OPERATIONAL AMPLIFIERS AND THEIR APPLICATIONS

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INTRODUCTION TO OPERATIONAL AMPLIFIERS

Functionally speaking, an operational amplifier is a device which, by means of negative feedback, is capable of processing a signal with a high degree of accuracy limited primarily only by the tolerances in the values of the passive elements used in the input and feedback networks.

Electronically, an operational amplifier is simply a high-gain amplifier designed to remain stable with large amounts of negative feedback from output to input.

General-purpose operational amplifiers, useful for linear amplification with precise values of gain, and for accurate integration and differentiation operations, have low output impedance and are DC-coupled, with the output DC level at ground potential.

The primary functions of the operational amplifier are achieved by means of negative feedback from the output to the input. This requires that the output be inverted (180° out of phase) with respect to the input. The conventional symbol for the operational amplifier is the triangle shown in Figure 1-a. The output is the apex of the triangle; the input is the side opposite the output. Negative feedback, through a resistor, capacitor, inductor, network or non-linear impedance, designated "Z_f" is applied from the output to the input as shown in Figure 1-b. The input to which negative feedback is applied is generally termed "− input"* or "− grid" (in the case of vacuum-tube operational amplifiers).

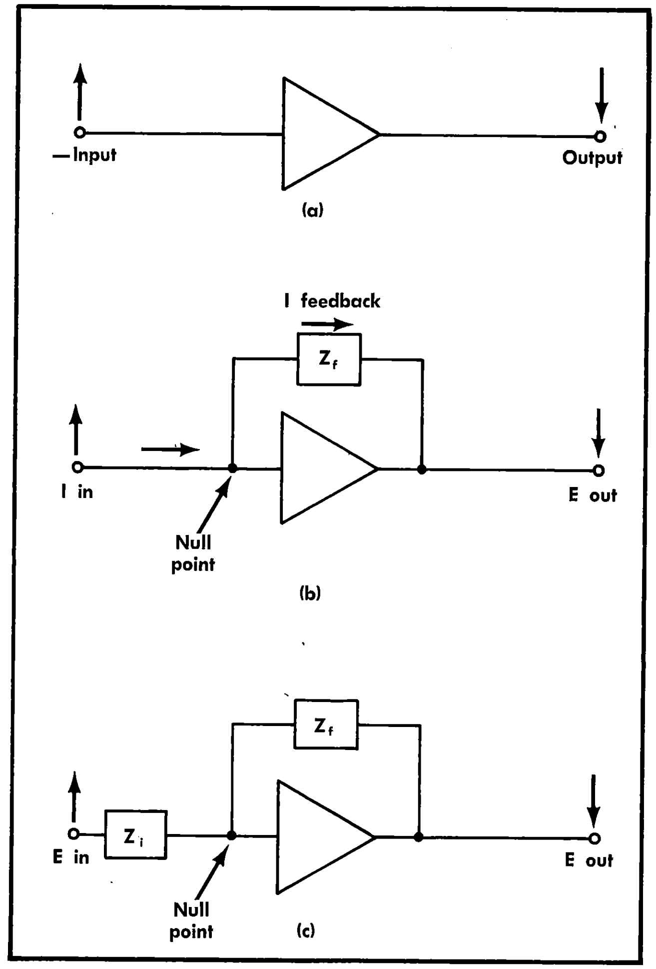


Figure 1. Conventional Operational Amplifier Symbols.
(a) The input is to the base of the triangular symbol, the output is from the apex opposite. The − input and output are out-of-phase (arrows).
(b) Feedback element Z_f provides the negative feedback to permit high-accuracy operations. The amplifier seeks a null at the input by providing feedback current through Z_f equal and opposite to the input current I_{in}. Output voltage is whatever is necessary to provide required balancing current through Z_f.
(c) Input element Z_i converts a voltage signal (E_{in}) to current, which is balanced by current through Z_f.

Operational Amplifier Seeks Current Null at − Input

An operational amplifier, using negative feedback, functions in the manner of a self-balancing bridge, providing through the feedback element whatever current is necessary to hold the − input at null (ground potential). See Figure 1-b. The output signal is a function of this current and the impedance of the feedback element.

The − input, held to ground potential by the feedback current, appears as a very low impedance to any signal source. Using resistive feedback, for instance, the input appears to be the resistance of the feedback element, divided by the open-circuit gain of the operational amplifier.

If current is applied to the − input, it would tend to develop voltage across the impedance of the feedback element, and move the − input away from ground potential. The output, however, swings in the opposite direction,

* **The operational amplifiers of the Tektronix Type O Plugin unit also provide access to a non-inverting input. Uses of this "+ input" or "+ grid" are discussed later.**

providing current to balance the input current and hold the − input at ground. If the impedance of the feedback element is high, the output voltage must become quite high to provide enough current to balance even a small input current.

Input Element Z_i Converts Input Signal to Current

Since we more often have to deal with voltage rather than current signals, an additional element is used in most operational amplifier applications, designated "Z_i" (input impedance). This is an impedance placed in series with the − input, converting into **current** that parameter of the input signal which we want to appear as voltage at the output (Figure 1-c).

If Z_i and Z_f are both resistors (Figure 2), the operational amplifier becomes a simple voltage amplifier, the gain of which is $-Z_f/Z_i$.

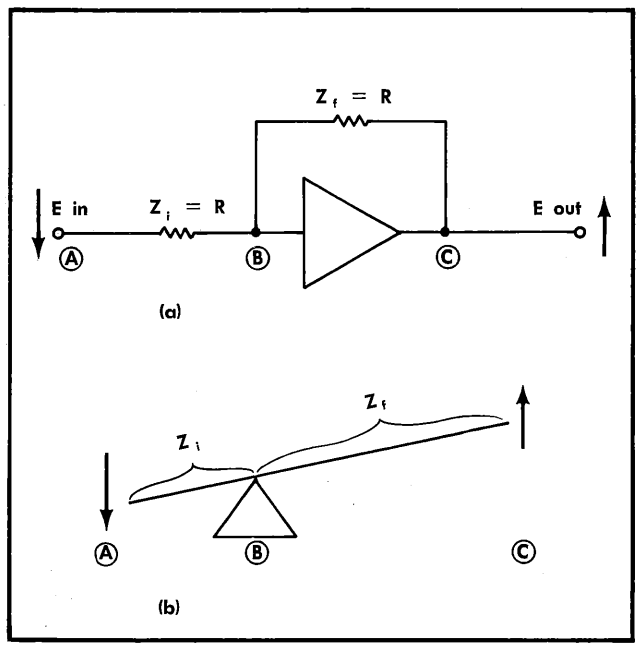


Figure 2.
(a) Operational amplifier using resistors for both Z_i and Z_f becomes fixed-gain linear amplifier. Gain is $\frac{-Z_f}{Z_i}$.
(b) "See-Saw" operation of operational amplifier. System appears to pivot about a fulcrum (the null point *B*) whose "location" is determined by Z_f/Z_i.

Let's examine the mechanism by which this works. Referring again to Figure 2, we apply a voltage to point *A*, causing current to flow through Z_i. Were it not for the operational amplifier, this current would also flow through Z_f and to ground through the low impedance at point C, making Z_i and Z_f a voltage divider, and raising the voltage at point *B*. However, the operational amplifier operates to hold the voltage at point *B* (the − input) at ground potential. To do this, it must supply at point C a voltage which will cause a current to flow through Z_f which will just balance the current in the opposite direction flowing through Z_i. When point *B* is thus held at ground potential, the voltage across Z_i is obviously equal to the applied voltage at *A*.

Output Voltage is Input Current × Impedance of Z_f

The current through Z_i is equal to the applied voltage at *A* divided by the impedance (in this case, resistance) of Z_i, or E_{in}/Z_i. This same value of current must flow in the opposite direction through Z_f in order to keep point *B* at ground. The voltage at point C, then, must be E_{in}/Z_i (which is the value of the current in Z_f) multiplied by Z_f. The output is inverted (of opposite polarity) from the input, so we say that $E_{out} = (-E_{in})\left(\frac{Z_f}{Z_i}\right)$, and the voltage gain of this amplifier configuration is seen to be $\frac{-Z_f}{Z_i}$.

See-Saw Operation

As indicated in Figure 2-b, the operational amplifier with resistive input and feedback elements acts in see-saw fashion, the amplifier moving the output end of the see-saw in response to any motion of the input end, causing the system to pivot about an imaginary fulcrum, which is the "sensing point" (− input). The distance from the near end to the sensing point or fulcrum corresponds to the Z_i or input resistor, and the distance from the fulcrum to the far end corresponds to Z_f. The motion of the far end depends on the motion of the near end and the ratio of the two distances. This analogy suggests that the operational amplifier may be used to solve dynamic problems in mechanical engineering, and so it can. One of the principal uses of operational amplifiers has been in the rapid solution of complex mechanical or hydraulic problems by means of electronic analogs of mechanical or hydraulic systems: operational amplifiers are the basic components of an analog computer.

As may be expected, simple linear voltage amplification by precise gain factors is, though useful, not by any means the limit of the operational amplifier's capabilities.

Capacitor as Z_i Senses Rate-of-Change

Remembering that an operational amplifier with a resistor as a feedback element responds with an output voltage equal to the product of the input current and the feedback resistance, let's consider what happens if a capacitor is used instead of a resistor as Z_i (Figure 3).

The current through a capacitor is proportional to the **rate-of-change** of the voltage across the capacitor. A steady state DC voltage across a capacitor (assuming an "ideal" capacitor) passes no current througn the capacitor, so no balancing current need be furnished by the output to hold the − input of the operational amplifier at ground. The output voltage then, is zero.

If the voltage at the input is changed, however, the **change** causes a current to flow through capacitor Z_i. The **amount** of current that flows is directly proportional to

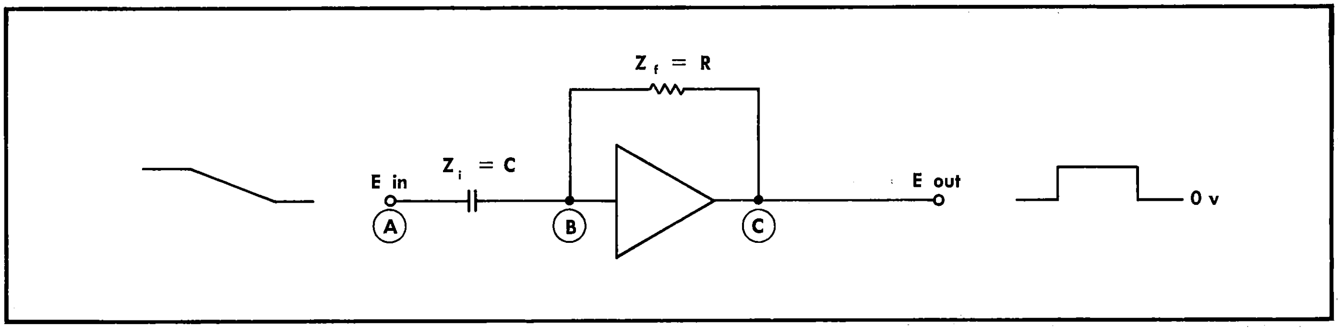


Figure 3. Operational Amplifier as Differentiator. Output is proportional to *rate-of-change* of input voltage. $E_{out} = \frac{-dE_{in}}{dt} \times RC$.

the capacitance of Z_i times the **rate of change** of the input voltage.

Let's assume that the potential at point *A* is + 100 v DC, and that we change it smoothly to + 95 v DC in five seconds. This represents a rate of change of one volt per second, the change taking place over a period of five seconds. If the value of Z_i is 1 μf, then, a current of − 1 microampere will flow through Z_i for those 5 seconds.

The operational amplifier will cause an equal and opposite current to flow in Z_f. If we select a value of 1 megohm for Z_f, the one microampere current necessary to balance the circuit will require + 1 v to appear at the output of the operational amplifier, during the time that 1 μa current flows through the capacitor.

This operation is **differentiation:** sensing the **rate-of-change** of an input voltage, and providing an output voltage proportional to that rate of change.

The actual relationship of output to input is this: $E_{out} = -\left(\frac{dE_{in}}{dt}\right)$ (RC), where the expression $\frac{dE_{in}}{dt}$ indicates the rate of change (in volts per second) of the input signal at any given instant, and *R* and *C* are Z_f and Z_i respectively.

In our example, we used a constant rate of change, and obtained a constant voltage level out. Had the rate been less even, the output signal would have demonstrated this dramatically with wide variations in amplitude. The differentiator senses both the rate and direction of change, and is very useful in detecting small variations of slope or discontinuities in waveforms.

Differentiator Has Rising Sine Wave Response Characteristic

In responding to sine-waves, the differentiator has a rising characteristic directly proportional to frequency, within its own bandwith limitations (see chart page 6). The output voltage is equal to (E_{in}) (2πf RC), and the output waveform is shifted in phase by − 90° from the input (the phase shift across the capacitor is actually + 90°, but the output is inverted, shifting it another 180°).

Capacitor as Z_f Senses Input Amplitude and Duration

If we interchange the resistor and capacitor used for differentiation, and use a resistor for Z_i and a capacitor for Z_f (Figure 4) we obtain, as might be expected, the exact opposite characteristics from those obtained above. While in differentiation we obtained an output voltage proportional to the rate of change of the input, by swapping the resistor and capacitor, the output signal becomes a rate of change which is proportional to the input voltage.

This characteristic allows us to use the operational amplifier for integration, since the instantaneous value of output voltage at any time is a measure of both the amplitude and duration (up to that time) of the input signal—to be exact, a sum of all the amplitudes, multiplied by their durations, of the input waveform since the start of the measurement.

Here's how integration works: Let's assume the conditions of Figure 4 (Z_i = 1 meg, Z_f = 1 μf), and an input signal level of zero volts. No current flows through Z_i, so the operational amplifier needs to supply no balancing current through Z_f. Suppose now we apply a DC voltage of − 1v to Z_i. This will cause a current of − 1 μa to flow in Z_i, and the operational amplifier will seek to provide a balancing current through Z_f. To obtain a steady current of + 1 μa through 1 μf, the operational amplifier will have to provide a continually rising voltage at the output, the rate of rise required being 1 volt per second. It will continue to provide this rate of rise until the input voltage is changed or the amplifier reaches its swing limit ("bottoms out"), or approaches its open-loop gain.

Now, this rate-of-rise, though helpful in understanding the mechanism by which the operational amplifier performs integration, is not the "answer" we seek from an integrator. The significant characteristic is the exact voltage level at a certain time, or after a certain interval.

Integrator Holds Final Level Until Reset

Before the amplifier reaches its output limit, suppose we remove the input voltage to Z_i. The output does not return to ground, but remains at the level it reached just before the signal was removed. The rate of rise has stopped because the necessity for providing + 1 μa through Z_f to maintain the null at the − input has been

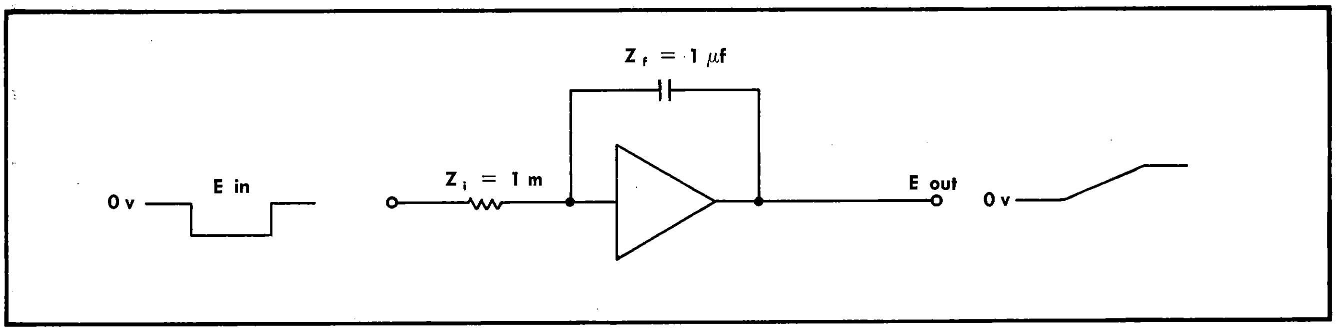


Figure 4. Operational Amplifier as Integrator. Output *rate of change* is proportional to input voltage. $\frac{dE_{out}}{dt} = \frac{-E_{in}}{RC}$, or $E_{out} = \frac{-1}{RC}\int E_{in}\,dt$. RC in the example here is 1 second. Output, then, is 1 volt per second per volt input, and—most important—the output level at anytime is *one volt per volt-second input.*

removed. With an ideal capacitor and amplifier, the output voltage would remain at the last level reached indefinitely, until an input signal of the opposite polarity was applied to Z_i, and a negative-going rate of change at the output was required to maintain the null at the − input.

If the positive input signal is greater than our original − 1 volt, it will take less time for the output voltage to reach zero than it originally took to rise. If the positive signal is smaller, it will take more time.

The absolute output level of the integrator at the end of some interval is the sum of the products of all the voltages applied to Z_i since the output was at zero, times the durations of these voltages, that sum divided by − RC.

Interpreting Answers Obtained From Integrator

The mathematical expression for the output level reached in a given interval of time (T_2 − T_1) is as follows:

$$E_{out} = \left(\frac{-1}{RC}\right)\int_{T_1}^{T_2} E_{in}\,dt$$

The integral sign indicates that the value to be used is the **sum** of all of the products (E_{in} X dt) shown, between the limits (T_1, T_2) noted. The expression "dt" indicates infinitely small increments of time.

It is not necessary, however, to understand and be able to manipulate expressions in integral calculus to understand and make use of an operational amplifier integrator.

The integrator provides a voltage output proportional to the net number of volt-seconds applied to the input. If the total volt-seconds of one polarity is equalled by those of the opposite polarity, the output level at the end of the selected interval will be zero. Let's look at some examples.

Simple Example of Data From Integrator

First, we'll assume the signal we want to integrate is a simple one-volt positive pulse of one second duration (Figure 5). The sum of all voltages times durations between T_1 and T_2 is one volt-second. Using 1 megohm and 1 microfarad for Z_i and Z_f, the operational amplifier output will fall at the rate of one volt per second $\left(\frac{-E_{in}}{RC}\right)$ for one second, reaching − 1 v when the pulse ends, and remaining at that level.

In reading this output level at T_2 we know that the input signal has amounted to 1 volt-second during the interval T_1 to T_2. Note also that a later observation, at T_3, gives the same answer, since E_{in} has been 0 between T_2 and T_3.

More Complex Cases

Now, take the more complicated case of the waveform in Figure 6-a. Its four voltage levels, of different duration, cause the integrator output to fall at four different rates, reaching a final level representing the total number of volt-seconds contained in the waveform. It should be apparent now that the integrator can measure the total volt-seconds contained in even the very complex waveform of Figure 6-b—something that would be difficult to measure by direct observation of the waveform. This type of operation is often referred to as "taking the area under the curve," since the area underneath a waveform plotted against time (i.e., the area bounded by T_1, T_2, the waveform and the line representing 0 volts) is the number of volt-seconds involved. Note, too, that we needn't wait for T_2 to obtain a reading: the instantaneous value of E_{out} at **anytime** is proportional to the input volt-seconds up to that time.

Using Different Values of R and C

In the cases we've used for illustration, RC was 1 ($10^6 \times 10^{-6}$), and the numerical value of the output voltage at the end of the integrating interval was the number of volt-seconds in the input waveform. Using other values of R and C requires some additional calculation. To find the actual input volt-seconds, multiply the output voltage by (− RC). Example: R is 200 k, C is .01 μf and the output voltage after the selected interval is − 2.5 volts. Multiplying − 2.5 by ($-2 \times 10^5 \times .1 \times 10^{-8}$) gives us 5×10^{-3}, or 5 millivolt-seconds, positive polarity. Note that because of the polarity-reversal in the amplifier, we multiply by (− RC), to obtain the proper sign in the answer.

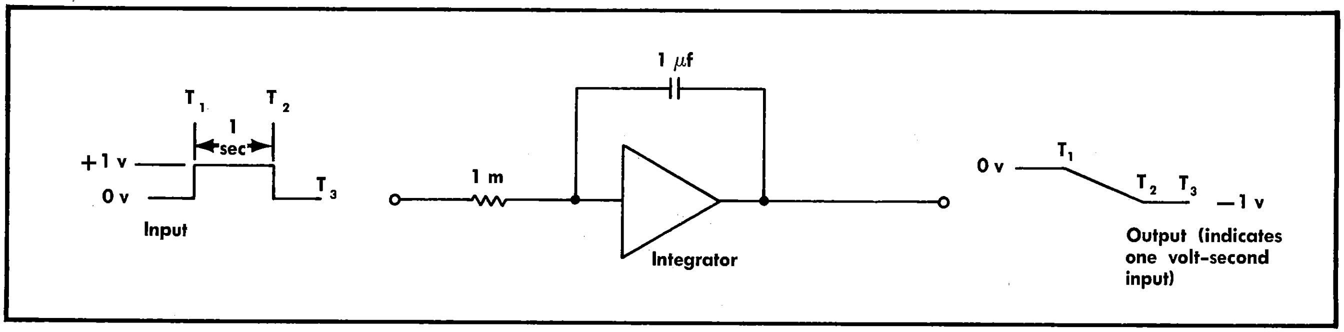


Figure 5. Simple case of integrating 1-volt-second pulse. Integrator does not improve measurement accuracy in so simple a case.

Measuring Ampere-Seconds (Coulombs)

To measure ampere-seconds, Z_i is omitted, and the current source is applied directly to the − input. The output level reached in a given time $(T_2 - T_1)$ is $\frac{-1}{C}\int_{T_1}^{T_2} I_{in}\,dt$.

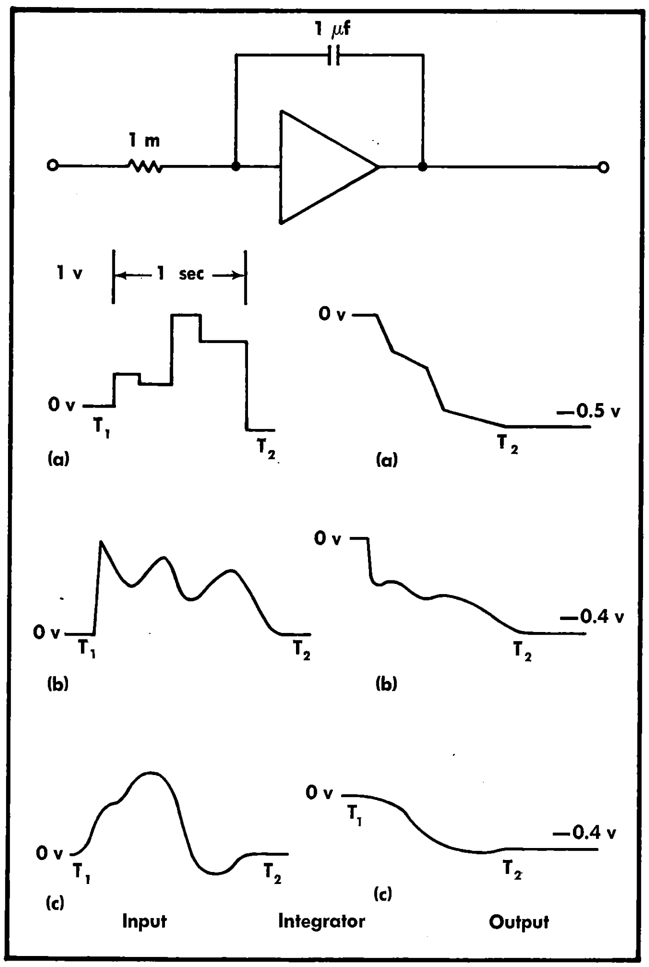


Figure 6. Integrating more complex waveforms to determine "area under the curve" between T_1 and T_2. Note that in (c) the negative portion of the input waveform *reduces* the net integral.

Integrator Response to + and − Signals

If a waveform to be integrated contains both positive and negative polarity portions during the integrating interval, the output will be proportional to the **difference** between the volt-seconds of each polarity, the integrator being an averaging device. If it's desired to add the two polarities instead of allowing them to be subtracted, it is necessary to precede the integrator with an "absolute-value amplifier" (full wave rectifier) which inverts one of the polarities.

Necessity to "Reset" Integrator After T_2

The "integrating interval" (T_1 to T_2) has been mentioned several times. Because we frequently deal with repetitive signals, and continued integration of a waveform which is not perfectly symmetrical with respect to zero volts will eventually drive the operational amplifier to its output voltage limit, it's desirable to have some way of returning the output to zero at or after T_2, the end of the desired interval.

For slow work, a pushbutton which can be used to discharge Z_f manually is usually sufficient. Other circuits which may be used to perform this function automatically are shown in the applications section, page 2-4. Where the integrating interval is quite short, RC networks may be placed around Z_f to return the output level to 0 v through a time constant much longer (e.g., 100 ×) than the integrating interval.

In the Type O Unit, the "Integrator LF Reject" switch positions perform this function whenever Z_f is set to a capacitive value.

Since the LF Reject circuit operates continually to return the integrator output to zero, it is necessary not only to keep the integrating interval short with respect to the LF Reject time-constant, but also to measure E_o before it has had a chance to decay, whenever these circuits are used. The value of resistors used in the circuit will also limit the maximum output obtainable for any given amplitude input $\left(Max\ \frac{E_{out}}{E_{in}} \approx \frac{R_r}{Z_i}\right)$, where R_r is the resistance of the LF Reject circuit.

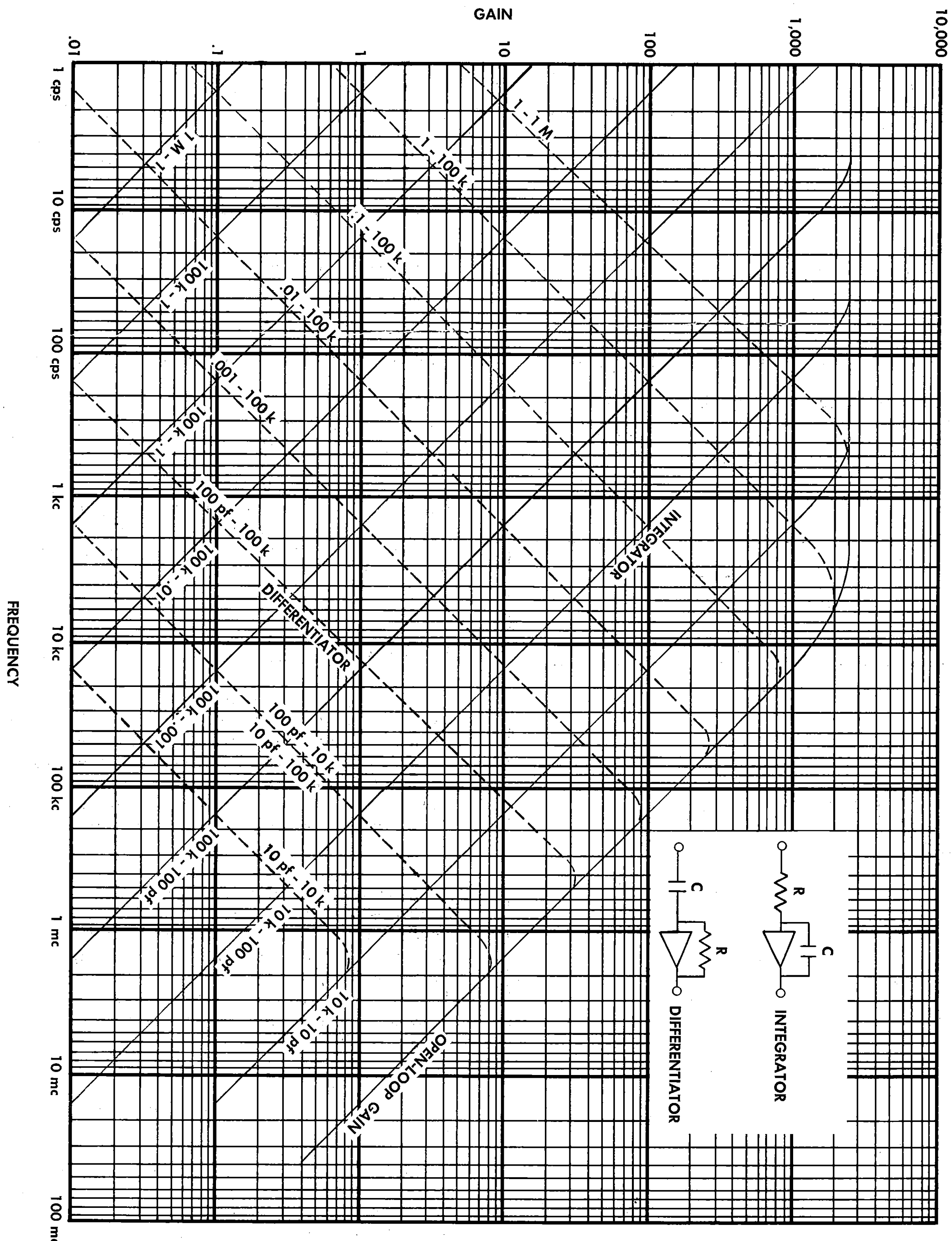


Figure 7. Average Gain-Frequency characteristics for integration and differentiation.

Reset or LF Reject Imperative When $Z_f = C$ is Small

Use of resetting or LF reject circuits is usually imperative when small values of C are used for Z_f, since the small amount of grid current which flows in the − input grid even in the absence of an input signal is sufficient to cause a relatively rapid rise in output voltage as the operational amplifier tries to hold the − input null with balancing current through Z_f.

Response of Integrator to Sine Waves

For sine-waves, the gain of the integrator varies inversely with frequency, the actual gain being $\frac{-1}{2\pi f\ RC}$, except as limited by the open-loop DC gain (at low frequencies) and the open-loop gain-bandwidth product at high frequencies (see Figure 7). At low frequencies, the gain becomes less than the formula would indicate, the effect becoming noticeable at approximately the point where the formula indicates a gain of approximately 1/3 the open loop gain. At high frequencies, the error becomes significant above approximately 1/10 of the open-loop gain-bandwidth product. Except as limited above, the integrator shifts the phase of the input sine wave by +90°.

Use of The + Input

Many operational amplifiers (including those in the Type O unit) provide access to a non-inverting input, referred to as the + grid or + input. A positive-going signal injected at this point produces a positive-going signal at the output. Conventional identification of + and − inputs is shown in Figure 8.

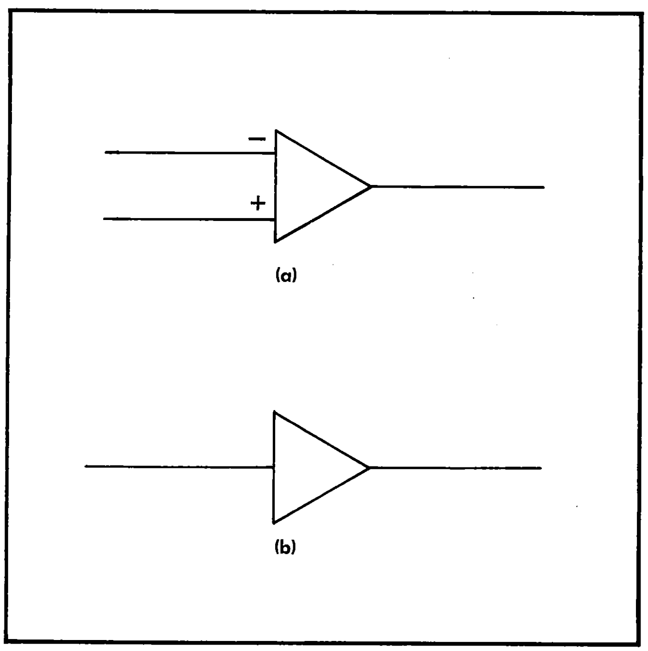


Figure 8. Identification (a) of + and − inputs of an operational amplifier. If only one input is shown (b), it is always assumed to be the − input.

