

# ANALOG DIALOGUE

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## Chopper Amplifier's PPM Stability Enables Electron Microscope To Scrutinize Individual Molecules\*

\*Based on informal conversations with Robert Libbey, Design Engineer RCA, Camden, N.J.

*Current source, based on ultra-stable operational amplifier, supplies lens current with 2.5 part-per-million stability, enables a new generation of electron microscopes to resolve better than 8 Angstroms. This article introduces electron microscope principles, illustrates the need for utmost current stability, and describes the current source design.*

The electron microscope "sees" by using an electron beam as an ultra-short wavelength "light source." Owing to the dramatic reduction in diffraction effects afforded by its short-wave illumination, the instrument can resolve a thousandfold more finely than the best optical instrument. Evolution of the electron microscope during the past few decades has opened entirely new realms of exploration and study, and has enabled researchers to examine individual molecules and viruses not previously discernible with even the most powerful optical microscope.

Electronic engineers, so familiar with a heated cathode's random emission, may be forgiven for believing electron beams to exemplify particle rather than wave motion. Yet during 1924, deBroglie paralleled Einstein's earlier unification of wave and corpuscular light theories by showing that a beam of charged particles may also be represented by a wave motion of definite and calculable wavelength. In fact, the wavelength of an electron beam accelerated through 50,000 volts is roughly 0.05 Angstroms, or  $5 \times 10^{-10}$  centimeters. This is approximately six orders-of-magnitude shorter than the  $4 \times 10^{-5}$  centimeters for visible light at the ultra-violet end of the spectrum.

Electron microscopes have evolved at an explosive pace compared with the optical microscope's 300 years of development since Anton Leeuwenhoek observed his first microbe. In 1926, Busch published theoretical calculations showing how an electrostatic field could focus an electron beam, thereby laying the foundations for geometrical optics. Bruche and Johansson built the first electrostatic instrument in 1932. Then also in 1932, Knoll and Ruska accelerated the pace by publishing the first experimental results obtained with an electron microscope employing magnetic lenses. The industry hasn't looked back since.

### RESOLUTION

Diffraction effects, caused by mutual interference between phase-related wave-fronts, ultimately limit a microscope's ability to distinguish separate points on any object being viewed. Diffraction is directly related to wavelength, as Abbe's equation:  $d = 0.62\lambda/n \sin \alpha$  demonstrates. This equation shows how resolution  $d$  improves as wavelength  $\lambda$  shrinks. The electron microscope's "light source," being several orders of magnitude shorter than visible light, improves resolution dramatically.

A good optical microscope can distinguish points separated by a distance roughly equal to half a wavelength. For a 4000 Angstrom light source, this works out to about 2000 Angstroms, or 0.0002 centimeters. The naked eye can resolve to about 0.02 centimeters. Compromises in the electron microscope between diffraction and spherical aberration effects yield an optimum resolution of about 8 Angstroms, or  $8 \times 10^{-10}$  centimeters. (Spherical aberration is minimized by a small aperture, diffraction is minimized by a large one.)

### ALSO IN THIS ISSUE

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## LENS COMPARISONS

Electron microscopes are virtually identical in fundamental principle to their optical counterparts, as the ray diagram, Fig. 1, shows. Lenses are arranged in conventional condenser, objective, and projection configurations for transmission microscopes of both types. Even so, the differences in practice are more marked than the similarities in principle.

An optical microscope's day-to-day performance depends primarily upon such basic factors as dimensional stability of lenses and structure. By contrast, an electron microscope's day-to-day resolving power is a much more delicate matter and depends upon far less tangible parameters than glass and metal configurations. Magnetic field geometry establishes the electron microscope's working efficiency, making magnetic stability, hence lens current constancy, analogous to the optical microscope's lens and structural stability.

The electron microscope exemplifies modern instrument trends towards increasing use of electrical rather than physical "dimensions." Sophisticated designs in many fields are using electronics to achieve new functions and improved accuracy. Examples occur in caesium cell frequency standards, particle accelerators, mass spectrometers, and doubtless many more. One commercial analytical balance even uses the pull of an accurately calibrated solenoid for fractional milligram resolution of sample weight. In all of these instances, utmost electrical stability is reflected directly in the instrument's usefulness. Because Analog Devices is participating with instrument manufacturers in the development of advanced electronic instrumentation, we've obtained RCA's permission to present the design and circuit details of the current source used in their new Model EMU4 microscope.

## CURRENT SOURCE

Until recently, electronic circuitry imposed a final limitation on the routine performance of commercial electron micro-

scopes. However, chopper stabilized amplifiers with better than  $0.5\mu\text{V}/^\circ\text{C}$  drift, and ultra-stable Zener diodes with comparable specifications, have elevated the performance of electronic circuits to at least the equal of the instrument's physical hardware. Oscilloscope users are aware that the focus knob on their instrument can stand a good deal of twiddling before the trace loses its definition. This slackness of focus is certainly not duplicated by an electron microscope capable of enlarging objects by 200,000 or more. At highest magnification levels a 10 PPM lens current drift drastically degrades resolving power and defeats the purpose of the original \$50,000 to \$500,000 expenditure.

## BASIC CIRCUIT

The basic current source circuit that supplies lens current for the EMU4 microscope is analyzed in Fig. 2. The analysis shows how errors due to drift and noise reflect in output current deviations.

Compared with other current sources<sup>1</sup>, this circuit has the advantage that its main amplifier operates with one input terminal grounded, simultaneously eliminating common mode errors and enabling a chopper stabilized operational amplifier to establish the basic stability. Too, the current booster circuit can be a simple emitter follower or Darlington pair enclosed within the feedback loop. Equivalent booster drift, when reflected back to the reference source, is then reduced by the chopper amplifier's  $10^8$  gain to negligible proportions compared with the chopper amplifier's own drift.

Penalties paid for these simplifications include a somewhat increased overall drift gain, and the need for a floating load. Since the lens coil is isolated from ground anyway, the floating load condition imposes no extra problems. Similarly, chopper amplifiers can be selected for ample margin of stability over the increased drift gain.

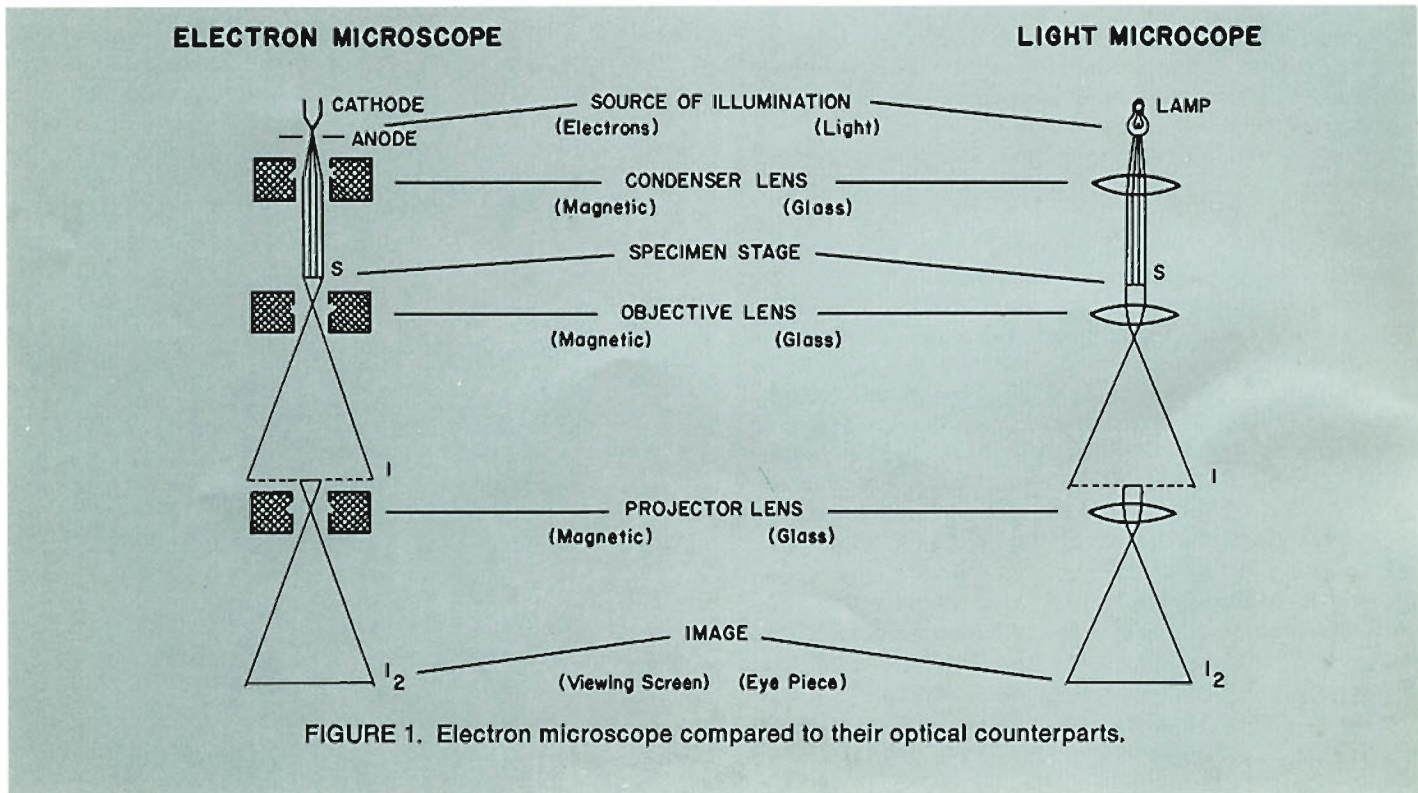


FIGURE 1. Electron microscope compared to their optical counterparts.

<sup>1</sup>Current Sources, by Bill Miller, Analog Dialogue No. 1 (Available on request from Analog Devices, Inc.)

**ACTUAL CIRCUIT**

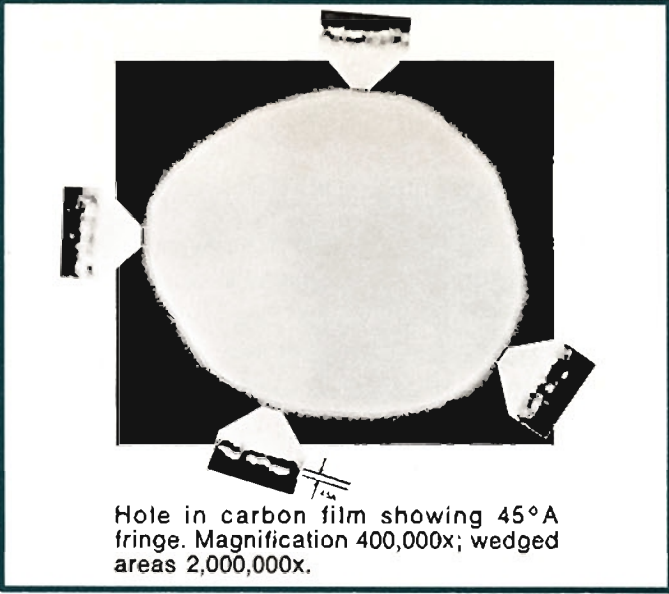
Owing to the relative wide range of voltage needed by the lens coil, the booster amplifier is not a simple emitter follower, but increases output voltage from the chopper amplifier's — 6 volts to roughly 50 volts at maximum coil current. Additionally, the Darlington configuration raises output current to nearly 300 milliamps.

