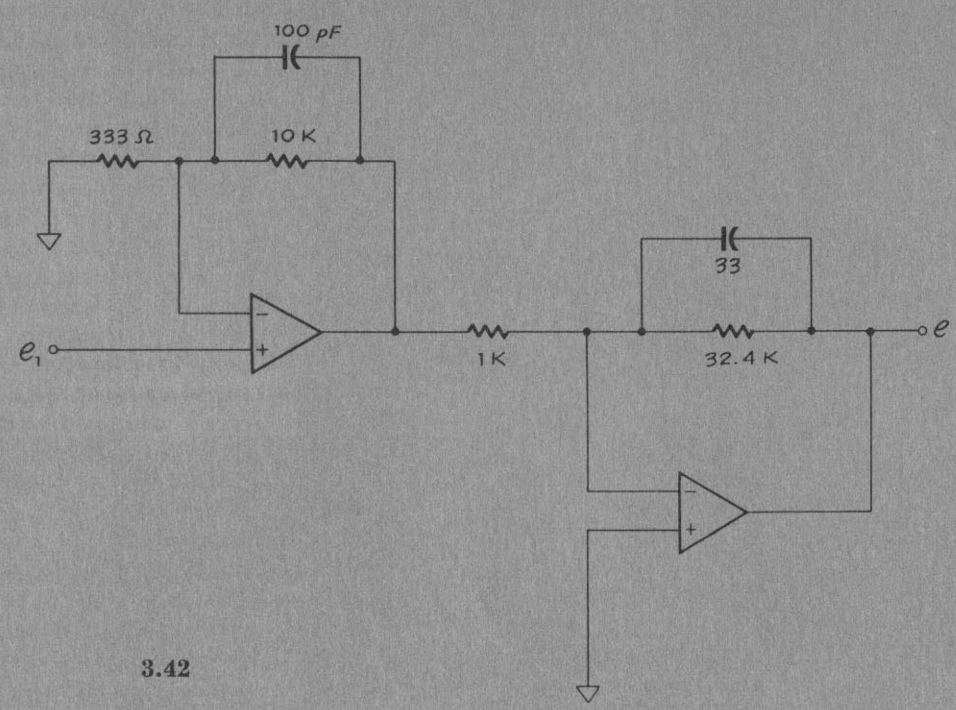
III.42 WIDE BAND, GAIN-OF-1000 AMPLIFIER. The wide bandwidth of this circuit is the result of cascading two amplifiers, producing a resultant gain-bandwidth which is about half the product of the two. It generally makes good sense to use the same amplifier type in both positions—however, if that is not done, for other reasons, the gain required of each stage should be directly proportional to the gain-bandwidth product of the amplifier associated with that stage, for optimum resultant bandwidth.

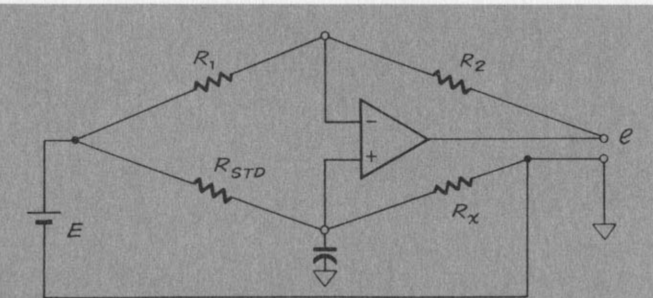
The circuit consists of a follower with gain (see II.2), appropriately rolled off for stability by means of a suitable feedback compensating capacitor, driving an inverting stage with gain, similarly compensated against instability. If the wiring is carefully done so as to minimize strays, the necessary roll-off capacitor should not significantly limit the bandwidth, particularly if the impedance levels are held to minimum, as shown. An outstanding advantage of this circuit is the fact that though low impedances are used in the feedback groups, the follower configuration used for the first stage provides very high input impedance, and the second stage is driven by the relatively low output impedance of the follower, and exhibits a reasonably low output impedance itself. Thus, with a total of only 6 external components, and 2 economical, standard amplifiers, we have constructed a precise, stable, low-noise, wideband (about 50 kHz) amplifier with a gain of 1000. Note that the amplifier response extends all the way down to DC, and that the noise and uncertainty levels, referred to the input, are of the order of microvolts, using the recommended amplifiers. This circuit will find wide use in instrumentation and computation circuits, and its utility can be further extended by the addition of a booster stage, when necessary.



3.42

III.42 I.7 I.8 I.9 I.10 I.11 I.12 I.13 I.14 I.15 I.16 I.17 I.18 II.1 II.2 II.3 II.4 III.35 III.36 III.38

III.43 WHEATSTONE BRIDGE—DEVIATION MEASUREMENT CIRCUIT. This application of the basic Adder-Subtractor circuit (II.5) has many advantages over a conventional Wheatstone bridge or deviation-bridge circuit. In this configuration, the bridge excitation voltage E is applied as a signal to both the Adder and the Subtractor networks, and the amplifier output voltage indicates the extent to which R_1/R_2 does not conform to R_{STD}/R_x. Note the following advantages: (1) The unknown, R_x, is grounded, which aids in guarding and shielding; (2) Provided that e does not saturate when R_x is removed or shorted, there is always some value of E within the linear region that will "balance" the bridge under those extreme conditions; (3) It is perfectly practical to drive very large current through R_x since only R_{STD} is involved in that path, but none of the instrument circuitry need carry those currents; (4) The read-out is grounded, which is a very considerable advantage if a sensitive or high impedance detector is to be used. (5) Unfortunately, the output e is linear only with R_2; hence a linear calibration can be achieved only if the unknown is placed in the R_2 instead of the R_x position, sacrificing some of the advantages previously mentioned. Note the use of a bypass capacitor across R_x, reducing the chance of stray coupling or pick-up.



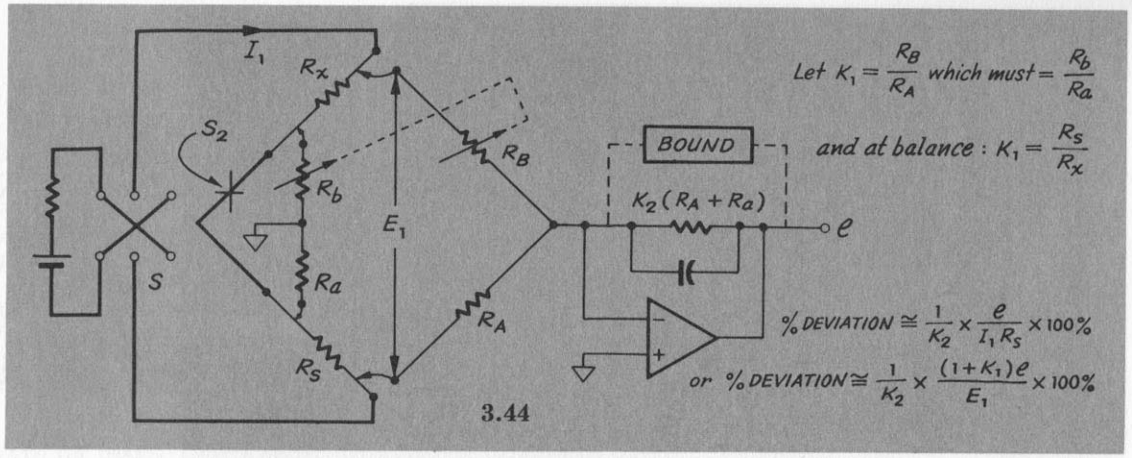
3.43

$$e = \left[\frac{R_x - \frac{R_2}{R_1} R_{STD}}{R_x + R_{STD}}\right] E$$

III.43 I.27 II.1 to II.4 II.41 III.44 to III.48 III.58 III.78 III.79

III.44 KELVIN-DOUBLE-BRIDGE DEVIATION INDICATOR (EXCITATION FLOATING).

A current-to-voltage transducer circuit (á là III.30) makes a very effective detector and linear deviation indicator for a Kelvin Double Bridge used for low-resistance (e.g., sub-ohm) measurements, provided that the DC bridge excitation source may float. The unknown four-terminal resistance, R_x, is measured by comparing it with a fixed or variable standard, R_S, of the same order of magnitude, with the switch closed. R_a and R_b compensate for lead and contact resistance, provided that they are precisely proportional to R_A and R_B. This proportion is proved if the bridge is balanced with switch S open.



