

Edited by Bill Travis

Video multiplexer uses high-speed op amps

Bruce Carter, Texas Instruments, Dallas, TX

VIDEO multiplexers route video from several sources to a single channel. Low-end consumer products use CMOS analog switches and multiplexers, such as the 4066 and 4051. Unfortunately, these devices have a series on-resistance that ranges from approximately 100Ω to 1 kΩ, a resistance that is not constant with video level and that appears in series with the signal. The traditional way of solving this problem is by buffering the analog-switch outputs with transistor stages. With this approach, the characteristics of the CMOS switch and the buffer

stage degrade video performance. However, if you forget the multiplexing action for a moment and consider just the

buffer-amplifier function, you will see that a better approach exists. It must present high enough input impedance to the switch that a 1-kΩ switch resistance is inconsequential and that variation in resistance of the switch with IRE (Institute of Radio Engineers) level produces no lumi-

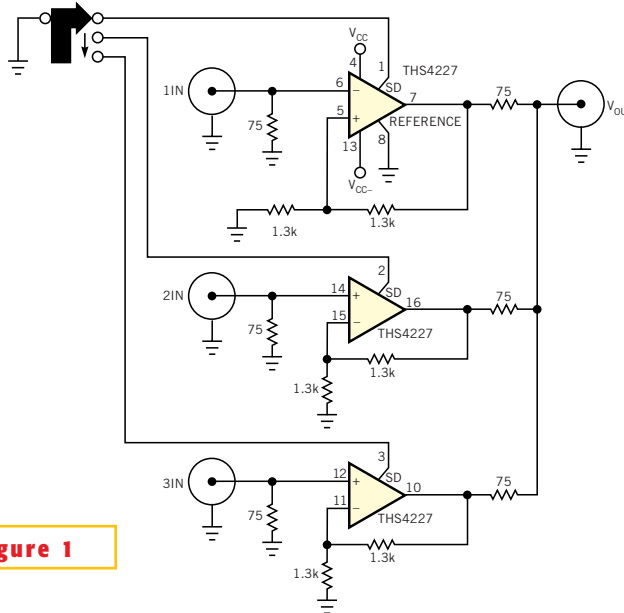


Figure 1

High-speed video op amps make ideal video multiplexers, devoid of video distortion or other artifacts.

nance shifts. **Figure 1** shows a configuration using high-speed op amps in a video-multiplexing application.

High-speed op amps have plenty of bandwidth for video applications. By using an op amp that has 20 or more times the video bandwidth, roll-off and phase shift at 6 MHz are negligible. An op amp has high input impedance in the noninverting mode. You can terminate it for 75Ω input impedance by connecting a simple resistor. Two equal resistors create a gain of two in the noninverting configuration. The gain compensates for a 75Ω back-termination resistor on the op amp's output. The overall stage gain is therefore one. Now, consider the multiplexing function. Some video op amps have a “power-down” feature. This feature allows disabling the output of the op amp, producing a 0V (0 IRE) black level on its output. Its output can therefore connect in parallel with the outputs of other op

amps. This feature enables the op amp to operate as a video multiplexer. The multiplexer in **Figure 1** shows a three-position, single-pole rotary switch. This switch could be a “break-before-make” or an electronic switching system, perhaps with an intelligent infrared interface.

To test the quality of signals passed through the video multiplexer, this design uses the Lucasfilm THX (www.thx.com) test patterns on one video input and a high-quality NTSC program source on another input. When the contrast/picture test goes through the video multiplexer as the active source, the presence of the op amp as a buffer has no effect on black and white levels. No bleeding or blooming

occurs. Any crosstalk results in a visible brightening of the center of the picture in the video-program source, but none occurs. You set the brightness level with the video multiplexer, not in the circuit. Then, you insert the video buffer into the signal path. You'll find that brightness level does not change. The brightness setup test is also an ideal way to test for crosstalk between two video channels. Crosstalk would show up on the black background as a “ghost” image of the program material on the inactive channel; however, none occurs. No color shifts appear in the SMPTE (Society of Motion Picture and Television Engineers) bars with or without the video multiplexer in the signal chain. The color-bar patterns would also produce color shifts in the other channel if crosstalk were a factor. Human skin is the toughest color to get “right,” and any change in skin tone arising from color crosstalk is apparent. No flesh-tone color shifts occur in the test. □

