SECTION 3-4: BRIDGE CIRCUITS

An Introduction to Bridges

This section discusses the fundamental bridge circuit concept. To gain greatest appreciation of these ideas, it should be studied along with those sections discussing precision op amp in Chapters 1. These sections can be read sequentially if the reader already understands the design issues related to precision op amp applications.

Resistive elements are some of the most common sensors. They are inexpensive, and relatively easy to interface with signal-conditioning circuits. Resistive elements can be made sensitive to temperature, strain (by pressure or by flex), and light. Using these basic elements, many complex physical phenomena can be measured, such as: fluid or mass flow (by sensing the temperature difference between two calibrated resistances), dew-point humidity (by measuring two different temperature points), etc.

◆ **Strain Gages**	**120Ω, 350Ω, 3500Ω**
◆ **Weigh-Scale Load Cells**	**350Ω - 3500Ω**
◆ **Pressure Sensors**	**350Ω - 3500Ω**
◆ **Relative Humidity**	**100kΩ - 10MΩ**
◆ **Resistance Temperature Devices (RTDs)**	**100Ω , 1000Ω**
◆ **Thermistors**	**100Ω - 10MΩ**

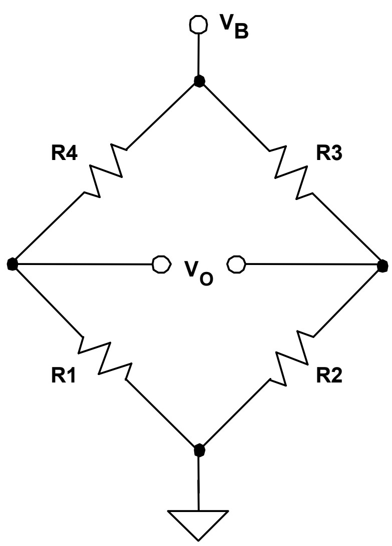
Figure 3.59: *Sensor resistances used in bridge circuits span a wide dynamic range*

Sensor element resistance can range from less than 100 Ω to several hundred kΩ, depending on the sensor design and the physical environment to be measured. Figure 3.59 indicates the wide range of sensor resistance encountered. For example, RTDs are typically 100 Ω or 1000 Ω. Thermistors are typically 3500 Ω or higher.

Resistive sensors such as RTDs and strain gages produce relatively small percentage changes in resistance, in response to a change in a physical variable such as temperature or force. For example, platinum RTDs have a temperature coefficient of about 0.385%/°C. Thus, in order to accurately resolve temperature to 1ºC, the overall measurement accuracy must be much better than 0.385 Ω when using a 100 Ω RTD.

Strain gages present a significant measurement challenge because the typical change in resistance over the entire operating range of a strain gage may be less than 1% of the nominal resistance value. Accurately measuring small resistance changes is therefore critical when applying resistive sensors.

A simple method for measuring resistance is to force a constant current through the resistive sensor, and measure the voltage output. This requires both an accurate current source and an accurate means of measuring the voltage. Any change in the current will be interpreted as a resistance change. In addition, the power dissipation in the resistive sensor must be small and in accordance with the manufacturer's recommendations, so that self-heating does not produce errors. As a result, the drive current must be small, which tends to limit the resolution of this approach.



THE WHEATSTONE BRIDGE:

$$V_O = V_B\left(\frac{R1}{R1+R4} - \frac{R2}{R2+R3}\right)$$

AT BALANCE,

$$V_O = 0 \text{ if } \frac{R1}{R4} = \frac{R2}{R3}$$

Figure 3.60: *The basic Wheatstone bridge produces an output null when the ratios of sidearm resistances match*

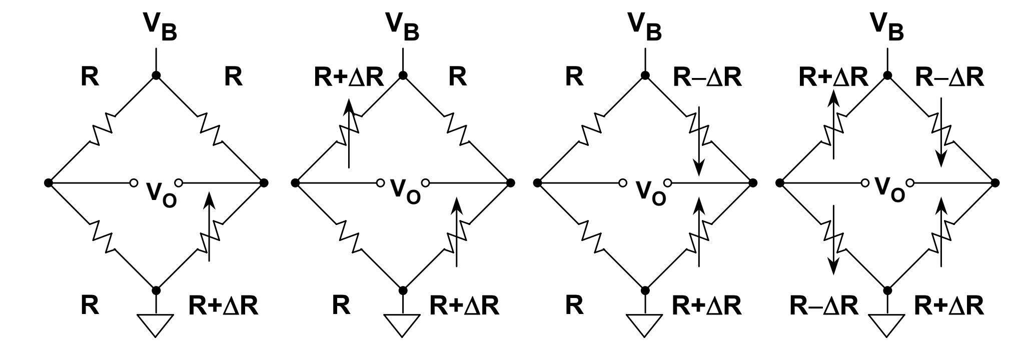
A *resistance bridge,* shown in Figure 3.60, offers an attractive alternative for measuring small resistance changes accurately. This is a basic Wheatstone bridge (actually developed by S. H. Christie in 1833), and is a prime example. It consists of four resistors connected to form a quadrilateral, a source of excitation voltage V_B (or, alternately, a current) connected across one of the diagonals, and a voltage detector connected across the other diagonal. The detector measures the difference between the outputs of the two voltage dividers connected across the excitation voltage, V_B. The general form of the bridge output V_O is noted in the figure.

There are two principal ways of operating a bridge such as this. One is by operating it as a null detector, where the bridge measures resistance indirectly by comparison with a similar standard resistance. On the other hand, it can be used as a device that reads a resistance difference directly, as a proportional voltage output.

When R1/R4 = R2/R3, the resistance bridge is said to be at a *null*, irrespective of the mode of excitation (current or voltage, AC or DC), the magnitude of excitation, the mode of readout (current or voltage), or the impedance of the detector. Therefore, if the ratio of

R2/R3 is fixed at K, a null is achieved when R1 = K·R4. If R1 is unknown and R4 is an accurately determined variable resistance, the magnitude of R1 can be found by adjusting R4 until an output null is achieved. Conversely, in sensor-type measurements, R4 may be a fixed reference, and a null occurs when the magnitude of the external variable (strain, temperature, etc.) is such that R1 = K·R4.

Null measurements are principally used in feedback systems involving electromechanical and/or human elements. Such systems seek to force the active element (strain gage, RTD, thermistor, etc.) to balance the bridge by influencing the parameter being measured.



	(A) Single-Element Varying	(B) Two-Element Varying (1)	(C) Two-Element Varying (2)	(D) All-Element Varying
V_O:	$\frac{V_B}{4}\left[\frac{\Delta R}{R+\frac{\Delta R}{2}}\right]$	$\frac{V_B}{2}\left[\frac{\Delta R}{R+\frac{\Delta R}{2}}\right]$	$\frac{V_B}{2}\left[\frac{\Delta R}{R}\right]$	$V_B\left[\frac{\Delta R}{R}\right]$
Linearity Error:	0.5%/%	0.5%/%	0	0

Figure 3.61: *The output voltage sensitivity and linearity of constant voltage drive bridge configurations differs according to the number of active elements*

For the majority of sensor applications employing bridges, however, the deviation of one or more resistors in a bridge from an initial value is measured as an indication of the magnitude (or a change) in the measured variable. In these cases, the output voltage change is an indication of the resistance change. Because very small resistance changes are common, the output voltage change may be as small as tens of millivolts, even with the excitation voltage V_B = 10 V (typical for a load cell application).

