# Revision

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# *Outline*

✓ Bipolar Junction Transistor (Lecture 02)

- Introduction – Forward-Active Mode Operation
- CE Current-Voltage Characteristics
- DC Analysis of Transistor Circuits
- Small-Signal Hybrid–$\pi$ Equivalent Circuit
- Small-Signal Voltage Gain
- Hybrid–$\pi$ Equivalent Circuit, including Early Effect

# *Outline*

✓ BJT Amplifiers' (Lecture 03)

- Common-Emitter (CE) Amplifiers
- Common-Collector (CC) or Emitter-Follower Amplifier
- Common-Base (CB) Amplifier

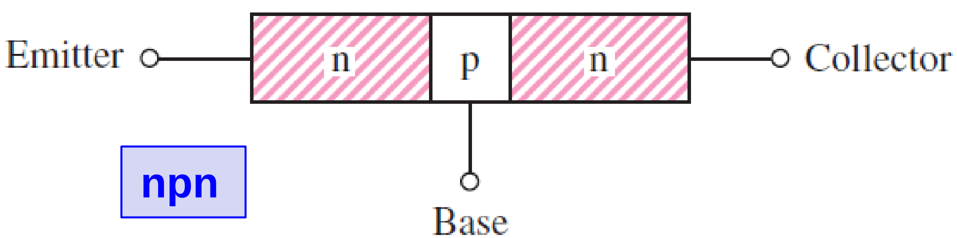
✓ Expanded Hybrid–$\pi$ Equivalent Circuit (Lecture 04)

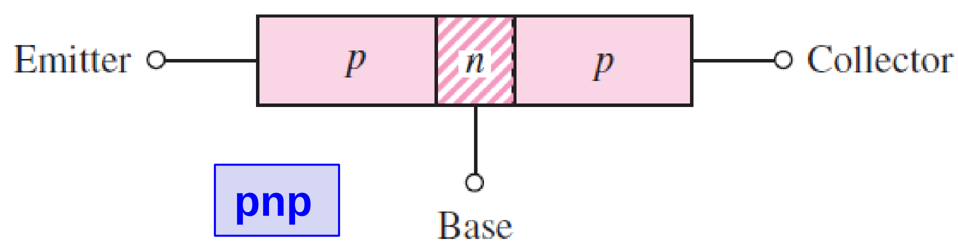
- Short-Circuit Current Gain
- Cutoff Frequency
- Miller Effect and Miller Cpacitance

# Basic Bipolar Junction Transistor (BJT)

Transistor principle is that *the voltage between two terminals controls the current through the third terminal*.

- ✓ BJT has 3 separately doped regions & contains 2 pn junctions. A single pn junction has two modes of operation – forward and reverses biases.
- ✓ Bipolar transistor with 2 pn junctions, therefore has 4 possible modes of operation, which is one reason for the versatility of the device.
- ✓ With 3 separately doped regions, the bipolar transistor is a 3 terminal device. Current in the transistor is due to the flow of both electrons and holes, hence the name *bipolar*.

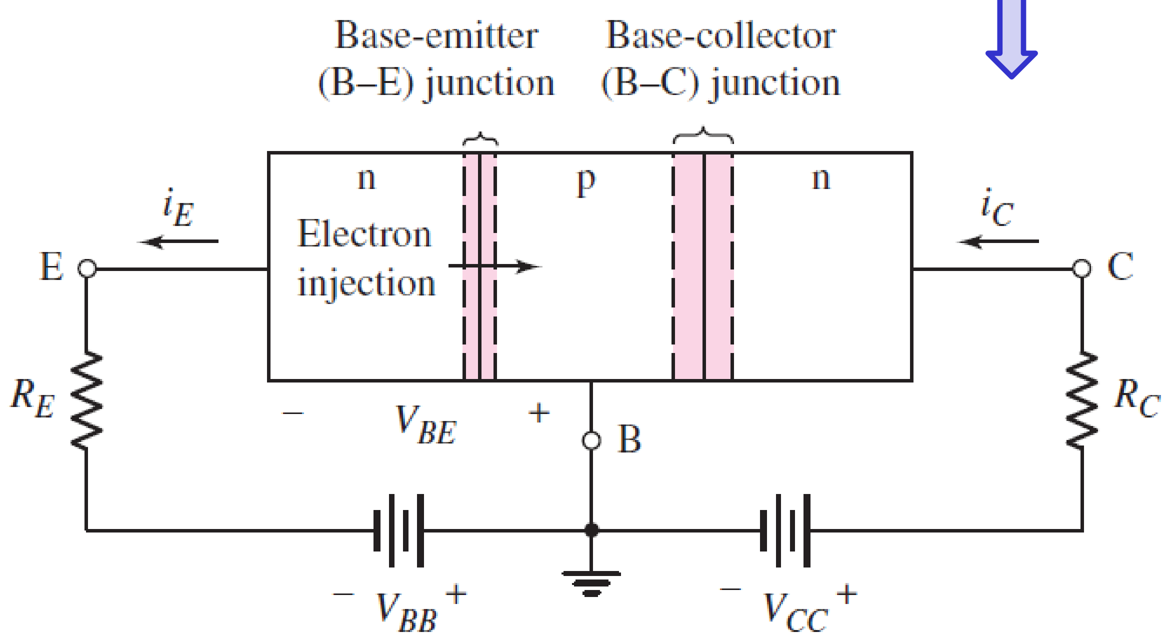


# *Forward-Active Mode Operation*

- ✓ Since the transistor has 2 pn junctions, 4 possible bias combinations may be applied to the device, depending on whether a forward or reverse bias is applied to each junction.
- ✓ For example, if the transistor is used as an amplifying device, the **base-emitter (B-E) junction** is forward biased and the **base-collector (B-C) junction** is reverse biased, in a configuration called *forward-active operating mode,* or simply called the *active region*.

Since B-E junction is forward-biased, electrons from the emitter are injected across B-E junction into the base, creating excess minority carrier concentration in base. Since B-C junction is reverse-biased, the electron concentration at the edge of junction is 0.



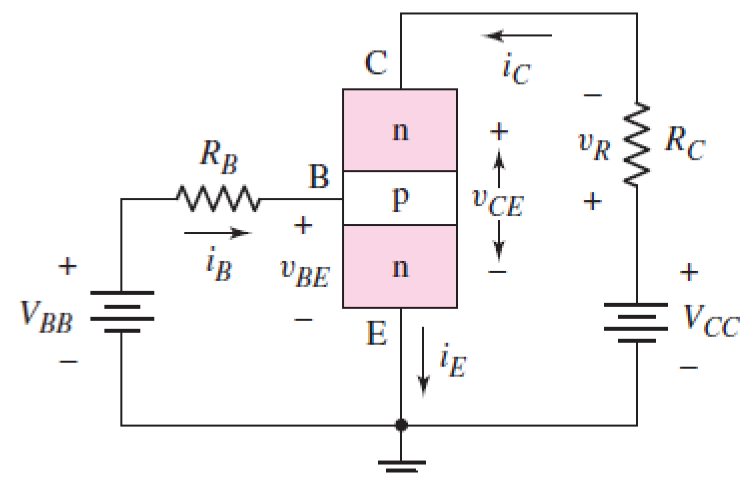
# Forward-Active Mode Operation

## Common-Emitter (CE) Current Gain:

- ✓ If bipolar transistor is considered as a single node, by using KCL, we get,

$$i_E = i_C + i_B$$

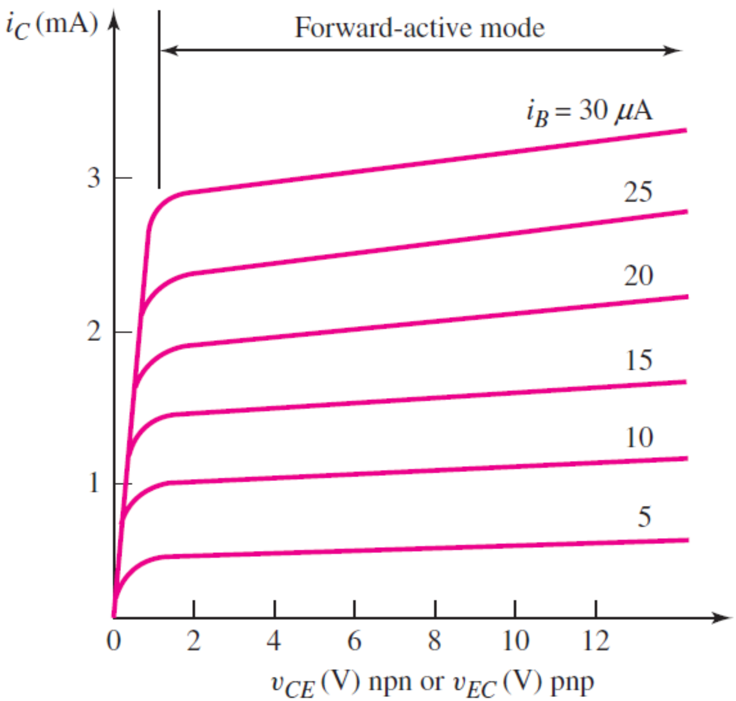
- ✓ If the transistor is biased in the forward-active mode, then $i_C = \beta i_B$.
- ✓ From the above equations, we can get the relationship, $i_E = (1+\beta)i_B$.
- ✓ Moreover, $i_C$ & $i_E$ are related as, $i_C = \left(\frac{\beta}{1+\beta}\right) i_E$
- ✓ Recall $i_C = \alpha i_E \rightarrow \alpha = \frac{\beta}{1+\beta}$

