INSTRUMENTATION AMPLIFIERS

BACKGROUND

The monolithic operational amplifier (op-amp) has, in recent years, surpassed the basic discrete transistor as an analog building block in almost all instrumentation applications. While there are still many application areas (particularly very high speed, high voltage, high current, and ultra low noise) in which the discrete transistor (or even the vacuum tube) predominates, the ease of application of the op-amp has vastly simplified analog design at the system level, at least at frequencies below 100 MHz or so. The origins of the op-amp are buried deep in early negative feedback techniques, but suffice it to say here that the term "operational amplifier" appears to have been coined in a paper by Ragazzini and his colleagues (1), although the term "computing amplifier" survived for a while. The op-amp consisted of a high-gain inverting amplifier that remained stable with 100% negative feedback. This enabled highly controlled closed-loop functions (particularly integrators) to be realized, which, in turn, formed the basis for many analog computing functions. Later op-amps included a noninverting input, modifying the transfer function to one of high gain appearing *differentially* between the two inputs.

Of course, differential amplifiers were not new; simple long-tailed-pair amplifiers with controlled gain date far back to the early days of electronics. But the op-amp was different in that the gain control came from purely external components (assuming high enough open-loop gain), potentially enabling precision gain control from purely passive components.

However, herein lies the dilemma: When feedback is applied around an op-amp, the inverting input becomes a low (ideally zero) impedance. The noninverting input, though, remains at high impedance. Many instrumentation (and other) systems require a precise differential amplifier with high (relative to signal source) impedance for *both* inputs, and this function is now known as an *instrumentation amplifier*. Obviously, simple op-amps require considerable modification to fulfill this function.

DEFINITION

An *instrumentation amplifier* is a precision amplifier with single-ended output and differential inputs and with precisely controlled gain for voltages appearing between its inputs. Ideally, voltages common to both inputs should not affect the output (this will lead to a discussion concerning common-mode rejection in due course). Additionally, both inputs are expected to have high impedance (relative to the source impedance), and this is normally expected to be symmetric, at least to a first order.

Unlike the op-amp, the origins of the term "instrumentation amplifier" are somewhat nebulous. Some early monolithic precision op-amps, notably the μA725 from Fairchild Semiconductor (1969) and the OP-07 from Precision Monolithics (1975), were referred to as instrumentation operational amplifiers. This was undoubtedly a marketing label intended to emphasize their precision input characteristics, necessary in many instrumentation applications; but they were still op-amps and not instrumentation amplifiers in the sense now generally accepted. The term "differential amplifier" has its origins lost to obscurity, but such terms as "high-accuracy differential amplifier" or "direct-coupled differential amplifier" appear regularly in texts prior to 1960, and these, of course, are terms that basically describe the quintessential instrumentation amplifier. In the mid-1960s the term *data amplifier* became commonplace to describe such amplifiers (2,3), and several companies were producing self-contained modular amplifiers at this time. The now commonly used term *instrumentation amplifier* was certainly in use by 1967 (4), and the two terms were concurrent for a while.

THE BASIC INSTRUMENTATION AMPLIFIER

Possibly the simplest way to achieve the instrumentation amplifier function is with a single op-amp configured for equal inverting and noninverting gains.

Referring to Fig. 1, the amplifier produces a gain between the inverting input (−IN) and the output, which can be derived from classical op-amp theory as follows:

$$G- = -\frac{R4}{R3}$$

The gain from the noninverting input (+IN) to the output can be described by

$$G+ = \left(\frac{R2}{R1+R2}\right)\left(\frac{R3+R4}{R4}\right)$$

So if the ratio $R1/R2$ is made identical to the ratio $R3/R4$, the resulting output gain is identical for both inputs but of opposite sign (assuming, for the moment, that $A1$ is ideal and neglecting any source impedance). This is tantamount to saying that the *common-mode* gain is zero. The absolute value of gain from either input to the output with respect to the other input is still given by

$$G = \frac{R4}{R3}$$

and this is the *differential-mode* gain. For simplicity, the term "gain" when used without qualification, will subsequently be assumed to be the differential-mode value. This circuit thus performs the basic functions of an instrumentation amplifier.

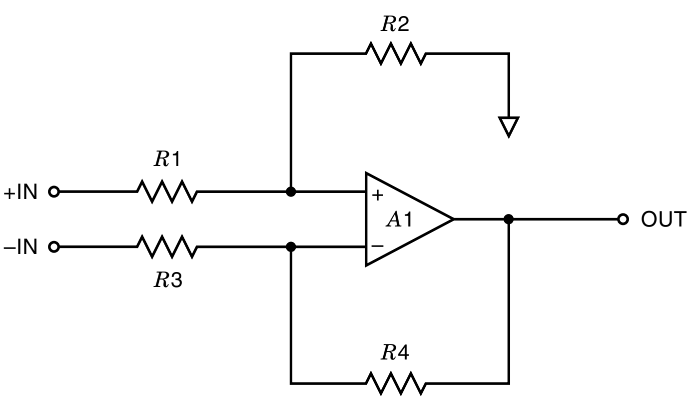

Figure 1. Basic instrumentation amplifier.

Although this circuit is commonly used where economy is a major concern, as will be shown, it lacks many of the features demanded in modern instrumentation systems. It is, however, a good starting point to illustrate some of the problems encountered in the design of instrumentation amplifiers in general.

GAIN ERROR

Errors in gain for the simple instrumentation amplifier of Fig. 1 are essentially those of the resistors themselves, unless the gain is unusually high or the op-amp has poor open-loop gain or common-mode rejection ratio (CMRR). For such a simple circuit, the effects of resistor mismatch cause much greater problems by common-mode rejection than by gain error. Small gain errors can often be tolerated or calibrated out, whereas in the case of poor common-mode rejection spurious signals end up mixed with the desired signal and may be impossible to remove. This situation is exacerbated greatly if the common-mode signal is comparable to (or even greater than) the desired one, which unfortunately is often the case (this is the major reason why an instrumentation amplifier is usually needed).

COMMON-MODE REJECTION RATIO

As already mentioned, if the inverting and noninverting gains are perfectly matched, then input voltages common to both of these inputs theoretically never appear at the output. This is, of course, the ideal case. In the case of the simple circuit of Fig. 1, mismatch in the ratios $R2/R1$ and $R3/R4$ results in different gains from each input to the output (neglecting the sign difference, which is obviously intentional). This means that signals common to both inputs will appear at the output to some degree. The accepted standard measurement of the extent to which this will occur is the CMRR. There are various ways to define the CMRR, but the classic definition is simply the differential-mode gain divided by the common-mode gain. This is a rather large number, and for this reason it is almost always expressed in (voltage) decibels:

$$\mathrm{CMRR(dB)} = 20\log_{10}\left(\mathrm{CMRR(ratio)}\right)$$

The sign of CMRR can also be confusing. Frequently, CMRR is expressed as something like -70 dB. Strictly speaking, this would imply a common-mode gain greater than the differential-mode gain, but (although it is possible to design an amplifier that would accomplish this) such figures should be treated as common-mode *acceptance* (the inverse of CMRR) when dealing with realistic instrumentation amplifiers. The CMRR of a simple instrumentation amplifier as depicted in Fig. 1 is obviously dependent on the accuracy to which the resistors can be matched. Using the symbol Δ_R, for resistor mismatch, and assuming no correlation in the matching, it can be shown that worst-case CMRR owing to this term alone can be expressed by

$$\mathrm{CMRR} = \frac{1+G}{4\Delta_R}$$

where G is the selected gain.

