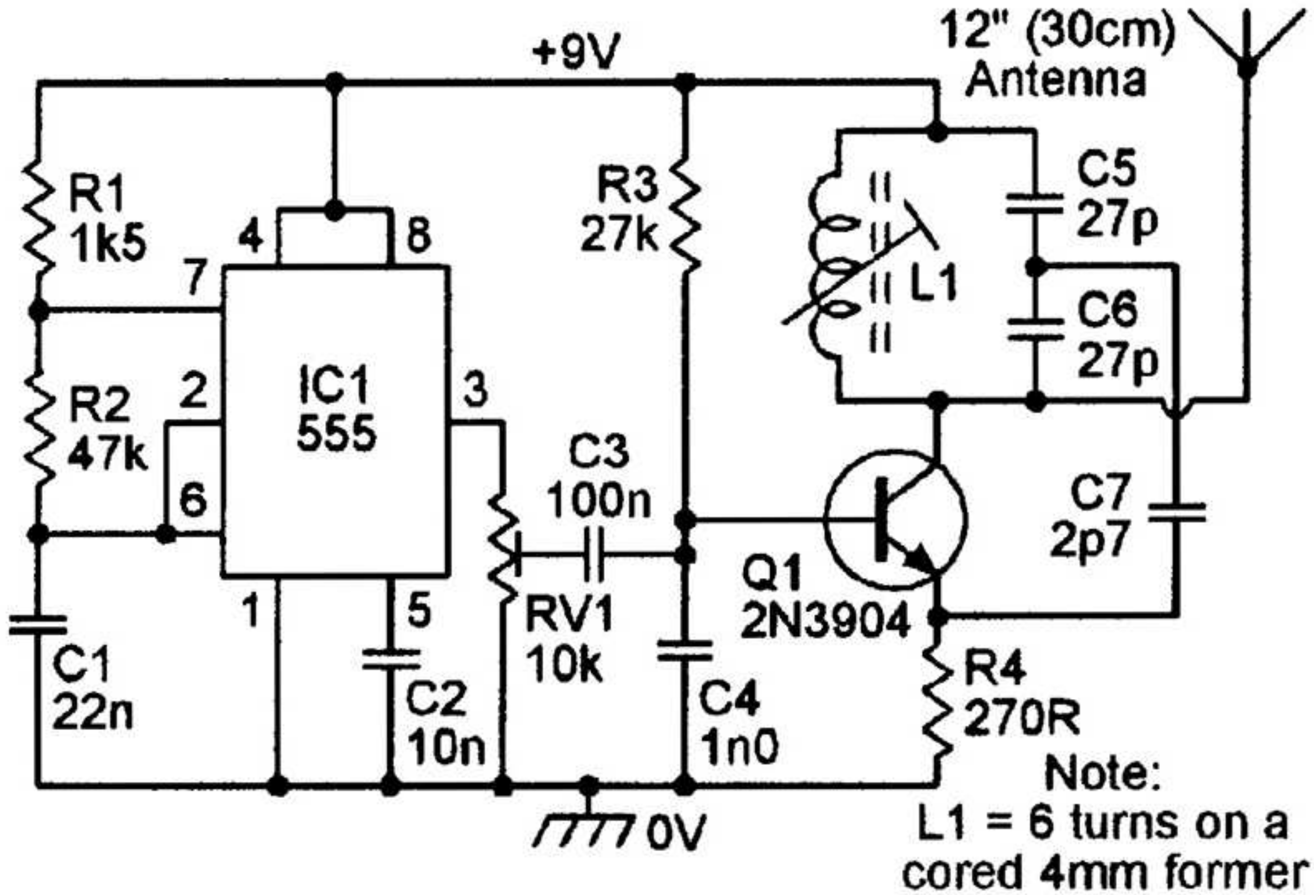


FIGURE 11. FM radio transmitter alarm.

The **Figure 12** circuit uses a two-wire electret microphone insert to pick up voice sounds, etc., which are amplified by Q1 and used to modulate the Q2 VHF oscillator; this circuit thus acts as an FM microphone or bug. In both circuits, the VHF oscillator is a Colpitts type, but with the transistor used in the common-base mode and C7 giving feedback from the tank output back to the emitter input.

