ANALOG DEVICES

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APPLICATION NOTE

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Errors and Error Budget Analysis in Instrumentation Amplifier Applications

by Eamon Nash

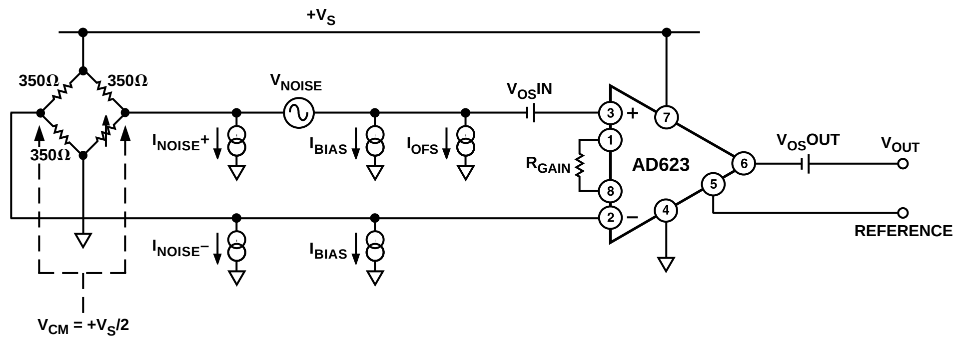

Figure 1. Error Sources in a Typical Instrumentation Amplifier

INTRODUCTION

This application note describes a systematic approach to calculating the overall error in an instrumentation amplifier (in amp) application. We will begin by describing the primary sources of error (e.g., offset voltage, CMRR, etc.) in an in amp. Then, using data sheet specifications and practical examples, we will compare the accuracy of various in amp solutions (e.g., discrete vs. integrated, three op amp integrated vs. two op amp integrated).

Because instrumentation amplifiers are most often used in low speed precision applications, we generally focus on dc errors such as offset voltage, bias current and low frequency noise (primarily at harmonics of the line frequency of either 50 Hz or 60 Hz). We must also estimate the errors that will result from sizable changes in temperature due the rugged and noisy environment in which many in amps find themselves.

Its also important to remember that the effect of particular error sources will vary from application to application. In thermocouple applications, for example, the source impedance of the sensor is very low (typically not greater than a few ohms even when there is a long cable between sensor and amplifier). As a result, errors due to bias current and noise current can be neglected when compared to input offset voltage errors.

RTO and RTI

Before we consider individual error sources, understanding of what we mean by RTO and RTI is important. In any device that can operate with a gain greater than unity (e.g., any op amp or in amp), the absolute size of an error will be greater at the output than at the input. For example, the noise at the output will be the gain times the specified input noise. We must, therefore, specify whether an error is referred to the input (RTI) or referred to the output (RTO). For example, if we wanted to refer output offset voltage to the input, we would simply divide the error by the gain, i.e.,

$$\text{Output Offset Error (RTI)} = V_{OS}OUT/Gain$$

Referring all errors to the input, as is common practice, allows easy comparison between error sizes and the size of the input signal.

Parts per Million—PPM

Parts per million or ppm is a popular way of specifying errors that are quite small. PPM is dimensionless so we must make the error relative to something. In these examples, it is appropriate to compare to the full-scale input signal. For example, the input offset voltage, expressed in ppm, is given by the equation:

$$\text{Input Offset Error (ppm)} = (V_{OS}/V_{IN\ FULL\ SCALE}) \times 10^6$$

Error Sources in Discrete and Integrated Instrumentation Amplifiers
Figure 1 shows the most common and prevalent error sources in discrete and integrated in amps. These error sources are detailed individually below.

Offset Voltage
Offset voltage results from a mismatch between transistor V_{BE}s in an amplifier's input stage. This voltage can be modeled as a small dc voltage in series with the input signal, as shown in Figure 1. Like the input signal, it will be amplified by the gain of the in amp. In the case of in amps with more than one stage (e.g., the classic three op amp in amp) the input transistors of the output stage will also contribute an offset component. However, as long as the output stage has gain of unity, as is generally the practice, the in amp's programmed gain will have no effect on the absolute size of the output offset error. However, for error computation, this error is usually referred back to the input so that its effect can be compared to the size of the input signal. This yields the equation:

$$\textit{Total Offset Error (RTI)} = V_{OS\_IN} + V_{OS\_OUT}/Gain$$

From this equation, it is clear that the effect of output offset voltage will decrease as the in amp's programmed gain increases.

Offset and Bias Currents
Bias currents flow into or out of the in amp's inputs. These are usually the base currents of npn or pnp transistors. These small currents will therefore have a defined polarity for a particular type of in amp.

