High Frequency BJT Model & Cascode BJT Amplifier
Gain of 10 Amplifier – Non-ideal Transistor

Gain starts dropping at > 1MHz.

Why!
Because of internal transistor capacitances that we have ignored in our low frequency and mid-band models.
Sketch of Typical Voltage Gain Response for a CE Amplifier

Low Frequency Band
Due to external blocking and bypass capacitors.
Internal C's o.c.

Midband
ALL capacitances are neglected, i.e.
External C's s.c.
Internal C's o.c.

High Frequency Band
Due to BJT parasitic capacitors $C_\pi$ and $C_\mu$.
External C's s.c.

$\left| A_v \right|(dB)$

$BW = f_H - f_L \approx f_H$

$GBP = \left| A_{v-mid} \right| \cdot BW$

$f_L$
$f_H$

$20 \log_{10} \left| A_v \right|(dB)$

3 dB
High Frequency Small-signal Model (Fwd. Act.)

Two capacitors and a resistor added to mid-band small signal model.

- base-emitter capacitor, $C_\mu$
- base-collector capacitor, $C_\pi$
- resistor, $r_x$, representing the base terminal resistance ($r_x \ll r_\pi$); ignored in hand calculations.
High Frequency Small-signal Model (Fwd. Act.)

\[ C_\mu = \frac{C_{\mu 0}}{(1 + \frac{V_{CB}}{\phi_{0cb}})^m} \]

\[ C_{je} = \frac{C_{je0}}{(1 + \frac{V_{EB}}{\phi_{0eb}})^m} \]

Non-linear, voltage controlled

\[ m = \text{junction grading coefficient 0.2 to 0.5} \]

\[ C_\pi = C_{de} + C_{je} \approx C_{de} + 2 C_{je0} \approx C_{de} \]

\[ C_{de} = g_m \tau_F = \frac{I_C}{V_T} \tau_F \]

\[ C_{\mu} \approx 2 C_{\mu 0} \]

\[ C_{de} = \text{base-charging (diffusion) cap} \]

\[ \tau_F = \text{forward-base transit time} \]

SPICE/MultiSim

- \( C_{JC} = C_{\mu 0} \)
- \( C_{JE} = C_{je0} \)
- \( TF = \tau_F \)
- \( RB = r_x \)
High Frequency Small-signal Model (IC)

$C_{CS} =$ collector-substrate depletion capacitance, ignored in hand calcs.
High Frequency Small-signal Model

The transistor parasitic capacitances have a strong effect on circuit high frequency performance!

- $C_\pi$ attenuates base signal, decreasing $v_{be}$ since $X_{C\pi} \rightarrow 0$ (short circuit) at very high-frequencies.
- As we will see later; $C_\mu$ is principal cause of gain loss at high-frequencies. At the base base-collector $C_\mu$ looks like a capacitor of value $k C_\mu$ between base-emitter, where $k > 1$ and may be $>> 1$.

- This phenomenon is called the **Miller Effect**.
Prototype Common Emitter Circuit

At high frequencies, large blocking and bypass capacitors are “short circuits”

High frequency small-signal ac model
Multisim Simulation

Mid-band gain @ 100 kHz

Gain @ 8.69 MHz

\[ A_{v-mid} = -g_m R_C = -204 \left( 46 \text{ dB} @ 180^\circ \right) \]
Introducing the Miller Effect

The feedback connection of $C_\mu$ between base-collector causes it to appear in the amplifier like a large capacitor $(1 - K)C_\mu$ between the base-emitter terminals. This phenomenon is called the “Miller effect” and capacitor multiplier “$1 - K$” acting on $C_\mu$ equals the CE amplifier mid-band gain, i.e.

