13.1

$\boxed{13.1}$ $1/\beta_\infty = 1 + 20/100 = 1.2$ V/V $= 1.584$ dB, regardless of v_I.

$v_I = 10$ V: $r_e = 26\ \Omega$, $r_o = 100$ kΩ, $R_1 = (10\ k\Omega) \| (100\ k\Omega) \| (2\ M\Omega) = 9.05$ kΩ, $R_2 = 26 + 4.3$ k$\Omega \simeq 4.33$ kΩ; $1/\beta_0 = R_2/R_1 = 0.478$ V/V $= -6.40$ dB; $f_z = 1/(2\pi \times 9.05\times10^{-3} \times 120\times10^{-12}) = 147$ kHz; $f_p = 367$ kHz.

$v_I = 10$ mV: $r_e = 26$ kΩ, $r_o = 100$ MΩ, $R_1 \simeq 10$ kΩ, $R_2 = 30.3$ kΩ; $1/\beta_0 = 9.7$ dB; $f_z = 133$ kHz; $f_p = 52.5$ kHz.

$v_I = 1$ mV: $r_e = 260$ kΩ, $R_2 = 264.3$ kΩ; $1/\beta_0 = 28.4$ dB; $f_z = 133$ kHz; $f_p = 6.0$ kHz.

$\boxed{13.2}$ Worst case is $v_I = 10$ V, when $R_1 = 9.05$ kΩ, $R_2 = 4.33$ kΩ, $f_z = 147$ kHz, $f_p = 367$ kHz;

$$T = a\beta \simeq \frac{10^6}{jf}\,\frac{9.05}{4.33}\,\frac{1 + jf/367\times10^3}{1 + jf/147\times10^3}$$

Using trial and error we find that $|T| = 1$ for $f = f_x = 893$ kHz, where $\measuredangle T = -103°$. Thus, $\phi_m = -103 + 180 = 77°$.

$\boxed{13.3}$ $-2.303(1 + R_2/R_1)0.026 = -2 \Rightarrow$ $R_2 = 32.4 R_1$. Use $R_1 = 1$ kΩ Q81 type, $R_2 = 32.4$ kΩ, 1%. For optimum logging range, (13.2) leave $R = 10.0\ \text{k}\Omega$. Then,

$$\frac{i_I}{I_i} = \frac{v_I/R}{V_{REF}/R_r} = \frac{v_I}{1\text{V}} \Rightarrow \frac{R}{R_r} V_{ref} = 1\text{V} \Rightarrow$$

$R_r = RV_{REF}/(1\text{V}) = 10^4 \times 6.95 = 69.8\ \text{k}\Omega$, 1%.

All remaining components remain the same.

$\boxed{13.4}$ (a) $i_{C1}/i_{C2} = (I_{S1}/I_{S2})e^{v_{B1}/V_T}$. Using $e^x = 10^{x/2.303}$ and letting $i_{C2} = i_O$, we get $i_O = I_o \times 10^{v_I/V_i}$, where

$$I_o = \frac{I_{S2}}{I_{S1}} i_{C1} = \frac{I_{S2}}{I_{S1}}\frac{V_{REF}}{R_r}, \text{ and}$$

$$V_i = -2.303\frac{R_1+R_2}{R_1}V_T.$$

(b) $I_o = 10\ \mu\text{A} \Rightarrow R_r = 6.95/0.01 = 698\ \text{k}\Omega$, 1%. Using $10^x = 2^{3.322x}$ and imposing

$$-1\text{V} = -\frac{2.303}{3.322}\left(1 + \frac{R_2}{R_1}\right)26\ \text{mV}$$ gives $R_2 = 54.49R_1$. Use $R_1 = 1\text{k}\Omega$ Q81 type, and $R_2 = 54.6\ \text{k}\Omega$, 1%.

(c) Assuming the -15V supply is well regulated and clean, connect a resistor R_3 between the base of Q_1 and -15 V to downshift v_{B1}. In part (a) v_{B1} varied over the range $-90.11\text{ mV} \le v_{B1} \le +90.11\text{ mV}$. To ensure the same range of variability, we now need, for $v_I = 0$,

(13.3)

$$\frac{0-(-0.09011)}{R_2}+\frac{0-(-0.09011)}{R_1}=\frac{-0.09011-(-15)}{R_3}$$

and, for $v_I = +10$ V,

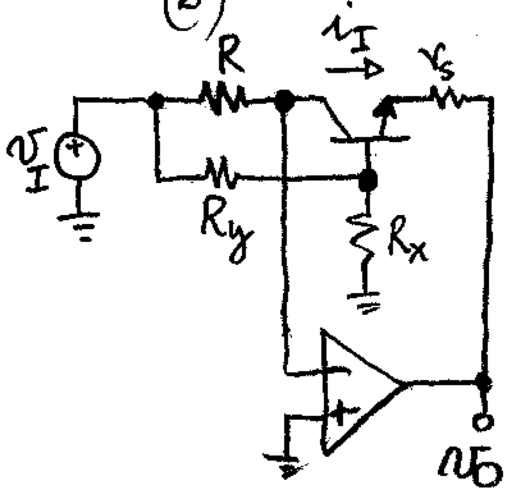
$$\frac{10-0.09011}{R_2}=\frac{0.09011}{R_1}+\frac{0.09011+15}{R_3}$$

letting $Q_1 = 1$ kΩ Q81 type and solving, we get $R_2 = 54.15$ kΩ (use 53.6 kΩ, 1%), and $R_3 = 162.46$ kΩ (use 162 kΩ, 1%).

13.5 (a) By Eq. (13.5) and KVL,

$v_O = -v_{BE} - v_{r_s} = -V_T \ln \frac{v_I}{RI_s} - \frac{r_s}{R} v_I$, or also $v_O = -V_T \ln \frac{i_I}{I_s} - r_s i_I$. For $i_I = 1$ mA, $r_s i_I = 1$ mV; for $i_I = 0.1$ mA, $r_s i_I = 0.1$ mV. The percentage input error p is such that $r_s i_I = (26\,\text{mV}) \ln(1+p)$. Substituting, we find $p(1\,\text{mA}) = 3.92\%$, $p(0.1\,\text{mA}) = 0.385\%$.

(b)



$$v_O = v_B - v_{BE} - r_s i_I$$

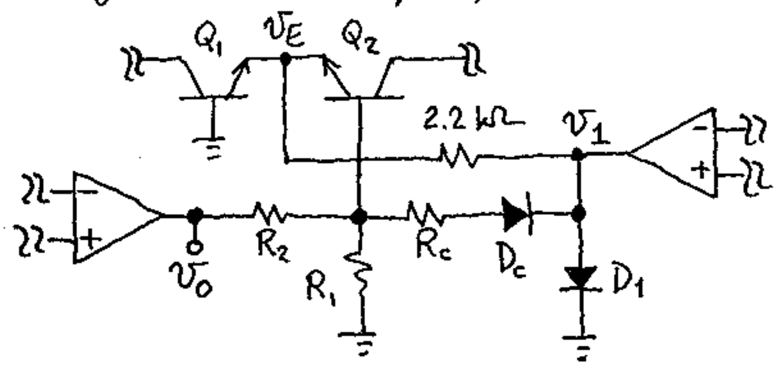
$$= \frac{R_x}{R_x+R_y} v_I - \left[V_T \times \ln \frac{v_I}{RI_s} + \frac{r_s}{R} v_I\right]$$

Imposing $\frac{R_x}{R_x+R_y} = \frac{r_s}{R}$, or $R_y/R_x = R/r_s - 1$

gives $v_O = -V_T \ln (v_I/RI_s)$, regardless of r_s.

