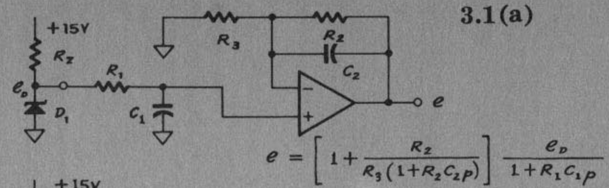
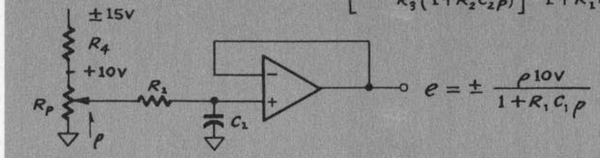
III.1 FIXED-VOLTAGE PRECISE REFERENCE. In Figure (a), R_z is adjusted to set the current through the temperature-compensated zener diode to a value at which it has essentially zero temperature coefficient. Noise and pickup are attenuated by the R_1C_1 filter, and the resulting voltage is amplified and buffered by a follower-with-gain circuit (see II.26). The capacitor C_2 is chosen primarily not for noise reduction, but to assure dynamic stability. The gain may be trimmed to compensate for the zener tolerance and achieve the precise voltage desired at the output. The resistors R_1, R_2 and R_3 should be low in value, in order that the amplifier input offset and noise currents do not contribute significant error. However R_1C_1 should be high enough for effective filtering, and $R_2 + R_3$ should not load the amplifier output excessively.

Figure (b) provides an adjustable output voltage, using as its precise reference a stable source such as the Philbrick PR-300 (or even PR-30) power supply, or circuit (a). R_4 may be trimmed so that the full-scale output (e.g., 10 V) is delivered precisely when $\rho = 1$. Since this circuit does not load the potentiometer, its dial calibration can be made accurate (say, $\pm 1\%$) and linear.



3.1(a)



3.1(b)

III.1
I.7
to
I.14
I.18
to
I.22
I.29
I.30
I.31
I.34
I.46

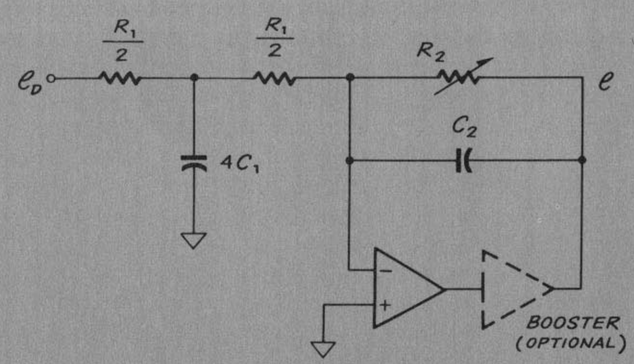
III.2 PRECISE VOLTAGE SOURCE. Instead of the differential amplifier circuits described in III.1, the single-ended amplifier circuit shown here may be preferable, especially since the amplifier may then be a chopper-stabilized design. The output voltage is directly proportional to R_2, which may, for convenience, be a decade resistor box (typical range 0 to 10 k) or a (10 k) rheostat may be used. In the latter case, the input resistance, R_1, will probably have to be trimmed to accommodate the resistance tolerance of the rheostat. The dial calibration will be accurate to the limits of resolution and linearity of R_2. The RC filter shown here may require a larger value of capacitance for the same attenuation than did that of the III.1 circuits.

The input voltage, e_D, may be derived from a resistor and compensated zener diode circuit, as in 3.1(a). However, to achieve maximum temperature stability, that resistor, R_z, must be smaller than it was in 3.1(a), because of the input current, e_D/R_1. Alternatively, e_D may be a stable power or reference supply.

Special attention to grounding is required if the full capabilities of this circuit are to be realized. Ground returns from any heavy loads should be returned directly to the amplifier's power common at the power supply. Also, the current flowing through the zener diode (if used) should not be allowed to flow through a high-quality ground system to a remote tie point. Instead, either tie the high-quality ground to the zener supply's power common at the zener diode, or bleed a balancing current from the negative side of the supply to the junction of the zener diode and high-quality ground. If the power supply is used as the reference, avoid corrupting the reference voltage with common-line resistive drops caused by amplifier quiescent and load currents. (This can be avoided by running separate power and reference wires between the supply and the amplifier circuit.)

