Edited by Bill Travis

Annual Design Ideas Grand Prize Winner

Frequent Design Ideas author takes home annual award

EDN is pleased to announce that Stephen Woodward's design entry is the grand-prize winner of *EDN*'s 2001 Design Ideas competition. Woodward will receive a $1500 check for his July 5 entry, "Quickly discharge power-supply capacitors," which we present again below in case you missed it the first time. Woodward is a frequent *EDN* contributor and previously won the annual award for his first submission to *EDN* in 1974.

The Design Ideas section is among the most popular features in *EDN*, because it embodies the traditional engineering challenge: An individual sees a well-defined and bounded problem; applies a mix of hardware, software, and algorithms; and produces an innovative solution to that problem. In many cases, the problem is unique or requires a solution that is complex and costly.

According to Woodward, "There's nothing quite like the satisfaction that comes from seeing your name in print attached to something that others find a useful idea. I've done over a hundred of these bite-sized articles, and the pleasure has never faded. The ideas that form the basis for my Design Ideas come from two basic sources. Some I develop specifically for publication. Others come from real-life applications. The circuit that won this year is one of these. The rapid and safe discharge of large power-supply filters is a problem I've been bothered by for years. The topology in this year's winning Design Idea is a solution I came up with only after long consideration. It has served me well in dozens of designs."

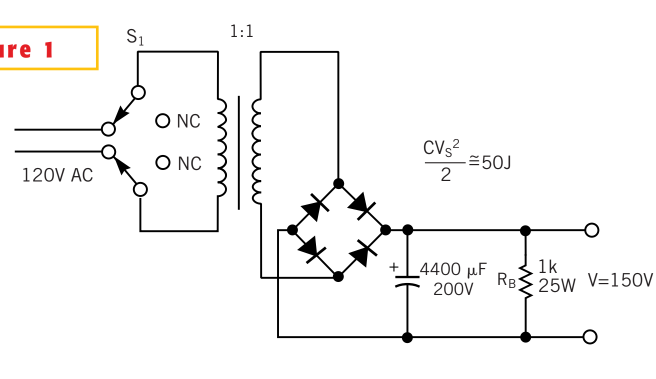
We congratulate Woodward on his achievement.

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.

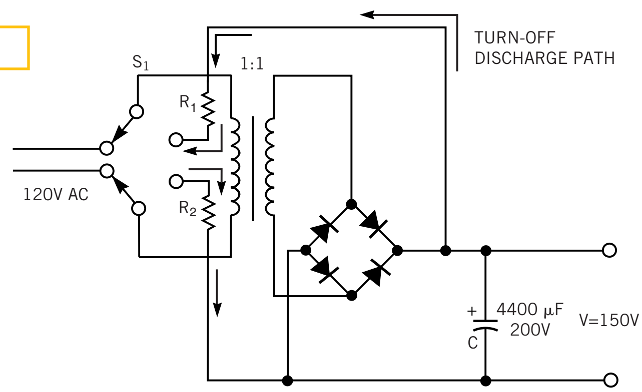
Figure 1



A bleeder resistor ensures safety but wastes much power.

Figure 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 that 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



Otherwise unused switch contacts can dump energy while not wasting 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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High-side current sensor has period output

Greg Sutterlin, Maxim Integrated Products, Wexford, PA

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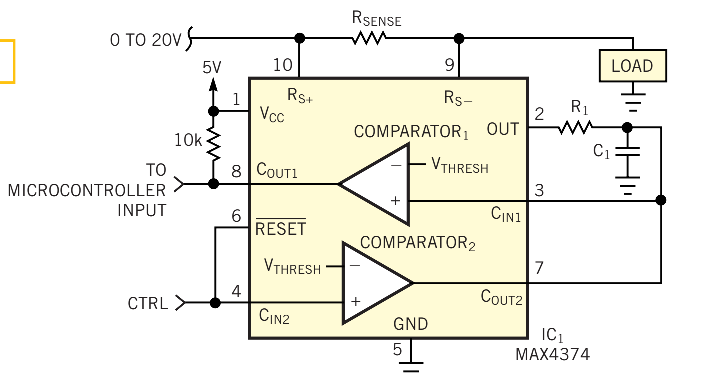
You use high-side current monitoring in many battery-powered products that require accurate monitoring of load current, charger current, or both. In applications for nonportable designs, high-side-current monitoring serves as a power-supply watchdog that can flag a failure in downstream devices. The monitoring can also eliminate hazardous conditions by preventing power-supply overloads. Further, high-side-current monitoring of motor/servo circuits can produce useful feedback in control applications. These applications require a device that converts high-side current directly to a digital signal (**Figure 1**). IC_1 is a low-cost, high-side-current-sense amplifier that converts high-side current to a proportional, ground-referenced voltage. Its two internal comparators (latching and nonlatching) implement a voltage-to-pulse converter that produces an output pulse width proportional to the measured current.