For off-the-shelf 0.1% tolerance resistors and a gain equal to 10, the CMRR could be as poor as 2750 or about 69 dB. This would be regarded as extremely meager in most precision applications. Clearly, a fine trim on any of the four resistors can be used to improve this parameter, and that is why integrated circuit instrumentation amplifiers are so commonly used where automated resistor adjustments (by means of a laser or other techniques) can be cost-effectively employed in their manufacture. Using such techniques, the critical resistor *ratios* can be trimmed to a tolerance of 0.005% or so, resulting in a CMRR in the preceding example of over 100 dB.

Even when the resistors are perfectly matched, common-mode errors in the op-amp itself cannot be overlooked. Common-mode rejection in the op-amp shows up directly as a CMRR error in the output multiplied by the gain. Modern precision monolithic op-amps frequently boast CMRR specifications of 110 dB or more. When the instrumentation amplifier is used at low gains, such figures can be considered negligible compared to errors such as resistor mismatch. At high gains, however, this may not be the case. The open-loop gain of the op-amp, ideally infinite, can also upset the behavior of both CMRR and gain error.

An often more serious consideration is the CMRR behavior with frequency. Both CMRR and gain roll-off with increases in frequency are typical of operational amplifiers with dominant-pole (or alternate) methods of compensation designed to be operated with good closed-loop stability and high feedback factors (often 100%). This effect is exacerbated in many precision op-amps that use multiple stages of gain, which, in turn, demand high levels of internal frequency compensation or, to put it another way, tend to have poor ac characteristics.

The most commonly encountered frequency of interest for CMRR is the fundamental frequency of the power-supply grid, which is nominally 60 Hz in the United States and Canada and 50 Hz in much of the remainder of the world. Additional harmonics generated by power transformers, rectifiers, and thyristor control systems produce radiated frequencies that can be far more important than initially expected, given that capacitive coupling tends to increase with frequency whereas op-amp precision tends to decrease with frequency. The subject of RF (radio frequency) susceptibility with frequencies up to several GHz is obviously an extreme (but often important) consideration in the evaluation of CMRR in an instrumentation amplifier.

A full dynamic analysis of factors affecting CMRR (and other effects, such as ac gain error, ac power supply rejection ratio [PSRR], and settling time) is considered beyond the scope of the present discourse. Suffice it to say, however, that even at 50 Hz, many precision op-amps can produce CMRR errors far worse than simple dc analysis would predict.

ADVANTAGES OF THE BASIC INSTRUMENTATION AMPLIFIER

One advantage of the simple amplifier of Fig. 1 is that only one op-amp is used. This is, of course, good not only for economic reasons but also because only one op-amp contributes to the overall error budget. A less obvious advantage is that the input voltage range is very high. Because of the attenuating effect of $R1$ and $R2$, the common-mode input range can extend beyond the input voltage range of the op-amp itself.

This is often useful where external signals approach or exceed the supply voltage available to the instrumentation amplifier.

SHORTCOMINGS OF THE BASIC INSTRUMENTATION AMPLIFIER

The most serious shortcoming of the configuration shown in Fig. 1 is that the input impedance is not very high (unless the resistors are made impractically large, with severe noise and bandwidth penalties). To make things worse, the analysis of the effects of the input impedance is highly dependent on the characteristics of the input signals.

The impedance at +IN is equal to $R1 + R2$, while at −IN it is equal to $R3$ because the feedback action of $A1$ forces a low impedance at its inverting input. For the usual arrangement, where $R2$ is equal to $R4$, and $R1$ is equal to $R3$ (to preserve symmetry at the op-amp inputs, for input bias current considerations among others), the input impedance is therefore considerably higher at +IN than at −IN. While this can be corrected by reducing the values of $R1$ and $R2$ (or better, by adding a resistor from +IN to ground), this is often a dangerous practice. For truly differential inputs with matched source impedances, the input impedance is not actually asymmetric (due to the fact that the inverting input of $A1$ follows part of the signal at +IN). Attempting to balance the absolute input impedances thus could cause severe CMRR errors when the inputs are driven differentially from well-matched source impedances. If the source impedances are not well matched (or at least not well defined), the usefulness of this configuration becomes highly questionable.

For the record, the common-mode input impedance turns out to be $\frac{1}{2}(R1 + R2)$ and the differential value is equal to $2(R1)$. These equations only hold for balanced differential sources.

Another nuisance is that the gain cannot be varied without simultaneously changing two resistors, and again tight matching must always be preserved if CMRR is not to suffer. However, if $R2$ and $R4$ are each split into two resistors, then an additional resistor connected between the midpoints can be used to increase the gain with minimal effect on CMRR. Now, of course, six matched resistors are required to obtain good CMRR. Such shortcomings are the main reasons to look to alternative instrumentation amplifier topologies.

Despite the limitations, the basic configuration of Fig. 1 is useful enough that at least two companies (Analog Devices and Burr-Brown) are producing this circuit in (laser-trimmed) monolithic form at the time of writing.

THE CLASSIC THREE-OP-AMP INSTRUMENTATION AMPLIFIER

Adding buffer amplifiers to both inputs will clearly remedy the input impedance problem of the aforementioned configuration, but a little more elaboration can solve the gain-setting problem as well. This leads to the classic three-op-amp configuration of Fig. 2. Once more, a definitive reference has proven to be elusive, but at the very least it dates back to a 1966 George Philbrick publication (5).

Referring to Fig. 2, $A3$, in conjunction with $R1$–$R4$, forms a differential amplifier identical to the circuit of Fig. 1. If R_G is omitted, $A1$ and $A2$ act as unity-gain buffers, removing the input impedance problems described previously. In the presence of R_G, the differential gain between the inputs and outputs of $A1$ and $A2$ becomes

$$G_{\text{diff}} = 1 + \frac{(R5 + R6)}{R_G}$$

The common-mode gain, however, remains at unity. Thus, when $R1$–$R4$ are carefully trimmed for optimum CMRR, the differential gain can be increased by reduction of R_G without affecting overall common-mode gain. Because of the way CMRR is defined (as the ratio of differential-mode gain to common-mode gain), the effective CMRR of the amplifier becomes proportional to G_{diff}. The overall gain of the amplifier will be the product of the gain of the second stage and G_{diff} or, (assuming $R1/R2 = R3/R4$),

$$G = \left(1 + \frac{(R5 + R6)}{R_G}\right)\left(\frac{R4}{R3}\right)$$

Thus the lower limit on gain is set by $R4/R3$.

The best distribution of the gain between the first and second stages is the subject of considerable compromise. By us-

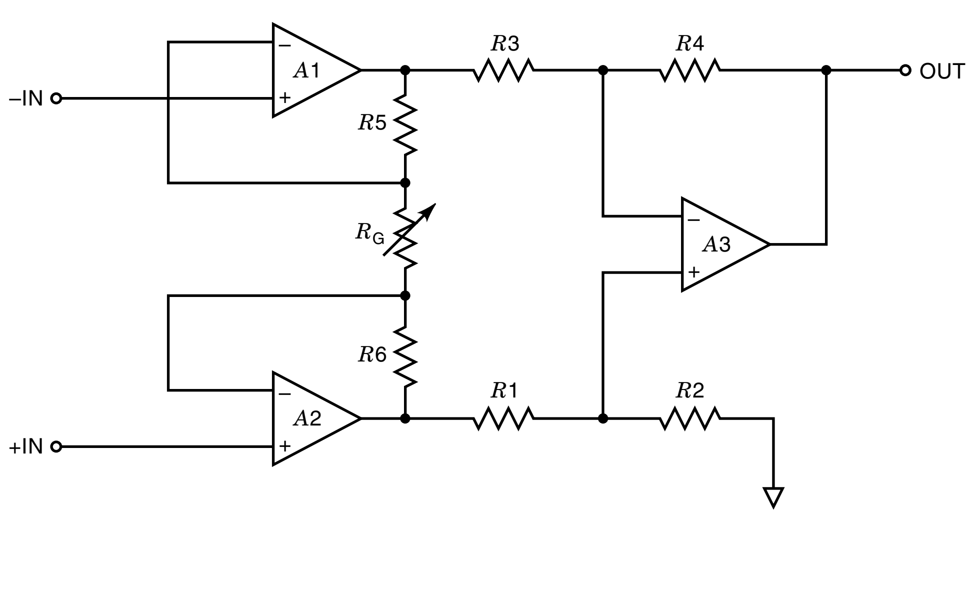

Figure 2. Three-op-amp instrumentation amplifier.

ing a gain below unity in the second stage ($R4 < R3$), the overall CMRR (for a given resistor mismatch) is increased, and the lower gain limit is extended. Unfortunately, this tends to put the gain burden on the op-amps $A1$ and $A2$, thus amplifying their input referred errors and reducing their bandwidth. Conversely, taking gain from the second stage ($R4 > R3$) can improve the overall bandwidth at the expense of CMRR (and amplification of the input referred errors of $A3$) and increases the limit on the lowest achievable gain.

