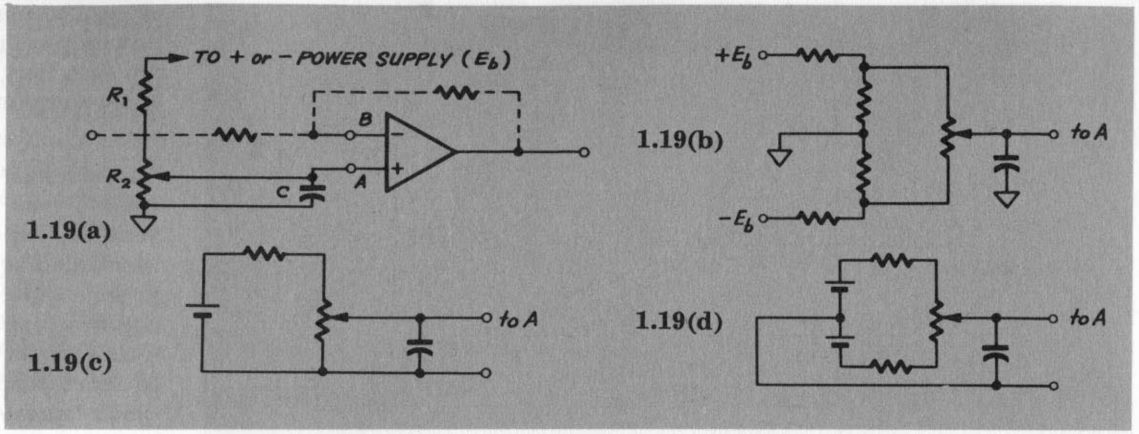
I.19 EXTERNAL VOLTAGE BIAS. All Philbrick solid-state Operational Amplifiers have either special terminals for external adjusting circuits, or built-in adjustment, the purpose of both being to achieve zero voltage offset. The internal adjustments, or the recommended external adjusting circuit when provided, should be used in preference to those shown here, to minimize temperature drift and assure proper internal operating points. Circuits (a) and (b) may be used at the positive input of a single-ended amplifier, (b) providing bipolar adjustment. Circuits (c) and (d), since they use cells, may be useful in differential amplifier applications, in which the bias supply must often "float."

Care should be exercised in component selection for all bias circuits, since drift in them will give rise to apparent amplifier drift.



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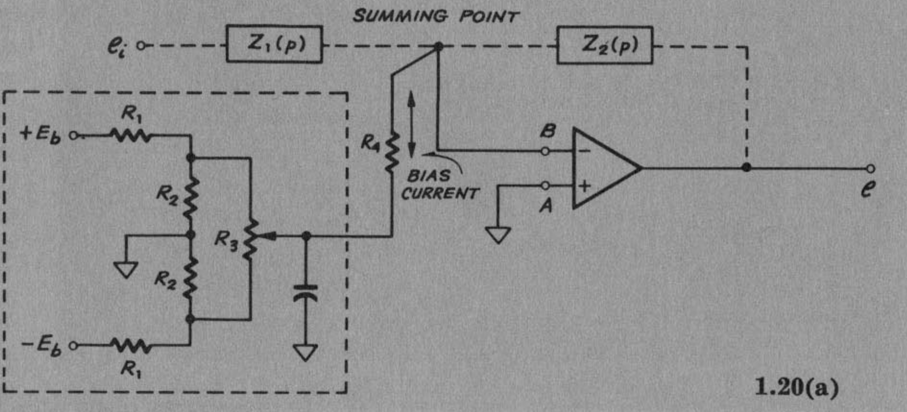
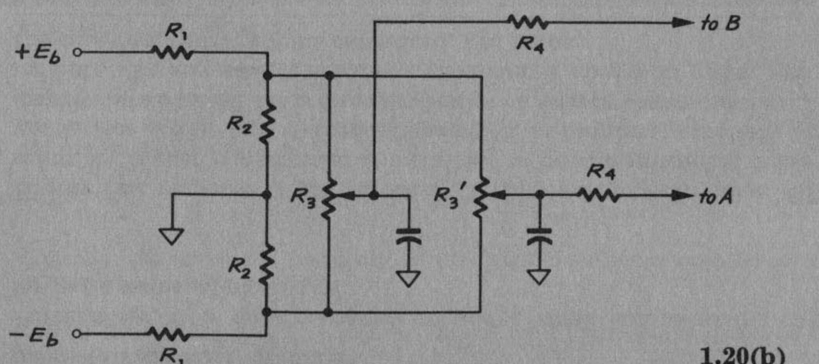
I.20 EXTERNAL CURRENT BIAS. Offset currents of either polarity may be supplied to terminal *B* in circuit (a) through a high resistance, R_4. (It should be noted that the R_1-R_2-R_3 circuit may be replaced by a potentiometer *only*, provided that the power supply provides a "clean" neutral terminal, for a discussion of which, see I.29 and I.30.)

