Manipulate current signals with op amps

Op amps can condition current signals as well as they can handle voltages. In applications like 2-wire instrumentation, V/F converters, and photodetectors, current carries the signal, and voltage-mode amplifiers only introduce unnecessary errors. Using op amps in current mode will improve your circuits' performance.

Jerald Graeme, *Burr-Brown Corp*

Op amps are widely used to handle signal voltages, but they also can carry out analogous tasks on current signals. In the 2-wire instrumentation that you'll commonly find in remote monitoring, and for many photodetectors and V/F converters, current—not voltage—carries the signal. By using op amps in current mode, you can tailor these signals to your needs without having to convert between voltage and current. The performance of these current-mode circuits is comparable to that of their more common voltage-mode counterparts. (For a review of the basic principles of current-mode op-amp circuits, see **box,** "Review of current-mode op-amp configurations").

Current-mode circuits can condition signals for current-output devices like photodiodes and 2-wire transmitters. For example, suppose you want to interface a photodiode with a V/F converter, which offers an economical means of A/D conversion. In such a circuit,

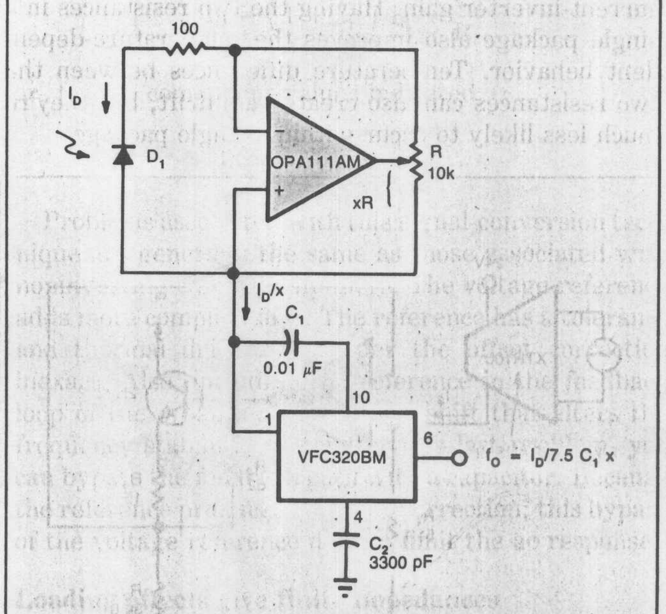

Fig 1—You can use a current amplifier to scale current for a charge-balance-type V/F-converter input. This approach lets you leave the signal in the current mode, thus avoiding the gain drift that a typical V/F converter's input resistor would produce.

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The performance of current-mode circuits is comparable to that of more common voltage-mode amplifiers.

gain thermal drift largely determines the conversion error. You could convert the current to voltage with an input resistor and then follow with the V/F converter, but many charge-balance-type V/F converters accept current inputs directly. These devices eliminate the input resistor, which causes much of the troublesome gain drift. When using a charge-balance-type V/F converter, you just need to scale the current from the photodiode to the input current range of the V/F converter.

You must, of course, scale that current in a manner that doesn't introduce additional gain-drift error. Otherwise, you'd lose the advantage of using a charge-balance-type V/F converter. One way to avoid introducing gain drift is to use a current-inverting op amp to scale the current. In **Fig 1,** a single potentiometer supplies two feedback resistances. Because the two resistances are part of the same potentiometer, their temperature coefficients match each other closely. The ratio of their resistances therefore shouldn't change with temperature, and only this ratio determines the current-inverter gain. Having the two resistances in a single package also improves the temperature-dependent behavior. Temperature differences between the two resistances can also create gain drift, but they're much less likely to occur within a single package.

Connecting **Fig 1**'s inverter, which is acting as a current-difference amplifier, to the diode source produces a simple scaling relation between the currents; the output current is I_D/x, where I_D is the current from the diode and x is the potentiometer's fraction in the positive feedback path. This design holds the voltage across the diode at nearly 0V—a condition that's required for maximum linear operation. Because the diode capacitance affects the feedback factor, the decoupling resistor added in series with the diode introduces a deviation from this zero bias.

When the scaled current is delivered to the V/F converter, the proportional output-signal frequency is

$$f_O = \frac{I_D}{7.5C_1x}.$$

The components in the circuit set the full-scale frequency to 10 kHz for a V/F-converter input current of 0.25 mA. The potentiometer setting determines the scaling to that full-scale current. For an accurate setting, your potentiometer should feature a high adjustment resolution.

