Voltage reference selection basics

Texas Instruments

Voltage references are key components in data conversion systems which enable the ADC and DAC to read accurate values and are used in various sensing applications.

While they are simple parts which often may only have two or three pins, there are numerous parameters that affect the performance of a reference, and careful consideration of all the parameters are required to select the proper one. This paper covers the differences between the shunt and series architectures, explains key parameters and special features, and shows how to properly calculate the error for a given reference in a given operating condition.

Choosing the best shunt or series reference for your application

Voltage references are a key building block in data conversion systems. Understanding their specifications and how they contribute to error is necessary for selecting the right reference for the application. **Figure 1** shows the application of a voltage reference in a simple analog-to-digital converter (ADC) and digital-to-analog converter (DAC). In each case, the reference voltage (V_{REF}) acts as a very precise analog 'meter stick' against which the incoming analog signal is compared (as in an ADC) or the outgoing analog signal is generated (as for a DAC). As such, a stable system reference is required for accurate and repeatable data conversion; and as the number of bits increases, less reference error can be tolerated. Monolithic voltage references produce an output voltage which is substantially immune to variations in ambient temperature as well as loading, input supply, and time. While many ADCs and DACs incorporate an internal reference, beyond 8 to 10 bits it is rare to find one with sufficient precision as high-density CMOS technologies commonly used for data converters typically produce low-

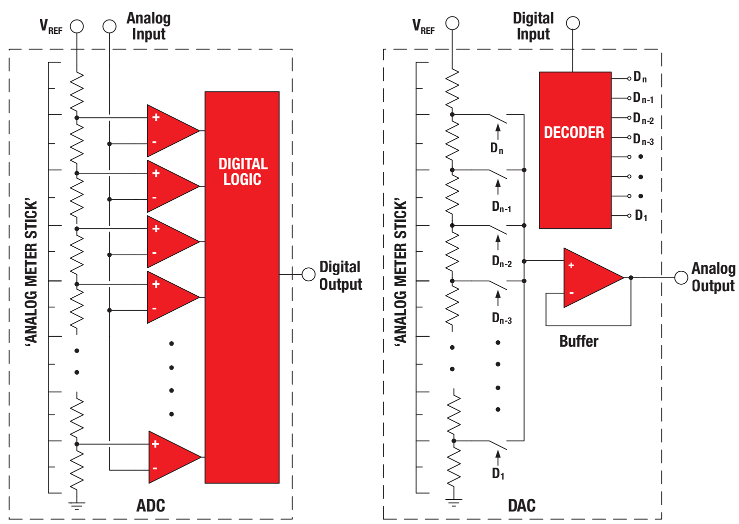


Figure 1. Simplified ADC/DAC diagrams.

quality references. In most cases, the internal reference can be overdriven by an external one to improve performance. Terms such as "high precision" and "ultra-high precision" are common in reference datasheets but do little to help designers in their selection. This article seeks to provide an explanation of common reference specifications, rank their relative importance and show how a designer can use them in some simple calculations to narrow the search.

Figure 2 shows the two available voltage references topologies: series and shunt. A series reference provides load current through a series transistor located between V_{IN} and V_{REF} (Q1), and is basically a high-precision, low-current linear regulator. A shunt reference regulates V_{REF} by shunting excess current to ground via a parallel transistor (Q2). In general, series references require less power than shunt references because load current is provided as it is needed. The bias current of a shunt reference (I_{BIAS}) is set by the value of R_{BIAS} and must be greater than or equal to the maximum load current plus the reference's minimum operating current (the minimum bias current required for regulation). In applications where the maximum load current is low (e.g. below 100 µA to 200 µA), the disparity in power consumption between series and shunt references shrinks. There is no inherent difference in accuracy between the two topologies and high- and low-precision examples are available in both varieties.

The advantages and disadvantages of the two architectures are summarized in **Table 1**. Generally, shunt references offer more flexibility (V_{IN} range,