Bias currents generate error voltages when they flow through source impedances. The bias current times the source impedance generates a small dc voltage which appears in series with the input offset voltage. However if both inputs of the in amp are looking at the same source impedance, equal bias currents will generate a small common-mode input voltage (typically in the µV range) that will be well suppressed by any device which has reasonable common-mode rejection. If the source impedances of the in amp's inverting and noninverting inputs are not equal, the resulting error will be greater by the bias current times the difference in the impedances.

We must also consider the offset current, which is the difference between the two bias currents. This difference will generate an offset type error equal to the offset current times the source impedance. Because either of the bias currents can be the greater, the offset current can be of either polarity.

Common-Mode Rejection
An ideal in amp will amplify the differential voltage between its inverting and noninverting inputs regardless of any dc offsets appearing on both inputs. So any dc offset appearing on both inputs ($+V_S/2$ in Figure 1) will be rejected by the in amp. This dc or common-mode component is present in many applications. Indeed, removing this common-mode component is often the primary function of an instrumentation amplifier in an application.

In practice, not all of the input common-mode signal will be rejected and some will appear at the output. Common-mode rejection ratio is a measure of how well the instrumentation amplifier rejects common-mode signals. It is defined by the formula:

$$CMRR\,(dB) = 20 \times \log\left[\frac{Gain \times V_{CM}}{V_{OUT}}\right]$$

We can rewrite this equation to allow calculation of the output voltage that results from a particular input common-mode voltage.

$$V_{OUT} = \frac{Gain \times V_{CM}}{\log^{-1}\left(\dfrac{CMRR}{20}\right)}$$

AC and DC Common-Mode Rejection
Poor common-mode rejection at dc will result in a dc offset at the output. While this can be calibrated away (just like offset voltage), poor common-mode rejection of ac signals is much more troublesome. If, for example, the input circuit picks up 50 Hz or 60 Hz interference from the mains, an ac voltage will result at the output. Its presence will reduce resolution. Filtering is a solution only in very slow applications where the maximum frequency is much less than 50 Hz/60 Hz.

Table I shows the output voltages of two in amps, the AD623 and the INA126, that result when a 60 Hz common-mode voltage of 100 mV amplitude is picked up by the input.

Table I. Effect of CMRR on Output Voltage of AD623 and INA126 for a 60 Hz, 100 mV Amplitude Common-Mode Input

Gain	V_{IN} (cm)	CMRR–INA126	CMRR–AD623	V_{OUT}–INA126	V_{OUT}–AD623
10	100 mV @ 60 Hz	83 dB	100 dB	70.7 µV	10 µV
100	100 mV @ 60 Hz	83 dB	110 dB	707 µV	31.6 µV
1000	100 mV @ 60 Hz	83 dB	110 dB	7.07 mV	316 µV

Noise

While offset voltages and bias currents ultimately lead to offset errors at the output, noise sources will degrade the resolution of a circuit. There are two noise sources in most amplifiers, voltage noise and current noise. As is the case with offset voltage and bias current, the degree to which these sources affect the resolution varies with the application.

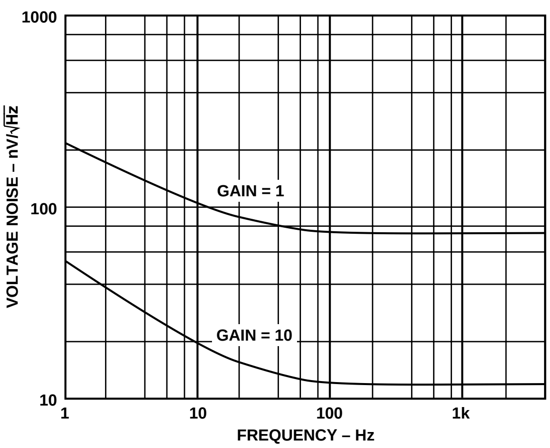

Figure 2. Voltage Noise Spectral Density of a Typical In Amp

The voltage noise spectral density of a typical in amp is shown in Figure 2 (a plot of current noise spectral density would have a similar characteristic). While the response is flat at higher frequencies (above about 100 Hz, the so-called 1/f frequency), the noise spectral density increases as the frequency approaches dc. To calculate the resulting RTI rms noise, the noise spectral density is multiplied by the square root of the bandwidth of interest. The bandwidth may be the bandwidth of the in amp at that particular gain or it may be something less. For example, if the output signal from the in amp is low pass filtered, the corner frequency of the filter will define the bandwidth of interest. Note that if the output of the instrumentation amplifier is being digitized in an analog-to-digital converter (ADC), any post filtering should also be factored into calculating the bandwidth of interest.

