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Voltage Reference Application and Design Note

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## What and Why is a Voltage Reference?

Conceptually, a voltage reference is a very simple device with only one purpose in its life. Quite simply, the purpose of a voltage reference is to generate an exact output voltage no matter what happens with respect to its operating voltage, load current, temperature changes or the passage of time.

The purpose of this Application Note is to provide an understanding of voltage references, their limits of error, and interesting application circuits using high accuracy voltage references. The Intersil X60008 voltage reference approaches the design in a new and exciting topology that provides extremely tight initial tolerance, low temperature drift, and unbelievably low operating current.

## Where is a Voltage Reference Used?

Voltage references are used to provide a very precise voltage for measurements to be made against. The accuracy of any measurement is only as good as the ability to compare it against a known standard.

High resolution A/D converters and D/A converters, digital meters, smart sensors with threshold detectors, servo systems, battery management, precision regulators and many other precision industrial control systems require a precision voltage reference at their core.

Voltage references can also be used to very accurately set another variable; for example, a laser diode might need to operate at a very precise current to generate the proper wavelength of light. A high accuracy constant current source can be designed using a voltage reference as shown in the Applications section.

## Difference from a Voltage Regulator

Voltage references and voltage regulators seem to be very similar devices; both are used to generate a regulated output voltage that is immune to changes in load current, input voltage, temperature, etc. A voltage regulator is intended to provide higher output current than a voltage reference. However, the voltage regulator is much less accurate than a voltage reference, the output noise is higher, and the long term stability is not specified in a voltage regulator. Additionally, the voltage regulator is put into a package that can withstand the heat that is generated by the power dissipated by the regulator.

## Types of Voltage References - Advantages and Disadvantages

There are different topologies to operate a voltage regulator, and there are many different techniques to generate highly accurate output voltages from a voltage reference. The Intersil X60008 voltage reference approaches the design in a new and exciting topology that provides extremely tight initial tolerance, low temperature drift, very low long term drift, and unbelievably low operating current.

The Intersil Floating Gate Analog (FGA) technology will be discussed, but first let's review the various operating modes and reference techniques.

## Series and Shunt Mode Operation

First, voltage references are designed to operate in either series mode or shunt mode as shown in Figure 1.

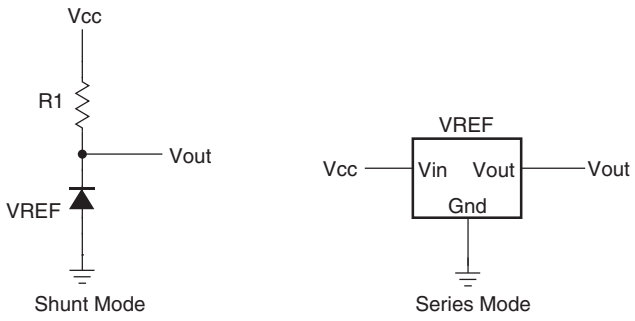



**FIGURE 1.**

Shunt mode references are typically less accurate than Series mode, but require lower operating current. They can be operated from very high input voltage (Vcc) because only the resistor R1 sees the high voltage. Shunt references can be used to generate negative reference voltages or a reference voltage that is floating between potentials.

Series mode references are typically much higher accuracy and lower noise than Shunt mode references. However, the supply voltage (Vcc) is limited to the absolute maximum rating of the device. Generally, a Series mode reference provides a positive output voltage with respect to ground; however, the extremely low supply current of the Intersil X60008 voltage reference allows clever circuit design tricks to be used to allow negative reference voltages or a reference voltage that is floating between potentials.

## Voltage Reference Design Techniques

Integrated voltage references are REALLY difficult to make! Most IC voltage reference designs depend on using nonlinear and highly process dependent characteristic of transistors to cancel the temperature coefficients of diodes or transistors. Extensive trimming of both initial tolerance and temperature drift is required, and, often these two parameters cannot be trimmed to satisfy both conditions. As we will soon see, the Intersil X60008 voltage reference depends only on the ability to force and measure a high accuracy potential. But first, let's review several popular voltage reference topologies – the Bandgap reference, the XFET™ reference, and the buried Zener reference. Finally, I will give an overview of the Intersil Floating Gate Analog technology.

## *Bandgap Voltage Reference*

Bandgap references are ideally suited for voltage reference applications that require low reference voltage (<5V), low operating current (<1mA), medium temperature drift (>20ppm/C), and versatile series/shunt mode operation.

With the exception of the Intersil Floating Gate Analog technology, all voltage reference topologies rely on the inherent temperature dependence of a transistor (bipolar or FET). A highly simplified block diagram of a Bandgap reference is shown in Figure 2.

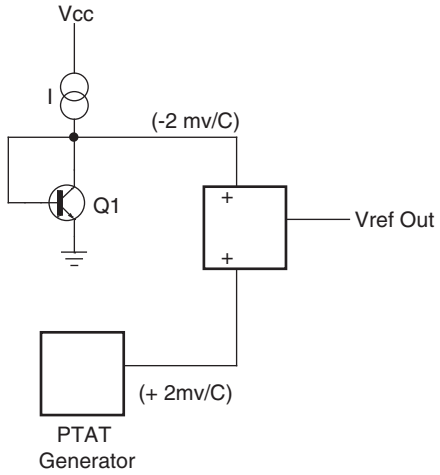



**FIGURE 2.**

Two voltage sources are generated; the first voltage source is the Vbe of a forward biased transistor with an output voltage of 0.7V with a -2mV/°C temperature coefficient. A second voltage source, the Proportional To Absolute Temperature (PTAT) generator, produces an output voltage with a +2mV/°C temperature coefficient. By operating two transistors at different current densities, a PTAT voltage is obtained. The two voltages are applied to a summing circuit so that the two temperature drifts cancel to yield a Vref Out voltage with zero temperature drift. Due to the magic of semiconductor junctions, the temperature drift cancellation requires an output voltage that is equal to the Bandgap voltage of silicon extrapolated to 0º Kelvin; this is approximately 1.24V depending on the fabrication process. As can be seen in the graph in Figure 3, zero TC only occurs at one point on the curve due to the nonlinear relationship of semiconductor junctions. For additional reading on the theory of Bandgap voltage references, see Notes 5 and 6.

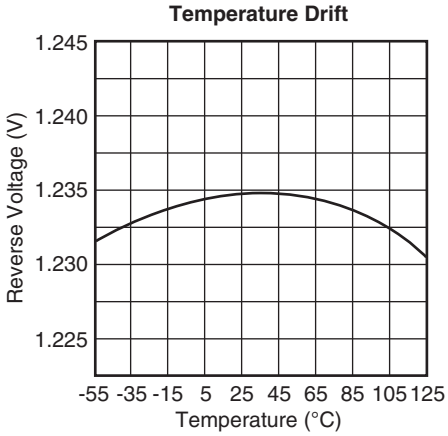



**FIGURE 3. (Note 2)**

It should be noted that this curve is highly idealized, and the curves taken from actual devices show a wide variation from unit to unit; this is shown in the graph in Figure 4 below for three typical IC voltage references with the same part number.

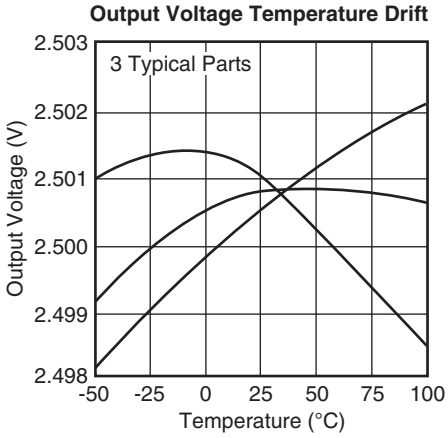



**FIGURE 4. (Note 3)**

A Bandgap reference shows the bow shaped curve as illustrated in Figure 3. The temperature drift can be lowered by adding a second order compensation term to achieve “curvature correction”. These curvature correction Bandgap references typically have an S-shaped TC curve as shown in Figure 5.

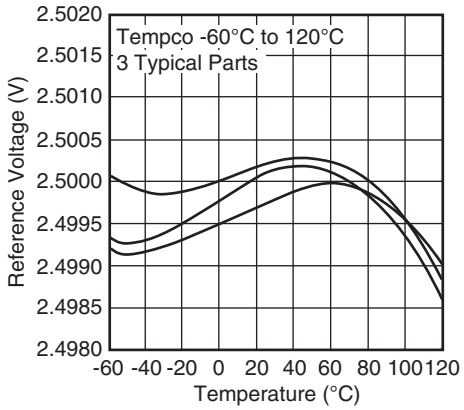



**FIGURE 5. (Note 3)**

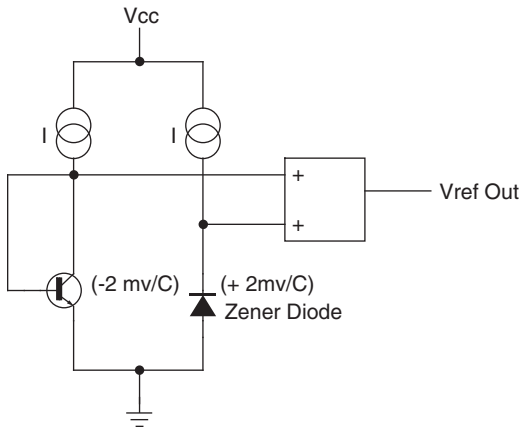



**FIGURE 6.**

### *XFET™ Voltage Reference*

XFET references are similar in principle to Bandgap references except they depend on Junction Field Effect Transistors (JFET) instead of bipolar transistors as used in the typical Bandgap reference. XFET references can offer lower noise and drift than Bandgap references while operating at lower supply current. For additional reading on the theory of XFET voltage references, see Note 7.

In summary, Bandgap references are usually used for voltage reference applications that require low reference voltage (<5V), low operating current (<1mA), medium temperature drift (>20ppm/C), and versatile series/shunt mode operation.

### *Buried Zener Voltage Reference*

Zener diodes have traditionally been used to make high quality voltage references. Prior to integrated circuits, discrete Zener diodes were used for voltage reference applications; “temperature compensated” Zener diodes such as 1N829 were used in discrete and hybrid circuits. When IC Zener diodes were first used they were very noisy and unstable due to surface contamination and crystal imperfections. It was found that by moving the Zener junction from the surface of the die to below the surface the noise and stability were greatly improved. These Zener diodes became known as “buried Zener diodes”, and have been the workhorse device for high quality voltage references.

Zener diodes in the 5 to 8 V range show a temperature drift that is approximately +2mV/°C. By combining a forward biased diode junction with a Zener diode, a voltage reference with zero TC could be obtained as shown in Figure 6.

However, there are lots of caveats that go along with this reference circuit:

1. The temperature drift of the two diodes is dependent on their bias current, and trimming is required to achieve low TC.
2. Often, the zero TC adjustment changes the absolute output voltage from its intended value so you can not obtain tight initial tolerance and low temperature drift at the same time.
3. The supply current is higher than a Bandgap reference since the Zener diode and forward biased diode are operated at a higher current to achieve low noise and zero temperature drift. Also, for maximum flexibility in setting the diode’s TC, two current sources are used to bias the diodes.
4. The supply voltage must be greater than the Zener diode voltage and the bias current source which makes the supply voltage greater than 7V.

In summary, buried Zener references are usually used for voltage reference applications that require low temperature drift and low noise. However, they require higher supply voltage (>7V), higher operating current (>1.5mA), and operate only in series mode unless external biasing is used.

### *Floating Gate Analog Technology Voltage Reference*

The Intersil Floating Gate Analog technology takes a radically different approach to make a high quality voltage reference. Instead of using the inherent temperature drift characteristics of transistors and diodes which are highly nonlinear, process dependent, and extremely inflexible, the Intersil Floating Gate Analog technology stores a precise voltage on Cstore, a floating gate. The floating gate voltage is buffered with a high quality CMOS amplifier as shown in the simplified diagram shown in Figure 7.

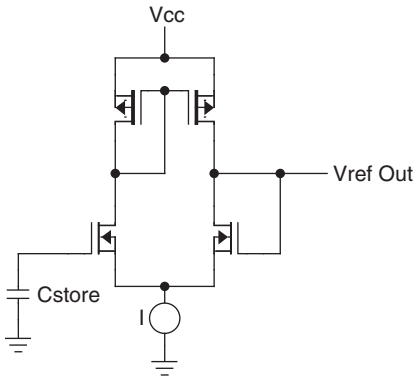



**FIGURE 7.**

The resulting voltage reference has excellent characteristics which are unique in the industry; very low temperature drift (1ppm/°C), high initial accuracy, and extremely low supply current (<1µA). Also, the reference voltage is not limited to “magic” voltages obtained from Bandgap references or buried Zener diodes to achieve temperature drift cancellation. In addition, there is no need for trimming via lasers, fusible links, or Zener zapping. Standard output voltage settings from 0.9000 to 5.000 are programmed as part of the standard manufacturing process as discussed in the following section.

To understand how the output voltage is programmed, refer to the simplified diagram shown in Figure 8:

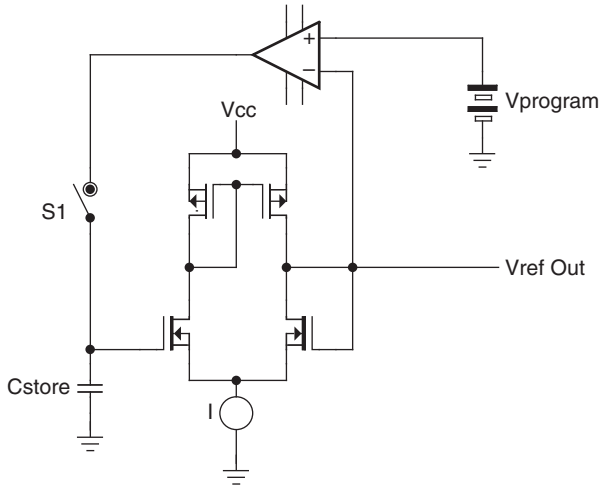



**FIGURE 8.**

During production testing, an external voltage, Vprogram, is applied to the device under test, and switch S1 is closed. A servo amplifier forces the Vprogram voltage onto the floating gate capacitor, Cstore. When the programming and test modes are complete, the programming voltage, Vprogram is removed, and switch S1 is opened leaving a charge on capacitor Cstore which is the desired reference voltage. Typically the trapped charge on the floating gate can remain without loss for greater than 10 years. As one might expect, the switch S1 is a very critical element and is highly simplified in this description. Switch S1 is really two tunnel diodes using a mechanism know as the Fowler-Nordheim Tunneling effect. For a complete description of the Intersil Floating Gate Analog technology, see Appendix A, “Precision Voltage Reference Using EEPROM Technology”, by Jim McCreary.

In summary, the Intersil Floating Gate Analog technology provides a voltage reference which has excellent characteristics which are unique in the industry; very low temperature drift (1ppm/°C), high initial accuracy (0.01%), and extremely low supply current (<1µA). Also, the reference voltage is not limited to “magic” voltages so it is possible to provide output voltage settings from 0.9000 to 5.000 that are programmed as part of the standard manufacturing process.

## Key Voltage Reference Specifications

### Absolute Initial Accuracy

1. **Absolute Initial Accuracy** defines the range of the reference’s output voltage with a defined input voltage, load current, and ambient temperature. The top grade voltage reference, X60008AIS8-50 has an output voltage range of 4.9995V to 5.0005V (5V ±.5mV or 5V ±.01%) with a 6.5V input, no load current, and 25°C operating temperature.

### Temperature Coefficient and Using the Box Method

2. **Temperature Coefficient (TC)** is a measure of the output voltage change with respect to changes in the operating temperature. The top grade voltage reference, X60008AIS8-50 has a TC of 1ppm/°C which makes it one of lowest temperature drift references in the industry. Older voltage references such as the LM399 or LTZ1000 achieve low TC by temperature stabilizing the device die with a heater. However, just the heater supply current (25ma) is typically 50,000 times higher than the X60008AIS8-50 supply current of .5µA!

   The limits stated for TC are governed by the method of measurement, and there are many ways to cheat when specifying TC in a voltage reference. For example, let’s consider the Temperature Drift curve shown in Figure 9 for a Bandgap reference. At -55°C the Reference Voltage is 1.232V and at +125°C the Reference Voltage is 1.231V. Therefore, the change in Reference Voltage is only 1mV for a change in temperature of 180°C or 5.5µv/°C; for a 1.235 V reference, this gives a TC of 4.4ppm/°C. Due to the parabolic shape of the TC curve, it is possible to calculate a TC of zero using this method.

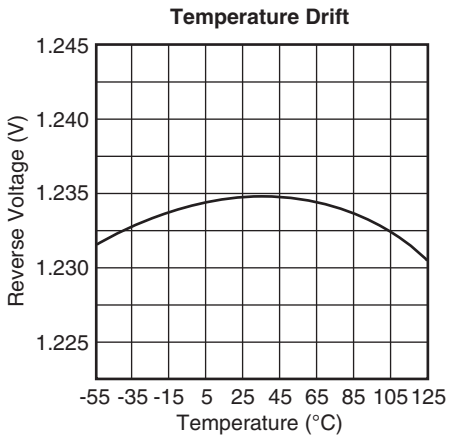



**FIGURE 9.**

The overwhelming standard for specifying the TC of a reference is the “Box Method” as shown in Figure 10.