The design procedure for such a circuit is first of all to establish a theoretical paper "Bode" design, then produce a simple prototype for error measurement and diagnosis. The next step is to add voltage and current amplifiers to the circuit, introduce long leads and control switches, then check for frequency and phase stability, as well as dc stability. Frequency-stabilizing capacitors are introduced to ensure operating stability under all operating conditions, including switching transients.

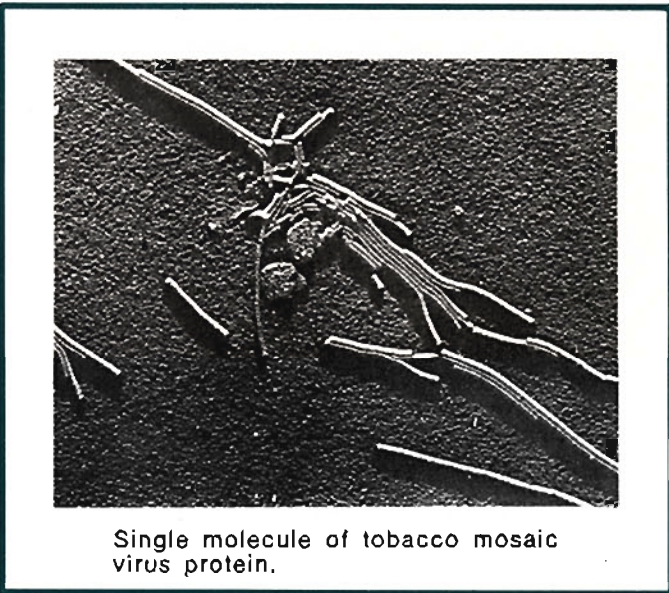
Feedback resistors should be low noise types (they should be checked by the "current noise" method), and must afford stability commensurate with the circuit's final specifications. Operating the resistors at low dissipation levels, and in a uniform temperature environment, are important design factors.

Temperature compensated Zeners operated in the proper temperature environment (but without constant temperature oven), provide about 1 ppm short-term stability. The amplifier drift, as shown in Fig. 3, can be well below this value for the 5°C temperature range.

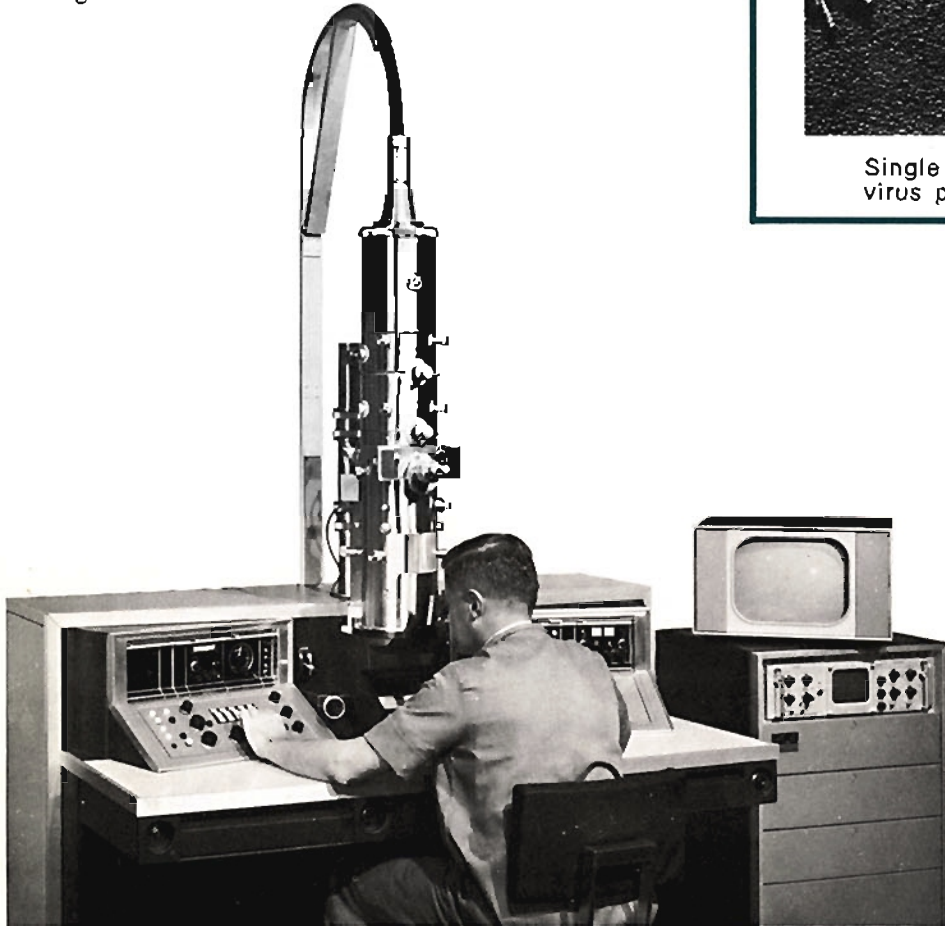
Overall performance of the circuit meets the worst-case current drift specifications of 2.5 parts-per-million deviation over the one-minute exposure interval needed by some image photographic processes. Generally speaking, a one-second exposure is adequate, making the 2.5 ppm drift specification ample. Long-term stability of the lens current output is about 10 ppm/hour, and its state-of-the-art repeatability enables operators to start the morning's work where they left off the previous evening.



Hole in carbon film showing 45°A fringe. Magnification 400,000x; wedged areas 2,000,000x.



Single molecule of tobacco mosaic virus protein.

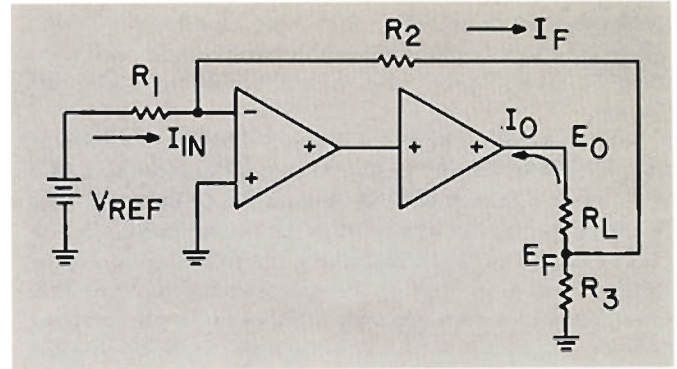


Heralding a new generation in electron microscopes — The RCA EMU-4.

FIGURE 2. ANALYSIS OF CURRENT SOURCE'S OUTPUT VOLTAGE AND CURRENT

Analysis for determining current source's output current and output voltage in terms of Reference voltage and circuit values.

The analysis is based on the equality between input current  $I_{in}$  and feedback current  $I_f$  that prevails during steady-state operation.



### VOLTAGE OUTPUT

The summing junction potential is very nearly zero (virtual ground principle), so that input current  $I_{in} = V_{ref}/R_1$ , and

$$\text{feedback current } I_f = E_o / \left( R_2 + \frac{R_3 R_L}{R_3 + R_L} \right)$$

$$\text{Since } I_{in} = I_f, V_{ref}/R_1 = E_o / \left( R_2 + \frac{R_3 R_L}{R_3 + R_L} \right)$$

The artificial thevenin voltage, designated  $E_f$ , can also be expressed in terms of amplifier output voltage by:

$$E_f = E_o \left( \frac{R_3}{R_L + R_3} \right)$$

This expression for  $E_f$  can be substituted in the identity  $I_{in} = I_f$ , enabling the amplifier output voltage to be defined in terms of reference voltage  $V_{ref}$ .

$$\text{Accordingly, } V_{ref}/R_1 = \left( \frac{E_o R_3}{R_3 + R_L} \right) \div \left( R_2 + \frac{R_3 R_L}{R_3 + R_L} \right)$$

Output voltage expressed explicitly becomes

$$E_o = \frac{V_{ref}}{R_1} \left( \frac{R_2 R_3 + R_2 R_L + R_3 R_L}{R_3} \right)$$

Conventionally, amplifier output voltage is expressed in terms of the ratio between feedback resistor  $R_2$  and input resistor  $R_1$ . The current source's output voltage can be manipulated into this conventional form too.

$$\begin{aligned} \text{Thus, } E_o &= V_{ref} \left( \frac{R_2}{R_1} \right) \left( 1 + \frac{R_L}{R_3} + \frac{R_L}{R_2} \right) \\ &= V_{ref} \left( \frac{R_2}{R_1} \right) \left( 1 + R_L / \text{Parallel combination of } R_2 \text{ and } R_3 \right) \end{aligned}$$

### CURRENT OUTPUT

The amplifier's output current flows through load resistor  $R_L$  then divides between the parallel combination of  $R_3$  and  $R_2$ , both of which, to all intents and purposes, have one terminal

$$\text{grounded. Consequently, } I_o = E_o / \left( R_L + \frac{R_2 R_3}{R_2 + R_3} \right)$$

Substituting the expression for output voltage in the above equation gives

$$I_o = V_{ref} \left( \frac{R_2}{R_1} \right) \left( 1 + \frac{R_L}{R_3} + \frac{R_L}{R_2} \right) \div \left( R_L + \frac{R_2 R_3}{R_2 + R_3} \right)$$

which can be simplified by algebraic juggling to

$$I_o \cong \frac{V_{ref}}{R_1} \left( 1 + \frac{R_2}{R_3} \right) \cong \frac{V_{ref}}{R_1} \left( \frac{R_2}{R_3} \right) \text{ for } R_2 \gg R_3$$

This expression shows that so long as  $R_2$  is much larger than  $R_3$ , output current varies linearly with the values of  $R_2$ . Further, output current is independent of the actual values of load resistance  $R_L$ , provided that this resistance is small enough to allow the amplifier to operate within its normal output voltage range.

### ERROR CONSIDERATIONS

Output current deviations are caused by voltage and current drifts within the amplifier, and also by the amplifier's voltage and current noise. An analysis of the output current deviations produced by these effects proceeds from the assumption that noise and drift errors can be referred to the input circuit as small deviations in reference voltage,  $V_{ref}$ . In effect, the amplifier's noise and drift may be represented as small voltage generators acting in series with the reference source, and having net voltage  $\Delta V_{ref}$  producing voltage errors amounting to  $\Delta V_{ref}$ .

