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Gas Chromatograph Uses Varactor Bridge Flame Detector Amplifier for Enhanced Performance

By **Harry Gill**, *Perkin-Elmer Corp.*

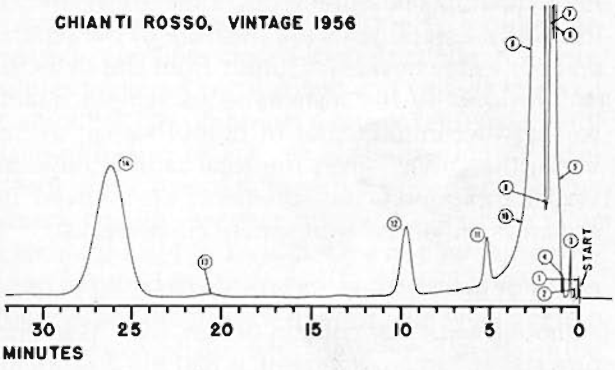
Current amplifier circuit based on parametric (varactor bridge) op amp increases gas chromatograph's useful sensitivity, stability, and dynamic range. New design furnishes 5×10^{-12} amp full scale output for recorder and integrator drive.

Electronic engineers have no monopoly on sophisticated instruments for measuring the different components of an unknown input. Gas chromatographs, which have evolved as rapidly since 1950 as microwave spectrum analyzers, enable research chemists to separate tenths of microlitres of any vaporizable sample into its individual constituents, and to measure the quantity of each sample-constituent with better than 2% accuracy.

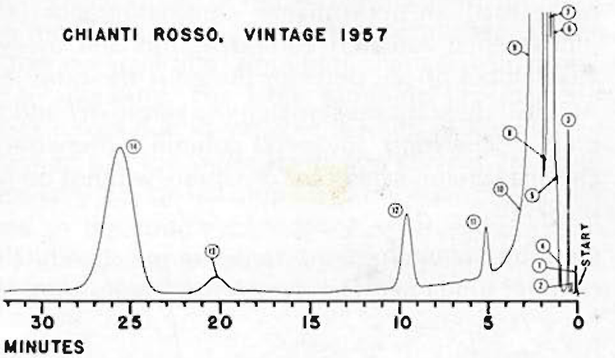
Not all gas chromatographs find their way into research laboratories or advanced chemical plants, as motorists convicted of dangerous driving can frequently testify. Many police departments use these instruments for qualitative measurements of blood-borne alcohol. In other novel applications, the gas chromatograph can distinguish vintage wine from last years' crop; detect from a sample of "minced earthworms" that pesticides wash into the soil and stay there (vide, Rachel Carson's SILENT SPRING); sniff the noxious fumes in an automobile's exhaust; or bolster an Englishman's conviction that teabags prevent Americans from ever tasting a civilized cup o' tea.

Chianti, vintage 1957, is introduced into the gas chromatograph. Comparison of differences in trace components on the chromatograms may be related to differences in relative quality of individual vintages.

CHIANTI ROSSO, VINTAGE 1956



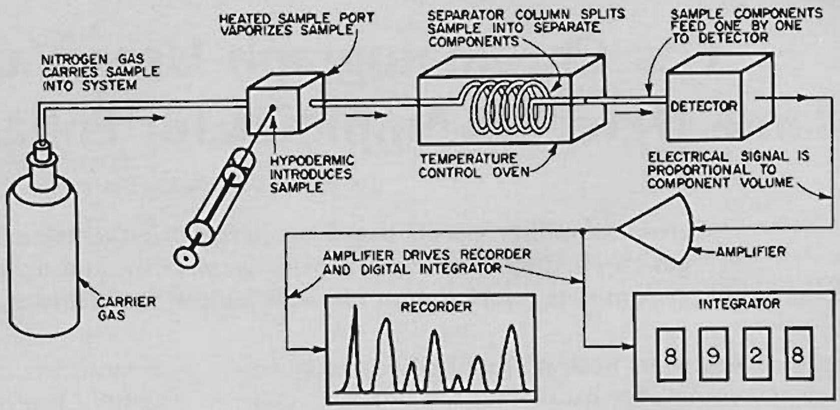
CHIANTI ROSSO, VINTAGE 1957



A gas chromatograph's resolution depends primarily upon the quality of the separator column used to segregate the sample into its individual components. Many different types of separator column can be "plugged in" to commercial chromatographs for a wide choice of analytical characteristics. Some of the more exotic separator columns are 300 foot coils, but more conventional types are wound into helixes that fit conveniently into the temperature controlled oven of most commercial instruments.

Figure 1

Samples introduced by hypodermic needle at heated sample port are vaporized then carried into the separator column by inert "carrier" gas. Adsorption or related process selectively delays different sample components, feeds them one-by-one to the detector. Gas Chromatograph technique can measure sample components down to 10^{-12} grams.



A simplified gas chromatograph block diagram is given in Fig. 1. Samples are introduced by hypodermic needle at the heated input port, where they are instantly vaporized, thence carried into the separator column by a stream of nitrogen, helium or other inert "carrier" gas. By selectively delaying the different sample components, the separator column feeds them one-by-one to the detector for component quantity measurements. Time delay between individual components is a measure of the separator column's effectiveness. Output from the detector is proportional to the instantaneous sample quantity, so that the time-integral of output signal, or "area under the curve," gives the total sample flow. Individual component measurements are derived from the areas under the appropriate curve-peaks.

DETECTOR

Although separator column design, with its temperature controlled environment, is half the development battle for high performance chromatographs (early fundamental research earned Martin and Synge a 1952 Nobel prize), detector design is the other half. An ideal detector must match the sensitivity and stability of the most advanced column, otherwise the chromatograph's inherent capability will not be fully realized.

Gas chromatographs may use thermal conductivity or other fundamental parameters for measuring sample concentration. However, the flame ionization detector generally gives highest sensitivity, detecting the presence of components weighing less than 10^{-12} gram.

Although gas chromatograph detector considerations might seem a somewhat specialized topic for this magazine's broad readership, in reality the problem of stable, noise-free, picoampere measurement recurs in many branches of industry and research.

DETECTOR PRINCIPLES

A hydrogen flame within the detector assembly, Fig. 2, ionizes sample components emerging from the separator column, and develops a proportional

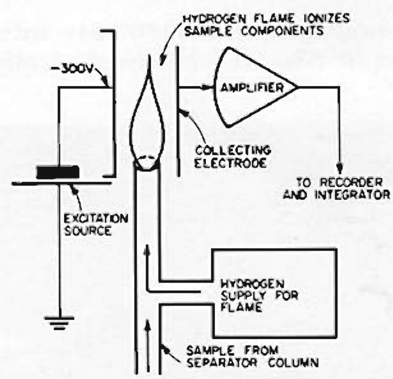


Figure 2

Flame ionization detector ionizes sample components with hydrogen flame, develops output current proportional to sample flow. Amplifier must resolve down to 2 x 10^{-14} ampere flame noise level, give accurate measurement for picoampere signals.

current flow between excitation electrode and collector plate. A high gain amplifier completes the external circuit and provides the drive signal necessary for recorder and integrator operation.

At highest sensitivity levels, detector resolution is limited by flame noise to about 5 x 10^{-14} amperes; consequently, there is little advantage in using amplifier systems with higher sensitivity or lower noise. On the other hand, amplifiers unable to resolve down to this level because of poor signal-to-noise ratio would be unable to respond to the full range of detector output, hence would cramp the chromatograph's wide output range.

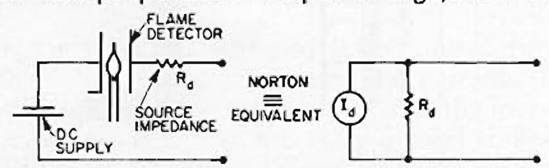
Past chromatography instruments have used electrometer tubes and post amplifiers to measure flame detector current. Warmup delays, aging, shift in operating point, noise, and humidity effects are some of the problems associated with such amplifiers. In particular, noise is the limiting factor with electrometer tubes, because it restricts dynamic range, and forms the weak link in the detector-amplifier-recorder-integrator chain. Consequently, we set out to design a new amplifier system that eliminated the electrometer tube, and which would raise overall chromatograph performance in the process.

A commercial varactor bridge operational amplifier solved the problem. This unit provided picoampere sensitivity, very low noise levels, and gave wider dynamic range than we could use. Additional virtues were its solid state 3 cu. inch P.C. mounting construction, high open-loop gain, high input impedance, modest warmup time, and ample output for direct recorder and integrator drive.

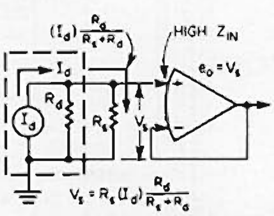
Key specifications for the selected varactor bridge amplifier are 2 pA max bias current and 0.1 pA/°C bias drift at 25°C; 0.01 pA noise from DC to 1 Hz; 10^{10} ohms and 10^{12} ohms differential and common mode input impedance; 10^6 gain; ±10 volt, 20mA output rating; and potential 10^9:1 dynamic range.

CHOICE OF CIRCUITS

Probably the most obvious method for measuring detector response is to convert the output current signal to a proportional output voltage, then use a high gain voltage amplifier to bring signals up to the right level. A circuit such as Fig. 3A uses a sampling resistor R_s to develop voltage V_s proportional to detector current; amplifier A then buffers this voltage to drive the recorder and integrator. A noninverting amplifier configuration has highest input impedance, hence places maximum input impedance in parallel with R_s and minimizes loading errors.

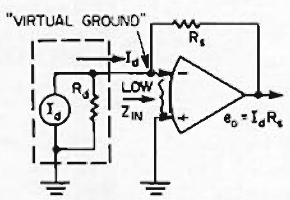


Detector (left) can be characterized by a Norton equivalent current source, I_d, in parallel with source impedance R_d (right).