A more subtle (but extremely important) effect, however, is the limit imposed on the common-mode input range. With no differential input, the common-mode input range is limited either by the input voltage range or the output voltage swing (whichever happens first) of $A1$ and $A2$. With modern "rail-to-rail" op-amps, this restriction can often be minimal (though most high-precision and high-speed op-amps still require considerable headroom, often a couple of volts or so). Even so, the power supplies dictate a limit on the output swing.

With a differential input applied, the output of either $A1$ or $A2$ swings positive about the common-mode input voltage, while the other swings negative. Thus the entire differential output voltage of the first stage directly subtracts from the available common-mode input range. This effect also limits the output swing of the overall amplifier in the case where the second stage is operated at a gain below unity.

As an extreme example, consider the case in which the supply voltages are ± 15 V and all op-amps are limited in range only by these supplies. Also consider the case in which the second stage is operated at a gain of 0.1 ($R3/R4 = 10$). With the outputs of $A1$ and $A2$ at opposite supply rails, the output swing will be only 3 V, and even if a 1 V output swing is all that is required, the common-mode input range will be reduced by 10 V, barely making a ± 10 V common-mode input range with symmetric swings for the input amplifiers.

A more practical example might be the case in which the second stage gain is set to unity and ± 10 V swings are required at the output. This output swing can now be achieved, but again the common-mode input range is still barely ± 10 V, even with ideal amplifiers. For realistic op-amps requiring 2 V of headroom, the common-mode input range is at best ± 8 V with a full differential signal applied. If this range is unacceptable, then gain must be provided in the second stage.

The symmetric swing referred to previously actually occurs when $R5$ is equal to $R6$. This is highly desirable because it maintains as much symmetry as possible in the two input op-amps. Making $R5$ and $R6$ dissimilar can cause offset problems (due to the input bias currents of the op-amps flowing through unequal impedances) and ac CMRR problems (because the effective closed-loop gains of the amplifiers are different). Purposely making $R5$ and $R6$ different values can position the common-mode input range closer to one supply or the other, but this is usually a poor solution to the problem.

Suffice it to say that most commercial implementations of this architecture (and there are many) have used symmetric values for $R5$ and $R6$ and a second stage gain of unity (occasionally greater). Even with these limitations, this is one of the most powerful configurations available for instrumentation amplifier realization.

REFERENCE INPUT

Because the instrumentation amplifier has a differential input and single-ended output, it is necessary to refer the output voltage to some reference potential. So far, this has been represented by a symbol commonly referred to as ground, which is a somewhat universal reference point in most analog systems. Often, however, the integrity of a global ground connection is highly questionable, particularly when high degrees of analog precision are sought. In some cases, it may be necessary to refer the output to some other potential; this is particularly true where only a single supply voltage is available and the so-called ground is actually one of the supplies (usually the negative one). In the circuit of Fig. 2 the ground connection is made via $R2$, but in fact this point can be used as a reference input since it has (close to) unity gain to the output. Most modern, commercially available instrumentation amplifiers therefore label this pin as reference rather than ground. In the interest of clarity, many of the figures in this article show the reference input as a ground symbol. For the purpose of this article, the ground symbol may be assumed to correspond to a reference input, which can be used to refer the output to a different point (or a precision Kelvin ground) inside the system.

SENSE INPUT

The overall feedback loop for an instrumentation amplifier is often completed externally to allow for Kelvin sensing of remote potentials where there exists the possibility of significant voltage drops along the connecting wires (due to finite loading of the amplifier output). Thus, some instrumentation amplifiers feature a sense input, which is almost always connected to the final amplifier output, either locally or remotely via force and sense wires.

Care must be used when taking advantage of this feature, however, since most instrumentation amplifiers can become unstable when presented with large amounts of capacitance at the output, a condition often created by long wires connected to the output (and sense) pins.

INPUT AND OUTPUT REFERRED ERRORS

This is a good point at which to introduce the concept of input and output referred errors. In virtually all variable-gain instrumentation amplifiers, most of the error terms actually have two components. One is called output referred, since it appears at the output independent of gain setting. Examples of this are the CMRR error caused by mismatch of resistors $R1$–$R4$ in Fig. 2 and all errors caused by op-amp $A3$. The other component is called input referred, and it appears at the output multiplied by the overall gain of the amplifier. Examples of this are almost all errors attributable to the input op-amps $A1$ and $A2$. Most instrumentation amplifier data sheets specify these terms (such as input offset voltage and output offset voltage) separately. Since the sign is generally unpredictable, at any particular gain the terms are usually presumed to add at the output (except for noise, where an RSS [root sum of squares] summation technique is generally applied).

In passing, it should be noted that it is not so much the individual errors of $A1$ and $A2$ that appear as the input referred error (again, except for noise) but the difference between them, since any systematic errors appear as a common-mode signal to the second stage. For this reason, a matched monolithic dual op-amp is usually used for $A1$ and $A2$ in the configuration of Fig. 2.

THE TWO-OP-AMP INSTRUMENTATION AMPLIFIER

The simple circuit of Fig. 1 attempts to balance the inverting and noninverting gains of an op-amp operated in a closed-loop configuration by attenuation of the signal at the noninverting input. Another alternative is to leave the noninverting input alone (thus maintaining its inherently high input impedance) and to use a second op-amp to balance the gains by amplifying the gain at the inverting input. Since the latter op-amp can be operated in a noninverting gain configuration, a high input impedance for both final instrumentation amplifier inputs can be preserved. The basic circuit is shown in Fig. 3. The op-amp $A2$ provides the differential function, while $A1$ amplifies the $-$IN input to equalize the gains between the $+$IN and $-$IN inputs. The incremental transfer gain from $+$IN to the output is given by

$$G+ = 1 + \frac{R4}{R3}$$

whereas the incremental transfer gain from $-$IN to the output is given by:

$$G- = -\left(1 + \frac{R2}{R1}\right)\left(\frac{R4}{R3}\right)$$

To obtain good CMRR, the absolute value of these gains must again be equalized, and this is achieved by making the ratio of $R2/R1$ equal to $R3/R4$, in which case the common-mode gain is theoretically zero and the differential gain is given by

$$G_{\text{diff}} = 1 + \frac{R4}{R3}$$

The resistor matching requirements for CMRR are very similar to the simple configuration of Fig. 1, except that the CMRR depends on $G/4\Delta R$ rather than $(1 + G)/4\Delta R$. Choosing $R2$ equal to $R3$ and $R4$ equal to $R1$ balances input bias current errors of the op-amps (but not their dynamic characteristics—more on this later). Of more importance, however, are the gain and common-mode input range limitations. The first

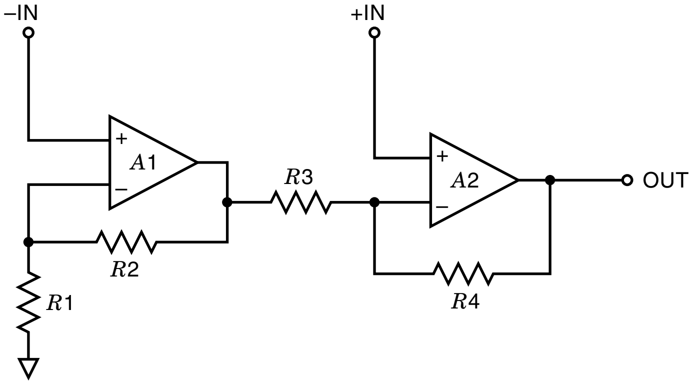

Figure 3. Two-op-amp instrumentation amplifier.