The limit of error of this device may be stated as the sum of the following error sources:

- The absolute accuracy to which the input current may be established.
- The stability, against both time and temperature, of e_D.
- The absolute accuracy to which R_2 is known, at the particular setting used.
- The amplifier input offset current which represents some fraction of the summing-point current, and creates a first-order error.
- The amplifier (residual) offset voltage, including drift and all other uncertainty factors (see I.14), which also constitutes a first-order error term (and is one of the factors affecting the accuracy with which the input current may be established).
- The finite-gain error factor—possibly significant for amplifiers having low gain.



$$e = -\frac{R_2}{R_1} \frac{e_D}{(1 + R_2C_2p)(1 + R_1C_1p)}$$

$$\left(\text{at DC,} \quad e = -\frac{R_2}{R_1} e_D\right)$$

3.2

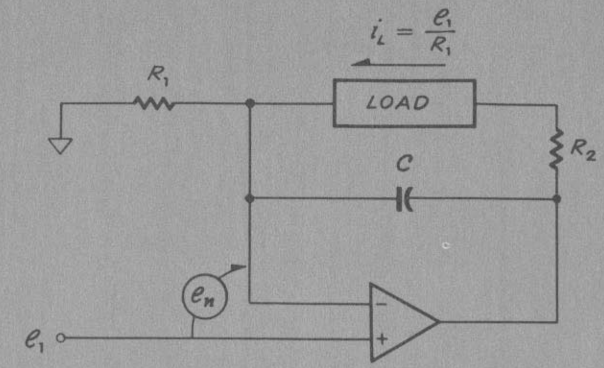
III.2
I.7
to
I.14
I.18
to
I.22
I.29
I.30
I.31
I.34
I.46
II.1
II.3
II.4
III.1
III.61

III.3 CURRENT PUMP I — FLOATING LOAD. This circuit is similar to the follower with gain 2.2(b) with the load replacing the feedback resistor. The amplifier is selected so that the noise and offset currents at each of its inputs cause error that is sufficiently small as to be negligible in the application. The high (low-frequency) gain of the amplifier assures that the voltage across the resistor R_1 is very nearly e_1, provided that the amplifier has not saturated. The current through R_1 must also flow through the load. (This current can be provided by a booster, if the amplifier cannot furnish enough.)

Special care is required to prevent the flow of heavy load current through the high-quality ground system to a remote tie point. It may be wise to put the tie point at the resistor R_1.

R_2 and C are selected to achieve stability. Neither is *always* required, but both are necessary when driving inductive loads. Also see 2.22(b), in which the load is a grounded-base transistor. There are innumerable active and/or non-linear loads for which this circuit may not be sufficiently stable. A more elaborate circuit will be required to drive current through such loads.

3.3



III.3
I.7
I.18
I.34
II.1
to
II.5
III.4
to
III.9
III.31
III.38
III.59
III.80

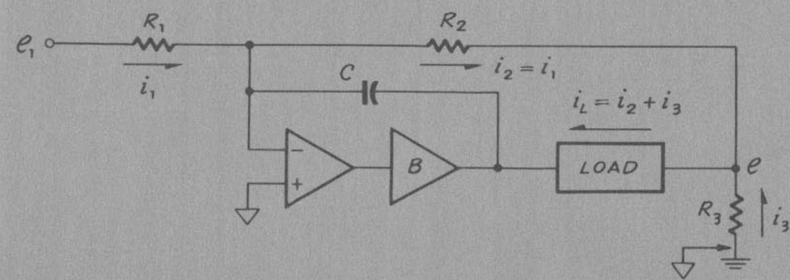
III.4 CURRENT PUMP II — FLOATING LOAD. In the current pump of circuit (a) the input current is again given by e_1/R_1 to the accuracy determined by the ratio of e_1 to e_n. The feedback current that flows through R_2 is equal to the signal input current i_1, to an accuracy determined by its ratio to the input error current of the amplifier. The voltage across R_3 is $-R_2 i_1$, subject to the errors just described, and the booster must furnish both i_2 and i_3 . . . through the load. Thus, the load current is proportional both to the input voltage and the conductance of R_1.

Once again, both load terminals are high, and the load may be complex and active, within the voltage and current capability of the amplifier and booster. This circuit can be used as a deflection-coil driver, in which case, C and R_A would be needed for dynamic stabilization. These components, in addition to the booster voltage-range and the deflection-coil inductance, limit the maximum rate of change of load current, which may be an important factor in deflection-system performance, particularly for a saw-tooth current wave form.