Amplifier input impedance must be very high compared with sampling resistor to minimize loading effect, also to reduce errors caused by temperature-induced input impedance variations.

Fig. 3A



Source impedance, R_d, is paralleled by amplifier's very low closed-loop input impedance. Consequently, R_d does not enter into the output equation.

Fig. 3B

Figure 3

Noninverting voltage amplifier (left), uses sampling resistor R_s to develop signal voltages in response to detector output current. Alternative circuit (right), uses no sampling resistor, eliminates common mode and impedance errors, converts detector output directly to voltage with superior stability and noise performance.

However, the most obvious amplification techniques are not necessarily the best! A current-to-voltage converter, or current amplifier, Fig3B, provides many advantages compared with the conventional voltage amplifier, even though both circuits can be based on the same operational amplifier. Instead of using sampling resistor R_s to convert detector current to equivalent voltage signals, the current amplifier can convert and amplify in one simple arrangement.

The primary advantage to the current amplifier is that feedback holds the amplifier input voltage at "virtual ground" so that input impedance is very low. Actually the circuits of Fig. 3A and 3B can be viewed as current meters. An ideal current meter would have zero impedance so that the measuring instrument would have no effect on the circuit being measured. While the circuit of Fig. 3B approaches this ideal, the circuit of Fig. 3A introduces a large impedance, R_s, into the circuit being measured, hence introduces measuring errors that depend upon the relative magnitudes of the sampling resistor, R_s, and source resistance R_d.

Any attempt to reduce measuring errors by reducing the magnitude of sampling resistor R_s requires an increase in amplifier closed-loop gain, which in turn amplifies the input voltage drift and voltage noise of the amplifier. By contrast, voltage-noise-gain and voltage-drift-gain of Fig. 3B is always unity*, independent of the size of R_s (which is now placed in the feedback circuit). Another disadvantage to the sampling technique of Fig. 3A is that the noninverting connection required to obtain very high input impedance introduces another source of error, namely, common mode rejection errors, which are not present in the single-ended circuit of Fig. 3B. Furthermore, the common mode impedance (which sets input impedance for the noninverting amplifier) changes with ambient temperature. For large sampling resistors this causes errors due to variable loading of the voltage amplifier. On the other hand, the current amplifier offers very low closed loop input impedance, as compared to the source impedance, so that the amplifier's impedance variation for this configuration has negligible effect on total accuracy.

$$*\left[\frac{(R_s + R_d)}{R_d} \doteq 1\right]$$

ACTUAL CIRCUIT

A simplified flame detector circuit, based on Analog Devices' Model 301 Varactor Bridge (parametric) amplifier* connected as a current-to-voltage converter, is shown in Fig. 4. The circuit's overall specifications include 5×10^{-12} amperes full-scale reading on the most sensitive range, 2.5×10^{-14} amperes maximum equivalent input noise, 5×10^{-14} amperes long-term drift, 7.5×10^{-15} amperes/°C drift over the range 15°C to 32°C, and 5×10^{-14} amperes offset change for line voltage variation from 105 to 130 volts.

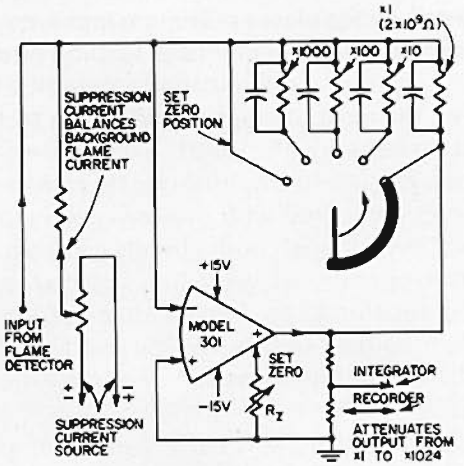


Figure 4

Current-to-voltage converter based on varactor bridge (parametric) op amp uses a 2×10^9 ohm feedback resistor for 5×10^{-12} amperes full-scale reading. Dynamic range is 10^5:1 on most sensitive scale; 10^6:1 on all others.

Four levels of current sensitivity, all developing up to 10 volts at the amplifier output, are selected by range switch S_1. The switch connects additional resistors in parallel with the 2,000 megohm highest-sensitivity feedback resistor to reduce net feedback resistance, hence sensitivity, in steps of ten. Output voltage, V_o, for a given input current, I_{in}, is simply related to feedback resistance, R_f, by $V_o = I_{in} \times R_f$. This gives a 2 megohm total feedback resistance for the least sensitive range. Parallel capacitors used with each feedback resistor ensure amplifier closed loop stability and give correct operating bandwidth for each sensitivity range. Amplifier output to the recorder can be attenuated by x1 to x1024 in multiples of 2.

The most sensitive scale yields a potential 100,000:1 dynamic signal range through its ability to handle input signal swings from 5×10^{-14} amperes (twice the noise level) to 5×10^{-9} amperes at maximum output attenuation. The higher input level develops an output voltage $I_{in} \times R_f$ equal to $(5 \times 10^{-9}) \times (2 \times 10^9) = 10$ volts, which is the maximum Model 301 output before saturation.

See article, page 6, L. R. Smith, "A Parametric Operational Amplifier," NEREM, Boston, 1966.

Key to the detector amplifier's temperature stability of 7.5×10^{-15} amperes/°C drift for normal laboratory environments is the temperature-compensated amplifier enclosure. The heated enclosure reduces ambient temperature variations by a factor of approximately 60, prevents humidity from affecting high value resistors and associated insulation, and stabilizes the varactor bridge amplifier.

The varactor bridge Model 301 amplifier draws a maximum bias current of 2 pA at 25°C, which doubles for every 10°C temperature rise. Since the bias current is in effect drawn through the feedback resistor R_f, variations in bias current (i.e., bias current drift) develop spurious amplifier outputs. Although initial offsets can be externally zero'd by adjusting R_z, bias-current increase from 2 pA to 4 pA for a 25°C to 35°C rise creates an "unzero'd" offset error of $(2 \times 10^{-12}) \times (2 \times 10^9) = 4$ millivolts. This is equal to 40% of the sensitive range's 5×10^{-12}A full scale recorder output. However, the amplifier enclosure effectively insulates against ambient variations, so that drift is less than 1.5% of full scale for the same 10°C temperature rise.

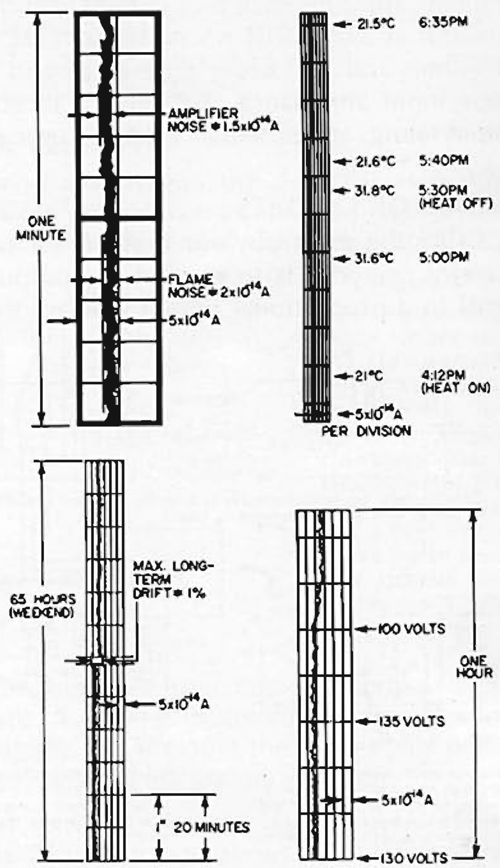


Figure 5

Actual measurements show how chromotograph amplifier easily meets specifications. Noise (top left) of 1.5×10^{-14} amps is well below flame noise level of roughly 2×10^{-14} amps; ambient temperature variation from 21°C to 31.8°C produces about 5×10^{-14} amp offset (top right). Long term stability test (bottom left), produces less than 1% drift during 65-hour weekend, while effect of supply voltage variation from 130 to 110 volts is approximately 5×10^{-14} amps offset (bottom right).

An auxiliary reference supply furnishes suppression currents needed to balance the flame detector's background current, and is adjustable from 1×10^{-14} to 1×10^{-8} amperes. Both suppression current supply and the amplifier's ±15 volt DC input are provided by regulated power supplies rather than batteries.

Actual amplifier performance falls well within the specifications, as the actual test charts, Fig. 5, show. In fact, measured amplifier noise was about 1.5×10^{-14} amperes, which is less than the 2.5×10^{-14} amperes usually cited for flame noise. Consequently, there is clearly little advantage in using elaborate and expensive low-noise techniques to reduce amplifier noise below its present level.

THE AUTHOR

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NEW PRODUCTS

Model 147A/B/C FET Operational Amplifier excels on every major specification.

Most FET operational amplifiers offer advantages of high input impedance and low bias current but at a sacrifice in other specifications, usually voltage drift and common mode rejection. The Model 147A/B/C is the first general purpose FET op amp which excels in every performance category.

Model 147, with voltage drift of only 2μV/°C (max) and input bias current of 15pA (max), approaches the performance of chopper stabilized amplifiers and offers advantages of smaller size, lower price, lower noise and the versatility of differential inputs. The 147 boosts the inherently poor CMR of FET's to 300,000 — a 10 to 100 fold increase over most FET amplifiers.

