Keywords: Linearization of Wheatstone Bridge, bridge circuits, RTD, temperature detection, temperature monitoring, analog front end for temperature sensing, Precision amplifier, Precision voltage reference, MAX44267, MAX6025, industrial, medical

APPLICATION NOTE 6144

# LINEARIZATION OF WHEATSTONE-BRIDGE

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*Abstract: This application note discusses the resistance-variable element in a Wheatstone bridge—the first choices for front-end sensors. We will examine its behavior and explain how to linearize the bridge circuit to optimize performance. The simplicity and effectiveness of a bridge circuit makes it very useful for monitoring temperature, mass, pressure, humidity, light, and other analog properties in industrial and medical applications.*

## Introduction

Wheatstone bridge circuits have been in the field for a very long time and still are among the first choices for front-end sensors. Whether the bridges are symmetric or asymmetric, balanced or unbalanced, you can accurately measure an unknown impedance using the bridge. In fact, the simplicity and effectiveness of a bridge circuit makes it very useful for monitoring temperature, mass, pressure, humidity, light, and other analog properties in industrial and medical applications.

The Wheatstone bridge has a single impedance-variable element that is inherently nonlinear away from the balance point. Bridge circuits are commonly used to detect the temperature of a boiler, chamber, or a process situated hundreds of feet away from the actual circuit. Usually a sensor element, typically a resistance temperature detector (RTD), thermistor, or thermocouple, is situated in the hot/cold environment to provide information about resistance change to temperature.

In the following discussion, we will consider this resistance-variable element in a Wheatstone bridge. We will examine its behavior and explain how to linearize the bridge circuit to optimize performance. Note finally, that when we speak generally about "bridges," this article is focused on circuit design for a Wheatstone bridge.

## Single Variable-Resistance Wheatstone Bridge

Resistance-variable Wheatstone bridge circuits perform most of the front-end tasks in a design. They use inexpensive, accurate discrete parts. By incorporating an RTD element, the bridge's inherent resistance variations are kept within the accepted linearity and tolerance limits, depending on the manufacturer of the RTD.

RTD devices have a very detailed data sheet characterizing their behavior with look-up tables and even transfer function equations down to four or more orders of error compensating terms. To ensure a high-precision system, designers must consider both the inherent nonlinearity of the RTD element and the Wheatstone bridge, then painfully calibrate the front-end, and linearize the front-end at the microcontroller side. Increasing the order of the equation in the microcontroller is going to improve the linearity. A typical bridge circuit (**Figure 1**) detects milliohms of changes in resistance (ΔR).

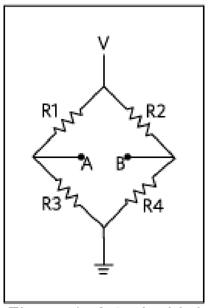


*Figure 1. A typical bridge with nodes A and B sensing output voltage from a change in resistance (ΔR).*

Assuming that R1 = R2 = R3 = R4 = R in Figure 1, the bridge is balanced with nodes A and B at a constant V/2 (volts) and with a differential voltage of 0V across VAB. When there is a change in resistance (ΔR) from R3, then the output differential voltage created is:

$$\left(\frac{V \times R3}{R1 + R3}\right) - \left(\frac{V \times R4}{R2 + R4}\right) = V_{AB}$$

$$V_{AB} = \left(\left(\frac{R3}{R1\left(1+\frac{R3}{R1}\right)}\right) - \left(\frac{R4}{R2\left(1+\frac{R4}{R2}\right)}\right)\right) \times V$$

$$V_{AB} = \left(\frac{\frac{R3}{R1}}{\left(1+\frac{R3}{R1}\right)} - \frac{\frac{R4}{R2}}{\left(1+\frac{R4}{R2}\right)}\right) \times V$$

$$V_{AB} = V \times \left[\frac{\frac{R3}{R1} - \frac{R4}{R2}}{\left(1+\frac{R3}{R1}\right)\left(1+\frac{R4}{R2}\right)}\right] \qquad \text{(Eq. 1)}$$

when R1 = R2 = R3 = R4 = R, the bridge is balanced.

