WORKING WITH OP-AMPS

Op-amp du jour—a little of this and a little of that.

RAY MARSTON

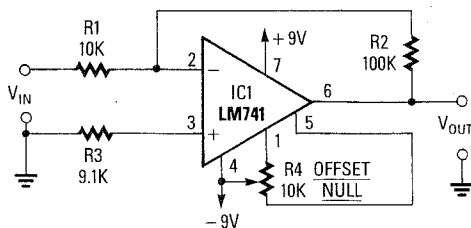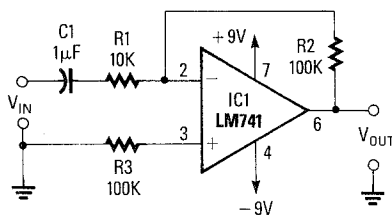FIG. 1—INVERTING DC AMPLIFIER with offset nulling and ×10 voltage gain.

NOTES:

$$A_V = \frac{R2}{R1}$$

$$V_{OUT} = -A_V \times V_{IN}$$

$$Z_{IN} = R1$$

$$R3 = R1\|R2$$

FIG. 2—INVERTING AC AMPLIFIER with ×10 voltage gain.

NOTES: R3 = R2

$$\text{CLOSED-LOOP BW} = \frac{F_T}{A_V}$$

F_T = OPEN-LOOP BANDWIDTH

A_V = CLOSED-LOOP GAIN

$$A_V = \frac{R1 + R2}{R1} = \frac{R2}{R1} + 1$$

FIG. 3—NON-INVERTING DC AMPLIFIER with offset nulling and ×10 voltage gain.

FIG. 4—NON-INVERTING VARIABLE-gain DC amplifier.

LET'S DESIGN CIRCUITS AROUND THE STAN-dard LM741 op-amp, powered by a ±9-volt supply. Starting with inverting and non-inverting congifurations, and how they work, we'll move on to voltage followers, biasing theory, adders and substractors, and phase splitters. We'll finish up with active filters. In practice, however, as long as the supply voltage stays the same, any op-amp can be substituted for the LM741 without modifying any of our circuits.

Inverting op-amp

Figure 1 shows an inverting DC-amplifier with an overall voltage gain (A_V) of 10. The voltage gain is determined by the resistance ratio R2/R1; by substituting a potentiometer for fixed resistor R2, the voltage gain can be made variable. The V_{OUT} formula shows the gain (A_V) to be negative because the output voltage is inverted 180 degrees from the input.

Offset nulling via R4 is used to adjust the output voltage to zero when the input voltage is zero. If the offset-nulling network is removed, the output voltage will be offset by an amount equal to the op-amp's input-offset voltage (typically 1 mV) multi-plied by the closed-loop voltage gain (A_V). For example, if the circuit has a gain of 100, the output may be offset by 100 mV when both inputs are grounded. For offset biasing stability, R3 should have a value equal to the parallel value of R1 and R2.

As shown in Fig. 2, by wiring a coupling capacitor in series with the input resistor R1, an AC amplifier is created. Notice that offset nulling is not needed and, for optimum biasing, R3 is given a value equal to R2.

In the small-signal AC amplifier, the closed-loop BW (bandwidth) is equal to the open-loop BW (F_T) divided by the gain (A_V). For example, when the F_T is 1 MHz and A_V equals 10, then the closed-loop BW equals 100 kHz. The F_T value for any op-amp is obtained from the manufacturer's specification sheet: The LM741's F_T is typically 1 MHz.