If the output is connected directly to the − input, the operational amplifier becomes a non-inverting gain-of-one voltage amplifier for a signal applied to the + grid, with very high input impedance and very low output impedance.

Non-Inverting Amplifier With Gain > 1

With less than 100% negative feedback (Figure 9), obtained by putting the − input on a voltage divider between the output and ground, gains of greater than one may be realized, the actual gain being $\frac{R_i + R_f}{R_i}$, or 2 where $R_i = R_f$.

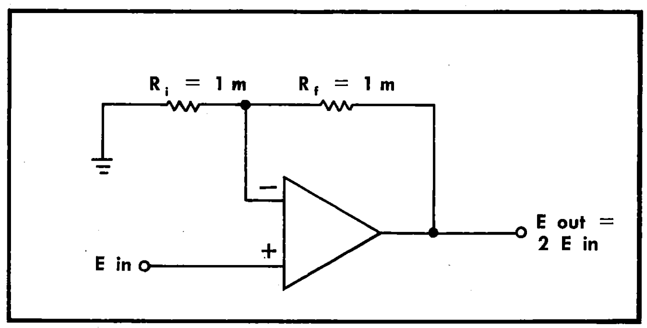


Figure 9. Gain of Two Using + Input. Very high input resistance (> 10^9 Ω) for signals on the order of 1 v amplitude is possible. Other values of gain may be obtained using different ratios of R_i and R_f.

Feedback applied to the + input from the output is positive feedback, which tends to raise the input impedance of the + input toward infinity as the amplitude of the feedback approaches the amplitude of the input signal. If the loop gain (feedback amplitude compared to signal amplitude) exceeds 1 for any frequency, the amplifier becomes unstable (negative input resistance) and will oscillate at that frequency. If the loop gain exceeds 1 at DC, the amplifier will swing to its output voltage limit and stay there. The + input is useful for applications combining positive and negative feedback, and for use of the operational amplifier as an oscillator, waveform generator or multivibrator. The + input may also be used to provide a balanced or differential input, in which the operational amplifier responds only to the instantaneous **difference** between the signals applied to the + and − inputs. Other uses are suggested in the applications section.

Operational Amplifier Limitations

In performing linear operations with an operational amplifier, it is necessary to recognize and allow for the limitations of the amplifier and technique used, to obtain accurate results. The chief limitations are:

1. Open-loop gain.
2. Gain-bandwidth product.
3. Grid current (chiefly of concern during integration).
4. Output current and voltage capability.
5. Signal source impedance.

Introduction

1. Open Loop Gain

The accuracy of all operations is ultimately limited by the open-loop gain of the amplifier, which determines how closely the amplifier is capable of holding the − input null. An amplifier with infinite gain would provide a null of exactly 0 volts, and the impedance at the − input (using feedback) would be exactly 0 ohms.

With finite gain, the − input does not quite null, and does not appear as 0 ohms. With an open-loop gain of A*, the − input moves $\frac{1}{A}$ times the output voltage swing, and appears as an impedance which is $\frac{Z_f}{1-A}$. If this voltage swing of $\frac{E_{out}}{A}$ is a significant fraction of the input signal E_{in}, or if the impedance $\frac{Z_f}{1-A}$ is a significant fraction of Z_i, there will be a definite output signal error in addition to the error introduced by the tolerances of Z_i and Z_f. The exact value of this error is $1 - \frac{A}{A - 1 - \frac{Z_f}{Z_i}}$.

So long as $\frac{Z_f}{Z_i}$ is small and A is large, the error is not serious. For instance, using the O-Unit's operational amplifiers (at low frequencies where $A = -2500$) in the simple fixed-gain amplifier mode with resistors for Z_i and Z_f, we see that the error for a gain of 1 ($Z_i = Z_f$) is only $1 - \frac{2500}{2502}$, or less than 0.1%. For a gain of 100, however, the error becomes $1 - \frac{2500}{2601}$, or almost 4%. (A gain-correcting resistor is automatically shunted across Z_i in the O-Unit when the internal Z_i resistor is set to 10 K and Z_f to 0.5 or 1.0 meg.

Using external components, similar precautions should be observed when high gain is required).

*NOTE

Common usage in the analog computer field assigns a negative number to the open-loop gain between the − input and output (and a positive number to the gain from the + input). Therefore, in calculating values from formulas involving A and the − input, it is necessary to keep in mind that A is a negative number, and the expression "1 − A" for instance, when A is − 2500, equals + 2501, not − 2499.

One simplification has been made in this article. Closed-loop gain, commonly expressed as

$$\frac{-Z_f}{Z_i}\left[\frac{1}{1 - \frac{1}{A}\left(1 + \frac{Z_f}{Z_i}\right)}\right]$$

has been reduced to

$$\frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}}\right]$$

It may also be written

$$\frac{-Z_f}{Z_i}\left[\frac{1}{1 - \frac{1 + Z_f/Z_i}{A}}\right]$$

if this seems to indicate the effect of A on accuracy more clearly.

Approximate Error Calculation Using C for Z_i or Z_f.

Since it is not easy to assign a single impedance value in the error formula for Z_f or Z_i when one of them is a capacitor, it is convenient to use the ratio E_{out}/E_{in}, representing the actually obtained voltage gain, to compute the approximate error. The error ϵ is found by this formula:

$$\epsilon = 1 - \left[\frac{\frac{E_{out}}{E_{in}} - A}{1 - A}\right]$$, or, more simply,

$$\epsilon = \frac{1 - \frac{E_{out}}{E_{in}}}{1 - A}$$, where A, as before, is the

open-loop gain, and E_{out}/E_{in} is the actually obtained voltage gain (Don't forget—both A and E_{out}/E_{in} are **negative numbers**). Example: where A is − 1000 and the observed E_{out}/E_{in} is − 50, the error has been 51/1001 or 5.095%. The output "− 50" represents, then, 94.905% of the correct value, and the correct value is − 50/0.94905, or almost − 52.7.

For convenience, you may want to rearrange the terms as shown below, to determine how large an output signal to allow, for a given input and an arbitrarily selected maximum error:

$$\frac{\text{Max } E_{out}}{E_{in}} = 1 - \epsilon(1 - A)$$

Using the Tektronix Type O Operational Amplifier for integration, for instance, to keep error due to amplifier gain below 1%, the output voltage during or at the end of the integrating interval should not exceed the average value of the signal being integrated by more than a factor of 1 − (.01 × 2501), or − 24, for low frequencies. The same limitation should be observed during differentiation.

The minimum open loop gain required by an operational amplifier to operate within a given error even at "zero" Z_f/Z_i is $A = \frac{(\epsilon - 1)}{\epsilon}$, where ϵ is the error expressed as a decimal fraction (.01 = 1%, 0.1 = 10%, etc.).

Where Z_f/Z_i is a finite number, the minimum open-loop gain required for a given maximum error is

$$A = \frac{(\epsilon - 1)(1 + Z_f/Z_i)}{\epsilon}.$$

The application of these formulae will be most useful in observing gain-bandwidth limitations, discussed below.

2A. Gain-Bandwidth Product:

The gain-factor A varies with frequency, and it's important to know what the effective value of A is for the frequencies or signal frequency components being used. In the Type O, the gain factor A is constant (− 2500) only to about 1 KC, dropping off to −1000 at about 15 KC, and reaching a value of −1 at approximately 15 Mc.

The error introduced by the gain factor, then, becomes greater with frequency, and for accurate measurements the allowable ratio of E_{out} to E_{in} must be reduced as higher-frequency information is processed.

Although the drop in gain at high frequencies in the open-loop bandwidth characteristic follows the same pattern as that of an integrator, it must be remembered that this response is obtained **without** input and feedback elements. The effect of this rolloff will **add** to the effect of the integrating components, altering their effect.

At a frequency approximately 1/10 of the open-loop gain-bandwidth product, the open-loop gain will be insufficient (on the order of 10 or so) to provide accuracy better than 9% even at "zero" closed loop gain, or 16.7% when Z_f/Z_i is 1, (i.e., $E_{out} = E_{in}$). Above 1/10 of the open-loop gain-bandwidth product, answers will be only approximate, although the data may be useful for frequencies as high as 1/3 of the open-loop gain-bandwidth product.

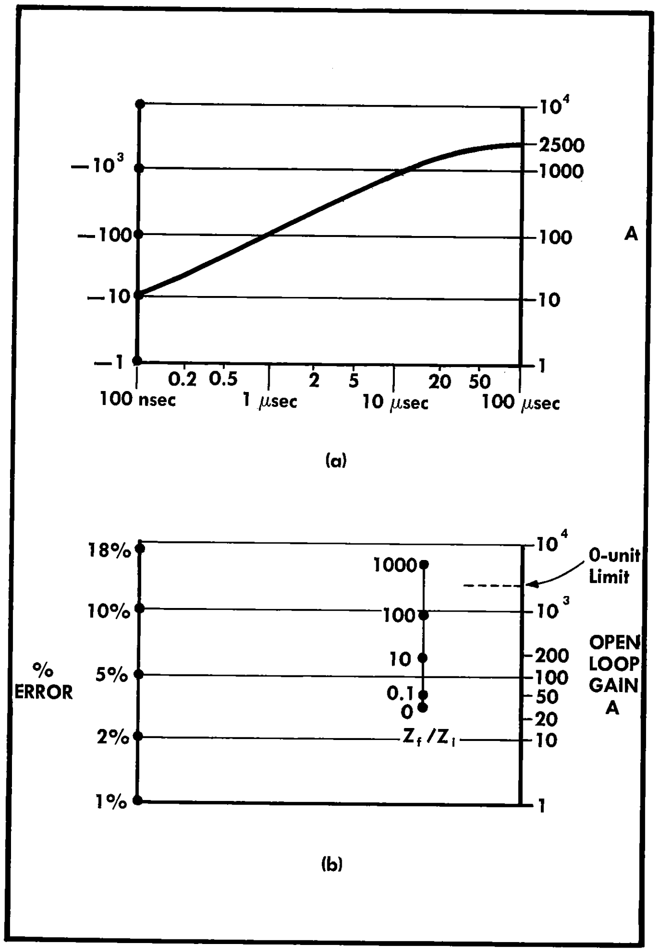


Figure 10.
(a) Variation in open-loop gain after application of signal, for O-Unit.
(b) Nomograph for determining A-FACTOR, ERROR and Z_f/Z_i. Given any two factors, the third may be found. (Lay straightedge across chart.)

2B. Gain-versus-Time Factor—Complex Waveshapes:

In working with pulse and complex waveforms, open-loop gain in terms of frequency is not too useful. Instead, the open-loop risetime characteristic, Figure 10(a) may be used to determine the time after the start of a signal at which the A-factor has reached a sufficiently high level to permit the desired accuracy. Figure 10(b) shows the A-factor required to support a given accuracy at a given attempted or "virtual" gain (Z_f/Z_i).

"Virtual gain" (roughly, E_{out}/E_{in}) in the case of integration or differentiation is the ratio between the RC time constant chosen and the time interval involved in the operation.

In the case of integration, virtual gain G_v will be: $G_v = \frac{-t}{RC}$, where t is the integrating interval—i.e., that span of time during which the integral continues to increase. The larger the values of integrating components, the smaller the virtual gain.

In the case of differentiation, the virtual gain will be: $G_v = \frac{-RC}{t}$, where t is that span of time during which the input signal has its steepest slope. The larger the values of differentiating components, the higher the virtual gain.

As can be seen from Figure 10(b), holding virtual gain to a value of one or so is a good general rule of thumb for accurate measurements.

NOTE

It should be kept in mind that the values of the internal 10 pf and 100 pf Z_f and Z_i components of the O-Unit have been adjusted under dynamic conditions, to compensate partially for the time-dependent errors indicated in Figure 10(a). For greatest measurement accuracy, standard waveforms involving a similar time interval and virtual gain as the signal to be measured should be used to determine the probable measurement error, or to trim the values of external components to provide direct readings for the particular waveform to be measured (comparison method). However, correction of this sort can be optimized for only a limited range of waveforms, and cannot extend the operating range of the system indefinitely.

3. Grid Current:

During integration, any grid-current flowing in the − input will be integrated along with the current through E_{in}, except when this current is bucked out through a DC path from output to input (in the Type O Unit "Integrator LF Reject" circuits are provided for this purpose).

The amount of grid-current flowing in the −input circuit may be determined by switching out any "LF Reject" circuit and measuring the length of time it takes the output signal to rise or fall 1 v with a capacitor as Z_f (no signal input). The grid current I_g is found by this formula $I_g = \frac{C}{t}$, where t is the time (in seconds) required for the output to move one volt.

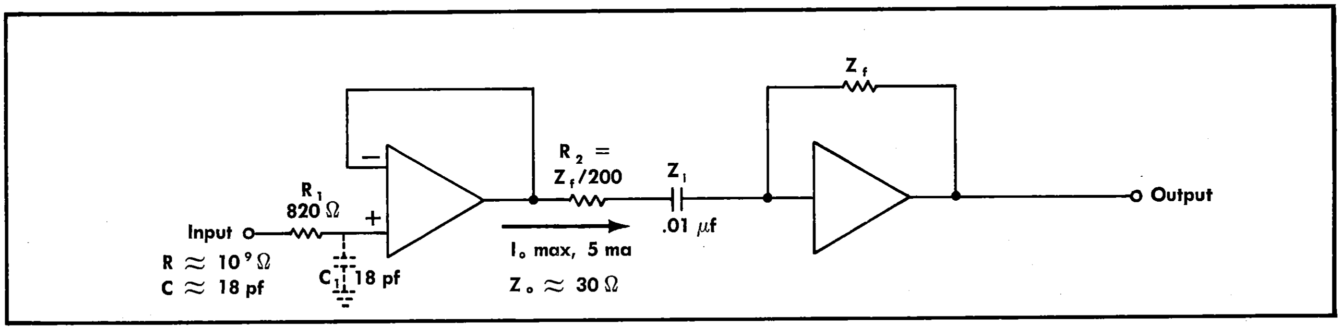


Figure 11. Operational amplifier connected as gain-of-one, non-inverting amplifier to drive low-input-impedance differentiator from high impedance signal source. If output current capability is 5 ma (as in the O-Unit), driver amplifier will reproduce faithfully an input rate-of-change as high as 0.5 v/µsec into .01 µf, several orders of magnitude in excess of the amount necessary to obtain usable output from the differentiator. Component R_1 combines with the input C of the first operational amplifier to compensate its response in this mode. R_2 limits the current to the second operational amplifier to prevent overdriving, and reduces noise components possibly introduced by first amplifier.

In the Type O unit, a grid current of 300 picoamperes at the −grid is normal.

It is not usually practical to try to adjust an operational amplifier for "zero" input current, since this condition is not as stable as a fixed value of grid-current appropriate to the input tube type and amplifier design. In low-frequency operational amplifiers using electrometer tubes as input elements, extremely low values of input grid current can be obtained with good stability. In wide-band units, higher values must be tolerated.

Once the grid current has been set to a known value, its effect on a given integrating operation can be computed. So long as the value of I_g is very small compared to the average value of the current through Z_i during the integrating interval, the effect of I_g can be largely ignored.

4. Output Current and Voltage Limits:

Any operational amplifier is limited in the amount of current and voltage it can deliver to its feedback network and any external load with good linearity. If these limits are exceeded during any part of an operation, the accuracy of that part of the operation, at least, will be impaired.

In the case of the O-Unit, maximum output is ±50 v and ±5 ma. At high speeds, the maximum rate-of-change at the output will be limited by the available current, and should not exceed 20 v per µsec, when loaded by the O-Unit's oscilloscope preamplifier (47 pf) and 10 pf of other loading (e.g., Z_f).

5. Input Signal Source Impedance:

A part of Z_i, the input element of the operational amplifier circuit, is the source impedance of the signal being processed. Linear operations using precision input and feedback components will be accurate only if the source impedance of the signal is very small compared to the impedance of the input component, or the value of the input or feedback component is trimmed to allow for the impedance of the signal source.

Where trimming of components is not practical, or the signal source impedance is not resistive and linear, the usual practice is to process the signal first through a gain-of-one, high input-impedance, low-output-impedance amplifier, such as that shown in Figure 11, to obtain a low-impedance signal source for the desired operation. In the case shown, the output impedance of the first amplifier is **too** low, making it capable of overdriving the second. A current-limiting resistor helps keep down noise as well as prevent overdriving.

Shunt Impedance Across − Input

Though we tend to think of the −input or −grid as a "virtual ground", and that impedances between this point and ground will have a negligible effect on the performance of the operational amplifier, this is only partially true. The true impedance of this point is $Z_f/(1 - A)$ and that instead of holding a perfect voltage null (as would be the case if A were infinite), its voltage excursions actually amount to E_{out}/A.

So long as A is large and Z_f has a fairly low value, an impedance across the −input which is large compared to Z_f or Z_i will have little effect on performance. However, in high-frequency work, where the effective value of A is low, more and more care must be exercised to assure that shunt impedances—particularly capacitive reactances, which become lower with increasing frequency—do not interfere with the operation (Figure 12).

The general expression for the closed loop gain of an operational amplifier $\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}}\right]$ may be modified as follows to show the effect of shunt impedance Z_s across the − grid:

$$\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i} - \frac{Z_f}{Z_s}}\right],$$

keeping in mind that A is a negative number. As you can see, unless Z_s is very high compared to Z_f, its effect on accuracy may become comparable to that of Z_f/Z_i.

The terms in the above equation can be rearranged to show the effect of Z_s as related to Z_i:

$$\frac{E_{out}}{E_{in}} = \frac{-Z_f}{Z_i}\left[\frac{A}{A - 1 - \frac{Z_f}{Z_i}\left(\frac{Z_s + Z_i}{Z_s}\right)}\right]$$

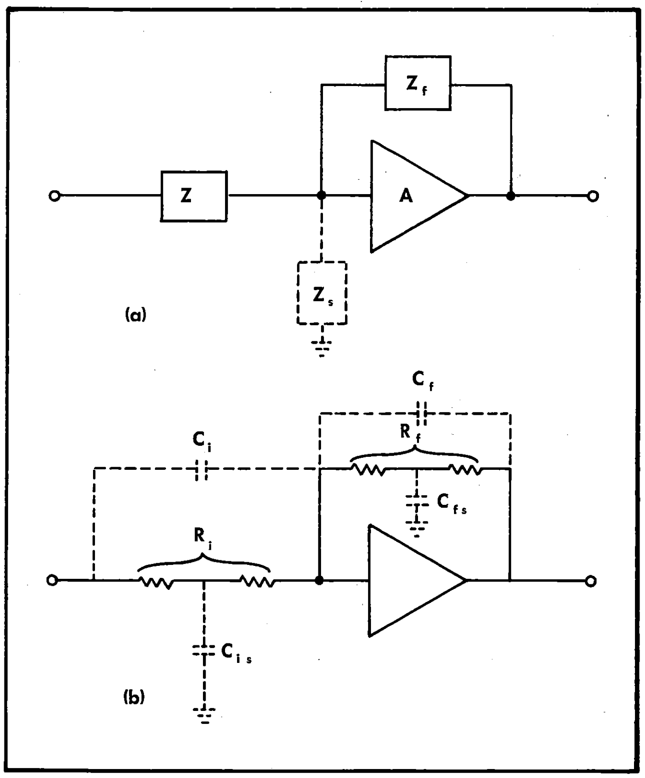


Figure 12.
(a) Shunt Impedance across − input. Where Z_f is large compared to Z_i and Z_f, and open-loop gain A is high, effect of Z_f is negligible.
(b) Where Z_i or Z_f is a resistor, and particularly if a large (> 100k) value, more serious errors may be caused by capacitance from the resistor body (highest impedance point) to ground, and, in the case of R_i during integration, end-to-end capacitance of R_i. Time constants involved in shunt capacitance C_{is} and C_{fs} are approximately RC/4.

Correcting For The Effects of Stray C

In high-speed work, the accuracy of operations will be affected by C_s during the start of an operation when the effective value of A is low, and also by the end-to-end and distributed capacitance to ground of the resistors used for Z_i and Z_f (Fig. 12b).

To correct for strays and the variation in A, the 100 pf and 10 pf values of Z_i and Z_f in the Type O operational amplifiers are factory adjusted under dynamic conditions, and no external compensation of these components is generally required. If it is intended to use values in this range externally, they should be padded or trimmed as necessary under conditions similar to those of the contemplated measurements.

The resistors used as Z_i and Z_f, however, can be given only partial compensation internally, since the optimum value of compensation varies with the application. For this reason, it is usually necessary in dealing with short-duration or high-frequency signals to add external compensation to R_f or R_i when these components are used in amplification and differentiation.

Figure 13 illustrates the corrections necessary to improve operational accuracy for each of the three basic operations.

Note, however, that except in the case of straight amplification (Fig. 13A), the compensation itself introduces possible errors which must be recognized and allowed for in interpretation of results.

Compensated Amplifier

In the case of amplification, selecting small values of capacitance (on the order of 2-25 pf) for C_1 and C_2, the closed-loop risetime can be made to approach the slope of the open-loop risetime (Fig. 9), providing a gain-bandwidth product about equal to the open-loop gain-bandwidth product. Without compensation, the amplifier may typically achieve perhaps only 1/20 of this figure.

Compensated Differentiator

Without compensation, the differentiator (Fig. 13B) may respond to a sudden change in dE_{in}/dt by overshoot, followed by sinusoidal ringing, due to the fact that excess output voltage must be developed to charge via R_f the input capacitance and the distributed stray capacitance of R_f itself, as well as provide the current needed to obtain a null at the − input. As soon as the strays are charged, however, the excess current through R_f upsets the null, and the output must swing in the opposite direction to re-establish the null and discharge the capacitance associated with R_f—hence the ringing. A small capacitance across R_f provides the current needed to establish the null at the start of the waveform without having to develop excess voltage across R_f.

Differentiator Compensation Limits Initial Accuracy

The presence of this capacitance, however, limits the output voltage maximum to approximately $\frac{-dE_{in}\,C_i}{C_2}$. After an abrupt change in the input waveform, then, when dE_{in} is small, but $\frac{dE_{in}}{dt} \times RC$ may be quite large, the output voltage limitation of $\frac{-dE_{in}\,C_i}{C_2}$ may result in a significant error. The solution in this case is to select a larger value of C_i and smaller values for R_f and C_2 (keeping the $R_f C_i$ time constant the same) to minimize the error, and keep its duration as short as possible.

Integrator Compensation Rarely Needed

Failure of the integrator to start integrating at the proper rate at the beginning of a fast-rise pulse or after a sharp step in the input waveform is usually due largely to the distributed stray capacitance to ground in R_i. This is infrequent; more commonly the error is in the opposite direction because of excessive capacitive coupling of the input waveform around R_i into the $-$grid directly, producing a step of approximately $\frac{-dE_{in}\,C_{in}}{C_f}$.

The first ($-$error) waveform in Fig. 13C was obtained by deliberately putting a ground plane near the center of a 9 Megohm R_i and carefully shielding the $-$grid. Removing the ground plane and shield produced the third ($+$error) waveform, using the same input signal (a rectangular pulse) and components.

Normally, the "undercompensated" effect would only occur when R_i is composed of several resistors in series, or when a high-value potentiometer is used as, or in series with, R_i.

The solution usually is to select a smaller value for R_i and a larger one for C_f, to maintain the same time-constant. Normally, if a signal source is capable of driving a large value of R_i with capacitive compensation, it is also capable of driving a smaller value of R_i without compensation.

Theoretically—as when a potentiometer is used in conjunction with R_i—it is possible to compensate the RC losses in R_i by shunting R_i with a series RC network of the proper time-constant, or by using a small value of R in series with C_f. In practice, these added components usually add nearly as many stray-C problems as they cure, and "compensation" of this sort is not recommended. Compensating with simple capacitance across R_i produces a "step" error at any abrupt transition, and usually an error of greater magnitude than the one to be corrected.

If R_i is a single component, an environmental "guard" driven by the input signal (e.g., a short piece of wire soldered to the input end of R_i and dressed near the body of the resistor) can make some correction, but its use requires more complete shielding of the $-$grid and the $-$grid end of R_i.

Using "Standard" Wavefoms For Comparison

The use of standard waveforms (pulses and ramps) with known parameters, is of considerable help in adjusting compensation and assuring best accuracy for critical measurements near the limits of the instrument's capabilities. For many purposes, such "standard" waveforms may be obtained by attenuation of the oscilloscope gate and sawtooth output waveforms. Selection of time and amplitude parameters close to those of waveforms to be measured will give best assurance against possible system errors.

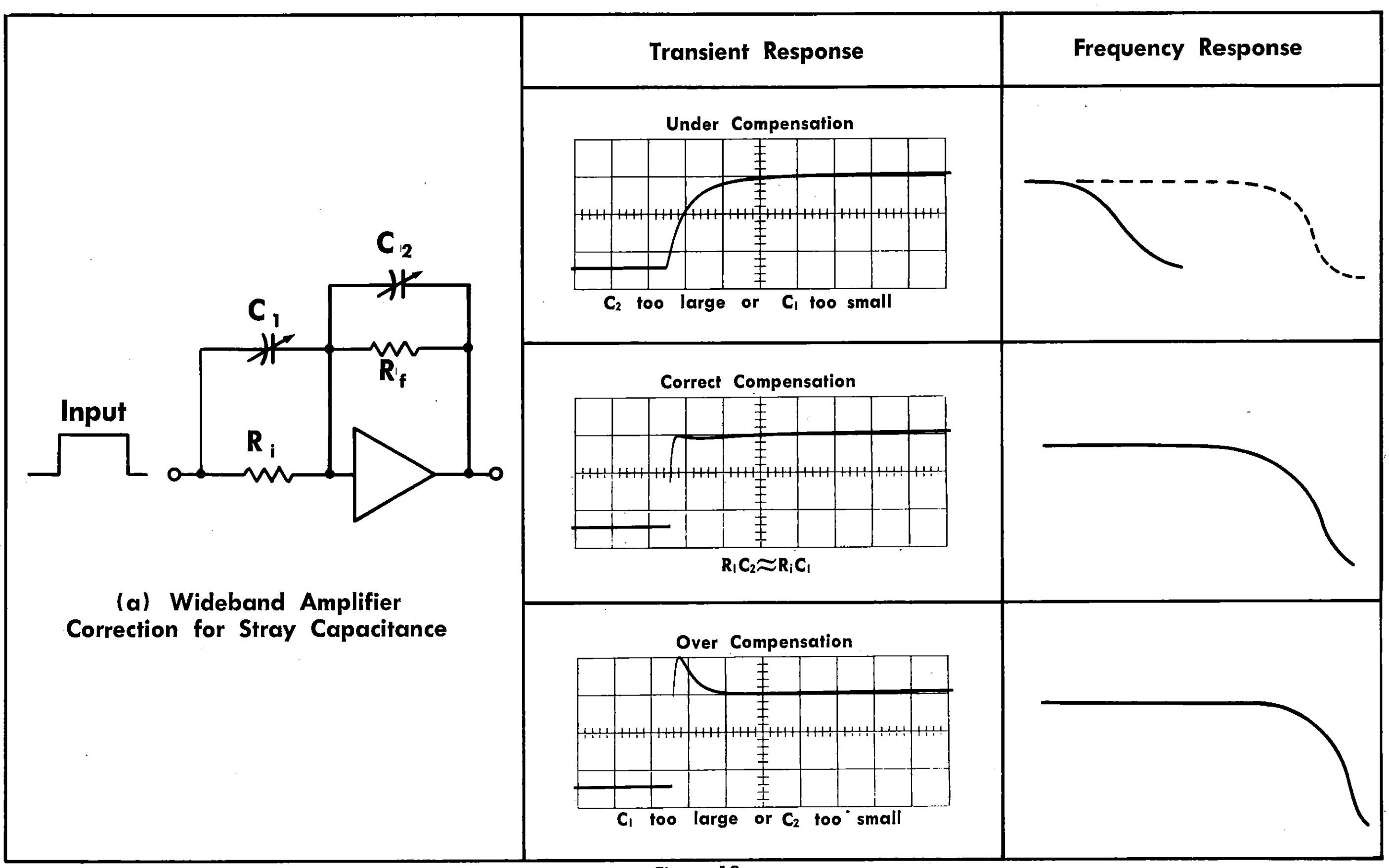


Figure 13.

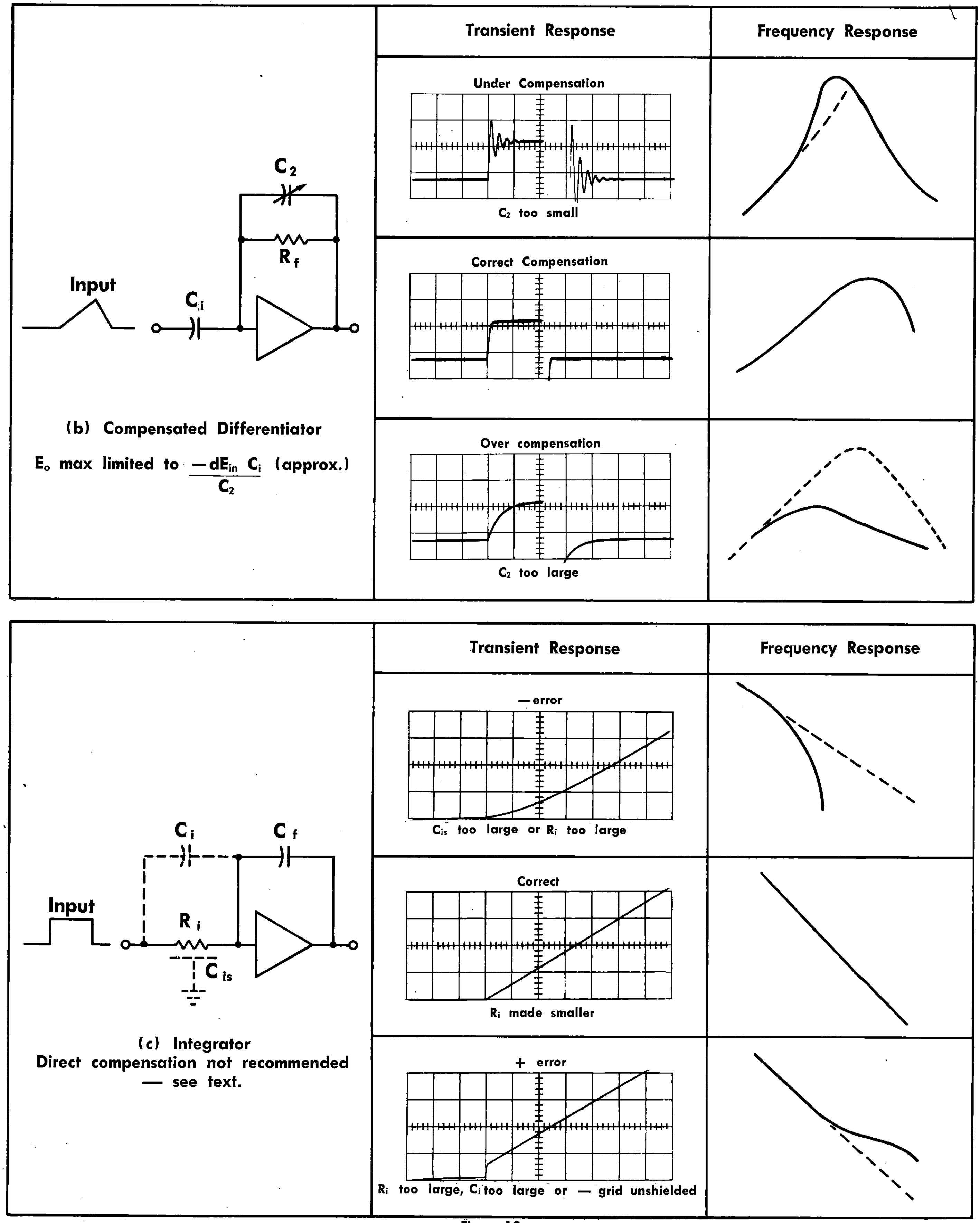
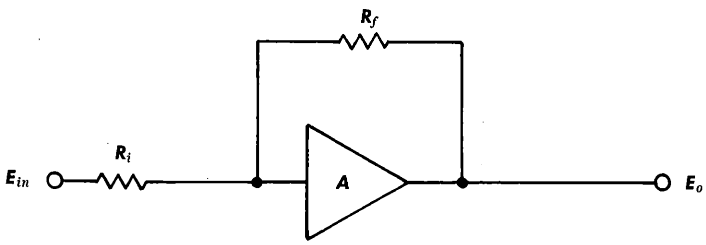

Transient Response
Frequency Response
Under Compensation
C2 too small
Correct Compensation
Over compensation
C2 too large
Input
C2
Rf
Ci
(b) Compensated Differentiator
Eo max limited to −dEin Ci / C2 (approx.)
Transient Response
Frequency Response
— error
Cis too large or Ri too large
Correct
Ri made smaller
+ error
Ri too large, Ci too large or — grid unshielded
Input
Ci
Cf
Ri
Cis
(c) Integrator
Direct compensation not recommended
— see text.

