SECTION 7-1

Voltage References

Reference circuits and linear regulators actually have much in common. In fact, the latter could be functionally described as a reference circuit, but with greater current (or power) output. Accordingly, almost all of the specifications of the two circuit types have great commonality (even though the performance of references is usually tighter with regard to drift, accuracy, etc.). In many cases today the support circuitry is included in the converter package. This is advantageous to the designer since it simplifies the design process and guarantees performance of the system.

Precision Voltage References

Voltage references have a major impact on the performance and accuracy of analog systems. A $\pm 5\,\text{mV}$ tolerance on a 5V reference corresponds to $\pm 0.1\%$ absolute accuracy which is only 10-bit accuracy. For a 12-bit system, choosing a reference that has a $\pm 1\,\text{mV}$ tolerance may be far more cost effective than performing manual calibration, while both high initial accuracy and calibration will be necessary in a system making absolute 16-bit measurements. Note that many systems make *relative* measurements rather than absolute ones, and in such cases the absolute accuracy of the reference is not as important, although noise and short-term stability may be.

Temperature drift or drift due to aging may be an even greater problem than absolute accuracy. The initial error can always be trimmed, but compensating for drift is difficult. Where possible, references should be chosen for temperature coefficient and aging characteristics which preserve adequate accuracy over the operating temperature range and expected lifetime of the system.

Noise in voltage references is often overlooked, but it can be very important in system design. Noise is an instantaneous change in the reference voltage. It is generally specified on data sheets, but system designers frequently ignore the specification and assume that voltage references do not contribute to system noise.

There are two dynamic issues that must be considered with voltage references: their behavior at start-up, and their behavior with transient loads. With regard to the first, always bear in mind that voltage references *do not power-up instantly* (this is true for references inside ADCs and DACs as well as discrete designs). Thus it is rarely possible to turn on an analog-to-digital converters (ADC) and reference, whether internal or external, make a reading, and turn off again within a few microseconds, however attractive such a procedure might be in terms of energy saving.

Regarding the second point, a given reference IC may or may not be well suited for pulse-loading conditions, dependent on the specific architecture. Many references use low power, and therefore low bandwidth, output buffer amplifiers. This makes for poor behavior under fast transient loads, which may degrade the performance of fast ADCs (especially successive approximation and flash ADCs). Suitable decoupling can ease the problem (but some references oscillate with capacitive loads), or an additional external broadband buffer amplifier may be used to drive the node where the transients occur.

Types of Voltage References

In terms of the functionality of their circuit connection, standard reference ICs are often only available in *series* or *three-terminal* form (V_{IN}, Common, V_{OUT}), and also in positive polarity only. The series types have the potential advantages of lower and more stable quiescent current, standard pre-trimmed output voltages, and relatively high output current without accuracy loss. *Shunt* or *two-terminal* (i.e., diode-like) references are more flexible regarding operating polarity, but they are also more restrictive as to loading. They can in fact eat up excessive power with widely varying resistor-fed voltage inputs. Also, they sometimes come in non-standard voltages. All of these various factors tend to govern when one functional type is preferred over the other.

Some simple diode-based references are shown in Figure 7-1. In the first of these, a current driven forward-biased diode (or diode-connected transistor) produces a voltage, $V_f = V_{REF}$. While the junction drop is somewhat decoupled from the raw supply, it has numerous deficiencies as a reference. Among them are a strong TC of about −0.3%/°C, some sensitivity to loading, and a rather inflexible output voltage; it is only available in 600 mV jumps.

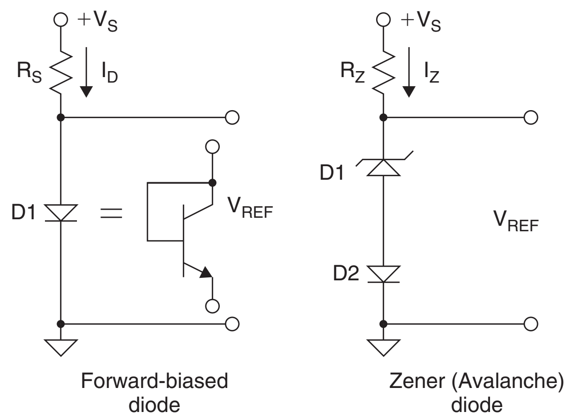

Figure 7-1: Simple diode reference circuits

By contrast, these most simple references (as well as all other shunt-type regulators) have a basic advantage, which is the fact that the polarity is readily reversible by flipping connections and reversing the drive current. However, a basic limitation of all shunt regulators is that load current must always be less (usually appreciably less) than the driving current, I_D.

In the second circuit of Figure 7-1, a zener or avalanche diode is used, and an appreciably higher output voltage realized. While true *zener* breakdown occurs below 5 V, *avalanche* breakdown occurs at higher voltages and has a positive temperature coefficient. Note that diode reverse breakdown is referred to almost universally today as *zener*, even though it is usually avalanche breakdown. With a D1 breakdown voltage in the 5–8 V range, the net positive TC is such that it equals the negative TC of forward-biased diode D2, yielding a net TC of 100 ppm/°C or less with proper bias current. Combinations of such carefully chosen diodes formed the basis of the early single package "temperature-compensated zener" references, such as the 1N821–1N829 series.

The temperature-compensated zener reference is limited in terms of initial accuracy, since the best TC combinations fall at odd voltages, such as the 1N829's 6.2 V. And, the scheme is also limited for loading, since for best TC the diode current must be carefully controlled. Unlike a fundamentally lower voltage (<2 V) reference, zener diode-based references must of necessity be driven from voltage sources appreciably higher than 6 V levels, so this precludes operation of zener references from 5 V system supplies. References based on low TC zener (avalanche) diodes also tend to be noisy, due to the basic noise of the breakdown mechanism. This has been improved greatly with *monolithic* zener types, as is described further below.

Bandgap References

The development of low voltage (<5V) references based on the bandgap voltage of silicon led to the introduction of various ICs which could be operated on low voltage supplies with good TC performance. The first of these was the LM109 (Reference 1), and a basic bandgap reference cell is shown in Figure 7-2.