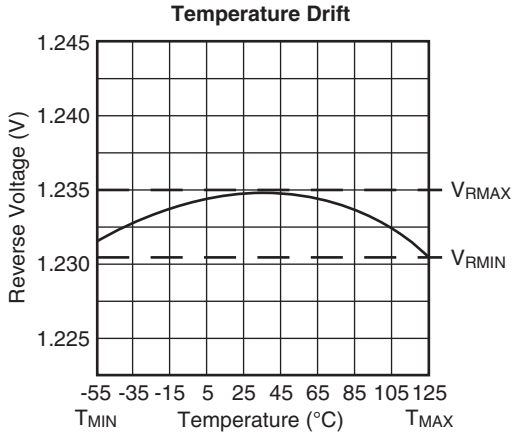



**FIGURE 10.**

In the Box Method, the reference voltage is measured throughout the temperature range from the minimum specified temperature (Tmin) to the maximum specified temperature (Tmax). The minimum reference voltage (Vrmin) and maximum reference voltage (Vrmax) is determined within the temperature range. If Vr is the reference output voltage at 25°C, the TC of the reference is calculated by:

$$TC = \frac{(Vrmax - Vrmin)/(Tmax - Tmin)}{Vr} \times 10^6 \text{ ppm/°C}$$

In the same example as above using the graph in Figure 9:

Tmin = -55°C
Tmax = +125°C
Vrmin = 1.231V
Vrmax = 1.235V
Vr = 1.235V

Notice the large difference between the first method of calculating TC which gives 4ppm/°C vs the standard Box Method which gives 18ppm/°C.

$$TC = \frac{(1.235 - 1.231)/(125 - (-55))}{1.235} \times 10^6 = 18 \text{ ppm/°C}$$

With both a Bandgap reference and buried Zener reference, the TC curve is a nonlinear relationship so the designer cannot infer a proportional TC relationship. Curvature corrected Bandgap references with the “s-shaped” curve may have a TC slope which exceeds the average specified TC by 2x or 3x. The TC characteristic of the Intersil X60008 is nearly a straight line with curvature of less than 0.5ppm/°C over the industrial temperature range of -40°C to +85°C. The combination of very low TC and a predictable TC slope is unique to the Intersil X60008 due to its floating gate topology. Figure 11 shows the flat slope TC curves for the X60008.

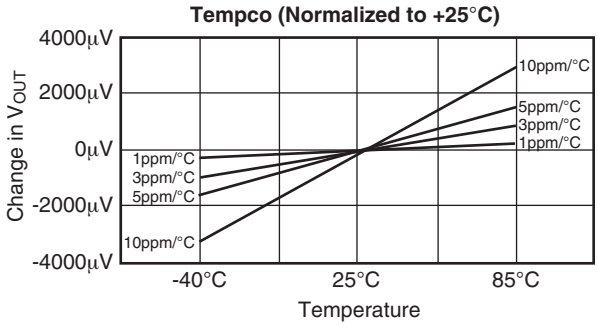



**FIGURE 11.**

In an A/D converter (ADC) or D/A converter (DAC) design the reference temperature drift is an error source in the full-scale accuracy. Over the full operating range, the total drift must be less than 0.5 LSB to maintain an accuracy consistent with the resolution of the ADC or DAC. The chart below shows the drift requirements for various system accuracies over a 0°C to +70°C temperature range.

| RESOLUTION (BITS) | 1/2 LSB FOR A 5 V FULL-SCALE (mV) | DRIFT REQUIRED (ppm/°C) |
|---|---|---|
| 8 | 9.77 | 27.90 |
| 10 | 2.44 | 6.98 |
| 12 | 0.61 | 1.74 |
| 14 | 0.15 | 0.44 |
| 16 | 0.04 | 0.11 |

### *Supply or Quiescent Current*

1. The **Supply Current** of a voltage reference is the current that is required to operate the voltage reference with no load current. In the case of the shunt reference, the supply current is the minimum current that must be allowed to flow into the device for proper operation. Generally, voltage references are designed with very low supply current to minimize self-heating effects which would degrade the TC of the device. The Intersil X60008 voltage reference is unique in the industry due to its incredibly low supply current of only 500nA - it is the first voltage reference to make possible continuous battery operation as discussed in the Applications section.

### *Output Noise; Wideband (10Hz-1kHz) and 0.1 - 10Hz*

2. **Voltage reference Output Noise** is generally specified as a peak-peak voltage in the 0.1 – 10Hz bandwidth which is useful for low frequency systems such as temperature measurement. Using a rms reading, output noise can also be specified in a higher frequency bandwidth such as 10Hz to 1kHz. Assuming the noise is truly random, the peak-peak noise can be estimated by multiplying the rms value by 6. For example, a voltage reference with 2μV, rms 10Hz to 1kHz noise will have a peak-peak noise of approximately 12μV.

   In reality, the best way to specify high frequency noise is to show a graph of noise voltage spectral density in nv/√Hz vs frequency. This allows the design engineer to calculate the reference noise based on the bandwidth of the system.

   The following equation can be used to calculate the required RMS noise voltage spectral density:

   $$En < Vref / (12 * 2^N * \sqrt{BW})$$

   Where En - Noise density (V/√Hz)
   Vref - Reference voltage
   N - Resolution
   BW - System bandwidth

   For example, for a 12-bit system with a 5V reference operating in an audio bandwidth of 100Hz to 20kHz:

   $$En < 5 / (12 * 2^{12} * \sqrt{(20kHz – 100Hz)}$$
   $$< 720 nv//\sqrt{Hz}$$

### *Noise Performance and Reduction*

The X60008 output noise voltage in a 0.1Hz to 10Hz bandwidth is typically 30 μVp-p. The noise measurement is made with a bandpass filter made of a 1 pole high-pass filter with a corner frequency at 0.1Hz and a 2-pole low-pass filter with a corner frequency at 12.6Hz to create a filter with a 9.9Hz bandwidth. Noise in the 10kHz to 1MHz bandwidth is approximately 400μVp-p with no capacitance on the output, as shown in Fig.12 below. These noise measurements are made with a 2 decade bandpass filter made of a 1 pole high-pass filter with a corner frequency at 1/10 of the center frequency and 1-pole low-pass filter with a corner frequency at 10 times the center frequency. Figure 12 also shows the noise in the 10kHz to 1MHz band can be reduced to about 50μVp-p using a 0.001μF capacitor on the output. Noise in the 1kHz to 100kHz band can be further reduced using a 0.1μF capacitor on the output, but noise in the 1Hz to 100Hz band increases due to instability of the very low power amplifier with a 0.1μF capacitance load. For load capacitances above 0.001μF, the noise reduction network shown in Figure 13 is recommended. This network reduces noise significantly over the full bandwidth. As shown in Figure 12, noise is reduced to less than 40μVp-p from 1Hz to 1MHz using this network with a 0.01μF capacitor and a 2kΩ resistor in series with a 10μF capacitor.

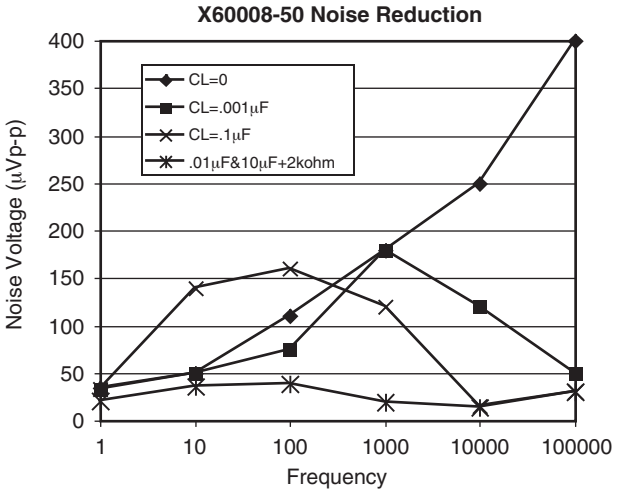



**FIGURE 12.**

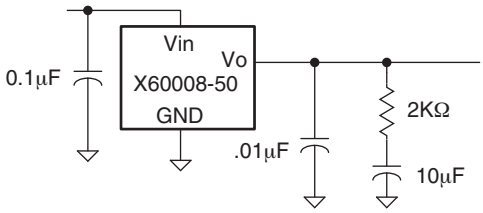



**FIGURE 13.**

Often external filtering may be required to reduce the reference noise to acceptable limits based on the system resolution and accuracy requirements. The filters for voltage references will be discussed further in the “Care and Feeding” section of this Application Note.

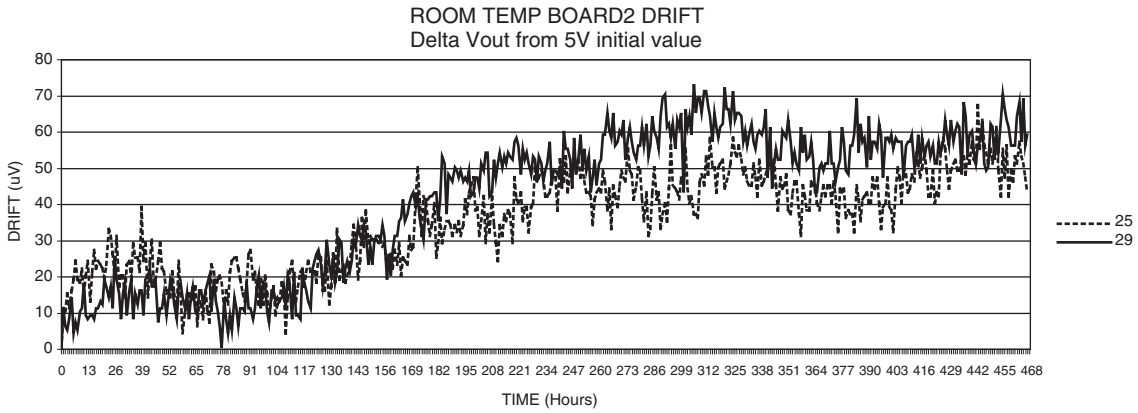



**FIGURE 14.**

### *Line Regulation*

1. **Line Regulation** specifies the amount the output voltage will change as the input voltage is varied over its limits. Line regulation is specified as μV/V or ppm/V. For example, the Intersil X60008 Line Regulation specification is 100μV/V, maximum; therefore, if the input voltage is changed by 2V from 6V to 8V, then the output voltage will change by 200μV maximum.

### *Load Regulation*

2. **Load Regulation** specifies the amount the output voltage will change as the load current is varied over its limits. Load regulation is specified as μV/mA or ppm/mA. Due to the output driver stage of a voltage reference, often the current sinking specification may be different than the output sourcing current specification. For example, the Intersil X60008 Line Regulation specification is 100μV/mA sinking current and 50μV/mA sourcing current.

### *Long Term Stability*

3. **Long Term Stability** tries to predict the amount the output voltage will change over an extended time period and is specified as ppm/1000 hours. The only accurate way to measure long term stability is to wait a very long time! There is no easy and accurate way to accelerate the long term stability of a voltage reference. Voltage reference vendors have tried to use elevated temperature to predict the effect of aging with the voltage reference. This technique leads to inaccurate and highly optimistic long term stability results as discussed in Reference Note 8.

   The long term stability for the X60008 is shown in Figure 14. Notice that after approximately 10 days (240 hours) of initial power-up drift, the output voltage variations stabilize at about 20μV peak-peak or 4ppm from the +5V nominal output voltage. Ongoing tests will provide further data on the long term stability for the X60008.

### *Thermal Hysteresis*

1. **Thermal Hysteresis** specifies the maximum change in room temperature output voltage after the voltage reference is cycled between two extreme temperature. For example, the Intersil X60008 is temperature cycled from 25°C to -40°C and returned to 25°C. The output voltage is measured and recorded. Then the temperature is cycled to +85°C and returned to 25°C. The output voltage is then measured and recorded again. The deviation between the first 25°C reading and the second 25°C reading is the thermal hysteresis expressed in ppm.

   Thermal hysteresis is a direct result of the device package with the smaller plastic SO packages exhibiting higher thermal hysteresis than the older TO39 metal can packages. For further discussion, see Reference Note 8.

### *Input Voltage Range*

1. **Input voltage range** specifies the range of the input voltage for proper device operation and meeting the device electrical specifications. In addition, there is an Absolute Maximum Rating which must not be exceeded to avoid damaging the device.

### *Dropout Voltage*

2. Dropout Voltage specifies the minimum differential voltage between the input voltage and reference output voltage to maintain the specified accuracy. Many high precision voltage references require 1V or greater of input voltage to output voltage differential. The Intersil X60008 dropout voltage is typically less than 150mV with an output current of 5mA.

## Advantages of Intersil X60008 Voltage Reference

The Intersil X60008 series of voltage references use the Floating Gate Analog (FGA) technology to create references with extremely tight initial tolerance, very low temperature drift, and ultra low supply current. As discussed earlier in this Application Note, the charge stored on a floating gate cell is set precisely in manufacturing. The reference output voltage is a buffered version of the floating gate voltage. The resulting reference device has excellent characteristics which are unique in the industry:

### Absolute Accuracy

1. **The Output Voltage Accuracy** is extremely tight at ±500µV or ±0.01% for a +5V output. The tight initial accuracy is set as part of the final manufacturing process and does not depend on traditional trimming techniques such as laser trims or fusible links.

### Low Temperature Drift

2. **The Temperature Drift** is extremely low at 1ppm/°C and is a well behaved monotonic slope with no second or third order inflections as shown by Bandgap references. The low temperature drift does not depend on traditional trimming techniques such as laser trims or fusible links.

### Low Power

3. **The Supply Current** achieves nanopower levels due to the FGA and CMOS technology. At room temperatures, the supply current is typically 500nA which is 1 to 3 orders of magnitude lower than competitive voltage references.

### Ability to Set any Output Voltage

4. **The Ability to set any Output Voltage** is certainly unique to the X60008 voltage reference. The reference voltage is not limited to magic voltages as required by Bandgap voltage or Zener diode voltages. Output voltages in the range of 0.9000V to 5.0000V can easily and accurately be set in manufacturing.

The process used in these voltage references is a floating gate CMOS process, and the buffer amplifier circuitry uses CMOS transistors for the amplifier and output stage. While providing excellent accuracy, low TC, and low power, there are limitations in output noise level due to the MOS device characteristics. The limitations are addressed with circuit techniques discussed in this Application Note.

## Care and Feeding of Voltage References

To many, voltage references would seem extremely easy to apply – after all, they only have three pins …. an input voltage, an output voltage, and a ground pin. Some references may have additional pins for trim, temperature output, or noise cancellation, but the X60008 series of references only have three pins. However, unless one is very careful with the details, things can go wrong very fast! Nothing should be left to chance in a high accuracy system.

Two schematics shown below for the same circuit illustrate good and bad design practices.

The Bad Schematic shown in Figure 15 is what is typically drawn using today's CAD software which has following issues:

1. There is no grounding scheme shown with the use of only one ground symbol.
2. There is no hint to where the grounds or critical connections should be tied together, and as shown at J1, the grounds could introduce digital noise into the analog circuits.
3. Decoupling capacitors are shown in a row with no indication about which component they are associated.
4. There is no flow to the schematic because the circuit blocks are not interconnected, and there is no left to right flow. While this schematic is shown on one page, Sometimes schematics of the circuit blocks are grouped on unrelated pages. The worse case was a schematic where each component was shown separately with only Net List names indicating connections!
5. The CAD Library symbol for the A/D converter makes no sense from a circuit function viewpoint since analog and digital functions are mixed together.

The Good Schematic is shown in Figure 16; the connections are identical to the Bad Schematic, but it is redrawn to illustrate proper connections and PCB routing.