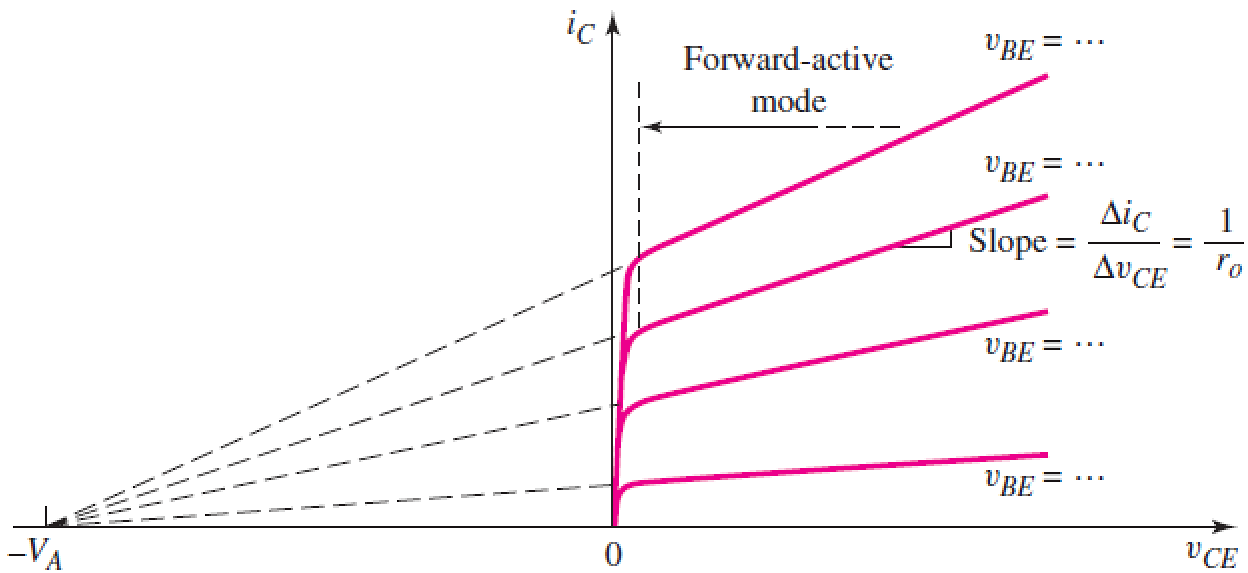



**Common-emitter configuration**

# CE Circuit Current-Voltage Characteristics

The collector current is plotted against the collector–emitter voltage, for various constant values of the base current.




**Exaggerated view plotted for constant values of B-E voltage**

The slope in these characteristics is due to an effect called base-width modulation that was first analyzed by J. M. Early – called *Early effect*.

# CE Circuit Current-Voltage Characteristics

- When the curves are extrapolated to zero current, they meet at a point on the negative voltage axis, at $v_{CE} = -V_A$. The voltage $V_A$ is positive, called *Early voltage*. Typical values are in the range $50 < V_A < 300\,V$.

- The linear dependence of $i_C$ vs $v_{CE}$ in the forward-active mode can be as,

$$i_C = I_S\, e^{v_{BE}/V_T}\left(1 + \frac{v_{CE}}{V_A}\right)$$

- Non-zero slope of the curves indicates that the *output resistance* $r_o$ looking into the collector is finite. This $r_o$ is determined from, $\frac{1}{r_o} = \left.\frac{\partial i_C}{\partial v_{CE}}\right|_{v_{BE=Const}}$

- Therefore, we can show that, $r_o \cong \frac{V_A}{I_C}$

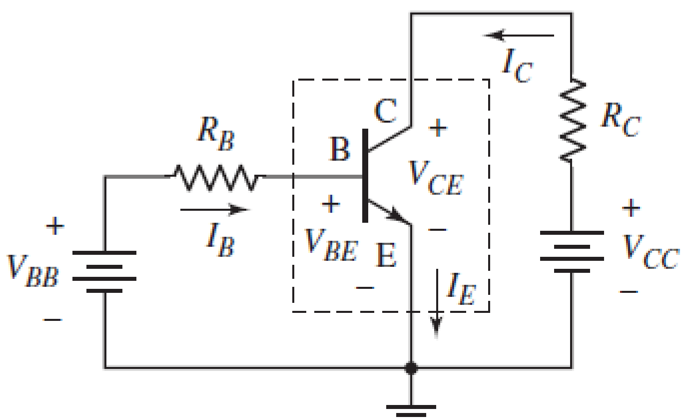
$I_C$ is the quiescent collector current when $v_{BE}$ is constant and $v_{CE} < v_A$.

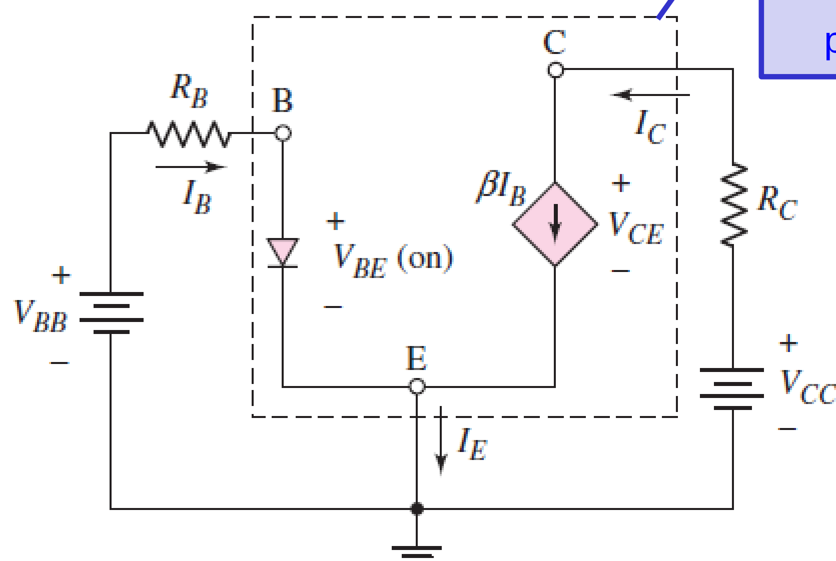
# DC Analysis of Transistor Circuits

- ✓ Now analyze and design the dc biasing of transistor circuits – an important part of designing bipolar amplifiers which is the focus of later syllabus.
- ✓ The *piecewise linear model of a pn junction* can be used for the dc analysis of bipolar transistor circuit.
- ✓ One of the basic transistor circuit configuration is *common-emitter (CE) circuit* – the *emitter terminal* is obviously at *ground potential*.

Piecewise linear transistor parameters

CE circuit with npn transistor




DC equivalent circuit

# DC Analysis of Transistor Circuits

- ✓ Assume that the B-E junction is forward-biased so that the voltage drop across that junction is cut-in or turn-on voltage $V_{BE}(on)$.
- ✓ When the transistor is biased in forward-active mode, the *collector current* is represented as a dependent current source, function of base current.
- ✓ Neglect the reverse-biased junction leakage current and Early effect.

The base current is, $I_B = \frac{V_{BB} - V_{BE}(on)}{R_B}$

It is implicit that $V_{BB} > V_{BE}(on) \rightarrow I_B > 0$. If $V_{BB} < V_{BE}(on)$, transistor is cut off & $I_B = 0$.

In the collector-emitter portion, we can write

$I_C = \beta I_B$, and $V_{CC} = I_C R_C + V_{CE}$ (or) $V_{CE} = V_{CC} - I_C R_C$

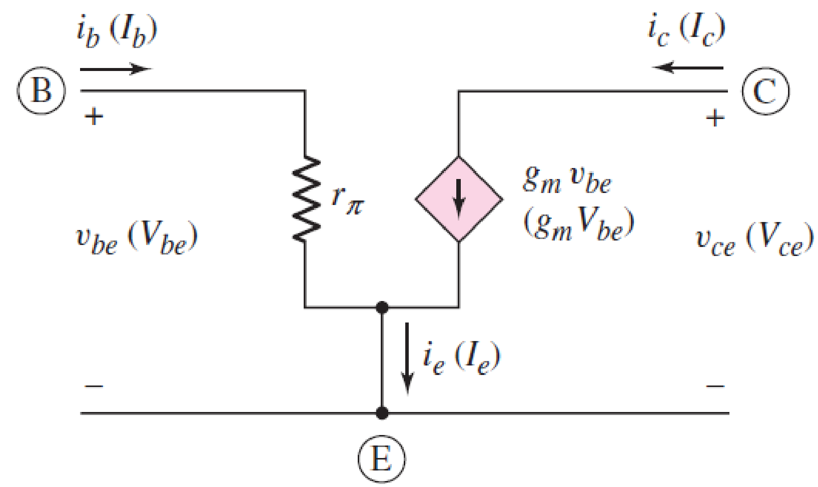
It is also implicit that $V_{CE} > V_{BE}(on)$, which means B-C junction is reverse biased and the transistor is biased in the forward-active mode.

# Hybrid–π Equivalent Circuit

We can now *develop an equivalent circuit model* of BJT – voltage controlled current source and explicitly includes input resistance looking into base ($r_\pi$).