For a single variable-resistance element, for R3 = R + ΔR and R1 = R2 = R4 = R:

$$V_{AB} = V \times \left(\frac{\frac{R+\Delta R}{R} - 1}{\left(1+\frac{R+\Delta R}{R}\right) \times 2}\right)$$

$$V_{AB} = V/4 \times \left(\frac{\Delta R}{\left(R+\frac{\Delta R}{2R}\right)}\right) \qquad \text{(Eq. 2)}$$

**Equation 2** suggests that increasing the constant supply voltage, V, to the bridge will increase the output voltage, i.e., the swing range across the bridge. This also suggests that having a dual supply across the four-legged resistance arrangement could be helpful not only to increase the range, but also to help maintain a 0V common-mode voltage across the AB nodes.

The voltage $V_{AB}$ is usually amplified by using a subsequent amplifying stage typically done with a differential amplifier. There is a caveat, however. Changing the common-mode voltage across $V_{AB}$ adds more error and complexity in the amplifying second stage, which is usually realized as a instrumentation quality differential amplifier. Therefore, a common-mode voltage centered around 0V is preferable and certainly easier to manage.

**Figure 2** illustrates the natural tendency of the bridge's single variable element: inherent nonlinearity in its transfer function.

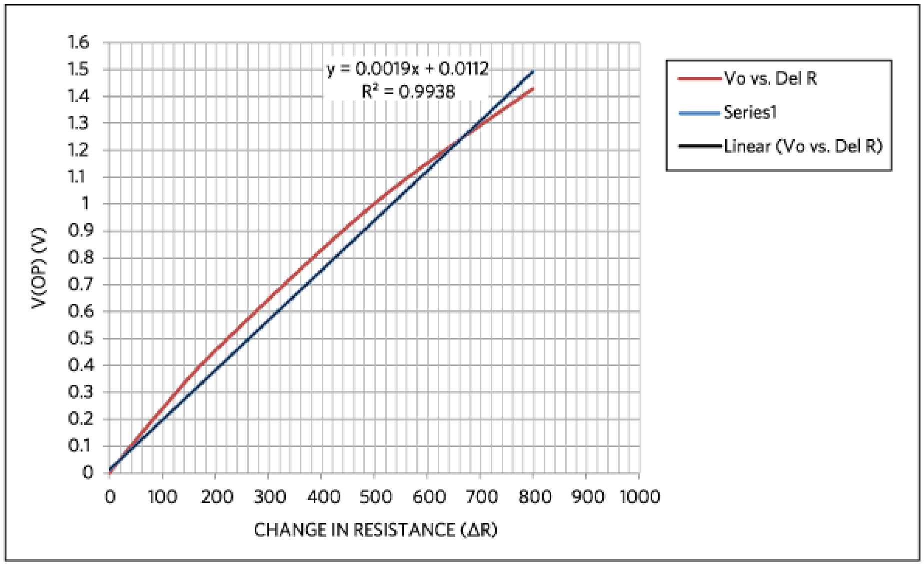


*Figure 2. V(A/B) vs the change in ΔR, the effect of the bridge's nonlinearity from 800Ω of resistance change. Trend line added for comparison.*

Look closer at the trend line in the figure. The curve's absolute deviation from an ideal straight line or, in fact, the linearity error is about 0.62%. This is obtained by comparing the curved trend line with the line of best fit, i.e., the straight line relative to the curve. In this way, we actually quantify the worst-case linearity error for the above curved line. In some cases 0.6% is certainly not acceptable, and this article illustrates a way to achieve better than 0.1% accuracy.

Besides the inherent nonlinearity of the bridge, the designer must also manage the nonlinearity of the temperature sensor element, RTD, or even thermistor as discussed in the prior section. When sensing the differential voltage across the nodes A and B, the instrumentation amplifier (**Figure 3**) has a common-mode voltage of V/2. The amplifier is usually a differential amplifier with four resistors or a three-op-amp instrumentation amplifier integrated in a single package.

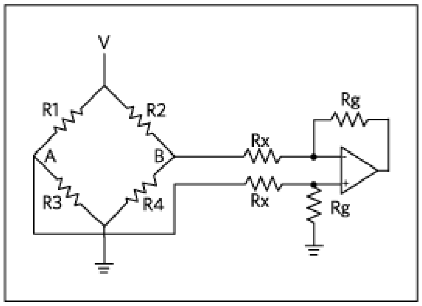


*Figure 3. An instrumentation amplifier, connected to the original bridge circuit in Figure 1.*

When a differential amplifier is used, the nodes A and B are connected to the amplifier's input gain-setting resistors, as shown in Figure 3. The choice of the op amp and the input resistors is important as this path directs current away from the bridge, hence affecting the accuracy.