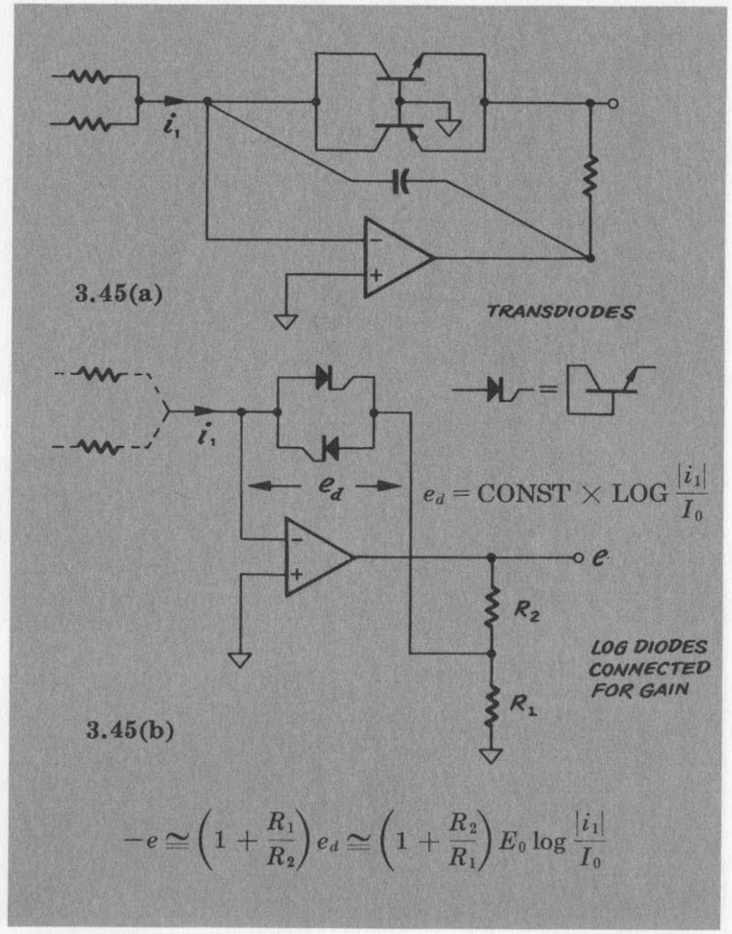
3.44

III.44
I.27
II.1
to
II.4
II.41
III.43
III.45
to
III.48
III.58
III.78

III.45 SIMPLE LOGARITHMIC-RESPONSE NULL-DEVIATION AMPLIFIER.

The current-to-voltage transducer of III.30 makes an excellent null detector, because of its high current sensitivity and low voltage drop, as noted earlier. In many applications, however, this high sensitivity may be a mixed blessing, because null detectors must often function in circuits that can and do operate "off-null" for long periods of time, causing saturation (or requiring bounding) and giving little or no indication, when driven to saturation or bounding, of just how far off null they are—thus rendering adjustment of the external circuit difficult, and "touchy," even near null. One cure for this situation is the modification of the response characteristic to a predictable *logarithmic* characteristic. The circuit shown here takes advantage of the inherently logarithmic voltage-current characteristic of transistors connected as diodes (see II.22, or the Technical Data Sheet on Model PL1 Logarithmic Transconductor) to achieve a response in which the output is proportional to the logarithm of the input current, as shown on the drawing. The output voltage is fed back through back-to-back transdiodes to the summing point. (It should be noted that the diode that is conducting, forms a bound circuit, such as that described in I.25.) The feedback circuit is so proportioned as to operate the diodes at currents in the range from less than 10^{-8} to 10^{-3} amperes, in which region their voltage/current characteristic most closely approaches a logarithmic relationship. For a given input current, then, the voltage drop across the diode will be $E_0 \log (i/I_0)$ (since the diode current must equal the input current in the ideal case). Thus the output voltage is also proportional to the logarithm of the input current. The divider shown in (b) but equally applicable to (a) produces closed-loop gain; in the degenerate case, in which $R_2 = 0$, the output voltage is simply equal to the forward voltage of the conducting diode. The response equation shown on the drawing is based upon the above assumptions, as well as the usual "ideal case"—negligible amplifier input current and null voltage at the summing point, infinite gain, constant temperature, etc.

Adding a resistor across the diodes modifies the effective diode characteristic at very low currents, limiting the maximum sensitivity, if desirable.



3.45(a)

3.45(b)

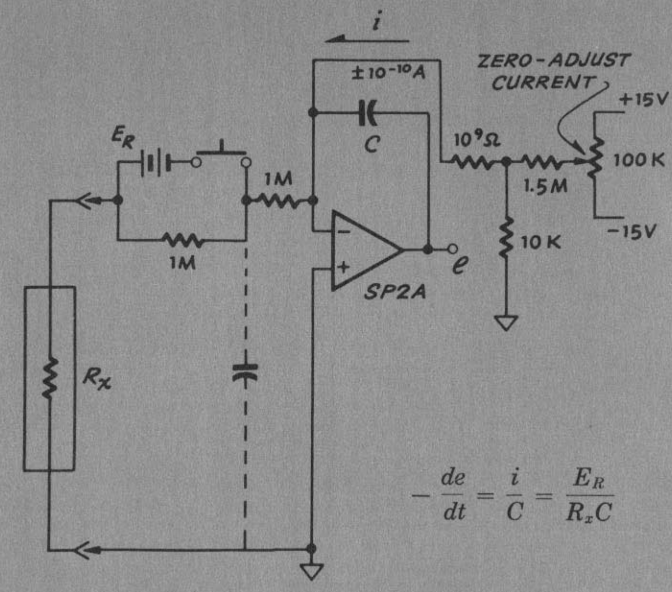
III.45
I.27
I.32
II.22
II.28
III.58
III.78

III.46 INSULATION RESISTANCE, TERAOHMS TO GROUND.

III.46 INSULATION RESISTANCE, TERAOHMS TO GROUND. This application of an integrator has far-ranging implications to the instrument circuit designer, being but one of many examples of the use of charging time to measure small currents. In this application, R_x is an extremely high resistance—of the order of teraohms (1 teraohm = 10^{12} Ω), measured between some test point and ground. After taking the usual precautions to prevent anomalous results—shielding, guarding and cleaning the resistance path in question—the circuit is connected to R_x, as shown, and the zero adjustment potentiometer is manipulated until there is no significant observable change in the output voltage with time. This condition reassures us that the net current into the integrator, including all of the usual pestilences—amplifier offset current, capacitor leakage, stray leakage, currents generated by thermal EMF's, etc.—is zero.

If we now close the push-button switch in series with E_R, an input current will flow that is almost exactly equal to E_R/R_x. (One may ignore the piddling megohm in series with the integrator input, compared to teraohms, of course.) This current will cause the output voltage to change at a rate determined by R_x and C, for a given value of E_R. Typical values are shown in the chart appended to the drawing. For nominal accuracies, a stop watch and a conventional meter are adequate, since the test may be allowed to run for a time corresponding to 10 volts before the amplifier saturates. The circuit is capable of much higher accuracy, however, and the inventive reader will no doubt be tempted to automate the procedure, using a relay for the pushbutton, a comparator to detect the exact moment at which the output arrives at a critical value, and an accurate time-interval meter to record the elapsed time.

Note the suggested use of a modest bypass capacitor across the integrator input, to reduce the effects of pick-up. Note also the use of the Tee network into the zeroing circuit to reduce the value of the series resistor to a practical 10^9 ohms. This will tend to increase the noise gain for voltage errors at the summing point. However, as long as E_R is of the order of volts (compared to sub-millivolts), the gigaohm shunt load at the amplifier input will not be significant. Remember that the accuracy of the measurement is directly affected by the absolute accuracy to which E_R is known and maintained, and by the amplifier's current offset drift.



$$-\frac{de}{dt} = \frac{i}{C} = \frac{E_R}{R_x C}$$

3.46

R_x	E_R	C	$\frac{de}{dt}$
10^{13}	100	100 pF	1V/10 sec
10^{14}	100	100 pF	1V/100 sec
10^{12}	10	100 pF	1V/10 sec
10^{11}	10	1000 pF	1V/10 sec

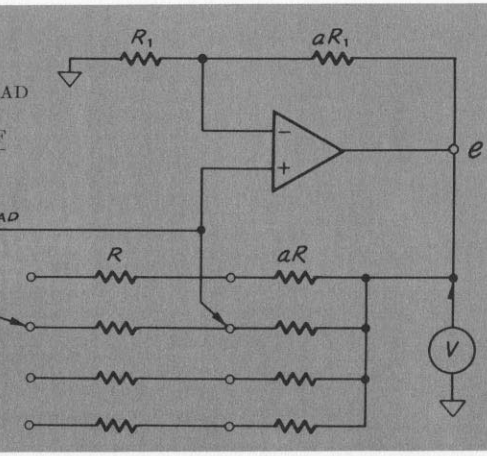
III.46
I.7
I.8
I.9
I.10
to
I.15
I.17
I.18
I.34
I.36
II.10
II.11
II.12
II.49
III.43
III.44
III.45
III.47
III.48

III.47 RESISTANCE MEASUREMENT, GROUNDED SAMPLE.