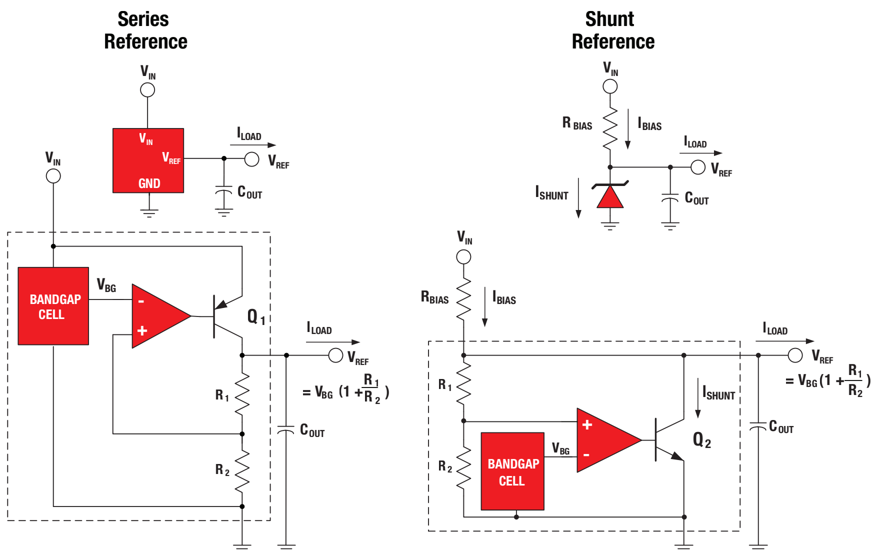


Figure 2. *Circuit symbols and simplified schematics of series and shunt architectures.*

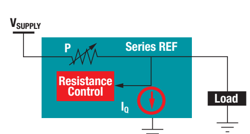
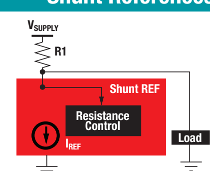
	Series References	Shunt References
Diagram		
Number of Terminals	At least 3	At least 2
Advantages	• Significantly lower power dissipation • Generally higher precision • Low I_Q • Low dropout	• Wide V_{IN} tolerant with proper resistor selection • Can be used to create negative or floating reference voltage • Inherent current sourcing and sinking
Disadvantages	• Limited max V_{IN}	• V_{IN} current fixed at max load • No shutdown mode
Key Markets	• Factory automation, grid, medical, test	• Isolated power supplies, adapters, automotive
TI Nomenclature	• LM41xx, REFxxxx	• LM40xx-N, LM(V)431, LM1/2/385, LM1/2/336 • TL43xLI, ATL43xLI, TL(V)43x, TLVH43x

Table 1. *Series vs shunt architectures.*

creation of negative or floating references) and better power supply rejection at the expense of higher power consumption while series references typically dissipate less power and can achieve better performance for extremely high-precision applications. The typical application diagram of data converters will often show a zener diode symbol representing the reference, indicating the use of a shunt reference. This is merely a convention, and in nearly all cases a series reference could be used as well.

Voltage reference specifications

As with all ICs, voltage references have standardized parameters for determining the right part for a design. The following are key specifications in order of importance that are pertinent to all suppliers.

1. Temperature coefficient

The variation in V_{REF} over temperature is defined by its temperature coefficient (TC, also referred to as “drift”) which has units of parts-per-million per degree Celsius (ppm/°C). It is convenient to represent the reference voltage temperature dependence as a polynomial for the sake of discussion:

$$V_{REF}(T) = V_{REF}\big|_{25^\circ C}\left(1 + TC_1\left(\frac{T}{25^\circ C}\right) + TC_2\left(\frac{T}{25^\circ C}\right)^2 + TC_3\left(\frac{T}{25^\circ C}\right)^3 + \ldots\right)$$

TC1 represents the first-order (linear) temperature dependence, TC2 the second-order, and so on. Higher than first-order terms are usually lumped together and described as the “curvature” of the drift.

The majority of monolithic references are based on a bandgap reference. A bandgap reference is created when a specific Proportional To Absolute Temperature (PTAT) voltage is added to the Complementary To Absolute Temperature (CTAT) base-to-emitter voltage of a bipolar transistor yielding a voltage at roughly the bandgap energy of silicon (~1.2 V) where TC1 is nearly zero. Neither the PTAT nor CTAT voltage is perfectly linear leading to non-zero higher-order TC coefficients, with TC2 usually being dominant. References designed for drifts less than 20 ppm/°C generally require special circuitry to reduce TC2 (and possibly higher-order terms), and their datasheets will often mention some form of “curvature correction.” Another common type of reference is based on a buried-zener diode voltage plus a bipolar base-to-emitter voltage to produce a stable reference voltage on the order of 7 V. The drift performance of buried-zener references is on par with that of bandgap references, although their noise performance is superior. Buried-zener references usually require large quiescent currents and must have an input supply greater than 7.2 V, so they cannot be used in low-voltage applications (V_{IN} = 3.3 V, 5 V, etc.).