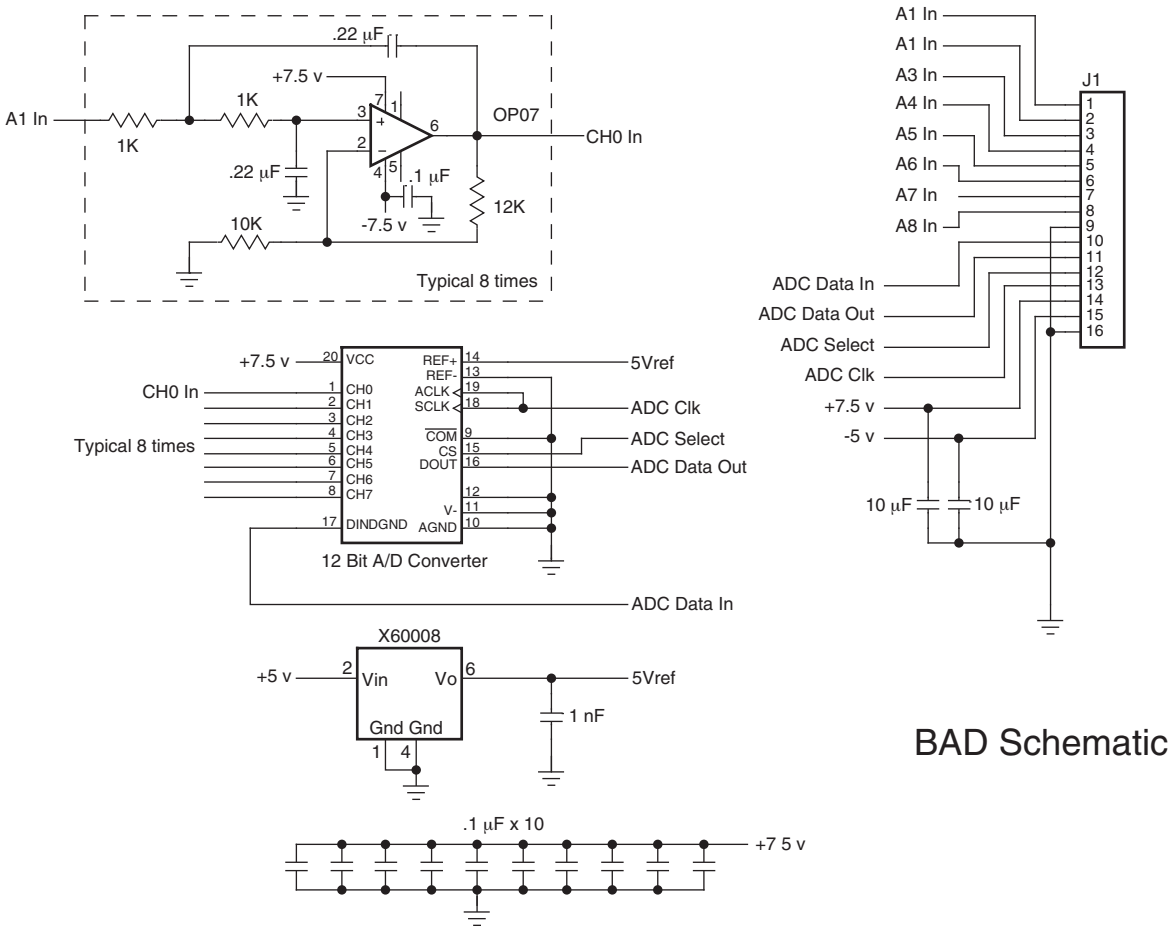



**FIGURE 15.**

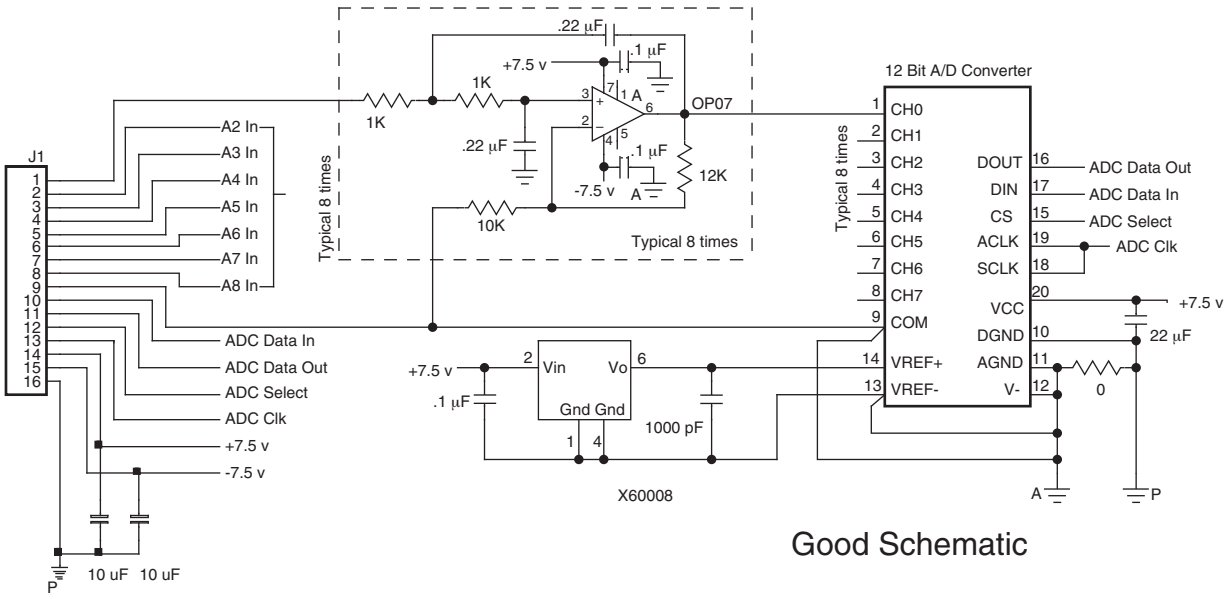



**FIGURE 16.**

1. The schematic flows from left to right; the input connector J1 is on the left, and the logic outputs are on the right side of the schematic.
2. A ground scheme is shown using a Power (or digital) Ground plane and an Analog Ground plane. The two ground planes are clearly designated with a “P” and “A”.
3. To be sure there is no digital noise introduced into the Analog Ground, the two grounds are tied together at only one point at the A/D converter. Furthermore, a 0Ω resistor is used to connect the two grounds; this ensures a separate Net for each ground so the PCB layout software or layout person does not arbitrarily connect the two grounds. The use of a 0Ω resistor is cheap insurance against a noisy and inaccurate analog system!
4. The Analog Ground pin (pin 9) is shown to connect directly to the COM pin of the A/D converter. Likewise, the X60008 Reference Ground pins (pins 1and 4) are shown to connect directly to the REF- pin of the A/D converter. There are no additional DC or AC currents in these lines which could cause DC errors or excessive noise.
5. Notice the use of diagonal lines on the COM pin and REF- pin of the A/D converter; these indicate a connection directly at the pin – not at a convenient point on the ground plane, but right at the pin.
6. The standard Library part for the A/D converter was not used in this schematic. Instead, a new library part was created to allow proper signal flow and connections. It should be noted that creating this new library part took several iterations and additional time to get it right, but it was worth the effort to create proper documentation.
7. While not apparent on this schematic, a design should never be based on ease of schematic entry!

### *Grounding and IR drops*

As discussed earlier in this application note, voltage references are used to provide a very precise voltage for measurements to be made against. The accuracy of any measurement is only as good as the ability to compare it against a known standard. If the PCB connections are made incorrectly, the most perfect voltage reference can still have errors resulting from poor grounding considerations and not understanding the impact of Ohm’s Law. Any analog or mixed signal PCB must have a well thought-out grounding scheme with multiple ground planes or traces. There must be no heavy DC current or AC current in the analog ground planes that connect the voltage reference to the system measurement point.

The Good Schematic shown in Figure 16 is an example of a proper voltage reference connection for a data acquisition system where the voltage reference ground pin is connected directly to the measurement point on the A/D converter.

Shown below in Figure 17 is a 10 Bit Adjustable, 0-5 Amp Active Load circuit that is useful for testing power supplies and DC/DC converters.

The details of this circuit will be discussed in the applications section, but notice the way that the voltage reference and digital pot connections are shown. To avoid errors caused by IR drops, the connections must be made directly at the leads of the .05Ω current sense resistor. Just 10mΩ of contact resistance or PCB trace resistance will cause a 20% error in the current setting as illustrated in Figure 18.

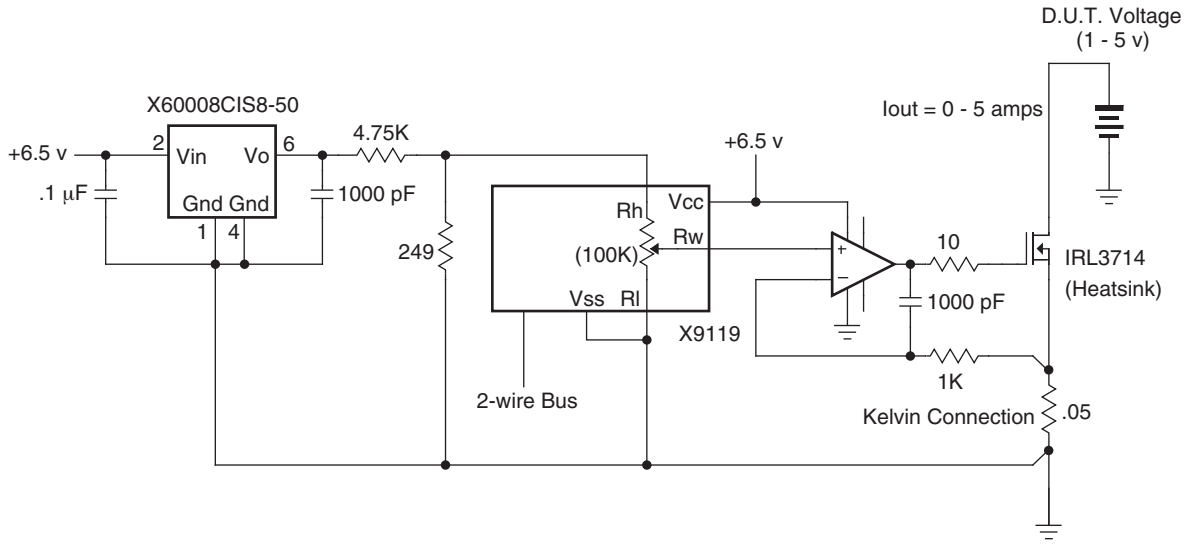



**FIGURE 17.**

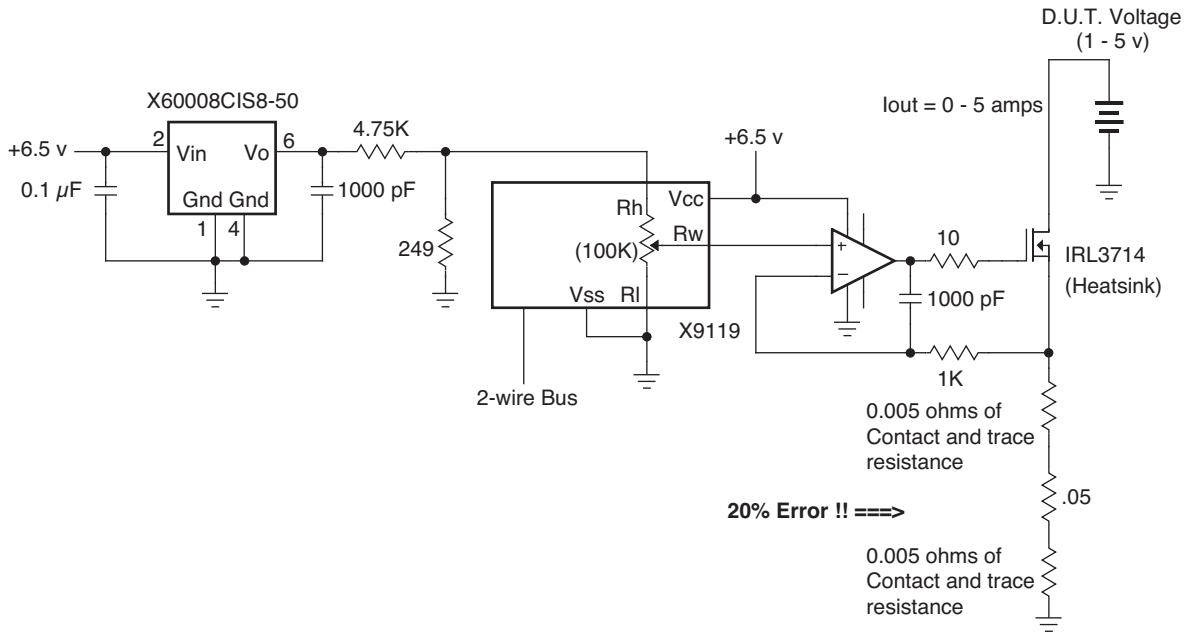



**FIGURE 18.**

Due to the outstanding initial accuracy of the X60008AIS8-50 of ±500µV, it takes very little PCB trace resistance to introduce errors that exceed the specifications of the X60008AIS8-50. The chart below shows the maximum PCB trace resistance at a given load current to maintain the X60008 accuracy. The chart is based on using 1oz. copper which has a typical sheet resistance of 0.5mΩ /square. The resistance of the PCB trace can be calculated by:

R = 0.5 mΩ/square * Length / Width

| LOAD CURRENT | MAXIMUM TRACE RESISTANCE | MAXIMUM LENGTH OF 20 MIL TRACE |
|---|---|---|
| (mA) | (mΩ) | (inches) |
| 10 | 50 | 2 |
| 50 | 10 | 0.4 |
| 100 | 5 | 0.2 |
| 500 | 1 | 0.04 |
| 1 amp | 0.5 | 0.02 |

Remote sensing as shown in Figure 19 is required if high accuracy at high load currents is required. Notice that remote sensing is employed on both the +5V Ref side of the 50Ω load and on the ground side of the load.

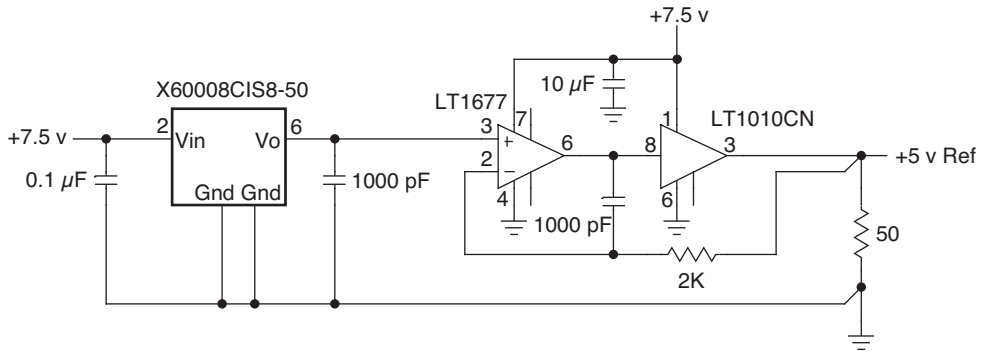



**FIGURE 19.**

### *PCB Mounting and Location on PCB*

For applications requiring the highest accuracy, the board mounting location of the voltage reference is critical. Placing the X60008 voltage reference in areas of the PCB subject to twisting can cause degradation in accuracy of the reference voltage due to die stresses. It is normally best to place the device near the edge of the board or the shortest side. The axis of bending is most limited in these locations.

The following techniques can be used to improve the accuracy:

1. Mechanically restraining the PCB with mounting screws and grommets in each corner of the PCB
2. Thicker boards and not using thin or flexprint PCB
3. Sots in the board around the voltage reference
4. Avoid trapping adhesives and solder flux under the package

### *Soldering Care*

The incredibly tight output voltage (±500µV) means that great care must be taken in the soldering process to avoid stress on the die. The key to soldering the parts onto a PCB is to not do it by hand one pin at a time with a high temperature soldering iron. Having one pin at high temperature while other pins are at low temperature puts very high uneven stress on the device and causes large (many mV) shifts in the X60008 output voltage.

The best way to solder the X60008 voltage reference is to IR solder it onto the board with the lowest temperature possible; +220 °C or below is recommended. Also, only put the X60008 voltage reference through the IR reflow process once, not multiple times, to minimize output voltage shift due to package stress during IR reflow. Therefore, on a PCB that uses components mounted to both the top side and backside of the board, make sure the X60008 soldering operation is the second run through the IR reflow process.

### *DNC Pins*

DNC or NC pins are truly **D**o **N**ot **C**onnect pins! On most high accuracy voltage references, post-package trimming is achieved with the use of these DNC pins. On some voltage references, the DNC pins are fusible links to trim the output voltage.

Incoming inspection or in-circuit PCB level tests that force voltages and currents into pins must be avoided at the DNC pins so as not to alter set voltages or damage the device due to electrical over-stress.

### *Voltage Reference Output Filters*

Output filters can be used to reduce the voltage output noise of the X60008; however, much care must be taken since it is very easy to “lose” 0.5mV in the filter stage. Testing the output noise of a high quality reference such as the X60008 is often a challenge since to test for 0.1 to 10Hz noise from a voltage reference, it is necessary to AC couple the output of the reference before it is applied to a high gain stage as shown in Figure 20.

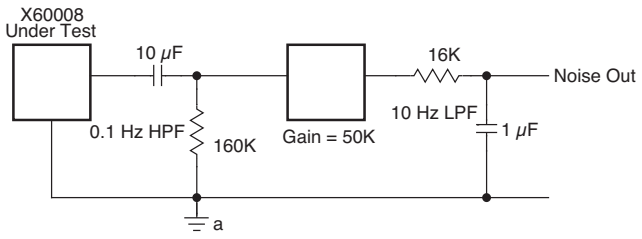



**FIGURE 20.**

Additionally, the high gain stage in a typical noise test circuit requires a very low voltage noise op amp; however, low voltage noise op amps exhibit high current noise which prevent the use of high resistance values in the 0.1Hz AC coupling filter. Appendix B describes a novel voltage reference test circuit which eliminates the need for a 0.1Hz AC and includes a peak to peak voltage detector. Also in Appendix B test data is shown for various output filter circuits.

To effectively attenuate noise in the 0.1Hz - 10Hz bandwidth, it is necessary to use large value capacitors and/or large value resistors. For example, for a one pole low pass filter with a 0.1Hz corner frequency it is necessary to use resistor and capacitor values as shown below:

| RESISTOR (kΩ) | CAPACITOR (µF) |
|---|---|
| 160 | 10 |
| 16 | 100 |
| 1.6 | 1000 |

While it seems attractive to use a 1000µF capacitor and 1.6kΩ resistor, the large leakage current of an electrolytic capacitor will generate an error voltage across the 1.6kΩ resistor. For example, the leakage current of a typical 1000µF capacitor is 100µA (Panasonic VS series); an error voltage of 100µA * 1600Ω = 160mV is created! Reducing the capacitor value and increasing the resistor value does not help because while the leakage current is reduced, the resistor value is increased! For example, for a 100µF capacitor in the same Panasonic VS series, the leakage current is reduced to 10µA, but the resistor value is increased to 16kΩ. The same 160mV of error is introduced!

Fortunately there is a solution obtained by bootstrapping two capacitors as shown in the filter circuit in Figure 21. The bootstrapping arrangement lowers the applied voltage across the capacitors and drops the capacitor leakage current to acceptable levels. This simple filter circuit reduces the 0.1Hz to 10Hz noise by at least 50%, and attenuates the higher frequency reference noise by the characteristic of a single pole low pass filter with a 0.1Hz corner frequency.

When designing noise filters for the X60008 voltage reference remember:

1. Beware of capacitor leakage current working against high value resistors which generate an error voltage.
2. Use a low offset voltage and low bias current op amp for the buffer amplifier.
3. Low frequency filters require a long time to settle to high accuracy levels; for example ten time constants are required for 0.01% settling time. A 0.1Hz filter requires 15 seconds to settle to 0.01%!