III.47 RESISTANCE MEASUREMENT, GROUNDED SAMPLE. For precise resistance measurements at relatively low voltage, over a very wide range of resistances, this interesting application of the Howland circuit is highly recommended. Note that range switching is accomplished by selecting resistance-ratio *sets*, the accuracy of which is always easier to establish and maintain than that of individual "standard" resistors. The same is true for the ratio set R_1/aR_1. Even wider ranges may be obtained by simultaneously switching the reference voltage . . . in decade steps, perhaps. Note that the ratio of the voltage across the sample to the output voltage is the divider ratio, a. If a is set to 100, for example, 10 millivolts across the sample will correspond to a 1 volt output, thus avoiding consideration of self-heating in all but the lowest resistance ranges.

$$e = (1 + a)e_{LOAD}$$

$$\frac{e_{LOAD}}{R_x} = \frac{E_{REF}}{R}$$

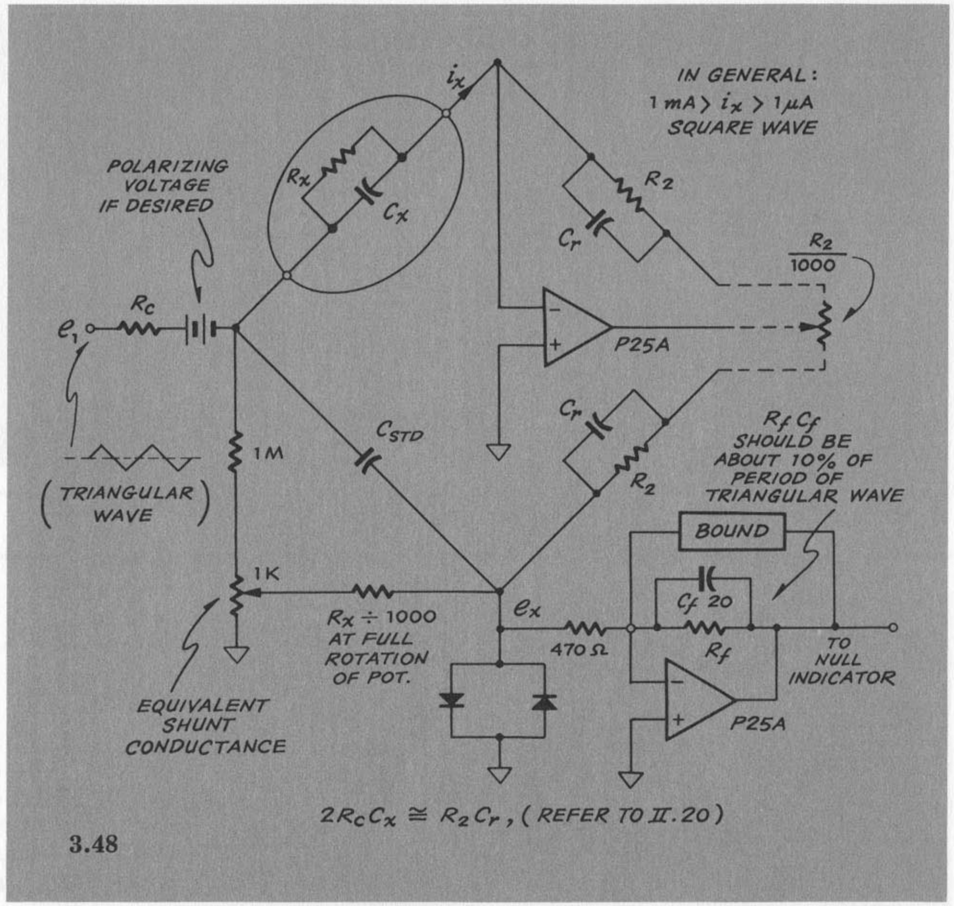


3.47

III.47
I.7
to
I.18
II.6
III.43
III.44
III.45
III.46
III.58
III.78

III.48 PRECISE CAPACITANCE-DEVIATION BRIDGE. There is no shortage of extremely precise capacitor bridges available to-day, but typically they use small voltage signals at relatively high frequency. This intentionally excludes such effects as voltage coefficient and polarization. Furthermore, the effects of shunt (leakage) resistance, series resistance; dielectric and other losses are lumped together in most bridges as the "dissipation factor," while the series inductance tends to lower the indicated capacitance somewhat. For measuring capacitors to be used in analog integrating circuits, there is no substitute for operating them under conditions similar to the anticipated use, e.g. driving ("integrating") them slowly over the voltage range (as shown here) with a constant current of perhaps 1 to 1000 microamperes, again selected to simulate the intended service.

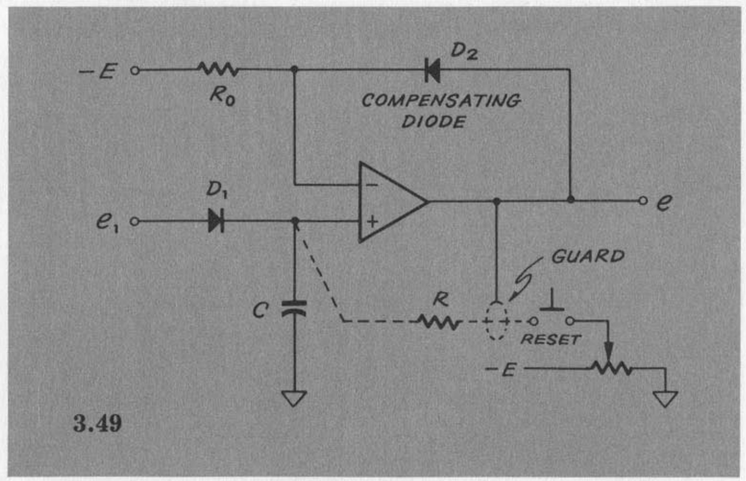
The excitation, e_1, applied to this bridge should be a triangular wave for maximum resolution (although a sine wave can serve well). A triangular wave provides approximately constant current, first positive, then negative. Thus the output of the amplifier is a fair square wave of voltage, and the null error voltage will be a small square wave (of sorts) for a small mismatch between the unknown capacitance and the standard. Tilt in the square wave implies a difference in shunt leakage, whereas a "bow" implies a difference in voltage coefficient between the unknown and the standard. Series resistance or inductance of the amounts found in practical capacitors for precision integrators will not affect the balance; neither, in general, will the shunt (leakage) conductance. The potentiometer, $R_2/1000$, can be used when a readout of the percentage deviation is desired, the extremes of the potentiometer representing about $\pm 0.1\%$ difference between C_x and C_{std}.



3.48

III.48
I.7
to
I.18
I.33
II.10
II.11
II.12
II.41
III.43
to
III.46
III.58
III.78

III.49 SIMPLE PEAK-READER AND MEMORY. This circuit may be set to any of a wide range of initial conditions (i.e., the stored charge on C), by adjusting the potentiometer. After reset, signals larger than the stored value will charge the capacitor, as rapidly as the source impedance permits. The capacitor will "remember" the *most positive* (or least negative) signal. The follower unloads the capacitor, provides unity gain, and low output impedance. If the diode is reversed, the circuit will remember the most negative signal applied to the input. Departures from the Ideal include: (1) reverse leakage in D1, whenever the signal is at a lower value than the peak value stored; (2) leakage (inherent in and stray) across the capacitor itself, as well as dielectric soakage; (3) amplifier input admittance and leakage current; (4) amplifier common-mode error; and (5) the forward drop (about 0.2 V) and effective source resistance of D1—the first causing a gradually-decreasing magnitude error; the second slowing down the rate of decrease, primarily because of the increase in charging-time-constant, at low diode-drop. R limits the discharge current. Diode D2 provides first-order compensation for the average drop across D1. Choose R_0 and E for zero net offset error for the most probable amplitude and duration of peaks.

