Edited by Bill Travis and Anne Watson Swager

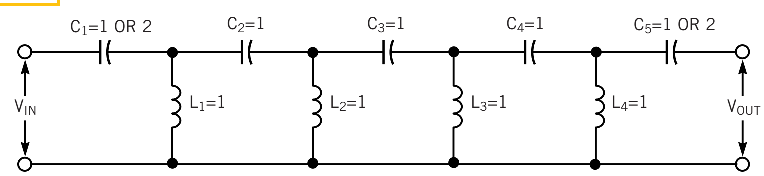
Highpass filters use modified equal-element design

Richard Kurzrok, Queens Village, NY

USING A MODIFIED equal-element design for a lumped-circuit lowpass filter has several advantages over the well-known equal-element design (**references 1** and **2**). The modified design exhibits superior passband performance with only modest degradation of stopband selectivity. Moreover, the modified design is simple and easy to manufacture. You can extend the modified equal-element design to highpass LC filters. Both equal-element and modified equal-element filters use the normalized highpass prototype (**Figure 1**). For the equal-element filter, the normalized value of the outside capacitors C_1 and C_5 is 1; for the modified equal-element filter, it's 2. The design reference frequency, at normalized frequency x=1.0, is not the 3-dB cutoff frequency for the filters in **Figure 1**. The 3-dB cutoff frequency occurs close to x=1.9. In contrast, in Butterworth and Chebyshev filters, x can equal 1.0 at the 3-dB cutoff frequencies. You use this different normalization method to calculate the values of the circuit elements.

We designed, assembled, and tested two nine-pole, modified equal-element filters. The filters used vector boards in die-cast aluminum boxes with BNCs. All

Figure 1



Doubling the values of the outside capacitors in an equal-element filter improves passband performance.

TABLE 1—MEASURED RESPONSE FOR 3.56-MHz MODIFIED HIGHPASS FILTER

Frequency (MHz)	Insertion loss (dB)	Frequency (MHz)	Insertion loss (dB)
2.9	39	6	0.4
2.95	34	6.5	0.25
3.1	29.8	7	0.2
3.3	19.8	8	0.2
3.5	9.8	10	0.1
3.6	4.5	15	0.15
3.7	1.2	20	0.1
3.8	1	25	0.15
3.9	1.7	30	0.2
4	2	40	0.2
4.2	1.7	50	0.25
4.5	0.8	60	0.35
5	0.25	80	0.2
5.5	0.4	100	0.4

TABLE 2—MEASURED RESPONSE FOR 11.17-MHz MODIFIED HIGHPASS FILTER

Frequency (MHz)	Insertion loss (dB)	Frequency (MHz)	Insertion loss (dB)
9	37.8	13	1.4
9.5	30	13.5	0.85
10	22	14	0.4
10.5	15	14.5	0.25
11	7	15	0.25
11.2	2.8	17.5	0.4
11.4	1	19	0.45
11.6	0.8	20	0.4
11.8	1.2	25	0.3
12	1.5	30	0.2
12.2	1.8	40	0.2
12.4	2	50	0.2
12.6	1.8	60 to 100	Less than 0.6
12.8	1.6		

capacitors were ±5% polypropylene units. We selected the filter design frequencies based on available capacitor values. The first filter had a reference frequency of 6.773 MHz. Assuming a ratio of 1.902-to-1, this figure corresponds to a cutoff frequency of 3.561 MHz. After denormalizing the filter to actual circuit values (as in **Reference 1**), we determined all inductor values, L_1 through L_4, to be 1.175 μH. The interior filter capacitors, C_2, C_3, and C_4, are equal to 470 pF. The filter's input and output capacitors, C_1 and C_5, are then equal to 940 pF. To obtain this value, we connected standard 820- and 120-pF capacitors in parallel. All the inductors used 15 turns of number 26 magnet wire wound on Micro Metals T37-2 toroids. **Table 1** shows the measured amplitude response.

The second filter had a reference frequency of 21.22 MHz. Assuming a ratio of 1.902-to-1, this figure corresponds to a cutoff frequency of 11.168 MHz. After denormalizing the filter to actual circuit values (as in **Reference 1**), we determined all inductor values, L_1 through L_4, to be 0.375 μH. The interior filter capacitors, C_2, C_3, and C_4, are equal to 150 pF. The filter's input and output capacitors, C_1 and C_5, are then equal to 300 pF. To obtain this value, we connected two standard 150-pF capacitors in parallel. All the inductors used 10 turns of number 26 magnet wire wound on Micro Metals T25-6 toroids. **Table 2** shows the measured amplitude response. Assuming inductor unloaded Q of approximately 100, the measured data shows good correlation with calculated values. You can cascade the modified equal-element highpass filter with a similar lowpass filter to obtain a bandpass filter of high bandwidth. This technique provides an alternative to using image parameters (**Reference 3**).