Load current variations caused by input voltage variations  $\Delta V_{ref}$  can be found by differentiating the output current expression with respect to reference or input voltage.

$$\begin{aligned} \text{Thus, } \frac{dI_o}{dV_{ref}} &= \frac{d}{dV_{ref}} \left[ \frac{V_{ref}}{R_1} \left( 1 + \frac{R_2}{R_3} \right) \right] \\ &= \frac{1}{R_1} \left( 1 + \frac{R_2}{R_3} \right) = \frac{I_o}{V_{ref}} \end{aligned}$$

$$\text{And because } \frac{dI_o}{dV_{ref}} \cong \frac{\Delta I_o}{\Delta V_{ref}}, \Delta I_o = \Delta V_{ref} \left( \frac{I_o}{V_{ref}} \right)$$

Or in other words, the proportionate current change is equal

$$\text{to the proportionate input voltage change: } \frac{\Delta I_o}{I_o} = \frac{\Delta V_{ref}}{V_{ref}}$$

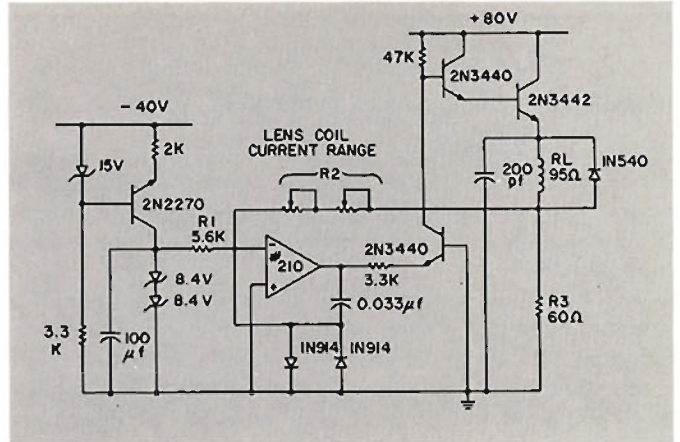
This means that a 10 part-per-million increase in reference voltage creates a 10 part-per-million increase in output current. (A numerical error analysis for a practical current source is worked out in Fig. 3)

**FIGURE 3. CURRENT SOURCE CIRCUIT SUPPLIES HIGHLY STABLE LENS CURRENT FOR RCA'S MODEL EMU4 ELECTRON MICROSCOPE.**

Short term stability of better than 2.5 ppm is derived basically from Model 210 chopper stabilized op amp. Booster circuit raises output voltage and current to as high as 50 volt 300mA from the amplifier's basic  $\pm 10$  volt, 20mA.

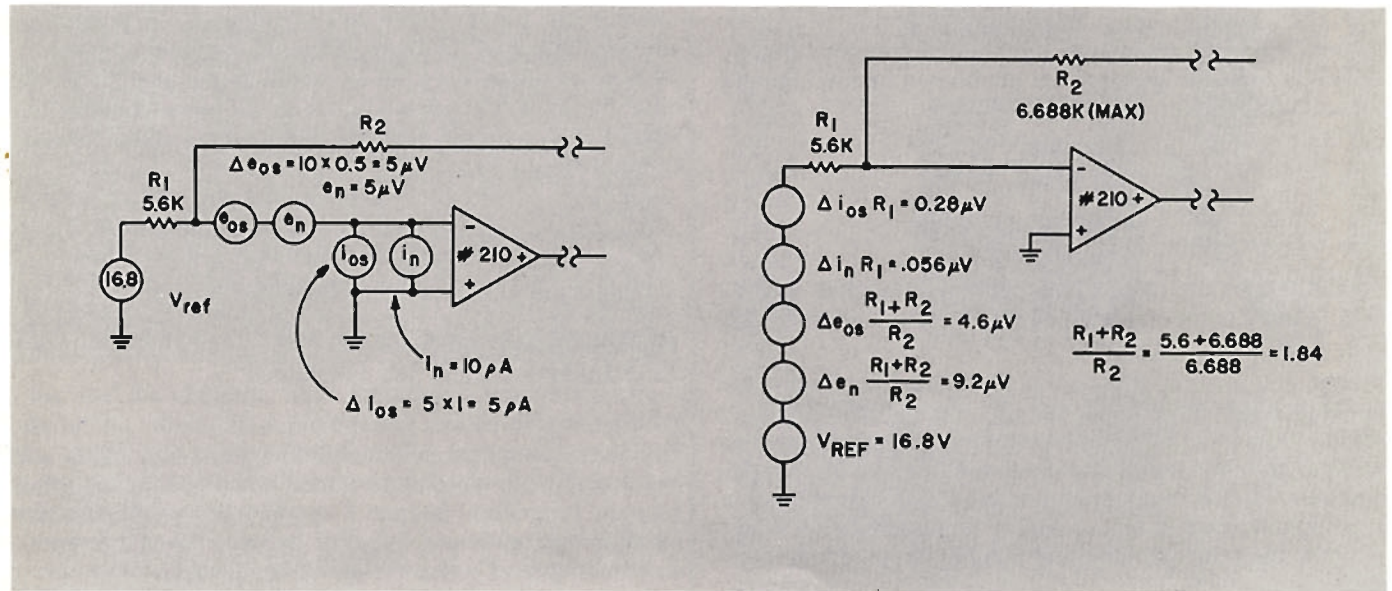
**CALCULATION OF ERRORS**

Numerical values for output current error are determined for a given temperature range from the worst-case drift and noise figures quoted in the amplifier's specification tables. For Model 210 these are  $0.5\mu\text{V}/^\circ\text{C}$  and  $1\text{pA}/^\circ\text{C}$  drift, and  $5\mu\text{V}$  and  $10\text{pA}$  peak-to-peak noise, respectively. Although noise is practically independent of temperature for small environmental ranges, the voltage and current offsets caused by a  $5^\circ\text{C}$  variation around room temperature amount to  $5 \times 0.5 = 2.5\mu\text{V}$  and  $5 \times 1 = 5\text{pA}$ . These values for noise and offset have been inserted into the equivalent current source circuit below (left). When referred to the reference source (instead of the actual amplifier terminals), the noise and offset values are modified by the potentiometric effect of circuit components, leading to the equivalent error voltage generators as shown in



the illustration below (right).

The sum of voltage and current offset errors, and also the sum of voltage and current noise errors, both as proportions of the reference voltage, give a measure of the output current error directly. The calculations are performed here for the actual electron microscope lens current supply (top).



**DRIFT ERRORS**

Sum of voltage and current offset errors due to  $5^\circ\text{C}$  temperature variation, are referred to input as: —

$$\Delta e_{os} \left( \frac{R_1 + R_2}{R_2} \right) + \Delta i_{os} R_1$$

$$\cong 4.6\mu\text{V} + 0.28\mu\text{V} \cong 4.63\mu\text{V}$$

Offset error as a proportion of circuit input error  $\Delta V_{ref}/V_{ref}$   
 $= 4.63 \times 10^{-6} / 16.8 = 0.28$  parts/million

**NOISE ERRORS**

Sum of voltage and current noise errors in the frequency range DC to 1Hz, are referred to the input as: —

$$e_n \left( \frac{R_1 + R_2}{R_2} \right) + i_{os} R_1$$

$$\cong 9.2\mu\text{V} + 0.56\mu\text{V} \cong 9.26\mu\text{V}$$

Noise error as a proportion of circuit input error  $\Delta V_{ref}/V_{ref}$   
 $= 9.26 \times 10^{-6} / 16.8 = 0.55\text{ppm}$

**CONCLUSION**

Drift errors for the  $5^\circ\text{C}$  temperature range, happen to be half the error caused by noise for the amplifier. Ambient temperature varies relatively slowly, so that for short-term exposure periods of a few minutes, noise will be the predominant cause

of lens current deviation.

The 0.83ppm amplifier error combined with the 1ppm Zener diode error provide better than the 2.5ppm design goal.

# Letters to the Editor

## MISUSE OF EXACT GAIN FORMULA

Dear Sir:

Somewhat over a year ago I noticed a continuing increase in the misuse of the exact gain formula in operational amplifiers. Attempts to bring this to the readers' attention were met with indifference.

In your June issue of ANALOG DIALOGUE the error has again been repeated. I refer on page 13 to the example using data from Figure 11. The author computes an amplification of -99, or 1% from ideal. Actually if he had taken into account the 90 degree phase lag,  $\Theta$ , of A at the selected frequency of 100 Hz on the 6 dB slope in figure 11, the true G is  $-100/\sqrt{1+(0.0101)^2}$  or 0.005% from ideal rather than 1%. Naturally, the effect of quadrature is less for departures of phase angle from 90 degrees. The standard formula (1) in the enclosed letter article can be modified into a working formula such as

$$G = -\frac{Z_f}{Z_i} \left[ \frac{1}{1 + \frac{1}{A} \cdot (\cos \Theta - i \sin \Theta) \cdot (1 + Z_f/Z_i)} \right]$$

I should like to see the enclosed letter article published in your journal. It has been approved for publication by the National Bureau of Standards.

Louis A. Marzetta

(Letter referenced above)

## INCORRECT USAGE OF EXACT CLOSED-LOOP VOLTAGE GAIN FORMULA IN OPERATIONAL AMPLIFIERS

No one who has been interested in operational amplifiers during the past year could avoid acquiring a voluminous folder of articles, booklets and reports on the subject. Most of the promotional literature contains tutorial material that is, for the most part, helpful. But a serious error in the application of a basic formula seems to have crept into so much of it that I submit this letter in the hope of bringing it to the reader's attention.

The usual format is to introduce the reader to the principles of operation by writing input and output loop expressions for the amplifier and its external impedances. Considering an ideal amplifier with infinite gain and bandwidth, the expression for closed-loop voltage amplification is given, as

$$\frac{V_{out}}{V_{in}} = -$$

$\frac{Z_{feedback}}{Z_{input}}$ . The next step is to define an operational amplifier with special emphasis on the 6 dB octave roll-off characteristic of amplification. Unfortunately, most of the authors did not keep this slope in mind when they copied the following expression, the exact closed-loop voltage amplification formula.

$$\frac{V_{out}}{V_{in}} = - \left( \frac{Z_f}{Z_i} \right) \frac{1}{1 + \frac{1}{A} \left( 1 + \frac{Z_f}{Z_i} \right)} \quad (1)$$

A is identified as the open-loop amplification, while  $Z_i$  and  $Z_f$  are the input and feedback impedances. The second bracketed group of terms is generally called the error factor. It expresses the degree of departure of the actual amplification from the ideal simple ratio of  $Z_f$  and  $Z_i$ . In a number of articles this formula has been augmented with numerical examples, convenient monograms and tables — almost all of which are in error for exact gain computation.

