Amplifier Frequency Response, Feedback, Oscillations; Op-Amp Block Diagram and Gain-Bandwidth Product

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Negative Feedback Example: Voltage Amplifier
Review Negative Feedback Advantages

- Improved input impedance
- Improved output impedance
- Improved linearity
- Improved frequency response
- However, gain reduced, must avoid oscillation
Negative Feedback and Voltage Amplifier

Voltage Amplifier with Negative Feedback

\[ A_F = \frac{v_o}{v_{in}} = \frac{A}{1 + AB} \quad R_{iF} = \frac{v_{in}}{i_{in}} = (1 + AB)R_{in} \]

\[ R_{oF} = \frac{R_o}{1 + AB} \]

AB is called the loop gain.

(see solutions to Prob. 10.35 for proofs)

[Diagram of a voltage amplifier with feedback network]

Fig. P10.35
© Bobrow
Voltage Amplifier: \( A_F \) dependence on \( A \)

\[ B = 1/20, \quad A_F = A/(1 + AB) = [A^{-1} + B]^{-1}: \]

<table>
<thead>
<tr>
<th>( A )</th>
<th>( A_F )</th>
</tr>
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<tbody>
<tr>
<td>200000</td>
<td>19.998</td>
</tr>
<tr>
<td>100000</td>
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<td>100</td>
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</tr>
<tr>
<td>50</td>
<td>14.3</td>
</tr>
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</table>

- \( A_{F_{\text{max}}} = 20 \). If \( A \gg A_{F_{\text{max}}} \), \( A_F \) is insensitive to \( A \). \( A_F \) is down \( \approx 3 \) dB from maximum when \( A = 50 \).

- Reduces distortion due to \( A \) nonlinearity, allows for variations in amplifier gain from device to device. *(What Black wanted back in the 1920’s for his telephone long-distance line amplifiers)*

- Suppose \( A \) is 200000 at low frequency (say 1 Hz) but falling with frequency like \( 1/f \) at high frequencies due to a built-in low-pass filter with \( f_c = 5 \) Hz. With feedback, the -3 dB bandwidth would be improved, since \( A_F \) remains high until \( A \) has fallen many orders of magnitude.
Phase-Shift Oscillator and $AB = -1$

**Sinusoidal oscillation when $AB = -1$**

$$A_F = \frac{v_o}{v_{in}} = \frac{A}{1 + AB}$$

**Example:** If $A = 200000$, $R_{in} = 2 \text{ M}\Omega$, $R_o = 75 \Omega$ and $B = \frac{1}{20}$, then $A = 20$, $R_{iF} = 20 \text{ G}\Omega$ and $R_{oF} = 7.5 \text{ m}\Omega$. 

$\text{AB} = -1$ assumed negative feedback. ("Barkhausen x Criterion")

Taking this into account, the amplifier gain $A = \frac{R_F}{R}$. (-1 there already in $-1$)

For the RC network,

$$B = \frac{V_i}{V_o} = \frac{\omega^3 R^2 C^3}{\omega RC (\omega^2 R\epsilon^2 - 5) + j(1 - 6\omega^2 R\epsilon^2)}$$

$B$ must be real for $AB = -1$ so

$$1 - 6\omega^2 R\epsilon^2 = 0 \Rightarrow \omega = \omega_0 = \frac{1}{\sqrt{6RC}}$$

$$B = \frac{\omega^2 R\epsilon^2}{(\omega_0^2 R\epsilon^2 - 5)} \quad \omega_0^2 R\epsilon^2 = \frac{1}{6}$$

$$\frac{\omega}{\omega_0} - 5 = \frac{1}{1 - 30} = -\frac{1}{29}$$

$$\text{AB} = -1 \Rightarrow \frac{R_F}{R} \left( -\frac{1}{29} \right) = -1$$

$$R_F = 29R$$

*German physicist, 1881-1956.*
Amplifier Low Frequency Limitations

Simplified treatment of coupling capacitors at input or output stages of amplifiers. (Both source and load impedances are important... and must be known for an exact answer. If one is much larger than the other, simplifications are possible.)