At emitter node, we have

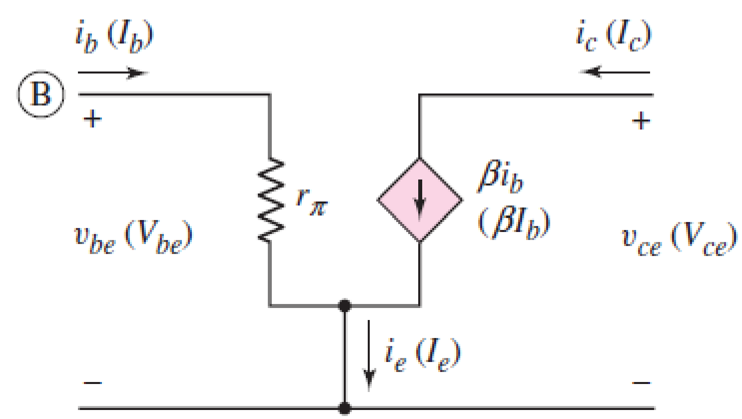
$$i_e = \frac{v_{be}}{r_\pi} + g_m v_{be} = \frac{v_{be}}{r_\pi}(1 + g_m r_\pi) = \frac{v_{be}}{r_\pi}(1 + \beta)$$




We can relate small-signal collector current to the small-signal base current as,

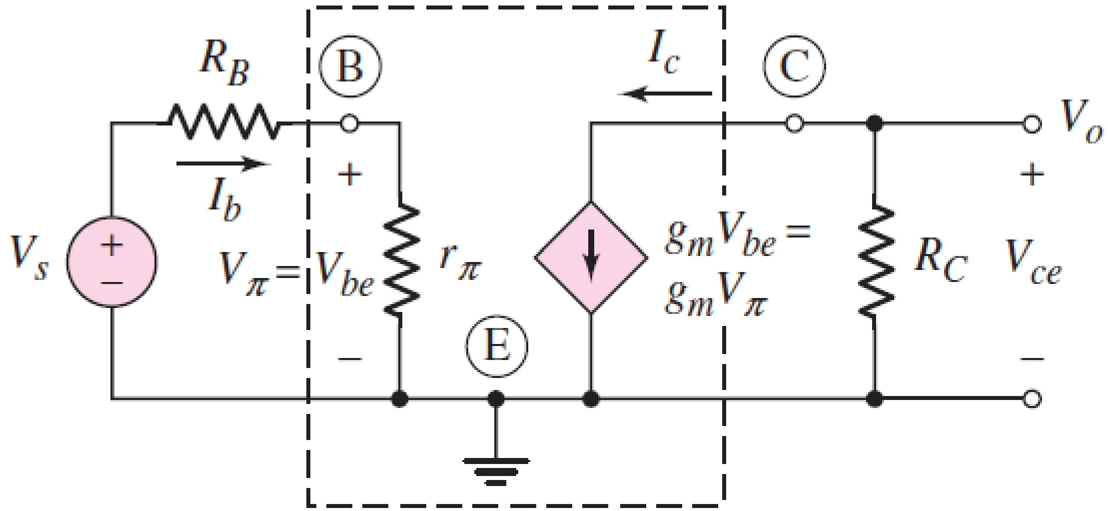
$$i_c = g_m v_{be} = g_m(i_b\, r_\pi) = (g_m r_\pi) i_b = \beta i_b$$

Consider, $g_m r_\pi = \frac{I_{CQ}}{V_T} \times \frac{\beta V_T}{I_{CQ}} = \beta$



# Small-Signal Voltage Gain

We can now *incorporate the small-signal hybrid–π* model into the ac equivalent circuit – replace the equivalent model of the transistor.




Small-signal voltage gain

$$A_v = \frac{V_o}{V_s} = \frac{V_{ce}}{V_s} = \frac{-(g_m V_\pi) R_C}{V_\pi / \left(\frac{r_\pi}{r_\pi + R_B}\right)} = -(g_m R_C)\left(\frac{r_\pi}{r_\pi + R_B}\right)$$

Note that, $V_\pi = \left(\frac{r_\pi}{r_\pi + R_B}\right) V_s$

Voltage divider rule

# Problem-Solving Technique

Since we are dealing with linear amplifier circuits, superposition applies, which means that we can perform the *dc* and *ac* analyses separately. The procedure is as follows:

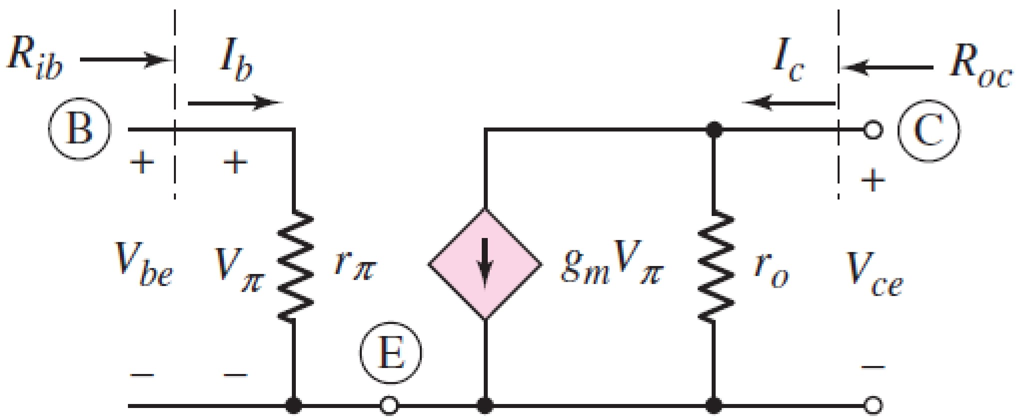
1. Analyze the circuit with *only the dc sources present*. This solution is the dc or quiescent solution, which uses the dc signal models. The transistor must be biased in the forward-active region in order to produce linear amplifier.

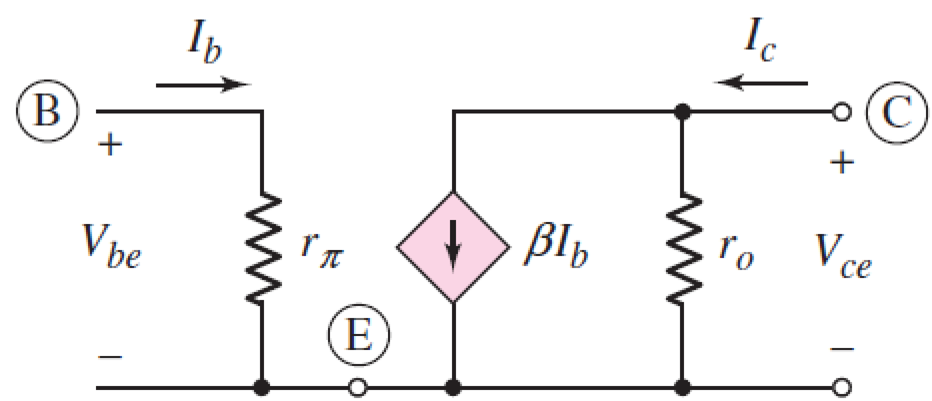
2. Replace each element in the circuit with its *small-signal model*. The small-signal model *hybrid*$-\pi$ applies to the transistor.

3. Analyze the small-signal equivalent circuit, *setting the dc source components equal to zero*, to produce the response of the circuit to the time-varying input signals only.

# Hybrid–π Equivalent Circuit with Early Effect

So far we have assumed that the collector current is independent of the collector-emitter voltage in the small-signal equivalent circuit. Recall Early Effect, where the collector current varies with collector-emitter voltage.




$i_C = I_S\, e^{v_{BE}/V_T}\left(1 + \frac{v_{CE}}{V_A}\right)$ & $r_o \cong \frac{V_A}{I_C}$ is *small-signal transistor output resistance*.

This resistance is an equivalent *Norton resistance*, which means that $r_o$ is in parallel with the dependent current sources.

**Input resistance, $r_\pi$ (ohms); current gain, $\beta$ (dimensionless)**
**output resistance, $r_o$ (ohms); transconductance, $g_m$ (mhos);**

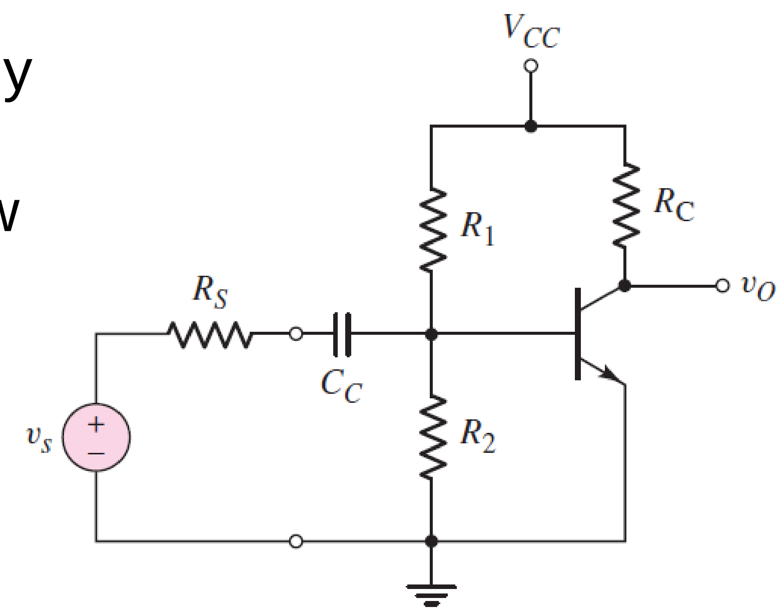
# Basic BJT Amplifiers (Lecture 03)

# Common-Emitter (CE) Amplifier

- ✓ Note that the *emitter is at ground potential* – hence called common-emitter.
- ✓ Signal from the signal source is coupled into the base of the transistor through the coupling capacitor $C_C$, which provides *dc isolation* between the amplifier and the signal source.
- ✓ The *DC transistor biasing* is established by $R_1$ and $R_2$, and is not disturbed when the signal source is capacitively coupled to the amplifier.

Assume that the signal frequency is 1) sufficiently high that any *coupling capacitance* acts as a perfect short circuit, and 2) is also sufficiently low that the *transistor capacitances* are neglected.

Neglect any capacitance effects within the transistor.