IC_1's Out pin charges C_1 via R_1. When C_1's voltage reaches 0.6V, $Comparator_1$ latches in the high-impedance state. The time required to charge C_1 to 0.6V is proportional to the measured current. $Comparator_2$, in conjunction with the Reset pin, initiates the conversion and removes the previously existing charge on C_1. The Reset and C_{IN2} pins, tied together and connected to a TTL-compatible microcontroller output, CTRL, control the conversion process. Normally, CTRL is

Figure 1



The duration of a negative-going pulse at C_{OUT1} is proportional to the current flowing through R_{SENSE}.

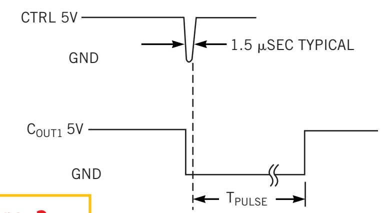


Figure 2

These waveforms illustrate the operation of the circuit in Figure 1.

high. The microcontroller starts a conversion by pulsing CTRL low, discharging C_1 and clearing the latch in comparator$_1$ (C_{OUT1} goes low.) The microcontroller now measures the time from the CTRL transition to the low-to-high transition at C_{OUT1} (**Figure 2**). The period begins at the low-to-high transition of CTRL and ends at the low-to-high transition of C_{OUT1}. As a function of the current levels of interest, you select R_1 and C_1 values to create pulse durations in the tens of milliseconds. As a result, the Out settling time of 20 µsec and the comparator propagation delays of 4 µsec have negligible effects on the measurement accuracy.

To derive an expression for the output pulse width, start with the relationship for an RC-charging circuit: $V_{THRESH}=\text{Out}(1-e^{-TPULSE/R_1C_1})$. For the Out-pin voltage, substitute the expression $I_{LOAD}\times R_{SENSE}\times A_V$ and solve for I_{LOAD}: $I_{LOAD}=V_{THRESH}/[R_{SENSE}\times A_V(1-e^{-TPULSE/R_1C_1})]$, where I_{LOAD}=measured current in amperes, V_{THRESH}=comparator threshold=0.6V, R_{SENSE}=current-sense resistor in ohms, A_V=gain of IC_1, and T_{PULSE}=time to charge C_1 to V_{THRESH} in seconds.

For example, selecting $R_1=1\ M\Omega$, $C_1=0.1\ \mu F$, $R_{SENSE}=0.075\Omega$, and $A_V=20$ produces a T_{PULSE} measurement of 0.022 sec in response to a 2A current. Thus, given a microcontroller timer port, an external interrupt, or simply an available microcontroller input, IC_1, and two external passive components implement high-side-current-to-digital conversion without the need for a discrete A/D converter.

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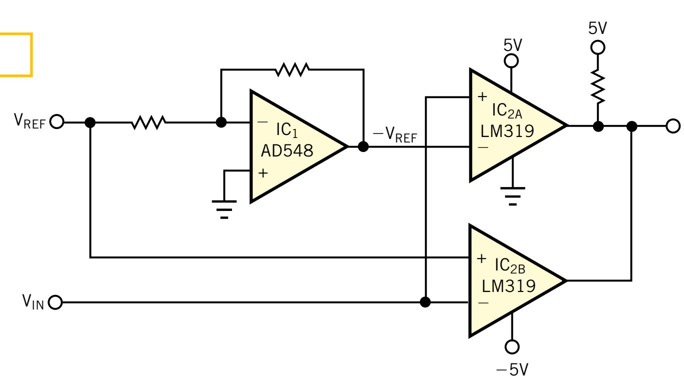
Absolute-value comparator touts accuracy, size

