Isolation Barrier Precisely
Passes Transducer Signals

by Jeffrey A. Greenwald

Through the judicious use of isolation ampli-fiers that incorporate transformer coupling, you can design circuits that condition and amplify low-level transducer signals. Such circuits can precisely measure thermocouple, resistance-temperature-detector and strain-gauge outputs while providing common-mode isolation in harsh environments.

Isolation amplifiers protect data-acquisition compo-nents from potentially destructive voltages present at remote transducers. These amplifiers are also useful when you need to amplify low-level signals in multi-channel applications. They can also eliminate measure-ment errors caused by ground loops. Amplifiers with internal transformers reduce circuit costs by eliminat-ing the need for an additional isolated power supply. And in many cases it's possible to connect a transducer to an isolation amplifier with only a few external com-ponents.

The transducers that you'll most often encounter in-clude thermocouples, resistance temperature detectors (RTDs), pressure sensors and strain gauges. These sensors' outputs usually require high amplification. The sensors may also require some sort of external excita-tion, and you may have to perform filtering or other

signal conditioning to yield a useful signal. In some cases, you'll have to offset the sensor's output to place it into the proper signal range for the data-acquisition hardware. You can perform all of these functions with rather simple circuits and precise isolation amplifiers.

The isolation amplifiers you use in these circuits must have high enough breakdown potential to withstand ground faults and thus prevent system destruction. In many applications, you may be called upon to mount pressure transducers and strain gauges on metallic pipes and objects that are at different electric poten-tials than those of your data-acquisition system. High isolation comes in handy to break ground loops, to reduce noise, or to otherwise allow you to measure strain safely in the presence of several hundred volts.

The isolation amplifier needs to be able to respond to out of range transients that may overload pressure transducers. The AD208, for example, recovers from 5V input transients at a gain of 1000 in 5 msec without damage. Overload often happens when a valve closes, restricting the flow of fluid in a pipe. Since pressure is proportional to force and inversely proportional to cross-sectional area, the pressure rapidly rises to sev-eral times the full-scale pressure. This effect is called the water-hammer effect. The same condition is occur-ring when your pipes make a banging noise after you have shut off the water in the sink.

Possibly the most common transducer that you'll use isolation amplifiers with is a thermocouple. Thermocou-ples are low-cost sensors, and can measure tempera-tures over a wide range. **Table 1** lists some common types and their ranges. Thermocouples comprise two

dissimilar metals. A phenomenon called the Seebeck effect generates a temperature-dependent microvolt signal when the two dissimilar metals join to form a closed-loop circuit.

Although thermocouples are inexpensive, they have one major disadvantage. In addition to the desirable thermocouple effect that you want to measure, undesirable thermocouple effects occur when you attach the thermocouple to the outside world, typically in the form of copper wire. If the thermocouple material is not also copper—it typically is not—additional thermocouple effects known as cold-junction errors occur.

For example, the two measurement terminals of a J-type thermocouple consist of iron (Fe) and constantan (Con) contacts. (Constantan is a nickel-copper alloy.) When you attach each side of the thermocouple to copper wire, two different metals join—iron and copper on one end and constantan and copper on the other—to create thermally induced voltages whose values are a function of ambient temperature. The induced voltages at the constantan-copper and iron-copper cold junctions are opposite in polarity to the Seebeck voltage created by the thermocouple.

You can model the two cold junctions as two voltage sources in series with the thermocouple (**Fig 1**). While the thermocouple is sensing temperature at a specific physical location, these cold junctions are functions of the ambient environment. The voltage at the measurement terminal, $V_M(T)$, is the difference between the
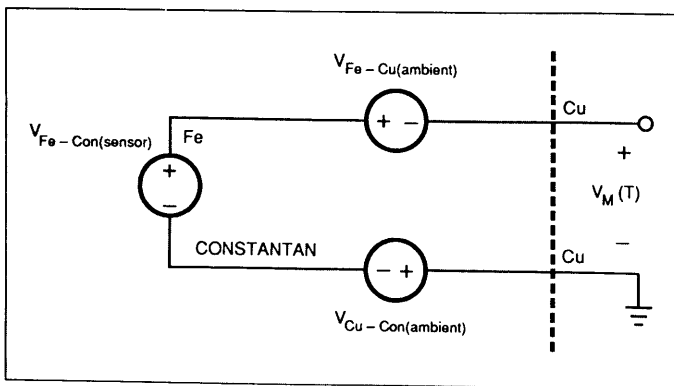
Fig 1—When you connect the two ends of a thermocouple to a dissimilar metal such as copper, you generate unwanted thermocouple effects called cold junctions. Voltage sources in series with the thermocouple's voltage model the effects of these junctions.