Non-inverting op-amp

Shown in Fig 3 is a non-inverting ×10 DC amplifier with offset compensation. The voltage gain is determined by [(R2/R1) + 1]. If R2 is given a value of zero (a short circuit), the gain falls to un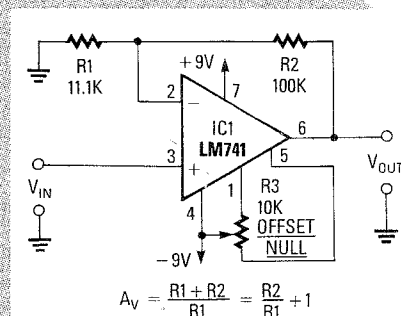ity. As shown in Fig. 4, the gain can be mad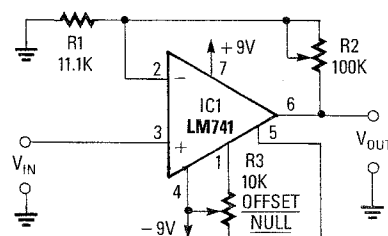e variable, over the range of 1 to 101, by replacing R2 with a potentiometer and connecting the wiper to the inverting input. The parallel value of R1 and R2 should ideally (for optimum biasing) be equal to the source resistance of the input signal. A major feature of the non-inverting op-amp is its high input-impedance, which can easily have a value of hundreds of megohms.

Figure 5 shows a ×10 non-inverting AC amplifier. Notice, here, that gain-control resistors R2 and R1 are isolated from ground via the bypass capacitor C2. At operating frequen-
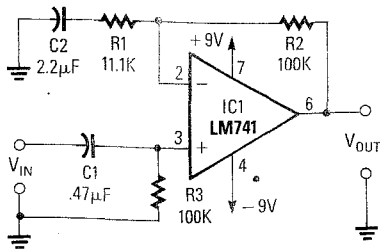
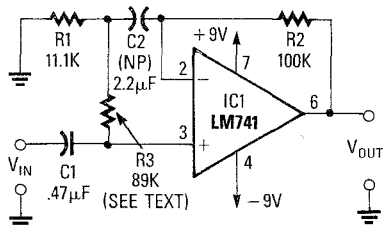**FIG. 5—NON-INVERTING ×10 AC amplifier with 100,000-ohm input impedance.**

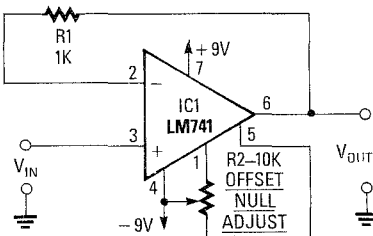**FIG. 6—NON-INVERTING ×10 AC amplifier with 50-megohm input impedance.**

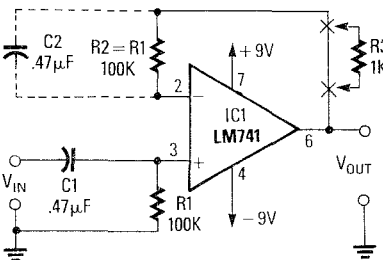**FIG. 7—DC VOLTAGE FOLLOWER with offset nulling.**

FIG. 8—AC-VOLTAGE FOLLOWER with 100,000-ohm input impedance.

cies, C2 has negligible impedance, so the AC-voltage gain is determined by the R2/R1 ratio; however, because there's no return path for output DC to ground, the inverting input is subject to virtually 100% DC negative feedback, that is, unity gain. But for unity gain, wouldn't resistor R2 need to be a direct short for 100% feedback? Not quite; let's see why.

Except for input-offset currents, no DC current flows through feedback resistor R2, so there's no voltage drop and, consequently, whatever voltage

appears at the output (pin 6) also appears at the inverting input (pin 2). It sounds surprising, but as far as DC is concerned, R2 looks like a short circuit from pin 6 to pin 2. For optimum input-current biasing, R3 should have a value equal to that of R2. Clearly, the non-inverting input impedance equals the R3 value.

