Edited by Bill Travis

Combine two 8-bit outputs to make one 16-bit DAC

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Inexpensive, 16-bit, monolithic DACs can serve almost all applications. However, some applications require unconventional approaches. This Design Idea design concerns circuitry I recently designed for a tunable-diode laser spectrometer for a Mars-exploration application. The control circuitry included two 16-bit DACs that interface to the radiation-hardened, 8051-variant 69RH051A microcontroller. Because of the intended space-flight-qualified specification, everything in the design had to consist solely of components from the NPSL (NASA parts-selection list). This restriction posed a challenge, because, at design finalization, the NPSL included no appropriate, flight-qualified, 16-bit DACs, and the budget included no funds for certification of new devices. I escaped from this impasse by exploiting two fortuitous facts: The update rate of the two DACs was only tens of hertz, and the 69RH-051A had a number of uncommitted, 8-bit, 14.5-kHz PWM outputs. These outputs made one 16-bit DAC; a second pair of PWM bits and an identical circuit made the other (**Figure 1**).

Hex inverter IC_1's V_{CC} rail connects to a precision 5V reference. The inverter's outputs are accurate analog square waves. The low-order PWM-signal output, PWM0, of the 8051 controls the V_3 square wave, and the high-order PWM output, PWM1, controls the V_1 square wave. R_2 and R_6 passively sum the two square waves in the ratio $R_2/R_6=3290/1$ million$=1/255$ to produce V_4, duplicating the 2^8 ratio of the 16-bit sum. This action makes the dc component of V_4 equal to $5V(REF)(PWM0+255PWM1)/256$. Thus, if you write the 0 to 255, high-order byte of a 0 to 65,535, 16-bit DAC setting to the CEX1 register of the 8051 and write the 0 to 255, low-order byte to CEX0, a corresponding 16-bit analog representation appears in the dc component of V_4. The accuracy of the R_2-to-R_6 ratio is the only limit on the monotonicity and accuracy of this circuit. For example, one part in 25,500=14.5 bits for 1%-tolerance R_2 and R_6 and a full 16 bits for 0.3% tolerance or better. But the story doesn't end there. Two problems remain.

The first problem is the extraction of V_4's desired dc component from all—or

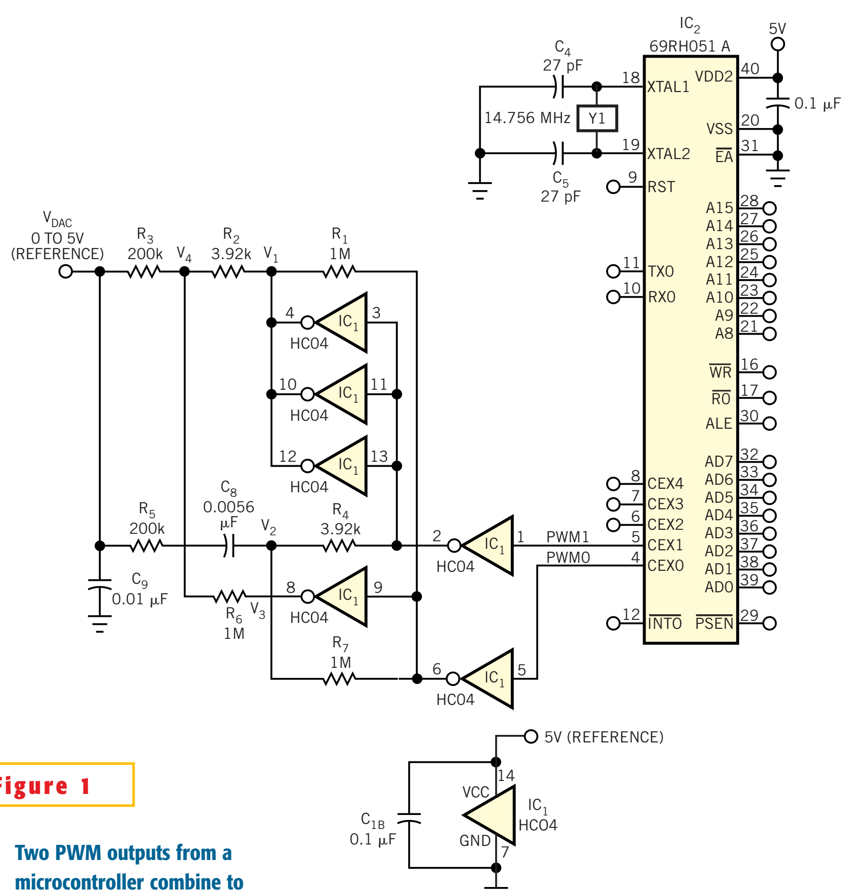

Figure 1

Two PWM outputs from a microcontroller combine to form a monotonic 16-bit DAC.