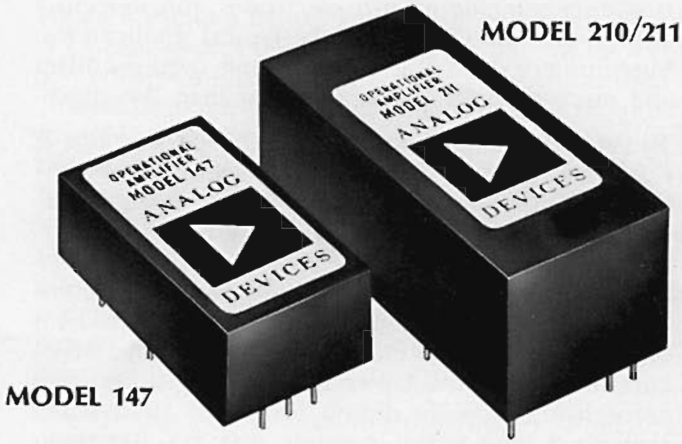
With 15pA bias current and 0.1pA noise, the 147 is excellent for measuring very low currents. With 10MHz and 10V/μsec response on both the inverting or noninverting inputs, the 147 is a good choice for sample/hold circuits, A-D and D-A convertors or wideband, high impedance ($10^{12}\Omega$, 3pF) noninverting amplifiers.

SPECIFICATIONS

Open Loop Gain, min.	10^6
Rated Output, min.	±10V @ 10mA
Unity Bandwidth	10MHz
Full Power Response, min.	150kHz
Slewing Rate, min.	10V/μsec
Common Mode Rejection	300,000
Input Impedance, C.M.	$10^{12}\Omega$, 3pF
Voltage Noise	3μV, p-p, DC to 1 Hz
Current Noise	0.1pA, p-p, DC to 1Hz
Size	2" x 1.2" x .62"

	Model A	Model B	Model C
Bias Current, max.*	30pA	15pA	15pA
Voltage Drift, max.			
(+10 to +60°C)	15μV/°C	5μV/°C	2μV/°C
(–25°C to +85°C)	15μV/°C	10μV/°C	5μV/°C
Price (1-9)	$110.	$120.	$135.

* At 25°C, doubles each 10°C



Model 210/211 Chopper Stabilized Op Amp's Drift is reduced with no increase in price.

Refinements in circuit design have netted two-fold reduction in voltage and current drift of now famous Model 210/211, making these units the industry's best buy in chopper stabilized operational amplifiers. For only $120, Model 211 offers maximum drift of 1μV/°C and 3pA/°C, while the Model 210 guarantees drift of 0.5μV/°C and 1pA/°C for $157. Other high performance specifications (see below) are retained as before.

SPECIFICATIONS

Open Loop Gain, min.	10^8
Rated Output, min.	±10V @ 20mA
Unity Bandwidth	20MHz
Slewing Rate, min.	100V/μsec
Overload Recovery	0.2μsec
Voltage Noise	5μV, p-p, DC to 1Hz
Size	2.8" x 1.3" x .95"

	Model 210	Model 211
Voltage Drift, max.	0.5μV/°C	1μV/°C
Current Drift, max.	1pA/°C	3pA/°C
Price (1-9)	$157.	$120.

A Parametric Operational Amplifier

Varactor bridge circuit achieves 10^{-12} amps input current and 10^{-14} amps current noise . . . a tenfold improvement over the best FET operational amplifiers.

By L. R. Smith

Analog Devices, Inc., Cambridge, Massachusetts

Presented at NEREM 1966

The 301 operational amplifier was designed to amplify currents in the picoampere range. Logarithmic and integrating circuits and a wide range of transducers including p-h electrodes, ion detectors and photomultiplier tubes are typical applications. Measuring or detecting small currents requires offset and noise currents that are smaller than the signal.

The parametric technique[1] was chosen because it can provide offset currents approaching zero without degrading other parameters desired in an operational amplifier. Chopper amplifiers have considerable low frequency noise and generally sacrifice either bandwidth and overload response or differential operation. Although metal oxide silicon field effect transistors and electrometer tubes have offset currents well below 1 picoampere, they suffer from zero drift especially during warm up. Field effect transistors have offset currents that are 100 times those possible with the 301.

The parametric amplifier offers a current noise improvement of as much as 1000 to 1 over conventional circuits. The noise performance is especially good in the sub-audio range where conventional circuits suffer from 1/f noise[2].

High performance differential operation comes almost automatically. Common mode parameters are limited only by the insulation properties of the materials used to build the input stage.

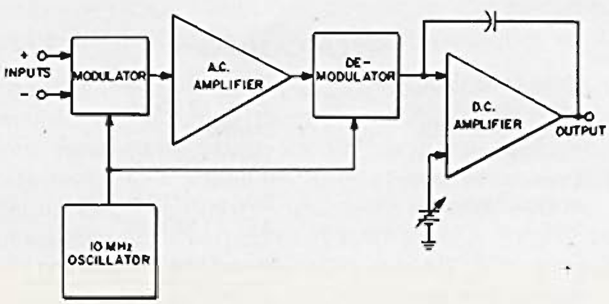


Figure 1. Block Diagram

A block diagram of the amplifier is shown in Figure 1. The parametric section is pumped by a low level 10MHz signal. The modulated output signal is amplified by a 3 stage AC amplifier. Negative feedback is used to stabilize the gain of the AC amplifier. The signal is then synchronously rectified by a diode ring demodulator. The signal is further amplified by a conventional differential transistor operational amplifier.

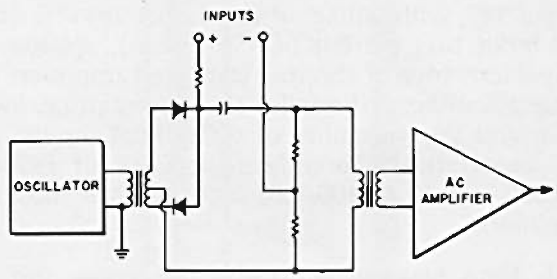
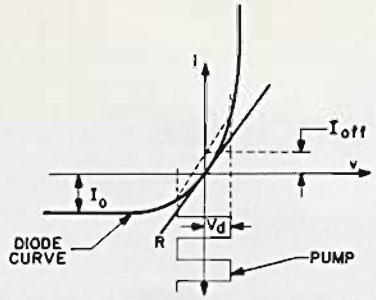


Figure 2. Parametric Section

Figure 2 shows the parametric section which is a balanced varactor diode capacitance bridge. The input signal unbalances the bridge, giving a suppressed carrier modulated output. The bridge has a power gain of 43 dB, resulting from the large ratio of DC input resistance to 10MHz output impedance.

As shown in Figure 3, offset current, I_{off}, is due to square law rectification in the varactors and can be made arbitrarily small by decreasing the pump amplitude, V_d. A compromise must be made between offset current and noise, since decreasing the pump amplitude reduces the parametric gain and increases the equivalent input noise voltage. One feature of this design is that only minor component value changes are needed to favor either offset current or noise performance. Normally the pump amplitude is fixed at a level which gives less than 2 picoampere offset current at 25°C. The offset current doubles for every 9.5°C increase in ambient temperature.



$$i = I_0\left(e^{\frac{V}{.026\eta}} - 1\right)$$

LET $\eta \approx 2$ AND EXPAND :

$$i \approx I_0\left[\frac{V}{.052} + \frac{1}{2}\frac{V^2}{(.052)^2} + + +\right]$$

$$R \approx \frac{.052}{I_0}$$

$\therefore$ FOR $R > 10^{+10}\Omega$ $I_0 < 5pA$

ASSUME SQUARE WAVE DRIVE :

$$I_{off} \approx \frac{I_0 V_d^2}{2(.052)^2} \approx \frac{V_d^2}{2 \times .052R}$$

$\therefore$ FOR $I_{off} < 1pA$ AND $R = 10^{+10}\Omega$

$$V_d < \sqrt{2 \times .052 \times 10^{+10} \times 10^{-12}}$$

$$V_d < 30\,mV.$$

Figure 3. Offset Current (I_{off}) vs Pump Amplitude (V_d)

The silicon varactor diodes are selected for high DC resistance at zero bias. This is important for obtaining low offset currents and high input resistance. The input resistance is typically 10^{10} at 25°C. The input resistance halves for every 9.5°C temperature rise.

The varactor diodes are matched to obtain less than 50µV/°C voltage offset temperature drift. Sufficient gain stabilization is provided in the carrier portion of the amplifier to allow trimming of the voltage offset to be performed after demodulation. Since the zero adjustment operates at DC instead of 10MHz the control may be placed at any convenient location.

The equivalent input current noise is due to Johnson noise arising from the zero biased diode resistance. At frequencies under 4Hz the current noise is approximately 0.0012pA/$\sqrt{Hz}$. Above 4Hz the current noise rises at a 6dB/octave rate due to the equivalent input noise voltage appearing across the input capacitance. The equivalent input noise voltage results from the noise of the first transistor in the AC stage. This amounts to 0.07µV/$\sqrt{Hz}$, above 1Hz. 1/f noise dominates below 1Hz. At sub-audio frequencies the amplifier has a 0.1dB noise figure at an optimum source impedance of 50M. This is a marked improvement over the 10 to 40dB sub-audio noise figure of more conventional circuits.

Since a 10MHz pump was used it has been possible to couple the varactor bridge with miniature ferrite transformers. The isolation thus obtained results in common mode rejection ratios of at least 160dB and common mode voltage capability of up to 300 volts. The transformer cores and much of the input circuitry is insulated with teflon so that the common mode resistance typically measures 6 x $10^{13}\Omega$. Thus when connected with series feedback the closed loop input impedance will generally be well above 10^{12}.

Other characteristics include: gain of greater than 5 x 10^5, gain bandwidth of 500kHz, and an output of ±10 volts at ±20mA. The amplifier is packaged as a fully encapsulated 3 cubic inch module, which can be mounted directly to a printed circuit board or inserted in a socket.