Figure 6 shows how to design for a high input impedance (typically 50 megohms). The low end of R3 is taken to ground via R1, and the AC-feedback signal appearing at the R3-R1 junction is virtually identical to that appearing on the non-inverting input. Consequently, near-identical signal voltages appear at both ends of R3, which thus passes negligible signal current. The apparent impedance of that resistor is increased to near infinity by that "bootstrap" action. In practice, however, the input impedance is limited to about 50 megohms because of the op-amp's socket and PC-board impedances. For optimum input-current biasing, the sum of R2 and R3 should equal R1.

Voltage follower

A voltage follower produces an output signal identical to the input signal. The circuit functions as a unity-gain non-inverting amplifier with 100% negative feedback, where the input impedance is very high and the output impedance is very low. Figure 7 shows a voltage follower with offset nulling; however, there would be an output error of only a few millivolts by eliminating the offset nulling entirely. Notice that, for optimum input-current biasing, feedback resistor R1 should have a value equal to the source resistance of the input signal.

The value of feedback resistor R1 can be varied over a wide range (from zero to 100,000 ohms) without greatly influencing the output accuracy. If an op-amp has a low open-loop BW, the

R1 value can usually be reduced to zero. However, many op-amps with a high open-loop BW tend towards instability when used in the unity-gain mode. To reduce the bandwidth and enhance op-amp stability, R1 should be 1000 ohms or greater.

Figure 8 shows an AC version of the voltage follower. Here, the input AC-signal is coupled through C1, and the non-inverting input is tied to ground via R1, which determines the input impedance. Ideally, feedback resistor R2 should have the same value as R1 for correcting input-current offsets. However, if R2's value is very high, it may reduce the bandwidth too much; that problem can be overcome by shunting R2 with C2. Further stability can be achieved for high open-loop BW op-amps by connecting R3 in series with R2.
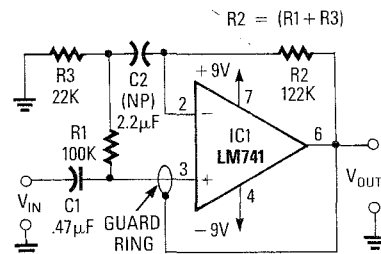
Figure 9 achieves an extremely

FIG. 9—AC VOLTAGE FOLLOWER with 50-megohm input impedance without guard ring, or 500 megohms with the guard ring.
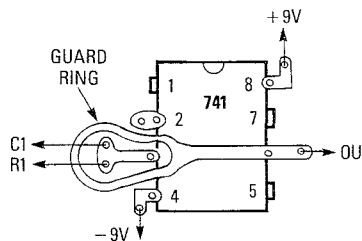

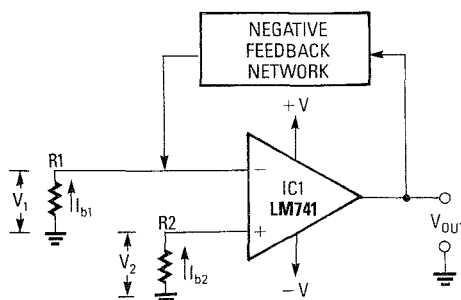**FIG. 10—GUARD RING ETCHED ON A PCB as viewed through the top of the PC board.**

NOTES: $I_{b1} = I_{b2} = I_b$

BIASING OUTPUT ERROR:

$V_{OUT} = I_b (R2 - R1) A_V$

$= A_V (V_2 - V_1)$

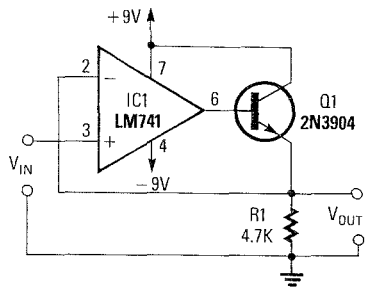**FIG. 11—INPUT BIASING of an op-amp.**

FIG. 12—UNI-DIRECTIONAL BOOSTED output-current drive in a DC voltage follower.

