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Design an RTD interface with a spreadsheet

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RTDs (resistance-temperature detectors) are the preferred sensor choices for designs requiring precision. Although RTDs are approximately linear over the limited temperature range of 0 to 100°C, these sensors exhibit a slight but progressively more nonlinear temperature-versus-resistance characteristic as the measurement range widens. Consequently, over an extended span, curve fitting is necessary if the system is to achieve a high level of precision. One way to obviate the nonlinear characteristic of an RTD sensor is to design analog hardware to perform the curve-fitting mathematics before any additional signal processing occurs. This approach is especially attractive if you can keep both cost and component count low and if a microprocessor-driven design is not feasible. With low component count comes the added benefit of a small PCB (printed-circuit-board) footprint.

The most popular RTDs are made from platinum with a resistance value of 100Ω at 0°C and a metal purity that allows them to follow a standard European curve with a positive-temperature coefficient, α, equal to 0.00385Ω/Ω/°C. Less popular but still common are RTDs with a slightly higher metal purity. These RTDs have α of

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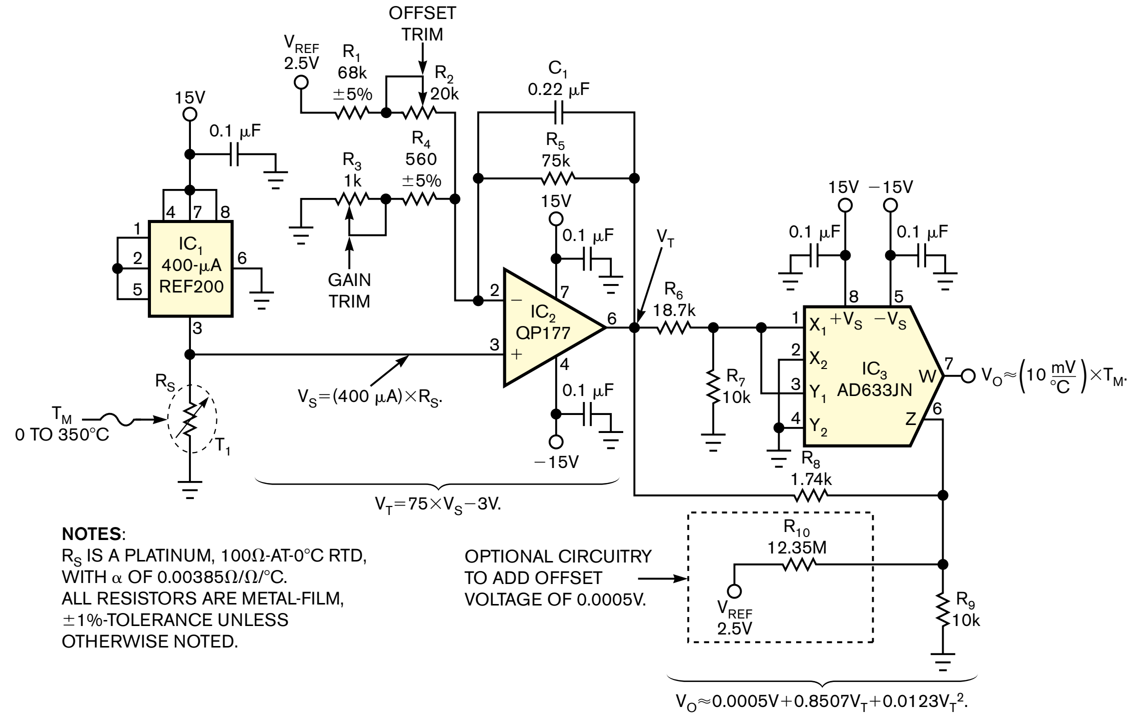

Figure 1 This RTD circuit uses a second-order polynomial to linearize the output of the sensor.

0.00392Ω/Ω/°C and follow the US curve. The circuit in **Figure 1** uses a standard RTD to measure temperature over the extended range of 0 to 350°C, an output voltage of 0 to 3.5V, and overall system accuracy greater than 0.5°C. The following linear **equation** expresses this sensor system:

$$V_O \approx \left(\frac{10\ \text{mV}}{°\text{C}}\right)T_M.$$

IC_1 is pin-configured to drive a constant current of 400 μA through the grounded sensor, T_1. Driving T_1 with this level of current—"zero-power" operation—keeps the worst-case power that the circuit dissipates in the sensor to less than 40 μW and reduces the self-heating errors to a second-order effect (**Reference 1**). Also, driving the RTD with a current source preserves its intrinsic nonlinearity and allows you to express the sensor's output voltage, V_S, as: 400 μA×R_S, where R_S is the resistance of the sensor.

