Current Loop Signal Conditioning: Practical Applications

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PRACTICAL APPLICATIONS

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ABSTRACT

This paper describes a variety of practical application circuits based on the current loop signal conditioning paradigm. Equations defining the circuit response are also provided. The constant current loop is a fundamental signal conditioning circuit concept that can be implemented in a variety of configurations for resistance-based transducers, such as strain gages and resistance temperature detectors. The circuit features signal conditioning outputs which are unaffected by extremely large variations in lead wire resistance, direct current frequency response, and inherent linearity with respect to resistance change. Sensitivity of this circuit is double that of a Wheatstone bridge circuit. Electrical output is zero for resistance change equals zero. The same excitation and output sense wires can serve multiple transducers. More application arrangements are possible with constant current loop signal conditioning than with the Wheatstone bridge.

INTRODUCTION

The Wheatstone bridge circuit has a long history of successfully being used to measure electrical resistance and small changes in that resistance. The variable resistance strain gage has used the Wheatstone bridge circuit in various forms for signal conditioning since its inception. An adaptation of the Wheatstone bridge includes multiple constant current excitation sources within and external to the bridge. A similar technique for minimizing the number of lead wires in multi-channel strain measurements also exists.

Current loop topology was developed to overcome the inherent difficulties of the Wheatstone bridge without the complexity arising from multiple excitation sources. An extension of this paradigm provides the ability to simultaneously measure temperature and strain by using thermocouple wire to connect a variable-resistance strain gage to the signal conditioning circuitry. The current loop is in daily use for strain gage signal conditioning at the NASA Dryden Flight Research Center.

This paper reviews the theory of the current loop paradigm and presents various possibilities for accomplishing the key voltage difference measurement function. Two loop current-regulation approaches are presented. In addition, a variety of circuit applications based on the current loop signal conditioning paradigm is described. Equations defining the circuit response are presented.
Key contributions of Allen R. Parker, Jr., who implemented the equations of the constant loop signal conditioning concept with practical circuitry and software, are gratefully acknowledged.

CURRENT LOOP THEORY

The current loop paradigm is a fundamental circuit concept that will operate with various electrical components and forms of excitation. Excitation possibilities include direct, alternating, and pulsed currents. Inductance and capacitance measurements are possible with alternating current excitation. The various forms of excitation each have advantages that indicate their selection for use in certain applications and environments. For simplicity, direct current excitation and resistive components are used in the illustrations and equations presented in this paper.

Single Remote Gage Resistance

Figure 1 diagrams the current loop signal conditioning paradigm and illustrates the theory that explains its operation for a single-gage resistance sensor. The unique part of the approach illustrated in figure 1 is the four-terminal voltage difference measuring system. The $R_w1$ through $R_w4$ are lead wire resistances with $R_w1$ and $R_w2$ carrying the constant excitation current, $I$. The gage is modeled by an initial resistance, $R$, in series with its resistance change, $\Delta R$. Note that if the sensing system for the voltage across the gage, $V_g$, has a sufficiently high input impedance, then

\[ V_{out} = I(\Delta R) \]

Figure 1. Current loop circuit for single-gage resistance.
no appreciable current will flow through \( R_{w3} \) and \( R_{w4} \). As a result, no significant voltage drop will occur across them. The \( R_{ref} \) is a reference resistor used to develop a voltage, \( V_{ref} \), which is subtracted from the voltage across the gage, \( V_g \).

The four-terminal, high-impedance voltage difference measuring system of figure 1 uses two terminals to sense \( V_g \) and two terminals to sense \( V_{ref} \). Equations 1 through 3 model the circuit and illustrate the benefit of this four-terminal voltage measurement in a single constant current loop.

\[
V_{out} = V_g - V_{ref} \tag{1}
\]

\[
V_{out} = I(R + \Delta R) - I(R_{ref}) \tag{2}
\]

When \( R_{ref} = R \),

\[
V_{out} = I(\Delta R) \tag{3}
\]

Note that \( R_w \) does not appear in equations 1 through 3; therefore, \( V_{out} \) is theoretically uninfluenced by any \( R_w \).

