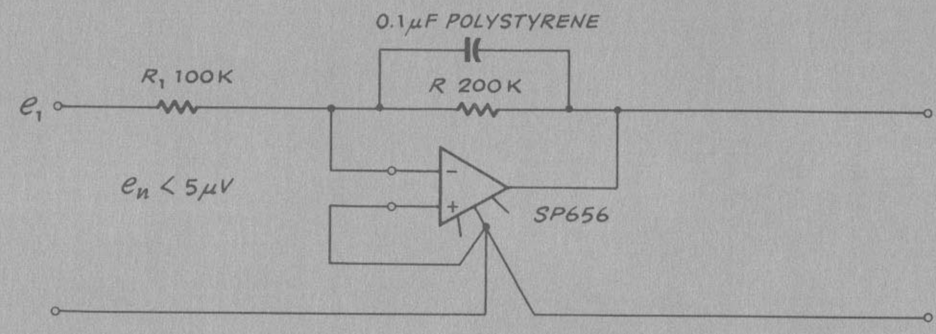
III.29 PRECISION DC AMPLIFIER. A low-drift (i.e., chopper-stabilized) single-ended amplifier, carefully grounded to preserve a true "high-quality" ground point common to input, output, and power circuits, if deliberately limited in bandwidth to reduce effective noise and ensure stability by a low-leakage (Polystyrene) feedback capacitor, will provide an extremely precise (1PPM) DC amplifier exhibiting, typically, less than 5 microvolts of null uncertainty. Under those circumstances, for DC and low-frequency signals, we may write that:

$$-e = \frac{R}{R_1} e_1 \tag{3-11}$$

to an accuracy almost entirely dependent on the absolute accuracy of the input and feedback resistors themselves. (At $e_1 = 5$ V, for example, e_n would be less than 1 PPM!)

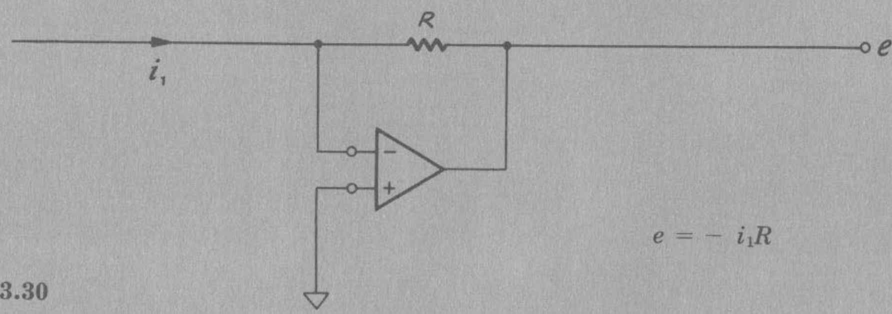


3.29

III.29
I.7
to
I.18
II.1
to
II.4
III.1
III.2
III.35
III.36
III.39
III.42
III.61

III.30 CURRENT-TO-VOLTAGE TRANSDUCER. In this simple powerful circuit, the output voltage is a direct function of the input current, $e = -i_1 R$. This "zero-drop" shunt provides an exceptionally inexpensive and convenient means of current measurement, since it not only introduces negligible voltage drop into the circuit, but its output voltage is developed at a low-impedance, high-energy level, capable of driving recorders, meter movements, or analog-to-digital converters.

For example, for 10^{-7} amperes full scale, $R = 10^7\ \Omega$ yields a one-volt output with at least 2-milliampere capability.



