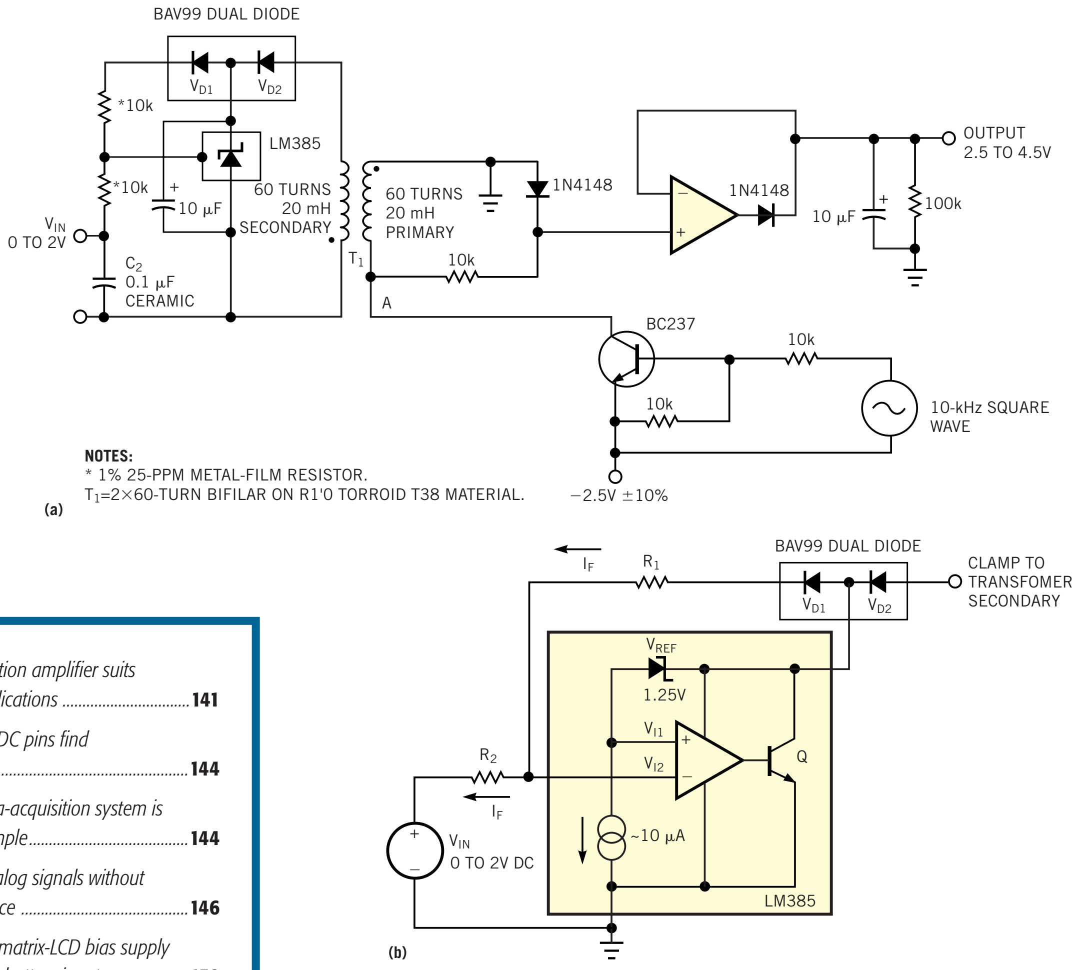
Edited by Bill Travis and Anne Watson Swager

Lost-cost isolation amplifier suits industrial applications

Andrew Russell, Philips Semiconductors, Hazel Grove, England

The circuit in **Figure 1a** is a low-cost isolation amplifier for instrumentation applications that provides as much as 500V of galvanic isolation between input and output. The amplifier uses only one small, low-cost transformer and with little modification lends itself to cost-effective multichannel applications. Input-to-output linearity is around 0.05% for a 2V input signal. The LM385 (National Semiconductor) low-power programmable reference diode, which operates in the shunt mode, and the dual planar BAV99 (Philips Semiconductors) diode are the major circuit components.