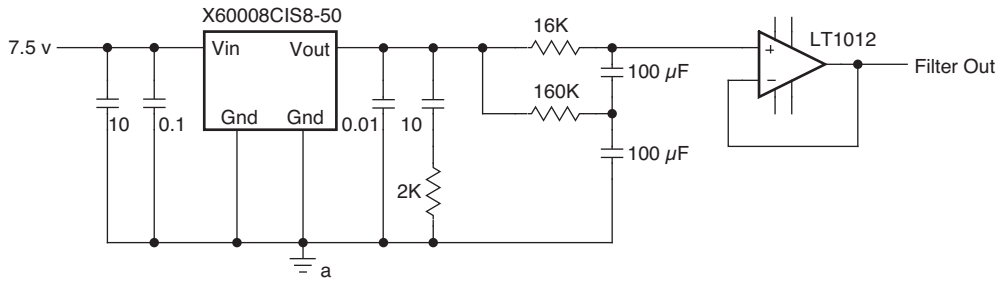



**FIGURE 21.**

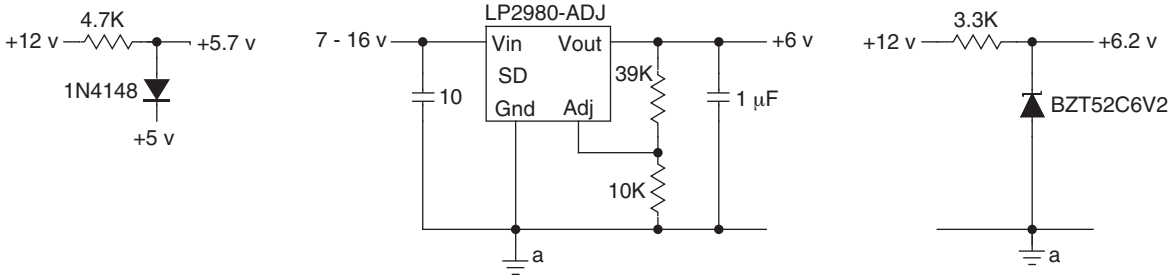



**FIGURE 22.**

### *Input Voltage Regulators*

If the input voltage comes from an unregulated source, it may be necessary to add a linear regulator before the X60008 to maintain rated accuracy of ±0.5mV. The Line Regulation specification for the X60008 is 100µv/V; if the input voltage ranges from 5V (minimum value) to +10V (maximum value), the output voltage could have an error of 500µV or 0.5mV. Any simple linear regulator can be used to stabilize the input voltage to acceptable levels. Several simple regulator circuits are shown in Figure 22.

It is also possible to operate the X60008 on either 3.3V or 5V if they are the only supply voltages in the system. Simple capacitor based charge pump circuits can be designed which use any digital clock in the system. The circuit shown below in Figure 23a allows operation in a 3.3V system; notice the additional resistor and capacitor to attenuate the switching noise and ripple. Figure 23b shows operating the X60008 from a single +5V supply. The 3.3V charge pump circuit uses Schottky diodes to minimize the diode losses to generate 6V from a 3.3V supply. The 5V charge pump circuit uses two silicon diodes to maximize the diode losses so the +10V maximum operating voltage specification of the X60008 is not exceeded.

Due to the low operating power of the X60008, it is also possible to “float” the operating voltage for the reference on virtually any voltage by simple charge pump circuits as shown above.

Operating on a negative supply or operating below a positive supply is also possible as shown in Figure 24. Operating below a positive supply can be useful in ground sourcing current sources as shown in the Applications Section.

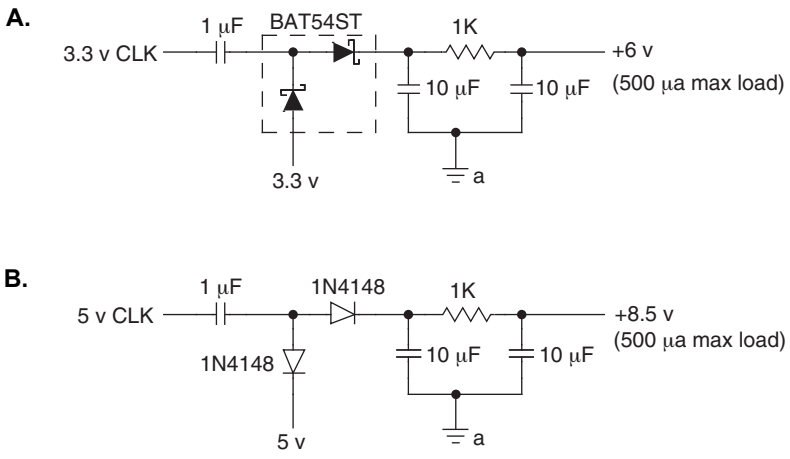



**FIGURE 23.**

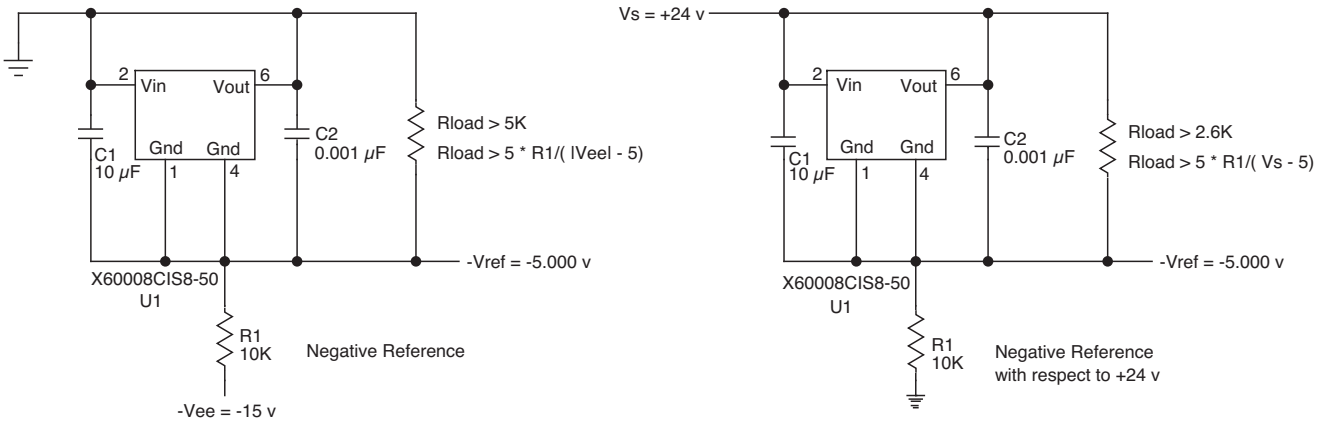



**FIGURE 24.**

### *Thermocouple Effects*

At any point in a circuit where dissimilar metals come in contact a small thermocouple voltage is developed. Fortunately, the copper lead frame of a surface mount device is the same copper material as PCB etch, and the thermocouple effect is minimized. However, there are many other places where thermocouples can be generated; for example, across a connector finger, across relay contacts, or even across a resistor! Yes, a poorly constructed resistor can show many µV/°C of thermocouple voltage. It has been found that external components (resistors, contacts, sockets, etc.) can create thermocouple voltages that exceed 10µV/°C. The top grade voltage reference, X60008AIS8-50, has a maximum TC of only 1ppm/°C which is a voltage change of only 5µV/°C. Therefore, without proper care, passive components can easily create errors that exceed the TC of the X60008AIS8-50.

It must be recognized that thermocouple voltages are developed by the difference in temperature between the two ends of dissimilar metal junctions, and not the absolute ambient temperature. If both ends of the metal junctions are isothermal (i.e., at the same temperature) there is no thermocouple voltage developed. Therefore, the first rule to avoid thermocouple effects is to eliminate hot spots on a PCB (i.e. linear voltage regulators). If hot spots cannot be avoided, then the two ends of metal junctions must be oriented so they are on isothermal lines on the PCB.

The second rule to minimize thermocouple effects is to balance the number of junctions in a loop so that the error voltages are canceled or become a common mode voltage that is reduced by the CMRR of the op amps in the signal chain. If the number of junctions are not balanced, then it may be necessary to create a junction by adding a series resistor that has no effect on circuit operation but balances the number of junctions.

## *Applications*

### *Auto-Calibrated 12 Bit Data Acquisition System*

The design of a true 12 bit data acquisition is extremely difficult due to the inaccuracies of the various components in the signal chain. Figure 25 shows a block diagram for a typical 8 input channel, 5V Full Scale, 12 bit data acquisition system; a quick error budget analysis for this system is shown below assuming 0 to 70°C operation.

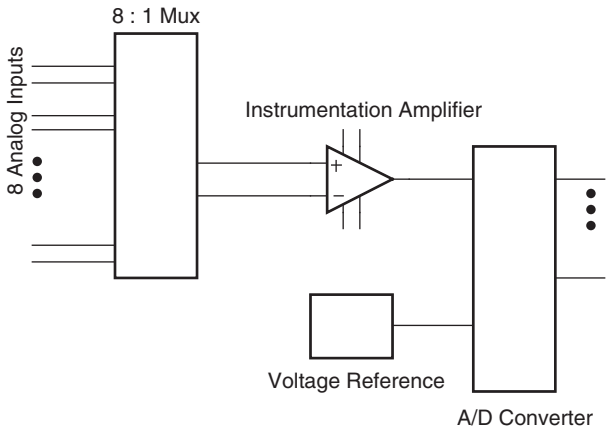



**FIGURE 25.**

| QUICK ERROR BUDGET ANALYSIS FOR FIGURE 25 | | |
|---|---|---|
| **COMPONENT** | **GAIN ERROR** | **OFFSET ERROR** |
| 8:1 Mux | 0 | 0 |
| Instrumentation Amp | ±0.03% | ±0.36mV ≥ ±0.007% |
| Instrumentation Amp TC | 50ppm/°C ≥ ±0.18% | 4.4µv/°C ≥ ±0.006% |
| Instrumentation Amp Gain Set | 0.1% | 0 |
| Instrumentation Amp Gain Set TC | 10ppm/°C ≥ ±0.04% | 0 |
| ADC + Voltage Reference | ±15 LSB ≥ ±0.4% | ±6 LSB ≥ ±0.15% |
| ADC + Voltage Reference TC | 45ppm/°C ≥ ±0.16% | |
| SubTotal of Errors | ±0.91% | ±0.16% |
| Total of Errors | | ±1.07% |

As one can see, for a simple 12 bit Data Acquisition System using a typical architecture with high performance Integrated Circuits the accuracy is a pretty miserable ±1%! And, this is not a comprehensive error analysis; for example, error for the MUX and the effects of linearity (INL and DNL) are not included.

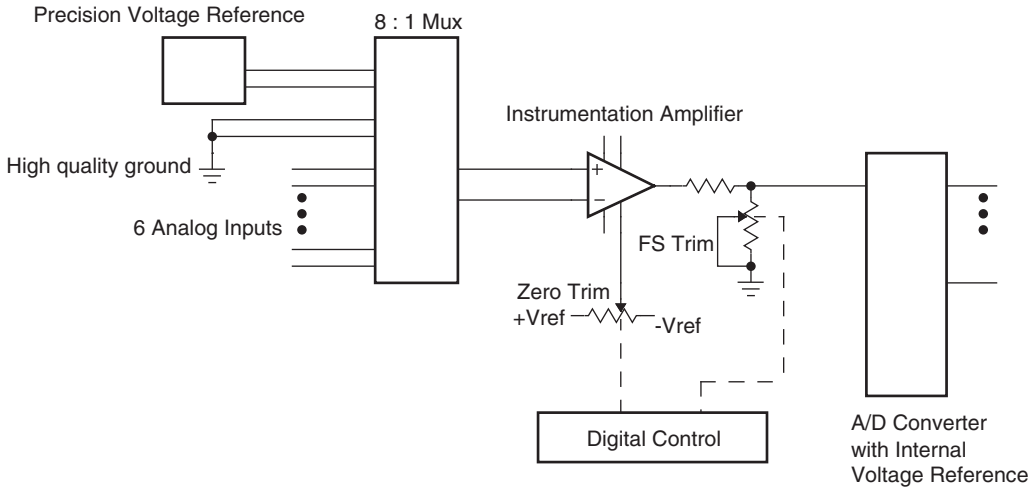



**FIGURE 26.**

The Integrated Circuit vendor for the A/D converter sidesteps this issue by including a calibration procedure in the data sheet for the device. Typical application circuits show adjustment pots for zero trim (offset) and full-scale trim (gain); the trims must be done at production test which does not take into account the temperature drift of the various components shown above. Of course, the adjustment pots can be replaced with digital pots such as the Intersil X9271 so the production test process can be automated. Zero or offset errors are easily tweaked out because with care there is a high quality zero volt source available; i.e., ground! It is more difficult to calibrate full scale voltage because a high quality source of full scale voltage is required; this is usually a calibration standard in the production test equipment. However, the temperature drift inaccuracies are not calibrated out once the system leaves the factory floor. Ideally, both the zero and full scale error would be calibrated before every high accuracy conversion or on a very regular basis that is more often than changes in the ambient temperature.

Figure 26 shows a revised architecture for the Auto Calibrated Data Acquisition System shown in Figure 25.

Two of the analog inputs are used as calibration source inputs for full scale adjust with a precision voltage reference and zero adjust with a high quality ground. Digital potentiometers (DCPSs) controlled by a microprocessor are used for zero trim and full scale trim. If non-volatile DCPS are used, the Auto Calibrated Data Acquisition System will maintain calibration even with the power removed. Calibration can be performed as often as required – before every high precision measurement, once a day, once an hour, etc. The user decides how often calibration is required.

The secret of this calibration scheme is the Precision Voltage Source since all the system full scale errors are referred to this device. The Intersil X60008AIS8-50 is an excellent choice for the reference since the initial accuracy is less than 0.4LSB for a +5V full scale range. Temperature drift of less than 0.3 LSB is maintained over a temperature range of 0°C to +70°C due to the low temperature coefficient of 1ppm/°C. Relatively inaccurate components can be used in the signal chain because their errors are calibrated out by the auto-calibration process.

The complete implementation of the Auto Calibrated Data Acquisition System is shown in the detailed schematic of Figure 27.

Two 4:1 differential MUXs (U1, U2) are used for 6 differential analog inputs. A high precision +5.000 V calibration source (X60008AIS8-50) is applied to S1A and S1B of U1 for full scale calibration. S2A and S2B of U1 are connected to a bias voltage set at +0.6 mV which is ∫ LSB of a 5V input range; this is used for zero calibration. The remaining 6 channels of U1 and U2 are used for the analog inputs. Additional analog MUXs could be added for more input channels. Likewise, if 8 analog inputs are required, a quad analog switch could be added to the DA and DB outputs with the calibration sources (+5V and Gnd) applied to their inputs.

The +5.000 V calibration source is obtained from a Intersil X60008AIS8-50 voltage reference. Input power for the X60008AIS8-50 is obtained by generating a +5.6V source with diode D1 and bias resistor R1. An output noise filter consists of C3, C4, and R2. The +0.6 mV bias voltage for zero calibration is obtained with R3 and R4. A -5V reference for the zero adjustment circuit is generated with an inverting op amp (U5B), R5, and R6.

An Instrumentation Amplifier (U4) converts the differential outputs from the MUX to a single ended output with a gain of 2.034. The gain of the IA is set by R9 such that:

$$\text{Gain} = (49.4k/R9) + 1$$

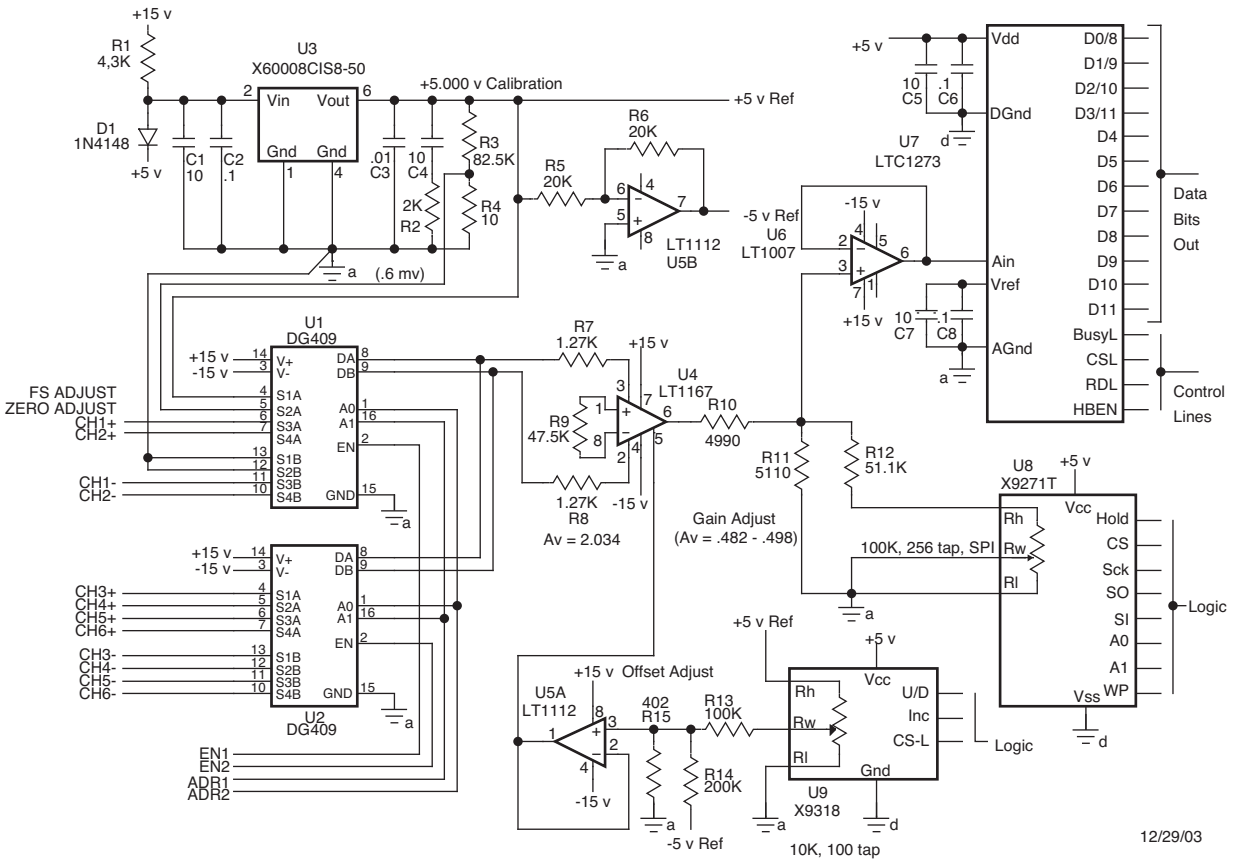



**FIGURE 27.**

The gain is set slightly higher than 2 so an adjustable 0.5 voltage divider can be used passively vary the overall gain by 1 ±1% to account for all the errors in the signal chain. Resistors R10, R11, R12, and D-Pot U8 are an adjustable voltage divider with a gain of 0.482 to 0.498; this allows the overall gain to be adjusted from 0.98 to 1.02 under the control of the D-Pot U8. Since the resistors are in the overall full scale feedback calibration loop, their tolerance or TC are not critical as long as there is adequate adjustment range.