A small difference between the initial gage resistance and the reference resistor will result in a correspondingly small output offset. This offset can be subtracted out in data reduction. This practice is standard procedure in strain-gage data reduction. Such subtraction is also commonly used with practical Wheatstone bridge circuits. The maximum output voltage change per unit of resistance change is achieved when using constant current excitation. Ignoring the second-order effects of the \( \Delta R \) term in the denominator of the equation for the Wheatstone bridge output gives

\[
e_o = (E_x/4)(\Delta R/R) \tag{4}
\]

where \( e_o \) is the output, and \( E_x \) is the excitation for Wheatstone bridge circuits. Because the \( E_x \) is \( 2V_g \) in a Wheatstone bridge circuit, the output in terms of the gage current and gage resistance change is

\[
e_o = I(\Delta R)/2 \tag{5}
\]

Note that the output available from the Wheatstone bridge is one-half of the output available from the constant current loop output (eq. 3).

**Multiple Remote Resistances**

The same reference resistor voltage drop, \( V_{ref} \), can be used as an input for more than one voltage difference function. This feature makes it practical to include more than one gage resistance in a single current loop. The key benefit of including multiple gages in the current loop is a reduction in the required number of lead wires.\(^5\) To make apparent strain corrections, the reference resistance can be a gage resistance to achieve temperature compensation and arithmetic calculations. Refer to the Apparent Strain Corrections sub-subsection for additional details.
Figure 2 illustrates three gage resistances, $R_{g1}$, $R_{g2}$, and $R_{g3}$, in a single loop. This configuration is applicable to the common technique of using a group of three strain gages installed near each other to estimate the magnitude and direction of principal strain. The advantages of the constant current loop are obtained with only six lead wires. That is three wires less than are required when using a Wheatstone bridge circuit for the same measurement requirement.

$$\frac{V_{i o}}{i} \cdot I$$

Figure 2. Strain-gage rosette measurement using current loop signal conditioning with six lead wires.

**Ratiometric Current Loop Measurements**

A single reference voltage used to derive system voltage reference levels can simplify and stabilize the measurement system. Figure 3 illustrates the ratiometric current loop which uses $V_{ref}$ to normalize the output voltage $V_{out}$. If the loop current should vary, then $V_{ref}$ will vary by the same amount. The resulting data values are the same regardless of the level of excitation current as long as no appreciable current variation occurs during the analog-to-digital conversion process. With ratiometric measurements,

$$\text{Data} = \frac{V_{out}}{V_{ref}} \quad (6)$$
When $R_{ref} = R$, 

$$\text{Data} = \frac{\Delta R}{IR_{ref}} \quad (7)$$

When $R_{ref} = R$, 

$$\text{Data} = \frac{\Delta R}{R} \quad (8)$$

independent of $I$. For this reason, ratiometric voltage measurements make excitation regulation theoretically unnecessary.

![Figure 3. Current loop system with ratiometric output.](image)

Practical ratiometric current loop voltage measurements can be accomplished in at least two ways. The $V_{ref}$ can be used as the reference input to the system analog-to-digital converter to achieve ratiometric current loop measurements. Alternatively, the measurement of $V_{out}$ can be numerically divided by a high-resolution measurement of $V_{ref}$ taken at essentially the same excitation current $I$ during which $V_{out}$ was measured.

In practice, the noise floor of the data varies with excitation level. Lower excitation levels necessarily result in a lower signal-to-noise ratio. If, however, excitation is maintained at a reasonable level, then the data output will be at least as precise and accurate from ratiometric measurements as with carefully regulated excitation. Current loop signal conditioning can be designed to operate without the expense of regulation circuitry over the useful life of a battery power supply.

For simplicity, the equations developed later in this paper assume constant current regulation. Those equations adapt directly to ratiometric measurements to yield the same data result with unregulated excitation.
VOLTAGE DIFFERENCE MEASUREMENT

The key function that makes possible the current loop paradigm is four-terminal voltage difference measurement. This measurement can be accomplished in many ways. The objective is to develop an output that is in direct proportion to the difference between two electrical potential differences. The resulting output must have appropriate stability and resolution for the intended application. Strain gage signal conditioning requires stability and resolution to within a few microvolts.

Several fundamental possibilities have been identified for accomplishing voltage difference measurement and are described in the following subsections. These possibilities develop a single potential difference output which is then observed with a conventional two-input voltmeter having suitable precision and stability. Other possibilities may also exist.

Potential Transport

A first potential difference can be transported from an inconvenient environment to another circuit location where it can be conveniently observed. This approach has found use in the "flying capacitor" multiplexer circuit.