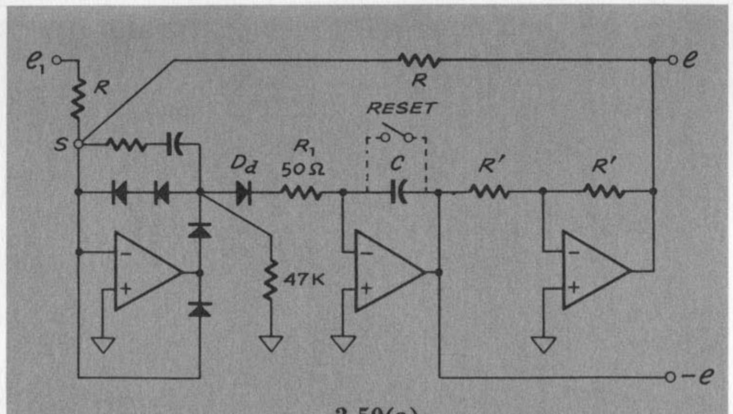


3.49

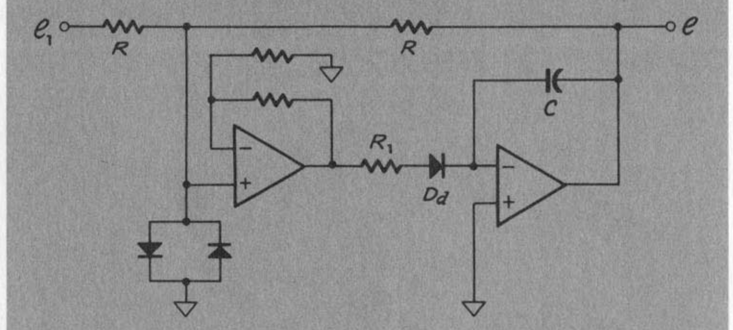
III.49
I.32
I.33
I.34
II.1
II.2
II.3
II.46
II.47
II.48
III.50
to
III.53
III.74
III.75
III.76

III.50 PRECISION PEAK-READER & MEMORY. This circuit can be broken down into a combination of three familiar elements: a precision electronic switch, of the type described in II.40, a track-and-hold memory (fast-charge integrator) of the type discussed in II.46, and the very familiar unity-gain inverter. The only new element introduced by this circuit configuration is the fact that all three of the circuits we have just enumerated are connected in a closed-loop configuration, with equal resistors (R) to form a unity-gain inverter with peak reading and peak holding characteristics. As the circuit is shown, it follows only negative peaks, but, by reversing all of the diodes in the switch, including D_d, the decoupling diode, it can be made to follow positive peaks. The circuit performs as follows: negative-going value of e_1 cause positive current rapidly to charge the integrating capacitor C (provided that R_1C is short compared to the fastest rate of change of e_1) until the positive output of the inverter following the integrator corresponds to the largest negative peak of e_1 (since the re-set switch was last operated), at which time the error voltage at summing point S goes to zero, and diode D_d is rendered non-conducting. The unity-gain inverter acts as a buffer on the integrator output, and also provides the necessary polarity reversal, so that the *major* unity-gain inverter (from e_1 to e, through R and R) will function correctly. When e_1 departs the peak value and becomes more positive—in circuit (a)—the precision switch remains open, and the integrator is prevented from discharging back through the switch by D_d. (The usual problems demand attention—leakage in C, input current in the integrator amplifier, closed-loop dynamic stability, and stray leakage.) Leakage in the decoupling diode is unimportant because the amplifier driving it will be observed to be the "perfect" halfwave rectifier whose output is zero for positive input. Assuming correct circuit design and most particularly, correct amplifier selection, this circuit will hold its peak for a period limited only by the capacitor's leakage.

A two-amplifier alternative is shown in (b).



3.50(a)



3.50(b)

III.50
I.32
I.33
I.34
II.1
II.2
II.3
II.36
II.37
II.38
II.39
II.40
II.46
II.47
III.49
III.51
III.52
III.53
III.74

III.51 SIMPLE PEAK-TO-PEAK READER. Positive values of e_1 charge C_1 through D_1, and are remembered. Negative voltage peaks charge C_2 through D_2, and are held there. Follower B input is then at the highest peak since C_1 was reset, and follower A output is at the deepest valley since C_2 was reset. The subtractor output is:

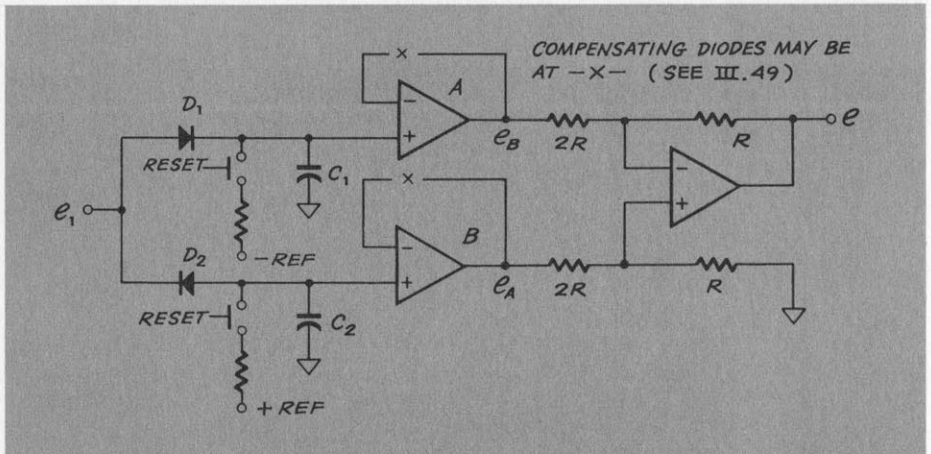
$$e = \tfrac{1}{2}(e_A - e_B) = -\tfrac{1}{2}(e_1 \text{ peak-to-peak}) \qquad (3\text{-}21)$$

Reversing the input connections of the subtractor will yield

$$e = \tfrac{1}{2}(e_B - e_A) = +\tfrac{1}{2}(e_1 \text{ peak-to-peak}) \qquad (3\text{-}22)$$

Diode and leakage errors prevent our calling this a precision device, but it has adequate accuracy for many purposes.

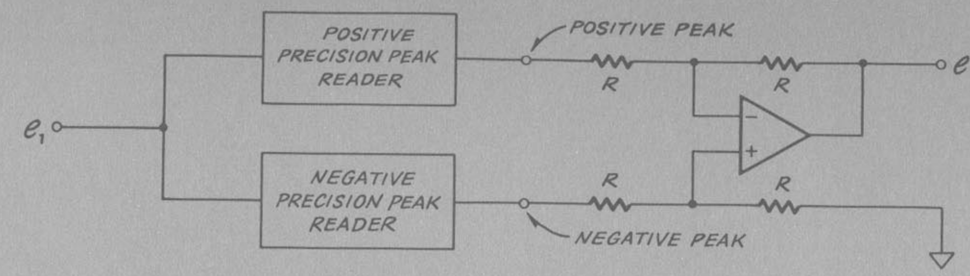
By closing the reset switches, e_B can be restored to the least value of e_1, and e_A to the maximum value of e_1, after which e_A and e_B may be driven by e_1 to the fullest range of values within the circuit ratings.



3.51

III.51
I.32
I.33
I.34
II.1
II.2
II.3
II.36
to
II.40
II.46
II.47
III.49
III.50
III.52
III.53
III.74

III.52 PRECISION PEAK-TO-PEAK READER-MEMORY. By combining two of the Precision Peak-Reader circuits described in III.50 and designing them with complementary diode polarities, so that one responds to and holds the highest positive peak, and the other the deepest "valley," and by combining these signals in a symmetrical Adder-Subtractor of unity-gain, we obtain an output equal to minus the maximum peak to peak amplitude, without reference to when the peaks occurred. Note that it is not absolutely necessary to use the Adder-Subtractor circuit—a voltmeter connected between the output of the positive peak reader and the negative peak reader will actually read the peak-to-peak value, as the difference between those two potentials. The voltmeter may not be grounded, however, and should be a true differential device.



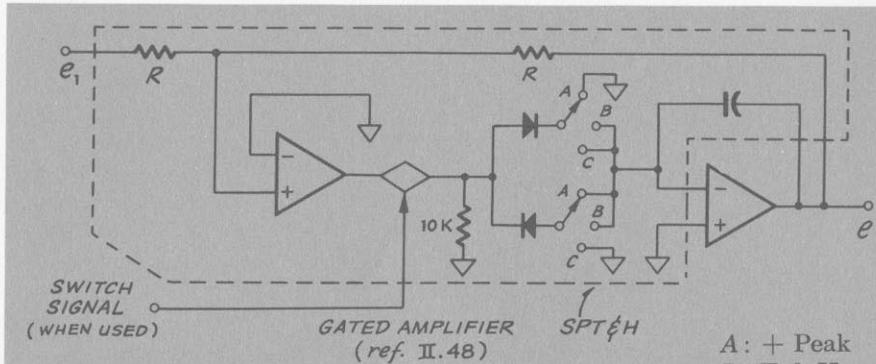
3.52

$$-e = E_{peak} - E_{valley}$$

III.52
I.32
I.33
I.34
II.1
II.2
II.3
II.36
to
II.40
II.46
to
III.53
III.74

III.53 TRACK—HOLD—PEAK READER. When the selector switch is in the *B* position, the circuit functions in exactly the same manner as the gated track-hold circuit of II.48. When the selector switch is in the *A* position, the integrator is not permitted to track positive-going input signals, because they are automatically shunted to ground through one of the diodes. Only negative-going signals are tracked, and then only the most negative signal is held, for the gated amplifier will receive a negative signal at its input (the resistive summing point) only when the input is more negative than the output is positive.

Similarly, when the switch is in the *C* position, positive-going peak input signals are tracked, and negative-going input signals are ignored. Note that there is a net inversion; the circuit really tracks and holds the *negative* of its input.