at least 15 or 16 bits=99.995%—of the undesired square-wave ac ripple. The R_3-C_9 lowpass filter does some of this work. If you make C_9 large enough, in principle, the filter could do the whole job. The reason this simple approach wouldn't work is that, to get such a large ripple attenuation of approximately 90 dB with a single-stage RC filter would require an approximately 300-msec time constant and a resultant 3-sec, 16-bit settling time. This glacial response time would be too slow even for this undemanding application. To speed things, the R_4, R_5, R_7, C_8 network synthesizes and then sums V_2: an inverse-polarity duplicate of V_4's 14.5-kHz ac component. This summation actively nulls out approximately 99% of the ripple. This nullifying action leaves such a small residue that an approximately 2-msec and, therefore, approximately 25-msec-settling-time R_3C_9 product easily erases it.

The other problem is compensation for the low, but still nonzero, on-resistance of the HC14 internal CMOS switches, so that the resistance doesn't perturb the critical R_2-to-R_6 ratio. This issue is of no particular concern for R_6, because the R_6-to-on-resistance ratio is greater than 10,000-to-1, making any associated error negligible. This situation is not the case for R_2, however, in which, despite the triple-parallel gates, the R_2-to-on-resistance ratio is approximately 300-to-1, which is small enough to merit attention. Load-cancellation resistor R_1 provides such attention. R_1 sums a current into the R_2 driving node that, because it is equal in magnitude but opposite in phase to the current through R_6, effectively cancels the load on the R_2 drivers. This process makes the combined on-resistance approximately 100 times less important than it otherwise would be. The result is a simple, highly linear and accurate voltage-output DAC with a respectable, if not blazingly fast, settling time of approximately 25 msec. And the most important result, in this case, was a parts list with an impeccable NPSL-compliant pedigree.□

LED driver provides software-controlled intensity

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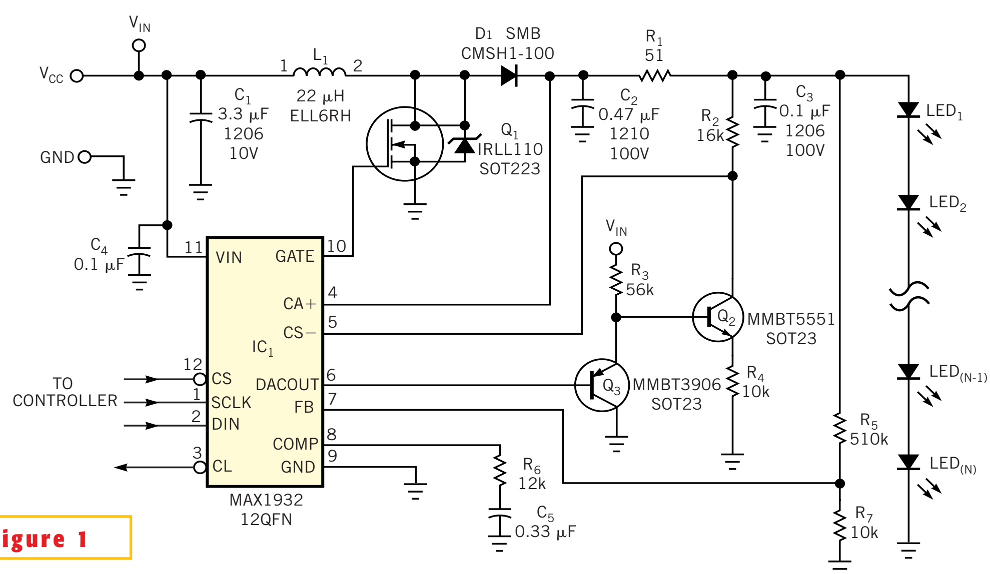
Recent advances in operating efficiency have expanded the use of LEDs from one of mere indicators to becoming driving forces in electronic lighting. Increased reliability and ruggedness (versus other lighting technologies) gives the LED a bright future indeed. Vendors in recent years have introduced many ICs for driving LEDs, but the problem of driving serial chains of LEDs has received less attention. One approach to that problem adapts a bias-supply IC for APDs (avalanche photodiodes) to provide adjustable-current, software shutdown, and logic indication of open-circuit faults (**Figure 1**). This design reconfigures the APD-bias IC, IC_1, to allow its low-voltage DAC output to modulate the high-voltage, current-sense feedback via a high-voltage-output transconductance stage comprising Q_2 and Q_3. These two complementary transistors provide first-order temperature-compensation sufficient for the application.