High pass network usually looks like this: \( R_o = \) source resistance.

\[ \begin{align*}
    &U \\
    &\quad \text{could be amplifier output stage or signal at input.} \\
    &\quad \text{or output stage or input impedance of amplifier.} \\
\end{align*} \]
Amplifier Low Frequency Limitations (continued)

A high-pass network usually looks like this: \( R_o = \text{source resistance} \).

\[ U = U_0 \cos(\omega t) \]

At high frequency, this is a resistive divider:

\[ V' = \frac{R_L}{R_o + R_L} U \]

In general,

\[ V' = V \left( \frac{R_L}{R_o + R_L} \right) \left( \frac{R_o + R_L}{R_o + R_L + j\omega C} \right) \]

\[ = \left( \frac{R_L}{R_o + R_L} \right) \left( \frac{R_o + R_L}{R_o + R_L + j\omega C} \right) \]

Thus this has a -3dB point relative to the value at high frequency where \( \frac{1}{j\omega C} = R_o + R_L \)

or

\[ f_c = \frac{1}{2\pi(R_o + R_L)C} \]
CE Amplifier

Terminology: \( C_c = C_{bc} \)  \( C_e = C_{be} \)

\[ f_T = \text{current gain bandwidth product} = \frac{1}{2\pi (C_c + C_e) R_e} \]

(a) Lower cutoff frequency: treat each part separately to see if any one part is dominant:

i) Input circuit (assume \( C_e \to \infty \))

\[ f_c = \frac{1}{2\pi \left( R_S + R_1 R_2 (R_2^{-1}) C_1 \right)} \]

\[ f_c = \frac{1}{2\pi \left( 1 \, \text{k} \Omega + 1 \, \text{M} \Omega \right) \times 1 \, \text{pF}} = 58 \, \text{MHz} \]

\[ R_T = h_{re} R_e = 22 \, \Omega \]
CE Amplifier

Terminology: $C_c = C_{bc}$, $C_e = C_{be}$

$\frac{f_T}{\omega} =$ current gain bandwidth product

$$f_T = \frac{1}{2\pi(C_c + C_e)R_e}$$

(a) Lower cutoff frequency: treat each part separately to see if any one part is dominant:

(ii) Output circuit:

$$f_c = \frac{1}{2\pi(R_c + R_L)C_c} = \frac{1}{2\pi \times 10k \times 4.7 \times 10^{-6}} = 3.3 \text{ Hz}$$
Amplifier High Frequency Limits

- Model for parallel (shunt) capacitances to ground in amplifier circuit:

\[ V = V_0 e^{jωt} \]

At low frequency, \( V' = \frac{R_L}{R_0 + R_L} V \).

The Thevenin equivalent of things to the left of the dotted line is

\[ V_{eq} = \frac{R_L V}{R_0 + R_L} \]

\[ R' = \frac{R_L || R_0}{R_L + R_0} \]

So \( f_C = \frac{1}{2\pi R'C_S} \).
CE Amplifier

Terminology: \( C_c = C_{bc} \) \( C_e = C_{be} \)

\[ f_T = \frac{1}{2\pi (C_c + C_e) R_e} \]

(a) Lower cutoff frequency: treat each part separately to see if any one part is dominant.

(iii) Emitter circuit at \( f_c \) due to \( C_e \) shunt capacitance, assuming \( C_1 \to \infty \) (ineffective below \( f_c \)):

\[ R_{equiv} = \text{resistance looking into emitter} = R_e + \frac{R_1 R_2}{h_{fe} + 1} \]

\[ f_c = \frac{1}{2\pi (R_{equiv} R_e) C_e} = \frac{1.13 \text{ kHz}}{} \]

Determines LF - 3dB point.
Amplifier High Frequency Limits

- Model for parallel (shunt) capacitances to ground in amplifier circuit:

\[ V = V_0 e^{j \omega t} \]

At low frequency, \( V' = \frac{R_L}{R_0 + R_L} V \).

The Thévenin equivalent of things to the left of the dotted line is

\[ V_t = \frac{R_L V}{R_0 + R_L} \]

So \( f_c = \frac{1}{2\pi R' C} \).

\( R' = \frac{R_L R_0}{R_L + R_0} \).

