Edited by Bill Travis and Anne Watson Swager

Circuit yields accurate absolute values

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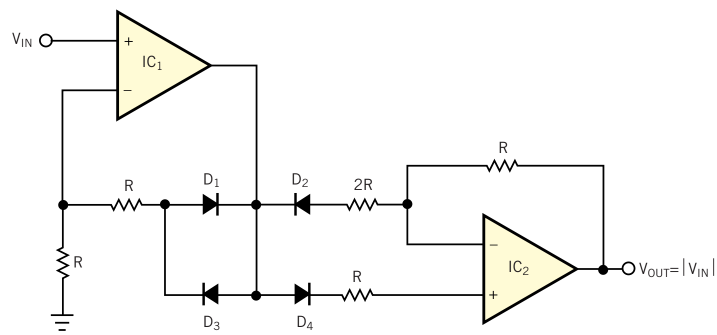
The circuit in Figure 1 delivers the absolute value of the input signal with an accuracy better than 10 ppm of the full-scale range. The circuit has low zero-crossing error. You can use it as an asynchronous demodulator, as a source for logarithmic amplifiers, or simply as a demonstration of the wonders of feedback. The circuit uses two op amps; five identical resistors, R; one double-value resistor, 2R; and four thermally matched diodes. When the input, V_{IN}, is positive, IC_1's output is $2V_{IN}+V_{D3}$ (the voltage drop across D_3). D_2 is reverse-biased, thus IC_2 behaves as a voltage follower, yielding $V_{OUT}=R(V_{IC1}-V_{D4})/2R$. Because the same amount of current (V_{IN}/R) flows in D_3 and D_4, assuming their characteristics are the same, $V_{D3}=V_{D4}$, and $V_{OUT}=V_{IN}$. When V_{IN} is negative, IC_1's output is $2V_{IN}-V_{D1}$. D_4 is reverse-biased, and IC_2 is an inverting amplifier, yielding $V_{OUT}=R(V_{IC1}+V_{D2})/2R$. Again, the current flowing in D_1 and D_2 is V_{IN}/R, and $V_{OUT}=-V_{IN}$. For good performance,

- The six resistors must match closely to guarantee symmetrical gain.
- The diode pairs, D_1-D_2 and D_3-D_4, must have tight thermal coupling to minimize errors at low input voltages. (It's best if the pairs are on the same chip.)
- The op amps must have low offset.

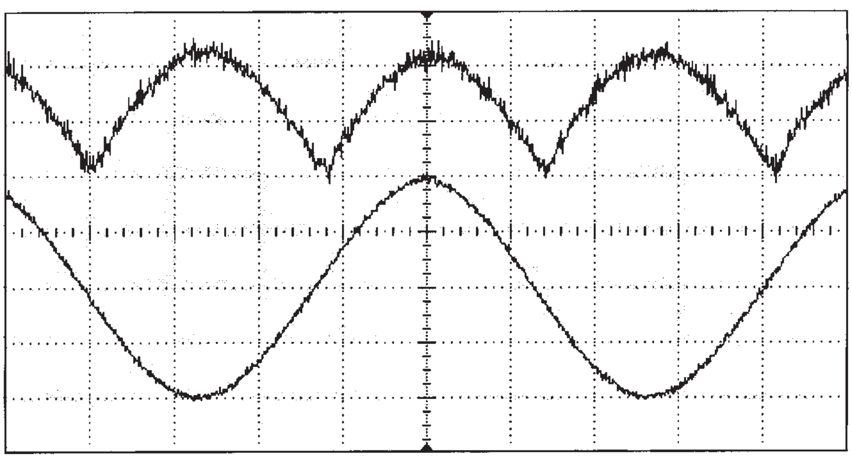
In a practical configuration, you can configure D_1 through D_4 using base-collector junctions of a monolithic transistor array, such as an MPQ6700. The resistors are 10- and 20-kΩ, 1% metal-film units. You can use optional 100Ω trimmers in series with the resistors in the circuit to trim for optimum performance. The op amps are OP27 devices, with their offset trimmed. After adjusting the op amps' offset and tweaking the resistors, the residual error is within 100 μV p-p over the 13V p-p operating range. **Figure 2** shows the behavior of the circuit with a 40-μV p-p input signal (bottom trace).

Figure 1



This full-wave rectifier circuit yields the accurate absolute value of the input signal.

Figure 2



Even with a 40-mV p-p signal, the circuit in Figure 1 yields an accurate absolute value.