Note that the part of circuit (a) that is contained within the dashed rectangle is the same as one of the voltage-bias circuits of 1.19(b); in fact, *any* of the bias circuits of I.19 may be used to develop current bias*, but it is less likely that the cell circuits would be used, since a higher voltage across the potentiometer permits the use of a larger value of R_4.

Circuit (b) may be used** to furnish adjustable, reversible bias currents to *both* inputs, provided either that the summing-point potential remains at or very near ground, or that the bias-supply "center-tap" may be floated and returned to a circuit point that is always at or near the summing-point potential, *or* that the variations in the summing-point potential are small compared to the voltage driving R_4 and R_4'.



1.20(a)



1.20(b)

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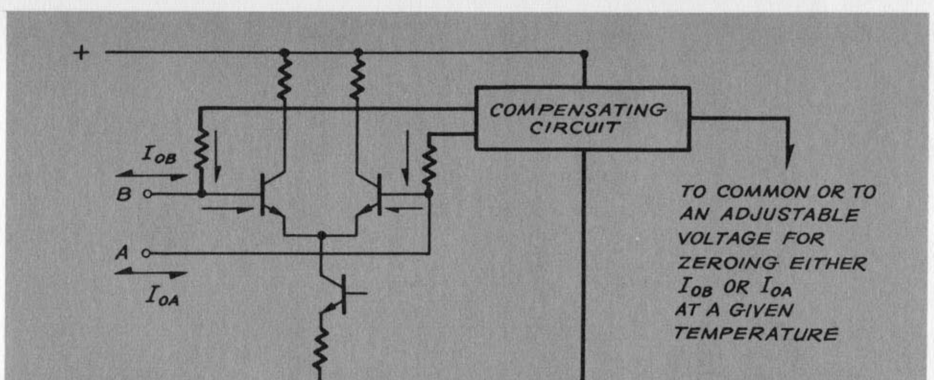
*In particular, the divider circuit of 1.19 (a) is well suited to current biasing of certain amplifiers having known, unipolar current offsets.

**At some small sacrifice in *CMR*, possibly, if the impedance unbalance introduced by the bias circuit is significant.

I.21 INTERNAL VOLTAGE & CURRENT BIAS. In most families of Philbrick Operational Amplifiers, there exist types in which voltage offset adjustments are self-contained. There are also types whose offset adjustment terminals are available, either for remote adjustment or for installation of fixed external resistors, in which bias adjustment is unnecessary or undesirable.

If very-nearly-*perfect* current biasing must be accomplished, it is usually accomplished externally, by connecting a recommended bias circuit at the amplifier's input terminals. However, there exist "compensated" amplifiers, in which current offsets are reduced to about 20% of the offsets in the prototype design of the family. In several of these types, an external terminal is available, to which a voltage bias may be applied to "zero" the offset current at either input, without loading that input.