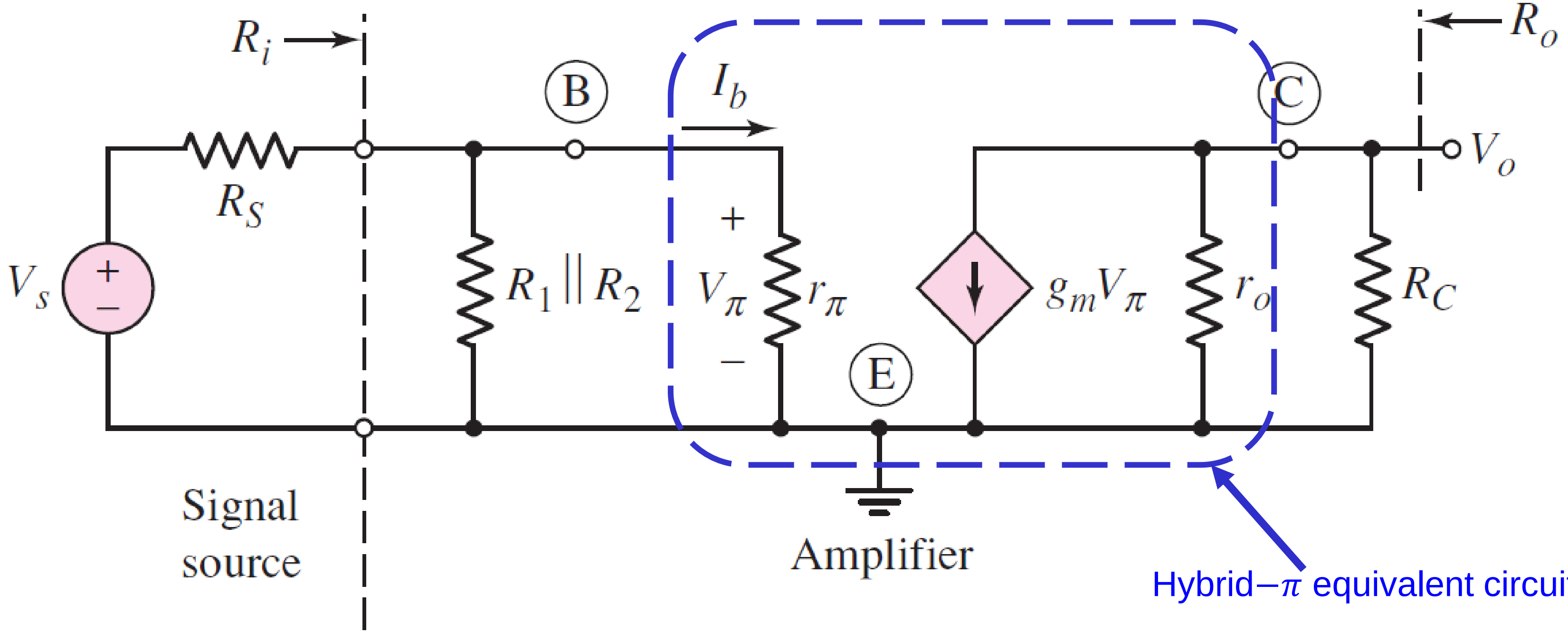

**Basic CE circuit with coupling capacitor**

# Common-Emitter – Small-signal Circuit

The output voltage can be written as, $V_o = -g_m V_\pi (r_o || R_C)$

The control voltage $V_\pi$ is found to be, $V_\pi = \frac{R_1||R_2||r_\pi}{R_1||R_2||r_\pi + R_S} \times V_s$

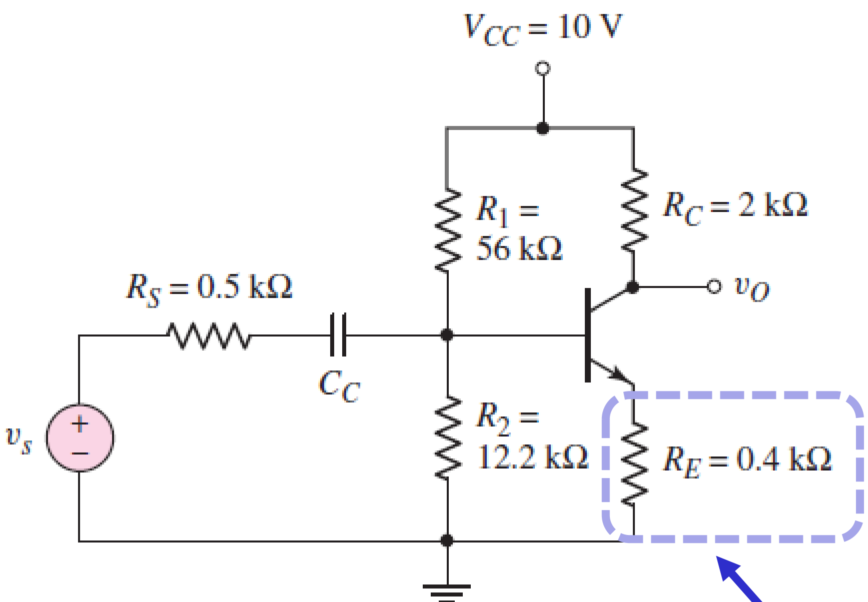
Thus, the small-signal voltage gain is, $A_v = \frac{V_o}{V_s} = -g_m (r_o || R_C) \frac{R_1||R_2||r_\pi}{R_1||R_2||r_\pi + R_S}$



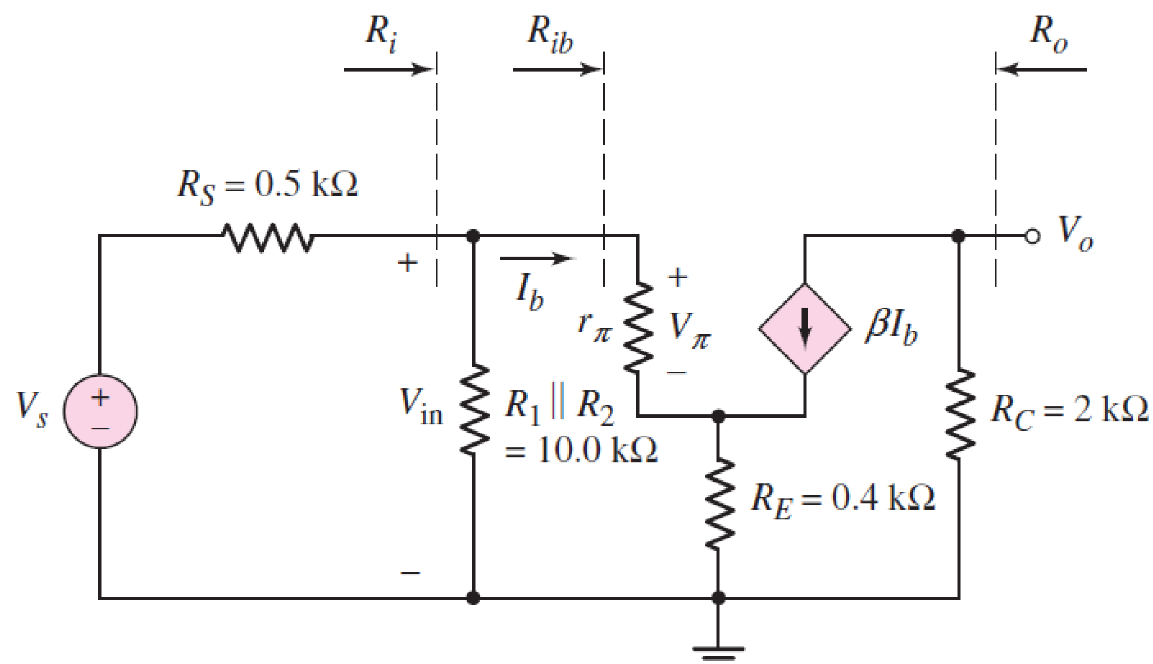

Small-signal equivalent circuit assuming coupling capacitor is a short circuit

# Common-Emitter with Emitter Resistor

- ✓ The earlier CE circuit is not very practical – voltage across $R_2$ provides the base-emitter voltage to bias the transistor in the forward-active region.
- ✓ A slight variation in the resistor value or in the transistor characteristics may cause the transistor to be biased in cutoff or saturation.
- ✓ *Although the emitter is not at ground potential, it is still called as CE circuit*.




Emitter resistor




**Small-signal equivalent circuit**

# Common-Emitter with Emitter Resistor

Note that current gain $\beta$ is used in the equivalent circuit & assume that Early voltage is infinite so the transistor output resistance $(r_o)$ can be neglected.

**Input resistance $R_{ib}$**: It is the input resistance looking into the base

Use KVL for the loop, $V_{in} = I_b r_\pi + (I_b + \beta I_b)R_E \rightarrow R_{ib} = \frac{V_{in}}{I_b} = r_\pi + (1+\beta)R_E$

**Voltage gain $A_v$**:

The output voltage is, $V_o = -(\beta I_b)\ R_C$

The input resistance to the amplifier, $R_i = R_1 || R_2 || R_{ib}$

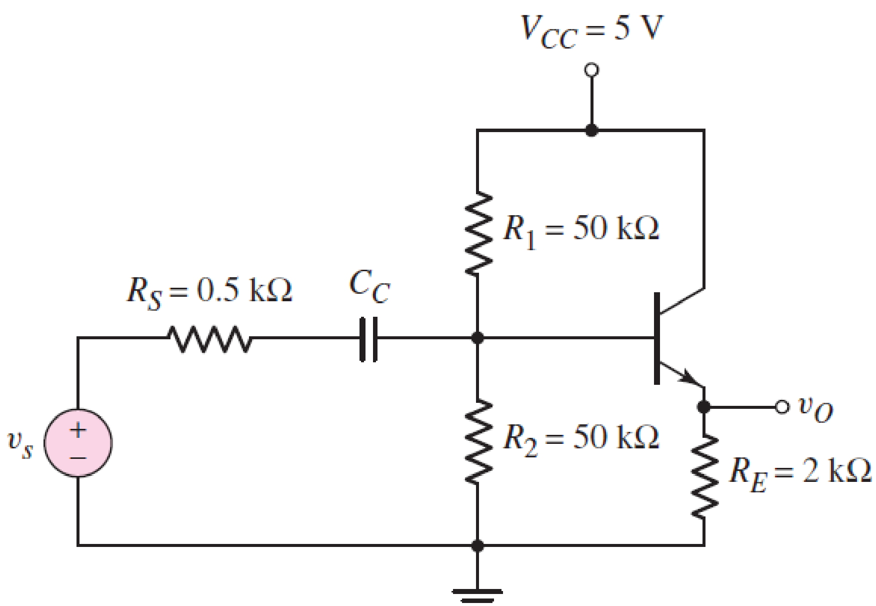
Moreover, $V_{in} = \left(\frac{R_i}{R_i + R_S}\right) V_s$