Figure 1



The APD driver, IC_1, provides high-voltage LED modules with software-adjustable intensity control.

Equations from the MAX1932 data sheet help you select components for the step-up dc/dc converter. The current-adjustment transfer function is:

$$I_{OUT}=\frac{V_{CL}-\frac{CODE\times 1.25V}{256}\times\frac{R_2}{R_4}}{R_1},$$

where V_{CL} is the current-limit threshold (2V), CODE is the digital code to the DAC in decimal format, and I_{OUT} is the desired output current. For this circuit, these conditions correspond to a full-scale output of 39 mA and a resolution of 150 µA. The three-wire serial interface that controls IC_1 allows you to shut down IC_1 by writing code 00hex to the DAC. The circuit also provides an output-voltage limit. If an LED fails open, the R_5-R_7 divider limits the output voltage, in this case, to 50V. Simultaneously, the CL pin goes high to indicate the open-fault condition.□

Improve roll-off of Sallen-Key filter

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The well-documented Sallen-Key active filter is a staple of analog design. This Design Idea shows a way to obtain better roll-off by adding just a few common passive components. **Figure 1a** shows a typical implementation of a three-pole, lowpass version. In operation, you adjust the ratio of capacitors C_1 and C_2 to give a peaked response for the two poles within the feedback loop. The peaked response compensates for the initial roll-off in the third pole formed by the R_3-C_3 section at the input. In **Figure 1b**, a twin-tee notch filter replaces the R_3-C_3 section at the input. The notch frequency, $F=1/(2\pi R_4C_4)$, is equal to approximately twice the desired cutoff frequency.

Select a value for R_4 that's approximately one-third to one-fourth the value of R_1, and then adjust R_4 as needed to allow use of standard capacitor values. The graph in **Figure 2** shows the improvement in the cutoff rate of the filter; the result is a quasi-elliptic response. A breadboard of the circuit in **Figure 1b** uses 5% parts. The measured results show good agreement with the Spice simulation. To take advantage of the faster roll-off, just scale the frequency and impedance to your application. The highpass dual of this circuit works as well as the lowpass version.□

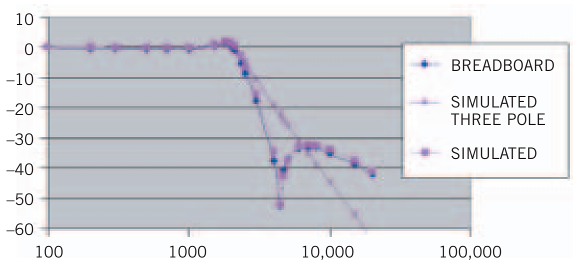

Figure 2 The improved cutoff rate of the filter results in a quasi-elliptical response.

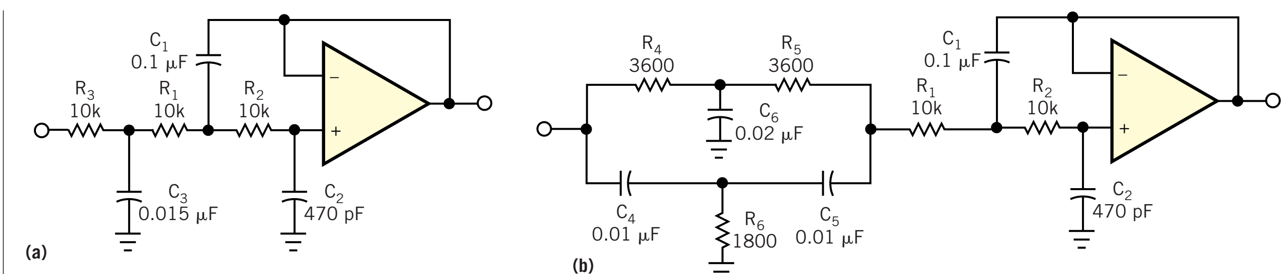

Figure 1 The addition of a twin-tee network (b) considerably improves the roll-off rate of the circuit (a).

AC-coupling instrumentation amplifier improves rejection range of differential dc input voltage

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The need for conditioning low-level ac signals in the presence of both common-mode noise and differential dc voltage prevails in many applications. In such situations, ac-coupling to instrumentation and difference amplifiers is mandatory to extract the ac signal and reject common-mode noise and differential dc voltage. This situation typically occurs in bioelectric-signal acquisition, in which metallic-electrode polarization produces a large random differential dc voltage, ranging from ±0.15V, which adds to low-level biological signals. Input ac-coupling is one approach to removing the differential dc content. But this technique requires adding a pair of capacitors and resistors to ac-couple the inputs of the difference amplifier. The manufacturing tolerances of these components severely degrade the CMRR (common-mode-rejection ratio) of the amplifier. If cost is not an issue, you could perform an initial trim, but this operation is useless for biological applications plagued by wide variations in electrodes and tissue impedances. The differential topology in **Figure 1** addresses these problems (**Reference 1**).