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Single resistor tunes lowpass filter

Vladimir Tepin, State Radioengineering University, Taganrog, Russia

ANY TUNABLE, second-order, active RC-filter section requires at least two thoroughly matched variable resistors. But the lowpass implementation in **Figure 1** provides for wide-range cutoff-frequency control using only a single variable resistor, R. In addition to the resistor, this filter comprises an operational amplifier, IC₂, which serves as a unity-gain buffer; two capacitors, C₁ and C₂; and a single-pole, double-throw analog switch, IC₁, driven by a periodic sequence of square-wave switching pulses applied to the SW input. Thanks to the high-fre-

quency periodic switching, you can simultaneously control the time constants of both C₁ and C₂ in their recharging processes using only R. The approximate voltage-transfer function of the filter, assuming that the switching frequency is much higher than the filter's cutoff frequency, is:

$$H(s) = \frac{1}{\left(\frac{s}{\omega_p}\right)^2 + \left(\frac{s}{\omega_p Q}\right) + 1}$$

where $\omega_p = 1/(R\sqrt{C_1 C_2 \Theta(1-\Theta)})$ is the pole frequency; $Q = \sqrt{\Theta(1-\Theta)}C_1/C_2$ is the quality factor; $\Theta = \tau/T$ is the online time ratio (duty cycle); π is the pulse width; and T is the switching period.

Obviously, controlling R results only in variations of pole frequency and does not affect the quality factor. So, you can tune this filter over a wide frequency range,

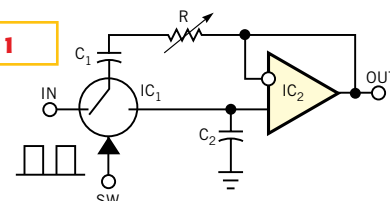


Figure 1

This second-order lowpass filter uses a single resistor to control cutoff frequency.

preserving its passband gain and ripple. You can achieve a stable value of Θ by using a binary counter. A high-resolution, digitally programmable potentiometer is probably the most appropriate choice for R in this filter. **Figure 2** shows the filter's frequency response, simulated in PSpice. This design tunes the cutoff frequency over 20 Hz to 20 kHz by varying the resistor value from 1.2 M Ω to 1.2 k Ω , with C₁=10 nF, C₂=1 nF, Θ =0.5, and a switching frequency of 500 kHz. Using this method, you can also implement a high-order lowpass filter by cascading second-order sections or by joining them to multiple-feedback structures. □

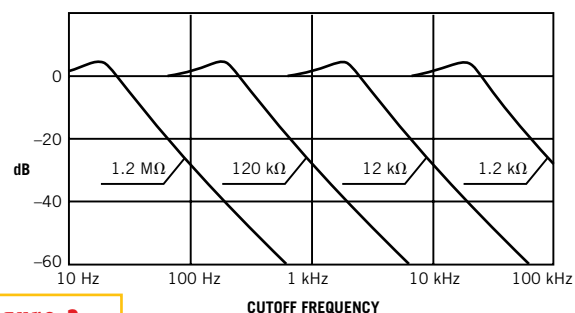


Figure 2

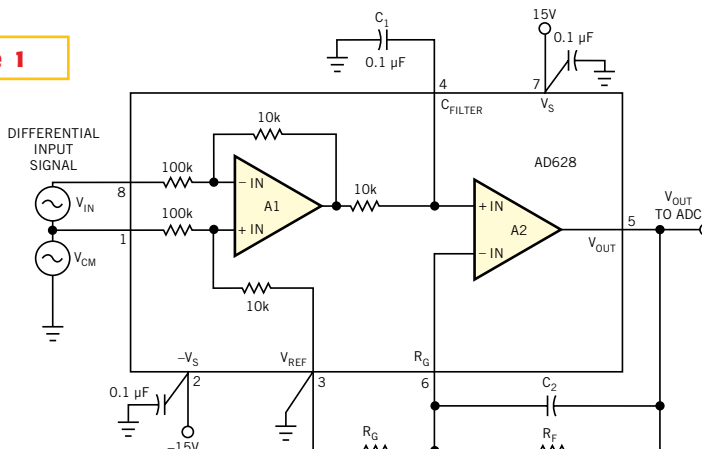
By varying one resistor, you can adjust the cutoff frequency over 20 Hz to 20 kHz.