Figure 1



An isolation amplifier for instrumentation applications provides as much as 500V of galvanic isolation and uses only one low-cost transformer (a). The clamp circuit includes dc blocking, which V_{D2} provides (b).

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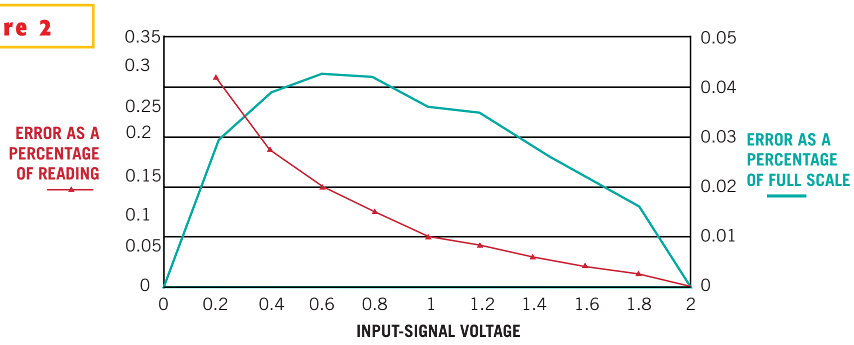
Figure 2



Linearity performance of the circuit is measurable in terms of percentage of reading and as a percentage of full scale.

To understand the circuit's operation, you have to first look at the clamp circuit (**Figure 1b**). The LM385 is a shunt regulator that consists of a control amplifier, a current-shunt transistor, and an internal precision 1.25V reference. Two external feedback resistors, R_1 and R_2, set the output voltage. In conventional LM385 applications, the collector of Q is the clamp, or current shunt point. However, in this design, the clamp has to include dc blocking, which V_{D2} provides because the circuit feeds V_{D2} with an ac signal from the transformer secondary. V_{D1}, which is inside the LM385 feedback loop, compensates for V_{D2}. Because V_{D1} and V_{D2} are thermally coupled inside a single package, changes in the forward voltage drop across V_{D2} due to temperature are mirrored across V_{D1}. The transfer function of the clamp circuit is $V_{CL}=2V_{REF}+V_{IN}$.

In the overall circuit, a symmetrical, 10-kHz square wave drives a low-cost BC237 npn transistor, which in turn drives the primary of T_1. In the forward mode, no secondary current flows due to the dc blocking action of V_{D2}. During this phase, primary magnetizing current, which the circuit converts to magnetic field energy, ramps from 0 to −650 μA. When the BC237 turns off due to the drive voltage on its base switching low, the inductive energy in the core dumps into the secondary, causing V_{D2} to be forward-biased and current to flow through the LM385 and back to the other side of the secondary winding. The anode voltage of the LM385 clamps at precisely $V_{IN}+2V_{REF}-V_{D1}$. However, the addition of V_{D2} into the clamp voltage at the anode of V_{D2} compensates for V_{D1}, resulting in a clamp voltage of $2V_{REF}+V_{IN}$. Note that the voltage drop across the diodes during clamping is different due to the large discrepancy in current between the two diodes; V_{D2} carries the peak clamp current, and V_{D1} conducts only the feedback current. However, the circuit largely compensates for the temperature-induced changes in forward voltage drop, which can be a major source of error. Although some mismatch in the thermal tracking of the diodes does occur due to the different forward currents in the diodes, this mismatch is small enough given the accuracies of the circuit that you can consider it a second-order effect.

The response time of the circuit at Point A is less than 3 msec for 10 to 90% and 90 to 10% input-signal steps. Note that the input signal must be capable of sinking the feedback current, I_F, which for the values in the circuit is approximately 65 μA. With 100-kΩ feedback resistors, the feedback current drops to approximately 8 μA. Drift is largely a function of feedback-resistor stability, LM385 temperature stability, and the thermal tracking of the diodes within the dual-diode package. Average current consumption of the circuit, excluding the peak-detector op amp, is approximately 150 μA. The noise and stability of the -2.5V supply that drives the transformer are not critical, and a simple zener regulator suffices. **Figure 2** depicts the linearity performance as both a percentage of reading and as a percentage of full scale.