Circuit (b) shows a version of this circuit, in which the input signal, i_1 is derived directly from a current source, in which case, the current pump operates simply as a current amplifier, the gain of which—i.e., the ratio of i_L to i_1—is given by:

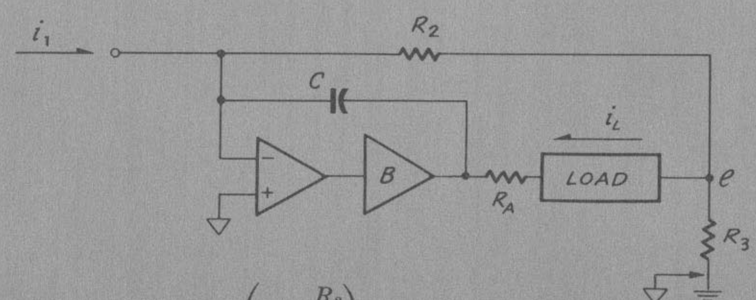
$$\frac{i_L}{i_1} = 1 + \frac{R_2}{R_3} \qquad (3\text{-}1)$$

Note that the power-common symbol is used for the termination of R_3. This connection is especially recommended when the drop caused by i_1 flowing through the path from high-quality ground to power common would be significant compared to e_1.



$$i_L = \frac{e_1}{R_1}\left(1 + \frac{R_2}{R_3}\right)$$

3.4(a)



$$i_L = i_1\left(1 + \frac{R_2}{R_3}\right)$$

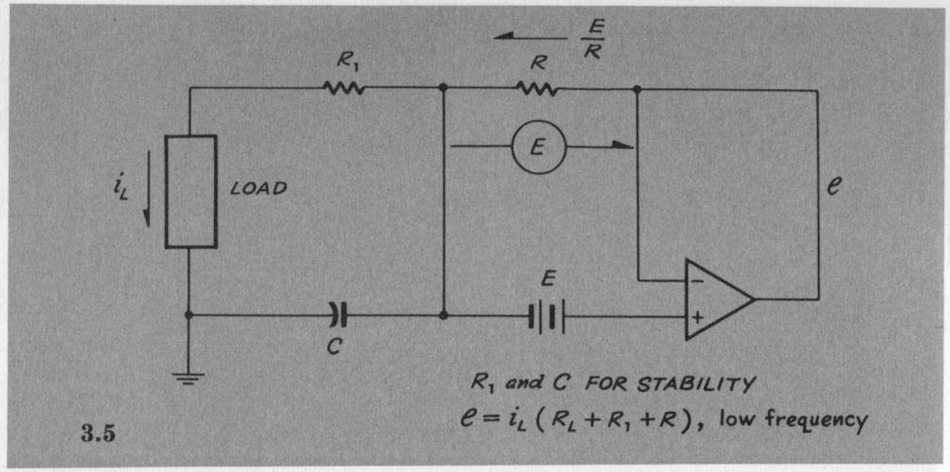
3.4(b)

III.4
I.7
I.18
I.34
II.1
to
II.5
III.3
III.5
to
III.9
III.31
III.38
III.59
III.80

pages 64 & 65

III.5 CURRENT PUMP III — GROUNDED LOAD. In this circuit, the voltage across R is equal to the reference voltage, E, within a margin of error given by the ratio e_n to E. The current through R is, therefore, E/R. Assuming that the amplifier input current is very small compared to E/R, the same current, E/R, flows through the load. Provided that the sum of e_L and E does not exceed the output voltage capability of the amplifier, and provided that i_L does not exceed its current capabilities, the circuit will drive the calculated current through any load impedance, including complex loads. Active loads can of course be driven but often tend to reduce the available voltage range, and may introduce a common-mode error factor.

This circuit has the convenience of grounding the load, Z_L, and the inconvenience of floating the reference source, E. Note that R is usually more convenient to adjust than E, to set the desired load current. R_1 and C are suggested for dynamic stability, particularly if the load is reactive.



