Edited by Brad Thompson

Power up a microcontroller with pre-power-down data

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It is sometimes necessary to retrieve data at power-up in the same way that they were at the last power-down, so that the product wakes up in the state it had before shutdown or to retrieve some measurement. One approach is to save critical variables into EEPROM or flash memory as soon as they change. This approach is generally not a good idea, because flash is typically limited to 100,000 write cycles, and EEPROM is typically limited to 1 million cycles. These numbers may seem large, but a product can easily reach them during their lifetimes.

Another approach is to use a battery to keep the microcontroller supplied so that it doesn't lose its RAM contents. This Design Idea presents an alternative option: detecting a power-down and triggering an interrupt routine that saves all the parameters in EEPROM or flash before the microcontroller supply falls below the operating threshold. **Figure 1** implements such an approach for a PIC18F6720 microcontroller.

One of the many features of this microcontroller is its low-voltage detection, which can trigger an interrupt when its LVD input goes below a threshold. You can set the threshold at 2.06V to 4.64V.

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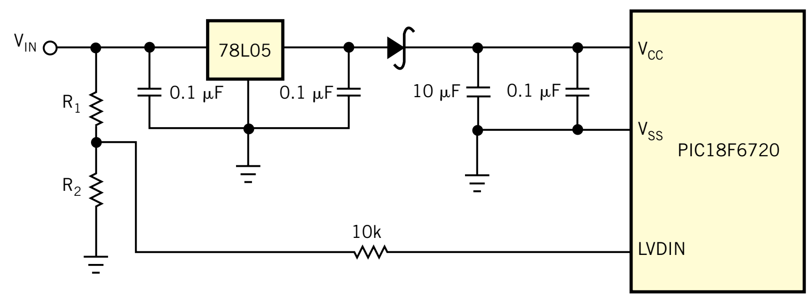

Figure 1 Use a power-down-detection scheme with a PIC18F6720 to save data before the microcontroller ceases to function.

The PIC18 microcomputer ceases functioning when its voltage supply is less than 4.2V. Because the EEPROM/flash-saving cycle is fairly time-consuming, the tactic is to monitor the voltage at the input of the 5V regulator to detect the power drop even before the microcomputer's supply starts to drop.

Select the LVD trip point inside the PIC18F6720 to be 1.22V, and calculate the required value of R_2/R_1 with the following equation:

$$\frac{R_1}{R_2} = \frac{V_{IN\_THRESHOLD}}{1.22} - 1,$$

where $V_{IN\_THRESHOLD}$ is the trip point below which a "data-save" function triggers. You should select this trip point to be as high as possible but not too high to avoid triggering on the ripples and noise on V_{IN-}.

Figure 2 shows the V_{IN} and V_{CC} waveforms when a power-down occurs. The ΔT represents the time allowed for saving data, which starts when the circuit detects the drop of V_{IN} and finishes when the voltage on the microcontroller goes below 4.2V, at which point it ceases to function. If the same 5V supply powers other devices , add a Schottky diode in series to ensure sufficient energy storage for the microcontroller to save the data. **Listing 1** in the Web version of this article at www.edn.com contains the assembly code that saves the data when a power-down occurs and retrieves the saved data at power-up.□

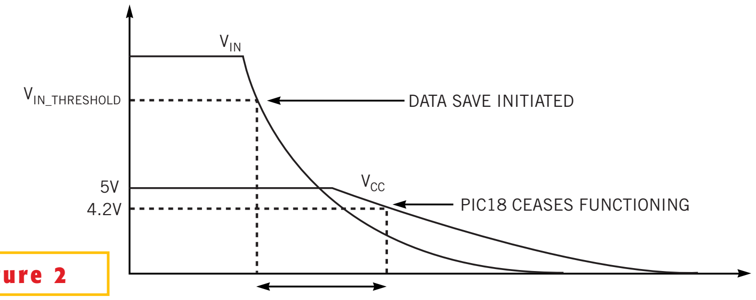

Figure 2 The V_{CC} and V_{IN} waveforms at power-down indicate the relationships in the sequence of events.