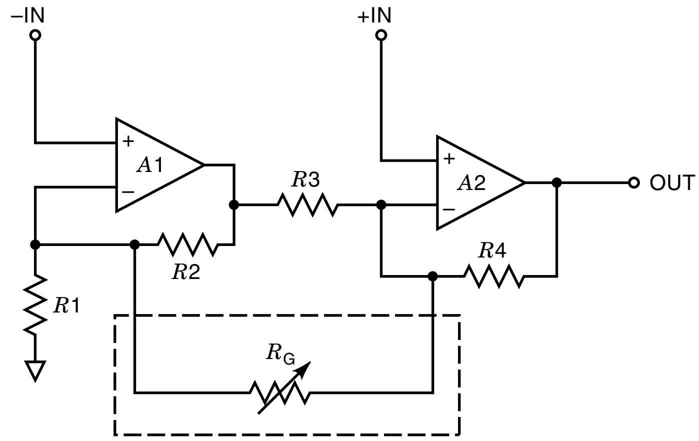

Figure 4. Modified two-op-amp instrumentation amplifier.

thing to be noted is that unity gain (or anything less) is impossible, because under these conditions the value of $R2$ must be infinite. Second, even at somewhat higher gains, it should be noted that $A1$ *always* amplifies the common-mode voltage, producing severe limitation of common-mode input range due to available swing at the output of $A1$. For this reason, it is rare to find this configuration used in an overall gain of less than five (the closed-loop gain of $A1$ asymptotically approaches unity as the overall gain is increased, progressively ameliorating this problem).

Another limitation concerns the ac characteristics. Like the circuit of Fig. 2, a matched dual monolithic op-amp used for $A1$ and $A2$ can contribute greatly to the goal of good dc precision. Unlike the circuit of Fig. 2, however, the op-amps operate under very different individual closed-loop conditions. This causes problems when trying to maintain a good CMRR versus frequency, particularly since the entire phase shift of $A1$ appears in the inverting path but is totally absent in the noninverting one. Some phase compensation techniques can be applied to help this situation, but for the most part if good ac characteristics are required, this is unlikely to be the topology of choice.

A final note concerns the variable-gain characteristics of the circuit of Fig. 3. Like the circuit of Fig. 1, any gain change requires that the ratio matching of $R2/R1$ to $R3/R4$ be left unchanged if CMRR performance is not to be degraded. Bearing in mind that there is a practical minimum gain for this configuration, there is a way to increase the gain without severe CMRR penalty. Figure 4 shows the modification. The addition of R_G between the inverting inputs of $A1$ and $A2$ modifies the gain equation to

$$G = \left(1 + \frac{R4}{R3}\right) + \left(\frac{2R4}{R_G}\right)$$

Using this technique, some commercial realizations of this configuration have had their usefulness greatly extended. Companies such as Linear Technology and Burr-Brown feature this configuration in their product portfolio (at the time of writing) in the form of monolithic integrated circuits.

AN ALTERNATIVE APPROACH USING THREE OP-AMPS

Another possibility using three op-amps is shown in Fig. 5 (6). Op-amp $A2$ does most of the work, with $A1$ (connected as

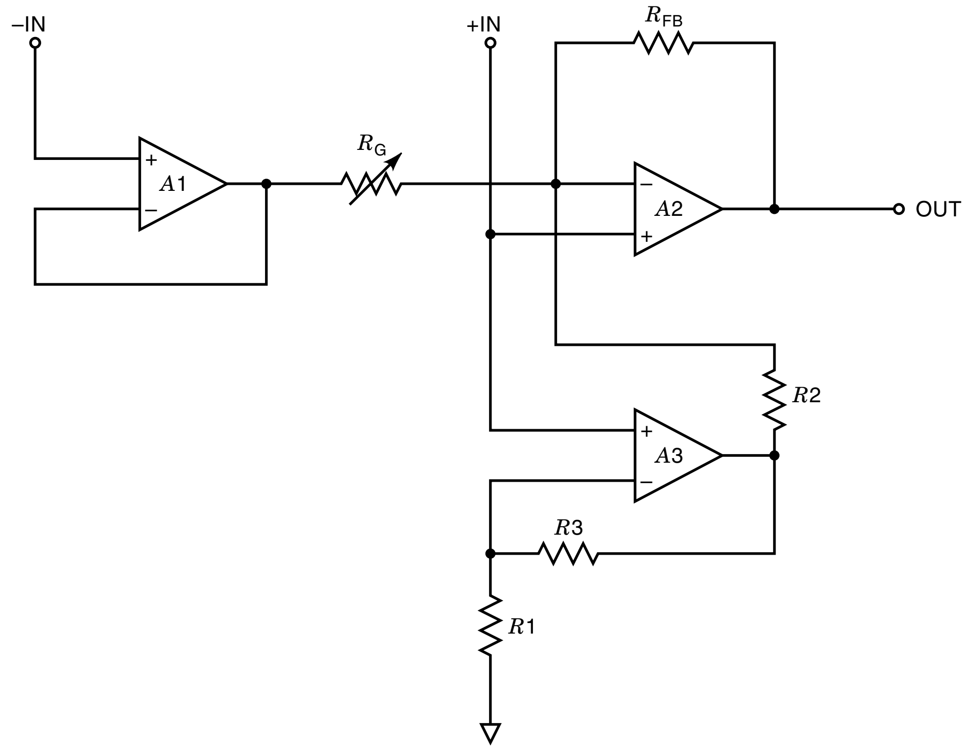

Figure 5. An interesting alternative three-op-amp configuration.

a unity-gain buffer) restoring a high impedance to the inverting input (−IN). In this case, however, the inverting and noninverting gains of $A2$ are equalized by using a third op-amp ($A3$) to provide active attenuation of the common-mode signal appearing at −IN. This common-mode signal is impressed across $R1$ and, by the action of $A3$, is injected as a nulling current back into the inverting input of $A2$. It can be shown that optimum CMRR is obtained when the ratio of $R3/R1$ is made equal to the ratio of $R2/R_{FB}$. For symmetry, generally $R1$ will be made equal to R_{FB} and $R2$ will be made equal to $R3$. This topology has some interesting characteristics:

1. CMRR can be trimmed by a fine adjustment on any of $R1$, $R2$, or $R3$ without affecting the gain of the amplifier.
2. The overall differential gain is simply R_{FB}/R_G. Thus gains from zero to any practical value are available by adjusting only one component (R_G). The CMRR (referred to output) is not affected by this gain adjustment.
3. The circuit is capable of a very wide common-mode input range. Op-amps $A1$ and $A2$ limit this range by their input/output swings in the usual fashion, of course, but the main limitation comes from the output swing of $A3$ since it amplifies the common-mode voltage by the factor $(1 + R3/R2)$. This limitation can be removed, however, by choosing the ratio of $R3/R2$ to be small enough that the other amplifiers become the limiting factor. The common-mode input range of the overall amplifier can thus approach the limits of the op-amps themselves.