In 2-wire-transmitter instrumentation, the most common output range runs from 4 to 20 mA, but you will also often encounter the 10- to 50-mA scale. Except

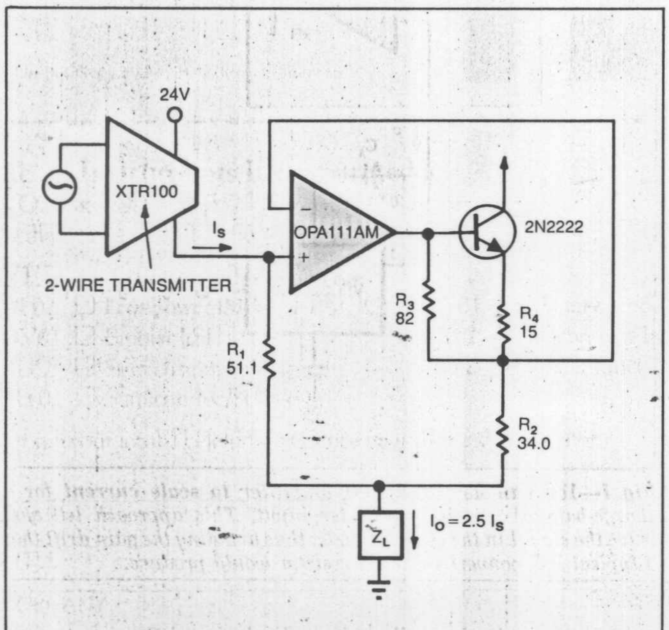

Fig 2—By boosting the op-amp output current, *a noninverting current amplifier can raise the 4- to 20-mA output of a 2-wire transmitter to the 10- to 50-mA scale.*

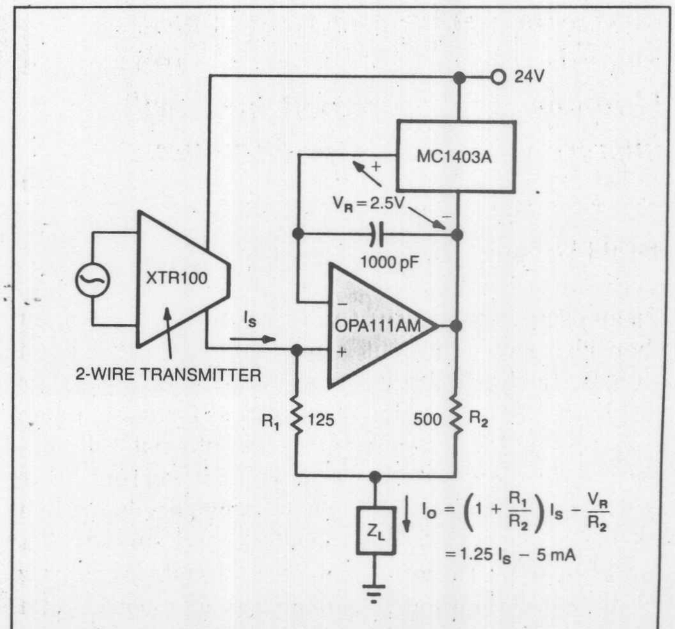

Fig 3—A voltage reference in a noninverting amplifier *circuit develops a cancellation current that removes the output offset of a 2-wire transmitter.*

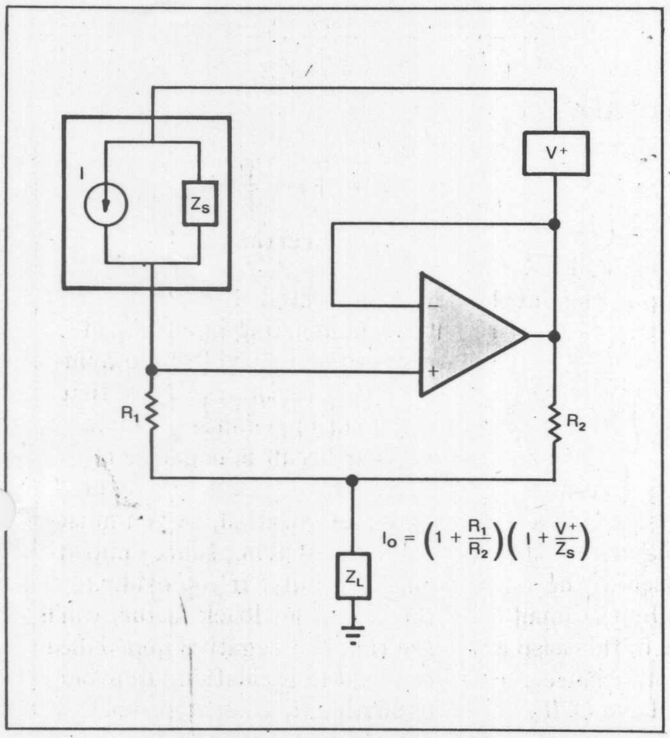

***Fig 4—Driving the source's power-supply common** from the op amp's output removes impedance loading errors from the noninverting current connection.*

for the high currents sometimes required, translations between the two ranges are excellent applications of the noninverting current-mode amplifier and its related attenuator. Op amps for the increased scale have to supply as much as 30 mA, and most op amps can't do this. However, those high currents don't have to be bipolar, so you can easily boost the output current to the level required.

As shown in **Fig 2,** you can boost the current and at the same time increase the op-amp current-limiting function to control the boosted current. This circuit transforms the 4- to 20-mA range into the 10- to 50-mA scale. An output emitter-follower boosts the current. Because the current available to R_3 from the op amp is limited, the voltage to the transistor base is also limited. That drive voltage can create so much demand for current in R_3 that the op amp's current limit is eventually reached, which in turn limits the transistor current. For the components shown, this limit is about 50 mA.