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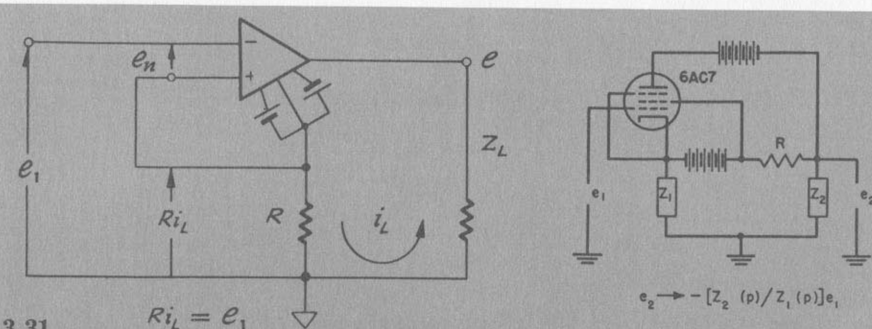
III.59
III.60
III.61
III.79

III.31 VOLTAGE-TO-CURRENT TRANSDUCER. This circuit establishes and maintains a precise proportionality between e_1, the input voltage, and i_L, the output current, such that

$$i_L = \frac{e_1}{R} \tag{3-12}$$

The proportionality is *independent* of Z_L, within the limitations of amplifier output voltage ratings. This proportionality arises from the fact that, so long as e_n is negligibly small with respect to e_1, and the amplifier input current is negligible with respect to i_L, the output voltage will drive current through Z_L and R so as to make $i_L R = e_1$.

Note that the power supply "floats" with respect to signal ground. If the chassis ground (which is also the AC ground) is tied to the signal ground and if there is not good electrostatic isolation between the primary and secondary, an error current will be induced to flow between chassis ground and power common returning through the load and amplifier.



3.31

NOTE: Z_L INCLUDES ALL PATHS FROM OUTPUT TERMINAL TO GROUND

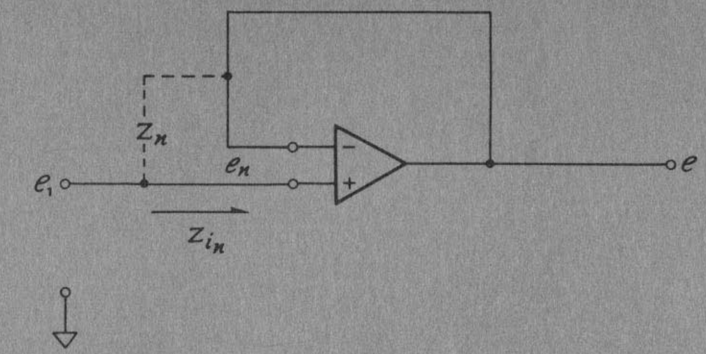
HISTORICAL NOTE: ONE OF GEORGE PHILBRICK'S EARLIEST OPERATIONAL AMPLIFIERS. ELECTRONICS, June 1948.

III.31
II.1
to
II.4
III.3
to
III.9
III.30
III.80

III.32 SIMPLE FOLLOWER. As described earlier, the simple voltage follower reproduces, to a very high accuracy, the input voltage e_1, without sign reversal—provided, of course, that e_n is very small compared to e_1. In the follower, the output is fed directly back to the negative input as degenerative feedback at high gain. Phase distortion and attenuation are negligible over the frequency range in which the amplifier gain is $\gg 1$. Input impedances from tens of megohms, to tens of gigohms and output impedances well below an ohm are easily obtained; as can be seen from the following expression, which relates input impedances to circuit and amplifier parameters, as shown in the circuit diagram:

$$Z_{in} = Z_{AG}{}^* \,||\, (Z_n \times \text{loop gain}) \tag{3-13}$$

*Ref. 1.18(c) . . . $Z_{AG} \,||\, Z_n \equiv \dfrac{Z_{AG}Z_n}{Z_{AG} + Z_n}$



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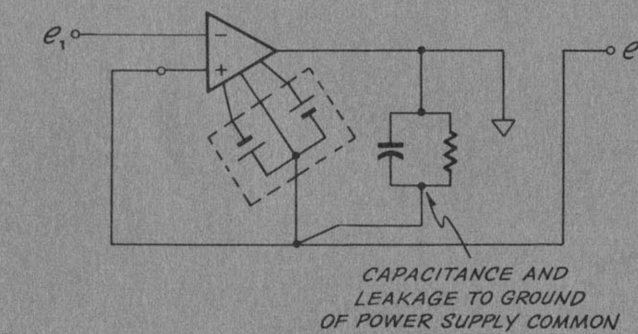
III.33
III.34

III.33 CHOPPER-STABILIZED FOLLOWER. Back in I.12, we described chopper stabilization of an Operational Amplifier, noting that the circuit usually employed left us with a single-ended amplifier. Such an amplifier would not lend itself to use as a conventional follower, such as that shown in III.32; however, by using the connection shown here, with the normal output circuit grounded, and the power-supply common return used as the output terminal, follower behavior is achieved. The capacitance and leakage resistance to ground of the power-supply common terminal are, of course, loading the output, but the output impedance is low enough, at least at low frequencies, to tolerate such loading.