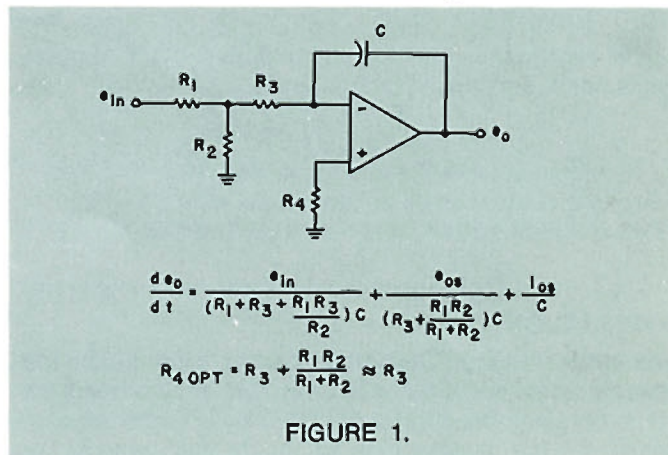
The formula is explicit for flat frequency response situations but it can lead to serious errors in calculating the precise amplification for operational amplifiers, the reason being that associated with an amplitude roll-off is a phase shift of 90 degrees for a 6 dB slope. At first glance one might underestimate the importance of the phase angle of A in the formula and expect to compute a sizeable error factor in the region of reduced amplification. In reality such an error factor involves a quadrature expression; therefore, the actual magnitude error may be very small. For example, calculate the error factor at the point where A has dropped to 100 on a 6 dB slope, in an operational amplifier with equal values for  $Z_i$  and  $Z_f$ . Neglecting the phase of A in the formula, the device would be thought to have an error factor of 2%. However, using the same formula, but treating A as a phasor quantity, the true error factor amounts to only 0.02%.

Louis A. Marzetta  
Measurement Engineering Division  
National Bureau of Standards  
Washington, D.C.

## INDIRECT METHOD OF ACHIEVING LARGE RC PRODUCT

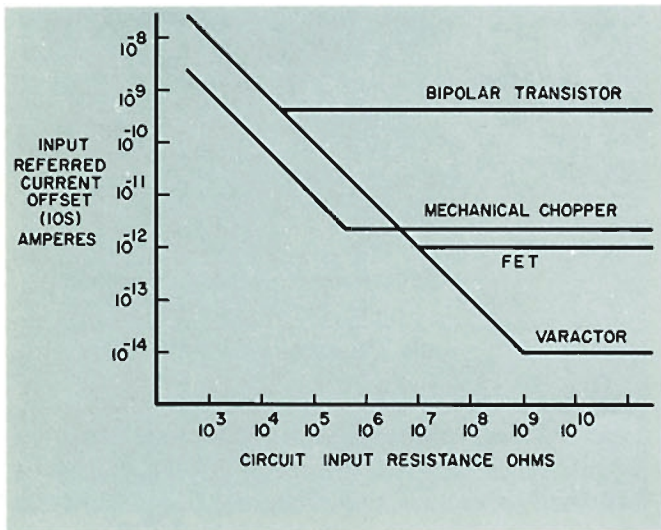
Dear Sir:

Ray Stata's article on Operational Integrators was timely and welcome. I wish, however, to extend his analysis to large time constant integrators. Because it is physically or economically impossible to achieve a large RC product directly, a Tee is commonly used to obtain the desired RC product.



As in the earlier analysis, C should be as large as possible. Assuming  $R_3$  cannot then be made large enough to achieve the necessary RC product, it is wise to choose  $R_3$  as large as possible, in order to reduce voltage offset errors. No particular advantage is gained by making  $R_1$  and  $R_2$  large, as their parallel value will generally be much smaller than  $R_3$  anyway.

Selection of a suitable amplifier can be made from the log-log plot of input referred current offset ( $I_{os}$ ) as a function of circuit input resistance. This plot is based on short term stability ( $\frac{1}{2}$  hour). Longer term stability (day) is about five times worse and temperature stability (per  $^{\circ}\text{C}$ ) is about two times worse.



Notice that for integrator input resistance values of ten megohms or less, about equal performance can be expected from either a FET or a varactor amplifier. For input resistance values less than several megohms, best performance is offered by mechanical chopper amplifiers.

In a laboratory environment only the time dependent offsets are important. At input resistances below the "break point" the voltage offset is the limiting offset, and above, the current offset. It is apparent that an open input integrator is the most practical means of measuring current offset.

A similar such graph can be made in terms of input referred voltage offset vs. circuit input resistance this form being especially suited to selection of scaling amplifiers or of scaling impedance levels.

Jim Davis (Graduate Student)  
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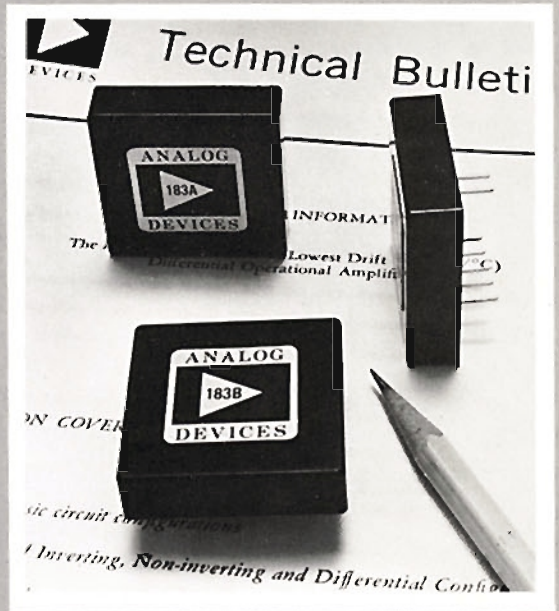
**Editors Note:** One should be aware that the voltage offset error using this method will be approximately  $1 + \frac{R_1}{R_2}$  times as large as it would be with the equivalent single resistor connected to  $e_{in}$ .

## the New Model 183 CHOPPERLESS Operational Amplifier

1.5 $\mu\text{V}/^{\circ}\text{C}$       \$65

3 $\mu\text{V}/^{\circ}\text{C}$       \$45

5 $\mu\text{V}/^{\circ}\text{C}$       \$35



### SPECIFICATIONS

Open Loop Gain	200,000 min.		
Rated Output	$\pm 10\text{V}$ @ 5mA		
Unity Gain Response	0.5MHz		
Full Power Response	5kHz		
Current Drift (Differential)	0.05nA/ $^{\circ}\text{C}$ max.*		
CMRR ( $\pm 10\text{V}$ )	100,000		
Warm Up Drift	20 $\mu\text{V}$ (20 minutes)		
Noise (dc to 1Hz)	1 $\mu\text{V}$ , peak to peak		
Offset Voltage	Model J	Model K	Model L
@ 25 $^{\circ}\text{C}$ , max.	3mV	.5mV	.5mV
vs. temp., max.*	5 $\mu\text{V}/^{\circ}\text{C}$	3 $\mu\text{V}/^{\circ}\text{C}$	1.5 $\mu\text{V}/^{\circ}\text{C}$
Price (1-9)	\$35	\$45	\$65

\*temperature range 10-60 $^{\circ}\text{C}$ ; for -25 -85 $^{\circ}\text{C}$  add \$5 and specify 183A/B/C

# Equivalent Circuits for Operational Amplifier Drift and Noise

by Richard Conant, Application Engineer, Analog Devices, Inc.

We recently received a letter arguing very logically that the equivalent circuit (Figure 1A) which we give in our application notes for drift and noise is incorrect. Instead the circuit of Figure 1B was put forth with good reasons as a more accurate representation. Since in some applications these two circuits give different results, one must be wrong assuming that we apply the same set of published drift specifications. The question then arises as to which circuit is correct and how do we decide which one is correct.

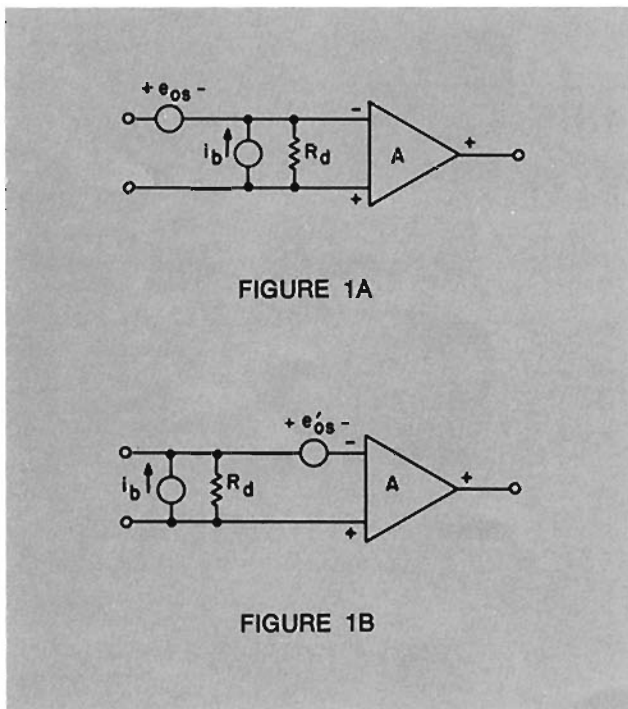


FIGURE 1A

FIGURE 1B

This controversy has come up a number of times in our applications department, since there are a number of equivalent circuits now published in the op amp literature which can in essence be reduced to either the circuit of Figure 1A or Figure 1B. Basically, the only difference is whether the equivalent voltage generator is located on the input or output side of the open loop input impedance  $R_d$ .

First let us compare the closed loop errors for these two circuits to see what difference we get. To simplify our discussion we shall consider only offset and drift errors but completely analogous results would apply to noise errors. Additionally,

we shall look at only the single ended case where the plus input is grounded but the same arguments would apply to differential or noninverting circuits as well. Figures 2A and 2B show the closed loop errors for the two equivalent circuits for the inverting amplifier configuration where  $e_{os}$  is the equivalent offset voltage generator,  $i_b$  is the equivalent bias current generator,  $R_d$  is the open loop input impedance and  $A$  is the open loop gain. The primed quantities for  $e_{os}$  and  $i_b$  indicate that the numerical values for the generators may differ due to the different equivalent circuit configurations.

We can better compare Equations 1 and 2 if we rewrite Equation 2 as follows:

$$e_o = \frac{e_i \left( \frac{R_f}{R_i} \right) + e'_{os} \left( \frac{R_i + R_f}{R_i} \right) - \left( i_b' + \frac{e'_{os}}{R_d} \right) R_f}{1 + 1/A + R_f/AR_d + R_f/AR_i}$$

Now we see that the only difference between Equations 1 and 2 is the term  $e'_{os}/R_d$ . That is the voltage generator in Figure 2B will generate an effective bias current through the finite impedance  $R_d$  which adds to the bias current  $i_b'$ . As a practical matter the only time this difference would show up is when the term  $e'_{os}/R_d$  is comparable to or larger than  $i_b'$ . For most transistor or FET input operational amplifiers, bias current ( $i_b'$ ) is much larger than voltage offset divided by  $R_d$ , that is,  $e'_{os}/R_d$ . However for chopper stabilized amplifiers  $R_d$  and  $i_b'$  are sufficiently small that the effect of  $e'_{os}$  cannot be ignored. Additionally, the difference between the two equivalent circuits becomes very striking for high frequency noise errors since shunt capacitance across  $R_d$  reduces this impedance at high frequencies.