This circuit is also called a "ΔV_{BE}" reference because the differing current densities between matched transistors Q1–Q2 produce a ΔV_{BE} across R_3. It works by summing the V_{BE} of Q3 with the amplified ΔV_{BE} of Q1–Q2, developed across R_2. The ΔV_{BE} and V_{BE} components have opposite polarity TCs; ΔV_{BE} is proportional to absolute temperature (PTAT), while V_{BE} is complementary to absolute temperature (CTAT). The summed output is V_R, and when it is equal to 1.205V (silicon bandgap voltage), the TC is a minimum.

The bandgap reference technique is attractive in IC designs because of several reasons; among these are the relative simplicity, and the avoidance of zeners and their noise. However, very important in these days of ever decreasing system supplies is the fundamental fact that bandgap devices operate at low voltages, i.e., <5V. Not only are they used for standalone IC references, but they are also used within the designs of many other linear ICs such as analog-to-digital converter (ADCs) and digital-to-analog converter (DACs).

However, the basic designs of Figure 7-2 suffer from load and current drive sensitivity, plus the fact that the output needs accurate scaling to more useful levels, i.e., 2.5V, 5V, etc. The load drive issue is best addressed with the use of a buffer amplifier, which also provides convenient voltage scaling to standard levels.

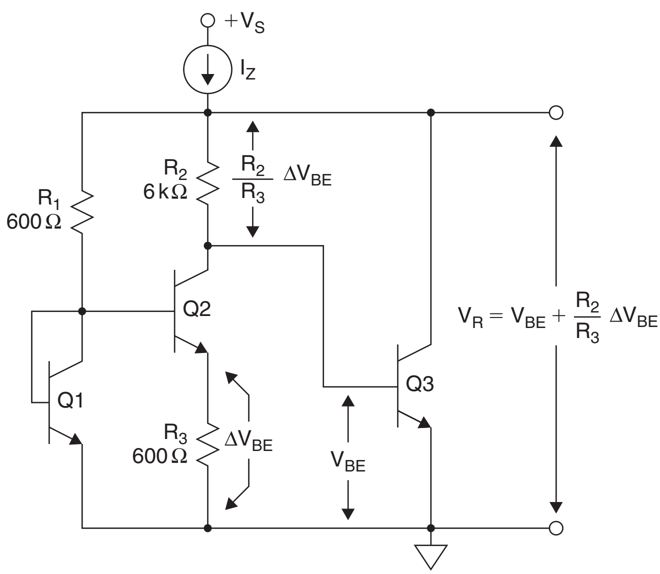

Figure 7-2: Basic bandgap reference

An improved three-terminal bandgap reference, the AD580 (introduced in 1974) is shown in Figure 7-3. Popularly called the "Brokaw Cell" (see References 2 and 3), this circuit provides on-chip output buffering, which allows good drive capability and standard output voltage scaling. The AD580 was the first precision bandgap-based IC reference, and variants of the topology have influenced further generations of both

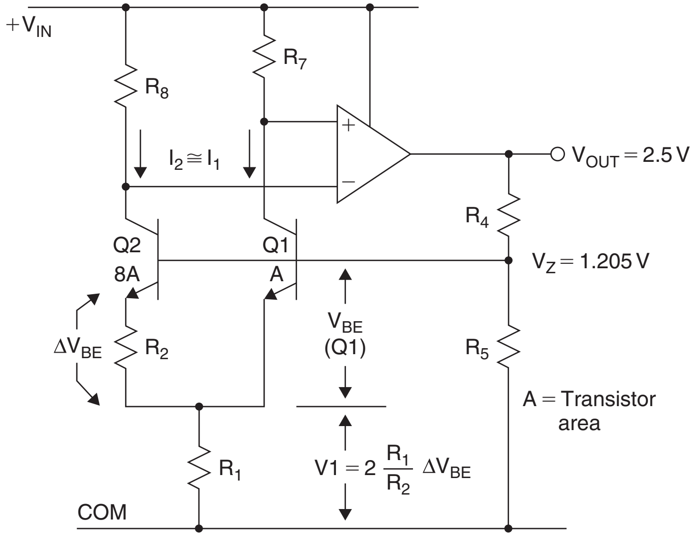

Figure 7-3: AD580 precision bandgap reference uses Brokaw Cell (1974)

industry standard references such as the REF01, REF02, and REF03 series, and more recent ADI bandgap parts such as the REF19x series, the AD680, AD780, the AD1582–AD1585 series, the ADR38x series, the ADR39x series, and recent SC-70 and SOT-23 offerings of improved versions of the REF01, REF02, and REF03 (designated ADR01, ADR02, and ADR03).

The AD580 has two 8:1 emitter-scaled transistors Q1–Q2 operating at identical collector currents (and thus 1/8 current densities), by virtue of equal load resistors and a closed loop around the buffer op amp. Due to the resultant smaller V_{BE} of the 8 × area Q2, R_2 in series with Q2 drops the ΔV_{BE} voltage, while R_1 (due to the current relationships) drops a PTAT voltage V1:

$$V1 = 2 \times \frac{R_1}{R_2} \times \Delta V_{BE} \tag{7-1}$$

The bandgap cell reference voltage V_Z appears at the base of Q1, and is the sum of V_{BE} (Q1) and V1, or 1.205 V, the bandgap voltage:

$$V_Z = V_{BE(Q1)} + V1 \tag{7-2}$$

$$= V_{BE(Q1)} + 2 \times \frac{R_1}{R_2} \times \Delta V_{BE} \tag{7-3}$$

$$= V_{BE(Q1)} + 2 \times \frac{R_1}{R_2} \times \frac{kT}{q} \times \ln \frac{J1}{J2} \tag{7-4}$$

$$= V_{BE(Q1)} + 2 \times \frac{R_1}{R_2} \times \frac{kT}{q} \times \ln 8 \qquad (7\text{-}5)$$

$$= 1.205\ \text{V}$$

Note that J1 = current density in Q1, J2 = current density in Q2, and J1/J2 = 8.