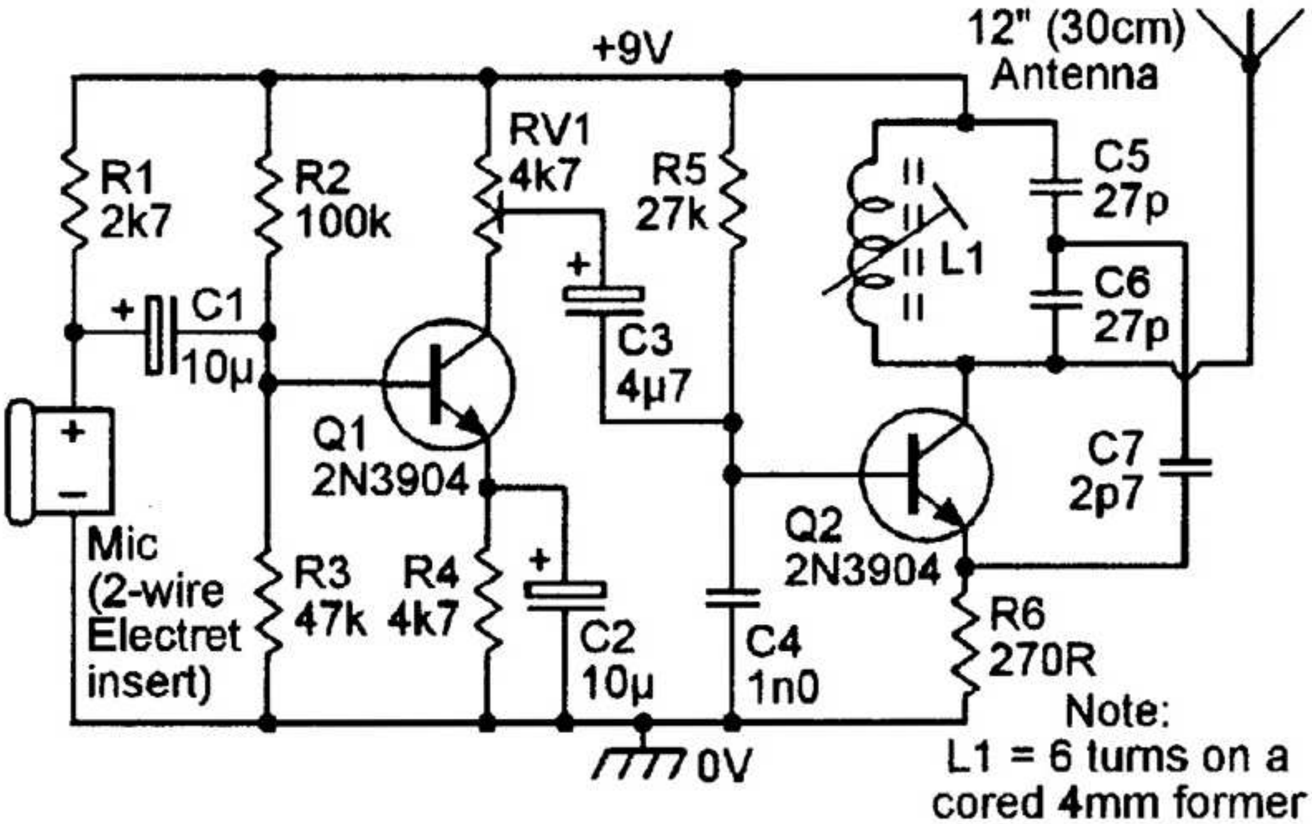


FIGURE 12. FM microphone/bug transmitter.

These two circuits have been designed to conform to American FCC regulations, and they thus produce a radiated field strength of less than 50 µV/m at a range of 15 meters (15 yards), and can be freely used in the USA. It should be noted, however, that their use is quite illegal in many countries, including the UK.

To set up these circuits, set the coil slug at its middle position, connect the battery, and tune the FM receiver to locate the transmitter frequency. If necessary, trim the slug to tune the transmitter to a clear spot in the FM band. RV1 should then be trimmed to set the modulation at a clean level.

TRANSISTOR AC VOLTMETERS

An ordinary moving-coil meter can be made to read AC voltages by feeding them to it via a rectifier and suitable multiplier resistor, but produces grossly non-linear scale readings if used to give FSD values below a few volts. This non-linearity problem can be overcome by connecting the meter circuitry into the feedback loop of a transistor common-emitter amplifier, as shown in the circuits of **Figures 13** to **15**, which (with the Rm values shown) each read 1 V FSD.

The **Figure 13** circuit uses a bridge rectifier type of meter network, and draws a quiescent current of 0.3 mA, has an FSD frequency response that is flat from below 15 Hz to above 150 kHz, and has superb linearity up to 100 kHz when using IN4148 silicon diodes or to above 150 kHz when using BAT85 Schottky types. R1 sets Q1's quiescent current at about treble the meter's FSD value, and thus gives the meter automatic overload protection.

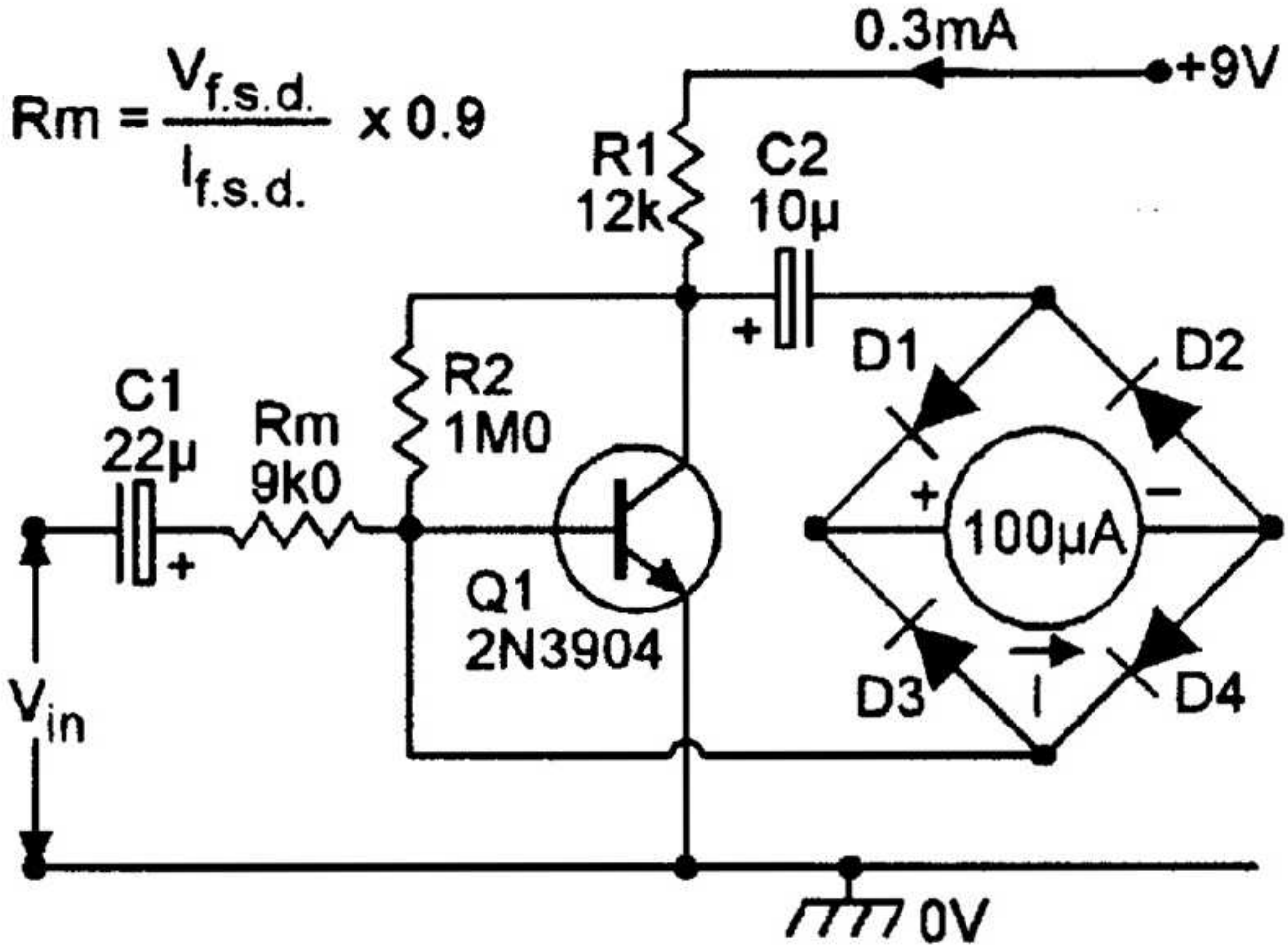


FIGURE 13. The frequency response of this 1 V AC meter is flat to above 150 kHz.