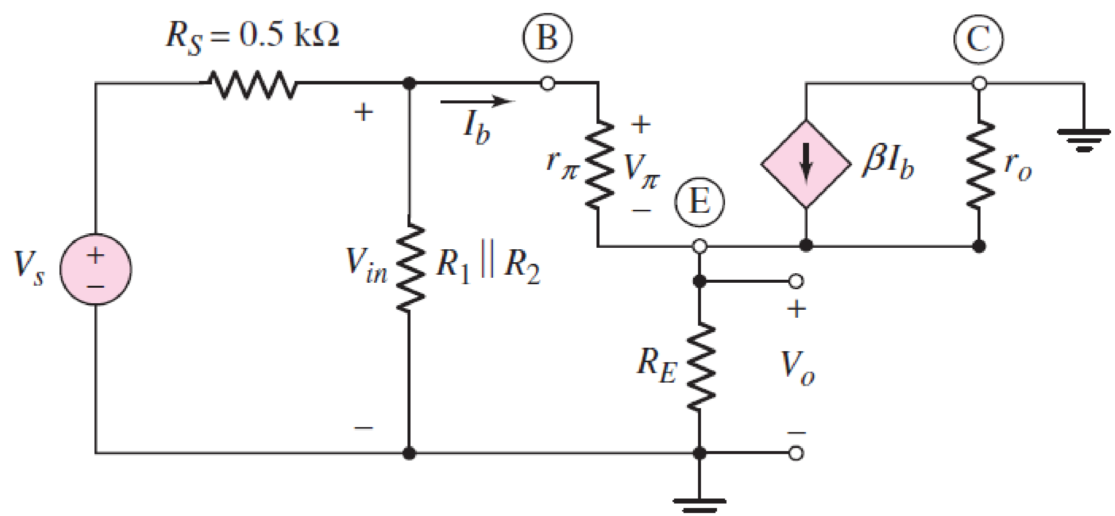
Therefore, $A_v = \frac{V_o}{V_s} = \frac{-(\beta I_b)\ R_C}{V_s} = -\beta R_C \left(\frac{V_{in}}{R_{ib}}\right)\left(\frac{1}{V_s}\right) = \frac{-\beta R_C}{r_\pi + (1+\beta)R_E}\left(\frac{R_i}{R_i + R_S}\right)$

If $R_i \gg R_S$ & $(1+\beta)R_E \gg r_\pi \rightarrow A_v \cong \frac{-\beta R_C}{(1+\beta)R_E} \cong \frac{-R_C}{R_E}$

# Common-Collector (CC) Amplifier

- ✓ The output signal is taken off of the emitter with respect to ground and the collector is connected directly to $V_{CC}$. Since $V_{CC}$ is at signal ground in the ac equivalent circuit (*see*) – named as common-collector (*Emitter follower*).
- ✓ Equivalent circuit – assume the coupling capacitor $C_C$ acts as a short circuit. Collector terminal is at signal ground & the transistor output resistance $r_o$ is in parallel with the dependent current source.




Small-signal equivalent circuit

# Common-Collector (CC) Amplifier

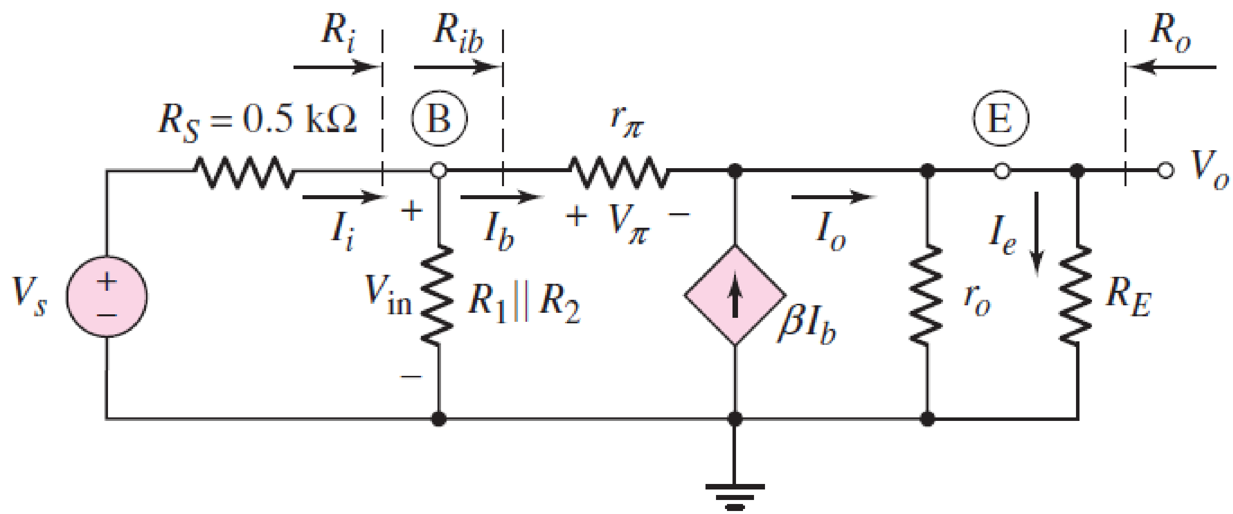
From the equivalent circuit,

$$I_o = (1+\beta)I_b$$

$$V_o = I_b(1+\beta)(r_o||R_E)$$

KVL for base-emitter loop,

$$V_{in} = I_b[r_\pi + (1+\beta)(r_o||R_E)]$$




**Small-signal equivalent circuit with all signal grounds connected together**

$$R_{ib} = \frac{V_{in}}{I_b} = r_\pi + (1+\beta)(r_o||R_E) \Longrightarrow$$ Input resistance looking into the base

We also write, $V_{in} = \left(\frac{R_i}{R_i+R_S}\right)V_s$ ; where, $R_i = R_1||R_2||R_{ib}$.

Small-signal voltage gain, $A_v = \frac{V_o}{V_s} = \frac{(1+\beta)(r_o||R_E)}{r_\pi+(1+\beta)(r_o||R_E)}\left(\frac{R_i}{R_i+R_S}\right)$

# Common-Collector (CC) Amplifier

Small-signal current gain, $A_i = \frac{I_e}{I_i}$

Using current divider rule, $I_b = \left(\frac{R_1||R_2}{R_1||R_2+R_{ib}}\right) I_i$

Since, $g_m V_\pi = \beta I_b$, then, $I_o = (1+\beta) I_b = (1+\beta)\left(\frac{R_1||R_2}{R_1||R_2+R_{ib}}\right) I_i$

Write the load current in terms of $I_o$ produces, $I_e = \left(\frac{r_o}{r_o+R_E}\right) I_o$

Therefore, small-signal current gain, $A_i = \frac{I_e}{I_i} = (1+\beta)\left(\frac{R_1||R_2}{R_1||R_2+R_{ib}}\right)\left(\frac{r_o}{r_o+R_E}\right)$

If we assume $R_1||R_2 \gg R_{ib}$ and $r_o \gg R_E$, then $A_i \cong (1+\beta)$

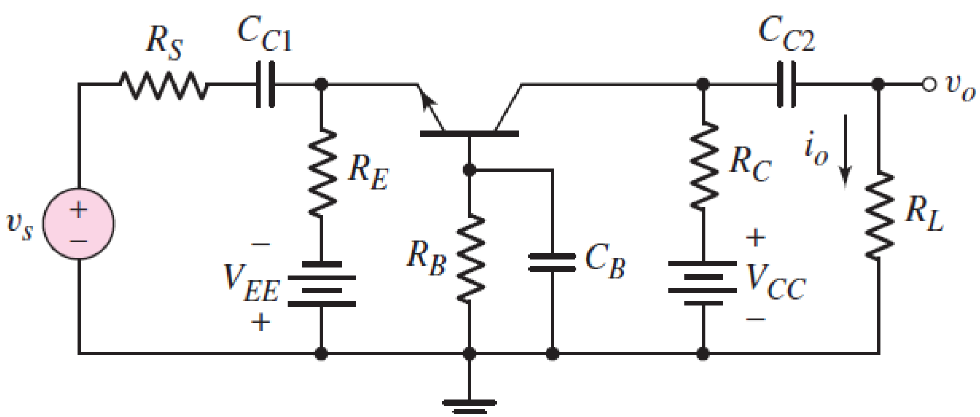
*Although small-signal voltage gain of CC amplifier is slightly less than 1, the small-signal current gain is normally greater than 1.*

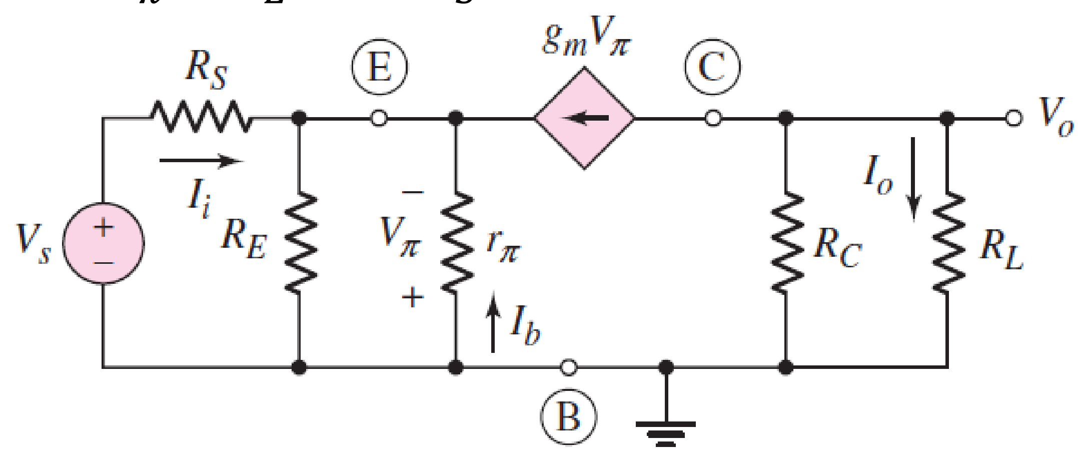
# Common-Base (CB) Amplifier

- ✓ Base is at signal ground & input signal is applied to emitter – *Common-Base*
- ✓ Assume the load is connected to the output through coupling capacitor $C_{C2}$.
- ✓ Assume output resistance $r_o$ to be infinite. The small-signal equivalent circuit of a CB configuration with hybrid$-\pi$ model is complex.