$$K = A_{v_{-\text{mid}}} = -g_m R_C$$

**NOTE:** CB and CC amplifiers do not suffer from the Miller effect, since in these amplifiers, one side of $C_\mu$ is connected directly to ground.
**Miller's Theorem**

\[ Z = \frac{1}{j 2\pi f C_{\mu}} \]

\[ V_o \approx V_{be} = -g_m R_C \]

\[ A_{v-mid} = \frac{V_o}{V_{be}} = -g_m R_C \]

\[ R_B \parallel r_\pi \gg R_S \]
Miller's Theorem

\[ I = I_1 \]
\[ I_2 = -I \]

\[ V_1 \]
\[ A_v V_1 \uparrow \] \[ V_2 \]

\[ I_1 = \frac{V_1 - V_2}{Z} = \frac{V_1 - A_v V_1}{Z} = \frac{V_1}{Z} \frac{1 - A_v}{1} \]

\[ I_2 = \frac{V_2 - V_1}{Z} = \frac{V_2 - \frac{1}{A_v} V_2}{Z} = \frac{V_2}{Z} \frac{1 - \frac{1}{A_v}}{1 - \frac{1}{A_v}} \]

\[ Z_1 = \frac{1}{j 2 \pi f C_\mu (1 - A_v)} \]

\[ Z_2 = \frac{V_2}{I_2} = \frac{V_2}{-I} = \frac{Z}{1 - \frac{1}{A_v}} \approx Z \text{ if } A_v >> 1 \]

Ignored in practical circuits

Kenneth R. Laker, update 01Oct14 KRL
Common Emitter Miller Effect Analysis

Determine effect of $C_\mu$:

Using phasor notation:

$$I_{Rc} = -g_m V_{be} + I_{C\mu}$$

or

$$V_o = I_{Rc} R_c = \left(-g_m V_{be} + I_{C\mu}\right) R_c$$

where

$$I_{C\mu} = \frac{1}{j 2 \pi f C_\mu} = V_{be} - V_o$$

$$I_{C\mu} = \left(\frac{V_{be} + g_m V_{be} R_c - I_{C\mu} R_c}{1 + g_m R_c \left(1 + j 2 \pi f R_c C_\mu\right)}\right) V_{be}$$

Note: The current through $C_\mu$ depends only on $V_{be}$!
Common Emitter Miller Effect Analysis II

From slide 13:

\[ I_{C\mu} = \frac{(1 + g_m R_C) j 2 \pi f C_{\mu}}{1 + j 2 \pi f R_C C_{\mu}} V_{be} \]

Miller Capacitance \( C_{eq} \):

\[ C_{eq} = (1 + g_m R_C) C_{\mu} = (1 - A_v) C_{\mu} \]

\[ V_{be} \approx (1 + g_m R_C) j 2 \pi f C_{\mu} V_{be} = j 2 \pi f C_{eq} V_{be} \]
\[ C_{eq} = (1 - A_v)C_\mu = (1 + g_m R_C)C_\mu \]
Common Emitter Miller Effect Analysis III

\[ C_{eq} = \left( 1 + g_m R_C \right) C_\mu \]

For our example circuit \( C_\mu = 2 \text{ pF} \):

\[ 1 + g_m R_C = 1 + 0.040 \cdot 5100 = 205 \]

\[ C_{eq} = (205) \cdot 2 \text{ pF} \approx 410 \text{ pF} \]
Simplified HF Model

\[ V_s' = V_s \frac{R_B || r_\pi}{R_B || r_\pi + R_S} \]

\[ R_S' = r_\pi || (R_B || R_S) \]

\[ R_L' = r_o || (R_C || R_L) \]

Thevenin Equiv.
Simplified HF Model

\[ R'_L = r_o \parallel (R_C \parallel R_L) \]
\[ R'_S = r_\pi \parallel (R_B \parallel R_S) \]
\[ V'_s = V_s \frac{R_B}{R_B \parallel r_\pi + R_S} \]

\[ C_{eq} = (1 + g_m R_C) C_{\mu} \]
\[ C_{tot} = C_{\pi} + C_{eq} \]
Simplified HF Model

\[ C_{\text{tot}} = C_{\pi} + (1 + g_m R_C) C_{\mu} \]

\[ V_{\text{be}} = \frac{1}{1 + j 2 \pi f C_{\text{tot}} + R_s' R_{\text{L}}} V_s' \]

\[ A_v(f) = \frac{V_o}{V_s} \approx -g_m R_L' \frac{1 + j 2 \pi f C_{\text{tot}} R_s'}{1 + j \frac{f}{f_H}} \]

\[ A_{v-mid} = -g_m R_L' \text{ and } f_H = \frac{1}{2 \pi C_{\text{tot}} R_s} \]
Multisim Simulation