In high frequency applications, low frequency noise generally can be neglected. So the RTI rms noise would simply be the product of the "flat" noise spectral density and the square root of the bandwidth. Note that the calculated rms noise must be converted to peak-to-peak by multiplying the rms value by 6.6[1]. For low frequency applications, data sheets usually specify peak-to-peak noise in the 0.1 Hz to 10 Hz frequency band. If high frequency noise is filtered at some point in the system, it can be neglected so that only the 0.1 Hz to 10 Hz noise is counted.

Because voltage and current noise are uncorrelated (i.e., are random and bear no relationship to one another), the overall error due to noise should not be simply the sum of all noise errors. It is more accurate to do a root-sum-square calculation of the total noise error.

$$\textit{Total noise} = \sqrt{\textit{Voltage Noise}^2 + R_{SOURCE} \times \textit{Current Noise}^2}$$

Linearity

This error will generally be specified in ppm (for a particular input span) in the data sheets of integrated instrumentation amplifiers. In the case of discrete in-amps, built from op amps, the nonlinearity is more difficult to calculate. Op amp data sheets generally do not specify linearity. Furthermore, even if the linearity of one op amp is known, it is necessary to estimate how the two or three op amps will interact to give an overall linearity specification. In many cases, the only option is to measure the linearity of the circuit by doing a dc sweep. The linearity in ppm will be given by the expression.

Nonlinearity (ppm) = (Maximum deviation of output voltage from ideal/gain/full-scale input) $\times 10^6$

Gain Error

The gain error of an integrated instrumentation amplifier will have two components, internal gain error and error due to the tolerance of the external gain setting resistor. While a precision external gain resistor will prevent the overall gain from degrading, there is little point in wasting cost on an external resistor which is much more accurate than the in amp's accuracy. It is also generally difficult to achieve exactly the desired gain (e.g., 10 or 100) when using standard resistor sizes.

It should be noted however that the choice of gain resistor can help to reduce the overall gain drift of the circuit. As an example, let's consider the AD623. Its gain is given by the equation:

$$\textit{Gain} = 1 + (100\ k\Omega / R_G)$$

The 100 kΩ value in this equation stems from two internal 50 kΩ resistors. These resistors have a temperature coefficient of –50 ppm/°C. By choosing an external gain resistor, which also has a negative temperature coefficient, the gain drift can be reduced.

Error Budget Analysis of Two Integrated In Amps: AD623 vs. INA126

Figure 3 shows the popular resistive bridge application. The bridge consists of four variable resistors. The bridge is excited by a single +5 V supply. Any change in the resistance of the variable resistors will generate a differential voltage (±20 mV full scale) which appears across the input of the in amp. The common-mode voltage of the differential signal is +2.5 V. This follows from the +5 V excitation voltage.

The gain of the in amp has been set so that the output signal swing is close to its maximum level but is still not clipping. Care should be taken not to saturate any of the internal nodes of the in amp. This is a function of the differential input voltage, the programmed gain and the common-mode level[2].

Table II shows the error budget calculations using the integrated in amps AD623 and INA126. All errors are referred to the input (i.e., are compared to the full-scale input voltage of 20 mV) and are then converted to parts per million (ppm) by multiplying the fractional error by 1×10^6. Alternatively, percentage errors are converted to ppm by multiplying by 1×10^4. Conversion between fractional, percentage and ppm is shown in Table III.

