Exercise 3.16

Because $v_D > 0.1 \text{ V}$, we have $i_D = i_s \exp(v_D/nV_T)$. Solving for $i_s$, we have $i_s = \frac{i_D}{\exp(v_D/nV_T)} = \frac{0.1 \times 10^{-3}}{\exp(0.6/0.026)} = 9.50 \times 10^{-15} \text{ A}$. Then for $v_D = 0.65$, we have

$$i_D = i_s \exp(v_D/nV_T) = 9.50 \times 10^{-15} \exp(0.65/0.026) = 0.684 \text{ mA}$$

and for $v_D = 0.7$ we have

$$i_D = i_s \exp(v_D/nV_T) = 9.50 \times 10^{-15} \exp(0.70/0.026) = 4.68 \text{ mA}$$

Exercise 3.20

$$C_j = \frac{C_{j0}}{[1 - (V_{DQ}/\phi_0)]^m} = \frac{5 \text{ pF}}{[1 - (V_{DQ}/0.8)]^{0.5}}$$

(a) $C_j = \frac{5 \text{ pF}}{[1 - (-5/0.8)]^{0.5}} = 1.86 \text{ pF}$

(b) $C_j = \frac{5 \text{ pF}}{[1 - (-50/0.8)]^{0.5}} = 0.627 \text{ pF}$

Problem 3.42

Any value of $R$ from about 100 $\Omega$ to 100 k$\Omega$ is suitable.
\[ v_{\text{test}} = R_1(i_b + r_{\pi}i_b/R_2) + r_\pi i_b \]
\[ i_{\text{test}} = \beta i_b + i_b + r_{\pi}i_b/R_2 \]
\[ R_o = \frac{v_{\text{test}}}{i_{\text{test}}} = \frac{r_\pi + R_1(1 + r_{\pi}/R_2)}{1 + \beta + r_{\pi}/R_2} = 126 \ \Omega \]

This circuit is sometimes used as a voltage reference (similar to a Zener diode regulator).

**Problem 4.64**

When the transistor is in the active region we have:

\[ V_o = V_{CC} - R_C \beta \frac{V_{in} - 0.7}{R_B} = 14.1 - 3V_{in} \]
(d) In part (b) the circuit acts as a linear amplifier.

Problem 4.65

\[ +12V \]
\[ I_{RC} \rightarrow R_C = 2.2k \Omega \rightarrow I_B \rightarrow V_0 \rightarrow R_B \rightarrow \]
\[ I_{RC} = 2.2k \Omega \]
\[ V_0 = 6V \]

For \( V_0 = 6 \) V we have \( I_{RC} = (12 - 6)/R_C = 2.73 \) mA and \( I_B = (6 - 0.7)/(22 \) k\( \Omega ) = 0.241 \) mA. Thus the maximum fanout is the largest integer that does not exceed \( I_{RC}/I_B = 11.33 \). Thus the maximum fanout is 11.

Problem 5.3

(a) Saturation because we have \( v_{GS} \geq V_{to} \) and \( v_{DS} \geq v_{GS} - V_{to} \).

(b) Triode because we have \( v_{DS} < v_{GS} - V_{to} \) and \( v_{GS} \geq V_{to} \).

(c) Cutoff because we have \( v_{GS} \leq V_{to} \).
Problem 6.12

- $V_{IL}$ is the highest input voltage guaranteed to be accepted as logic 0.
- $V_{IH}$ is the lowest input voltage guaranteed to be accepted as logic 1.
- $V_{OL}$ is the highest logic-0 output voltage produced (provided that the input voltages are higher than $V_{IH}$ or lower than $V_{IL}$).
- $V_{OH}$ is the lowest logic-1 output voltage produced (provided that the input voltages are lower than $V_{IL}$ or higher than $V_{IH}$).
- $N_M = V_{OH} - V_{IH}$
- $N_M = V_{IL} - V_{OL}$
- $I_{OH}$ is the current that the output is capable of sourcing when the output is high.
- $I_{OL}$ is the maximum current that the output can sink when the gate output is in the low state.
- $I_{IL}$ is the worst-case (maximum magnitude) input current, provided that the input voltage is in the acceptable logic-0 input range.
- $I_{IH}$ is the worst-case input current for a high input.