Figure 13.

Amplification	Notes

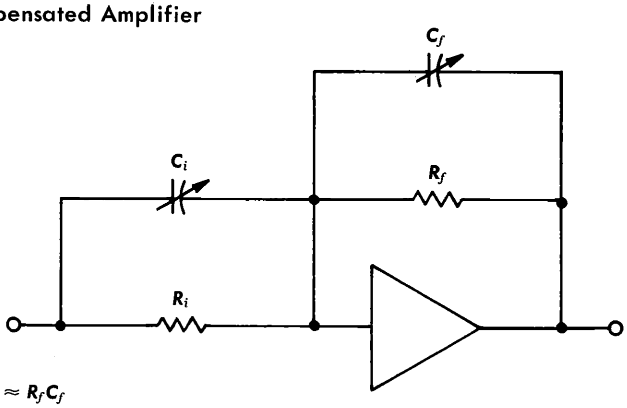
Basic Amplifier



$$E_o = \frac{-E_{in}R_f}{R_i}\left(\frac{A}{A - 1 - \frac{Z_f}{Z_i}}\right)$$ (*A* is negative)

Gain of the amplifier is determined by the ratio R_f/R_i. If open loop gain A is very large compared to R_f/R_i, $E_o \approx \frac{-R_f}{R_i} E_i$.

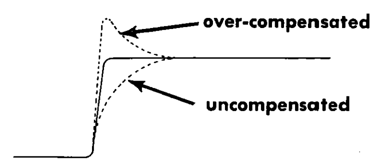
Compensated Amplifier



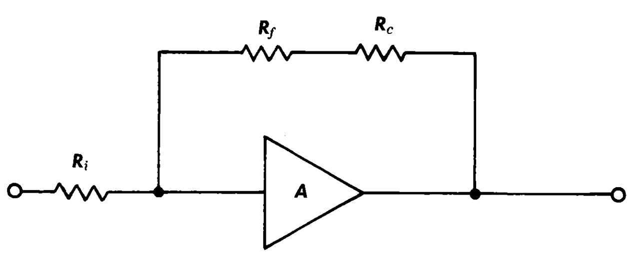
$R_iC_i \approx R_fC_f$

High frequency response and risetime are considerably enhanced when stray capacitance is compensated. C_i and C_f should be kept as small as possible (5-30 pf).

Step-function response:



Gain-Corrected Amplifier†



Unless open loop gain $A \gg \frac{R_f}{R_i}$, gain accuracy may be preserved only by trimming value of R_f or R_i.

Gain-correction resistance † is necessary if accuracy is required at high gain. Error* for a given R_f/R_i is

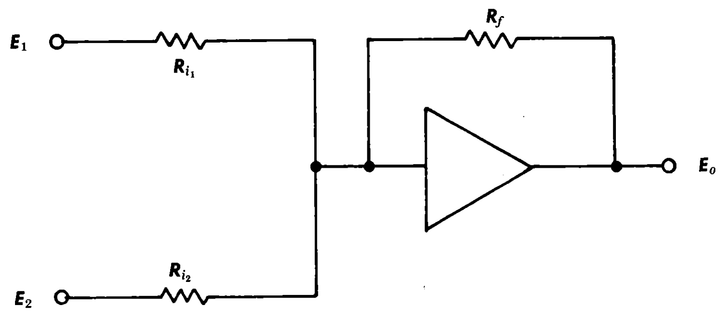
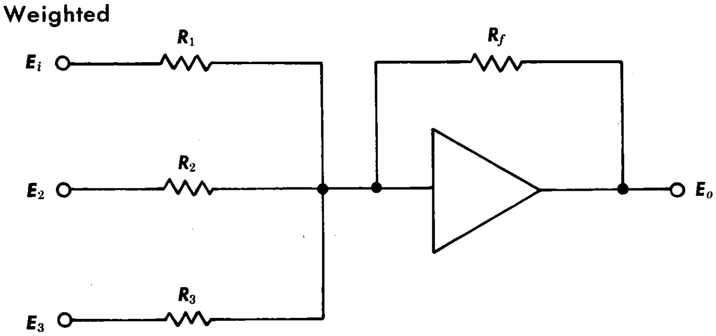
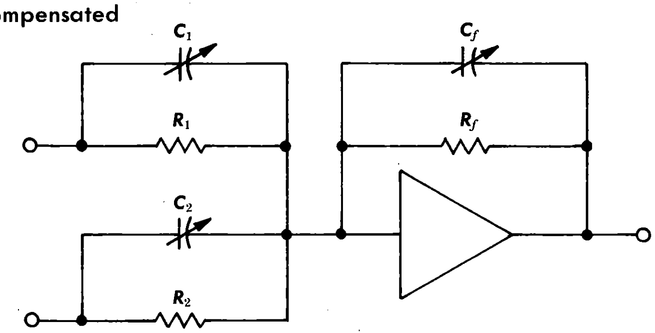
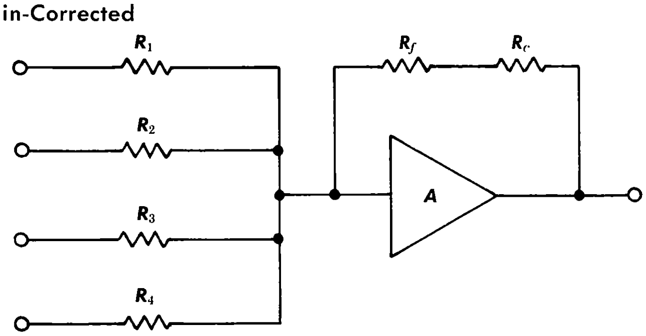
$$\left[1 - \left(\frac{A}{A - 1 - \frac{R_f}{R_i}}\right)\right] \times 100\%. \text{ (A is negative)}$$

$$\text{Correcting resistor } R_c = -R_f\left(\frac{1 + \frac{R_f}{R_i}}{A + \frac{R_f}{R_i}}\right)$$

Correction may also be made by shunting R_i.

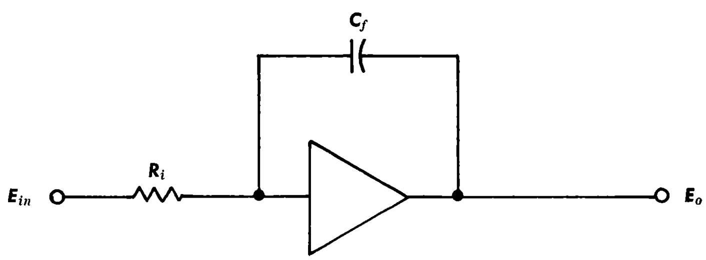
†In Type O, when internal Z_f and Z_i components are used, gain compensation is automatically switched in whenever required.

***In Type O, error is usually negligible at low frequencies if $R_f/R_i < 25$**

Summation	Notes
Basic Circuit  $E_o = -(E_1 + E_2)$ if $R_{i_1} = R_{i_2} = R_f$	If all resistors are of equal value, gain is unity for each input. For increased gain, R_f may be increased.
Weighted  $E_o = -R_f\left(\frac{E_1}{R_1} + \frac{E_2}{R_2} + \frac{E_3}{R_3}\right)$	Gain for any input is $-R_f/R_i$, permitting easy component selection for operations such as $(0.3E_1 + 7E_2 + 4E_3)$.
Compensated  $R_1C_1 \approx R_2C_2 \approx R_fC_f$	The summing amplifier cannot be quite as well compensated as simple amplifier because of input loading. Inputs may be separately adjusted if opposite input is grounded during adjustment, and process is repeated to eliminate interaction. Compensation of 3- and 4-input summing networks becomes increasingly more difficult and less practical.
Gain-Corrected  Unless $\left(\frac{R_f}{R_1} + \frac{R_f}{R_2} + \frac{R_f}{R_3} + \frac{R_f}{R_4}\right) \ll A$, R_f should be increased beyond nominal value to preserve accuracy.	Gain error for any input will be $\left(1 - \frac{A}{A - 1 - \left(\frac{R_f}{R_1} + \frac{R_f}{R_2} + \frac{R_f}{R_3} + \frac{R_f}{R_4}\right)}\right) \times 100\%$ where A is open-loop gain (a negative number) Value of correcting resistor R_c should be $-R_f\left(\frac{G+1}{G+A}\right)$ where $G = \left(\frac{R_f}{R_1} + \frac{R_f}{R_2} + \frac{R_f}{R_3} + \frac{R_f}{R_4}\right)$

Integration	Notes

Basic Circuit

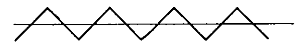


$$E_o = \frac{-1}{R_i C_f} \int E_{in} dt$$

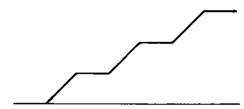
Notes:

Gain (sinewaves) $= \dfrac{-1}{2\pi\, f\, R_i\, C_f}$ except as limited by open loop gain.

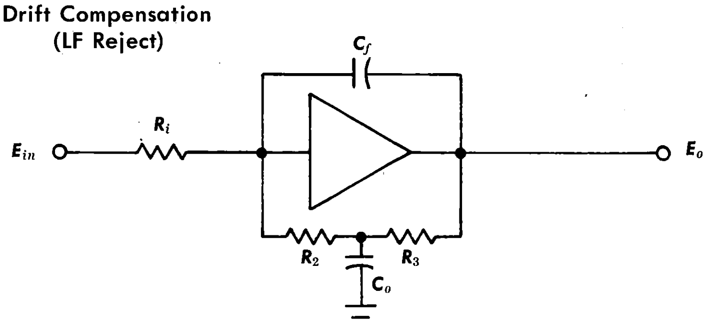
Square-wave response (for input symmetrical about O_v):



Square-wave response, unidirectional input:



Drift Compensation (LF Reject)



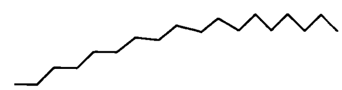
Reset time-constant $T_r = (R_2 + R_3)C_f$ **LF Reject time-constant $T_L = \dfrac{R_2\, R_3\, C_o}{R_2 + R_3}$**

Notes:

Response to dc and low-frequency information (including input grid current) is limited by LF REJECT circuit. DC response is limited to $\dfrac{-E_{dc}(R_2 + R_3)}{R_i}$.

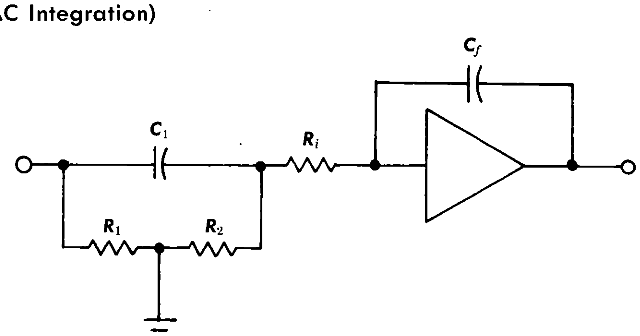
Capacitor C_o delays "dc" feedback by a value approaching T_L, thus improving short-term accuracy.

Square-wave response (Input unidirectional):



For accuracy, keep integrating interval $< 0.3\sqrt{R_2 R_3 C_o C_f}$

DC Rejection (AC Integration)



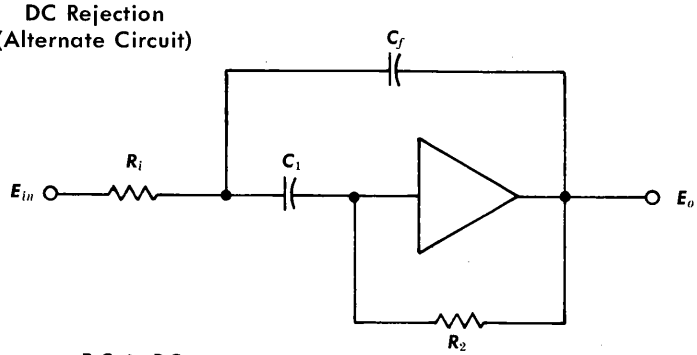
$C_1 \gg C_f$ **Integrating interval $\ll R_i\, C_1$**

Notes:

Input blocking capacitor C_1 prevents dc level of E_{in} from being integrated. If dc level is high, resistors R_1 and R_2 ($\ll$ capacitor leakage) minimize effect of capacitor leakage current.

This circuit does not reject input grid-current, but may be used in conjunction with LF REJECT Circuit above if limitations of each circuit are observed.

DC Rejection (Alternate Circuit)

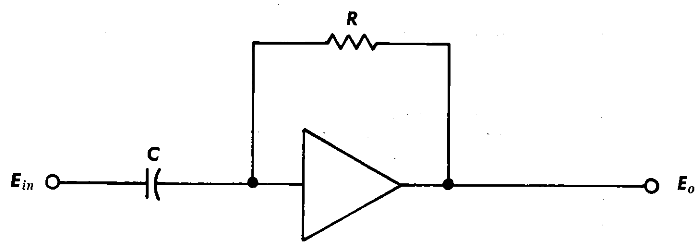
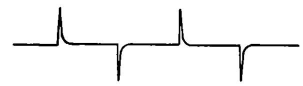
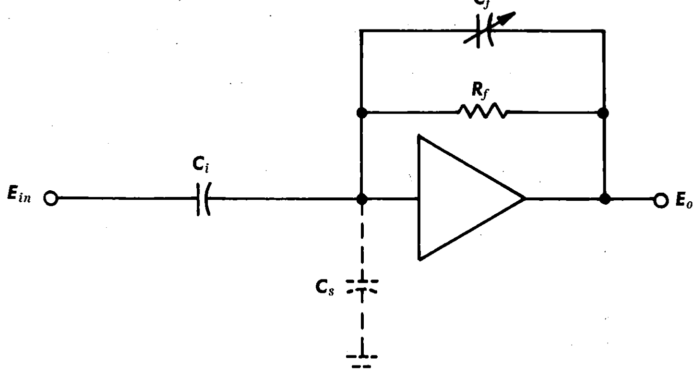
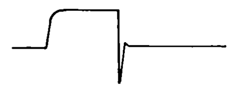
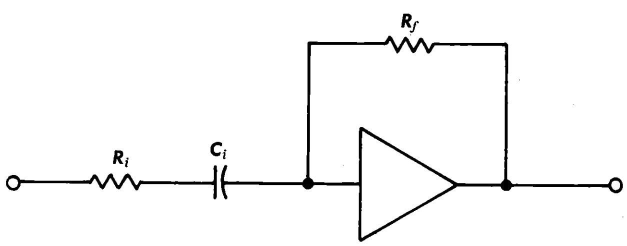


$R_2 C_1 \gg R_i C_f$

Notes:

This circuit combines the dc signal rejection and grid-current rejection of the two circuits above, but with fewer components.

For operational accuracy, integrating interval should be kept less than $0.3\sqrt{R_i\, C_f\, R_2\, C_1}$

Differentiation	Notes
Basic Circuit  $E_o = -RC\left(\frac{dE_{in}}{dt}\right)$	Gain (sinewaves) = $-2\pi fRC$, except as limited by open-loop gain-bandwidth product. **Squarewave response**  **Sawtooth response** 
Frequency Compensation 	Lag in current through R_f may produce ringing after sudden change of dE/dt. Adjust C_f using oscilloscope sawtooth of dE/dt comparable to that to be measured: **Sawtooth response:** **Uncompensated**  **Compensated** 
Noise suppression  $R_i \ll R_f$	Series resistance limits response of differentiator to wideband noise components above the frequency range of interest. Output swing is limited to $-dE_{in}$ R_f/R_i, so R_i should be on the order of $R_f/100$ if possible, to preserve accuracy in range of interest.

<table>
<tr><th>Voltage-Current Conversion</th><th>Notes</th></tr>
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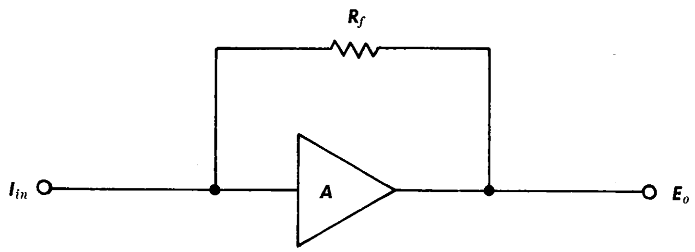
Current-to-Voltage



$E_o = -I_{in}R_f\left(\frac{A}{A-1}\right)$

</td><td>

Input current limitation is $\frac{E_o\text{max}}{R_f}$ or 0.5 $I_o$max, whichever is smaller. With finite value of A,

$$E_o = -I_{in}R_f\left(\frac{A}{A-1}\right) \text{ (A is negative).}$$

Unless A is very small or source impedance of I_{in} is low, $E_o \approx -I_{in} R_f$. Source impedance of I_{in} should be $\gg \frac{R_f}{A-1}$

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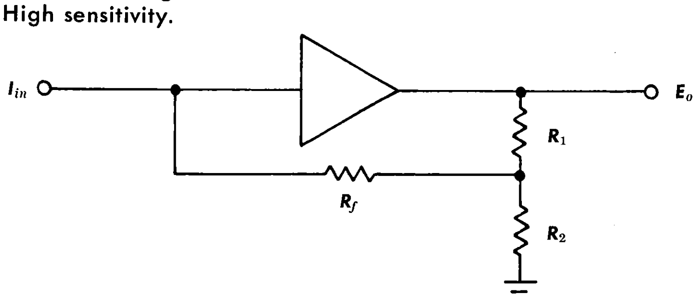
<tr><td>

Current-to-Voltage, High sensitivity.



Effective value of R_f is $R_f\left(\frac{R_1 + R_2}{R_2}\right)$; $E_o \approx -I_{in}R_f\left(\frac{R_1 + R_2}{R_2}\right)$

</td><td>

At low frequencies, improvement up to 50-1 in sensitivity may be practical, for a given value of R_f. Loss of input current to ground through R_f and R_2 causes

error $\epsilon = \frac{KR_f}{KR_f + \left(R_f + \frac{R_1}{K}\right)(1 - A)}$.

Division of feedback causes error $\epsilon = \frac{K-1}{A-1}$, where $K = \frac{R_1 + R_2}{R_2}$ and A = open loop gain (negative). R_2 may be trimmed to correct errors.

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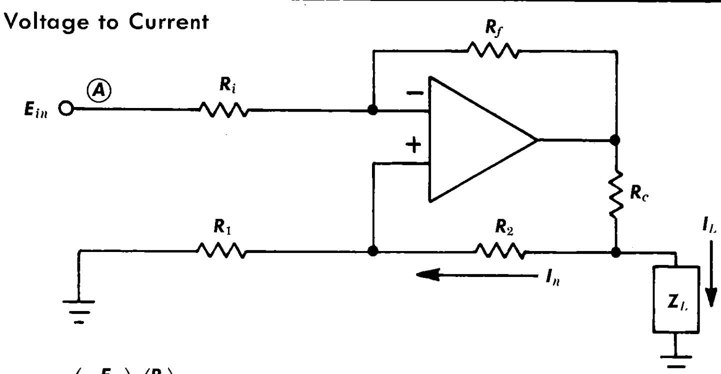
Voltage to Current



$I_L \approx \left(\frac{-E_{in}}{R_c}\right)\left(\frac{R_f}{R_i}\right) - I_n$ $R_1 + R_2 \gg Z_L < R_c$; $R_f/R_i = R_2/R_1$

</td><td>

With positive feedback, − grid seeks null at + grid level rather than ground. Circuit operates to keep drop constant across R_c; variations in Z_L control the + grid level. As value of Z_L approaches or exceeds R_c, effective input impedance at A drops, due to shift of − grid null point. $Z_{eff} = \frac{R_i}{1 + \frac{Z_L}{R_C}\left(\frac{R_f}{R_i + R_f}\right)}$ when $R_f/R_i = R_2/R_1$. Ideally, signal source impedance should be low, R_c larger than Z_L, and R_i large enough so that $Z_{eff} \gg$ Signal source Impedance. Practically, R_1 may be padded to correct for these sources of error.

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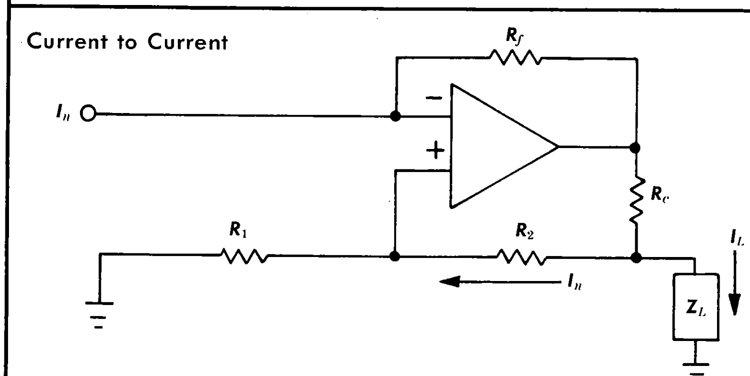
Current to Current



$I_L \approx \frac{-R_f}{R_c} I_{in} - I_n$. $R_1 + R_2 \gg Z_L < R_c$

</td><td>

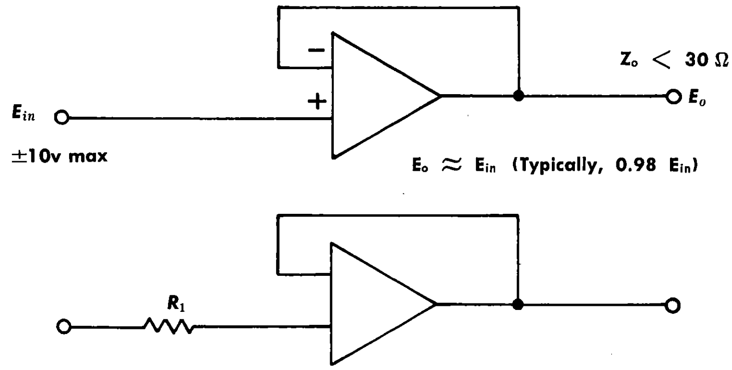
In the ideal case, source impedance of I_{in} is very large, and shift of − grid level has negligible effect. In practice, R_c and $\frac{R_1}{R_1 + R_2}$ must be selected in relation to Z_{source} and Z_L so that the system is stable and I_L will be constant for all expected values of Z_L. Voltage at − grid is $\left(\frac{-R_f I_{in} R_L}{R_c}\right)\left(\frac{R_1}{R_1 + R_2}\right)$. Ideally, $R_c > Z_L$, $R_2/R_1 = R_f/Z_{source}$, and $I_n \ll I_L$. Practically, R_2/R_1 should be made ≥ 1

</td></tr>
</table>

Notes

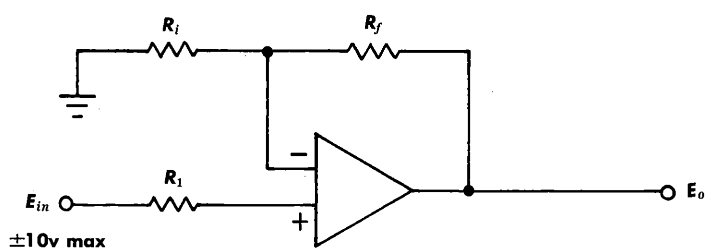
Amplification—High Input Impedance

Unity Gain Follower Amplifier



Notes: Input resistance is set by grid current. Typical value of $0.15nA$ at + Grid provides input resistance of $R = \frac{E_{in}}{I_G} = 7G\Omega$ for 1-volt signal, 7 MegΩ for 1 mv signal. Input C ≈ 18-25 pf. Output is not inverted. For wide-band compensation, select R_1 for optimum transient response (Typ.220-820Ω).

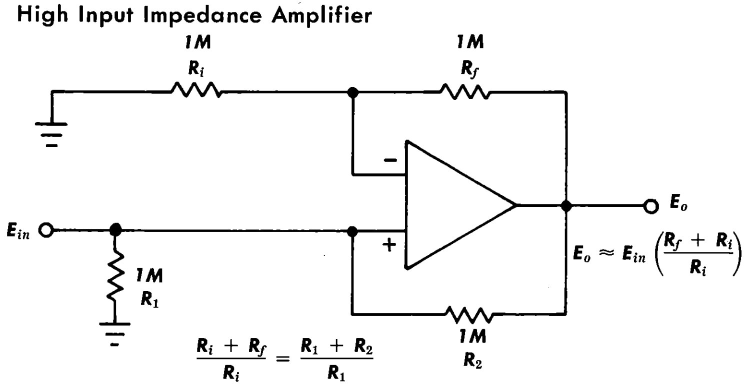
Non-Inverting Amplifier with Gain



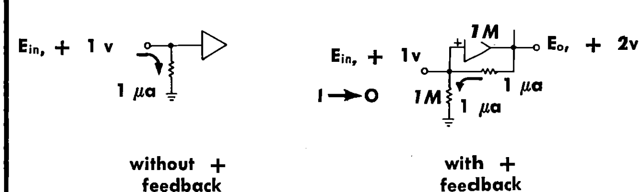
$$R_{in} = \frac{E_{in}}{I_g} \qquad C_{in} \approx 18\text{-}25 \text{ pf} \qquad E_o \approx E_{in}\left(\frac{R_f + R_i}{R_i}\right)$$

Notes: Feedback divider may be used to obtain gain > 1, or to correct unity gain circuit for gain of 1.00. For gain of 1.00, select $R_f = R_i\left(\frac{1}{G} - 1\right)$ where G is the observed gain without correction. R_i may be any convenient value > 10k.

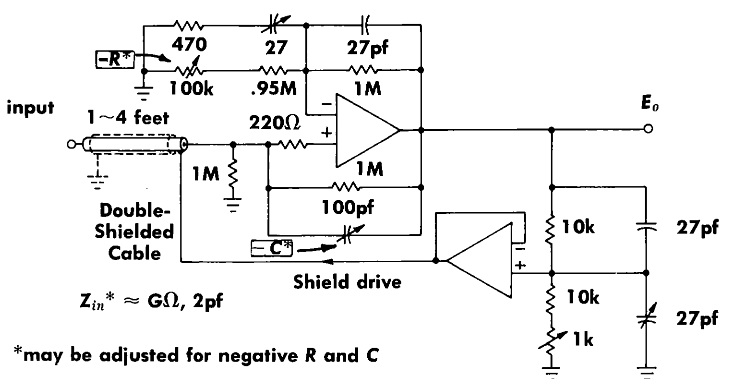
High Input Impedance Amplifier



Notes: Negative feedback via R_f sets gain; positive feedback via R_2 raises effective value of R_1, also decreases effective input C. The amplifier supplies most of the current required to drive input impedance.



Negative Impedance Amplifier

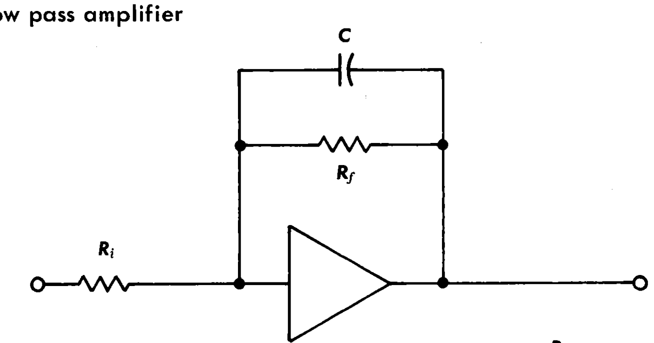


*may be adjusted for negative *R* and *C*

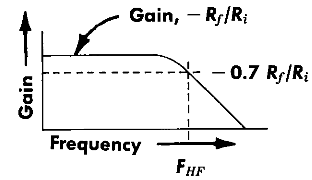
Notes: Capacitive compensation is added to the above circuit. The second amplifier drives the coax shield in-phase with the signal, further reducing input C. Negative resistance and negative capacity may be achieved by adjusting $-R$ and $-C$ control. With values shown, gain × 2, Bandwidth, ≈ 5 mc using Type O unit, from low-Z signal source. From high-Z signal source, bandwith may be as low as $\frac{1}{2\pi RC}$ where R and C are the Z_{in} settings.

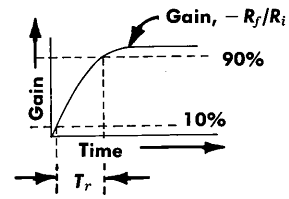
Bandpass Amplifiers	Notes

Low pass amplifier



L.F. Gain $= -\frac{R_f}{R_i}$





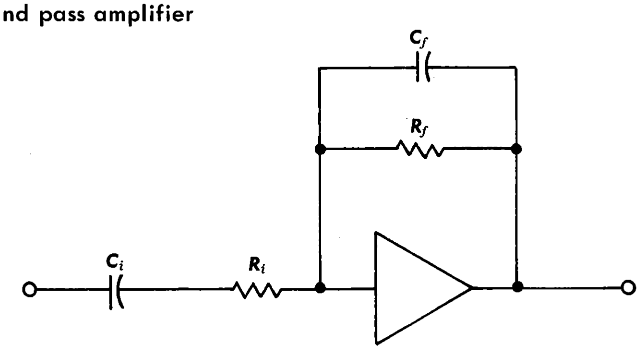
$f_{HF} = -3$db point

$= \frac{1}{2\pi R_f C_f}$ CPS

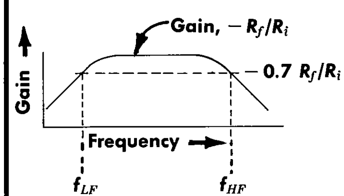
Rolloff above this point $\sim 6_{db}$/octave.

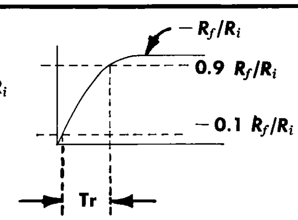
Risetime $T_r = 2.2 R_f C_f$
Response to step-function.