A second function that's useful when processing 2-wire transmitter signals is the removal of the 4- or 10-mA output offset. This offset supplies the quiescent current of transmitter circuitry, but you must remove the offset to extract the signal that's created in response to the input. To remove the offset, you can develop a counteracting offset in a current amplifier, which also rescales the remainder signal.

Fig 3 illustrates this approach. In this case, a voltage reference develops the offset cancellation. The output of the 2-wire transmitter supplies a noninverting current amplifier that converts the transmitter's output into a 0- to 20-mA signal. For this current amplifier, the inclusion of a constant voltage drop (V_R) modifies the usual voltage-follower feedback of the op amp. The voltage normally impressed on R_2 decreases by V_R, and the voltage reference develops a negative component of output current that cancels the 4-mA offset.

Just removing the offset would leave a somewhat cumbersome 0- to 16-mA output range, but you can adjust the ratio of the feedback resistors to convert the output range to 0 to 20 mA. From the combination of the offset cancellation and gain setting comes an output

$$I_O = \left(1 + \frac{R_1}{R_2}\right) I_S - \frac{V_R}{R_2}$$

or, for the component values indicated,

$$I_O = 1.25 I_S - 5 \text{ mA}.$$

Problems associated with this signal-conversion technique are generally the same as those associated with noninverting current amplifiers. The voltage reference adds more complications. The reference has a tolerance and thermal drift that render the offset correction inexact. Also, including the reference in the feedback loop of the op amp adds a phase shift that alters the frequency stability. To counter this last problem, you can bypass the feedback path with a capacitor. Because the reference produces only a dc correction, this bypass of the voltage reference doesn't limit the ac response.

Loading effects give finite impedances

Loading effect is one characteristic of current-mode circuits that doesn't have a parallel in voltage-mode circuits. For the ideal voltage amplifier, the input impedance is infinite and the output impedance is zero, which prevents loading errors at both ports. An ideal current amplifier would present zero impedance at its input and an infinite one at its output. For actual current-mode circuits, however, the impedances seen at the two ports scale with the load and source imped-

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Review of current-mode op amp configurations

The designs of current-mode op-amp circuits parallel those of voltage-mode inverting and noninverting amplifiers. **Fig A** shows two basic current-mode op-amp configurations. The load, Z_L, shown in these figures and the diagrams in the main text, could be a series of current-conditioning circuits. In these circuits, as in the op amps' voltage-mode counterparts, two resistors control the feedback around the op amp and set the gain.

The parallel with voltage-mode amps continues in the gain expressions: For the inverting devices the gain is the negative of the ratio of the two feedback resistors; for the noninverting devices, the gain is 1 plus the ratio. As determined by the two resistors, feedback forces these current-amplifier circuits to accept input currents and supply correspondingly scaled output currents.

In the inverting current amplifier **(Ref 1)** of **Fig Aa**, the input signal (I_I) drives the inverting op amp's input, causing the amplifier's output to swing and conduct I_I through R_1. The negative feedback that occurs creates a controlled voltage (V_1) across R_1 and impresses this voltage on resistor R_2 as well. Thus, the current established in R_2 is a scaled replica of the input-signal current. As delivered to the load (Z_L), however, I_O's polarity is the opposite of I_I's because

$$I_O = -\frac{V_1}{R_2},$$

which has a negative, or inverted, relationship to the developed V_1. The result is

$$I_O = -\left(\frac{R_1}{R_2}\right) I_I.$$

The noninverting current amplifier **(Fig Ab)** operates in a similar fashion. The voltage developed on R_2 replicates the voltage developed by the input current on R_1, but in this case a voltage follower is the cause. The noninverting drive of R_2 adds a proportional current to I_I and delivers the sum of the two currents to the load:

$$I_O = \left(1 + \frac{R_1}{R_2}\right) I_I.$$

In addition to gain equations, key performance characteristics of these circuits include impedance transformations, dc errors, and bandwidths. The circuits' impedance transformations are the same as those of a common-emitter transistor. The reflected input impedances are the load impedances multiplied by the current gain, and similar reflected output impedances are those of the sources divided by that gain. At the inputs, impedances are

$$Z_I = -\left(\frac{R_1}{R_2}\right) Z_L \text{ (inverting)}$$

$$Z_I = \left(1 + \frac{R_1}{R_2}\right) (R_1 + Z_L)$$

(noninverting).

As indicated, R_1 also affects the noninverting input impedance, so you should try to minimize that resistance. Note that the input impedance of the inverting circuit is negative because of the gain inversion of that configuration, so you must check the driving source impedance carefully. If you evaluate the circuit feedback factor, you'll see that the negative impedance can lead to oscillation. To avoid oscillation, the net impedance in the input circuit should remain positive; therefore,

$$Z_S > \left(\frac{R_1}{R_2}\right) Z_L,$$

where Z_S is the source impedance of the input current source. The negative input impedance results from the inherent positive feedback of the circuit, which is required for the gain inversion.