However, because of the presence of the R_4/R_5 (laser-trimmed) thin-film divider and the op amp, the actual voltage appearing at V_{OUT} can be scaled higher, in the AD580 case 2.5 V. Following this general principle, V_{OUT} can be raised to other practical levels, such as e.g., in the AD584, with taps for precise 2.5, 5, 7.5, and 10 V operations. The AD580 provides up to 10 mA output current while operating from supplies between 4.5 and 30 V. It is available in tolerances as low as 0.4%, with TCs as low as 10 ppm/°C.

Many of the recent developments in bandgap references have focused on smaller package size and cost reduction, to address system needs for smaller, more power efficient and less costly reference ICs. Among these are several recent bandgap-based IC references.

The AD1580 (introduced in 1996) is a shunt mode IC reference which is functionally quite similar to the classic shunt IC reference, the AD589 (introduced in 1980) mentioned above. A key difference is the fact that the AD1580 uses a newer, small geometry process, enabling its availability within the tiny SOT-23 package. The very small size of this package allows use in a wide variety of space limited applications, and the low operating current lends itself to portable battery powered uses. The AD1580 circuit is shown in simplified form in Figure 7-4.

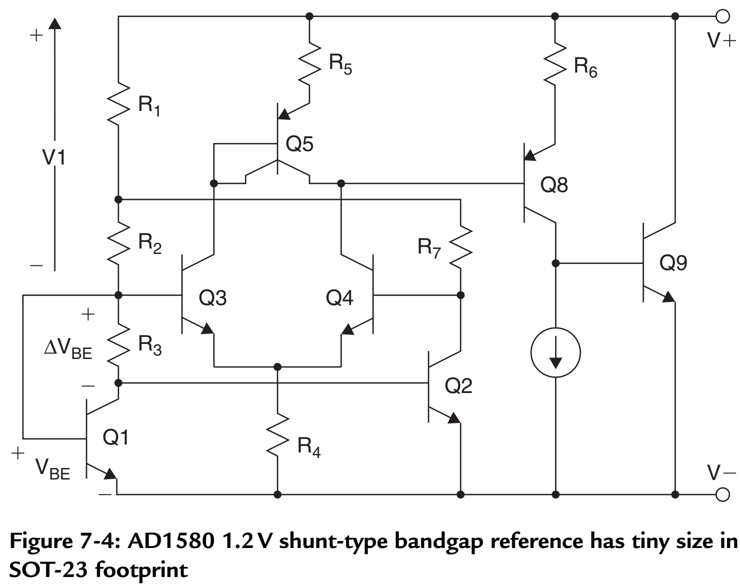

Figure 7-4: AD1580 1.2 V shunt-type bandgap reference has tiny size in SOT-23 footprint

In this circuit, like transistors Q1 and Q2 form the bandgap core, and are operated at a current ratio of 5 times, determined by the ratio of R_7 to R_2. An op amp is formed by the differential pair Q3–Q4, current mirror Q5, and driver/output stage Q8–Q9. In closed-loop equilibrium, this amplifier maintains the bottom ends of R_2–R_7 at the same potential.

As a result of the closed-loop control described, a basic ΔV_{BE} voltage is dropped across R_3, and a scaled PTAT voltage also appears as V1, which is effectively in series with V_{BE}. The nominal bandgap reference voltage of 1.225 V is then the sum of Q1's V_{BE} and V1. The AD1580 is designed to operate at currents as low as 50 µA, also handling maximum currents as high as 10 mA. It is available in grades with voltage tolerances of ±1 or ±10 mV, and with corresponding TCs of 50 or 100 ppm/°C.

The circuit diagram for the series, shown in Figure 7-5, may be recognized as a variant of the basic Brokaw bandgap cell, as described under Figure 7-3. In this case Q1–Q2 form the core, and the overall loop operates to produce the stable reference voltage V_{BG} at the base of Q1. A notable difference here is that the op amp's output stage is designed with push–pull common-emitter stages. This has the effect of requiring an output capacitor for stability, but it also provides the IC with relatively low dropout operation.

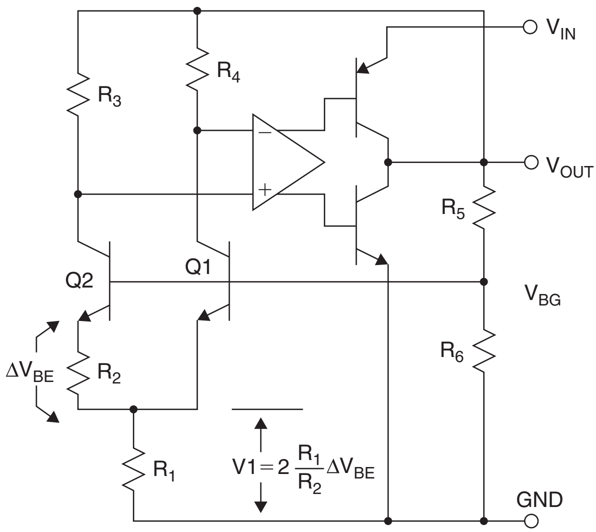

Figure 7-5: AD1582–AD1585 2.5–5 V series type bandgap references

The low dropout feature means essentially that V_{IN} can be lowered to as close as several hundred millivolts above the V_{OUT} level without disturbing operation. The push–pull operation also means that this device series can actually both sink and source currents at the output, as opposed to the classic reference operation of sourcing current (only). For the various output voltage ratings, the divider R_5–R_6 is adjusted for the respective levels.

The AD1582 series is designed to operate with quiescent currents of only 65 µA (maximum), which allows good power efficiency when used in low power systems with varying voltage inputs. The rated output current for the series is 5 mA, and they are available in grades with voltage tolerances of ±0.1% or ±1% of V_{OUT}, with corresponding TCs of 50 or 100 ppm/°C.

Because of stability requirements, devices of the AD1582 series must be used with both an output and input bypass capacitor. Recommended worst-case values for these are shown in the hookup diagram of Figure 7-6. For the electrical values noted, it is likely that tantalum chip capacitors will be the smallest in size.