The Ref pin of U4 is used for zero adjustments since there is a 1:1 correspondence between the Ref pin voltage and output voltage. Resistors R13, R14, R15, and D-Pot U9 provide an adjustable voltage of -10mV to +10mV under the control of U9. Since the Ref pin of U4 should driven by a low impedance to maintain the high common mode rejection ratio, op amp U5B is to buffer the ±10mV source.

The output of the adjustable 0.5 voltage divider is buffered with U6 since the A/D converter, U7, requires a low impedance and fast settling driver. The A/D converter is a 12 bit, 300Ks analog to digital converter with 12 bit parallel data outputs.

The calibration scheme is simple.

The zero calibration must be done first by selecting the +0.6 mV bias voltage on S2 of U1. Adjustment D-Pot U9 is incremented or decremented until there is an even flicker of output codes 0000 0000 $0000_b$ and 0000 0000 $0001_b$ (i.e., the LSB flickers evenly) from the A/D converter.

The full scale calibration is done second by selecting the +5.000V calibration source, X60008AIS8-50 on S1 or U1. Adjustment D-Pot U8 is varied until there is an even flicker of output codes 1111 1111 $1110_b$ and 1111 1111 $1111_b$ (i.e., the MSB flickers evenly) from the A/D converter. To the purist, the analog input voltage should be set for 5V – 1.5LSB = 4.99817 V for full scale calibration. However, to maintain the accuracy of the X60008AIS8-50, it was decided to accept a full scale input voltage of +5.000V + 0.5LSB = +5.0006V.

It must also be noted that the +5.000 V calibration source, X60008AIS8-50 and zero source could also be applied to signal conditioning before the MUX stage thus calibrating out any error in the signal conditioning circuitry. The auto-calibration scheme must be considered a closed feedback loop and any errors inside the feedback loop are reduced by the gain (i.e., adjustment range) of the feedback loop.

### *Auto-Calibrated 12 Bit Digital to Analog Converter*

The design of a true 12 bit Digital to Analog converter, like the design of a true 12 bit data acquisition system, is extremely difficult due to the inaccuracies of the various components in the signal chain. Figure 28 shows a block diagram for a typical 5V Full Scale, 12-bit Digital to Analog converter; a quick error budget analysis for this system is shown below assuming 0-70°C operation.

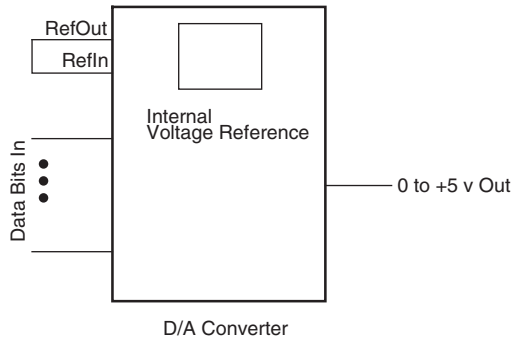



**FIGURE 28.**

| QUICK ERROR BUDGET ANALYSIS FOR FIGURE 28 | | |
|---|---|---|
| **COMPONENT** | **GAIN ERROR** | **OFFSET ERROR** |
| D/A Converter + Reference | ±0.2% | ±2LSB = ±.05% |
| D/A Converter + Reference TC | ±30ppm/°C ≥ ±0.15% | ±3 ppm/°C ≥ ±0.015% |
| SubTotal of Errors | ±0.35% | ±0.065% |
| Total of Errors | | ±0.42% |

As you can see, for a simple off the shelf high performance Integrated Circuit 12 bit D/A converter using a typical architecture, the accuracy of ±0.42% is nowhere near 12 bit resolution of ±0.02%.

Like the previous example for a 12 bit A/D converter, the Integrated Circuit vendor for the D/A converter (different manufacturer!) sidesteps this issue by including a calibration procedure in the data sheet for the device. Typical application circuits show adjustment pots for zero trim (offset) and full-scale trim (gain); the trims must be done at production test which does not take into account the temperature drift of the various components shown above. Of course, the adjustment pots can be replaced with digital pots such as the Intersil X9271 so the production test process can be automated. However, high precision voltage measurements must be made for both zero volts for offset errors and +5V for full scale errors which require calibration measurement standards in the production test equipment. However, the temperature drift inaccuracies are not calibrated out once the system leaves the factory floor. Ideally, both the zero and full scale error would be calibrated on a very regular basis that is more often than changes in the ambient temperature.

Figure 29 shows a revised architecture for the Auto Calibrated Digital to Analog Converter shown in Figure 28.

The output voltage from the D/A converter is compared against a high quality ground for zero trim and a high accuracy precision voltage for full scale trim. Digital potentiometers (DCPs) controlled by a microprocessor are used for zero trim and full scale trim. If non-volatile DCPs are used, the Auto Calibrated Digital to Analog Converter will maintain calibration even with the power removed. Calibration can be performed as often as required – before every high precision measurement, once a day, once an hour, etc. The user decides how often calibration is required.

The secret of this calibration scheme is the Precision Voltage Source since all the system full scale errors are referred to this device. The Intersil X60008AIS8-50 is an excellent choice for the reference since the initial accuracy is less than 0.5 LSB for a +5.0V full scale range. Temperature drift of less than 0.3 LSB is maintained over a temperature range of 0°C to +70°C due to the low temperature coefficient of 1ppm/°C. Relatively inaccurate components can be used in the signal chain because their errors are calibrated out by the auto-calibration process.

The complete implementation of the Auto Calibrated Digital to Analog Converter is shown in the detailed schematic of Figure 30.

U1 is a 12 bit parallel input D/A converter with an internal 10V reference; laser trimmed resistors on the 20V Span and 10V Span pins allow the user to set many output voltage ranges. A Bipolar Offset pin provides the ability to generate bipolar output ranges such as ±5V, or this pin can be used for zero or offset adjustments. For a complete description of the pin settings, refer to the AD767 data sheet. While the AD767 has an internal voltage reference and factory trimmed span resistors, the overall error is ±0.4% as illustrated in the Error Budget shown above. Figure 30 shows an auto-calibration technique to eliminate zero and full scale error sources. The D/A converter output voltage is compared against a +5.000V calibration voltage, X60008AIS8-50 and a high quality ground for full scale and gain adjustments via digital potentiometers U7 and U8.

The +5.000V calibration source is obtained from a Intersil X60008AIS8-50 voltage reference. Input power for the X60008AIS8-50 is obtained by generating a +5.6V source with diode D2 and bias resistor R8. An output noise filter consists of C5, C6, and R9. A +0.6 mV bias voltage for zero calibration is obtained with R10 and R11. A -5V reference for the zero adjustment circuit is generated with an inverting op amp (U4A), R12, and R13.

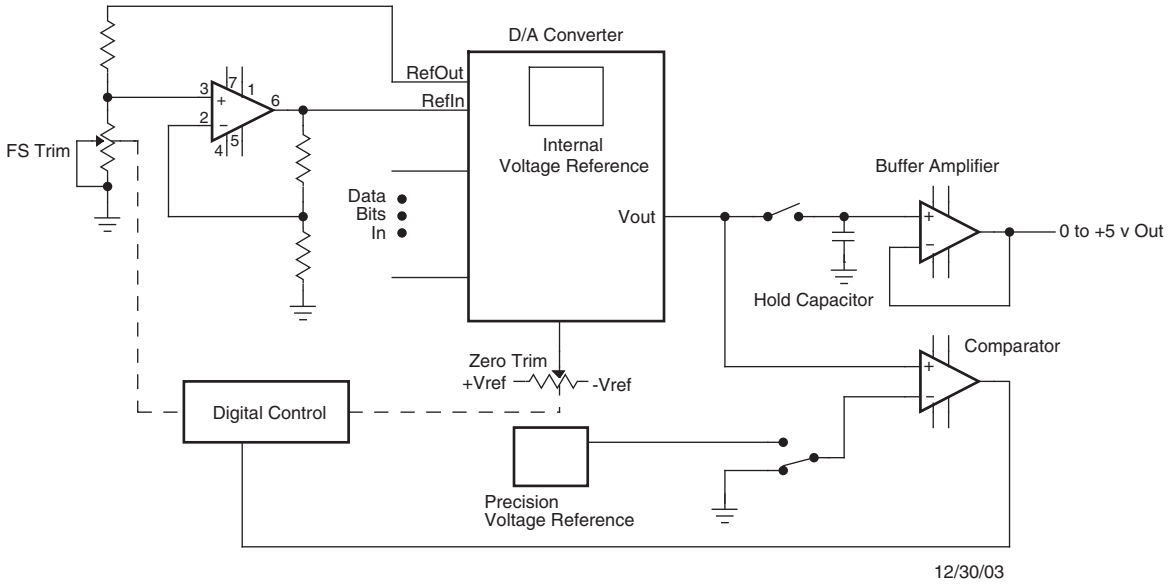



**FIGURE 29.**

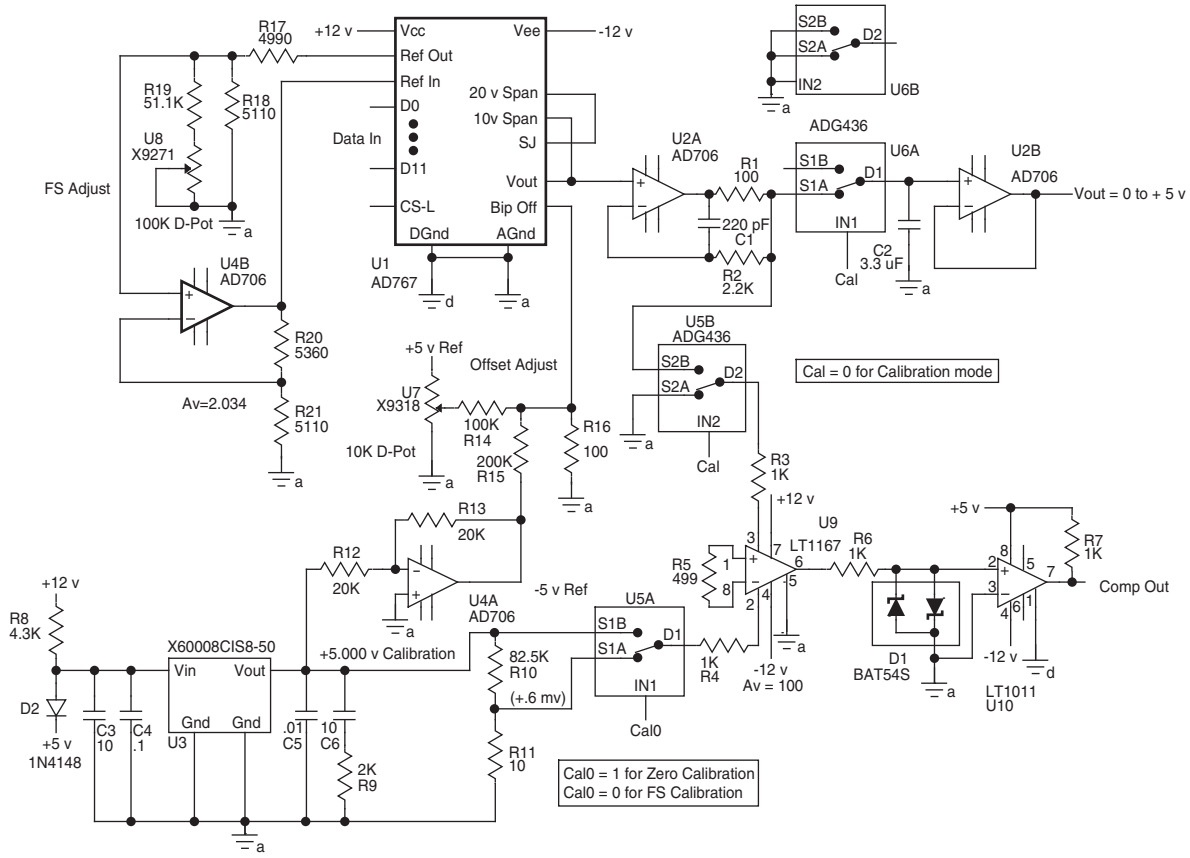



**FIGURE 30.**

The output voltage from the D/A converter, U1 is buffered with unity gain op amp (U2A) which is configured to drive a capacitive load, C2. Resistor R1 provides isolation from the capacitive load; resistor R2 and capacitor C1 are for loop compensation. Capacitor C2 is a Hold capacitor because during the calibration mode, switch S1A of U6A disconnects the D/A converter output voltage from the system output voltage, Vout. During the calibration mode, capacitor C2 stores and holds the previous D/A converter output voltage so that the system output voltage Vout stays constant. Op amp U2B buffers the hold capacitor voltage. To minimize charge injection error from the opening of S1A of U6A, a large capacitor value is used for C2. So, during normal run mode, switch S1A of U6A is closed, and the D/A converter output voltage is stored on capacitor C2. The system output voltage Vout tracks the D/A converter output voltage.

Instead of using manual adjustment potentiometers for the zero and full scale adjustments, digital potentiometers (DCPs) are used under digital control. Zero or offset errors are adjusted to zero by using the Bipolar Offset pin of U1. D-Pot U7, resistors R14, R15, and R16 varies the Bipolar Offset pin voltage by ±10 mV.

Full scale or gain adjustments are made by modifying the Ref Out voltage of U1 by ±1% before it is applied to the Ref In pin. The AD7676 data sheet recommends using a 100Ω potentiometer for this adjustment; however, this is difficult with a D-Pot because the Ref Out pin is at +10V. Also, the wiper resistance of a D-Pot (40-200Ω) prevents it from being used in varying a low resistance. Therefore, a scheme similar to the Auto-Calibrated Data Acquisition System is used. The Ref Out voltage is divided down by two with R17 and R18; D-Pot U8 and R19 allow the divided to be adjusted by ±1%. Then, the reference voltage is amplified back up by a gain of 2.034 amplifier circuit, U4B, R20, and R21.

During the calibration mode, S1A of U6A is opened and the system output voltage Vout remains at the previous D/A converter output voltage. Switch S2A of U5B is closed which applies the buffered D/A converter output voltage to an ultra low offset voltage comparator made up from Instrumentation Amplifier U9 and comparator U10. Since the offset voltage of the comparator circuit is a direct error source, it is necessary to minimize its offset voltage; unfortunately, the lowest offset voltage comparator (LT1011) still has a maximum offset voltage of 3mV worse case. To reduce the effect of U10's offset voltage, a gain of 100 difference amplifier, U9 is used; the overall effect of the comparator's offset voltage is attenuated by 100 to only 30μV. To minimize the voltage swing applied to the comparator, a Schottky diode clamp consisting of R6 and D1 is used so the input only swings ±0.3V. The output from the comparator, Comp Out, goes to a digital control circuit which increments or decrements the DCPS until a zero crossing is detected by the comparator.

There are two calibrations that must be made; first, the zero or offset calibration is performed. S1A of U5A is closed so that +0.6mV is applied to the comparator circuit input; the D/A converter input code is set to 0000 0000 $0000_b$. D-Pot U7 is adjusted until there is a transition on the Comp Out signal indicating that the D/A converter output voltage is +0.6 mV. Second, full scale or gain calibration is performed. S1B of U5A is closed so that +5.000 V is applied to the comparator circuit input; the D/A converter input code is set to 1111 1111 $1111_b$. DCP U8 is adjusted until there is a transition on the Comp Out signal indicating that the D/A converter output voltage is +5.000 V. To the purist, the +5.000 V calibration voltage applied to the comparator circuit should be set for 5V – 1.5LSB = 4.99817 V for full scale calibration. However, to maintain the accuracy of the X60008, it was decided to accept a full scale input voltage of +5.000V + 0.5LSB = +5.0006V.

The auto calibration scheme provides a closed feedback around the D/A converter, its internal reference voltage, and the buffer amplifier, U2A. The only remaining error at the system output voltage, Vout is the offset voltage of U2B (100μV) and the charge injection loss of switch S1A of U6A (90μV).

### *Digital Voltmeter Integrated Circuit Voltage Reference*

Integrated circuits for digital voltmeters are available from several semiconductor manufacturers; an example of such an IC is the MAX1494 from Maxim Integrated Products. This is a 4 1/2 digit single chip A/D converter with LCD drivers and is intended for digital voltmeters, digital panel meters, etc. A 4 1/2 digit meter can resolve 1 part in ±19,999 or ±.005% (±50ppm) on a 0.2V or 2V full scale range. The device has an internal voltage reference of +2.048V, or the user can supply an external reference. However, when even the simplest error analysis is performed, the inaccuracy of the IC is apparent. For example, from the data sheet, the gain error is ±0.5% using an external reference of 2.048V. The internal reference is specified at 2.048V ±2% with a temperature drift of 40ppm/°C. If the internal reference was to be used for measurements, the overall error of the reading would be ±2.5%.