Also the choice of resistors affects the bridge performance, as even 0.1%-tolerant resistors used with the amplifier provide only 60dB of common-mode rejection.

## Linearize the Bridge Output Without the Instrumentation Amplifier

From the previous discussion it seems logical to have dual supplies across the resistor bridge to increase the dynamic range, and to have the sensing nodes centered around the 0V common mode. The advantage of this design is that the transfer function from the node B is going to be linear with a change in resistance. The range of output swing from the bridge is doubled versus the output from the circuit in Figure 1.

The circuit implementation in **Figure 4** uses two op amps to replace the more complex instrumentation amplifier. Now the linearized bridge output avoids the unnecessary current paths created by the differential amplifier. This circuit eases the design process compared to the circuit in Figure 3. The only issue here is having positive and negative supplies to the amplifiers which are providing twice the swing range. The added benefit is improved common-mode rejection performance, as the second amplifier operates comfortably around 0V.

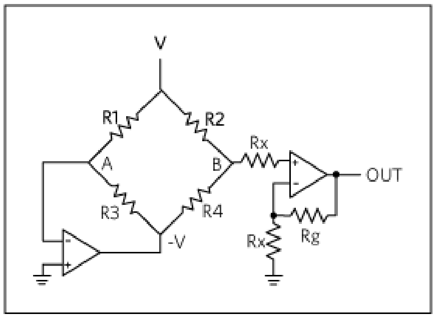


*Figure 4. This circuit replaces the complex instrumentation amplifier (Figure 3) with twoop amps.*

From Figure 4, node A sees GND as it is the summing node of amplifier 1. Thus, a constant current of $I_X = \frac{V}{R1}$ is forced through the R1|R3 branch, producing an equal and opposite voltage on the other side of the bridge with -V. When the single variable-resistance R3 changes (from R3 to R ±ΔR), then Ix (the change in current due to change in the resistance) flowing through this resistance produces a voltage V ±ΔV. A factor of this ΔV is manifested across node B by the balancing of the resistance bridge (for a balanced bridge, of course), as the current forced through resistor branch R2|R4 is equal to (V+ - (V- + ΔV))/(R3 + R4). Since node B is centered at 0V common mode, the voltage produced across node B is going to be gained by a noninverting amplifier. Furthermore, filtering can be done on this gain stage to optimize the bandwidth making the noise level acceptable for the application.

At balance, when R1 = R2 = R3 = R4 = R, the voltage at nodes A and B is:

$$\frac{V-(-V)}{R1+R3} \times R3 = \frac{V-(-V)}{R2+R4} \times R2 = 0$$

When a single variable-resistive element (R3) changes by ΔR (R3 + ΔR):

$$Ix = \frac{V}{R1}$$

1 × x(R3 + ΔR) = (-V) + (1 × x(-ΔR))

Since node A is at ground, the voltage at node B is:

$$\frac{V-(-V-\Delta V)}{R2+R4} \times R2$$

$$\frac{2V+\Delta V}{R2+R4} \times R2 = \left[\frac{2V}{R2+R4} \times R2\right] + \left[\frac{\Delta V}{R2+R4} \times R2\right]$$

We know that $\left[\frac{2V}{R2+R4} \times R2\right]$ term is = 0. Hence the voltage at node B is

$$V_B = \left[\frac{\Delta V}{R2+R4} \times R2\right] = \left[\frac{\Delta V}{1+\frac{R4}{R2}}\right]$$

And in a balanced bridge:

$$\frac{R4}{R2} = \frac{R3}{R1}$$

The same equation can be written as:

$$V_B = \left[\frac{\Delta V}{R2+R4} \times R2\right] = \left[\frac{\Delta V}{1+\frac{R3}{R1}}\right]$$

When using a balanced bridge with ratio $\frac{R4}{R2} = \frac{R3}{R1} = 1$, and we know that ΔV = Ix × ΔR and $Ix = \frac{V}{R1}$, then the equation becomes:

$$V_B = \left[\frac{V}{2}\right] \times \frac{\Delta R}{R1} \qquad \text{(Eq. 3)}$$

And simplistically putting R1 = R2 = R3 = R4 = R, then:

$$V_B = \left[\frac{V}{2}\right] \times \frac{\Delta R}{R}$$

At the output of the noninverting op amp, the equation then is:

$$V_B = -\left(\frac{V}{2} \times \frac{\Delta R}{R1}\right) \times \left(1 + \frac{R_g}{R_b}\right) \qquad \text{(Eq. 4)}$$

This suggests that the output from the second op amp is inverting in nature.