Buried Zener References

In terms of the design approaches used within the reference core, the two most popular basic types of IC references consist of the bandgap and buried zener units. Bandgaps have been discussed, but zener-based references warrant some further discussion.

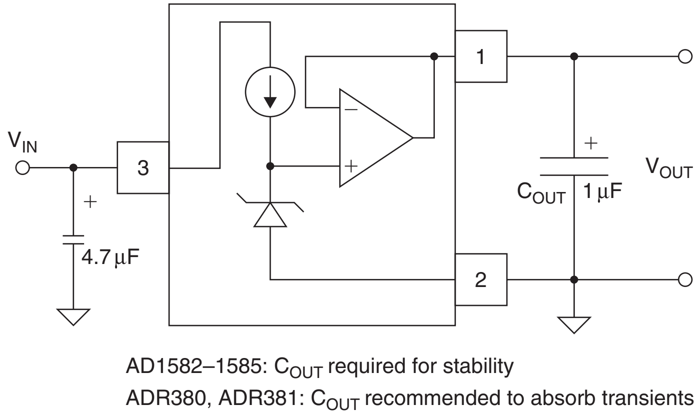

Figure 7-6: AD1582–AD1585 series connection diagram

In an IC chip, surface operated diode junction breakdown is prone to crystal imperfections and other contamination, thus zener diodes formed at the surface are more noisy and less stable than are *buried* (or subsurface) ones (Figure 7-7). ADI zener-based IC references employ the much preferred buried zener. This improves substantially upon the noise and drift of surface-mode operated zeners (see Reference 4). Buried zener references offer very low temperature drift, down to the 1–2 ppm/°C (AD588 and AD586) (Figure 7-8), and the lowest noise as a percent of full-scale, i.e., $100\,\text{nV}/\sqrt{\text{Hz}}$ or less. On the downside, the operating current of zener type references is usually relatively high, typically on the order of several milliamperes. The zener voltage is also relatively high, typically on the order of 5 V. This limits its application in low voltage circuits.

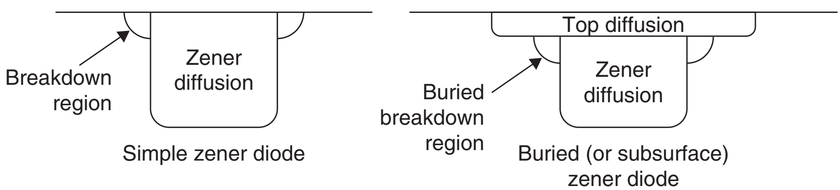

Figure 7-7: Simple surface zener versus a buried zener

An important general point arises when comparing noise performance of different references. The best way to do this is to compare the ratio of the noise (within a given bandwidth) to the DC output voltage. For example, a 10 V reference with a $100\,\text{nV}/\sqrt{\text{Hz}}$ noise density is 6 dB more quiet in relative terms than is a 5 V reference with the same noise level.

XFET® References

A third and relatively new category of IC reference core design is based on the properties of junction field effect (JFET) transistors. Somewhat analogous to the bandgap reference for bipolar transistors, the JFET-based reference operates a pair of JFET transistors with different pinchoff voltages, and amplifies the differential output to produce a stable reference voltage. One of the two JFETs uses an extra ion implantation, giving rise to the name XFET® (eXtra implantation junction Field Effect Transistor) for the reference core design.

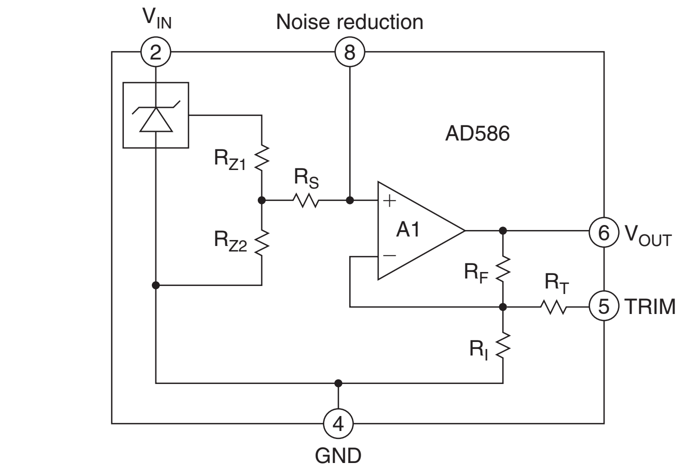

Figure 7-8: Typical buried zener reference (AD586)

The basic topology for the XFET reference circuit is shown in Figure 7-9. J1 and J2 are the two JFET transistors, which form the core of the reference. J1 and J2 are driven at the same current level from matched current sources, I1 and I2. To the right, J1 is the JFET with the extra implantation, which causes the difference in the J1–J2 pinchoff voltages to differ by 500 mV. With the pinchoff voltage of two such FETs purposely skewed, a differential voltage will appear between the gates for identical current drive conditions and equal source voltages. This voltage, ΔV_P, is:

$$\Delta V_P = V_{P1} - V_{P2} \tag{7-6}$$

where V_{P1} and V_{P2} are the pinchoff voltages of FETs J1 and J2, respectively.

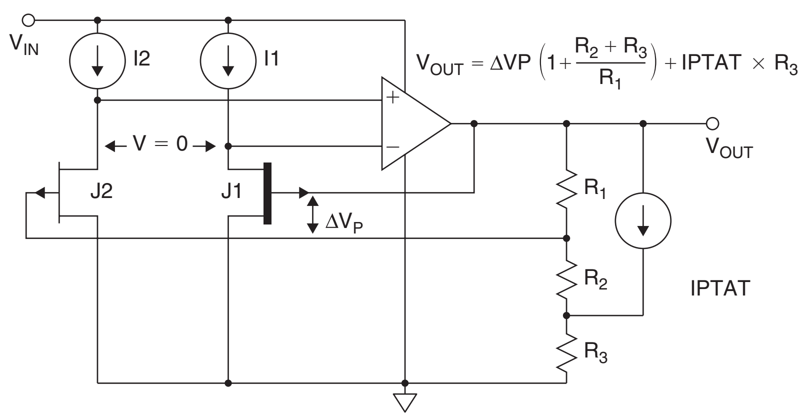

Figure 7-9: XFET® reference simplified schematic

Note that, within this circuit, the voltage ΔV_P exists between the *gates* of the two FETs. We also know that, with the overall feedback loop closed, the op amp axiom of zero input differential voltage will hold the sources of the two JFETs at same potential. These source voltages are applied as inputs to the op amp, the output of which drives feedback divider R_1–R_3. As this loop is configured, it stabilizes at an output voltage from the R_1–R_2 tap which does in fact produce the required ΔV_P between the J1–J2 gates. In essence, the op amp amplifies ΔV_P to produce V_{OUT}, where

$$V_{OUT} = \Delta V_P \left(1 + \frac{R_2 + R_3}{R_1}\right) + (I_{PTAT})(R_3) \qquad (7\text{-}7)$$