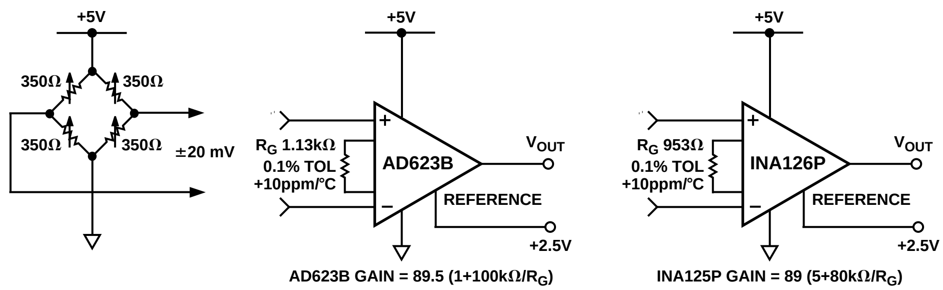

Figure 3. Amplifying the Differential Voltage from a Resistive Bridge Transducer

Table II. Error Budget Analysis: AD623 vs. INA126

Error Source	AD623B Circuit Calculation	INA126P Error Calculation	Total Error AD623 (ppm)	Total Error INA126 (ppm)
ABSOLUTE ACCURACY at T_A = +25°C				
Input Offset Voltage, mV	100 µV/20 mV	250 µV/20 mV	5,000	12,500
Output Offset Voltage, mV	500 µV/89.5/20 mV	Not Applicable	279	Not Applicable
Input Offset Current, nA	2 nA × 350 Ω/20 mV	2 nA × 350 Ω/20 mV	35	35
CMR, dB	105 dB→5.6 ppm × 2.5 V/20 mV	83 dB→71 ppm × 2.5 V/20 mV	700	8875
Gain	0.35% + 0.1%	0.5% + 0.1%	4500	6000
		Total Absolute Error	**10514**	**27410**
DRIFT TO +85°C				
Gain Drift, ppm/°C	(50 + 10) ppm/°C × 60°C	(100 + 10) ppm/°C × 60°C	3600	6600
Input Offset Voltage, mV/°C	1 µV/°C × 60°C/20 mV	3 µV/°C × 60°C/20 mV	3000	9000
Input Offset Current, pA/°C	5 pA/°C × 350 Ω × 60°C/20 mV	10 pA/°C × 350 Ω × 60°C/20 mV	5.25	10.5
Output Offset Voltage Drift, mV/°C	10 µV/°C × 60°C/ 89.5/20 mV	Not Applicable	335	Not Applicable
		Total Drift Error	**6940**	**15610**
RESOLUTION				
Gain Nonlinearity, ppm of Full Scale	50 ppm	20 ppm	50	20
Typ 0.1 Hz–10 Hz Voltage Noise, mV p-p	1.5 µV p-p/20 mV	0.7 µV p-p/20 mV	75	35
		Total Resolution Error	**125**	**55**
		Grand Total Error	**25859**	**43075**

Table III. Conversion Between ppm, Fractional Error and Percentage Error

% Error	Fractional Error	ppm Error
10	0.1	100000
1	0.01	10000
0.1	0.001	1000
0.01	0.0001	100
0.001	0.00001	10
0.0001	0.000001	1

Table II shows that the predominant error source is static errors (e.g., offset voltage, etc.). In many applications where some form of calibration is available, these errors can be removed. With the addition of some kind of ambient temperature measurement, this calibration can be extended to compensate for drift of static errors. It is more difficult to compensate for errors in resolution caused by the nonlinearity and noise of the in amp. Note that errors due to current noise have been neglected.

These errors are quite small and become insignificant when they are quadratically summed with the voltage noise.

Additional resolution errors that occur due to external interference cannot be characterized here. Significant in this area is the degradation in resolution that will be caused by common-mode pickup on the differential inputs of 50 Hz or 60 Hz interference (from lights or any equipment running on the mains). This will result in the 50 Hz/60 Hz hum being visible on the in amp's output. Obviously, high common-mode rejection, not just at dc but also over frequency, will help to minimize this interference. The common-mode rejection over frequency of the AD623 is shown in Figure 4. This shows for example, that the CMRR at 1 kHz, for a gain of 10, is still over 80 dB, more than sufficient for most applications.

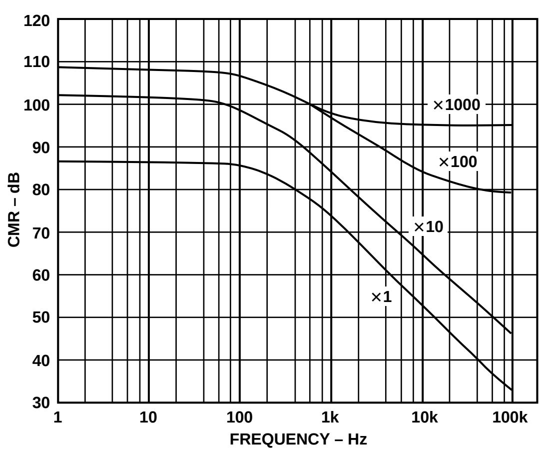

Figure 4. AD623 CMR vs. Frequency, +5 V Single Supply, V_{REF} = 2.5 V, Gain = 1, 10, 100, 1000

Make vs. Buy: A Typical Application Error Budget

The example in Figure 5 serves as a good comparison between the errors associated with an integrated and a discrete in amp implementation. Again, we have a ±20 mV signal we want to amplify. Using a dual op amp and a precision resistor network, a two op amp in amp can be implemented.