The obvious solution would be to use an external 2.048V reference instead of the internal reference. However, the external reference must be very low power so the overall power budget is not jeopardized by the additional current of the reference device. The Intersil X60008 voltage reference provides all the desired functionality: very high initial accuracy of ±0.01%, low temperature drift of 5 ppm/°C, and extremely low supply current of 0.5 μA. By using the X60008 as the external reference as shown in Figure 31, the only error of the DVM measurement system is the gain error of ±0.5%.

If higher accuracy is required for the application, an auto-calibration scheme similar to the Auto-Calibrated Data Acquisition System described earlier in the Application Note could be used which would reduce the measurement error to less than ±0.02%.

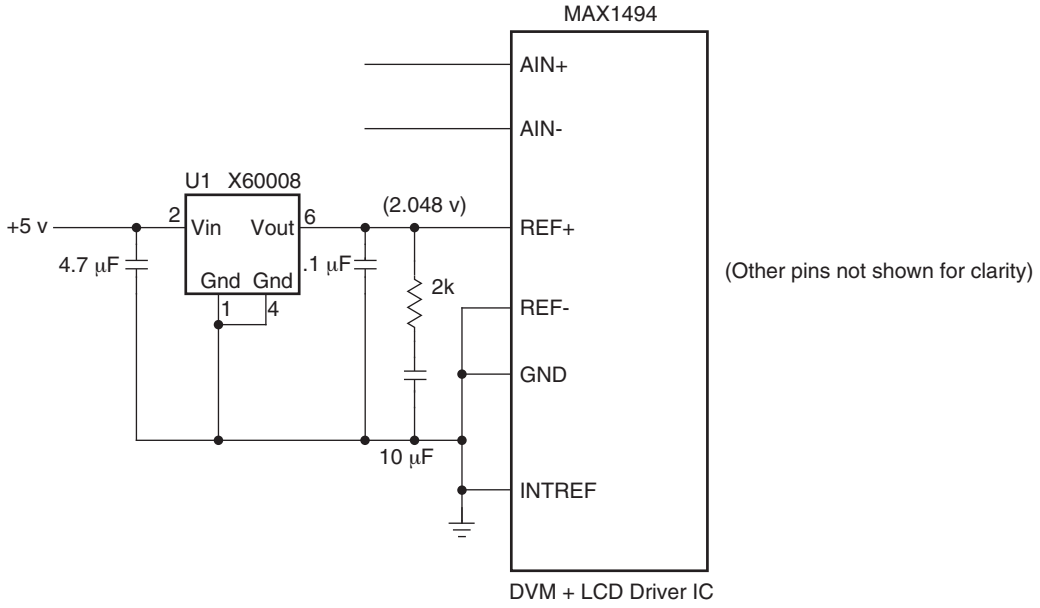



**FIGURE 31.**

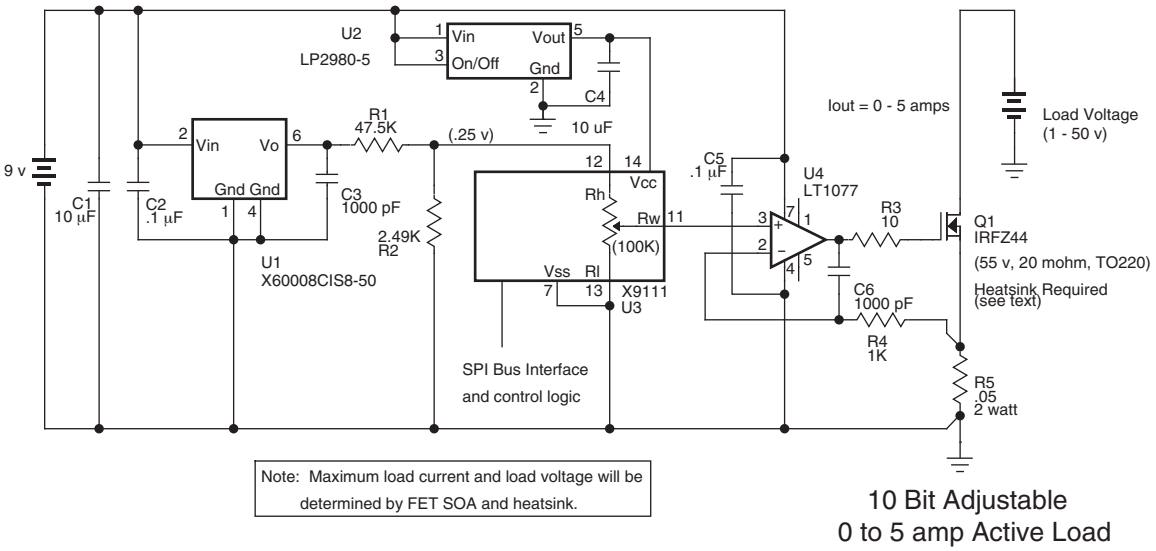



**FIGURE 32.**

### *0 to 5 Amp, 1 to 50 V Active Load 10 bit Resolution*

Often when testing power supplies and DC/DC converters it is helpful to have a programmable current source for an active load. Figure 32 shows a current source circuit that is capable of 0 to 5 Amp adjustable output current with 10 bits resolution (5mA) with 50V compliance voltage. The maximum simultaneous voltage and current will be limited by FET SOA considerations, the heatsink and available airflow. The active load is powered from a standard 9V battery; due to the very low operating current of the X60008 voltage reference, the battery current is less than 650µA for long battery life.

The Intersil X60008, U1 provides a precise +5.000V to a voltage divider R1 and R2 which generates +0.25V. The output voltage from the wiper terminal of a Intersil digital potentiometer (D-Pot), U3 varies from $0V_{DC}$ to $+0.25V_{DC}$ in 250µV increments under SPI bus control. Operating power for U3 is provided by a low power linear regulator with a +5V output. Op amp U4, FET Q1, and resistor R5 are a current source circuit where the output current is:

Iout = N/1023 * 0.25V/0.05Ω
= N/1023 * 5 amps
where N = Wiper Control Register (WCR) value in decimal

Resistor R3 provides gate isolation to prevent oscillation in the FET. Resistor R4 and capacitor C6 provide loop compensation for the current source circuit.

As mentioned earlier, the maximum output current and load voltage are limited to the SOA of the FET as shown below:

| LOAD VOLTAGE (Vdc) | OUTPUT CURRENT (Amps) | OUTPUT POWER (Watts) |
|---|---|---|
| 3.3V | 5 Amps | 16.5 |
| 5.0V | 5 Amps | 25 |
| 12V | 5 Amps | 60 |
| 25V | 2 Amps | 50 |
| 50V | 1 Amp | 50 |

In addition the FET must be attached to a large heatsink and equipped with a fan to obtain maximum output power and limiting the FET junction temperature to a safe temperature. For example, if the circuit is operating at full output power (60 watts) and room temperature (25°C), the heatsink and airflow must have a thermal resistance sink to air ($\theta$sa) of less than 1°C/Watt to keep the junction temperature of the FET at less than 125°C.

### *4 to 20 mA. Two Wire Current Transmitter*

In industrial control systems and process control systems, 4-20mA current loops are widely used to transmit analog process data over long distances on a factory floor or in a manufacturing complex. Current loops are used due to their noise immunity and suffer no loss of signal due to wire drops (IR losses). Over the years, current loops have proven to be a reliable and economical way to transmit analog data over an already installed copper wire. There are two types of 4-20mA current transmitters – two wire and three wire. Two wire transmitters work via a single pair of wires by modulating the current in the loop depending on a process variable input (temperature, pressure, voltage, etc.). They steal their operating power from the loop power source as shown in Figure 33A. A three wire 4-20mA current transmitter uses an external power supply for its operating voltage and may sink or source current depending on its configuration as shown in Figure 33B and Figure 33C.

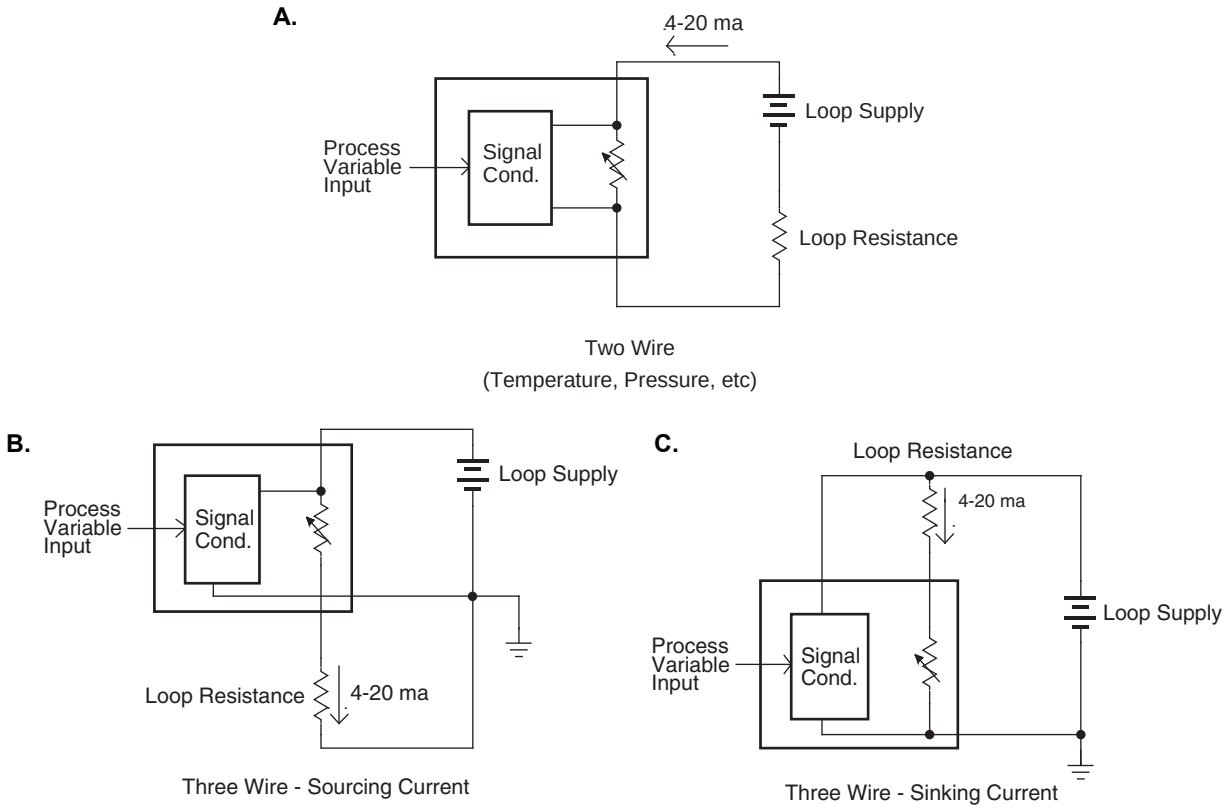



**FIGURE 33.**

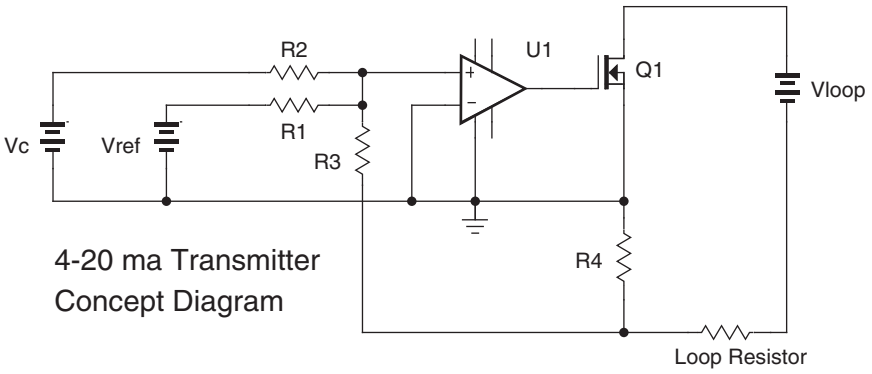



**FIGURE 34.**

The conceptual schematic diagram shown in Figure 34 is for a two wire 4-20mA transmitter.

Operation of the circuit can best be understood by summing the currents in the non-inverting input of U1:

Vc/R2 + Vref/R1 + (-Iloop)*R4/R3 = 0

Solving for Iloop:

Iloop = R3/R4 * (Vref/R1 + Vc/R2)

Summing the currents at the inverting input of U1 is not a mistake! Q1 provides an additional 180° of phase shift; therefore, for negative feedback, it is necessary to close the loop at the inverting input of U1 not the non-inverting input.

Notice the loop current is sensed by R4 so the quiescent current of U1, Vref, etc. are inside the feedback loop and cause no output current error. The only current that is not accounted for is the current in R3; typically this would be less than 0.1% FS error.

A complete schematic for the 2 wire 4-20mA current transmitter is shown in Figure 35.

You will notice the same components and reference designations as shown in the concept diagram! The Intersil X60008 voltage reference provides a precise +5.000V for Vref. Since the input voltage of the X60008 is limited to +10V maximum, a low dropout linear regulator U2 is used to regulate the loop voltage of 7 to 30V to 5.6V. The advantage of using a 5.6V regulator is that it provides a well stabilized source of power if additional signal conditioning is required; this will be demonstrated in the next application circuits for temperature measurements. The output current from the +5.6V bias supply is limited to approximately 3mA since the total current “stolen” from the loop must be less than 4mA. Due to the exceptionally low supply current of the X60008 voltage reference, the use of a FET for Q1 (instead of a bipolar transistor), and the low supply current of U3, the total quiescent current of the loop electronics (U1, U2, U3, etc.) is less than 250µA! This leaves 3.75mA available for signal conditioning circuitry. If lower

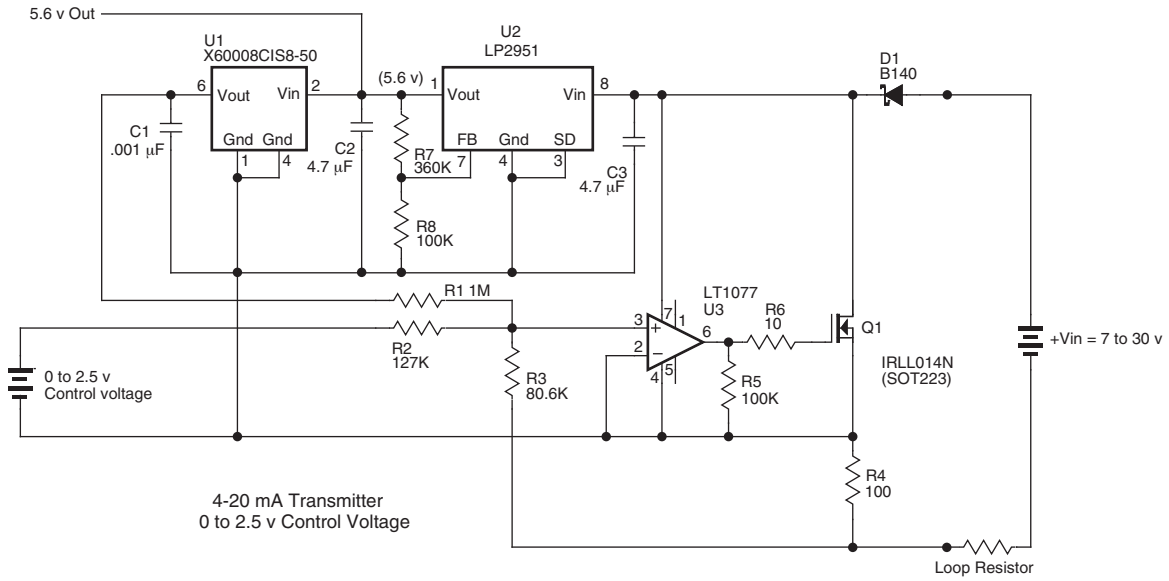



**FIGURE 35.**

temperature drift is required for A/D converters or other industrial sensors, the X60008AIS8-50 could also be used for less than 1ppm/°C temperature drift.

The following circuits show several examples of using this 4-20mA current transmitter to monitor temperature with an Integrated Circuit temperature sensor (Figure 36) and a PT100 RTD (Figure 37).

