EECS 142

Integrated Circuits for Communication

Lecture 4: Review of MOS and BJT Technology for High-Speed Applications

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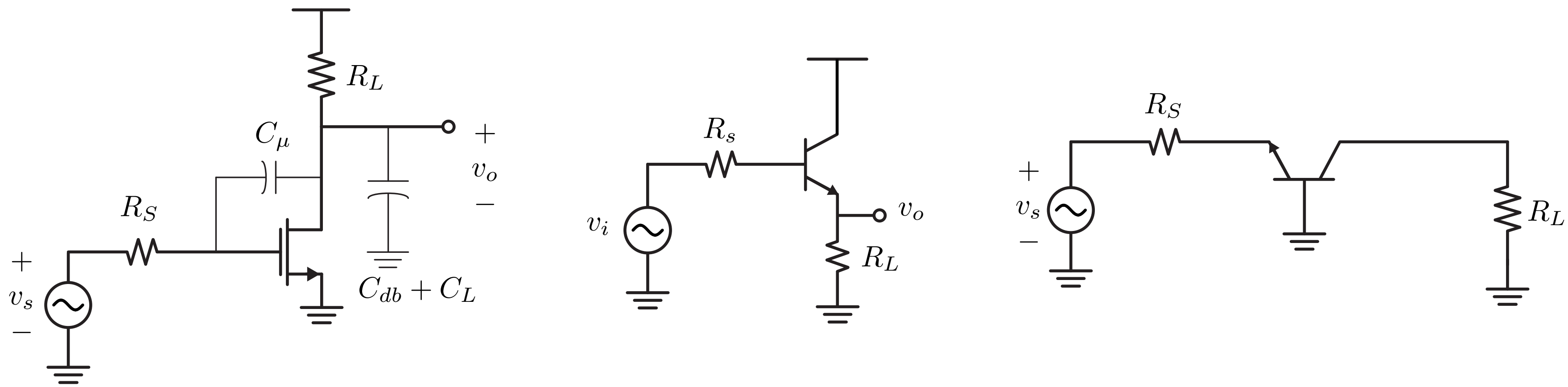
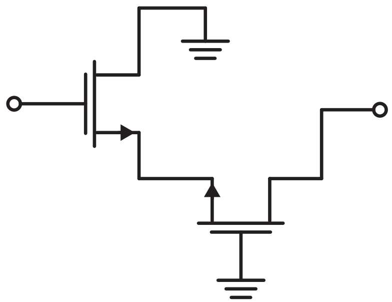
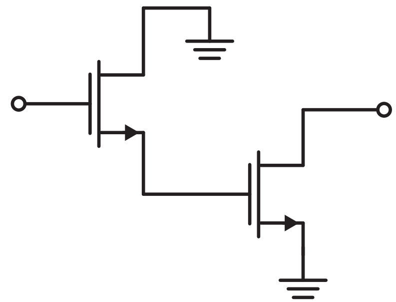
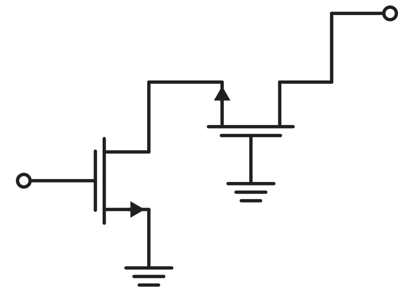
Basic Single Stage Amplifiers