As can be noted, this expression includes the basic output scaling (leftmost portion of the right terms), plus a rightmost temperature dependent term including I_{PTAT}. The I_{PTAT} portion of the expression compensates for a basic negative temperature coefficient of the XFET core, such that the overall net temperature drift of the reference is typically in a range of 3–8 ppm/°C.

The XFET architecture offers performance improvements over bandgap and buried zener references, particularly for systems where operating current is critical, yet drift and noise performance must still be excellent. XFET noise levels are lower than bandgap-based bipolar references operating at an equivalent current, the temperature drift is low and linear at 3–8 ppm/°C (allowing easier compensation when required), and the series has lower hysteresis than bandgaps. Thermal hysteresis is a low 50 ppm over a −40 to +125°C range, less that half that of a typical bandgap device. Finally, the long-term stability is excellent, typically only 50 ppm/1,000 hours.

Figure 7-10 summarizes the pro and con characteristics of the three reference architectures: bandgap, buried zener, and XFET.

Bandgap	Buried zener	XFET®
<5 V supplies	>5 V supplies	<5 V supplies
High noise @ High power	Low noise @ High power	Low noise @ Low power
Fair drift and long term stability	Good drift and long term stability	Excellent drift and long term stability
Fair hysteresis	Fair hysteresis	Low hysteresis

Figure 7-10: Characteristics of reference architectures

Though modern IC references come in a variety of styles, series operating, fixed output positive types do tend to dominate. They may or may not be low power, low noise, and/or low dropout, or available within a certain package. Of course, in a given application, any one of these differentiating factors can drive a choice, thus it behooves the designer to be aware of all the different devices available.

Figure 7-11 shows the typical schematic for a series type IC positive reference in an 8-pin package. (Note that "(x)" numbers refer to the standard pin for that function). There are several details which are important. Many references allow optional trimming by connecting an external trim circuit to drive the references' *trim* input pin (5). Some bandgap references also have a high impedance PTAT output (V_{TEMP}) for temperature sensing (pin 3). The intent here is that no appreciable current be drawn from this pin, but it can be useful for non-loading types of connections such as comparator inputs, to sense temperature thresholds, etc.

Some references have a pin labeled "noise reduction." This may cause some confusion. A capacitor connected to this pin will reduce the noise of the reference cell itself; this cell is typically followed by an internal buffer. The noise of this buffer will not be affected.

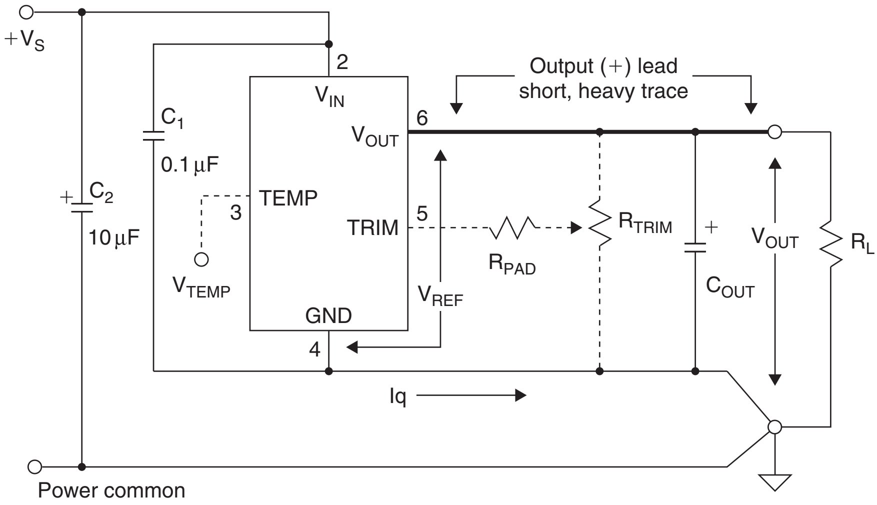

Figure 7-11: Standard positive output three-terminal reference hookup (8-pin DIP pinout)

All references should use decoupling capacitors on the input pin (2), but the amount of decoupling (if any) placed on the output (pin 6) depends on the stability of the reference's output op amp with capacitive load. Simply put, there is no hard-and-fast rule for capacitive loads here. For example, some three-terminal types *require* the output capacitor for stability (i.e., REF19x and AD1582–AD1585 series), while with others it is optional for performance improvement (AD780, REF43, ADR29x, ADR43x, AD38x, AD39x, ADR01, ADR02, ADR03). Even if the output capacitor is optional, it may still be required to supply the energy for transient load currents, as presented by some ADC reference input circuits. The safest rule then is that you should use the data sheet to verify what are the specific capacitive loading ground rules for the reference you intend to use, for the load conditions your circuit presents.

Voltage Reference Specifications

Tolerance

It is usually better to select a reference with the required value and accuracy and to avoid external trimming and scaling if possible. This allows the best TCs to be realized, as tight tolerances and low TCs usually go hand-in-hand. Tolerances as low as approximately 0.04% can be achieved with the AD586, AD780, REF195, and ADR43x series, while the AD588 is 0.01%. If and when trimming must be used, be sure to use the recommended trim network with no more range than is absolutely necessary. When/if additional external scaling is required, a precision op amp should be used, along with ratio-accurate, low TC tracking thin-film resistors.