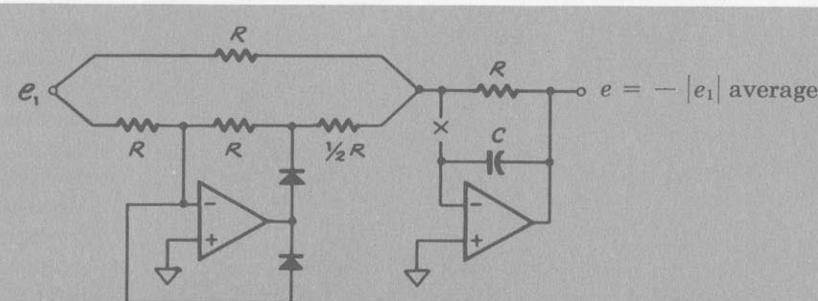


A: + Peak
B: T & H
C: − Peak

3.53

III.53
I.32
I.33
I.34
II.1
II.2
II.3
II.36
to
II.40
II.46
III.52
III.74

III.54 PRECISE AC AVERAGE READER. Refer back to II.42, Circuit (b)—the Precise Absolute Value Circuit. This circuit is identical to it, except that the capacitor *C* "averages," or smooths, the output voltage into a DC value almost exactly equal to the rectified average value of e_1. There are no appreciable diode errors of any kind in this "perfect" rectification. Furthermore, if the circuit is opened at point "x," the output will remain at whatever value it had at the instant of opening . . . at least, until the capacitor has discharged appreciably. A word of caution—unless the averager has a sufficiently long time-constant, the value that it "remembers" will depend upon the particular instant in the cycle of e_1 at which the circuit is opened. On the other hand, if the circuit has a long enough time-constant to eliminate ripple, its response to a change in the "envelope" magnitude of e_1 will be sluggish, and will require many cycles of e_1 to settle to a new value. This means that the "true average" exists only when the waveform of e_1 has been at equilibrium for a very long time . . . which may never happen.



3.54

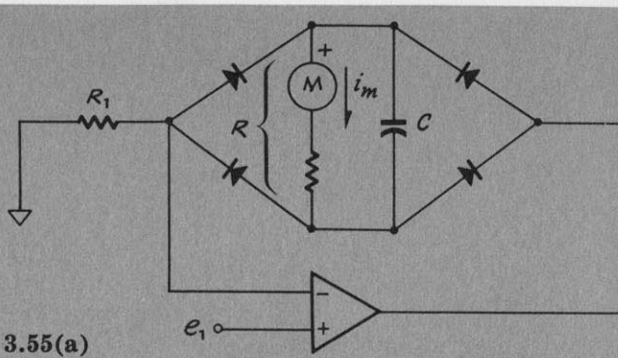
Averaging Time Constant = $T = RC$

III.54
II.10
II.11
II.12
II.41
III.55
III.63
III.65
III.81

pages 88 & 89

III.55 PRECISE AC MILLIVOLTMETER AND "ZERO-DROP" MILLIAMMETER. In circuit (a), a follower-with-gain is provided with a rectifier bridge as its feedback element. The amplifier forces the voltage across R_1 to track e_1 perfectly . . . ignoring the usual Departures From The Ideal. In order to do this, it must drive a current equal to e_1/R_1 through the bridge. By the courtesy of the rectifier diodes, this alternating current is rectified into DC, averaged by the capacitor C in accordance with the time-constant given on the drawing and displayed on the DC meter M. The rectification is precise (perfect) because the output voltage of the amplifier accommodates any and all of the rectifier diode idiosyncrasies in its determination to track e_1. This circuit, when equipped with a reasonably sensitive meter, may have a remarkably high sensitivity, and yet present nearly infinite impedance to the signal, compared to a conventional millivoltmeter.

Circuit (b) approaches the ideal of a zero-energy AC milliammeter, because the drop across its terminals is virtually zero . . . except for the usual Departures. The input current must flow directly into the summing junction, and, provided that neither the amplifier nor the diode clamps draw any significant current, the current through the bridge that forms the feedback network must have the same value. The return path for this current is through the amplifier output terminal, and back to the source, through the amplifier "common." Provided that the amplifier output current rating is not exceeded, and that the output voltage swing required of the amplifier, as it forces the bridge current to track the input current, does not exceed its rating, the meter will read the true and precise rectified value of I_{in}, "averaged" by the inertia of the meter movement. Since this circuit has negligible drop, it is practical to convert it to a precise *low*-impedance millivoltmeter, by connecting in series with either input terminal, a resistor of appropriate size. Even at 10 mV full-scale, the null voltage error should be negligible compared to the error in the meter movement. For a high impedance millivoltmeter, however, circuit (a) is preferable.

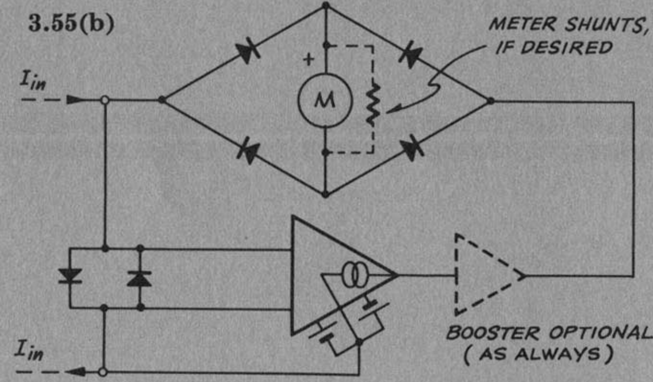


3.55(a)

$$\text{Meter Current } i_m = \left|\frac{e_1}{R_1}\right| \frac{1}{(1 + RCp)}$$

Time Constant $= RC$

3.55(b)

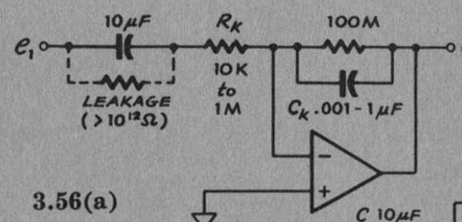


III.55
II.1
II.2
II.3
II.4
III.54
III.63
III.65
III.81

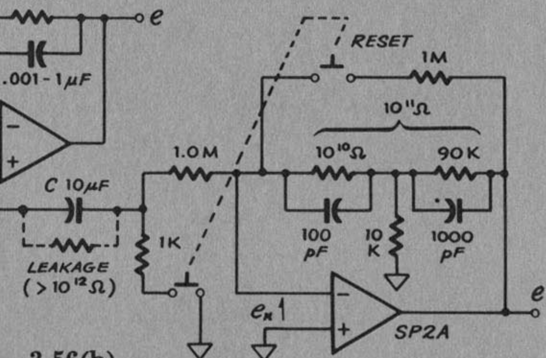
III.56 LONG-TERM DIFFERENTIATOR. Circuit (a) is a practical differentiator with a characteristic time of 1000 seconds. R_k and C_k are stabilizing elements (see II.18, et al) and the $10^{12}\Omega$ leakage is well tolerated (as are amplifier input current and stray leakage) by the 100 M feedback. To fix the order of magnitude, note that $de_1/dt = 1$ mV/sec. produces $e = 1$ volt!

Circuit (b) adds to our study of this extreme circuitry the need for reset (shown in the simplest form) and the means for extending the time-constant far above 1000 seconds. If $10^{12}\Omega$ is the best we can do in limiting the leakage, and if some "room" must be left for i_{in} and strays, then a feedback-network resistance of $10^{10}\Omega$ is just about all we dare use; however, the Tee Network (90 K/10 K) will buy us yet another order of magnitude, so that $T = 10^6$ seconds—not quite two *weeks!* Think of it: 1 *microvolt*/second in, for one volt out!

There is no merit in increasing C—the leakage will rise just as rapidly. Reducing C will only partly help, because then e_n/R will begin to dominate. The optimum value lies between 1.0 and 10 μF.



3.56(a)



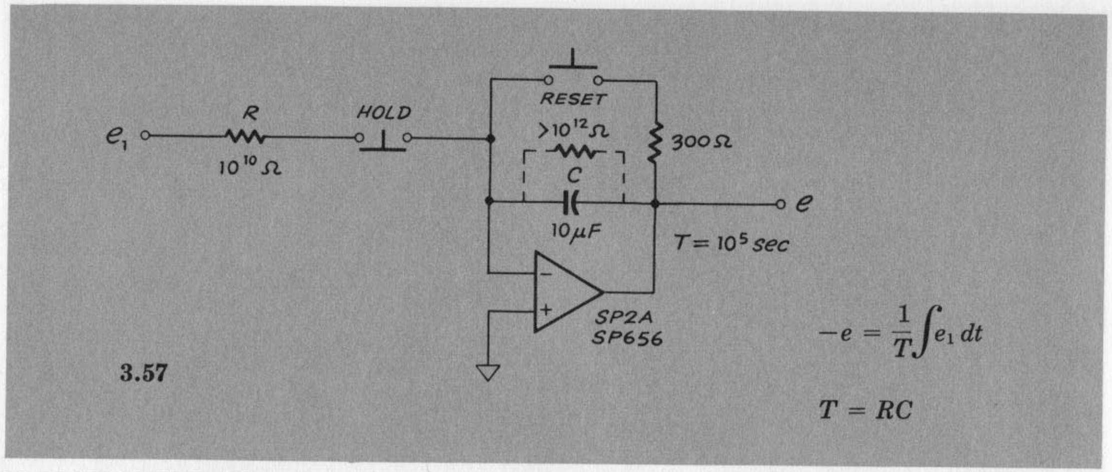
Reset it to existing level of e_1, after which differential changes may be measured.

