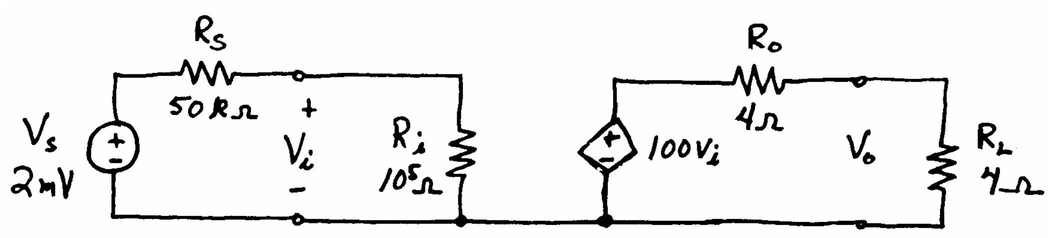
CHAPTER 11

P11.4* The equivalent circuit is:



$$A_v = \frac{V_o}{V_i} = A_{voc}\frac{R_L}{R_o + R_L} = 100\frac{4}{4+4} = 50$$

$$A_{vs} = \frac{V_o}{V_s} = A_{voc}\frac{R_i}{R_i + R_s}\frac{R_L}{R_o + R_L}$$

$$= 100\frac{100\times 10^3}{100\times 10^3 + 50\times 10^3}\frac{4}{4+4}$$

$$= 33.33$$

$$A_i = A_v\frac{R_i}{R_L} = 50\frac{10^5}{4} = 1.25\times 10^6$$

$$G = A_v A_i = 62.5\times 10^6$$

P11.13* With the switch closed, we have:

$$V_o = 100 \text{ mV} = A_v\frac{R_L}{R_o + R_L}V_s \qquad (1)$$

With the switch open, we have:

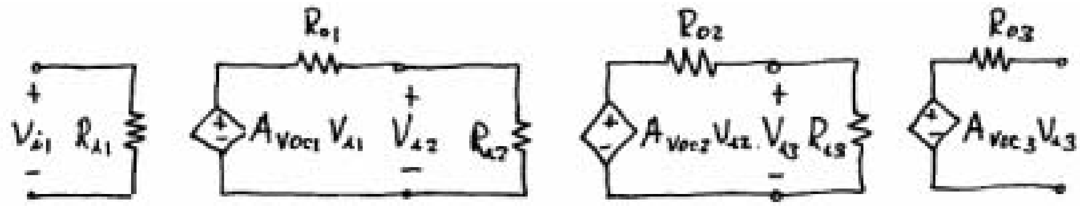
$$V_o = 50 \text{ mV} = A_v\frac{R_{in}}{R_{in} + 10^6}\frac{R_L}{R_o + R_L}V_s \qquad (2)$$

Dividing the respective sides of Equation (2) by those of Equation (1), we obtain:

$$0.5 = \frac{R_{in}}{R_{in} + 10^6}$$

Solving, we that that $R_{in} = 1\ \text{M}\Omega$.

P11.17* The equivalent circuit for the cascade is:



We have:

$$R_i = R_{i1} = 2\ \text{k}\Omega$$

$$R_o = R_{o3} = 3\ \text{k}\Omega$$

$$A_{oc} = A_{oc1}A_{oc2}A_{oc3}\frac{R_{i2}}{R_{i2}+R_{o1}}\frac{R_{i3}}{R_{i3}+R_{o2}}$$

$$= 100(200)(300)\frac{4000}{4000+1000}\frac{6000}{6000+2000}$$

$$= 3.6\times 10^{6}$$

P11.22* The two 15-V sources deliver power:

$$P_1 = (15\ \text{V})\times(1\ \text{A}) = 15\ \text{W}$$

$$P_2 = (15\ \text{V})\times(2\ \text{A}) = 30\ \text{W}$$

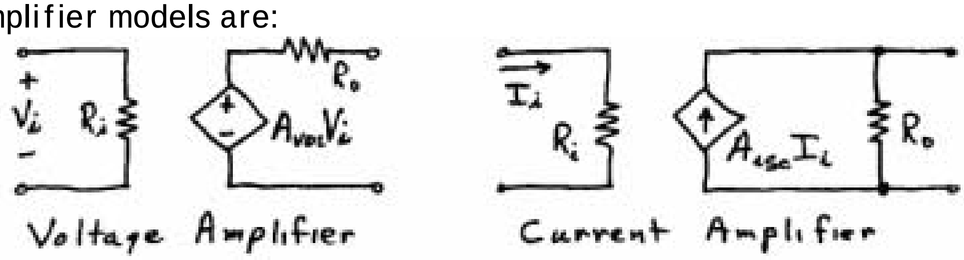
On the other hand, the 5-V source absorbs power:

