ANALOG DEVICES

APPLICATIONS MANUAL

Models 201, 202, 203 & 210

Chopper Stabilized Operational Amplifiers

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Analog Devices, Inc., Cambridge, Mass.
1967

1.0 INTRODUCTION

1.1 Description

Analog Devices' 200 Series chopper stabilized amplifiers utilize a low frequency AC amplifier, modulator, and demodulator to improve the drift characteristics of a conventional differential operational amplifier while retaining the wideband characteristics of the unstabilized amplifier (with the exception of differential inputs). The modulator (chopper) and demodulator are completely solid state, driven by an internal solid state chopper drive. A high speed overload recovery circuit is included as standard on all amplifiers. Outputs are completely short circuit protected.

The philosophy of operation consists of dividing the incoming frequency spectrum into two parts, such that all frequencies above a few cycles per second are amplified in a conventional manner by the main channel, while the very low frequency input signals are chopped, amplified by an AC amplifier, and demodulated before being applied to the main channel of the amplifier. This effectively divides the drift of the main channel amplifier (referred to the input) by the gain of the chopper amplifier, with the result that the predominant contributions to drift are those of the low level modulator.

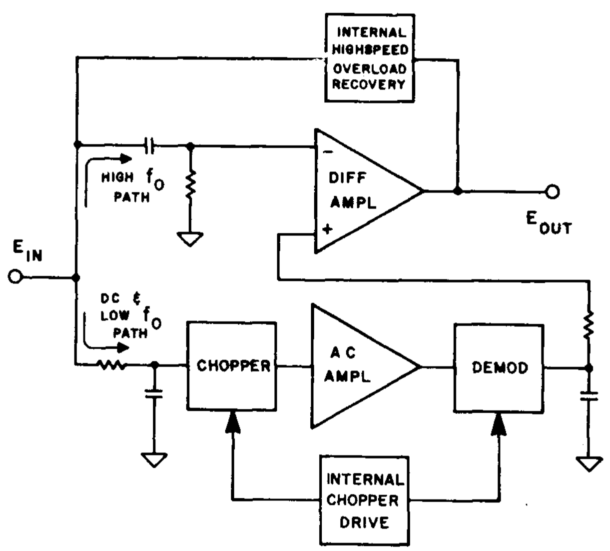


FIGURE 1 - CHOPPER STABILIZED AMPLIFIER BLOCK DIAGRAM

1.2 Basic Operation

The block diagram is shown in figure 1. DC and very low frequency components of the input signal are directed by a low pass filter to the chopper, are chopped, amplified, demodulated, filtered and applied to one input of the main channel. Higher frequency input signals are routed by the high pass filter directly to the main channel. The two frequency spectrums are re-combined by summing at the different inputs of the main channel. Proper phase relationship of the low frequency signals is accomplished by a phase reversal through the AC section and then by a second phase reversal via the (+) input of the main channel to be in phase with the higher frequency components applied directly to the (-) input of the main channel.

1.3 Chopper Stabilized vs. Differential Op Amps

With recent improvements in unstabilized operational amplifiers, a discussion of the advantages of chopper stabilized amplifiers may be of value. Voltage drift of unstabilized units is available to 3μV/°C and reported to be possible to less than 1μV/°C. This does not compare badly with drifts of 0.2μV/°C to 1.0μV/°C for stabilized amplifiers. However, offset current and current drift are much better in stabilized amplifiers; 50pA and .5pA/°C versus 1nA and 50pA/°C and only over a limited temperature range at that for unstabilized amplifiers.

Less obvious advantages of stabilized amplifiers are the short time required for warm up to within specifications, and the immunity to thermal gradients. Differential amplifiers have a warm up period of about 20 minutes, during which time initial offset voltage may change by several hundred microvolts. There is also great sensitivity to externally generated thermal gradients and in some cases internally generated gradients due to changing load conditions. These thermal drifts may be well in excess of those produced by changes in ambient temperature, particularly when operated over a narrow range of ambient temperatures.

A further advantage of stabilized amplifiers is that long term offset stability at constant temperature is excellent-as little as 10 μV over a several month period-while initial offset voltage of differential, unstabilized amplifiers may drift several hundred microvolts over the same period.

1.4 Special Features

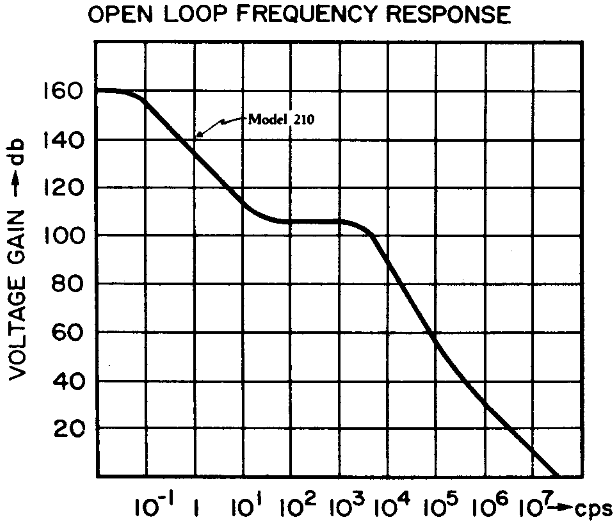
Special features of Analog Devices' 200 Series amplifiers include:

1. An internal chopper drive that eliminates an externally generated 60 Hz drive signal and the AC noise pick up usually accompanying this.
2. An integral zener, resistor, diode clamping network that keeps the amplifier from saturation and returns the output to the linear region after an input overload in less than a microsecond.
3. Output is short circuit proof with short circuit current limited to about 100 ma for all models including Model 201.