3.56(b)

III.56
I.7
to
I.18
I.33
I.34
I.36
II.18
to
II.21
III.57

III.57 LONG TERM INTEGRATOR—MEMORY. This circuit takes full advantage of the very low offset current and voltage of a superior (Philbrick) modern (Philbrick) solid-state Operational Amplifier. Even at 10 volts full-scale, the input current is 10^{-9} amperes . . . yet an accuracy of better than 1.0% may be anticipated—with a characteristic time of more than a day! (By the way, the limiting factor is capacitor leakage, *not* the amplifier offsets!)

Other *caveats:* Beware of stray leakage (select good insulation, and guard critical paths); noisy ground returns (see I.30); and stray coupling (see I.31).

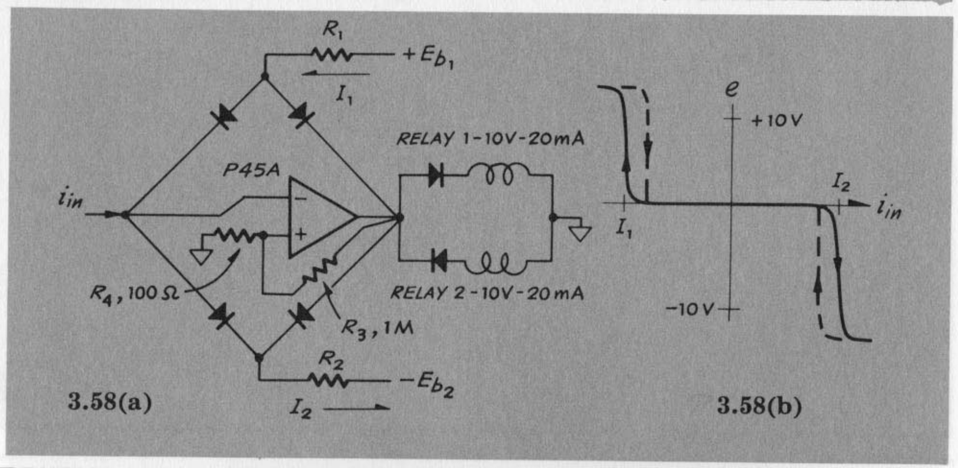


3.57

III.57 I.7 to I.18 I.33 I.34 I.36 II.10 II.11 II.12 III.56 III.73

III.58 NULL DEVIATION WITH 3-GRADE SORTING. This circuit has three possible states: (1) neither relay actuated; (2) relay 1 actuated; (3) relay 2 actuated. In state (1), normally-closed contacts on each relay, connected in series, "choose" the middle-grade-bin actuator—a condition corresponding to values of i_{in} that are smaller in magnitude than either I_1 or I_2. The output/input relationship is like that of the Dead Zone circuit of II.44 until i_{in} exceeds I_1 or I_2 in magnitude; after which, the loop opens and a rapid transition occurs in the output voltage, actuating relay 1 or relay 2. This output/input relationship is shown in (b).

R_3 and R_4 provide "latching" through hysteresis, as shown by the dashed lines in (b).



3.58(a) 3.58(b)

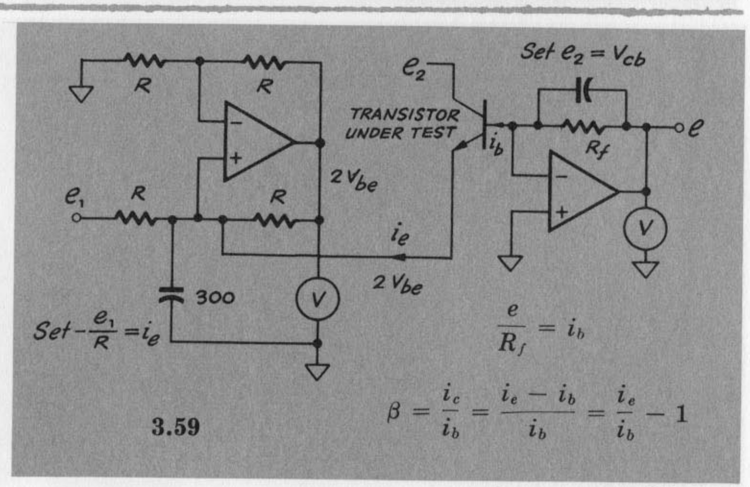
III.58 I.25 I.27 II.1 to II.4 II.30 II.41 III.43 III.44 III.45 III.47 III.48 III.78

III.59 PICOAMPERE TRANSISTOR PARAMETER TEST SET. Here we see the practical application of two familiar circuits to the testing of transistor parameters at picoampere current levels. The circuit to the right of the transistor, which is nothing more than the current-to-voltage transducer described in III.30, permits us to measure very low base currents conveniently and accurately, with a voltmeter that reads its output, e. The base current is then known:

$$i_b = \frac{e}{R_f} \qquad (3\text{-}23)$$

The emitter current is accurately established and maintained by the use of the Howland Circuit (III.6), so proportioned that the voltage read on the voltmeter is exactly twice V_{be}. (Note that the amplifier in the base circuit operates to maintain the base essentially at ground potential.) The emitter current is programmed by the input voltage to the Howland Circuit, according to the relationship $i_e = e_1/R$. V_{cb} is set by adjusting e_2. Note that this configuration provides a convenient means of measuring the "DC Beta" of high-gain transistors, with negligible approximation error:

$$\beta = \left(\frac{e_1}{e}\right)\left(\frac{R_f}{R}\right) - 1 \qquad (3\text{-}24)$$



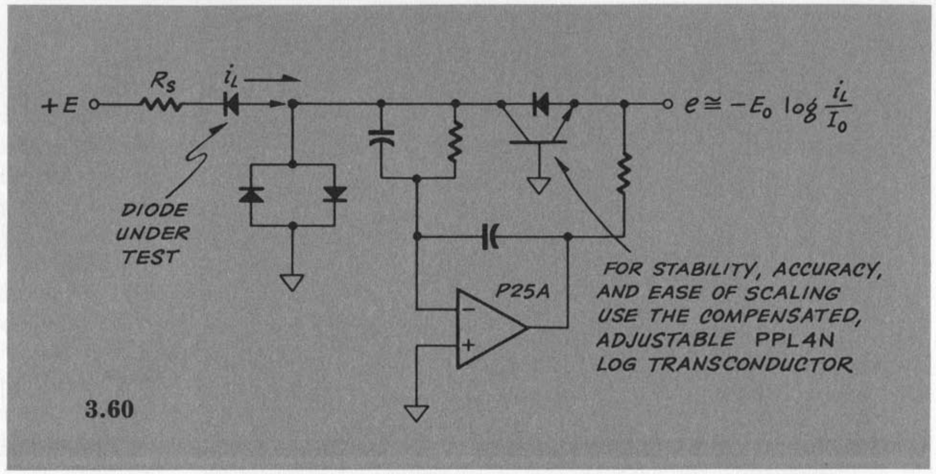
3.59

III.59 II.1 to II.5 III.3 III.4 III.5 III.6 III.7 III.8 III.30 III.60 III.61

pages 90 & 91

III.60 DIODE-TRANSISTOR LEAKAGE MEASUREMENT. The logarithmic-response null amplifier circuit of III.45 may be used to advantage in the measurement of very small diode or transistor leakage values. (A logarithmic characteristic is desirable since leakage values frequently vary, from diode to diode, over many decades. To protect against the occasional shorted or leaky diode, back-to-back diode clamps are shunted across the input to the circuit, and R_s is used to limit the current fed to them. A compensated, adjustable, logarithmic transconductor may be used here for accuracy, temperature independence, and gain adjustability.* The amplifier input terminal is deliberately decoupled from the diode clamps, to insure that they, and not the amplifier itself, limit first in the event of a shorted or very leaky diode. This circuit may be used to drive a recorder, a digital voltmeter, a conventional panel meter, or a limit comparator circuit (II.41 and III.58). It is well suited to automatic sorting of diodes or transistors.

*Philbrick SPL4-N/P are recommended.