In multichannel-isolation applications, you can delete the peak detector stage and feed Point A, or the primary winding, directly into a high-speed, multichannel ADC. For a 10-kHz drive frequency, the clamped waveform tops are typically approximately 20 to 40 μsec long. Sampling should take place at some fixed time after the rising edge, such as 25 to 30 μsec, because the amount of current that shunts through the LM385 decreases during the clamping period as the magnetic energy in the core decays. The LM385 and the associated rectifier diode, V_{D2}, have a dynamic resistance that depends on the current that each device is conducting. The secondary winding resistance times the clamp current also gives rise to a further error term. These two errors combine and are reflected as a slope on the clamped waveform tops appearing on the primary winding when you view the signal at Point A with a scope. For this reason, you need to sample this waveform at some fixed point after the rising edge of the clamped portion of the waveform. Otherwise, errors can result in the readings taken from one sampling event to the next.

In this design, slight overshoot on the rising edge is too fast to cause any problems on the peak detector, but this overshoot would cause a problem at the input to a high-speed ADC. Again, sampling at some fixed period after the rising edge will obviate any problems.

Note that the output clamp voltage measured at Point A varies from 2.5 to 4.5V because of the initial 2.5V offset that stems from the $2V_{REF}$ term in the transfer equation. You have to remove this offset through a subsequent offset removal circuit or by simply subtracting the offset value from the reading in software when using a high-speed ADC. The use of software calibration techniques makes this a viable option in a production environment. (DI #2474)

To Vote For This Design,
Circle No. 301

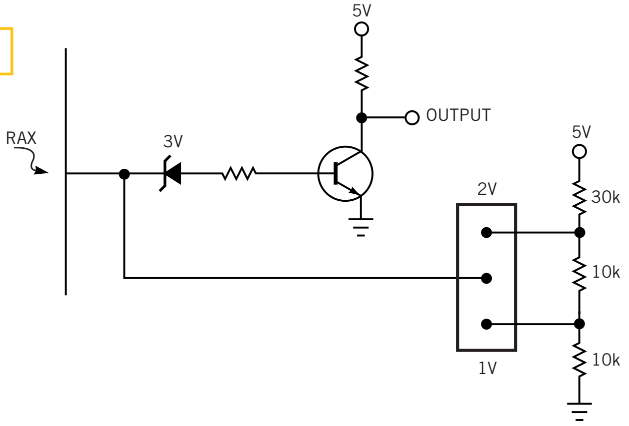
Unused µC ADC pins find second life

Kannan Natarajan, Mediatronix Private Limited, Kerala, India

Some PIC µCs, such as the 12C67X and 16C7X, have more than one analog input channel. If you don't need all the available channels, you can use the unused channels as general-purpose I/O. For example, you can use unused ADC pins for power-on status reading and as an output in normal operation.

In **Figure 1**, the jumper selects a mode—the battery charge voltage, for example—by PIC software at power-on. At power-on, software configures the RAX line as an analog input and reads the voltage—for example, 1V at Digit 0 and 2V at Digit 1. The same RAX line also drives the base of a transistor through a 3V zener diode. At power-on, the voltage at the RAX input is insufficient to turn on the transistor. However, after using power-on-initialization software, you configure RAX as a digital output. Now, the low-impedance high/low voltage at RAX can override the bias voltage at the jumper to turn the transistor on and off.

Figure 1



An unused PIC µC input acts as an input for power-on status reading and as an output to drive an external transistor on and off.

You can easily extend this method to multiple jumpers and BCD switches. The only condition is that the voltage that the jumpers determine should be less than 3V and that impedance should be high. (DI #2475)

To Vote For This Design,
Circle No. 302

Keyboard data-acquisition system is cheap and simple

Tom Lyons Fisher, Juniata College, Huntingdon, PA

The most important criteria of a data-acquisition system for college science laboratories are simplicity and price rather than precision or speed. The data-acquisition system in **Figure 1** offers adequate precision of less than 0.5% and speed of 1 Hz to replace the outdated laboratory chart recorder in student laboratories. You can install the system in 5 sec, operate it with a single toggle switch, and construct it for approximately $40. The only additional equipment necessary is a computer running Excel and an ATbus (not Universal serial bus), keyboard.