In many bridge applications, there may not just be a single variable element, but two, or even four elements, all of which may vary. Figure 3.61 above shows a family of four voltage-driven bridges, those most commonly suited for sensor applications. In the four cases the corresponding equations for V_O relate the bridge output voltage to the excitation voltage and the bridge resistance values. In all cases we assume a constant voltage drive, V_B. Note that since the bridge output is always directly proportional to V_B, the measurement accuracy can be no better than that of the accuracy of the excitation voltage.

In each case, the value of the fixed bridge resistor "R" is chosen to be equal to the nominal value of the variable resistor(s). The deviation of the variable resistor(s) about the nominal value is assumed to be proportional to the quantity being measured, such as strain (in the case of a strain gage), or temperature (in the case of an RTD).

The *sensitivity* of a bridge is the ratio of the maximum expected change in the output voltage to the excitation voltage. For instance, if V_B = 10 V, and the fullscale bridge output is 10 mV, then the sensitivity is 1 mV/V. For the four cases of Figure 3.61, sensitivity can be said to increase going left-right, or as more elements are made variable.

The *single-element varying* bridge of Figure 3.61A is most suited for temperature sensing using RTDs or thermistors. This configuration is also used with a single resistive strain gage. All the resistances are nominally equal, but one of them (the sensor) is variable by an amount ΔR. As the equation indicates, the relationship between the bridge output and ΔR is not linear. For example, if R = 100 Ω and ΔR = 0.1 Ω (0.1% change in resistance), the output of the bridge is 2.49875 mV for V_B = 10 V. The error is 2.50000 mV – 2.49875 mV, or 0.00125 mV. Converting this to a % of fullscale by dividing by 2.5 mV yields an end-point linearity error in percent of approximately 0.05%. (Bridge end-point linearity error is calculated as the worst error in % FS from a straight line which connects the origin and the end point at FS, i.e., the FS gain error is not included). If ΔR = 1 Ω, (1% change in resistance), the output of the bridge is 24.8756 mV, representing an end-point linearity error of approximately 0.5%. The end-point linearity error of the single-element bridge can be expressed in equation form:

Single-Element Varying Bridge End-Point Linearity Error ≈ % Change in Resistance ÷ 2

It should be noted that the above nonlinearity refers to the nonlinearity of the bridge itself and not *the sensor*. In practice, most sensors themselves will exhibit a certain specified amount of nonlinearity, which must also be accounted for in the final measurement.

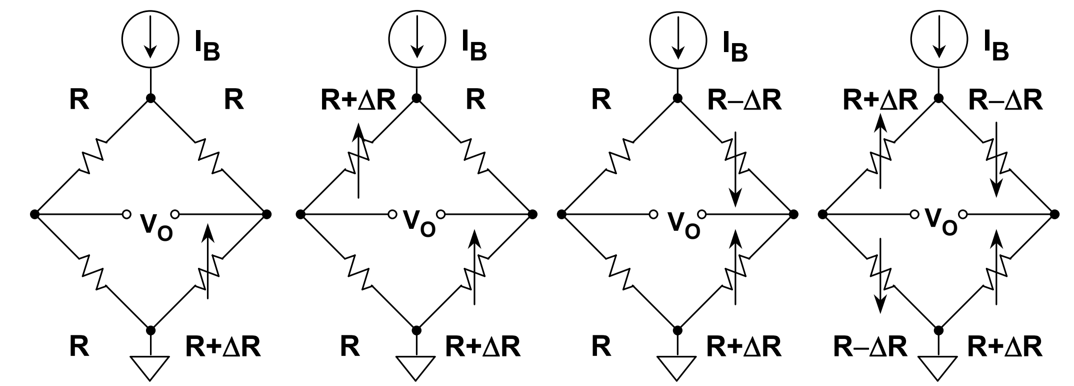
In some applications, the bridge nonlinearity noted above may be acceptable. But, if not, there are various methods available to linearize bridges. Since there is a fixed relationship between the bridge resistance change and its output (shown in the equations), software can be used to remove the linearity error in digital systems. Circuit techniques can also be used to linearize the bridge output directly, and these will be discussed shortly.

There are two cases to consider in the instance of a *two-element varying* bridge. In Case 1 (Figure 3.61B), both of the diagonally opposite elements change in the same direction. An example would be two identical strain gages mounted adjacent to each other, with their axes in parallel.

The nonlinearity for this case, 0.5%/%, the same as that of the single-element varying bridge of Figure 3.61A. However, it is interesting to note the sensitivity is now improved by a factor of 2, vis-à-vis the single-element varying setup. The two-element varying bridge is commonly found in pressure sensors and flow meter systems.

A second case of the two-element varying bridge, Case 2, is shown in Figure 3.61C. This bridge requires two identical elements that vary in *opposite* directions. This could correspond to two identical strain gages: one mounted on top of a flexing surface, and one on the bottom. Note that this configuration is now linear, and like two-element varying Case 1, it has twice the sensitivity of the Figure 3.61A configuration. Another way to view this configuration is to consider the terms R+ΔR and R–ΔR as comprising two sections of a linear potentiometer.

The *all-element varying* bridge of Figure 3.61D produces the most signal for a given resistance change, and is inherently linear. It is also an industry-standard configuration for load cells constructed from four identical strain gages. Understandably, it is also one of the most popular bridge configurations.



	(A) Single-Element Varying	(B) Two-Element Varying (1)	(C) Two-Element Varying (2)	(D) All-Element Varying
V_O:	$\frac{I_B R}{4}\left[\frac{\Delta R}{R + \frac{\Delta R}{4}}\right]$	$\frac{I_B}{2}\left[\Delta R\right]$	$\frac{I_B}{2}\left[\Delta R\right]$	$I_B\left[\Delta R\right]$
Linearity Error:	0.25%/%	0	0	0

Figure 3.62: *The output voltage sensitivity and linearity of constant current drive bridge configurations also differs according to the number of active elements*

Bridges may also be driven from constant current sources, as shown in Figure 3.62, for the corresponding cases of single, dual, dual, and four active element(s). As with the voltage-driven bridges, the analogous output expressions are noted, along with the sensitivities.

Current drive, although not as popular as voltage drive, does have advantages when the bridge is located remotely from the source of excitation. One advantage is that the wiring resistance doesn't introduce errors in the measurement; another is simpler, less expensive cabling. Note also that with constant current excitation, all bridge configurations are linear except the single-element varying case of Figure 3.62A.

In summary, there are many design issues relating to bridge circuits, as denoted by Figure 3.63 below. After selecting the basic configuration, the excitation method must be determined. The value of the excitation voltage or current must first be determined, as this directly influences sensitivity. Recall that the fullscale bridge output is directly proportional to the excitation voltage (or current). Typical bridge sensitivities are 1 mV/V to 10 mV/V.