The temperature coefficient can be specified over several different temperature ranges, including the commercial temperature range (0 to 70°C), the industrial temperature range (-40 to 85°C), and the extended temperature range (-40 to 125°C). There are several methods of defining TC, with the “box” method being used most often. The box method calculates TC using the difference in the maximum

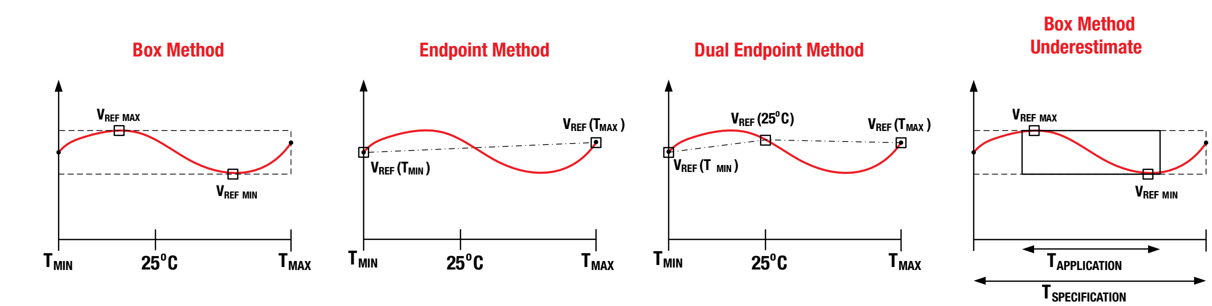


Figure 3. *Different methods for TC calculation.*

and minimum V_{REF} measurements over the entire temperature range whereas other methods use the values of V_{REF} at the endpoints of the temperature range (T_{MIN}, T_{MAX}).

Neither method is ideal. The weakness of the endpoints method is the failure to account for any curvature in the drift (TC2, TC3, etc.). Calculating the incremental TC from room temperature to both the minimum and maximum temperatures improves the situation as information on TC2 can be garnered using three data points rather than two. While the box method is more accurate than using endpoints, it may underestimate TC if the temperature range of the application is smaller than the range over which the TC is specified.

$$TC_{BOX} = 10^6 \cdot \left(\frac{V_{REF\_MAX}|_T - V_{REF\_MIN}|_T}{V_{REF}|_{25^\circ C}} \right) \cdot \left(\frac{1}{T_{MAX} - T_{MIN}} \right)$$

$$TC_{ENDPOINTS} = 10^6 \cdot \left(\frac{V_{REF}|_{TMAX} - V_{REF}|_{TMIN}}{V_{REF}|_{25^\circ C}} \right) \cdot \left(\frac{1}{T_{MAX} - T_{MIN}} \right)$$

2. Initial accuracy

The initial accuracy of V_{REF} indicates how close to the stated nominal voltage the reference voltage is guaranteed to be at room temperature under stated bias conditions. It is typically specified as a percentage and ranges from 0.01% to 1% (100-10,000 ppm). For example, a 2.5 V reference with 0.1% initial accuracy should be between 2.4975 V and 2.5025 V when measured at room temperature. The importance of initial accuracy depends mainly on whether the data conversion system is calibrated. Buried-zener references have very loose initial accuracy (5-10%) and will require some form of calibration.

3. 0.1-10 Hz peak-to-peak noise

The internally-generated noise of a voltage reference causes a dynamic error that degrades the signal-to-noise ratio (SNR) of a data converter, reducing the estimated number of bits of resolution (ENOB). Datasheets provide separate specifications for low- and high-frequency noise. Broadband noise is typically specified as an rms value in microvolts over the 10 Hz to 10 kHz bandwidth. Broadband noise is the less troublesome of the two as it can be reduced to some degree with a large V_{REF} bypass capacitor. Broadband noise may or may not be important in a given application depending on the bandwidth of the signal the designer in interested in. Low-frequency V_{REF} noise is specified over the 0.1 Hz to 10 Hz bandwidth as a peak-to-peak value (in µV or ppm). Filtering below 10 Hz is impractical, so the low-frequency noise contributes directly to the total reference error. Low-frequency noise is characterized using an active bandpass filter composed of a 1st-order high-pass filter at 0.1 Hz followed by an nth-order low-pass filter at 10 Hz. The order of the low-pass filter has a significant effect on measured peak-to-peak value. Using a 2nd-order low pass at 10 Hz will reduce the peak-to-peak value by 50 to 60% compared to a 1st-order filter.