Figure 4 shows a flying-capacitor-based current loop circuit which uses this approach. In the development of the current loop concept, potential transport was the first approach identified which

![Figure 4. The flying capacitor circuit for developing a voltage difference measurement.](image-url)
provided sufficient stability and resolution for strain gage signal conditioning. The $V_{\text{ref}}$ is transported to appear across a capacitor in series with $R_{w4}$. Then, $V_{\text{out}}$ is observed as the voltage difference between $V_g$ and $V_{\text{ref}}$. This circuit includes switches to accomplish various calibration and data validity assurance functions. Excitation defeat, output short, and shunt calibration can be added to all circuit examples.\textsuperscript{5}

**Current Transport**

An electrical current can be modulated to carry information. This approach is used in the data current and current-summing amplifier circuits which are described in the following sub-subsections.

**Data current circuit**

A voltage difference measurement can be accomplished when a "data current" is routed through a load resistance located where its voltage drop can be observed in series opposition to a second voltage level. A conventional voltmeter then indicates the desired voltage difference.

Figure 5 shows a current loop circuit using a data-current-based voltage difference measuring circuit. Here, the voltage developed across $R_{d1}$ is caused to equal $V_g$ by the operational amplifiers OA1 and OA2 and the current regulator pass element, Q1. This operation develops data current

![Figure 5. Data current circuit for developing a voltage difference.](image-url)
\[ ID = \frac{V_g}{R_d} \quad (9) \]

Amplifier OA1 is connected to cause the voltage sensed through \( R_{w3} \) to appear at the top end of \( R_{d1} \). The input of OA2 causes the voltage sensed through \( R_{w4} \) to appear at the bottom end of \( R_{d1} \) by turning on Q1 to cause the voltage drop across \( R_{d1} \) to equal \( V_g \). The \( R_h \) provides a loop voltage drop to allow enough “headroom” to permit Q1, the data current pass element, to operate unsaturated.

The voltage drop across \( R_{d2} \) is equal to \( V_g \) when \( R_{d2} \) is equal to \( R_{d1} \). The \( V_{out} \) is then the desired voltage difference between \( V_g \) and \( V_{ref} \). Amplification is available in this circuit when \( R_{ref} \) and \( R_{d2} \) are proportionally greater than \( R \) and \( R_{d1} \). The output of this circuit is

\[ K = \frac{R_{ref}}{R} = \frac{R_{d2}}{R_{d1}} \quad (10) \]

\[ V_{out} = KV_g - V_{ref} \quad (11) \]

\[ V_{out} = KI\Delta R \quad (12) \]

**Current-summing amplifier circuits**

Operational amplifiers connected to perform precision analog arithmetic can develop an output proportional to the difference in two input potential differences. The following sub-subsections use a summing amplifier and an instrumentation amplifier as examples to illustrate these possibilities.

**Summing amplifiers.** Figure 6 illustrates a classic analog subtraction circuit. This circuit uses operational amplifiers in a summing configuration to develop an output proportional to the voltage difference between two sets of floating inputs. Amplifiers OA1 through OA4 act as buffers to present a high impedance at their four inputs to the circuit nodes where the two voltage drops, \( V_g \) and \( V_{ref} \), are sensed. Amplifier OA4 is unnecessary when its input is from a low-impedance point, such as a power supply output. Input summing resistances, \( R_i \), and gain-setting resistances, \( R_o \), are each matched resistance sets. If the \( R_i \) resistors were directly connected to \( V_g \) and \( V_{ref} \), then significant currents could be diverted from the current loop to the voltage difference measuring system, hence the need for buffer amplifiers OA1 through OA4. Absence of buffer amplifiers could cause the output to be unacceptably influenced by \( R_{w1} \) through \( R_{w4} \). Amplification is available in this circuit in proportion to \( R_o/R_i \). The output of this circuit is

\[ V_{out} = (V_g - V_{ref})(R_o/R_i) \quad (13) \]

\[ V_{out} = I\Delta R(R_o/R_i) \quad (14) \]

**Instrumentation amplifiers.** Figure 7 illustrates subtraction by means of an instrumentation amplifier circuit. When operating at unity gain, an instrumentation amplifier produces an output voltage equal to the voltage difference between its input terminals. This output voltage is developed with respect to the point at which the output sense terminal is connected. By this means, the input level can be replicated at another point in the circuit to appear in series opposition to a second voltage. By connecting the sense terminal to the bottom of the reference resistor, the voltage between the instrumentation amplifier output and the most positive end of reference resistor
is \( V_g - V_{ref} \), the desired voltage difference output. Gain is also available in this circuit when the instrument amplifier gain is adjusted to equal \( R_{ref}/R \). The INA114 instrumentation amplifier is an appropriate choice for this purpose because of its low output-referenced errors.