IC_2 initially signal-conditions the sensor's output by first scaling the output voltage and then offsetting the result so that V_T is slightly larger than the 3.5V output at 350°C and that V_T equals 0V at 0°C. Adding gain and offset before linearization places less of a burden on the curve-fitting circuitry and helps to meet the system's precision specification. The combination of C_1 and R_5 implements a lowpass filter with a pole at approximately 10 Hz to remove power-supply noise. The following term describes the performance of IC_2 and its accompanying circuitry: $V_T=75V_S-3V$.

Next, an Excel spreadsheet creates the nonlinear-mathematical relationship between the voltage, V_T, and the system output, V_O (**Table 1**). The spreadsheet features 17 temperature entries—starting at 0°C, increasing in increments of 25°C, and ending at 400°C—for the measured temperature. Using a data set that extends beyond the intended measurement range of 350°C can reduce end errors in nonlinear systems. Values for R_S—which you derive from a standard RTD-resistance-versus-temperature table—and the **equations** allow you to compute V_S and V_T. The V_T and V_O columns are the input and output signals, respectively, for the linearization circuitry; you chart them using Excel's XY-scatter feature. You can use Excel's Trendline feature to create the following **equation**, the mathematical representation of the curve-fitting circuitry you need to linearize the sensor's output: $V_O=0.0005V+0.8597V_T+0.0123V_T^2$. IC_3 and four 1%-tolerant resistors or, optionally, five resistors implement a second-order polynomial: $V_O=a+bV_T+cV_T^2$, where a is the offset term, b is the linear coefficient, and c is the square-term coefficient.

TABLE 1 EXCEL-SPREADSHEET DATA

Measured temperature (°C)	R_S (Ω)	V_S (V)	V_T (V)	V_O (V)
0	100	0.04	0	0
25	109.73	0.0439	0.292	0.25
50	119.4	0.0479	0.582	0.5
75	128.99	0.0516	0.87	0.75
100	138.51	0.0554	1.155	1
125	147.95	0.0592	1.439	1.25
150	157.33	0.0629	1.72	1.5
175	166.62	0.0666	1.999	1.75
200	175.86	0.0703	2.276	2
225	185.01	0.074	2.55	2.25
250	194.1	0.0776	2.823	2.5
275	203.1	0.0812	3.093	2.75
300	212.05	0.0848	3.362	3
325	220.91	0.0884	3.627	3.25
350	229.72	0.0919	3.892	3.5
375	238.88	0.0956	4.166	3.75
400	247.09	0.0988	4.413	4

The curve-fitting-circuit design begins by first wiring the four inputs of IC_3 to create a positive square term that is scaled at the chip's output by an internal scale factor of ⅒V. Then, comparing terms, you find that the coefficient, c, must equal 0.0123. Because R_6 and R_7 form a voltage divider that attenuates the signal, V_T, you can express the coefficient with the following **equation**:

$$c=\frac{1}{10}\left(\frac{R_7}{R_6+R_7}\right)^2.$$

Select a value for R_7—10 kΩ for this design—and then use the preceding **equation** to find the value for R_6.

Resistors R_8, R_9, and, optionally, R_{10} form a passive adder to create the offset term, a, and the linear coefficient, b. You apply the output of the passive adder directly to the Z input, Pin 6 of IC_3, which adds the offset and linear terms to the square term to form the system response at Pin 7. Again comparing these terms, note that the offset term must equal 0.0005V. The offset term is only 0.5 mV, and eliminating it would add an error of approximately 0.05°C, so you can initially neglect it. Then, because the linear term's coefficient, b, must equal 0.8507, you first select a suitable value for R_9 and use the following **equation** to solve for R_8: $b=R_9/(R_8+R_9)$.

If you wish to design the optional circuitry and include the offset term, which is part of the passive adder, choose a stable 2.5V reference for V_{REF}, calculate the parallel combination of $R_8//R_9=R_{EQ}$ (the equivalent resistance of R_8 in parallel with R_9), and solve for R_{10} using the following voltage-divider **equation**: $a=(R_{EQ}/(R_9+R_{EQ}))V_{REF}$.