Simple circuit provides precision ADC interface

Moshe Gerstenhaber and Charles Kitchin, Analog Devices, Wilmington, MA

REAL-WORLD MEASUREMENT requires the extraction of weak signals from noisy sources. High common-mode voltages are often present even in differential measurements. The usual approach to this problem is to use an op amp or an instrumentation amplifier and then perform some type of low-pass-filtering to reduce the background noise level. The problems with this traditional approach are that a discrete op-amp circuit has poor common-mode rejection, and its input voltage range is always lower than the power-supply voltage. When you use a differential signal source with an instrumentation-amplifier circuit, using a monolithic IC can

Figure 1



An instrumentation-amplifier IC provides precision drive and lowpass filtering for an ADC input.

greatly improve common-mode rejection. However, a standard instrumentation amplifier cannot handle sources greater than the power-supply voltage or signals riding on high common-mode voltages. Instrumentation amplifiers using a single external gain resistor also suffer from gain drift. In addition, lowpass filtering requires the use of a separate op amp along with several external components. This approach uses up valuable board space. The circuit of **Figure 1** overcomes all of these performance limitations on one μ SOIC.

An AD628 precision-gain-block IC is configured as a differential-input amplifier and a two-pole lowpass filter. This circuit can extract weak signals riding on common-mode voltages as high as $\pm 120\text{V}$. The precision-gain block directly drives an ADC. A separate V_{REF} pin is available for offsetting the AD628 output signal so that it is centered in the middle of the ADC's input range. Although **Figure 1** indicates $\pm 15\text{V}$, the circuit can operate with ± 2.25 to $\pm 18\text{V}$ dual supplies. The V_{REF} pin can also allow single-supply operation; for this purpose, you simply bias V_{REF} at $V_S/2$. The gain block has two internal amplifiers: A1 and A2. Pin 3 connects to ground, thus operating amplifier A1 at a gain of 0.1. The output of A1 directly drives the positive input of amplifier A2. The first pole of the lowpass filter is a

TABLE 1—COMPONENT VALUES FOR 10V P-P FULL-SCALE INPUT FOR A TWO-POLE LOWPASS FILTER

	200 Hz	−3-dB corner frequency 1 kHz	5 kHz	10 kHz
Capacitor C_2	0.01 μF	0.002 μF	390 pF	220 pF
Capacitor C_1	0.047 μF	0.01 μF	0.002 μF	0.001 μF

Note: Output is 5V p-p; $R_F=49.9\text{ k}\Omega$, and $R_G=12.4\text{ k}\Omega$.

TABLE 2—COMPONENT VALUES FOR 20V P-P FULL-SCALE INPUT FOR A TWO-POLE LOWPASS FILTER

	200 Hz	−3-dB corner frequency 1 kHz	5 kHz	10 kHz
Capacitor C_2	0.02 μF	0.0039 μF	820 pF	390 pF
Capacitor C_1	0.047 μF	0.01 μF	0.002 μF	0.001 μF

Note: Output is 5V p-p, $R_F=24.3\text{ k}\Omega$, and $R_G=16.2\text{ k}\Omega$.