(13.4)

13.6 The compensation network for the case of the logarithmic amplifier becomes:



The voltages across R_c and the 2.2-kΩ resistance are

$$v_{R_c} = v_{B2} - v_{D_c} - v_1$$

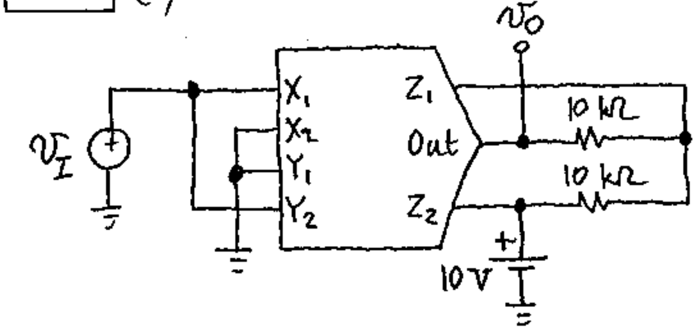
$$v_{2.2\,k\Omega} = v_{B2} - v_{BE2} - v_1$$

As long as $v_{D_c} \cong v_{BE2}$ we have $v_{R_c} \cong v_{2.2\,k\Omega}$, so $i_{R_c} = v_{R_c}/R_c \cong v_{2.2\,k\Omega}/R_c = [(2.2\,k\Omega)/R_c] \times (i_I + I_{REF})$. The current i_{R_c} shifts v_{B2} and, hence, v_E, by the amount $\Delta v_E = \Delta v_{B2} = -(R_1 // R_2) i_{R_c} = -(R_1//R_2)\frac{2.2\,k\Omega}{R_c}(i_I + I_{REF})$; the bulk resistance r_s of Q_1 causes the shift $\Delta v_E = -r_s i_I$. Imposing $-(R_1//R_2)\frac{2.2\,k\Omega}{R_c} i_I = -r_s i_I$ will compensate for the error due to r_s. This requires using $R_c = (R_1//R_2)(2.2\,k\Omega)/r_s$, and also recalibrating R_2 and R_r slightly.

For the circuit shown, use $R_c = (1//15.7) \times 10^3 \times 2.2 \times 10^3 / 0.5 = 4.12$ MΩ.

(13.5)

13.7 (a)



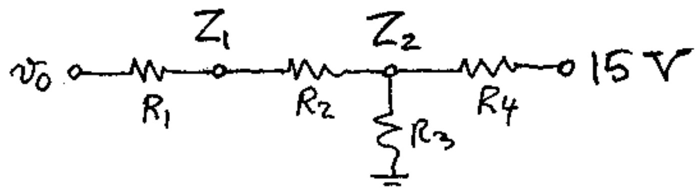
$v_{Z_1} = \frac{1}{2}(v_O + 10\text{ V})$; $v_{Z_2} = 10\text{ V}$; $v_{Z_1} - v_{Z_2} = \frac{1}{2}v_O - 5\text{V}$.

By Eq. (13.21) with $k = 1/10$,

$$\frac{1}{2}v_O - 5 = \frac{1}{10}(10\cos\omega t - 0)(0 - 10\cos\omega t) = -10\cos^2\omega t$$

$$\Rightarrow v_O = 10 - 20\,\frac{1+\cos 2\omega t}{2} = 10\cos 2\omega t \text{ V.}$$

(b) Connect as follows:

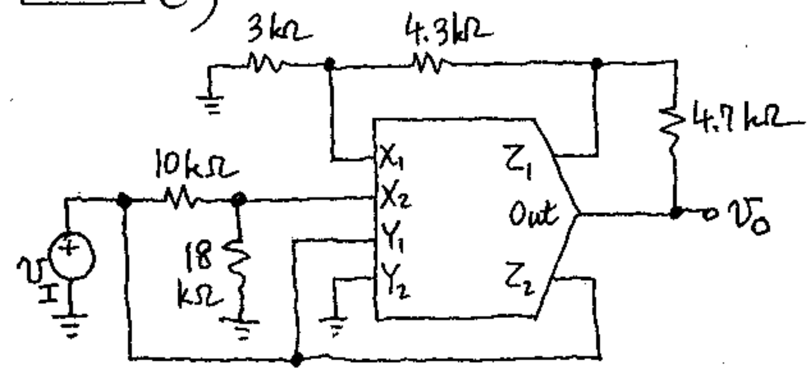


We want $v_{Z_1} - v_{Z_2} = \frac{1}{2}v_O - 5\text{V}$. The Thévenin equivalent of the circuit to the right of R_2 is a source $15R_3/(R_3+R_4)$ with a series resistance $R_3 \| R_4$, so

$$v_{Z_1} - v_{Z_2} = R_2\,\frac{v_O - 15R_3/(R_3+R_4)}{R_1 + R_2 + (R_3\|R_4)} = \frac{1}{2}v_O - 5\text{V}.$$

Use $R_3 = 20.0\text{ k}\Omega$, $R_4 = 10.0\text{ k}\Omega$, $R_2 = 10\text{ k}\Omega$, $R_1 = 3.32\text{ k}\Omega$, all 1%.

13.8 (a)



By Eq. (13.21),

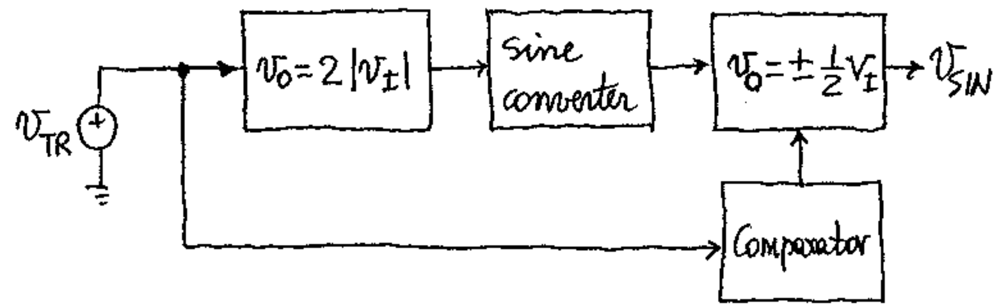
$$10\left(\frac{7.3}{12}v_O - v_I\right) = \left(\frac{3}{12}v_O - \frac{18}{28}v_I\right)\times\left(v_I - 0\right),\ \text{or}$$

$$v_O = \frac{840v_I - 54v_I^2}{511 - 21v_I} \cong 10\sin\left(\frac{v_I}{10}90°\right) = 10\sin 9v_I.$$

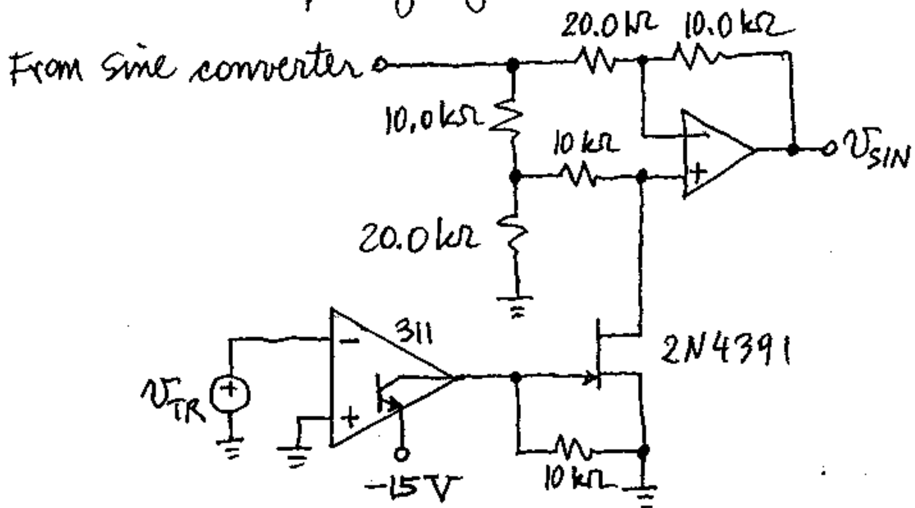
v_I (V)	v_I (°)	v_O (actual)	v_O (ideal)	% error
10	90	9.967	10	−0.330
5	45	7.020	7.071	−0.725
0	0	0	0	0
10/3	30	4.989	5	−0.227
20/3	60	8.625	8.660	−0.400
5/3	15	2.626	2.588	+1.470
25/3	75	9.673	9.659	+0.141

(b) As shown in the accompanying diagram, we use an absolute-value circuit to map the ±5 V range of v_{TR} to the range of 0 to 10 V needed by the sine converter, and then we use a programmable amplifier with a gain of +0.5 V/V when $v_I > 0$, and a gain of −0.5 V/V when $v_I < 0$.