Teno Cipri, Engineering Expressions Consulting, Sunnyvale, CA

A TYPICAL WINDOW COMPARATOR uses two comparators and a single op amp to determine whether a voltage is inside or outside a boundary region. **Figure 1** shows a typical implementation. IC_1 is an inverting op amp with a gain of −1. V_{REF} and $-V_{REF}$ create the window limits. When V_{IN} becomes more positive than V_{REF}, the output of IC_{2A} goes low. When V_{IN} becomes more negative than $-V_{REF}$, the output of IC_{2B} goes low. If V_{IN} is lower than V_{REF} and greater than $-V_{REF}$, both outputs of IC_2 remain high. The LM319 and AD548 come in 14- and eight-pin DIPs, respec-

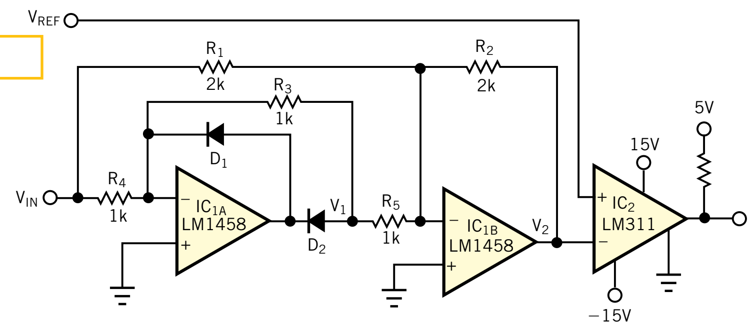
Figure 1



This traditional window-comparator circuit suffers from limited input range.

tively. The LM319 accommodates separate input and output power supplies. The input supply can be ±15V referenced to analog ground, and the output can use a logic supply referenced to logic ground. The circuit's input limit is ±2.5V because of the maximum differential-input limit of the LM319. If you think the input will exceed a ±5V differential voltage between V_{REF} and V_{IN} or $-V_{REF}$ and V_{IN}, then you must incorporate a clamping network with many additional discrete components. The circuit in **Figure 2** overcomes these limitations.

Figure 2



An absolute-value comparator circuit offers a wide input range and improved dc performance.

To increase the maximum differential-input voltage, you can use an LM311, but it is available only in eight-pin packages, so the circuit would require three eight-pin packages. To reduce the chip count, the circuit in **Figure 2** uses an absolute-value amplifier driving a single LM311 comparator, IC_2. Although at first glance the circuit in **Figure 1** may look simpler than the one in **Figure 2**, you can save pc-board area and improve performance by using a dual amplifier in a single eight-pin DIP and an LM311, also in an eight-pin DIP. In **Figure 2**, when the absolute value of V_{IN} exceeds V_{REF}, the output of comparator IC_2 goes low. When V_{IN} is positive, IC_{1A} inverts the signal, and the voltage at R_5 is equal to $-V_{IN}$. The current flowing through R_5 is −2 times that flowing through R_1, and the output of IC_{1B} is equal to V_{IN}. When V_{IN} is negative, D_2 blocks the output of IC_{1A}, which is clamped to the forward voltage of D_1. Because the inverting input of IC_{1A} and IC_{1B} are both at virtual ground, no current flows through R_3 and R_5. With IC_{1A} effectively out of the circuit, IC_{1B}'s gain is −1, and the output voltage is positive. The inverting-input voltage of IC_2 is always at a positive value. This circuit is symmetrical for positive and negative voltages. The following expressions define V_2: $V_2 = |V_{IN}| = -V_{IN} - (-2V_{IN})$ for positive inputs, and $V_2 = -V_{IN}$ for negative inputs.