Mid-band gain @ 100 kHz

\[ A_{v\text{-mid}} = -g_m R_C = -204 (46 \text{ dB} @ 180^\circ) \]

\[ C_{tot} = 12 \text{ pF} + (205) \cdot 2 \text{ pF} \approx 422 \text{ pF} \]

\[ f_H \approx f_{-3\text{dB}} = \frac{1}{2\pi C_{tot} R_S} \approx \frac{1}{2\pi \cdot 422 \text{ pF} \cdot 50 \Omega} \approx 7.54 \text{ MHz} \]

Gain @ 8.69 MHz
Frequency-dependent “beta”

\[ I_b = \left( \frac{1}{r_x} + j2\pi f (C_{\pi} + C_{\mu}) \right) V_b \]
where \( r_x \approx 0 \) (ignore \( r_x \))

The relationship \( i_c = \beta i_b \) does not apply at high frequencies \( f \geq f_H \! \)!

Using the relationship \(-i_c = f(V_b)\) – find the new relationship between \( i_b \) and \( i_c \). For \( i_b \) (using phasor notation \( I_x = I_x (jf) \) & \( V_x = V_x (jf) \) for frequency domain analysis):

@ node B': \( I_b = \left( \frac{1}{r_{\pi}} + j2\pi f C_{\pi} \right) V_b \)

short-circuit current
Frequency-dependent “beta”

\[ I_b = \left( \frac{1}{r_{\pi}} + j2\pi f (C_\pi + C_\mu) \right) V_b \]  

@ node C:  \[ I_c = (g_m - j2\pi f C_\mu) V_b \]  

Leads to a new relationship between the \( I_b \) and \( I_c \):

\[ \beta(jf) \to h_{fe} \]

\[ \beta(jf) = \frac{I_c}{I_b} = \frac{g_m - j2\pi f C_\mu}{\frac{1}{r_{\pi}} + j2\pi f (C_\pi + C_\mu)} \]

(ignoring \( r_o \))
**Frequency Response of “beta”**

\[
\beta(jf) = \frac{g_m - j2\pi f C_\mu}{1 + j2\pi f (C_\pi + C_\mu)}
\]

Multiplying N&D by \( r_\pi \)

\[
\beta(jf) = \frac{(g_m - j2\pi f C_\mu) r_\pi}{1 + j2\pi f (C_\pi + C_\mu) r_\pi}
\]

Factor N to isolate \( g_m \)

\[
(g_m - j2\pi f C_\mu)\frac{r_\pi}{g_m} = \frac{\beta}{1 - j2\pi f C_\mu}
\]

\[
\beta(jf) = \frac{\beta}{1 + j2\pi f (C_\pi + C_\mu) r_\pi}
\]

For \( f = f_{\text{low}} \) s.t. \( 2\pi f_{\text{low}} \frac{C_\mu}{g_m} \ll 1 \leq \frac{1}{10} \)

And \( 2\pi f_{\text{low}} (C_\pi + C_\mu) r_\pi \ll 1 \leq \frac{1}{10} \)

\[
\beta(jf) = g_m r_\pi = \beta
\]
Frequency Response of “beta” cont.

\[
\beta(jf) = \frac{(1 - j2\pi f \frac{C}{g_m})g_m r_\pi}{1 + j2\pi f (C_\pi + C_\mu) r_\pi} = \left(1 - j\frac{f}{f_z}\right)\frac{1 - j\frac{f}{f_\beta}}{1 + j\frac{f}{f_\beta}} g_m r_\pi = \left(1 - j\frac{f}{f_z}\right)\left(1 - j\frac{f}{f_\beta}\right) \beta
\]

Hence, the lower break frequency or \(-3dB\) frequency is \(f_\beta\)

\[
f_\beta = \frac{1}{2\pi(C_\pi + C_\mu) r_\pi} = \frac{g_m}{2\pi(C_\pi + C_\mu) \beta}
\]

the upper: \(f_z = \frac{1}{2\pi C_\mu / g_m} = \frac{g_m}{2\pi C_\mu}\)

where \(f_z \gg f_\beta\)
Frequency Response of “beta” cont.