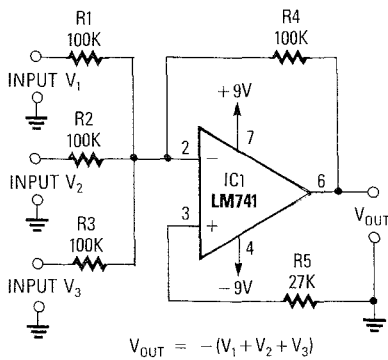FIG. 13—BI-DIRECTIONAL BOOSTED output-current in a DC voltage follower.

$$V_{OUT} = -(V_1 + V_2 + V_3)$$
FIG. 14—UNITY-GAIN DC ADDER.

Biasing accuracy

In Figs. 1 through 9, great emphasis was placed on selecting component values for optimum input biasing. Figure 11 shows a circuit for testing the input-offset bias. Both inputs are tied to ground via resistors R1 and R2; equal input bias currents, I_{b1} and I_{b2}, are drawn through those resistors, thereby generating equal voltage drops. Because a zero-differential voltage appears across the input, a biasing error of zero volts will appear at the output. If, on the other hand, R1 and R2 do not have equal values, or the input bias currents are slightly different, the voltage drops across each resistor will differ; that will produce an input differential voltage that will be amplified by A_V to produce an output-offset voltage. How significant is that output-voltage error, really?

In practice, a bipolar op-amp such as the LM741 has a typical I_b value of about 200 nA (0.2 μA), producing a voltage drop of 0.2 mV across a 1000-ohm resistor. FET-input op-amps have typical I_b values of about 0.02 nA, producing a voltage drop of a mere 0.02 μV across a 1000-ohm resistor. Therefore, in Fig. 11, if the R1 and R2 values differ by as much as 10,000 ohms, a LM741 op-amp will produce a biasing output error of only 2 mV in a unity-gain voltage follower, or 20 mV for a ×10 amplifier. If a FET-input op-amp is used in place of the LM741, the biasing output error of the voltage follower will be a mere 0.2 μV, and for the ×10 amplifier, a mere 2 μV.

It follows that Figs. 1 through 9 can accept considerable latitude in their biasing component values. With that in mind, let's look at more circuits.

Current-boosted follower

Most op-amps can provide maximum output currents of only a few milliamperes. However, as shown in Fig. 12, the current-driving capacity of a voltage follower can be easily increased by wiring an emitter-follower transistor to the output. Notice that the base-emitter junction is wired into the negative-feedback loop, which virtually eliminates the effects of junction non-linearity. That technique is not used in Fig. 13, which
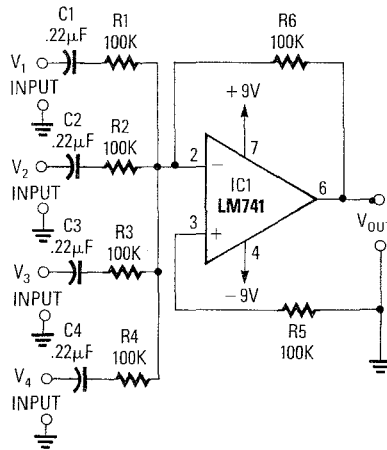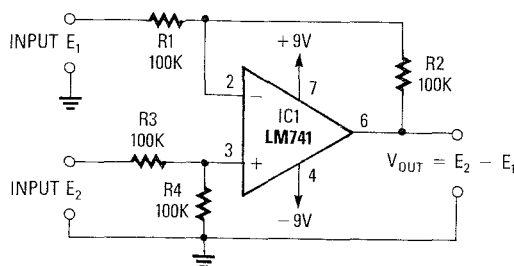

FIG. 15—4-INPUT AUDIO MIXER.

$$V_{OUT} = E_2 - E_1$$
FIG. 16—DIFFERENTIAL AMPLIFIER, or subtracter.