Figures 14 and **15** show pseudo full-wave and ghosted half-wave versions of the above circuit. These have a performance similar to that of **Figure 13**, but with better linearity and lower sensitivity. D3 is sometimes used in these circuits to apply slight forward bias to D1 and D2 and thus enhance linearity, but this makes the meter pass a standing current when no AC input is applied. The diodes used in these and all other electronic AC meter circuits shown in this article should be either silicon (IN4148, etc.) or (for exceptionally good performance) Schottky types; germanium types should not be used.

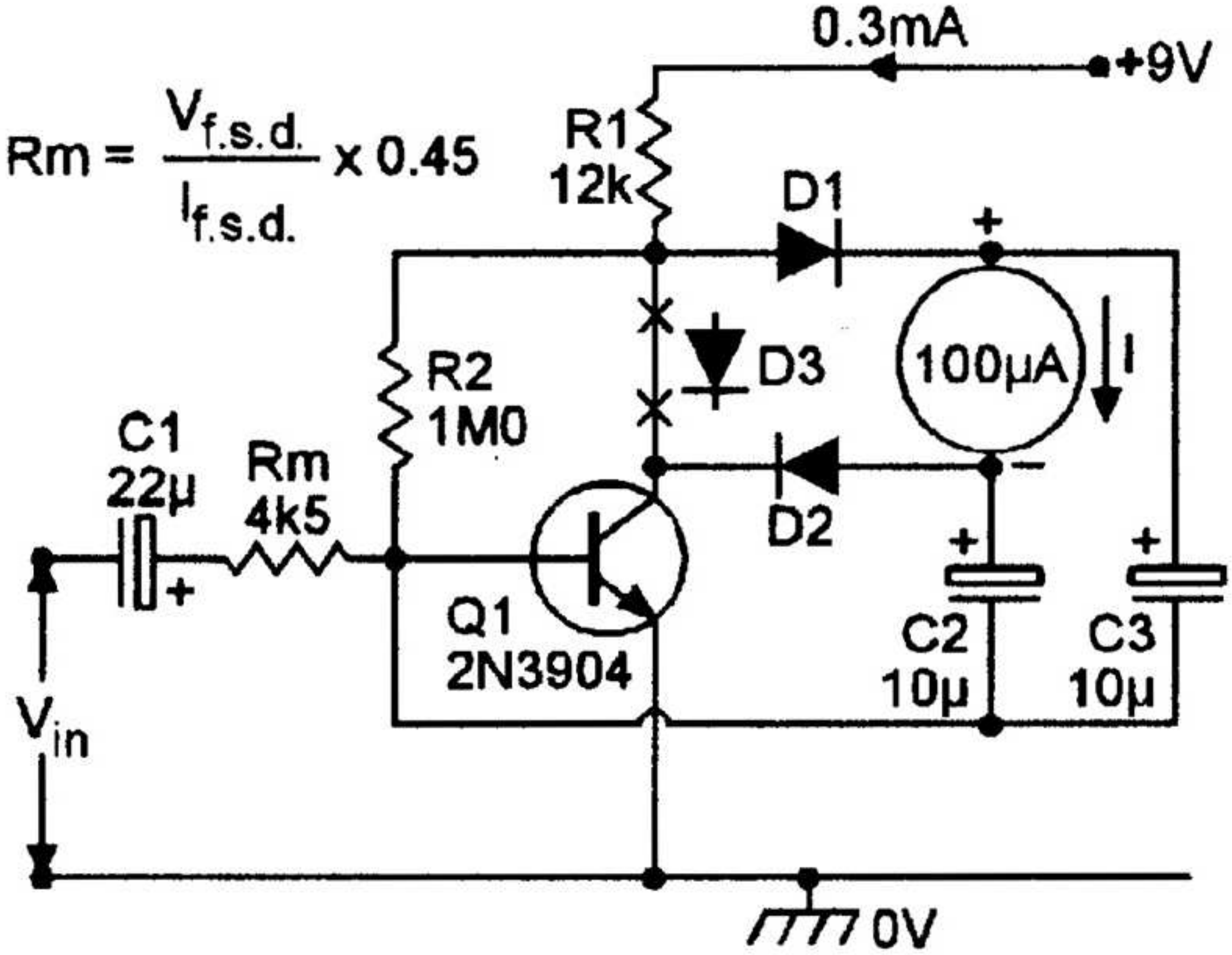


FIGURE 14. Pseudo full-wave version of the 1 V AC meter.