The principle of this ac-coupled instrumentation amplifier is to maintain the mean output voltage at 0V. To do so, you insert an autozero feedback loop, comprising IC_4, R_{FB}, and C_{FB}, in a classic three-op-amp instrumentation amplifier. This feedback loop produces a frequency-dependent transfer function:

$$\frac{V_{OUT}}{V_{IN1}-V_{IN2}}=\left(1+\frac{2R_2}{R_1}\right)\frac{jR_{FB}C_{FB}\omega}{1+jR_{FB}C_{FB}\omega}.$$

Consequently, the ac-coupled instrumentation amplifier behaves as a highpass filter with a –3-dB cutoff frequen-

(continued on pg 92)

cy from the equation $f=1/2\pi R_{FB}C_{FB}$. At first glance, you might think that the output-autozeroing behavior of the ac-coupled instrumentation amplifier is perfect. Unfortunately, the output autozeroing capability of this circuit is strongly limited. You can determine this limitation by expressing the output voltage as a function of the input signals and the integrator's output voltage, V_Z: $V_{OUT}=(1+2R_2/R_1)(V_{IN1}-V_{IN2})+V_Z=A_D(V_{IN1}-V_{IN2})+V_Z$, where V_{OUT} is the output voltage. In this expression, $A_D=1+2R_2/R_1$ is the differential gain in the passband. At dc, the output voltage is 0V as long as the integrator's output does not reach its saturation voltage, $V_{Z(MAX)}$. Therefore, setting the output voltage at 0V in the above expression yields the maximum differential-input dc voltage that this circuit can handle:

$$\Delta V_{IN(MAX)}=(V_{IN1}-V_{IN2})(MAX)=\pm\frac{V_{Z(MAX)}}{\left(1+\frac{2R_2}{R_1}\right)}=\pm\frac{V_{Z(MAX)}}{A_D}.$$

Consider, for instance, the typical performance and constraints of a portable biotelemetry system: differential gain of 1000, ±5V split power supplies, and op amps with rail-to-rail output-voltage swing. In this system, the application of the formula for ΔV_{IN} yields a maximum differential-input dc voltage of only ±5 mV. This limited performance is unacceptable for biological applications, in which you encounter differential-input dc voltages of ±0.15V. The ac-coupled instrumentation amplifier in **Figure 2** overcomes this limitation, thanks to the addition of "active feedback," which includes voltage divider R_3-R_4 and the associated buffer amplifier, IC_5. With this arrangement, the following equations give the new transfer function and high-pass cutoff frequency, respectively.

$$A_D=\frac{V_{OUT}}{V_{IN1}-V_{IN2}}=\left(1+\frac{2R_2}{R_1}\right)\left(1+\frac{R_4}{R_3}\right)\frac{jR_{FB}C_{FB}\omega}{1+jR_{FB}C_{FB}\omega}.$$

$$f_C=\frac{\left(1+\frac{R_4}{R_3}\right)}{2\pi R_{FB}C_{FB}}.$$

The expression for the output voltage as a function of the input signal and the integrator's output voltage, V_Z, becomes:

$$V_{OUT}=\left(1+\frac{2R_2}{R_1}\right)\left(1+\frac{R_4}{R_3}\right)(V_{IN1}-V_{IN2})+\left(1+\frac{R_4}{R_3}\right)V_Z=A_D(V_{IN1}-V_{IN2})+\left(1+\frac{R_4}{R_3}\right)V_Z.$$

In this expression, $A_D=(1+2R_2/R_1)(1+R_4/R_3)$ is the new differential gain in the passband.

At dc, the output voltage remains 0V as long as the integrator's output does not reach its saturation voltage, $V_{Z(MAX)}$. Therefore, setting the output voltage at 0V in the new expression for output voltage yields the new maximum differential-input dc voltage and differential gain. They are, respectively:

$$\Delta V_{IN(MAX)}=(V_{IN1}-V_{IN2})(MAX)=\pm\frac{V_{Z(MAX)}}{A_D}\left(1+\frac{R_4}{R_3}\right).$$

$$A_{D(MAX)}=\frac{V_{Z(MAX)}}{\Delta V_{IN(MAX)}}\left(1+\frac{R_4}{R_3}\right).$$

In the above equations, the additional term, $1+R_4/R_3$, is the gain of the active-feedback stage.