Problem 6.16

$$P_{dynamic} = fC_L (V_{SS})^2 = (400 \times 10^6) (100 \times 10^{-15})^2 = 0.36 \text{ mW}$$

Problem 6.20

$$N_M = V_{OH} - V_{IH} = 4.5 - 3 = 1.5 \text{ V}$$
$$N_M = V_{IL} - V_{OL} = 1.5 - 1 = 0.5 \text{ V}$$
Problem 6.23

Because the switch is open for \( V_I = 2 \) V we conclude that \( V_{IL} = 2 \) V. Similarly because the switch is closed for \( V_I = 3.5 \) V we conclude that \( V_{IH} = 3.5 \) V. Given that \( I_{OL} = -1 \) mA we can compute the largest output voltage in the low state by solving the circuit with the switch closed as shown below.

Writing a current equation for the circuit with the switch closed we have

\[
\frac{V_{OL}}{250} + \frac{V_{OL} - 5}{1000} = 1 \text{ mA}
\]
Problem 6.26

(a) With the switch closed, we have

\[ V_{\text{OL}} = 5 \cdot \frac{250}{1000 + 250} = 1 \text{ V} \]

With the switch open, the output voltage is \( V_{\text{OH}} = 5 \text{ V} \).

(b) The figure shows the waveform for the high-to-low transition.

The output voltage is given by

\[ V_0(t) = 1 + 4 \exp(-t/\tau) \]

where \( \tau \) is the time constant of the circuit. The time constant is the product of the capacitance and the Thévenin resistance of the gate with the switch closed. The Thévenin resistance is the parallel combination of the 1 kΩ and 250 Ω resistances which is 200 Ω. Thus we have \( \tau = 200 \cdot 0.4 = 0.4 \text{ ns} \). \( t_{\text{PLH}} \) is the time required for the output voltage to make half of the transition (at which time we have \( V_0 = 3 \text{ V} \)). Thus we can write:

\[ 3 = 1 + 4 \exp(-t_{\text{PLH}}/\tau) \]

Solving we find \( t_{\text{PLH}} = \tau \ln(2) = 0.277 \text{ ns} \).

Equation 6.23 in the book applies for the low-to-high transition with \( R_D = 1 \text{ kΩ} \). Thus we have

\[ t_{\text{PLH}} = -R_D \ln(0.5) = 0.6931R_D C = 1.386 \text{ ns} \]

Finally we have
\[ t_{PD} = \frac{1}{2}(t_{PHL} + t_{PLH}) = 0.832 \text{ ns} \]

(c) The static power dissipation with the output high is zero. With the output low the current taken from the power supply in steady state is \((5 \text{ V})/(1000 + 250) = 4 \text{ mA}\). Thus the power dissipation in the low output state is \(5 \text{ V} \times 4 \text{ mA} = 20 \text{ mW}\).
Problem 6.38

To improve (reduce) the switching intervals of a resistor-pull-up NMOS inverter we should (a) decrease $R_L$; (b) increase $W$; (c) decrease $L$; and (d) increase $V_{DD}$.

Problem 6.39

Using Equations 6.22 and 6.23, we have

$$t_r = 2.20R_D C = 44 \text{ ns}$$

$$t_{PLH} = 0.6931R_D C = 13.9 \text{ ns}$$

The simulation is stored in the file named P6_39. A plot of the transient response is shown below. With the cursor we can verify that $t_{PLH} = 13.9 \text{ ns}$ and determine that $t_{PHL} = 1.3 \text{ ns}$.

Problem 6.48

$$K_n = \left( \frac{W}{L} \right)_n \times \frac{K_P n}{2} = 75 \mu A/V^2$$

$$K_p = \left( \frac{W}{L} \right)_p \times \frac{K_P p}{2} = 75 \mu A/V^2$$
Problem 6.38

To improve (reduce) the switching intervals of a resistor-pull-up NMOS inverter we should (a) decrease $R_D$; (b) increase $W$; (c) decrease $L$; and (d) increase $V_{DD}$.

Problem 6.39

Using Equations 6.22 and 5.23, we have

$$t_x = 2.20R_D C = 44 \text{ ns}$$

$$t_{PLH} = 0.6931R_D C = 13.9 \text{ ns}$$

The simulation is stored in the file named P6.39. A plot of the transient response is shown below. With the cursor we can verify that $t_{PLH} = 13.9 \text{ ns}$ and determine that $t_{PHL} = 1.3 \text{ ns}$.