NOTES:

WHEN, $\dfrac{R1}{R2} = \dfrac{R3}{R4}$

THEN, $A_V = \dfrac{R2}{R1}$

high input impedance, where R1 is "bootstrapped" from the op-amp output via C2, causing R1's impedance to increase to near infinity—resulting in about a 50-megohm input impedance. The input impedance is still limited by the leakage impedances of the op-amp's IC socket and PCB, which can also be bootstrapped and raised to near-infinite values by using a printed circuit "guard ring" surrounding the op-amp's input pin. In that case, Fig. 9 would have a 500-megohm input impedance when using a LM741 op-amp, or even greater if a FET-input op-amp is used. Figure 10 shows a guard ring etched on a PCB.
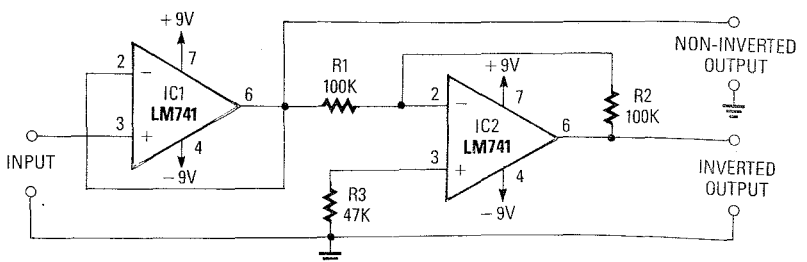

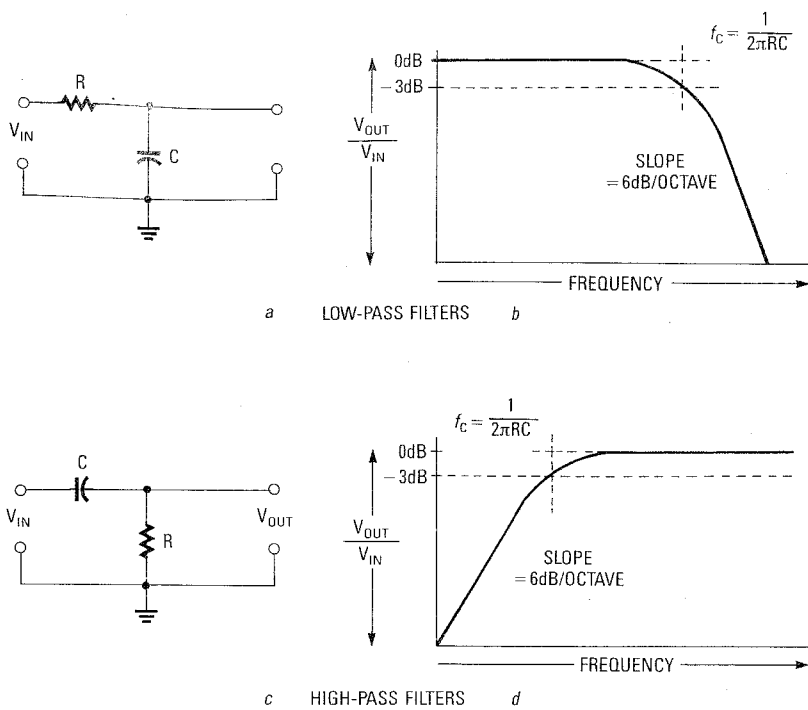FIG. 17—BALANCED PHASE-SPLITTER.

RADIO-ELECTRONICS

FIG. 18—1st-ORDER RC FILTERS FREQUENCY RESPONSE CURVES: (a) is a passive low-pass filter, and (b) is its frequency response; (c) is a passive high-pass filter, and (d) is its frequency response.
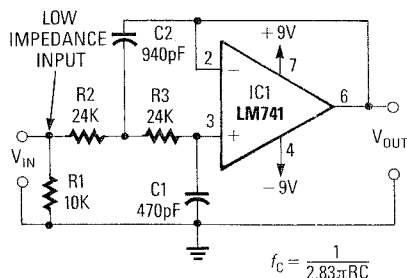
$$f_c = \frac{1}{2\pi RC}$$

a LOW-PASS FILTERS b

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

c HIGH-PASS FILTERS d

FIG. 19—2ND-ORDER LOW-PASS 10-KHZ unity-gain active filter.

$$f_c = \frac{1}{2.83\pi RC}$$

FIG. 20—2ND-ORDER LOW-PASS 10-KHZ equal-value-component active filter.