Temperature in °F is monitored in Figure 4 with an IC temperature sensor that provides an output voltage of 10mV/°F over a temperature range of +5°F to +200°F. The loop current will vary from 4mA to 20mA as the temperature changes as shown in the equation below:

Iloop = 4 + 0.0806 * T (mA)

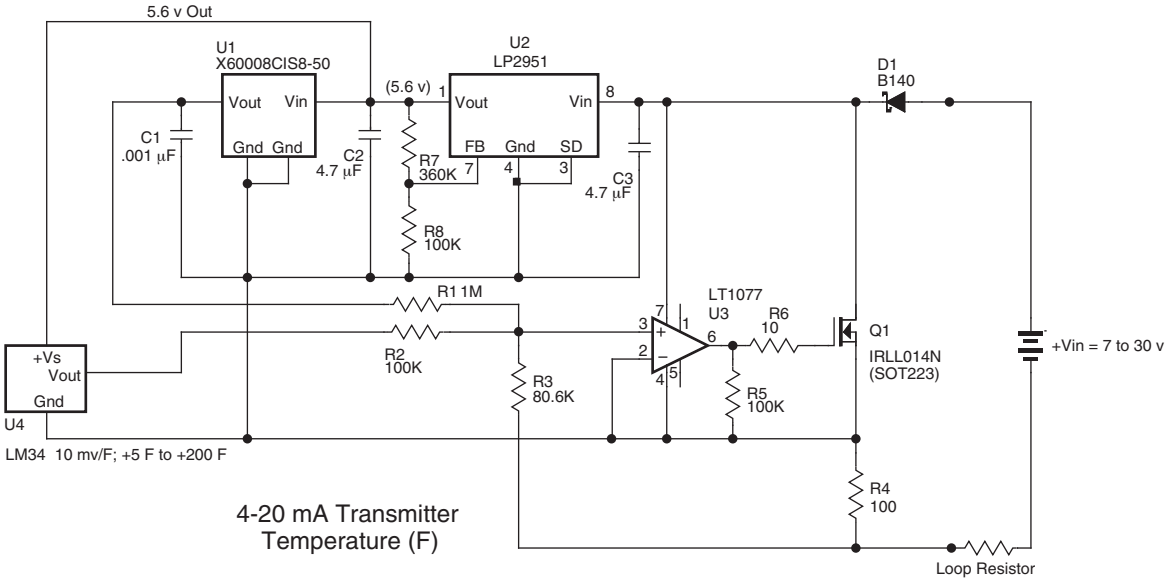



**FIGURE 36.**

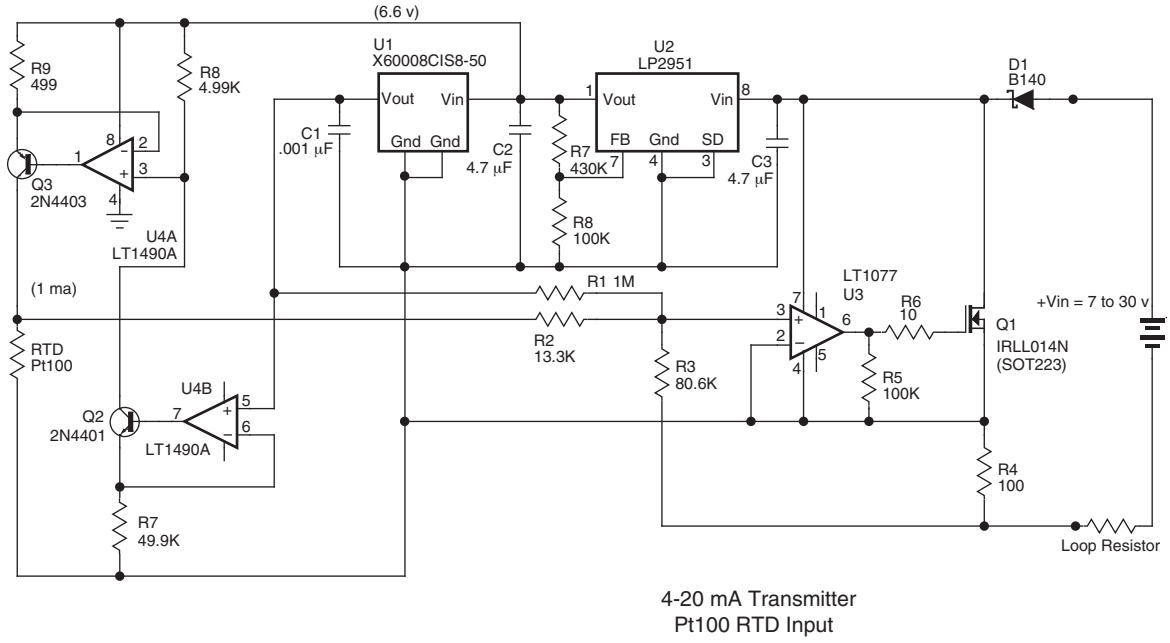
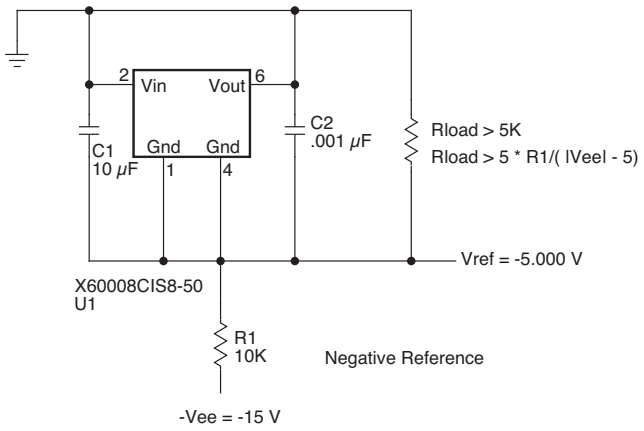



**FIGURE 37.**

Figure 37 shows interfacing the 4-20mA transmitter to a Pt100 RTD that is commonly used in high precision temperature measurements. The resistance of the RTD ($\alpha$ = 0.00392, American standard) is 100Ω at 0°C and varies from 60Ω for -100°C to 267Ω for +450°C. The RTD has an excitation current of 1mA from a constant current source circuit. The Intersil X60008 voltage reference performs double duty in this circuit; it provides the reference voltage for the 4mA offset current, and it provides a reference voltage for a 100µA current source consisting of U4B, R7, and Q2. The 100µA output current is converted to 500mV by R8; this voltage is referenced to the 6.6V bias supply from U2. A precision 1mA current source consisting of U4A, Q3, and R9 provides the excitation current for the RTD. The output voltage from the RTD will vary from 60mV to 267mV so the loop current will vary from 7.8mA at -100°C to 20mA at +450°C. Notice that the bias voltage from U2 was increased to 6.6V to allow for headroom in the current source circuit.

### *Negative Output Voltage Or Standing the Reference on its Head!*

Even though the Intersil X60008 Voltage Reference is a positive 5.000V output device, it is also possible to operate it on a negative voltage for a -5.000V output as shown in Figure 38.




**Note:** Extreme care should be taken in not shorting Vref to Vee. This would cause 15V across $V_{IN}$ to GND which will damage the device.

**FIGURE 38.**

While it would seem that this configuration violates the X60008 dropout voltage specification of 300mV (i.e., the input voltage must be 300mV greater than the output voltage), it must be recognized that this circuit is operating in a shunt mode. In the shunt mode, only the bottom output driver FET is active; the top side output driver FET is not active and therefore not required for circuit operation. The only requirement for proper operation is that the voltage divider made up of R1 and Rload always allow 5V between the Vout pin and Gnd pin. Thus, the load resistor Rload must satisfy the following equation:

Rload > 5V * R1 / (|Vee| -5)

For the example in Figure 38:

Rload > 5V * 10k / (|-15 V| - 5) = 5kΩ

Notice that the X60008 Voltage Reference never sees the -Vee voltage; only R1 sees the voltage, and -Vee can be any voltage as long as R1 can handle it! For example, -Vee could be -1000V if R1 = 1MΩ and Rload > 5kΩ!

Perhaps even more useful, is a voltage reference that hangs below a positive supply rail; a good example is the Precision, Low Noise Current Source circuit shown in this Application Note. Figure 39 shows the configuration for a “negative” reference voltage that is referenced to a positive supply.

In this circuit, the voltage reference output voltage is -5.000V with respect to +24V. This circuit will work with any supply voltage greater than 5V; the only limit is the amount of voltage that R1 can withstand. If the load current from the voltage reference is very small (i.e., op amp bias current), the resistor R1 can be a VERY large value since it must only supply the quiescent current of the X60008 and the leakage current in C1. For example, in this circuit (assuming no leakage current in C1), R1 could be as large as 48MΩ! While an interesting observation, this large resistor value is not practical because it would take 8 minutes to charge up C1!

Another interesting circuit that is possible due to the ultra low supply current of the X60008 is the ability to hang the reference voltage between any voltage, and actually put the 5.000V reference voltage into any point in a circuit as shown in Figure 40.

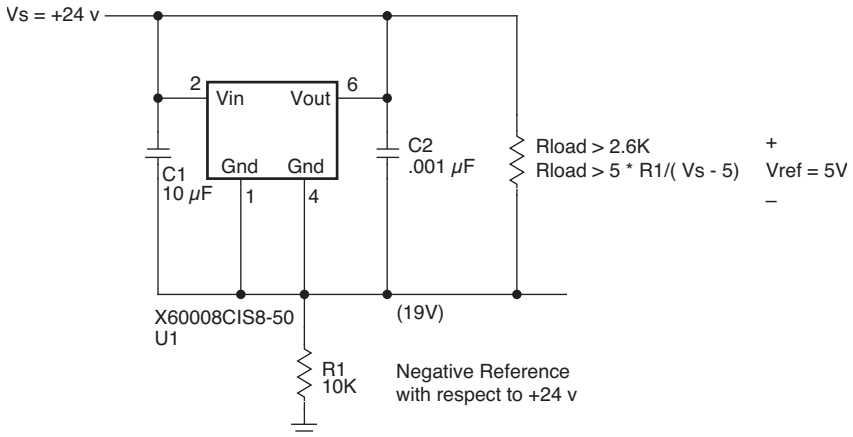



**FIGURE 39.**

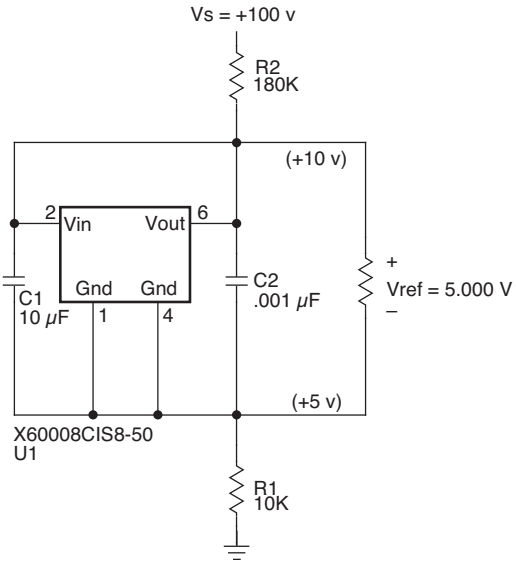



**FIGURE 40.**

### *Operation from a Battery or Capacitor*

The ultra low supply current (less than 0.5µa) of the Intersil X60008 voltage reference makes it a natural part for battery operated systems such as handheld DVMs, portable medical monitors, etc. However, its low supply current also allows the X60008 to be operated on a battery that is buried on a PCB and NEVER turned off thus eliminating any warm-up drift effects and enhancing long term stability. The circuit shown in Figure 41 operates from two 12.5mm coin cells to provide an input voltage of 6V.

Since the specified batteries (Panasonic BR1225-1HC) are rated at 50mAh, the battery life exceeds 10 years as shown below:

Life = 50mAh /0.5µA = 100,000 hours ➔ 4166 days ➔ 11 years

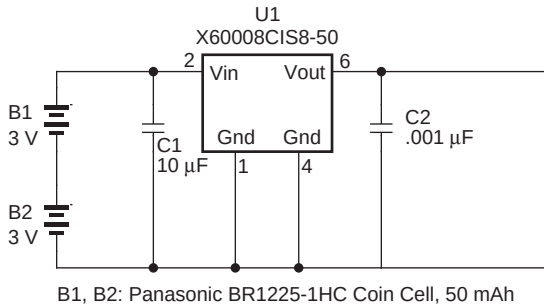



**FIGURE 41.**

Of course, any load current on the output of the X60008 will decrease the battery life since it will be added to the supply current, and it takes very little load current to drastically reduce the battery life. For example, just monitoring the reference voltage with a Digital Voltmeter with a 10MΩ input resistance will increase the total battery current to 1µA (0.5µA + 5V/10M), and the battery life will be cut in half!

It must also be noted that this voltage reference circuit is a truly floating voltage reference – there is no “Ground” on this circuit so the output voltage can be placed anywhere in a circuit.

It is also possible to operate the X60008 with only a capacitor supplying the input voltage during short – medium term power supply interruptions such as brown-outs or changing a battery as shown in Figure 42.

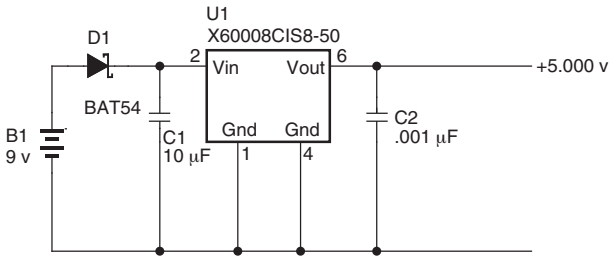



**FIGURE 42.**

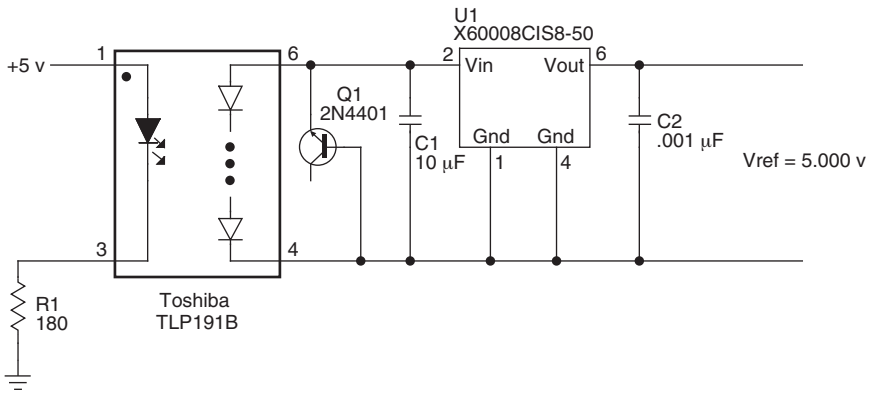



**FIGURE 43.**

Since I(t) = C * dv/dt for a capacitor, hold-up time dt = C/I * dv. In this circuit, hold-up time is 10µF /0.5µA* (9-5.3) = 74 seconds. This is easily enough time to change the battery B1 and not require a new warm-up period for the voltage reference. Caution must be exercised in attempting to increase the hold-up time by increasing the capacitance of C1. Unfortunately, the leakage current of a high value capacitor (>100µF) is much higher than the quiescent current of the X60008 so the capacitor leakage current will dominate. For example, a 1000µF/10V capacitor has a leakage current of 64µA; the hold-up time for this capacitor would be 58 seconds. Ideally, a film capacitor would be used for the lowest leakage current, but high value film capacitors are not readily available, are very expensive, and are very large.

Therefore, using a capacitor as the input voltage source for the X60008 should only be considered in short to medium hold-up time applications. Also, do not forget to take into account the voltage reference load current as discussed in the battery application above.

### *Optically Isolated Voltage Reference*

Another truly unique application for the Intersil X60008AIS8-50 voltage reference is an optically isolated voltage reference that can float at any point in a circuit with up to 2500 vac, rms isolation voltage. Once again, the ultra low supply current of the X60008AIS8-50 makes this possible as shown in Figure 43.

The heart of this circuit is the optically coupled photodiode array, Toshiba TLP191B which will generate an open circuit output voltage of 7 V and a short circuit current of 24µA when the input LED is operating at 20mA. The Thevenin equivalent of this voltage source is 7V with a Thevenin resistance of 290kΩ. Even though this is a very low voltage source, it is enough to supply the 0.5µA that is required by the X60008AIS8-50 to operate. In fact, Q1 is operated in reverse base-emitter breakdown (an excellent low current 7 V “Zener”) to clamp the X60008AIS8-50 input voltage to less that 10V; tests showed that the output voltage from the TLP191 exceeded 10V with no load on the X60008AIS8-50. Since the X60008AIS8-50 load current must by sourced by the photodiode array, the load current must be less than 20µA or greater than 250kΩ. Also, turn-on time for this circuit is 2 seconds (or less) since capacitor C1 must be charged by the 24µA from the photodiode array.

This demonstrates the “World’s First Isolated 1ppm/°C Voltage Reference”.

## *Appendix A*

“Precision Voltage Reference Using EEPROM Technology”, by Jim McCreary. Available at www.Intersil.com

## *Appendix B*

### *0.1 to 10Hz Noise Test Box with Peak to Peak Detector*

Testing the 0.1 to 10Hz output noise of a high quality reference such as the X60008 is often a challenge since to test for 0.1 to 10Hz noise, it is necessary to AC couple the output of the reference before it is applied to a high gain stage as shown in Figure 44.

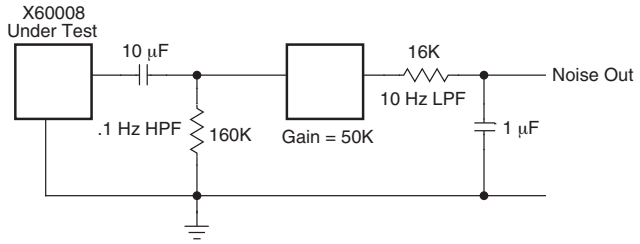



**FIGURE 44.**

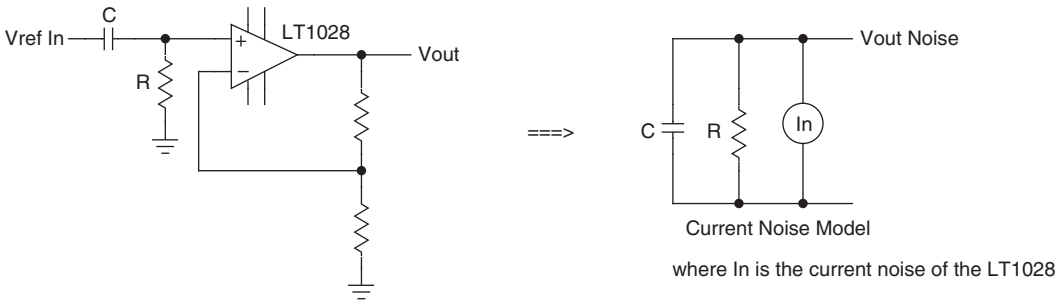



**FIGURE 45.**

Additionally, the high gain stage in a typical noise test circuit requires a very low voltage noise op amp; however, low voltage noise op amps exhibit high current noise which prevent the use of high resistance values in the 0.1Hz AC coupling filter. Appendix B describes a novel voltage reference test circuit which eliminates the need for a 0.1Hz AC coupling network and includes a peak to peak voltage detector. Also in Appendix B test data is shown for various output filter circuits.