Using Bode plot concepts, for the range where: \( f < f_\beta \)

\[
\beta(jf) = g_m r_\pi = \beta
\]

For the range where: \( f_\beta < f < f_z \) s.t. \(|1 - jf/f_z| \approx 1\)

We consider the frequency-dependent numerator term to be 1 and focus on the response of the denominator:

\[
\begin{align*}
&f_\beta < f < f_z \\
&\beta(jf) \approx \frac{g_m r_\pi}{1 + j\frac{f}{f_\beta}} = \frac{\beta}{1 + j\frac{f}{f_\beta}}
\end{align*}
\]
Frequency Response of “beta” cont.

Neglecting numerator term:
\[
\beta(jf) = \frac{g_m r_{\pi}}{1 + j \frac{f}{f_B}} = \frac{\beta}{1 + j \frac{f}{f_B}}
\]

And for \(f/f_B >> 1\) (but \(< f/f_z\)):  
\[
|\beta(jf)| \approx \frac{\beta}{f/f_B} = \beta \frac{f_B}{f}
\]

Unity gain bandwidth:  
\[
|\beta(jf)| = 1 \Rightarrow \beta \frac{f_B}{f}
\]

\[
f_T = \frac{\omega_T}{2\pi} = \beta f_B
\]

BJT unity-gain frequency or GBP
Frequency Response of “beta” cont.

\[ \beta = 100 \quad r_\pi = 2500 \, \Omega \quad C_\pi = 12 \, pF \quad C_\mu = 2 \, pF \quad g_m = 40 \cdot 10^{-3} \, S \]

\[ \omega_\beta = \frac{1}{(C_\pi + C_\mu) r_\pi} = \frac{10^{12} \cdot 10^{-3}}{(12 + 2) \cdot 2.5} = 28.57 \cdot 10^6 \, rps \]

\[ f_\beta = \frac{\omega_\beta}{2 \pi} = \frac{28.57}{6.28} \cdot 10^6 \, Hz = 4.55 \, MHz \]

\[ f_T = \beta f_\beta = 455 \, MHz \]

\[ \omega_z = \frac{g_m}{C_\mu} = \frac{40 \cdot 10^{-3} \cdot 10^{12}}{2} \, Hz = 20 \cdot 10^9 \, rps \]

\[ f_z = \frac{\omega_z}{2 \pi} = 3.18 \cdot 10^9 \, Hz = 3180 \, MHz \]
“beta” Bode Plot

$|\beta(jf)|$ (dB) vs. Frequency

$|\beta(jf)|$ vs. Frequency

Kenneth R. Laker, update 01Oct14 KRL
Insert 1 ohm resistors – we want to measure a current ratio.

\[ \beta(j\omega) = \frac{I_c}{I_b} = \frac{g_m - j2\pi f C_\mu}{\frac{1}{r'} + j2\pi f (C_\pi + C_\mu)} \]
Simulation Results

Low frequency $|\beta(f_f)|$

Unity Gain frequency about 440 MHz

Theory: $f_T = \beta f_\beta = 455 MHz$
The Cascode Amplifier

- A two transistor amplifier used to obtain simultaneously:
  1. Reasonably high input impedance.
  2. Reasonable voltage gain.
  3. Wide bandwidth.

- None of the conventional single transistor designs will satisfy all of the criteria above.
- The cascode amplifier will satisfy all of these criteria.
- A cascode is a CE Stage cascaded with a CB Stage.