This circuit (as might be anticipated) is not without drawbacks. The most significant of these is due to the fact that the errors of $A3$ are effectively amplified by the ratio $R_{FB}/R2$ (this is an output referred error, not affected by gain setting). So attempts to maximize the common-mode input range in this manner tend to produce higher output offset voltages and noise. Also, $A1$ and $A2$ operate in very different closed-loop conditions, making it difficult to maintain good CMRR at high frequencies. Finally, the inverting input (−IN) is loaded by only one op-amp, whereas the noninverting input (+IN) is loaded by two op-amps, making the input characteristics somewhat asymmetric, especially in the case where the op-amps have significant input bias currents. Nevertheless, this configuration has been used (with some modification) to produce a monolithic instrumentation amplifier capable of operating from a single 5 V supply, where the common-mode input voltage range extends all the way to the negative supply rail (7).

VENTURING BEYOND OP-AMP DESIGN TECHNIQUES

The examples presented so far have relied on configurations formed from conventional op-amps and resistors. True, all of the aforementioned circuits have been integrated in monolithic form at some time or another; but they all could also be effectively produced using commercially available off-the shelf components. In parallel with developments along these conventional lines, the monolithic integrated circuit industry has produced many topologies for the realization of the instrumentation amplifier function, most of which would be difficult, and certainly uneconomic, to produce outside the environment of a monolithic integrated circuit. The remainder of this article focuses on some of the more important techniques used to produce monolithic instrumentation amplifiers.

CURRENT-FEEDBACK TECHNIQUES

The traditional design approach of using conventional operational amplifiers with feedback consisting of resistive components is nowadays often referred to as *voltage feedback*. In the mid-1980s a new term started to appear in op-amp literature: *current feedback*. A current-feedback op-amp differs from a conventional one in that its inverting input is internally held at a low impedance; the displacement current in the compensation capacitor is ultimately derived from the current flowing in the feedback network (8). This enables such op-amps to have very high slew-rate characteristics. Also, to the extent that the inherent input impedance of the inverting input is less than that of the feedback network, such op-amps maintain a more constant bandwidth as the closed-loop gain is increased than their conventional counterparts (which tend to have a fixed gain-bandwidth product). The drawback here is that such op-amps have intrinsically imbalanced input stages and cannot approach the precision of more conventional types, despite many ingenious schemes to balance them up. Actually, current-feedback is an offshoot of what used to be called *cathode feedback* in the vacuum-tube days and is not a fundamentally new technique from a circuit theory viewpoint.

Applying current feedback to an instrumentation amplifier is actually much easier than in the case of a general-purpose op-amp (although all configurations described so far can be implemented using current-feedback op-amps). This is because the instrumentation amplifier is an inherently dc balanced structure with well-defined feedback components.

Figure 6 shows a current-feedback approach to the configuration of Fig. 2 (9), now produced in integrated circuit form by several manufacturers. Essentially, this consists of the classic three-op-amp design preceded by a preamplifier consisting of $Q1$ and $Q2$ and their associated load resistors, $R5$ and $R6$. Feedback (via the resistors R_{FB}) is now returned directly to the emitters of the input pair rather than to inherently high-impedance op-amp inputs, as in previous examples. The bias currents for $Q1$ and $Q2$ are not set by current sources $I1$ and $I2$ (as might first be thought) but rather are provided from the outputs of $A1$ and $A2$, in a common-mode feedback loop controlled by V_{bias}. Since these currents must flow through the feedback resistors (R_{FB}), $I1$ and $I2$ are added in order to center the common-mode swing at the outputs of $A1$ and $A2$ (these current sources are sometimes omitted when the input stage currents are small enough to produce negligible voltage drops across the feedback resistors).

The gain equations for this arrangement are identical to that of the example presented in Fig. 2, at least under dc conditions. However, to the extent that the dynamic impedance at the emitters of $Q1$ and $Q2$ is lower than the value for R_G, the latter component does not greatly attenuate the overall ac feedback, resulting in an approximately constant bandwidth (rather than a constant gain-bandwidth product typical of voltage-feedback configurations) as R_G is varied. The configuration does slow down at higher gains (as R_G becomes comparable to or less than the input transistors' dynamic emitter impedance), but it can still offer a considerable practical improvement in overall bandwidth and settling time.

Other advantages stem from the fact that only two transistors comprise the input stage, rather than the four necessary for two conventional op-amps. This leads to reduced input-referred errors (particularly noise). A minor disadvantage

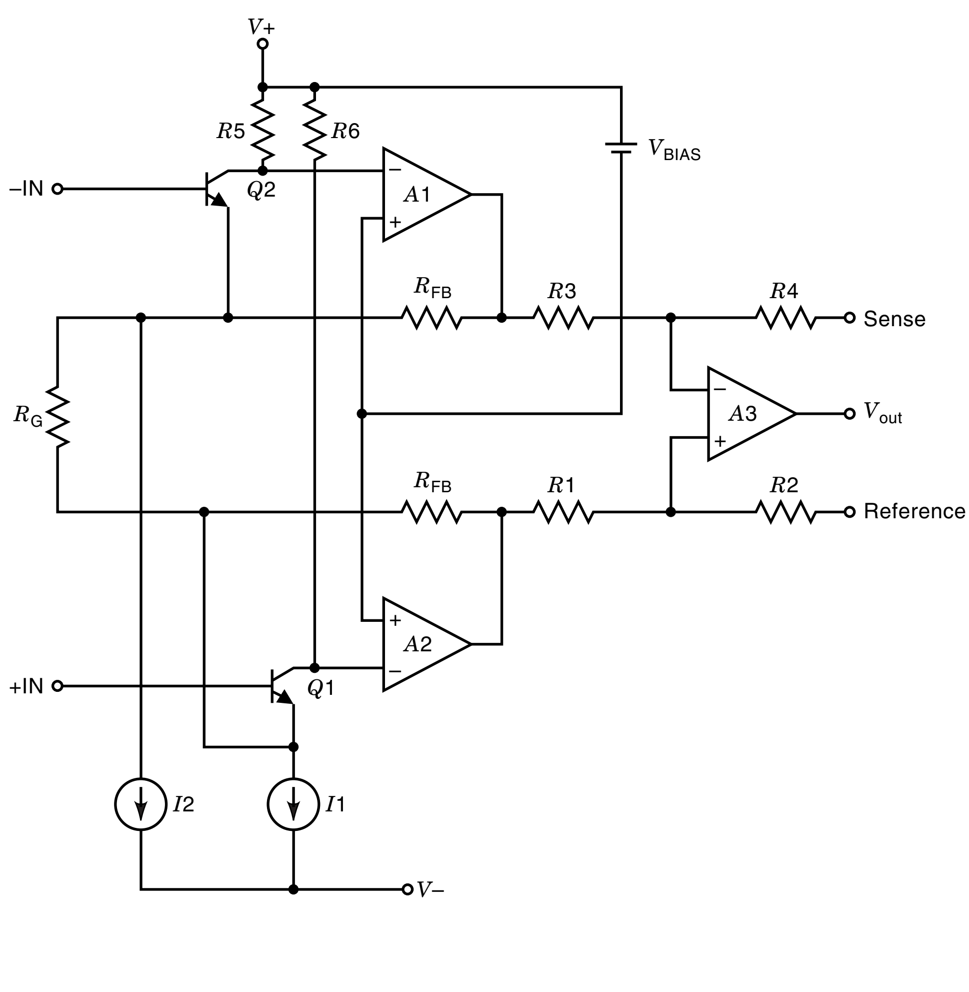

Figure 6. Current-feedback version of Fig. 2.

comes from the fact that the uncorrelated portion of the error currents in $I1$ and $I2$ (and the input transistors) has to flow through the feedback resistors and appears as an increase in output referred offset and noise. With careful design, however, this effect can be kept small. Similar current feedback configurations can be devised for the other voltage-feedback topologies presented previously.