3.5

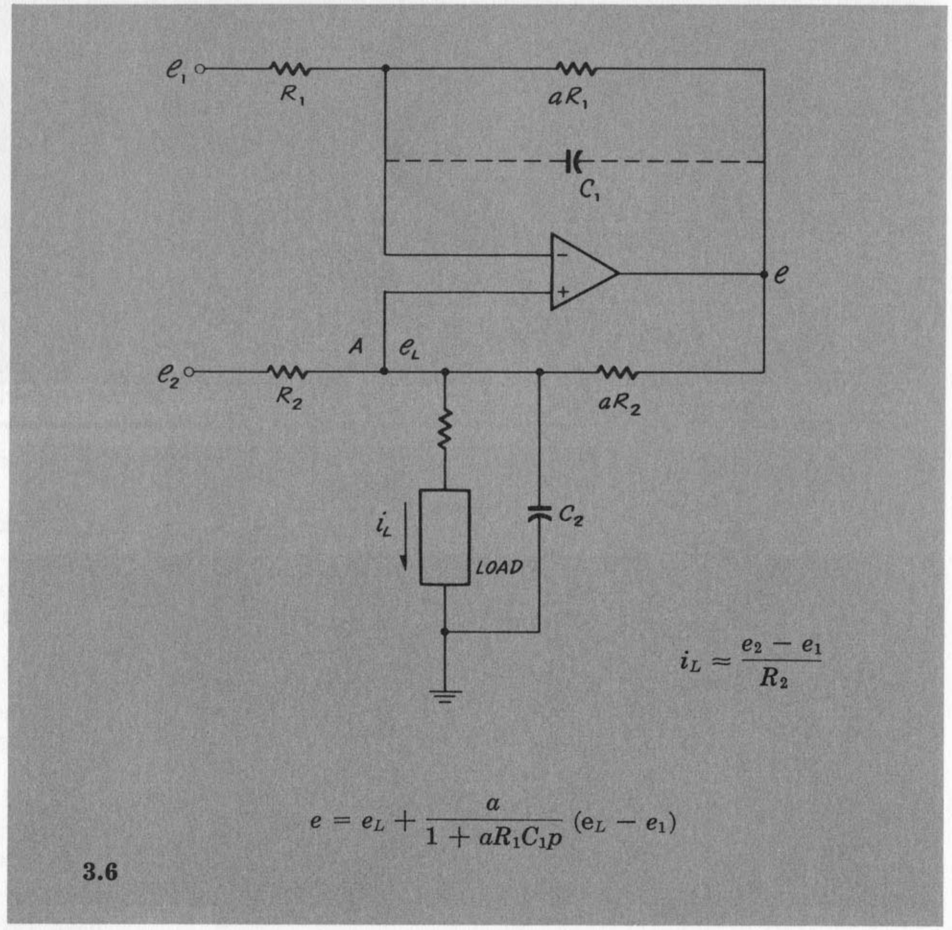
III.5
I.7
I.18
I.34
II.1
to
II.5
III.3
III.4
III.6
to
III.9
III.31
III.38
III.59

III.6 CURRENT PUMP IV — GROUNDED SOURCE, GROUNDED LOAD. The "Howland" circuit is made, as are most current sources, to approach infinite impedance by the application of positive feedback. The current through the load is given by:

$$i_L = \frac{e_2 - e_1}{R_2} \tag{3-2}$$

to an accuracy determined by the ratio of e_n to e_L, and by the ratio of the amplifier input current to i_L. In this circuit, both input voltages may be developed with respect to ground, and one side of the load may be conveniently grounded. Further, within the voltage and current capabilities of the amplifier, as with all of these circuits, both positive and negative currents may be driven through the load, depending only on the polarities of e_1 and e_2.

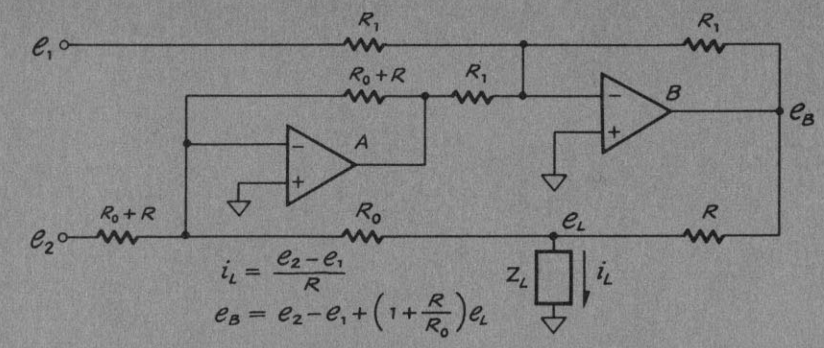
Note that e_2 supplies the short-circuit load current, whereas e_1 need only supply the current to a network that may usually be made to have a relatively high impedance; therefore, when and if only one signal needs to be used, and when the available signal polarity is appropriate, it is preferable to ground e_2 and use the e_1 terminal as the current-determining input terminal.