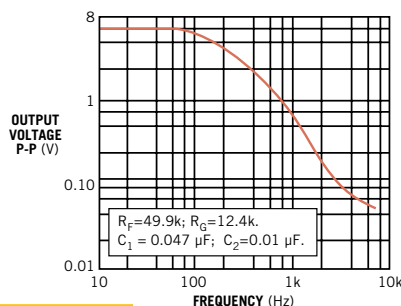


Figure 2

The frequency response of the two-pole filter in the circuit of **Figure 1** shows a sharp roll-off between the corner frequency and approximately 10 times the corner frequency.

function of the internal 10-k Ω resistor at the output of A1, and an external capacitor, C_1 . The gain of A2 is a function of external resistors R_F and R_G . An external RC time constant in the feedback of A2

creates the second pole. This time constant comprises capacitor C_2 across resistor R_F . Note that this second pole provides a more rapid roll-off of frequencies above its RC “corner” frequency ($1/(2\pi RC)$) than does a single-pole lowpass filter. However, as the input frequency increases, the gain of amplifier A2 eventually drops to unity and does not decrease. So, the ratio of R_F/R_G sets the voltage gain of amplifier A2 at frequencies below its −3-dB corner and unity gain at higher frequencies.

Figure 2 is a graph of the filter's output versus frequency using components to provide a 200-Hz, −3-dB corner frequency. Note the sharp roll-off between the corner frequency and approximately 10 times the corner frequency. Above this point, the second pole starts to become less effective, and the rate of attenuation is close to that of a single-pole response. **Tables 1** and **2** provide typical component values for various −3-dB corner frequencies and two full-scale input ranges. The values are rounded off to match standard resistor and capacitor values. Capacitors C_1 and C_2 must be high-Q, low-drift units; avoid low-grade disc ceramics. High-quality NP0 ceramic, Mylar, or polyester-film capacitors offer the best drift characteristics and settling time. □

Buck regulator operates without a dedicated clock

Robert Bell, National Semiconductor, Phoenix, AZ

MOST SWITCHING REGULATORS rely on a dedicated clock oscillator to determine the switching frequency of operation. A dedicated oscillator circuit within the power controller usually generates the clock signal. A class of hysteretic switching regulators can actually operate at a relatively fixed frequency without a clock, even with changing

input-line and output-loading conditions. **Figure 1** shows a simplified buck regulator operating in continuous-conduction mode. (The inductor current always remains positive.) The output voltage, V_{OUT} , is equal to DV_{IN} , where D is the duty-cycle ratio of buck switch Q_1 and V_{IN} is the input voltage. The duty cycle, D , at fixed-frequency operation is T_{ON}/T_S ,

where T_{ON} is the on-time of Q_1 and T_S is the switching-frequency period, $1/F_S$. Some rearranging and substitution leads to the expression $D=V_{\text{OUT}}/V_{\text{IN}}=T_{\text{ON}}/(1/F_S)=T_{\text{ON}}F_S$.

Now, look at a regulator circuit, which, rather than using a fixed clock and a PWM, uses a circuit that turns on Q_1 for a time, T_{ON} , that's inversely proportional

to the input voltage, V_{IN} . **Figure 2** shows a regulator based on this principle. This regulator does not contain a clock oscillator, yet it remains at a fixed operating frequency even while the input voltage varies from 14 to 75V. The two main regulation blocks within this regulator are the on-timer and the regulation comparator. The comparator monitors the output voltage. If the output voltage is lower than the target value, the comparator enables the output switch, Q_1 , for a period of time that the on-timer determines. The time period of the on-timer is $T_{ON} = KR_{ON}/V_{IN}$, where K is a constant (1.3×10^{-10}), R is a configuration resistor, and V_{IN} is the in-

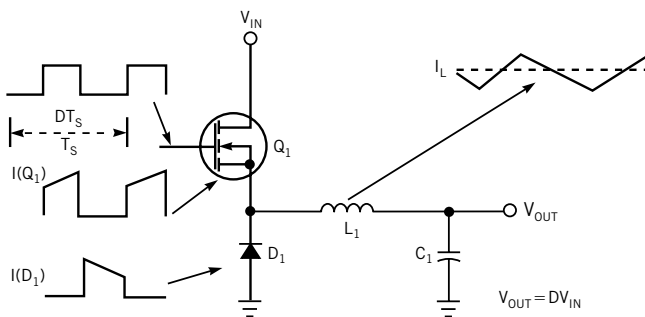


Figure 1

In a typical buck regulator, the output is the switching duty cycle times the input.