Another almost philosophical point can be made in comparing the circuits of Figures 2A and 2B. That is offset errors for the equivalent circuit in Figure 2A can be predicted without any knowledge of the value for input impedance  $R_d$  (except for the second order gain reduction due to reduced loop gain). The primary effect of finite



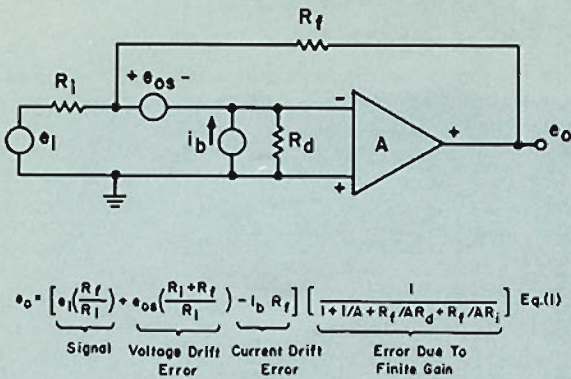


FIGURE 2A. Equivalent circuit and error equation.

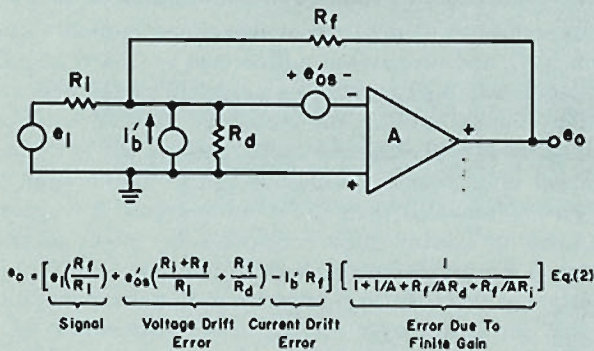


FIGURE 2B. Equivalent circuit and error equation.

$R_d$  is to increase gain errors and to reduce loop gain. One could then theoretically eliminate errors due to  $R_d$  by adding enough open loop gain at the output of the amplifier. On the other hand offset errors for the circuit of Figure 2B depend on the value for  $R_d$ . While offset voltage and bias current are usually specified as maximum values,  $R_d$  is given as a typical value since it is difficult to measure this parameter with better than order of magnitude accuracy.

We still have not answered the question of which is the correct equivalent circuit and why. The subtle point here is not which circuit is right or wrong but rather which circuit corresponds to the published specifications on the data sheet. That is either circuit can be correct and one can translate back and forth between the equivalent circuits by solving the equations:

$$e_{os} = e'_{os}$$

$$i_b = i'_b + e'_{os}/R_d$$

The question of which set of parameters to use is resolved by examining the test circuit with which the offset and/or noise parameters are measured. Figure 3 shows the test circuit for measuring offset, drift and noise used by most operational amplifier manufacturers.

To measure offset voltage, switch  $S_1$  is shorted and  $R_f$  is selected so that the term  $i_b R_f$  is small compared with  $e_{os} (R_i + R_f)/R_i$  in Equation 1. The output voltage is measured and the offset voltage is then computed from the equation  $e_o = e_{os} (R_i + R_f)/R_i$ .

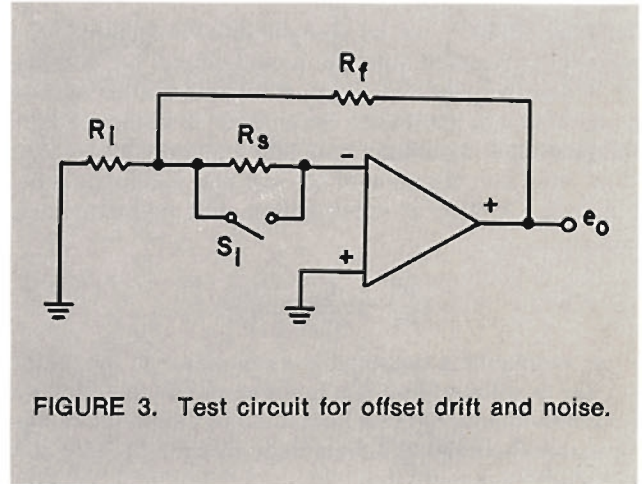


FIGURE 3. Test circuit for offset drift and noise.

One way to measure bias current would be to increase the value of  $R_f$  (still assuming  $S_1$  is shorted) until the term  $i_b R_f$  in Equation 1 is large compared to  $e_{os}(R_i + R_f)/R_i$ , to measure  $e_o$  and then to compute for  $i_b$  from  $e_o = i_b R_f$ . It turns out to be more practical to use the same values for  $R_i$  and  $R_f$  as for measuring  $e_{os}$  and to switch in a resistor  $R_s$  whose value is such that  $i_b R_s \gg e_{os}$ . Then you measure  $e_o$  and solve

$$e_o = (i_b R_s)(R_i + R_f)/R_i$$

In either case notice that the value measured for  $i_b$  corresponds to the equivalent circuit of Figure 2A. That is, compared to Figure 2B, this measurement includes the contributions of bias current drift  $i'_b$  and offset voltage  $e'_{os}/R_d$ . Or in other words,  $i_b = i'_b + e'_{os}/R_d$ . In conclusion, not only does the equivalent circuit of Figure 2A correspond to test circuits and therefore the published specifications but also this equivalent circuit makes more sense from the point of view of solving applications problems. That is, offset and noise errors are clearly divided into two categories; offset voltage errors ( $e_{os}$ ) which are independent of the values chosen for the external feedback components, and bias current errors which are directly proportional to the value chosen for the external feedback components. Additionally offset errors can be predicted without any knowledge of the open loop input impedance  $R_d$ .

# HOW TO SELECT R & C FOR MINIMUM INTEGRATOR DRIFT

by Richard Conant, Application Engineer, Analog Devices, Inc.

The transfer function for the integrator circuit in Figure 1 is  $e_o/e_i = 1/RCS$  where R is the input resistance, C is the feedback capacitance and 1/S is the LaPlace operator denoting integration. This article will show that for a given amplifier, output drift rate due to offset voltage ( $e_{os}$ ) and bias current ( $i_b$ ) can be minimized by a proper selection of R and C. For a more complete discussion of integrators, the reader is referred to an article entitled "Operational Integrators" in Volume I, number 1, of Analog Dialogue.

The term  $(1/RC)$  can be thought of as the "gain" of the integrator in terms of volts per second output, per volt input. For a desired value of "gain" R and C can be arbitrarily chosen although the lower limit on R is often set by the minimum input impedance required.

Offset referred to the input of an integrator is equal to the sum of  $e_{os}$  and  $i_b R$ . This corresponds to a drift rate at the output equal to

$$\frac{de_o}{dt} = \frac{e_{os}}{RC} + \frac{i_{os}}{C}$$

If this expression is examined it can be seen that the contribution due to voltage offset is a function of gain  $(1/RC)$  while the contribution due to bias current can be minimized by selecting a large C. In fact if C were large enough, this term would be negligible compared to  $e_{os}/RC$ .

Figure 2 is a plot of output drift rate vs. capacitance for various gains for several different operational amplifiers given in Figure 3. Here it is assumed that the initial offsets have been balanced and that we are examining the output drift rate only as a function of temperature. The curves are plotted on a log scale in a manner analogous to a Bode diagram.

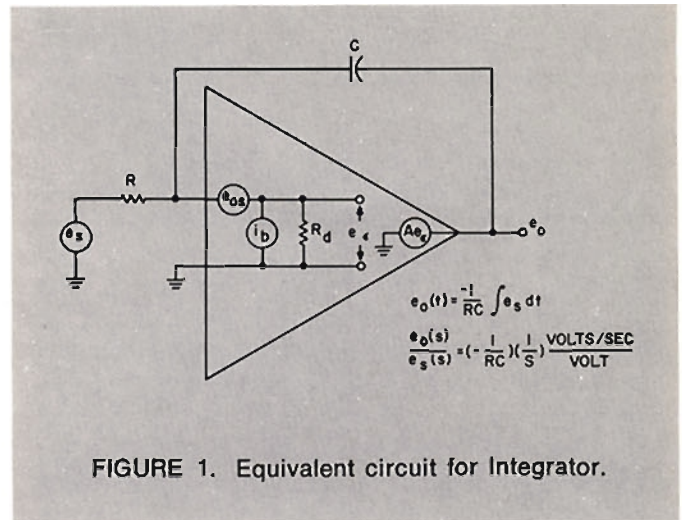


FIGURE 1. Equivalent circuit for Integrator.

For low values of C, drift rate is due almost entirely to bias current  $i_b/C$ , and decreases at 6dB/octave as C increases. For high values of C, drift rate reaches a constant value determined by offset voltage,  $e_{os}/RC$ . The breakpoint between these two conditions is set by the capacitance for which  $e_{os}/RC = i_b/C$ .

A number of interesting conclusions can be drawn from the plot. First, to minimize drift rate it can be argued that an optimum value for C occurs at the breakpoint. For larger values of C there is no net reduction in drift rate but only the disadvantages of a smaller input impedance, R, and a more expensive and physically larger size capacitor. (Remember RC is assumed to be constant.) It may not be practical to use the optimum value of C for a given amplifier because of two factors;

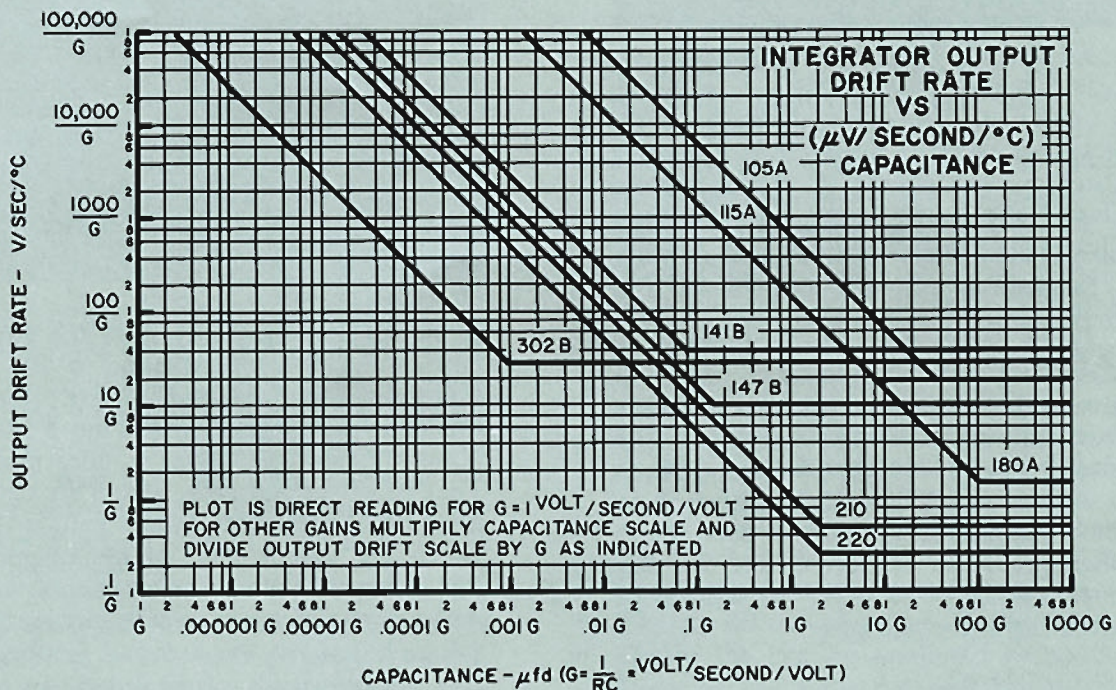


FIGURE 2. Integrator Output Drift Rate vs.  $(\mu V/\text{Second}/^\circ C)$  capacitance.