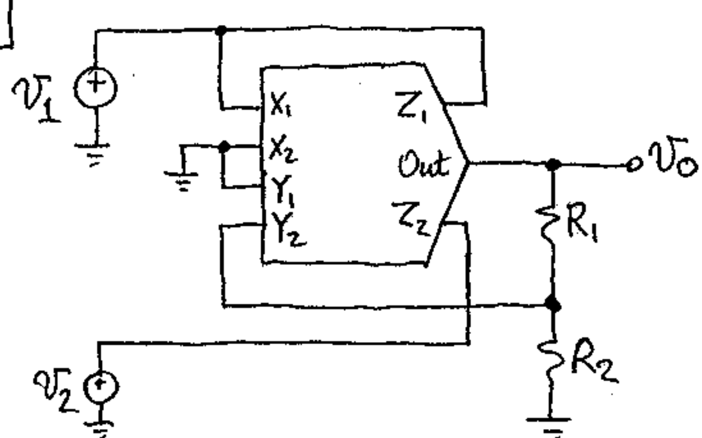
13.7



For the absolute-value circuit use Fig. (9.30) with $R_1 = R_2 = 2R_3 = R_4 = R_5/2 = 10.0\ \text{k}\Omega$. For the sine converter use part (a) above. For the polarity detector and programmable gain amplifier use the accompanying circuit.



13.9

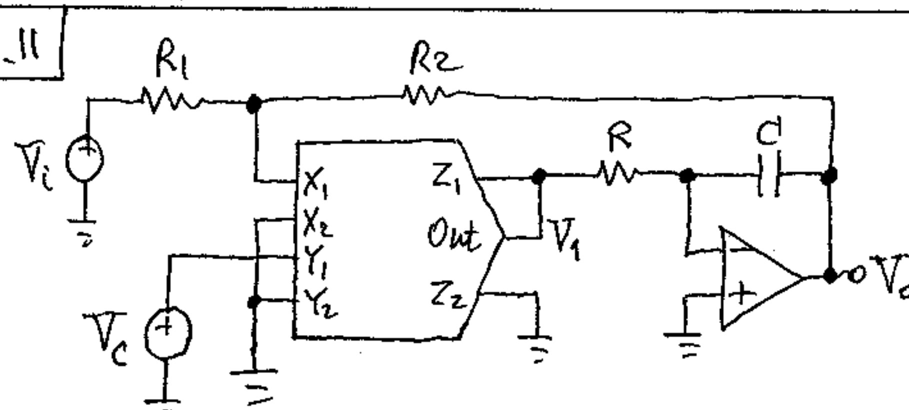


$$v_1 - v_2 = \frac{1}{10}(v_1 - 0)\times\left(0 - \frac{R_2}{R_1+R_2}v_O\right) \Rightarrow v_O = \left(1+\frac{R_1}{R_2}\right)10\,\frac{v_2 - v_1}{v_1}.$$

Use $R_1 = 18\ \text{k}\Omega$, $R_2 = 2\ \text{k}\Omega$.

13.10 $v_{X_1} = v_{Z_2} = 10 - R20/[R+R(1+\delta)] = 10\delta/(2+\delta)$, indicating a nonlinear dependence on δ, as we know. Now, since $v_{X_2} = v_{Y_2} = 0$ and $v_{Y_1} = v_{Z_1} = v_O$, Eq. (13.21) gives: $v_O - v_{X_1} = (1/10)(v_{X_1} - 0)(v_O - 0)$, that is, $v_O = 10X_1/(10 - v_{X_1})$. Substituting the above expression for v_{X_1} yields $v_O = 5\delta$, a linear dependence on δ.

13.11



$$V_o = -[1/(sRC)]V_1, \quad V_1 = \left(\frac{R_2}{R_1+R_2}V_i + \frac{R_1}{R_1+R_2}V_o\right)\frac{V_c}{10}.$$

$$H = \frac{V_o}{V_i} = -\frac{R_2}{R_1}\frac{1}{1+jf/f_0}, \quad f_0 = \frac{V_c}{2\pi(1+R_2/R_1)10RC}.$$

For a dc gain of 20 dB use $R_1 = 10\,k\Omega$, $R_2 = 100\,k\Omega$.

$V_c = 10\,V \Rightarrow f_0 = 10 \times 100 = 1\,kHz = \dfrac{10}{2\pi \times 11 \times 10RC}$.

Use $C = 1$ nF, $R = 14.3\,k\Omega$.

13.12 Let V_1 be the output of g_{m1}, and let I_1 and I_2 be the outputs of g_{m1} and g_{m2}. We then have

13.9

$$V_o = V_i + \frac{1}{sC_1} I_2 = V_i + \frac{g_{m2}}{sC_1}(V_1 - V_o)$$

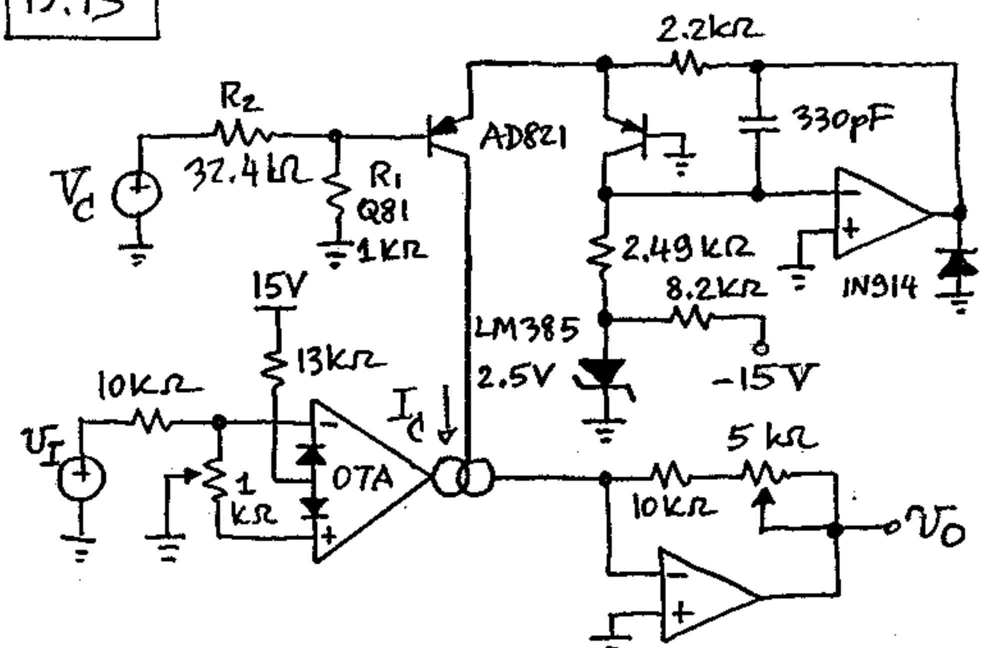
$$V_1 = \frac{1}{sC_2} I_1 = \frac{g_{m1}}{sC_2}(V_i - V_o)$$. Eliminating V_1,

$$\left(1 + s^2 \frac{C_1 C_2}{g_{m1} g_{m2}}\right) V_i = \left(s^2 \frac{C_1 C_2}{g_{m1} g_{m2}} + s\frac{C_2}{g_{m1}} + 1\right) V_o$$

$$\frac{V_o}{V_i} = \frac{1-(\omega/\omega_0)^2}{1-(\omega/\omega_0)^2 + (j\omega/\omega_0)/Q}, \quad \omega_0 = \sqrt{C_1 C_2 / g_{m1} g_{m2}}$$

$$Q = \sqrt{(C_1/C_2)(g_{m1}/g_{m2})}$$. Notch response.