The circuit shown at the right has temperature-compensated adjustable current bias, and is typical of compensated ("C"-type) Philbrick amplifiers.



1.21

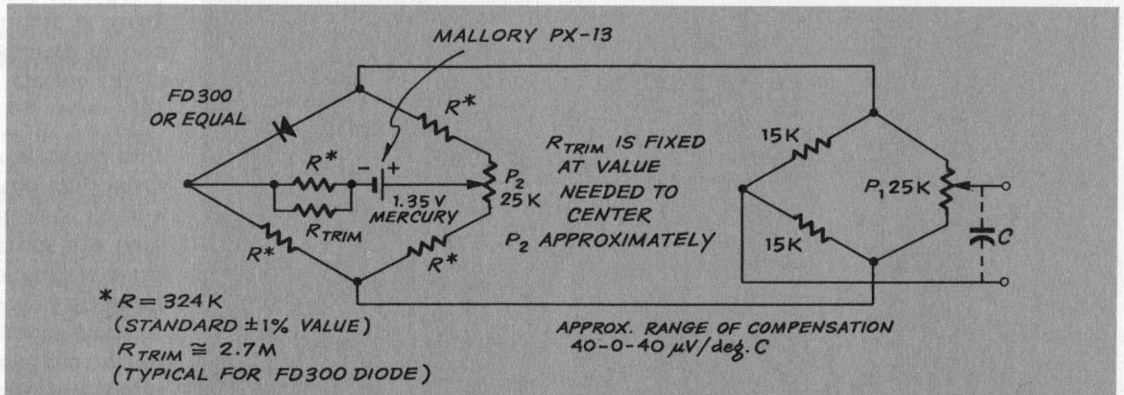
I.21 I.8 I.19 I.20 I.22 to I.24

I.22 TEMPERATURE-COMPENSATING BIAS CIRCUITS. The inexpensive circuit shown in (a) is designed to be inserted in series with either input to an operational amplifier for which bias (zero-adjustment) has already been provided.

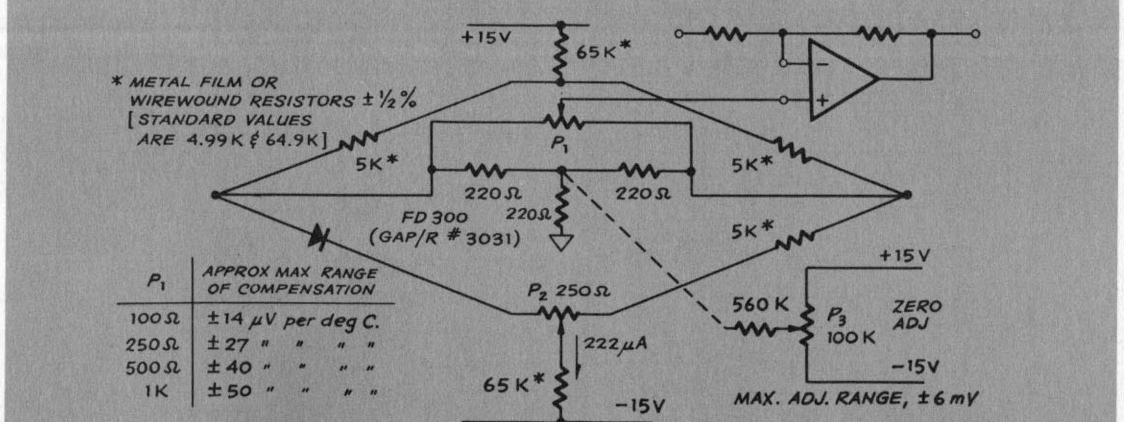
First, P_2 is adjusted so that, at room temperature, manipulation of P_1 has no effect, indicating that zero volts must now exist across the 15kΩ bridge. Presumably, at any other temperature, an offset will have developed in the amplifier. By re-zeroing the output, using only P_1, the entire circuit is exactly zeroed, at least at two temperatures. Because this circuit can be made physically small, its capacitance to ground may be kept low, and it can usually be connected in series with either input (floating). The 50-cent mercury cell recommended should last 15 years.