Where are these shunt capacitances?
Common Emitter Amplifier HF Limits

- At high frequencies, must consider BE and BC diode capacitances

![Diagram of a CE Amplifier]

Terminology:

\[ f_T = \frac{1}{2\pi(C_C + C_E)R_e} \]

\[ h_{re} = 100 \]

\[ C_C = 5 \text{ pF} \]

\[ R_e = 22 \Omega \]

\[ f_T = 150 \text{ MHz} \]
Shunt Capacitances In Small-Signal AC Models

Simple BJT Model at High Frequencies:

```
Base-Emitter Diode Capacitance
```
```
Base-Collector Diode Capacitance
```

Simple HF JFET Model:

```
C_{gs}, C_{gd} \approx 1\text{ pF}
C_{ds} \approx 1\text{ pF}
```

- How to deal with $C_c$ (or $C_{gd}$) which connects input and output?
Use Miller’s Theorem to Split $C_c$ (or $C_{gd}$)

Miller's Theorem:

If $V_2/V_1 = k$

(as for input and output nodes of a voltage amplifier)

$z_1 = \frac{z_1'}{1-k}$

$z_2 = \frac{z_2'}{1-k}$, $z_2 = \frac{k z_2'}{k-1}$

If $z_1' = 1/j\omega C'$ (Capacitor) and $k = -A$

then $C_1 = (A+1)C'$, $C_2 = \frac{A+1}{A}C'$.

Apply to HF BJT model in a common emitter amplifier with gain = -A:

- Input circuit (be) and output circuit (ce) are now separated
Miller’s Theorem Proof

Given: \( \frac{V_2}{V_1} = k \)

**Node 1 (write in terms of \( v_1 \)):**

\[
I_1 = \frac{V_1 - V_2}{Z'} = \left(1 - k\right)\frac{V_1}{Z'}
\]

\[
\frac{V_1}{I_1} \equiv Z_1 = \frac{Z'}{1-k}
\]

**Node 2 (write in terms of \( v_2 \)):**

\[
I_2 = \frac{V_2 - V_1}{Z'} = \frac{V_2}{Z'} \left(1 - \frac{1}{k}\right)
\]

\[
\frac{V_2}{I_2} \equiv Z_2 = \frac{Z'}{1-\frac{1}{k}} = \frac{kZ'}{k-1}
\]

QED
Common Emitter Amplifier Input Stage

HF BJT model with Miller's thm:

CE Amplifier Input stage after applying Miller's thm (high freq. case)

\[ f_c = \frac{1}{2\pi R_g C_g} \]

\[ R_g = R_s (1 + \frac{1}{A}) R_1 \]

\[ C_g = C_e + (1 + A) C_e \]

(Following text notation - g must stand for "lg" (low pass) network)
Common Emitter Amplifier Output Stage

HFE BJT model with Miller's gain:

\[ V_{out} = \frac{g_m V_{be}}{A + \frac{1}{A}} C_c \]

Output Stage

Norton:

\[ R_T = R_C R_L \]

\[ V_{out} = \frac{g_m V_{be}}{A + \frac{1}{A}} C_c \]

Note that the output circuit displays a constant gain-bandwidth product (assuming LF extends to \( \omega_0 \))

\[ B_W = \frac{A}{g_m t} \]

\[ A = \frac{R_T}{R_C} = g_m t \gg 1 \]

\[ (A_T = -A) \]

Gain-bandwidth product \( A \times B_W = \frac{R_T}{R_C} \times \frac{1}{2 \pi b_T \left( \frac{A + 1}{A} \right) C_c} = \frac{1}{2 \pi C_c} \]

If the input circuit BW does not limit (i.e., \( \gg \) this BW)

Gain BW, then we get this constant gain-BW product.
FET HF Model and Analysis

Exact same analysis applies to FETs, except the FET model is slightly different:

\[ \text{Diagram of FET model} \]

- \( C_{gs}, C_{gd} = 1-3 \text{ pF} \)
- \( C_{ds} \leq 1 \text{ pF} \)

Again, split \( C_{gd} \) into (via Miller):
- \( (A+1)C_{gd} \) in input circuit
- Parallel to \( C_{gs} \), \( \frac{(A+1)}{A}C_{gd} \) in output circuit \( \parallel \) to \( C_{ds} \)
9.54 (b) Upper cutoff frequency