The system is simple to operate. After you install it between the keyboard and the PC, the keyboard functions normally until you close the toggle switch, which puts the circuit into "acquire" mode. The system then bypasses the keyboard and "types" data into an Excel spreadsheet column at the rate of 1 point/sec. When you switch off the system, the circuit finishes sending the current data before returning control to the keyboard. The slow sampling rate gives Excel time to replot an entire column of data and thus appear to be charting in real time.

The central IC in the circuit is the PIC12C671-04 µC, IC$_2$, which has an onboard 8-bit ADC. The circuit configures this µC to receive an analog voltage through the A/D pin. Because laboratory instruments output 1V full scale and the ADC's internal reference is set for 5V full scale, a rail-to-rail single-supply op amp, IC$_1$, provides a gain of 5. The op amp's feedback circuit also acts as a low-pass filter. The system has acceptable offset of -1.2 bits and displays excellent linearity; the coefficient of determination, R^2, equals 0.99998.

Because no pins are available for external clocking, the circuit allows the PIC µC to run at approximately 4 MHz using its internal RC oscillator. However, this oscillator is not a sufficiently accurate timebase for even 8-bit precision, so an external 16.384-MHz oscillator-divider, IC$_4$, produces a 64-kHz waveform that feeds into the Tmr0 pin of the PIC. The combination of a divide-by-256 prescaler and the appropriate period loaded into Timer 0 provide an accurate 1-sec interrupt.

The PIC μC "types" to the computer by outputting signals that emulate the keyboard via the Clk and Dat lines of IC$_2$. These pins duplicate the wired-OR electrical characteristics of the keyboard interface. When the data-acquisition system is active, the keyboard must not connect to the computer. The circuit fulfills this requirement using analog switches inside IC$_3$ in the keyboard clock and data lines. The μC controls these switches using the Sw signal.

The keyboard line powers the entire circuit, and the circuit shields the handle of the spst switch as a protection from static electricity. You can download the source program for the PIC from *EDN*'s Web site, www.ednmag.com. Click on "Search Databases" and then enter the Software Center to download the file for Design Idea #2478. You can then compile under MPLAB 4.12 or use the .hex file. (DI #2478)

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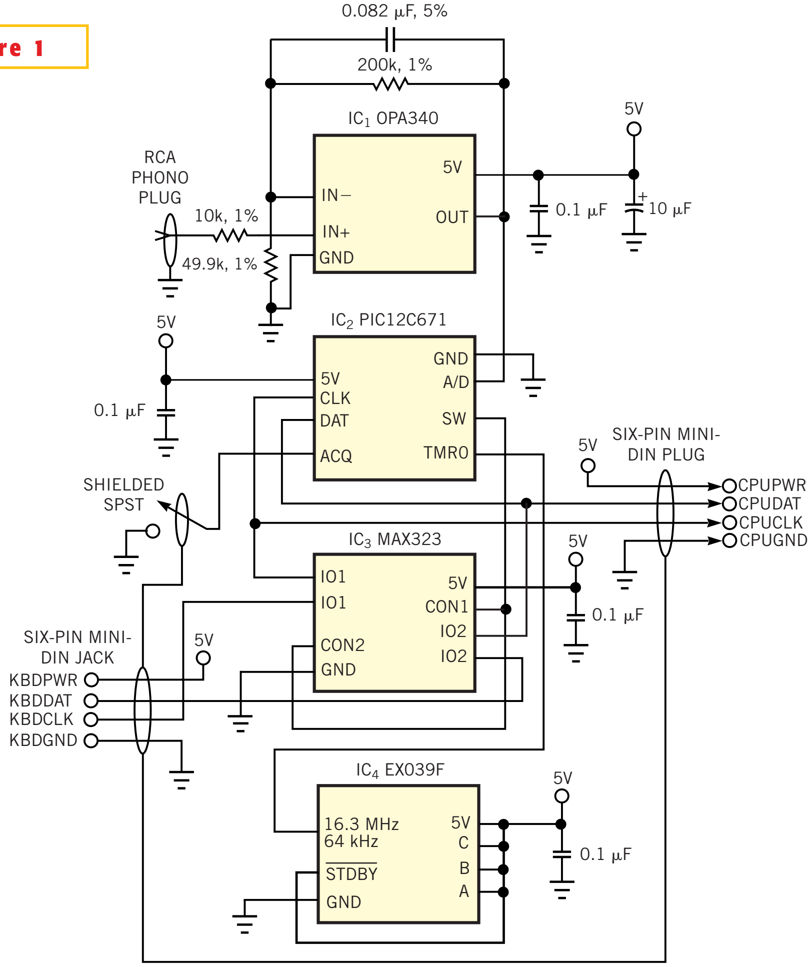
Figure 1