Chopper-stabilized followers generally have exceptionally low attenuation at or near "DC," and very low drift. The input impedances are high at low frequencies, too. Typical values are: gain; 1.0000000 ($\pm$0.1 PPM attenuation at "DC"); input impedance: $>10^{12}$ ohms.

Note that this circuit is an inverted version of III.32.



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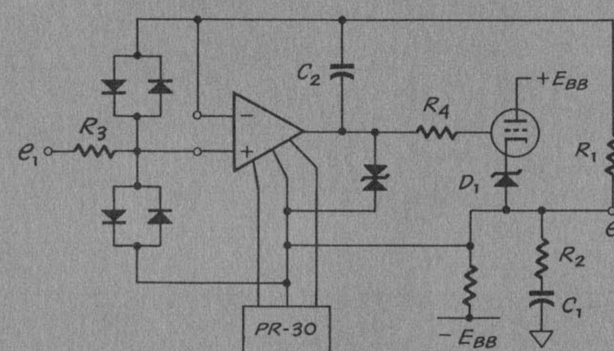
III.32
III.34

III.34 PRECISION HIGH-VOLTAGE FOLLOWER. Combining the best of two worlds, this circuit makes use of the wide dynamic range of a vacuum-tube cathode-follower circuit, and the high gain of a solid-state Operational Amplifier to produce a precise follower with high voltage range capability. The power supply of the Operational Amplifier rides along with the output voltage e. Diode pairs protect the amplifier input, and back-to-back zeners protect its output, in the traditions established in I.27. A principal limitation will be imposed by the insulation of the amplifier and power supply from the ground; another is dynamic stability (the amplifier's gain and the triode's μ are cascaded). R_2–C_1, C_2, and R_4 are introduced to stabilize the circuit against RF parasitic oscillations. R_1 and R_3 are current limiters, and are not large enough in value, compared to the input impedances they precede, to affect gain. D_1 establishes a convenient zero-signal (quiescent) bias level for the triode, and the Operational Amplifier output automatically seeks an appropriate quiescent level to satisfy the demands of the feedback around the circuit, while overcoming the contact-potential vagaries of the triode.



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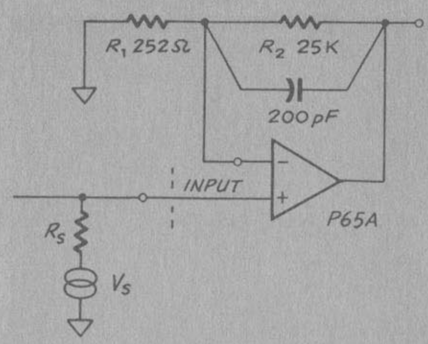
I.23
II.2
III.32
III.33

III.35 LOW-NOISE PRECISION PRE-AMPLIFIER. In this application of a follower with gain (see II.2) a precise gain of 100 is achieved, using moderately low impedances for optimum response and signal-to-noise ratio. Very precise, low-noise performance can be obtained with low impedance sources such as thermopiles or bridge circuits, as the data in figure 3.35 spell out. For source impedances of 10 to 100 kΩ, very good noise figures can be obtained (3 to 6 db, typically). The amplifier noted will perform very well, but for optimum noise figure an amplifier with field-effect transistors (i.e., model P25A) is recommended.