Table 1—Common thermocouples and their characteristics

Thermo-couple type	Maximum temperature range at 5V output (°C)	Maximum V_{IN} (mV)	Isolation amplifier's gain setting (V/V)	Suggested resistor values with reference to Fig 2		
				R_F (kΩ)	R_{TRIM} (kΩ)	R_Z (MΩ)
E	900	68.783	72.69	6.98	0.5	2
J	750	42.283	118.25	11.5	1	2
K	1250	50.633	98.75	9.53	1	2
R	1450	16.741	298.6	28.7	2	2
S	1450	14.973	333.9	32.4	2	2
T	350	17.816	280.6	27.4	2	2

voltage at the thermocouple and the sum of the induced voltages, as follows:

$$V_M(T) = V_{\text{Fe-Con(sensor)}} - V_{\text{Cu-Con(ambient)}} - V_{\text{Fe-Cu(ambient)}}.$$

Because the two cold junctions have a back-to-back, copper-to-copper connection, you can combine them in a single iron-constantan voltage source. You can then rewrite the voltage at the measurement terminal:

$$V_M(T) = V_{\text{Fe-Con(sensor)}} - V_{\text{Fe-Con(ambient)}}.$$

Canceling this cold-junction effect requires an opposing voltage in series with the thermocouple. This offsetting voltage should generate zero volts at zero degrees C and have a positive temperature slope equal to the thermocouple type you're using. In effect, you need a temperature sensor that measures the cold-junction ambient temperature and produces an offsetting voltage.

One method to measure and compensate for the cold-junction is to use a monolithic cold-junction compensator in series with the thermocouple (**Fig 2**). IC_1 generates a temperature-dependent output of 10 mV/°C. An internal laser-trimmed resistor network scales this signal to provide the correct slopes for most commonly used thermocouple types. You can access the scaled compensation voltage on one of four taps for various standard thermocouples: 60.9 μV/°C (E-type), 51.7 μV/°C (J-type), 40.6 μV/°C (K-type), and 6 μV/°C (R- and S-type).

The isolated ±8V dc supply at pins 5 and 6 of the isolation amp can supply as much as ±5 mA to the cold-junction compensator. However, because IC_1 only draws a maximum of 150 μA, there is sufficient power left over if you need to add more circuitry, such as an

input filter. In turn, the transformer-coupled isolation amplifier obtains its power from the power-clock driver, IC_3. The clock driver produces a 25-kHz, $15V_{p-p}$ square wave, which can drive up to 16 fully loaded isolation amplifiers.

You obtain the maximum sensitivity from the thermocouple circuit by amplifying the low-level input signal to the isolation amp's full-scale output range. The following equation yields the necessary gain from a few known values: the output voltage span of the thermocouple that corresponds to the thermocouple's temperature extremes ($V_{TC-HI} - V_{TC-LO}$); and the input-voltage requirements of your data-acquisition electronics, $V_{HI} - V_{LO}$.

$$Gain = (R_F/R_I + 1)$$

$$= (V_{HI} - V_{LO})/(V_{TC-HI} - V_{TC-LO}).$$

From the values in the **table**, you can see that a J-type thermocouple has a maximum temperature of 750°C, corresponding to a maximum output of 42.283 mV. Thus, the gain required to amplify the 0- to 42.283-mV signal to a 0 to 5V output span is $(5 - 0V)/(42.283$ mV $- 0$ mV$)$, or 118.25.

You can determine the optimal feedback resistor value, R_F, by first setting R_I to 100Ω, and then solving for R_F using the equation $R_F = (Gain - 1)R_I$, which yields $R_F = 117.25 \times 100$. Note that in **Fig 2**, R_F is the sum of R_F' and R_{TRIM}. To obtain a bipolar adjustment range of ±5% for the gain, choose $R_F' = 0.95R_F$, and $R_{TRIM} = 0.1R_F$.