This negative input impedance causes a problem for remote monitoring. In remote monitoring systems, operation in the current mode reduces noise pickup on long lines. However, these long lines also develop significant stray capacitances that shunt the source to lower its net received impedance. To counteract this potential stability compromise, you may have to add a decoupling resistor in series

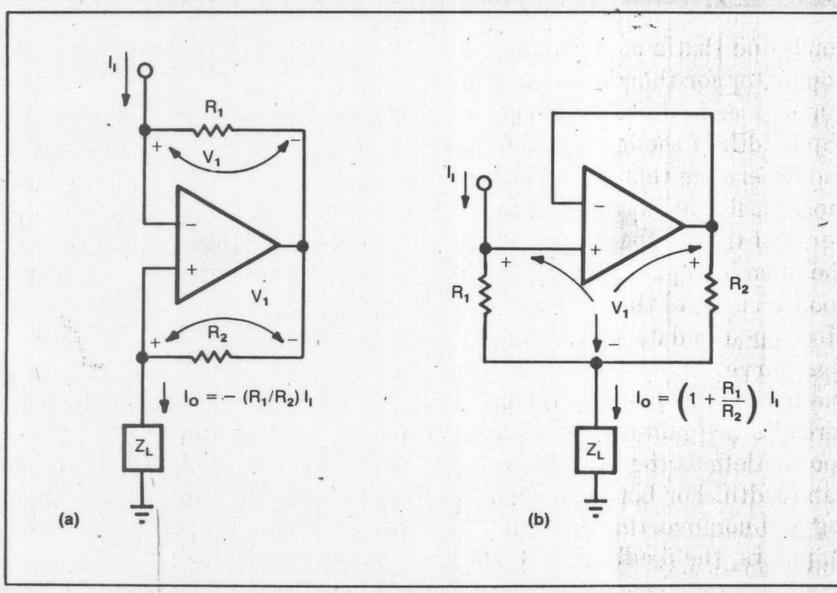

***Fig A—These op-amp circuits amplify current** instead of voltage. However, both the inverting (a) and the noninverting (b) current-mode configurations parallel similar voltage-amplifier designs.*

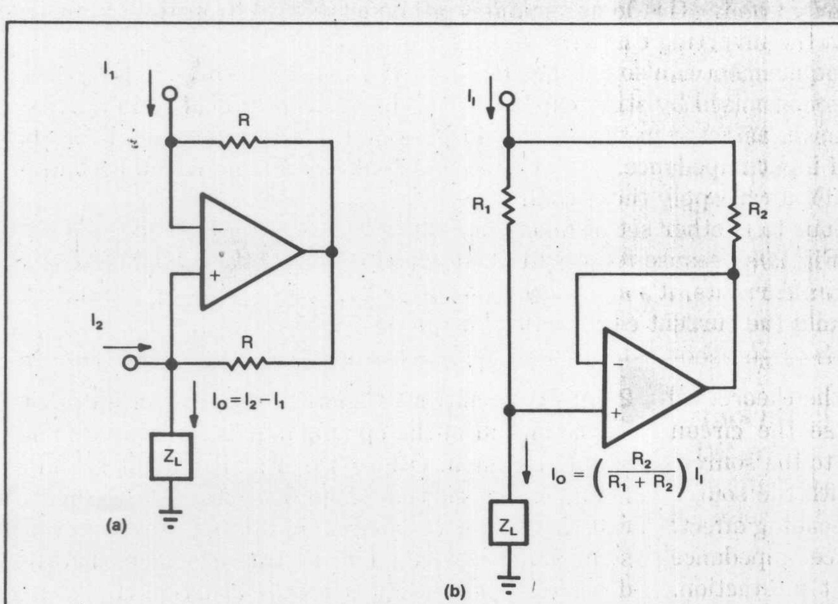

***Fig B—Shifting connections in Fig A's circuits** produce current amplifiers that are equivalent to difference amplifiers (a) and that can attenuate current (b).*

with the circuit input, or you may have to effect a capacitive bypass of the load.

Analogous to emitter followers

Continuing with the emitter-follower analogy, the impedance that you see at the output of these circuits is that of the source divided by the current gain, or

$$Z_O = -\left(\frac{R_2}{R_1}\right) Z_S \text{ (inverting)}$$

$$Z_O = \left(\frac{R_1 + R_2}{R_1}\right) Z_S$$

(noninverting).

Limitations of the op amp's input impedance, open-loop gain, and common-mode rejection can alter these impedances and the reflected input impedances. Except when approaching the frequency limit of the op amp, the stated equations give adequate approximations for these parameters.

The input offset voltage and input bias currents of the op amp cause the dc errors of these current amplifiers. These dc errors appear as output error currents that equal

$$I_{\epsilon O} = \left(\frac{V_{OS}}{R_2} + \left(\frac{R_1}{R_2}\right) I_{B-}\right) - I_{B+}$$

(inverting)

$$I_{\epsilon O} = \frac{V_{OS}}{R_2} - \left(1 + \frac{R_1}{R_2}\right) I_{B+}$$

(noninverting).

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These output offset currents are related to the op amp's input offset voltage when that voltage is divided by a gain-setting resistor and the circuit current gain and multiplied by at least one of the input bias currents.