**CALIBRATION APPROACHES**

A means for calibrating the overall measuring system end-to-end with respect to input resistance changes is a desirable operational feature. Fortunately, there is no need to parallel a remote \( R_g \) to achieve a useful calibration for sensitivity to individual loop resistance changes because current is the same in all parts of the loop. The circuitry carrying loop current is indicated by heavy lines in figures 1 through 15. If the desired output is the difference between two remote resistance changes, then paralleling one of these resistances may be necessary for a useful calibration. Calibration by changing the reference voltage and gage current are described next.

**Changing Reference Voltage**

Figures 1, 2, 4, 5, 6, and 7 show a calibration circuit that changes the reference voltage by a predictable amount, \( \Delta V_{cal} \). This circuit consists of a calibration resistor, \( R_{cal} \), which is electrically paralleled with the reference resistor, \( R_{ref} \), while the calibration switch is closed. This connection reduces the apparent resistance of \( R_{ref} \) by \( \Delta R_{cal} \) as calculated from

\[
\Delta R_{cal} = R_{ref} - (R_{ref})(R_{cal})/(R_{ref} + R_{cal})
\]
Figure 7. Instrumentation amplifier circuit for developing a voltage difference measurement.

The mechanical strain simulated by a $\Delta V$ calibration is

$$\text{Strain} = \frac{\Delta R_{\text{cal}}}{(GF R_g)} = \frac{R_{\text{ref}}}{R_g (R_{\text{ref}} + R_{\text{cal}})}$$

(16)

For convenience in reducing strain gage data when $R_g = R_{\text{ref}}$,

$$\text{Strain} = \frac{R_{\text{ref}}}{GF (R_{\text{ref}} + R_{\text{cal}})}$$

(17)

Because the same current, $I$, flows in all parts of the current loop, an apparent reduction $\Delta R_{\text{cal}}$ in $R_{\text{ref}}$ appears in the system output as a voltage change, $\Delta V_{\text{cal}}$, as though there had been an equivalent increase, $\Delta R_{\text{cal}}$, in $R_g$. Thus, $R_{\text{cal}}, R_g$, and $R_{\text{ref}}$ define a reliable overall measurement system sensitivity factor when a change in system output is caused by paralleling $R_{\text{ref}}$ with $R_{\text{cal}}$.

**Changing Gage Current**

Several new opportunities for circuit features develop when the voltage $V_{\text{ref}}$ across $R_{\text{ref}}$ is controlled to be constant in the feedback loop which regulates excitation current. As an example, figure 8 illustrates the change in excitation current, $\Delta I_{\text{cal}}$, calibration technique.
The constant current regulator operates by forcing sufficient current through the loop to cause the voltage drop across $R_{ref}$ to equal the reference source. This operation maintains the loop current at

$$I = \frac{V_{ref}}{R_{ref}}$$  \hspace{1cm} (18)

Connecting $R_{cal}$ in parallel with $R_{ref}$ causes a calibration current increment, $\Delta I_{cal}$, to additionally flow in the constant current loop.

$$\Delta I_{cal} = \frac{V_{ref}}{R_{cal}}$$  \hspace{1cm} (19)

The output indication $\Delta V_{cal}$ is a function of the gage resistance $R_g$ and $\Delta I_{cal}$. Note that from the voltage difference measuring system perspective, $\Delta V_{cal}$ could have been developed by either a change in gage resistance, $\Delta R_{cal}$, or by a change in excitation current, $\Delta I_{cal}$. Equation 20 defines this equivalence.

$$\Delta V_{cal} = \Delta I_{cal} R_g = I \Delta R_{cal}$$  \hspace{1cm} (20)

Substitution shows that

$$\frac{\Delta R_{cal}}{R_g} = \frac{R_{ref}}{R_{cal}}$$  \hspace{1cm} (21)
Note that the denominator of equation 21, which models the current change calibration, does not include $R_g$ as does equation 16 for voltage change calibration. By definition, strain is developed from resistance measurements by means of a gage factor, $GF$, calibration. That is,

$$GF(\text{strain}) = \Delta R/R$$

(22)

As a result, the strain simulated by a $\Delta I$ calibration is

$$\text{Strain} = \frac{R_{ref}}{GF R_{cal}}$$

(23)

This result is interesting in that the data shift caused by paralleling $R_{ref}$ with $R_{cal}$ provides the system sensitivity to $\Delta R/R$ without prior knowledge of $R_g$.