1. Consider input (base) circuit with Miller effect w.r.t. \( C = C_{\text{B}} \). We need to know mid-band gain, \( A = -A_{\text{V}}(\text{midband}) = \frac{R_{\text{C}}/r_{\text{e}}}{R_{\text{e}}} \) for this circuit

\[
A = \frac{2.5 \times 10^3}{22 \times 10^3} = 114
\]

\[
C_{\text{g}} = 2\pi fB = \frac{1}{R_{\text{g}} C_{\text{g}}}
\]

where \( R_{\text{g}} = R_{\text{S}}/R_{\text{P}}/R_{\text{T}} \) and \( R_{\text{P}} = R_{\text{I}}/R_{\text{Z}} \) (assumes \( R_{\text{B}} \) bypassed)

and \( C_{\text{g}} = C_{\text{e}} + (1+A)C_{\text{c}} \) (Miller effect)

Now we need \( C_{\text{e}} \) but we have

\[
C + C_{\text{e}} = \frac{1}{2\pi f_{\text{c}} R_{\text{e}}} = \frac{1}{2\pi \times 150 \text{ MHz} \times 22 \Omega} = 48 \text{ pF}
\]

\[
C_{\text{e}} = \frac{48 \text{ pF} - 5 \text{ pF}}{5 \text{ pF}} = 43 \text{ pF}
\]

\[
C = 43 \text{ pF} + 115 \times 5 \text{ pF} = 618 \text{ pF}
\]

\[
C_{\text{c}} = \frac{1}{2\pi L_{\text{g}} C_{\text{g}}} = \frac{1}{2\pi \times 639 \Omega \times 618 \text{ pF}} = 403 \text{ kHz}
\]
Consider output (collector) circuit with Miller effect.

\[ \omega_T = 2\pi f_B = \frac{1}{R_T C_T} \text{ where} \]
\[ R_T = R_c H R_L \text{ and } C_T = \frac{1 + \frac{A}{A}}{A} C_c \left( = \frac{A + 1}{A} C_B \right) \]
\[ f_c = \frac{1}{2\pi \times 2.5 \times 10^3 \times \frac{115}{114} \times 5 \times 5 \times 10^{-12}} = 12.6 \text{ MHz} \]

Thus the input circuit dominates and the

\[ Hf \quad f_c = 403 \text{ kHz}. \]

(c) Bandwidth = 403 kHz - 1 kHz = 402 kHz.
MOSFET Amplifier Example

\[ g_m = 0.8 \text{ m}\Omega \]
\[ C_{gs} = 4 \text{ pF} \]
\[ C_{gd} = 2 \text{ pF} \]
\[ C_d = 1 \text{ pF} \]
\[ R_d = 10 \text{ k}\Omega \]
(see Fig. 9.37)

(i) Lower cutoff frequency:
   i) Input circuit (assume \( C_s \to \infty \))

\[
\frac{1}{f_c} = \frac{1}{2\pi(R_s + R_{i1}R_2)C_1} = \frac{1}{2\pi(100 + 50000) \cdot 1 \text{ \mu F}}
\]

\[ f_c = 0.3 \text{ Hz} \]

ii) Output circuit: assume \( C_s \to \infty \) and take \( R_d \) into account (see p. 614):

\[
\frac{1}{f_c} = \frac{1}{2\pi(R_d \pi_1 + R_L)C_2} = \frac{1}{2\pi \times 9.29 \text{ k}\Omega \cdot 1 \text{ \mu F}}
\]

\[ f_c = 17 \text{ Hz} \]
Effect of $C_s$ on Low Frequency Response

(iii) Source circuit HF $f_C$ due to $C_s$ shunt, taking $r_d$ into account (see pp. 615–616):

$$f_C = \frac{1}{2\pi R_{OS} C_s} \left\{ \text{where } R_{OS} = R_s \parallel J_m \parallel r_d, \right.$$  

$$= \frac{1}{2\pi \times 7695 \times 1 \mu F}$$

$$f_C = 207 \text{ Hz}. \quad \text{determines LF -3dB point.}$$

Simple HF JFET Model:

- $C_{gd}$
- $C_{gs}$
- $C_{ds}$
- $r_d$ and $i_f$ necessary (esp. MOSFET)
- May neglect for JFET

$C_{gs}$, $C_{gd}$ &lt; 1–3 pF
$C_{ds}$ &lt; 1 pF
Upper Corner Frequency: Input Stage

Simple HF JFET Model:

\[ f_C = \frac{1}{2\pi \times 100 \times 10 \times 9.7 \text{ pF}} = 164 \text{ MHz} \]

(b) Upper cutoff frequency: See pp. 621-623. The FET circuit is related to the equivalent HF circuit of the BJT common emitter amplifier (see prob. 9.54 solution):