ACTIVE FEEDBACK

In the sense used here, *active feedback* involves the use of an active voltage-to-current converter as a feedback element, instead of the resistor, which until now has been shown as the basic feedback component. Confusingly, the term *current feedback* has also been employed for this type of feedback, and there is no real accepted standardization in common usage (maybe this article will encourage such a standard). At least I feel I have defined my own nomenclature (with some historical justification), but for the record I am one of the many whose past publications have (unintentionally) contributed to the present state of confusion.

CONCEPTUAL ACTIVE-FEEDBACK INSTRUMENTATION AMPLIFIERS

In the basic circuit of Fig. 7, two identical differential bipolar transistor pairs (they could be field-effect transistors in theory) are degenerated by R_{G1} and R_{G2} (assumed to be equal for now, as are the current sources $I1$ through $I4$). The degeneration resistors are large enough to ensure that any desired differential input voltage (impressed between +IN and −IN) will retain some sensible current in both of the input transistors, $Q1$ and $Q2$. Both differential pairs are summed into differential loads consisting of $R1$ and $R2$, which are further sensed by op-amp $A1$.

A negative feedback loop is provided from the output of $A1$ to the second differential pair (note that the base of $Q3$ is grounded). When the loop is closed (frequency compensated by capacitor C_c), the output voltage will thus be closely equal to the differential input, independent of the common-mode input voltage (10). The nonlinearities of the two transistor pairs nominally cancel under this arrangement; thus the circuit forms a unity-gain amplifier with common-mode rejection limited only by second-order nonidealities in the transistors themselves. The most important nonideality (for CMRR purposes) is the mismatch of output impedance of the two transistor pairs. Without resort to any kind of trimming, this can be a large number indeed, leading to a high CMRR.

The drawback of this arrangement is that it is difficult to adapt it to variable gain. At first glance, making R_{G1} a variable component does the trick (and will result in variable gain), but under these circumstances the nonlinearities of the transistor pairs no longer cancel (except at unity gain), resulting in cubic distortion products that are likely to be unacceptable at higher gains.

Another idea is to place a resistive attenuator between the output of $A1$ and the base of $Q4$. While this technique will preserve linearity, it has the unfortunate effect of amplifying both input and output referred errors. If the gain range is small and the corresponding input range is well defined, a very good amplifier can result, since the transconductance of both stages can be optimized. For wide gain ranges, however, the result is a noisy instrumentation amplifier with poor output referred errors and loss of bandwidth at high gain settings.

One possible method of producing a more usable instrumentation amplifier from this general idea is shown in Fig. 8. In this configuration, the nonlinearities of the transistor pairs are corrected by enclosing them in localized operational amplifier feedback loops. This enables the gain to be varied (without distortion penalties) by adjustment of the ratio of R_{G1} to R_{G2}. Some merging of the op-amp functions is possible, and a monolithic design based on this idea has been reported in the literature (11).

One interesting aspect of this configuration is that the major feedback loop (completed by op-amp $A1$) is almost totally divorced from the input section. This effectively means that as R_{G1} is reduced to increase gain (assuming R_{G2} is left alone), the loop bandwidth does not change. Thus the overall amplifier bandwidth tends to remain constant even as the gain is varied over a wide range (gains of zero to any practical upper limit are possible with this arrangement and with most other active-feedback configurations).

Unfortunately, this configuration has developed a rather complex input stage, which tends to produce high levels of input referred errors (particularly input offset voltage, input offset drift, and input referred noise). Other schemes that do not need such a complex input stage have been developed in

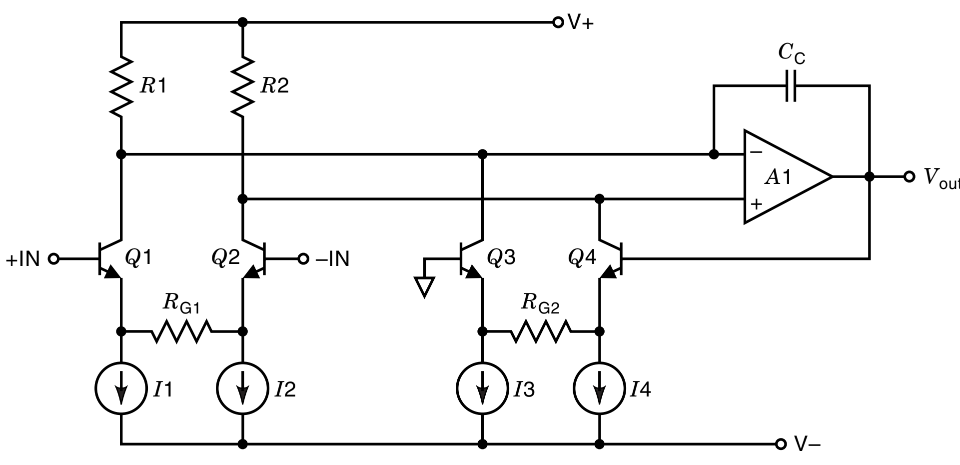

Figure 7. Conceptual active-feedback instrumentation amplifier.

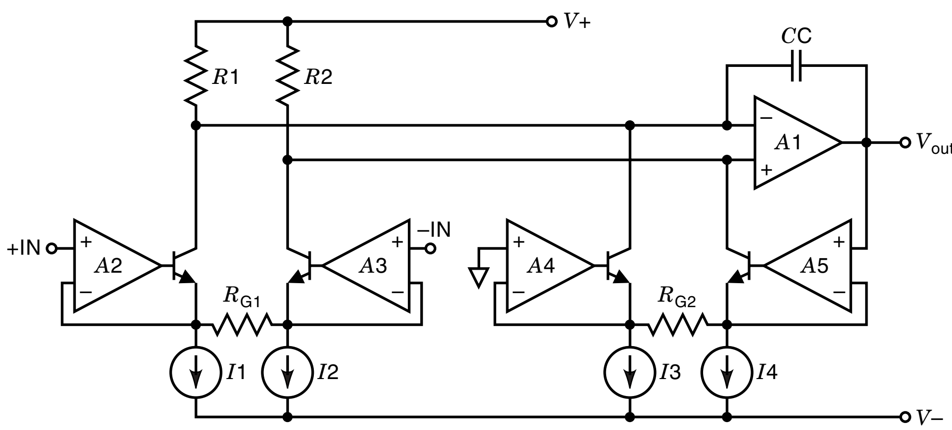

Figure 8. Active-feedback instrumentation amplifier with variable gain.

order to alleviate this situation. Before continuing, however, a note concerning output referred errors is in order. Active-feedback amplifiers, in general, tend to have high output referred errors (their major drawback compared to more conventional techniques using op-amps and resistors). This is because the active voltage-to-current converters used to provide the feedback have much larger offset and noise components than simple resistors. The high degree of gain flexibility (usually from zero upward), intrinsically high common-mode rejection (without trimming), and potentially high speed characteristics of the active-feedback instrumentation amplifier come with a penalty in terms of precision at low gains. As a crude generalization, modern active-feedback instrumentation amplifiers offer significant advantages in terms of speed and CMRR at most gains but tend to be poor in most other respects at gains below 50 or so. At lower gains (certainly below 10), more conventional techniques are likely to provide better overall performance and, in conjunction with the current-feedback approach, are likely to be competitive in terms of speed.