put voltage. If you now substitute T_{ON} in the previous buck-regulator equations, an interesting result occurs: $V_{OUT}/V_{IN} = F_s KR_{ON}/V_{IN}$. If you solve for F_s , you obtain $F_s = V_{OUT}/KR_{ON}$. Because V_{OUT} remains regulated and the K and R_{ON} terms

are constants, the switching frequency also remains constant. The constant-frequency relationship holds true provided that the inductor current remains continuous. At lighter loading, the current in the inductor becomes discontinuous. (The inductor

current is zero for some portion of the switching cycle.) At the onset of discontinuous operation, the switching frequency begins to decrease. This reduction is a desirable feature to maintain high efficiency as the load decreases, because switching losses greatly decrease at lower switching frequencies. You derive the switching frequency in discontinuous mode as follows: The peak inductor current $I_p = V_{IN} T_{ON}/L = V_{IN} KR_{ON}/LV_{IN} = KR_{ON}/L$, where L is the output-inductor value. The output power is $P_{OUT} = V_{OUT}^2/R_{OUT} = LI_p^2 F/2 = FK^2 R_{ON}^2/L$. Solving for F : $F = (V_{OUT}^2 L)/(R_{OUT} K^2 R_{ON}^2)$. As you can see, the switching frequency varies inversely with the output resistance, R_{OUT} .

Fixed-frequency operation without an oscillator offers a low-cost, easy-to-implement step-down regulator. You needn't worry about any loop-compensation or stability issues. The transient response is fast because the circuit has no bandwidth-limiting feedback components.

Depending on the inductor value and loading, the operating frequency remains constant for most of the output-power range. A desirable reduction in operating frequency occurs at low loading levels. □

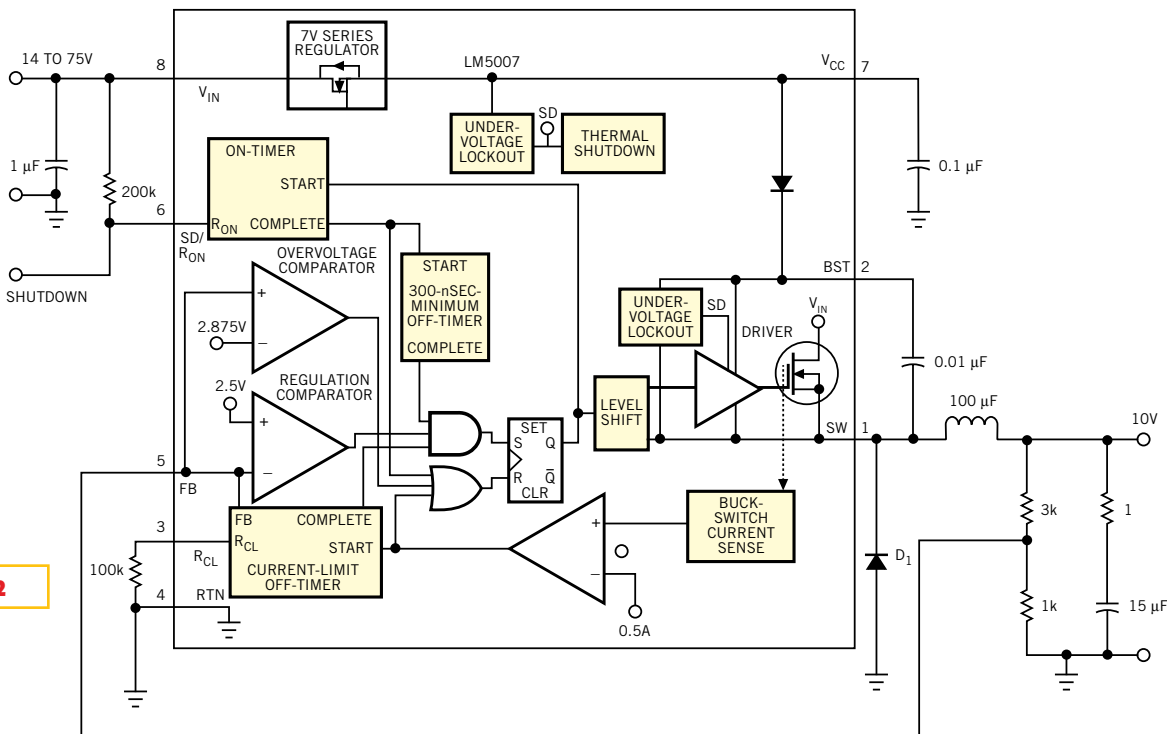


Figure 2

In this buck regulator, the switching frequency remains constant over a wide range of input voltages.