To calibrate this circuit, replace the sensor with a precision decade box. Set the decade box to simulate 0°C and adjust the offset trim of R_2 for an output of 0V at Pin 7 of IC_3. Next, set the decade box to simulate 350°C and adjust the gain trim of R_3 for an output of 3.5V. Repeat this sequence of trim steps until both points are fixed. The circuit in **Figure 1**—which includes optional circuitry—exhibits a worst-case measurement error at 250°C and 2.504V of 0.16%, or 0.4°C. Testing the circuit without the optional circuitry—the reference voltage and R_{10}—shows no discernible improvement in precision.**EDN**

REFERENCE

1 "IC Generates Second-Order Polynomial," *Electronic Design*, Aug 5, 1993, www.elecdesign.com/Articles/Index.cfm?AD=1&ArticleID=11502.

Isolated supply powers DVM module

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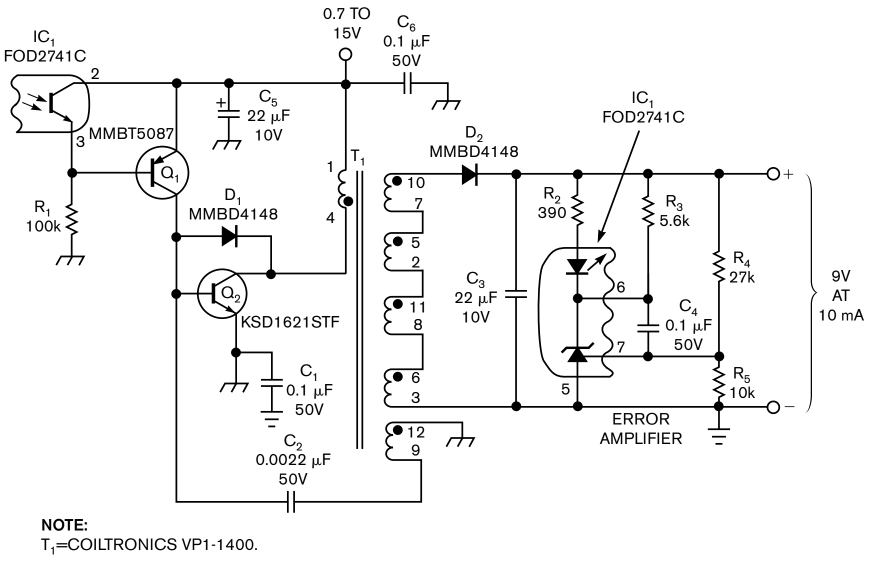
Low-cost DVM (digital-voltmeter) modules are economical and can significantly reduce design time for instrumentation. Yet, these modules also involve a significant number of design challenges. For example, their inputs are not isolated from the power supply, so you must add an isolated power supply. This task can both consume critical design time and add to system costs. Additionally, many uses for the modules require one- to four-cell-battery operation, and the modules require approximately 9V, translating to operation from 0.7 to 6V if you use new batteries until they are fully discharged. This wide input range also means that you should regulate the power-supply output.

DVM modules also have low parts count, and you can implement them using off-the-shelf components. Optionally, the modules can operate with input voltages as low as 0.25V if you replace the silicon transistors with germanium devices. However, germanium transistors are relatively expensive, so use them only in applications requiring low-input-voltage operation.

The power-supply design in **Figure 1** is a blocking oscillator that operates as a flyback converter with fixed on-time and variable off-time. The variable off-time regulates how often the transformer charges and delivers power to the load. The blocking oscillator consists of NPN transistor Q_2, transformer T_1, and capacitor C_2. The conductance of PNP transistor Q_1 controls the off-time of the oscillator in conjunction with C_2. The output of the transformer conducts to the energy-storage capacitor, C_3, through diode D_2 during transformer flyback. The error amplifier and optocoupler, IC_1, monitors the voltage across C_3. When the voltage at resistive divider R_4-R_5 exceeds 2.5V, the optocoupler conducts more and reduces the conduction of transistor Q_1, increasing the time required for the next power cycle.**EDN**



NOTE:
T_1=COILTRONICS VP1-1400.