Although large excitation voltages yield proportionally larger fullscale output voltages, they also result in higher bridge power dissipation, and thus raise the possibility of sensor resistor self-heating errors. On the other hand, low values of excitation voltage require more gain in the conditioning circuits, and also increase sensitivity to low level errors such as noise and offset voltages.

- **Selecting Configuration (1, 2, 4 - Element Varying)**
- **Selection of Voltage or Current Excitation**
- **Ratiometric Operation**
- **Stability of Excitation Voltage or Current**
- **Bridge Sensitivity: FS Output / Excitation Voltage 1mV / V to 10mV / V Typical**
- **Fullscale Bridge Outputs: 10mV - 100mV Typical**
- **Precision, Low Noise Amplification / Conditioning Techniques Required**
- **Linearization Techniques May Be Required**
- **Remote Sensors Present Challenges**

Figure 3.63: *A number of bridge considerations impact design choices*

Regardless of the absolute level, the stability of the excitation voltage or current directly affects the overall accuracy of the bridge output, as is evident from the V_B and I_B terms in the output expressions. Therefore stable references and/or *ratiometric* drive techniques are required, to maintain highest accuracy.

Here, ratiometric simply refers to the use of the bridge drive voltage of a voltage-driven bridge (or a current-proportional voltage, for a current-driven bridge) as the reference input to the ADC that digitizes the amplified bridge output voltage. In this manner the absolute accuracy and stability of the excitation voltage becomes a second order error. Examples to follow illustrate this point further.

Amplifying and Linearizing Bridge Outputs

The output of a single-element varying bridge may be amplified by a single precision op-amp connected as shown in Figure 3.64. Unfortunately this circuit, although attractive because of relative simplicity, has poor overall performance. Its gain predictability and accuracy are poor, and it unbalances the bridge due to loading from R_F and the op amp bias current. The R_F resistors must be carefully chosen and matched to maximize common mode rejection (CMR). Also, it is difficult to maximize the CMR while at the same time allowing different gain options. Gain is dependent upon the bridge resistances and R_F. In addition, the output is nonlinear, as the configuration does nothing to address the intrinsic bridge non-linearity. In summary, the circuit isn't recommended for precision use.

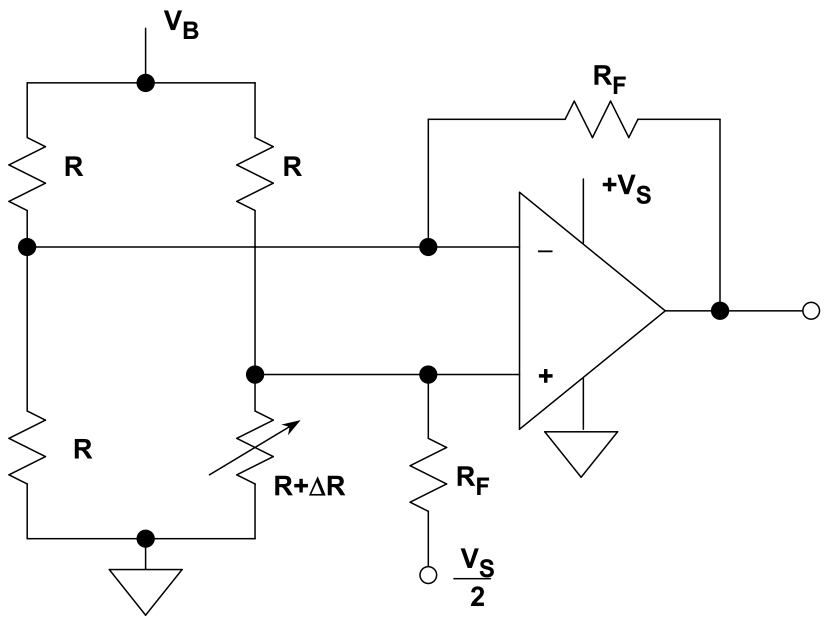


Figure 3.64: *Using a single op amp as a bridge amplifier*

However, a redeeming feature of this circuit is that it is capable of single supply operation, with a solitary op amp. Note that the R_F resistor connected to the non-inverting input is returned to $V_S/2$ (rather than ground) so that both positive and negative ΔR values can be accommodated, with the bipolar op amp output swing referenced to $V_S/2$.

A much better approach is to use an *instrumentation amplifier* (in-amp) for the required gain, as shown in Figure 3.65. This efficient circuit provides better gain accuracy, with the in-amp gain usually set with a single resistor, R_G. Since the amplifier provides dual, high-impedance loading to the bridge nodes, it does not unbalance or load the bridge. Using modern in-amp devices with gains ranging from 10-1000, excellent common mode rejection and gain accuracy can be achieved with this circuit.

However, due to the intrinsic characteristics of the bridge, the output is still nonlinear (see expression). As noted earlier, this can be corrected in software (assuming that the in-amp output is digitized using an analog-to-digital converter and followed by a microcontroller or microprocessor).

The in-amp can be operated on either dual supplies as shown, or alternately, on a single positive supply. In the figure, this corresponds to $-V_S = 0$. This is a key advantageous point, due the fact that all such bridge circuits bias the in-amp inputs at $V_B/2$, a voltage range typically compatible with amplifier bias requirements. In-amps such as the AD620 family, the AD623, and AD627 can be used in single (or dual) supply bridge applications, provided their restrictions on the gain and input and output voltage swings are observed.

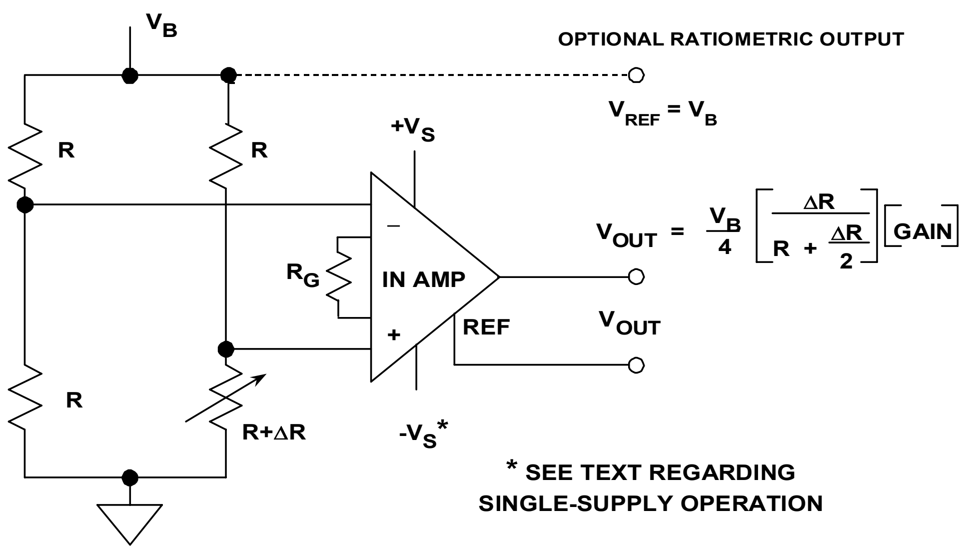


***Figure 3.65**: A generally preferred method of bridge amplification employs an instrumentation amplifier for stable gain and high CMR*

The bridge in this example is voltage driven, by the voltage V_B. This voltage can optionally be used for an ADC reference voltage, in which case it also is an additional output, V_{REF}.