The new expressions for $\Delta V_{IN(MAX)}$ and $A_{D(MAX)}$ clearly show the advantages of **Figure 2**'s ac-coupled instrumentation amplifier with active feedback: For an identical differential gain, you can extend the polarization-voltage range, $\Delta V_{IN(MAX)}$, by a factor equal to the gain of the active-feedback stage. Conversely, for a given polarization-voltage range, $\Delta V_{IN\ (MAX)}$, you can increase the differential gain by the gain of the active-feedback stage.

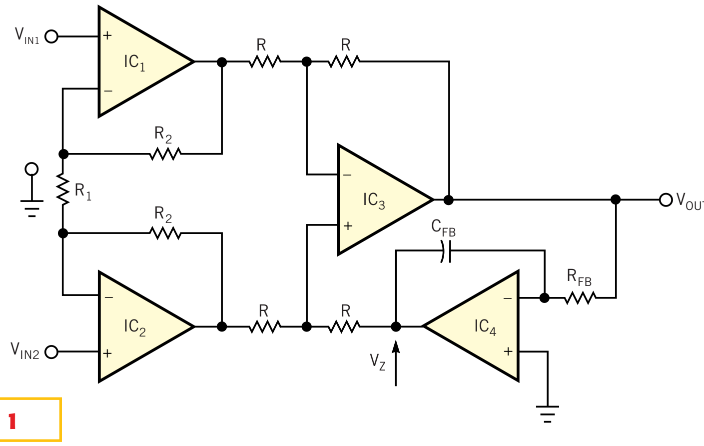
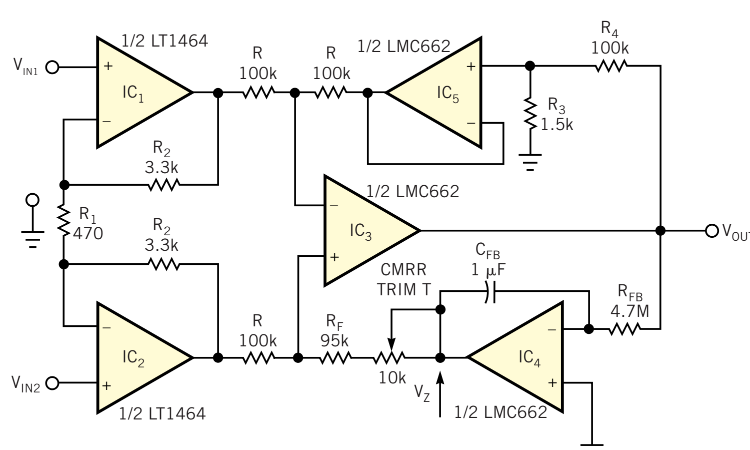


Figure 1

This ac-coupled instrumentation amplifier accommodates only ±5-mV maximum input.



NOTE: R AND R_F ARE ±1%; OTHERS ARE ±5%

Figure 2

This instrumentation amplifier can accommodate a differential-input range of ±0.34V.

The only drawback of this topology is apparent in the expression for f_C, the highpass cutoff frequency. You multiply this frequency by the gain of the active-feedback stage. Therefore, to maintain a given cutoff frequency, you must multiply the time constant by a factor equal to the active-feedback stage gain. This factor can be an issue in processing signals whose spectrum includes low-frequency components. In such applications, R_{FB} and C_{FB} can reach prohibitive values. Consequently, you must make a trade-off between the time constant and the active-feedback stage gain. The component values in **Figure 2** are a typical example of such a trade-off: The values are for an EEG (electroencephalogram) amplifier with ±5V split power supplies. The amplifier has a differential gain of 1000 and a highpass cutoff frequency of 2.3 Hz, and it can handle a differential-input dc-voltage range of ±0.34V.

To obtain this performance, you set the active-feedback stage gain and the differential-amplifier gain, respectively, to 67.6 and 15. With these gain values, the noise performance of the ac-coupled instrumentation amplifier of **Figure 2** is similar to that of a classic instrumentation amplifier. This situation occurs because the autozeroing and active-feedback stages, IC_4 and IC_5, are after the input differential stage, IC_1 and IC_2. Consequently, the gain of the differential stage roughly divides their respective noise contributions, which are therefore negligible. You can use several low-noise op-amps for IC_1 and IC_2. For portable biotelemetry applications, the LT1464 is a good compromise for input-noise density, noise-corner frequency, input-bias current and current drain. (Respectively: $V_{NOISE}=26\ nV/\sqrt{Hz}$, $f_C=9$ Hz, $I_{BIAS}=0.4$ pA, and $I_{CC}=230\ \mu A$.)