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

produces significant cross-over distortion as the output swings past zero volts. The distortion can be eliminated by suitably biasing Q1 and Q2; in which case, the circuit can form the basis of a good hi-fi amplifier.

Adders and subtractors

Figure 14 shows a unity-gain DC-voltage adder, which gives an inverted output voltage equal to the sum of the three input voltages. The current flowing in R4 is equal to the sum of the currents in R1 through R3. If required, the circuit can be made to give a voltage gain greater than unity by simply increasing the value of feedback resistor R4. Figure 15 is a 4-input audio mixer using AC input-coupling capacitors.

Figure 16 shows a unity-gain DC differential amplifier, also called a subtracter, where the output equals the difference between the two input-

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

FIG. 21—4TH-ORDER 10-KHZ low-pass filter.

NOTES: IC1 $A_{V1} = \frac{R2}{R1} + 1$

IC2 $A_{V2} = \frac{R4}{R3} + 1$

signal voltages; if the component values are chosen such that R2 = R4 and R1 = R3, then the voltage gain, A_V, equals R2/R1.

Balanced phase-splitter

A phase-splitter has two outputs that are identical in both amplitude and form, but one output is phase-shifted by 180 degrees relative to the other. Figure 17 shows a unity-gain DC phase splitter. Here, IC1 acts as a unity-gain non-inverting amplifier, while IC2 acts as a unity-gain inverting amplifier that provides the 180 degrees phase-shifted output.

Active filters

Filter circuits are used to reject unwanted frequencies while passing frequencies that the designer wants. The low-pass RC filter in Fig. 18-a, passes low-frequency signals, but rejects high-frequency signals. Once the output falls to 3 dB at a "break" or "cross-over" frequency, as Fig. 18-b shows, the output continues to rolloff at a rate of 6 dB/octave (20 dB/decade) as the frequency increases. For example, a 1-kHz low-pass filter gives 3-dB rejection at 1 kHz, 9-dB rejection at 2 kHz, and, roughly, a 15-dB rejection to a 4-kHz signal (or 20 dB
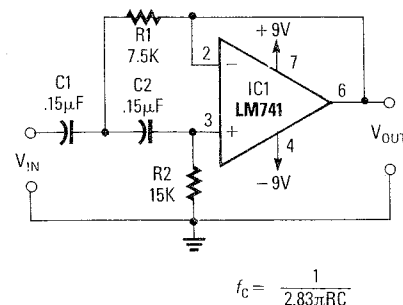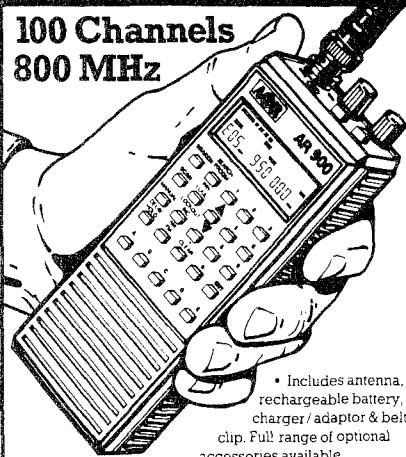

$$f_c = \frac{1}{2.83\pi RC}$$

FIG. 22—2ND-ORDER 100-HZ HIGH-PASS unity-gain active filter.