LED driver combines high speed, precision

Richard Cappels, Mesa, AZ

APPPLICATIONS such as turbidity measurement and densitometry require cleanly pulsed light sources with stable amplitudes. The LED driver in **Figure 1** illuminates retinal neurons in a biological experiment that has similar requirements. For a given LED at a given current, the intensity is stable, so switching a stable current is a simple and effective way to obtain the needed function. The circuit provides current pulses to the LED with rise and fall times lower than 500 nsec and overshoot lower than 7%. You can make the current computer-programmable by replacing the potentiometer with a DAC. The circuit comprises an adjustable, regulated current source (IC₁ and Q₂), an overdriven differential amplifier (Q₃ and Q₄) acting as a switch, and a level shifter (Q₁) to shift the TTL input signal to levels needed to drive the differential pair.

Voltage at the wiper of R₆ results in an equal voltage across R₉ because of feedback to the op amp. Because transistor Q₂ has a high alpha, most of the emitter current that produces the voltage across R₉ comes from the collector of Q₂. Because alpha varies little with temperature, this current remains stable. Transistors Q₃ and Q₄ constitute a differential pair. Depending on which transistor is conducting, the emitter of one or the other sources current to the collector of Q₁. When the base of Q₃ becomes several hundred millivolts more positive than the base of Q₄, current from Q₂ shunts to the 5V power supply. No current flows through Q₄, so the LED is off. When the base of Q₃ is less positive than that of Q₄, current from Q₂ passes through the LED. The all-or-nothing switching action re-

sults from the large differential voltage across the bases. Similar to the case of Q₂, the collector current in Q₄, when it is conducting, is a high and stable percentage of the emitter current. The constant load the emitters of Q₃ and Q₄ present to the current source enables the current source to operate continuously, allowing the use of a low-bandwidth op amp.

Q₁ is a common-base amplifier connected in a manner essentially the same as a TTL-input stage with the exception of C₃, the 1000-pF capacitor across base resistor R₄. When the input signal is greater than 2V, the base of Q₁ remains at 2.5V, and the collector of Q₁ rises enough to ensure that Q₄ and the LED conduct no current. When the input is below 0.4V, Q₁'s emitter voltage is low enough and the base current through R₄ is high enough to saturate Q₁. This action

holds the base of Q₃ low enough to ensure that all the collector current from Q₂ passes through Q₄ and the LED. When the input signal swings positive again, the energy stored in C₃ develops a reverse bias across Q₁'s emitter-base junction to quickly deplete the stored charge, resulting in a rapid turn-off. Make sure that you don't exceed the power rating of Q₄. Take the current and collector-to-emitter voltage into account. Using transistors in TO-92 packages and an LED that drops 2V at 50 mA, the circuit in **Figure 1** operates at temperatures greater than 55°C with a jumper in place of R₁₂. If you need higher currents or use smaller transistor packages, you may find it necessary to use a finite resistor for R₁₂ to lower the dissipation in Q₄ to a safe level. □