3.60

III.60
I.25
I.27
II.1
to
II.5
II.22
to
II.31
II.41
III.45
III.58
III.59
III.61

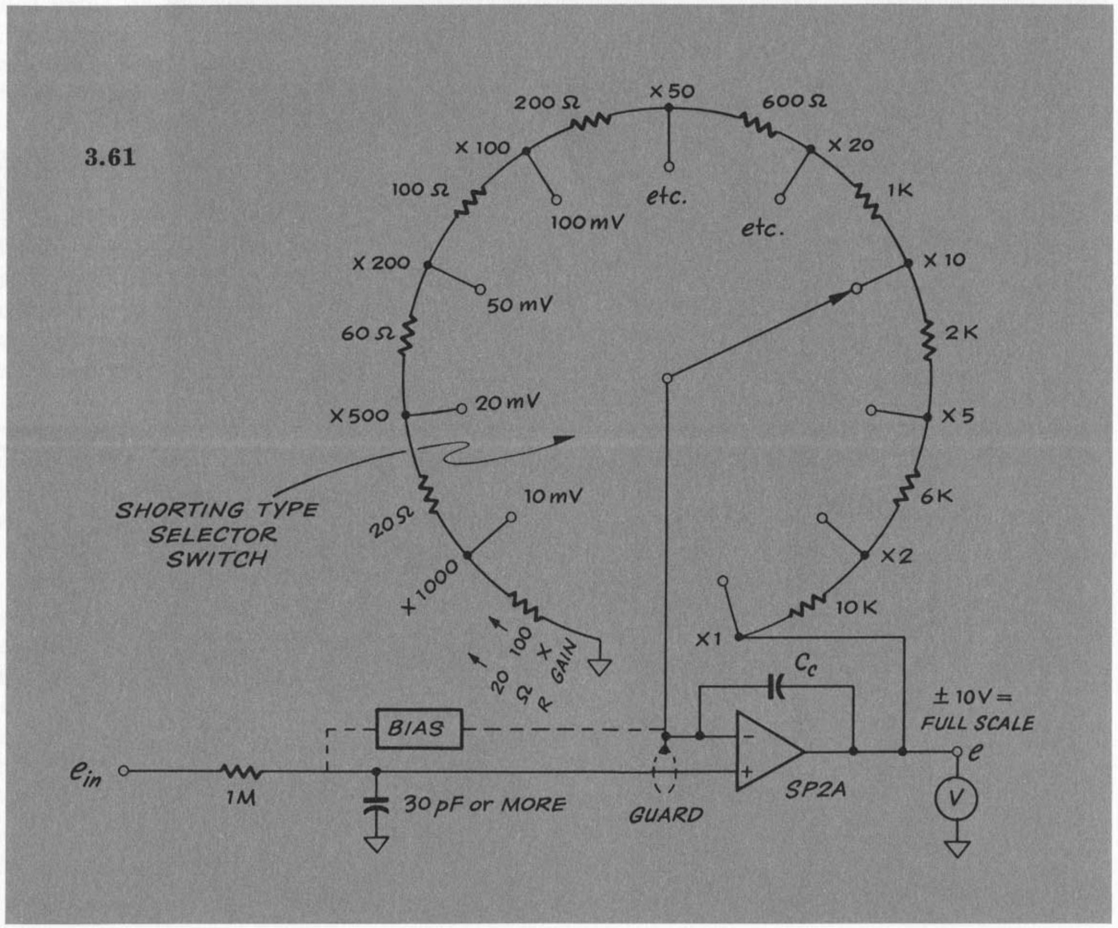
III.61 ELECTROMETER—MILLIVOLTMETER—AMPLIFIER. In this version of the Follower-With-Gain circuit, a ±10 volt output with nearly 2 mA of current capability is developed from full-scale input voltages as small as ±10 millivolts, while drawing less than 1 picoampere from the signal source. This exceptional sensitivity is made even more attractive by extremely high input impedance—ranging from at least 30,000 megohms (in the highest-sensitivity ranges) to millions of megohms (in the low-sensitivity ranges). The input impedance is actually the common-mode impedance of the positive input—in parallel with the differential impedance multiplied by the loop gain. With this order of input impedance (and appropriate care in construction), this circuit may certainly be assigned to the "electrometer" class, drawing currents of the order of 10^{-11} to 10^{-15} A. The input filter is provided to reduce noise and pickup, and both it and the compensating capacitor, C_c, are made as large as feasible to reduce the noise bandwidth without significantly slowing down the signal.

The "bias" current may be provided by a floating (cell) bias circuit (I.19, I.22) in series with a *very high* resistance, or by slightly offsetting the voltage bias, thus allowing it to develop an offset current through the amplifier's differential input resistance.

$$\left(\text{e.g.,}\ \frac{10\ \text{mV}}{10^{10}\ \Omega} = 10^{-12}\ \text{A}\right)$$

For less demanding applications, amplifiers of more modest performance at lower cost may be profitably employed.

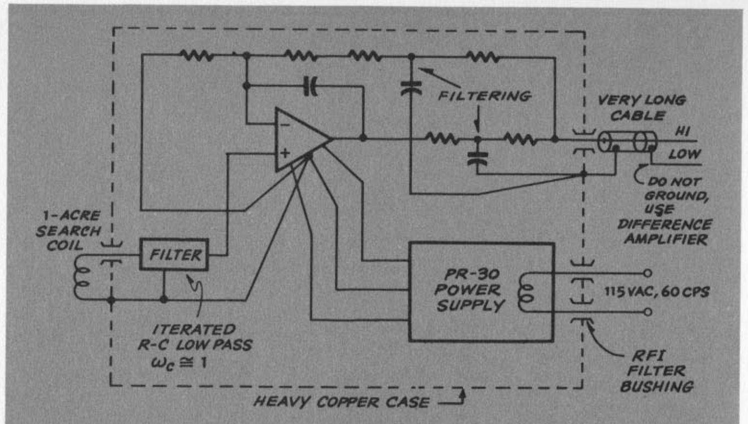
3.61



III.61
I.7
to
I.18
I.34
I.49
I.50
II.2

III.62 LOW-LEVEL MAGNETIC FIELD MEASUREMENT. This unusual application illustrates the low-level capabilities of Operational Amplifiers and shows the precautions necessary when extremely small signals must be detected in the presence of large ground currents and stray coupling. The earth's magnetic field varies very slowly. Even a 10,000-turn search coil 6 feet in diameter will produce only 1.28 microvolts rms for a peak flux excursion of 10 microgauss (1 Gamma) at 0.01 kHz. The only saving grace is the fact that the bandwidth can be spoken of in terms of cycles per hour; which permits us to use filters to reduce the effective noise level, moving power line disturbances many octaves away.

Note particularly: (1) the heavy copper case, for electrostatic shielding, and for some "shorted turn" effect on the incident power-frequency magnetic flux; (2) the "RFI filter bushings" used to lead signals in and out of the enclosure. Typical designs are the Sprague Type IJX 179 for the power lines and Type IJX 152 for the signal lines; (3) the use of an ungrounded coaxial cable for the output lead. If the cable is buried in the ground, it should serve as the signal return lead, and no other ground connections should be made. If the cable is not buried, the case should be grounded. Note that the input drives the positive terminal, and the amplifier is connected as a follower-with-gain.

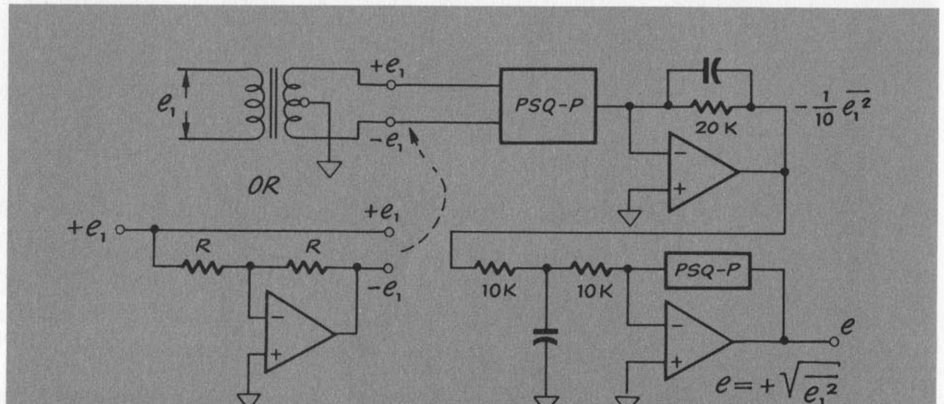


3.62

III.62 I.7 to I.18 I.28 to I.31 I.50 II.1 to II.4 III.22 III.23 III.24

III.63 TRUE RMS VALUE READER. Here we feed the signal and its inverted value into a voltage-squaring circuit employing the PSQ-P Quadratic Transconductor (see II.24 et al), averaging the squared output to obtain the *mean* square of the input waveform. We show alternate forms of obtaining e_1 and $-e_1$ from e_1 alone: a center-tapped transformer, and a unity-gain inverter.