13.13

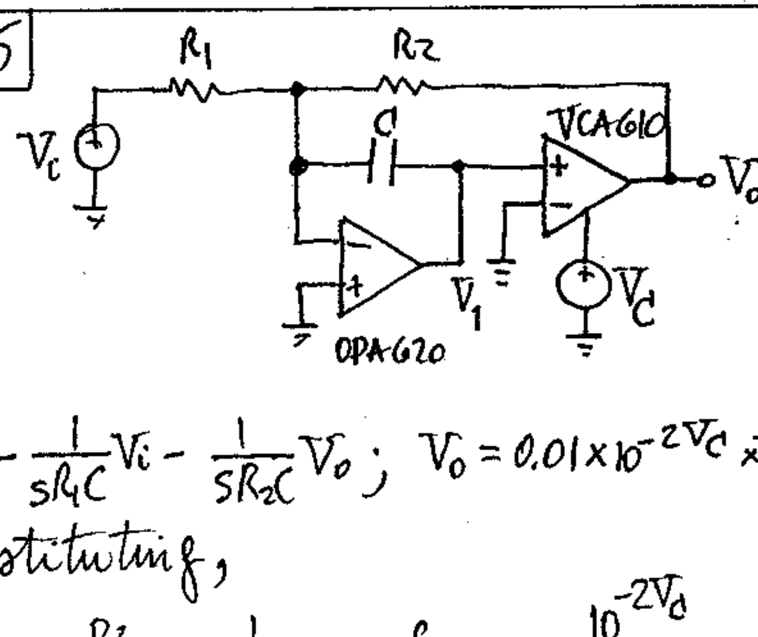


For a sensitivity of 1 oct/V we need $2.303(1+R_2/R_1)0.026 = 2$, that is, $R_2 = 32.4 R_1$. Use $R_1 = 1\ \mathrm{k\Omega}$ Q81 type, and $R_2 = 32.4\ \mathrm{k\Omega}$, 1%. To calibrate, set $v_I = 0$, $V_C = 0$, and adjust the 1-kΩ pot for $v_O = 0$. Then, set v_I to a 1-kHz, 5-V amplitude sinewave, and adjust the 5-kΩ pot so that v_O has a 5-V amplitude.

(13.10)

13.14 Use the circuit of Fig. 13.19 with the exponential V-I converter of Fig. 13.17. For a sensitivity of 1 octave/volt, use $R_2 = 32.4\ k\Omega$, 1% in Fig. 13.17. Since for $V_c = 0$ the V-I converter gives $I_C = 1\ mA$, which is then split in half by the AD821 pair of Fig. 13.19, we must have, by Eq. (13.30), $20\ kHz = 0.5\ mA/(2 \times \pi \times 12.2 C)$ or $C = 326\ pF$. Use two 330-pF capacitors, and make the 2.49-kΩ resistor of Fig. 13.17 adjustable until $f_0 = 20$ kHz with $V_c = 0$.

13.15



$$V_1 = -\frac{1}{sR_1C}V_i - \frac{1}{sR_2C}V_o\ ; \quad V_o = 0.01 \times 10^{-2V_c} \times V_1.$$

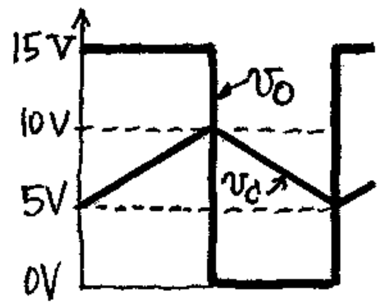
Substituting,

$$\frac{V_o}{V_i} = -\frac{R_2}{R_1}\frac{1}{1+jf/f_0}, \quad f_0 = \frac{10^{-2V_c}}{200\pi R_2 C}.$$

Unity dc gain $\Rightarrow R_1 = R_2$. For $V_c = -2V$ we want $f_0 = 1\ MHz$, or $10^6 = 10^4/(200\pi R_2 C)$.
Let $C = 10\ nF$. Then $R_1 = R_2 = 1.58\ k\Omega$, 1%.

(13.11)

13.16 (a) $v_O = 5$ V ⇒ DISCH pin floating ⇒ OTA inputs are $v_P = 2.5$V and $v_N = 1.36$ V ⇒ OTA's input stage saturates such that OTA sources the control current I_C to the capacitor.



$v_O = 0$ V ⇒ DISCH pin shorted to ground ⇒ $v_P \cong 0$ V and $v_N = 1.36$ V ⇒ OTA now sinks I_C from capacitor. Thus, v_C is a triangular wave alternating between $1/3\ V_{CC}$ and $2/3\ V_{CC}$. $C\Delta V = I\Delta t \Rightarrow C(10-5) = I_C T/2 \Rightarrow f_0 = 1/T = I_C/10C$.

(b) $C = I_C/10f_0 = 10^{-3}/(10\times10^5) = 1$ nF. Generate I_C with the exponential V-I converter of Fig. 13.17, where $R_2 = 32.4$ kΩ in order to achieve the desired sensitivity. To calibrate, make the 2.49 kΩ resistor in Fig. 13.17 adjustable, and set it so that $f_0 = 100$ kHz with $V_C = 0$ V.

13.17 With $F(s) = 1$ we have $v_e(s) = K_d K_a \theta_d(s)$ $= (10^4/\pi 10^4)\theta_d(s) = \theta_d(s)/\pi$, or $\theta_d(s) = \pi v_e(s)$.

(a) $\theta_d(t) = \pi \times 0.2\,[1 - e^{-t/(100\,\mu s)}]u(t)$
$= 36°\,[1 - e^{-t/(100\,\mu s)}]u(t)$.

(b) $\theta_d(t) = 19.33° \cos(2\pi 2500 t - 57.52°)$.

13.18 (a) With $R_2 = 0$ we get $F(s) = 1/(1 + s/\omega_p)$, $\omega_p = 1/R_1C = 100$ rad/s. Then, $T(j\omega) = 1/[(j\omega/10^4)(1 + j\omega/100)]$. Using trial and error, we find that $|T| = 1$ for $\omega_x = 997.5$ rad/s, where $\measuredangle T = -174.3°$, so $\phi_m = 5.7°$, an inadequate margin.

(b) For $\phi_m = 45°$ impose $\omega_p = K_v = 10^4$ rad/s. The actual crossover frequency and phase margin are then $\omega_x = 7{,}860$ rad/s, and $\phi_m = 51.8°$.

13.19 $R_1C = 1/100$, $R_2C = 1/10^3$. Let $C = 0.1\,\mu F$. Then, $R_1 = 100\,k\Omega$, and $R_2 = 10\,k\Omega$. $T(j\omega) = [1 + j\omega/10^3]/[(j\omega/10^4)(j\omega/10^2)]$. $|T| = 1$ for $\omega_x = 1272$ rad/s, and $\phi_m = 51.8°$.