1) Input

\[ f_C = \frac{1}{2\pi RC} \]

\[ f_C = \frac{1}{2\pi \times 100 \times 10 \times 9.7 \text{ pF}} = 164 \text{ MHz} \]

\[ R_g = R_s || R_1 || R_2 = 100 \Omega \]

\[ C_g = C_{gs} + (1 + A)C_{gd} \]

\[ A = g_m [R_d || R_0 || R_L] \]

\[ = 0.8 \text{ m} \Omega \times 2.31 \text{ k} \Omega \]

\[ = 1.85 \text{ k} \Omega \]

\[ C_g = 4 \text{ pF} + 2.85 \times 2 \text{ pF} = 9.7 \text{ pF} \]
Output Stage Upper Corner Frequency

Simple HF JFET Model:

\[ f_c = \frac{1}{2\pi R_T C_T} \]

where

\[ R_T = R_d || R_D || R_L = 2.3 \text{ k}\Omega \]

\[ C_T = C_{ds} + \frac{1+\alpha}{\alpha} C_{gd} \]

\[ = 1 \text{ pF} + \frac{2.85 \times 2 \text{ pF}}{1.85} \]

\[ = 4.1 \text{ pF} \]

\[ f_c = 16.8 \text{ MHz} \]

\[ (c) \quad f_{C_{HIGH}} - f_{C_{LOW}} = 16.8 \text{ MHz} \]

\[ 9.5^T (b) \quad \text{(ii) Output circuit, again related to CE BJT analysis,} \]

\[ f_c = \frac{1}{2\pi R_T C_T} \]

\[ = \frac{1}{2\pi \times 2.3 \text{ k}\Omega \times 4.1 \text{ pF}} \]

\[ = 16.8 \text{ MHz} \]

\[ \text{determine HF cutoff} \]

\[ (c) \quad f_{C_{HIGH}} - f_{C_{LOW}} = 16.8 \text{ MHz} \]
Ways to Improve Amplifier HF Response

- Reduce Miller effect
  - Common Base amplifier (see solution to Problem 9.21)
  - Differential Amp using non-inverting input with inverting input grounded
- Cascode circuit – similar to above
Common Base Detector Amplifier

- No Miller effect since $c_c, c_c$ grounded at base; fast if use fast BJT (small $c_c, c_c$ etc.)
**Can We Understand Amplifier Operation?**

(this page is supplementary material – not on final)

- This is an amplifier for short pulses of width \(\sim 1\) ns.
- Pulse response is covered in 116B (i.e., beyond the scope of this course) but we can understand its operation based on what we have learned so far in Physics 116A plus basic physics.
- We want the output pulse to have a fast risetime (sharp leading edge).
- If \(R_s \approx 50\) \(\Omega\), the input emitter circuit has \(f_c \approx 2\) GHz so the BJT delivers a short **current pulse** at the collector which follows the input voltage: \(i_c(t) \approx \alpha v_{in}(t)/r_e\).
- The collector current is integrated: \(\int pulse i_c(t) \, dt = Q = C v'\) to charge the combined capacitance \(C = 2 C_c\) of the input BJT and the first BJT in the Darlington pair, producing the rapidly rising leading edge of \(v'\) and the output pulse.
  - Integration occurs because the time constant of the BJT collector circuit is \(\tau = RC = 20\, k\Omega \times 2\, pF = 40\) ns, much longer than the input pulse width (assume the base current of the Darlington input transistor is negligible).
- The pulse height of the output pulse is proportional to the charge of the input pulse.
- The output is thus expected to look like the sketch. The tail can be shortened using a “speedup” RC network following the emitter follower output (not shown).
Check Simple Model With SPICE

(this page is supplementary material – not on final)

- Uses MSH10 RF amplifier BJT:
  \( c_c = 1 \text{ pf} \)
  \( c_e = 1.5 \text{ pf} \)

- SPICE BJT model for MSH10 is available from Fairchild website:
  www.fairchildsemi.com

- \( v_{in} = V(5) \) (red)
  \( v_{out} = V(8) \) (green)

- Good agreement with simple model
Other Possibilities to Avoid Miller Effect

- Note the common base circuit lurking in both
BiFET Op-Amp Simplified Diagram

Three-stage amplifier:

- The differential amplifier and common emitter amplifier use the large Thévenin equivalent AC resistance of a current source along with the input resistance of the following stage to achieve large gain. See text, Sec. 10.2.