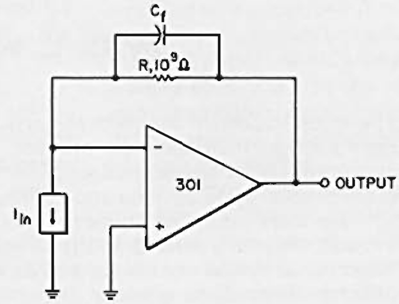


Figure 4. Current to Voltage Converter

Figure 4 shows a typical application of the amplifier. A current source transducer feeds a current to voltage converter. The output is one millivolt per picoampere of input. A feedback capacitor, C_f, is required to eliminate a phase lag due to the 500pF input capacitance of the amplifier and the 10^9 megohm feedback resistor. A 5 to 50 pF value, depending on overshoot requirements, will usually be sufficient. Shielding is usually required to eliminate 60Hz pickup and the effect of a moving charged body (such as a vibrating b+ lead). Graphite coated low noise coaxial cable has proven effective in reducing cable triboelectric effects.

The DC parametric amplifier described extends the performance of operational amplifiers down to the picoampere level. With further development of varactor diodes with higher gap potentials, such as gallium arsenide[3], offset currents of several femtoamperes should be possible. The low sub-audio noise of the 301 operational amplifier can be further improved if higher offset currents can be tolerated.

[1] Hoge, R. R. "A Sensitive Parametric Modulator For DC Measurements," IRE International Conventional Record, Vol. 8, Part 9, p. 34, 1960.
[2] J. R. Biard, "Low Frequency Reactance Amplifier," IEEE, February, 1960.
[3] J. Halpern, R. H. Rediker, "Low Reverse Leakage Gallium-Arsenide Diodes," Proceedings of the IRE, Vol. 48, No. 10, pp. 1780-1781, October, 1960.

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The Editor

Operational Amplifier Principles

by C. V. Weden, Fairchild Instrumentation*

The most important function of early, high quality, dc amplifiers was to perform mathematical *operations* such as multiplication, addition, integration, and differentiation. These amplifiers became known appropriately as *operational amplifiers*, and they were used principally in analog computers. Now, with modern solid-state devices and improved assembly and packaging technique, amplifiers of this description are more economical, more reliable, and more compact than their vacuum-tube predecessors. As a result, they fill a wide diversity of needs in signal generation and signal conditioning, active filtering, measurement, and control as well as in the traditional computing functions. The list of practical applications continues to grow as the amplifiers make use of new semiconductor devices such as matched differential pairs and improved junction- and MOS field-effect transistors.

Since more and more engineers are facing problems that can be solved best by operational amplifiers, we have prepared this note as a brief refresher on the fundamental principles involved. Most of the basic material can be found in textbooks, and several good articles on the state-of-the-art have appeared in various journals during the past year or two.

1. INTRODUCTION

The operational amplifier is a stable, high gain, dc-coupled amplifier which is usually used with a large amount of negative feedback. In this manner the *functional amplifier circuit* is made relatively insensitive to circuit loading and the effects of temperature and time on amplifier parameters. To a good approximation, the characteristics of the amplifier in a given circuit are the characteristics of the *external feedback elements* alone, over which the designer can exercise the degree of control warranted by the application. Furthermore, by the choice of feedback elements, the designer can use a given amplifier type for dozens of different functional circuits.

To illustrate the versatility of the operational amplifier, consider the generalized functional circuit in Figure 1. The triangle symbolizes the phase-inverting amplifier *per se* of high gain A, and Z_1, Z_f are the external feedback elements.

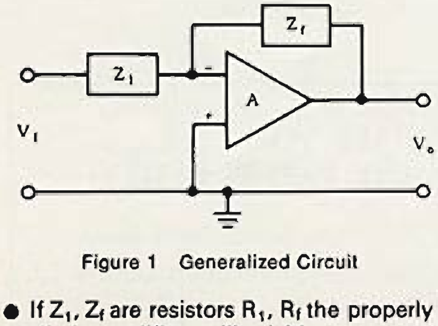


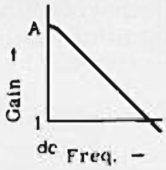
Figure 1 Generalized Circuit

- If Z_1, Z_f are resistors R_1, R_f the properly applied amplifier will yield an output voltage V_o equal to the input signal V_1 *multiplied* by the ratio R_f/R_1.
- If Z_1 is a resistor and Z_f is a capacitor, the circuit will *integrate* the input voltage in accordance with the externally established RC time constant.
- If Z_f is a transistor, the circuit will yield the *logarithm* of the input.

In addition to these linear and nonlinear operating modes, the operational amplifier is useful in the *switching* mode as a limit detector or comparator, and for circuit isolation and impedance matching.

These amplifiers will respond to *ac signals*, but the output capacitance of the transistors causes the amplifier gain to fall off as the frequency is increased. This effect is usually compensated in the multistage amplifier to produce a smooth roll-off of 6 db per octave, which renders the amplifier stable from oscillation under any value of *closed-loop* gain. (Fig. 2)

Figure 2
Open-Loop Response



The *ideal operational amplifier* would exhibit infinite input impedance so it would not load any source, and zero output impedance so it could drive any load. It would have infinite gain and bandwidth, and the output would be determined exactly by the input signal and the properties of the feedback circuit. The degree to which practical amplifiers approach the ideal is largely a matter of cost.

The principal errors that distinguish the practical operational amplifier from the ideal are the inherent *dc offset* voltage and current, which produce non-zero output with zero signal input, the *drift* of these offsets with temperature and time, and the higher frequency disturbance of similar nature which we call *noise*. Offsets, drift, and noise are *basic measurement errors*, and collectively they limit the signal resolution of the amplifier. A number of schemes have been developed to reduce their effect.

Theme and Variations

Today's operational amplifier market comprises amplifiers identified as *differential*, *chopper-stabilized*, *FET-input*, and *linear integrated circuit*, to name those in most popular usage.* At the heart of all of these variations is a basic dc amplifier with a balanced input stage (Fig. 3) for first order temperature compensation.

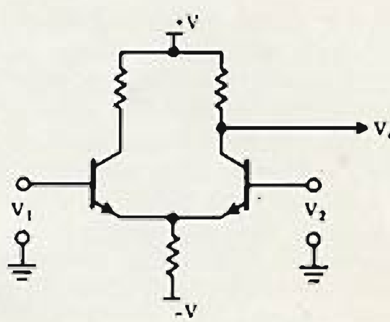


Figure 3 Differential Input Stage

* Editor's Note: Author omits varactor bridge op amps, see page 6.

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The *differential amplifier* is essentially this balanced input followed by additional stages, usually single-ended, for gain and impedance matching. Its output is a function of the difference in two inputs, hence it is particularly useful in amplifying signals from remote or otherwise isolated sources. In addition, it can be connected for non-inverting gain, which results in extremely high input impedance.

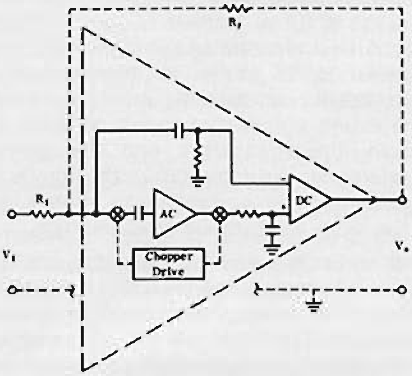


Figure 4 Chopper Stabilized Amplifier

The *chopper-stabilized amplifier* utilizes an ac-coupled "preamp" in conjunction with the external feedback loop to virtually null out the dc offsets and drift of the "main" dc amplifier. Referring to Figure 4, dc and low-frequency signals are modulated, amplified by an ac-coupled carrier amplifier, then demodulated, filtered, and fed to the differential-input of the main amplifier. Higher frequency signals are capacitively coupled directly to the dc amplifier. By careful proportioning of the circuits, the response of the preamp is superimposed on the response of the dc amplifier. The resulting over-all gain is the product of the dc gain of the "chopper channel" and the gain of the dc amplifier, over the full bandwidth of the dc amplifier. (Fig. 5)

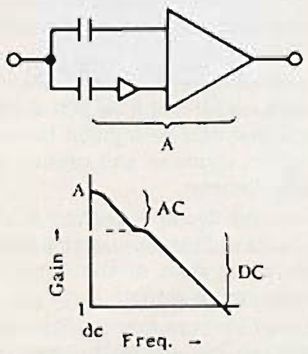


Figure 5 Chopper Amplifier Response

This technique is the most successful in reducing dc offsets and drift, and amplifiers are available with initial offsets less than ±20 microvolts and ±20 picoamperes and drift less than 0.5 μV/°C and 1 μV/week. Other types have approached this order of thermal drift *or* offset current, but where a single-ended amplifier can be used, this combination of specifications is most economically achieved by chopper-stabilization. Furthermore, the chopper-stabilized amplifier is unsurpassed for *long-term stability*.

Present-day chopper-stabilized amplifiers use only photo-resistor or transistor modulators (choppers), with the chopper drive generated internally by a multivibrator. The early mechanical chopper is still used, but short life, costly replacement, and the ac excitation voltage required spelled its demise as suitable electronic switches became available.

The *FET-input amplifier* is similar to the differential amplifier described above except for the use of field-effect transistors in the input stage. The features of this type amplifier are the extremely low input offset current and high input impedance associated with the FET devices. These characteristics make it especially useful for low drift integrators, sample and hold circuits, and commutating multiplexers.