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ADC enables temperature-compensated weigh-scale measurements

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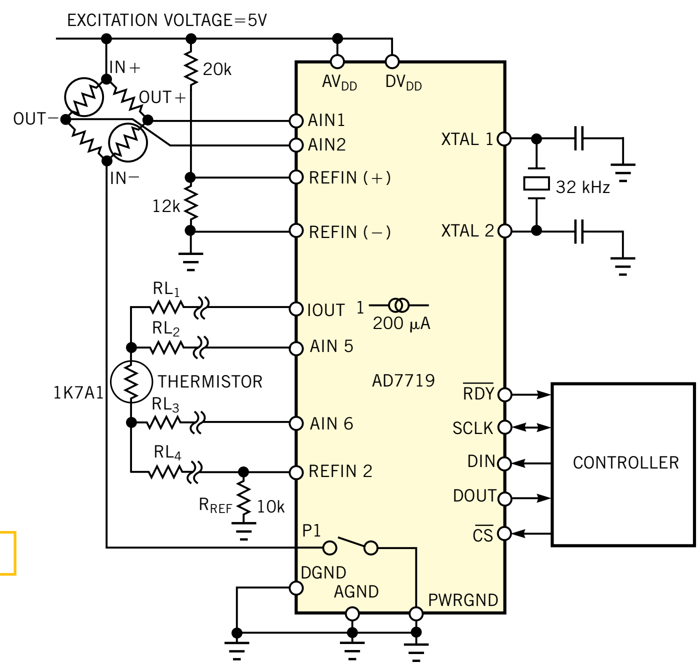
You can provide temperature compensation in weigh-scale applications by simultaneously measuring both the temperature of the bridge and the primary output of the bridge transducer. Traditionally, an integrated multiplexer connects multiple input variables to a single sigma-delta ADC. Each time the multiplexer switches the input, the ADC must flush the digital filter of all data pertaining to the previous channel. Before the new data becomes valid, the system must account for the settling time and latency, reducing the maximum throughput rate. For example, with an ADC containing a second-order sigma-delta modulator and a third-order digital filter, the output-settling time for a step input is three times the period of the data rate. Switching from the primary to the secondary channel can reduce the primary channel's throughput by a factor of six when you need to monitor primary and secondary variables together. In many cases, you can monitor the secondary variable only intermittently and thus minimize the reduction in throughput. **Figure 1** shows a solution to the throughput problem that uses the two independent channels of an AD7719 dual sigma-delta ADC.

Figure 1



Two independent ADCs eliminate the throughput limitations of multiplexed measuring systems.

The ADCs convert in parallel, so you can simultaneously measure both the bridge output and the bridge temperature. The output data from both measurements is available in parallel, thereby removing the latency associated with multiplexed data-acquisition systems. The main channel monitors the bridge transducer, and the secondary channel monitors the bridge temperature. The bridge transducer develops a differential output voltage between the Out(+) and Out(−) terminals. A bridge sensitivity of 3 mV/V produces a full-scale output of 15 mV when a 5V excitation source powers the bridge. The ADC's reference voltage can assume any value between and including the supply voltages, so you can use the bridge-excitation voltage to provide the reference to the ADC. A resistive divider, however, allows you to use the full dynamic range of the input. This implementation is fully ratiometric, so variations in the excitation voltage do not introduce errors in the system.

The resistor values of 20 and 12 kΩ in **Figure 1** yield a 1.875V reference voltage for the AD7719, with a 5V excitation voltage. The main-channel (programmable) gain is 128, resulting in a full-scale input span of the ADC equal to the full output span of the transducer. A low-side switch disables the transducer to save power in standby mode. The AD7719 features factory calibration, and its signal chain uses a chopping scheme to reduce gain and offset drifts, eliminating the need for field calibration. A key requirement in weigh-scale applications is the ability to reject line-frequency components (50 and 60 Hz). You can achieve simultaneous 50- and 60-Hz rejection by programming the AD7719 for an output data rate of 19.8 Hz. With a gain of 128, the ADC achieves 13-bit resolution at this data rate. You can increase the resolution by reducing the update rate or by providing additional digital filtering in the controller.

The secondary channel of the AD7719 monitors the bridge temperature with the aid of a thermistor. An on-chip current source excites the thermistor and generates the reference voltage for the AD7719. As a result, excitation signals do not affect performance, and the configuration is fully ratiometric. The circuit uses a four-wire force/sense configuration to reduce the effects of lead resistance. Lead resistance of the drive wires shifts the common-mode voltage but does not degrade the performance of the circuit. Lead resistance of the sense wires is immaterial because of the high impedance of the AD7719's analog inputs. The reference-setting resistor, R_{REF}, must have a low temperature coefficient. The AD7719 achieves 16-bit resolution in the secondary channel, using a 19.8-Hz update rate. The thermistor determines the operating range of the circuit. The maximum voltage on the auxiliary input is REFIN 2 or 2V. With a Betatherm 1K7A1 thermistor (www.betatherm.com) and 200-μA excitation current, the operating range is −26 to +70°C.