3.6

III.6
I.7
I.18
I.34
II.1
to
II.5
II.28
II.29
III.3
III.4
III.5
III.7
III.8
III.9
III.31
III.38
III.59

III.7 CURRENT PUMP V, CHOPPER-STABILIZED. When it is desired to gain the advantage of chopper stabilization in constructing a current pump of the Howland Configuration, one must deal with the problem that chopper-stabilized amplifiers do not offer the necessary differential-input configuration. The circuit shown here is the functional equivalent of the Howland Circuit. Resistor R performs the current monitoring function, and the small drop across it is compensated for by increasing the feed-back resistor in amplifier B by the same amount, so that amplifier B has slightly higher than unity gain. This current pump is capable of extremely accurate control of the current through the load, and is usually limited in accuracy by the resistors involved, rather than by amplifier drift or uncertainty. Its speed of response may be very nearly the same as that of the circuit of III.6, all other things being equal. Since current pumps have essentially "infinite" internal impedance, if a voltage is applied to e_L (e_1 and e_2 at zero), its source will see a very high impedance, and the circuit will behave as a stabilized follower with gain having "push-pull" output at points A & B.



3.7

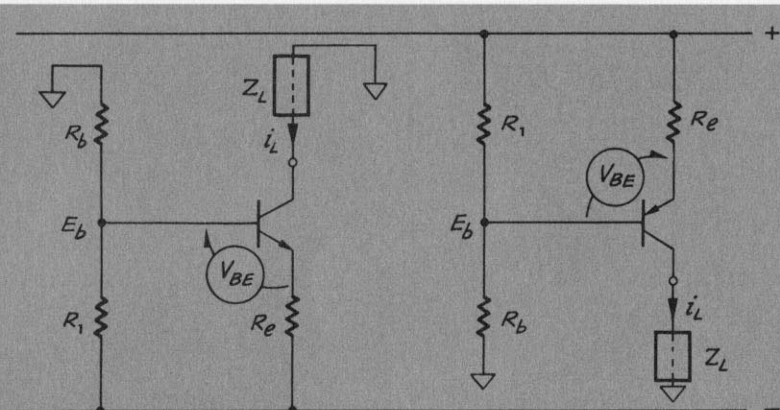
III.7
I.7
I.18
I.34
III.3
to
III.9
III.31
III.38
III.59

III.8 CURRENT PUMP—SINGLE TRANSISTOR. In the interest of objectivity, we feel it only fair to introduce, at this point, a pair of circuits in which some kind of current pump action may be achieved without benefit of an Operational Amplifier. These circuits are not eminently high-performance current pumps, and it frequently turns out that they are used in conjunction with Operational Amplifiers . . . which explains, perhaps, our gallant objectivity! Many applications will be found for these circuits in configurations that require relatively constant current, but in which current variations cause, at the most, second-order errors.

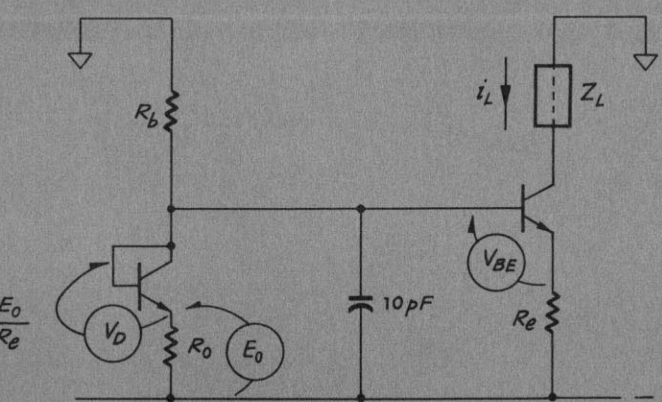
The circuits shown to the right behave identically in a complementary manner: as emitter-followers with the load to be driven connected into the collector-return circuit. The current level is established by setting the base voltage of the transistor, E_b, to an appropriate value—first by the resistance divider shown, or by furnishing an external signal of the required value. If E_b is large compared to the base-emitter drop of the transistor, then the current through R_e is approximately given by E_b/R_e. If the current gain, β, of the transistor is high enough, we may neglect the emitter-base current, and thus state that i_L is also approximately equal to E_b/R_e. The base-emitter voltage drop can be partially compensated if $R_b > R_1$ by adding a diode in series with R_1, thus improving accuracy and providing first-order temperature compensation.