TYPICAL PERFORMANCE

DC Gain = 100.0
Bandwidth = 25 kHz (3 db down)
Linearity = 0.1% or better
Output Impedance < 20Ω below 1 kHz
Voltage Noise: 1.1 μV RMS (160 Hz–16 kHz)
(Referred to input, for $R_s \leqslant 1$ kΩ) 2.0 μV RMS (¼ Hz–25 kHz)
Noise Figure (R_s = 20 kΩ): 5 db (¼ Hz–25 kHz)
Input Impedance: 50 MΩ ‖ 6 pF
Long-Term Drift: 50 μV per week or better at constant temperature

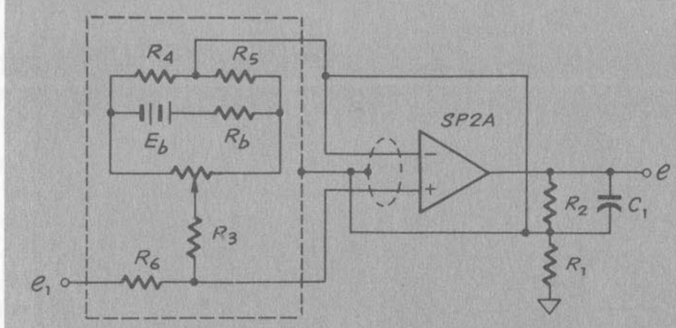
3.35

III.35 I.7 to I.18 I.50 II.2 III.36 III.42 III.61

III.36 SOLID STATE ELECTROMETER. The low-noise circuit just discussed in III.35 may be suited to electrometer service by adding the circuit modifications shown here:

- E_b, R_3, R_4, and R_5 make up a reversible current-offset-biasing circuit "bootstrapped" to extremely high impedance compared to R_6, so that the gain loss is negligible.
- C_1 is made large enough to narrow the bandwidth to a few cps, for lowest "white-noise" effect.
- R_b is chosen for optimum current bias sensitivity.
- Guarding of the input lead is provided, and returned to the negative input terminal.

Typical circuit values and performance are shown. For best performance, this circuit is not recommended for applications in which the operating temperature exceeds 50°C, and the lower the temperature, the lower the offset current and current noise.



$$e = \left(1 + \frac{R_2}{R_1}\right) e_1$$

$E_b = 1.3$ V

(Pot) ‖ $(R_4 + R_5) = 100$ kΩ

$\Delta E = \pm 6.5$ mV

For $\Delta I = \pm 2 \times 10^{-12}$A

$$R_3 = \frac{\Delta E}{\Delta I} = \frac{6.5 \times 10^{-3}}{2 \times 10^{-12}} \simeq 3 \text{ kM}\Omega$$

$R_6 = 10$ MΩ (protects e_1 source against faults)

3.36

III.36 I.7 to I.18 I.34 I.49 I.50 II.1 to II.4 III.29 III.35 III.38 III.42 III.61

III.37 NEUTRALIZING INPUT CAPACITANCE. In circuits employing the follower-with-gain configuration, the effects of capacitance to ground at the input terminal, whether due to input-cable shielding, strays, or amplifier input shunt capacitance, or any combination of the three, may be neutralized almost perfectly by connecting a capacitor, C_n, as a regenerative-feedback path from output to input, as shown here.

The undesired shunt input capacitance, C_s draws a current:

$$i_s = C_s p e_A \qquad (3\text{-}14)$$

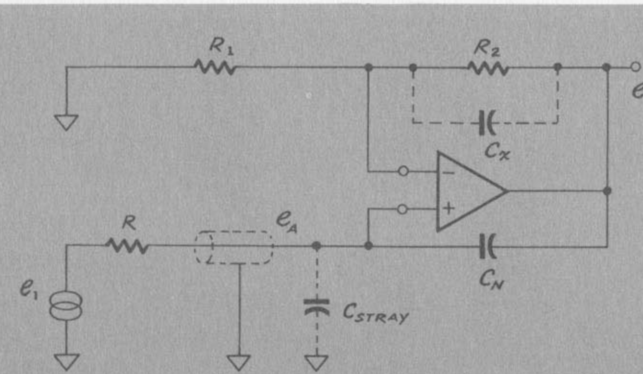
For perfect neutralization, C_n should draw an equal but opposite current:

$$i_n = C_n p(e - e_A) = i_s = C_s p e_A \qquad (3\text{-}15)$$

Thus C_n should be selected so that, for e_n negligible compared to e_1,

$$C_n = C_s \left(\frac{R_1}{R_2}\right) \qquad (3\text{-}16)$$

If C_n is too large, oscillation could occur, but C_x will prevent that. Note that much the same scheme may be used to "neutralize" leakage and other shunt input resistances, but the value of the neutralizing resistor becomes so large that a Tee network (see I.26) may be necessary.