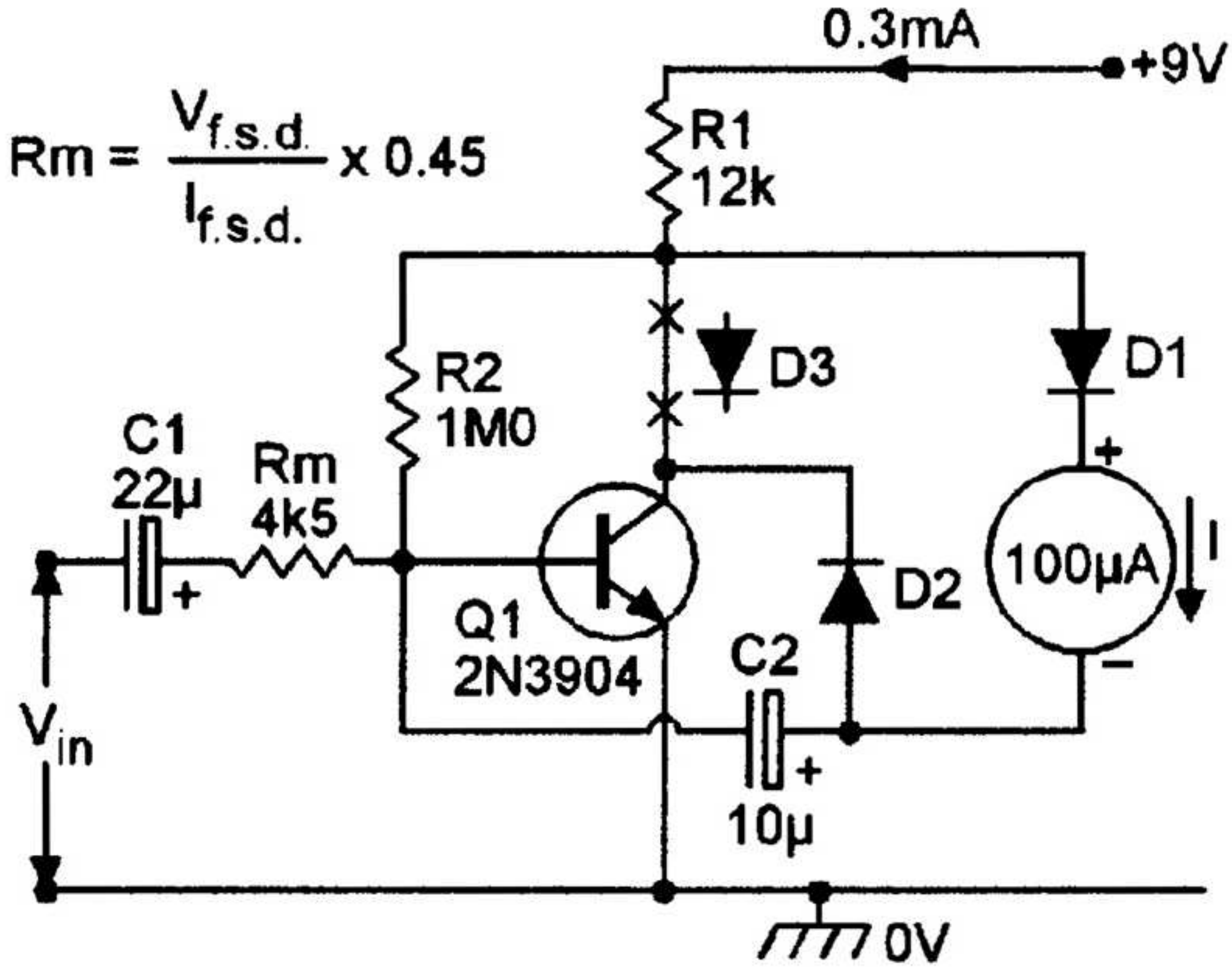


FIGURE 15. Ghosted half-wave version of the 1 V AC meter.

In the circuits in **Figures 13** to **15**, the FSD sensitivity is set at 1 V via Rm, which can not be reduced below the values shown without incurring a loss of meter linearity.

The Rm value can, however, safely be increased, to give higher FSD values, e.g., by a factor of 10 for 10 V FSD, etc.

If greater FSD sensitivity is wanted from the above circuits, it can be obtained by applying the input signal via a suitable pre-amplifier, i.e., via a +60 dB amplifier for 1 mV sensitivity, etc. **Figure 16** shows this technique applied to the **Figure 13** circuit, to give an FSD sensitivity variable between 20 mV and 200 mV via RV1. With the sensitivity set at 100 mV FSD, this circuit has an input impedance of 25 K and a bandwidth that is flat within 0.5 dB to 150 kHz.

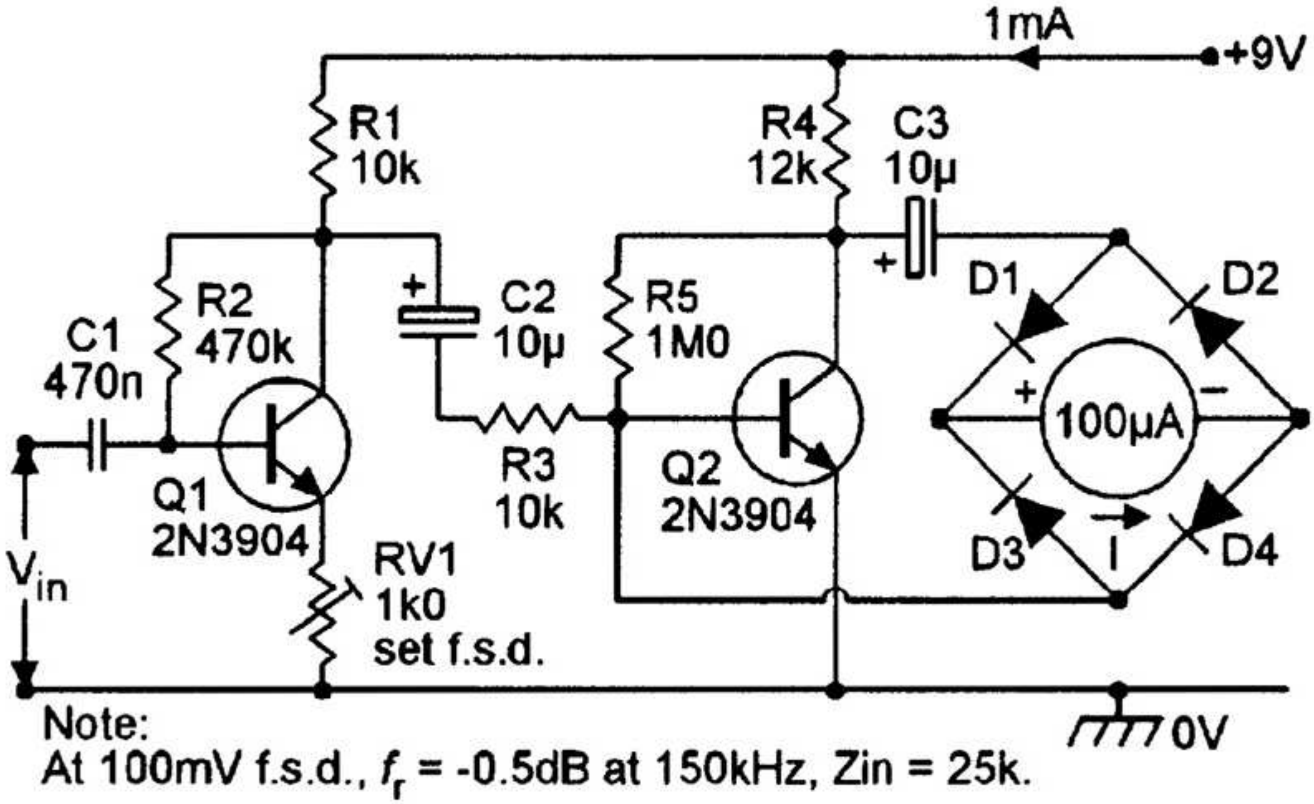


FIGURE 16. This AC voltmeter can be set to give FSD sensitivities in the range 20 mV to 200 mV.