Problem 6.48

$$K_n = \left( \frac{W}{L} \right)_n \times \frac{KP_n}{2} = 75 \ \mu A/V^2$$

$$K_p = \left( \frac{W}{L} \right)_p \times \frac{KP_p}{2} = 75 \ \mu A/V^2$$
Because the inverter is designed with \( K_p = K_n \) the transfer characteristic is symmetrical and \( V_O = V_{DD}/2 \) when \( V_I = V_{DD}/2 \). For all cases the transistors operate in the saturation region. For \( V_I = V_{DD}/2 \) we have

\[
I_{DD} = i_{Dn} = K_n (V_{GSn} - V_{ton})^2 (1 + \lambda V_{DSn})
\]

\[
= K_n (V_{DD}/2 - V_{ton})^2 (1 + \lambda V_{DD}/2)
\]

\[
= 75 \times 10^{-6} (V_{DD}/2 - 1)^2
\]

<table>
<thead>
<tr>
<th>( V_{DD} ) (V)</th>
<th>( I_{DD} ) (( \mu )A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>18.75</td>
</tr>
<tr>
<td>5</td>
<td>168.8</td>
</tr>
<tr>
<td>10</td>
<td>1200</td>
</tr>
</tbody>
</table>

For \( V_I = 0 \), we have \( I_{DD} = 0 \) for all values of \( V_{DD} \).

**Problem 6.49**

In this case the PMOS delivers current to the short circuit. The PMOS has \( V_{GS} = -V_{DD} \), \( V_{DS} = -V_{DD} \) and \( K_p = 75 \mu \)A/V\(^2 \). Thus the current is

\[
i_{short} = i_{Dp} = K_p (V_{GS} - V_{top})^2 = K_p (-V_{DD} + 1)^2
\]

<table>
<thead>
<tr>
<th>( V_{DD} ) (V)</th>
<th>( I_{DD} ) (mA)</th>
<th>( P ) (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>0.3</td>
<td>0.9</td>
</tr>
<tr>
<td>5</td>
<td>1.2</td>
<td>6.0</td>
</tr>
<tr>
<td>10</td>
<td>6.08</td>
<td>60.8</td>
</tr>
</tbody>
</table>
The small-signal equivalent circuit for the CMOS inverter is:

Because the transistors are identical except for polarity, the \( g_m \) and \( r_d \) values are equal. From the circuit, we have

\[
\frac{dV_o}{dV_i} = \frac{V_o}{V_i} = -g_m r_d
\]

For the NMOS transistor we have

\[
i_D = K(v_{GS} - V_{to})^2 (1 + \lambda v_{DS})
\]

\[
g_m = \left. \frac{\delta i_D}{\delta v_{GS}} \right|_{Q-point} = 2K(v_{GS} - V_{to}) (1 + \lambda v_{DS}) \bigg|_{Q-point}
\]

\[
= 2K(V_{DD}/2 - V_{to}) (1 + \lambda V_{DD}/2)
\]

\[
1/r_d = \left. \frac{\delta i_D}{\delta v_{DS}} \right|_{Q-point} = \lambda K(v_{GS} - V_{to})^2 \bigg|_{Q-point}
\]

\[
= \lambda K(V_{DD}/2 - V_{to})^2
\]

\[
\frac{dV_o}{dV_i} = -g_m r_d = - \frac{2K(V_{DD}/2 - V_{to}) (1 + \lambda V_{DD}/2)}{\lambda K(V_{DD}/2 - V_{to})^2}
\]

\[
= \frac{-2 (1 + \lambda V_{DD}/2)}{\lambda (V_{DD}/2 - V_{to})}
\]
Problem 6.57

From Equations 6.29 and 6.30 we see that the switching times are inversely proportional to W/L. Thus we should increase the W/L ratios of the transistors by a factor of 1.25. (In practice increasing W/L may also increase C_L and an even larger factor may be needed. This can be determined by trial and err using SPICE.)

Problem 6.58

We can compute the switching times using Equations 6.29 and 6.30 which are:

\[
t_{PHL} = \frac{C_L V_{DD}}{W L n \left(K_P (V_{DD} - V_{ON})^2 \right)} \quad t_{PLH} = \frac{C_L V_{DD}}{W L p \left(K_P (V_{DD} - |V_{TOP}|)^2 \right)}
\]

<table>
<thead>
<tr>
<th>t_{PHL} (ns)</th>
<th>t_{PLH} (ns)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(a) 4.17</td>
<td>4.17</td>
</tr>
<tr>
<td>(b) 4.17</td>
<td>0.417</td>
</tr>
<tr>
<td>(c) 0.417</td>
<td>4.17</td>
</tr>
</tbody>
</table>

Problem 8.9

(a) Refer to Figure P8.9 in the book. Writing a current equation at the output node, we obtain

\[sC(V_o - A_v V_{in}) + (V_o - V_{in})/R = 0\]

Solving for the gain we obtain:

\[A(s) = \frac{V_o}{V_{in}} = \frac{1 - RC_s}{1 + RC_s}\]

(b) We have a pole at \(s = -1/(RC)\) and a zero at \(s = 1/(RC)\).