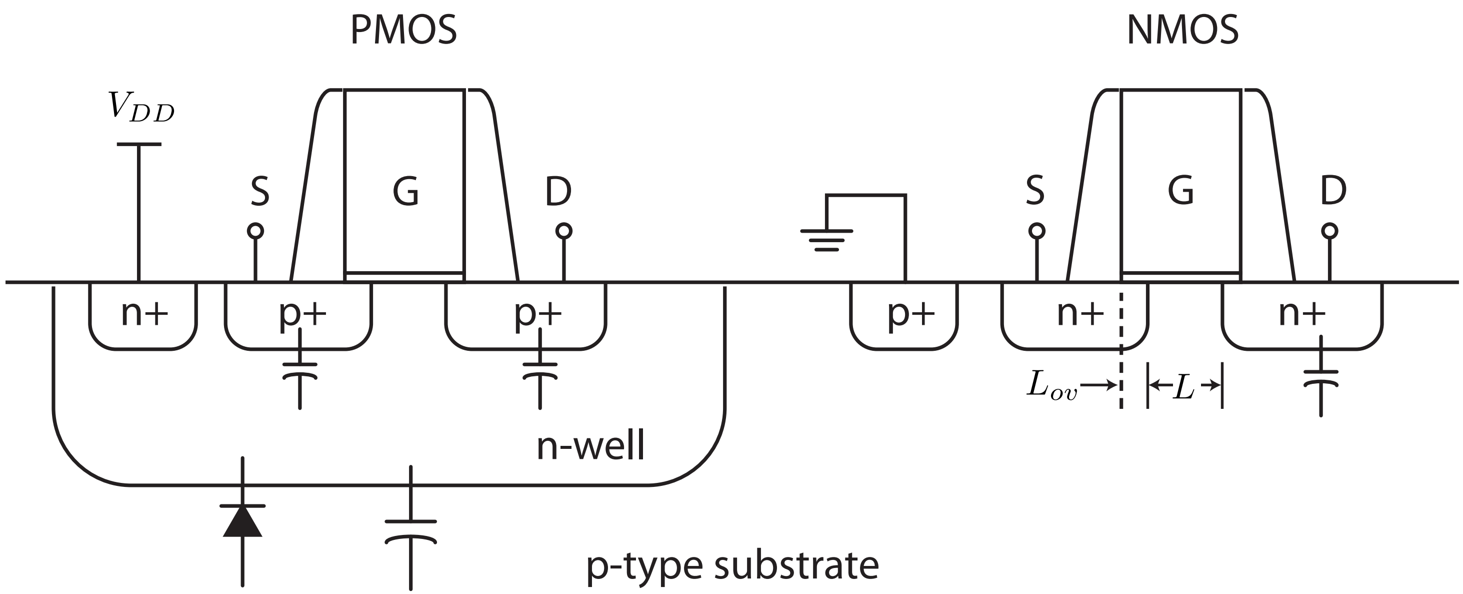
- The CE/CS amplifier has a small bandwidth due to the Miller feedback. We must keep the source resistance low.
- The CC/CD (or follower) is very wideband but only provides current gain.
- The CB/CG amplifier is also wideband (no Miller), but only offers voltage gain. Has small input impedance (sometimes good).
- The CE/CS offers the best power gain and noise figure, but bandwidth limitations are an issue.

Wideband Two-Stage Amplifiers

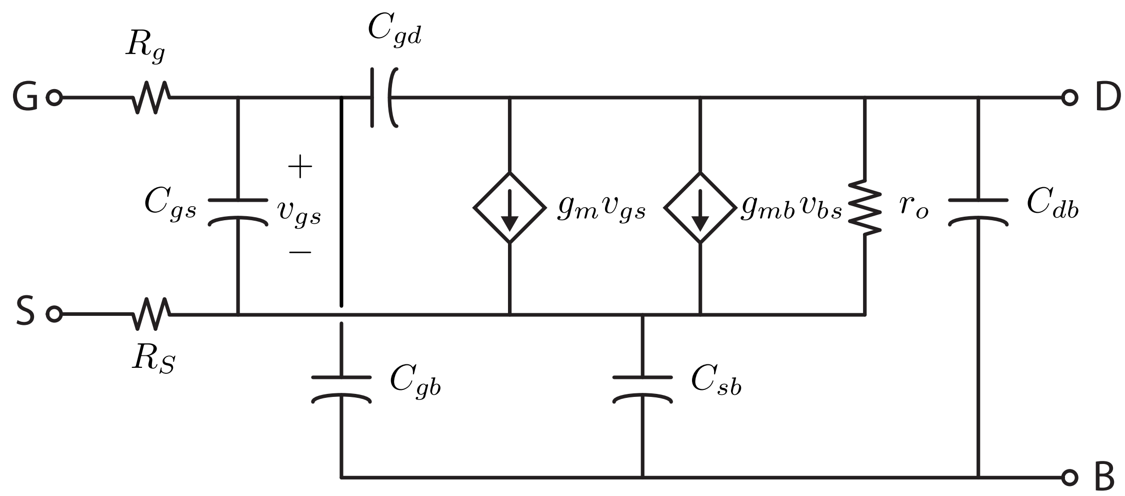
- A source follower driving a common-source amplifier buffers the high source impedance and drives the common source amplifier with a low source impedance.
- A source follower driving a common gate amplifier boosts the input impedance. This is essentially a differential pair driven single-endedly.
- A common-source amplifier drives a common-gate amplifier, or a cascode amplifier. Miller effect is minimized by lowering the gain of the common-source stage.