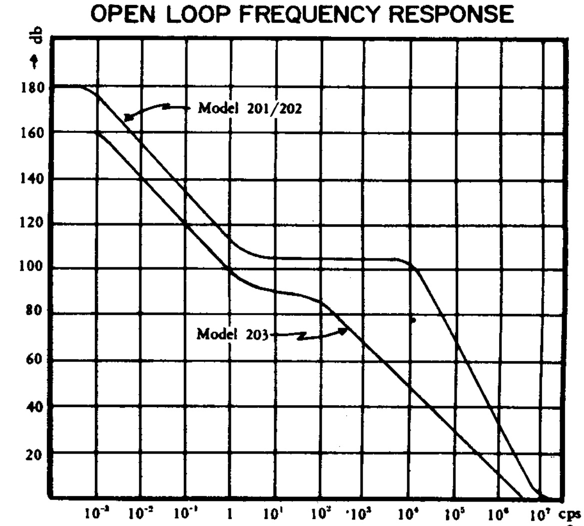
2.0 SELECTION GUIDE

Specifications (typical at 25°C and ±15VDC unless otherwise noted)	Model 201	Model 202	Model 203	Model 210
OPEN LOOP GAIN, dc, rated load	10^9	10^9	10^8	10^8
RATED OUTPUT Voltage, p to p, min. Current, min.	 ±11 V 100ma	 ±11 V 20 ma	 ±11 V 20 ma	 ±10 V 20 ma
FREQUENCY RESPONSE Unity Gain, Small Signal Full Power Response Slewing Rate Overload Recovery	 10mc 500KC 30V/μsec 0.5μsec	 10 mc 500 KC 30V/μsec 0.5μsec	 2 mc 20 KC 1.2V/μsec 5μsec	 20 mc 500 KC 100V/ μsec 0.2μsec
INPUT OFFSET VOLTAGE Initial Offset, 25°C, max. Average vs. Temp. (-25 to 75°C) max. (+10 to 85°C) max. (-25 to 10°C) max. vs. Supply Voltage vs. Time	 ±20μV 0.2μV/°C -- -- 0.4μV/ % 1μV/day	 ±20μV 0.2μV/°C -- -- 0.4μV/ % 1μV/ day	 ±20μV 0.2μV/°C -- -- 0.4μV/ % 1μV/ day	 ±100μV -- 1μV/°C 2μV/°C 10μV/% 1μV/ day
INPUT OFFSET CURRENT Initial Offset, 25°C, max. Average vs. Temp. (-25 to 75°C) max. (-25 to 85°C) max. vs. Supply Voltage	 ±50pa 0.5pa/°C -- 1pa/ %	 ±50pa 0.5pa/°C -- 1pa/ %	 ±50pa 0.5pa/° C -- 1pa/ %	 ±100pa -- 2pa/°C 10pa/%
INPUT CHARACTERISTICS Input Impedance, dc, open loop Voltage Noise, DC to 1cps, p to p 5cps to 50KC, rms Current Noise, DC to 1cps, p to p	 220KΩ 25 μV 10 μV 20 pa	 220KΩ 25μV 10μV 20pa	 220KΩ 10μV 10μV 10pa	 500KΩ 5μV 10μV 10pa
POWER SUPPLY Voltage Current, quiescent	 ± 15 to 16 VDC + 15, - 25 ma	 ± 15 to 16 VDC + 13, - 20 ma	 ± 15 to 16 VDC + 13, -20 ma	 ± 15 to 16 VDC + 40 , -10 ma
TEMPERATURE RANGE Specifications Operating Storage	 -25 to 75°C -40 to 75°C -55 to 75°C	 -25 to 75°C -40 to 75°C -55 to 75°C	 -25 to 75°C -40 to 75°C -55 to 75°C	 -25 to 85°C -40 to 85°C -55 to 100°C
PRICE (1-9) (10-24)	$270. $256.	$235. $224.	$215. $205.	$157. $148.

OPEN LOOP FREQUENCY RESPONSE



OPEN LOOP FREQUENCY RESPONSE



3.0 AMPLIFIER INSTALLATION & APPLICATION CONSIDERATIONS

3.1 Pin Connections

The pin configuration and wiring diagrams shown in figures 2 and 3 indicate how the amplifier should be connected for proper operation. Note that the input return terminal must be externally connected to power supply common unless an external balance pot is required (see below). To prevent ground loops, a separate wire should go directly from the input return to power supply common. The small dot in the

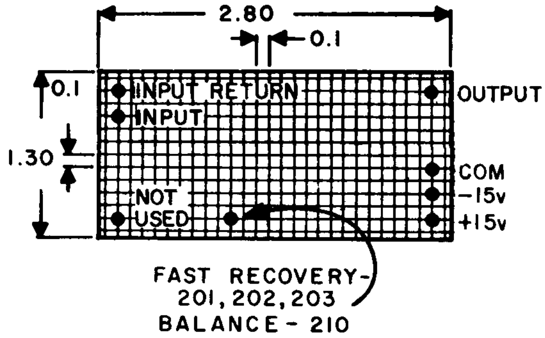


FIGURE 2 - PIN LAYOUT
(BOTTOM VIEW)

bottom corner of the amplifier triangle symbol denotes a connection to the power supply common. The input return for all figures (after figure 4) is implied to be connected externally to the power supply common. Where the fast overload recovery feature is desirable on Models 201,202, and 203, the fast overload terminal must be externally shorted to the output terminal. Model 210 has the fast over-load recovery circuit always operative (internally connected):

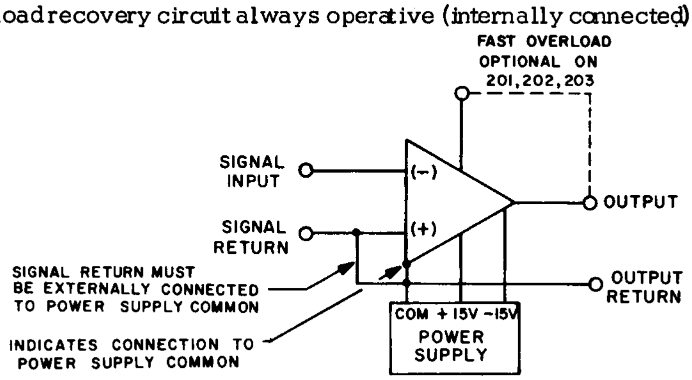


FIGURE 3 - CONNECTIONS

The amplifiers can be damaged by applying incorrect polarity of supply voltage. Be sure polarities are correct.

Remember that the exceptionally low drift and input current specification can be degraded by improper external wiring. Leakage resistance from the summing junction to an external voltage potential is particularly critical. For example, a leakage resistance of 10^{12}ohms between the summing junction and the 15VDC supply voltage results in 15pA input current, which is greater than the typical initial offset current of the amplifier. For critical applications such as high quality integrators, leads to the summing junction should be isolated from external voltage potentials by ground potential shielding. Moreover, thermocouple voltages can generate offset voltage drift comparable to that of the amplifier and precautions should be taken to minimize this source of offset potential.

3.2 External Zero Offset Adjustments

Typical initial offset voltage on the 201, 202, and 203 is $\pm 10\mu V$ with a maximum specification of $\pm 20\mu V$. In most circuits such low offset does not require an external balance adjustment. However, in critical applications the circuit in figure 4 can be used to zero the amplifier. The range of adjustment for the resistor values given is about

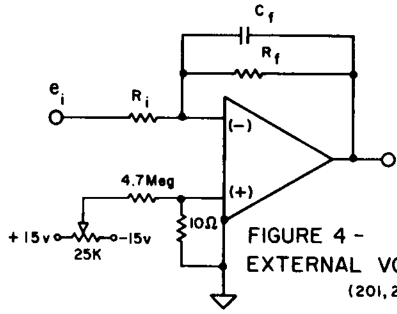


FIGURE 4 - EXTERNAL VOLTAGE BALANCE
(201,202,203)

$\pm 30\ \mu V$. The 4.7 megohm resistor can be decreased for a greater range adjustment where it is desirable to zero suppress an external offset voltage. The input return should never be raised above a few millivolts unless the factory is consulted for the possible effects on other operating specifications, moreover, impedances greater than 100 ohms should never be used in the input return lead. Notice that the stability of the ±15VDC bias voltages and of the voltage divider resistors are not particularly critical since, for example, a 1% change of $30\mu V$ is only $0.3\mu V$.

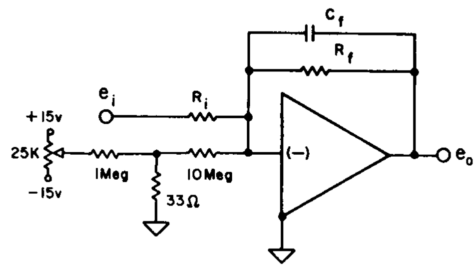


FIGURE 5 - EXTERNAL CURRENT BALANCE

For precision integrator or low offset current applications, the initial offset current can be zeroed with the circuit shown in figure 5 as an alternative to the voltage biasing of figure 4. The adjustment range of the bias circuit in figure 5 is ±50pA.

For a fixed source impedance, voltage or current biasing have an equivalent effect and either technique can be used to zero both initial voltage and current offsets. However, for a change source impedance in critical applications, as in the case of opening and shorting the input or in multiplexing the input to various source impedances, then both voltage and current biasing must be used to zero offset voltage and current independently.

Model 210 has a maximum initial offset voltage of $\pm 100\mu V$ but requires only an external 50K pot connected to the balance terminal for zeroing. If initial voltage zero is not required, the balance terminal of the amplifier should be left open. The circuit of figure 5 may also be used with the 210 to zero initial offset current if desired, except the adjustment range should be increased to ±100pA by increasing the 33Ω resistor to 68Ω.