When the switch is in the acquire position, this data-acquisition-system circuit bypasses the keyboard and puts data into an Excel spreadsheet column at the rate of 1 point/sec.

AC-couple analog signals without a dc bias source

Joseph Luis Sousa, Linear Technology Corp, North Chelmsford, MA

A conventional ac-coupling circuit in a single-supply system comprises a series capacitor and a shunt resistor to ground. Unfortunately, the negative peaks of the input signal can exceed the −0.3V input operating-range limits of an ADC, such as the LTC1402 serial 12-bit sampling ADC. To avoid going below ground, the circuit must return the shunt input resistor to a midsupply voltage source. This problem is classic with all single-supply ADCs. You can use fully differential analog inputs, such as those of the LTC1402 2.2M-sample/sec, 12-bit, serial ADC, to ac-couple an analog signal without this midsupply bias voltage.

The ADC inputs derive the common-mode dc operating voltage directly from the input signal. The circuit has two requirements: The analog input signal must remain between 0V and the 5V supply voltage, and the ac transients must remain below the ±2V bipolar input range of the ADC. In **Figure 1a**, R$_1$ and the grounded C_1 at the A_{IN-} input of the ADC cancel the low-frequency signals and provide the basic ac-coupling function. R_2 and its shunt capacitor, C_2, at the ADC's A_{IN+} input cancel the sampling current bias offset. The optional C_3-R_3 46-MHz lowpass network isolates the ADC input from sampling-glitch-sensitive circuitry.

The frequency response for the values in the circuit has a low-cutoff pole at 1 kHz and low-frequency stopband rejection in excess of −60 dB, as set by the common-mode-rejection specification of the ADC, independent of RC-component-match accuracy (**Figure 1b**). The LTC1402 accepts wide bandwidth, full-scale, 4V p-p signals as great as 80 MHz. This ac-coupling circuit adds no distortion to the input signal. You can couple a 1.1-MHz Nyquist frequency sine wave into the ADC while keeping the THD below −82 dB.(DI #2479)