CMOS Cross Section



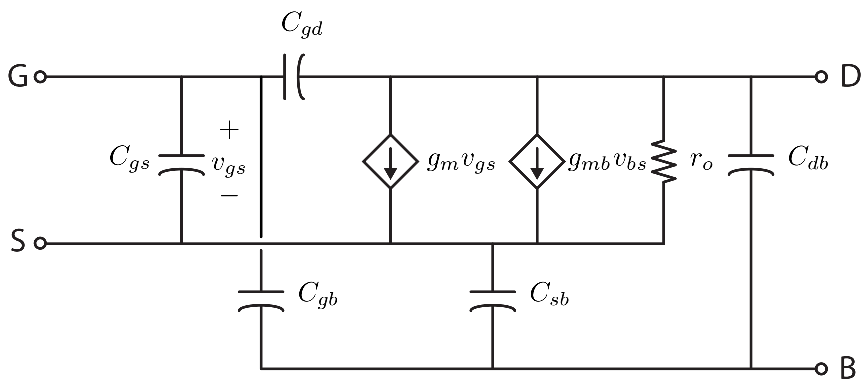
- Modern short channel CMOS process has very short channel lengths ($L < 100\,\mathrm{nm}$). To ensure gate control of channel, as opposed to drain control (DIBL), we employ thin junctions and thin oxide ($t_{ox} < 5\,\mathrm{nm}$).
- Due to lithographic limitations, there is an overlap between the gate and the source/drain junctions. This leads to overlap capacitance. In a modern FET this is a substantial fraction of the gate capacitance.

FET Small-Signal Model

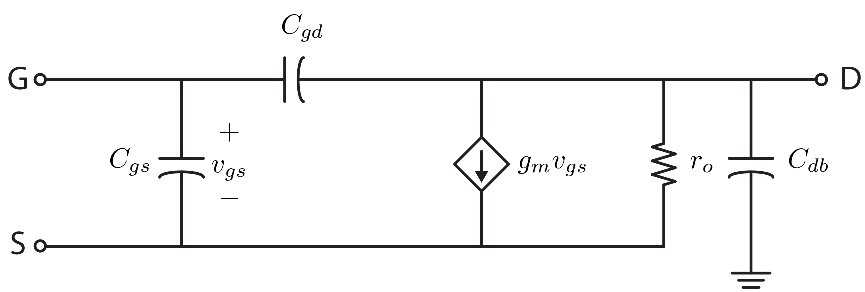


- The junctions of a FET form reverse-biased pn junctions with the substrate (well), or the body node. This is another form of parasitic capacitance in the structure, C_{db} and C_{sb}.
- At low frequency, $R_{in} \sim \infty$. There is external gate resistance R_g due to the polysilicon gate and R_s due to junction resistance.
- In the forward active (saturation) region, the input capacitance is given by $C_{gs} = \frac{2}{3} C_{ox} \cdot W \cdot L$
- r_o is due to channel length modulation and other short channel effects (such as DIBL).

Simplified FET Model

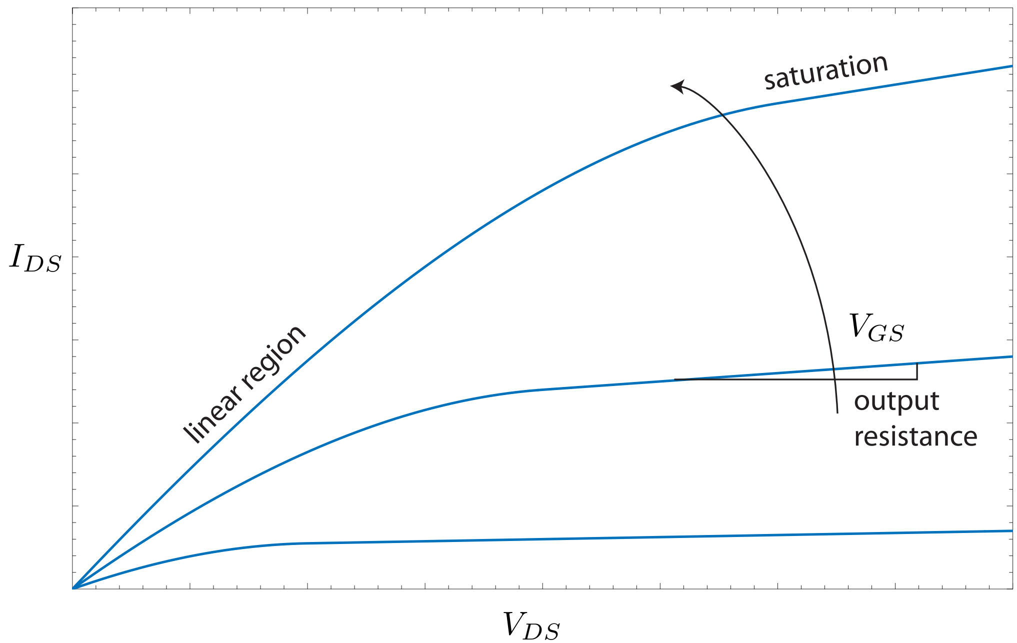


Layout parasitics increase the capacitance values!