(Historical Note: the cascode amplifier was a cascade of grounded cathode and grounded grid vacuum tube stages – hence the name “cascode,” which has remained in modern terminology.)
The Cascode Amplifier

Comments:
1. $R_1$, $R_2$, $R_3$, and $R_C$ set the bias levels for both $Q1$ and $Q2$.
2. Determine $R_E$ for the desired voltage gain.
3. $C_{in}$ and $C_{byp}$ are to act as “open circuits” at dc and act as near “short circuits” at all operating frequencies $f > f_{min}$.
## CE and CB Amplifier Feature Review

<table>
<thead>
<tr>
<th>Feature</th>
<th>CE Amplifier</th>
<th>CB Amplifier</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage Gain ((A_v))</td>
<td>moderate ((-R_c/R_e))</td>
<td>High ((R_c/(R_s + r_e)))</td>
</tr>
<tr>
<td>Current Gain ((A_i))</td>
<td>High ((\beta))</td>
<td>low (about 1)</td>
</tr>
<tr>
<td>Input Resistance</td>
<td>High ((R_B</td>
<td></td>
</tr>
<tr>
<td>Output Resistance</td>
<td>High ((R_c</td>
<td></td>
</tr>
</tbody>
</table>

Important for **Cascode**
ASSUME: \( Q1 = Q2 \); e.g. \( \beta \)'s identical, and both Forward Active
Cascode Small Signal Analysis

1. Show reduction in Miller effect
2. Evaluate small-signal voltage gain

OBSERVATIONS

a. The emitter current of the CB Stage is the collector current of the CE Stage. (This also holds for the dc bias currents.)

\[ i_{e1} = i_{c2} \]

b. The base current of the CB Stage is:

\[ i_{b1} = \frac{i_{e1}}{\beta + 1} = \frac{i_{c2}}{\beta + 1} \]

c. Hence, both stages have about same collector current \( i_{cl} \approx i_{c2} \) and same \( g_m, r_e, r_{\pi} \).

\[ I_{C1} \approx I_{C2} \]
Cascode Small Signal Analysis cont.

The input resistance $R_{in1}$ to the CB Stage is the small-signal “$r_{e1}$” for the CB Stage.

$$i_{bl} = \frac{i_{e1}}{\beta + 1} = \frac{i_{c2}}{\beta + 1}$$

The CE output voltage, the voltage drop from Q2 collector to ground, is:

$$v_{c2} = v_{el} = -r_\pi i_{bl} = -\frac{r_\pi}{\beta + 1} i_{e1} = -\frac{r_\pi}{\beta + 1} i_{c2}$$

Therefore, the CB Stage input resistance is:

$$R_{in1} = \frac{v_{el}}{-i_{e1}} \approx \frac{r_\pi}{\beta + 1} = r_{e1}$$

Miller Effect

Much reduced

$$A_{vCE-Stage} = \frac{v_{c2}}{v_s} \approx -\frac{R_{in1}}{R_E} \approx -\frac{r_e}{R_E} < 1 \Rightarrow C_{eq} = (1 + \frac{r_e}{R_E})C_\mu < 2C_\mu$$
Cascode Small Signal Analysis - cont.

Now, find the CE collector current in terms of the input voltage $v_s$:

Recall $i_{c1} \approx i_{c2}$

$$i_{b2} \approx \frac{v_s}{R_S \parallel R_B + r_\pi + (\beta + 1) R_E}$$

$$i_{c2} = \beta i_{b2} \approx \frac{\beta v_s}{R_S \parallel R_B + r_\pi + (\beta + 1) R_E} \approx \frac{\beta v_s}{(\beta + 1) R_E}$$

for bias insensitivity: $(\beta + 1) R_E \gg R_S \parallel R_B + r_\pi$

OBSERVATIONS:

1. Voltage gain $A_v$ is about the same as a stand-alone CE Amplifier.
2. HF cutoff is much higher than a CE Amplifier due to the reduced $C_{eq}$. 