Figure 1 This isolated flyback supply powers a digital-voltmeter module from 0.7 to 15V.

IC performs delayed system reset upon power-up

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In most applications, the $\overline{\text{MR}}$ (manual-reset) pin usually connects to a switch to create a manual-reset signal to the supervisory chip. Subsequently, after a predetermined time-out-active period, it goes back to the high state in an active-low reset. A manual reset is a good feature for most applications; however, it requires human intervention to create the reset. In some applications, a manual reset could be a hassle because you must perform it each time the system powers up.

Further, applications involving embedded microprocessors can require the reset output to hold high—that is, inactive—for a certain period of time before you can apply the reset, or active low. The circuit in **Figure 1** proves useful during power-up when there is no need to press the reset button once the device powers up, because reset occurs automatically with the predetermined hold time before you apply the reset-low signal.

The circuit employs a reset-supervisory chip with the $\overline{\text{MR}}$ pin and active-low output, $\overline{\text{RESET}}$. Normally, the $\overline{\text{MR}}$ input has an internal pullup resistor with a value of 20 to 50 kΩ. During power-up, this $\overline{\text{MR}}$ internal resistor charges up capacitor C_1 to the maximum value to V_{DD} at the positive side. To create an $\overline{\text{MR}}$ reset input to the supervisory chip, its $\overline{\text{MR}}$ input must receive an active-low ground signal, requiring transistor Q_1 to turn on. The turn-on-time period depends on the RC-time constant of R_1 and C_2. These two components determine when Q_1 turns on and thus provide an adjustable hold time for the $\overline{\text{RESET}}$ output to hold high. To increase the hold time, simply increase the RC-time constant of R_1 and C_2.

The supervisory reset chip asserts its $\overline{\text{RESET}}$ output only when the voltage at the $\overline{\text{MR}}$ pin exceeds the threshold-trigger voltage and the supervisor's internal reset period has elapsed. This time-out period filters any short input-voltage transients. Because of Q_1's turn-on, the negative side of C_1 becomes grounded. Because the positive side of C_1 cannot instantly change its polarity, it pulls low and slowly charges up again through the internal pullup resistor of the $\overline{\text{MR}}$ input. When it reaches the threshold voltage of the reset chip, it then asserts the reset once it reaches the time-out period of the chip. The selection of C_1 is not critical. However, its value should be sufficiently large—0.1 to 10 µF, for example—that the RC time constant for C_1 and the internal pullup resistor are large enough. This value ensures that C_1 holds the voltage low at $\overline{\text{MR}}$ for at least 1 µsec.

The transistor remains on after C_2 charges toward the biased voltage of Q_1. At the next power-up or when you manually reset the circuit by pressing the pushbutton switch, the transistor discharges C_2. Once this action happens, Q_1 turns off. R_1 charges up the negative side of C_1 to the supply voltage, V_{DD}. Because the positive side of capacitor C_1 cannot change instantly, it appears to be charged to $2V_{DD}$. However, the protection diode, D_1, clamps C_1's voltage to just V_{DD} plus the diode's turn-on voltage. The cycle repeats once C_2 charges enough to again turn on Q_1. **EDN**

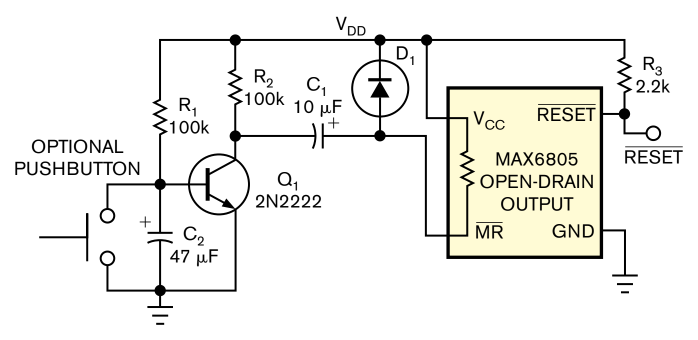

Figure 1 This simple circuit automatically resets a microcontroller upon power-up.