A theoretical analysis using the LT 1464's noise parameters shows that under worst-case conditions, the input-noise voltage should not exceed 11 μV rms. Tests on prototypes confirm this prediction; the tests effectively measure input-noise voltages of 3 to 6 μV rms. To sum up, an ac-coupled instrumentation amplifier with active feedback is well-suited for applications requiring high differential gain, a capability for handling large differential-input dc voltages, and low-noise performance.□

Reference

1. Stitt, Mark, "AC-Coupled Instrumentation and Difference Amplifier," Burr-Brown, AB-008, May 1990.

Simplify computer-aided engineering with scientific-to-engineering conversion

Alexander Bell, Infosoft International, Rego Park, NY

The simple yet useful formula in this Design Idea enables conversion from scientific format (for example, 2.2^{-9}), which is typical for CAE (computer-aided-engineering), double-precision output values, into "human-friendly" engineering format (for example, 2.2 nF). The engineering format is more suitable for bills of material and other electrical and electronic-engineering documents and specifications.

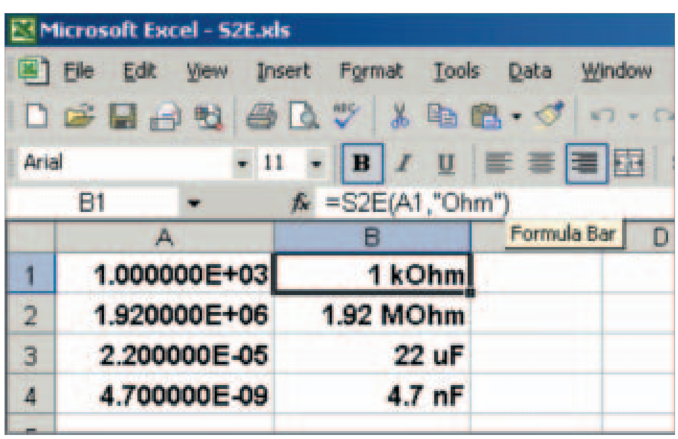


Figure 1 The formula appears in the formula bar, taking the first numeric parameter from the column to the left. The unit of electrical resistance, "ohm," is a second parameter.

The formula is rather straightforward. It takes two parameters. The first is the numerical value, and the second one specifies the unit of measurement—ohms, farads, or henries, for example. Alternatively, it could be of any random text, including an empty string, "." The formula calculates the mantissa/order of magnitude and returns the text string, formatted in compliance with commonly accepted electrical-engineering practice. **Listing 1**, available at www.edn.com, shows details of the formula. The tricky part of the formula is the conversion to a decimal type after the formula calculates the ratio of two log values (**Reference 1**). This step ensures the correct order-of-magnitude calculations in cases in which the mantissa of the input value is close or equal to one.

The formula is in VB; you could use it in any VB/VBA-backed software applications. In this example, the function is in a code module of an MS Excel file (**Figure 1**). You could also use it as an Excel Add-In (.xla) or a "pure-VB," compiled-DLL component. You can download **Listing 1** and the Excel file from the Web version of this Design Idea at www.edn.com. The input numerical value in this formula has double-precision accuracy, and its range spans from approximately -1.79^{308} to $+1.79^{308}$, which is sufficient for any practical engineering calculations. Note that the maximum value is even bigger than the famous "googol," which is represented by 100 digits.□