After filtering, the output of the averager is then fed to a square-root circuit, which employs the PSQ-N Transconductor (II.23, et al) so that the output e, of this last circuit is the square root of the mean square of e_1, otherwise known as the true RMS value of e_1 . . . over a frequency range extending from zero well up into the audio spectrum. The square root and elementary filter circuits shown may be replaced by an integrator (II.10) for true continuous-accumulation (total energy) measurement.

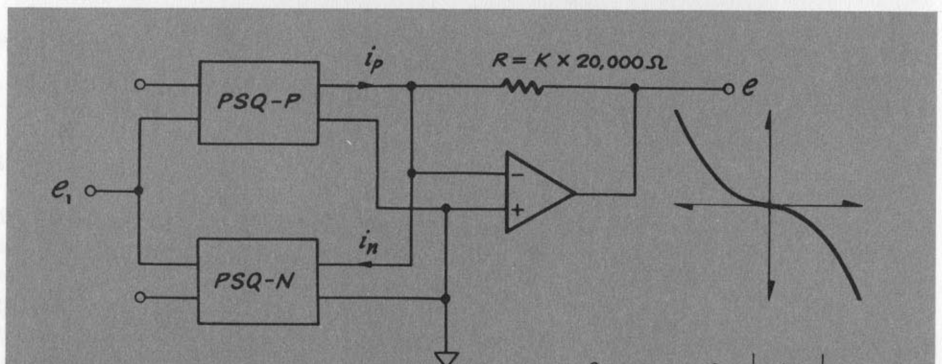


3.63

III.63 II.22 to II.32 III.54 III.55 III.65 III.81

III.64 ODD-VALUE SQUARER. The output of this circuit is proportional to the negative of the product of the input signal and the absolute value of the input signal—always in either the second or the fourth quadrant, as shown in the graph to the right. The constant K is proportional to the value of R. (For the standard Philbrick Quadratic Transconductors shown, $K = 1.00$ when $R = 20{,}000$ ohms.) As for the circuit, the so-called "ab-square," it consists of our old friend the current-to-voltage transducer, driven by the output current of one or the other of the Quadratic Transconductors. Positive values of e_1 will produce negative output voltages, and negative values of e_1 will produce positive output voltages. The current provided by the transconductor is proportional to the square of e_1, as is the resultant output voltage. The inverse of this function—the odd-value root, or "ab-root"—can be generated by interchanging the input and feedback elements.

Applications include: simulation of bidirectional hydraulic turbulent friction, synthesis of quadratic nonlinear conductance, creation of a graduated null or graduated dead-zone, or (when connected in the feedback path for odd-value rooting) synthesis of quadratic impedance.



3.64

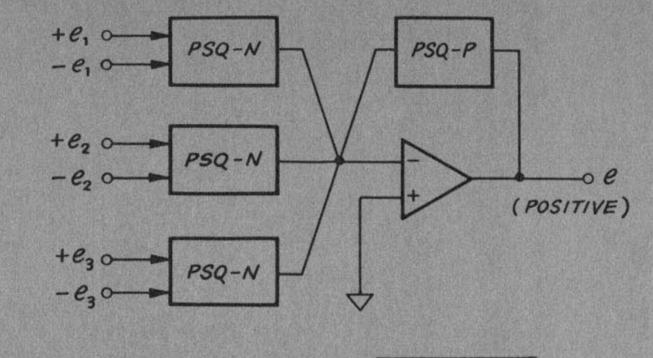
III.64 II.22 to II.32

III.65 PRECISE SQUARE ROOT OF THE SUM OF SQUARES. Each of the three input Quadratic Transconductors produces a current proportional to the *square* of the voltage applied to it. The currents are equally weighted, and their sum, proportional to the *sum of the squares*, is drawn from the summing point. The amplifier output, e, must force an equal current into the summing point, through the Quadratic Transconductor used as the feedback network. The required value of e is (see II.24) proportional to the *square-root* of the feedback current. Thus, e is proportional to the square-root of the sum of the squares of e_1, e_2, e_3, etc. *Now then,* please note that if the constants of proportionality applied to e_1, e_2, and e_3, are all the same as the one applied to e, and all are weighted the same, then the value of e is *equal* to the square-root of the sum of the squares, with *no* constant of proportionality required!

By the way, for problems yielding a negative output, that is:

$$e = -\sqrt{e_1^2 + e_2^2 + e_3^2} \qquad (3\text{-}25)$$

interchange the roles of PSQ-P's and PSQ-N's.



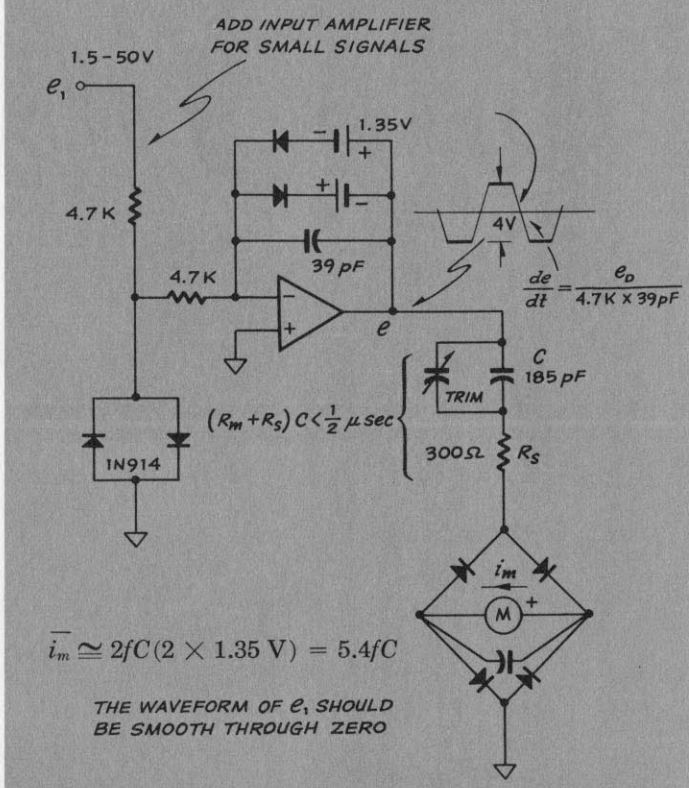
3.65 $e = \sqrt{e_1^2 + e_2^2 + e_3^2}$

III.65
II.22
to
II.32
III.54
III.55
III.63
III.81

III.66 FREQUENCY METER. This circuit exploits the high open-loop gain of an Operational Amplifier to generate a meter reading that is closely proportional to the frequency of an incoming signal, but independent of the signal amplitude, over amplitudes ranging from about ±1.5 V to ±50 V for frequencies from DC to 100 kHz. The amplifier is output-bounded by a symmetrical bi-directional feedback network consisting of diodes in series with mercury cells, the series combination being such as to bound the output at about ±2 volts (4 V, peak-to-peak). If the input voltage is large enough to be limited by the input-clamping diodes, an essentially-constant voltage is maintained across them. This clamped voltage drives a constant current through the 4.7 kΩ input resistor, causing a current of the same magnitude to flow through the feedback capacitor, producing an output ramp that increases in magnitude until bounded. The input resistors and clamping diodes also protect the amplifier against excessive input signal voltage, enabling it to handle a wide dynamic range of input signals.

The peak-to-peak amplitude of the trapezoidal output waveform sketched on the drawing is a function only of the parameters of the feedback bounding circuit, and the slope of the rise and the fall is essentially constant, depending only on the voltage-drop of the input-clamping diode. Therefore, as frequency increases, the ratio of ramp time (rise time and fall time) to bound time (the flat portion) increases, and the waveform approaches, in the limit, a triangular wave.

The amplifier drives a meter, through a full-wave-bridge rectifier, via a capacitor, C. Because of the low resistance of the meter and its associated diodes, the capacitor effectively differentiates the amplifier output voltage, passing a current proportional to the trapezoidal slope during the rise and fall portions of the waveform, and passing zero current during the bounded (flat) portion. Thus, the result of applying the trapezoidal wave-form to the rectifier circuit is to charge the capacitor C to the peak applied voltage, on each half-cycle of the driving signal. The charge current is a "spike" or pulse, with an *energy* content that is independent of the amplitude or frequency of the input signal. One such pulse of charge is fed to the meter for each half-cycle of the incoming wave, so that the average current through the meter in any reasonable averaging period is directly proportional to the number of those pulses that arrive each second, which is a true and linear function of frequency . . . and frequency only.



$\overline{i_m} \cong 2fC(2 \times 1.35\ \text{V}) = 5.4fC$

3.66

III.66
I.25
I.27
II.18
II.43
III.54
III.67