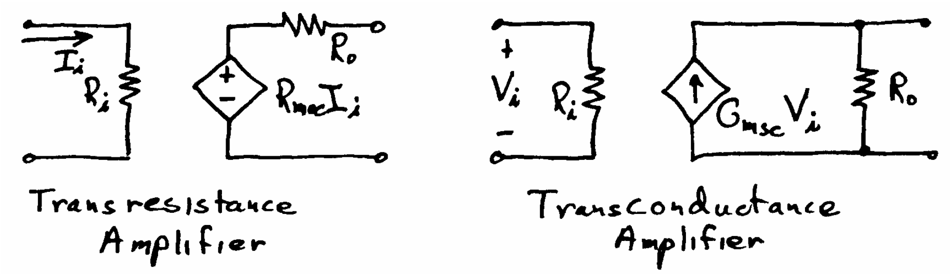
$$P_3 = (5\ \text{V})\times(-1\ \text{A}) = -5\ \text{W}$$

Thus, the new power supplied to the amplifier is:

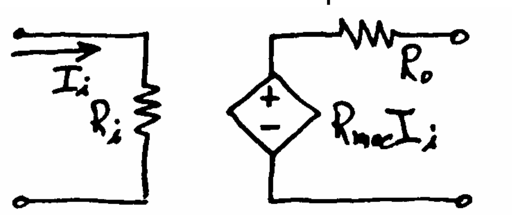
$$P_s = P_1 + P_2 + P_3 = 40\ \text{W}$$

P11.25 The voltage gain A_{oc} is measured under open-circuit conditions.
The current gain A_{isc} is measured under short-circuit conditions.
The transresistance gain R_{moc} is measured under open-circuit conditions.
The transconductance gain G_{msc} is measured under short-circuit conditions.
The amplifier models are:





P11.34 The circuit model for the amplifier is:

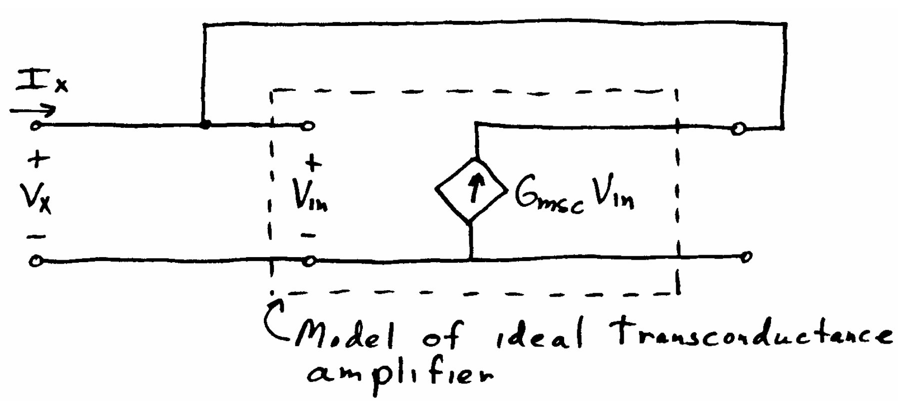


$$A_{voc} = \frac{V_{ooc}}{V_i} = \frac{R_{moc} I_i}{R_i I_i} = \frac{R_{moc}}{R_i} = \frac{200\ \text{k}\Omega}{10\ \text{k}\Omega} = 20$$

$$A_{isc} = \frac{I_{osc}}{I_i} = \frac{R_{moc} I_i / R_o}{I_i} = \frac{R_{moc}}{R_o} = \frac{200\ \text{k}\Omega}{2\ \text{k}\Omega} = 100$$

$$G_{msc} = \frac{I_{osc}}{V_i} = \frac{R_{moc} I_i / R_o}{R_i I_i} = \frac{R_{moc}}{R_i R_o} = \frac{200\ \text{k}\Omega}{(10\ \text{k}\Omega)(2\ \text{k}\Omega)} = 0.01\ \text{S}$$

P11.40 The equivalent circuit is:



$$R_x = \frac{V_x}{I_x} = \frac{V_{\text{in}}}{-G_{msc}V_{in}} = -\frac{1}{G_{msc}} = -10\ \Omega$$

Thus if G_{msc} is positive, the circuit behaves as a negative resistance.