Provided that the available power supply can furnish the load voltage drop and E_b, leaving a volt or two for the emitter-to-collector voltage, and provided that the ratings of the transistor are not exceeded, this circuit, in all its approximate glory, functions as a current source of some utility. It is reasonably independent of the nature of the load—simple, complex, or active—and is certainly economical and straightforward.

It is doubtful whether an order of accuracy of better than 0.05% is worth attempting with this circuit.



3.8 (a)



3.8 (b) FIRST-ORDER TEMPERATURE-COMPENSATED, NPN

III.8
I.45
II.34
II.35
II.43
II.44
III.3
to
III.7
III.9

III.9 CONSTANT-CURRENT REGULATOR. In this circuit, a unipolar derivative of the Howland circuit, a follower with gain is used to add or subtract a correcting current to or from (larger) unregulated current derived directly from the power supply through R_3. It must be appreciated that this circuit depends for its accuracy on the stability of the +15 volt supply with respect to signal ground . . . as well as the accuracy and stability of R_3,* and the accuracy and stability of the other resistors in the feedback networks.

Let us deal first with the diodes connected across the amplifier input terminals. They function merely as input signal clamps, preventing excessive input voltage in the event that the load is disconnected or suddenly becomes a very high impedance.

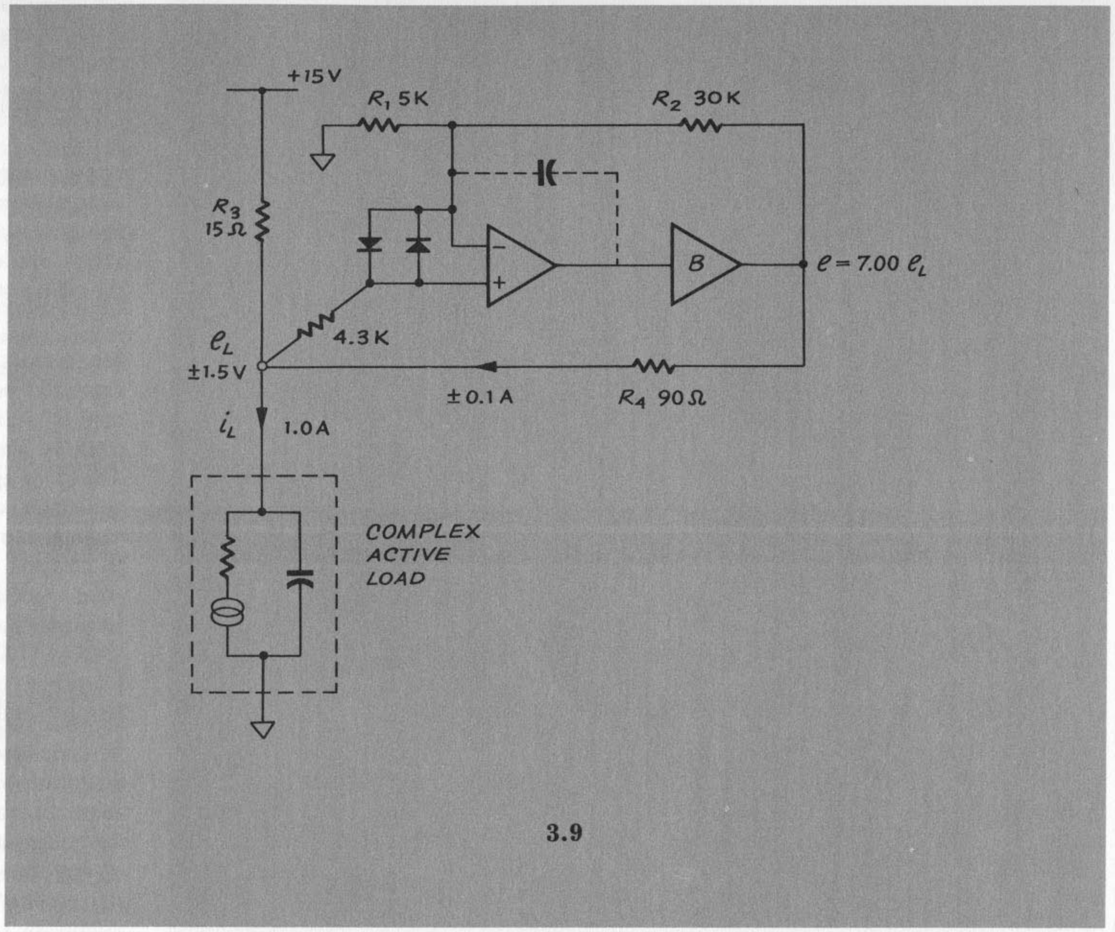
If $R_2 = aR_1$, and $R_4 = aR_3$, then it can be shown that in the ideal case the load current is simply given by $i_L = 15/R_3$.