1. The required value for C may be so large and expensive that it would be wiser to choose an amplifier with lower bias current for a required output drift rate.
2. The resulting input impedance for the optimum value of C would be so low that an amplifier with lower bias current should be selected to keep the input impedance up.

Note that the plots in Figure 2 can be made to apply for any value of gain (that is  $G = 1/RC$ ). Another general point can be made in comparing the plots at different gains. That is at higher gains ( $G = 10$ ), offset voltage tends to be more important in choosing an amplifier for a given drift rate whereas for lower gains ( $G = .1$ ) bias tends to be more important in choosing an amplifier for lowest drift rate. It is also quite apparent that the best amplifier choice in terms of output drift rate vs. price depends very much on the required gain and the capacitance to be used.

	Model and Type	$\frac{\Delta i_b}{\Delta t}$ (pa/°C)	$\frac{\Delta e_{os}}{\Delta t}$ ( $\mu$ V/°C)	\$ (1-9)
105A	Bipolar Transistor	700	20	16.00
115A	Bipolar Transistor	150	20	18.00
180A	Bipolar Transistor	150	1.5	80.00
141B	Field Effect Transistor	3	40	30.00
147B	Field Effect Transistor	1.5	10	115.00
302B	Varactor Bridge	.025	30	135.00
210	Chopper Stabilized	1	0.5	157.00
220	Chopper Stabilized	0.5	0.25	165.00

Figure 3. Comparison of Amplifier Types for Integrator Use.

## Solid State OP-AMP Measures Kilovolts With 0.05% Accuracy

By T. P. Kohler and E. H. Hudspeth, General Electric Co., Syracuse, N.Y.

Current drawn by an electrostatic voltmeter in high-voltage tests lowers the measured voltage and consequently limits the accuracy of the measurement. The 2% accuracy of a conventional electrostatic voltmeter causes a significant 200-volt error in a 10-kilovolt measurement. A charge transfer circuit that precisely divides the high voltage and protects the divided voltage from current drain during test keeps measurement error within .05%.

The voltage division is accomplished by transferring the charge on a small, high-voltage capacitor to a large, low-voltage capacitor. The large capacitor is the feedback capacitor of an operational amplifier. Any current drawn by the voltmeter during test is replaced on the capacitor by the amplifier.

The voltage to be measured,  $V_1$ , charges capacitor  $C_1$ , when the switch  $S_1$ , is moved to position A. Corona losses are avoided with rounded contacts on both the swinger and position A. By using a glass, oil-filled capacitor for  $C_1$ , the error introduced by dielectric losses is eliminated.

To measure the voltage under load conditions switch  $S_2$  is moved to the load position. The value of  $R_1$  is determined by a calculation when the desired load current is known.

When  $S_1$  is moved to position B, the charge on  $C_1$  is completely discharged into the input of the operational amplifier. The resistor  $R_2$ , is placed in series with amplifier to slow the current flow and insure response of the amplifier. The output current of the operational amplifier — equal to the input current — accumulates on  $C_2$  and charges it to a voltage,  $V_2$ , that is related to  $V_1$  by

$$V_1 = V_2 \frac{C_2}{C_1}$$

The voltage,  $V_2$ , is measured by a digital voltmeter. Since the discharge of  $C_2$  during the measurement is replenished by the operational amplifier,  $V_2$  remains permanently accurate. The voltage  $V_2$  is removed from the capacitor by closing switch  $S_3$ , thus allowing the engineer to make further measurements. Since the  $C_2/C_1$  ratio is involved in the calculation of  $V_1$ , it is precisely determined by placing a standard voltage on  $C_1$ , and measuring the voltage on  $C_2$  after current transfer.

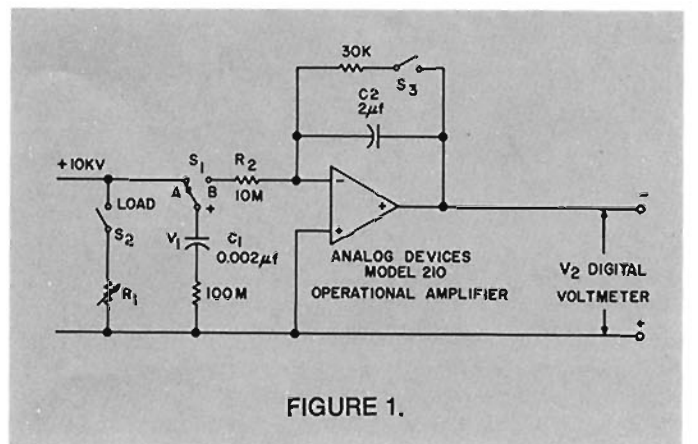


FIGURE 1.

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# Ideal rectifier uses equal-value resistors

ED Dec. 20, 1967 {Edited}  
Follow to Previous "Ideal Rectifier" Article

By Allan G. Lloyd, Project Engineer, Avion Electronics, Inc., Paramus, N.J.

Operational amplifiers can be combined with diodes and resistors to perform nearly ideal rectification, satisfying  $e_o = |e_i|$ . The diodes are situated within the feedback loop, so that the diode forward voltage drop is reduced at the output by the feedback factor. Figure 1a shows a commonly used full-wave rectifying circuit which gives a positive output,  $e_o$ , for bipolar input at  $e_i$ . This circuit has the following disadvantages:

- Output  $e_o$  due to plus input at  $e_i$  is obtained by bucking out the plus current from  $R_5$  with the negative (rectified) current through  $R_3$ . The tolerance of this difference voltage can be three times resistor tolerance, since it is obtained by subtraction.
- Non-zero input impedance. If a summing junction is required at the input node, another operational amplifier is required.
- Increased drive requirements. The input at  $e_i$  drives two amplifiers in parallel.
- Unequal resistor values.  $R_3$  is half of  $R_5$ .

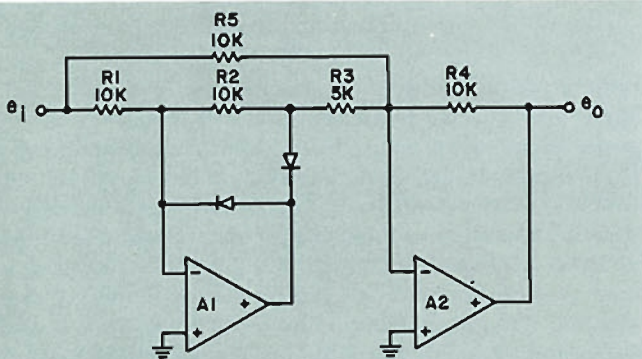


FIGURE 1a.

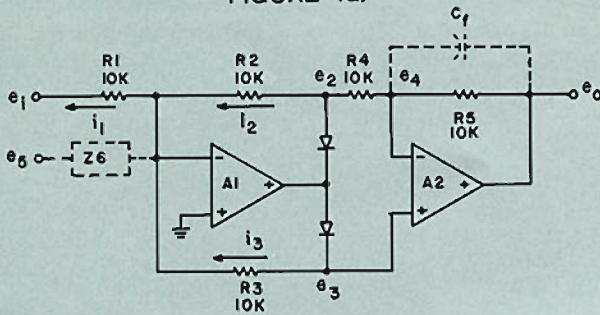


FIGURE 1b.

An improved 'ideal' full-wave rectifier (b) is achieved by modifying a standard circuit (a).

Figure 1b shows an "ideal" rectifier configuration which has the following advantages:

- All resistors are of the same (arbitrary) value.
- Output voltage tolerance is an additive function of resistor tolerance. Equal-value resistors are easy to select for high accuracy.
- Zero input impedance. The single summing junction at  $n$  allows extra isolated inputs to be connected, as shown dotted for  $e_5$  and  $Z_6$ .

It may be necessary to connect a small capacitor,  $C_f$ , in parallel with  $R_5$  in order to prevent high-frequency oscillations. Rectified voltages  $e_2$  and  $e_3$  are unequal in this circuit.

It turns out that the equal-resistor version is a special case and that, in general, the resistors can be unequal as long as they satisfy Eq. 3 below, which is derived in the following manner.

The voltage gain for plus input is:

$$G(+)=e_o/e_i=(R_2/R_1)(R_5/R_4). \quad (1)$$

This is the gain of two tandem-connected inverting operational amplifiers.

For minus input there are two feedback paths to the summing junction of A1, one from each amplifier out (Figure 1b). Mathematically this is expressed as:

$$i_1=i_2+i_3$$

or

$$e_i/R_1=e_o/(R_2+R_4+R_5)+e_3/R_3;$$

since:

$$e_3=e_4=e_o(R_2+R_4)/(R_2+R_4+R_5),$$

then:

$$\begin{aligned} 1/G(-) &= e_i/e_o \\ &= R_1[1/(R_2+R_4+R_5) \\ &\quad + (R_2+R_4)/R_3 \\ &\quad \times (R_2+R_4+R_5)]. \end{aligned} \quad (2)$$

This is equated with  $1/G(+)=[(R_1/R_2)(R_4/R_5)]$  to ensure equal output for plus and minus inputs. The resulting identity is solved for  $R_2/R_3$ :

$$R_2/R_3=(R_4/R_5)+[(1-R_2/R_4)/(1+R_2/R_4)]. \quad (3)$$

Inspection of Eq. 3 shows that if  $R_2=R_4$  the second term drops out and  $R_3=R_5$ . Furthermore, for unity gain,  $R_2=R_3$  and  $R_2=R_3=R_4=R_5$ ; the result is the circuit shown in Figure 1b. If current gain is required from A1,  $R_2/R_4 > 1$  and Eq. 3 is used to calculate  $R_3$  after  $R_2, R_4$ , and  $R_5$  have been selected. If the calculated value of  $R_3$  is negative,  $R_5$  is reduced and  $R_3$  is recalculated.

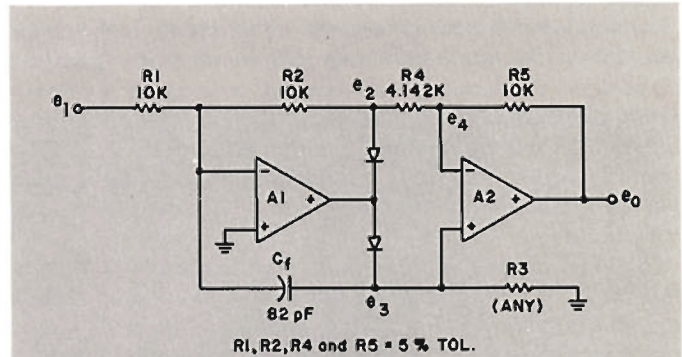


FIGURE 2. Degenerate form of ideal rectifier uses only four precision resistors.