AC MILLIVOLTMETER CIRCUITS

A one-transistor AC meter cannot be given an FSD sensitivity greater than 1 V without loss of linearity. If greater sensitivity is needed, two or more stages of transistor amplification must be used. The highest useful FSD sensitivity that can be obtained (with good linearity and gain stability) from a two-transistor circuit is 10 mV, and **Figure 17** shows an excellent example that gives FSD sensitivities in the range 10 mV to 100 mV (set via Rx). It uses D1 and D2 in the "ghosted half-wave" configuration, and its response is flat within 0.5 dB to above 150 kHz; the circuit's input impedance is about 120 K when set to give 100 mV FSD sensitivity (Rx = 470 Ohms); when set to give 10 mV sensitivity (Rx = 47 Ohms), the input impedance varies from 90 K at 15 kHz to 56 K at 150 kHz.

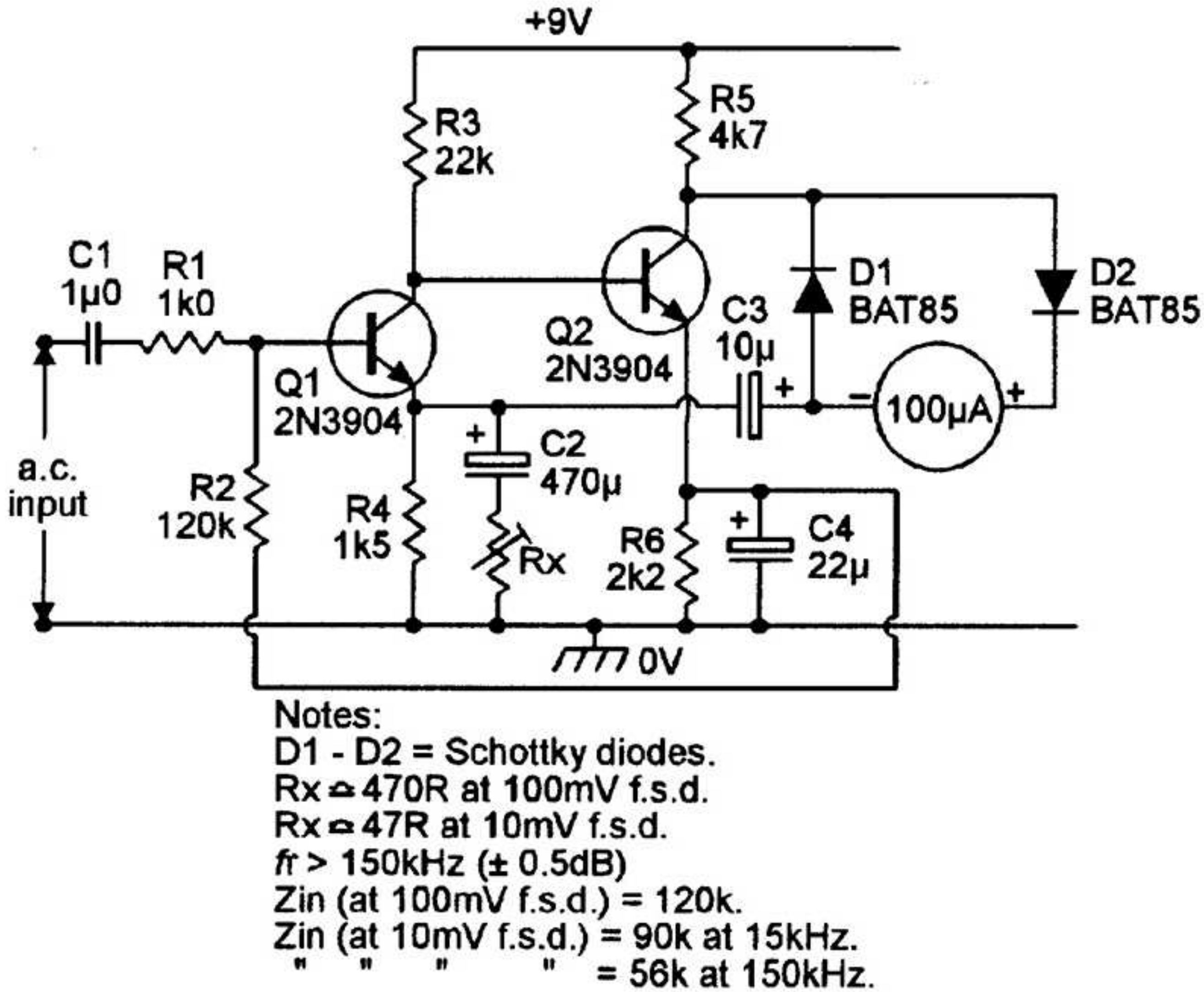


FIGURE 17. Wideband AC millivoltmeter with FSD sensitivity variable from 10 mV to 100 mV via Rx.

Figure 18 shows a simple x10 pre-amplifier that can be used to boost the above circuit's FSD sensitivity to 1 mV; this circuit has an input impedance of 45 K and has a good wideband response. Note, when building highly sensitive AC millivoltmeters, great care must be taken to keep all connecting leads short, to prevent unwanted RF pickup.