In addition, when selecting feedback resistors that minimize circuit oscillation, you should also consider the error created by the input offset voltage. Only the ratio of the feedback resistors sets the current gain, so you still have freedom in choosing the absolute resistances. However, the lower the value of R_2, the greater the output error current from V_{OS}. Yet if you make that resistor too large, it will drop too much voltage in the circuit. For noninverting circuits, that drop is in series with the load voltage, while for the inverting case, the drop forces a voltage swing at the op amp's output.

Determination of ac errors caused by the op amp is less direct but follows that of conventional analysis. You won't encounter the higher-frequency errors associated with gain, input impedance, output impedance, and common-mode rejection at the low frequencies at which most current-loop instruments operate. When you do see these high-frequency errors, you'll find that a common denominator for their frequency of importance is the closed-loop bandwidth of the circuit. You can determine that bandwidth most easily by plotting the inverse of the feedback factor on the open-loop gain-magnitude response curve of the op amp. Most op-amp data sheets supply that curve. The point at which the inverse feedback factor intercepts the gain-magnitude response defines the closed-loop bandwidth. For both the inverting and noninverting current amplifers, the feedback factor is

$$\beta \cong -\frac{R_2}{R_2 + Z_L} \text{ for } Z \gg R_1.$$

In the inverting case, the $Z_S \gg R_1$ requirement can sometimes be compromised by stray capacitance, as noted in the discussion of input impedance.

You can apply these configurations to another set of functions **(Fig B)**. Because it's easy to sum currents, it's also easy to build the current equivalent of a voltage-difference amplifier. You supply one current directly to the load but invert the other, so that

$$I_O = (I_2 - I_1).$$

Errors and application considerations of this connection remain basically the same as those for the inverting current amplifier. What's more, the noninverting current amplifier offers a feature not found in the voltage-mode version: noninverting gains of less than unity. Because the current-mode version is actually a bilateral circuit, input and output can reverse roles, as is the case in **Fig Bb**'s circuit. In that case, the circuit operates as a current attenuator with the following output:

$$I_O = \left(\frac{R_2}{R_1 + R_2}\right) I_I.$$

The input/output switch produces another change in operation because it includes R_1 in the feedback factor, which becomes

$$\beta \cong -\left(\frac{R_1 + R_2}{R_1 + R_2 + Z_L}\right)$$

$$\text{for } Z_S \gg R_1 \text{ or } R_2.$$

ances, so these impedances don't follow the theoretical ideal. The impedances are finite because the circuit transfers the load voltage swing directly to the source. In the source, the load voltage reacts with the source impedance to create a current change or loading effect.

If you could make the effective source impedance very large, you could remove that input interaction. What's more, the reflected output impedance would remain very high. In the noninverting case, you can implement this increase in source impedance by simply bootstrapping the source's power supply, as shown in **Fig 4.** The current source represented there could be a 2-wire transmitter or another current-output device.

The output of the op amp drives the source's power-supply common. Otherwise, **Fig 4**'s circuit is simply a noninverting current amplifier with the voltage held constant across the source. Z_S doesn't have load voltage swings impressed on it in this case, so the circuit doesn't experience any associated current changes. The circuit therefore produces a very high effective source impedance. A factor approximating the parallel combination of the op amp's open-loop gain and common-mode-rejection ratio boosts the impedance. At higher frequencies, of course, both the open-loop gain and the

You must scale current in a way that doesn't introduce additional gain-drift error, or you lose the benefits of using a charge-balance-type V/F converter.

common-mode-rejection ratio decrease, but you'll see a dramatic improvement in the lower ranges.

You can find a lesser but similar improvement in an analogous circuit incorporating an inverting current amplifier, where the swing on feedback resistor R_1 remains across the source. In this case, the improvement factor is Z_L/R_2.

Nonlinear circuit-mode functions

You can implement several nonlinear circuit functions in the current mode, including those of the peak detector, limiter, and precision rectifier. In the current-loop instrumentation of process control, where the key indicators are peak excursions, you'll frequently need a peak detector. **Fig 5** shows a current-operated implementation of a peak detector. The circuit consists of an input gating amplifier (IC_1), a holding capacitor, and an output buffer amplifier (IC_2). Around this combination, you can see a feedback loop like that of an inverting current amplifier. As with the inverting amplifier, R_1 and R_2 provide the feedback. A FET switch produces the reset.

The input-current polarity and magnitude determine circuit operation. That current must reach a negative level sufficient to override R_1's feedback current before the capacitor voltage can change. Otherwise, the capacitor maintains a constant voltage on the combination of R_2 and the load, and the capacitor supplies a constant output current. That current is

$$I_O = -\left(\frac{R_1}{R_2}\right) I_{I(MAX)},$$

where $I_{I(MAX)}$ is the last recorded peak.

To change the output current, the input current must overcome the condition established by the most negative previous input level. Until the input current does this, R_1 supplies a current to the input circuit that forward-biases D_1. This current drives the output of IC_1 below the voltage of the capacitor that is reverse-biasing D_2.