Power MOSFET is core of regulated-dc electronic load

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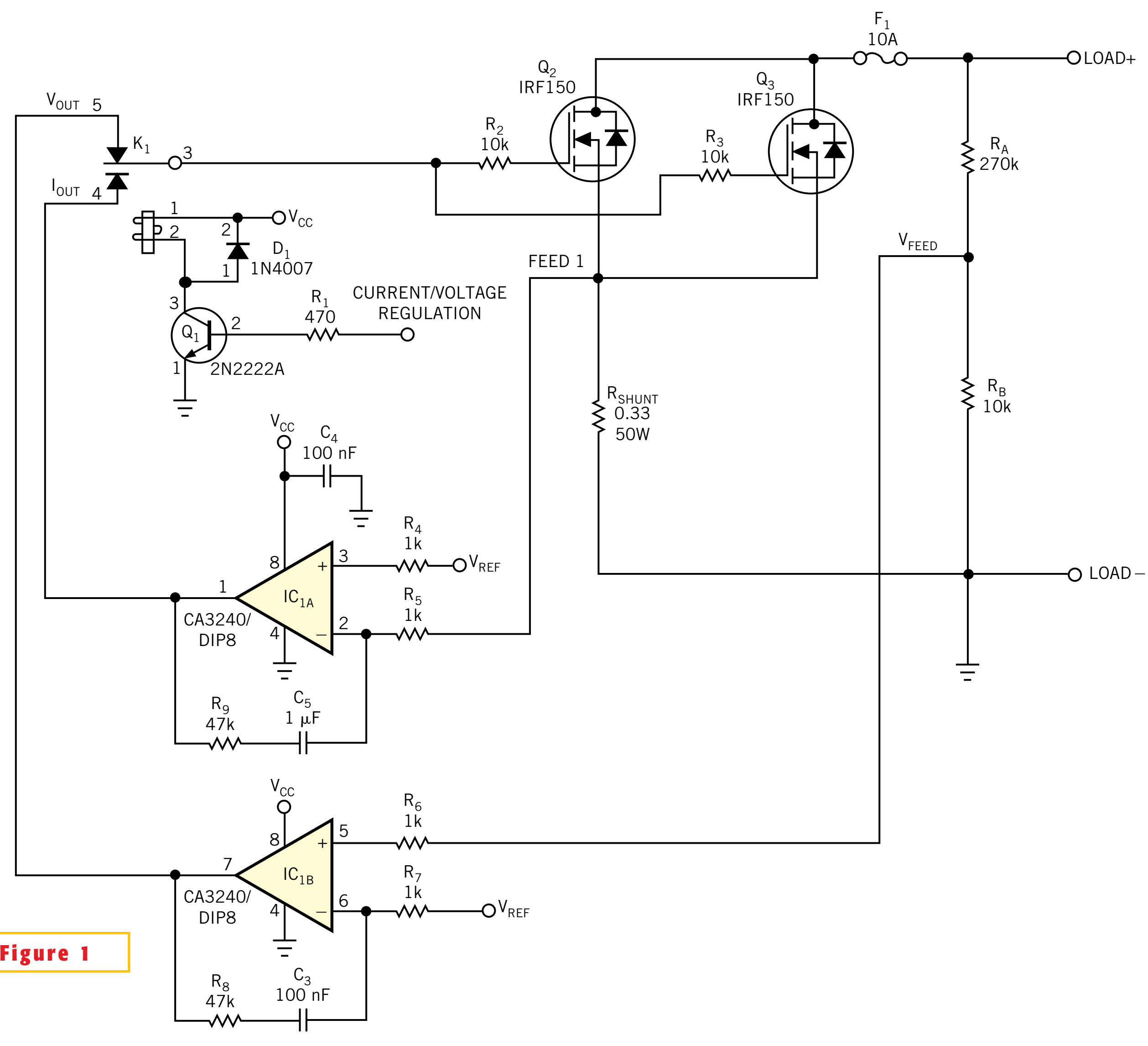
Designers use electronic dc loads for testing power supplies and sources, such as solar arrays or batteries, but commercial ones are often expensive. By using a power MOSFET in its linear region, you can build your own (**Figure 1**). It uses two simple feedback loops to allow the transistors to work as a current drain in current-regulation mode or as a voltage source in voltage-regulation mode. Designers use current-regulation mode when they are characterizing voltage sources, in which the power source must deliver current value that is set in the electronic load. They use voltage-regulation mode with current sources because it forces the sources to operate at a voltage that the load sets.