Band pass amplifier



Mid-band Gain $= -\frac{R_f}{R_i}$



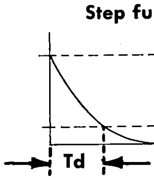


Tr = 2.2 $R_f C_f$
Step function response

$f_{LF} = \frac{1}{2\pi R_i C_i}$

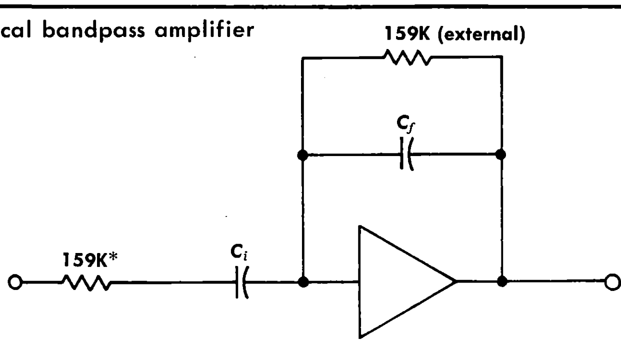
$f_{HF} = \frac{1}{2\pi R_f C_f}$

Rolloff, ~ 6db/Octave.



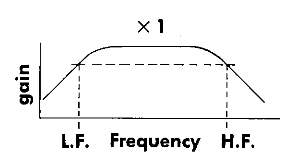
Td = $R_i C_i$

Typical bandpass amplifier



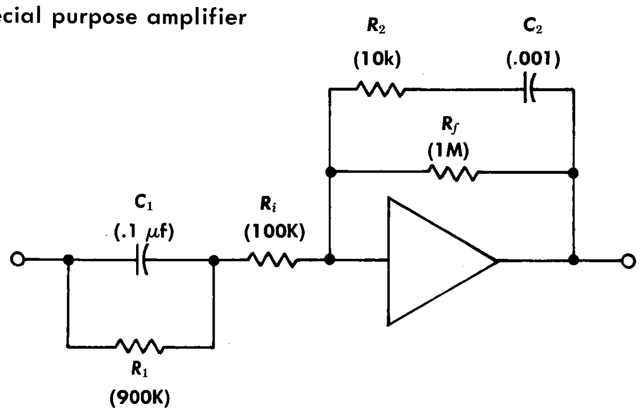
C_i and C_f selected by switch: 10pf – 1 μfd

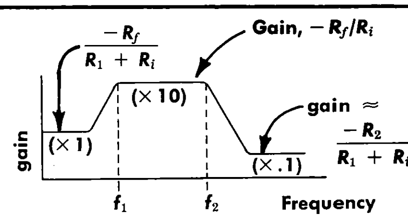
*Including signal source impedance



L.F.	C_i	H.F.	C_f
1cps	1μf	100kc	10pf
10c	.1	10kc	.0001μf
100c	.01	1kc	.001
1kc	.001	100c	.01
10kc	.0001	10c	.1
100kc	10pf	1c	1μf

Special purpose amplifier





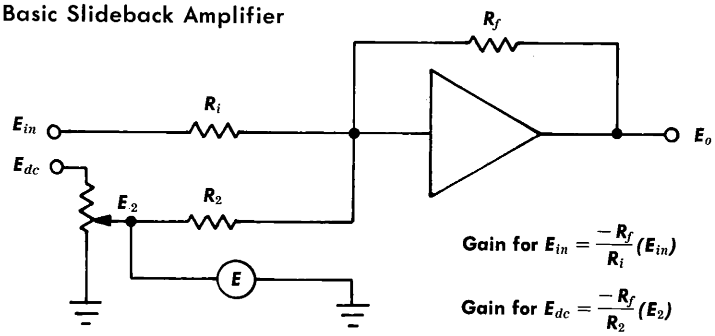
(15.9 cps) (159 cps)

$$f_1 = \frac{1}{2\pi R_i C_1} \text{ cps}$$

$$f_2 = \frac{1}{2\pi R_f C_2} \text{ cps}$$

<table>
<tr><th>Summation—Slideback Techniques</th><th>Notes</th></tr>
<tr><td>

Basic Slideback Amplifier



Gain for $E_{in} = \frac{-R_f}{R_i}(E_{in})$

Gain for $E_{dc} = \frac{-R_f}{R_2}(E_2)$

$E_o \approx -R_f\left(\frac{E_{in}}{R_i} + \frac{E_2}{R_2}\right)$

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DC offset voltage may be added to signal to permit DC coupled observation or measurement of a small signal at an elevated DC level. If $R_2 < R_i$, offset voltage source may be smaller than level to be offset. Loading effect of R_2 will be negligible if $\frac{R_f}{R_i} + \frac{R_f}{R_2} \ll A$. For error-correction techniques, see page 1-2. Use of low $\frac{R_f}{R_2}$ gain, and voltmeter at point E permits high-resolution DC and AC measurements by slideback-null technique.

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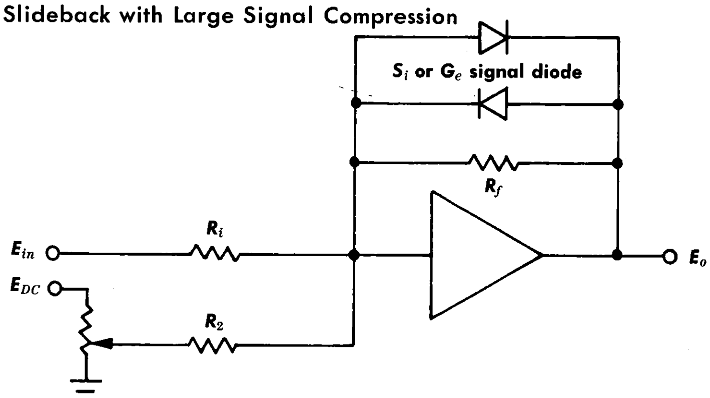
Slideback with Large Signal Compression



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For observation of a small signal superimposed on a large one, compression of the large signal is desirable. Amount of compression will depend on diodes. Silicon diodes (e.g. 6045, 6061) will provide a linear output window of about ±250 mv. Germanium diodes (e.g., T12G, T13G, 6075) provide continuous compression (no linear "window"). Additional diodes or resistors in series with diodes shown may be added to alter compression.

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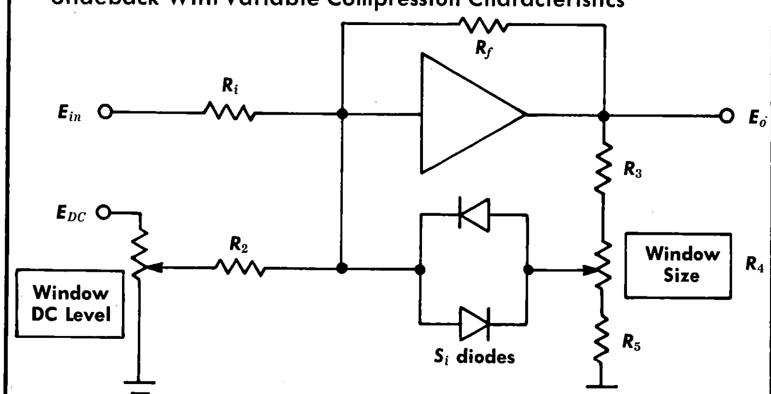
Slideback With Variable Compression Characteristics



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"Window Size" Control also affects compression slope outside of "window" as well as size of window. Output divider ($R_3 + R_4 + R_5$) should not be less than 20K impedance, and gain should be kept low, for accuracy. Output inside the "window" will be linear over the range $E_o = \pm 0.25K$, where K is R_3-R_4-R_5 divider ratio.

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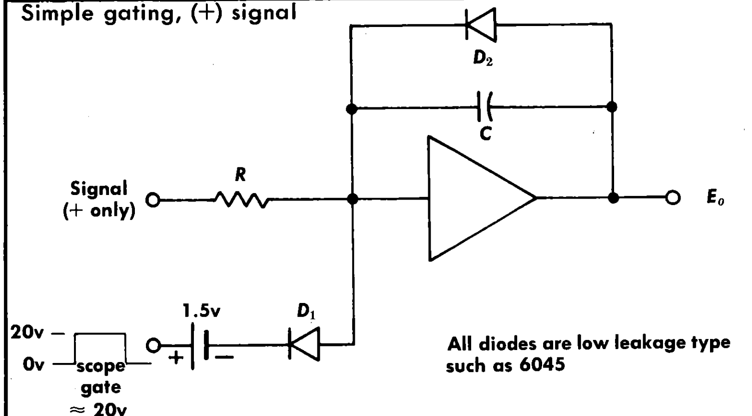
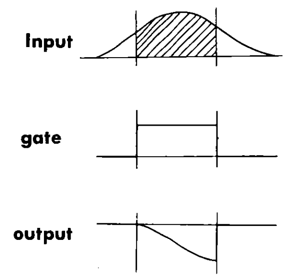
Integration — Gated Operation	Notes

Simple gating, (+) signal

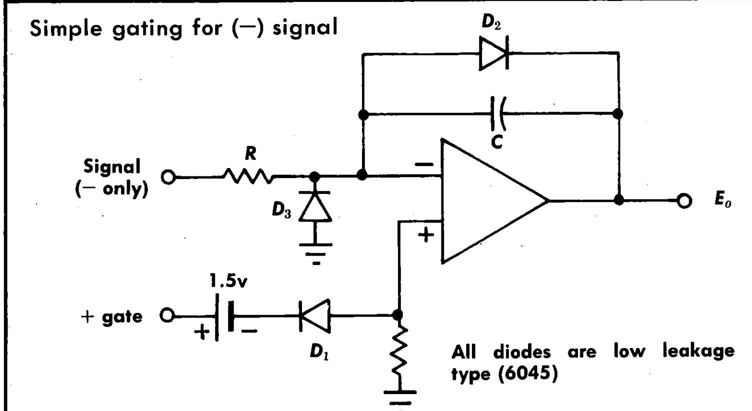


Gated integrator accepts positive signals only. When gate is zero, output is clamped by D_2 to approx. + 0.5v

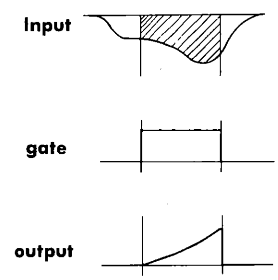
Small mercury cell may be used to obtain − 1.5v "off" level of gate, or oscilloscope may be modified (see app.)



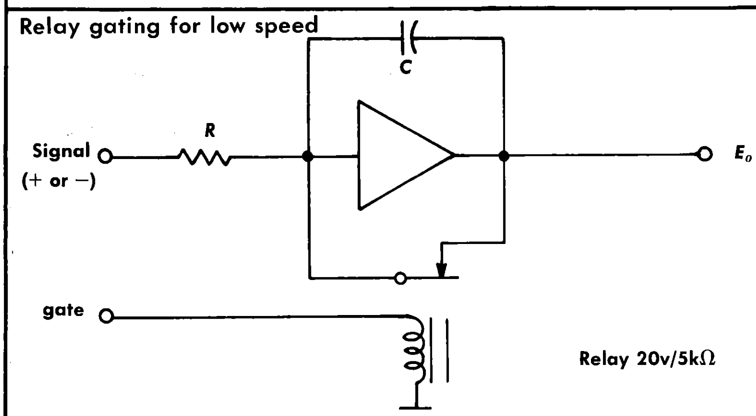
Simple gating for (−) signal



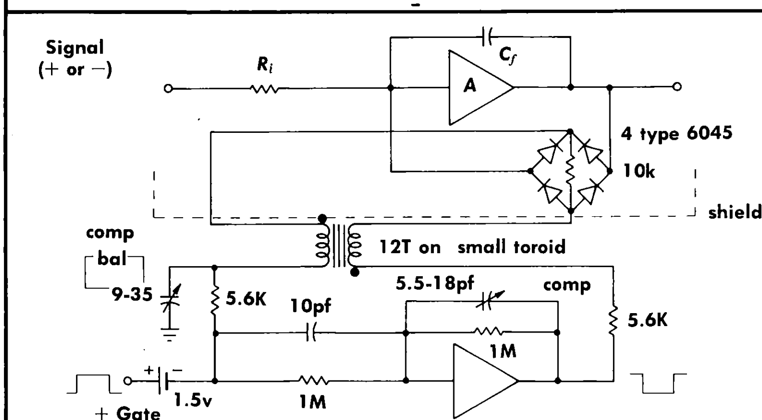
Negative signal can be gated by rearranging the circuit. The gate will drive + grid. When gate is zero, output is clamped to approx. − 0.5v.



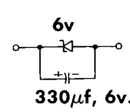
Relay gating for low speed



For very slow speed applications, relay can serve for simple gating. Input can be + or −.



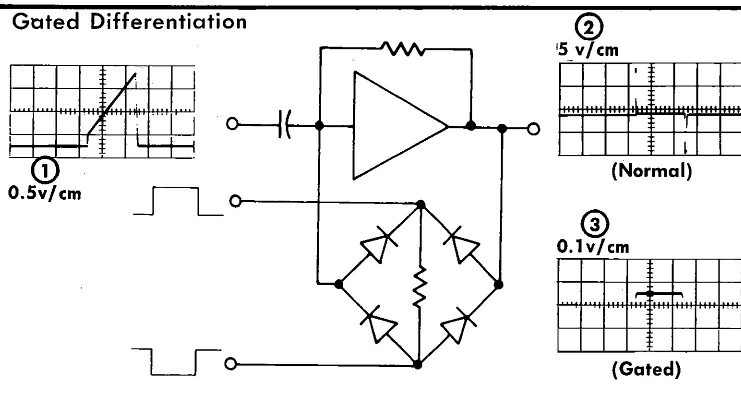
This arrangement uses diode bridge and gating amplifier B. It can accept + or − input signals. Output signal should not exceed the gating voltage. Using 535, 45, "A" gate, any portion of curve may be integrated, using sweep delay facility. For single-shot or low duty-cycle work, battery, or oscilloscope + Gate circuit modification must be used to keep Gate level negative during "off" time. If "off" time is less than 9 × "on" time, battery may be replaced by capacitor and zener diode.



Off time ≈ 2 sec max for value shown

<table>
<tr><th>Differentiation</th><th>Notes</th></tr>
<tr><td>

Gated Differentiation



for gate-driving circuit, see opposite page

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① Waveform to be differentiated.

② Differentiation without gating. Steps at start and end of ramp produce large transient which may overload scope input, make measurement or ramp linearity difficult.

③ Gated operation. Transients are eliminated, allowing accurate evaluation of ramp.

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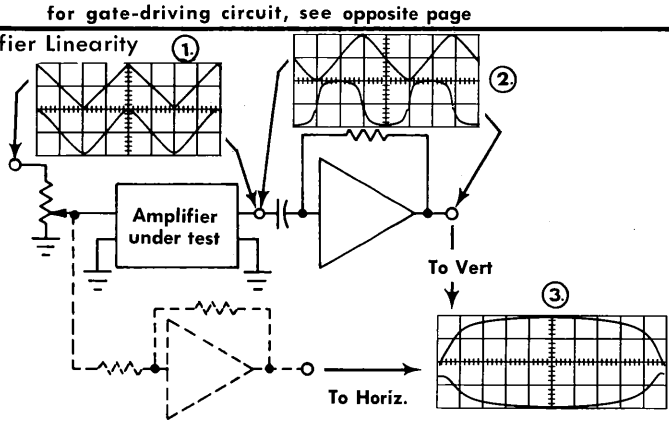
Amplifier Linearity



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① Input to Amp. under test.
Output of Amp. under test.

② Output of Amp. under test.
Same, differentiated.

③ Vert: Differentiator Output.
Horiz: Input to Amp. Under test.

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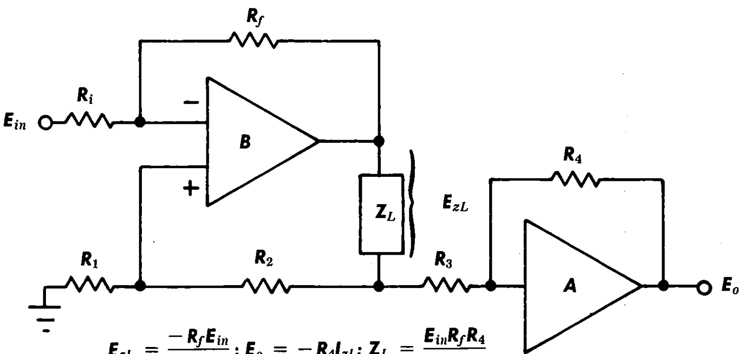
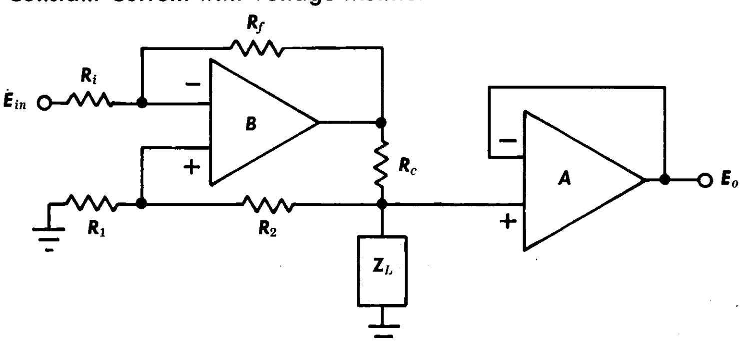
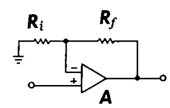
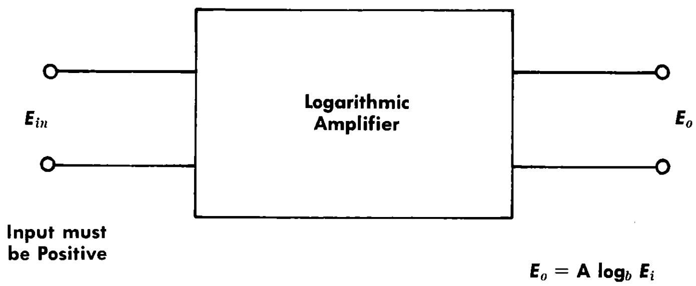
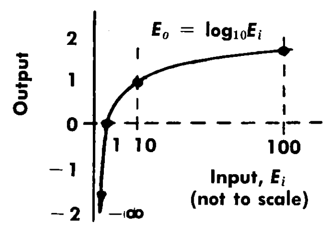
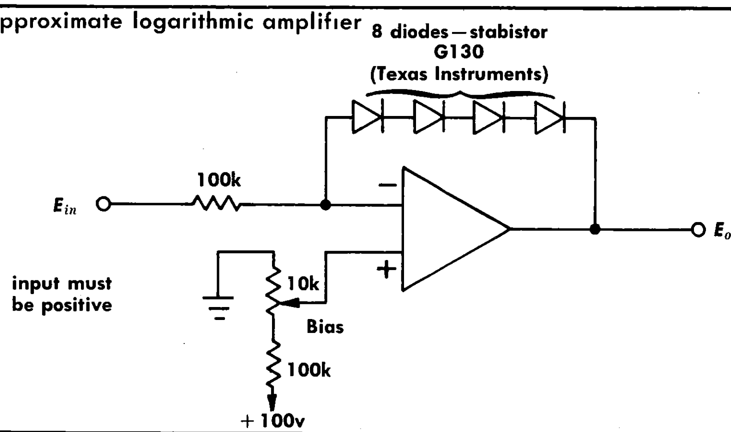
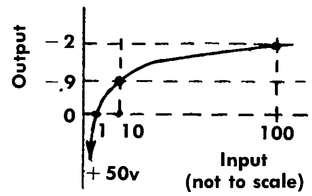
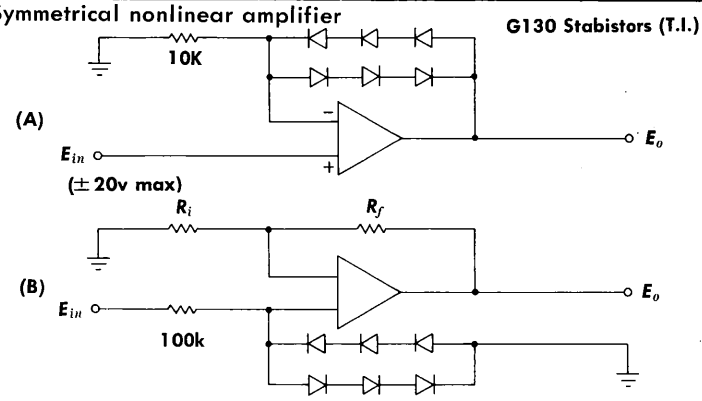
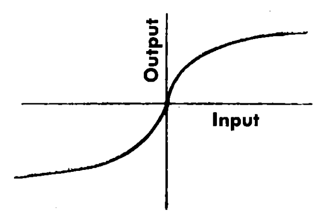
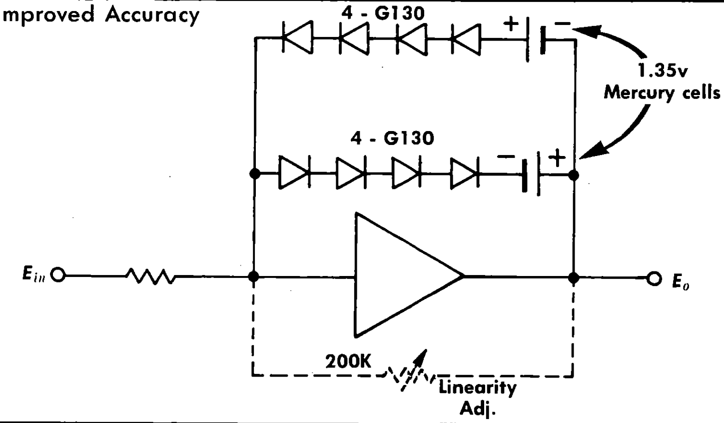
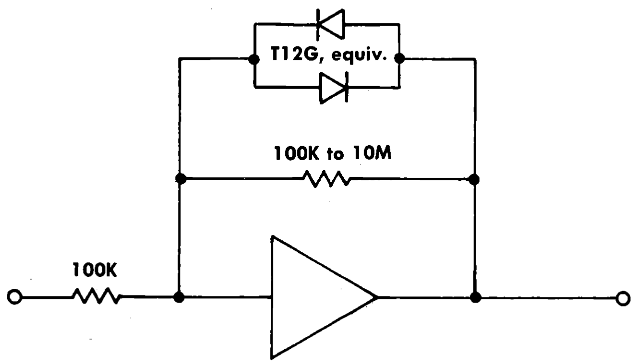
Voltage-Current Conversions	Notes
Constant Voltage with Current Monitor  $E_{zL} = \frac{-R_f E_{in}}{R_i}$; $E_o = -R_4 I_{zL}$; $Z_L = \frac{E_{in} R_f R_4}{E_o R_i}$	A practical circuit for component-testing, where Z_L need not be grounded. R_2/R_1 should be equal to R_f/R_i, and $R_3 < Z_L \ll R_1 + R_2$. R_1 may be trimmed or padded to correct tolerance or system errors. For Type O, E_{ZL} and E_o should not exceed 50v; I_{ZL} should not exceed 5ma.
Constant Current with Voltage Monitor  $\frac{R_2}{R_1} = \frac{R_f}{R_i}$; $R_c > Z_L \ll R_1 + R_2$ $I_L = \frac{-R_f E_{in}}{R_i R_c}$; $E_o \approx E_{Z_L}$	Monitor amplifier A has very high (GΩ) input impedance, so has negligible effect on current through Z_L. A small ($\approx 2\%$) correction may be necessary for gain of A where high accuracy is required, thus:  $R_f = R_i\left(\frac{1}{\text{observed Gain}} - 1\right)$ **Example: Observed gain, 0.98. Selecting 100K for R_i, then $R_f = 100K\left(\frac{1}{0.98} - 1\right) = 2.04K$.**

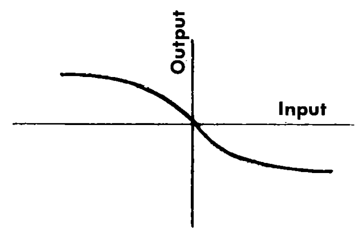
Low-Frequency Applications ("Log" diodes)	Notes
Thēoretical logarithmic amplifier  	 An advantage of a true log amplifier is capability of accepting a wide range of input signal levels. Since input can not be negative value, for ac measurements, a symmetrical amplifier is recommended.
Approximate logarithmic amplifier  	 Simulating a true log amplifier. Bias is adjusted to give zero volt output when input is +1v. When input is zero, amplifier saturates to ≈ 50 volts. Many germanium diodes can be used such as T12G. For more dynamic range and controlled characteristics, stabistors are recommended.
Symmetrical nonlinear amplifier  	 Circuit (A) provides very high (≈ GΩ) input resistance for low level applications. E_{in}, ±20v max. Circuit (B) permits output level to be adjusted by selection of R_f and R_i.
Improved Accuracy  	Small, hearing-aid type mercury cells may be used for bias to obtain better accuracy, especially at low levels. For best shelf-life disconnect batteries when not in use. Further improvement for a given application is obtained with linearity adjustment pot, ≈ 200K.

Wideband Amplifiers (Signal Diodes)	Notes

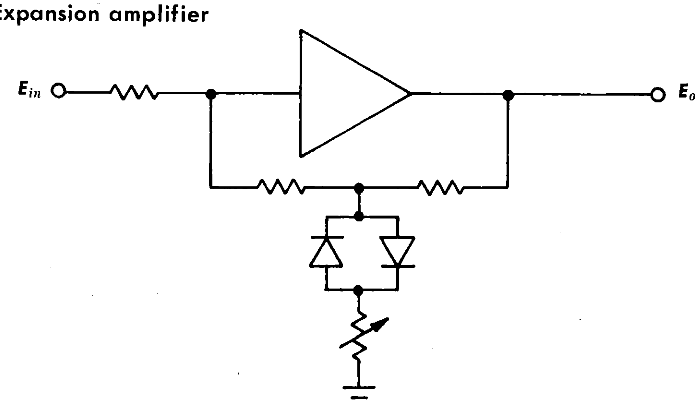
Simple nonlinear amplifier

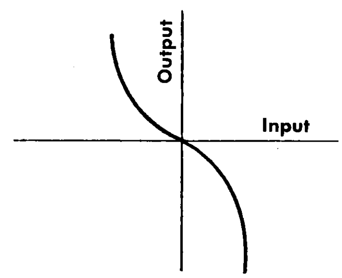


If Germanium diodes are placed in the feedback circuit, this amplifier acts as compression type amplifier.

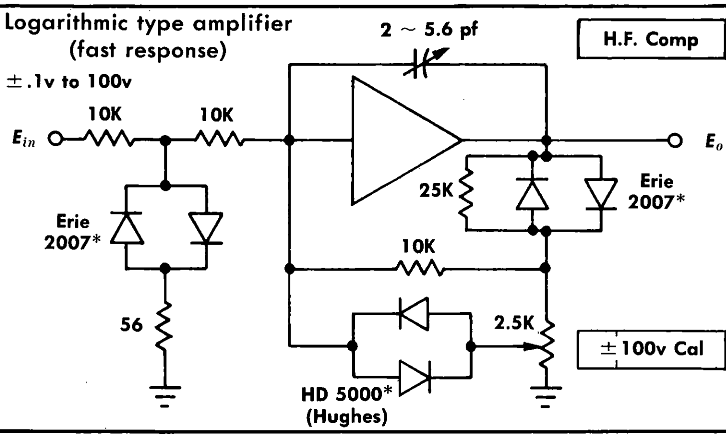


Expansion amplifier

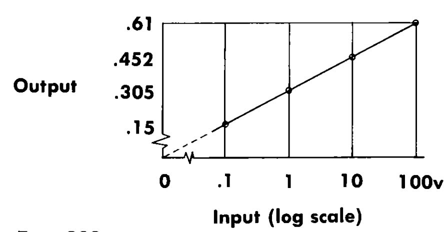




Logarithmic type amplifier (fast response)

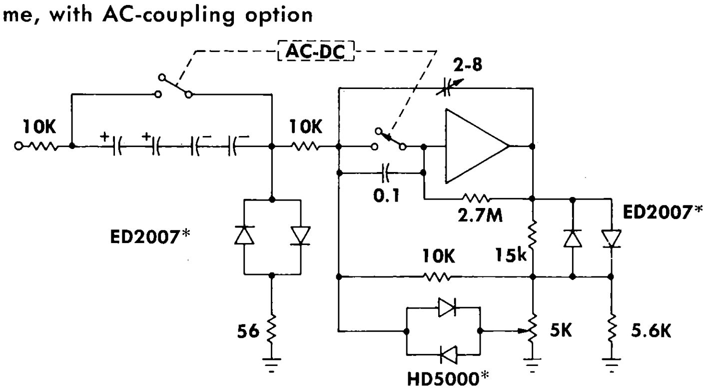


E_o .305 + .15 $\log_{10} E_{in}$

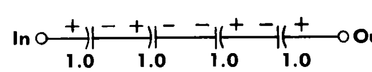


$T_r \approx$ 200 nsec
$T_f \approx$ 300 nsec
For 10v signal.
Diode pairs must be carefully selected.