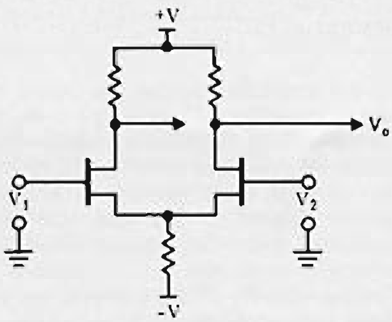


Figure 6 FET Input Stage

Newer types using MOS-FET's achieve input offset current on the order of a picoampere, which makes them attractive for charge amplifiers and electrometer applications. This feature is a trade-off with offset *voltage* and voltage drift.

The *linear integrated circuit* is a differential amplifier formed by monolithic semiconductor technology on a single silicon die. It is similar to the discrete-component differential amplifier except that the user must supply the frequency compensating elements (one or two resistors and capacitors) which are required to complete stability from oscillation. The principal advantage of today's linear IC is *small size*. This means more circuit functions per unit area of pc card, for example, and smaller equipment size for a given degree of complexity. The IC amplifiers have the potential for economy and reliability.

2. SPECIFICATIONS

Within all of the amplifier categories described above, a variety of products is available to help the user match the amplifier to his application without buying capability he doesn't need. The table (on the next page) shows the parameters usually specified and the ranges of gain and bandwidth, input performance, and output ratings available with amplifiers now on the market.

Since the operational amplifier is used in a large variety of closed-loop circuits, it is customary to normalize the specifications by using *open-loop* parameters where applicable and to refer closed-loop parameters to the input terminal. The latter process is accomplished simply by dividing the output figures by the closed-loop gain.

Definition of Terms

There is fairly general agreement in the industry on the definition of terms used to specify the characteristics and ratings of operational amplifiers. Some noteworthy conflicts exist, however, and there may be significant differences in the conditions under which certain specifications apply. The following list should help orient the uninitiated user.

- *DC Voltage Gain, open-loop* is the gain of the amplifier without external feedback. For good performance this parameter should be on the order of 100 times or more greater than the *closed-loop gain* desired in the functional amplifier circuit (= R_f/R_1). Under these conditions the closed-loop gain is virtually independent of the open-loop gain. Defining *loop gain* as the ratio of open-loop gain to closed-loop gain, the error by which closed-loop gain is diminished from R_f/R_1 is given by 1/loop gain.
- *Gain-Bandwidth Product* is a constant in amplifiers with the usual 6 db/octave *roll-off*, numerically equal to the frequency at which the gain has "rolled off" to unity. An amplifier with a unity-gain *cross-over frequency* of 1 MHz, for example, will have an open-loop gain of 10^4 at a signal frequency of 100 Hz. In *general-purpose* units the gain is down to unity in the frequency range of 0.5 to 5 MHz. In *wideband* units the unity-gain frequency is in the range of 10 to 100 MHz.
- *Gain Roll-Off* is the rate at which the open-loop gain falls off with increasing signal frequency. The optimum roll-off is 6 db per octave (= 20 db per decade), which is an expression of constant gain-bandwidth product. This is the response associated with a maximum phase shift of —90°, for which the amplifier will be

Continued from page 9

Comparative Specifications of Currently Available Operational Amplifiers

Specifications	General Purpose Differential	Chopper-stabilized	FET-Input	Integrated Circuit
DC Voltage Gain, open-loop:	10^4-10^6	10^6-10^8	10^4-10^6	10^3-10^5
Gain-Bandwidth Product:	0.4-10MHz	0.3-10 MHz	1-30 MHz	1-10 MHz
Slewing Rate Limit:	0.1-3 V/μs	0.6-50 V/μs	0.5-20 V/μs	0.1-30 V/μs
Initial Input Offset:				
Voltage:	0.3-1 mV	10-200 μV	1-5 mV	1-5 mV
Current:	1.5-500 nA	10-2000 pA	1-50 pA	100-500 nA
Drift vs. Temperature:				
Voltage:	5-100 μV/°C	0.5-10 μV/°C	5-100 μV/°C	3-50 μV°C
Current:	0.2-5 nA/°C	0.5-20 pA/°C	20-100 pA/°C	0.1-5 nA/°C
Drift vs. Time:				
Voltage:	10-100μV/day	1 μV/wk	5-100μV/day	—
Input Noise:				
0.1-1.0 Hz:	—	5-10 μV p-p	—	—
Wideband:	1-20 μV rms	10-150μV rms	2-30μV rms	—
Input Impedance, open-loop:				
Differential:	100-500 k	0.5-1 Meg	10^{10}-10^{12} Ω	10-500 k
Common Mode:	10-500 Meg	—	10^{10}-10^{12} Ω	—
Input Voltage (maximum):				
Differential:	5-15 V	—	—	1-8 V
Common Mode:	3-20 V	—	8-11 V	1-11 V
Common Mode Rejection:	50-100 db	—	60-74 db	60-100 db
Output Voltage:	10-20 V	10-150 V	10-11 V	3-12 V
Output Load Current:	1-20 mA	2-100 mA	2-20 mA	1-10 mA
Output Capacitive Loading:	0.0005-10μf	0.0005-0.02μf	—	—

free from any tendency to oscillate under all values of closed-loop gain.

● *Slewing Rate Limit* is the maximum time rate of change of output voltage for a step input. "Step response" is sometimes given as *velocity limit* or as *risetime*. Slewing rate is usually specified in volts per microsecond. It is related to the maximum frequency *f* at which full output voltage *V* can be obtained by $\Delta V/\Delta t = 2\pi fV$.

● *Input Offset Voltage* is the input voltage required to zero the dc component of the output with zero input signal and zero source impedance. Offset is an error inherent to some degree in all practical amplifiers. The offset usually can be adjusted to zero by applying a small voltage derived from the amplifier's power supplies. *Initial* offset signifies that the parameter is determined before any external balance adjustments are made. Offset, drift, and noise (following) combine to add a variable error to the input signal. The total error must be small compared to the minimum signal to be amplified with reasonable accuracy, or linearity.

● *Input Offset Current* is the input current required to zero the dc output current with zero signal and infinite source impedance. Offset current has the effect of an error voltage drop across R_1, the series input (or "summing") resistance. In the inverting amplifier the closed-loop *input impedance* (following) is equal to R_1; therefore, high input impedance is a trade-off with offset error. In the differential amplifier offset current is the *average* of the currents into the two input terminals with zero output. Most linear IC's and some discrete-component amplifiers identify this parameter as *input bias current*, and specify the *difference* in the two currents as *offset*. Since the differential current may be 10-25% of the average, it is important to note this conflict in terminology. In some applications it is the differential input current that constitutes the error, but in many circuits, especially those involving integration, it is the current into either input that is the limiting factor.

● *Drift vs. Temperature* is the slowly varying change in offset voltage and offset current due to a change in temperature. Voltage drift vs. temperature is usually specified in μV/°C, but it is not necessarily linear over the entire operating temperature range of the amplifier.

● *Drift vs. Time* refers to the similar effect of time on offset voltage and current.

● *Drift vs. Supply Voltage* is a measure of the effect of changes in supply voltages on the offsets. This parameter is properly specified in μV per percent change in supply voltage from the rated value, but it is sometimes given in μV per volt.

● *Input Noise* is the normalized value of any output disturbance not contained in the input signal. Noise is generated in the transistors and resistors of the amplifier, and it may be coupled from power lines and rf sources. To properly specify noise one must define the source resistance R_1 as well as the bandwidth. "Wideband" noise voltage includes frequencies up to the range of 1 kHz to 1 MHz, usually specified in rms volts. Some amplifiers are also specified for peak-to-peak noise in the range of 0.1 to 1 Hz.

A basic feature of the differential amplifier is its ability to amplify signals impressed across its input terminals, rejecting signals that appear between the two inputs together and the ground reference. We may define the former as *differential* or *normal mode* signals and the latter as *common mode* signals.

● *Input Impedance, open-loop* is the complex impedance seen looking into the input terminals of the amplifier without external feedback. In the conventional (bipolar transistor)-input differential amplifier Z_{in} depends on the current gain h_{FE} and the intrinsic base resistance r_e of the input devices; it is typically in the range of 200 k to 500 k. The extremely high (typically 10^{11} ohms) Z_{in} of the FET-input amplifier, as well, results from the geometry and bulk properties of the FET's. In the chopper-stabilized amplifier, Z_{in} is proportional to the *on* and *off* resistance of the chopper device, typically 1 Meg at dc.

● *Common-Mode Input Impedance* of a differential amplifier is the impedance between the two input terminals together and ground. It is determined largely by the characteristics of the input transistors and the near-infinite impedance of the current source used in the common emitter lead of the input stage.

● *Differential Input Voltage* rating is the maximum voltage that can be applied across the input terminals of a differential amplifier without causing damage to the amplifier.

● *Common-Mode Input Voltage* rating of a differential amplifier is the maximum voltage that can be applied between the two inputs together and ground without causing damage.

● *Common-Mode Rejection* is the ratio of the gain of the amplifier to a differential signal to the gain of the amplifier to a common-mode signal.

● *Output Voltage* rating is the maximum output voltage which the amplifier will develop in the linear operating region; i.e., before the onset of saturation.

● *Output Load Current* rating is the maximum current that the amplifier will deliver to, or accept from, a load. This rating includes the amount, however small, which is caused to flow in the feedback loop.

- *Output Capacitive Loading* is the maximum capacitance that can be placed on the output of the amplifier at unity gain without increasing the phase shift to the point of inducing oscillation. The limiting value increases in direct proportion with the closed-loop gain. The traditional way to accommodate higher loading capacitance is to isolate the load from the amplifier output terminal with a resistance. With some amplifiers the resistance required is only a few ohms, in which case the error and dissipation introduced are minimal.
- *Output Impedance, open-loop* is the complex impedance seen looking into the output terminals of the amplifier with no external feedback. In closed-loop operation the output impedance is equal to the open-loop impedance divided by the loop gain. If the open-loop impedance is not more than a few hundred ohms and the loop gain is high enough for good gain accuracy and stability, the closed-loop impedance will be on the order of an ohm or less, which can be neglected in most applications.