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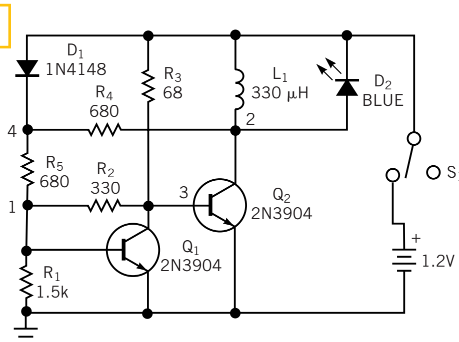
Single cell lights any LED

Al Dutcher, Al Labs, West Deptford, NJ

The circuit in **Figure 1** allows you to light any type of LED from a single cell whose voltage ranges from 1 to 1.5V. This range accommodates alkaline, carbon-zinc, NiCd, or NiMH single cells. The circuit's principal application is in LED-based flashlights, such as a red LED in an astronomer's flashlight, which doesn't interfere with night vision. White LEDs make handy general-purpose flashlights. You can use the circuit in **Figure 1** with LEDs ranging from infrared (1.2V) to blue or white (3.5V). The circuit is tolerant of the varying LED voltage requirements and delivers relatively constant power. It provides compensation for varying battery voltage. The circuit is an open-loop, discontinuous, flyback boost converter. Q_2 is the main switch, which charges L_1 with the energy to deliver to the LED. When Q_2 turns off, it allows L_1 to dump the stored energy into the LED during flyback.

Figure 1



A simple circuit provides drive from a single cell to an LED of any type or color.

Q_1, an inverting amplifier, drives Q_2, an inverting switch. R_4, R_5, and R_2 provide feedback around the circuit. Two inversions around the loop equal noninversion, so regeneration (positive feedback) exists. If you replace L_1 with a resistor, the circuit would form a classic bistable flip-flop. L_1 blocks dc feedback and allows it only at ac. Thus, the circuit is astable, meaning it oscillates. Q_2's on-time is a function of the time it takes L_1's current to ramp up to the point at which Q_2 can no longer stay in saturation. At this point, the circuit flips to the off state for the duration of the energy dump into the LED, and the process repeats. Because inductors maintain current flow, they are essentially current sources as long as their stored energy lasts. An inductor assumes any voltage necessary to maintain its constant-current flow. This property allows the circuit in **Figure 1** to comply with the LED's voltage requirement.

Constant-voltage devices, such as LEDs, are happiest when they receive their drive from current sources. The LED in **Figure 1** receives pulses at a rapid rate. The inductor size is relatively unimportant, because it determines only the oscillation frequency. If, in the unlikely case the inductor value is too large, the LED flashes too slowly, resulting in a perceivable flicker. If the inductor value is too small, switching losses predominate, and efficiency suffers. The value in **Figure 1** produces oscillation in the 50-kHz neighborhood, a reasonable compromise. D_1 provides compensation for varying cell voltage. By the voltage-division action at Node 4, D_1 provides a variable-clipping operation. The higher the supply voltage, the higher the clipping level, and the result is correspondingly less feedback. Q_1 inverts this clipping level to reduce the turn-on bias to Q_2 at higher cell voltages. We chose 2N3904s, but any small-signal npn works. Q_2 runs at high current at the end of the charging ramp. Internal resistance causes its base-voltage requirement to rise. The R_2-R_1 divider at Q_1's base raises the collector voltage to match that requirement and thus controls Q_2's final current.

The LED's drive current is a triangular pulse of approximately 120 mA peak, for an average of approximately 30 mA to a red LED and 15 mA to a white one. These levels give a reasonable brightness to a flashlight without unduly stressing the LED. The supply current for the circuit is approximately 40 mA. A 1600-mAhr NiMH AA cell lasts approximately four hours. L_1 must be able to handle the peak current without saturating. The total cost of the circuit in **Figure 1** is less than that of a white LED. You can use higher current devices and larger cells to run multiple LEDs. In this case, you can connect the LEDs in series. If you connect them in parallel, the LEDs need swamping (ballast) resistors. You can also rectify and filter the circuit's output to provide a convenient, albeit uncontrolled, dc supply for other uses.