Similarly, you can use R_Z to set the offset adjustment range. The following equation relates this resistance to the gain, feedback-resistor value, isolation voltage, V_{ISO}, and two offset voltage values. One of these offset values, V_{OS-RTI}, is the referred-to-input offset; the other, V_{OS-RTO}, is the referred-to-output offset.

$$R_Z = V_{ISO}R_F/(V_{OS-RTO} + (Gain \times V_{OS-RTI}).$$

The AD208 isolation amplifier, for example, typically has a V_{ISO} of 8V, a V_{OS-RTO} of 15 mV max, and a V_{OS-RTI} of 250 μV max. Using these values and those for the gain and R_F, you can solve for $R_Z \approx 2$ MΩ.

Perform calibration first

For the circuit to achieve maximum accuracy, you'll need to calibrate the circuit in **Fig 2** to minimize offset and gain errors. You can adjust the offset by first shorting pin 1 of IC_2 to pin 2, the Input Common. (Or you can insert the thermocouple in an ice-point refer-
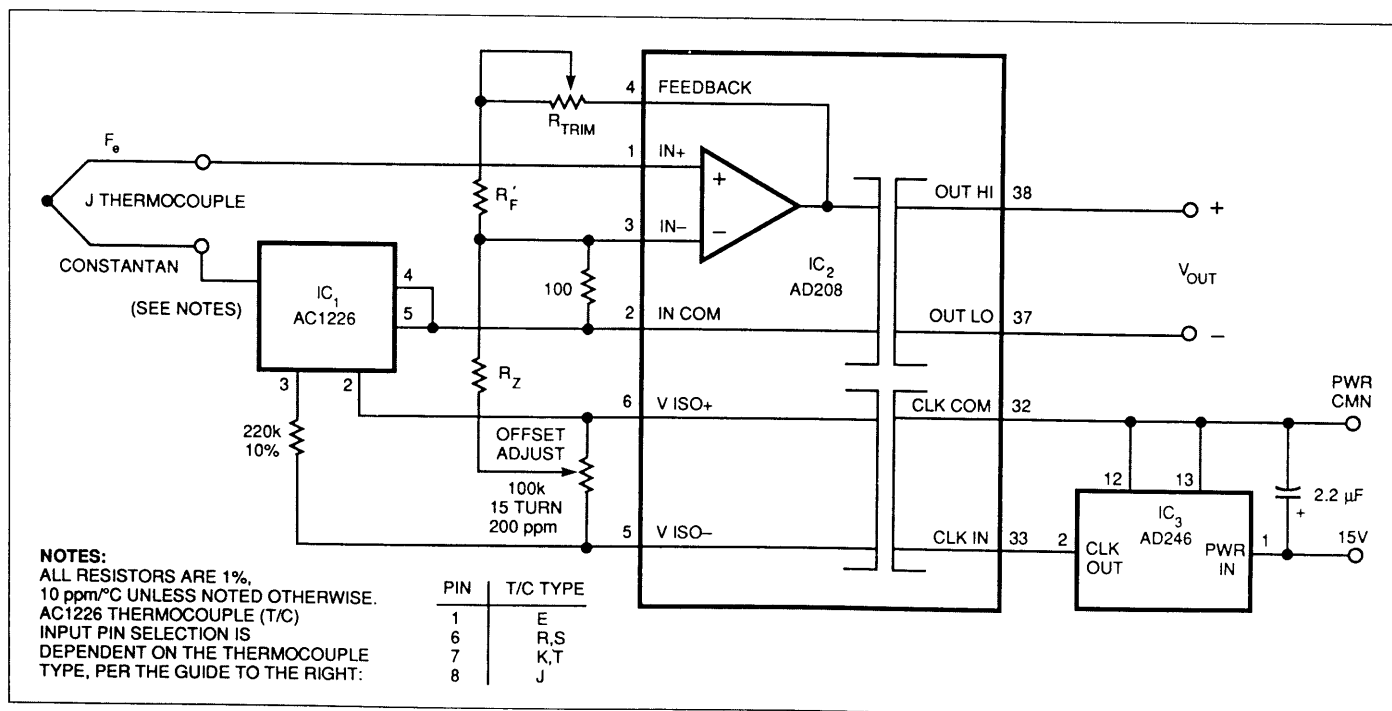
Fig 2—To compensate for cold-junction effects, this circuit inserts temperature-dependent compensation, generated by IC_1, between the thermocouple and an isolation amplifier.