- You’re probably more familiar with the simplified FET model without gate and source resistance. These resistances are small but are important when we analyze the power gain of the device.
- If we ground the bulk node, such as a common source amplifier, we can eliminate a lot of clutter since the g_{mb} generator is shorted.

MOS Device Characteristics



- For a long channel FET, before "pinch-off", the device drain current responds nearly linearly with V_{DS}, hence the term "linear region".

$$I_{DS} = \mu C_{ox} \frac{W}{L} \left((V_{GS} - V_T) V_{DS} - \frac{V_{DS}^2}{2} \right)$$

- Beyond pinch-off, when $V_{DS} = V_{GS} - V_T$, the current *saturates* and remains essentially constant. This is the saturation region.

$$I_{DS} = \frac{1}{2} \mu C_{ox} \frac{W}{L} \left((V_{GS} - V_T)^2 (1 + \lambda V_{DS}) \right)$$

Transconductance

- The transconductance in saturation is given by

$$g_m = \frac{dI_{DS}}{dV_{GS}} = \mu C_{ox}\frac{W}{L}(V_{GS} - V_T)(1 + \lambda V_{DS})$$

$$g_m = \frac{2I_{DS}}{V_{GS} - V_T} = \frac{2I_{DS}}{\sqrt{\frac{2I_{DS}}{\mu C_{ox}\frac{W}{L}}}}$$

$$g_m = \sqrt{2\mu C_{ox}\frac{W}{L}I_{DS}} \propto \sqrt{I_{DS}}$$

- The variation of current in saturation is due to the output impedance of the device. Short channel devices have much stronger current variation.
- We also find that the current variation with V_{GS} is weaker than quadratic and the drain current is therefore lower than predicted by the long channel equations.
- This is partly due to the threshold voltage variation and reduced mobility.

Unity Gain Frequency

- The short circuit current gain of a device is given by

$$G_i = \frac{i_o}{i_i} \approx \frac{g_m}{j\omega(C_{gs} + C_{gd})}$$

- The frequency of unity gain ω_T is given by solving $|G_i| = 1$

$$\omega_T = \frac{g_m}{C_{gs} + C_{gd}}$$

- This frequency plays an important role in the frequency response of high speed amplifiers. Often there is a gain-bandwidth tradeoff related to ω_T

$$G \times BW = \omega_T$$

MOSFET Unity Gain Frequency

- For a long-channel MOSFET we have the following relationship

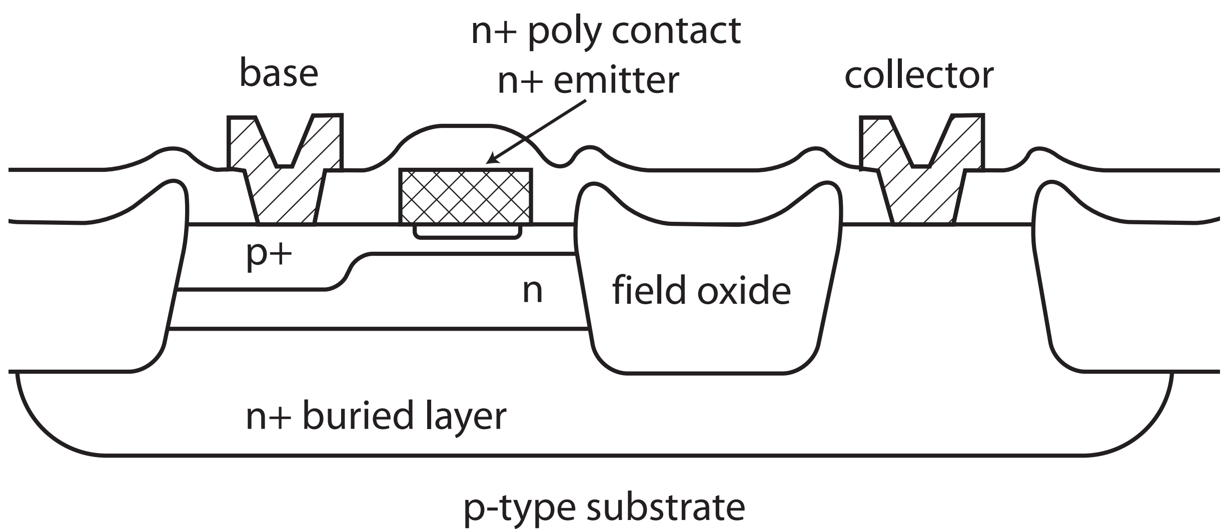
$$\omega_T = \frac{g_m}{C_{gs} + C_{gd}} \approx \frac{\mu C_{ox}\frac{W}{L}(V_{GS} - V_T)}{\frac{2}{3}W \cdot LC_{ox}}$$