3.3 Single Ended Input

The 200 Series, like most chopper stabilized operational amplifiers, have single ended (inverting only) inputs. This means that one side of the input signal must be common to the power supply ground and the output ground.

Special circuitry is required to perform non-inverting and differential input amplification. Refer to paragraphs 5.3 and 5.4 for circuits and considerations.

3.4 Closed Loop Stability

The high frequency performance of Models 201, 202, and 210 is characterized by a useful small signal bandwidth of up to 10mc, slewing rates of 30 to 100 V/ µsec and open loop gain greater than 60db at 100KC. Note that gain bandwidth product at unity gain is 10mc, but increases for a closed loop gain of 100 to 80mc. These specifications are achieved in part by internal phase compensation networks that attentuate the open loop gain at nearly 12 db/octave. "Fast rolloff" operational amplifiers require a small feedback capacitor across the feedback resistor in most circuits to provide proper phase margin for stability. For optimum bandwidth, it is suggested that a square wave be applied to the closed loop circuit, and the value of C_f adjusted for the desired transient response at the output. For a range of gains and operational resistances, C_f will vary from one or two pf to perhaps hundreds of pf.

Maximum available closed loop bandwidth is determined by the intersection of the closed loop gain curve (more accurately 1/feedback factor) with the open loop gain curve. The approximate value of C_f may be calculated by selecting a frequency of about 7/10ths of the above intersection frequency as the break frequency (-3db point) of R_fC_f. For R_f less than one megohm, closed loop bandwidth will then be $\omega_{3db} = 1/R_fC_f$.

Maximum closed loop bandwidth of Models 201 and 202, for example, is 5mc at unity gain (the reduction from 10mc is due primarily to the feedback factor of $\beta = 1/2$ for the unity gain inverting configuration). For a closed loop gain of 100, maximum closed loop bandwidth is 800KC. If wide bandwidth is not required, it is suggested that C_f be increased to limit the bandwidth to the minimum value commensurate with system requirement so as to reduce noise.

The Model 203, identical to the Model 202 except for lower frequency response, lower slewing rate, and 6db/octave attenuation, should be considered where wide bandwidth is not required (as in a low frequency integrator), or a high ly reactive load is to be driven.

3.5 Improving Bandwidth & Slewing Rate of Inverting Amplifiers

When circuit design reasons compel large value of feedback resistance for inverting amplifiers, and bandwidth is then limited either by the internal feedback capacitance of approximately 1pf or by an external stabilizing capacitor, Cf, in conjunction with R_f, an improvement in bandwidth may be obtained by shunting R_i with a small capacitor and resistor in series (figure 6).

where $R_iC_i \approx R_fC_f$. R_r should be chosen as perhaps $R_i/10$. A frequency response test or square wave transient response test is suggested for determining optimum values of C_i and R_r.

The published slewing rate specification is determined with

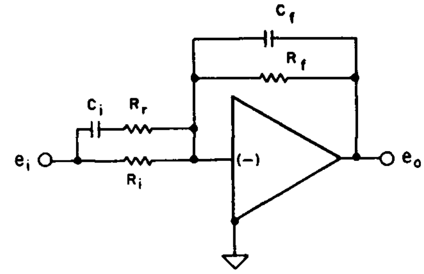


FIGURE 6- IMPROVING BANDWIDTH

small values of input and feedback resistors. Input capacitance of chopper stabilized amplifiers is on the order of one hundred picofarads. This capacitance, or any feedback capacitance, may limit the slewing capability of an amplifier to less than the specification if large values of operational resistors are used. The above technique of adding a small capacitor across the input resistor may also improve slew rate as well as bandwidth.

3.6 Capacitance Loads

Load capacitance, in conjunction with the output impedance of the amplifier can cause oscillations. Figure 7 shows an isolating circuit suitable for most applications with capacitance loads. Connecting the feedback resistance to point A causes less interaction between R (20-100 ohms) and the other circuit values, but results in a higher output impedance (=R). Connecting the feedback resistance to point B achieves extremely small low frequency output impedance since R is inside the feedback loop.

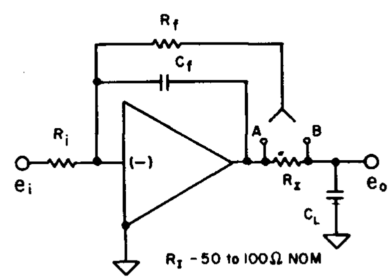


FIGURE 7-ISOLATING CAPACITANCE LOADS

3.7 Long Leads

All lead connections, including those to the power supply, should be made as short as possible. If it is necessary to have long leads to the feedback network, connect the stabilizing capacitor, C_f, directly from the output terminal to the input terminal.

3.8 Overcoming Large Feedback Resistors

Low level DC amplifier circuits many times require feedback resistors in the 10 to 1000 megohm range. High value, stable, precision resistors are not readily available. The circuit in figure 8 shows one way to circumvent the use of very large resistance values. The gain for this circuit is,

$$\frac{e_o}{e_i} = - \left(\frac{R_1 \| R_f + R_2}{R_1 \| R_f}\right)\left(\frac{R_f}{R_i}\right)$$

which for $R_f >> R_1$ reduces to

$$\frac{e_o}{e_i} = - \left(\frac{R_1 + R_2}{R_1}\right)\left(\frac{R_f}{R_i}\right)$$

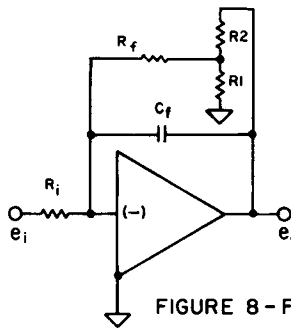


FIGURE 8 - FEEDBACK VOLTAGE DIVIDER

For example, if a closed loop gain of 1000 is required with a summing impedance of 1 megohm, we could select the following circuit values, R_i = 1 megohm, R_1 = 100 ohms, R_2 = 100K ohms and R_f = 1 megohm.

One disadvantage to this technique arises in that voltage drift and noise are increased. Voltage drift referred to the output in the regular configuration for the example above would be

$$\Delta e_o = \left(1 + \frac{1000M}{1M}\right) e_d \Delta T = 1001 e_d \Delta T$$

and for the three resistor network case would be

$$\Delta e_o = \left(\frac{R_1 + R_2}{R_1}\right)\left(\frac{R_f + R_i}{R_i}\right) e_d \,\Delta T$$

$$= \left(\frac{.1K + 100K}{.1K}\right)\left(\frac{1M + 1M}{1M}\right) e_d \Delta T$$

$$= 2002 e_d \Delta T$$

This effect can be minimized by using as small a ratio as practical for $(R_1 + R_2)/R_1$.

3.9 Noise

Noise at the output of an operational amplifier is the sum of various noise components. Input noise voltage is often all that is specified on an amplifier data sheet. This information is sufficient to derive the noise characteristics for a closed loop circuit having gains of 100 to 1000 and very low operational resistances, but additional calculations must be made for other cases. The sources of noise in an operational amplifier circuit are:

1. Input voltage noise of amplifier
2. Input current noise of amplifier
3. Thermal noise of input resistor
4. Thermal noise of feedback resistor
5. Current noise of input resistor
6. Output voltage noise
7. Power supply coupled noise
8. Pick up and RFI

The bandwidth, or frequency characterization of the noise sources, and the degree of correlation between the sources must be also determined for a complete noise analysis.