3.37

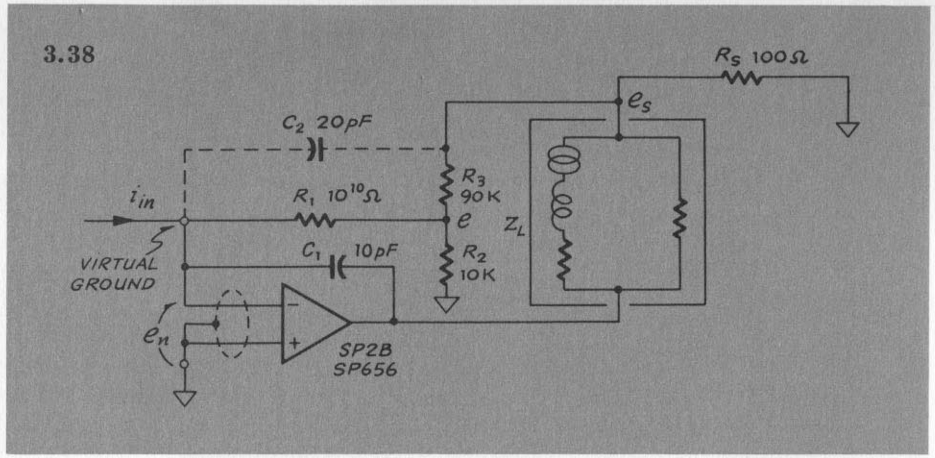
III.37 III.71 III.72

III.38 LOW-NOISE PRECISION CURRENT AMPLIFIER. This is the circuit of choice when the input current is referenced to ground and the load is floating. (If the load may not float, a circuit such as 3.7 must be used.)

In this rather more complex version of the current-to-voltage transducer of III.30, the current through the load, Z_L, is monitored by a low-resistance shunt, R_s, after which that signal e_s is fed back through a Tee network (see I.26) made up of R_3, R_2, and R_1, to the summing point, which is at a virtual ground. The load may be complex and active, within the limitations of the output voltage rating of the booster, but it must be amenable to "floating" off ground by e_s and R_s. With high-performance amplifier (such as either of those noted) the DC current gain of this circuit is given by:

$$\frac{i_L}{i_{in}} = \left(\frac{R_1}{R_s}\right)\left(1 + \frac{R_3}{R_2}\right) \tag{3-17}$$

For dynamic stability, feedback capacitance C_1 or C_2 (or both) is usually in order.