A new circuit results if  $R_3$  is allowed to go to infinity and the noninverting input of A2 is returned to ground through a resistor of arbitrary value (Figure 2). This is an ideal rectifier circuit that requires only three precision resistors (four, if  $R_1$  is counted in). If a further constraint that three of these resistors must be equal is imposed, the resistor ratios are:

$$R_1=R_2=R_5=R;$$

$$R_4=(2^{1/2}-1)R;$$

$$|G|=2^{1/2}+1.$$

(continued on next page)

In general:

$$R_2 > R_4;$$

$$R_4/R_5 = (R_2/R_4 - 1)/(R_2/R_4 + 1);$$

$$|G| = (R_2/R_1)(R_5/R_4)$$

$$= (R_2 + R_4 + R_5)/R_1.$$

The gain is fixed at  $1 + 2\frac{1}{2}$  for both plus and minus inputs. One difficulty with the circuit of Figure 2 is that the forward path for negative inputs consists of the full open-loop gain of A1 and A2 in tandem. A value for  $C_f$  of 82pF was sufficient to stabilize an operating breadboard made with low-cost commercial units.

Allan G. Lloyd

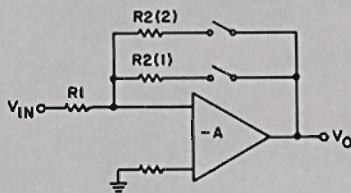
## Modified feedback simplifies programmable voltage supply

By S. K. Bbola, Applications Engineer, Transatron Electronic, Ltd., Maidenhead, England.

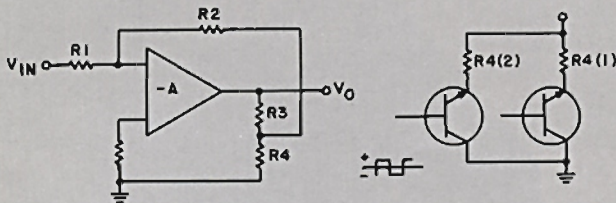
In a programmable voltage supply, using operational amplifiers, the output voltage is usually altered by varying  $R_2$  with series switches, generally relays (Figure a).

When transistors are used as switches, they require isolated transformer drive, adding to size and cost. To avoid this, the feedback network can be modified as shown in Figure b and output is varied by programming  $R_4$  only. The transfer function is given by:

$V_o/V_{in} = (R_2/R_1) + (R_3/R_4) + (R_2/R_1)(R_3/R_4)$ , assuming large input resistance and large open-loop gain. Proper values of  $R_4$  are switched in by using 2N2432 as switches, to give required output voltages.



a



b

Transistor switching to program a power supply which uses operational amplifiers is possible when a standard circuit (a) is modified as shown (b).

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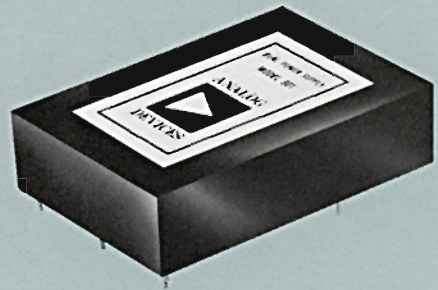
# NEW PRODUCT

## Model 901

### DUAL OP-AMP POWER SUPPLY

#### Features

- Small Size — 2.5" x 3.5" x .87"
- Dual Output —  $\pm 15V$  @ 40mA each
- Regulation — 0.2% max.
- Operates directly from 115VAC, 60Hz
- Epoxy Cast Module for PC Mounting
- Short Circuit Protected
- Low Price — \$49 (1-9); \$39 (10-24)



The Model 901 Dual Power Supply was designed specifically to power operational amplifiers operating from  $\pm 15VDC$ . The unit connects directly to 115VAC line power and produces two fixed 15 volt outputs at 40mA each. The circuit is epoxy cast into a module that can be soldered directly onto printed circuit boards or that can be plugged into an optional mating socket. Hold down nuts are provided to mechanically secure the module in either configuration.

#### SPECIFICATIONS (typical @ 25°C unless noted)

Input Voltage	105 to 125 VAC, 50 to 400 Hz, 7VA max.
Dual Output Voltage (fixed) Accuracy	$\pm 15V$ @ 0 to 40 mA $\pm (15.0$ to $15.3)V$ , $-15V$ within $\pm 1\%$ of $+15V$
Temperature Coefficient, max.	0.03%/°C
Warm Up Drift	$\pm 0.25\%$ (37 mV)
Regulation	
Line (105 to 125 VAC), max.	0.1%
Load (0 to 100%), max.	0.2%
Ripple, max.	2 mV rms; 10mV p-p
Output Impedance	2 ohms @ 10 kHz
Short Circuit Protection	Either output to common indefinitely 0 to 71°C.
Operating Temperature	0 to 71°C derate 1.5 mA/°C above 60°C derate 0.5 mA/°C below 15°C
Storage Temperature	-25 to +85°C
Surface Temperature Rise	20°C above ambient at full load
Weight	9.5 ounces
Price	\$49. (1-9); \$39. (10-24)

Specifications subject to change without notice

# OP-AMP MOUNTING TECHNIQUES

Analog Dialogue receives a great number of inquiries each month regarding recommended mounting techniques for various A.D. amplifiers.

One fear frequently expressed is that the amplifier may be destroyed by making solder connections directly to the pins of the amplifier. As shown in Figure 1 these pins extend well into the amplifier package, and the internal connections are made at the remote end of the pin. Consequently, heat generated by normal soldering processes, including dip, flow, wave, or hand soldering will cause no damage whatever.

Another difficulty is sometimes encountered in the insertion of the amplifier into pre-drilled circuit boards. The pin spacing on all A.D. amplifiers is held to .005", but combined with the tolerances of the drilled P.C. board, insertion difficulties may arise. Although some users solve this problem by drilling over-

sized holes, this occasionally leads to faulty solder connections. Reports from the field indicate that many customers effectively eliminate these problems by the use of pin receptacles. (Figure 2.) These devices are commercially available from AMP, Incorporated (#1-380758-0) and Cambion (#338-1-03). These receptacles have sufficient "float" built into them to accommodate the tolerances usually encountered in P.C. board applications. Additionally, some users have found that because the receptacles can be soldered in place and the amplifier inserted later, problems of logistics, rework, and retrofit are relieved. The spring-loaded receptacles are re-usable and will secure the amplifier for most industrial and commercial environments. Where extreme vibration or demanding specifications require absolute security, any A.D. amplifier can be supplied with molded-in screw inserts (Figure 3) on special order.

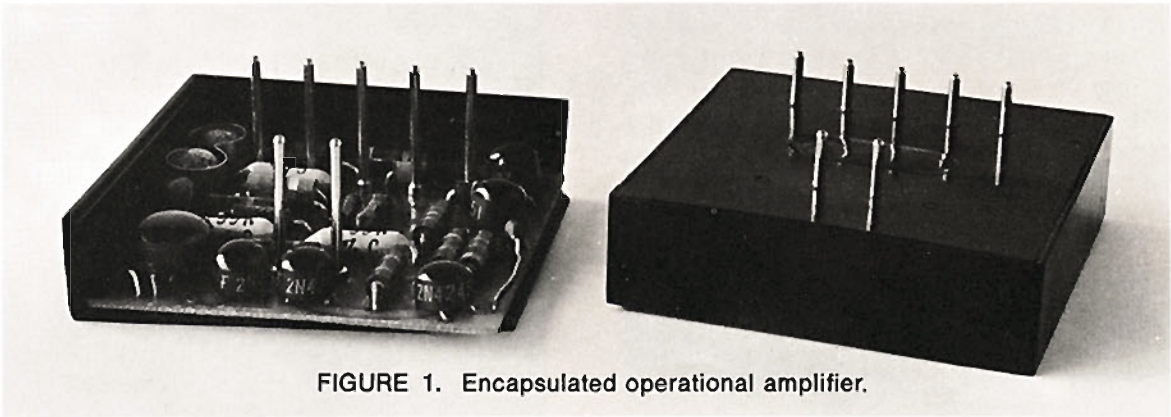


FIGURE 1. Encapsulated operational amplifier.

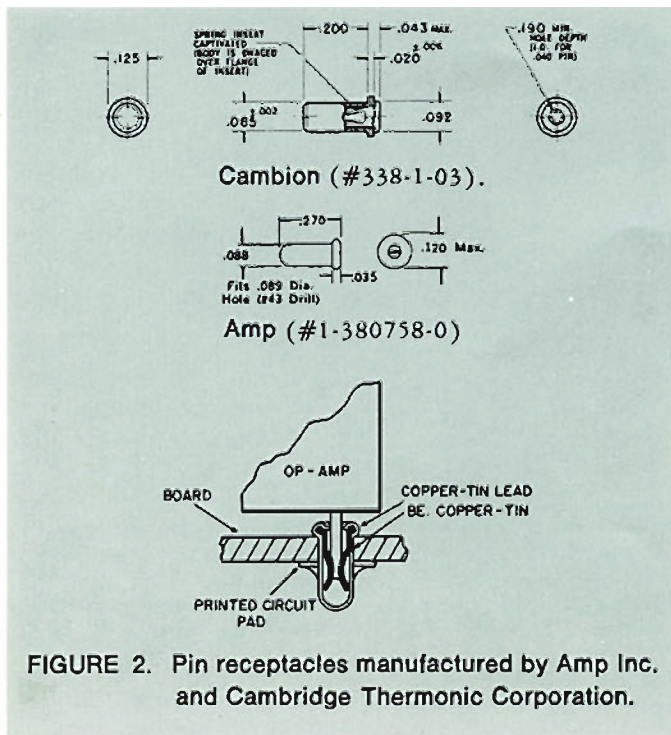


FIGURE 2. Pin receptacles manufactured by Amp Inc. and Cambridge Thermonic Corporation.