Various techniques are available to linearize bridge outputs, but it is important to distinguish between the linearity of the bridge equation (discussed earlier), and the sensor response linearity to the phenomenon being sensed. For example, if the active sensor element is an RTD, the bridge used to implement the measurement might have perfectly adequate linearity; *yet the output could still be nonlinear,* due to the RTD device's intrinsic nonlinearity. Manufacturers of sensors employing bridges address the nonlinearity issue in a variety of ways, including keeping the resistive swings in the bridge small, shaping complementary nonlinear response into the active elements of the bridge, using resistive trims for first-order corrections, and others. In the examples which follow, what is being addressed is the linearity error of the bridge configuration itself (as opposed to a sensor element within the bridge).

Figure 3.66 shows a single-element varying active bridge circuit, in which an op amp produces a forced bridge null condition. For this single-element varying case, only the op amp feedback resistance varies, with the remaining three resistances fixed.

As used here, the op amp output provides a buffered, ground referenced, low impedance output for the bridge measurement, effectively suppressing the $V_B/2$ CM bridge component at the op amp inputs.

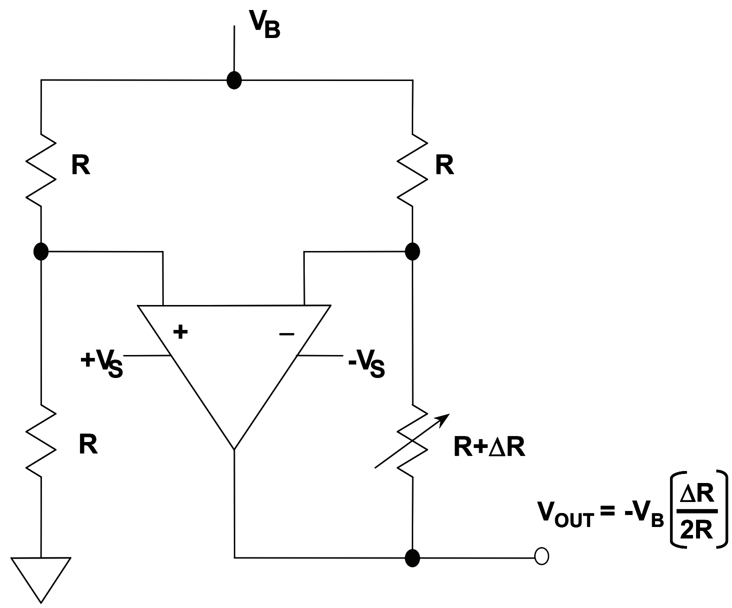


Figure 3.66: *Linearizing a single-element varying bridge (Method 1)*

The circuit works by adding a voltage in series with the variable resistance arm. This voltage is equal in magnitude and opposite in polarity to the incremental voltage across the varying element, and is linear with ΔR. As can be noted, the three constant "R" valued resistances and the op amp operate to drive a constant current in the variable resistance. This is the basic mechanism that produces the linearized output.

This active bridge has a sensitivity gain of two over the standard single-element varying bridge (Figure 3.62A, again). The key point is that the bridge's incremental resistance/ voltage output becomes linear, even for large values of ΔR. However, because of a still relatively small output signal, a second amplifier must usually follow this bridge. Note also that the op amp used in this circuit requires dual supplies, because its output must go negative for conditions where ΔR is positive.

Another circuit for linearizing a single-element varying bridge is shown in Figure 3.67. The top node of the bridge is excited by the voltage, V_B. The bottom of the bridge is driven in complementary fashion by the left op amp, which maintains a constant current of V_B/R in the varying resistance element, R + ΔR. Like the circuit of Figure 3.66, the constant current drive for the single-element variable resistance provides the mechanism for linearity improvement. Also, because of the fact that the bridge left-side center node is ground-referenced by the op amp, this configuration effectively suppresses CM voltages. This has the virtue of making the op amp selection somewhat less critical. Of course, performance parameters of high gain, low offset/noise, and high stability are all still needed.

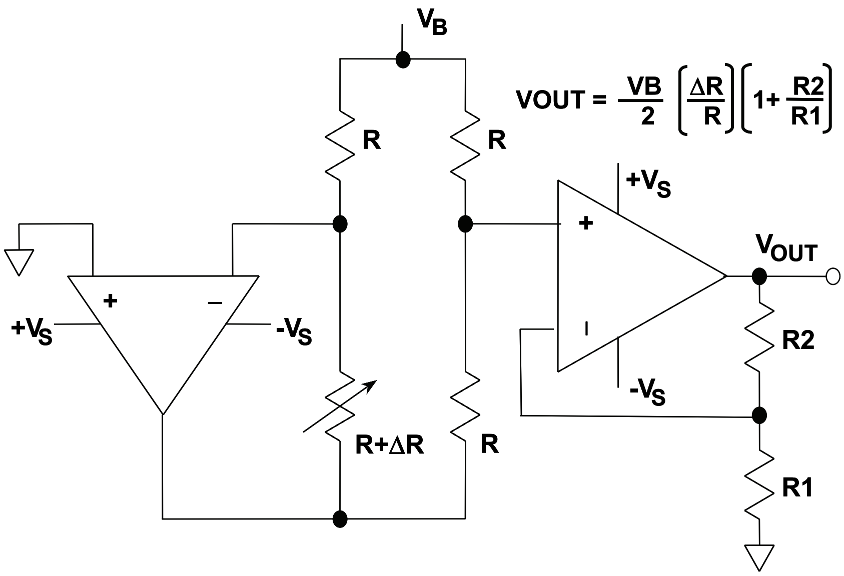


Figure 3.67: *Linearizing a single-element varying bridge (Method 2)*

The output signal is taken from the right-hand leg of the bridge, and is amplified by a second op amp, connected as a non-inverting gain stage. With the scaling freedom provided by the second op amp, the configuration is very flexible. The net output is linear, and has a bridge-output referred sensitivity comparable to the single-element varying circuit of Figure 3.66.

The circuit in Figure 3.67 requires two op amps operating on dual supplies. In addition, paired resistors R1-R2 must be ratio matched and stable types, for overall accurate and stable gain. The circuit can be a practical one using a dual precision op amp, such as an AD708, the OP2177 or the OP213.

A closely related circuit for linearizing a voltage-driven, *two-element* varying bridge can be adapted directly from the basic circuit of Figure 3.67. This form of the circuit, shown in Figure 3.68, is identical to the previous single-element varying case, with the exception that the resistance between V_B and the op amp (+) input is now also variable (i.e., both diagonal $R + \Delta R$ resistances vary, in a like manner).