3. FEEDBACK AMPLIFIER ANALYSIS

To guide the designer in the use of an operational amplifier to fulfill his requirements, it may be helpful to review just what we achieve with the high intrinsic voltage gain and negative feedback, and how closed-loop performance is affected by the finite open-loop parameters of the practical amplifier.

Closed-Loop Gain

Let us start by developing an expression for the *transfer function* of a basic feedback amplifier, which we commonly call the *closed-loop gain*. (Fig. 7)

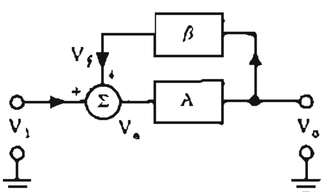


Figure 7 Basic Feedback Amplifier

A is the open-loop gain of our operational amplifier, assumed to be phase inverting so that if V_1 is positive, V_0 will be negative.

If A were precisely known and controllable, we might satisfy a number of voltage-gain applications without the feedback loop. Usually this is not the case, so we sample the output V_0 and feed back a portion β to the input terminal Σ. Here the feedback voltage V_f is summed with the input signal V_1. Terminal Σ is defined as the *summing junction*, and since V_f must be a negative number with respect to V_1, this terminal is often identified as the *negative* or *inverting* input of the amplifier. β is known as the feedback factor.

The input error voltage is

$$V_e = V_1 + V_f, \tag{1}$$

and by definition, $V_f = \beta V_0$.

Also,
$$V_e = \frac{V_0}{A}. \tag{2}$$

Substituting,
$$\frac{V_0}{A} = V_1 + \beta V_0$$

$$V_0(1-A\beta) = AV_1$$

and
$$\boxed{\frac{V_0}{V_1} = G = \frac{A}{1-A\beta}}, \tag{3}$$

where G is defined as the closed-loop gain. One sees that if the quantity $A\beta >> 1$,

then
$$\boxed{G \simeq -\frac{1}{\beta}}, \tag{4}$$

or the *closed-loop gain* is approximately equal to the reciprocal of the feedback factor, which *is independent of the highly variable open-loop gain*. The minus sign denotes phase inversion in the amplifier.

The quantity Aβ is defined as *loop gain*, which involves both the amplifier itself (A) and the external feedback loop used with it (β). It is usually the loop gain that determines how close to the ideal a given amplifier circuit will appear, e.g., whether the simplified relationship of Eq. (4) is valid or not.

Example: Find the closed-loop gain of an amplifier with an open-loop gain A = −10,000 if we feed back 10 percent of the output, i.e. β = 0.1.

$$G = \frac{A}{1-A\beta} = \frac{-10\,000}{1+1000} \simeq -9.99$$

Using the same operational amplifier (A = —10 000) but with β = 0.01, we find

$$G = \frac{-10\,000}{1+100} \simeq -99,$$

and again with β = 0.001,

$$G = \frac{-10\,000}{1+10} \simeq -900.$$

Notice that in the above examples $G \simeq -1/\beta$ would have given us —10, —100, and —1000, respectively, and that as the loop gain Aβ becomes smaller, G deviates more and more from $-1/\beta$. Indeed, the gain error is given by $1/(A\beta)$. If $A\beta >> 1$, one sees further from Eq. (3) that

$$\boxed{A\beta \simeq \frac{A}{G}}, \tag{5}$$

i.e. the loop gain may be determined by the *ratio* of open-loop gain to closed-loop gain.

Returning to Eq. (2) we see that as A approaches infinity, the error voltage V_e approaches zero, i.e., the summing junction is maintained close to the ground reference. With $A = -10^7$ and an output of ±20 volts, for example, V_e will be ±2 microvolts. Using the approximation that $V_e = 0$, which usually is valid, it follows that the error current into (or out of) the amplifier will be zero. The current through the feedback loop, then, is equal to the current through the source.

Gain Stability

We have stated that the principal effect of the negative feedback loop is to stabilize the voltage gain of the amplifier. This effect can be shown quantitatively by differentiating G with respect to A:

From Eq. (3)
$$G = \frac{A}{1-A\beta},$$

then
$$\frac{dG}{dA} = \frac{1}{(1-A\beta)^2},$$

and
$$\boxed{\frac{dG}{G} = \left(\frac{1}{1-A\beta}\right)\frac{dA}{A}} \tag{6}$$

Therefore, the relative change in closed-loop gain is approximately equal to the relative change in open-loop gain divided by the loop gain.

Example: Find the variation in closed-loop gain using an amplifier with nominal open-loop gain of —10 000 subject to 20 percent deviation, and β = 0.01.

$$\frac{dG}{G} = \left(\frac{1}{1-A\beta}\right)\frac{dA}{A} = \left(\frac{1}{1+100}\right)0.20 \simeq 0.002$$

i.e., G will vary by only 0.2 percent.

Where do we Buy a β ?

The feedback factor β usually takes the form of a precision voltage divider, and the input signal voltage V_1 is made to appear as a current source by feeding it through a series resistance: (Fig. 8)

Operational Amplifier Principles

Continued from page 11

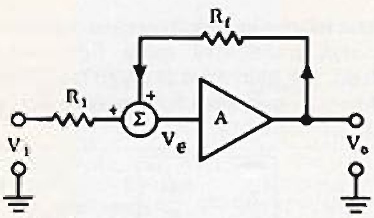


Figure 8 Amplifier With Feedback

By superposition,

$$V_e = \frac{R_f}{R_1 + R_f} V_1 + \frac{R_1}{R_1 + R_f} V_o.$$

Substituting $V_e = V_o/A$,

$$\left(1 - \frac{AR_1}{R_1 + R_f}\right) V_o = \frac{AR_1}{R_1 + R_f} V_1.$$

and,

$$\frac{V_o}{V_1} = G = \frac{\dfrac{AR_f}{R_1 + R_f}}{1 - \dfrac{AR_1}{R_1 + R_f}}$$

Simplifying and factoring out the ratio $-R^f/R_1$,

$$\boxed{G = -\frac{R_f}{R_1}\left[\frac{1}{1 - \dfrac{R_1 + R_f}{AR_1}}\right]} \quad (7)$$

Now Eq. (3) can be manipulated into the form

$$G = -\frac{1}{\beta}\left[\frac{1}{1 - \dfrac{1}{A\beta}}\right] \quad (8)$$

from which, by comparison with Eq. (7), it can be deduced that

$$\frac{AR_1}{R_1 + R_f} = A\beta = \text{loop gain.}$$

Again, if the loop gain is much larger than unity, Eq. (7) reduces to

$$\boxed{G \simeq -\frac{R_f}{R_1}} \quad , (9)$$

which is consistent with Eq. (4) that the closed-loop gain is independent of the open-loop gain. If it is not true that $A\beta >> 1$ we must use the more precise expression, Eq. (7).

Example: Find the closed-loop gain of an amplifier with $A = -10\,000$ if $R_1 = 10$ k, $R_f = 10$ Meg. From Eq. (7),

$$G = -\frac{10\,\text{Meg}}{10\,\text{k}}\left[\frac{1}{1 - \dfrac{10\text{ k} + 10\text{ Meg}}{-10\,000\,(10\text{k})}}\right]$$

$$= 1000\left(\frac{1}{1 + 0.1}\right)$$

$$= -910$$

Note again that the voltage divider ratio, Eq. (9), would have predicted $G = -1000$. A closed-loop gain of -1000 could be achieved, of course, by adjusting R_f/R_1 to approximately 1100, but in this example the loop gain is only 10, which is not adequate to insure a high degree of gain stability. If a nominal closed-loop gain of -1000 is required, better design practice would be to increase $A\beta$ by selecting another amplifier with higher open-loop gain. With $A = -10^6$, for example, the loop gain is increased to 1000 and the error due to finite open-loop gain is reduced to 0.1 percent.

Frequency Response

It is important to realize that the high open-loop gain we have been using is available only at dc and very low frequencies. At higher frequencies the gain is attenuated markedly, due largely to the effects of transistor output capacitance. To insure stable operation, the operational amplifier usually is compensated to provide a smooth roll-off of gain with increasing frequency.

The amplifier therefore looks like an RC lag network which attenuates the dc open-loop gain A; the effect of frequency on A may be seen by analyzing the response of this network: (Fig. 9)

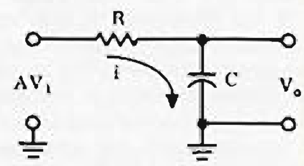


Figure 9 RC Lag Network

The current i through R and C is

$$i = \frac{AV_1}{R + X_c} = \frac{AV_1}{R + \dfrac{1}{j\omega C}}$$

where $\omega = 2\,\pi f$ is the frequency in radians, and j is the complex operator. Therefore,

$$V_o = \frac{i}{j\omega C} = \frac{A\,V_1}{1 + j\omega CR}$$

and

$$\frac{V_o}{V_1} = A(\omega) = 1 + \frac{A}{j\omega CR} \quad (10)$$

where $A(\omega)$ is the open-loop gain as a function of frequency.

The frequency response of a lag network is shown in Figure 10.

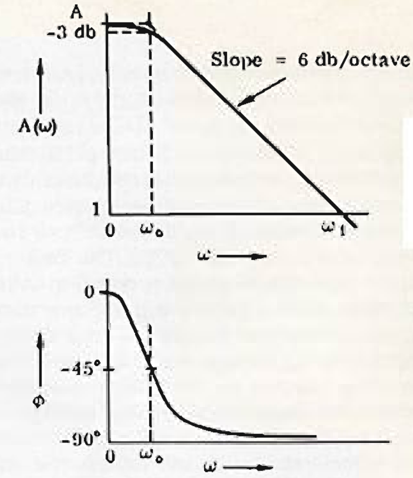


Figure 10 Response of RC Network

This display of gain and phase shift vs. frequency is known as a Bode plot.