Figure (b) shows a similar circuit, intended for non-floating use, and operated from the amplifier power supply. First, at room temperature, P_2 is adjusted so that manipulation of P_1 has no effect. At some new temperature, typically toward the extreme of the desired range of operation, the amplifier is re-zeroed, by manipulation of P_1 only. The temperature coefficient of the resistors used in these circuits is unimportant, provided that it is repeatable, but the resistors must be stable with time, humidity, etc. Selective assembly or trimming to accomplish rough balance of the bridges (with P_2 centered) would obviate the need for ±½% tolerances on starred resistors.

Both of these circuits will reduce the effect of a 25°C temperature change to the effect of approximately 1°C change, when used with most Philbrick amplifiers.



1.22(a)



P_1	APPROX MAX RANGE OF COMPENSATION
100Ω	±14 μV per deg C.
250Ω	±27 " " " "
500Ω	±40 " " " "
1K	±50 " " " "

1.22(b)

I.22 I.8 I.19 to I.21 I.23 I.24

I.23 FOLLOWER BIAS CIRCUITS. Here are some biasing methods for use when Operational Amplifiers are used as voltage followers (see II.2). Circuit (a) has high input impedance with no attenuation, but has the capacitance of the bias circuit shunted across the input. Circuit (b) also provides high input impedance, but transfers the shunt capacitance to the low-impedance output. Both are unnecessary in amplifiers having internal voltage bias.

In (c), no cell is required, but the follower is left with a gain slightly higher than unity, due to the attenuation of feedback (see II.2). This can be avoided by using a constant-current source in place of R_1.

In (d) we show a scheme for *current* biasing. R_3 is very high —perhaps 10^7–10^9 Ω—and lowers the input impedance only slightly.

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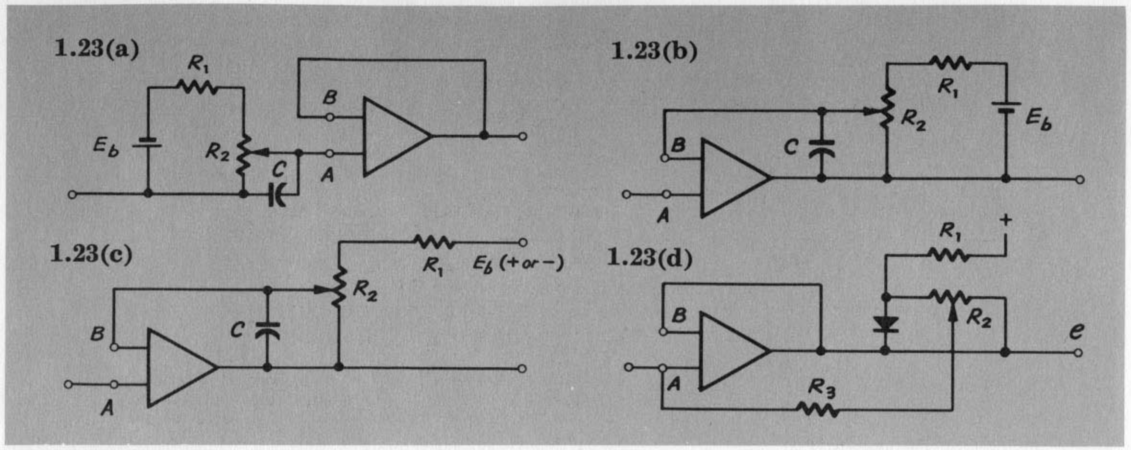

I.24 BIASING THE OUTPUT VOLTAGE. Sometimes it is necessary, or at least convenient, to develop the output signal, e, around an operating point other than zero. This is not at all difficult when the bias-plus-range of the output is still within the output range of the amplifier—in summing-amplifier circuits, one simply supplies an additive bias having appropriate scale factor as an input signal. However, it is not well known that, so long as the *range of output swing* (due to signals) and the maximum required value of output current are both within the amplifier's ratings, the output of the amplifier circuit may have a precise initial setting at a value of voltage many times that of the amplifier's maximum rating, *without using a booster*. The practical technology involved may be worth discussion.