In current mode, R_{SHUNT} senses I_{LOAD}, and the resultant voltage feeds back to the inverting input of op amp IC_{1A}. Because the dc gain of this amplifier is high in the linear-feedback operating range, the inverting input stays equal to the

Figure 1



Using MOSFETs and a relay, this electronic load can operate in both current- and voltage-regulation modes.

noninverting input, which corresponds to V_{IREF}. The amplifier establishes its output value to operate MOSFETs Q_2 and Q_3 in a linear region and, therefore, dissipate the power from the source. The value of the source current is proportional to the current-loop reference, V_{IREF}, and is $I_{LOAD}=V_{IREF}/R_{SHUNT}$. Set V_{IREF} using a resistive voltage divider connected to a stable reference, or use the output of a D/A converter from a PC-based I/O card for flexible configuration.

Voltage-operating mode is similar, but now the sensed variable is the output voltage, which voltage divider R_A/R_B attenuates, so that the electronic load can operate at higher voltages than the op-amp supply voltage. The sensed voltage feeds back to the noninverting input of IC_{1B}, and the MOSFETs again operate in the linear region. Load voltage $V_{LOAD}=V_{VREF}\times(R_A+R_B)/R_B$.

The dual-op-amp CA3240, IC_1, can operate with an input voltage below its negative supply rail, which is useful for single-supply operation, but you can use any op amp if you have a symmetrical supply. Relay K_1 switches operating mode through a digital control line driving Q_1. The MOSFET is critical; you can add the IRF150 devices this design uses in parallel to increase the current-handing capabilities due to their positive-temperature coefficient, which equalizes the current flowing in the parallel MOSFETs. With the two MOSFETs in the circuit, the load handles 10A, and power consumption is greater than 100W, so using a heat sink and small fan is a good idea.

This circuit is useful for characterizing photovoltaic modules, which have two source modes. With this circuit and a PC-based setup, the I-V characteristic of a photovoltaic module from Helios Technology (www.heliostechnology.com) shows a region above V_{MPP} (voltage in the maximum point), at which a sharp transition corresponds to a voltage source (**Figure 2**). At voltages below V_{MPP}, the photovoltaic modules look like a current source. It is normally difficult to characterize this flat region of the curve with a simple current-mode electronic load, because the voltage output is sensitive to small variations in current, and thus a constant-voltage mode load is a better choice.□

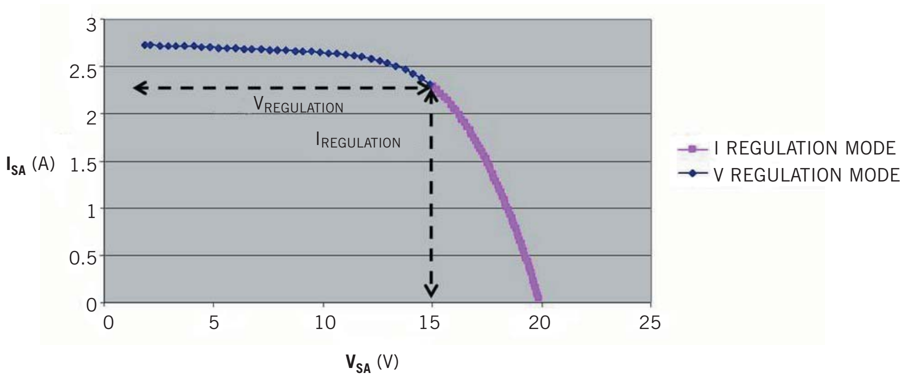


Figure 2 The I-V characteristics of a photovoltaic module, using the electronic load, show the special attributes of these power sources.

Use PSpice to model distributed-gap cores

Jeff Fries, GE Transportation Systems Global Signaling, Grain Valley, MO

PSpice software lets you create magnetic-core models that simulate nonlinear magnetic devices (**Figure 1**). These simulations are useful for observing hard-to-measure magnetic parameters such as core flux density, especially when you cannot quickly procure a sample device. The required inputs to the PSpice magnetic-core model are initial permeability of the core material, data points from the B-H magnetization curve, and the physical properties of the core, such as magnetic-path length, cross-sectional area, and air-gap length.