For the same applied voltage V_B, this form of the circuit has twice the sensitivity, which is evident in the output expressions. A dual supply op amp is again required, and additional gain may also be necessary.

The two-element varying bridge circuit shown in Figure 3.69 uses an op amp, a sense resistor, and a voltage reference, set up in a feedback loop containing the sensing bridge. The net effect of the loop is to maintain a constant current through the bridge of $I_B = V_{REF}/R_{SENSE}$. The current through each leg of the bridge remains constant ($I_B/2$) as the resistances change, therefore the output is a linear function of ΔR. An in-amp provides the additional gain.

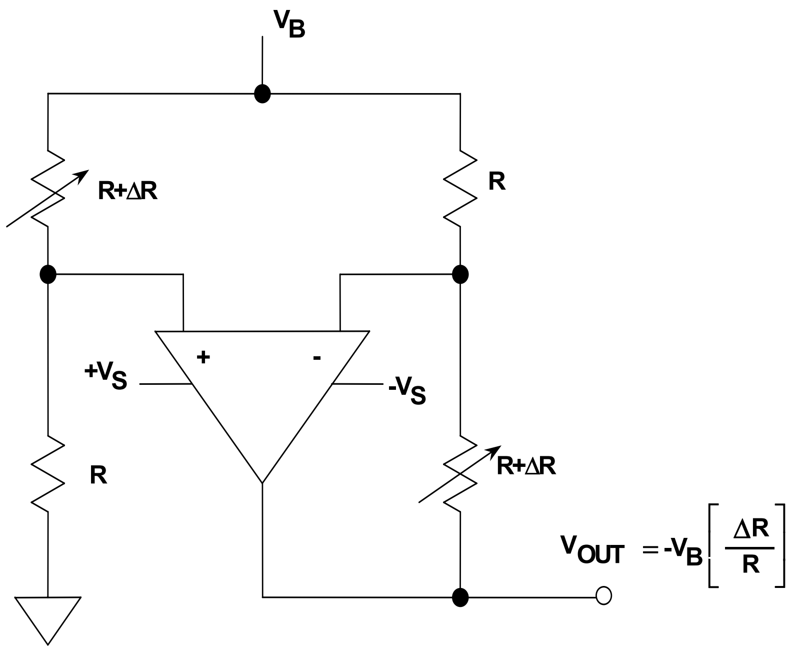


Figure 3.68: *Linearizing a two-element varying voltage-driven bridge (Method 1)*

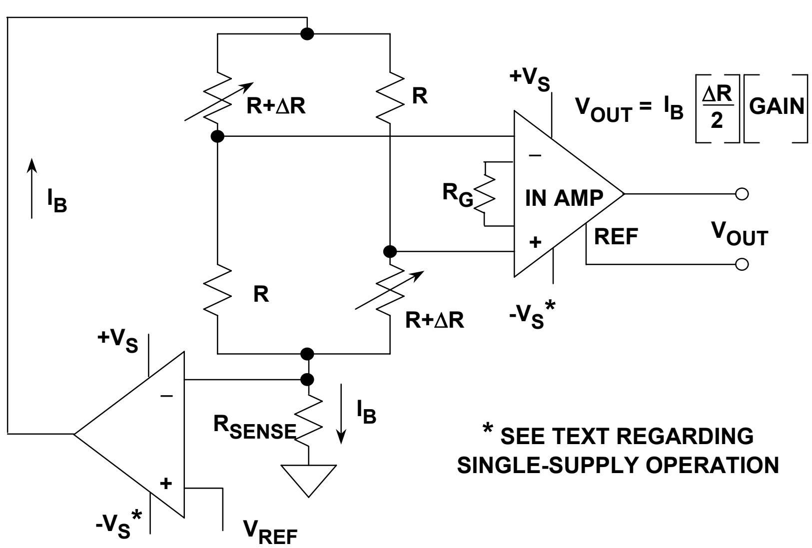


Figure 3.70: *Linearizing a two-element varying current-driven bridge (Method 2)*

This circuit can be operated on a single supply with the proper choice of amplifiers and signal levels. If ratiometric operation of an ADC is desired, the V_{REF} voltage can be used to drive the ADC.

Driving Remote Bridges

Wiring resistance and noise pickup are the biggest problems associated with remotely located bridges. Figure 3.71 shows a 350 Ω strain gage, which is connected to the rest of the bridge circuit by 100 feet of 30 gage twisted pair copper wire. The resistance of the wire at 25ºC is 0.105 Ω/ft, or 10.5 Ω for 100 ft. The total lead resistance in series with the 350 Ω strain gage is therefore 21 Ω. The temperature coefficient of the copper wire is 0.385%/ºC. Now we will calculate the gain and offset error in the bridge output due to a +10ºC temperature rise in the cable. These calculations are easy to make, because the bridge output voltage is simply the difference between the output of two voltage dividers, each driven from a +10 V source.

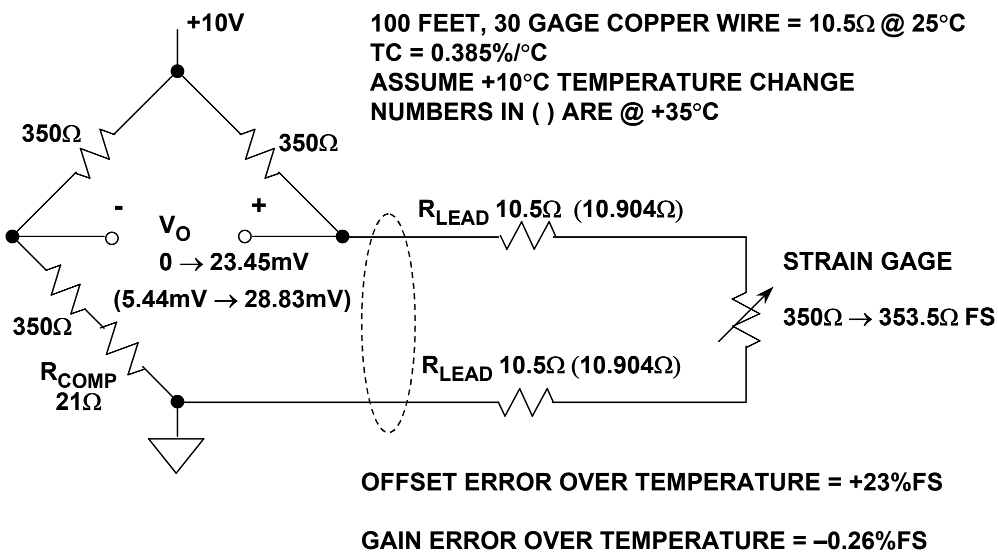


Figure 3.71: *Wiring resistance related errors with remote bridge sensor*

The fullscale variation of the strain gage resistance (with flex) above its nominal 350 Ω value is +1% (+3.5 Ω), corresponding to a fullscale strain gage resistance of 353.5 Ω which causes a bridge output voltage of +23.45 mV. Notice that the addition of the 21 Ω R_{COMP} resistor compensates for the wiring resistance and balances the bridge when the strain gage resistance is 350 Ω. Without R_{COMP}, the bridge would have an output offset voltage of 145.63 mV for a nominal strain gage resistance of 350 Ω. This offset could be compensated for in software just as easily, but for this example, we chose to do it with R_{COMP}.