13.20 Passive filter:

$$H(s)=\frac{K_v F(s)}{s+K_v F(s)}=\frac{K_v(1+s/\omega_z)/(1+s/\omega_p)}{s+K_v(1+s/\omega_z)/(1+s/\omega_p)}$$

$$=\frac{K_v+K_v s/\omega_z}{s+s^2/\omega_p+K_v+K_v s/\omega_z}=\frac{1+s/\omega_z}{s^2/K_v\omega_p+s(1/K_v+1/\omega_z)+1}$$

$$=\frac{N(s)}{(s/\omega_n)^2+2\zeta(s/\omega_n)+1},\quad \omega_n=\sqrt{K_v\omega_p}\ ;$$

$$2\zeta s/\omega_n=s(1/K_v+1/\omega_z)\Rightarrow \zeta=\frac{\omega_n}{2\omega_z}\left(1+\frac{\omega_z}{K_v}\right);$$

$$N(s)=1+\frac{2\zeta s}{\omega_n}-\frac{s}{K_v}=1+\frac{s}{\omega_n}\left(2\zeta-\frac{\omega_n}{K_v}\right).$$

Active filter:

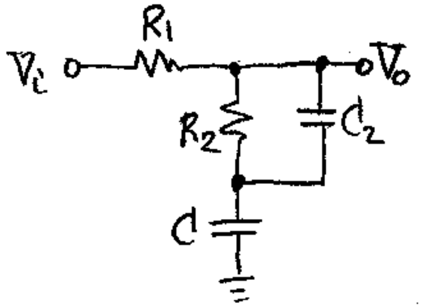
$$H(s)=\frac{k_v(1+s/\omega_z)/(s/\omega_p)}{s+K_v(1+s/\omega_z)/(s/\omega_p)}=\frac{K_v+K_v s/\omega_z}{s^2/\omega_p+K_v+s K_v/\omega_z}$$

$$=\frac{1+s/\omega_z}{s^2/K_v\omega_p+s/\omega_z+1}=\frac{1+2\zeta(s/\omega_n)}{(s/\omega_p)^2+2\zeta(s/\omega_n)+1},$$

$$\omega_n=\sqrt{K_v\omega_p},\quad \zeta=\frac{\omega_n}{2\omega_z}.$$

13.21 $K_v=0.2\times1\times2\pi10^6=2\pi10^5\ s^{-1}$; $\omega_n\cong\omega_0/100=2\pi10^4$ rad/s. $R_1C=1/\omega_p=K_v/\omega_n^2=1/2\pi10^3$ s; $R_2C=1/\omega_z=2\zeta/\omega_n=1/Q\omega_n=1/\pi10^4$ s. Let $C=$ 10 nF. Then, $R_1=15.8$ kΩ and $R_2=3.16$ kΩ.

13.22 Let $Z_2=R_2\|(1/sC_2)=R_2/(1+sR_2C_2)$. Then,



$$F(s)=\frac{V_o}{V_i}=\frac{Z_2+1/sC}{R_1+Z_2+1/sC}=\frac{1+sR_2(C+C_2)}{s^2R_1R_2CC_2+s(R_1C+R_2C+R_2C_2)+1}.$$

(13.14)

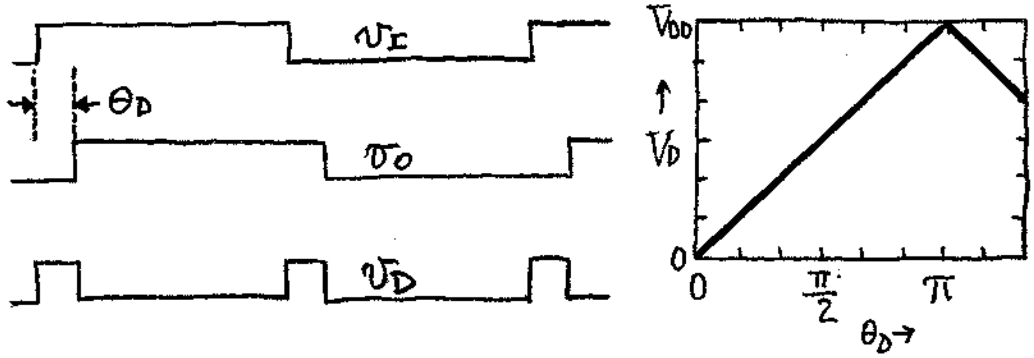
Substituting $C_2 = C_1/10$, $R_2 = 1/366C_1$, and $R_1 = 1/25C_1 - 1/366C_1$ we get

$$F(s) = \frac{1+s/332.7}{1+s^2/98208+s/24.83}.$$ Consequently,

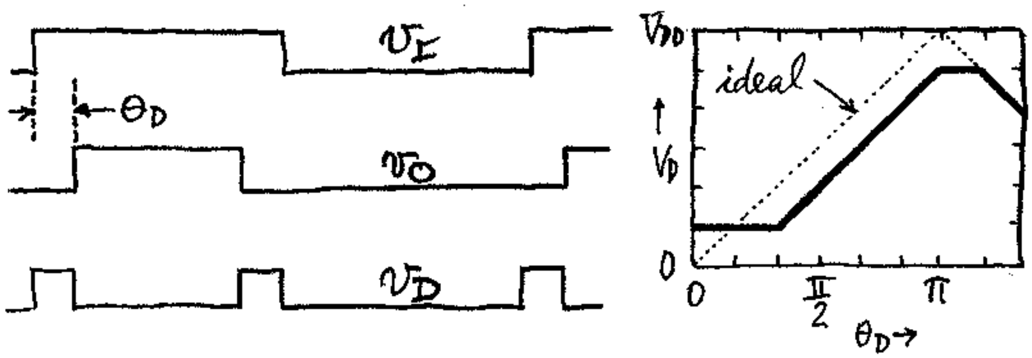
$$T(j\omega) = K_v \frac{F(j\omega)}{j\omega} = \frac{10^4(1+j\omega/332.7)}{j\omega\left[1-\omega^2/98208+j\omega/24.83\right]}.$$

By trial and error we find $\omega_x = 797$ rad/s, and $\phi_m = 57.7°$; ω_x increases by 40 rad/s, and ϕ_m decreases by 8.3°.

$\boxed{13.23}$ Case $D_I = D_O = 1/2$:

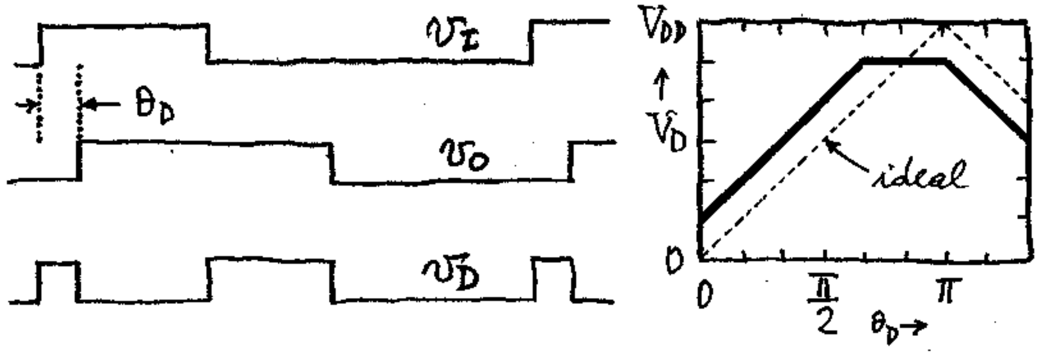


Case $D_I = 1/2$, $D_O = 1/3$:

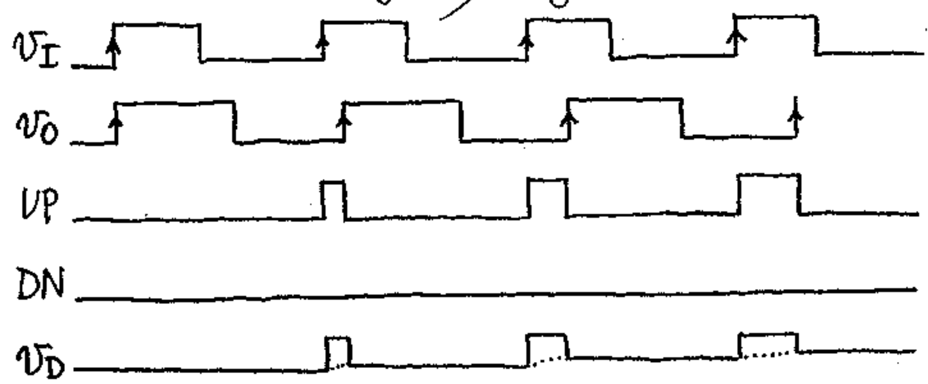


(13.15)

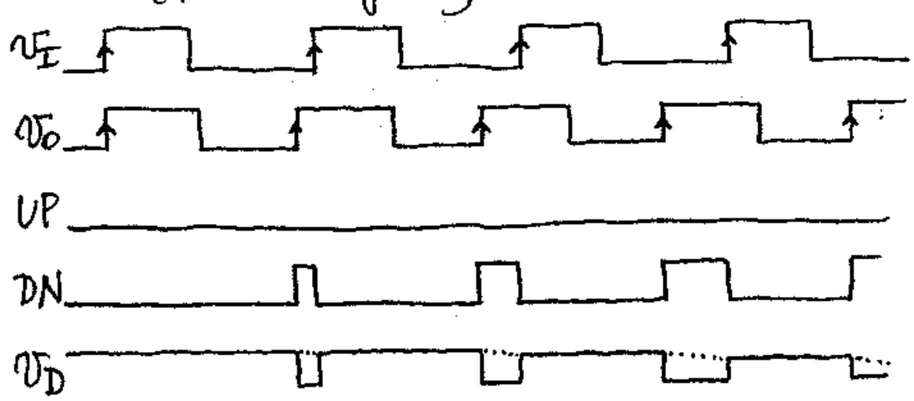
Case $D_I = 1/3,\ D_O = 1/2$:



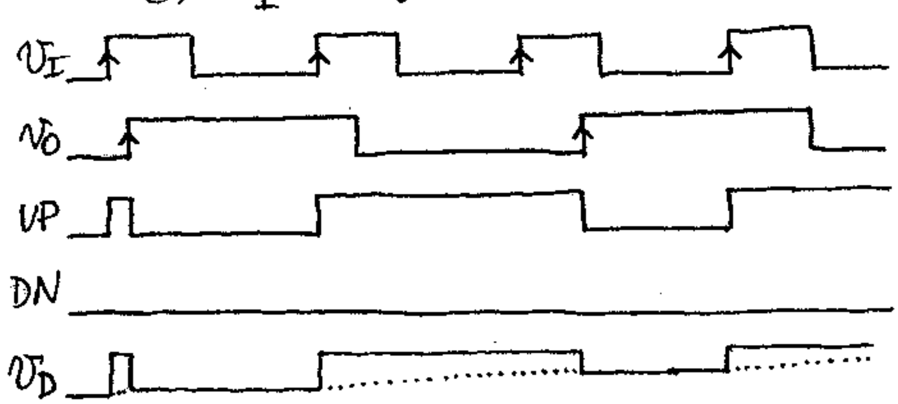
13.24 (a) ω_I slightly higher than ω_O:



(b) ω_I slightly lower than ω_O:

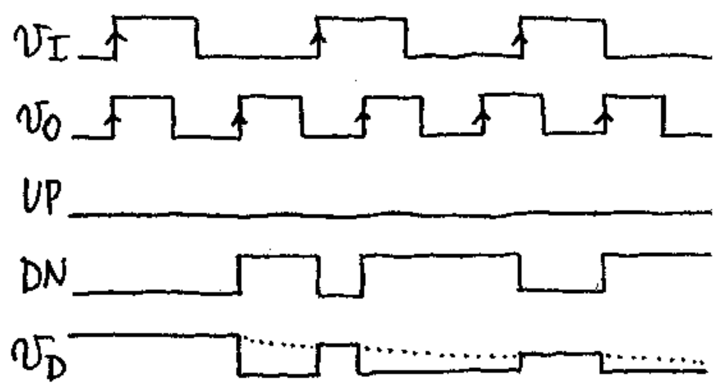


(c) $\omega_I \gg \omega_O$:



(13.16)

(d) $\omega_I \ll \omega_O$



13.25 (a) $K_d = 5/\pi$; $K_a = 1$; $K_o = 2\pi \times 5\times 10^6 = \pi 10^7$;
$K_v = (5/\pi)\times 1 \times \pi 10^7 = 5\times 10^7\ s^{-1}$.

$\omega_n = \pi 10^4$ rad/s. Using Eqs. (13.46),

$\omega_p = \omega_n^2/K_v = (\pi 10^4)^2/(5\times 10^7) = 19.74$ rad/s

$\Rightarrow (R_1+R_2)C = 1/19.74$

$\zeta = 1/2Q = 1/(2\times 0.5) = 1 = \pi 10^4\left(\frac{1}{2\omega_z} + \frac{1}{2\times 5\times 10^7}\right)$

$\cong \pi 10^4/(2\omega_z) \Rightarrow \omega_z = \frac{\pi}{2}10^4 = 15{,}708$ rad/s

$\Rightarrow R_2C = 1/15{,}708$. Pick $C = 100$ nF. Then,

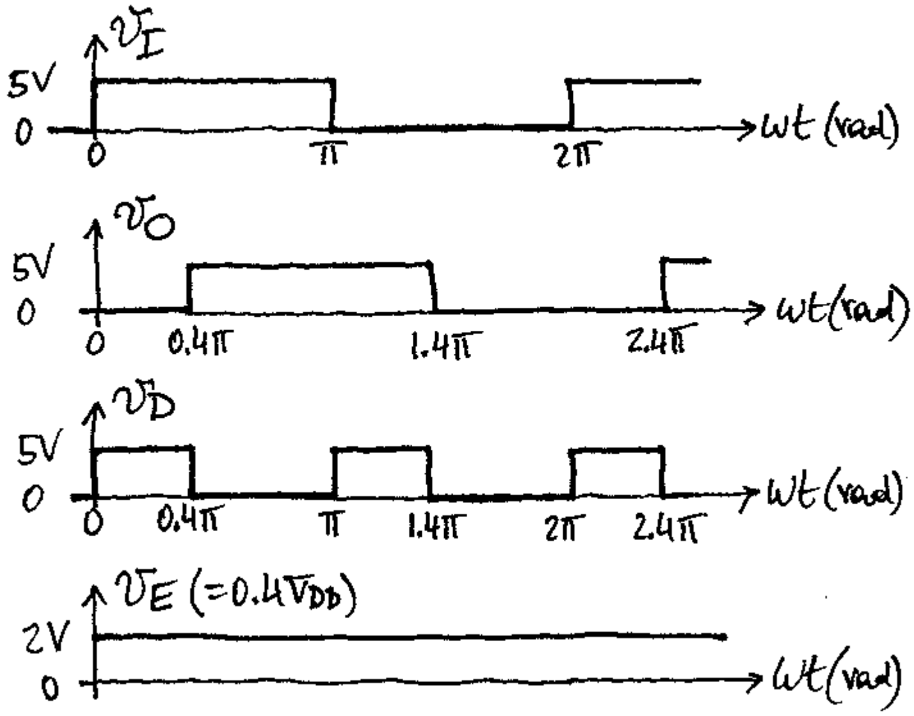
$R_2 = 637\ \Omega$ (Use $634\ \Omega$, 1%), and $R_1 = 507\ k\Omega$ (use $511\ k\Omega$, 1%). Use $C_2 = 10$ nF.