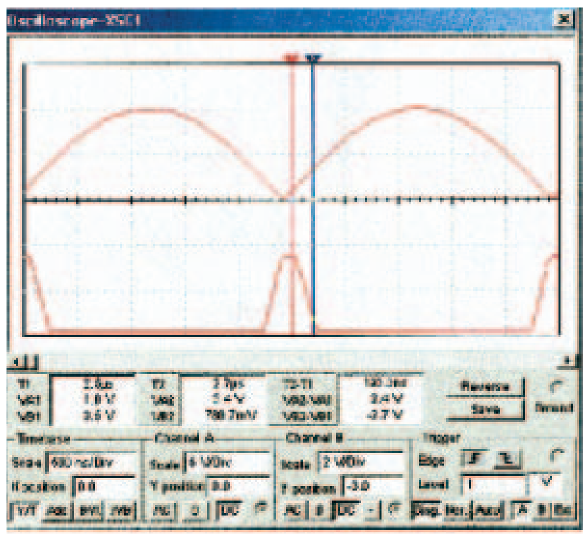
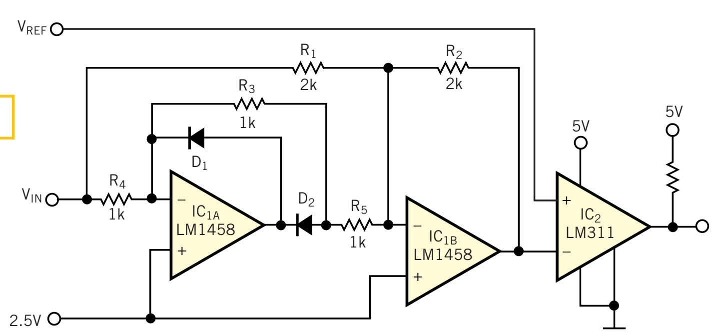


Figure 3

Channel A is the rectified (absolute-value) output of IC_{1B}; Channel B is the output of the comparator.

The dual comparators of **Figure 1** can have slightly different thresholds. You can select the dual op amps in **Figure 2** for input offsets below 1 mV; doing so allows the circuit in **Figure 2** to offer improved dc performance. Another advantage of the circuit in **Figure 2** is the fact that you need change only R_2 to set the gain of the circuit. Most comparators have offset voltages of several millivolts, so scaling up the input voltage improves accuracy by increasing the signal/offset ratio. The circuit in **Figure 1** would require the addition of another op amp to achieve this goal. The simulation in **Figure 3** shows the circuit response. Channel A is the output of IC_{1B}. Channel B is the output of the comparator with V_{REF} set at 1V. Marker 1 corresponds to the 1V threshold, and Marker 2 corresponds to logic low at the output of the comparator. The circuit in **Figure 4** is another variation of the circuit that uses a single 5V supply. It works for input signals of 0 to 5V and V_{REF} of 2.5 and 5V. The 0 and 5V inputs result in the maximum value of 5V at the output of IC_{1B}. With V_{IN} at 2.5V, the output of IC_{1B} assumes the minimum value of 2.5V.

Figure 4



You can modify the circuit in Figure 2 to work with a single supply.

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Programmable oscillator uses digital potentiometers

Alan Li, Analog Devices, San Jose, CA

Digital potentiometers are versatile devices; you can use them in many filtering and waveform-generation applications. This Design Idea describes an oscillator in which setting the resistance of two digital potentiometers independently programs the oscillation amplitude and frequency. **Figure 1** shows a typical diode-stabilized Wien-bridge oscillator that generates accurate sine waves from 10 to 200 kHz. In this classic oscillator circuit, the Wien network comprising R, R′, C, and C′, provides positive feedback, and R_1 and R_2 provide negative feedback. R_2 is the parallel combination of R_{2A} and R_{2B} in series with R_{DIODE}. To establish sustainable oscillation, the phase shift of the loop should be 0° when the loop gain is unity. In this circuit, you can determine the loop gain, A(jω)β(jω) by multiplying the amplifier gain by the transfer function V_P/V_O. With R=R′ and C=C′, the loop gain is

$$A(s)\beta(s)=\frac{1+R_2/R_1}{3+sRC+1/sRC}. \quad (1)$$

Substituting s=jω and rearranging the real and imaginary terms yields

$$A(j\omega)\beta(j\omega)=\frac{1+R_2/R_1}{3+j(\omega RC-1/\omega RC)}. \quad (2)$$