Note that $\Delta I$ calibration involves precise currents flowing through $R_{ref}$ and $R_g$. A small systematic error can exist when a $\Delta I$ offset adjustment circuit is also in use. This error is typically ignored, but it is simple to remove at the “balance” condition (zero electrical output from the voltage difference measuring system). In this situation, the magnitude of $R_g$ instead $R_{ref}$ in equation 21 is used.

OFFSET ADJUSTMENTS

Offset adjustments should be derived from the excitation current level. This derivation will cause excitation level variations to result in percent-of-reading sensitivity errors rather than in additional percent-of-full-scale offset drifts. Offset adjustment by changing the reference voltage and by changing the gage current are described next.

Changing Gage Current

Figure 9 illustrates a $\Delta I$ offset adjustment circuit. Magnitude of the offset adjustment is limited by $R_{offset}$. Resistor $R_{offset}$ is connected between the positive end of $V_{ref}$ and a potential $V_{offset}$, with a magnitude and polarity adjustable between zero and $2V_{ref}$. This circuit provides a variable bipolar offset current, $\pm I_{offset}$, which increases or decreases $I_g$.

$$\pm I_{offset} = \pm V_{offset}/R_{offset}$$

(24)

The offset current is applied additively to the gage current, $I_{gage}$, to cause the gage voltage, $V_g$, to approach $V_{ref}$, the voltage drop across the reference resistor.

$$V_g = (I_{ref} + I_{offset})R_g$$

(25)

Changing Reference Voltage

Figure 10 illustrates a $\Delta V$ offset adjustment circuit with $\Delta I$ calibration. The offset level is applied additively to the reference voltage before it is sensed by the voltage difference measuring system. This approach does not affect the level of calibration output from shunting $R_{ref}$ with $R_{cal}$. Offset authority is established by the ratio of output-to-input offset amplification resistances, $R_{oo}$ to $R_{io}$. 
Figure 9. Current loop circuit with $\Delta I$ offset adjustment.

Figure 10. Current loop circuit with reference voltage offset adjustment.
APPLICATIONS

Examples shown in figures 1 through 10 sense the voltage drop, \( V_g \), directly across a gage resistance. This connection causes the output of current loop signal conditioning to be uninfluenced by any lead wire resistance as long as the voltage difference signal conditioning system and the current regulator operate within their ability to reject common mode voltages and to deliver a constant current to the set of resistances in the loop.

In addition, these figures provide separate outputs for each gage resistance in the current loop. Arranging two or more gages in a current loop circuit such that gage resistance changes add, subtract, or both, to develop a single output can be useful. Half- and full-bridge arrangements of the Wheatstone circuit combine gage outputs in this manner. Current loop signal conditioning provides more analog computation opportunities than the Wheatstone bridge. The application examples that follow show how additional computations can be accomplished.

Minimizing Conductor Quantity

If lead wire resistances are consistent enough, then acceptable results may be obtained by using fewer lead wires. Three-wire connections to one-fourth- and one-half-bridge Wheatstone bridge circuits always depend on consistent lead wire resistance.

Three-wire connection of gage resistances in a current loop is accomplished by including a lead wire resistance with each monitored resistance (\( R_w \) with \( R_g \) and \( R_w \) with \( R_{\text{ref}} \)) (fig. 11). As long as \( R_w \) and \( R_{\text{ref}} \) remain identical, they can vary without their changes being observed. All other benefits of current loop signal conditioning remain available in this situation.

Equations 26 and 27 describe three-wire gage connections.

\[
V_{\text{out}} = I \left[ (R + \Delta R + R_w) - (R_{\text{ref}} + R_w) \right]
\]  \hspace{1cm} (26)

When \( R_w = R_w \) and \( R_{\text{ref}} = R \),

\[
V_{\text{out}} = I (\Delta R)
\]  \hspace{1cm} (27)

This result is the same as in equation 2. Lead wires with resistances that vary identically will induce no more than a constant offset in the output indication.

Using Analog Computations

Analog computations are possible in current loop circuits by including the voltage drops of additive gages in the direct \( (V_g) \) input and the voltage drops of subtractive gages in the inverting \( (V_{\text{ref}}) \) input of a voltage difference measurement circuit. In this situation, calibration by shunting a remote resistance in the circuit may be necessary. Shunting remote gage resistances through their sense lead wires is necessary because no “local” reference resistance is sensed by the voltage difference measuring circuit. Single and multiple loop computations are discussed in the following sub-subsections.
Single loop

A variety of analog computations can be implemented within a single loop. These computations are accomplished by using one or more remote gages in a current loop to develop $V_g$, $V_{ref}$, or both. Note also that $V_s$ for one voltage difference measuring circuit can be used as $V_{ref}$ for another circuit. This feature makes the constant current loop an extremely versatile circuit for analog computations based on changes in remote gage resistances.