Kenneth R. Laker, update 01Oct14 KRL
Common Emitter Stage

\[ C_{tot} = C_\pi + (1 + g_m R_C) C_\mu \]

Cascode Stage

\[ C_{tot} = C_\pi + \left(1 + \frac{r_e}{R_E}\right) C_\mu \]

\[ < 2 C_\mu \]
**Cascode Biasing**

1. Choose $I_{E1}$ – make it relatively large to reduce $R_{in1} = r_e = V_T/I_{E1}$ to push out HF break frequencies.

2. Choose $R_C$ for suitable voltage swing $V_{C1}$ and $R_E$ for desired gain.

3. Choose bias resistor string such that its current $I_I$ is about 0.1 of the collector current $I_{C1}$.

4. Given $R_E$, $I_{E2}$ and $V_{BE2} = 0.7 \, V$ calc. $R_3$.

5. Need to also determine $R_1$ & $R_2$.

\[ \alpha I_{E2} = I_{C2} = I_{E1} = \frac{1}{\alpha} I_{C1} \Rightarrow I_{C1} = \alpha^2 I_{E2} \approx I_{E2} \]
Since the CE-Stage gain is very small:

a. The collector swing of Q2 will be small.

b. The Q2 collector bias \( V_{C2} = V_{B1} - 0.7 \, V \).

6. Set \( V_{B1} - V_{B2} = 1 \, V \Rightarrow V_{CE2} = 1 \, V \)

Since \( V_{CE2} = 1 \, V > V_{CE2sat} \) Q2 forward active.

7. Next determine \( R_2 \). Its drop \( V_{R2} = 1 \, V \) with the known current.

\[
R_2 = \frac{V_{B1} - V_{B2}}{I_1}
\]

\[
V_{CE2} = V_{C2} - V_{Re} = V_{C2} - (V_{B2} - 0.7 \, V)
\]

\[
= V_{B1} - 0.7 \, V - V_{Re}
\]

\[
= V_{B1} - 0.7 \, V - V_{B2} + 0.7 \, V
\]

\[
= V_{B1} - V_{B2} = 1 \, V
\]
Cascode Biasing - cont.

8. Then calculate $R_3$.

$$R_3 = \frac{V_{B2}}{I_1}$$

where $V_{B2} = 0.7V + I_E R_E$

Note: $R_1 + R_2 + R_3 = \frac{V_{CC}}{I_1}$

9. Then calculate $R_1$.

$$R_1 = \frac{V_{CC}}{0.1 I_C} - R_2 - R_3$$
Cascode Bias Summary

SPECIFIED: $A_v$, $V_{CC}$, $V_{C1}$ (CB collector voltage);

SPECIFIED: $f_H = \Rightarrow C_{tot}(r_e) = \Rightarrow I_E$ (or $I_C$) $\Rightarrow r_e$.

DETERMINE: $R_C$, $R_E$, $R_1$, $R_2$ and $R_3$.

SET: $V_{B1} - V_{B2} = 1V \Rightarrow V_{CE2} = 1V$

\[ I_{C2} = I_{E1} \approx I_{C1} \approx I_{E2} = I_C \]

\[ \text{STEP1: } R_C = \frac{V_{C1}}{I_C}, \quad R_E = \frac{R_C}{|A_v|} \]

\[ \text{STEP2: } R_2 = \frac{V_{B1} - V_{B2}}{I_1} = \frac{1V}{0.1I_C} \]

\[ \text{STEP3: } R_3 = \frac{V_{B2}}{I_1} = \frac{0.7V + I_E R_E}{0.1I_C} \]

\[ R_1 + R_2 + R_3 = \frac{V_{CC}}{I_1} = \frac{V_{CC}}{0.1I_C} \]

\[ \text{STEP4: } R_1 = \frac{V_{CC}}{0.1I_C} - R_2 - R_3 \]
Cascode Bias Example

Typical Bias Conditions

\[ V_{CC} - I_C R_E = 1.7 \]

\[ I_E2 \approx I_C2 = I_{E1} \approx I_C1 \Rightarrow I_C1 \approx I_E2 \]
Cascode Bias Example cont.

1. Choose $I_{E1}$ – to set $r_e$.
Let $r_e = 5\Omega \Rightarrow I_{E1} = 0.025\,V/r_e = 5\,mA$

2. Set desired gain magnitude. For example if $A_V = -10$, then $R_C/R_E = 10$.

3. Since the CE stage gain is very small, $V_{CE2}$ can be small, i.e. $V_{CE2} = V_{B1} - V_{B2} = 1\,V$. 
Cascode Bias Example cont.