You define the phase angle of the loop gain as

$$0=\tan^{-1}\left[\frac{\mathrm{Im}|A(j\omega)\beta(j\omega)|}{\mathrm{Re}|A(j\omega)\beta(j\omega)|}\right]. \quad (3)$$

You force the imaginary term to zero to set the phase shift to 0°. As a result, the oscillation frequency becomes

$$\omega_0=\frac{1}{RC} \text{ or } f_0=\frac{1}{2\pi RC}, \quad (4)$$

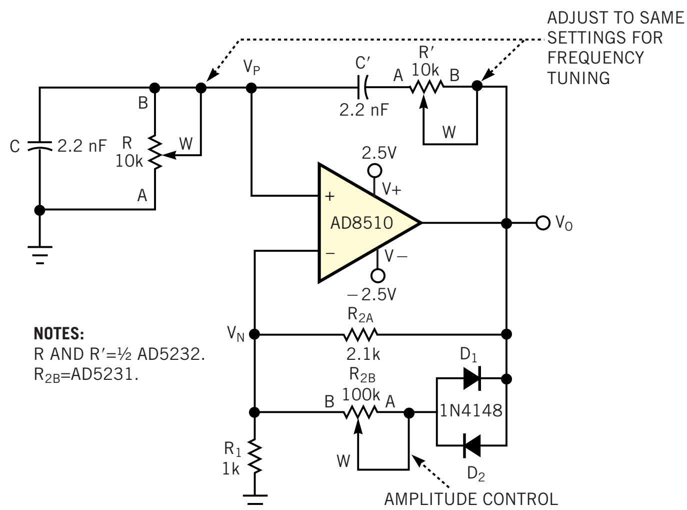
where R is the programmable resistance:

$$R=\frac{256-D}{256}R_{AB}. \quad (5)$$

D is the decimal equivalent of the digital code programmed in the AD5232, and R_{AB} is its end-to-end resistance. To sustain oscillation, the bridge must be in balance. If the positive feedback is too great, the oscillation amplitude increases until the amplifier saturates. If the negative feedback is too great, the oscillation amplitude damps out. As **Equation 2** shows, the attenuation of the loop is 3 at resonance. Thus, setting $R_2/R_1=2$ balances the bridge. In practice, you should set R_2/R_1 slightly higher than 2 to ensure that the oscillation can start. The alternating turn-on of the diodes makes R_2/R_1 momentarily smaller than 2, thereby stabilizing the oscillation. In addition, R_{2B} can independently tune the amplitude, because $2/3(V_O)=I_DR_{2B}+V_D$.

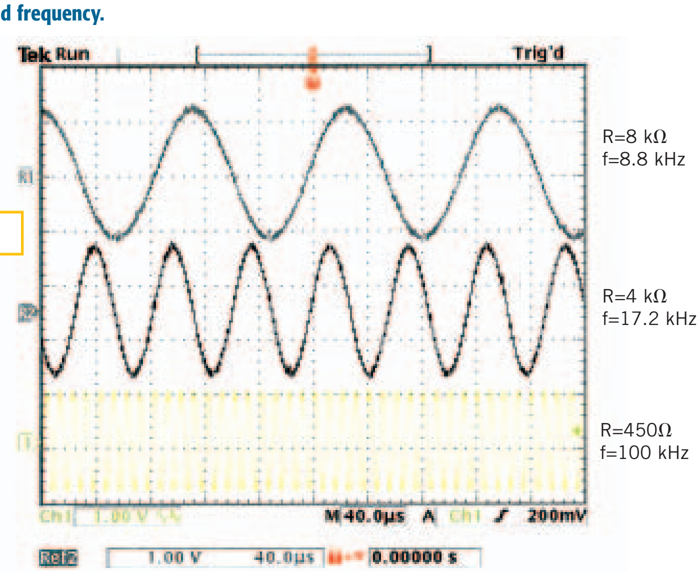
You can short-circuit R_{2B}, which yields an oscillation amplitude of approximately ±0.6V. V_O, I_D, and V_D are inter-

Figure 1



This Wien-bridge oscillator uses digital potentiometers to provide independent settings of amplitude and frequency.