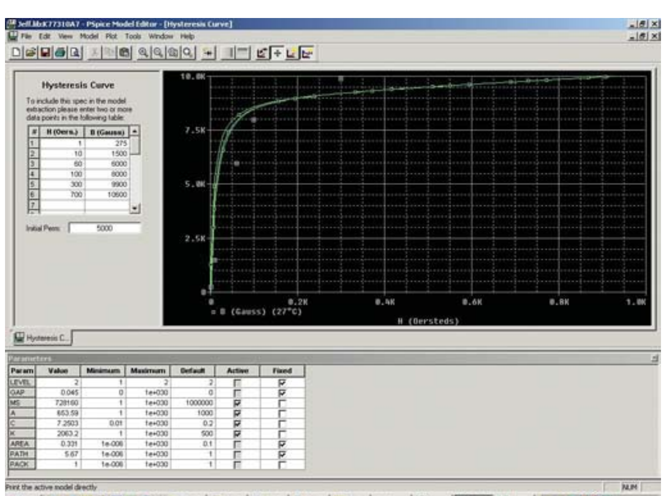


Figure 1 You use a screen capture from PSpice Model Editor to create a magnetic-core model.

All of the needed inputs for the magnetic-core model are typically available from core manufacturers' data sheets. However, in the case of a distributed gap with powder cores such as MPP or KoolMu, you need to determine the equivalent air gap to model the core using PSpice, because it relies on the air-gap length as input data to the model. Using the conservation of flux and manipulating Ampere's Law for a magnetic circuit with an air gap result in: $1/U_E=(1/U_I)+(L_G/L_E)$, where U_E is the effective permeability of the core, U_I is the initial permeability of the core material, L_G is the length of the gap in centimeters, and L_E is the magnetic-path length of the core in centimeters. Assuming that the initial permeability, U_I, of the core is high, which is typical of distributed-gap cores, then the term $1/U_I$ drops out, and you can rearrange the equation to solve for the gap length as $L_G=L_E/U_E$. Using the magnetic-path length, L_E, and effective permeability, U_E, that the core manufacturer's data sheet specifies, calculate the equivalent air-gap length of the distributed gap-core for use in the PSpice model.

As an example, take the KoolMu 77310-A7 toroidal powder core from Magnetics Inc (www.mag-inc.com). Because the data sheet does not specify the initial permeability of the KoolMu core, arbitrarily use 5000. (This parameter is insignificant in the model due to the air gap.) Use the magnetization curve for the KoolMu material and mark the data points in **Table 1**.

Physical data for the 77310-A7 core shows a magnetic path length of 5.67 cm, cross-sectional area of 0.331 cm^2, and effective permeability of 125. From this data, you calculate the effective air-gap length of 0.045 cm. Enter this data into PSpice for the core model.

A quick and easy way to verify the accuracy of the model is to create an inductor in PSpice using your magnetic-core model. Place the inductor in a series-tuned RLC circuit (**Figure 2**). Using PSpice, run an ac sweep of the circuit, and use a probe to find the resonant frequency, f_{RES}. Using the resonant frequency, you can calculate the measured inductance of the PSpice model as $L_{MEAS}=1/(4\times\pi 2\times f_{RES}2\times C)$. If your magnetic-core model is correct, this should be close to the expected inductance calculated as $L_{EXP}=(N^2)\times A_L$, where N is the number of turns, and the core data sheet typically supplies the inductance factor, A_L.□

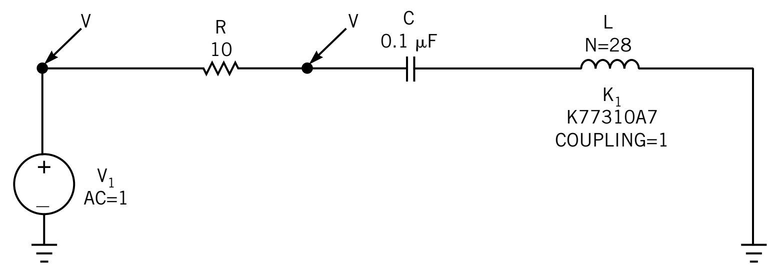


Figure 2 You can use this RLC-resonant circuit to measure the modeled inductance in PSpice, using an ac sweep analysis.

TABLE 1—DATA POINTS FOR KOOLMU CORE

B (Gauss)	H (Oersteds)
275	1
1500	10
6000	60
8000	100
9900	300
10,600	700

Precision divide-by-two analog attenuator needs no external components

Moshe Gerstanhaber and Chau Tran, Analog Devices, Wilmington, MA

Many modern A/D converters offer only a 5V input range, and using these converters with a ±5V or larger input signal gives the designer a problem: how to discard half of a good analog signal without introducing errors and distortion. To solve the problem, you can use an attenuator comprising two operational amplifiers and two resistors (**Figure 1**). However, this approach can reduce a system's performance by introducing gain errors due to amplifier offset and drift and resistor mismatch.