When I_I does reach a sufficiently negative level, it reverse-biases D_1 and drives the output of IC_1 positive, thereby forward-biasing D_2. The feedback loop now goes through IC_1, forcing IC_2 to conduct the input current through R_1. In this tracking state, the capacitor voltage increases, which in turn increases the output current. As the magnitude of I_I drops below this last level, the new capacitor voltage lets the feedback current of R_1 again forward-bias D_1 and reverse-bias

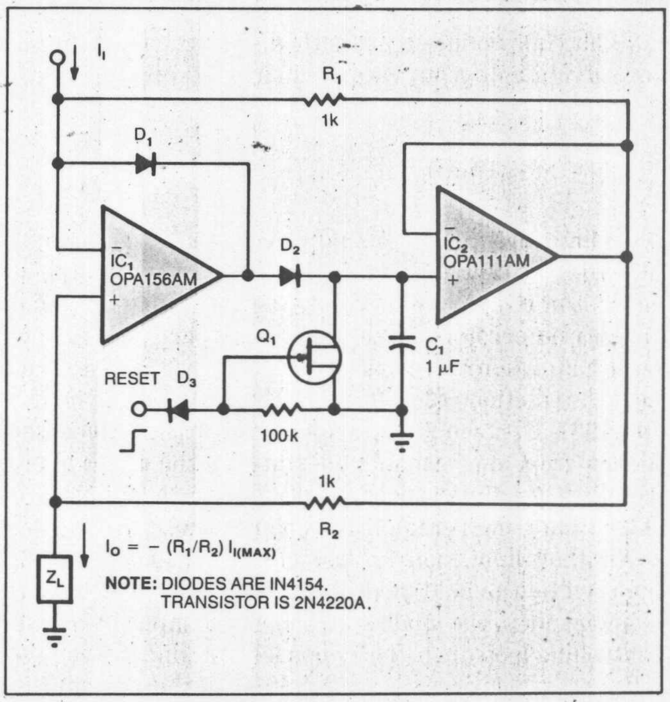

Fig 5—By having the input current open and close the feedback loop that determines the voltage on the holding capacitor, this circuit can detect and retain the peak current.

D_2, thereby resuming the holding state.

To begin a new measurement interval, a positive transition at the FET gate resets the capacitor. Normally, you would want to hold the gate negative to keep the FET switch off. At the beginning of a reset cycle, the current conducted by the FET discharges the capacitor. This discharge forward-biases D_2 and reduces the current through R_1 to the point where D_1 reverse-biases. The tracking mode then resumes, and the op amp overrides the discharging of the FET. Thus, reset never continues below the level corresponding to the input that's present.

Errors affect the peak detector

DC errors, diode switching times, and output impedance affect the accuracy of this peak detector. Amplifier IC_1 injects an input offset current of

$$I_{\epsilon I} = I_{B-}$$

and an output offset of

$$I_{\epsilon O} = \frac{V_{OS}}{R_2} - I_{B-}.$$

One function that's useful when processing 2-wire transmitter signals is the removal of the 4- to 10-mA output offset.

The input bias current, IC_2, and the FET gate leakage current (I_G) produce a parasitic discharge of the capacitor, giving an output rate of change, or droop (D), of

$$D = \frac{I_G + I_{B+}}{C}.$$

To limit droop, you should choose for IC_1 an op amp like the OPA111AM, which minimizes the input bias current in IC_2. The input offset voltage of that amplifier creates no error because it becomes part of the feedback loop when the capacitor is charging. Speed is the main attraction of IC_1, which in this case is the OPA156A. This op amp's ideal operation drives the diodes from one conduction state to the other more rapidly. The finite speed of the OPA156A determines the circuit's high-frequency performance.

Another limitation on the circuit's performance is imposed by the hold mode's open feedback loop. In this feedback loop, the diode states keep the input-to-output controlling loop open. This open loop produces a voltage follower, which transfers the voltage of the capacitor to R_2 and the load. If other factors don't introduce load-voltage variations, the load receives the intended constant current from the follower. If some other influence intervenes, however, the current delivered will be in error. In other words, the circuit has a low output impedance when operating in the hold mode and may require bootstrapping.

If you measure the load voltage with another voltage follower, you can transfer that voltage and drive the normally grounded terminal of the capacitor. This technique eliminates the influence of load voltage on output current.

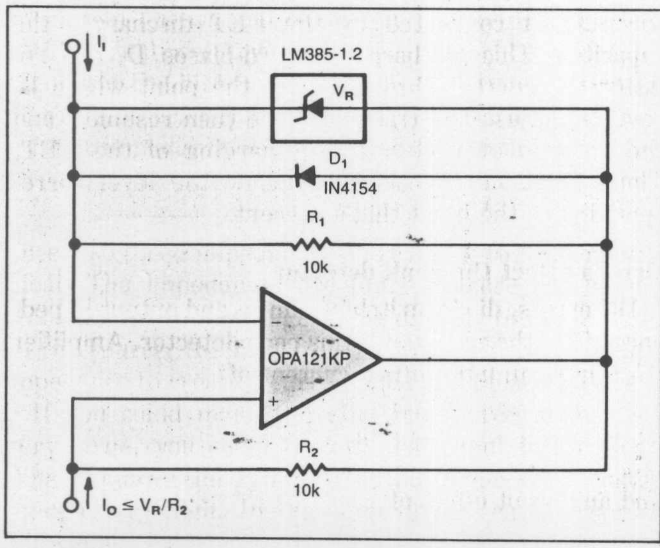

Fig 6—A voltage reference in the feedback of this current inverter can clamp the output current to a level set by the voltage reference.