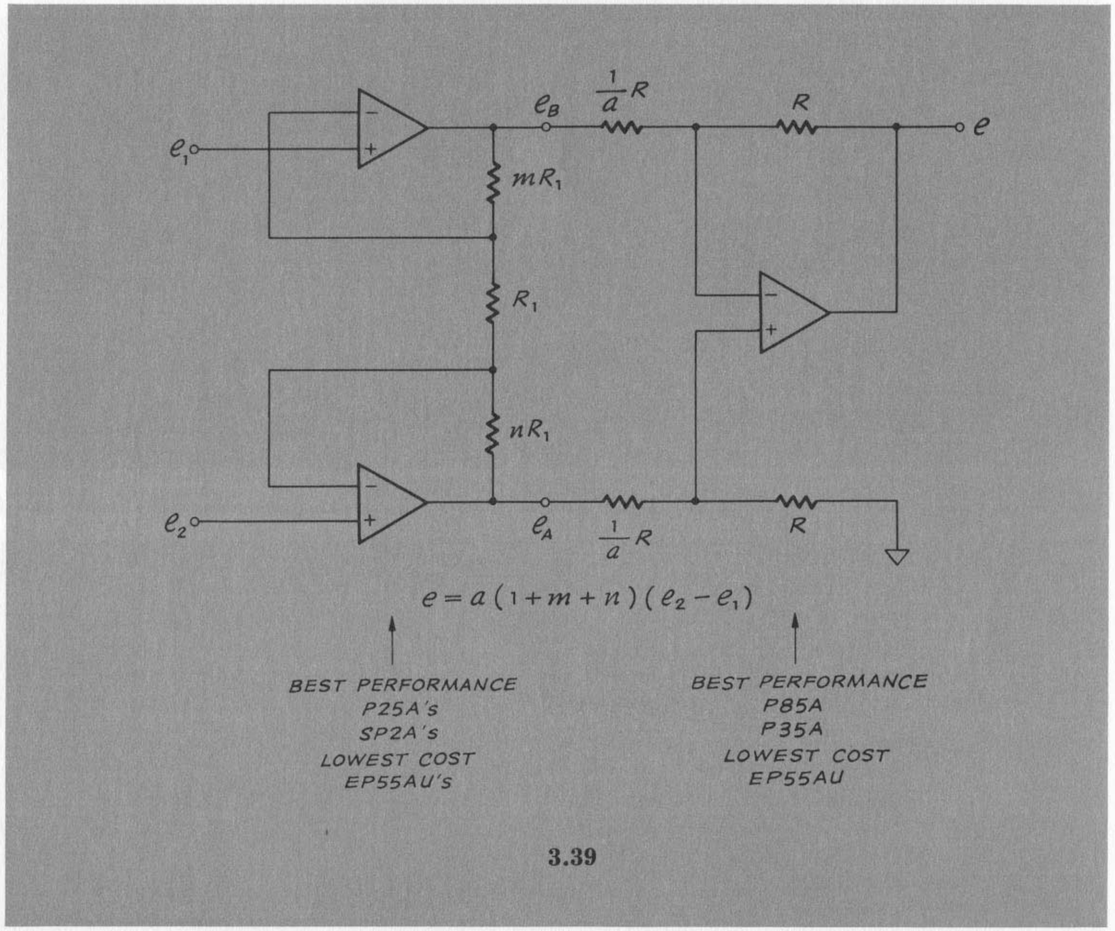
III.38
I.7
to
I.18
I.34
I.49
I.50

III.39 VERY-HIGH-IMPEDANCE DIFFERENTIAL INPUT AMPLIFIER. This circuit combines the low output impedance of a symmetrical subtractor (see II.5) and the very high input impedance and *CMRR* of a dual-amplifier balanced differential input stage. The output stage is straightforward, and requires no special comment other than the observation that, since the circuit is being driven from the low impedance outputs of the dual-differential-stage amplifiers, the impedance levels of the feedback and divider networks in the subtractor may be kept quite low, for good stability and bandwidth. The input stage is interesting in that (unlike a pair of followers-with-gain, which magnify both common-mode and differential signals) this cross-coupled circuit passes the common-mode signal *at unity gain*, but amplifies the differential signal with a gain of $(1 + m + n)$.

The limit of common-mode error is established by the degree to which the error voltages, e_n, for the two amplifiers: (a) approach zero, and (b) track one another.

The input impedances—both common-mode and differential—are extremely high, as might be expected of follower configurations, so that the circuit is almost completely insensitive to unbalances in source impedance of e_1 and e_2, and the input terminals are very close to "passive"—i.e., there is negligible interaction between the input terminals.

This circuit, equipped with the recommended amplifiers, constitutes an extremely high-performance differential-to-single-ended transition, with excellent gain, bandwidth and stability. It may be supplemented, of course, with a booster, if need be.