Small-signal output voltage, $V_o = -(g_m V_\pi)(R_C || R_L)$

KCL at the emitter node gives, $g_m V_\pi + \frac{V_\pi}{r_\pi} + \frac{V_\pi}{R_E} + \frac{V_S - (-V_\pi)}{R_S} = 0$



Small-signal equivalent circuit

# Common-Base (CB) Amplifier

Since $\beta = g_m r_\pi$, the above equation is $V_\pi \left(\frac{1+\beta}{r_\pi} + \frac{1}{R_E} + \frac{1}{R_S}\right) = -\frac{V_s}{R_S}$

Then, $V_\pi = -\frac{V_s}{R_S}\left[\left(\frac{r_\pi}{1+\beta}\right) || R_E || R_S\right]$

Substitute the control voltage $V_\pi$ in the output voltage equation, which results

Small-signal voltage gain, $A_v = \frac{V_o}{V_s} = g_m \left(\frac{R_C || R_L}{R_S}\right)\left[\left(\frac{r_\pi}{1+\beta}\right) || R_E || R_S\right]$

If $R_S \to 0$, the voltage gain becomes,

$$A_v = g_m (R_C || R_L)$$

For CB circuit, the small-signal voltage gain is usually greater than 1.

# Common-Base (CB) Amplifier

Small-signal current gain, $A_i = \frac{I_o}{I_i}$

Write KCL at the emitter node, we get, $I_i + \frac{V_\pi}{r_\pi} + g_m V_\pi + \frac{V_\pi}{R_E} = 0$

Solving for $V_\pi$ gives, $V_\pi = -I_i \left[ \left( \frac{r_\pi}{1+\beta} \right) || R_E \right]$

The load current, $I_o = -(g_m V_\pi) \left( \frac{R_C}{R_C + R_L} \right)$

Therefore, the small-signal current gain can be written as

$$A_i = \frac{I_o}{I_i} = g_m \left( \frac{R_C}{R_C + R_L} \right) \left[ \left( \frac{r_\pi}{1+\beta} \right) || R_E \right]$$

$$A_i = \frac{I_o}{I_i} = \frac{g_m r_\pi}{1+\beta} = \frac{\beta}{1+\beta} \quad \text{if } R_E \to \infty \;\&\; R_L \to 0$$

Small-signal current gain is slightly less than 1.

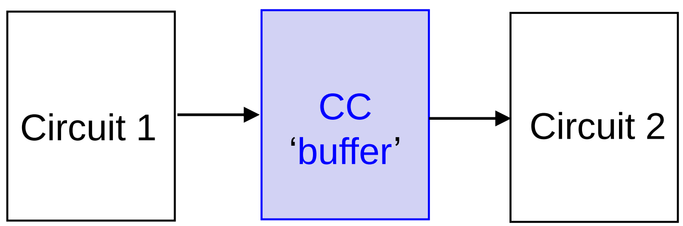
# Comparison of Three Amplifiers

Characteristics of the three BJT amplifier configurations

| Configuration | Voltage gain | Current gain | Input resistance | Output resistance |
|---|---|---|---|---|
| Common emitter | $A_v > 1$ | $A_i > 1$ | Moderate | Moderate to high |
| Emitter follower | $A_v \cong 1$ | $A_i > 1$ | High | Low |
| Common base | $A_v > 1$ | $A_i \cong 1$ | Low | Moderate to high |

*CC circuit* has very high input resistance, low output resistance, and $A_v \cong 1$.

✓ It is often used to isolate two circuits from each other, so circuit 2 does not draw current from circuit 1 – useful as *buffer*.




High input impedance means it draws very little current from circuit 1 & is also able to drive circuit 2 easily from its low output impedance & high current gain. A voltage gain of nearly unity means the signal from circuit 1 is passed onto circuit 2 unchanged.

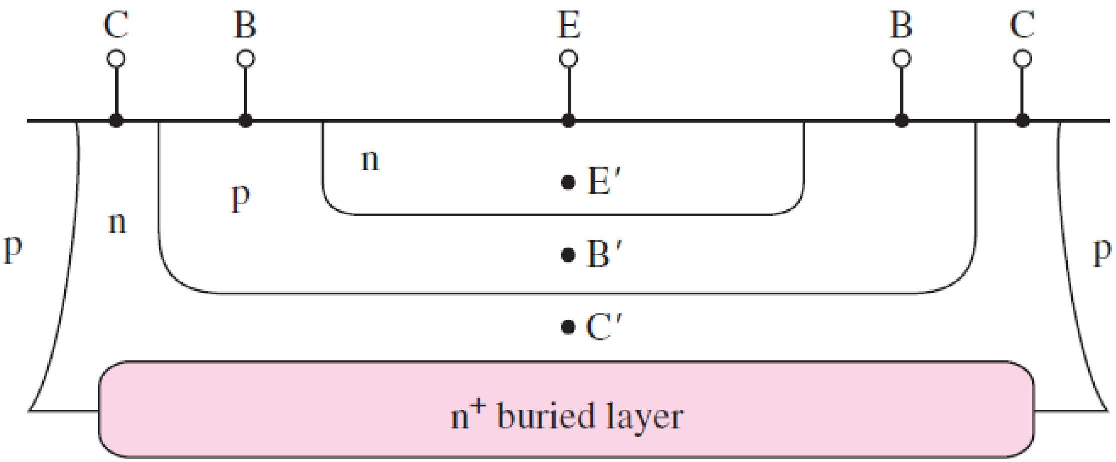
# Frequency Response of Amplifier Circuits (Lecture 03)

# *Expanded Hybrid−π equivalent Circuit*

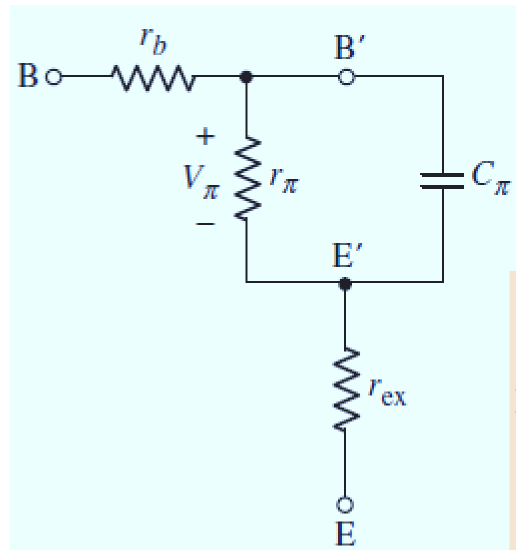
See the cross section of an npn bipolar transistor for hybrid−$\pi$ model.

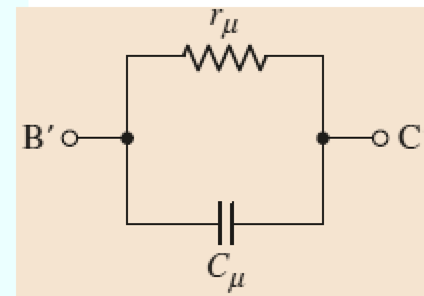
✓ $C, B$, and $E$ terminals are external connections to the transistor, and $C', B'$ & $E'$ are idealized internal collector, base, and emitter regions.

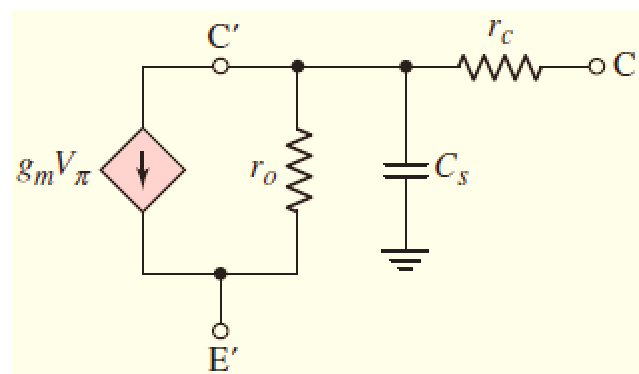
$r_b$ is base series resistance between $B$ & $B'$.
$C_\pi$ is forward biased junction capacitance.
$r_\pi$ is forward biased diffusion resistance.
$r_{ex}$ is emitter series resistance $E'$ & $E$.