A thermocouple relies on the Seebeck effect to generate a temperature-dependent micro-volt signal when two dissimilar metals join to form a closed-loop circuit.

ence—a 0°C environment.) Then, adjust the offset potentiometer until the amplifier's output measures 0V. You can calibrate the gain by applying the full-scale thermocouple voltage (42.283 mV for a J-type thermocouple) to the input of IC_2. Then adjust the gain-trim potentiometer until the dc signal at the output of the isolator is 5V. Because the offset and gain may interact with one another, you may have to recalibrate for zero offset following the gain adjustment.

After removing the calibration source, the cold-junction compensator should cause no more than ±2°C of residual error. The temperature stability of the circuit in **Fig 2** (referred to output) is 212.2 µV/°C, yielding a sensitivity to ambient temperature of 0.032 µC/°C. Gain drift is 321 µV/°C, delivering a sensitivity to ambient temperature of 0.048 µC/°C.

Measure temperature more precisely

RTDs offer better stability and linearity than thermocouples, but are limited to temperatures below 850°C. The most common RTDs are made of platinum, because platinum is highly stable at elevated tempera-

tures. Platinum RTDs typically have a 100Ω resistance and a temperature coefficient of +0.385%/°C at 0°C (European DIN Standard 43 760).

Because the RTD's resistance is the temperature-dependent parameter, the signal-conditioning circuit must supply a current to excite the sensor and then measure the voltage drop across the RTD. The circuit must apply relatively low excitation currents to avoid self-heating errors. As a result of the low excitation current and the low sensitivity of the RTDs, you'll need to employ a low-drift, high-gain amplifier to handle the RTD's millivolt output signal. Finally, to obtain the high-measurement accuracy that is possible with RTDs, you need to eliminate any initial offset voltage at the reference temperature, usually 0°C. You'll also need to reduce errors created by lead resistance in the wire that connects the RTD to the measurement electronics.

Fig 3's circuit is one way to achieve these goals. This measurement circuit combines excitation-current sources, lead-wire compensation and amplification for a 100Ω RTD. The LT1009, a 2.5V reference, amplifier

NOTE: ALL RESISTORS ARE 1%, 10 ppm/°C UNLESS NOTED OTHERWISE.
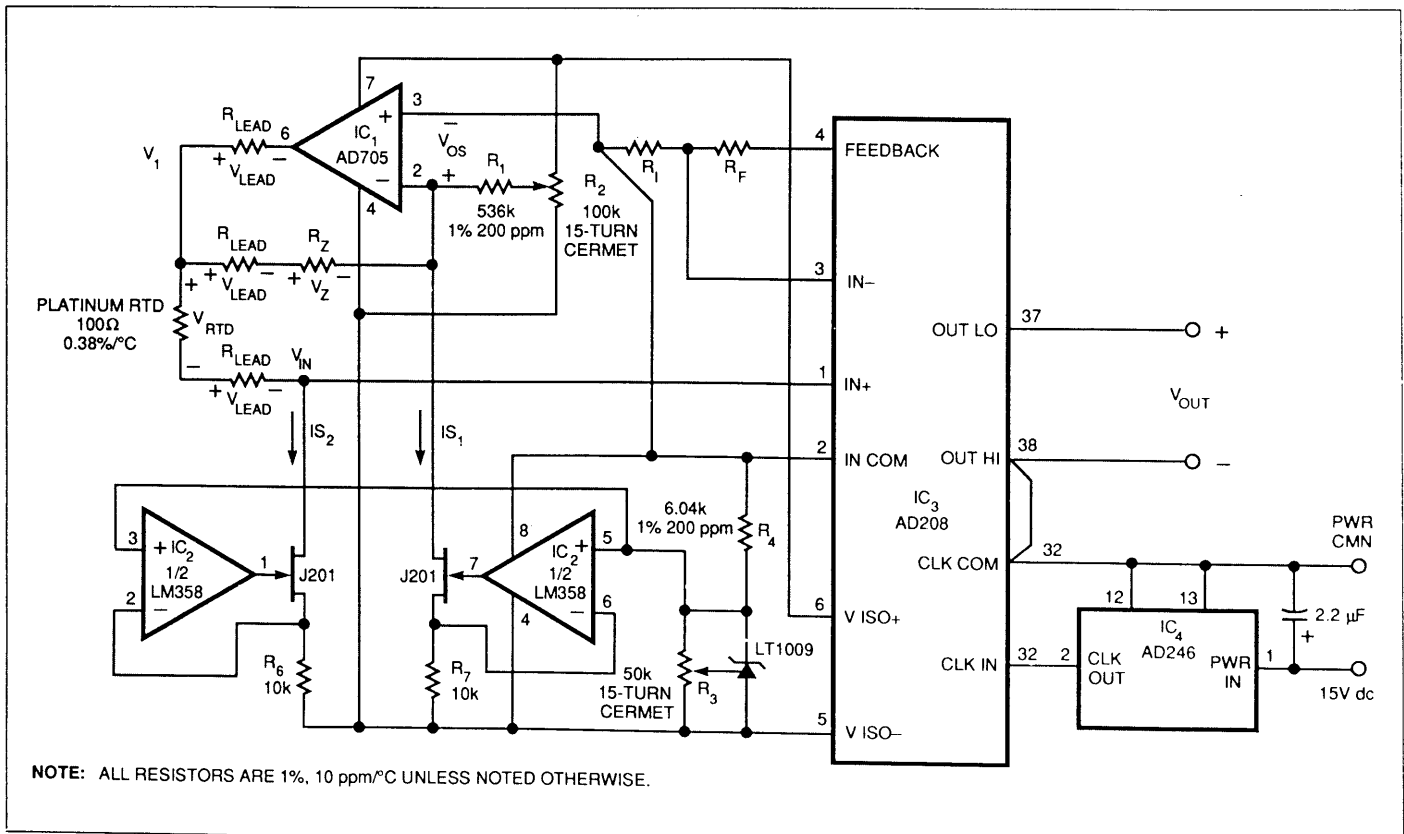
Fig 3—Dual currents and a ground-servo amplifier, IC_1, allow this circuit to excite and provide zero suppression to a 100Ω, platinum RTD over a 0 to 500°C range.