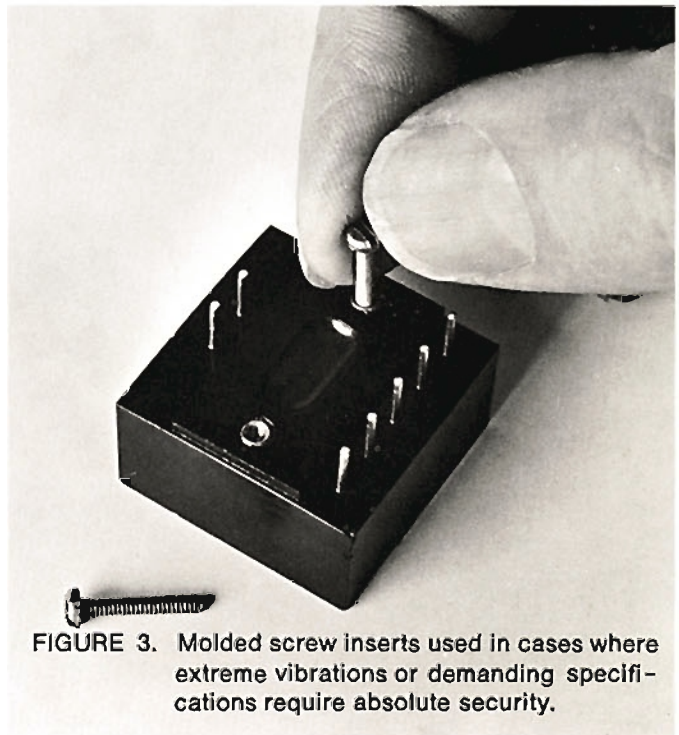


FIGURE 3. Molded screw inserts used in cases where extreme vibrations or demanding specifications require absolute security.

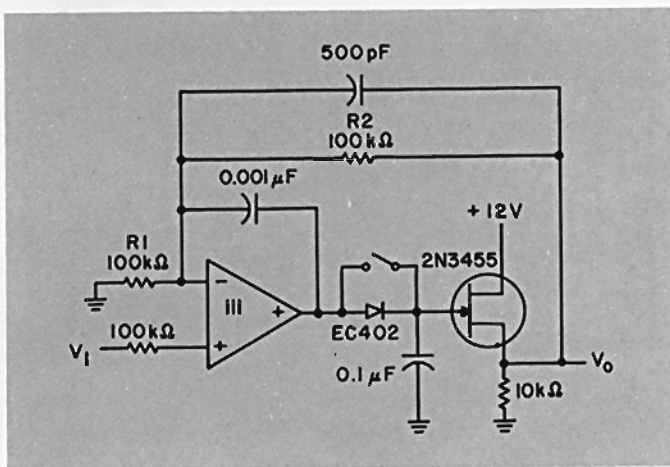
# Precision peak-reading circuit

By E. E. Wheeler, PIRA, (The Research Association for the Paper & Board, Printing & Packaging Industries) Leatherhead, Surrey.

Sir, — A recent laboratory requirement involved the capture of the peak value of a positive pulse of varying duration. The standard method of doing this is by charging a capacitor *via* a diode and sensing the voltage across the capacitor with a high-input-impedance device, such as a valve. Disadvantages of such a system are primarily due to the diode. On applying a pulse, this has initially a low forward resistance but as the voltage across the storage capacitor rises the diode forward resistance increases. In a test the following results were obtained from an SGS Fairchild EC402 diode:

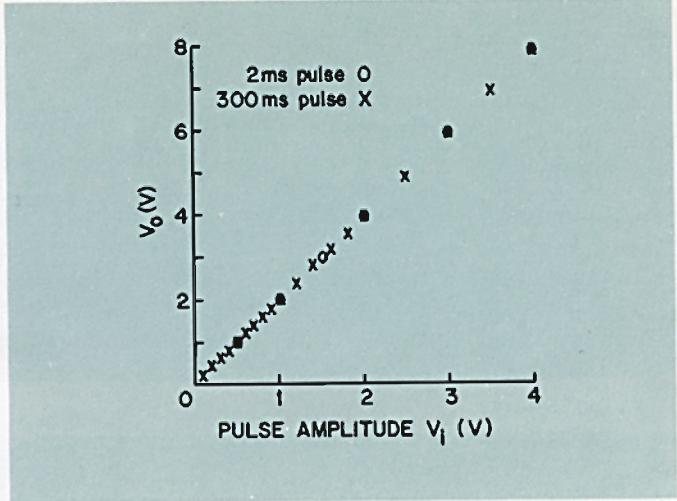
Forward current, mA,	Forward voltage, V	Forward resistance, $\Omega$
1.0	0.65	650
0.52	0.62	1200
0.25	0.59	2360
0.10	0.55	5500
0.05	0.53	10600
0.025	0.51	20400
0.0125	0.485	38800
0.006	0.46	76700
0.003	0.435	145000
0.0015	0.41	274000
0.0005	0.365	730000
0.00025	0.345	1380000

The diode forward characteristic causes the peak-reading circuit to be sensitive to pulse duration, a short pulse giving a lower reading than a long pulse of the same amplitude. The diode characteristic also necessitates a meter with a nonlinear scale, particularly at the low end.



The illustrated circuit overcomes the above disadvantages, giving, for pulses of 2 to 300ms, an output that is not a function of pulselength and providing a linear  $V_0/V_1$  relationship. Performance was not investigated beyond these limits or for  $V_1$  greater than 4V. The graph shows the linearity and independence of pulselength that was achieved.

The gain of the circuit is given by  $(R_1 + R_2)/R_1$ . The 500pF feedback capacitor brings results from 2ms pulses into line with those from 300ms pulses. The operational amplifier is an



Analog 111, the diode is a low-reverse-leakage type and the field-effect transistor is by Amelco. The circuit will hold the peak value of an applied pulse for an hour or more if due attention is paid to insulation. Closing the switch resets the system to zero. A 20000 $\Omega$ /V meter was used to record  $V_0$ , but in the final application an emitter follower is connected to the field-effect transistor and the feedback loop taken from the emitter-follower output.

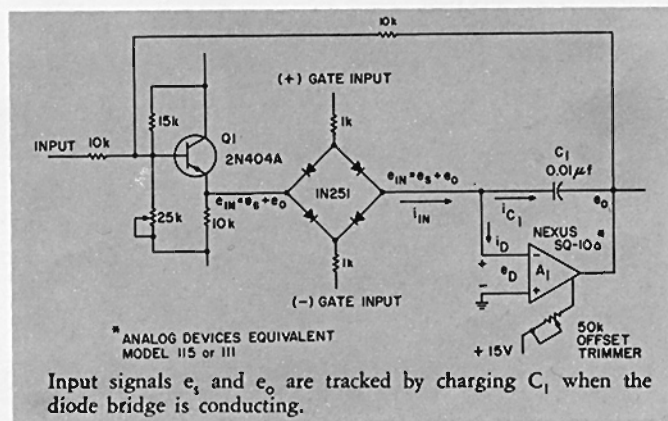
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## Signal is sampled and held for 1 minute

By Richard A. De Perna

Neurophysiology Laboratory, Massachusetts General Hospital, Boston

A low-cost operational amplifier and a single transistor combine to form a \$20 track and hold circuit for use with low sampling rates. This application requires sampling of a 30-hertz sinewave once per cycle with a 0.4 millisecond sample width; therefore, an off-the-shelf, low performance operational amplifier was used. By substituting more expensive operational amplifiers in the integrator, performance can be upgraded. Input signal  $e_i$  and the feedback signal  $e_o$  are fed to the base of emitter follower  $Q_1$ . When gate voltages of +1.5 and -1.5 volts are applied to the +gate and -gate terminals respectively, the diode bridge conducts and capacitor  $C_1$  charges through the low output impedance of  $Q_1$ . In this mode the feedback path through the 10-kilohm resistor causes the output to track the input.



Input signals  $e_i$  and  $e_o$  are tracked by charging  $C_1$  when the diode bridge is conducting.

(continued on page 16)

When the voltages at the gate inputs of the diode bridge are reversed, the forward path of the diode bridge is opened and the circuit behaves as a conventional integrator in the hold mode. Thus the capacitor remains charged to the value of the output voltage immediately prior to switching. The 50-kilohm offset trimmer balances out the offset current of the operational amplifier to prevent rundown of the output voltage  $e_o$  during the hold mode.

Operation of the circuit is possible with sample widths down to 0.4 milliseconds or track widths of several times this value. In this application the network functions as a sample and hold circuit because with this gate width it is just possible to charge

the capacitor between the most negative and positive output values. With larger gate widths the output will track the input after the initial charging time. The data can be held for periods of up to one minute. The minimum sample width is determined by the charging time constant of the capacitor. Although the time constant may be reduced with a smaller capacitor, this makes the offset adjustment more critical. The circuit becomes more sensitive to the time and temperature variations of the offset current of operational amplifier  $A_1$ . Usually,  $C_1$  is chosen as the largest value of capacitance that provides satisfactory operation with the desired sample and track width.

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### WORTH READING

*Linear Integrated Circuits Applications Handbook*  
Fairchild Semiconductor  
313 Fairchild Dr.  
Mountain View, Calif.

A prolific publication to "provide significant insight into the advantage and limitations inherent in monolithic construction and how these relate to specific systems (and circuit) design problems, and serve as a convenient reference for all users of linear integrated circuits." Many applications ideas. (188 pages)

*Foggy "Specs" Blur Designs*  
Electronics, October 16, 1967  
McGraw-Hill Building  
330 W. 42nd St., N.Y., N.Y. 10036

A very good discussion by Mr. Harris of Zeltex, Inc. which points out some of the more subtle limitations of op amp specs particularly as related to IC's. Buyer beware. (4 pages)

*Ramp and Triangular Waveform Generator*  
Electronic Instrument Digest, Sept. 1966  
Milton S. Kiver Publication, Inc.  
222 West Adams Street  
Chicago, Illinois

Circuits are discussed to achieve ramp and triangular waveforms using op amps. Sources of error are discussed. (4 pages)

*Transistor Leakage Current Test Circuits*  
Electronic Design, July 19, 1967  
850 Third Avenue, N.Y., N.Y. 10022

Ronald Mann of T.I. defines 10 different transistor leakage currents and describes how these specifications can be measured using simple op amp circuits. (7 pages)

*Current Dividers Convert Digital Signals into Analog Voltages*  
Electronics, November 14, 1967  
McGraw-Hill Bldg., 330 W. 42nd St.  
New York, N.Y. 10036

Article describes how to design D-A converters using op amps and active current dividers in lieu of precision resistors. (7 pages)

*Analyzing Linear IC's*  
Motorola Monitor, Vol. 5, No. 2  
P.O. Box 13408  
Phoenix, Arizona 85002

Article analyzes philosophy of basic differential amplifier design used in operational amplifiers. Goes on to discuss circuit configurations used in Motorola's line of linear IC op amps. Very, very good reading. (8 pages)

*Experimental Investigation of Neutron and Gamma Radiation Effects on Semiconductor Differential Amplifier*  
U.S. Army Missile Command  
Control Systems Branch  
Redstone Arsenal, Alabama 35809  
Report #RG-TR-67-4, Mar. 1967  
by Pat H. McIngvale

Several commercially available op amps were subjected to radiation and experimental data is reported. Objective was to determine the usefulness of these devices in a radiation environment.

*Equivalent Circuits Estimate Damage from Nuclear Radiation*  
Electronics, Oct. 30, 1967  
McGraw-Hill Building  
330 W. 42nd Street  
New York, New York 10036

Presents model to predict effects of radiation on simple one transistor. Article together with bibliography and references gives a starting point to explore radiation effect on solid-state amplifiers.