Some manufacturers use up to 8th-order filters, so a designer should read the datasheet notes carefully when comparing references. From a design perspective, the 0.1 Hz to 10 Hz noise is mainly due to the flicker (1/f) noise of the devices and resistors in the bandgap cell, and therefore scales linearly with V_{REF}. For example, a 5 V reference will have twice the peak-to-peak noise voltage as the 2.5 V option of the same part. Reducing the noise requires considerably higher current and larger devices in the bandgap cell, so very low noise references (<5 µVPP) often have large quiescent currents (hundreds of microamps to milliamps) and tend to be in larger packages. Buried zener references have the best noise performance available because no gain is required to generate the output voltage. Bandgap cells typically have a closed loop gain of 15 V/V to 20 V/V, causing device and resistor noise to be amplified.

4. Thermal Hysteresis

Thermal hysteresis is the shift in V_{REF} due to one or more thermal excursions and is specified in parts-per-million. A thermal excursion is defined as an excursion from room temperature to a minimum or to a maximum temperature and finally back to room temperature (for example 25ºC to -40ºC to 25ºC or 25ºC to 125ºC to 25ºC). The temperature range (commercial, industrial, extended) and definition of thermal excursion may vary by manufacturer, making direct comparison difficult. Thermal excursions over a wider temperature range typically lead to a larger shift in V_{REF}. Even if the temperature range of the application is narrow, the thermal excursion when soldering the IC to the PCB and any subsequent solder reflows will induce thermal hysteresis.

The main cause of thermal hysteresis is thermo-mechanically induced die stress and therefore is a function of not only the temperature excursion, but also the package, molding compound, die-attach material, as well as the IC layout itself.

As a rule of thumb, references in larger molded packages tend to exhibit lower thermal hysteresis. Thermal hysteresis is not tested in production and datasheets only provide a typical shift value.

5. Long Term Stability

Long-term stability describes the typical shift in V_{REF} after 1000 hours (6 weeks) of continuous operation under nominal conditions. It is meant to give the designer a rough idea of the stability of the reference voltage over the life of the application. The prevailing wisdom is that the majority of the shift in V_{REF} occurs in the first 1000 hours as longterm stability is related logarithmically with time. A six-week test time is not feasible in production, so long-term stability is characterized on a small sample of parts (15 to 30 units) at room temperature and the typical shift is specified. Once a reference is soldered down on a PCB, changes in the board stress can also cause permanent shifts in V_{REF}. Board stress dependence is not currently captured in datasheets, so the designer should locate the reference on a portion of the PCB least prone to flexing. Different packages will have different sensitivity to stress; metal cans are largely immune, and surface-mount plastic packages become progressively more sensitive the smaller the package (for example, the same die will perform better in an SO-8 than an SC70 package).

6. Load Regulation

Load regulation is the measure of the variation in V_{REF} as a function of load current and is specified either as a percentage or in parts-per-million per milliamp (ppm/mA). It is calculated by dividing the relative change in V_{REF} at minimum and maximum load currents by the range of the load current.

Load regulation depends on both the design of the reference and the parasitic resistance separating it from the load, so the reference should be placed

$$LOAD\_REG(ppm/mA) = 10^6 \cdot \left(\frac{V_{REF}\ I_{ILOAD\_MAX} - V_{REF}\ I_{ILOAD\_MIN}}{V_{REF}\ I_{ILOAD\_MIN}} \right) \left(\frac{1}{I_{LOAD\_MAX} - I_{LOAD\_MIN}} \right)$$

as close to the load as the PCB layout will allow. References with pins for both forcing and sensing V_{REF} provide some immunity to this problem. The impedance of the reference input is large enough (>10 kΩ) on many data converters that load regulation error may not be significant. Maximum load current information can be found in ADC/DAC datasheets specified as either a minimum reference pin resistance (R_{REF}) or a maximum reference current (I_{REF}). In situations where the reference is buffered with a high-speed op amp, load regulation error can usually be ignored. The dual of load regulation for shunt references is the ‘change in reverse breakdown voltage with current’ that specifies the change in V_{REF} as a function of the current shunted away from the load. It is calculated with the same equation as load regulation where load current is replaced with shunted current (I_{SHUNT}). The amount of shunted current depends on both the load current and the input voltage so the ‘change in reverse voltage with current’ specification also indicates line sensitivity.