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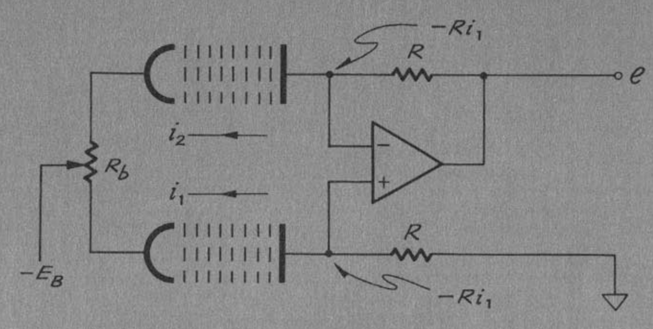
I.16
I.18
I.48
II.1
II.2
II.3
II.4
II.5

III.40 PHOTOMULTIPLIER-DIFFERENTIAL CURRENT AMPLIFIER. The Adder-Subtractor is very well suited to the conversion of the difference between small currents to a husky voltage at manageably-low impedance levels. In this example, the small currents are supplied by photomultiplier tubes operated from a power supply of hundreds to thousands of volts. The response of this circuit is linearly and accurately proportional to the difference between the two photomultiplier currents. Since the feedback networks cannot have more than about 10 volts across them, and probably have less than 5, the amplifier end of the circuit operates comfortably near ground, compared to $-E_B$; thus, the ultor voltages cannot vary by any large percentage with the current through them, so that the sensitivity will remain relatively constant, even with sizeable current differences. R_B is a zeroing control, and is meant to provide "dark-current" balancing. It may be used to set an *arbitrary* zero level at any standard level of light intensity. Typically, with $R = 10^8$ ohms, this circuit will drive a 1,000-ohm, 1 mA recorder full-scale for a current difference of only 10 nanoamperes.



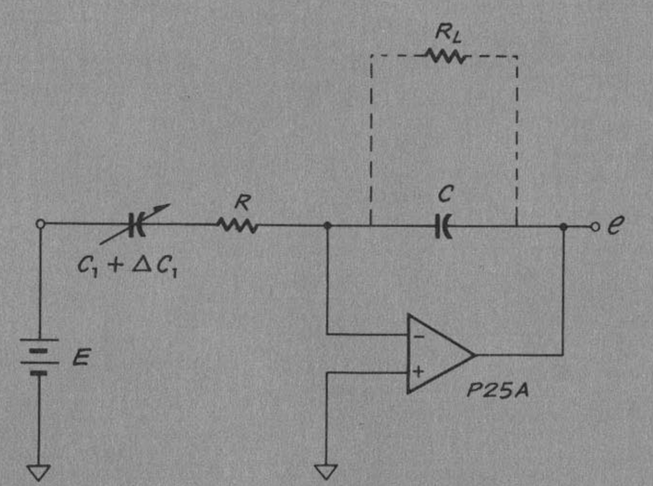
3.40 $e = R(i_2 - i_1)$

III.40
II.5
III.36
III.39
III.61

III.41 CHARGE AMPLIFIER. Capacitive transducers are often employed to accomplish the conversion of some mechanical, thermal, chemical, etc. phenomenon to electrical information. They respond to physical stimuli by providing varying electrical charge, which can often be considered as a constant voltage source in series with a variable capacitor, or vice versa. Typical of these is the capacitance microphone, represented in our drawing as a battery, E, in series with a capacitance $C_1 + \Delta C_1$. An Operational Amplifier, connected as shown, provides an ideal transition from change of capacitance to change of output voltage . . . unloading, linearizing, and providing appreciable output current and voltage capabilities.

A change in capacitance must cause a change in charge, and that charge is displaced into the summing point, which must be balanced by an equivalent displacement of charge across the feedback capacitor, C, caused in turn by a change in the output voltage, e. The response equation shown on the drawing relates the output voltage both to a change in capacitance, and to a change in the equivalent series voltage, E. Note that either or both—a change in C_1, or a change in E—will cause a change in output voltage, because *either will cause a change in charge* . . . hence, the name "Charge Amplifier." Piezoelectric transducers do not have an impressed voltage but do produce a dQ output. Compare this circuit with that of the current-to-voltage transducer of III.30: resistors convert current to voltage; capacitors convert charge to voltage.