Assume that the cable temperature increases +10ºC above nominal room temperature. This results in a total lead resistance increase of +0.404 Ω (10.5 Ω×0.00385/ºC×10ºC) in each lead. *Note: The values in parentheses in the diagram indicate the values at +35ºC.* The total additional lead resistance (of the two leads) is +0.808 Ω. With no strain, this additional lead resistance produces an offset of +5.44 mV in the bridge output. Fullscale strain produces a bridge output of +28.83 mV (a change of +23.39 mV from no strain).

Thus the increase in temperature produces an offset voltage error of +5.44 mV (+23% fullscale) and a gain error of –0.06 mV (23.39 mV – 23.45 mV), or –0.26% fullscale. Note that these errors are produced solely by the 30 gage wire, and do not include any temperature coefficient errors in the strain gage itself.

The effects of wiring resistance on the bridge output can be minimized by the 3-wire connection shown in Figure 3.72. We assume that the bridge output voltage is measured by a high impedance device, therefore there is no current in the sense lead. Note that the sense lead measures the voltage output of a divider: the top half is the bridge resistor plus the lead resistance, and the bottom half is strain gage resistance plus the lead resistance. The nominal sense voltage is therefore independent of the lead resistance. When the strain gage resistance increases to fullscale (353.5 Ω), the bridge output increases to +24.15 mV.

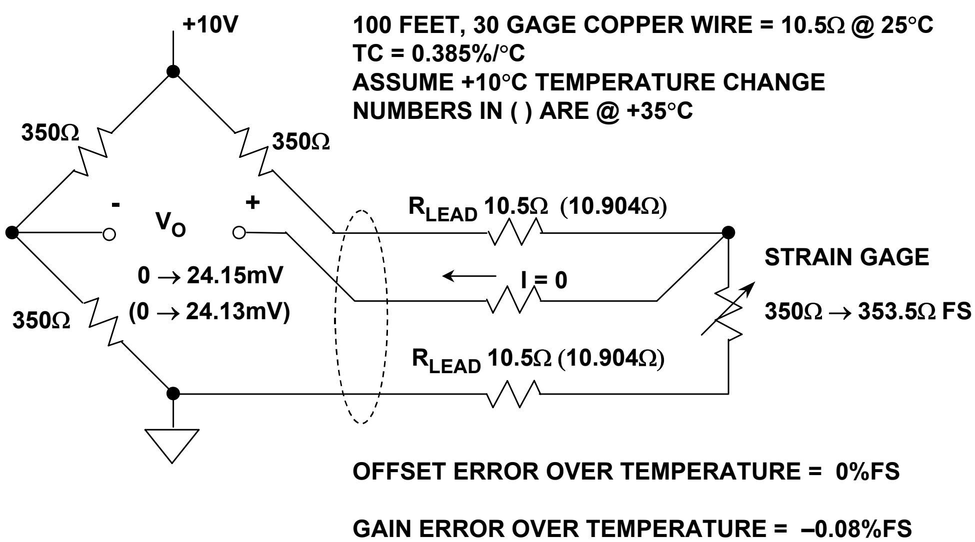


Figure 3.72: *Remote bridge wiring resistance errors are reduced with 3-wire sensor connection*

Increasing the temperature to +35ºC increases the lead resistance by +0.404 Ω in each half of the divider. The fullscale bridge output voltage decreases to +24.13 mV because of the small loss in sensitivity, but there is no offset error. The gain error due to the temperature increase of +10ºC is therefore only –0.02 mV, or –0.08% of fullscale. Compare this to the +23% fullscale offset error and the –0.26% gain error for the two-wire connection shown in Figure 3.72.

The three-wire method works well for remotely located resistive elements which make up one leg of a single-element varying bridge. However, all-element varying bridges generally are housed in a complete assembly, as in the case of a load cell. When these bridges are remotely located from the conditioning electronics, special techniques must be used to maintain accuracy.

Of particular concern is maintaining the accuracy and stability of the bridge excitation voltage. The bridge output is directly proportional to the excitation voltage, and any drift in the excitation voltage produces a corresponding drift in the output voltage.

For this reason, most all-element varying bridges (such as load cells) are six-lead ssemblies: two leads for the bridge output, two leads for the bridge excitation, and two *sense* leads. To take full advantage of the additional accuracy that these extra leads allow, a method called Kelvin or 4-wire sensing is employed, as shown in Figure 3.73 below.

In this setup the drive voltage V_B is not applied directly to the bridge, but goes instead to the input of the upper precision op amp, which is connected in a feedback loop around the bridge (+) terminal. Although there may be a substantial voltage drop in the +FORCE lead resistance of the remote cable, the op amp will automatically correct for it, since it has a feedback path through the +SENSE lead. The net effect is that the upper node of the remote bridge is maintained at a precise level of V_B (within the capability of the op amp used, of course). A similar situation occurs with the bottom precision op amp, which drives the bridge (-) terminal to a ground level, as established by the op amp input ground reference. Again, the voltage drop in the –FORCE lead is relatively immaterial, because of the sensing at the –SENSE terminal.

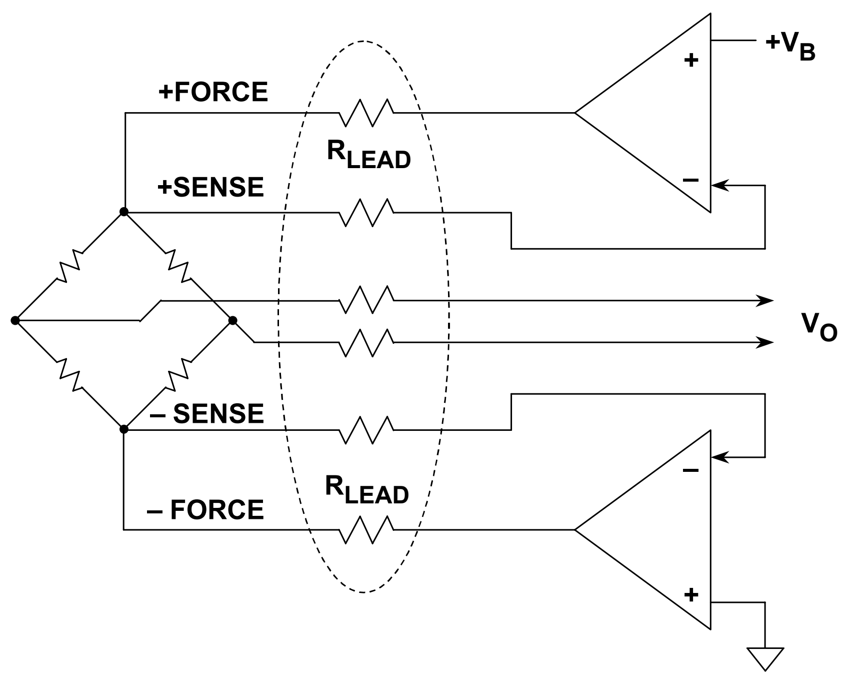


Figure 3.73: *A Kelvin sensing system with a 6-wire voltage-driven bridge connection and precision op amps minimizes errors due to wire lead resistances*

In both cases, the sense lines go to high impedance op amp inputs, thus there is minimal error due to the bias current induced voltage drop across their lead resistance. The op amps maintain the required excitation voltage at the remote bridge, to make the voltage measured between the (+) and (-) sense leads always equal to V_B.