7. Line Regulation

Line regulation applies only to series voltage references and is the measure of the change in the reference voltage as a function of the input voltage. The importance of line regulation depends on the tolerance of the input supply. In situations where the input voltage tolerance is within 10% or less, it may not contribute significantly to the total error.

$$LINE\_REG = 10^6 \cdot \left(\frac{V_{REF}\,|_{VIN\_MAX} - V_{REF}\,|_{VIN\_MIN}}{V_{REF}\,|_{VIN\_MIN}} \right) \cdot \left(\frac{1}{V_{IN\_MAX} - V_{IN\_MIN}} \right)$$

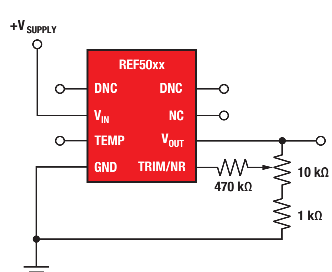
The extension of line regulation over frequency is the Power Supply Rejection Ratio (PSRR). PSRR is rarely specified but typical curves are usually provided in the datasheet. As with line regulation, the importance of PSRR depends on specifics of the input supply. If V_{IN} is noisy (generated with a switching regulator, sensitive to EMI, subject to large load transients), PSRR may be critical. The analogous specification for shunt references is the reverse dynamic impedance, which indicates the sensitivity of V_{REF} to an AC current. Noise on the supply powering a shunt reference is converted to a noise current through R_{BIAS}. Some shunt reference datasheets will specify the reverse dynamic impedance at 60 Hz and 120 Hz, and nearly all will provide a plot of reverse dynamic impedance versus frequency.

8. Special Features

In applications where power consumption is crucial, a series reference is usually the right choice. The quiescent current of most series references ranges from 25 µA to 200 µA, although several are available with I_Q<1 µA. Low quiescent current generally comes at the expense of precision (TC and initial accuracy) and higher noise. Some series references can also be disabled via an external ENABLE / SHUTDOWN pin causing the quiescent current to drop to a few microamps or less when V_{REF} is not needed. A power-saving mode is not possible in shunt references. Additionally, series references can have dropout voltages less than 200 mV, allowing them to be used at lower input voltages. Shunt references can also be used at low voltages, but the bias current may vary widely with changes in V_{IN} due to the small RBIAS resistor required.

References do not require many external passive components but proper selection can improve performance. A bypass capacitor on V_{REF} substantially improves PSRR (or reverse dynamic impedance in the case of a shunt reference) at higher frequencies. It will also improve the load transient response, and reduce high-frequency noise. Generally speaking, the best performance is achieved with the largest bypass capacitor allowed. The range of allowable bypass capacitors depends on the stability of the reference, which should be detailed along with ESR restrictions in the component selection section of the datasheet. When using a large bypass capacitor (>1 µF) it may be advantageous to bypass it with a smaller value, lower-ESR capacitor to reduce the effects of the ESR and ESL. The reverse dynamic impedance of a shunt reference varies inversely with the amount of current shunted. If noise immunity is more important than power consumption in a given application, a smaller R_{BIAS} may be used to increase I_{SHUNT}.

Some series references have a TRIM/NOISE REDUCTION (NR) pin to further enhance performance. By using a series of resistors shown in **Figure 4**, the TRIM/NR pin can be used to adjust output voltage by up to ±15 mV. The pin can also be used to create a low pass filter to decrease overall noise measured on V_{OUT} by using a capacitor as shown in **Figure 5**. Note that increasing the capacitor size will continue to improve noise performance, but also increases startup time.



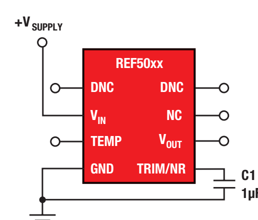


Figure 4. *V_{OUT} adjustment using the TRIM/NR pin (left).*

Figure 5. *Noise reduction using the TRIM/NR pin (right).*

9. Other Considerations

Voltage references are becoming increasingly popular in the automotive space and as reliability becomes more critical, the need for AEC-Q100 qualified parts increases as well. The AEC-Q100 qualification was created by the Automotive Electronics Council and requires specific quality and testing procedures to ensure a device's performance. On top of the general qualification, there are also grades from 0-4 which indicate the temperature range of the qualified device. At the time of this article, the most common is grade 1 where the operating temperature range of the device spans from -40 to 125°C. It should be noted that many manufacturers offer voltage references in the extended temperature range of -40 to 125°C and mention that the device is "suitable for automotive applications" but this does not mean that the device is AEC-Q100 grade 1-qualified. Therefore, designers for automotive applications should be cautious of devices that state suitability for automotive applications unless it is specifically specified in the datasheet as AEC-Q100 qualified.