**Cross section of npn transistor for hybrid−$\pi$ model**



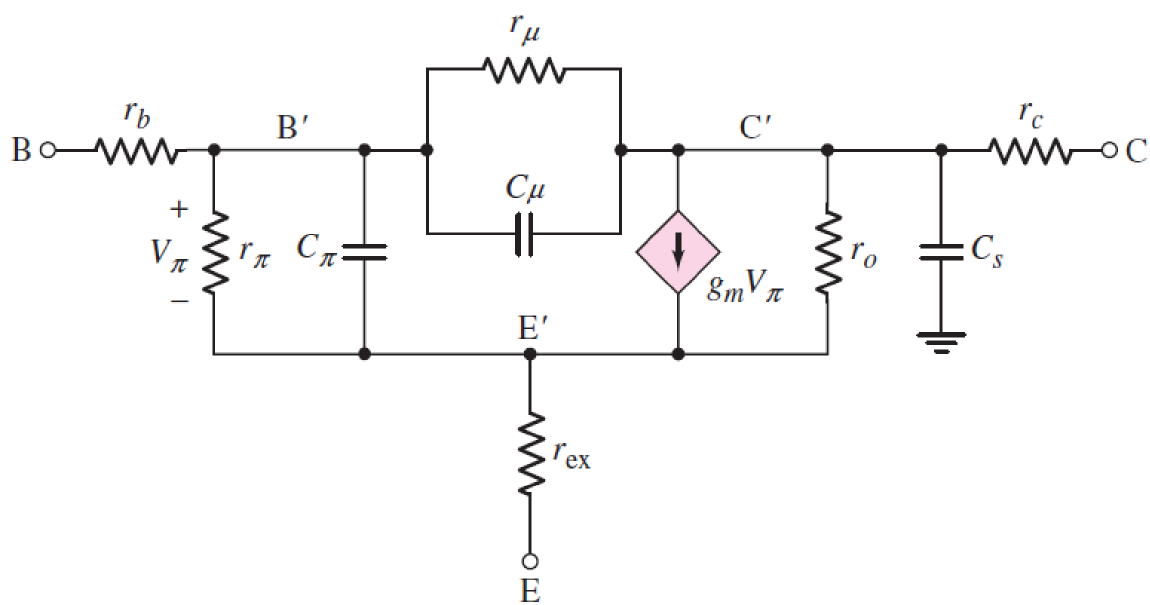



# Expanded Hybrid–π equivalent Circuit

$r_c$ is collector series resistance between $C$ & $C'$.
$C_s$ is junction capacitance of reverse-biased collector-substrate junction.
$g_m V_\pi$ is collector current controlled by internal base-emitter voltage.
$r_o$ is inverse of output conductance $g_o$ and is due primarily to Early effect.

$r_\mu$ is reverse-biased diffusion resistance (range of megohms & neglected).
$C_\mu$ is reverse-biased junction capacitance (normally < $C_\pi$, however, *cannot be neglected due to Miller effect*).



# Short-Circuit Current Gain

Neglect the parasitic resistances $r_b, r_c, r_{ex}$, & $r_\mu$, and substrate capacitance $C_s$.

Write KCL at the input node, $I_b = \frac{V_\pi}{r_\pi} + \frac{V_\pi}{\frac{1}{j\omega C_\pi}} + \frac{V_\pi}{\frac{1}{j\omega C_\mu}} = V_\pi\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]$

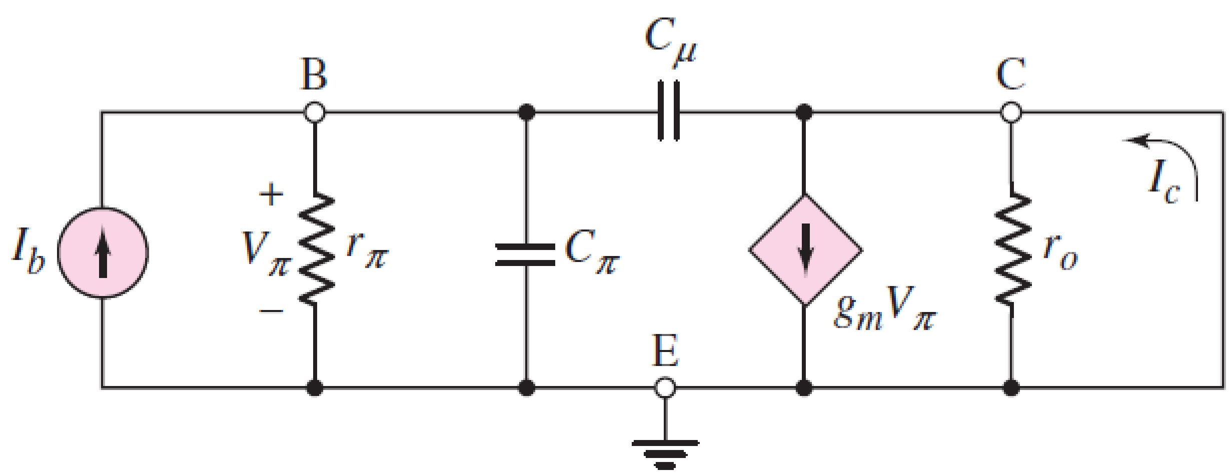
Note that $V_\pi \neq I_b r_\pi$, since a portion of $I_b$ is now shunted through $C_\pi$ & $C_\mu$. From KCL at the output node,

$$\frac{V_\pi}{\frac{1}{j\omega C_\mu}} + I_c = g_m V_\pi \;\rightarrow\; I_c = V_\pi(g_m - j\omega C_\mu) \;\rightarrow\; V_\pi = \frac{I_c}{(g_m - j\omega C_\mu)}$$

Substitute $V_\pi$ in $I_b$ results as,

$$I_b = I_c \times \frac{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]}{g_m - j\omega C_\mu}$$

$$\therefore\; A_i = \frac{I_c}{I_b} = \frac{(g_m - j\omega C_\mu)}{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]}$$



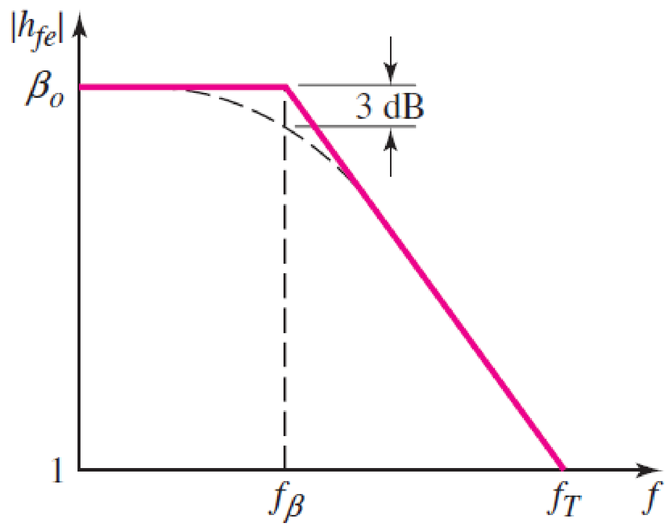

**Simplified hybrid−$\pi$ equivalent circuit**

# Short-Circuit Current Gain

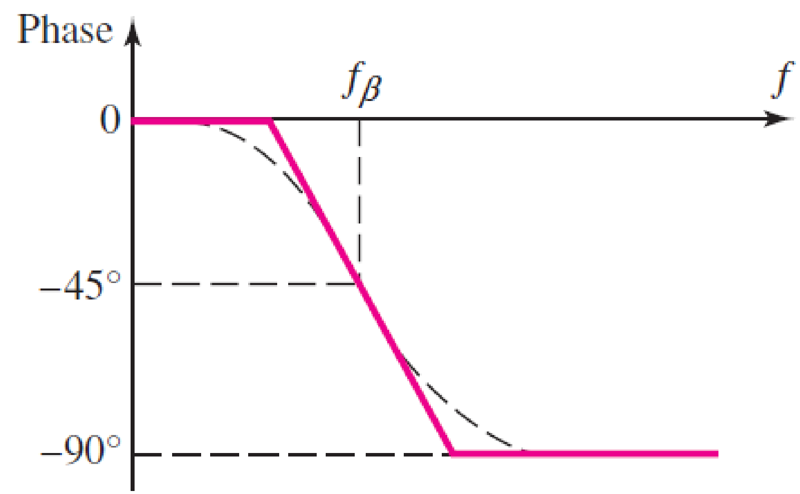
$$A_i = h_{fe} = \frac{I_c}{I_b} = \frac{(g_m - j\omega C_\mu)}{\left[\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)\right]} \cong \frac{g_m r_\pi}{1 + j\omega r_\pi (C_\pi + C_\mu)} \quad \text{if } \omega C_\mu \ll g_m$$

See the Bode plot of short-circuit current gain magnitude. The corner frequency, beta cutoff frequency ($f_\beta$), is given by,

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




**Magnitude plot**




**Phase plot**

# Cutoff Frequency

Note that the magnitude of current gain decreases with increasing frequency and reaches to 1 at $f_T$ (cutoff frequency). We can write $A_i$ in the below form:

$$A_i = h_{fe} = \frac{\beta_o}{1 + j\left(\frac{f}{f_\beta}\right)} \rightarrow \left|h_{fe}\right| = \frac{\beta_o}{\sqrt{1 + \left(\frac{f}{f_\beta}\right)^2}}$$

At $f_T$, $\left|h_{fe}\right| = 1$ and normally $\beta_o \gg 1, f_T \gg f_\beta$. Therefore, equation becomes

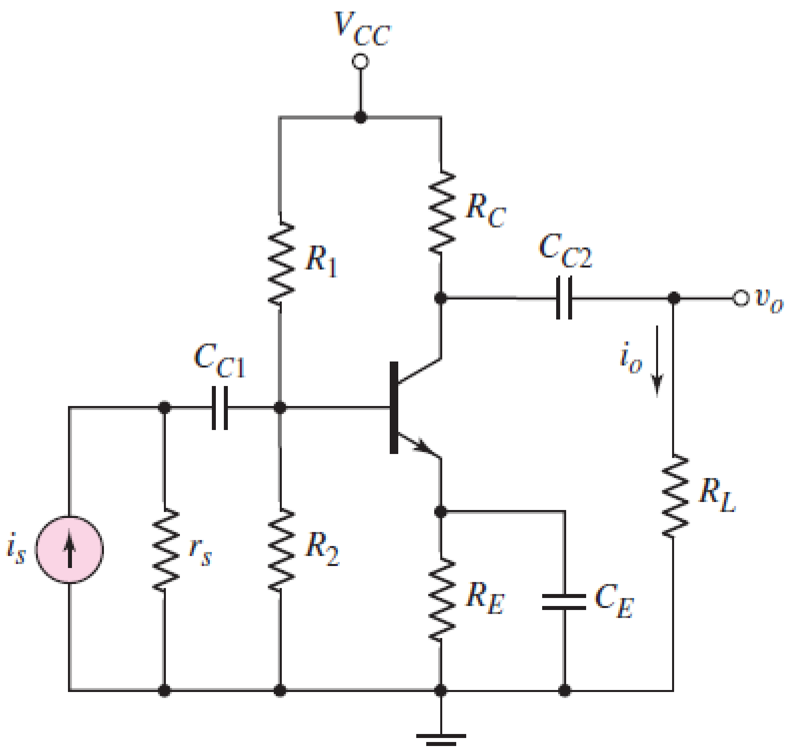
$$1 \cong \frac{\beta_o}{\sqrt{\left(\frac{f}{f_\beta}\right)^2}} = \frac{\beta_o f_\beta}{f_T} \rightarrow f_T = \beta_o f_\beta = \beta_o \frac{1}{2\pi r_\pi (C_\pi + C_\mu)} = \frac{g_m}{2\pi (C_\pi + C_\mu)}$$