Same, with AC-coupling option

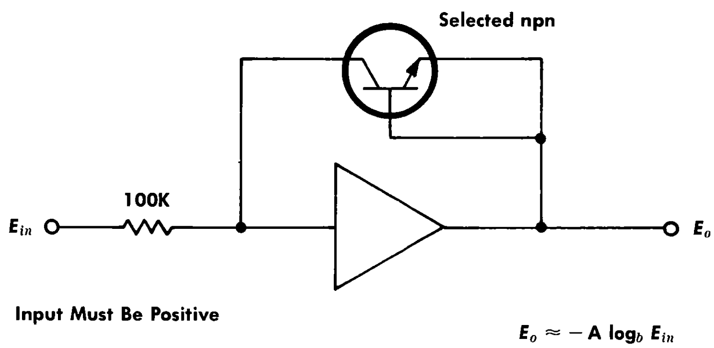
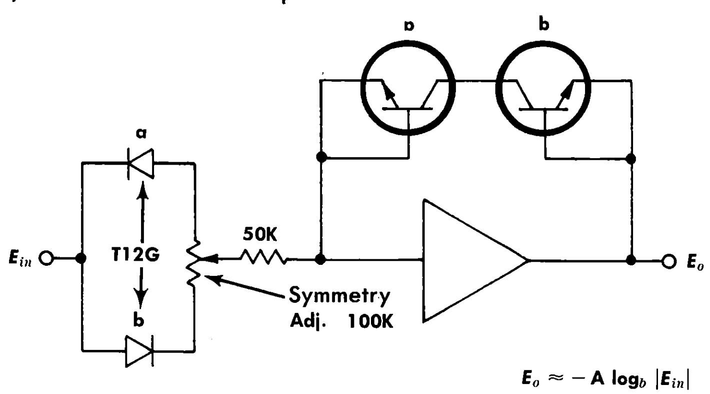
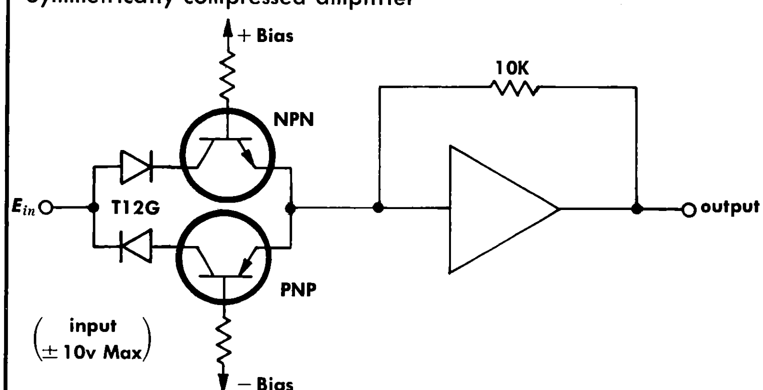
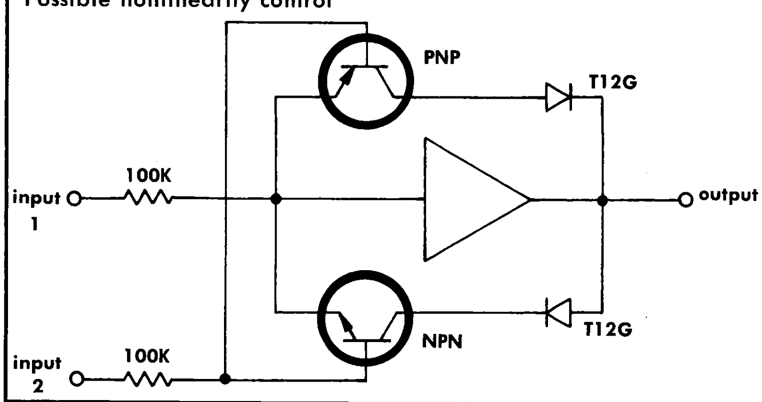


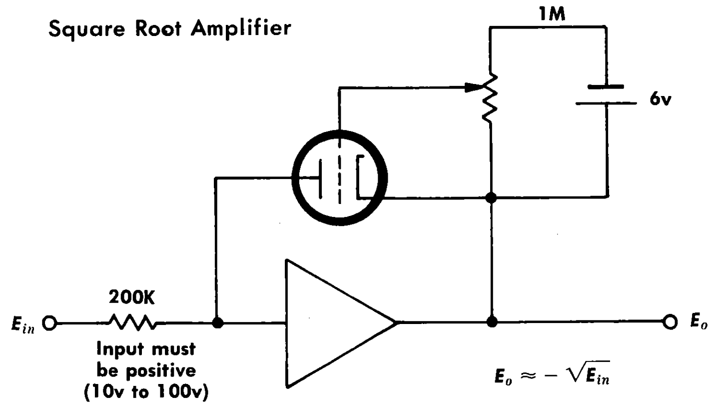
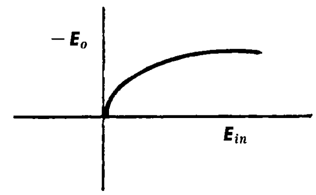
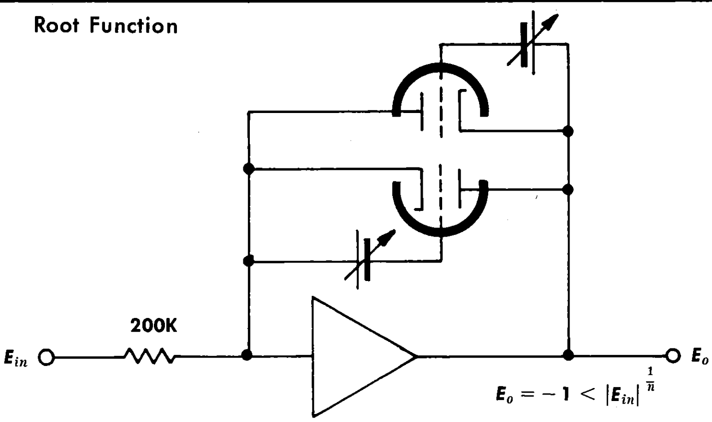
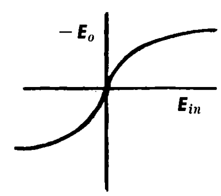
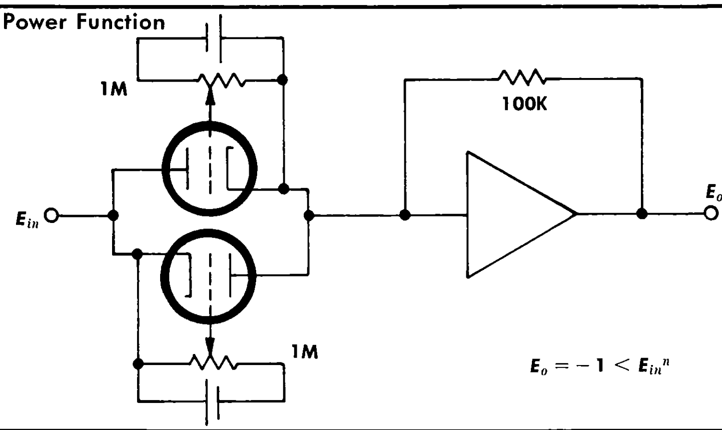
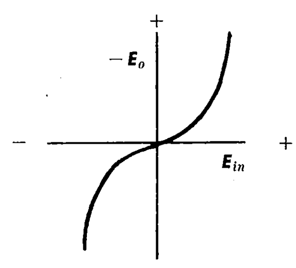
Input coupling can be made to have a long time-constant by using tantalum electrolytic capacitors in series, back-to-back, thus:



The tantalum capacitors can accept a small reverse bias, so two 0.5 μf 100v, or four 1.0 μf, 50v units can be made to provide 0.25 μf, 100v service. This entire adapter is available (assembled and calibrated) from Tektronix, under *P/n* 013-0067-00.

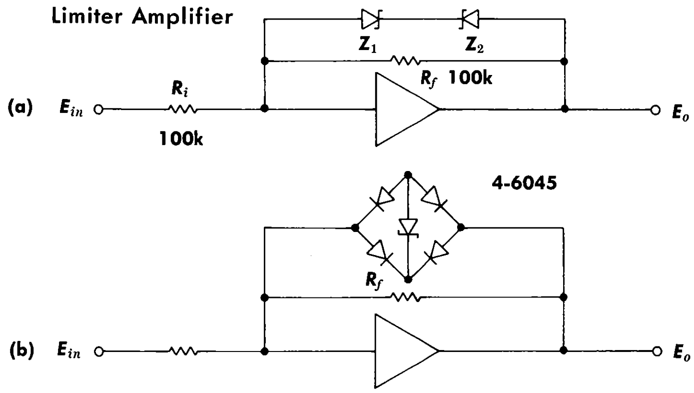
*Selected and matched. See appendix.

Transistors As Passive Nonlinear Elements	Notes
Approximate logarithmic 	Certain NPN transistors exhibit logarithmic voltage-current breakdown characteristics in the reverse direction. This arrangement can accept only a positive input signal. Selection of NPN transistors may make this type of logarithmic amplifier impractical since the use of certain type diodes can give more predictable logarithmic characteristics. (only advantage is high level output)
Symmetrical nonlinear amplifier 	To accept positive and negative signal two transistors (matched) are used. Signals are same as listed above.
Symmetrically compressed amplifier 	
Possible nonlinearity control 	

Vacuum Triodes As Nonlinear Elements	Notes
Square Root Amplifier 	 Certain low-μ triodes such as 12AU7 or Nuvistor 7586 may be placed in the feedback circuit to obtain square root characteristics.
Root Function 	 Two triodes placed back to back in the feedback circuit. By adjusting bias, the root may be varied from approximately .9 to 10.
Power Function 	 If the triodes are placed in the input circuit, power function may be obtained. The power may be varied from .9 to 10 approximately.

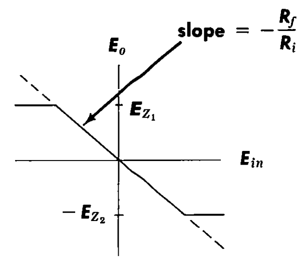
Zener Diodes As Nonlinear Elements	Notes

Limiter Amplifier

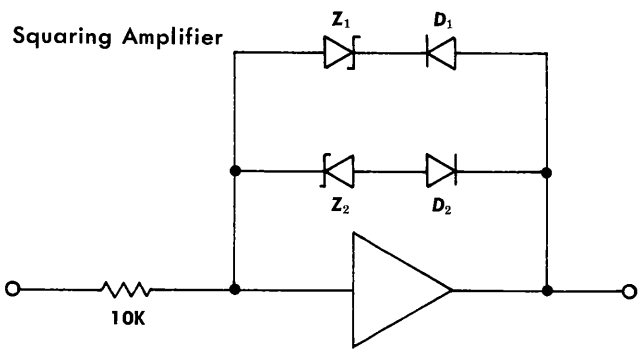


(a) If back to back zener diodes are inserted in the feedback circuit, the output voltage will be limited by the zener breakdown voltage.

(b) For perfect symmetry, use one Zener diode and diode bridge as shown. Low-C, low-leakage diodes allow more accurate, faster operations.

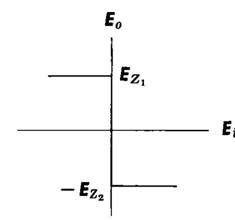


Squaring Amplifier

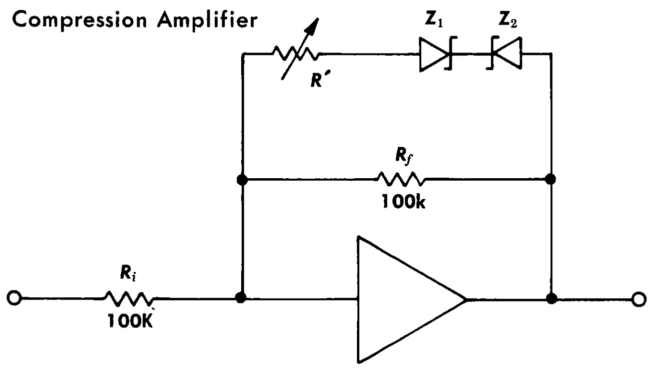


If the feedback resistor in the above amplifier is removed, the amplifier output gives very high gain near zero signal. In this arrangement the zener diodes are coupled with low capacity diodes to minimize zener diode capacitance.

For perfect symmetry use diode bridge as above. (Additional series diodes are then unneeded.)

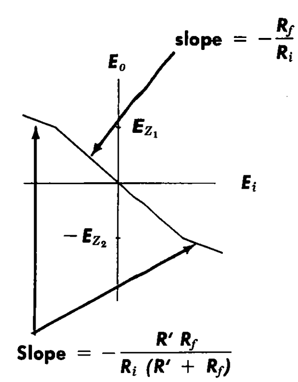


Compression Amplifier

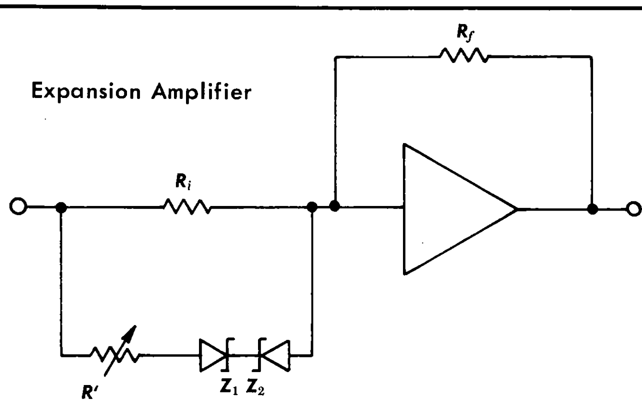


A variable resistor in series with the zener diode in the limiter amplifier can control the amount of compression for a large signal.

For symmetry, use bridge arrangement.

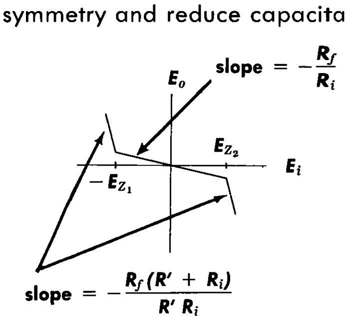


Expansion Amplifier



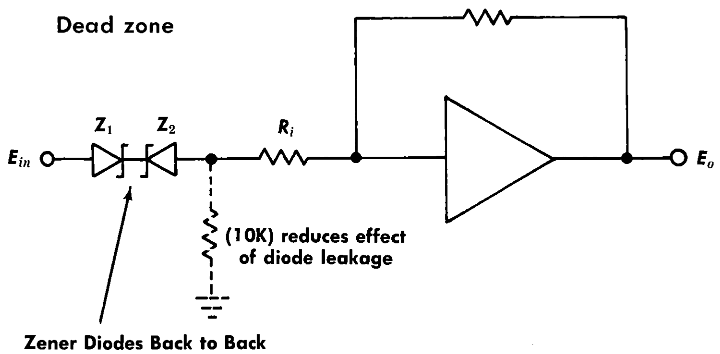
Input and output circuits are reversed in this amplifier to obtain higher gain for large signal input.

Again, the diode bridge arrangement may be used to provide symmetry and reduce capacitance.

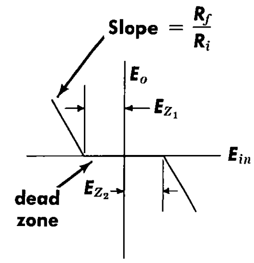


Zener diodes—Simulation of Nonlinear Mechanical Phenomena

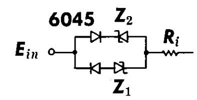
Dead zone



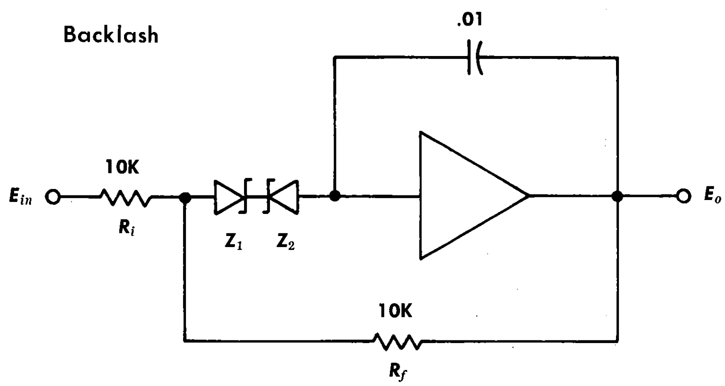
Simulating a dead zone



Bridge arrangement (p 3-5) may be used where symmetry, capacitance or leakage are a problem. For asymmetrical operation with minimum leakage and C, use input circuit shown:



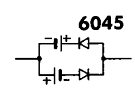
Backlash

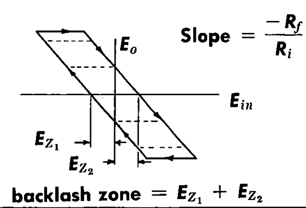


Simulation of backlash:

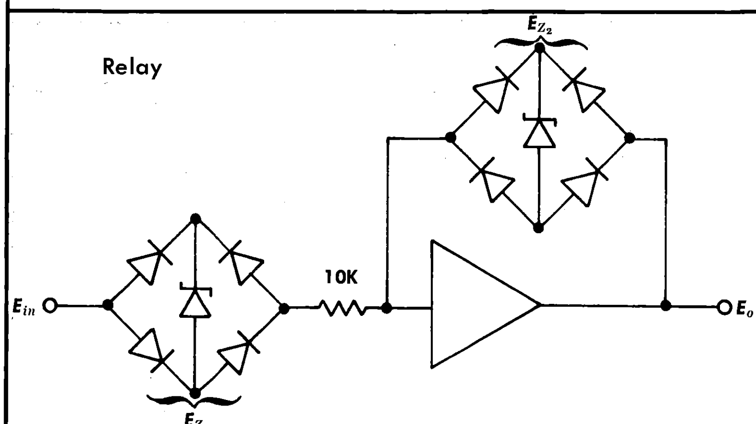
The capacitor serves as a memory to hold the output constant as E_{in} moves through backlash zone (changes direction).

For low-leakage memory, use small (hearing-aid) batteries and low leakage diodes



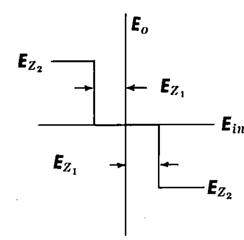


Relay



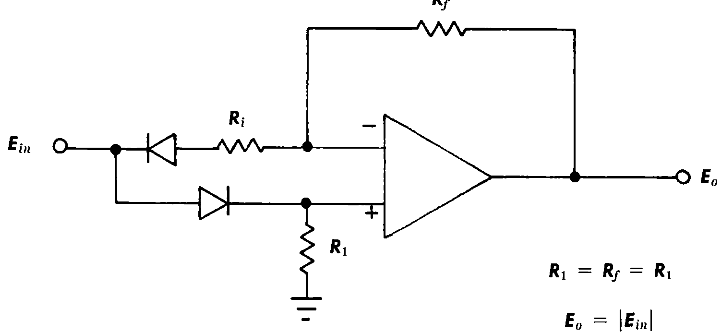
Representation of polarized double-throw relay which has equal pull-in and pull-out current.

Asymmetrical systems may be simulated by using Zener diodes and signal diodes in series as in the "Dead Zone" notes above.

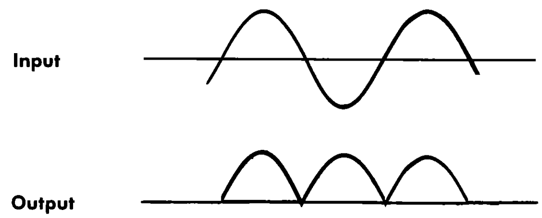


Absolute Value Amplifiers — Notes

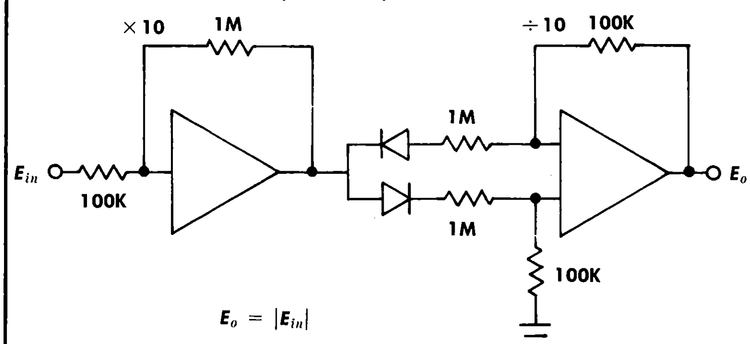
Full-wave rectification



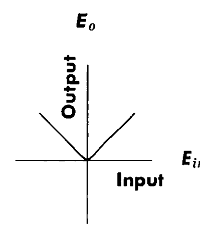
Output gives full-wave rectification of input. Output is always positive sign. Diodes contribute ≈ .3 voltage drop.



Full-wave rectification (low level)



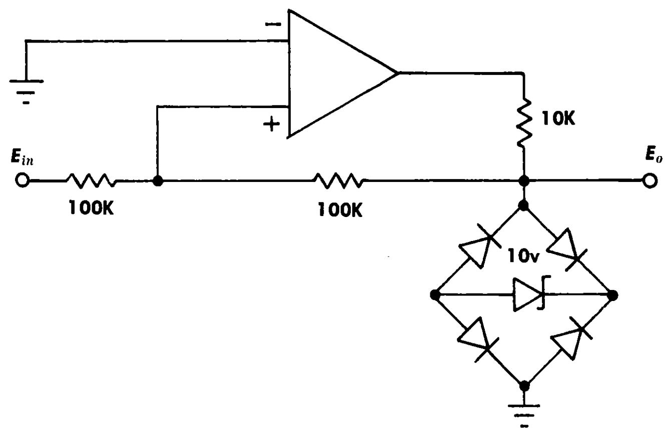
By inserting an amplifier ahead of full-wave rectifier, the effect of diode voltage drop is reduced by the factor of amplification. This technique may be applied to low level full-wave rectification.



Notes

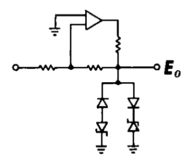
Switching Circuits	Notes

Bistable Comparator (± input)

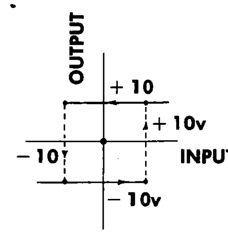


The Bistable Comparator incorporates a zener diode as a reference. The input voltage is compared against the reference diode; when the input voltage becomes equal to the zener voltage, output will flip to 10v of the same polarity as input.

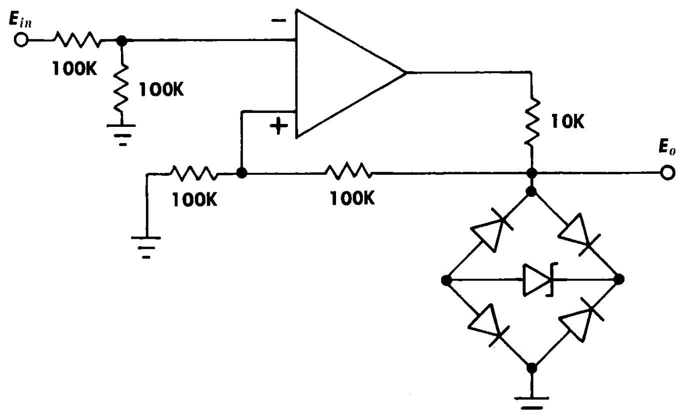
Use of different references for + and − comparisons is also possible:



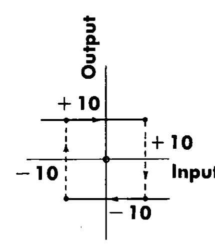
Use of decoupling diodes improves switching time.



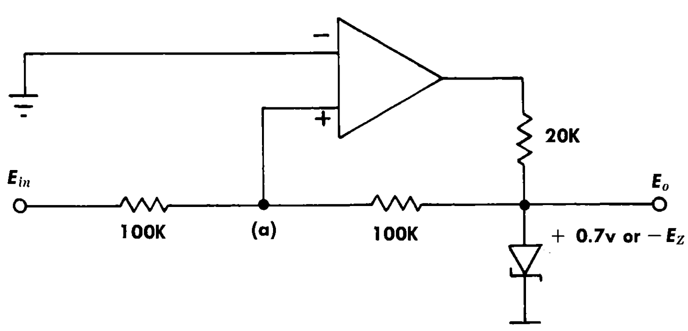
Bistable Comparator, inverting.



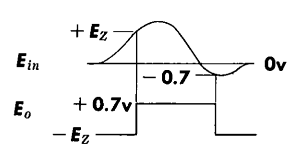
Similar to above, but output will flip to opposite polarity.



Bistable Comparator, unidirectional

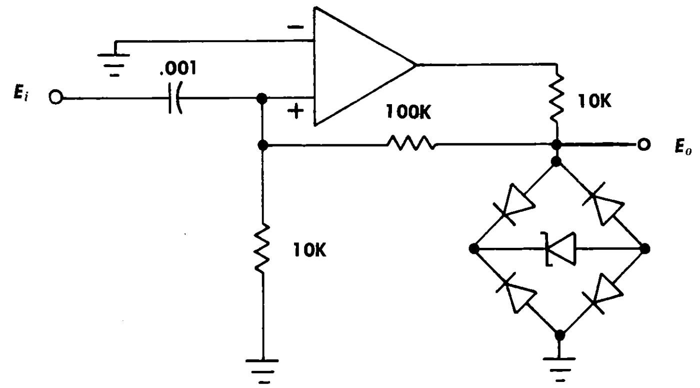


Single zener diode flips comparator when $E_{in} = +E_z$ or − 0.7v. (Reset level governed by forward resistance of zener). Reset can be made to occur at 0v if − grid is returned to about + 0.7v. Comparator point will become E_z − 0.7v. Bias may also be injected at (a) for same result.

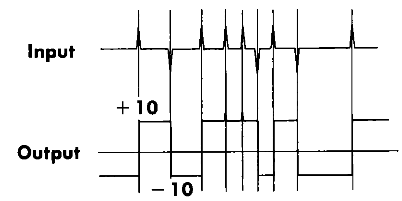


Bistable Multivibrators	Notes

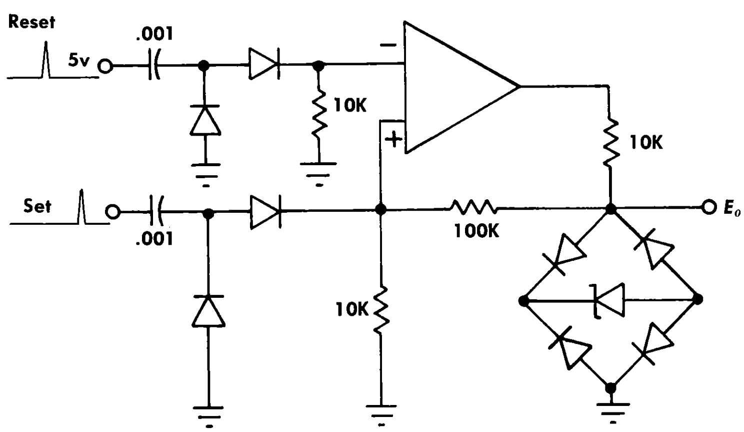
Bistable Multivibrator
(single input)



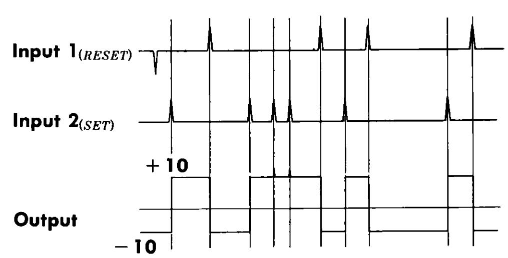
Output flips to ±10v anytime an opposite-polarity pulse arrives. Once flipped, additional pulses of the same polarity have no effect:



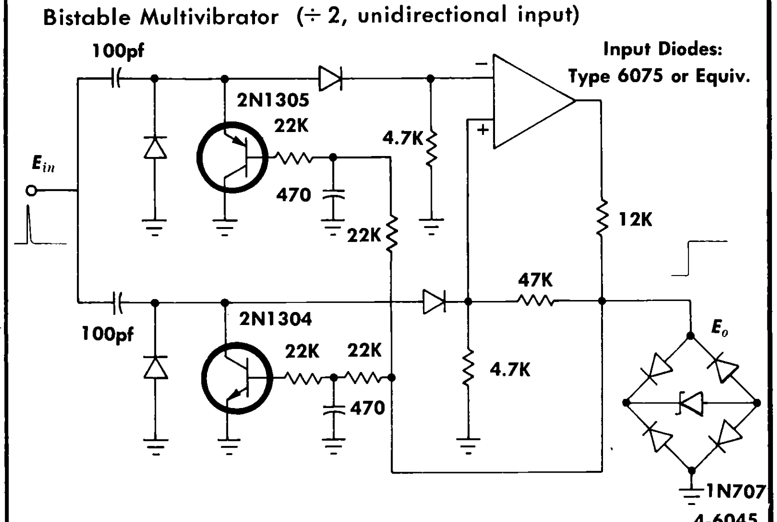
Bistable Multivibrator
(two inputs)



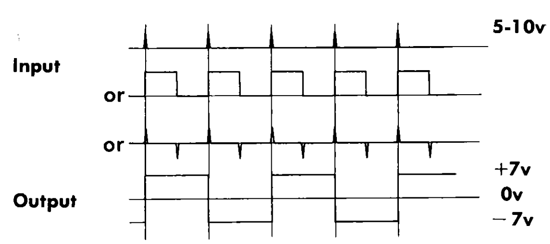
AC (pulse) toggle switch action can be obtained.



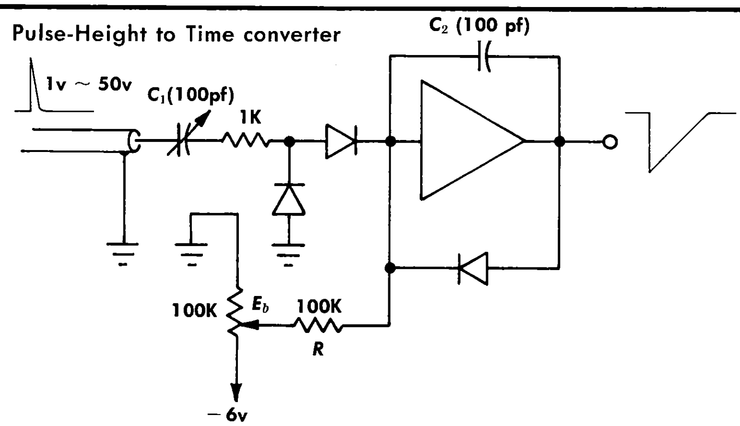
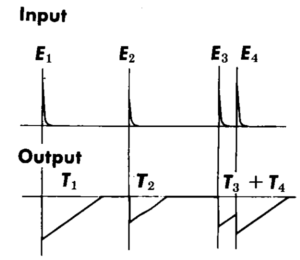
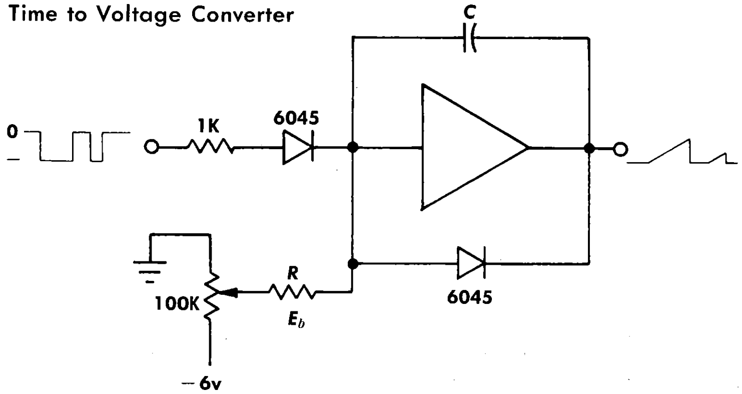
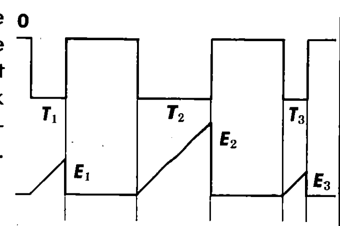
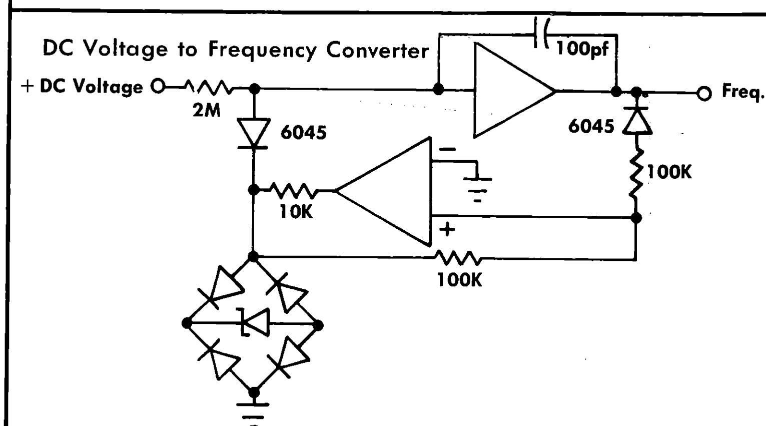
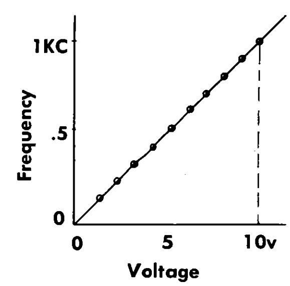
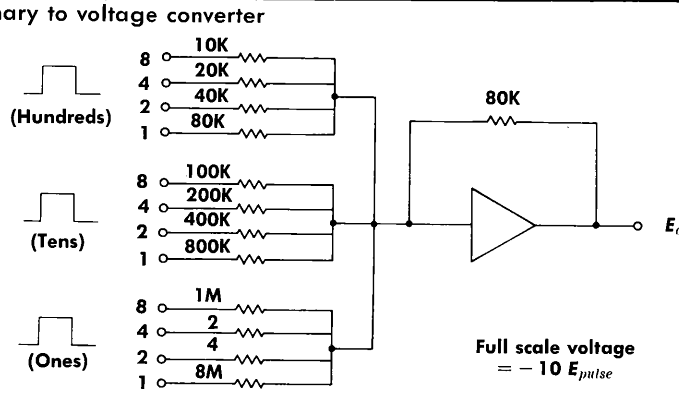
Bistable Multivibrator (÷ 2, unidirectional input)

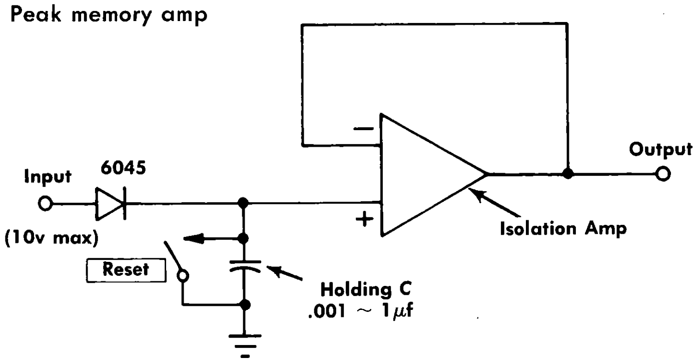
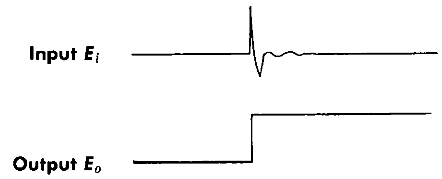
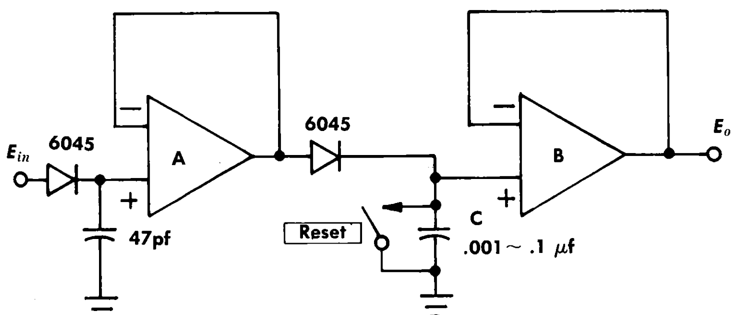
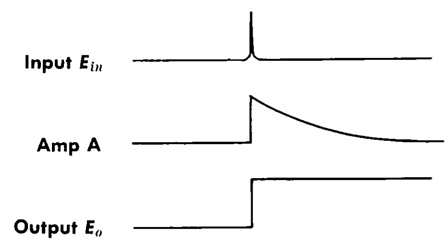
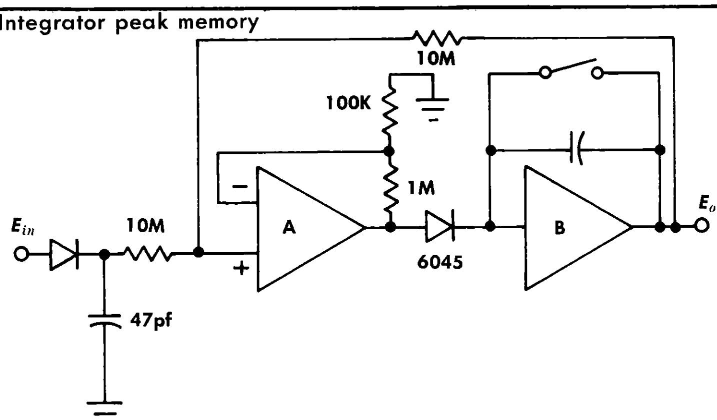


Transistors perform steering function to provide ÷ 2 action. Lag time-constant in transistor base circuits is selected for application.