Due to the popularity of voltage references, many manufacturers have produced "identical" products and sometimes even release them as the same part number as the original manufacturer for part recognition. While the main performance characteristics are identical, there are times when there are subtle differences which could be of concern to a designer depending on their application. For example, the LM4040 offered from Texas Instruments has a wideband noise value of 35 μV_{RMS} for a 2.5V output. The same part from another supplier has a wideband noise value of 350 µVRMS which is 10x the value even though the initial accuracy and temperature coefficient are the same. Other differences that are prevalently found are operating temperature range and current consumption. A designer should be careful when considering secondary or alternate parts with the same part number and should do a thorough comparison of the performance to ensure that the key parameters are indeed the same.

Voltage reference selection begins with satisfying the application operating conditions, specifically: nominal V_{REF}, V_{IN} range, current drive, power consumption, and package size. Beyond that, a reference is chosen based on the accuracy requirements of a given data converter application. The most convenient unit for understanding how the reference error affects accuracy is in terms of the least significant bit (LSB) of the data converter. The LSB in

$$\mathrm{LSB(ppm)} = 10^6 \cdot \left(\frac{1}{2}\right)^{\mathrm{NOB}} \qquad \mathrm{NOB:Number\ of\ Bits}$$

Bits	LSB (ppm)
8	3906
10	977
12	244
14	61
16	15

Table 2. *LSB values in PPM for common data converter resolutions.*

units of parts-per-million is simply one million divided by two raised to the number of bits power. **Table 2** provides the LSB values for common resolutions.

ADCs / DACs have their own sources of error including integral nonlinearity (INL), differential on linearity (DNL), and gain and off set error. If we consider the case of the more common unipolar data converter, voltage reference error is functionally equivalent to a gain error. INL and DNL gauge the nonlinearity of a data converter, on which the reference voltage has no effect. The gain and off set errors can be understood conceptually by recognizing that ADCs / DACs have two reference voltages: V_{REF} and GND in the case of a unipolar data converter, and V_{REF} and $-V_{REF}$ for bipolar data converters. The offset error is the deviation in the output (in bits for an ADC and voltage for a DAC) from the ideal minus full-scale (MFS) value when a MFS input is applied. The MFS reference voltage is GND so V_{REF} error has no effect. The gain error is the deviation from the ideal positive full-scale (PFS) output for a PFS input, minus the offset error. The PFS reference voltage is V_{REF}, so any shift in the reference voltage equates to a gain error. The types of shifts in a V_{REF} that affect PFS are errors such as initial accuracy, temperature coefficient, and long term drift. As such, the reference error can cause loss of dynamic range for input signals near PFS, which is also where it has the most pronounced effect on accuracy. Therefore when designing a system it is necessary to account for these errors for signals that approach PFS. The effect of reference error on a mid-scale (MS) input signal is half that for a PFS input, and is negligible for inputs near MFS. For example, a worst-case reference error of 8 LSB would result in a loss of 3 bits of accuracy for a PFS input, 2 bits of lost accuracy at mid-scale, and no loss of accuracy at MFS. For example, in a system where dynamic range is the priority, it will be important to use the largest acceptable voltage reference to minimize the accuracy loss from PFS. If the designer has no idea what kind of reference error they can live with, matching the worst-case reference error to the maximum gain error is a reasonable starting point. In systems where error contributors are statistically independent, and consequently add together as a root mean squared sum, balancing the error contributions represents the optimal case. Otherwise, the error will tend to be dominated by one variable and the accuracy of the other variable(s) is essentially wasted.

In calculating the total error in V_{REF} it is helpful to separate the specifications where a maximum value is guaranteed (TC, initial accuracy, load regulation, line regulation) and those where only a typical value is provided (0.1 Hz to 10 Hz noise, thermal hysteresis, and long-term stability). Other than initial accuracy, the guaranteed specifications are all linear coefficients and their contribution to the total error can be calculated based on the operating ranges of the reference (temperature range, load current, and input voltage).

$$\text{ERROR}\ |_{\text{TEMP}} = \text{TC} \cdot (T_{\text{MAX}} - T_{\text{MIN}})$$
$$\text{ERROR}\ |_{\text{LOAD}} = \text{LOAD\_REG} \cdot (I_{\text{LOAD\_MAX}} - I_{\text{LOAD\_MIN}})$$
$$\text{ERROR}\ |_{\text{LINE}} = \text{LINE\_REG} \cdot (V_{\text{IN\_MAX}} - V_{\text{IN\_MIN}})$$
$$\text{ERROR}\ |_{\text{GUARANTEED}} = \text{ERROR}\ |_{\text{INITIAL\_ACCURACY}} + \text{ERROR}\ |_{\text{TEMP}} + \text{ERROR}\ |_{\text{LOAD}} + \text{ERROR}\ |_{\text{LINE}}$$