The break frequency ω_o occurs when $X_c = R$; then $\omega_o = 1/(RC)$ and

$$\boxed{A(\omega) = \frac{A}{1 + j\dfrac{\omega}{\omega_o}}} \quad (11)$$

The magnitude of $A(\omega)$ then is

$$A(\omega) = \frac{A}{\left[1 + \left(\dfrac{\omega}{\omega_o}\right)^2\right]^{1/2}} \quad (12)$$

and the phase shift ∅ is

$$\varnothing = -\tan^{-1}\frac{\omega}{\omega_o} \quad (13)$$

That $A(\omega)$ is down 3 db from A at the break frequency may be shown by evaluating Eq. (12) at $\omega = \omega_o$:

Then

$$\frac{A(\omega)}{A} = \frac{1}{(1 + 1^2)^{1/2}} = 0.71 = -3\text{ db.}$$

Also,

$$\varnothing_o = -\tan^{-1} 1 = -45°.$$

In chopper-stabilized amplifiers the principal break frequency is less than 1 Hz, therefore it is of interest to examine Eq. (12) at $\omega >> \omega_o$. This yields

$$\frac{A(\omega)}{A} = \frac{\omega_o}{\omega},$$

or

$$A(\omega)\,\omega = A\,\omega_o \simeq \text{constant.}$$

Since $\omega_o \simeq 0$ we may interpret ω as *bandwidth*. The practical amplifier is characterized by a reasonably constant *gain-bandwidth product*. This figure is numerically the same, then, as the unity-gain cross-over frequency ω_1 at which $A(\omega) = -1$,

i.e., $$\boxed{A(\omega)\,\omega = -\omega_1} \quad (14)$$

The nearly constant slope of the response may be described as a *gain roll-off* of 6 db per octave, or 20 db per decade. This is simply a statement that the gain is down by a factor of 2 when the frequency doubles, or the gain is down by 10X if the frequency is increased by 10X.

Example: Find the closed-loop gain of an amplifier with a dc open-loop gain, $A = -10^7$, gain-bandwidth product $f_1 = 1$ MHz, if $R_1 = 10k$, $R_f = 1$ Meg, and the signal frequency $f = 100$ Hz.

From Eq. (14) $A(\omega) = -\dfrac{1\ \text{MHz}}{100\ \text{Hz}} = -10^4$.

Now from Eq. (7), using $A(\omega)$ instead of A,

$$G = -\frac{R_f}{R_1}\left[\frac{1}{1-\dfrac{R_1+R_f}{A(\omega)\,R_1}}\right]$$

$$= -\frac{1\ \text{Meg}}{10k}\left[\frac{1}{1-\dfrac{10k+1\ \text{Meg}}{-10^4\,(10k)}}\right]$$

$$= -100\left(\frac{1}{1+0.01}\right)$$

$$= -99.$$

The important point here is that *at a frequency of 100 Hz, the applicable open-loop gain is $A(\omega)$* and not the dc open-loop gain given on the amplifier specification sheet.

Let us put this example on the Bode plot, as shown in figure 11.

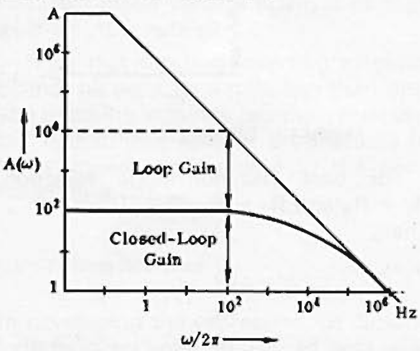


Figure 11 Open and Closed Loop Response

Since gain is plotted on a logarithmic scale (db), the loop gain (= open-loop gain/closed-loop gain) appears as the difference between the open-loop gain $A(\omega)$ and the closed-loop gain. The loop gain $A(\omega)\beta = 100$, and the closed-loop gain G is approximately 100X with an error of 1/100.

Note that the closed-loop response is flat until it approaches the roll-off, whereupon it merges with the open-loop response. Beyond this point the effect of the negative feedback loop is no longer present.

Certain amplifiers are compensated with a higher break frequency to extend the usable gain-bandwidth capabilities of the amplifier to higher signal frequencies.

The unity gain-bandwidth of the amplifier is relatively fixed, however, so the gain must roll off at a faster rate, e.g. at 12 db/octave. This rate of closure with the closed-loop gain line is accompanied by additional phase shift, unfortunately, with the result that the amplifier is no longer stable under all conditions of closed-loop operation.

Another consequence of the attenuation of $A(\omega)$ is increasing input error voltage, V_e. Recall that an amplifier with $A = -10^7$ and $V_o = \pm 20$ volts, for example, would have an input error of ± 2 microvolts at dc. If the signal frequency is 100 Hz, however, and $A(\omega) = -10^4$, then V_e would be ± 2 millivolts.

Transient Response

The closed-loop response of an amplifier to a pulse input or step function is an exponential with the operational time constant. The time required to reach a given proportion of the final output voltage is proportional to the closed-loop gain and inversely proportional to the unity-gain bandwidth.

The inherent bandwidth of the amplifier, then, along with the roll-off compensation, affects the basic step response of the amplifier. This response is usually specified as *slewing rate limit* $\Delta V/\Delta t$, or maximum rate of change of output voltage, in volts per microsecond.

Occasionally this response is specified as the maximum frequency f_m at which the peak output voltage V_m can be realized:

$$\frac{\Delta V}{\Delta t} = 2\,\pi f_m\,V_m,$$

then

$$f_m = \frac{\Delta V/\Delta t}{2\pi V_m}.$$

For sine-wave signals the limiting factor is usually not slewing rate, but gain-bandwidth product and the available loop gain.

Input and Output Impedance

The infinite input impedance and zero output impedance of the ideal amplifier are approached in the real world as shown in the table of comparative specifications. Open-loop input impedance ranges from about 200 k to 10^{12} ohms, output impedance from a few ohms to a few thousand ohms.

Finite input impedance Z_{in} does not have a direct effect on closed-loop gain, but if the summing resistance R_1 is comparable to or greater than Z_{in}, gain accuracy is degraded through a reduction in the loop gain. Finite output impedance Z_o has a similar effect, but usually negligible if Z_o is less than a few hundred ohms.

With feedback, the open-loop figures change considerably. In the basic inverting amplifier, the closed-loop input impedance is substantially the summing resistance R_1. In the non-inverting connection of the differential amplifier, the closed-loop input impedance is equal to the open-loop impedance times the loop gain. This effective high-impedance feature of the differential amplifier is often utilized to accommodate high-impedance sources without loading.

Closed-loop output impedance is equal to the open-loop impedance *divided* by the loop gain, which can yield a very small value. With a loop gain of 1000, or higher, therefore, it matters little whether the open-loop output impedance is 5 ohms or 500 ohms.

4. APPLICATIONS

Typical Circuits

The endless collection of circuits which make effective and convenient use of the operational amplifier attest to its versatility. Many examples are described in the technical and trade literature, and a full recounting is beyond the scope of this paper.

Some of the more frequently used operational amplifier circuits are shown below.

Summing Amplifier

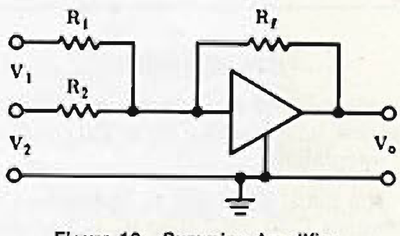


Figure 12 Summing Amplifier

This circuit illustrates summing and inverting voltage gain.

$$V_o = -R_f\left(\frac{V_1}{R_1} + \frac{V_2}{R_2}\right).$$

Operational Amplifier Principles

Continued from page 13

where the negative sign indicates phase inversion of V_O with respect to the input.

Note that if $V_2 = 0$, then $V_O = -\frac{R_f}{R_1} V_1$

the closed-loop input impedance is R_1 for source V_1.

Inverting Gain, High Impedance

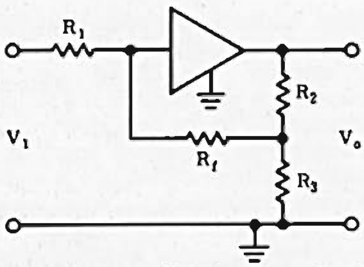


Figure 13 Inverting, High Impedance Amplifier

In this circuit V_O is divided and then a portion fed back to provide high closed-loop gain and high input impedance ($= R_1$) without the need for excessively high feedback resistance.

$$V_o \simeq -\frac{R_2}{R_3}\left(\frac{R_f}{R_1} V_1\right)$$

If V_1 is a reference voltage source and R_2 is a floating load, this circuit acts as a constant current supply, delivering large output current I_O with low current drain from the reference source.

$$I_O = \frac{V_{ref}}{R_1}\left(\frac{R_f + R_3}{R_3}\right)$$

Integrator

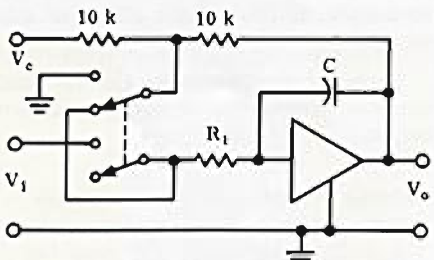


Figure 14 Integrator

Here an initial reference state must be defined, which shows up as the constant of integration.