P11.47 To sense the short-circuit current of a sensor, we need an amplifier with very low input resistance (compared to the Thévenin resistance of the sensor). For the load current to be independent of the variable load resistance, we need an amplifier with very high output resistance (compared to the largest load resistance). Thus, we need a nearly ideal current amplifier.

P11.55* We are given

$$v_{\text{in}}(t) = 0.1\cos(2000\pi t) + 0.2\cos(4000\pi t + 30^\circ)$$

and

$$v_o(t) = 10\cos(2000\pi t - 20^\circ) + 15\cos(4000\pi t + 20^\circ)$$

The phasors for the 1000-Hz components are $\mathbf{V}_{\text{in}} = 0.1\angle 0^\circ$ and $\mathbf{V}_o = 10\angle -20^\circ$. Thus the complex gain for the 1000-Hz component is

$$A_v = \frac{\mathbf{V}_o}{\mathbf{V}_{\text{in}}} = \frac{10\angle -20^\circ}{0.1\angle 0^\circ} = 100\angle -20^\circ$$

Similarly, the complex gain for the 2000-Hz component is

$$A_v = \frac{\mathbf{V}_o}{\mathbf{V}_{\text{in}}} = \frac{15\angle 20^\circ}{0.2\angle 30^\circ} = 75\angle -10^\circ$$

P11.58* We are given that the gain of the amplifier as a function of frequency is:

$$A(f) = \frac{1000}{[1 + j(f / f_B)]^2}$$

For $f << f_B$, the gain magnitude is approximately 1000. This is the midband gain. As frequency increases, the gain magnitude decreases. At the half-power frequency f_{hp} the gain magnitude is $1000 / \sqrt{2}$. Thus we can write

$$|A(f_{hp})| = \frac{1000}{\sqrt{2}} = \frac{1000}{1 + (f_{hp} / f_B)^2}$$

Solving, we find $f_{hp} = f_B\sqrt{\sqrt{2} - 1} \cong 0.6436 f_B$

P11.67 (a) Applying the voltage-division principle, we have

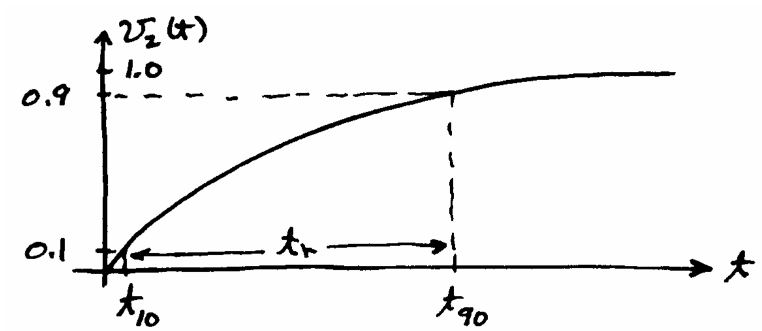
$$A = \frac{\mathbf{V}_2}{\mathbf{V}_1} = \frac{1/(j\omega C)}{R + 1/(j\omega C)} = \frac{1}{1 + j\omega RC}$$

Defining $B = 1/2\pi RC$, we have $A = \dfrac{1}{1 + j(f/B)}$

where B is the half-power bandwidth. The gain magnitude approaches zero as frequency becomes large.

(b) The transient response is:

$$v_2(t) = [1 - \exp(-t/RC)]u(t)$$



We have $0.1 = 1 - \exp(-t_{10}/RC)$ and $0.9 = 1 - \exp(-t_{90}/RC)$

Solving, we obtain:

$$t_{10} = -RC\ln(0.9)$$

$$t_{90} = -RC\ln(0.1)$$

$$t_r = t_{90} - t_{10} = RC\ln(9)$$

(c) Combining the results of parts (a) and (b), we obtain

$$t_r = \frac{\ln(9)}{2\pi B} \cong \frac{0.35}{B}$$

This is the basis for the rule-of-thumb given in Equation 11.11.

P11.68 (a) Applying the voltage-division principle, we have

$$A = \frac{\mathbf{V}_2}{\mathbf{V}_1} = \frac{R}{R + 1/(j\omega C)} = \frac{1}{1 + 1/j\omega RC}$$

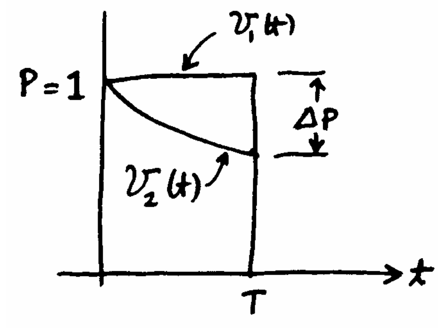
Defining, $f_L = 1/2\pi RC$, we have

$$A = \frac{1}{1 - j(f_L/f)}$$

(b) $f_L = 1/2\pi RC$ is the half-power frequency. The gain magnitude approaches unity as frequency becomes large. At dc, the gain is zero.