Broadband noise at the output (exclusive of pick up and RFI) for a circuit with a high closed loop gain and small operational impedances will be approximately the broadband input voltage noise times the closed loop gain. Note that the noise is specified for a particular noise bandwidth (also called a "brick wall" or ideal filter bandwidth). The noise measurement must be made with this filter inserted between the amplifier and the measuring device. If a single pole filter is used in place of the " brick wall " filter, a -3db point of about 63% of the noise bandwidth must be chosen to compensate for the 6db/octave slope of the single pole filter instead of the infinite slope of the ideal filter.

A brief indication of the calculations will be illustrated. The circuit of figure 9 will be used for the analysis.

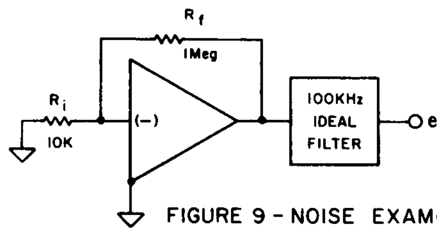


FIGURE 9 - NOISE EXAMPLE

Assume an input voltage noise of 10μV RMS over a noise bandwidth of 100KC. Output noise from this source will be 1 mV RMS. Broadband current noise for the amplifier can be about 1 nA RMS; with 1 megohm feedback resistor, output noise will be 1mV RMS from this source.

Assuming the input voltage and current generators to have a correlation coefficient of 0.3 (a typical value for these amplifiers), the output noise is

$$\sqrt{(1mV)^2 + 2 \times .3 \times 1mV \times 1mV + (1mV)^2}$$

$$= 1.61\text{mV RMS RTO}$$

Thermal noise of any impedance Z is

$$E_{th(rms)} = \sqrt{4KT\Delta fR}$$

where $\sqrt{4KT} = 1.3 \times 10^{-10}$ at 25°C

Δf is the ideal filter bandwidth over which the noise is measured.
R is the real part of the impedance Z.

Thermal noise of the input resistor (referred to the output) is

$$E_{th(rms)} = 100 \times 1.3 \times 10^{-10}\sqrt{100KC \times 10K\Omega}$$

$$= .4\text{mV RMS RTO}$$

This will be an uncorrelated noise source and so will add in quadrature to the other noise sources.

We now have $\sqrt{1.61^2 + .4^2} = 1.66$mV RMS RTO

Thermal noise of R_f is 40μV RMS RTO which will be insignificant in the present example.

Resistor current noise is noise generated in an conventional resistor when current passes through it. The spectral density has a 1/f characteristic with wirewound and some metal film resistors having a coefficient of perhaps .1μV RMS per volt applied per decade of frequency and noisy carbon composition or film resistors having as much as 10μV RMS per volt per decade. The current noise due to a noisy input resistor in the present circuit considered from 1cps to 100KC (5 decades) would be

$$E_{c(rms)} = 100 \times \frac{10\mu V}{V} \times .1V \times \sqrt{5}$$

$$= .22mV \text{ RMS RTO}$$

The .1V is full scale input voltage; current noise is a linear function of input and would obviously contribute no noise with zero input. The noise at full scale would now be 1.68mV RMS RTO.

This particular calculation shows equal effects due to amplifier input voltage and current noise with much reduced effects due to other causes. Other cases may show a much different proportion of effects, for example, resistor current noise would show a marked change in respect to other contributions if a narrower bandwidth were used.

3.10 Multiplexing

When multiplexing (sometimes called scanning or commutating) into a chopper stabilized amplifier with millivolt level signals, certain errors arise that can be minimized by certain precautions. Intermodulation with the chopper drive frequency will show up as signal output errors which are possible to minimize by choosing a scan rate that will not interact with the chopper frequency. Chopper frequency for Models 201, 202, and 203 is about 35Hz and for Model 210 about 150Hz. Harmonics of these frequencies must also be considered when choosing the scan rate.

On special order the Model 201, 202, and 203 may be obtained with provisions for externally driving the chopping oscillator at the scanning frequency or sub-multiples of the scanning frequency. This will eliminate intermodulation noise. The driving frequency must be 35Hz ±1%.

High speed multiplexing, that is, scan rates in the kilocycle region, will cause a shift in the baseline for any chopper stabilized amplifier. This is due to inability of the amplifier output to follow the input switched risetime, generating spikes at the summing junction, and subsequent nonlinear amplification by the chopper amplifier causing base line shift. This problem can be minimized by slowing input rise time, by selecting an amplifier with greater slew rate and full frequency output, or by reducing scan rate.

3.11 Overload Recovery

An internal fast overload recovery circuit is included in these amplifiers to prevent the amplifier output from saturating. Thus the long delays following output saturation which is characteristic of most chopper stabilized amplifiers is avoided.

The internal overload protection circuit is shown below in figure 10. Notice that the fast overload circuit must be externally connected on Models 201, 202, and 203 while it is permanently connected internally on the Model 210.

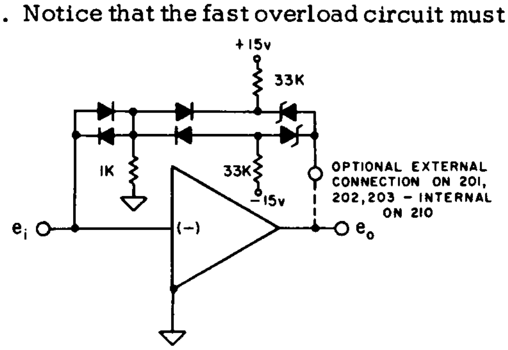


FIGURE 10 - INTERNAL OVERLOAD RECOVERY

Rated output voltage for the Model 201, 202, and 203 is ±11 volts <u>without</u> the overload recovery circuit, but it can be as low as ± 9 volts with this circuit connected due to zener tolerances. Minimum output of ± 10 volts with the overload circuit connected can be obtained on special order.

The circuit provides additional feedback current when the amplifier output exceeds the zener plus diode voltage drops. The maximum amount of current that can be fed back is the amplifier's rated output current less the load current, therefore the input overload current must not be allowed to exceed this.

The proper operation of this circuit depends on the following instructions:

1. Power supply voltage must be a minimum of ±15 volts for proper operation.
2. All input overload current must be supplied through the feedback network from the amplifier output in order to maintain linear operation. If the input current plus the load current exceeds the amplifier's rated output current, then the overload recovery circuit will not function properly. For example, a 1000 ohm load resistor draws 10ma. Therfore, with the 202, maximum input overload current cannot exceed 10ma. Input current is normally limited by R_i, but some special circuits (for example, non-inverting amplifiers) may require additional current limiting to retain overload recovery characteristics; refer to paragraph 5.3 for overload protection circuits for non-inverting amplifiers.
3. If the input resistor, R_i, does not adequately limit overload current, an additional stage of silicon diode limiting may be employed as in figure 11 (two diodes in series may be required, R_p can be about 100 to 200Ω).

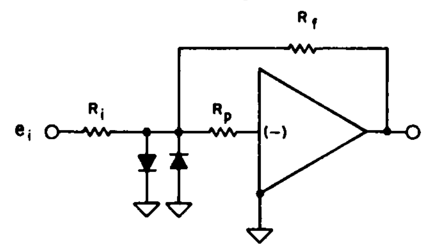


FIGURE 11 - OVERLOAD CURRENT LIMITING

Maximum overload transient voltages across the input terminals should be limited to ±50 volts. Due to dissipation limiting, maximum steady state voltage overload to the input should not exceed ±15 VDC.