Reference

1. Bell, Alexander, "What's wrong with INT(LOG) in VBA?" *Access-VB-SQL Advisor*, October 2002, pg 65.

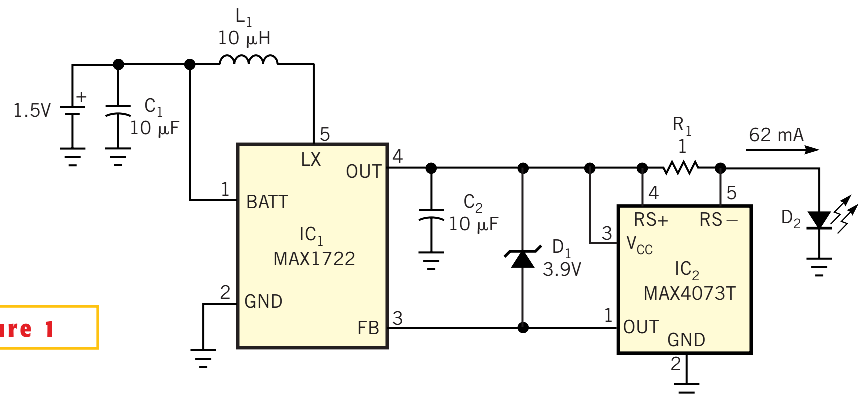
1.5V battery powers white-LED driver

Steve Caldwell, Maxim Integrated Products, Chandler, AZ

Although white LEDs are common in a variety of lighting applications, their 3 to 4V forward-voltage drop makes low-voltage applications challenging. Charge pumps and other ICs are available for driving white LEDs, but they generally don't work with the low supply voltage of 1.5V in single-cell-battery applications. The low-voltage circuit of **Figure 1** provides a current-regulated output suitable for driving white LEDs. The boost converter, IC_1, can supply load currents to 62 mA with input voltages as low as 1.2V, making it suitable for use with a 1.5V, single-cell battery. Because a white LED draws negligible load current until the output voltage rises above 3V, the boost converter can start with input voltages as low as 0.8V.

By deriving feedback from a high-side current-sense amplifier, IC_2, the circuit allows current regulation without sacrificing efficiency. IC_2's 1.8-MHz bandwidth also eliminates instability in the feedback loop. IC_2 amplifies the voltage across R_1 with a gain of 20. This high gain boosts efficiency by enabling use of a small-valued current-sensing resistor. You can calculate the value of R_1 from the desired output current: $R_1=1.235V/(20\times I_{OUT})$. For 1.5V input and 62-mA output, the circuit efficiency of **Figure 1** is approximately 80%. Zener diode D_1 provides overvoltage protection at the output. When the output voltage rises above the sum of the zener voltage (V_Z) and IC_1's 1.235V feedback voltage (V_{FB}), the feedback voltage (Pin 3) rises and causes IC_1 to stop switching. Thus, for an open-circuit output, the output voltage is regulated at V_Z+V_{FB}.□

Figure 1



Powered from a single-cell battery, this circuit provides a regulated output current suitable for driving a white LED.

Simple V_{COM} adjustment uses any logic-supply voltage

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All TFT (thin-film-transistor) LCD panels require at least one appropriately tuned V_{COM} signal to provide a reference point for the panel's backplane. The exact value of V_{COM} varies from panel to panel, so the manufacturer must program the voltage at the factory to match the characteristics of each screen. An appropriately tuned V_{COM} reduces flicker and other undesirable effects. Traditionally, the V_{COM} adjustment used mechanical potentiometers or trimmers in the voltage-divider mode. In recent years, however, panel makers have begun looking at alternative approaches because mechanical trimmers can't provide the necessary resolution for optimal image fidelity on large panels. They also require a physical adjustment that technicians on the assembly line usually perform. This adjustment is not only time-consuming, but also prone to field failures arising from human error or mechanical vibration.

A simple alternative to achieving the increasing adjustment resolution for optimal panel-image fidelity is to replace the mechanical potentiometer with a digital potentiometer. Using digital potentiometers, panel makers can automate the V_{COM}-adjustment process, resulting in lower manufacturing cost and higher product quality. Unfortunately, many panels operate at higher voltages, and the choice of available supply voltages is limited. The system implementation for a 5V supply is straightforward (**Figure 1**). Without a 5V supply, the circuit can become more complex.

This Design Idea shows a simple way

TABLE 1—OUTPUT-VOLTAGE RANGE

R_2 tolerance (%), scale	R_2 (kΩ)	V_{COM} (V)	Step size (mV)
–30, zero	0	3.5	3.9
–30 mid	3.5	4.0	
–30, full	7	4.5	
30, zero	0	3.3	6.8
30, mid	6.5	4.2	
30, full	13	5.1	

that you can use any available logic supply to power the potentiometer providing the V_{COM} adjustment. The 6- or 8-bit AD5258/59 nonvolatile digital potentiometer demonstrates this approach. An I²C serial interface provides control and stores the desired potentiometer setting into the EEPROM. The AD5259 uses a 5V, submicron CMOS process for low power dissipation. It comes in a space-saving 10-pin MSOP, an important feature in low-cost, space-constrained applications. For systems that have no 5V supply, many designers would be tempted to simply tap off the potentiometer's series-resistor string at the 5V location. This approach is not viable, because, during programming (writing to the