Apparent strain corrections. These corrections are accomplished by developing $V_{ref}$ from the voltage drop across an "unstrained" gage in the same temperature environment as one or more strain-sensing gages. Figure 12 shows how apparent strain corrections are done without developing errors from lead wire resistances. Here, shunting $R_{cal}$ across the remote unstrained gage provides a simultaneous calibration output for each of the strain-sensing voltage difference measuring systems.

Unlike the Wheatstone bridge, a single unstrained gage in a current loop can provide temperature compensation for several independent strain-sensing gages, for example, a strain gage rosette. This circuit minimizes gage and lead wire quantity in a circuit that is insensitive to lead wire resistance changes. If wire resistances $R_{w1}$ and $R_{w2}$ remain alike, then only three lead wires may be required.
Wheatstone bridge computation similarities. Figure 13 illustrates how a set of four gages can be connected in a current loop such that their resistance changes add and subtract in a manner similar to Wheatstone bridge circuits. Two gages comprise the additive and two gages comprise the subtractive voltage-sensing segments of the current loop. Note that any number of gages could have been included to expand the analog computation equation. When each gage has the same initial resistance, the output from this circuit is

\[
V_{out} = I(\Delta R_{g_1} + \Delta R_{g_2} - \Delta R_{g_3} - \Delta R_{g_4}) \tag{28}
\]

Gage resistance labels in figure 13 do not reflect the adjacent positions they would have in a four-arm Wheatstone bridge arrangement. Resistance changes in opposite Wheatstone bridge arms are additive; such changes in adjacent arms subtract. If \( R_{w1} \) and \( R_{w2} \) are sufficiently alike, then the current loop equivalent of the Wheatstone bridge can be achieved with only three lead wires.

Multiple loops

Figure 14 illustrates how multiple current loop channel outputs can be combined to achieve a single output that is independently influenced by each current loop. This circuit accomplishes the calculations for combining measurement channels to implement loads equations. The output from this circuit is

\[
V_{out} = RF(V1/R11 + V2/R12 + V3/R13 - V4/R14 - V5/R16 - V6/R16) \tag{29}
\]
Figure 13. Analog computation using four gages.

Figure 14. Multiple loop computation circuit.
Modifying Existing Wheatstone Bridge Systems

A substantial capital investment already exists in measurement systems that use Wheatstone bridge-based signal conditioning. Converting these existing systems to current loop operation is possible and practical.

Figure 15 shows a NASA-designed circuit modification to the existing Dryden Flight Research Center, Thermostructural Laboratory, Edwards, California, data acquisition system. This modification converts the signal conditioning from Wheatstone bridge to current loop by replacing the Wheatstone bridge completion and calibration circuitry. No other hardware or software changes were required to include current loop signal conditioning.

Several hundred channels of this circuit are now in daily operational use. All component values are identified. The designated INA114 instrumentation amplifier component is critical. The three operational amplifiers which it contains are essentially identical. As a result, this component has exceptionally low output-referred errors.

![Figure 15. Modification to convert existing equipment from Wheatstone bridge to current loop circuitry.](image)

CONCLUSIONS

The constant current loop is a fundamental signal conditioning circuit concept that can be implemented in a variety of configurations. Current loop signal conditioning circuits can be insensitive
to changes in the resistance of any lead wire. The current change calibration identifies sensitivity to strain by developing an output that is directly proportional to the gage resistance at the time of calibration. Adapting existing Wheatstone bridge-based measurement systems to current loop operation can be practical. More application arrangements are possible using the constant current loop than using the Wheatstone bridge.

REFERENCES


This paper describes a variety of practical application circuits based on the current loop signal conditioning paradigm. Equations defining the circuit response are also provided. The constant current loop is a fundamental signal conditioning circuit concept that can be implemented in a variety of configurations for resistance-based transducers, such as strain gages and resistance temperature detectors. The circuit features signal conditioning outputs which are unaffected by extremely large variations in lead wire resistance, direct current frequency response, and inherent linearity with respect to resistance change. Sensitivity of this circuit is double that of a Wheatstone bridge circuit. Electrical output is zero for resistance change equals zero. The same excitation and output sense wires can serve multiple transducers. More application arrangements are possible with constant current loop signal conditioning than with the Wheatstone bridge.