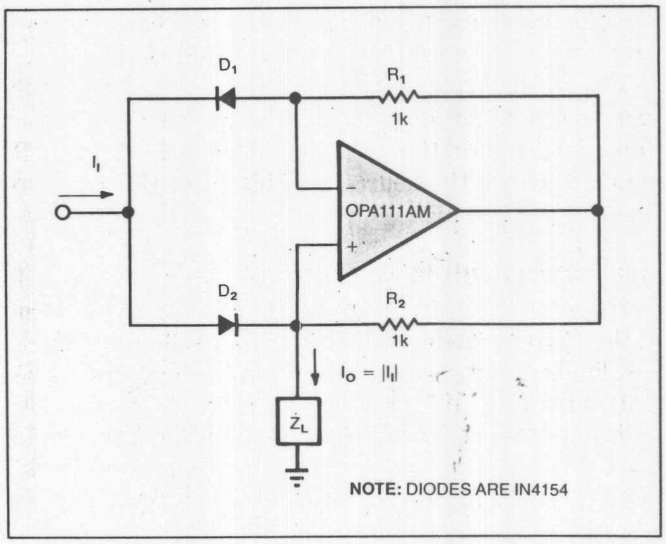

Fig 7—The ease with which currents switch diodes simplifies absolute-value conversion in the current mode.

Current-mode op amps can clamp currents

A second nonlinear application for current-mode op amps is in limiters, or clamps. By restricting the feedback voltage of a current inverter, a clamp imposes an upper limit on the output current. In **Fig 6**'s circuit, positive input currents develop related voltages on R_1, up to the operating voltage (V_R) of the voltage reference. At V_R, the reference turns on and conducts any further increase in current, thus shunting it away from R_1. Because the op amp transfers the voltage on R_1 directly to R_2, the output current delivered by the latter is limited to

$$I_O \le \frac{V_R}{R_2}.$$

For the elements shown, this limit is 120 μA.

Negative input currents will also result in a limited output because negative inputs turn on D_1, whose purpose is to protect the voltage reference against excessive forward current, thereby reducing the current through R_1.

The behavior of a real clamp only approximates the sharpness, frequency response, and noise level of an ideal clamp. The minimum quiescent current needed to

Loading effects are one characteristic of current-mode circuits that don't have a parallel in voltage-mode designs.

bias internal circuitry limits the turn-on sharpness of the reference. For the reference shown, that current must reach 8 μA before you can establish a stable limit voltage.

The reference circuitry also has a bandwidth compromise imposed by its amplifier stages, so its series impedance increases with frequency. At 1 kHz, the impedance can approach 100Ω. The impedance will therefore increase the clamp level unless R_1 is large enough to make this impedance negligible. Because the preferred voltage references are of the band-gap variety, noise is unavoidable. The elements shown create an output noise of approximately 60 nA rms.

Op amps determine absolute values

A third nonlinear function easily implemented in the current mode is the absolute-value operation, carried out by a variety of voltage-operated circuits (see **Refs 2** through **4**). In **Fig 7**'s circuit, the diodes create a gate that directs an input current to either the input of a current inverter or straight to the output load. A positive input current forward-biases D_2 and passes directly to the output load. D_1 is reverse-biased by this current polarity, and the diode disconnects the inverting input of the current amplifer from the signal path.

A negative input current reverse-biases D_2, blocking that direct path, but forward-biases D_1. Therefore, the current ends up in the current inverter's input, where its polarity reverses before it arrives at the load. Thus, both polarities of input current result in a positive output current with unity gain magnitude for the current inverter, or

$$I_O = |I_I|.$$

Error in this gain magnitude causes most of the inaccuracy in the current-mode absolute-value function. The error results primarily from inequalities between the two circuit resistors and from their temperature-dependent tracking. If the two resistors differ, then the gain magnitude received by the negative currents won't be the same as the unity transfer of the positive currents. DC errors from the op amp produce additional errors. As in the case of the general inverting current amplifier, these errors generate an offset current. That error current is