First, an approximate approach: In the unity-gain inverter of circuit (a), a battery or zener diode, E_B, is used to bias the output. Either polarity of bias may be used; we show a negative operating point here. This circuit is easily realized, but drift in the battery or in the zener voltage could cause significant error, since it would be external to the feedback loop. If the load impedance is low, the zener or battery impedance will also cause a loss in gain.

Circuit (b) is better. The feedback is connected around E_B, effectively eliminating its drift, and the *dynamic* performance is unaffected by the presence of the zener; that is, the circuit provides an output precisely determined by the inputs and the feedback network, and the sole function of E_B is to maintain the amplifier output within the amplifier's working range. As an example of this, assume that the normal rated output voltage range of the amplifier is ± 10 V, and that $E_B = 10$ V. The amplifier will then function as a precise unity-gain inverter for total effective inputs of 0 to +20 V, producing outputs (e) of 0 to −20 V . . . while e', the unbiased output, is varying from *approximately* +10 to −10 V.

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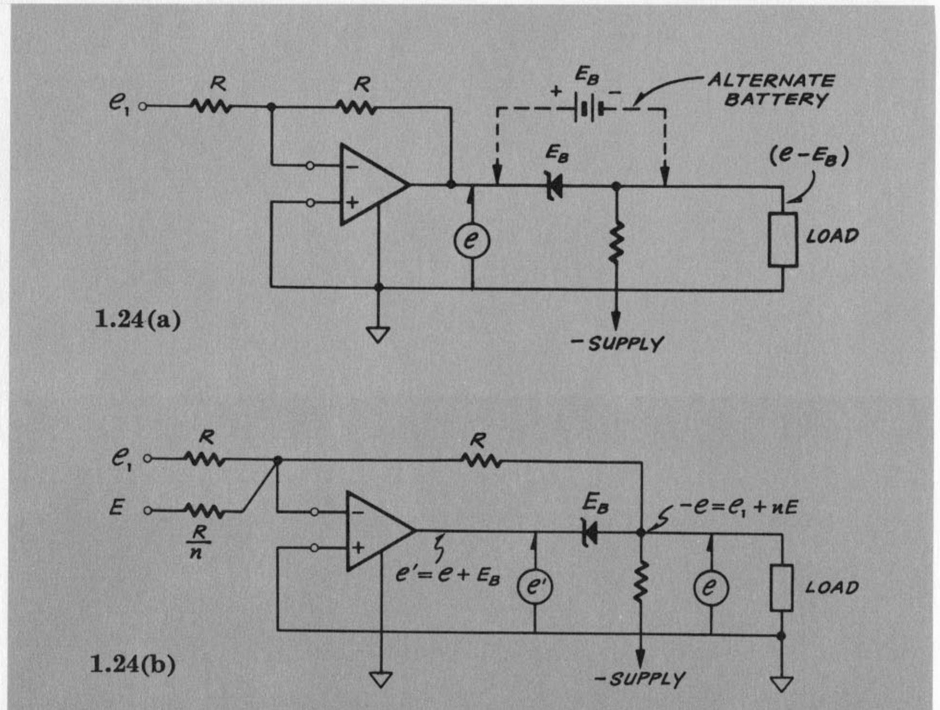
I.19
to
I.24



choose $E_B \cong nE$, if e_1 has symmetrical positive and negative amplitudes.