**Figure 5** shows the transfer function and its nonlinearity from the Figure 4 implementation.

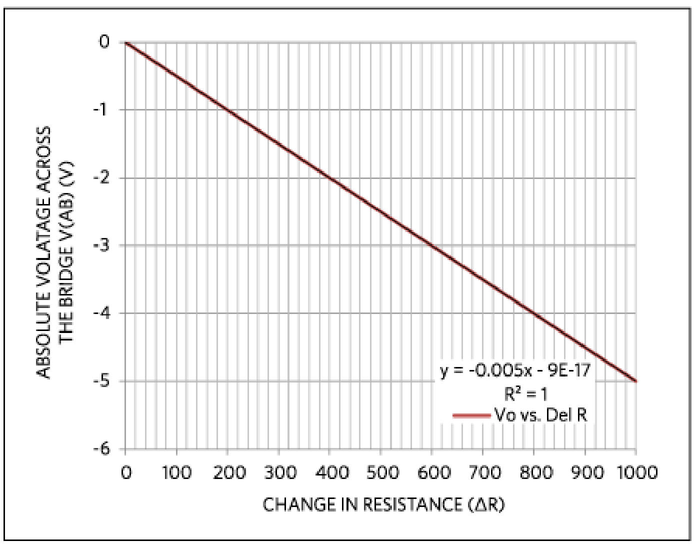


*Figure 5. Bridge output vs. change in resistance. Data are based on the design in Figure 4.*

The absolute deviation from an ideal straight line, i.e., the linearity error, in Figure 5 is less than 0.02%. Improvement in absolute nonlinearity means that the full-scale error or the relative error is also going to improve.

There are no interacting resistive branches, so precision matching of resistors is not required. The variation of the Rx and Rg will only provide a gain error, which can be calibrated at the same as the RTD device.

The above data suggest that this approach can be a viable implementation for 12-, 14-, 16-, even 18-bit applications. The design is simple and very little calibration is needed by the microcontroller. This circuit has, in fact, been widely used in the field for many years.

To implement the Figure 4, circuit you need a dual-supply voltage for the front-end. This negative supply also needs additional board space and components, a requirement that often may not be a viable option if this is the only place in the entire system where the negative supply is needed. Low- offset voltage, low-offset drift, and low-noise performance are additional requirements for a high-precision bridge sensor.

### Implementing the Bridge Design with a Dual Op Amp

What if the amplifier used in Figure 4 needs only one power supply? The MAX44267 operates from a single supply and is capable of outputting bipolar voltages. Unlike other single-supply amplifiers which need headroom above ground, the MAX44267 provides a true-zero output, making it a great fit for bridge sensors (**Figure 6**). The MAX44267 integrates charge-pump circuitry that generates the negative voltage rail in conjunction with external capacitors. This allows the amplifier to operate from a single +4.5V to +15V power supply, but it is as effective as a normal dual-rail ±4.5V to ±15V amplifier.