(c) Solving for transient response, we obtain:

$$v_2(t) = \exp(-t/RC) \text{ for } 0 < t < T$$



Thus, we have:

$$\Delta P = 1 - \exp(-T/RC) \tag{1}$$

$$\text{Percentage tilt} = \frac{\Delta P}{P} \times 100\% = [1 - \exp(-T/RC)] \times 100\%$$

However,

$$\exp(-T/RC) = 1 - \left(\frac{T}{RC}\right) - \left(\frac{T}{RC}\right)^2 - \left(\frac{T}{RC}\right)^3 - \cdots$$

For $T << RC$

$$\exp(-T/RC) = 1 - \left(\frac{T}{RC}\right)$$

Using this to substitute in Equation (1), we have

$$\text{Percentage tilt} = \frac{T}{RC} \times 100\%$$

(d) Combining the results of parts (b) and (c), we obtain

$$\text{Percentage tilt} \cong 2\pi f_L T \times 100\%$$

This result is precise only for a first-order circuit for which $RC >> T$ but it provides a useful rule-of-thumb for other high pass filters provided that the percentage tilt is small (i.e., less than 10%).

P11.74 Substituting the input into the equation for the transfer characteristic, we obtain:

$$v_o(t) = 20\cos(200\pi t) + 2.4\cos^2(200\pi t) + 3.2\cos^3(200\pi t)$$

Applying the trigonometric identities suggested in the problem statement, we obtain

$$v_o(t) = 1.2 + 22.4\cos(200\pi t) + 1.2\cos(400\pi t) + 0.8\cos^3(600\pi t)$$

Thus the amplitude of the desired output term is V_1 = 22.4, the amplitude of the second harmonic is V_2 = 1.2 V, and the amplitude of the third harmonic is V_3 = 0.8 V. The amplitudes of higher order terms are zero. Then using Equations 11.17, 11.18, and 11.19, we have

$$D_2 = \frac{V_2}{V_1} = \frac{1.2}{22.4} = 0.05357$$
$$D_3 = \frac{V_3}{V_1} = \frac{0.8}{22.4} = 0.03571$$
$$D_4 = \frac{V_4}{V_1} = \frac{0}{10} = 0$$
$$D = \sqrt{D_2^2 + D_3^2 + D_4^4 + \cdots} = \sqrt{(0.05357)^2 + (0.03571)^2}$$
$$= 0.06438$$
$$= 6.438\%$$

P11.78* With the input terminals tied together, the differential signal is zero and the common-mode signal is:

$$v_{icm} = (v_{i1} + v_{i2})/2 = 10 \text{ mV rms}$$

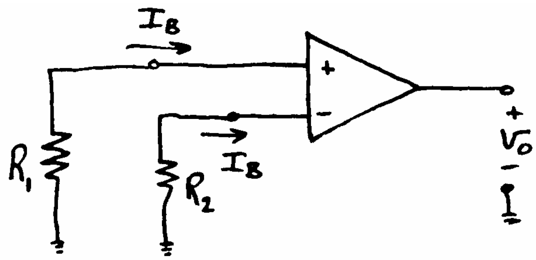
The common-mode gain is:

$$A_{cm} = v_o / v_{icm} = 20/10 = 2$$

The common-mode rejection ratio is:

$$CMRR = 20\log\frac{|A_d|}{|A_{cm}|} = 20\log\frac{500}{2} = 47.96 \text{ dB}$$

P11.82* The circuit is:



The bias currents are equal. However, because the resistances may not be equal, a differential input voltage is possible. In the extreme case, one resistor could have a value of 1050 Ω and the other could have a value of 950 Ω. Then the differential input voltage is:

$$\begin{aligned} v_{id} &= R_1 I_{bias} - R_2 I_{bias} \\ &= (R_1 - R_2) I_{bias} \\ &= \pm 100\ \Omega \times 100\ \text{nA} \\ &= \pm 10\ \mu\text{V} \end{aligned}$$

Thus, the extreme values of the output voltage are:

$$v_o = A_d v_{id} = \pm 5\ \text{mV}$$

If the resistors are exactly equal, then the output voltage is zero.