IC$_1$, dual LM358 amplifiers, IC$_2$, and a pair of low Vgs$_{off}$ n-channel JFETs such as the J201 compose two low-power current sources. One current source, IS$_2$, excites the RTD. The other source, IS$_1$, supplies current to offset the RTD such that the voltage at the isolation amplifier's input is zero at 0°C. This zero-suppression technique saves you from having to reduce the gain sensitivity of the RTD circuit to keep the signal within the operating range of the isolation amplifier.

IS$_2$ excites the RTD with a low current, 250 μA, to minimize self-heating. By adjusting the trimming potentiometer, R$_3$, you can change the excitation current's range by as much as ±5%. The circuit implements zero suppression using a simple ground-servo amplifier comprising IC$_1$ and R$_Z$. (R$_Z$ equals the RTD's 100Ω value at 0°C). The ground-servo amplifier provides an effective means of providing zero suppression without having to use an expensive differential amplifier. It also compensates for the finite lead resistance by remotely sensing the voltage at the RTD.

The current, IS$_1$, develops a voltage, V$_1$, equal to

$$V_1 = V_{OS} + V_Z + V_{Lead},$$

where V$_{OS}$ is IC$_1$'s offset voltage.

IS$_2$ in turn generates the voltage, V$_{IN}$, seen by IC$_3$'s input amplifier. The value of V$_{IN}$ equals

$$V_{IN} = V_1 - V_{RTD} - V_{Lead}.$$

Substituting the previous equation for V$_1$ in this equation for V$_{IN}$ yields

$$V_{IN} = V_{OS} + V_Z - V_{RTD}.$$

The trimming potentiometer, R$_2$, connected between ±V$_{ISO}$ and the large injection resistor, R$_1$, nulls any residual offset. To minimize errors created by offset drift in IC$_1$, choose a low-power, low-offset, and low-drift amplifier for this function.

The sensitivity of a 100Ω platinum RTD with a 250-μA current excitation is approximately 95 μV/°C. Therefore, setting the isolation amplifier's gain to 105 will result in an output sensitivity of 10 mV/°C for a temperature range of 0 to 500°C. With this amplification, the amplifier's output will span a 0 to −5V output range. By connecting IC$_3$'s OUT HI pin to ground (CLK COM) and using OUT LO as the output, the circuit inverts the output range to 0 to 5V.

The gain equation of the circuit in **Fig 3** is the same as the equation for **Fig 2** except for the additional excitation current term, IS$_2$, as follows:

$$Gain = (R_F/R_I + 1)$$

$$= (V_{HI} - V_{LO})/IS_2(V_{TC-HI} - V_{TC-LO}).$$

If you let R$_I$ equal 100Ω, R$_F$ should be 10.5 kΩ for a 0 to 500°C temperature range.