NOTE: be sure amplifier can supply worst case output current at all levels.

Note: Be sure amplifier can supply worst-case output current at all levels.

I.25 BOUND CIRCUITS—WHY & HOW. "Bounding" means restriction of the output voltage to some assigned maximum, even when the input signal exceeds its assigned full-scale value (either polarity). In practice, this is generally accomplished by use of a supplementary and non-linear feedback circuit, such as those shown here. Note that the ideal "bound" circuit is *dormant* until full scale is reached, at which time it takes over the job of supplying the feedback current necessary to balance the loop, at no significant increase in *e*.

So far as the amplifier is concerned, the output voltage *could* be restrained by a non-linear load, such as a zener diode connected across the normal load, or it could be allowed to go to its own limit, probably without permanent damage. However, after an amplifier "saturates", i.e., has gone off to a limit, it can require many seconds—even minutes—before it recovers its former zero adjustment.† This is particularly true of chopper-stabilized designs. (The time required for the actual recovery of the zero setting, to within, say, the typical value of short-term drift, may well take 10,000 times as long as the "recovery time-constant" often quoted for an amplifier.)

The use of "bound" circuits effectively eliminates the problem of recovery from overload by *preventing* both the occurence of overload and its consequences.

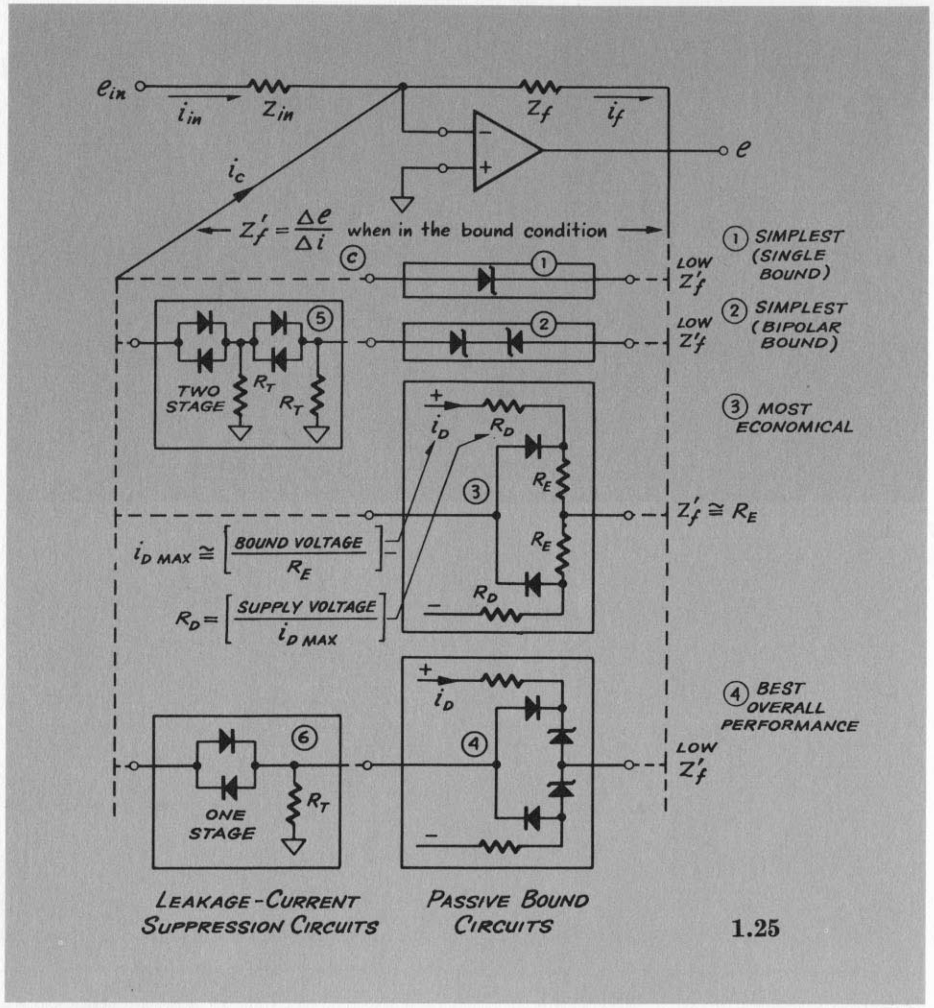
The ideal bound circuit would have an infinite "dormant" impedance *below* the critical value of bound signal, and zero "active" impedance above that value . . . in fact, it would resemble in those respects an ideal zener diode. Since, for almost all applications, both upper and lower bound limits must be established (i.e., both positive-going and negative-going signals must be limited) a pair of such ideal zener diodes would have to be used in the bound circuit. Since most applications specify symmetrical bound limits, we usually strive to make the voltage at which the bound circuit begins to conduct the same for the upper bound as it is for the lower bound.