Figure 2



These three frequencies reflect three digital-potentiometer settings.

dependent variables. With proper selection of R_{2B}, the circuit can reach equilibrium such that V_O converges. However, R_{2B} should not be large enough to saturate the output. In this circuit, R_{2B} is a separate 100-kΩ digital potentiometer. As the resistance varies from the minimum value to 35 kΩ, the oscillation amplitude varies from ±0.6 to ±2.3V. Using 2.2 nF for C and C′, and a 10-kΩ dual digital potentiometer with R and R′ set to 8, 4, and 0.7 kΩ, you can tune the oscillation frequency to 8.8, 17.6, and 100 kHz, respectively (**Figure 2**). The frequency error is ±3%. Higher frequencies are achievable with increased error; at 200 kHz, the error becomes ±6%. Two cautionary notes are in order: In frequency-dependent applications, you should be aware that the bandwidth of the digital potentiometer is a function of the programmed resistance. You must therefore take care not to violate the bandwidth limitations. In addition, the frequency tuning requires that you adjust R and R′ to the same setting. If you adjust the two channels one at a time, an unacceptable intermediate state may occur. If this problem is an issue, you can use separate devices in daisy-chain mode, enabling you to simultaneously program the parts to the same setting.

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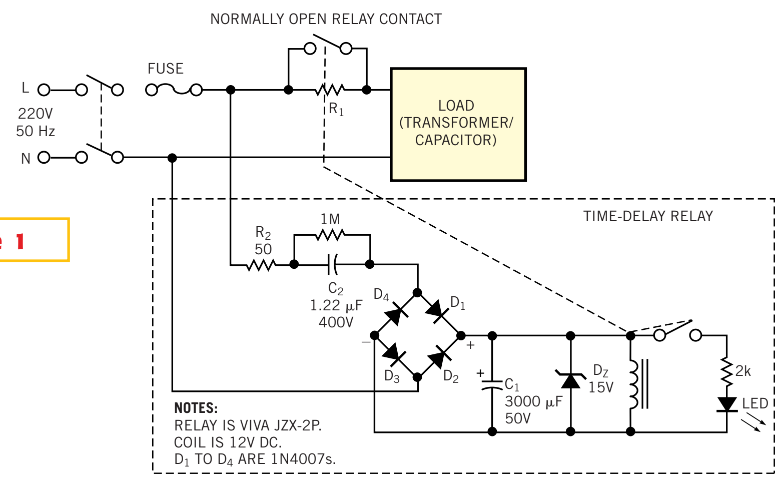
Time-delay relay reduces inrush current

P Seshanna, Assumption University, Bangkok, Thailand

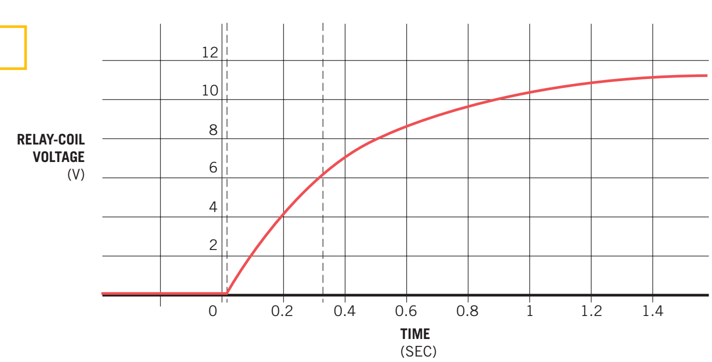
A transformer switching onto a line can sometimes cause a circuit breaker to trip or a fuse to blow. This phenomenon occurs even if the transformer presents no load, such as when the secondary is open. The problem arises because of the heavy magnetizing inrush current in the transformer. The amplitude of the current depends on the instant on the ac waveform at which the transformer becomes energized. The inrush current is at its maximum value if the transformer becomes energized when the ac waveform goes through its zero-crossing point. A similar situation exists when a capacitor in a power-factor-improvement bank switches onto the line. In this case, the inrush current is at its maximum when the ac waveform goes through its peak value. Normally, a mechanical contactor effects the switching without any control of the instant of switching. The inrush current dies down exponentially to the normal operating value of the load within a few cycles. If the breaker trips from the initial inrush current, you close the breaker to re-energize the circuit and hope for the best. You can limit inrush current by inserting a series resistor, R_1, during switching and then shorting this resistor after the transient period (**Figure 1**).