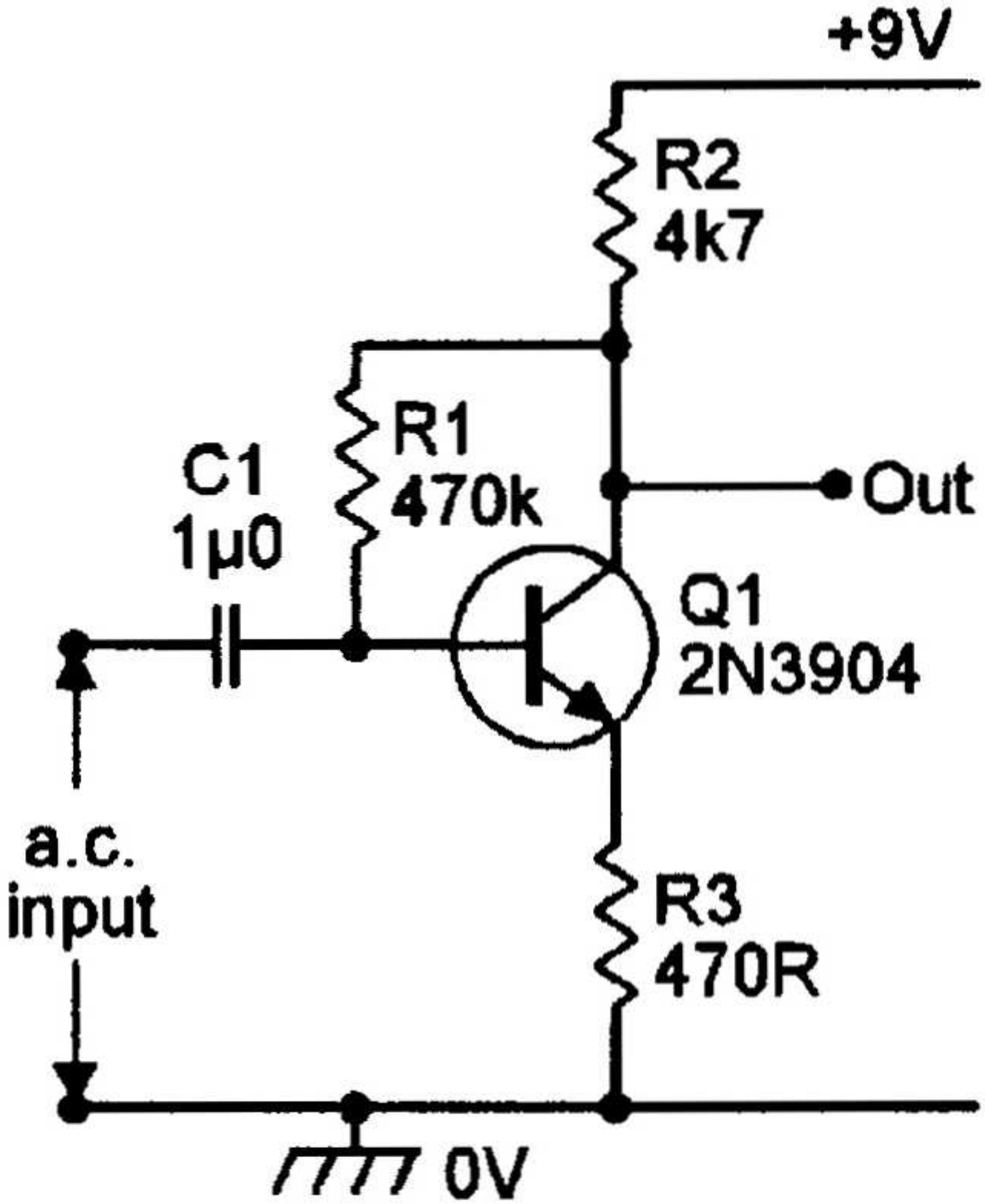


FIGURE 18. This x10 wideband pre-amplifier is used to boost an AC millivoltmeter's sensitivity.

A wide-range AC volt/millivolt meter can be made by feeding the input signals to a sensitive AC meter via suitable attenuator circuitry. To avoid excessive attenuator complexity, the technique of **Figure 19** is often adopted; the input is fed to a high-impedance unity-gain buffer, either directly (on "mV" ranges) or via a compensated 60 dB attenuator (on V ranges), and the buffer's output is fed to a basic 1 mV FSD meter via a simple low-impedance attenuator, which in this example has 1-3-10, etc., ranging.