- Note C1 makes use of the Miller Effect to achieve a large effective capacitance for a dominant low-pass filter.

Current Mirror Circuit

BJTs have identical characteristics

\[
I_{Q_{BE}} = I_{C_2} = \frac{V_{CC} - V_{BE_1}}{R}
\]
Effect of Dominant Low-Pass Filter

\[
A = \frac{A_0}{1 + \frac{j\omega}{\omega_c}}
\]

\[
A_F = \frac{A}{1 + A_0 B}
\]

\[
= \frac{A_0 / (1 + j\omega / \omega_c)}{1 + A_0 B / (1 + j\omega / \omega_c)}
\]

\[
= \frac{A_0}{(1 + A_0 B) \left[ 1 + \frac{j\omega / \omega_c}{1 + A_0 B} \right]}
\]

\[
= \frac{A_0 F}{1 + j\omega / \omega_{c_F}}
\]

Finally,
\[
A_{OF} \omega_{c_F} = \frac{A_0 \omega_c}{1 + A_0 B} \left( 1 + A_0 B \right) = A_0 \omega_c
\]

Upper corner frequency is multiplied by \((\text{loop gain} + 1)\)≈loop gain

- The product of gain and bandwidth is constant.
- The maximum phase shift is 90° (won’t oscillate for resistive B).
- High frequency performance is compromised.
**Bode Plot: \( A_F \) and Bandwidth**

\[ \text{On Bode Plot} \quad (A_0 = 20 \text{ kHz} = 106 \text{ dB}) \]

- \( f_c = 5 \text{ Hz} \)
- \(-3 \text{ dB} @ 5 \text{ Hz} \)

When \( A \gg 1 \) at low freq,

\[ A_F = \frac{A_0}{1 + A_0B} \approx \frac{A_0}{A_0B} \]

\[ A_0(\text{dB}) = A_F(\text{dB}) + (A_0B)\text{dB} \]

**At low frequencies** in the example above,

\( A_F(\text{dB}) = 26 \text{ dB} \)

\[ A_F \approx \frac{A}{AB} \text{ as long as } A \gg \frac{1}{B} \text{ and } A_F(\text{dB}) \approx A(\text{dB}) - (AB)(\text{dB}) \]
741 Specifications

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<tr>
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<td>typ. 80-100 dB</td>
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</tbody>
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- Output protected against short-circuits
- Input offset voltage can be balanced out with external pot
- See Sec. 10.2 for details
No dominant pole: has 3 low-pass filters in series. Amplifier phase shift >180° with significant gain and can oscillate with a resistive feedback network.

- No large $C_1$: Could make amplifier with BW of several MHz.
- Considerable gain left when phase shift equals 180 degrees at 12.5 MHz.
- Not fool-proof: A unity gain voltage follower would oscillate.

Figure 15-29 Open-loop gain and phase-shift characteristics of the $\mu$A702A.
BJT CE Large Signal Performance

Common Emitter Amplifier

- The maximum output voltage swing is set by BJT cutoff and saturation
- Start with the BJT curves of $I_C$ vs. $V_{CE}$ for various values of $I_B$, locate Q point
- Draw straight line through Q point with slope $dI_C/dV_{CE}$ for midband AC signals (AC Load Line) to determine useful range
- For AC, $v_c = -R_Ci_c$ so AC load line slope $= i_c/v_c = -1/R_C$ in this case. Output voltage swing follows AC load line.
CE Amplifier: DC and AC Load Lines

- Max. symmetrical voltage swing when Q point centered on AC load line
- At Q, no input, BJT power dissipation $p \approx V_{CE}I_C = 4 \text{ V} \times 2 \text{ mA} = 8 \text{ mW}$
- If the Q point is centered, the average power dissipated by the BJT is max. with no AC input – actually less when producing a signal. (See sec. 9.4 in text for details)
- Base bias chain keeps both BJTs just at cutoff (or slightly “on”) at Q point – \( \approx \) No BJT power dissipated if no input signal.

- AC input causes one or the other BJT to provide the output.

- Maximum average BJT power now \( 0.1V_{CEq}i_C(sat) \) – much more efficient use of BJTs and power – useful for driving low impedance loads at high power.