Note— a subtle point is that the lower op amp will need to operate on dual supplies, since the drive to the -FORCE lead will cause the op amp output to go negative. Because of relatively high current in the bridge (~30 mA), current buffering stages at the op amp outputs are likely advisable for this circuit.

Although Kelvin sensing eliminates errors due to voltage drops in the bridge wiring resistance, the basic drive voltage V_B must still be highly stable since it directly affects the bridge output voltage. In addition, the op amps must have low offset, low drift, and low noise. Ratiometric operation can be optionally added, simply by using V_B to drive the ADC reference input.

The constant current excitation method shown in Figure 3.74 below is another method for minimizing the effects of wiring resistance on the measurement accuracy. This system drives a precise current I through the bridge, proportioned as per the expression in the figure. An advantage of the circuit in Figure 3.74 is that it only uses one amplifier.

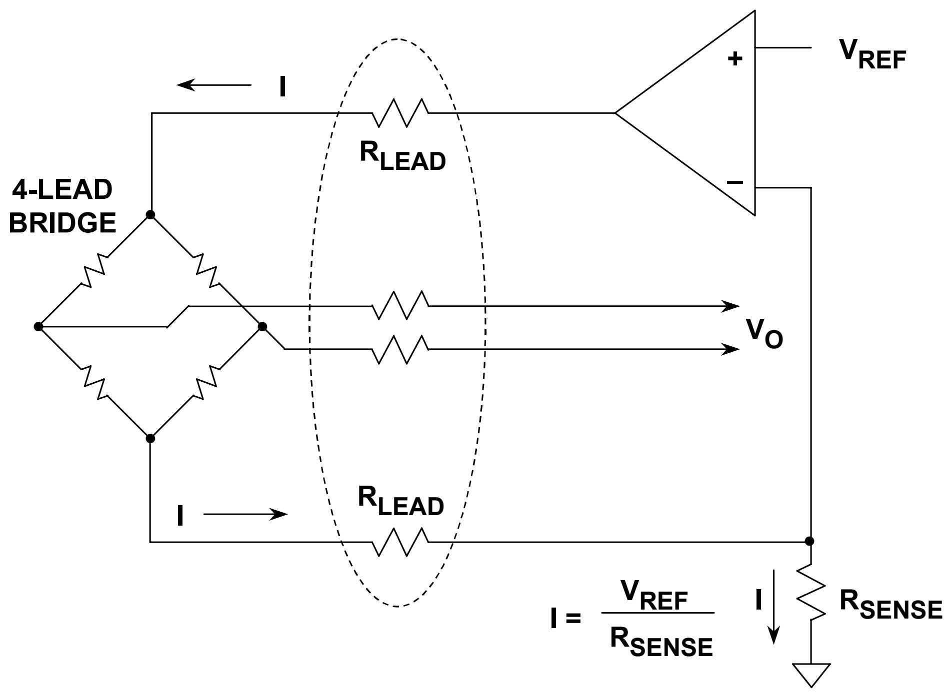


Figure 3.74: *A 4-wire current-driven bridge scheme also minimizes errors due to wire lead resistances, plus allows simpler cabling*

However, the accuracy of the reference, the sense resistor, and the op amp all influence the overall accuracy. While the precision required of the op amp should be obvious, one thing not necessarily obvious is that it may be required to deliver appreciable current, when I is more than a few mA (which it will be with standard 350 Ω bridges). In such cases, current buffering of the op amp is again in order.

Therefore for highest precision with this circuit, a buffer stage is recommended. This can be as simple as a small transistor, since the bridge drive is unidirectional.

System offset minimization

Maintaining an accuracy of 0.1% or better with a fullscale bridge output voltage of 20 mV requires that the sum of all offset errors be less than 20 µV. Parasitic thermocouples are cases in point, and if not given due attention, they can cause serious temperature drift errors. All dissimilar metal-metal connections generate voltages between a few and tens of microvolts for a 1ºC temperature differential, are basic thermocouple fact-of-life.

Fortunately however, within a bridge measurement system the signal connections are differential, therefore this factor can be used to minimize the impact of parasitic thermocouples.

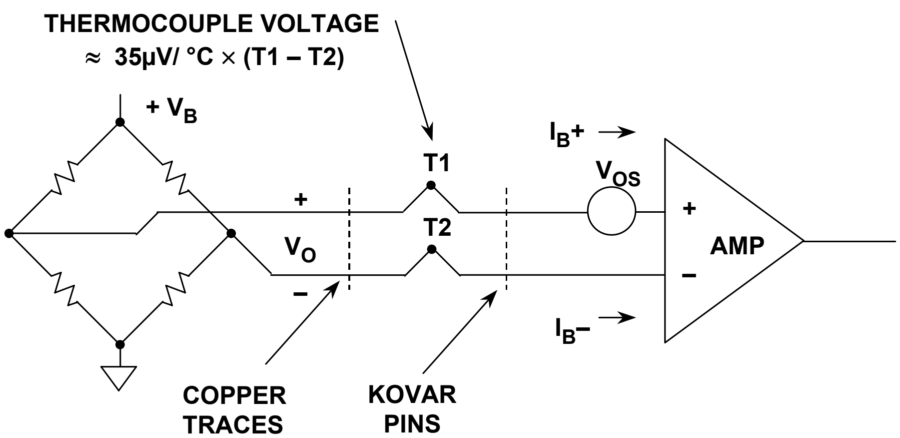


Figure 3.75: *Typical sources of offset voltage within bridge measurement systems*

Figure 3.75 shows some typical sources of offset error that are inevitable in a system. Within a differential signal path, only those thermocouple pairs whose junctions are actually at different temperatures will degrade the signal. The diagram shows a typical parasitic junction formed between the copper printed circuit board traces and the kovar pins of an IC amplifier.

This thermocouple voltage is about 35 µV/ºC temperature differential. Note that this package-PC trace thermocouple voltage is significantly less when using a plastic package with a copper lead frame (recommended). Regardless of what package is used, all metal-metal connections along the signal path should be designed so that minimal temperature differences occur between the sides.

The amplifier offset voltage and bias currents are further sources of offset error. The amplifier bias current must flow through the source impedance. Any unbalance in either the source resistances or the bias currents produce offset errors. In addition, the offset voltage and bias currents are a function of temperature.

High performance low offset, low offset drift, low bias current, and low noise precision amplifiers such as the AD707, the OP177 or OP1177 are required. In some cases, chopper-stabilized amplifiers such as the AD8551/AD8552/AD8554 may be a solution.

AC bridge excitation such as that shown in Figure 3.76 below can effectively remove offset voltage effects in series with a bridge output, V_O.