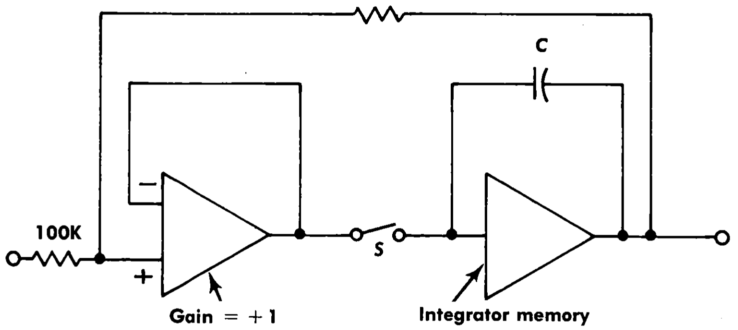
Transistors may be any moderately fast complementary pair (see appendix).

Digital Conversions	Notes
Pulse-Height to Time converter 	If a positive pulse is applied to the input, it will charge C_1 and this charge is also dumped into C_2, causing the output to step down. This voltage is discharged by integrator.  $T \cong \frac{RC_1}{E_b} E_{pulse}$
Time to Voltage Converter 	A negative gate opens the diode gate allowing the integrator to integrate at a constant rate. The peak voltage is thus proportional to the time period.  $E \cong \frac{E_b}{RC} T$
	This will give conversion factor of approximately 1 kc per 10 volts input. The conversion accuracy is in the neighborhood of $\pm 1\%$. 
Binary to voltage converter 	Simple Binary (BCD) to voltage conversion is accomplished by summing input pulse voltages through the properly weighted input resistors. If input pulse is -1.0 volt, the full scale voltage will be $+9.99$ volts. Gain error $= 1 - \frac{A}{A - 1 - 16.65}$ if BCD signal source (esp. 'hundreds' signal) has low impedance.

Peak Memory Circuits	Notes
Peak memory amp 	 The holding capacitor is charged through a low-leakage diode. Since the leakage of the diode (1nanoamp) and grid current (.5 namp) are low, voltage E_o can be held a relatively long time.
Fast response peak memory amp 	 To reduce input loading effect, two amplifiers may be cascaded.
Integrator peak memory 	Integrator is used as memory to improve DC accuracy.

Sample-Hold Amplifier

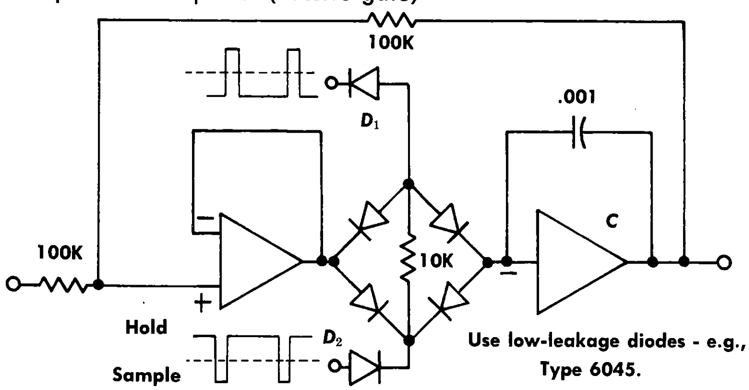
Sample hold



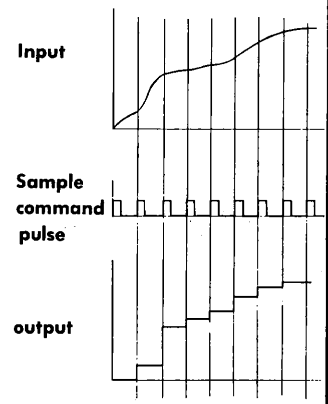
(Illustration of sample hold circuit.)

This circuit is often called **integrator memory:** When the switch is closed the capacitor is charged by the first amplifier. When the switch opens the integrator holds the voltage until the switch is closed again.

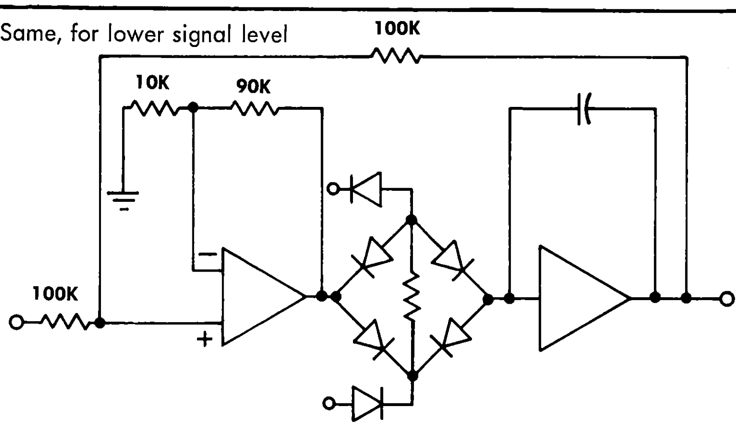
Sample Hold amplifier (Passive gate)



Passive gate provides output of $E_{in} - 2\ E_d$ for each sample, where E_d is forward drop of gate diodes. Advantage of passive gate is that "sample" command + and − pulses need not be perfectly balanced since D_1 and D_2 simply disconnect during "sample" time.

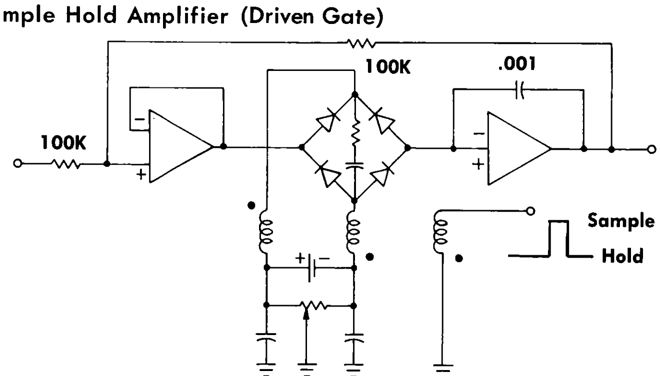


Same, for lower signal level

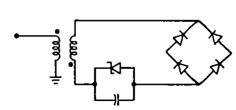


If diode forward drop is significant in the circuit above, gain may be added to input amplifier to minimize this effect.

Sample Hold Amplifier (Driven Gate)



As above, but battery provides "hold" voltage; balanced "sample" pulses obtained via pulse transformer. Battery voltage must be larger than expected input voltage; "sample" command pulse larger than battery voltage. Battery may be eliminated if sampling rate is high and hold time is not long:

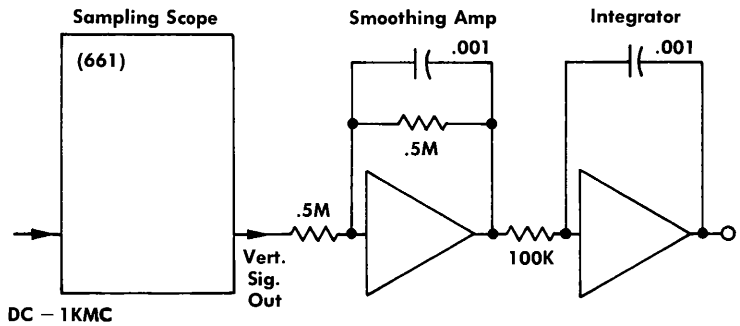


$E_{Zener} > E_{in}$ max.

Notes

Processing of Equivalent-Time Signals In Real Time. — Notes

Sampling Integration

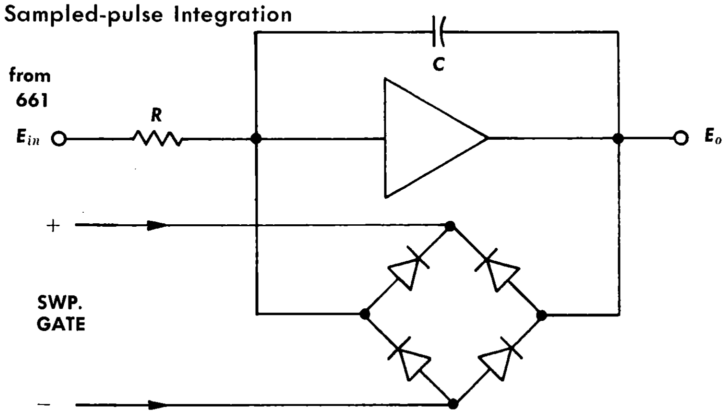


$$G_{(s)} \approx \frac{1 - e - \tau S}{S(\tau_1 S + 1)} \qquad \frac{1}{\tau_2 S + 1} \qquad \frac{K}{S}$$

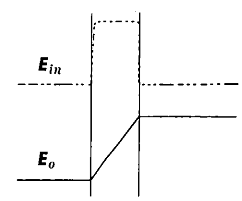
Sampled-data contain many higher frequency components as well as desired lower frequency signals. In many operations, it may be necessary to insert a smoothing amplifier before doing a linear operation. It is possible to integrate a repetitive signal to approximately 1 kmc using sampling scope vertical signal out.

For differentiation and integration it may be necessary to drive sampling oscilloscope horizontal with slow external ramp to force a fixed real-time equivalent-time scaling factor (See Appendix).

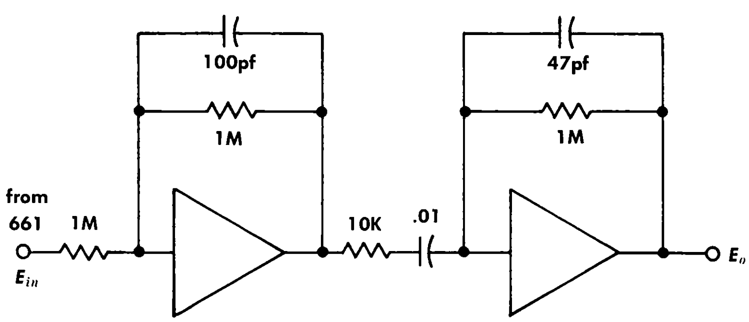
Sampled-pulse Integration



Sampled-pulse may be integrated to obtain area of the pulse in volt sec. Proper scaling factor to convert real time to sampling time is necessary; real: equivalent time ratio must be held constant. For details of gating circuit, see gated integrator (page 2-4).

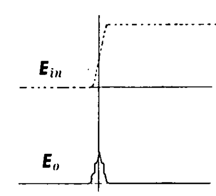


Sampled-pulse Differentiation

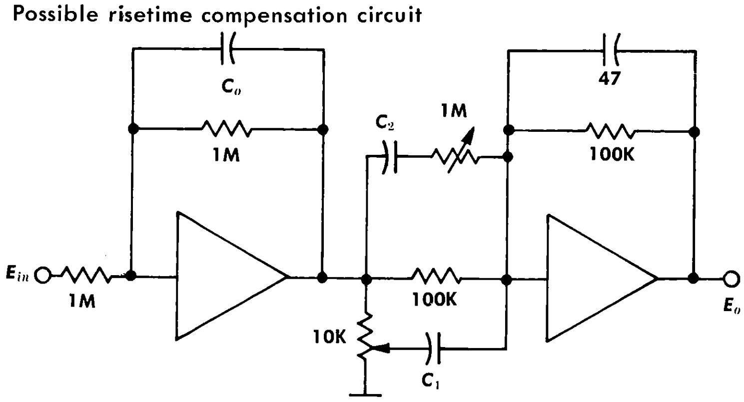


Differentiation of sampled pulse may be performed after passing the sampled signal through a smoothing amplifier.

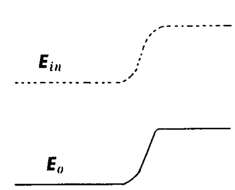
Additional H. F. limiting CIRCUIT is necessary in differentiation. Here again, time-sealing factor must be held constant in sampling oscilloscope.



Possible risetime compensation circuit



This arrangement can provide possible risetime improvement, however this type of compensation is applicable only to a given frequency and a given sweep speed.

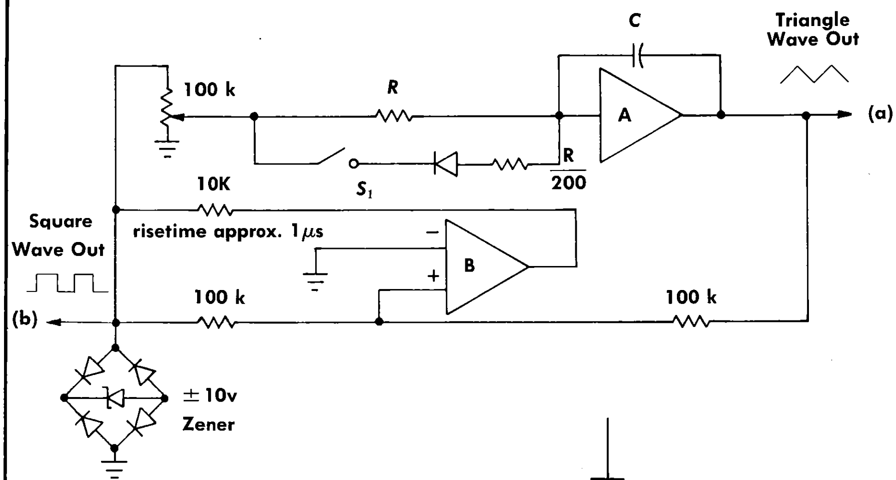


Notes

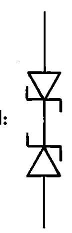
L. F. Function Generators	Notes

Triangle, Square, Sawtooth Pulse Generator

Combining an integrator and a bistable comparator, precise low frequency waveforms can be generated. Operational amplifier B is working as a bistable comparator which switches when input reaches + or − 10 volts for ± 10 volts square wave output.

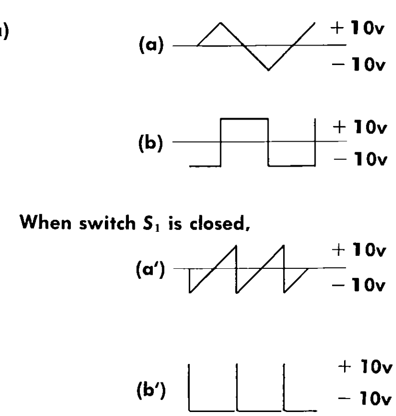


Two Zener diodes may be used:



or ± Zener.

Normal waveforms at (a) and (b) are



$f = \frac{1}{4RC}$

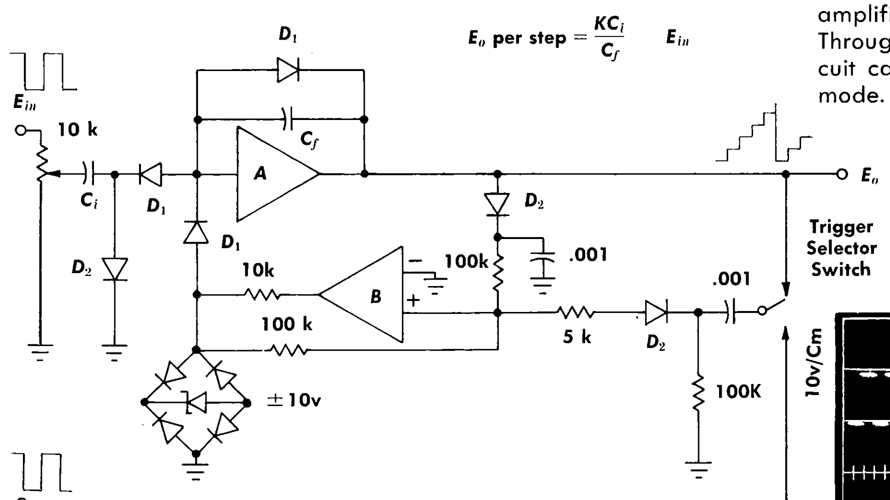
$f = \frac{1}{2RC}$ (when switch S_1 is closed)

Stair-step Generator

This arrangement provides a convenient laboratory stair-step generator useful in applications such as pulse rate countdown, transistor base drive, or pulse-controlled time base.

Amplifier A operates as a pulse integrator and amplifier B functions as a bistable comparator. Through use of the trigger selector switch, the circuit can operate in either a repetitive or triggered mode.

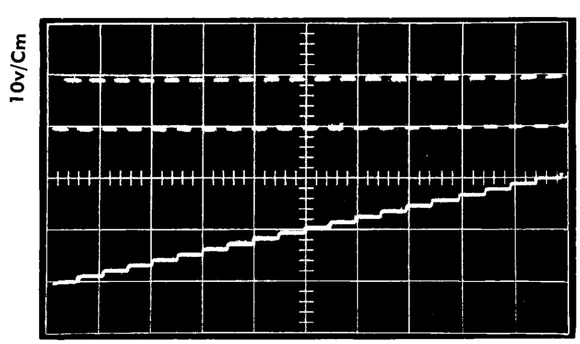
E_o per step $= \frac{KC_i}{C_f}$ E_{in}



Ext. Trig. (Single trigger should be − going.)

D_1 = very low leakage silicon diodes (Type 6045 or equiv.)

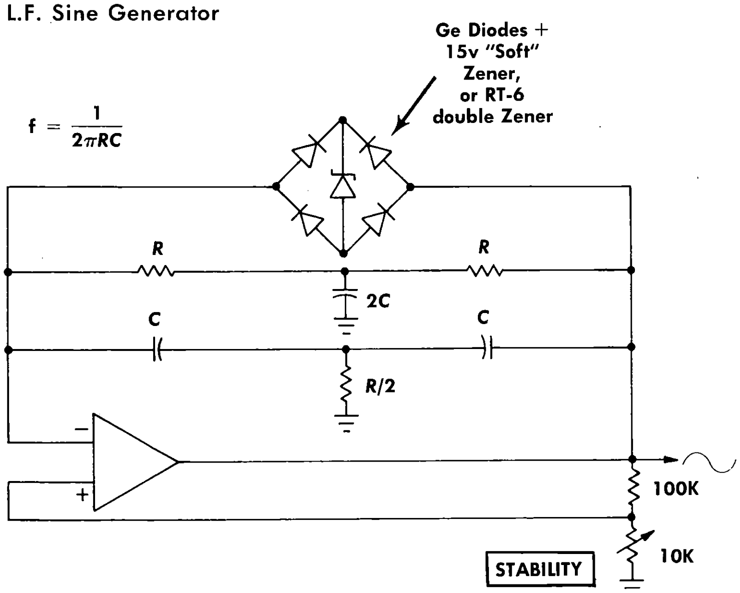
D_2 = Any signal diodes (T12G or equiv.)



STAIR-STEP Generator 5 v/cm Sweep 2 msec/cm

Sine Wave Generators	Notes
L.F. Sine Generator	

$f = \frac{1}{2\pi RC}$



A very low frequency sinewave can be generated with twin T filter network in the feedback loop of an operational amplifier. Positive feedback is provided to sustain oscillation. Amplitude is controlled by the nonlinear impedance of zener diode in the negative feedback. Adjust STABILITY control for stable output and best waveshape.

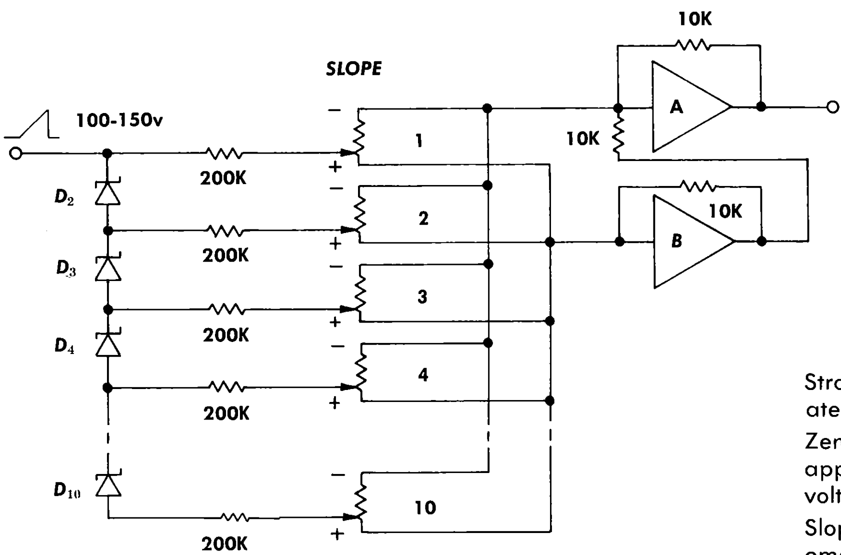
(c)


$f = \frac{1}{2\pi RC}$

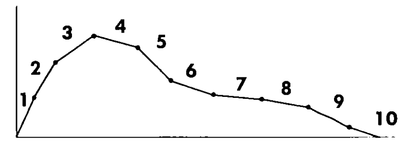
Zener type should be selected for a gradual ("soft") knee. If an output amplitude of 4v p-p is adequate, Type RT-6 double Zener may be used to replace bridge circuit. Reject Zener diodes (High dynamic impedance) often work quite well.

Arbitrary Function Generators	Notes

Segments function generator



9 Diodes, D_2 to D_{10} are 12 to 15v zener diodes For 100v sawtooth, use 8-10v zener diodes.

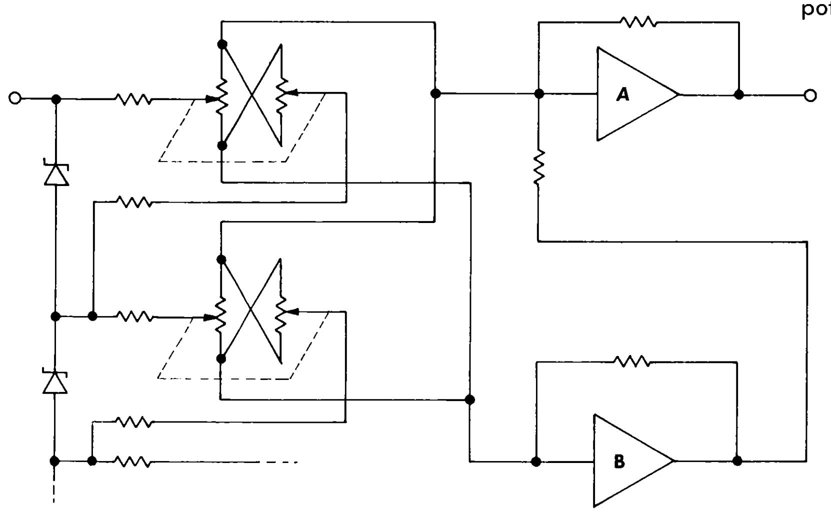


Straight line approximated function can be generated with the aid of zener diodes and potentiometers.

Zener diodes are normally cut-off. As sawtooth is applied zener diodes start conducting at even voltage intervals.

Slope of each segment is adjusted by the potentiometers. Note that no external voltages are required, except the sweep sawtooth.

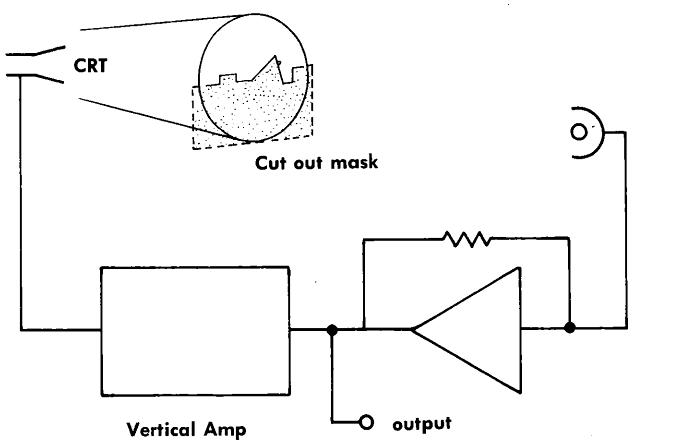
Segments function generator (Independent Segments)



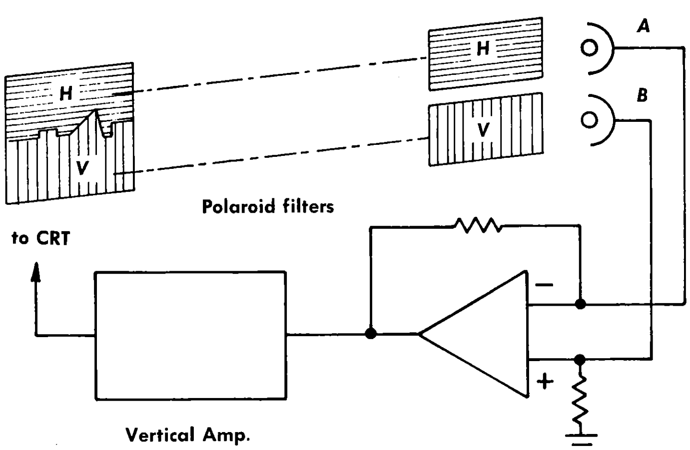
Segment slopes in the above circuit are additive. To obtain more independent adjustment of segments 2-10, cancelling signal is fed back to earlier-segment potentiometers.

Arbitrary Function Generators—Optical Techniques

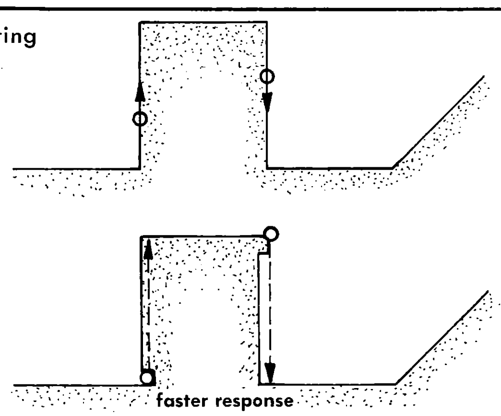
Simple System



Differential Fast System



Mask Cutting



Notes

A light sensitive element is placed in front of the CRT and the mask. If the beam is above the mask, a negative signal will be produced to force the beam to move down. Normally the beam is positioned above the mask so that beam always tries to go up; however, the light feedback does not allow the beam to show too much. As a result the beam follows the cut out mask. DC coupled signal can be obtained at the output of operational amplifier.

When using the simple system the spot size has to be made small to obtain good resolution, but when it is operated at a low light level it will suffer in speed.

The differential system incorporates polaroid filter masks and two light sensing elements. Substantial improvement in speed and resolution may be expected.

When the beam spot is in H plane the light is sensed by light sensor A. B will not receive light since the filter is out of phase. Thus the signal from A will force the beam to move down. If the beam spot is in V plane it, likewise, will be forced upward by B.

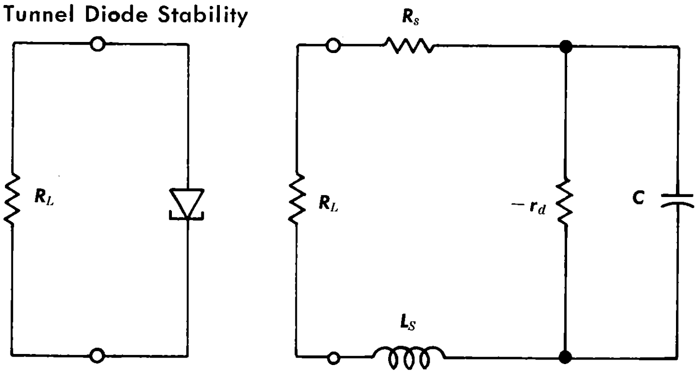
The light is picked up differentially and the beam is moved push-pull action. With a cadmium sulfide photocell (Clairex CL-404) response is limited to the millisecond region. Faster response may be obtained from vacuum phototubes, phototransistors or photomultipliers, at increased cost. System is ultimately limited by oscilloscope delay line, phosphor risetime and decay, etc.; 1 μsec response is probably the practical limit.

Suggested application: integration or differentiation of photographed or drawn waveforms or shapes.

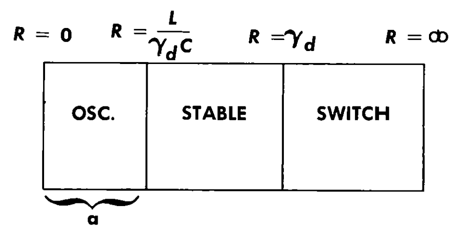
The mask can be under cut to improve response time. The polaroid filter may be cut by using a hot resistance wire; the temperature can be adjusted by controlling current.