3.12 Turn On Transient

If the ±15 VDC power supplies do not come on at exactly the same instant, the output of the amplifier will saturate during initial turn on and require as much as thirty seconds to recover on the Models 201, 202, and 203 and about two seconds on the Model 210. This is normal operation for these amplifiers, and, as a matter of fact, for any other chopper stabilized amplifier. Once the amplifier has recovered from the initial turn on transient, it will recover very rapidly from overload due to overdriving the input.

3.13 Power Supply Voltage

Rated specifications depend on the supply voltages being in the range from ± 15 to 16 VDC. As the supply voltage drops below ±15VDC the overload circuit will not function properly and below 13 to 14 volts the amplifiers operating specifications are degraded.

Model 205 is available for operation from ±12 volt power supplies; specifications are similar to the 202 with reduced output swing. Models are available on special order to operate at other supply voltages.

The amplifiers are relatively insensitive to gradual power supply changes or low frequency variations. However, there is sensitivity to supply voltage AC ripple and transients. For best operation, ripple voltage should be no more than a few millivolts and voltage transients should be decoupled as much as possible.

Low power supply output impedance at high frequencies may be lost through long lead lengths, and local decoupling of the power leads may be required. All models have capacitors connected internally across the power supply terminals to give some local decoupling.

3.14 Accessories

A plug in socket, Model AC 1002, (figure 12) is available which can be used for mounting the amplifier or for aiding in evaluation. A noise shield, Model AC1106 (figure 13), is also available for reducing the effects of noise pick up in areas of high noise and RFI generation. The AC1106 also serves as a hold down bracket in high shock or vibration environments.

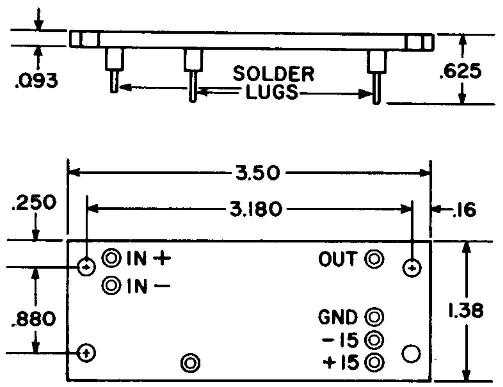


FIGURE 12 — AC1002 MATING SOCKET

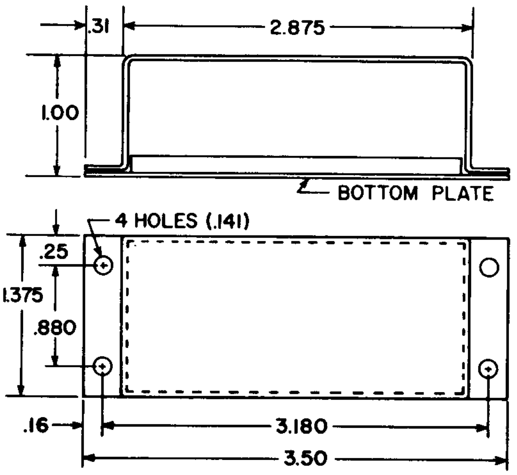


FIGURE 13 — AC1106 NOISE SHIELD & HOLD DOWN BRACKET

A circuit simulator, Model AC 1100 is available to aid in conveniently breadboarding and evaluating various closed loop circuits. The Model AC1100 is made up of a universal plug in socket and a versatile arrangement of 3/4 inch spaced banana jacks for connecting input, output, and power supply voltages and for plug in of feedback components mounted on banana plugs.

4.0 DEFINITION AND MEASUREMENT OF INPUT OFFSET DRIFT

4.1 Offset Drift Sources

An equivalent circuit for offset drift is shown in figure 14.

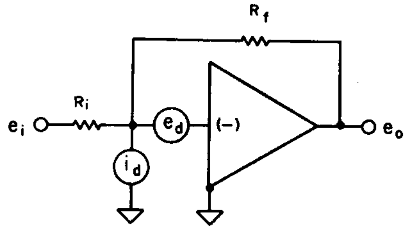


FIGURE 14 - EQUIVALENT INPUT DRIFT GEN.

The drift generators e_d and i_d include the effects of temperature changes, supply voltage changes, time, and internal gradients due to self-heating. Voltage source drift referred to the output is given by

$$\Delta e_o = e_d \left(1 + \frac{R_f}{R_i}\right)$$

and current source drift by

$$\Delta e_o = i_d R_f$$

e_d is determined by calculating the sum of three components; the expected variation in ambient temperature in degrees C times the temperature coefficient of voltage drift in µV/°C, the change in supply voltage in % times the coefficient of voltage drift per % change in supply, and the drift of voltage offset with time.

i_d is determined in a similar way. Total worst case output drift due to the above six drift coefficients is

$$\Delta e_o = e_d \left(1 + \frac{R_f}{R_i}\right) + i_d R_f$$

which, if referred to the input, becomes

$$\Delta e_i = e_d \left(1 + \frac{R_i}{R_f}\right) + i_d R_i$$

where e_d and i_d are implied to each be the sum of three components.

In addition to the above sources of drift within the amplifier, the drift characteristics of the external voltage and current offset adjustment networks must be investigated for extremely critical applications. The change in value of the potentiometers and resistors with time and temperature must be calculated and added to the previous drift calculation.

For an additional analysis of drift sources, refer to "Operational Amplifiers - Part IV" published by Analog Devices.

4.2 Voltage and Current Drift Measurements

The measurement of extremely low levels of voltage and current drift impose a severe problem on instrumentation. A method suitable for amplifying input voltage offsets to levels which can be easily measured or recorded is shown in figure 15. The low value of input resistor reduces the effect of input current drift to a value negligible in relation to the voltage drift. Sensitivity of the circuit is 1mV output per microvolt of drift e_d.

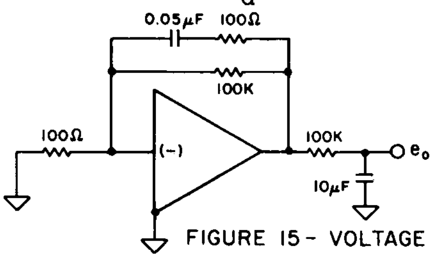


FIGURE 15 - VOLTAGE DRIFT

Figure 16 shows a circuit for measuring input current offset. The 10 megohm resistor produces a voltage due to current offset which is much larger than the effect of voltage offset of the amplifier. Sensitivity of the circuit is 1mV output per picoamp of drift, i_d.

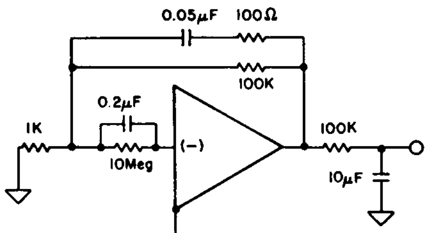


FIGURE 16 - CURRENT DRIFT

5.0 APPLICATIONS AND CIRCUITS

5.1 Inverting Amplifier

The conventional inverting amplifier is illustrated in figure 17. Scaling is accomplished by varying ratio R_f/R_i.

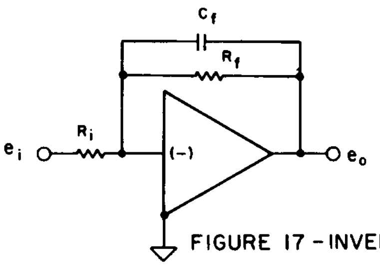


FIGURE 17 - INVERTING AMPLIFIER

Gain of greater or less than unity is practical with typical range from .1 to 5,000. Widest bandwidth and fastest slew rate is obtained when using smallest practical resistor values. See paragraph 3.4 for determining value of C_f. Note that output must drive R_f as well as external load, therefore, output current available for load is reduced by E_{fs}/R_f.