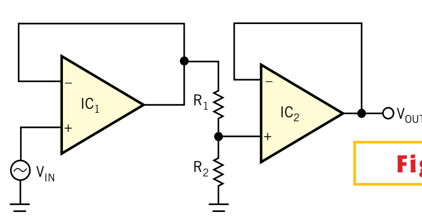


Figure 1 A conventional active voltage divider requires two op amps, IC_1 and IC_2, and two resistors, R_1 and R_2, that form a 2-to-1 voltage divider.

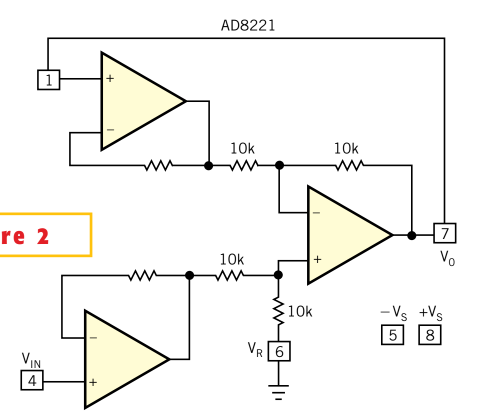


Figure 2 You can use an instrumentation amplifier to halve an analog signal's amplitude. All resistors are internal to the IC.

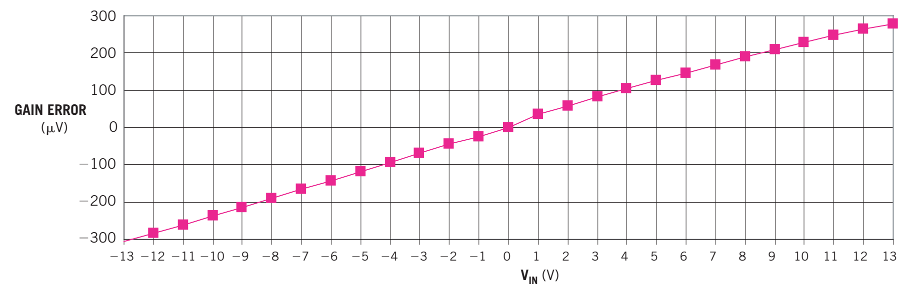
Figure 2 shows an alternative circuit that provides a precision gain of one-half with low offset, low drift, and low input-bias currents and that uses an AD8221 instrumentation amplifier.

The amplifier's output, V_O, equals the difference between the two inputs, V_{IN+} and V_{IN-}: $V_O=(V_{IN+})-(V_{IN-})$. Connecting the amplifier's output to its inverting input and substituting V_O for V_{IN-} yields: $V_O=(V_{IN+})-(V_O)$, or $V_O=(V_{IN+})$.

Thus, the circuit provides a precision gain of one-half with no external components and, in this configuration, is unconditionally stable. The performance plots of **figures 3** and **4**, respectively, show a gain error of less than 300 μV and a maximum nonlinearity error of about 1 ppm over a 26V input-voltage range.

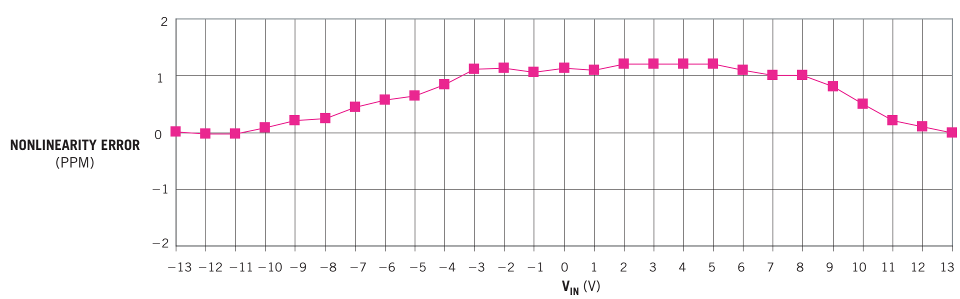
To introduce an offset voltage, V_{OS}, that equals half of a reference voltage ($V_{OS}=V_R/2$), connect the AD8221's reference input (Pin 6) to voltage V_R. To bias the attenuator's output at half of the positive- or negative-power-supply voltage, connect the reference pin to the appropriate power supply.□

Figure 3



The circuit in Figure 2 introduces a full-scale gain error of less than 300 μV over a 26V input-voltage range.