The concept is simple, and can be described as follows. The net bridge output voltage is measured under the two phased-sequence conditions, as shown. A first measurement (top) drives the bridge at the top node with excitation voltage V_B. This yields a first-phase measurement output V_A, where V_A is the sum of the desired bridge output voltage V_O and the net offset error voltage E_{OS}.

In the second measurement (bottom) the polarity of the bridge excitation is then reversed, and a second measurement, V_B, is made. Subtracting V_B from V_A yields $2V_O$, and the offset error term E_{OS} cancels as noted from the mathematical expression in the figure.

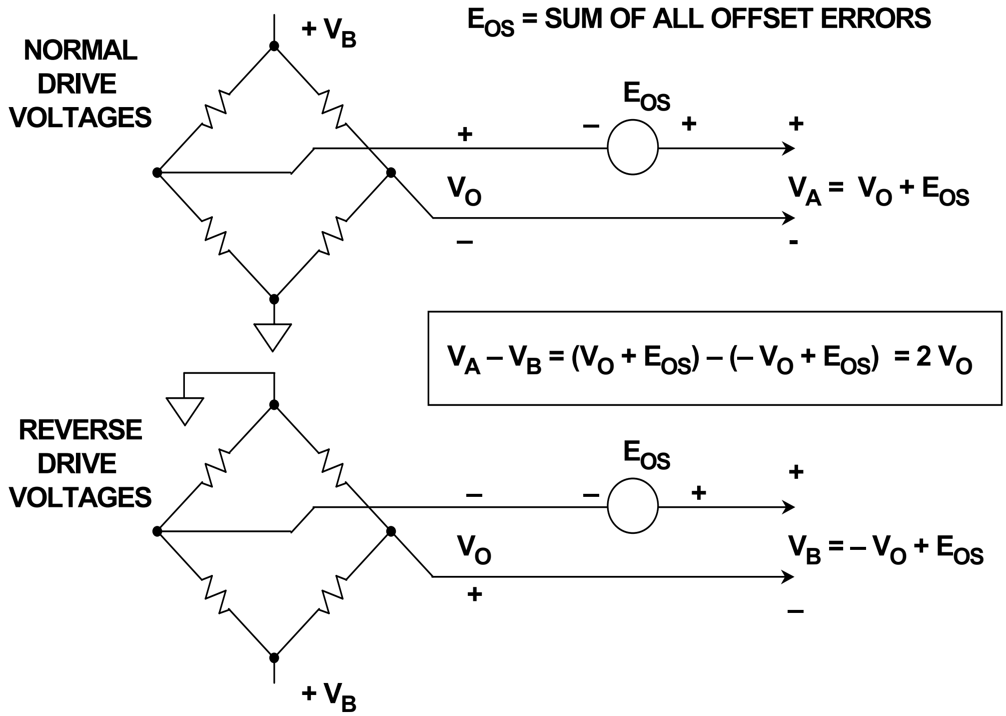


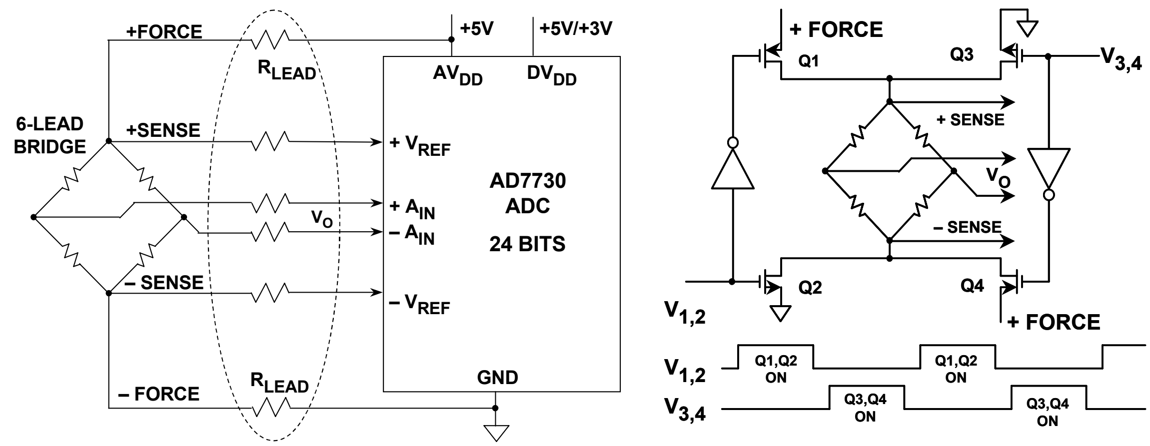
Figure 3.76: *AC bridge excitation minimizes system offset voltages*

Obviously, a full implementation of this technique requires a highly accurate measurement ADC such as the AD7730 (see Reference 5) as well as a microcontroller to perform the subtraction.

Note that if a ratiometric reference is desired, the ADC must also accommodate the changing polarity of the reference voltage, as well as sense the magnitude. Again, the AD7730 includes this capability.

A very powerful combination of bridge circuit techniques is shown in Figure 3.77, an example of a high performance ADC. In Figure 3.77A is shown a basic DC operated ratiometric technique, combined with Kelvin sensing to minimize errors due to wiring resistance, which eliminates the need for an accurate excitation voltage.

The AD7730 measurement ADC can be driven from a single supply voltage of 5 V, which in this case is also used to excite the remote bridge. Both the analog input and the reference input to the ADC are high impedance and fully differential. By using the + and – SENSE outputs from the bridge as the differential reference voltage to the ADC, there is no loss in measurement accuracy if the actual bridge excitation voltage varies.



(A) DC excitation **(B) AC excitation (simplified)**

Figure 3.77: *Ratiometric DC or AC operation with Kelvin sensing can be implemented using the AD7730 ADC*

To implement AC bridge operation of the AD7730, an "H" bridge driver of P-Channel and N-Channel MOSFETs can be configured as shown in Figure 3.77B (note — dedicated bridge driver chips are available, such as the Micrel MIC4427). This scheme, added to the basic functionality of the AD7730 configuration of Figure 3.77A greatly increases the utility of the offset canceling circuit, as generally outlined in the preceding discussion of Figure 3.76.

Because of the on-resistance of the H-bridge MOSFETs, Kelvin sensing must also be used in these AC bridge applications. It is also important that the drive signals be non-overlapping, as noted, to prevent excessive MOSFET switching currents. The AD7730 ADC has on-chip circuitry which generates the required non-overlapping drive signals to implement this AC bridge excitation. All that needs adding is the switching bridge as noted in Figure 3.77B.

The AD7730 is one of a family of sigma-delta ADCs with high resolution (24 bits) and internal programmable gain amplifiers (PGAs) and is ideally suited for bridge applications. These ADCs have self- and system calibration features, which allow offset and gain errors due to the ADC to be minimized. For instance, the AD7730 has an offset drift of 5 nV/ºC and a gain drift of 2 ppm/ºC. Offset and gain errors can be reduced to a few microvolts using the system calibration feature.

REFERENCES: BRIDGE CIRCUITS

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