Fig 3's RTD circuit requires offset and gain calibration similar to that of the thermocouple-conditioning circuit. The easiest calibration method is to substitute a known resistance in place of the RTD. By replacing the RTD with a 100Ω resistor you can simulate a temperature input of 0°C. Then you can adjust the offset-trimming potentiometer, R$_2$, so that IC$_3$'s output is 0V. Then exchange the 100Ω resistor with one that simulates the RTD at 500°C (280.9Ω, DIN 43 760 standard), and trim the reference voltage until the IC$_3$'s output equals 5V.

A 100Ω platinum RTD that has a 250-μA excitation current and is amplified by a gain of 105 V/V will display a referred-to-output (RTO) sensitivity of 10.1 mV/°C. The circuit's RTO offset and gain drift is 433.2 μV/°C and 360 μV/°C, respectively. This offset and gain drift yields a total sensitivity to ambient temperature changes of 0.079°C/°C.

Condition strain-gauge signals

Another popular class of transducers used in fluid-flow monitoring (and in stress and strain measurements) are pressure transducers and strain gauges. Most of these transducer types consist of resistive elements whose resistance fluctuates in response to mechanical forces. These resistors, like RTDs, need an excitation signal and zero suppression. A common method of implementing these transducers is by placing them into one or more legs of a Wheatstone bridge. A differential amplifier then amplifies the small signal imposed on a large common-mode dc voltage.

Fig 4 shows a simple and accurate method of exciting and conditioning a strain gauge using the bridge approach—and without having to use a costly differential amplifier. This circuit connects a 100-mV strain gauge to the ±5V excitation signal derived from a precision voltage reference, IC$_1$. This reference contains a 10V buried-zener reference, half-bridge completion resistors (R$_1$ and R$_2$), and four amplifiers. Amplifier A$_1$

performs as a ground-servo amplifier for zero suppression, and output amplifiers A_2 and A_3 add remote-sense capability. **Fig 4** configures IC_1 for $\pm 5V$ operation with emitter-follower current boosters Q_1 and Q_2 to power the strain gauges. The current boosters can supply as much as 10 mA of load current into a 500Ω strain gauge. This current is limited by power limitations of the booster transistors, the power clock, and the saturation flux of the power transformer.

The ground-servo amplifier, A_1, and half-bridge completion resistors, R_1 and R_2, turn an otherwise differential measurement into a single-ended one. A_1 senses the reference voltage via pins 11 and 10 and forces its midpoint to zero. Thus, it forces the bridge to operate symmetrically around ground. If you need to make a remote measurement in cases where the lead resistance in the bridge can cause significant measurement errors, this circuit runs the sense leads of A_2 and A_3 to the remote strain gauge. A_2 and A_3 drive the bases of Q_1 and Q_2 such that the remote sense point will be $V_{REF} \pm 5V$.

Fig 4 amplifies a 100-mV, ground-referenced transducer signal by 50 to produce a $\pm 5V$ output range. As in the previous circuits, you determine the isolation amp's gain by using the standard noninverting gain formula: $R_F/(R_1 + 1)$.

If you're using a 500Ω strain gauge that requires more than 10 mA, you'll need to add more power to the circuit. You can derive useful power from the power clock by using a simple half-wave rectifier, filter, and transformer comprising T_1, D_1, D_2, and 4.7-μF capacitors. The power clock's $15V_{p-p}$, 25-kHz square-wave output provides the necessary current to supply 20 mA dc from the $\pm 8V$ output of the unregulated, isolated power supply. Because the AD588 voltage reference requires a maximum of 10 mA, 10 mA are available to the strain gauge.