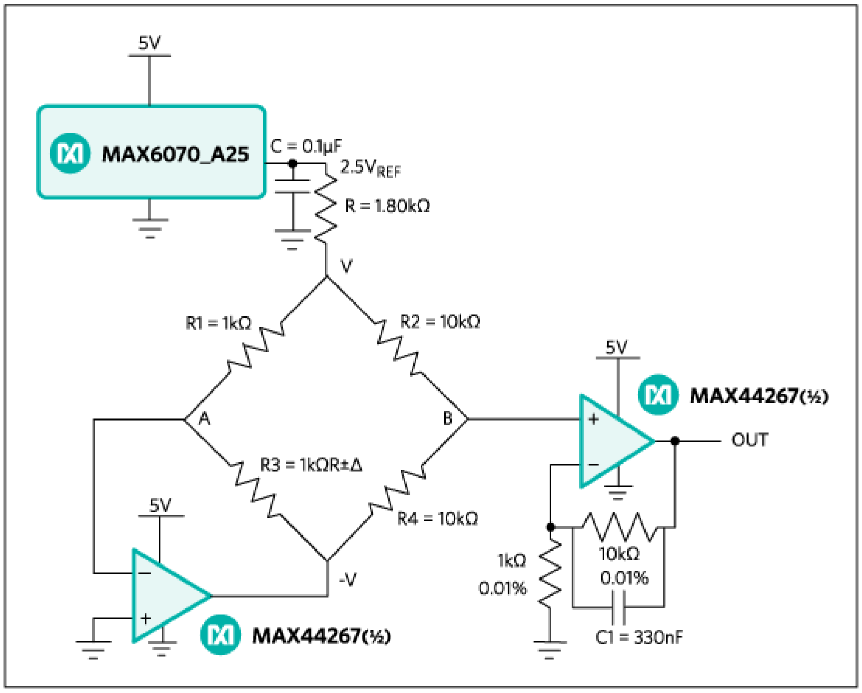


*Figure 6. The MAX44267 precision, low- noise, low-drift, dual op amp offers a true-zero output from a single supply.*

The MAX44267 can be implemented in the Figure 6 circuit with just one supply voltage (positive supply, $V_{CC}$). The integrated negative $V_{SS}$ generator or charge pump generates a negative supply voltage. This architecture provides a good advantage to the designer because it eliminates the need for negative supply regulators and reduces board layout space and cost.

**Figure 7** includes the MAX6070_A25 voltage reference to generate a $2.5V_{OUT}$ reference. A dual op amp (again, the MAX44267) is used with the resistance bridge where R1 = R3 = 1kΩ and R2 = R4 = 10kΩ. An additional 1.8kΩ is used in series to reduce the amount of current flowing through the bridge and to reduce the power dissipation. The V(+) node becomes one-third the voltage reference's output at a balanced condition. This is followed with second-stage amplification with a gain of 11 at the OUTA node.