The sources of error, beyond the resistor contributions and the power supply stability mentioned above are as follows: Input current to the amplifier; Voltage offset and drift; and CME and finite-gain error factors.

The booster current ratings should be selected to match the requirements of the load, and the output voltage capability of the booster must be such that it can furnish that current through R_4, when R_4 is made equal to aR_3.

The 4.3 kΩ resistor shown in series with the positive input terminal in our example, is so proportioned (with respect to R_1 and R_2) as to reduce the effect of current offset. It also prevents excessive current in the clamping diodes.

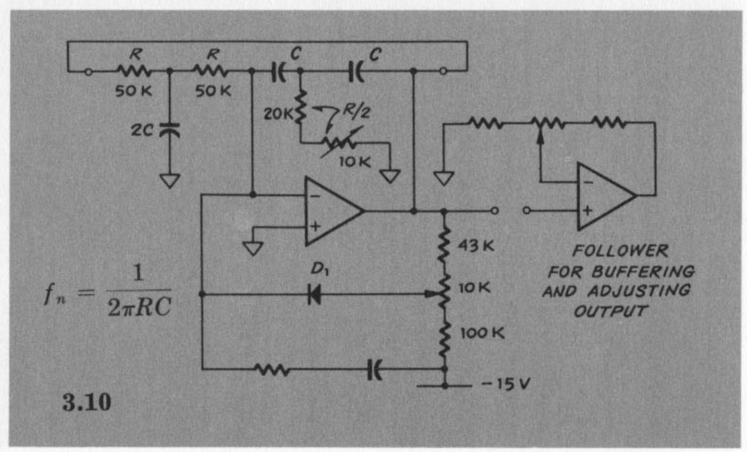
*Establishing high accuracy and holding high stability in a 15Ω 15-watt resistor may not be as easy as it sounds. Extreme derating helps, of course.



3.9

III.9
I.34
I.46
II.1
II.2
II.3
II.4
II.5
III.3
III.4
III.5
III.6
III.7
III.8

III. 10 SINE WAVE OSCILLATOR—TWIN-TEE. This circuit is almost identical to the double integrator of II.14. At some frequency, when $R/2$ is exactly 25 kΩ, the phase shift around the feedback loop is exactly 180°, so that the total phase shift, including the amplifier's sign inversion, is 360°, and the circuit is zero damped. A small increase in $R/2$ will encourage oscillatory buildup. (Note that the accuracy of frequency and zero-ness of damping are encouraged if f is well within the amplifier bandwidth—see I.17 and I.41.) Without the damping network at the lower right of the diagram, sustained oscillations would build up between the extremes of saturation of amplifier—producing a distorted output. Beyond a selected threshold, a damping path through the biased diode, D_1, maintains a stable, sinusoidal oscillation at a selected, adjustable value of peak-to-peak output. Adjusting the threshold will adjust the output. The voltage at which the damping circuit limits the amplitude is a direct function of power supply voltage, and is temperature-dependent because the diode forward voltage drop is. The series RC circuit from summing point to power supply can be proportioned for rapid startup yet minimal interference with normal operation. The output impedance of this circuit is, unfortunately, high near f_n; therefore, this circuit would benefit from the addition of a follower or inside-the-loop booster. The follower choice allows adjustable output, as shown.



3.10

III.10
I.7
I.17
I.18
I.26
I.40
to
I.43
II.12
III.11
to
III.14