References

1. Kurzrok, Richard, "Equal-element filter improves passband performance," *EDN*, March 15, 2001, pg 123.
2. Taub, JJ, "Design of Minimum Loss Bandpass Filters," *Microwave Journal*, November 1963, pg 67.
3. Kurzrok, Richard, "Wideband filter uses image parameters," *EDN*, Oct 26, 2000, pg 174.

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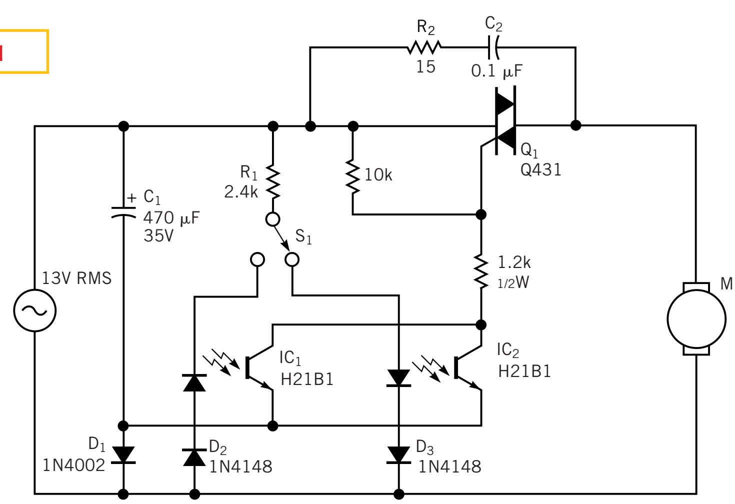
Position detectors provide motor-control logic

Steve Pomeroy and Russell Hedges, Elgar Corp, San Diego, CA

In the circuit of **Figure 1**, assume that a brush-type dc motor must drive a load back and forth between two endpoints on a lead screw. Optical sensors determine end of travel, and an SPDT switch selects to which end to send the load. The sensors themselves supply all the necessary directional logic, and a triac powers the motor with the necessary polarity of half-wave pulses from the 13.5V-ac input. Starting with the load parked at the south end, when you set switch S_1 to north, the ac input connects to the LED in the north-side sensor, IC_1, through current-limiting resistor R_1 and reverse-polarity-protection diode D_2. The phototransistor output from IC_1 then supplies firing pulses to the gate of triac Q_1 during the northbound half-cycle, and the load proceeds toward that detector.

Figure 1



The optocouplers provide both position-control logic and triac gate drive to the motor.

Similarly, IC_2 drives the motor during the other half-cycle to push the load south, and stops when it reaches the south endpoint. This scheme works even when you change the direction switch while the load is in motion. Power for the gate drives comes from the ac input and the half-wave rectifier comprising D_1 and C_1. You might need snubber R_2 and C_2 if the motor's inductance causes spurious firings of Q_1 during undesired half-cycles. The motor's stall current is approximately 2A. You can easily scale the design for larger or smaller motors.

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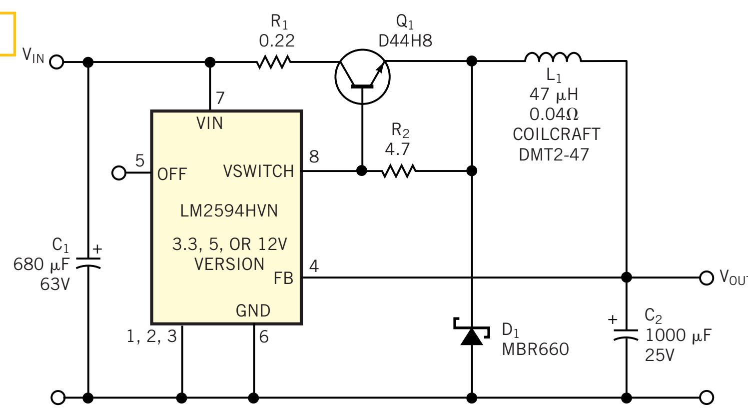
Bipolar transistor boosts switcher's current by 12 times