PRECISION ACTIVE-FEEDBACK INSTRUMENTATION AMPLIFIERS

The basic idea for an active-feedback amplifier with a precision input stage is depicted in Fig. 9. Transistors $Q1$ and $Q2$ are biased at a quiescent point determined by the standing currents from a highly linear voltage-to-current (V-to-I) converter. Feedback provided by op-amp $A1$ to the V-to-I converter controls the differential characteristics. If the open-loop gain is high, the feedback loop will force the currents in $Q1$ and $Q2$ to be equal—regardless of differential input—while the output currents of the converter are not equal due to the presence of the gain-setting resistor, R_G. Because the input transistors are operated under identical conditions, the differential input voltage is directly forced across R_G, so the differential output currents of the V-to-I converter are $2(\Delta V_{in}/R_G)$. Since the V-to-I converter is presumed to be linear, the output voltage is now equal to the input voltage multiplied by the product of R_G and half the differential transconductance of the converter, independent of the common-mode input voltage because of the converter's inherently high output impedance.

The output referred errors (as in previous examples) are those of the active V-to-I converter and can be comparatively high. The input errors, though, can closely approach the intrinsic errors of $Q1$ and $Q2$ alone, yielding a theoretical input stage precision about as good as anything available on a monolithic integrated circuit.

The structure of the V-to-I converter is obviously critical to the performance of this topology, and various methodologies have been used from time to time. With the inherent advantage of high CMRR without trimming, it is not surprising that the first fully integrated monolithic instrumentation amplifier would be an active-feedback design (12) (which became commercially available as the Analog Devices part number AD520). Four years later, a much improved design (13) was introduced (the AD521). Figure 10 shows the basic topology. Amplifier block $A2$ adjusts the current sources $I3$ and $I4$ to maintain constant currents in the input transistors, $Q3$ and $Q4$. Under these circumstances, the differential input voltage is accurately forced across the resistor R_G. The difference in $I3$ and $I4$ is now simply twice the input voltage divided by R_G. The current sources $I1$ and $I2$ are slaved to $I3$ and $I4$, so their difference is exactly the same. The amplifier $A1$ forces half of this difference to appear across the resistor R_s. The gain is now simply R_s/R_G. In the example presented, both R_s and R_G were external components, although there is not much

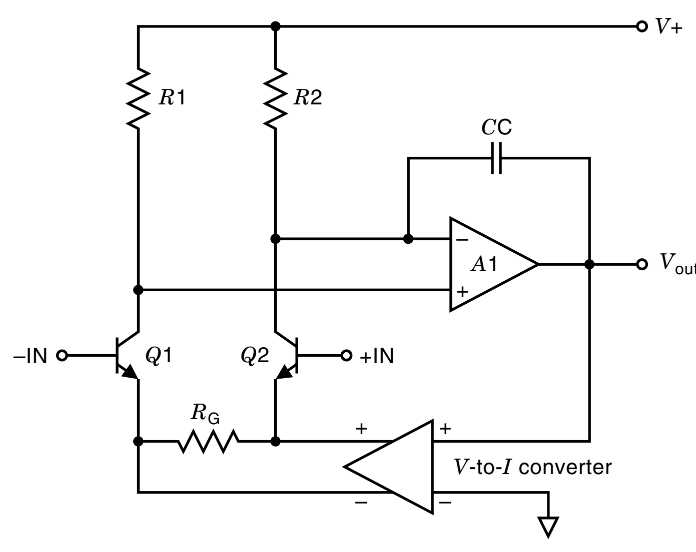

Figure 9. Conceptual topology for a precision active-feedback instrumentation amplifier.

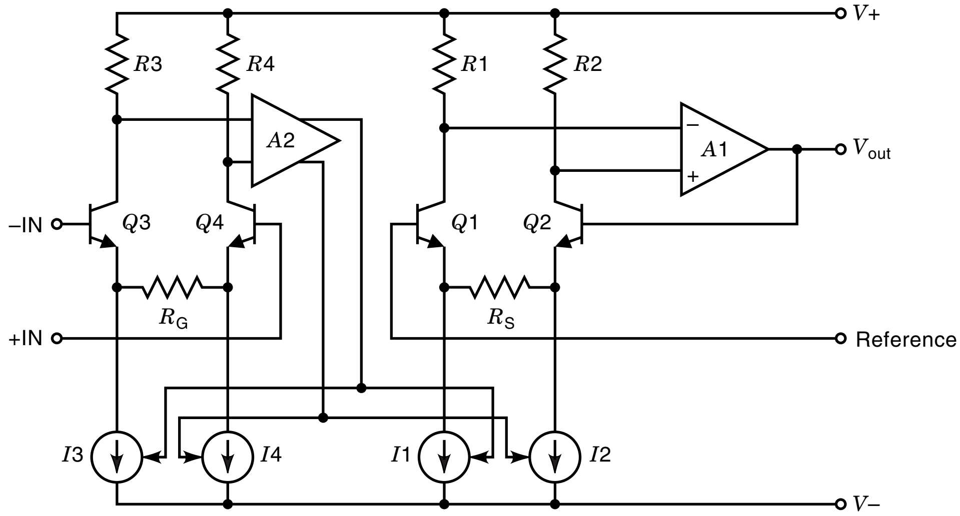

Figure 10. Active-feedback instrumentation amplifier using a parallel *V*-to-*I* converter.

flexibility in the choice of R_s because of limitations of the maximum values of the internal current sources.

The need for two amplifier blocks and parallel-connected controlled current sources tends to exacerbate output referred errors in such designs, and one method of alleviating this is to place the *V*-to-*I* converter in series with the input stage. This is not a trivial task, since the resulting *V*-to-*I* converter must reside entirely in the space left between the extremes of the common-mode input range and one of the supplies (usually the negative). One method of achieving this is shown in Fig. 11, first implemented by National Semiconductor (14).

The *V*-to-*I* converter is composed of *A*2, *A*3, *Q*3, and *Q*4 with *I*1 and *I*2 providing emitter bias current for the input stage, *Q*1 and *Q*2. To prevent negative common-mode excursions at the input from causing *Q*3 and *Q*4 to saturate, the output is attenuated and level shifted by resistors *R*3 and *R*4. Balance is restored to the other side of the *V*-to-*I* converter by an identical network consisting of *R*5 and *R*6. *A*2 and *A*3 can now be made extremely simple because no level shifting is required and any systematic offset will cancel between the two op-amps. One drawback of this technique is that mismatches in the ratios of resistors *R*3/*R*4 and *R*5/*R*6 can cause severe degradation of the negative power-supply rejection.

The effective differential transconductance of the *V*-to-*I* converter is twice the inverse of R_{G2}, and similarly the effective input stage transconductance is twice the inverse of R_{G1}. The overall transfer function is now given by

$$G = \left(\frac{R4}{R3 + R4}\right)\left(\frac{R_{G2}}{R_{G1}}\right)$$

For the example of Ref. 14 (National part number LM363), the gains are selected by pin-strapping internal resistors to