Input impedance equals R_i; output impedance is usually a fraction of an ohm at DC, rising to values around 100 ohms at high frequencies.

5.2 Summing Amplifier

The summing amplifier may be used for scaling at the same time.

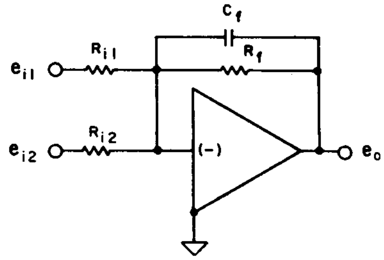


FIGURE 18 - SUMMING AMPLIFIER

$$e_o = -\left(\frac{R_f}{R_{i1}}\right) e_{i1} - \left(\frac{R_f}{R_{i2}}\right) e_{i2}$$

Bandwidth of a multiple input summing amplifier is less than a single input inverting amplifier even though all impedances may be equal, because of the reduction in feedback factor (β). Feedback factor is the attenuation ratio of the operational impedances, which for a two input summing amplifier is

$$\beta = \frac{R_{i1} \,||\, R_{i2} \,||\, R_{IN}}{R_{i1} ||R_{i2}|| R_{IN} + R_f}$$

Output noise voltage and drift are increased over inverter values also for the same reason. In the drift case,

$$e_d \left(1 + \frac{R_f}{R_{i1}||R_{i2}}\right)$$

is the expression for output drift of a two input summing amplifier which may be compared with $e_d(1+R_f/R_i)$ for an inverter.

5.3 Non-Inverting Amplifiers (High Input Impedance)

The single ended input (inverting only) of chopper stabilized amplifiers make special circuitry a requisite for obtaining a non - inverting output from a single amplifier. The general condition is input, output, and power supply grounds cannot all be common. Stated another way: either the input, the output, or the power supply must be floating (isolated) with respect to the remaining two circuit grounds (which two may usually be common).

When an isolated or floating power supply is required, the foremost consideration is the degree of isolation between input power and output power. Resistance values of 1000 megohms and an effective capacitance range of .01 to .1pf are possible in a well built, double shielded transformer, and the floating power supply should have specifications of this magnitude. Some applications may not require the degree of isolation of others, however, the unguarded or effective capacitance and leakage resistance should be known before attempting to use a power supply in a floating configuration.

Voltage Follower (Power Supply Floating)

Here, the regular output of the amplifier is grounded and used for the input and output returns. The output is taken from the floating power supply common terminal. Input impedance is typically 1000 megohms at DC (depending on the isolation of the floating power supply). Bandwidth is full amplifier small signal bandwidth. Offset current compensation if required is in conventional manner (see figure 5) except that 50 K pot is connected across floating power supplies.

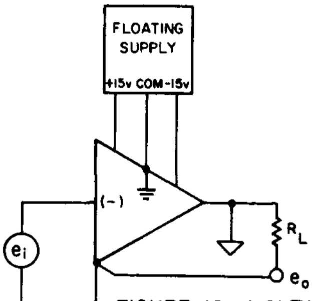


FIGURE 19 - VOLTAGE FOLLOWER

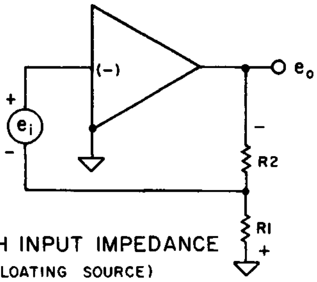


FIGURE 20- HIGH INPUT IMPEDANCE
(FLOATING SOURCE)

Gain of the circuit in figure 20 is $(R_1 + R_2) / R_1$.

Source must be truly floating or improper operation will result. Input impedance is >100 megohms at DC typically. (Actually, for low source impedance, input impedance at DC = $R_{in} \times A\beta$ where A is the DC gain, β is $R_1/(R_1+R_2)$ and R_{in} is the amplifier open loop input resistance).

Offset current may be compensated in the usual manner by injecting current into the amplifier input.

Gain (at DC) is limited to greater than or equal to unity. For gains of less than unity, the inverting connection must be used, or an input voltage divider is required.

Grounded Source (Floating Output and Power Supply)
If a grounded source must be amplified, and the output load and power supply can be floated, the same circuit as the preceeding may be used giving the advantages of high input resistance and high speed.

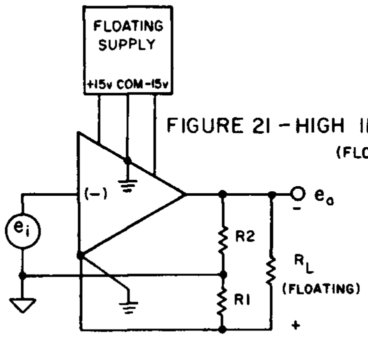


FIGURE 21 - HIGH INPUT IMPEDANCE
(FLOATING LOAD)

Two Amplifier High Input Impedance Circuits

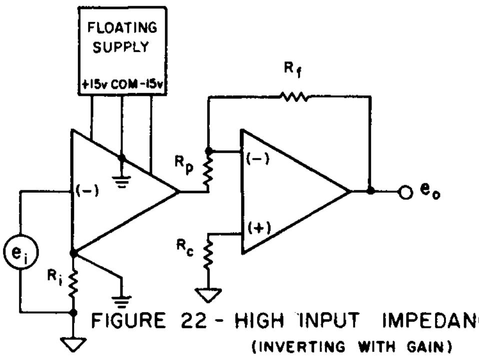


FIGURE 22 - HIGH INPUT IMPEDANCE
(INVERTING WITH GAIN)

Characteristics of the circuit of figure 22 are high input impedance, wide bandwidth, and high gain if desired together with a grounded load. Gain is the ratio of R_f/R_i, R_p is a small value for overload current limiting and R_c is the usual parallel equivalent of R_p and R_f for offset drift compensation.

The input voltage, e_i, appears across R_i producing a current e_i/R_i which must flow out of the amplifier output, since the power supply is floating. This current flows into the summing junction of the output amplifier and appears across R_f, thus the gain is

$$e_o = \frac{e_i}{R_i} \times R_f = e_i \times \frac{R_f}{R_i}$$

Models 106 and 108 will be satisfactory for the output amplifier in most applications since drift of the output is divided by R_f/R_i referred to the input, e_i. It is important to keep resistor values very low; preferably R_f should be 2K or lower although output current limitations set a limit on the size.

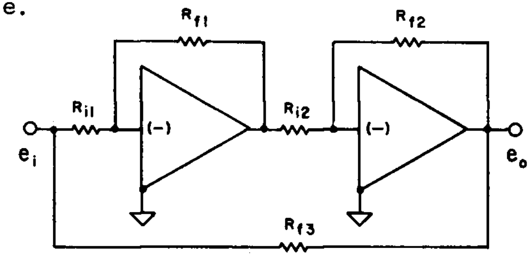


FIGURE 23 - HIGH INPUT IMPEDANCE
(BOOTSTRAPPED INPUT - GROUNDED SUPPLY)

The circuit of figure 23 provides a high input impedance because of the positive feedback to the input through R_{f3}. Gain may be equally divided between the two amplifiers or may be taken mostly in the first amplifier therefore reducing the drift requirements of the second amplifier, in this case R_{f3} is chosen $\geq (R_{f1} \times R_{f2})/R_{i2}$.

Maximum input impedance occurs when R_{f3} is exactly equal to $(R_{f1} \times R_{i2})/R_{i2}$. Source impedance must be small with respect to R_{f3} to prevent oscillations from occurring.