Wayne Rewinkel, National Semiconductor Corp, Santa Clara, CA

The circuit in **Figure 1** uses a minimal number of external parts to raise the maximum output current of a 0.5A buck switching-regulator IC to more than 6A. The circuit accommodates input voltages of 15 to 60V and delivers output voltages of 3.3, 5, or 12V, depending on your choice of IC. **Figure 2** provides a graph of conversion efficiency for the three standard output voltages, plotted over a range of input voltages extending to 60V. The circuit is useful in applications requiring higher input voltage, higher current, or both than is available from standard ICs. The LM2594HVN is a buck regulator that switches an internal 0.5A device at 150 kHz. This current suffices to feed the base of Q_1 and the bias resistor, R_2. The function of R_2 is to quickly turn off Q_1, a fast npn switch with a beta greater than 10 at 6A. The purpose of R_1 may not be obvious without some knowledge of the internal workings of the 2594. Its value is such to produce sufficient voltage drop at peak current so Q_1 begins to saturate. The saturation causes Q_1's beta to drop, and, as the transistor's base current rises to more than 0.5A, the 2594 drops into its pulse-by-pulse limited-protection mode, followed by a reduction in clock frequency if the overload is severe.

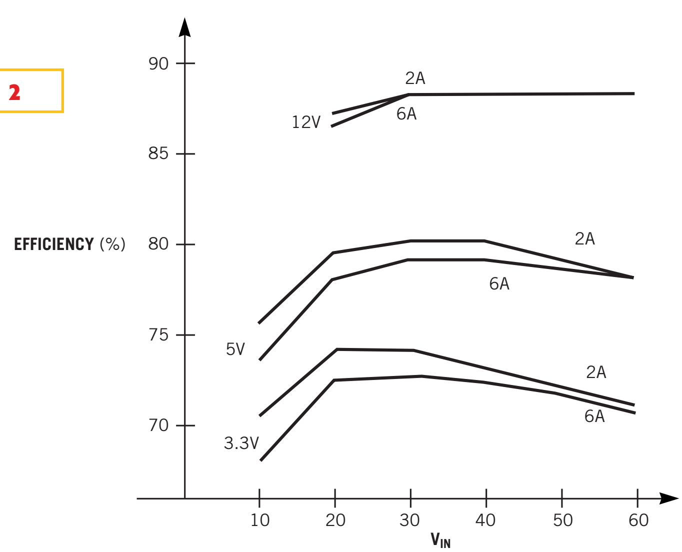
This design example uses through-hole components, because low-ESR capacitors and inductors in through-hole form are inexpensive and easy to find. Worst-case line and load conditions cause Q_1 and D_1 to dissipate 3W each, so you must choose a heat-sink size to keep the temperature rise within acceptable limits. A heat-sink rating of 6 to 7°C/W can accommodate both devices for operation to 85°C ambient temperature. The capacitors are low-ESR types from Nichicon's PL series (www.nichicon-us.com). R_2 dissipates less than 0.25W, but, at full load current, R_1 can dissipate 1W at high V_{IN} and more than 5W at low V_{IN}. You should locate R_1 away from the regulator IC to minimize heating. The DIP version of the 2594 dissipates as much as 0.5W at high V_{IN}; you should solder leads 1, 2, 3, and 6 to a ground-plane area greater than 2 in.^{2} to avoid thermal shutdown. If you don't need the IC's on/off feature, then you should also solder Pin 5 to the ground plane.

Figure 1



Less expensive than a "brick" converter, this circuit accommodates high input voltages and output currents.

Figure 2



The circuit of Figure 1 delivers good efficiency for 15 to 60V inputs.

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Negative resistor cancels op-amp load

Elliott Simons, Maxim Integrated Products, Sunnyvale, CA

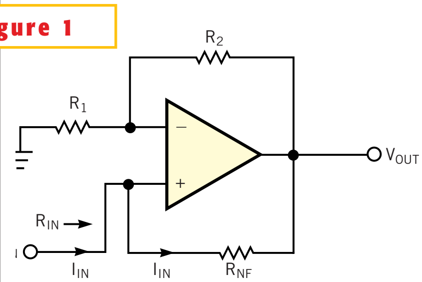
Accurate op amps have high open-loop gain, low offset voltage and current, low voltage and current noise, and low distortion. However, they often lack the ability to provide high output currents while maintaining all the other high-accuracy specifications. In other words, high-accuracy op amps have a problem driving low-impedance loads. One solution to the problem is to "cancel" the load. If your resistive load is RΩ and you connect it in parallel with a negative resistor of −RΩ, the resistance of the parallel combination is infinite. The circuit of **Figure 1** can generate negative resistance at its input: $R_{IN}=-R_{NF}(R_1/R_2)$. You derive this value as follows:

$V_{OUT}=V_{IN}(1+R_2/R_1)$;

$I_{IN}=(V_{IN}-V_{OUT})/R_{NF}=-(V_{IN}/R_{NF})(R_2/R_1)$; and

$R_{IN}=V_{IN}/I_{IN}=-R_{NF}(R_1/R_2)$.