One microcontroller pin drives two LEDs with low quiescent current

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The basis for this Design Idea is a circuit that uses three resistors and a microcontroller-I/O pin to work as input high impedance or output to independently drive two LEDs (**Reference 1**). The idea sounded good for this design, mainly because of the lack of spare I/O pins in the microcontroller and the simplicity of the implementation. Unfortunately, you cannot use the circuit in battery-powered designs because it exhibits a current leakage on the order of 2 mA even with both LEDs off. This Design Idea modifies that circuit, using only one I/O pin to drive the two LEDs but with a low current drain (**Figure 1**). Although the circuit uses a couple of diodes and a resistor, the price and the component count are low.

THE LEDs START DIMMING AT APPROXIMATELY 4V WITH A CURRENT OF 80 µA AND ARE FULLY ON WITH 4.4V AT A CURRENT OF 1 mA.

The basis for the operation of both circuits is the nonlinear characteristic of a diode, in which current grows exponentially with the voltage applied across it. To describe the operation, suppose that the microcontroller pin is configured as an input, leaving the pin in high impedance. In the first circuit, assume that LEDs need a voltage of approximately 1.5V to turn on and that the small-signal-diode voltage drop is approximately 0.6V (**Figure 1a**). So, to turn on both LEDs, you theoretically need 4.2V. In practice, the LEDs start dimming at approximately 4V with a current of 80 µA and are fully on with 4.4V at a current of 1 mA. With 3.3V, leakage current is merely 2.41 µA. The nominal voltage for this circuit can be slightly lower than 3.3V, but, in that case, you should use Schottky diodes.

The second circuit is for supply voltages greater than 5V (**Figure 1b**). Using the values in the **figure**, the LEDs start dimming with 7V at 74-μA current and are fully on with 8.5V at 1 mA, remaining off for a 5V supply at 1.53 μA.

To turn on the LEDs, you must configure the microcontroller's I/O pin as an output; an output value of one turns on the lower LED, and a value of zero turns on the upper LED. If both LEDs must appear to be on, your program can cycle the port pin between one and zero with a frequency greater than 50 Hz. To calculate the value of the resistor in both cases, the following formulas apply: $R=(3.3V-V_D-V_{LED})/I_{LED}$ (**Figure 1a**), and $R=(V_{CC}-V_Z-V_{LED})/I_{LED}$ (**Figure 1b**), where I_{LED} is the desired LED-on current, V_D is the voltage across the diode when an I_{LED} current flows through it, V_Z is the zener-diode voltage, and V_{LED} is the forward voltage across the LED when an I_{LED} current flows through it. You should use a Schmitt trigger or an analog input for the I/O pin to avoid excessive current draw. **EDN**

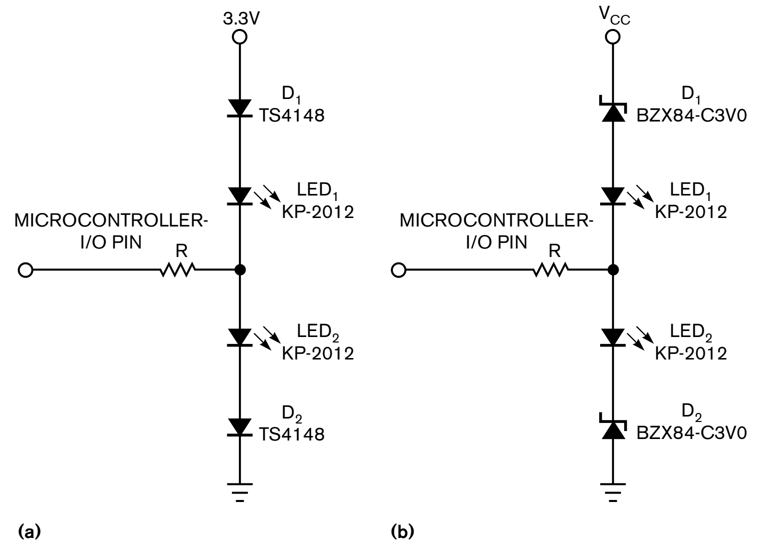

Figure 1 In these simple circuits, the LEDs need a voltage of approximately 1.5V (a) or more than 5V (b) to turn on. The circuits allow a battery-powered microcontroller to control two LEDs with only one I/O pin.

REFERENCE

1 Pefhany, Spehro, "Circuit Controls Two LEDs With One Microcontroller Port Pin," *Electronic Design*, April 1, 2002, http://electronicdesign.com/Articles/Index.cfm?AD=1&ArticleID=1683.