Overload Current Limiting

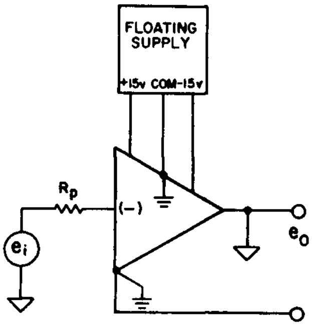


FIGURE 24 - OVERLOAD CURRENT LIMITING

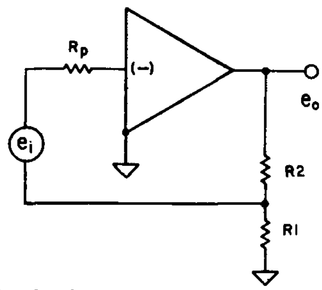


FIGURE 25 - OVERLOAD CURRENT LIMITING
(FLOATING SOURCE)

The voltage follower and non-inverting configurations require a current limiting resistor at the input (figures 24 and 25) for proper operation of the high speed overload circuit.

R_p should be the smallest value possible commensurate with anticipated overload to avoid additional phase shifts . Again, the extra diode limiter may be of value (figure 26) .

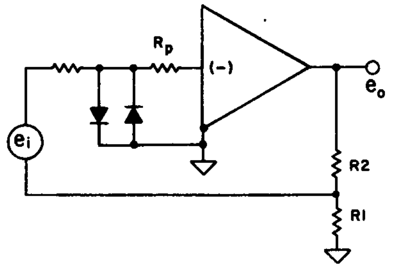


FIGURE 26 - ADDITIONAL CURRENT LIMITING
(DIODE LIMITER)

5.4 Differential Input Configuration

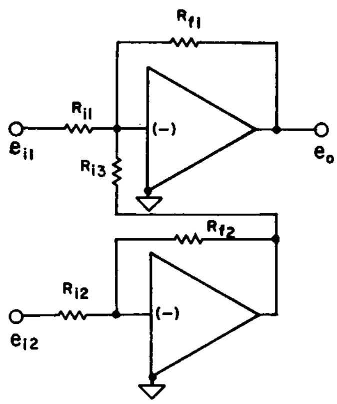


FIGURE 27- DIFFERENTIAL INPUTS

Figure 27 shows an obvious method of forming a differential connection, and is probably the simplest and easiest to use. An extension of the voltage follower is shown in figure 28. The circuit is suitable for gains of unity to 50 or so, but higher gains become noisy, since the input noise of all three amplifiers is additive (at quadrature for uncorrelated noise sources).

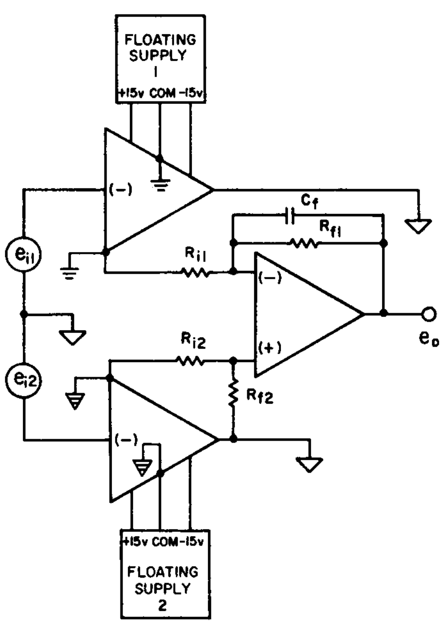


FIGURE 28- HIGH INPUT IMPEDANCE DIFFERENTIAL INPUT

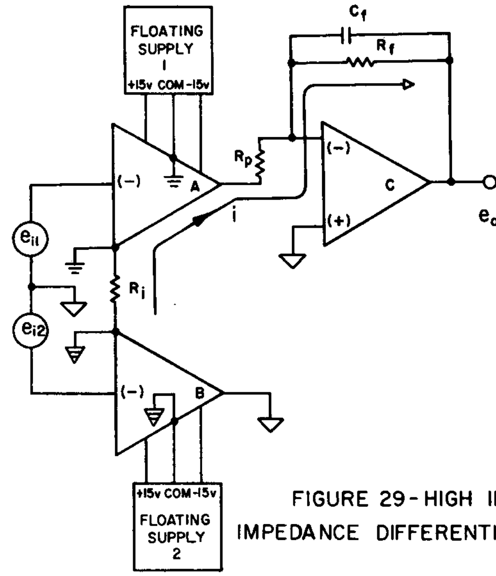


FIGURE 29- HIGH INPUT IMPEDANCE DIFFERENTIAL INPUT

The circuit of figure 29 is suitable for gains of unity to several thousand. Operation is as follows: the voltage $e_{i1} - e_{i2}$ appears across R_i (as if the two input amplifiers were voltage followers), the current $(e_{i1} - e_{i2})/R_i$ flows out of A's output into the summing junction of C and through R_f (R_p is only for overload protection). Therefore gain is R_f/R_i. Input impedance is hundreds of megohms, being limited usually by the isolation of the floating power supplies. Amplifier C can be any non-stabilized differential amplifier with suitable bandwidth and stability characteristics (Model 203 for A and B and Model 108 for C are suggested for low noise, moderate bandwidth).

With Models 202 or 210, bandwidths of hundreds of kilocycles may be obtained with extremely high input resistance at DC. For the overload protection to be operative, small resistors must be incorporated in series with each input. Care must be exercised in stabilization of this circuit due to the extremely large forward gain of the total configuration.

5.5 Integrating Amplifier

The exceptional voltage and current drift specifications, together with high open loop gain make the 200 Series amplifiers particularly well suited for use as integrators. Model 203 is designed especially for integrating applications and is recommended in all cases except where the bandwidth of Models 202 or 210 is required or possibly where the lower cost and more modest performance of the Model 210 is acceptable.

The specifications of importance for an integrator are voltage offset and drift, current offset and drift, open loop gain, and input resistance. Drift rate of the integrator is obviously related to voltage and current drifts; additional droop of a held value (and non-linearity during integration) results from the finite values of input resistance and DC gain. The measure of droop (in addition to offset contributions) is calculated by assuming a resistor in parallel with the integrating capacitor equal to R_{IN} × DC gain and computing the discharge rate.

The main causes of integrator error are voltage and current offsets and change of offset. Initial offset can be adjusted to zero, but drift (change of initial offset) with time, temperature, and supply voltage, cannot be compensated for and must be specified sufficiently low to maintain system accuracy. Voltage offset and current offset at any particular temperature can be compensated for at the same time with either a voltage offset trim or current offset trim as long as R_i is constant. If R_i must change in value (be switched in value for example) or if integration is stopped by open circuiting the source ($R_i \to \infty$) instead of grounding the input, then voltage and current offsets must be adjusted independently with separate trim pots for most accurate results.

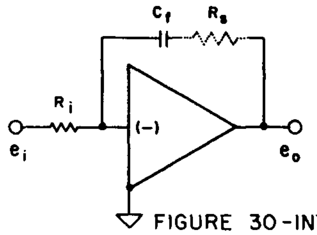


FIGURE 30-INTEGRATING AMPLIFIER

The conventional circuit is shown in figure 30. For a constant held input voltage, output rate of change is $1/R_iC_f$ volts/second per volt of input signal.

Voltage drift rate due to voltage offset is

$$\frac{\Delta e_o}{\Delta t} = \frac{e_d}{R_iC_f}$$ and for current offset is

$$\frac{\Delta e_o}{\Delta t} = \frac{i_d}{C}$$, where e_d and i_d are defined in 4.1.