Drift

The XFET and buried zener reference families have the best long-term drift and TC performance. The XFET ADR43x series have TCs as low as 3 ppm/°C. TCs as low as 1–2 ppm/°C are available with the AD586 and AD588 buried zener references, and the AD780 bandgap reference is almost as good at 3 ppm/°C.

The XFET series achieves long-term drifts of 50 ppm/1,000 hours, while the buried zener types come in at 25 ppm/1,000 hours. Note that where a Figure is given for long-term drift, it is usually drift expressed in ppm/1,000 hours. There are 8,766 hours in a year, and many engineers multiply the 1,000 hour Figure by 8.77 to find the annual drift—this is not correct, and can in fact be quite pessimistic. Long-term drift in precision analog circuits is a "random walk" phenomenon and increases with the *square root* of the elapsed time (this supposes that drift is due to random micro-effects in the chip and not some overriding cause such as contamination). The 1 year Figure will therefore be about $\sqrt{8.766} \approx 3$ times the 1,000 hour Figure, and the 10 year value will be roughly 9 times the 1,000 hour value. In practice, things are a little better even than this, as devices tend to stabilize with age.

The accuracy of an ADC or DAC can be no better than that of its reference. Reference temperature drift affects full-scale accuracy as shown in Figure 7-12. This table shows system resolution and the TC required to maintain ½ LSB error over an operating temperature range of 100°C. For example, a TC of about 1 ppm/°C is required to maintain ½ LSB error at 12 bits. For smaller operating temperature ranges, the drift requirement will be less. The last three columns of the table show the voltage value of ½ LSB for popular full-scale ranges.

Bits	Required drift (ppm/°C)	½ LSB Weight (mV) 10, 5 and 2.5 V Full-scale ranges		
		10 V	5 V	2.5 V
8	19.53	19.53	9.77	4.88
9	9.77	9.77	4.88	2.44
10	4.88	4.88	2.44	1.22
11	2.44	2.44	1.22	0.61
12	1.22	1.22	0.61	0.31
13	0.61	0.61	0.31	0.15
14	0.31	0.31	0.15	0.08
15	0.15	0.15	0.08	0.04
16	0.08	0.08	0.04	0.02

Figure 7-12: Reference temperature drift requirements for various system accuracies (½ LSB criteria, 100°C span)

Supply Range

IC reference supply voltages range from about 3 V (or less) above rated output, to as high as 30 V (or more) above rated output. Exceptions are devices designed for low dropout, such as the REF19X, AD1582–AD1585, ADR38X, ADR39X series. At low currents, the REF195 can deliver 5 V with an input as low as 5.1 V (100 mV dropout). Note that due to process limits, some references may have more restrictive maximum voltage input ranges, such as the AD1582–AD1585 series (12 V), the ADR29x series (15 V), and the ADR43x series (18 V).

Load Sensitivity

Load sensitivity (or output impedance) is usually specified in µV/mA of load current, or mΩ, or ppm/mA. While Figures of 70 ppm/mA or less are quite good (AD780, REF43, REF195, ADR29X, ADR43X), it should be noted that external wiring drops can produce comparable errors at high currents, without care in layout. Load current dependent errors are minimized with short, heavy conductors on the (+) output and on the ground return. For the highest precision, buffer amplifiers and Kelvin sensing circuits (AD588, AD688, ADR39x) are used to ensure accurate voltages at the load.

The output of a buffered reference is the output of an op amp, and therefore the source impedance is a function of frequency. Typical reference output impedance rises at 6 dB/octave from the DC value, and is nominally about 10 Ω at a few hundred kilohertz. This impedance can be lowered with an external capacitor, provided the op amp within the reference remains stable for such loading.

Line Sensitivity

Line sensitivity (or regulation) is usually specified in μV/V, (or ppm/V) of input change, and is typically 25 ppm/V (−92 dB) in the REF43, REF195, AD680, AD780, ADR29X, ADR39X, and ADR43X. For DC and very low frequencies, such errors are easily masked by noise.

As with op amps, the line sensitivity (or power supply rejection) of references degrades with increasing frequency, typically 30–50 dB at a few hundred kilohertz. For this reason, the reference input should be highly decoupled (LF and HF). Line rejection can also be increased with a low dropout pre-regulator, such as one of the ADP3300 series parts.

Noise

Reference noise is not always specified, and when it is, there is not total uniformity on how. For example, some devices are characterized for peak-to-peak noise in a 0.1–10 Hz bandwidth, while others are specified in terms of wideband RMS or peak-to-peak noise over a specified bandwidth. The most useful way to specify noise (as with op amps) is a plot of noise voltage spectral density ($nV/\sqrt{Hz}$) versus frequency.

Low noise references are important in high resolution systems to prevent loss of accuracy. Since white noise is statistical, a given noise density must be related to an equivalent peak-to-peak noise in the relevant bandwidth. Strictly speaking, the peak-to-peak noise in a Gaussian system is infinite (but its probability is infinitesimal). Conventionally, the figure of 6.6 × RMS is used to define a practical peak value—statistically, this occurs less than 0.1% of the time. This peak-to-peak value should be less than ½ LSB in order to maintain required accuracy. If peak-to-peak noise is assumed to be 6 times the RMS value, then for an N-bit system, reference voltage full-scale V_{REF}, reference noise bandwidth (BW), the required noise voltage spectral density E_n ($V/\sqrt{Hz}$) is given by:

$$E_n \le \frac{V_{REF}}{12 \times 2^N \times \sqrt{BW}} \tag{7-8}$$

For a 10 V, 12-bit, 100 kHz system, the noise requirement is a modest 643 $nV/\sqrt{Hz}$. Figure 7-13 shows that increasing resolution and/or lower full-scale references make noise requirements more stringent. The 100 kHz bandwidth assumption is somewhat arbitrary, but the user may reduce it with external filtering, thereby reducing the noise. Most good IC references have noise spectral densities around 100 $nV/\sqrt{Hz}$, so additional filtering is obviously required in most high resolution systems, especially those with low values of V_{REF}.