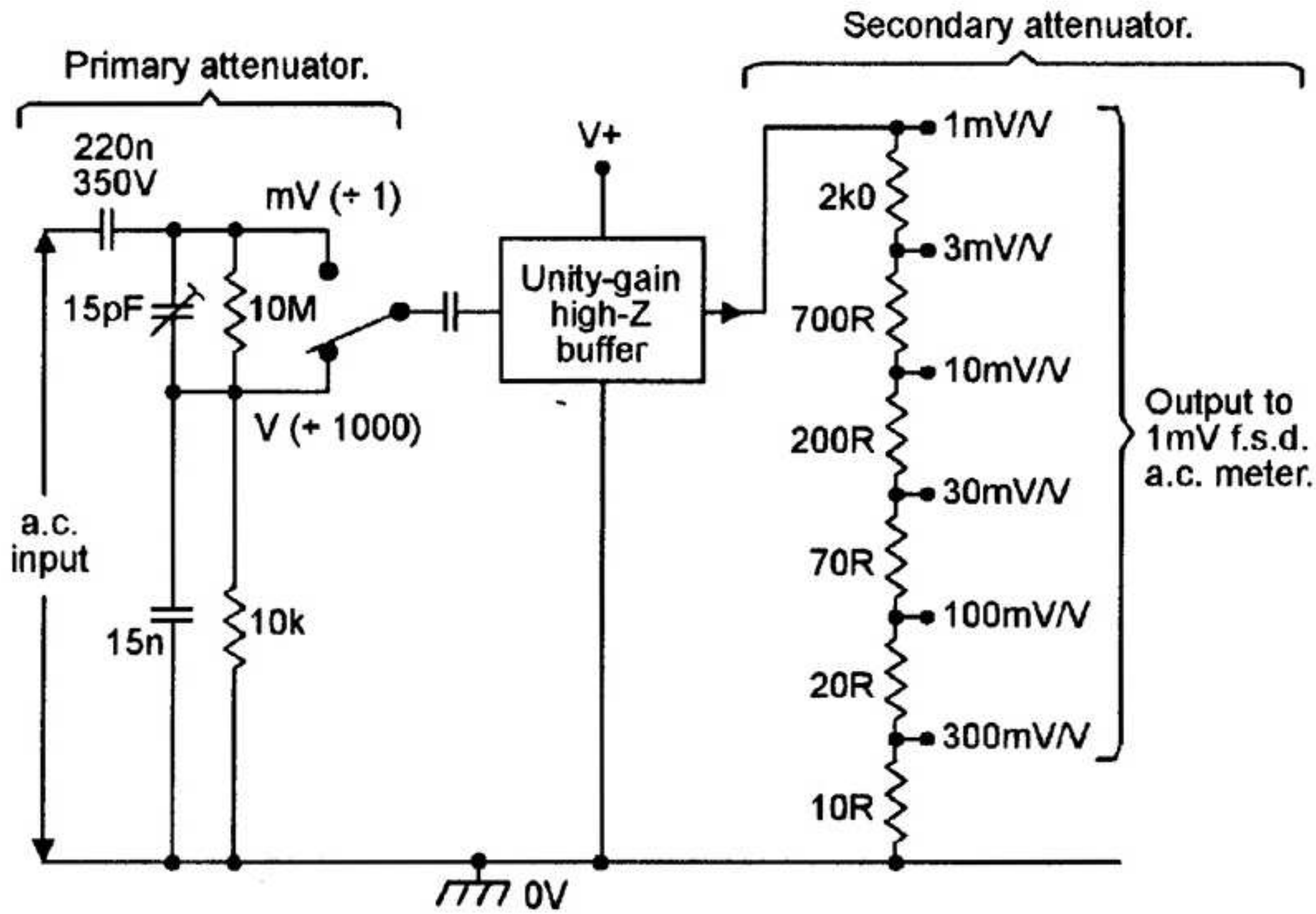


FIGURE 19. Basic multi-range AC volt/millivolt meter circuit.

Note, *when using this circuit that its input-to-unity-gain-buffer-output frequency response is virtually flat over the (typical) frequency range 20 Hz-150 kHz when used on the "mV" ranges, and that the primary attenuator's 15 pF trimmer must — when initially setting up the circuit — be adjusted on test to obtain the same frequency response on the basic "V" range.*

Figure 20 shows a useful variation of the above technique. In this case, the input buffer also serves as a x10 amplifier, and the secondary attenuator's output is fed to a meter with 10 mV FSD sensitivity, the net effect being that a maximum overall sensitivity of 1 mV is obtained with a minimum of complexity.

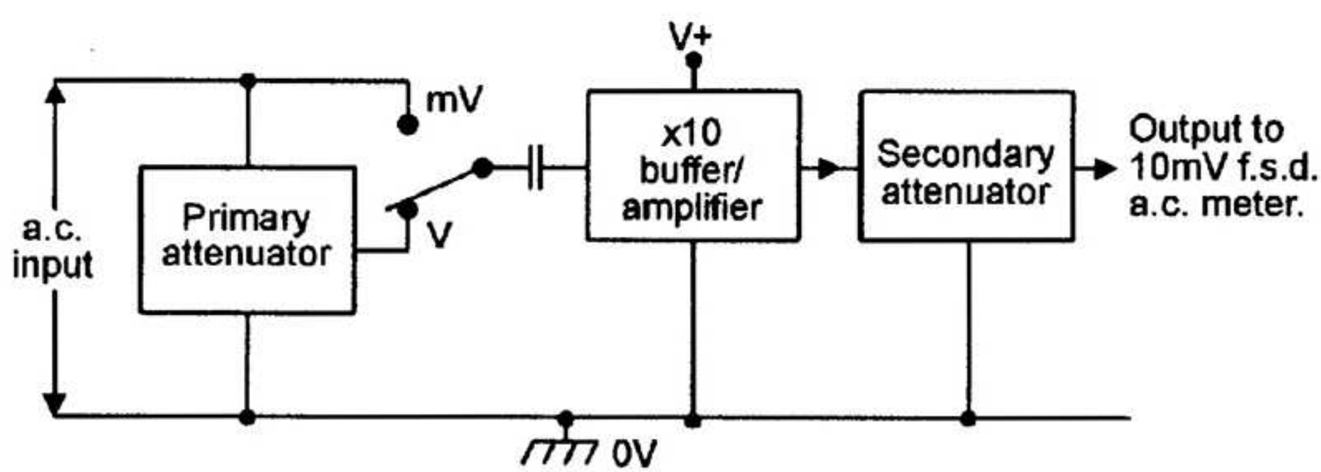


FIGURE 20. A useful AC volt/millivolt meter circuit variation.

Figures 21 and **22** show input buffers suitable for use with the above types of multi-range circuits. The **Figure 21** design is that of a unity-gain buffer; it gives an input impedance of about 4.0 M.