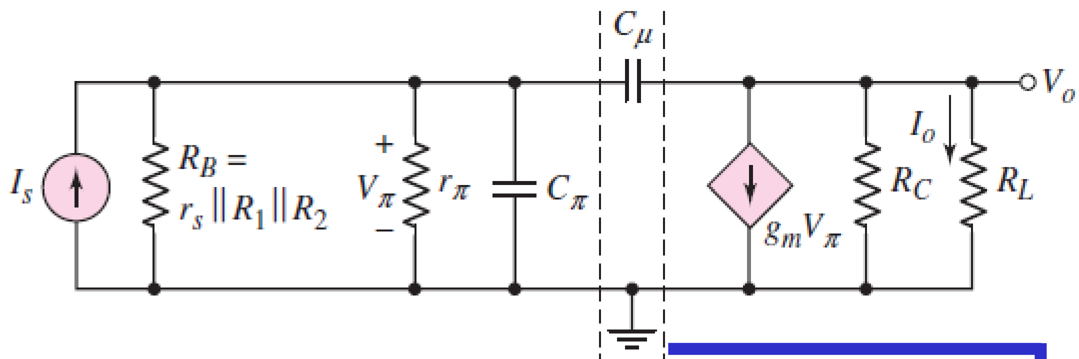
Frequency $f_\beta$ is called the bandwidth of transistor, therefore, $f_T$ is more commonly called as unity-gain bandwidth (or) gain bandwidth product.

# *Miller Effect and Miller Capacitance*

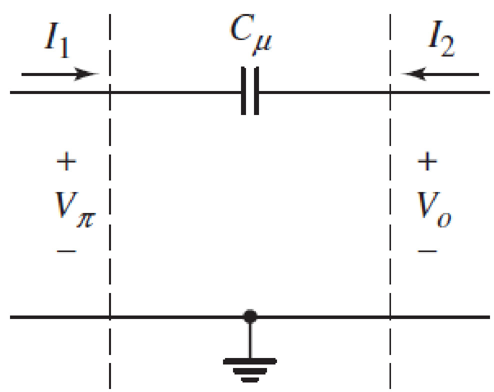
As mentioned earlier, the capacitance $C_\mu$ cannot really be ignored – *Miller effect, or feedback effect*, is a multiplication effect of $C_\mu$ in circuit applications.

✓ Assume the frequency is sufficiently high for the coupling and bypass capacitors to act as short circuits – $C_\mu$ connects output back to the input.




**Small-signal hybrid−$\pi$ equivalent circuit**

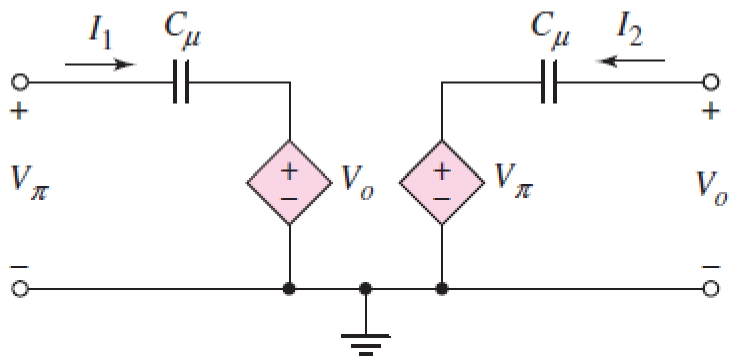



$$V_\pi = I_1\left(\frac{1}{j\omega C_\mu}\right) + V_o$$

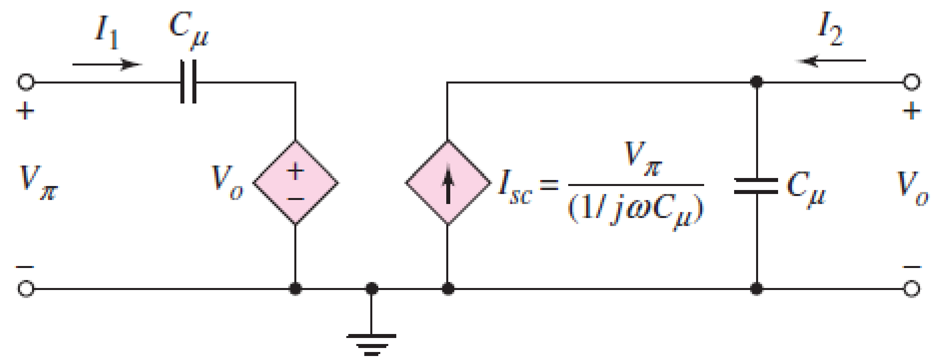
$$V_o = I_2\left(\frac{1}{j\omega C_\mu}\right) + V_\pi$$

# Miller Effect and Miller Capacitance

- ✓ Now form a two-port equivalent circuit and then convert the Thevenin equivalent circuit on the output side to Norton equivalent circuit.

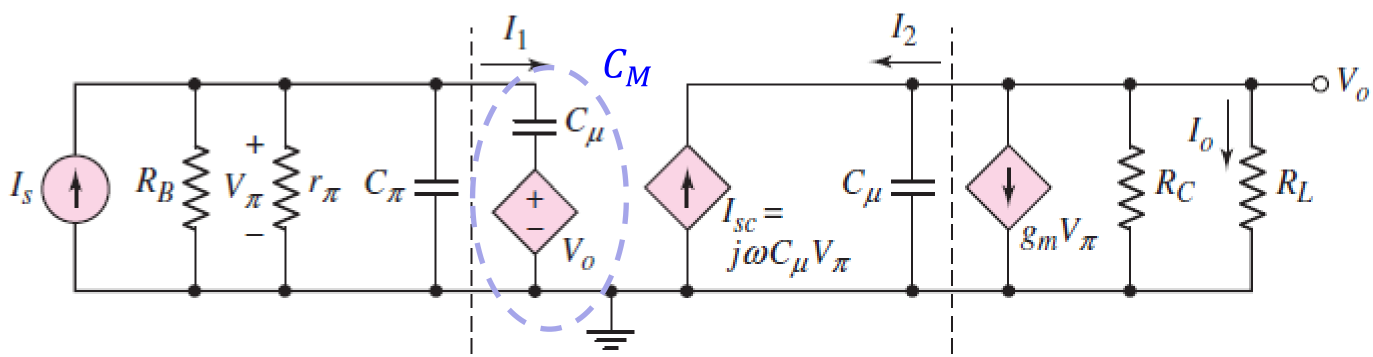



**Two-port equivalent circuit of $C_\mu$: Thevenin**




**Two-port equivalent circuit of $C_\mu$: Norton at output**

- ✓ Again, reconsider the original equivalent circuit and now replace the circuit segment between the dotted lines with the above Norton circuit.




**Small-signal circuit, including two-port equivalent model of $C_\mu$**

# Miller Effect and Miller Capacitance

Recall the expression for $V_\pi$ and obtain the equation for $I_1$ as follows:

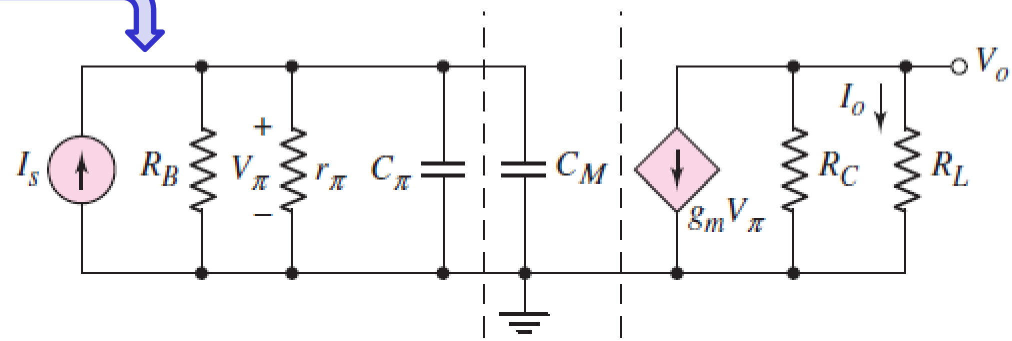
$$I_1 = \frac{V_\pi - V_o}{\frac{1}{j\omega C_\mu}} = j\omega C_\mu (V_\pi - V_o)$$

The output voltage is, $V_o = -g_m V_\pi (R_C||R_L)$

Therefore, $I_1$ becomes as, $I_1 = j\omega C_\mu[1 + g_m(R_C||R_L)]V_\pi$

The circuit segment between the dotted lines can be replaced by an equivalent capacitance called Miller capacitance as, $C_M = C_\mu[1 + g_m(R_C||R_L)]$ Note that the multiplication effect of $C_\mu$ is the Miller effect.

Consider the frequency of operation is very much smaller such that 1) $I_{sc}$ is negligible compared to $g_m V_\pi$ source, 2) $C_\mu$ will be much greater than $R_C||R_L$, therefore $C_\mu$ can be considered as an open-circuit.




The input capacitance is now $C_\pi + C_M$.

# *See you in the Final Exam (Jan 03, Friday@2 PM)*

# The End