Example using REF3425

$$\text{ERROR}\ |_{\text{TEMP}} = (6\ \text{ppm}/^{\circ}\text{C}) \cdot (55\,^{\circ}\text{C} - 0^{\circ}\text{C}) = 300\ \text{ppm}$$
$$\text{ERROR}\ |_{\text{LOAD}} = (40\ \text{ppm/mA}) \cdot (0.5\ \text{mA} - 0\ \text{mA}) = 20\ \text{ppm}$$
$$\text{ERROR}\ |_{\text{LINE}} = (15\ \text{ppm/V}) \cdot (5.5\ \text{V} - 4.5\ \text{V}) = 15\ \text{ppm}$$
$$\text{ERROR}\ |_{\text{GUARANTEED}} = (0.05\% = 500\ \text{ppm}) + 300\ \text{ppm} + 20\ \text{ppm} + 15\ \text{ppm} = 835\ \text{ppm}$$

In calibrated systems, initial accuracy can be dropped from the equation. The above calculation represents the worst case, and most of the time a reference will perform better than the guaranteed maximums (especially when it comes to line and load regulation where the maximum may be more a function of the testing system due to the very low signal-to-noise ratio of the measurement). It is worth noting that the statistical methods through which the guaranteed maximum specifications are calculated vary by manufacturer, so comparing datasheets may not tell the full story. If the designer wants to estimate the average reference error, they can take the rms sum of the individual error sources rather than just adding them up. For simplicity, one can convert from peak to peak to rms value by using the equation peak-to-peak = 6.6x rms. This is because 99.99% of the error will fall within 6.6 standard deviations of the total error distribution. In most cases, the TC error will be dominant, so TC error by itself gives a good indication of average reference performance.

Datasheets only provide typical values for thermal hysteresis and long-term stability,

$$\mathrm{ERROR}\ |_{\mathrm{THERMAL\_HYSTERESIS}} \approx (\mathrm{Typ.\ Thermal\ Hysteresis})$$

$$\mathrm{ERROR}\ |_{\mathrm{LONG-TERM\_STABILITY}} \approx (\mathrm{Typ.\ Long\text{-}Term\ Stability})$$

$$\mathrm{ERROR}\ |_{\mathrm{LF\_NOISE}} = 10^{6} \cdot \left(\frac{0.1-10\ \mathrm{Hz\ Peak-to-Peak}}{V_{\mathrm{REF}}} \right)$$

$$\mathrm{ERROR}\ |_{\mathrm{TOT}} = \mathrm{ERROR}\ |_{\mathrm{GUARANTEED}} + \mathrm{ERROR}\ |_{\mathrm{THERMAL\_HYSTERESIS}} + \mathrm{ERROR}\ |_{\mathrm{LONG-TERM\_STABILITY}} + \mathrm{ERROR}\ |_{\mathrm{LF\_NOISE}}$$

Example using REF3425

$$\mathrm{ERROR}\ |_{\mathrm{THERMAL\_HYSTERESIS}} = 30\ \mathrm{ppm}$$

$$\mathrm{ERROR}\ |_{\mathrm{LONG-TERM\_STABILITY}} = 25\ \mathrm{ppm}$$

$$\mathrm{ERROR}\ |_{\mathrm{LF\_NOISE}} = 5\ \mathrm{ppm}$$

$$\mathrm{ERROR}\ |_{\mathrm{TOT}} = 835\ \mathrm{ppm} + 30\ \mathrm{ppm} + 25\ \mathrm{ppm} + 5\ \mathrm{ppm} = 895\ \mathrm{ppm}$$

but both are likely to vary a great deal unit to unit. The typical specification is not very helpful in estimating worst-case error without knowing the standard deviation of the distribution. Many times this information can be obtained by calling the manufacturer. Otherwise, a conservative, albeit crude, approach would be to multiply the typical specification by three or four to get a ballpark estimate of the worst-case shift. This is assuming that the standard deviation of the distribution is on the order of the mean value and designing for a two or three standard deviation worst case. The loss of resolution due to noise is harder to predict and can only really be known by testing a reference in the application.

Low-frequency noise should be very consistent on a unit-to-unit basis and no 'sand-bagging' of the typical value is required. Over a 10-second window, one can expect the V_{REF} to shift by an amount equal to the 0.1 Hz to 10 Hz peak-to-peak specification.