(b) $f_O = 10^7 + (v_E - 2.5)5\times 10^6$; $f_O = 7.5$ MHz $\Rightarrow v_E = 2$ V.

$\theta_I - \theta_O = v_E/K_d = 2/(5/\pi) = 1.257$ radians.

The waveforms are as follows:

13.17



13.26 $\omega_p = 553$ rad/s, $\omega_z = 22.5\times10^3$ rad/s, $\zeta = 1/\sqrt{2}$, $K_o = 1.122\times10^6$ rad/s, $K_v = 1.786\times10^6$ s^{-1}, $\omega_n = \sqrt{\omega_p K_v} = 3.143\times10^4$ rad/s; $2\zeta - \omega_n/K_v = 1.397$. Substituting into Eq. (13.46a),

$$H(s) = \frac{s/(2.25\times10^4)+1}{[s/(3.143\times10^4)]^2 + s/(2.22\times10^4)+1}$$

Computing at $s = j\omega_m = j2\pi\times10^3$ we get $H(j2\pi10^3) = 1.0373\angle{-0.82^\circ}$. Then,

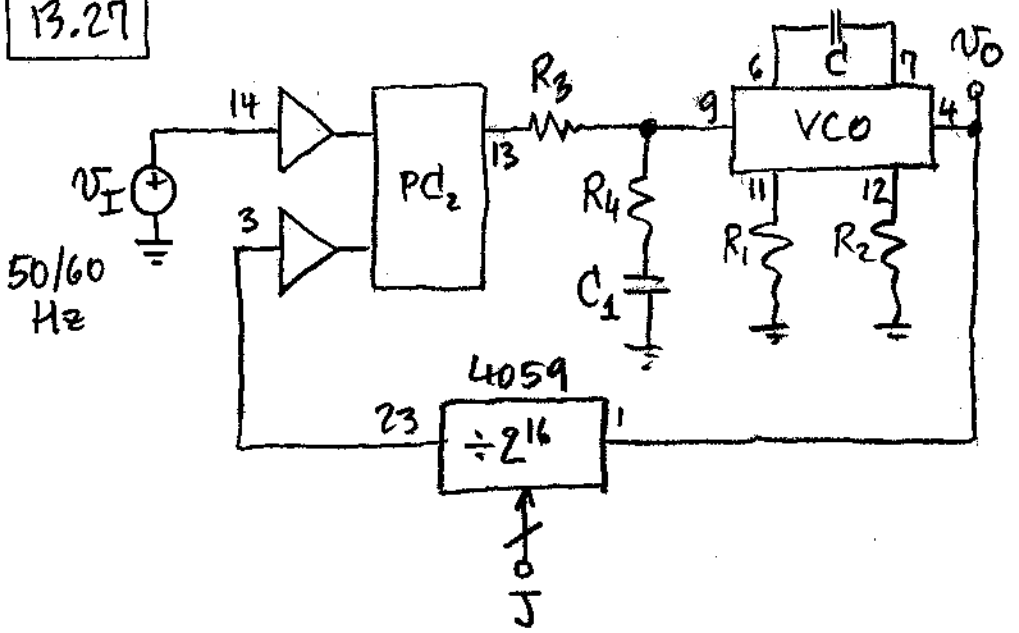
$$v_e(t) = \frac{|\omega_i|}{K_o} 1.0373\cos(2\pi10^3 t - 0.82^\circ).$$

Substituting $|\omega_i| = 2\pi\times10\times10^3$ rad/s and $K_o = 1.122\times10^6$ rad/s we finally get

$$v_e(t) = (58.09\ \text{mV})\cos(2\pi10^3 t - 0.82^\circ).$$

(13.18)

13.27



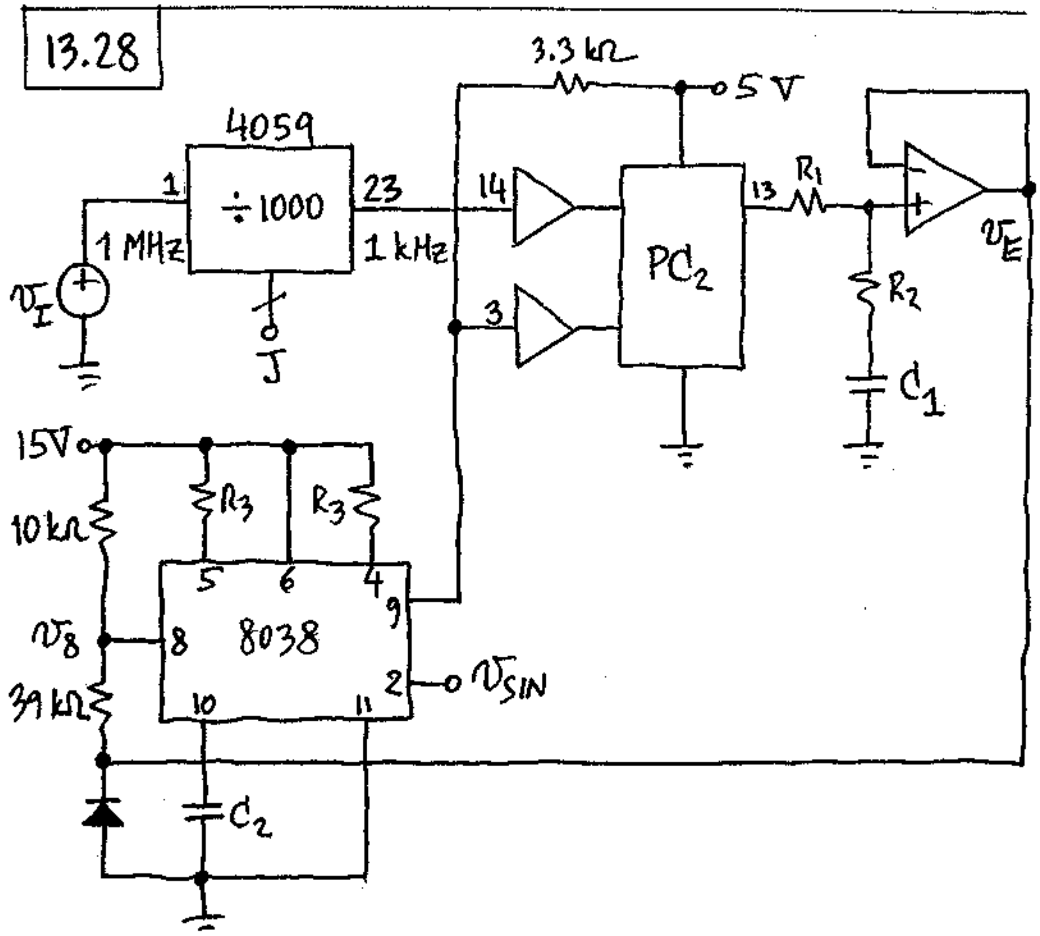
Design for $f_I = 55$ Hz. Then, $f_0 = 55 \times 2^{16} = 3.60$ MHz. Arbitrarily impose $2f_R = 2$MHz, so that $K_o = 2\pi \times 2 \times 10^6/(3.9-1.1) = 4.49$ (Mrad/s)/V. To meet these specifications, the PLL Design Program suggests using $R_1 = 22$ kΩ, $R_2 = 39$ kΩ, and $C = 100$ pF.