$$I_{eO} = \left(\frac{V_{OS}}{R} + I_{B-}\right) - I_{B+}.$$

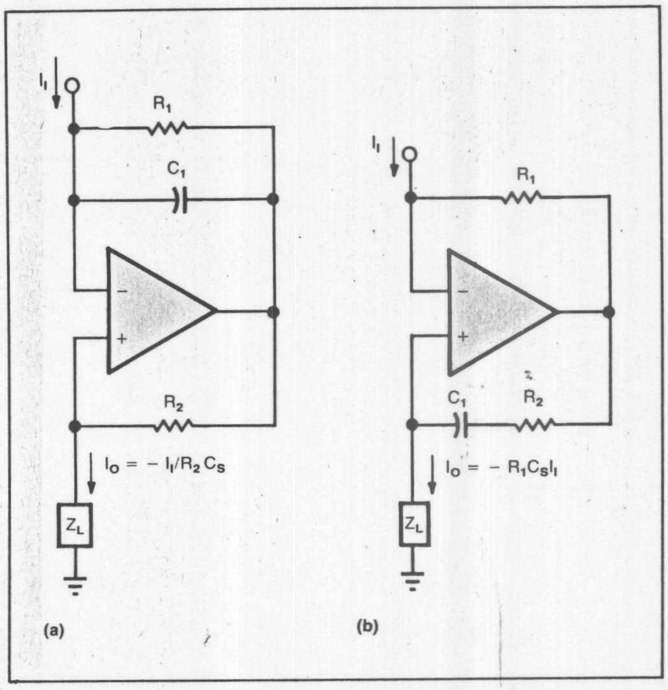

Fig 8—These designs translate integration (a) and differentiation (b) into current-mode functions from their voltage-mode counterparts.

Because diode switching doesn't occur within an op amp's feedback loop in **Fig 7**, the op amp's higher-frequency error figure marks an improvement over that of voltage-mode amplifiers. When driving feedback diodes **(Ref 2)**, high-gain op amps require significant slewing and settling times, but no diodes are operating in feedback loops in **Fig 7**. Instead, the driving capabilities of the signal's current source determine the diode switching time. Up to the much higher frequency limit imposed by the op amp's bandwidth, the absolute-value conversion will be fast if the current source is fast.

You can also adapt integrators and differentiators to the current mode. The configurations shown in **Fig 8** operate much as their voltage-mode counterparts do; they even retain two of the same fundamental limitations. In the integrator of **Fig 8a**, the capacitor accumulates charge from the flow of an input current and develops a voltage equal to the time integral of that current. The op amp impresses that voltage on resistor R_2. The voltage converts into an integration-related output current of

$$I_O \cong -\frac{I_I}{R_2C_S} \text{ for } f \gg \frac{1}{2\pi R_1C}.$$

You can implement several nonlinear circuit functions in the current mode, including those of the peak detector, limiter, and precision rectifier.

As in the voltage-mode version of this circuit, you must reset the capacitor **(Ref 2)**, which even charges without application of an input signal because of the op amp's input bias current. That parasitic charging produces an output dc error current of

$$I_{\epsilon O} = \left(\frac{V_{OS}}{R_2} + \frac{I_{B-}}{R_2 C_S}\right) - I_{B+}.$$

The second term of the error expression, I_{B-}, is the time integral of a constant. Consequently, with no input signal present, the output increases continuously until it reaches a final value of $R_1 I_{B-}$.

Resistor R_1 also restricts the integrator. You need this stop resistor to keep the negative-feedback factor greater than the positive-feedback factor. To do so, you must choose R_1 so that

$$Z_1 < \left(\frac{Z_S}{Z_L}\right) R_2, \text{ where } Z_1 = \frac{R_1}{1 + R_1 C_S}.$$

In the differentiator of **Fig 8b**, an input current creates a voltage on R_1 that will also appear on C once the capacitor charges. Again, the capacitor needs a reset. Another limitation found on both the voltage- and current-mode versions of this circuit is that, while charging, the capacitor conducts a current that approximates the time derivative of the voltage created on R_1. R_2 causes this approximation. You need a stop resistor to avoid oscillation at high frequencies **(Ref 3)**, because the positive-feedback factor would otherwise approach unity. You can prevent this oscillation when

$$Z_2 > \left(\frac{Z_L}{Z_S}\right) R_1, \text{ where } Z_2 = R_2 + \frac{1}{C_S}.$$

Within the bounds of that approximation, the capacitor-charging current delivered to the load is determined using the following expression:

$$I_O \cong -R_1 C_S I_I \text{ for } f \ll \frac{1}{2\pi R_2 C}.$$

The ac-coupling effect of the capacitor reduces the dc errors in this differentiator. The capacitor decouples the op-amp input's offset voltage and the current drawn by the op amp's inverting input, so that $I_{\epsilon O} = I_{B+}$.

Circuit gain determines the impedances reflected at the inputs and outputs of these integrators and differentiators in the same manner as for the inverting amplifier. At the inputs, the impedances are those of the loads, multiplied by the current gain, resulting in

$$Z_I = -\frac{Z_L}{R_2 C_S} \text{ (integrator)}$$

$$Z_I = -R_1 C_S Z_L \text{ (differentiator).}$$

The source impedances, divided by the circuit gains, are at the output:

$$Z_O = -R_2 C_S Z_S \text{ (integrator)}$$

$$Z_O = -\frac{Z_S}{R_1 C_S} \text{ (differentiator).}$$

The reactive nature of these impedances suggests that you must take care to maintain frequency stability. You can achieve this stability primarily by selecting the appropriate stop resistors. Choosing the correct resistors produces a negative net feedback factor, which maintains stability. For the best frequency stability, you should also decouple sources resistively and bypass loads capacitively. **EDN**

References

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Author's biography

Jerald Graeme manages the instrumentation components design department at Burr-Brown Corp (Tucson, AZ), where he directs a linear integrated circuit development group. He received his MSEE from Stanford University and his BSEE from the University of Arizona. He enjoys photography, scuba diving, and woodworking.