The errors associated with each implementation are detailed in Table IV and show the integrated in amp to be more precise, both at ambient and over temperature. It should be noted that the discrete implementation is quite a bit more expensive (by about 100% in this example). This is primarily due to the cost of the low drift precision resistor network.

Note, the input offset current of the discrete in amp implementation is the maximum difference in the bias currents of the two op amps, not the offset currents of the individual op amps. Also, while the values of the resistor network are chosen so that the inverting and noninverting inputs of each op amp see the same impedance (about 350 Ω), the offset current of each op amp will add an additional error which must be characterized.

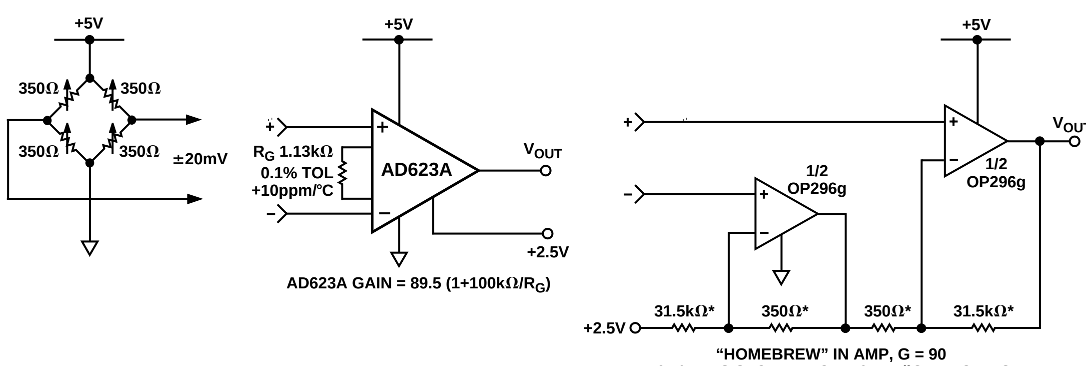

Figure 5. Make vs. Buy

Table IV. Make vs. Buy Error Budget

Error Source	AD623A Circuit Calculation	"Homebrew" Circuit Calculation	Total Error AD623–ppm	Total Error Homebrew–ppm
ABSOLUTE ACCURACY at T_A = +25°C				
Total RTI Offset Voltage, mV	(200 μV+[1000 μV/89.5])/20 mV	(300 μV × 2)/20 mV	10,559	30,000
Input Offset Current, nA	2 nA × 350 Ω/20 mV	100 nA × 350 Ω/20 mV*	35	1750
Internal Offset Current (Homebrew Only)	Not Applicable	16 nA × 350 Ω/20 mV*		280
CMR, dB	105 dB→5.6 ppm × 2.5 V/20 mV	(0.1% Match × 2.5 V)/ 90/20 mV	700	1388
Gain	0.35% + 0.1%	0.1% Match	4500	1000
		Total Absolute Error	**15794**	**34418**
DRIFT TO +85°C				
Gain Drift, ppm/°C	(50 + 10) ppm/°C × 60°C	50 ppm/°C × 60°C	3600	3000
Total RTI Offset Voltage, mV/°C	(2 μV/°C+[10 μV/°C/89.5]) × 60°C/20 mV	(7 μV/°C × 60°C)/20 mV	6335	21,000
Input Offset Current, pA/°C	5 pA/°C × 350 Ω × 60°C/20 mV	Not Applicable*	5.25	
Internal Offset Current (Homebrew Only)	Not Applicable	(120 pA/°C × 350 Ω × 2) × 60°C/20 mV		252
		Total Drift Error	**9940**	**24252**
RESOLUTION				
Gain Nonlinearity, ppm of Full Scale	50 ppm	20 ppm	50	20
Typ 0.1 Hz–10 Hz Voltage Noise, mV p-p	1.5 μV p-p/20 mV	(0.8 μV p-p × $\sqrt{2}$)/20 mV	75	56.57
		Total Resolution Error	**125**	**77**
		Grand Total Error	**25859**	**58747**

*Error over temperature. bias current of OP296 is only specified as a maximum value over temperature (i.e., no value specified at + 25°C).

REFERENCES

1. *Analog-Digital Conversion Handbook, Third Edition*, pp 550–553, by the Engineering Staff of Analog Devices, Inc., edited by Daniel H. Sheingold, Prentice Hall, Englewood Cliffs, NJ 07632.

2. AD623 Single Supply, Rail-to-Rail, Low Cost Instrumentation Amplifier, data sheet, p 15.

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