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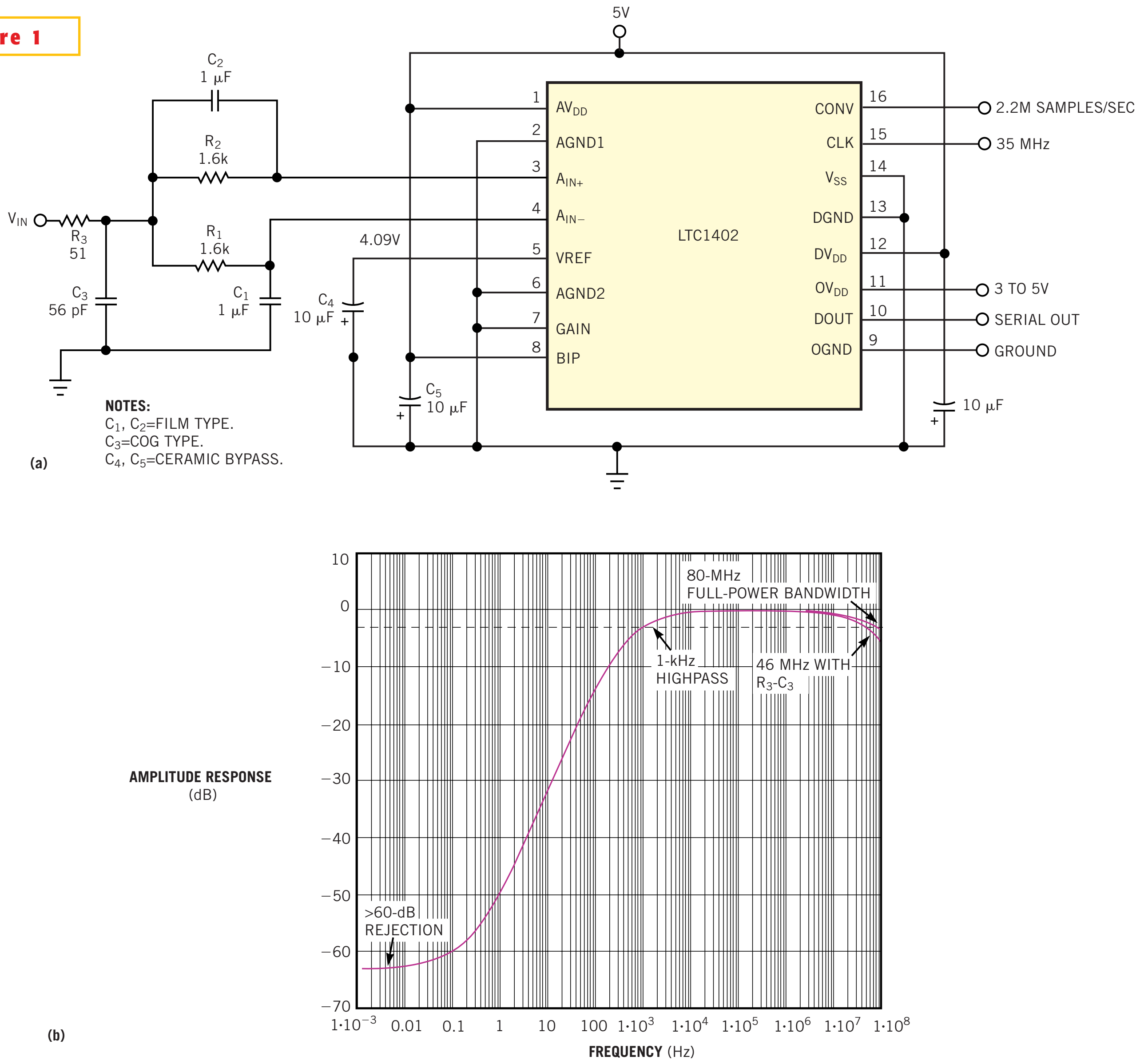
Figure 1



You can use fully differential analog inputs, such as those of the LTC1402 ADC, to ac-couple an analog signal without this midsupply bias voltage (a). The circuit's frequency response includes a low-cutoff pole at 1 kHz and low-frequency rejection of 260 dB (b).