To effectively block noise in the 0.1Hz - 10Hz bandwidth, it is necessary to use large value capacitors and/or large value resistors. For example, for a one pole high pass filter with a 0.1Hz corner frequency it is necessary to use resistor and capacitor values as shown below:

| RESISTOR (kΩ) | CAPACITOR (µF) |
|---|---|
| 160 | 10 |
| 16 | 100 |
| 1.6 | 1000 |

While it seems attractive to use a 1000µF capacitor and 1.6kΩ resistor, the large leakage current of an electrolytic capacitor will generate a voltage across the 1.6kΩ resistor that will gained up by the amplifier that will send the amplifier output into saturation. For example, the leakage current of a typical 1000µF capacitor is 100µA (Panasonic VS series); an error voltage of 100µA * 1600Ω = 160mV is created! Reducing the capacitor value and increasing the resistor value does not help because while the leakage current is reduced, the resistor value is increased! For example, for a 100µF capacitor in the same Panasonic VS series, the leakage current is reduced to 10µA, but the resistor value is increased to 16kΩ, and the same 160mV is introduced.

Additional noise will be introduced into the measurement by the current noise of the op amp in the gain stage developing a noise voltage across the resistor as shown in Figure 45. It is interesting to note that the frequency characteristics of the resulting noise voltage will be shaped by the high pass filter network. At high frequency, the capacitor's impedance is very low and there is no resulting voltage noise. Unfortunately, the voltage noise resulting from the current noise is shaped by the same filter characteristic as the measurement!

To eliminate the problems associated with the AC coupling network (i.e., an error voltage is introduced and noise voltage is introduced), the simplest solution is to throw away the culprit – the AC coupling filter. But now you are faced with the problem of applying a reference voltage of +5V to a gain stage of 50,000; the output of the op amp will try to go to 250,000Vdc which of course is not very practical! But, you can subtract off the DC component of the reference voltage and leave only the noise voltage of the reference as shown in the concept diagram in Figure 46.

An instrumentation amplifier is used to subtract the Voltage Reference Under Test from a Precision Low Noise Voltage Reference with voltage noise that is much less than the noise of the Voltage Reference Under Test. There are no filters in the measurement path at this point so there are no errors associated with RC filter networks; all of the 0.1Hz to 10Hz frequency shaping is done after the high gain stage of the instrumentation amplifier.

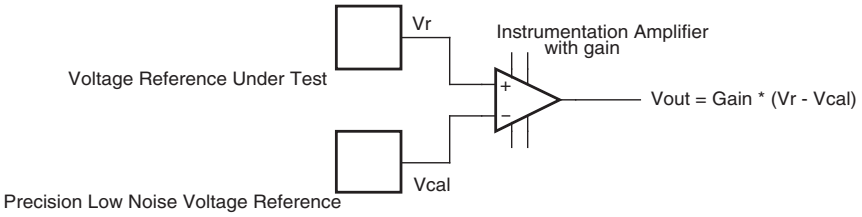



**FIGURE 46.**

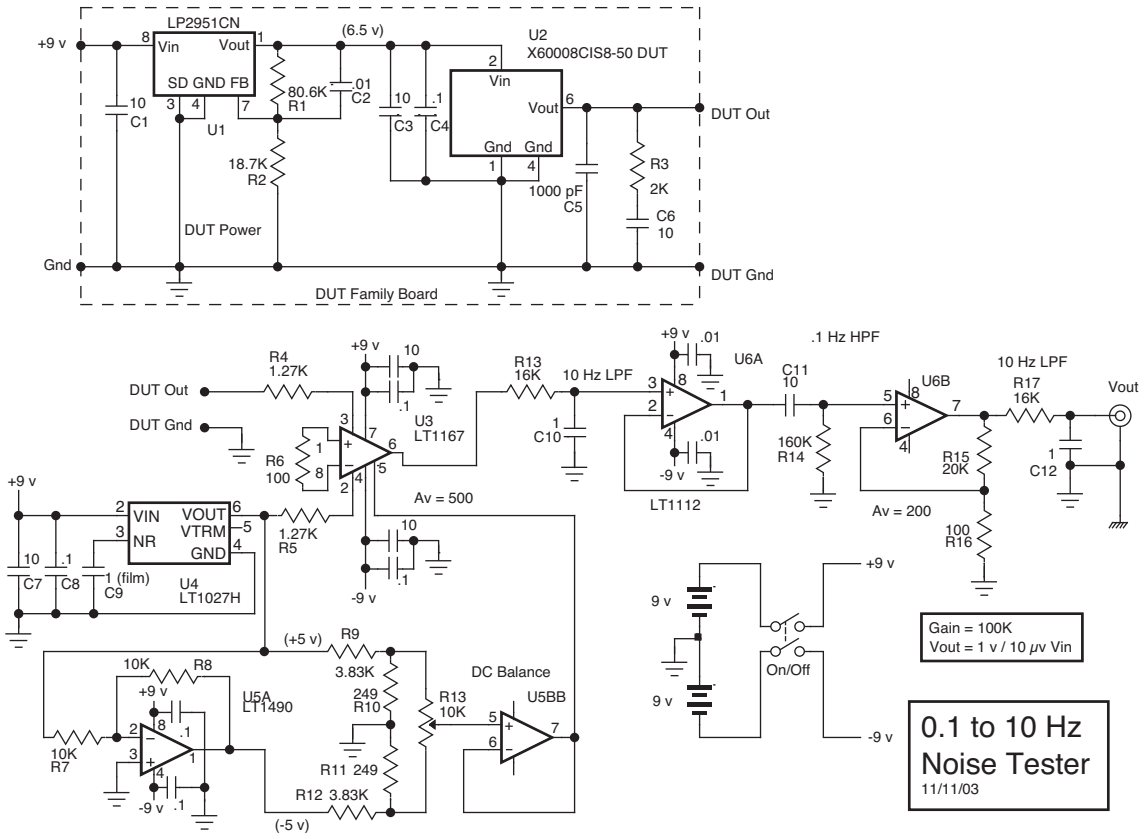



**FIGURE 47.**

Figure 47 shows the complete schematic for the 0.1 to 10Hz Noise Tester.

Overall gain for the tester is 100K so that an input 10µV yields 1V at the output. Operating power for the Noise Tester is from two 9V “transistor radio” batteries to ensure complete isolation from the AC power line, and no ground loops when an oscilloscope is connected to the output. The Noise Tester is mounted in a metal chassis with a tight cover to eliminate pick-up to the AC power line or RF signals such as radio or TV transmitters. Every integrated circuit is bypassed with a 10µF and 0.1µF capacitor. The X60008 under test (DUT) is mounted on a Family Board so different reference configurations can be easily tested. A linear regulator U1 sets the DUT input voltage to 6.5V as set by R1 and R2.

The output voltage from the DUT is applied to the non-inverting input of instrumentation amplifier U3. A low noise reference U4 is connected to the inverting input of the IA so that the low noise reference output voltage is subtracted from the DUT output voltage leaving only the noise voltage of the DUT. R6 sets the gain of the IA at 500. A DC balance circuit (U5, etc.) is shown to null out the DC difference between the DUT and low noise reference. In reality, the output tolerance of both the X60008 voltage reference and LT1027 low noise reference was so low that the DC balance circuit was not really needed, but it is shown here for completeness.

One pole of a 10Hz low pass filter is set with R13 and C10; U6A is a low bias current, low noise unity gain buffer. R14 and C11 are a one zero high pass filter with a corner frequency of 0.1Hz. U6B provides a gain of 200 so the total gain is 500 x 200 = 100,000.

R17 and C12 provide a second pole for the 10Hz low pass filter. The second zero for the 0.1Hz high pass filter is set by observing the Vout waveform for 10 seconds.

Voltage reference noise tests are performed with this Noise Tester the same as any other noise test box. A short BNC cable connects the Noise Tester to a storage oscilloscope input. The DUT is allowed to warm up for approximately 5 minutes to allow the DUT and test box to stabilize after turning on switch S1. Ten second test runs are made with the storage scope to measure the peak to peak waveform over the 10 second interval.

However, this test method is somewhat subjective because the peak to peak noise reading must be made off the oscilloscope screen; it is made even more difficult if a storage scope is not available for these readings. Figure 48 shows a pair of peak detector circuits to store the maximum peak voltage (U1, D1,

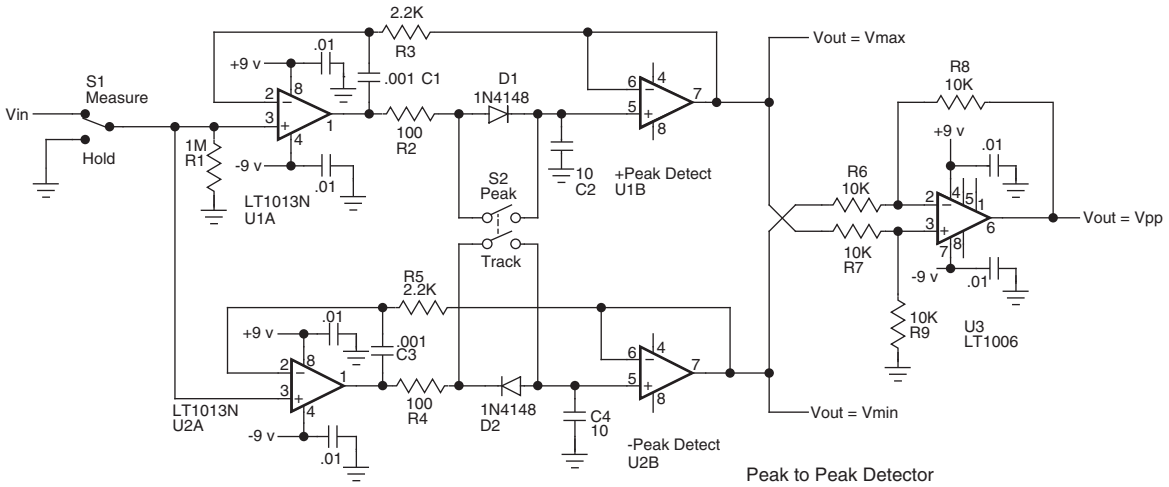



**FIGURE 48.**

C2) and minimum peak voltage (U2, D2, C4). A difference amplifier (U3, R6-R9) subtracts the maximum reading and minimum reading for a DC output voltage that is equal to the peak-peak noise voltage. The peak detector circuits use isolation resistors (R2, R4) and local loop compensation (R3, C1, R5, C3) to drive the heavy capacitive load of C2 and C4. In the test box, mechanical switches were used for S1 and S2; analog switches could also be used if logic control is required.

To use the Peak to Peak Detector circuit with the Noise Tester, the output of the Noise Tester (Figure 47) is connected to Vin of Figure 48. When switch S1 is placed in the Measure position, and S2 is placed in the Track position the output voltage of U1B and U2B tracks the input voltage. To measure the peak noise voltage, S2 is placed in the Peak position, and the output voltage of U1B (Vmax) and U2B (Vmin) detects and holds the maximum and minimum values on C2 and C4. After 10 seconds, switch S1 is moved to the Hold position, and the output voltage, Vout is measured with a DVM or oscilloscope. The voltage at Vout represents the peak to peak noise voltage in a 0.1Hz to 10Hz bandwidth for the Voltage Reference Under Test.

### *Test Results*

Various configurations of X60008 Voltage Reference and Filters were tested using the Noise Tester and Peak to Peak Detector. In each test, five test runs were made and the results averaged as shown in the following tables.

1. The noise test system and strategy were verified by using a LT1027 as the DUT so that a LT1027 was tested against a LT1027. The measured peak to peak output noise was less than 3µV indicating the test system was working properly.

| TEST SET-UP CHECK: | | | | | | |
|---|---|---|---|---|---|---|
| TEST CONDITION: LT1027 WITH 1µF NR AND 4.7µF COUT | | | | | | |
| PART ID # | GRADE | VOLTS PP (MV) | | | | AVERAGE |
| | | 3.0 | 2.7 | 2.5 | 2.6 | 2.7 |

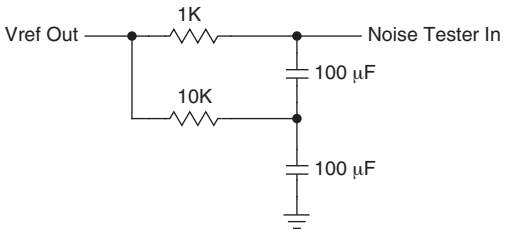
2. A sample of ten X60008 Grade C and Grade D were tested using an output network of 0.01µF in parallel with 10µF + 2K.

| TEST CONDITION: OUTPUT FILTER NONE | | | | | | | |
|---|---|---|---|---|---|---|---|
| PART ID # | GRADE | VOLTS PP (MV) | | | | | AVERAGE |
| 5 | D | 39 | 43 | 37 | 46 | 39 | 40.8 |
| 10 | D | 23 | 28 | 21 | 23 | 27 | 24.4 |
| 12 | D | 38 | 34 | 34 | 35 | 32 | 34.6 |
| 14 | D | 23 | 20 | 24 | 22 | 18 | 21.4 |
| 8 | D | 30 | 29 | 30 | 27 | 31 | 29.4 |
| 9 | C | 24 | 25 | 24 | 30 | 28 | 26.2 |
| 6 | C | 22 | 27 | 22 | 24 | 27 | 24.4 |
| 13 | C | 34 | 29 | 29 | 34 | 30 | 31.2 |
| 16 | C | 29 | 29 | 33 | 28 | 31 | 30.0 |
| 17 | C | 30 | 30 | 26 | 28 | 25 | 27.8 |

3. A sample of two X60008 Grade D were tested using an output network of 0.01µF

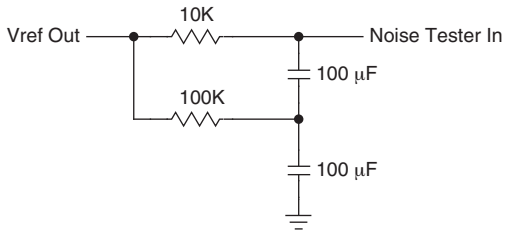
| TEST CONDITION: OUTPUT NETWORK =.01µF; OUTPUT FILTER NONE | | | | | | | |
|---|---|---|---|---|---|---|---|
| PART ID # | GRADE | VOLTS PP (µV) | | | | | AVERAGE |
| 5 | D | 36 | 38 | 38 | 40 | 36 | 37.6 |
| 8 | D | 29 | 32 | 31 | 31 | 33 | 31.2 |

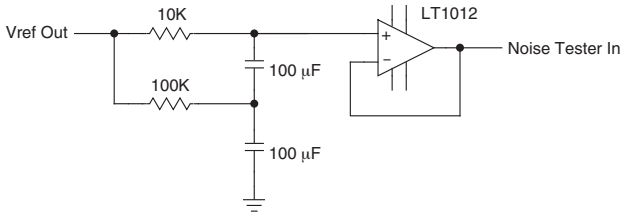
4. A sample of two X60008 Grade D were tested using an output network of 0.01µF in parallel with 10µF + 2K. The following filter was used.




| PART ID # | VOLTS PP (µV) | | | | | AVERAGE |
|---|---|---|---|---|---|---|
| 5 | 40 | 35 | 32 | 44 | 35 | 37.2 |
| 10 | 20 | 21 | 22 | 18 | 22 | 20.6 |

5. A sample of two X60008 Grade D were tested using an output network of 0.01µF in parallel with 10µF + 2K. The following filter was used.




| PART ID # | VOLTS PP (µV) | | | | | AVERAGE |
|---|---|---|---|---|---|---|
| 5 | 18 | 17 | 20 | 19 | 19 | 18.6 |
| 10 | 14 | 13 | 10 | 13 | 11 | 12.2 |
| 12 | 17 | 15 | 13 | 14 | 19 | 15.6 |
| 14 | 10 | 11 | 10 | 10 | 9 | 10.0 |
| 8 | 12 | 17 | 15 | 9 | 12 | 13.0 |

6. A sample of two X60008 Grade D were tested using an output network of 0.01µF in parallel with 10µF + 2K. The following filter was used.

| PART ID # | VOLTS PP (µV) | | | | | AVERAGE |
|---|---|---|---|---|---|---|
| 5 | 14 | 11 | 10 | 15 | 10 | 12.0 |
| 10 | 9 | 13 | 10 | 11 | 12 | 11.0 |
| 12 | 12 | 11 | 9 | 15 | 8 | 11.0 |
| 14 | 10 | 10 | 9 | 10 | 7 | 9.2 |
| 8 | 11 | 10 | 8 | 11 | 13 | 10.6 |

As you can see from the Summary below, a noise reduction of 50% to 75% is achieved with the use of an output filter as shown in Test 6 above.

| SUMMARY: | AVERAGE VOLTS PP (µV) | | | | | |
|---|---|---|---|---|---|---|
| PART ID # | TEST | 5 | 10 | 12 | 14 | 8 |
| No Filter | 2 | 40.8 | 24.4 | 34.6 | 21.4 | 29.4 |
| 1.6 Hz Filter | 4 | 37.2 | 20.6 | | | |
| .16 Hz Filter | 5 | 18.6 | 12.2 | 15.6 | 10.0 | 13.0 |
| .16 Hz with LT1012 Buffer | 6 | 12.0 | 11.0 | 11.0 | 9.2 | 10.6 |

## Footnotes

1. Linear Technology Corp. LT1004 datasheet
2. Linear Technology Corp. LT1460 datasheet
3. Linear Technology Corp. LT1461 datasheet
4. Bob Widler, *New Developments in IC Voltage Regulators*, IEEE Journal of Solid State Circuits, Vol. SC-6, February 1971.
5. Paul Brokaw, *A Simple Three-Terminal IC Bandgap Voltage Reference*, IEEE Journal of Solid State Circuits, Vol. SC-9, December 1974.
6. Roya Nasraty, *XFET™ References*, Analog Dialog.
7. Linear Technology Corp., Design Note 229, *Don't Be Fooled by Voltage Reference Long Term-Drift and Thermal Hysteresis*, Jon Wright.
8. Mies van der Robe, *NY Herald Tribune*, June 28, 1959

## List of Figures

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