The initial condition is established by charging C to $-V_c$; then V_1 is switched in to integrate.

$$V_o = \frac{-1}{R_1 C}\int_0^t V_1\,dt - V_c$$

If $V_c = 0$, $R_1 = 100\,k$, $C = 1\,\mu f$, then $V_O = -10\int_0^t V_1\,dt$. The integration ramp will continue until V_O reaches the saturation level of the amplifier or until it is otherwise terminated. The method of switching from *initial condition* to *integrate* isolates the summing junction from the disturbing effects of switching transients.

To measure integrator drift, $\Delta V/\Delta t$, one may set the initial V_c, then set $V_1 = 0$ and observe any deviation from the initial offset, $\frac{\Delta V}{\Delta t} = \frac{i}{C}$, where i is the input offset current of the amplifier. In practical applications the integration time *t* may vary from a few milliseconds to several hours so the limit of acceptable input current will vary accordingly.

Voltage Comparator

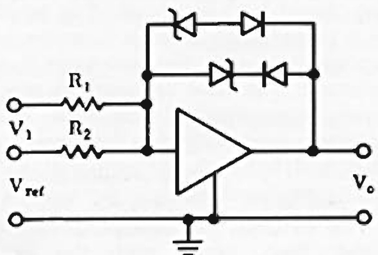


Figure 15 Voltage Comparator

V_{ref} is of opposite polarity to the unknown V_1, and R_1, R_2 provide scaling if necessary. V_O detects the coincidence of V_1 to the absolute value of V_{ref}.

Overload Clamps

Excessive input voltage will saturate the amplifier in the simple voltage-gain applications. Since recovery from the overload in a chopper-stabilized amplifier may take several seconds after removal of the offending signal, it is often prudent to prevent saturation. In simplest form the output may be clamped with a pair of back-to-back Zeners across the feedback resistor, limiting V_O to $\pm V_z$, the Zener voltage. During normal operation the amplifier is in the linear mode and the gain is determined by $-R_f/R_1$. But when the V_O reaches V_z the Zeners conduct, increasing the negative feedback and preventing V_1 from driving V_O into saturation. Recovery time is reduced to milliseconds, but one must be aware that any leakage current through the diodes can introduce significant error.

The following circuit illustrates an improved clamp which reduces the leakage problem.

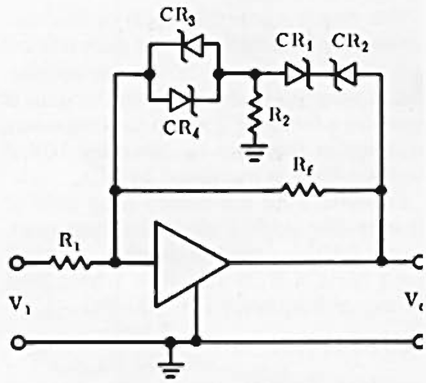


Figure 16 Overload Clamp Circuit

In normal amplifier operation R_2 ab sorbs the leakage of the Zeners CR_1 an CR_2; the minute drop across R_2 cause only a very small leakage through diod CR_3 and CR_4 into the summing junctio

Differential Voltage Gain

This circuit shows the basic invertin amplifier connection for a differenti input, single-ended output:

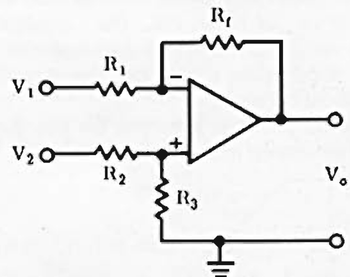


Figure 17 Differential Amplifier

For best common-mode rejectio $R_1 = R_2$ and $R_3 = R_f$.
Then,

$$V_O = -\frac{R_f}{R_1}(V_1 - V_2)$$

Non-Inverting Gain

The differential amplifier may be connec ed in the following manner to deliver output voltage in phase with the inpu

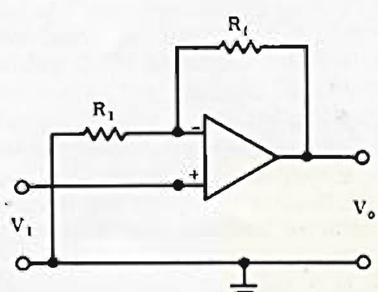


Figure 18 Non-Inverting With Gain

$$V_o = +\left(\frac{R_f + R_1}{R_1}\right) V_1$$

An important feature of this circuit is that it maintains high input impedance even with low values of R_1 and R_f. (Closed-loop Z_{in} = open-loop Z_{in} times the loop gain.)

If we remove R_1, we have the unity-gain voltage follower, often used as an isolation buffer. Here virtually no current flows into the summing junction, hence through the feedback loop. Theoretically, therefore, R_f may have any finite value, but offset current errors and the effects of stray capacitance are minimized with $R_f = 0$.

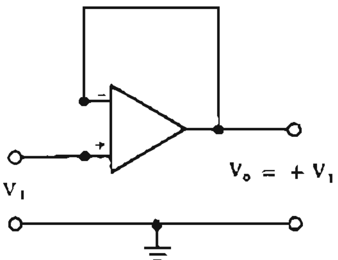


Figure 19 Voltage Follower

Since V_1 is the *common-mode input voltage*, the full output voltage capability of the amplifier can be utilized only if the common-mode voltage rating is as high as the output rating.

Note that the non-inverting amplifier cannot be used for a gain less than one. The inverting amplifier has this versatility, but it does not achieve the increase in input impedance possible in the non-inverting mode.

Gain-Setting Resistors

In developing the expression for closed-loop gain we were concerned only with the ratio of R_f to R_1. The magnitude of these resistors is not completely arbitrary, and the following considerations may aid the selection process.

First, since input current offset, drift, and noise are proportional to R_1, this resistance should be kept reasonably small. The optimum range of R_1 lies between the bounds for amplifier specification compliance and source loading.

The feedback resistor R_f should be low enough to draw a current of at least one microampere ($= V_o/R_f$) in order to firmly establish the voltage divider. In addition, The stray capacitance of very high resistance values tends to degrade the frequency response. On the other hand, R_f must be large enough to produce the desired gain with a reasonable value of R_1.

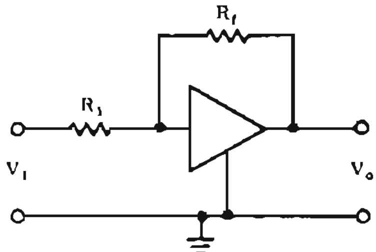


Figure 20 Gain Setting Resistors

With these considerations observed, it is usually possible to adjust R_1 and R_f to standard values that will yield the desired gain. In general it is feasible to use resistors between about 1 k and 20 Meg to achieve closed-loop gains up to about 1000X. Usually metal-film resistors are necessary in order to preserve the temperature and time stability of the amplifier.

Amplifier Selection

The first decision to make in selecting an amplifier is whether a *single-ended* or a *differential* input is required. For operation in the normal linear mode this is often a question of the physical disposition of the source and load or other factors affecting the compatibility of grounds. If the load is isolated from ground, a differential output is required as well.

Secondly, we must consider the *voltage level and impedance of the source* in relation to the offset, drift, and noise specifications. It may be feasible to balance out the initial offsets, but the drift and noise of the amplifier, of course, must be low enough to permit resolution of the minimum signal level. In some applications, such as those involving integration, the input offset current is the determining parameter.

Next, we need to select an amplifier with an open-loop gain that is high enough for the application. "High enough for the application" means that the resulting *loop gain* (open-loop gain/closed-loop gain) will yield satisfactory stability and gain accuracy. Remember that in calculating loop gain the specified dc open-loop gain may be used only for dc and very low frequency signals. At frequencies above a few Hz the effective open-loop gain is found by dividing the specified gain-bandwidth product by the signal frequency.

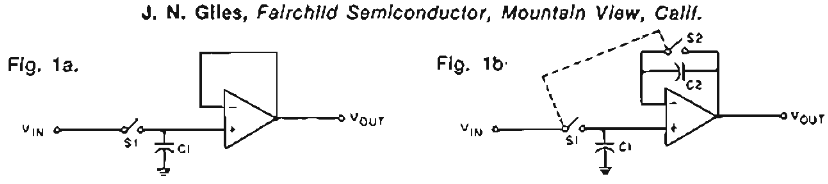
In certain applications it is necessary to match the output capability of the amplifier to the voltage and current required to drive the load. In critical applications the input specifications dominate the choice, since the output can be boosted as required with a fairly simple buffer.

Other considerations include package size and style, operating temperature and compatibility with other environmental conditions, power supply constraints, and cost. The relative importance of these items depends on the requirements of the whole system concerned as well as the specific function of the amplifier.

by C.V. Weden, Fairchild Instrumentation

Capacitor improves sample-and-hold circuit

J. N. Giles, *Fairchild Semiconductor, Mountain View, Calif.*



Marked improvement in voltage-holding ability of a sample-and-hold circuit is possible when a capacitor is added (b) to the conventional circuit (a).

Conventional sample-and-hold circuits using operational amplifiers have the general form of Fig. 1a. The voltage to be held is sampled through switch *S1* and stored on capacitor *C1*. The amplifier functions as a high-input-impedance, unity-gain buffer between the voltage on the capacitor and the outside world. The charge on the storage capacitor leaks off at a rate determined by the amplifier input bias current and the shunt resistance to ground.

The addition of capacitor *C2*, equal to *C1*, between the output and the inverting input of the amplifier (see Fig. 1b) improves the decay time of the circuit by better than a factor of ten. The circuit operates as before, except that leakage across *C1* is now compensated for by an equivalent leakage across *C2* such that the output voltage remains almost constant, depending on the degree of match between the two input bias currents and the capacitors. The output drift can even be adjusted to zero by trimming one of the capacitors to compensate for the small difference in bias currents.

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