Tunnel Diode Studies—Curve Tracing	Notes

Tunnel Diode Stability

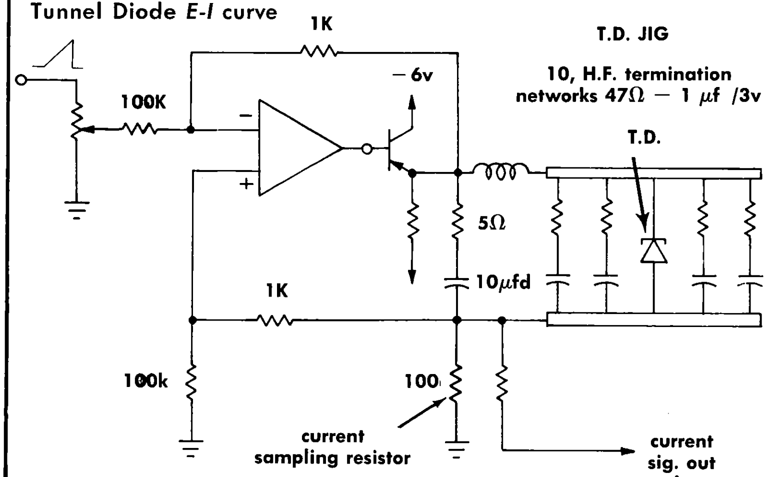


Tunnel diode stability criteria are given



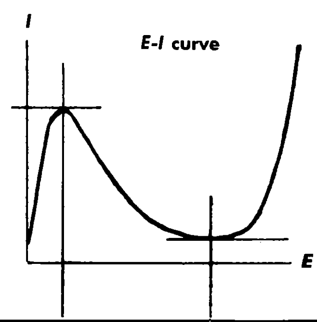
a. (can be made conditionally stable at low frequency)

Tunnel Diode *E-I* curve

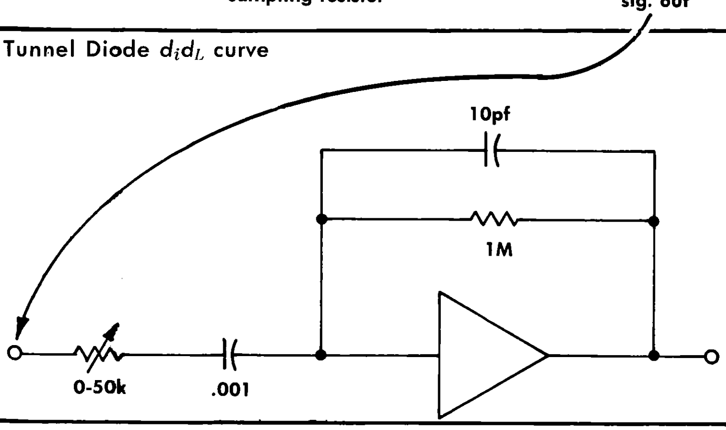


Operational amplifier with additional E.F. can drive tunnel diode with very low impedance.

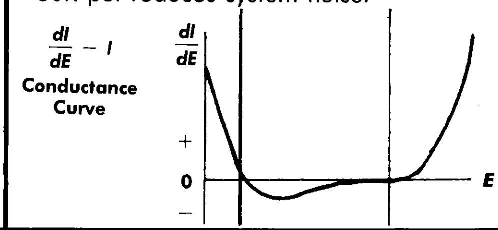
T.D. jig stabilizes diode at high frequency. Thus tunnel diode may be driven by very slow ramp.



Tunnel Diode $d_i d_L$ curve

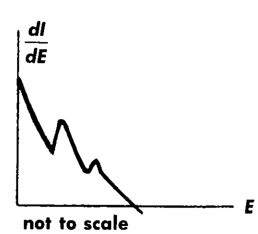


Linear drive of tunnel diode allows differentiation of current to obtain dI/dE curve, plotted against E (oscilloscope sawtooth). Input to differentiator is current signal output from curve-tracer above. 10pf capacitor compensates the differentiator; 50K pot reduces system noise.



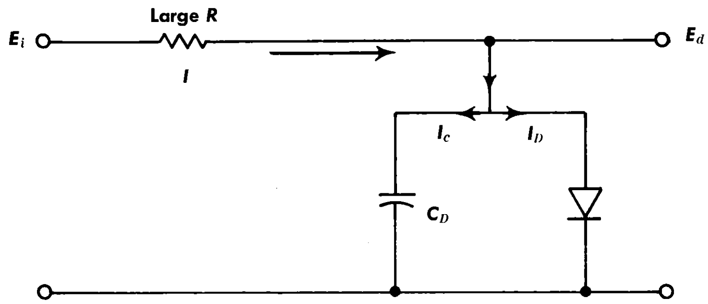
Tunnel diode quantum phenomenon may be observed at very low temperature.

Some tunnel diodes in liquid Helium indicate wiggles on the conductance curve.



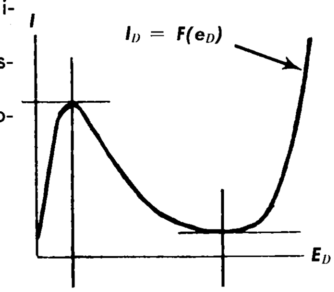
Tunnel Diode Studies, Switching characteristics	Notes

Tunnel Diode Switching



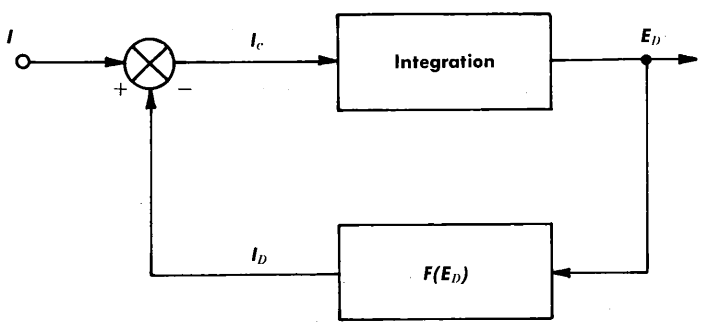
C_D = Equivalent diode capacity

Assumptions:
1) Lead inductance negligible.
2) Diode capacitance assumed constant.
3) Constant current applied to the diode.



I = applied current, I_D = diode current.
I_c = current to charge diode capacity

Block Diagram



Summing all the current at the node of E_D, we have

$$I - I_c - I_D = 0 \qquad \dots\dots 1$$

$$I_c = I - I_D \qquad \dots\dots 1a$$

$$I = \frac{E_i - E_D}{R} \qquad \dots\dots 2$$

$$I_c = C_D \int \frac{dE_D}{dt} \qquad \dots\dots 3$$

$$I_D = F(E_D) \qquad \dots\dots 4$$

Integrating (3)

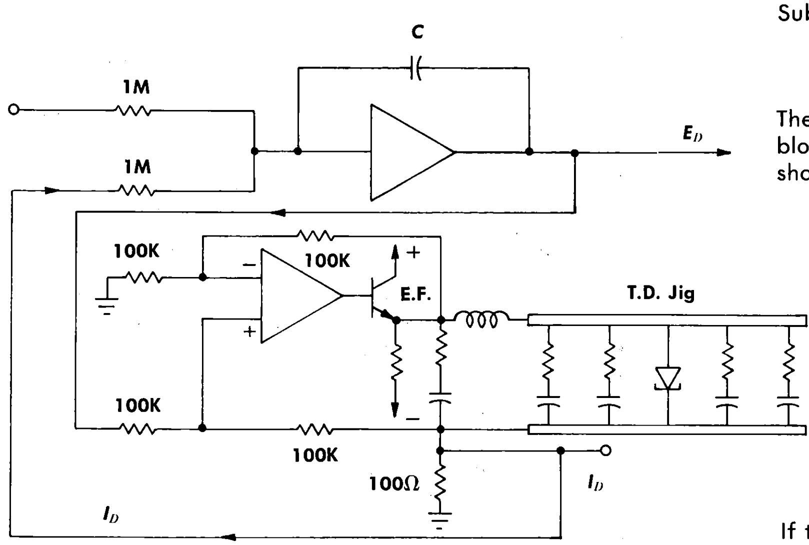
$$E_D = \frac{1}{C} \int I_c \, dt \qquad \dots\dots 5$$

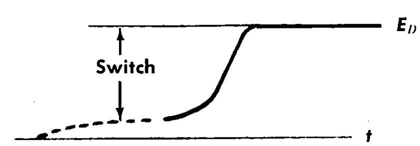
Substituting eq. (1a)

$$E_D = \frac{1}{C} \int (I - I_D) \, dt \dots\dots 5a$$

The equations (4) and (5a) are represented by the block diagram, and operational amplifier circuit is shown at left.

Analog Solution of Risetime*



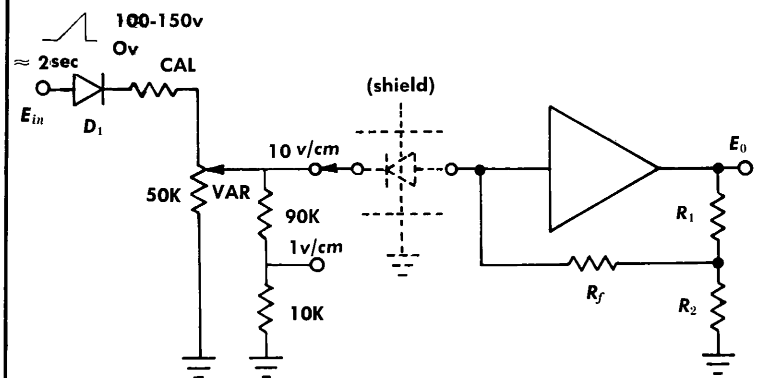


If the integrating capacitor C is chosen 0.01 μfd for diode capacitance, C_D, 1 pf; the scaling factor will be 1 nsec real time equal to 10 msec analog time.

*It is possible to compute diode capacitance from diode switching time.

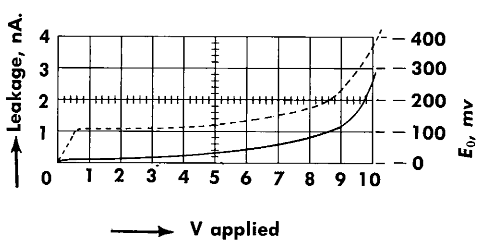
Voltage-Current Conversions	Notes

Semiconductor Leakage & Capacitance Measurement—Precalibrated Drive.



$$E_0 \approx -I_L R_f\left(\frac{R_1 + R_2}{R_2}\right).$$

For $R_f = 10M$, $R_1 = 90K$, $R_2 = 10K$
$E_0 = 100mv/nanoampere$



——— 200 msec/cm (Leakage)
– – – – 2 msec/cm (Capacitance + Leakage)

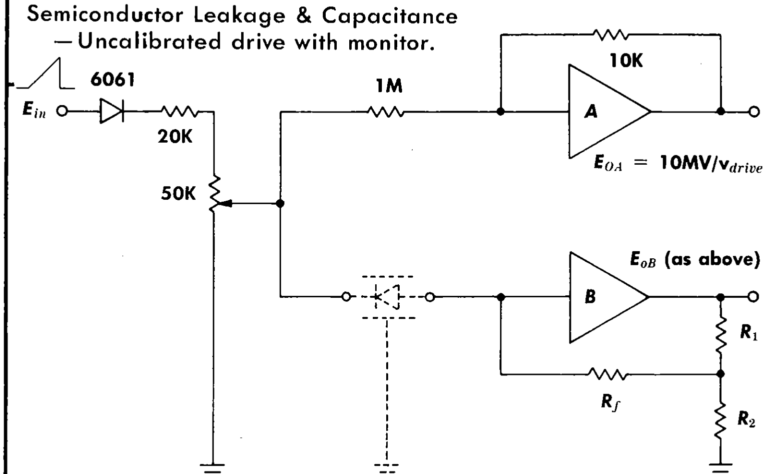
Vertical displacement of trace at faster sweep indicates junction capacitance

$$C = I\,dt/de = \frac{\text{Trace displacement} \times \text{Time/Cm}}{\text{Horizontal Volts/Cm}}$$

In example shown, $C = \frac{1nA \times 2\text{ msec}}{1v} = 2\text{ pf.}$

Indication will be in error (especially at start of trace) if sawtooth attenuator is not compensated or if sweep is operated at a high rep rate.

Semiconductor Leakage & Capacitance —Uncalibrated drive with monitor.

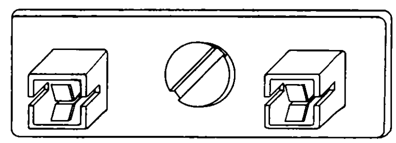


Notes

Oscilloscope sawtooth provides both horizontal deflection and calibrated drive to junction under test. Slow sweep (e.g., 200 msec/cm) provides vertical deflection proportional to leakage; vertical displacement of trace at faster sweep (e.g., 2 msec/cm) is proportional to capacitance.

Divider $R_1 - R_2$ permits Nanoampere sensitivity with reasonable values of R_f. Diode D, (use low-capacitance type) prevents slight negative quiescent level of sawtooth from forward-biasing junction under test.

Construction of test jig is quite critical. Leakage and capacitance paths must be to **ground,** not terminal-to terminal. Tektronix P/n 432-032 diode-holder may be used if mounted to ground plane with large binder-head screw, thus:



Both test jig and R_f should be shielded for high-sensitivity measurements.

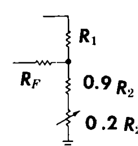
System errors:
Use of divider R_1R_2 to ground causes some input current to flow through R_f and divider to ground (negligible if $K < 20$). Actual error is

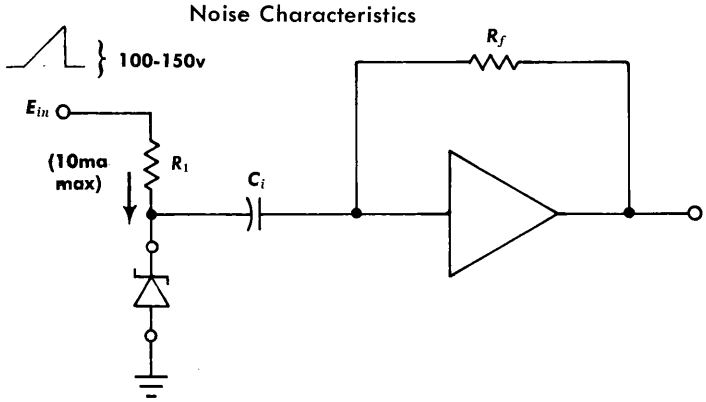
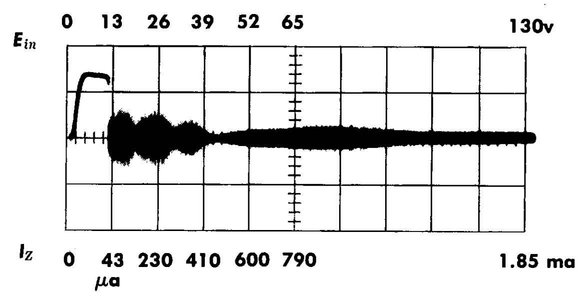
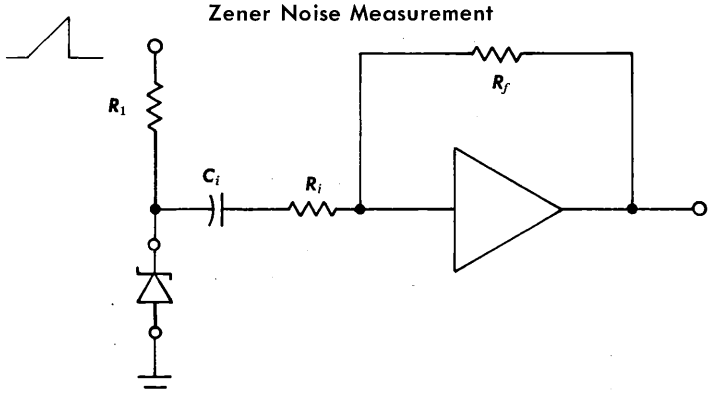
$$\frac{KR_f}{KR_f + \left(R_f + \frac{R_1}{K}\right)(1 - A)} \quad \text{where } K = \frac{R_2 + R_1}{R_2}.$$

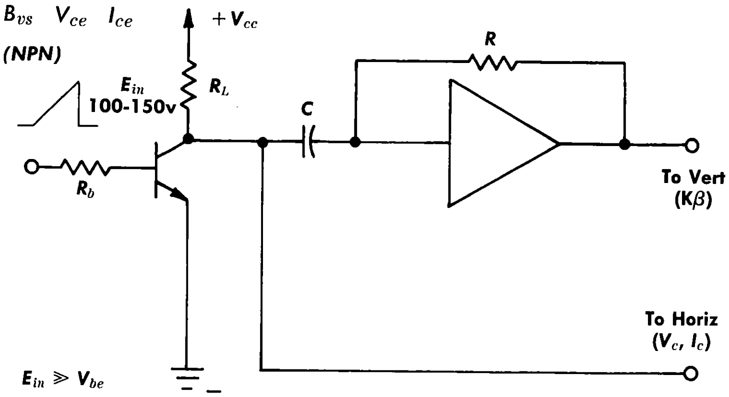
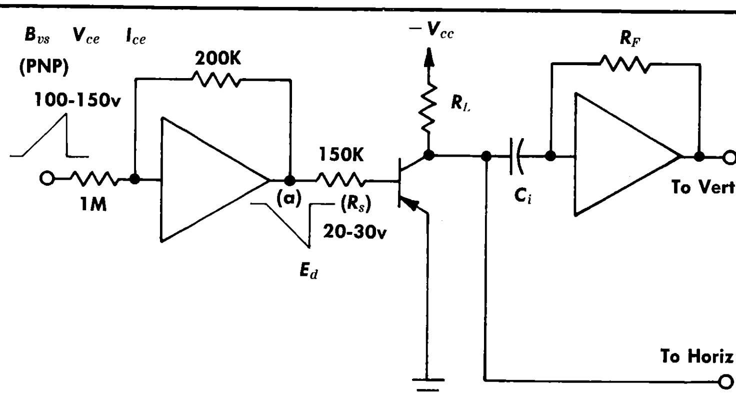
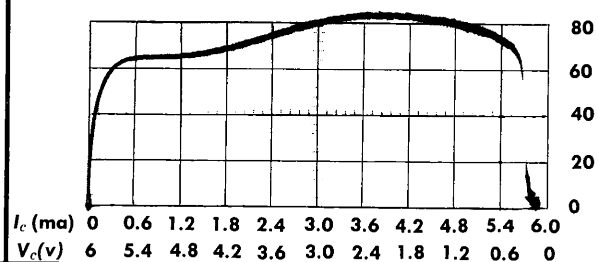
Error due to division of feedback voltage is $\frac{K-1}{A-1}$

(Open loop gain A is **negative**).

R_2 may be reduced in value and padded to correct accumulated system and tolerance errors:



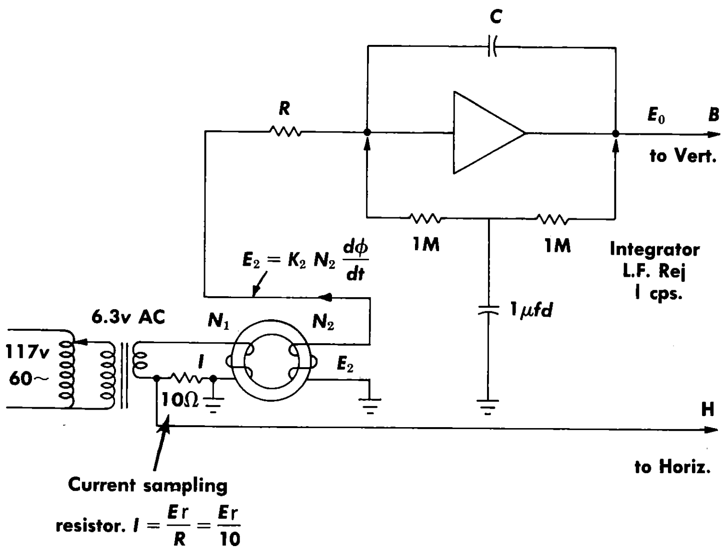
Zener Diode Tests	Notes
Noise Characteristics  	Variation of Zener noise with current may be evaluated by driving Zener with current sawtooth, differentiating output. Zener current $= \frac{(E_{in} - E_z)}{R_1}$ Example: Motorola 1/4 M10Z10: Vertical, 1V/cm; Horizontal, 13v/cm. R_1 = 70K, C_i = .001, R_f = 200K. Noise minimum at 500μa. (Not all zener diodes show these minima). 
Zener Noise Measurement  	Actual (p-p) amplitude of Zener noise is obtained by adding R_i. $E_o \approx \frac{R_f}{R_i}(E_{noise})$, for mid-band components.

Transistors	Notes
B_{vs} V_{ce} I_{ce} (NPN) 	Differentiation of collector waveform as base current is linearly increased provides vertical deflection proportional to β. Horizontal deflection represents all V_{ce}, I_{ce} points along R_L load-line. $$\beta = \frac{-E_o R_s\,(10T)}{R_f\,C_i\,R_L\,E_d}$$ Where E_d is the voltage of E_{in} at $10T$, and T is the sawtooth time/cm setting.
B_{vs} V_{ce} I_{ce} (PNP) 	For PNP, base drive must be inverted. Beta is calculated as above. Example: 2N970, V_{cc} 6v, R_L 1k, E_d, 30v, Time/Cm 0.1 msec/cm, R_f, 500K, C_i, .001μf, vertical sensitivity 2 v/cm. 

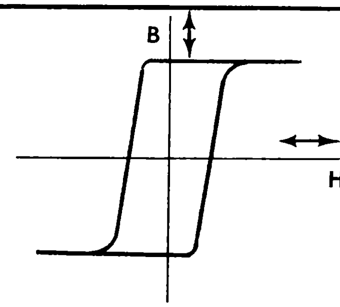
Notes

B-H Loop Plotting	Notes

B-H Curve Plotting



Notes



Secondary voltage E_2 is $E_2 = K_2\ N_2 \frac{d\phi}{dt}$.

If we integrate E_2, the output of the integrator E_o will be proportional to flux ϕ:

$$E_o = \frac{-1}{RC} \int K_2 N_2 \frac{d\phi}{dt}\, dt = \frac{-K_2 N_2 \phi}{RC}$$

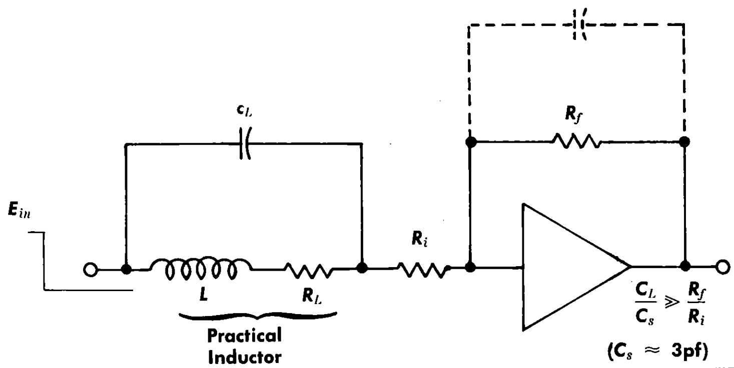
Since flux density $B = \phi/A$, where A is cross-sectional area of core, $B = \frac{-RC\, E_o}{K_2 N_2 A}$. Magnetizing force $H = \frac{K_1 N_1\ I}{\mathcal{L}}$ where $\mathcal{L}$ is the flux path length. With 10Ω current sampling resistor in primary,

$$I = \frac{E_r}{10} \text{ and } H = \frac{K_1 N_1 E_r}{10\mathcal{L}}$$

If area A and length $\mathcal{L}$ are given in square meters and meters respectively, $K_1 = K_2 = 1$.

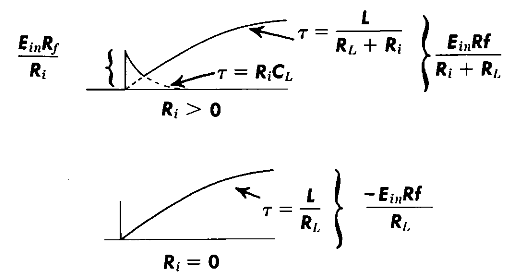
INDUCTOR MEASUREMENTS	Notes
Inductor Evaluation	Evaluation of step response indicates effective R_L and C_L components: τ is 0-63% or 100-37%.

Inductor Evaluation



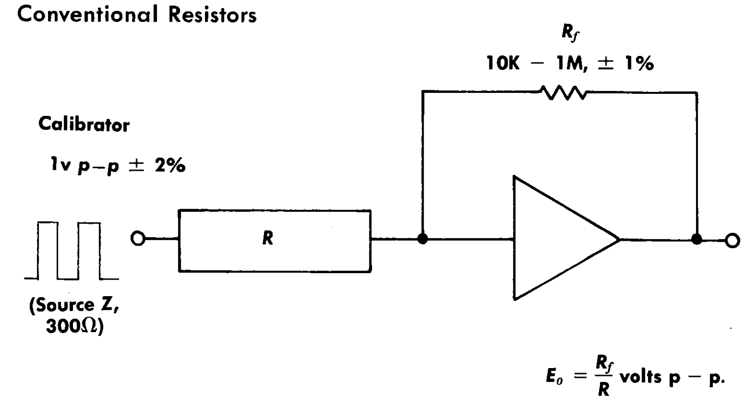
Notes

Evaluation of step response indicates effective R_L and C_L components:
τ is 0-63% or 100-37%.



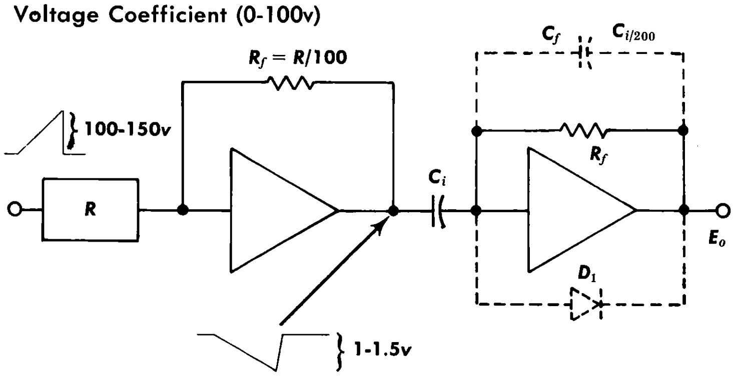
Resistance Measurements — Notes

Conventional Resistors



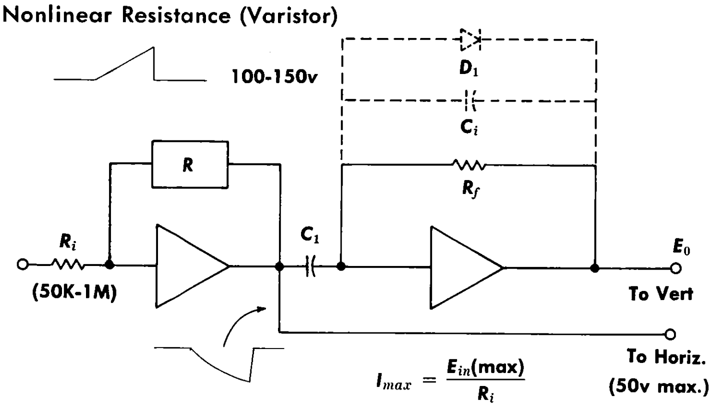
For output 0-10v, measurement accuracy is essentially the tolerance of the calibrator voltage and resistor used for R_f. Using oscilloscope calibrator and O-Unit internal R_f, accuracy is 3% or better for values 50K-10M. For values < 50K, drive unknown with X1 amplifier.

Voltage Coefficient (0-100v)

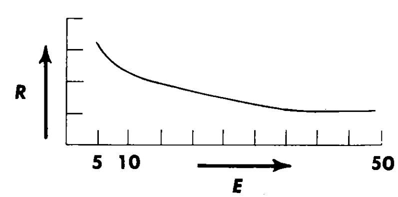


Resistor voltage coefficient may be determined by differentiating a ramp, the linearity of which is controlled by resistor under test. C_f may be used to limit noise; Diode D_1 prevents sawtooth flyback from causing overload.

Nonlinear Resistance (Varistor)

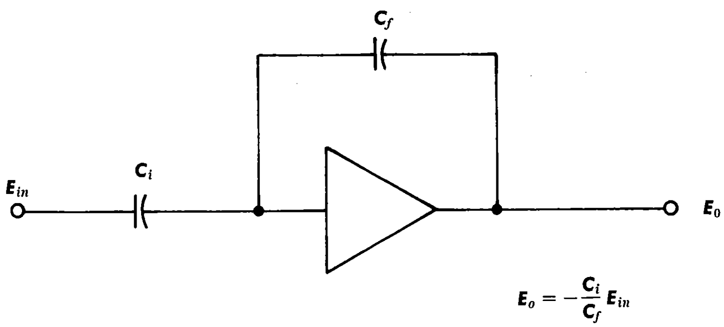


First operational amplifier applies linear ramp of current to varistor used as R_f. Voltage across varistor is used for horizontal deflection. Output of differentiator is proportional to R. (Use slow sawtooth) Diode D_1 prevents flyback overloads; capacitor C_1 ($\ll C_i$) may be used to reduce noise.



Capacitance Measurement (1pf–10μf)

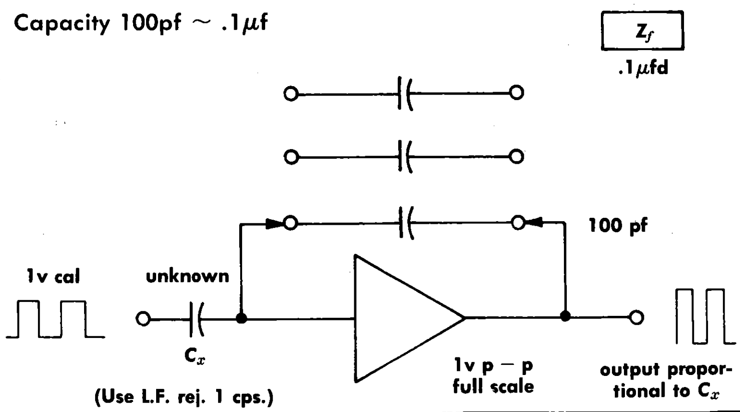
Basic idea



Notes

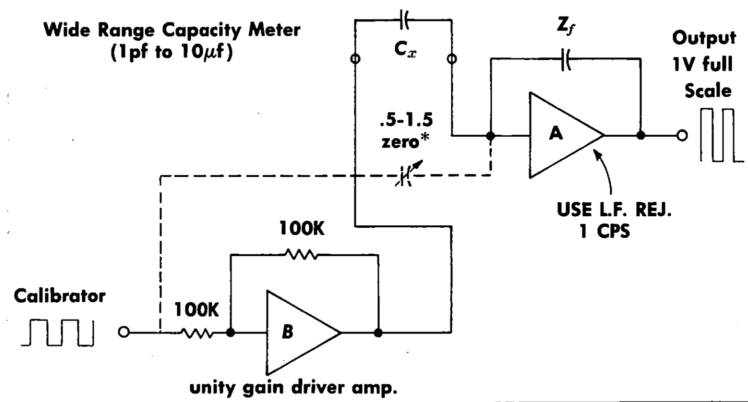
When capacitors are placed in input and feedback, it will make an amplifier. Gain of this amplifier is $-C_i/C_f$, ratio of input C to the feedback C. Since the output voltage of the amplifier is proportional to the C_i, it is possible to measure capacitance. Theoretically, this amplifier should have dc response but drift and leakage limit the low frequency response. For high frequency, charging current limits the response — use smaller C.

Capacity 100pf ~ .1μf



Unknown capacitor is compared to the precision internal feedback capacitor. When unknown C_x is same as feedback C_f the output voltage is 1 volt. The output voltage can be monitored by scope or meter. The calibrator impedance is too high to drive higher capacitance; therefore, use of isolation amplifier is recommended as shown below.