Consider an alternative

Keep in mind that these voltage-output circuits require an external, unity-gain buffer amplifier, if you plan to drive a load that is less than 1 mΩ. As an option to all of the previous signal-conditioning circuits, you can convert the output voltage of the AD208 isolation
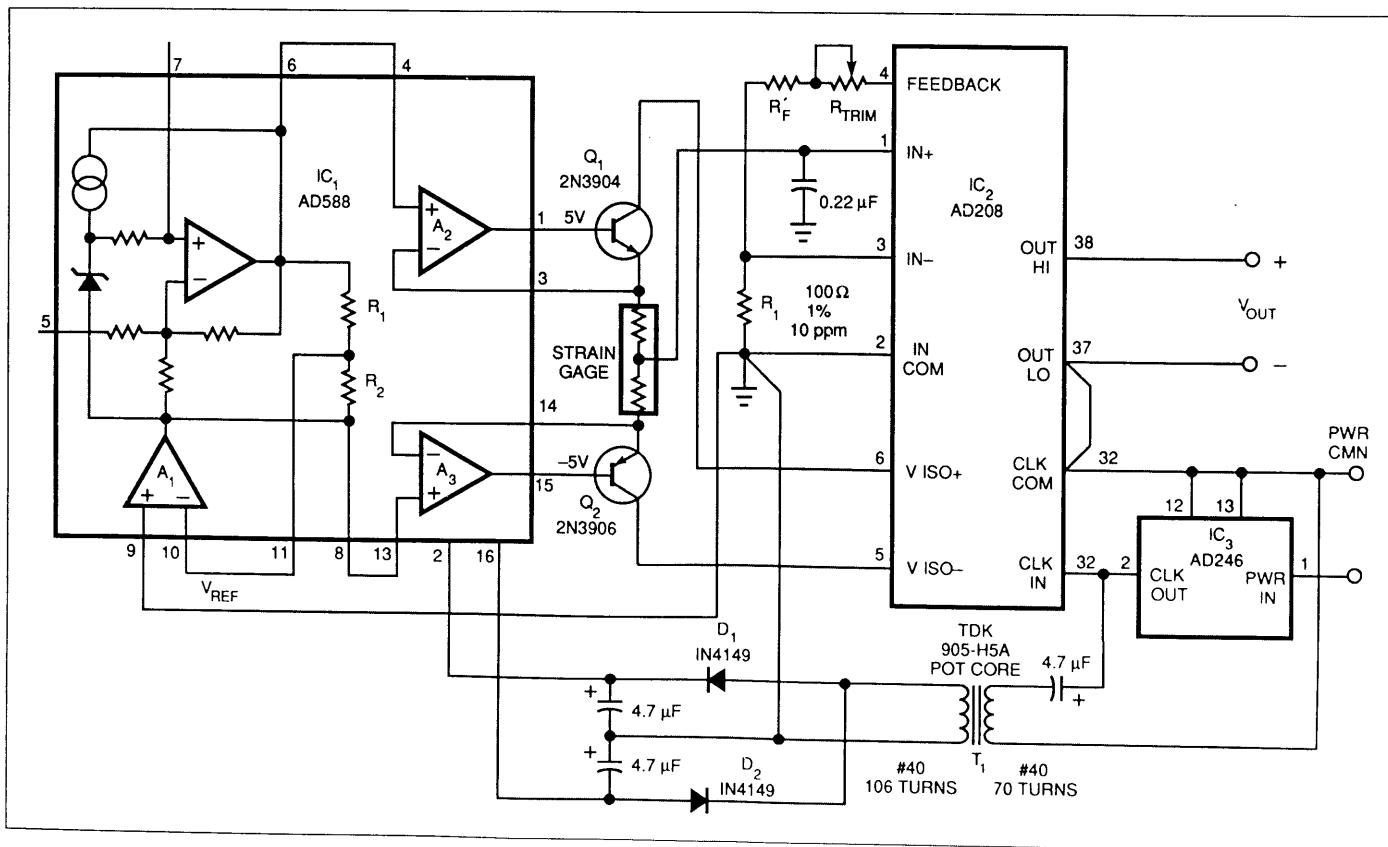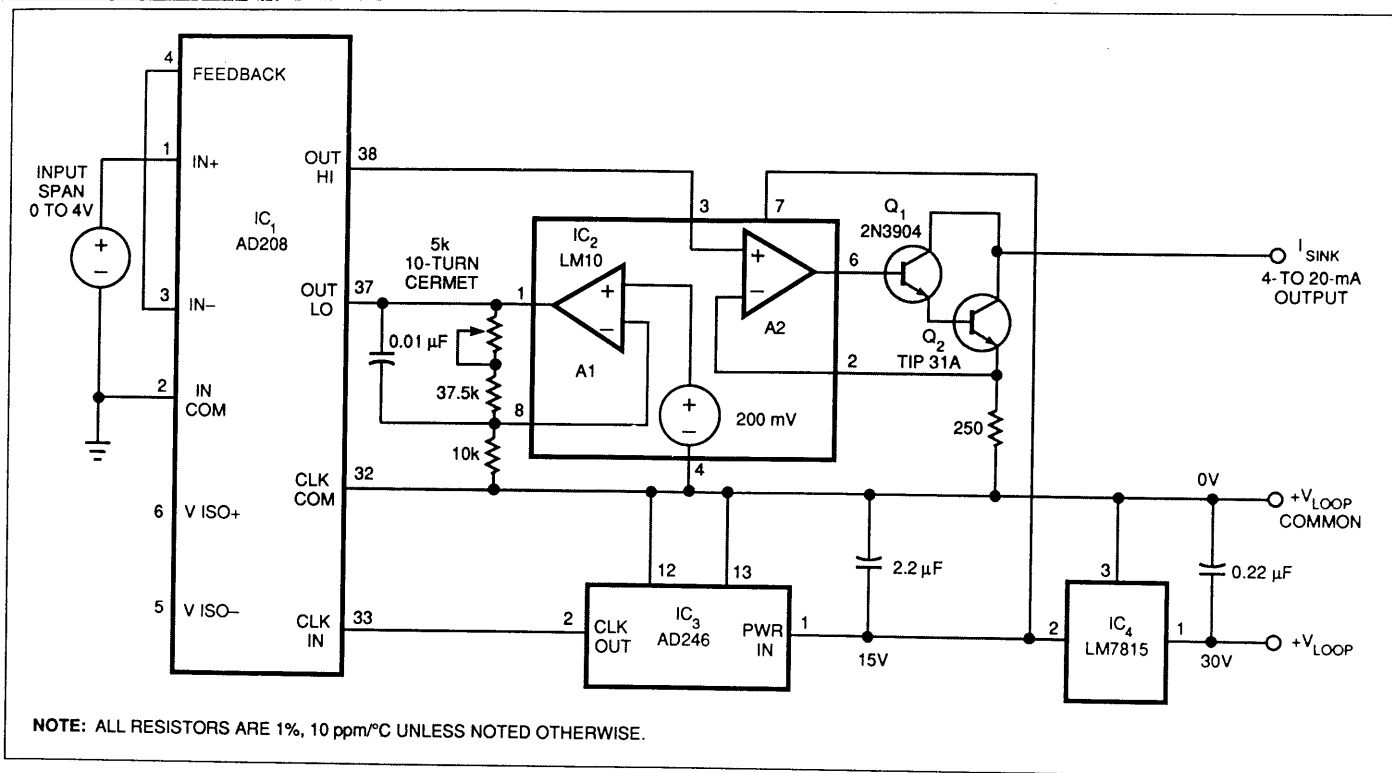