Figure 1



A time-delay-relay circuit can eliminate annoying circuit-breaker trips and blown fuses.

Figure 2



Only after 330 msec does the time-delay relay short-circuit the resistor in series with the load.

Lately, designers have been inserting negative-temperature-coefficient thermistors in series with some loads, such as switch-mode power supplies. This device presents a high resistance at the instant of switching, thus limiting the inrush current. After a few cycles, the resistance of the thermistor drops to a low value, allowing normal operation of the load. In contrast, the circuit in **Figure 1** physically inserts a resistor in series with the load to limit the inrush current and then short-circuits the resistor after a time delay. You can adapt the circuit to any size load by suitably selecting the series resistor and the relay-contact rating. A drawback of negative-temperature-coefficient thermistors is their limited joule heat-absorption capacity. The circuit in **Figure 1** works directly from the ac line to which the load drawing inrush current is connected.

The steady-state dc-current requirement of the relay coil determines the values of the other circuit components. You select capacitor C_2 such that the average value of the rectified current, I_{AVE}, is equal to the current the relay coil requires. The coil resistance should be smaller than the capacitive reactance of C_2 at line frequency. Under these conditions, the average rectified current is approximately $I_{AVE}=V(2\pi fC_2)/1.11$, where V is the rms value of the line voltage (220V), f is the line frequency (50 Hz), and C_2 is the required capacitor value. Once you know the relay current, you can select capacitor C_2 and the bridge diodes. The value of capacitor C_1 determines the delay time.

The voltage across C_1 rises exponentially with a time constant, $\tau=R_LC_1$ (**Figure 2**). If you know the relay's pickup voltage and its coil resistance, R_L, you can choose the required value for C_1. It is easy to see that when you close the main switch, the circuit simultaneously energizes the load drawing inrush current and the time-delay relay. A constant average-current source drives the capacitor/relay combination, and the dc voltage rises exponentially. When this voltage reaches the pickup voltage of the relay, the relay's normally open contact across the series resistor closes, thereby short-circuiting the resistor. When you open the main switch, the voltage across the relay coil drops, again exponentially. When this voltage reaches the dropout voltage of the relay, the contact opens. The resistor is again in series with the load and ready for the next switching operation. The pickup voltage of the 12V relay in the test is approximately 6V, and the contact-closure time is 330 msec, as the dashed line in **Figure 2** shows.

The important design considerations are as follows:

- The normal operating voltage of the relay must be less than 10% and must be lower than the ac-line voltage.
- C_2 determines the average operating current through the relay.
- The relay's contact rating must be adequate to meet the load-current requirement. The relay in **Figure 1** is small and has 12V-dc coil rating and 220V-ac, 5A contact rating. The measured coil resistance is approximately 160Ω.

The function of the 50Ω resistor, R_2, in **Figure 1** is to limit the switch-on surge current into the time-delay-relay circuit. The zener diode, D_Z, limits the voltage rise across C_1 to 15V in the event of a relay-coil open circuit. You can use the circuit in **Figure 1** in the laboratory for energizing a 220V, 1-kVA transformer for use in experiments.

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Optimize linear-sensor resolution

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

A wide variety of sensors and transducers of physical phenomena work via the mechanism of variable resistance. These devices sense temperature, light, pressure, humidity, conductivity, and force, for example. Such sensors measure the physical parameter of interest by reading out the inherent parameter-sensitive resistance or conductance. All resistive sensors share the need for interface circuitry that provides a suitable source of excitation current and appropriate gain and offset of the resulting parameter-dependent voltage. The circuit in **Figure 1** and the design equations provide a universal solution for an application-specific, optimum-resolution, ratiometric interface of almost any resistive sensor. The technique works with any unipolar ADC with an externally accessible full-scale reference. When you configure it properly, the interface circuit selectively maps the range of interest of sensor output resistance (R_{MIN} to R_{MAX}) onto the full-scale span of the ADC. The circuit thus optimally uses available resolution that might otherwise be wasted on sensor resistances that lie outside the range of a given application.