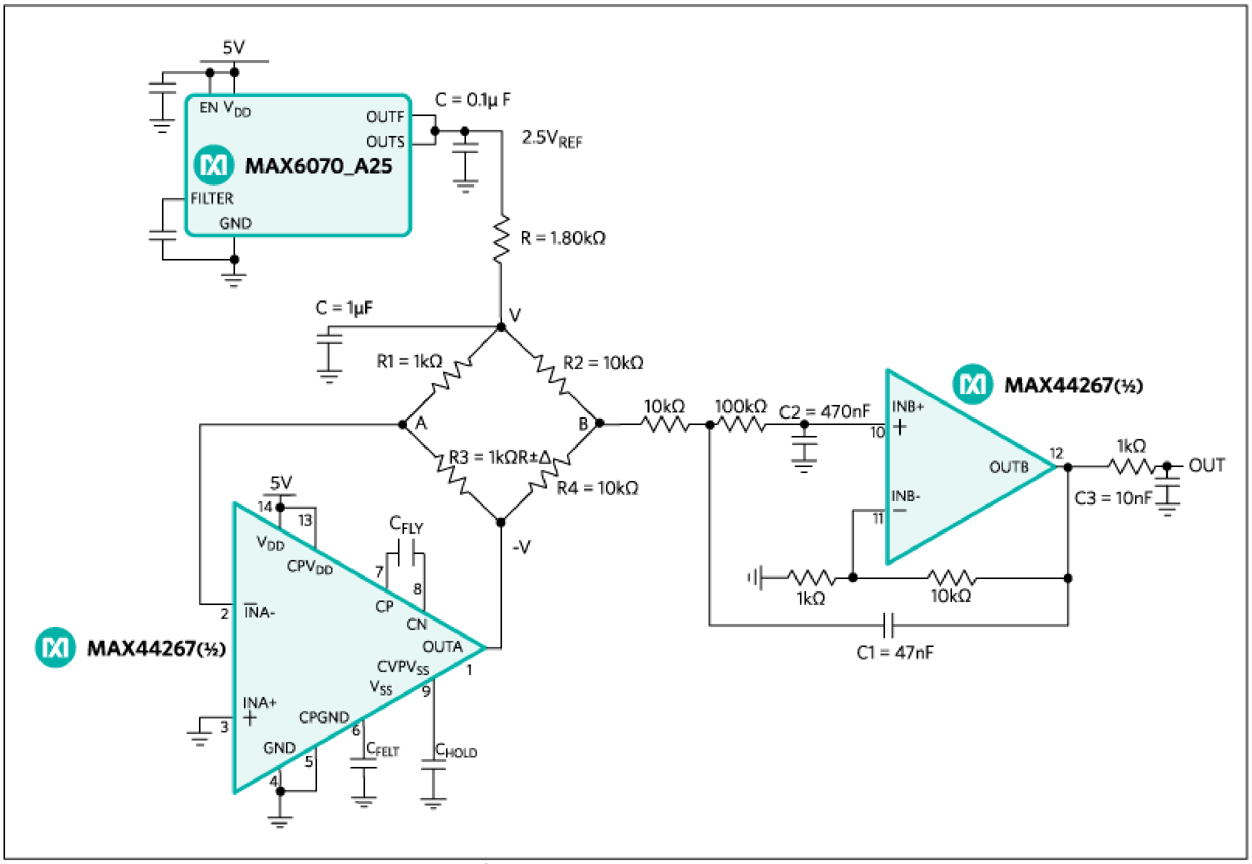


*Figure 7. Because the MAX44267 op amp operates from a single supply.*

A Fluke® RTD calibrator was used as the temperature-dependent resistance element (as PT1000) in place of R3; a temperature change from -50°C to +155°C is evaluated. For the given temperature change using a PT1000, the change in resistance (±R) is about 800Ω and an equivalent range of 325mV is effected (see Equation 4). Because amplifier 2 has an internal negative supply, it can accommodate this swing (-242mV to -83mV) at its input below ground, and it provides an output gain of 11.

Figure 7 utilizes a Sallen-Key filter in the second stage to filter the input signal to the required bandwidth (50Hz used in this case). Full-scale error accuracy within ±0.05% is obtained from the bridge output at node B without any calibration or trim. In this way, the transfer function of the bridge circuit is made linear; improved performance of the front-end circuit is realized using MAX44267 in the subsequent section.

## Test Measurements

1. **Figure 8** shows the absolute bridge voltage output versus change in resistance (a linearity curve output), under 0.02%.

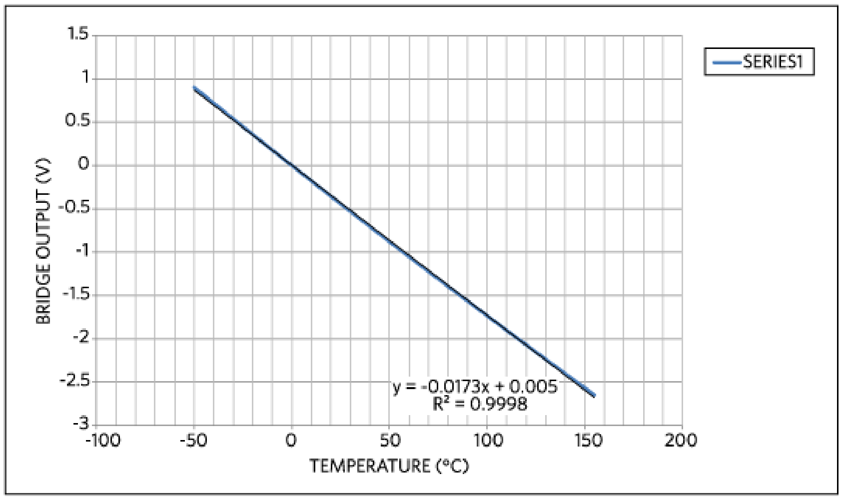


*Figure 8. The transfer curve shows the absolute voltage output vs. temperature for the Figure 7 circuit.*

2. **Figure 9** shows the gain error plot as percentage versus full-scale. The error curve shows small wriggles, in order of 0.002%, that are a combination of manual data plotting and measurement setup noise.