To minimize total drift for any given time constant, R_iC_i observe that voltage drift is dependent only on the product R_iC_f, and cannot be minimized except by increasing the time constant or choosing a better amplifier. However, current drift can be minimized by choosing C as large as possible.

A small resistor, R_s, in series with C_f, may be required for proper stabilization with wide bandwidth amplifiers.

5.6 Photomultiplier Tube Amplification And Current To Voltage Converters

Photocells, photomultipliers, ionization chambers and a host of other scientific instruments have extremely high effective source impedances and low output currents. Devices of this nature approach an ideal current generator in performance and can be treated as such for choosing circuit approaches.

An instrumentation method usual in the past has been to insert a large value resistor in series with the output of the current source, and measure the voltage drop across this resistor with an amplifier having very high input impedance and very low offset current.

A circuit configuration that in many instances will provide superior results is shown in figure 31.

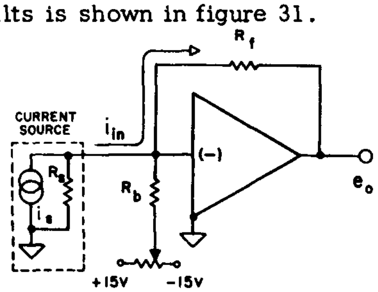


FIGURE 31- CURRENT to VOLTAGE CIRCUIT
(PHOTOMULTIPLIER ect.)

This is the familiar inverting amplifier used for current to voltage conversion. A current i_{in} is produced by the source, (and neglecting for the moment R_b) almost all of i_{in} will flow through the feedback resistor, R_f, since the input current to actuate the amplifier is extremely small. The output voltage will then be $i_{in} \times R_f$. The transfer function for the circuit is in the form of a transimpedance, $e_o/i_{in}=R_f$. The function of R_b is to offset any initial current from the source, for example, dark current in a photomultiplier tube, and to compensate for the small offset current of the amplifier. Notice that R_b does not load the current source, for the input to the amplifier is a current summing junction and therefore a virtual ground.

Advantages of this arrangement compared to a high input impedance voltage follower amplifying the signal produced across a load resistor are as follows:

1. Linearity of most transducers is improved; current sources show maximum linearity working into a voltage node. If linearity errors arising from large values of load resistance are attempted to be improved by reducing the load resistor and increasing voltage gain of the amplifier, then voltage drift and voltage noise are increased by the amount of the voltage gain.
2. Common mode voltage errors of a voltage follower circuit are eliminated.
3. Initial offset currents much larger than the dynamic range of the signal input are easily zeroed by choosing a value of $R_b < R_f$.
4. Input impedance changing with time and temperature can produce appreciable error in the voltage follower.

Noise and drift at the output will be due primarily to current noise and drift. Input voltage noise and drift are not amplified in a transimpedance and in a practical case, the output contribution due to the equivalent input voltage noise and drift generators will be at most twice the input values.

Output current noise and drift are as usual $i_n R_f$ and $i_d R_f$ respectively. For best signal to noise characteristics, a feedback capacitor should be included to limit the bandwidth to that required for system performance.

5.7 Precision Voltage Source

A precision variable (or fixed) voltage reference source is illustrated by figure 32.

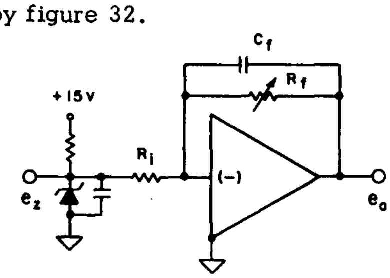


FIGURE 32- PRECISION VOLTAGE SOURCE

This arrangement provides a linear relation between R_f and output voltage so that a calibrated variable resistance will produce a calibrated output voltage. A temperature compensated zener reference can be used for precision stability or a conventional 5.6 volt zener for moderate stability requirements. Various switching schemes can be devised to obtain bi - polar outputs. R_i is usually between a few thousand ohms and a hundred thousand ohms. C_f should be a low leakage capacitor as large as practical. Model 203 is recommended for this application. See paragraph 3.6 if large capacitive loads must be driven.

5.8 Current Sources

For current source applications it should be noted that the Model 201 has a hefty 100ma output rating while the Models 202, 203, and 210 all produce 20ma.

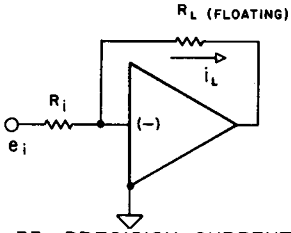


FIGURE 33 - PRECISION CURRENT SOURCE

Figure 33 shows a precision current source with a requirement that the load must be floating.

Current through R_L is e_i/R_i. Limitations on size of R_L are maximum voltage drop across R_L and maximum current through R_L to be within amplifier output specifications.

By using a floating power supply, a precision current source may be constructed as in figure 34.

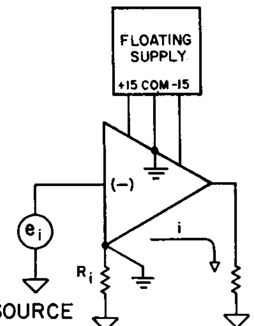


FIGURE 34 - CURRENT SOURCE
(FLOATING SUPPLY)

The input voltage, e_i, appears across R_i, producing the current i. Because of the floating power supply this current in turn must flow through R_L for a return path. Note that if e_i is about 10 volts, the output can swing from 0 to +20 volts, giving a greater range than other techniques.

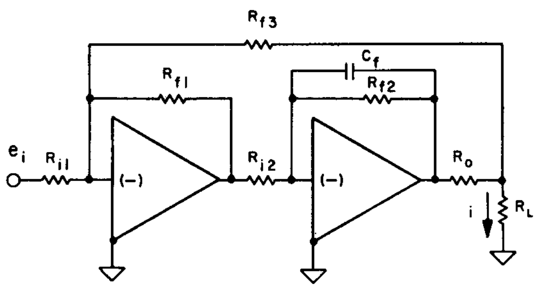


FIGURE 35 - CURRENT SOURCE
(GROUNDED SUPPLY)

The circuit of figure 35 produces a current of

$$\frac{R_{f3} \times E_1}{R_{i1} \times R_o}$$

through R_L with the requirement that $R_{f3} >> R_L$.

Precisely this current flows through R_o, but a small amount must be channeled through R_{f3}, therefore the current in R_L is low by the amount in R_{f3}.

If Models 201, 202, or 210 are used, feedback capacitors will be required as usual. With Model 203, a feedback capacitor on the second amplifier will reduce noise and provide additional stability for the circuit. Suggested values are $R_{i1} = R_{f1} = R_{i2} = R_{f2} = R_{f3} =$ 100K and $R_o = 100\ \Omega$ for 10ma output per volt input or R_o=1K for 1ma output per volt input.

5.9 Bridge Amplifier (Floating Bridge Supply)

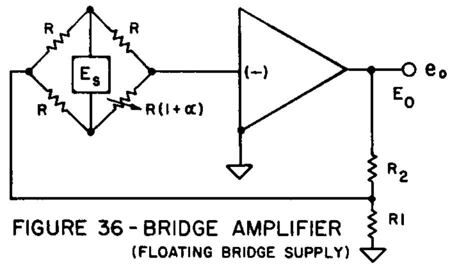


FIGURE 36 - BRIDGE AMPLIFIER
(FLOATING BRIDGE SUPPLY)

Figure 36 illustrates the use of a non-inverting high input impedance configuration for amplifying a low level bridge signal. Output for a single active arm is

$$e_o = \left(\frac{R_1 + R_2}{R_1}\right) \left(\frac{E_s}{2(2+\alpha)}\right)$$