Some references, e.g., the AD587 buried zener type, have a pin designated as the *noise reduction pin* (see data sheet). This pin is connected to a high impedance node preceding the on-chip buffer amplifier. Thus an externally connected capacitor C_N will form a lowpass filter with an internal resistor, to limit the effective noise bandwidth seen at the output. A 1 μF capacitor gives a 3 dB bandwidth of 40 Hz. Note that

Bits	Noise density (nv/√Hz) for 10, 5 and 2.5 V Fulls-cale ranges		
	10 V	5 V	2.5 V
12	643	322	161
13	322	161	80
14	161	80	40
15	80	40	20
16	40	20	10

Figure 7-13: Reference noise requirements for various system accuracies (½ LSB/100 kHz criteria) scaled references

this method of noise reduction is by no means universal, and other devices may implement noise reduction differently, if at all. Also note that it does not affect the noise of the buffer amplifier.

There are also general-purpose methods of noise reduction, which can be used to reduce the noise of any reference IC, at any standard voltage level. Note that the DC characteristics of the reference filter will affect the accuracy of the reference.

A useful approach when a non-standard reference voltage is required is to simply buffer and scale a basic low voltage reference diode. With this approach, a potential difficulty is getting an amplifier to work well at such low voltages as 3 V. A workhorse solution is the low power reference and scaling buffer shown in Figure 7-14. Here a low current 1.2 V two-terminal reference diode is used for D1, which can be either a 1.200 V ADR512, 1.235 V AD589, or the 1.225 V AD1580. Resistor R_1 sets the diode current in either case,

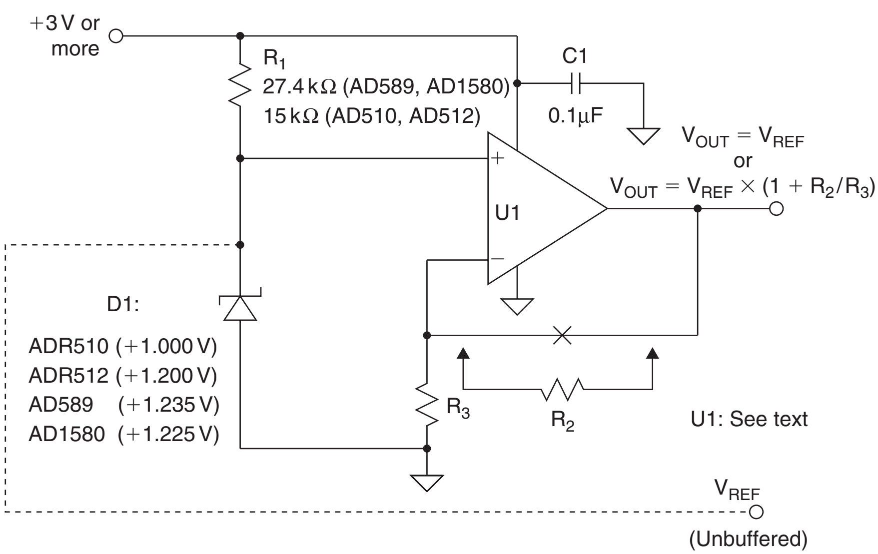

Figure 7-14: Rail-to-rail output op amps allow greatest flexibility in low dropout references

and is chosen for the diode minimum current requirement at a minimum supply of 2.7V. Obviously, loading on the unbuffered diode must be minimized at the V_{REF} node.

The amplifier U1 both buffers and optionally scales up the nominal 1.0 or 1.2V reference, allowing much higher source/sink output currents. Of course, a higher op amp quiescent current is expended in doing this, but this is a basic tradeoff of the approach.

In Figure 7-14, without gain scaling resistors R_2–R_3, V_{OUT} is simply equal to V_{REF}. With the use of the scaling resistors, V_{OUT} can be set anywhere between a lower limit of V_{REF}, and an upper limit of the positive rail, due to the op amp's rail–rail output swing. Also, note that this buffered reference is inherently low dropout, allowing a +4.5V (or more) reference output on a +5V supply, for example. The general expression for V_{OUT} is shown in the Figure, where V_{REF} is the reference voltage.

Voltage Reference Pulse Current Response

The response of references to dynamic loads is often a concern, especially in applications such as driving ADCs and DACs. Fast changes in load current invariably perturb the output, often outside the rated error band. For example, the reference input to a sigma–delta ADC may be the switched capacitor circuit shown in Figure 7-15. The dynamic load causes current spikes in the reference as the capacitor C_{IN} is charged and discharged. As a result, noise may be induced on the ADC reference circuitry.

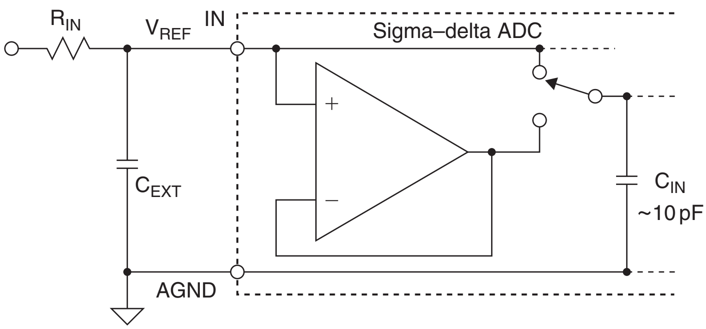

Figure 7-15: Switched capacitor input of sigma–delta ADC presents a dynamic load to the voltage reference

Although sigma–delta ADCs have an internal digital filter, transients on the reference input can still cause appreciable conversion errors. Thus it is important to maintain a low noise, transient free potential at the ADC's reference input. Be aware that if the reference source impedance is too high, dynamic loading can cause the reference input to shift by more than 5 mV.

A bypass capacitor on the output of a reference may help it to cope with load transients, but many references are unstable with large capacitive loads. Therefore it is quite important to verify that the device chosen will satisfactorily drive the output capacitance required. In any case, the converter reference inputs should always be decoupled—with at least 0.1 µF, and with an additional 5–50 µF if there is any low frequency ripple on its supply.