FIG. 23—2ND-ORDER 100-HZ HIGH-PASS
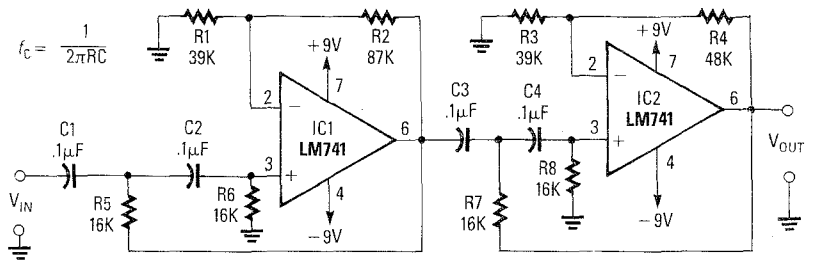equal-value-component active filter.

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

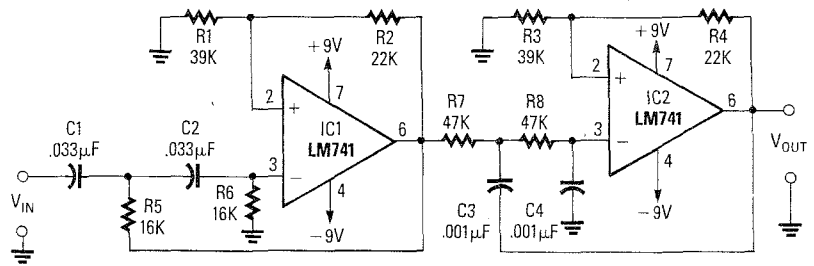FIG. 24—4TH-ORDER 100-HZ HIGH-PASS FILTER.

FIG. 25—300-HZ TO 3.4-KHZ SPEECH FILTER with 2nd-order response.

the feedback network of an op-amp
making what are known as "active"
filters. Let's look at practical designs.

Figure 19 shows a Butterworth 2nd-
order low-pass filter with a break fre-
quency at 10 kHz. That design gives
unity gain within its passband. To
change the break frequency, simply
change either the R or the C value. A
major disadvantage of Fig. 19 is that
the value of C2 must be precisely
twice the value of C1 for correct oper-
ation and, in practice, that can result
in some rather odd component values.

to a 10-kHz signal). Notice that the
octave of 1 kHz is 2 kHz, and the
octave of 2 kHz is 4 kHz; each octave
causes an added attenuation of 6 dB.

The high-pass RC filter in Fig.
18-*c*, passes high-frequency signals
but rejects low-frequency signals.
After the output is 3-dB down at a
break frequency, as Fig. 18-*d* shows,
the output continues to roll-off at a
rate of 6 dB/octave.

Each of the two filter circuits uses a
single RC stage, and is known as the
"1st order" filter. If we cascade "n"
filter stages, it would be known as an
"nth order" filter, and would have an
output slope beyond the break fre-
quency of (n × 6 dB)/octave.

Unfortunately, simple RC filters
cannot be simply cascaded, because
each filter section would interact with
each other, resulting in poor perfor-
mance. All of that can be overcome
by incorporating the same filter into

Figure 20 shows an alternative 2nd-
order 10-kHz low-pass filter, which
overcomes that design snag by using
equal capacitor values. Notice here
that the op-amp is designed to give a
voltage gain (4 dB) via R2 and R1.

Figure 21 shows how equal-value-
component filters can be cascaded to
make a 4th-order low-pass filter with a
slope of 24 dB/octave.

Figures 22 and 23 shows a unity-
gain and equal-value component ver-
sions, respectively, of 2nd-order 100-
Hz high-pass filters. Figure 24 shows
a 4th-order 100-Hz high-pass filter.
The operating frequency can be
changed by increasing either the re-
sistor or capacitor values to reduce the
break frequency. Finally, Fig. 25
shows how a high-pass and low-pass
filter can be wired in series to make a
300-Hz to 3.4-kHz speech filter that
gives 12-dB/octave rejection to all sig-
nals outside of that range.   **R-E**