Fig 4—Amplifier A_1, part of a programmable voltage reference, senses the reference midpoint and forces the bridge to operate around that value. T_1 and its associated components add power if needed.

Resistance temperature detectors (RTDs) offer better stability and linearity than thermocouples, but have a narrower operating-temperature range.

Fig 5—This 4- to 20-mA current-loop circuit isolates the offset adjustment at the OUT LO pin, from the gain adjustment at the input.

amplifier to a 4- to 20-mA current (**Fig 5**). IC_1's common-mode output range, OUT LO – CLK COM, allows you to offset the OUT LO pin, thereby providing the offset necessary for the voltage-to-current conversion. IC_2's A_1 amplifies the 200-mV reference by 5, and drives OUT LO to 1V. You can then select resistor values to fix the gain of the AD208 such that the normal-mode output range is 0 to 4V. One requirement of this circuit is that the output common-mode voltage plus the normal-mode voltage must be less than or equal to 6V.

IC_1's OUT HI drives a voltage-to-current converter comprising A_2, Q_1, Q_2, and a 250Ω resistor. This converter generates the requisite 4- to 20-mA output with as much as 25V of compliance for a 30V supply. The circuit connects Q_1 and Q_2 in a Darlington configuration to keep base-current errors to a minimum. Q_2 must dissipate as much as 600 mW over your desired operating temperature range. Therefore, a medium-power transistor such as a TIP 31A works best. You can trim the 4-mA offset by adjusting the gain of A_1 and then independently adjusting the span at the input side of IC_1. This independent adjustment minimizes the interaction between offset and input span commonly encountered in other circuits.

IC_4, an LM7815 voltage regulator, preregulates a 30V input and supplies 15V to the power oscillator. If you are using more than a couple of isolation amplifiers in a multichannel application, you need to be aware that the AD208 isolation amplifier, in combination with the AD246 oscillator, represents a reactive load to your preregulator. Higher-than-average currents can flow into IC_3 and can force your preregulator to go into current limiting.