The circuit works as follows: R_1 sources the sensor excitation/bias current, I_B. The op amp boosts and offsets the resulting sensor voltage, I_BR_T, as a function of the R_1-R_2-R_3 network. When $R_T=R_{MIN}$, $V_{OUT}=0V$, and, when $R_T=R_{MAX}$, $V_{OUT}=V_{REF}$. Thus, $R_T=(V_{OUT}/V_{REF})(R_{MAX}-R_{MIN})+R_{MIN}$. Positive feedback via R_4 cancels the effect of voltage variation across R_1 and thus maintains constant-current excitation of the sensor throughout the R_{MIN} to R_{MAX} range. You select R_1 through R_4 as follows: First, select a value for I_B. Sometimes, sensor limitations determine an appropriate value for I_B. Self-heating errors, for example, may limit the maximum excitation current you can apply to temperature sensors such as thermistors and RTDs. But if the given sensor is indifferent to the magnitude of I_B, then you'll obtain optimum tolerance of op-amp offset and gain errors with

$I_B=V_{REF}/(R_{MAX}+R_{MIN})$. With I_B known, you can compute the required gain: $G=V_{REF}/(R_{MAX}-R_{MIN})$. Then,

$R_1=(Z-R_{MAX})(1+1/G)$.

$R_2=Z-R_1$.

$R_3=(G-1)(R_1R_2/Z)$.

$R_4=GR_1$, where $Z=V_{REF}/I_B$.

For a practical example of how this circuit can be handy in a real application, consider a system in which you use a 100Ω platinum RTD to sense temperatures in the range of 25 to 50°C with 0.1° resolution. The corresponding resistance range is 109.73 to 119.4Ω. The 0.39Ω/°C temperature coefficient would require at least 12 bits of conversion resolution without scale expansion. But you can make an 8-bit ADC suffice using the circuit in **Figure 1** with the values shown. The calculations are as follows: I_B=1 mA limits self-heating power to an acceptable 120 μW. Assuming V_{REF}=2.5V, Z=2.5V/1 mA=2500Ω. G=2.5/9.67/1 mA=258.5. It therefore follows that

$R_1=(2500-119.4)(1+1/258.5)=2390\Omega$.

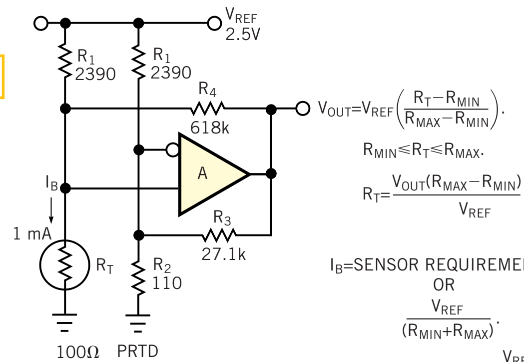
$R_2=2500-2390=110\Omega$.

$R_3=(257.5)(2390\times110/2500)=27{,}100\Omega$.

$R_4=258.5\times2390=618{,}000\Omega$.

$R_T=9.67(V_{OUT}/V_{REF})+109.73$.

Figure 1



$V_{OUT}=V_{REF}\left(\frac{R_T-R_{MIN}}{R_{MAX}-R_{MIN}}\right)$.

$R_{MIN}\le R_T\le R_{MAX}$.

$R_T=\frac{V_{OUT}(R_{MAX}-R_{MIN})}{V_{REF}}+R_{MIN}$.

I_B=SENSOR REQUIREMENTS,
OR

$\frac{V_{REF}}{(R_{MIN}+R_{MAX})}$.

$\Delta V=I_B(R_{MAX}-R_{MIN})$; $G=\frac{V_{REF}}{\Delta V}$; $Z=\frac{V_{REF}}{I_B}$.

$R_1=(Z-R_{MAX})\left(1+\frac{1}{G}\right)$.

$R_2=Z-R_1$

$R_3=(G-1)\left(\frac{R_1R_2}{Z}\right)$.

$R_4=GR_1$.

NOTES:
R_{MIN}=109.73Ω (25°C).
R_{MAX}=119.40Ω (50°C).

This circuit allows you to optimize resolution for sensors using resistive elements.

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