Specs:
\[ V_{CC} = 12 \, V \quad V_{C1} = 7 \, V \quad r_e = 5 \, \Omega \quad |A_v| = \frac{R_C}{R_E} = 10 \]
\[ \beta = 100 \]

Determine \( R_C \) for \( V_{C1} = 7 \, V \).

\[ I_{E1} = 0.025 \, V / r_e = 5 \, mA \]
\[ R_C = \frac{V_{CC} - V_{C1}}{I_{E1}} = \frac{5 \, V}{5 \times 10^{-3} \, A} = 1000 \, \Omega \]
\[ R_E = \frac{R_C}{|A_v|} = \frac{R_C}{10} = 100 \, \Omega \]
Cascode Bias Example cont.

\[
V_{CC} = 12 \quad R_C = 1 \, k\Omega \quad r_e = 5 \, \Omega \quad R_E = 100 \, \Omega
\]
\[
\beta = 100
\]

Make current through the string of bias resistors \( I_1 = 0.1 \, I_C = 0.5 \, mA \).

\[
R_1 + R_2 + R_3 = \frac{V_{CC}}{I_1} = \frac{12}{5 \cdot 10^{-4}} = 24 \, k\Omega
\]

Calculate the bias voltages (base side of Q1, Q2):

\[
V_{R1} = V_{CC} - 1.7V - I_C R_E = 12V - 1.7V - 0.5V = 9.8V
\]
\[
V_{R2} = V_{BL} - V_{B2} = 1V
\]
\[
V_{R3} = V_{B2} = 0.7 + I_C R_E = 0.7 + 5 \cdot 10^{-3} \cdot 100 = 1.2V
\]
Cascode Bias Example cont.

\[ V_{B2} = 5 \cdot 10^{-4} \quad R_3 = 1.2 \, V \]

\[ R_3 = 2.4 \, k\, \Omega \]

\[ V_{B1} - V_{B2} = 5 \cdot 10^{-4} \quad R_2 = 1.0 \, V \]

\[ R_2 = 2 \, k\, \Omega \]

Recall: \( R_1 + R_2 + R_3 = 24 \, k\, \Omega \)

\[ R_1 = 24000 - 2.400 - 2000 = 19.6 \, k\, \Omega \]

Recall for ac: \( R_B = R_2 \parallel R_3 \)

\[ R_B = 2 \, k\, \Omega \parallel 2.4 \, k\, \Omega = 1.1 \, k\, \Omega \]

\[ V_{CC} = 12 \, V, \quad R_C = 1 \, k\, \Omega, \quad V_{B2} = 1.2 \, V, \]

\[ I_C = 5 \, mA, \quad R_E = 100 \, \Omega, \quad V_{B1} - V_{B2} = 1.0 \, V \]
\[ \beta = 100 \]
\[ r_e = 5 \Omega \Rightarrow I_C = 5 mA \]
\[ V_{C1} = 7 V \]
\[ |A_v| = \frac{R_C}{R_E} = 10 \]
\[ R_1 = 19.6 k \Omega \]
\[ R_2 = 2 k \Omega \]
\[ R_3 = 2.4 k \Omega \]
\[ R_C = 1 k \Omega \]
\[ R_E = 100 \Omega \]

**NOTE:** \( R_B = R_2 \parallel R_3 = 1.09 k \Omega \ll \beta R_E = 10 k \Omega \)

\[ f_H = \frac{1}{2 \pi C_{\text{tot}} R_S} \]
\[ C_{\text{tot}} = C_\pi + (1 + \frac{r_e}{R_E}) C_\mu \]
\[ = C_\pi + 1.05 C_\mu \]
\[ f_{H\text{cascode}} = 225.8 \text{ MHz} \]

If \( C_\pi = 12 \text{ pF} \)
\[ C_\mu = 2 \text{ pF} \]
\[ C_{\text{tot}} = 14.1 \text{ pF} \]

For CE with \(|A_v| = 10\)
\[ f_{H\text{CE}} = 94 \text{ MHz} \]