Since some references misbehave with transient loads, either by oscillating or by losing accuracy for comparatively long periods, it is advisable to test the pulse response of voltage references which may encounter transient loads. A suitable circuit is shown in Figure 7-16. In a typical voltage reference, a step change of 1 mA produces the transients shown. Both the duration of the transient and the amplitude of the ringing *increase* when a 0.01 µF capacitor is connected to the reference output.

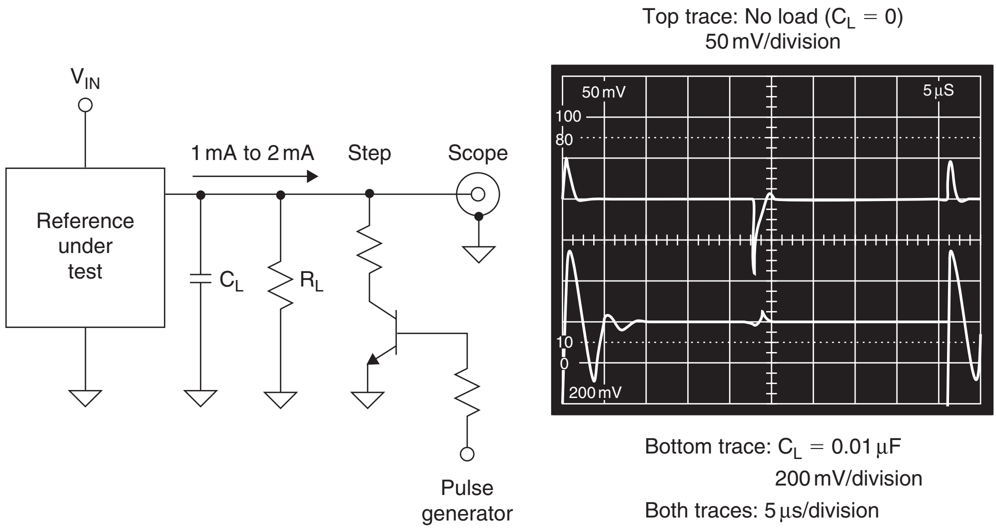

Figure 7-16: Make sure reference is stable with large capacitive loads

As noted above, reference bypass capacitors are useful when driving the reference inputs of successive approximation ADCs. Figure 7-17 illustrates reference voltage settling behavior immediately following the "Start Convert" command. A small capacitor (0.01 µF) does not provide sufficient charge storage to keep the reference voltage stable during conversion, and errors may result. As shown by the bottom trace, decoupling with a ≥1 µF capacitor maintains the reference stability during conversion.

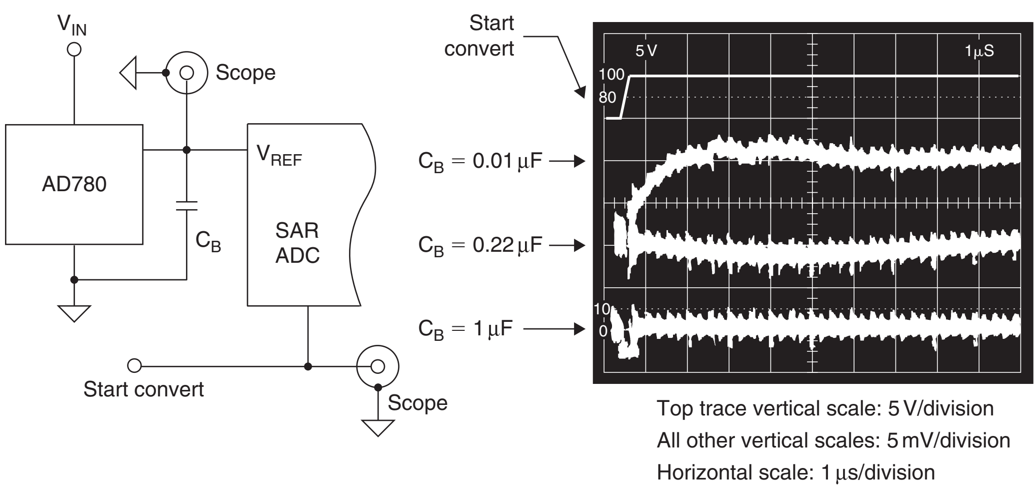

Figure 7-17: Successive approximation ADCs can present a dynamic transient load to the reference

Where voltage references are required to drive large capacitances, it is also critically important to realize that their turn-on time will be prolonged. Experiment may be needed to determine the delay before the reference output reaches full accuracy, but it will certainly be much longer than the time specified on the data sheet for the same reference in a low capacitance loaded state.

Low Noise References for High Resolution Converters

High resolution converters (both sigma–delta and high speed ones) can benefit from recent improvements in IC references, such as lower noise and the ability to drive capacitive loads. Even though many data converters have internal references, the performance of these references is often compromised because of the limitations of the converter process. In such cases, using an external reference rather than the internal one often yields better overall performance. For example, the AD7710 series of 22-bit ADCs has a 2.5 V internal reference with a 0.1–10 Hz noise of 8.3 μV RMS (2,600 nV/$\sqrt{\text{Hz}}$), while the AD780 reference noise is only 0.67 μV RMS (200 nV/$\sqrt{\text{Hz}}$). The internal noise of the AD7710 series in this bandwidth is about 1.7 μV RMS. The use of the AD780 increases the effective resolution of the AD7710 from about 20.5 bits to 21.5 bits.

There is one possible but yet quite real problem when replacing the internal reference of a converter with a higher precision external one. The converter in question may have been trimmed during manufacture to deliver its specified performance with a relatively inaccurate internal reference. In such a case, using a more accurate external reference with the converter may actually introduce additional gain error! For example, the early AD574 had a guaranteed uncalibrated gain accuracy of 0.125% when using an internal 10 V reference (which itself had a specified accuracy of only ±1%). It is obvious that if such a device, having an internal reference which is at one end of the specified range, is used with an external reference of exactly 10 V, then its gain will be about 1% in error.

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