Simple active-matrix-LCD bias supply operates from battery input

Michael Shrivathson, National Semiconductor Corp, Santa Clara, CA

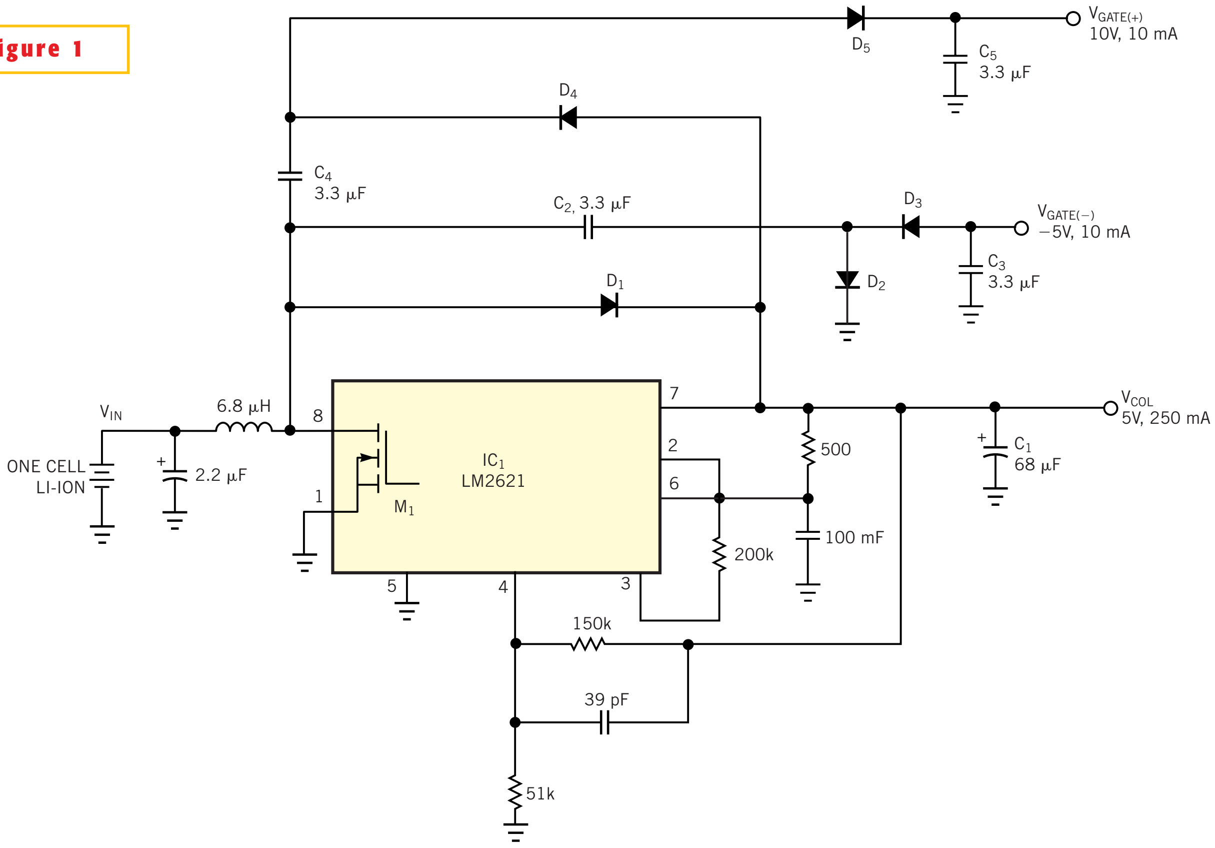
Many active-matrix-LCD applications need multiple voltages for thin film-transistor (TFT) bias. Typically, three voltages are necessary: 5V for the column driver; a positive voltage, such as 10V; and a negative voltage, such −5V, for the TFT gate drive, or row driver. For handheld electronic devices, a battery must produce these voltages. The most popular batteries in these devices are two-cell NiCd alkaline or one-cell lithium-ion batteries.

Figure 1 shows a simple, cost-effective way of providing these bias voltages. A step-up regulator, IC_1, forms the heart of the circuit. This regulator switches at a constant frequency of 1 MHz and a fixed duty cycle of 70%. IC_1 steps up the input voltage to 5V by storing the energy in the inductor when the internal MOSFET, M_1, is on and transferring this energy to C_1 when M_1 is off. IC_1's hysteretic gated-oscillator control scheme achieves the regulation.

C_2, C_3, D_2, and D_3 form a charge-pump inverter to provide an output of approximately −5V. When M_1 is off, C_2 connects in parallel with C_1 through D_1 and D_2. Thus, C_2 charges to V_{COL}, or 5V. When M_1 turns on, C_2 connects in parallel with C_3 through M_1 and D_3. Because of the polarity of this connection, C_3 charges to approximately $-V_{COL}$, or −5V.

C_4, C_5, D_4, and D_5 form a charge-pump doubler that provides an output of 10V. When M_1 is on, C_4 connects in parallel with C_1 through D_4 and M_1. Thus, C_4 charges to V_{COL} (5V). When M_1 turns off, C_1 and C_4 connect in series through D_1 and D_5, and this series pair connects in parallel with C_5. Thus, C_5 charges to approximately two times V_{COL}, or 10V.

Figure 1



A step-up regulator, IC_1; a charge-pump inverter comprising C_2, C_3, D_2, and D_3; and a charge-pump doubler comprising C_4, C_5, D_4, and D_5 produce the three voltages necessary for active-matrix-LCD applications.