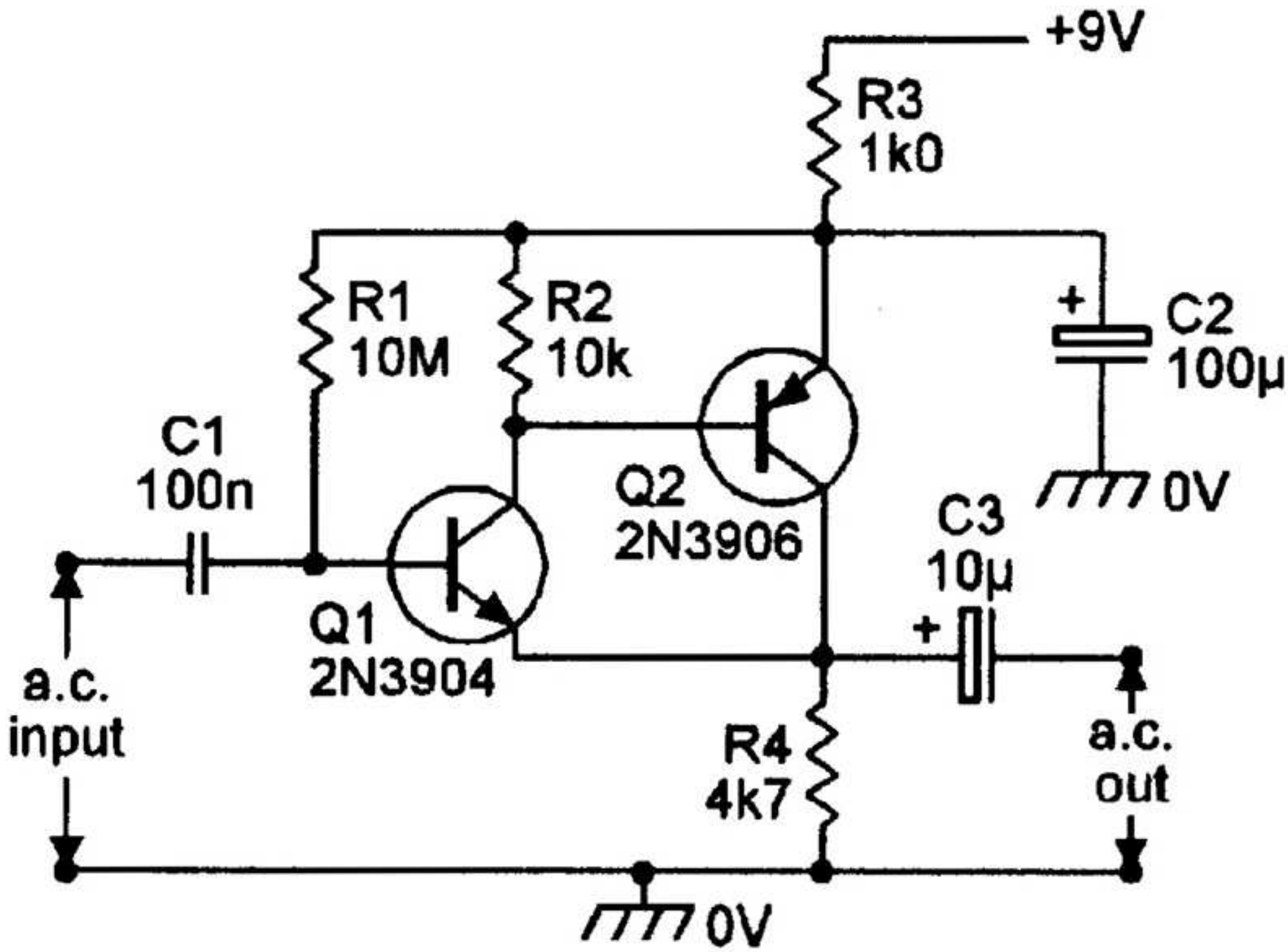


FIGURE 21. Unity-gain input buffer.

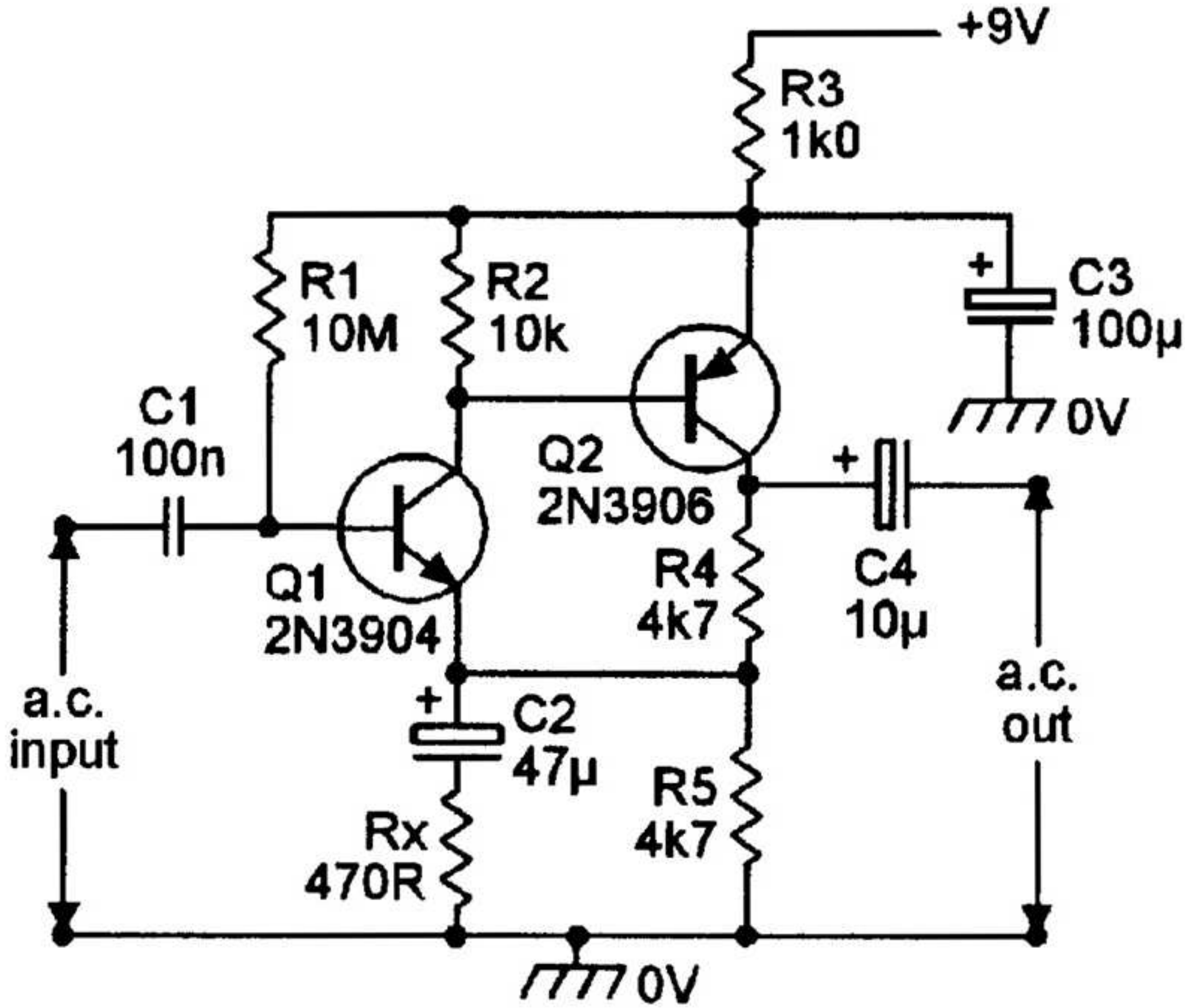


FIGURE 22. Buffer with x10 gain.

The **Figure 22** buffer gives a x10 voltage gain (set by the R1/Rx ratio) and has an input impedance of 1.0 M. **NV**