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Lowpass filter uses only two values

Richard M Kurzrok, Queens Village, NY

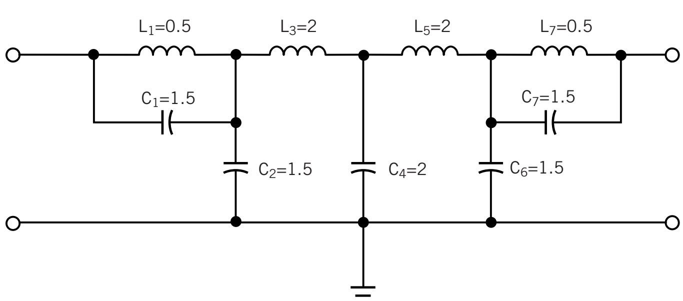
In recent years, image-parameter design of LC filters has received new consideration (**references 1** and **2**). The composite lowpass filter uses interior constant-k full sections terminated by m-derived half-sections. For best passband response, you usually select m to equal 0.6. However, m=0.5 can still give useful filter performance while reducing the number of component values. A low number of component values is advantageous for low-cost manufacturing. The design technique is also applicable to highpass and wideband filters (**references 3** and **4**). **Figure 1** shows a schematic of the composite lowpass filter. The filter uses four inductors of two different values and five capacitors of two different values. **Figure 2** shows the schematic of an equivalent composite lowpass filter. This filter uses judicious combinations of components in series, parallel, and series-parallel. The filter uses eight inductors and 14 capacitors of only one value each.

TABLE 1—AMPLITUDE RESPONSE FOR FILTER IN FIGURE 2

Frequency (MHz)	Insertion loss (dB)	Frequency (MHz)	Insertion loss (dB)
1	0.1	4	5.2
1.5	0.1	4.1	12.7
2	0.15	4.2	20.8
2.5	0.1	4.3	31.2
2.9	0.2	4.4	Greater than 40
3.2	0.25	5	Greater than 40
3.4	0.3	10	Greater than 40
3.6	0.4	15	Greater than 40
3.7	0.45	20	Greater than 40
3.8	0.6	25	Greater than 40
3.9	1.4	30	Greater than 40

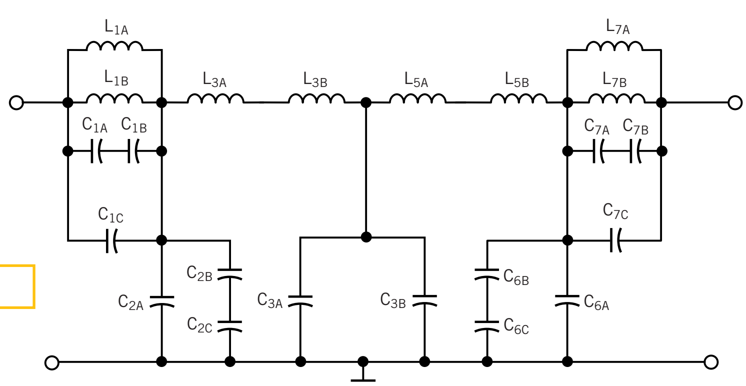
Table 1 gives test results for the lowpass filter of **Figure 2.** We constructed the filter on Vector board in a die-cast aluminum enclosure with BNC input and output connectors. The 3-dB cutoff design frequency is 3.88 MHz with source and load impedances of 50Ω. All capacitors are polypropylene units of 820 pF±5%. All inductors are 2.05 μH and made of 20 turns of number 26 AWG magnet wire on Micro Metals T37-2 toroids. The toroidal inductors, with unloaded Q exceeding 100, provide low passband-insertion loss. Surface-mount inductors with unloaded Q of 10 to 20 yield higher passband losses, which are acceptable in many applications.

Figure 1



A composite lowpass filter uses four component values.

Figure 2



NOTE:
ALL NORMALIZED INDUCTORS AND CAPACITORS EQUAL 1.

By judicious connection of components, this filter uses only one value each for the inductors and capacitors.

References

1. Kurzrok, Richard M, "Low cost lowpass filter design using image parameters," *Applied Microwave & Wireless*, February 1999, pg 72, and correction May 1999, pg 12.
2. Kurzrok , Richard M, "Update the design of image-parameter filters," *Microwaves & RF*, May 2000, pg 119.
3. Kurzrok, Richard M, "Filter design uses image parameters," *EDN*, May 25, 2000, pg 111.
4. Kurzrok, Richard M, "Wideband filter uses image parameters," *EDN*, Oct 26, 2000, pg 174.