Once the worst-case reference error in parts-per-million is estimated, it can be converted into LSB for different data converter resolutions using the values in **Table 2**. The worst-case accuracy loss at positive full-scale and mid-scale can then be calculated taking the log base-2 of the number of LSB of error.

$$\text{ERROR } I_{TOTAL}\ (\text{LSB}) = \frac{895\ \text{ppm}}{\left(\text{ppm}/\text{LSB}\right)} = 0.9\ \text{LSB (10 bit)} = 3.6\ \text{LSB (12 bit)} = 14.6\ \text{LSB (14 bit)}$$

$$\text{Worst Case Lost Accuracy} = \log_2\left(\text{ERROR } I_{TOTAL}\ (\text{LSB})\right)$$

$$\text{Worst Case Lost Accuracy (PFS)} = 0\ \text{bit (10 bit)} = 1.8\ \text{bits (12 bit)} = 3.8\ \text{bits (14 bit)}$$

$$\text{Worst Case Lost Accuracy (MS)} = 0\ \text{bit (10 bit)} = 0.8\ \text{bits (12 bit)} = 2.8\ \text{bits (14 bit)}$$

If the average rather than the worst case is considered, the rms sum of reference error contributors can be taken (replacing the maximum values for the typical values).

$$\text{ERROR } I_{RMS} = \sqrt{(300)^2 + (500)^2 + (20)^2 + (15)^2 + (30)^2 + (25)^2 + (5)^2} = 585\ \text{ppm}$$

$$\text{ERROR } I_{RMS}\ (\text{LSB}) = \frac{585\ \text{ppm}}{\left(\text{ppm}/\text{LSB}\right)} = 0.6\ \text{LSB (10 bit)} = 2.4\ \text{LSB (12 bit)} = 9.6\ \text{LSB (14 bit)}$$

$$\text{Worst Case Lost Accuracy} = \log_2\left(\text{ERROR } I_{TOTAL}\ (\text{LSB})\right)$$

$$\text{Worst Case Lost Accuracy (PFS)} = 0\ \text{bit (10 bit)} = 1.3\ \text{bits (12 bit)} = 3.3\ \text{bits (14 bit)}$$

$$\text{Worst Case Lost Accuracy (MS)} = 0\ \text{bit (10 bit)} = 0.3\ \text{bits (12 bit)} = 2.3\ \text{bits (14 bit)}$$

Conclusion

Using the voltage reference specifications to perform the above analysis, a designer should be able to predict the typical and worst-case accuracy loss due to reference error in their data conversion system. Repeating this exercise for several different references should provide the designer with more insight into what reference specifications are most critical in their application. These often include operating temperature range for temperature drift, initial accuracy if calibration is not possible, and low frequency noise which often cannot be filtered. Knowing which parameter is the dominant error factor helps to narrow down the reference options significantly and makes choosing the right reference easier.

Resources

1. "Voltage Reference Selection Basics" Power Designer, issue #123 by David Megaw
2. Voltage Reference e-book Tips and tricks for designing with voltage references
3. For online resources for reference designs, technical documents, and selector wheel, visit the Voltage Reference landing page ti.com/vref

   For online resources such as reference designs and other technical resources, visit the Voltage Reference landing page (ti.com/vref)
4. Try the interactive selection tool to help instantly choose the proper reference for an ADC from the broad TI portfolio, see the WEBENCH® Series Voltage Reference Selector Tool at ti.com/webenchvref

Appendix

Unit Conversion

Parts-per million (ppm): Parts-per million is a common units notation that is often used to describe ratios of very small numbers. ppm stands for 1 part per million such as percent stands for 1 part per hundred. It is often very useful to do all error calculations in ppm due to its unit neutral nature but not all electrical characteristics come in ppm. Here are some common conversions examples that are used to get ppm.

Example 1: 0.2% accuracy to ppm

1. Convert percent to a ratio

$$\frac{Percent}{100} = \frac{0.2\%}{100} = 0.002$$

2. Convert ratio into ppm

$$ratio * 1000000 = 0.002 * 1000000 = 2000\ ppm$$

Example 2: 30 µV deviation to ppm of a 3 V source

1. Convert 30 µV to V

$$\frac{30\ \mu V}{1000000} = 0.00003\ V$$

2. Divide by 3 V to get the ratio based on a 3 V source

$$\frac{0.00003\ V}{3\ V} = 0.00001$$

3. Convert the ratio into ppm

$$ratio * 1000000 = 0.00001 * 1000000 = 10\ ppm$$

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