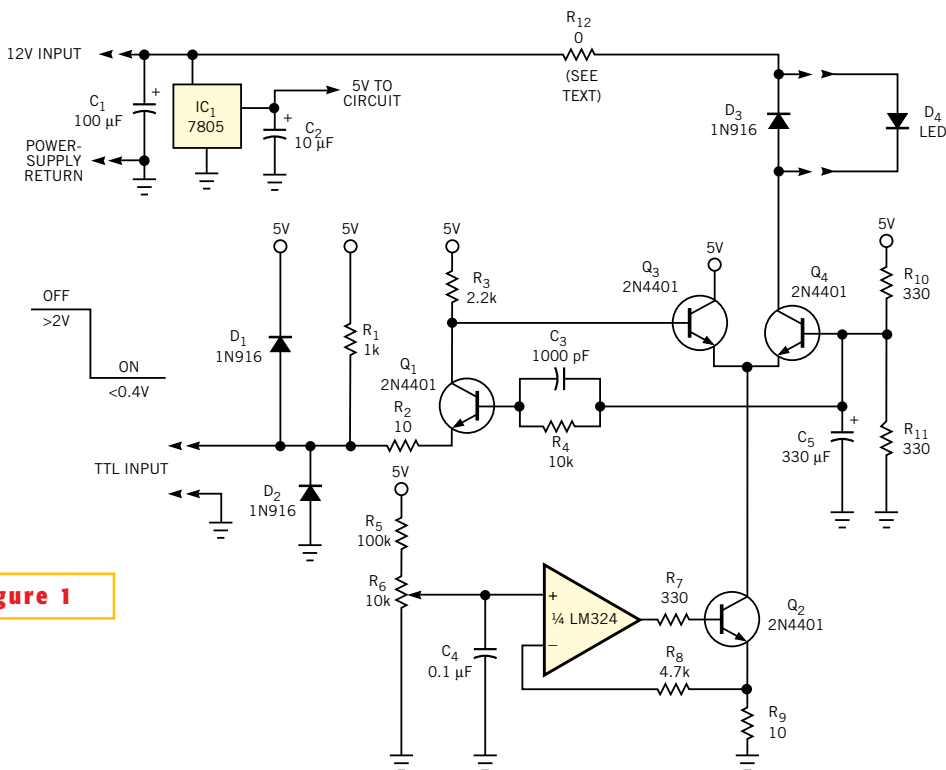


Figure 1

This circuit delivers a stable, precision dose of current to an LED.

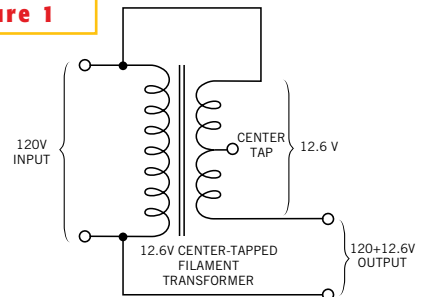
Filament transformer adjusts line voltage

EE Barnes, Cochranville, PA

THE PROBLEM: I couldn't use my Heathkit oscilloscope in a house I lived in during the 1960s because my lab was too far from the power-line input to the house, and the line drop through

the house was substantial. Depending on the time of day, the screen would shrink to perhaps half the normal display size. I checked the line voltage, and it was down to just approximately

Figure 1



You can use a small, inexpensive filament transformer to buck or boost the ac-line voltage.

100V. I lacked the funds to buy a high-wattage Variac to deal with the problem.

The solution: I had a couple of 12.6V filament transformers, rated at 3 or 4A. I simply connected one of these in my lab, with the primary winding across the ac line (**Figure 1**). Then, I connected the secondary winding such that one side connected to the ac line, and the other side provided the new, boosted ac line. Because the transformer had a center tap, I could adjust the line voltage in 6.3V steps. The beauty of this approach is that the transformer handles only the incremental power from the slight boost in voltage. And the technique uses less space and is less expensive than using a Variac.

Note that, by changing the polarity of the filament transformer's output, you can decrease rather than increase the ac output. This fact could come in handy in situations in which the line voltage is too high, causing incandescent-lamp burn-out. Reduction in lamp life is a function of approximately the 13th power of the overvoltage (**Reference 1**). For the long, skinny, and expensive European incandescent lamps that some bathrooms use as a vertical light source, the lamp-life reduction can be significant. You can buck, or subtract, the line voltage to increase the lamps' life. Even at nominal line voltage, you can use the method to drop the voltage to an expensive or particularly inaccessible incandescent lamp. □

REFERENCE

1. Fink, Donald and Christiansen, Donald, *Electronic Engineers' Handbook*, 1975, McGraw-Hill, pg 11-6.