(c) \(A(f) = \frac{1 - j(f/f_b)}{1 + j(f/f_b)}\) in which \(f_b = 1/(2\pi RC)\).

The magnitude and phase plots are shown on the next page.
Problem 9.5

For $A = 1000$, we have $A_f = A/(1 + A\beta) = 9.9$.

$$x_o(t) = A_f x_s = 9.9\cos(\omega t)$$

$$x_f(t) = \beta x_o = 0.99\cos(\omega t)$$

$$x_i(t) = x_s - \beta x_o \approx 0.01\cos(\omega t)$$

For $A = 10^4$, we have $A_f = A/(1 + A\beta) = 9.99$.

$$x_o(t) = A_f x_s = 9.99\cos(\omega t)$$

$$x_f(t) = \beta x_o = 0.999\cos(\omega t)$$

$$x_i(t) = x_s - \beta x_o \approx 0.001\cos(\omega t)$$

As $A$ approaches $\infty$, $x_i(t)$ approaches zero.
Problem 9.8

(a) \( A_f = A/(1 - A\beta) \) In this amplifier, the closed-loop gain magnitude is less than the open-loop gain magnitude if \((1 - A\beta)\) is larger than unity. This happens if either \(A\) or \(\beta\) (but not both) assume negative values. (We assume that both \(A\) and \(\beta\) are real numbers.)

(b) If \(A\) is negative and \(\beta\) is positive, \(x_f\) should be added to \(x_S\) (as in Figure P9.8 in the book) to attain negative feedback.

(c) If both \(A\) and \(\beta\) are negative, \(x_f\) should be subtracted from \(x_S\) (as in Figure 9.1 in the book) to attain negative feedback.
For the configuration of Figure P9.8, we have $A_f = \frac{A}{1 - A \beta}$. Substituting values, we have

$A = 10$ and $A_f = 100$ for $0 < x_i < 1$ or $0 < x_o < 10$

$A = 5$ and $A_f = 9.09$ for $-2 < x_i < 0$ or $-10 < x_o < 0$

Thus for $x_s = 0.1\sin(\omega t)$, the positive peak of the output is $100 \times 0.1 = 10$ V and the negative peak is $9.09 \times (-0.1) = -0.909$. A sketch of the output waveform is

![Graph showing positive and negative peaks]

With positive feedback, the ratio of the positive peak to the negative peak is 11.0, whereas the ratio is 2 without feedback. Notice that the output waveform is more distorted for the
amplifier with positive feedback than for the amplifier without feedback.
Problem 9.59

(a) transient response = Aexp(-t)

(b) transient response = Asin(1000t) + Bcos(1000t)

(c) transient response = Aexp(-3t) + Bexp(-4t)

(d) transient response = Aexp(-t)sin(1000t) + Bexp(-t)cos(1000t)

in which A and B are constants that depend on initial conditions, the excitation and the zeros.
Problem 9.65

The poles are given by Equation 9.50:

\[ s = \frac{1}{2} (2\pi f_1 + 2\pi f_2) \pm \frac{1}{2} \sqrt{(2\pi f_1 + 2\pi f_2)^2 - 16\pi^2 f_1 f_2 (1 + A_0 \beta)} \]

Substituting \( f_1 = 1000, f_2 = 500, A_0 = 1000, \) and \( \beta = 0.1, \) we eventually obtain

\[ s = -4712 \pm j 44.4 \times 10^3 \]
Problem 9.81

For the system of Figure P9.81a we have:

\[ A_{fa} = \frac{A_1 A_2}{1 + \beta A_1 A_2} \]

For the system of Figure P9.81b, we have:

\[ A_{fb} = \frac{A_1}{1 + \beta_1 A_1} \times \frac{A_2}{1 + \beta_2 A_2} \]

Evaluating, we find the values given in the table:
<table>
<thead>
<tr>
<th></th>
<th>$A_1 = 100$</th>
<th>$A_1 = 90$</th>
<th>% change</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A_{fa}$</td>
<td>99.01</td>
<td>98.90</td>
<td>0.11%</td>
</tr>
<tr>
<td>$A_{fb}$</td>
<td>82.64</td>
<td>81.82</td>
<td>1.0%</td>
</tr>
</tbody>
</table>

Thus, overall feedback is better for obtaining precision values of closed-loop gain.