(use integrator L.F. rej. at 1 cps.)



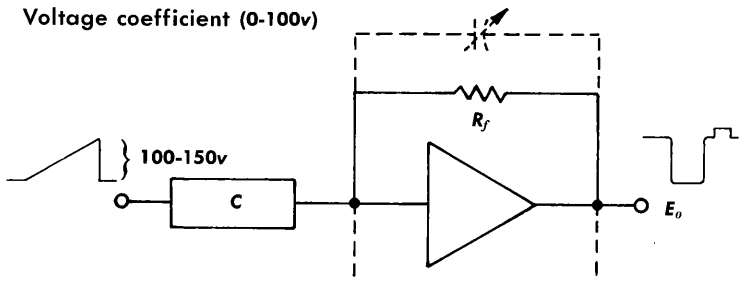
Careful arrangement enables capacitance measurement for values of 1pf to 10μf.

RANGE	Cal.	Z_f
0 - 10μf	10mv	.1
0 - 1μf	.1v	.1
0 - .1	.1v	.01
0 - .01	.1v	.001
0 - .001	1v	.001
0 - 100pf	10v	.001
0 - 10pf*	10v	.0001

*** to compensate stray capacity, adjust neutralizing C for zero C output indication.**

Dielectric Tests	Notes

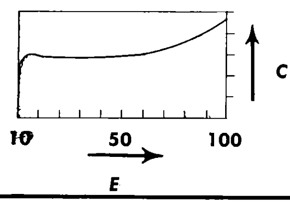
Voltage coefficient (0-100v)



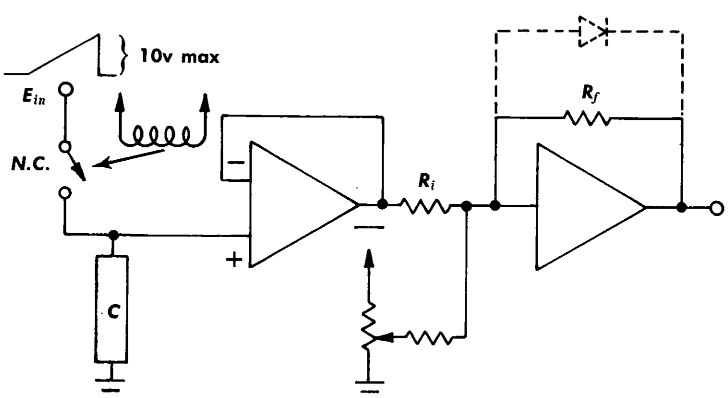
$$C = \frac{-E_o}{R_f(dE_{in}/dt)}$$

Voltage coefficient of capacitors from 1 pf to 250 μf may be evaluated by proper selection of sweep time/cm and R_f, using oscilloscope sawtooth. Horizontal calibration will depend on individual oscilloscope. Sweep repetition rate must be made very low (long hold-off) for large value capacitors.

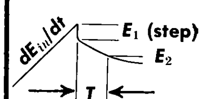
(Display Inverted)

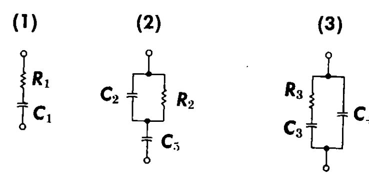


Internal Resistance (Series)



Reed relay is opened just before sawtooth reaches peak. Internal resistance in capacitor will cause negative step following relay opening. Slideback arrangement allows detailed analysis of step.





$E_{in} \gg (E_1 + E_2)$
Ramp duration $\gg \tau$
Measure τ to 0.37 E_2

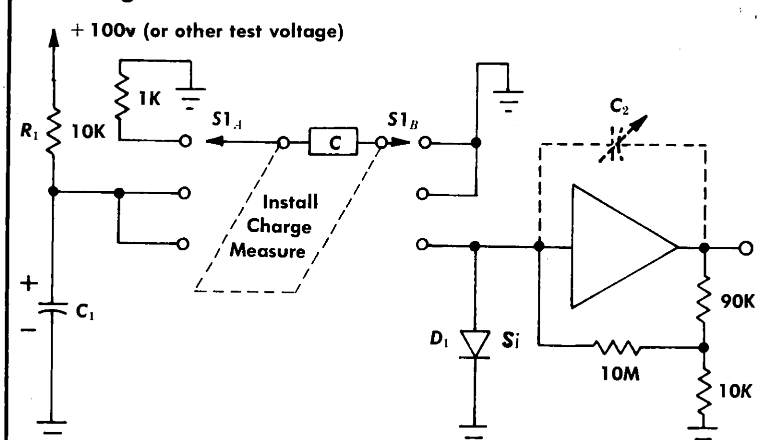
(1) $E_1 = \frac{dE_{in}}{dt} R_1C_1$, $\tau = 0$

(2) $E_2 = \frac{dE_{in}}{dt} R_2C_5$, $\tau = R_2C_2$

(3) $E_2 = \left(\frac{dE_{in}}{dt}\right)\left(\frac{RC_3^2}{C_4}\right)$, $\tau = R\left(\frac{C_3C_4}{C_3 + C_4}\right)$

If E_2 is due to condition 3 (Soak), capacitor will show buildup of open-circuit voltage after rapid discharge.

Leakage



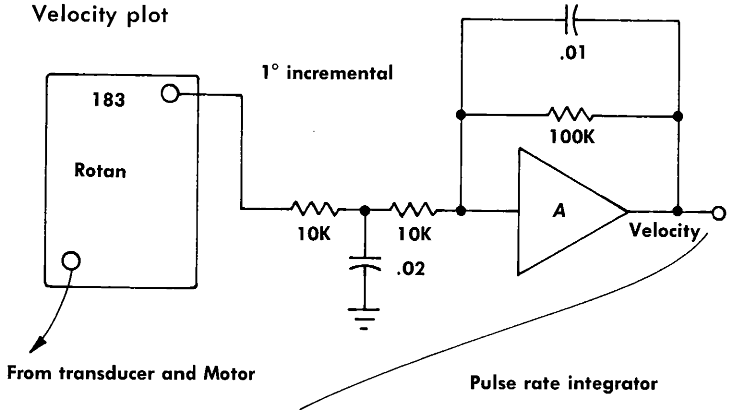
"Charge" and "measure" positions may be made spring-return for safety.

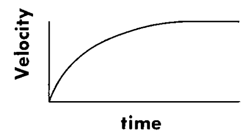
High-sensitivity measurements are made possible by divided feedback. For values shown, sensitivity is 1v output per 10 nA leakage. 50 mv resolution = 500pA, or $0.2 \times 10^{12}\Omega$ for 100v test voltage. Precautions must be taken to avoid excessive noise. Diode D_1 and Resistor R_1 protect against breakdown. C_2 reduces noise.

Notes

Velocity Acceleration Studies	Notes

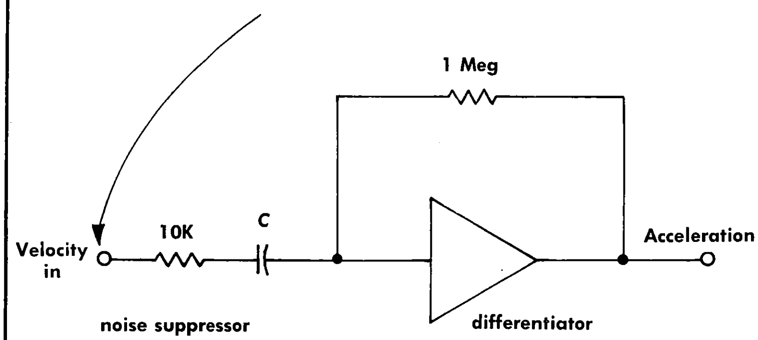
Velocity plot

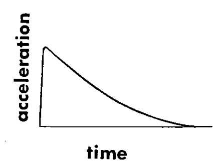




Rotational velocity may be obtained by integrating pulse rate, using the ROTAN 1° incremental pulse.

Acceleration plot

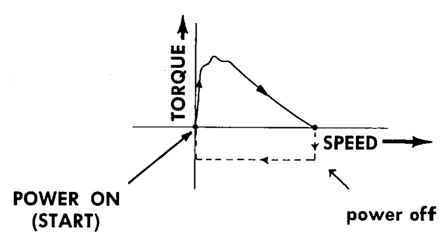




Velocity signal is differentiated to obtain acceleration. 10K resistor is added to suppress noise.

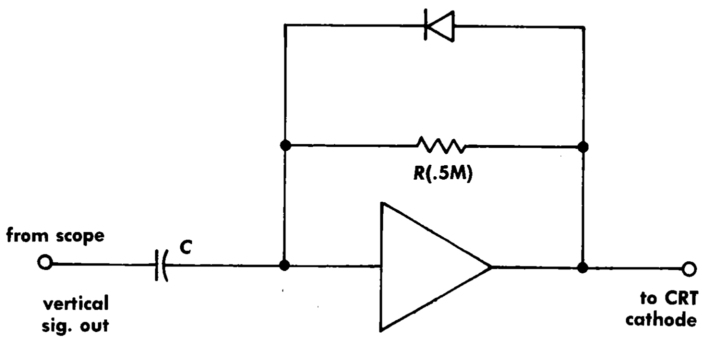
Speed—Torque Plot

X Y plot of velocity and acceleration gives a speed—torque plot. A typical motor characteristic may be obtained by starting and stopping the motor and observing the torque speed characteristics.



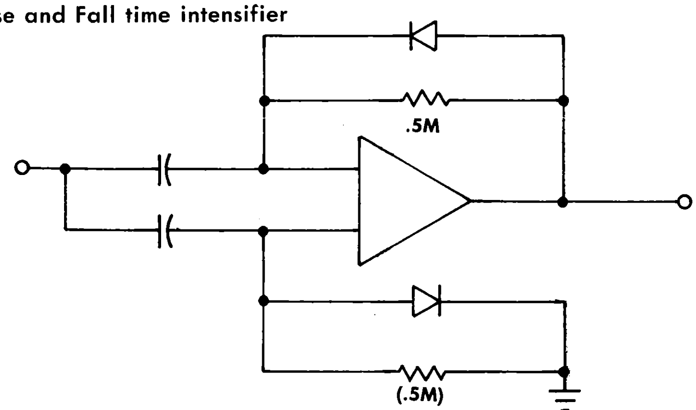
Risetime Beam Intensifier	Notes

Risetime intensifier

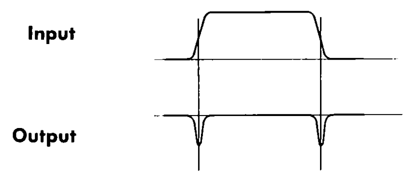


A differentiator can be used as a beam intensifier in certain experimental set ups. Scope vertical signal out can provide signal to the differentiator and the output is fed to the CRT cathode. A word of caution; vert. signal output should not be monitored by the same scope. Possible oscillation may result.

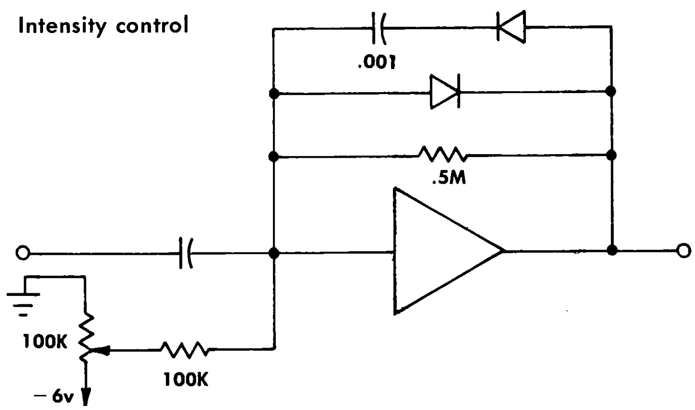
Rise and Fall time intensifier



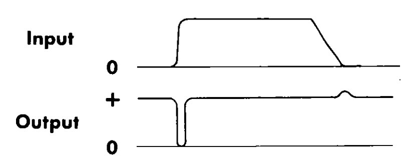
Applying signal to the both positive and negative grid, both rise and fall time can be intensified.



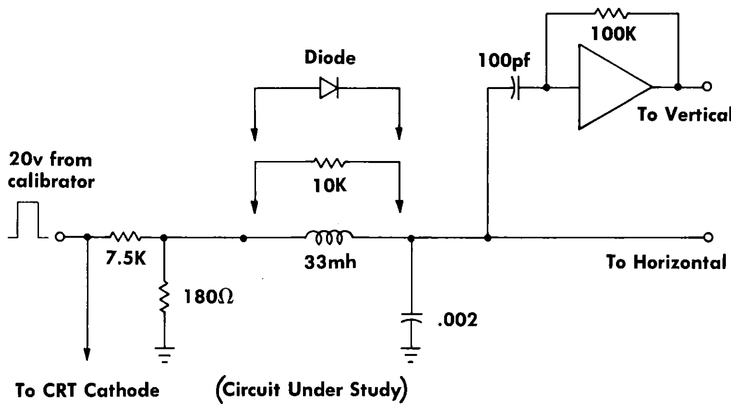
Intensity control



If the cathode is over-driven a serious defocusing may result. Thus some means of maximum intensity limiter is desired. In this circuit biased diodes limit the output swing to prevent defocusing.



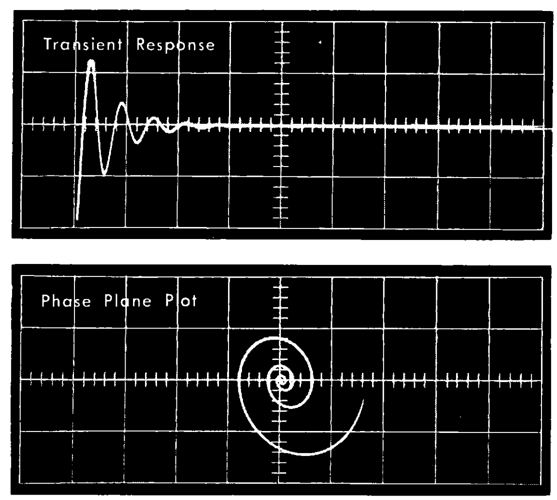
Phase—Plane Plot	Notes



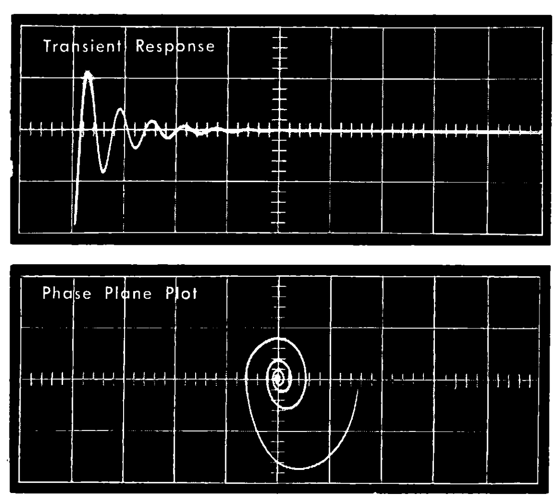
In studying system stability of control systems or feedback systems, transient response is often employed. If a non-linear element is involved in a system, stability criteria usually become quite complex, and analysis is normally limited to linear approximations.

A phase plane plot can overcome some of this limitation since it can present the stability, as well as the nature of non-linearity.

As an illustration, the effect of resistive and diode damping on an LCR circuit is compared in a transient response display and a phase plane plot. Circuit values are chosen such that the diode does not conduct very heavily and it is very difficult to distinguish non-linearity in the transient response display. However, when the original (transient response) signal is plotted against the differentiated signal, the non-linear effect is easily observed. (See photos below.)



10 k Damping

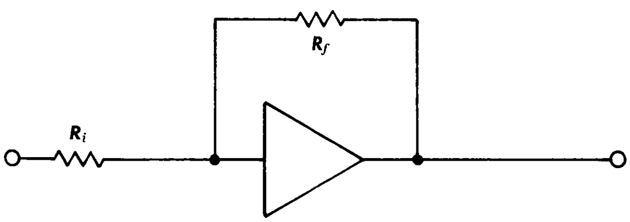


Diode Damping

Notes

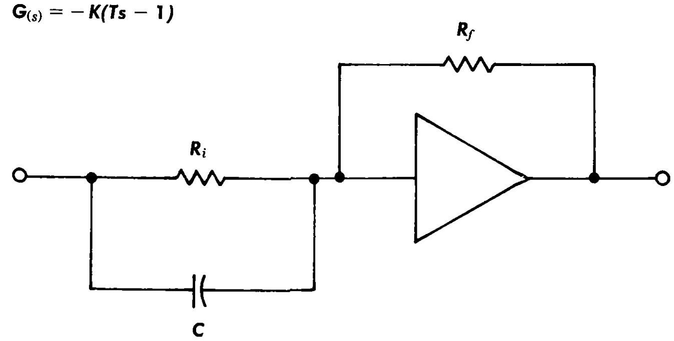
Transfer Functions, E_o/E_{in}.

$G_{(s)} = -k$



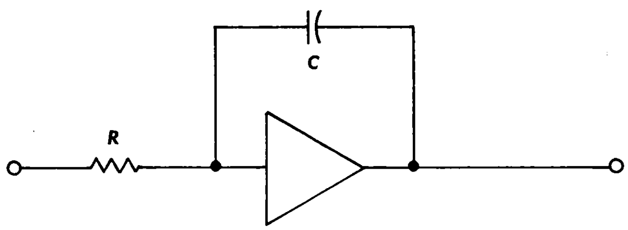
$k = \frac{R_f}{R_i}$

$G_{(s)} = -K(Ts - 1)$



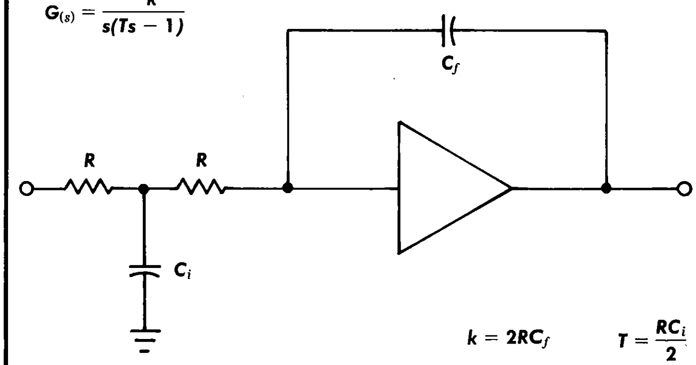
$k = \frac{R_f}{R_i}$ $T = R_iC$

$G_{(s)} = \frac{-k}{s}$



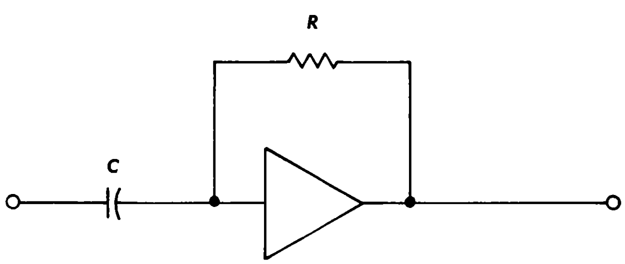
$k = \frac{1}{RC}$

$G_{(s)} = \frac{-k}{s(Ts - 1)}$



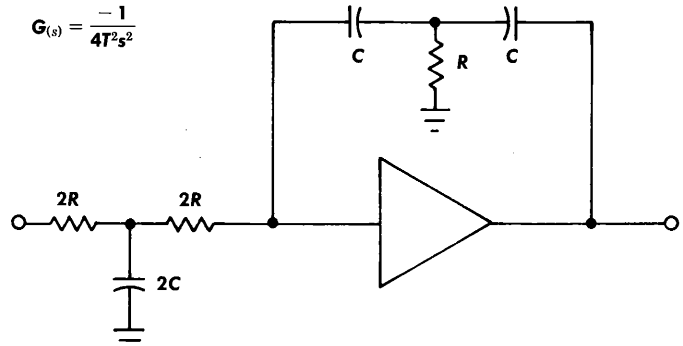
$k = 2RC_f$ $T = \frac{RC_i}{2}$

$G_{(s)} = -ks$



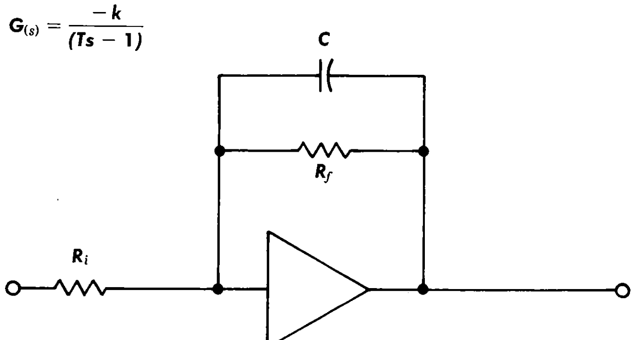
$k = RC$

$G_{(s)} = \frac{-1}{4T^2s^2}$



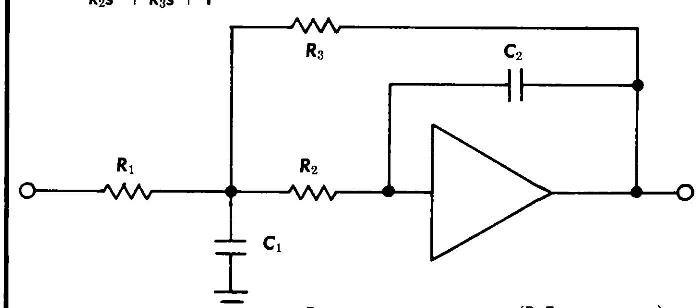
$T = RC$

$G_{(s)} = \frac{-k}{(Ts - 1)}$



$k = \frac{R_f}{R_i}$ $T = R_fC$

$G_{(s)} = \frac{-k_1}{k_2s^2 + k_3s + 1}$



$k_1 = \frac{R_3}{R_1}, k_2 = R_2R_3C_1C_2, k_3 = \left(\frac{R_2R_3}{R_1} + R_3 + R_2\right)C_2$

Notes

APPENDIX

Semiconductors:

Some of the semiconductors referred to in these notes are non-registered types selected for special characteristics and available only from the original manufacturers.

In a few cases (Types 6045, 6061, 6075), the diodes referred to are made to Tektronix specifications, and may be procured in small quantities from Tektronix if suitable equivalents are not locally available.

Germanium, General Purpose Signal Diodes:

	PIV	Fwd ma	Leakage	Procure
T12G (Transitron, Raytheon) HD2607 (Hughes) CGD-1075 (Clevite)	75v	20 ma	30 μa@10v	Locally
Type 6075	25v	40 ma	10 μa@ 6v	Tektronix (152-0075-00)
ED-2007 (Erie Resistor Corp.)	15v	5 ma	10 μa@ 6v	Locally
DR-746 (General Instruments) HD2948 (Hughes)	10-25v	40 ma	10 μa@ 6v	Locally
T13G (Transitron)	25v	50 ma	2 μa@10v	Locally

Silicon Signal Diodes (Low leakage or high voltage):

	PIV	Fwd ma	Leakage	Procure
HD5000 (Hughes)	10v	12 ma	0.2 μa@ 5v	Locally
6045	125v	40 ma	.002 μa@ 9v	Tektronix (152-0045-00)
6061 RD2594 (Rheem) ED2927 (Erie) CSD2800 (Clevite)	200v	100 ma	5 μa@175v	Tektronix (152-0061-00) or Locally
G-130 (Texas Instruments) "Log Diode", "Stabistor".				Locally

Zener Diodes:

Voltage	Current	Use	Tolerance	Mfr.	Procure
±6v*	25 ma	RT6*	±1v	Hoffman	Locally
6v	60 ma	1N753	5.6-6.8v	Texas Instrs.	Locally
7v	30 ma	1N707	6.2-8.0v	Hoffman	Locally
10v	25 ma	¼M10Z10	±10%	Motorola	Locally
15v	15 ma	1N718A	±5%	Pacific	Locally

*Double Zener

Transistors:

		Procure
2N1304, 2N1305	General Instrument, Raytheon, Sylvania, Texas Instrs.	Locally
2N706B, 2N970	(Motorola, Fairchild) may be substituted for {2N1304, 2N1305} in ÷ 2 Multivibrator circuit. For Higher speed operation.	Procure locally
2N1302, 2N1303	May also be used in ÷ 2 Multivibrator.	

Selected Diodes:

The wideband logarithmic amplifier requires careful selection and matching of diodes to obtain calibrated symmetrical output.

If unidirectional operation only is needed, only half the diodes are needed—selected on the basis of the criteria below. If symmetrical operation is needed, the selected diodes must be further matched in pairs.

Circuit Location, Type	Selection Criteria (DC) @ I_{fwd}	Select for Forward Drop	Matching Criteria: Match Forward Drop Within:
Input ED-2007 (Erie Resistor Corp.)	10 μa	0.10-0.13v	.004v
	10 ma	0.35-0.50v	.01 v
Feedback ED-2007	10 μa	0.10-0.13v	.004v
	1 ma	0.25-0.30v	.01 v
Feedback HD-5000 (Hughes)	10 μa	0.25-0.40v	.10 v
	100 μa	0.35-0.50v	.01 v

Typically, it takes about 10 ED-2007's to obtain two matched pairs and 5 HD-5000's to obtain one matched pair. *

Precision Resistors:

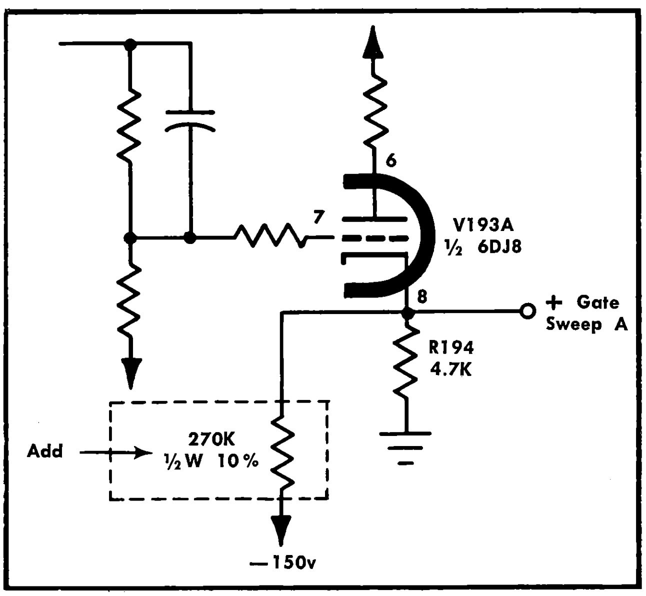
Precision carbon-film and metal film resistors in 1% and closer tolerance are available from most electronics suppliers at a cost of 50 cents to $1.50 each. Use of smallest sizes (⅛w, ¼w, etc.) compatible with power requirements is recommended for high-speed work. Use of large "meter" type wirewound precision resistors is not recommended for fast work because of excessive capacitance and (in some cases) inductance.

Oscilloscope Modifications:

A slight modification of your oscilloscope may facilitate use of the Type O-Unit for some applications. Extensive modification should not be undertaken without consulting your local field engineer, however.

+Gate Output Modification:

For gated-integrator operation, it may be desirable to have the +Gate output from your oscilloscope drop to a negative DC level of 2 or 3 volts between sweeps, to avoid the use of a battery or integrating network to obtain this negative voltage. It is possible to modify your instrument as shown in the following schematic.
To modify the output, add a resistor between the cathode of the +Gate output cathode follower and − 150v, the value of the resistor about 60 times the value of the cathode resistor already in the circuit.
Example:
Type 545A or 535A, +Gate, Sweep A:



*Yield figures are for diodes manufactured in 1963 and earlier. Yield figures for more recent production approach zero for the ED-2007 because of modified forward characteristics in this type. Instead of HD5000 diodes, 1N3605 diodes (G.E.) may be used with some modification of circuit values. There is no entirely adequate commercially available substitute for the early ED-2007, but by selecting resistor values and accepting some reduction in operational accuracy, adequate performance for many applications may be obtained from selected ED-2007's of current manufacture or from other similar germanium diodes such as the T13G.

DC Calibrator Output:

In most Tektronix oscilloscopes manufactured since 1960, the square-wave calibrator pentode section is a Type 6AU6, and so is individually removable (not so in earlier instruments using the triode-pentode Type 6U8). Removal of this tube causes the calibrator output to fly up to its precalibrated value and hold this DC level. Reference

DC levels from + 0.2 mv to + 100 v may thus be obtained from the front-panel calibrator output connector for use in operational amplifier applications. Current capabilities of this circuit are limited, however. Although up to 5 ma may safely be drawn from the + 100 v position, the high impedance divider may be damaged if the 50 v position is heavily loaded. Accurate calibrated output will be available only into high (meg ohm) impedances.

Negative reference voltages up to − 50 v may be obtained by processing the + DC voltage from the modified calibrator through a − 1 amplifier (one operational amplifier set up with R_i = 1 M, R_f = 1 M). Care should be exercised to avoid driving the amplifier with more than 50 v or overloading the output (5 ma max).

Where currents in excess of 5 ma are required, external supplies should be used, to avoid overloading the regulated DC supplies in the oscilloscope or damaging internal components.

Obtaining Time Scaling Factors From Sampling Oscilloscopes

In integrating or differentiating waveforms obtained from the vertical signal output of sampling oscilloscopes, it is essential to preserve a known and stable real-time: equivalent time ratio, or scaling factor. The operations performed by the operational amplifier are in real time; the information contained in the sampling oscilloscope output is in equivalent time.

Unless some means is employed to **force** a fixed real time: equivalent time ratio, the ratio will be dependent on the repetition rate of the signal being sampled, up to about 100 KC; above this point by the sampling-rate limiter in the sampling oscilloscope. An erratic pulse repetition or sampling rate may cause significant errors.

To force a fixed relationship between real and equivalent time, it is only necessary to drive the external horizontal input (external scan) of the sampling oscilloscope with the real-time sawtooth of the oscilloscope used with the operational amplifier. The real-time sawtooth should be properly attenuated so that it provides the **same display** on both the sampling and the real-time oscilloscopes when the sampling oscilloscope vertical signal output is displayed on the real-time oscilloscope. The real-time sweep should be made slow enough to provide about 100 samples/cm or more on the sampling oscilloscope, particularly during differentiation. The horizontal scaling factor under these circumstances becomes the ratio of time/cm switch settings of the sampling and real-time oscilloscopes.

It is not necessary to trigger the real-time oscilloscope sweep; it may be run at any convenient repetition rate. The retrace and hold-off of the real-time sweep will not be blanked on the sampling oscilloscope, so free-running the real-time sweep at a slow time/cm setting is generally most satisfactory.

Vertical scaling factors require no special techniques. However, since accumulated calibration errors and loading effects may be serious in some cases, it's usually advisable to obtain a 1:1 scaling factor as follows:

1. Apply several cm of calibrator signal to the sampling oscilloscope input.
2. Using the same value Z_i as will be used in the intended operation, set the operational amplifier for $\times(-1)$ amplification ($Z_f = Z_i$).
3. Adjust the real-time oscilloscope to display the same number of centimeters as the sampling oscilloscope Volts/Cm and Calibrator settings indicate **should** be displayed on the sampling oscilloscope. The vertical scaling factor is now 1:1. Z_f may now be reset to the correct value for the intended operation. Since DC levels are critical in integration, the sampling oscilloscope DC offset (and/or vertical position or DC balance controls, depending on the make and model) should be carefully adjusted so that the vertical signal output is at 0 v DC when the signal waveform is at the reference level. The accuracy of this setting may be confirmed during adjustment of the vertical scaling factor, when the operational amplifier is set for $\times(-1)$ operation.

Bibliography:

Much valuable information concerning theory, construction and applications of operational amplifiers will be found in the following:

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Notes