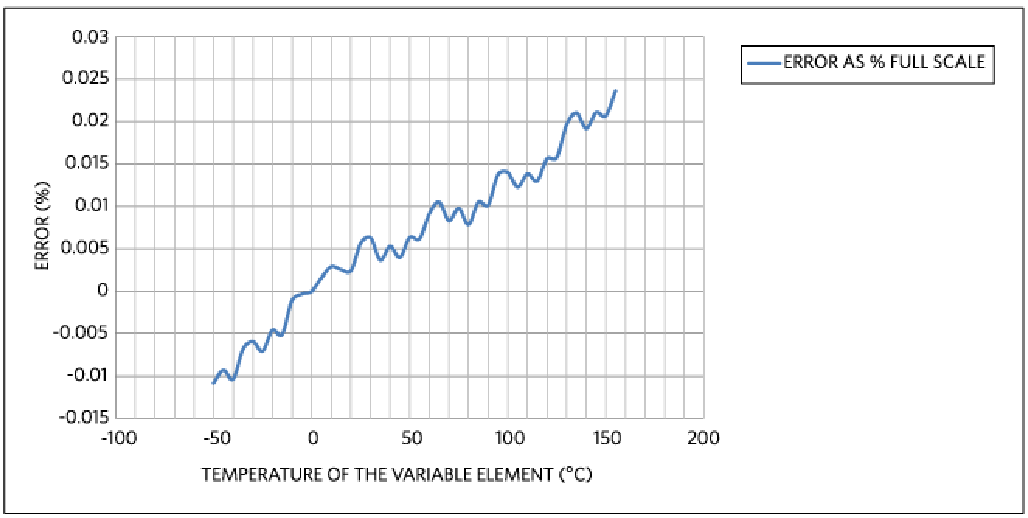


*Figure 9. Error % vs. full scale for the circuit in Figure 7.*

3. **Figure 10** shows the voltage noise density of the bridge plus amplifier: 115nV/Ã Hz at 1kHz and 500nV/Ã Hz under 50Hz. A 50 Hz filter was implemented in the second stage to remove the sensitivity to line noise.

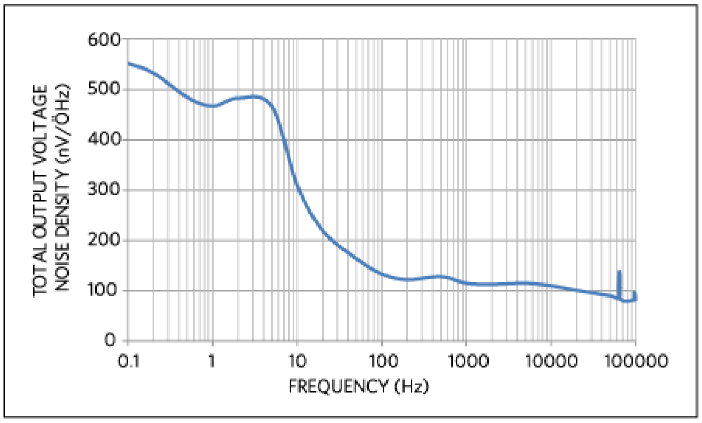


*Figure 10. Output noise density vs. frequency for the circuit in Figure 7.*

4. **Figure 11** shows the voltage noise($V_{P-P}$) of the bridge plus amplifier, 0.1Hz to 10Hz, 6μ$V_{P-P}$.

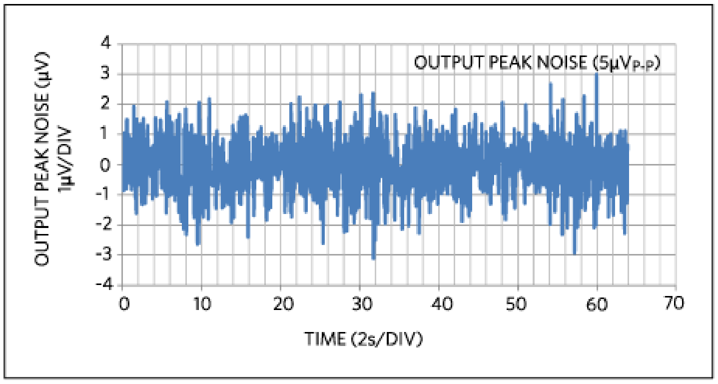


*Figure 11. Peak-to-peak voltage noise from 0.1Hz to 10Hz for the circuit in Figure 7.*

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| Related Parts | | |
|---|---|---|
| MAX44267 | +15V Single-Supply, Dual Op Amp with ±10V Output Range | |
| MAX6070 | Low-Noise, High-Precision Series Voltage References | Free Samples |

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