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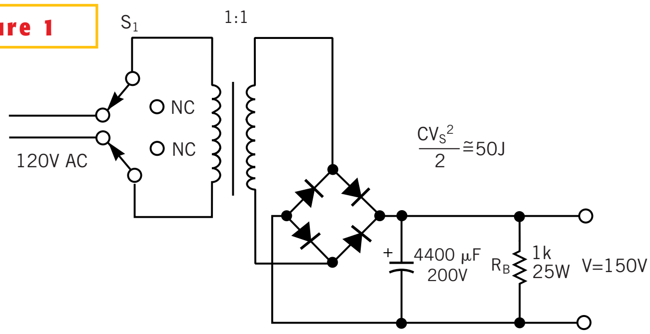
Quickly discharge power-supply capacitors

Stephen Woodward, University of North Carolina, Chapel Hill, NC

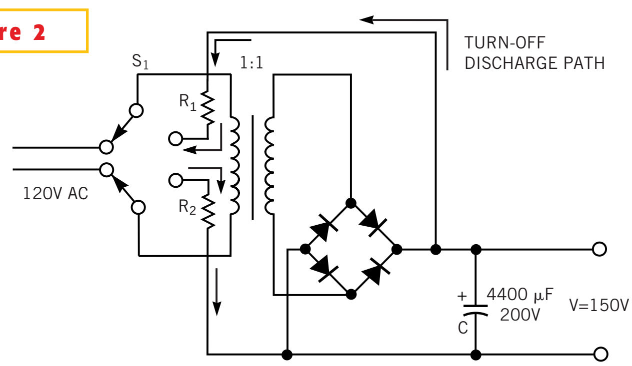
A PERENNIAL CHALLENGE in power-supply design is the safe and speedy discharge, or "dump," at turn-off of the large amount of energy stored in the postrectification filter capacitors. This energy, $CV^2/2$, can usually reach tens of joules. If you let the capacitors self-discharge, dangerous voltages can persist on unloaded electrolytic filter capacitors for hours or even days. These charged capacitors can pose a significant hazard to service personnel or even to the equipment itself. The standard and obvious solution to this problem is the traditional "bleeder" resistor, R_B (**Figure 1**). The trouble with the R_B fix is that power continuously and wastefully "bleeds" through R_B, not only when it's desirable during a capacitor dump, but also constantly when the power supply is on. The resulting energy hemorrhage is sometimes far from negligible.

Figure 1



A bleeder resistor ensures safety but wastes much power.

Figure 2



Otherwise unused switch contacts can dump energy while not wasting power.

Figure 1 offers an illustration of the problem, taken from the power supply of a pulse generator. The $CV^2/2$ energy stored at the nominal 150V operating voltage is $150^2 \times 4400\ \mu F/2$, or approximately 50J. Suppose you choose the R_B fix for this supply and opt to achieve 90% discharge of the 4400-µF capacitor within 10 sec after turning off the supply. You then have to select R_B to provide a constant RC time no longer than 10/ln(10), or 4.3 sec. R_B, therefore, equals 4.3 sec/4400 µF, or approximately 1 kΩ. The resulting continuous power dissipated in R_B is 150^2/1 kΩ, or approximately 23W. This figure represents an undesirable power-dissipation penalty in a low-duty-cycle pulse-generator application. This waste dominates all energy consumption and heat production in what is otherwise a low-average-power circuit. This scenario is an unavoidable drawback of bleeder resistors. Whenever you apply the 10%-in-10-sec safety criterion, the downside is the inevitable dissipation of almost half the $CV^2/2$ energy during each second the circuit is under power.

Figure 2 shows a much more selective and thrifty fix for the energy-dump problem. The otherwise-unused off-throw contacts of the DPDT on/off power switch create a filter-capacitor-discharge path that exists only when you need it: when the supply is turned off. When the switch moves to the off position, it establishes a discharge path through resistors R_1 and R_2 and the power transformer's primary winding. The result is an almost arbitrarily rapid dump of the stored energy, while the circuit suffers zero power-on energy waste. Use the following four criteria to optimally select R_1, R_2, and S_1:

- The peak discharge current, $V/(R_1+R_2)$, should not exceed S_1's contact rating.
- The pulse-handling capability of R_1 and R_2 should be adequate to handle the $CV^2/2$ thermal impulse. A 3W rating for R_1 and R_2 is adequate for this 50J example.
- The discharge time constant, $(R_1+R_2)C$, should be short enough to ensure quick disposal of the stored energy.
- S_1 must have a break-before-make architecture that ensures breaking both connections to the ac mains before making either discharge connection, and vice versa. Otherwise, a hazardous ground-fault condition may occur at on/off transitions.

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