In practice, the performance we have just described is approached with varying degrees of fidelity by circuits that range from simple to complex, and from inexpensive to quite costly, with performance roughly related to the cost and complexity invested. In the diagram to the right, we have shown and annotated some practical bound circuits, about which the following brief comments may be instructive:

- The simple zener bounds of circuits (1) and (2) frequently have need of leakage-current decoupling; for example, by means of diode-resistor Tee networks*, such as (5) and (6). This is especially true if i_c must be minimized.
- If Z'_f must be very low in the "active" mode, the zener bound is the preferred circuit. This is especially the case when the allowable excursion in *e* (when bounded) must be held to very small values.
- In designing decoupling circuits, such as (5) or (6), select diodes that have relatively high forward drop at about a microampere of forward current. This will enhance the isolation effect of the Tee.
- Shielding and guarding the low-potential terminals of the Tee may be required to achieve *extremely* low dormant-mode leakage.
- Circuits (3) and (4) have less reverse leakage than (1) and (2), but circuit (3) allows a greater excursion in *e* in the "active" mode, and circuit (4) is the most expensive. The combination of (4) and (6) has superb characteristics; (4) and (5) combine to approach the ideal; (2) and (5) approach the ideal at low voltage.

†Non-chopper-stabilized types will recover to within thermal-offset range in well under one millisecond after even prolonged overload.

*See I.26 for a more complete discussion of the Tee network. The decouplers shown in (5) and (6) are extensions of the resistive Tee, wherein diodes are used to achieve thresholds.



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I.26 TEE NETWORKS.

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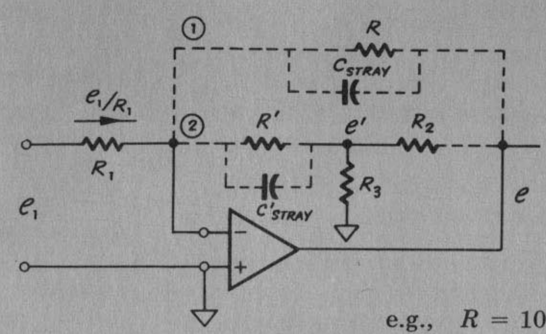
I.25
I.31

It is often desirable (and sometimes necessary) to replace a large feedback resistor by a 3-terminal resistance network, so proportioned that it has the same *feedback* effect, but higher accuracy, lower cost, and *far less susceptibility* to stray-leakage and stray-capacitance errors, because the tee network establishes much lower feedback impedance levels, end-to-end and to ground.* In the figure, e' must be equal to the product of input current and R'. This can occur only if e is equal to the product of e' and the inverse of the attenuation ratio of the divider formed by R_2 feeding R_3 and R' in parallel. Hence large values of R can be simulated by a 3-terminal network using practical values of resistance.

*We also use nonlinear Tee networks to reduce "dormant" leakage in bound circuits (I.25).