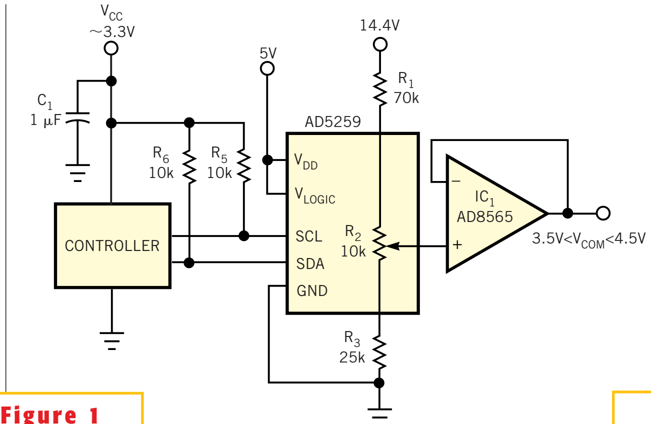

Figure 1

A digital potentiometer makes it easy to adjust V_{COM} to the desired value.

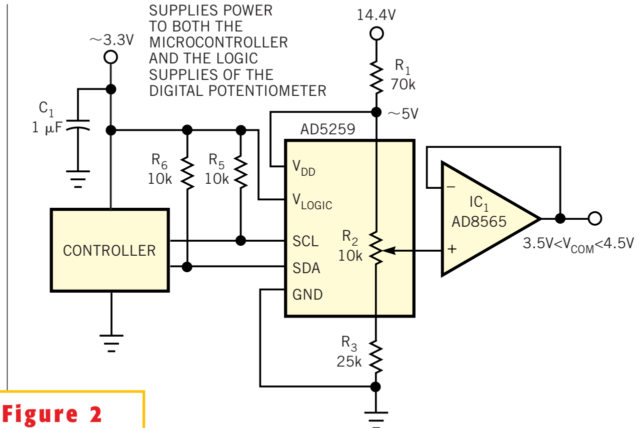

Figure 2

A separate V_{LOGIC} pin makes it possible to derive the V_{DD} supply from the potentiometer's resistor string.

EEPROM), the AD5259's V_{LOGIC} pin typically draws 35 mA. It cannot draw this current level through R_1 because the voltage drop would be too large. For this reason, the AD5259 has a separate V_{LOGIC} pin that can connect to any available logic supply. In **Figure 2**, V_{LOGIC} uses the supply voltage from the microcontroller that is controlling the digital potentiometer. Now, V_{LOGIC} draws the 35-mA programming current, and V_{DD} draws only microamps of supply current to bias the internal switches in the digital potentiometer's internal resistor string. If the panel requires a higher V_{COM} voltage, you can add two resistors to place the op amp in a noninverting gain configuration.

The digital potentiometer has ±30% end-to-end resistance tolerance. Assuming that the tolerances of R_1, R_3, and V_{DD} are negligible compared with those of the potentiometer, you can achieve the range of output values that **Table 1** shows. Assume that the desired value of V_{COM} is 4V±0.5V, with a maximum step size of 10 mV. As **Table 1** shows, the circuit in **Figure 2** guarantees an output range of 3.5 to 5.4V with a step size within ±10 mV. And, despite the ±30% tolerance of R_2, the midscale V_{COM} output meets the target specification. Also, because the digital potentiometer's logic supply matches the microcontroller's logic levels, the microcontroller can read data back if desired. **Figure 3** shows a block diagram of the digital potentiometer.□

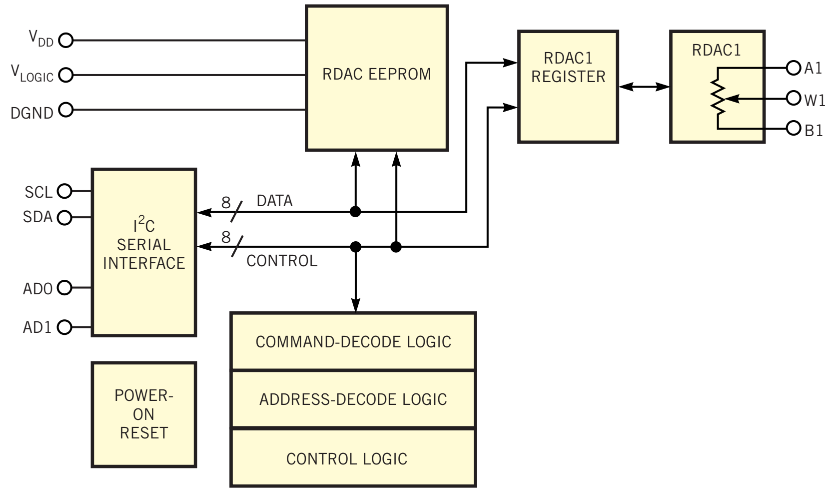

Figure 3 This block diagram shows the digital potentiometer's inner workings.