Note that this circuit has the desirable property of being virtually independent of shunt capacitance across the input, since such capacitance is connected from the summing point to ground, and has across it only the residual null voltage, which should be negligibly low in practical application. This independence of input capacitance permits the use of long shielded cables between the transducer and the amplifier, without significantly affecting the accuracy of the conversion from charge to voltage. Leakage resistance in parallel with C, on the other hand, must be deliberately sustained, in order to prevent the amplifier output from drifting to saturation. (See Integrators, particularly II.12.) It will be noted that the sensitivity is inversely proportional to the value of the feedback capacitance; therefore, the smallest value that will be large compared to "strays" will yield the highest predictable sensitivity. One must then be sure that at the lowest frequency of interest, X_C will be small compared to R_L, and that the amplifier's offset current will be sufficiently small to prevent saturation with the required value of R. Amplifiers employing field-effect transistors are recommended for this application.



$$Q = C_1E = -Ce$$
$$dQ = -Cde$$
$$-de = E\frac{dC_1}{C} + \frac{C_1}{C}dE$$

3.41

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I.18
II.10
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II.12
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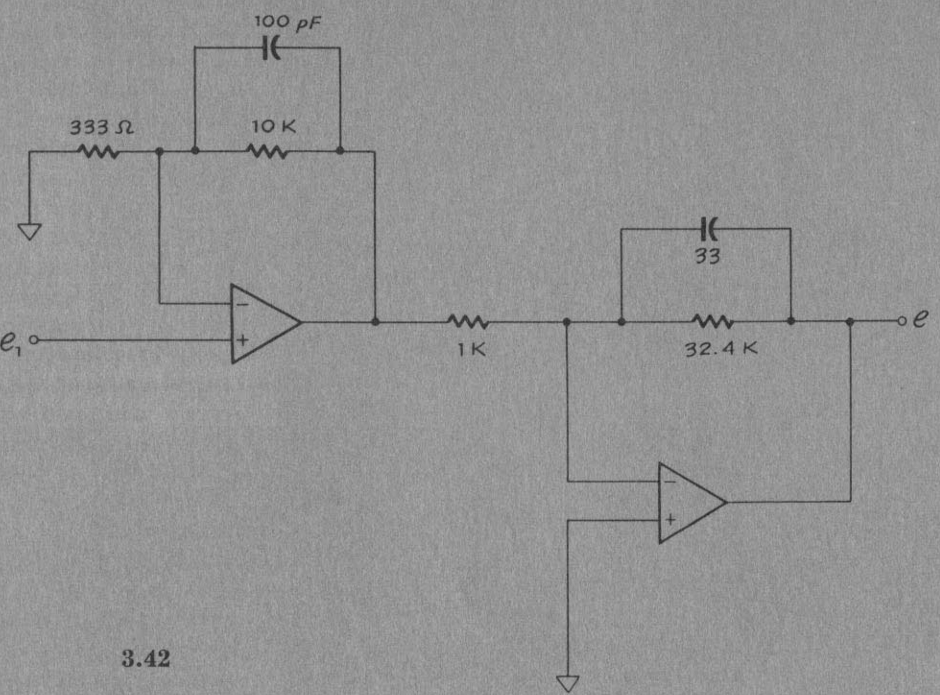
II.20
II.21

pages 82 & 83

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.



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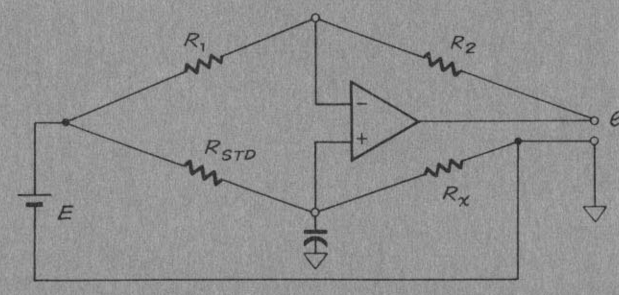
III.35
III.36
III.38

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.



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

3.43

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to
II.4
II.41
III.44
to
III.48
III.58
III.78
III.79