Using PC_2, we get $K_d = 5/4\pi$. Then, $K_V = (5/4\pi) \times 4.49 \times 10^6/2^{16} = 27.2\ s^{-1}$.

Choose $\omega_n = \omega_I/20 = 2\pi 55/20 = 17$ rad/s, and $\zeta = 1/\sqrt{2}$. Then, $\omega_p = \omega_n^2/K_V = 11$ rad/s, and $\omega_z = 21.5$ rad/s. Let $C_1 = 1\ \mu$F. Then, $R_4 = 1/\omega_z C_1 \cong 47$ kΩ, and $R_3 = 1/\omega_p C - R_4 = 91$ kΩ.

(13.19)

13.28



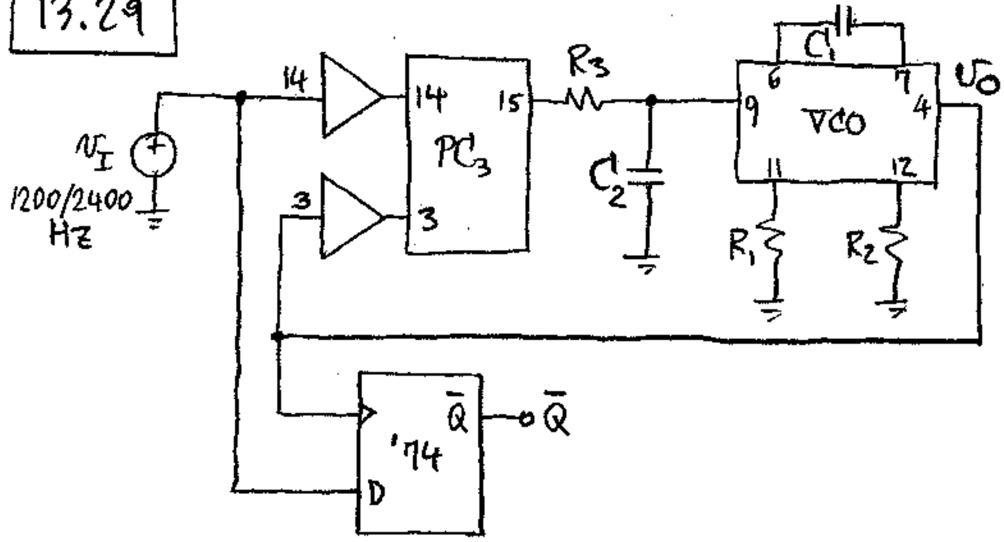
Use a 4059 counter configured as a modulo-1000 counter to divide the 1-MHz input down to a 1-kHz reference signal for the phase comparator. The other input to the phase comparator is obtained from the 8038's open-collector output using the 3.3-kΩ pullup resistor. Buffer the filter with a FET-input op amp to avoid loading. $0\text{ V} \le v_E \le 5\text{ V} \Rightarrow 12\text{ V} \le v_8 \le 13\text{ V}$. Let $f_0 = 1$ kHz for $v_E = 2.5\text{V}$, i.e. for $v_8 = 12.5\text{ V}$. Impose $i_{R_3} = 0.1$ mA, so $R_3 = (15-12.5)/0.1 = 24.9$ kΩ. Using $C_2\Delta V = I\Delta t$ with $\Delta V = 2\times$

13.20

$(15/3) = 10$ V, $I = 0.1$ mA, $\Delta t = 1/f_0 = 1$ ms gives $C_2 = 10$ nF. The purpose of the diode is to protect the 8038 against possible excessively negative swings of v_E at power turn-on.

To design the filter, observe that $K_d = 5/4\pi$ V/rad. Moreover, $\Delta v_E = 5$ V $\Rightarrow \Delta v_{R_3} = 1$ V $\Rightarrow \Delta i_{R_3} = 1/(25\ \text{k}\Omega) = 40\ \mu\text{A} \Rightarrow \Delta f_0 = (40\ \mu\text{A})/(10\ \text{V} \times 10\times10^{-9}\ \text{F}) = 400$ Hz. So, $K_o = \Delta f_0/\Delta v_E = 400/5 = 80$ Hz/V $= 160\pi$ (rad/s)/V. Consequently, $K_v = (5/4\pi)160\pi = 200\ \text{s}^{-1}$. Arbitrarily impose $\omega_n = 2\pi$ rad/s, so $\omega_p = \omega_n^2/K_v = 0.2$ rad/s; and $\zeta = 1/\sqrt{2}$, so $\omega_z = (2\zeta/\omega_n - 1/K_v)^{-1} = 4.54$ rad/s. Let $C = 3.3\ \mu$F. Then, $R_2 = 68$ kΩ, $R_1 = 1.5$ MΩ.

13.29

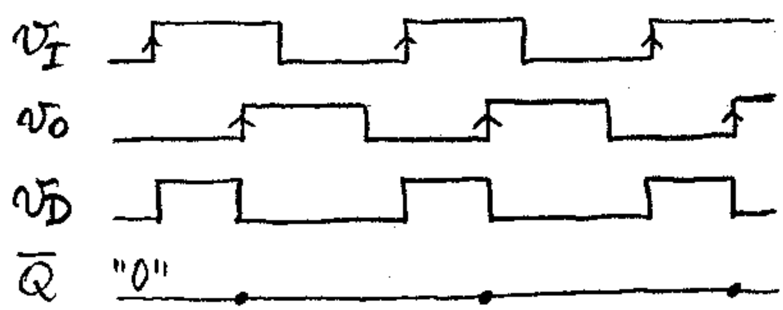


PLL Design program $\Rightarrow C_1 = 33$ nF, $R_1 = 91$ kΩ, $R_2 = 1.3$ MΩ. $R_3C_2 = 1/2\pi f_H \Rightarrow C_2 = 33$ nF, $R_3 = 2$ kΩ.

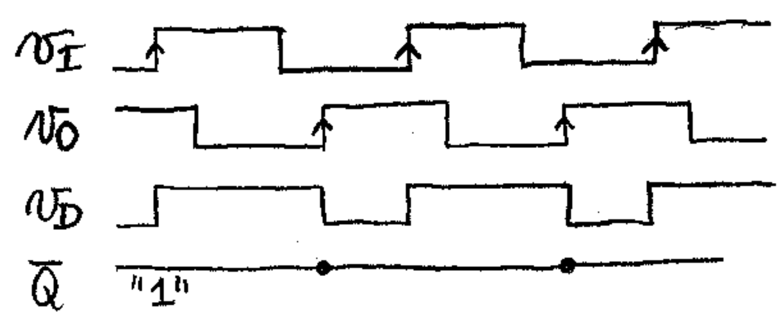
(13.21)

With reference to Fig. 13.26 we can write $f_0 = Av_E + B$; substituting $f_0(v_E = 1.1) = 1800-1000$ and $f_0(v_E = 3.9) = 1800+1000$, we get $f_0 = (5000 v_E + 100)/7$. The control voltages required to make the VCO oscillate at $f_0 = 1200$ Hz and $f_0 = 2400$ Hz are, respectively, $V_{EL} = 83/50 = 1.66$ V, and $V_{EH} = 83/25 = 3.32$ V. According to Fig. 13.29b, the corresponding phase differences are $\theta_{DL} \cong 120°$ and $\theta_{DH} \cong 240° = -120°$.

Case $f_I = 1200$ Hz ($V_E = 1.66$ V):



Case $f_I = 2400$ Hz ($V_E = 3.32$ V):



The use of PC_3 simplifies the D-flip-flop timing considerably.