Figure 1

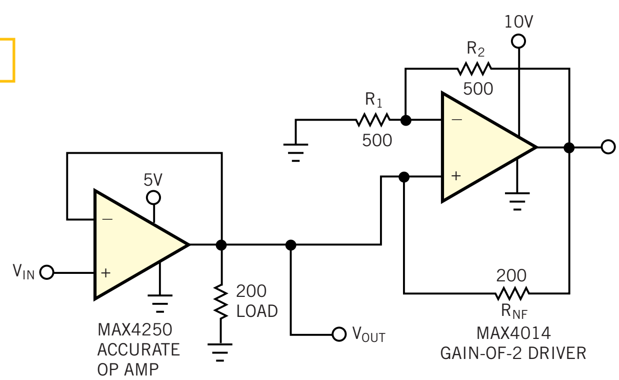


This circuit exhibits negative resistance at its input.

Figure 2 presents a practical application of the concept. The first op amp is an accurate, unity-gain buffer, and the second op amp is a high-current, high-bandwidth, gain-of-2 driver. Because $R_1=R_2$ in this negative-resistance stage, its input resistance is $-R_{NF}=-200\Omega$, which matches the magnitude of the accurate buffer's load resistance. If these magnitudes match perfectly, the buffer sees an open circuit at its output. The buffer drives the positive input of the second amplifier, and the second amplifier, via its negative-resistance input, drives the load. Gain error, output-current limits, and resistor mismatches limit the minimum resistance the circuit can drive, but driving a 200Ω load is easy. That load is an order of magnitude lower than the load the unassisted accurate amplifier can handle without suffering degradation of performance. Note that the second op amp's gain error, offset voltage, and offset current do not affect the first op amp's accuracy. The step response of this circuit is well-behaved and exhibits no ringing.

The negative-resistance approach works equally well with dual-supply op amps, because the negative-resistance portion can both source and sink current. If the driver op amp does not have built-in gain-setting resistors, you can set its noninverting gain closer to unity, thereby allowing both op amps to share a power supply. This approach limits the output swing of the accurate op amp, but that restriction may be acceptable in a given application. To ensure full bandwidth for the accurate op amp, the driver op amp should have much higher bandwidth.

Figure 2



Connecting a negative resistance in parallel with the load enables a precision op amp to drive 200V.

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Op amps make JFET circuits repeatable

Glen Brisebois, Linear Technology Corp, Milpitas, CA

Because they use practically no bias current (a useful feature in itself), JFETs also have practically no current noise. This feature means that you can use JFETs in very-high-resistance circuits and obtain good noise performance. JFETs are also fast devices, with garden-variety types specified in the hundreds of megahertz. On the flip side, JFETs are difficult to exploit in a manufacturing environment because they have widely varying dc specifications. With a simple resistor-based bias network, the same part number can give results that differ by several volts from device to device. One way to use JFETs in a repeatable and manufacturable manner is to use the topology that **Figure 1** shows. The pur-

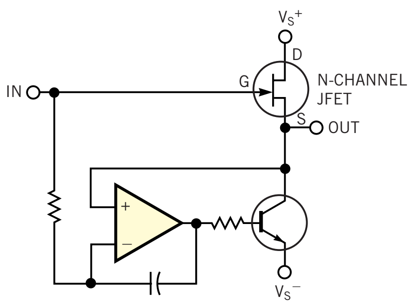

Figure 1

The op amp biases the JFET at I_{DSS}, with V_{GS} 50V.