This circuit provides 250 mA at the 5V output, V_{COL}, with 3% accuracy. The ac ripple is less than 100 mV. The circuit regulates the 10V output, $V_{GATE(+)}$, with 5% accuracy, and this output can provide 10 mA. The ac ripple at the 10V output is approximately 30 mV. The circuit regulates the −5V output, $V_{GATE(-)}$, with 6% accuracy and provides as much as 10 mA of output current. The ac ripple voltage at this output is 40 mV. A minimum load of 25 mA at the V_{COL} output ensures sufficient charge-pump action and thus maintains $V_{GATE(+)}$ and $V_{GATE(-)}$ at their nominal values. The efficiency of this circuit varies from 75 to 82% when operating from a one-cell lithium-ion battery. (DI #2477)

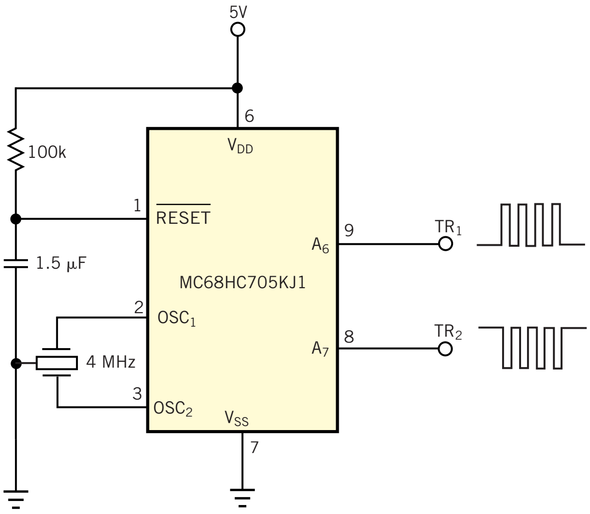
To Vote For This Design,
Circle No. 305

µC generates a frequency burst

Abel Raynus, Armatron International Inc, Melrose, MA

Pulse-sonar applications require generating bursts of a given frequency, duration, and repetition rate. Traditionally, the burst generator comprises a crystal oscillator with pulse modulation. But the easiest and cheapest way to generate the bursts is by using an inexpensive 8-bit µC, such as the 68HC705KJ1 and 68HC705J1A (Motorola) and do the whole job using software. You can get additional benefits by outputting two signals in opposite phase to feed the ultrasonic transducer directly or via a push-pull buffer (**Figure 1**). Note that only two µC pins are necessary for burst generation. You can use the rest of the pins for different purposes.

Figure 1



The easiest way to generate bursts for pulse sonar applications is to use a single µC and do the whole job in software.

The highest frequency that the µC can generate depends on the value of the highest oscillator frequency, f_{OSC}, that the manufacturer specifies and the structure of the instruction set, namely the quantity of machine cycles the µC takes to execute an instruction. With f_{OSC}=4.00 MHz, the mentioned µCs can generate a maximum frequency of 58.8 kHz. This value is a good match for sonar projects because most of the ultrasonic transducers, working in an air medium, have a standard resonant frequency of 40 kHz. To lower the frequency from 58.8 to 40.0 kHz requires a simple delay of 4 µsec using nop and brn instructions.

The constant value in the counter "Number" determines the burst duration. With one 8-bit counter, the burst duration can range from 0.1 to 3.2 msec. If a longer burst is necessary, you can add one or two more counters. If you choose a duration of 1 msec, as in this case, the value to put into the counter is

$$\text{NUMBER} = \frac{\text{BURST DURATION}}{\text{HALF OF PERIOD}} = \frac{1000\ \mu\text{SEC}}{12.5\ \mu\text{SEC}} = 80.$$

How you program the burst, repetition rate depends on the timer structure of the µC. For µCs with 16-bit programmable timers, the best way is to use either timer-overflow or output-compare functions. For µCs with multifunction timers, only the first eight timer stages are usable. Thus, timer overflow occurs every 0.51 msec, which is too short for a repetition period. So, you can use either real-time interrupt or, as in this case, organize a pacemaker based on the timer-overflow-interrupt. This design generates a burst every time the counter T rolls over from $FF to $00 with a repetition period of 131 msec. You can download the accompanying programs from *EDN*'s Web site, www.ednmag.com. Click on "Search Databases" and then enter the Software Center to download the files for Design Idea #2480. (DI #2480)

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