1.26

$$R'C'_{\text{STRAY}} \ll RC_{\text{STRAY}} \quad \text{if } R' \ll R$$

$$-e' = \frac{R'}{R_1}e_1$$

$$-e = \frac{R}{R_1}e_1 = \frac{R'}{R_1}\left(1 + \frac{R_2}{R'} + \frac{R_2}{R_3}\right)e_1$$

$$\text{Hence,} \quad R_3 = \frac{R'R_2}{R - R' - R_2}$$

$$\text{e.g.,} \quad R = 10^9\Omega,\ R' = R_2 = 10^6\Omega,\ \therefore R_3 = 1.002 \times 10^3\Omega$$

$$\text{Noise gain} \cong 1 + \frac{R}{R_1} + \frac{R}{R'},\ (R_3 \ll R')$$

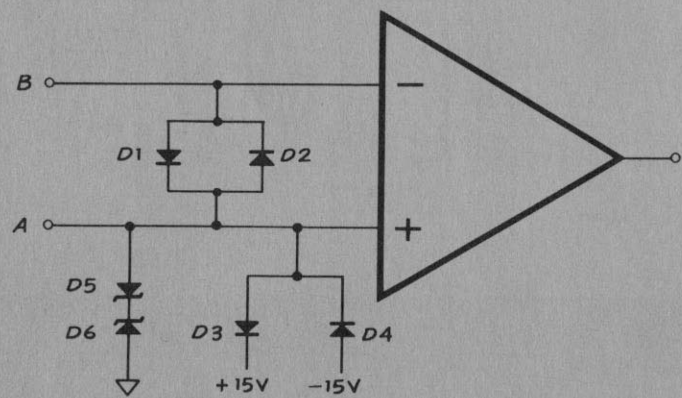
I.27 FAULT-PROTECTION, INPUT AND OUTPUT.

I.27
I.25

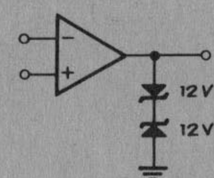
The bound circuits described in I.25 serve to restrict the excursion of the amplifier output voltage when the input signal exceeds its assigned "full-scale" value. They do *not* automatically restrict the flow of resulting fault *currents* within the amplifier to safe values. When very high voltages or currents can exist in the presence of a fault, additional protective measures should be taken.

Automatic limitation of amplifier input voltage can be accomplished as shown in diagram (a) by connection of a pair of diodes D_1-D_2 across the input. In most circuits, a few hundred ohms in series with each active input will not affect accuracy, and will protect the diodes from excessive currents, in the event that a "stiff" voltage supply of excessive magnitude is connected across the input. Protection against excessive common-mode voltage in 10-volt amplifier circuits is obtained by adding an additional pair of diodes to each input, as are D_3 and D_4 biased to positive and negative 15 VDC, as indicated in the diagram, or by a pair of 12-volt (±10%) zener diodes (D_5 and D_6) for each input, connected from the input to ground.

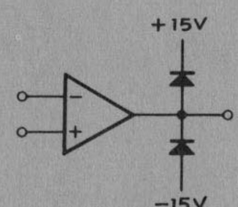
Protection of the output circuit against inadvertent load short-circuits is an inherent part of Philbrick amplifier design. Shorting the output to ground does no harm at normal operating temperatures (should be avoided in solid-state units operating above 65°C), and shorts to either supply voltage can be tolerated for brief periods. Prolonged shorts should be avoided, particularly in high-output units, to prevent damage to or alteration of component characteristics by excessive heating. A pair of 12-volt zener diodes between output and ground as in diagram (b), or reversed diodes to the ±15 VDC supplies in diagram (c) will provide insurance against damage from accidental connection of the output to higher-voltage sources. In all cases, diodes should be low-leakage types, with appropriate current and wattage ratings for worst-case fault conditions.



1.27(a)



1.27(b)



1.27(c)