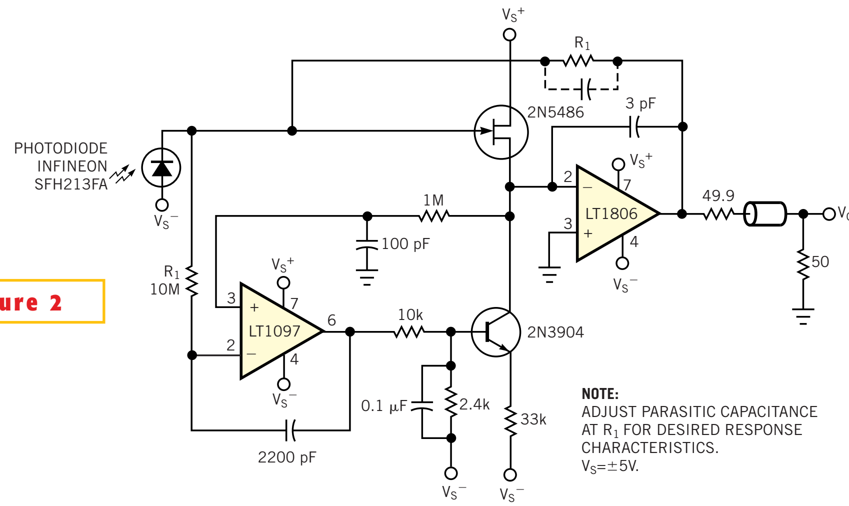

Figure 2

This fast, high-gain photodiode amplifier uses Figure 1's scheme to bias the JFET.

pose of the op amp is to bias the JFET at $V_{GS}=0V$ and, therefore, at $I_D=I_{DSS}$. It meets this goal by increasing the current in the bipolar transistor until $V_{GS}=0V$ and $I_D=I_{DSS}$.

In this condition, the JFET operates at its highest gain (g_m) and lowest voltage-noise condition. The JFET operates as a follower with zero offset. The only requirement for the op amp is that it have ultralow bias current. A variety of op-amp types satisfy this criterion, including JFET-input op amps, such as the LT1462; superbeta-input op amps, such as the LT1097; and micropower op amps, such as the LT1494. **Figure 2** shows an implementation of the topology of **Figure 1**, using an inexpensive, fast 2N5486 JFET. This device specifies I_{DSS} at 8 to 20 mA at room temperature. The LT1097 maintains the gate-source voltage at 0V by adjusting the JFET's drain current. The source of the JFET connects to the inverting input of the 325-MHz, low-noise LT1806 op amp. R_F closes the loop back to the JFET's gate. In this application, the circuit serves as a transimpedance amplifier for a fast photodiode.

Selecting a high value, 10 MΩ for R_1 maintains low noise gain, but you could reasonably reduce it to a few times larger than R_F. The values of the other resistors and capacitors in the LT1097 loop to attenuate the noise and shape the noise bandwidth of the slow loop. Measurements show the output-noise spectral density is $9\ nV/\sqrt{Hz}$ with $R_F=0\Omega$, so resistor noise dominates with R_F greater than approximately 10 kΩ. **Table 1** shows the rise time and bandwidth achieved for several transimpedance gains (as set by R_F). To obtain optimum speed characteristics, you make "parasitic-capacitance adjustments" (the capacitor with broken lines in **Figure 2**) by adjusting the proximity of R_F's leads to its body. **Figure 3** shows the time-domain pulse response with $R_F=1$ MΩ. Connecting two 499-kΩ resistors in series improves the response.

TABLE 1—RESULTS FOR VARIOUS R_F WITH 1.2V OUTPUT STEP

R_F	10 to 90% rise time (nsec)	3-dB bandwidth (MHz)
100 kΩ	64	6.8
200 kΩ	94	4.6
499 kΩ	154	3
1 MΩ	263	1.8

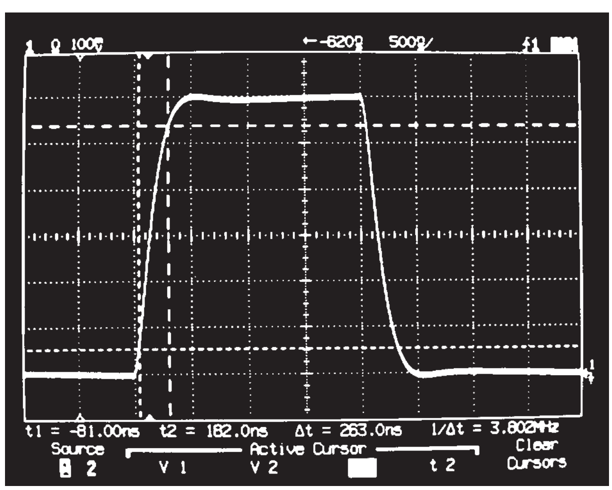

Figure 3

The circuit of Figure 2 exhibits clean pulse response with little overshoot or ringing.

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