Figure 4



The circuit in Figure 2 introduces a maximum nonlinearity error of about 1 ppm over a 26V input-voltage range.

Quartz crystal-based remote thermometer features direct Celsius readout

Jim Williams and Mark Thoren, Linear Technology Corp

Although quartz crystals have served as temperature sensors, designers haven't taken advantage of the technology because few manufacturers offer the sensors as standard products (**references 1** and **2**). In contrast to conventional resistance- or semiconductor-based sensors, a quartz-based sensor provides inherently digital-signal conditioning, good stability, and a direct digital output that's immune to noise and thus ideally suited to remote-sensor placement (**Figure 1**, pg 100).

An economical and commercially available quartz temperature sensor, Y_1 and IC_1, an LTC-485 RS485 transceiver in transmitter mode, form a Pierce crystal oscillator. The sensor, an Epson HTS-206, presents a nominal frequency of 40 kHz at 25°C and a temperature coefficient of −29.6/ppm/°C (**Reference 3**). The transceiver's differential-line-driver outputs deliver a frequency-coded temperature signal over a twisted-pair cable at distances as far as 1000 ft.

A second LTC-485, IC_2, in receiving mode, accepts the differential data and presents a single-ended output to IC_3, a PIC-16F73 processor that converts the frequency-coded temperature data and presents the temperature in Celsius format on LCD_1. You can view and download the conversion program's source code in the Web version of this Design Idea at www.edn.com.□

References

1. Benjamin, Albert, "The Linear Quartz Thermometer—A New Tool for Measuring Absolute and Differential Temperature," *Hewlett-Packard Journal*, March 1965.

2. Williams, Jim, "Practical Circuitry for Measurement and Control Problems," Application Note 61, August 1994, Linear Technology Corp.

3. HTS-206 specifications, Epson Corp, www.eea.epson.com.

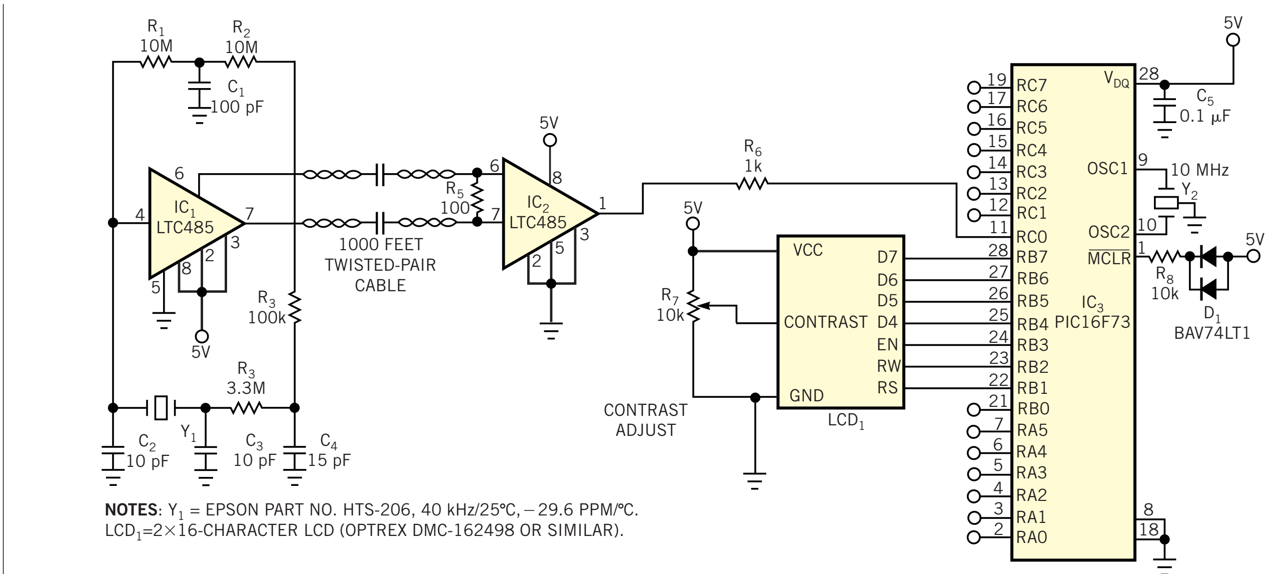


Figure 1 A quartz-crystal temperature sensor provides Celsius-temperature readouts accurate to 2% over a −40 to +85°C range.