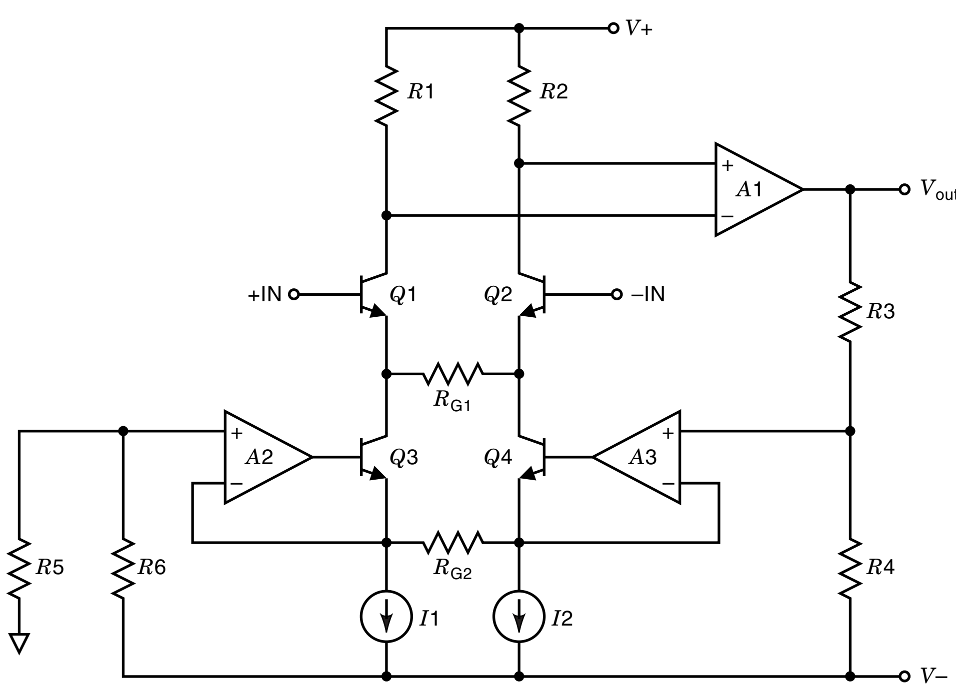

Figure 11. Active-feedback instrumentation amplifier using a serially connected *V*-to-*I* converter.

provide gains of 10, 100, or 1000 (fixed-gain versions with gains of 10, 100, or 500 are also provided).

Other implementations of this architecture (15,16) (Analog Devices part numbers AMP-01 and AMP-05) have left R_{G1} and R_{G2} as external components with the ratio of ($R3$ + $R4/R4$) internally set at 20. This results in an overall differential gain equal to $20(R_{G2}/R_{G1})$ and, like most configurations of the active-feedback amplifier, allows a user-defined gain setting from zero to any sensible upper limit.

GLOSSARY OF FREQUENTLY ENCOUNTERED TERMS

Gain

The gain specification relates to the transfer function of the instrumentation amplifier. Typically, gain can be fixed, adjustable by pin-strapping, digitally selectable, or controlled by an external resistor (R_G). In the latter case a transfer equation will normally be provided.

Gain Range

The gain range is the overall range over which the gain equation is considered valid. At the lower end it is generally limited by the type of configuration used, while at the upper end it is often theoretically unlimited. Practically, at very high gains, all instrumentation amplifiers eventually exhibit errors that render them unusable. When an upper figure appears on a data sheet, it is usually the point above which the device manufacturer is not willing to provide any guaranteed specifications.

Gain Error

The number given by the gain error specification (usually expressed as a percentage) describes the maximum deviation from the gain equation. This is, for convenience, often quoted (and tested) at several fixed gains, with the user left to interpolate between them. Note that when an external resistor appears in the gain equation, the absolute tolerance of this component also appears as part of the gain error.

Nonlinearity

The instrumentation amplifier is assumed in simple theory to have a linear transfer characteristic from the differential input to the output. Obviously, in practice some nonlinearity will exist, and this represents an error that is nearly impossible to trim out or compensate for.

Nonlinearity is generally expressed as the peak deviation from a straight line superimposed on the plot of the output voltage as the input voltage is varied (at a particular gain setting) over a range wide enough to sweep the output over a specified excursion range. It is normally expressed as a percentage of the *maximum output excursion,* not the value of the output excursion at any other point. As such, nonlinearity is thus generally defined as an output referred error, and specifications will usually be quoted at a selection of representative gains.

The manner in which the straight line is defined can also cause some confusion. A line drawn between zero and theoretical full scale is probably the obvious one, but one argument suggests that this would produce a nonlinearity specification that includes other error terms (particularly offset voltage and gain error), which are generally specified separately (and in theory can be calibrated out). If the resulting line is drawn between measured zero and full scale, complications can occur when positive and negative signals are accommodated, since the resulting line drawn between the positive and negative full-scale output may not pass through zero. Generally, a line drawn between two arbitrary points (such as zero and theoretical full scale) gives rise to the term *end point nonlinearity,* whereas a line skewed to pass through two or more measured points with minimum peak error is termed best-fit nonlinearity. Careful reading of a data sheet specification is necessary to determine the effect that a given nonlinearity specification will have on an actual system.

Offset Voltage

The offset voltage of a practical instrumentation amplifier consists of two terms: the input offset voltage and the output offset voltage. The input offset voltage is an error voltage that appears at the output multiplied by the selected (differential) gain of the amplifier. Therefore, it tends to predominate when the amplifier is configured for high gain applications. The output offset is an error term always present at the output, regardless of gain setting. Theoretically, it is defined as the error voltage at the output when the gain is set to zero, though this has to be extrapolated for many configurations where zero gain setting is impossible. The output offset voltage is obviously most troublesome at low overall amplifier gains.

Input (Bias) Current, Offset Current, and Input Impedance

The input current is simply the current drawn by one or both of the inputs when the amplifier is operated in its normal region (sometimes expressed as an average, or as a maximum of the two). This is often called an input bias current because such currents (at least in a bipolar junction transistor amplifier) are the base currents of the input transistors necessary to maintain them at their selected bias point. For amplifiers with field-effect transistor inputs, the input current generally reflects leakage currents associated with details of their fabrication.

The input offset current is the difference between the two input currents, of paramount importance when balanced source impedances are used. This is a measure of how well the two input currents are matched on a particular device.

Related specifications include differential and common-mode input impedance. The former represents the change in input offset current when a differential voltage is applied; it is normally defined as the reciprocal of the change in input offset current times the differential voltage used to induce it. The common-mode input impedance is normally defined as the reciprocal of the change in the sum of the two input currents multiplied by the common-mode voltage used to induce it. For most modern instrumentation amplifiers, the effect of both input impedance terms is usually negligible compared to the overall values of the input currents.

Common-Mode Rejection Ratio

The CMRR is a measure of the change in output voltage when both inputs are changed by equal amounts. These specifications are usually given for both a full-range input voltage change and a specified source imbalance. Because CMRR is usually a large number, it is usually expressed in decibels. For example, a CMRR of 10,000 could be expressed as 80 dB. Because CMRR generally consists of both input and output referred components but is always specified referred to the input, it will normally (apparently) increase with gain. For this reason, CMRR is almost always specified at several representative gain settings.

Common-Mode Input Voltage Range

The common-mode input voltage range represents the maximum excursion common to both inputs over which the CMRR specifications are guaranteed. For some instrumentation amplifiers, this is a function of differential input voltage, and often the input voltage range will be expressed by an equation rather than a fixed value. Another way to express the input voltage range is to specify a maximum excursion for either input, since for some amplifiers (particularly active-feedback types) this is a more realistic definition.

Power-Supply Rejection Ratio

The power-supply rejection ratio (PSRR) is a measure of the change in output voltage either when both power supplies are changed by equal amounts (in opposite directions, to remove any CMRR component) or when each supply is varied independent of the other (of course, there is only one supply to be varied in the case of a single-supply amplifier). Like CMRR, PSRR is often expressed in decibels, generally consists of both input and output referred components, is normally specified referred to the input, and will normally increase with gain.

Settling Time

Settling time is defined as that length of time required for the output voltage to approach and remain within a certain tolerance of its final value. It is usually specified for a fast full-scale input step and includes output slewing time. Since several factors contribute to the overall settling time, fast settling to 0.1% does not necessarily mean proportionately fast settling to 0.01%. In addition, settling time is not necessarily a function of gain. Some of the contributing factors include slew rate limiting, underdamping (ringing), and thermal gradients (long tails).

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INSTRUMENTATION FOR PLASMAS. See Fusion reactor instrumentation.

INSTRUMENTATION FOR POWER. See Power meters.

INSTRUMENTATION FOR RADIATION MONITORING. See Radiation monitoring.