- Canceling common factors we have

$$\omega_T = \frac{3}{2}\frac{\mu}{L^2}(V_{GS} - V_T)$$

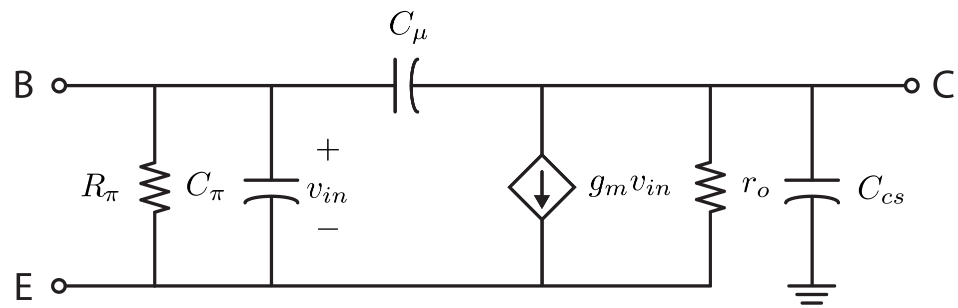
- We see that ω_T is bias dependent. The strong L^2 length dependence only holds for long-channel devices. Short channel devices, in the limit of velocity saturated operation, reduce to $1/L$ dependence.
- Note that $\mu = f(V_{GS} - V_T)$ due to reduced mobility at the surface of the transistor.

Bipolar Device Cross Section



- Most transistor "action" occurs in the small npn sandwich under the emitter. The base width should be made as small as possible in order to minimize recombination. The emitter doping should be much larger than the base doping to maximize electron injection into the base.
- A SiGe HBT transistor behaves very similarly to a normal BJT, but has lower base resistance r_b since the doping in the base can be increased without compromising performance of the structure.

Bipolar Small-Signal Model



- The core model is similar to a FET small-signal model. The resistor r_π, though, dominates the input impedance at low frequency. At high frequency, C_π dominates.
- C_μ is due to the collector-base reverse biased diode capacitance. C_{cs} is the collector to substrate parasitic capacitance. In some processes, this is reduced with an oxide layer.
- C_π has two components, due to the junction capacitance (forward-biased) and a diffusion capacitance

$$C_\pi = C_{bej} + C_{diff}$$

Bipolar Exponential

- Due to Boltzmann statistics, the collector current is described very accurately with an exponential relationship

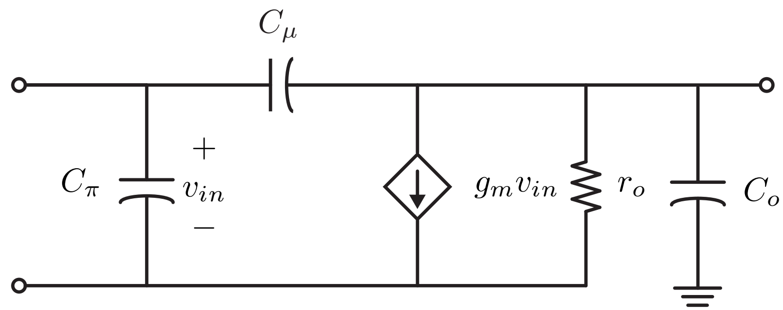
$$I_C \approx I_S e^{\frac{qV_{be}}{kT}}$$

- The device transconductance is therefore proportional to current

$$g_m = \frac{dI_C}{dV_{be}} = I_S \frac{q}{kT} e^{\frac{qV_{be}}{kT}} = \frac{qI_C}{kT}$$

- where $kT/q = 26\,\mathrm{mV}$ at room temperature. Compare this to the equation for the FET. Since we usually have $kT/q < (V_{gs} - V_T)$, the bipolar has a much larger transconductance for the same current. This is the biggest advantage of a bipolar over a FET.

Generic Equivalent Circuit



- The generic figure above represents both a FET and a bipolar at high frequency.
- Notice that this model holds when $r_\pi \gg \frac{1}{\omega C_\pi} = X_\pi$.
- Since

$$\frac{r_\pi}{X_\pi} = \omega r_\pi C_\pi = \omega \frac{\beta_0}{g_m} C_\pi \approx \beta \frac{\omega}{\omega_T}$$

- Say $\beta_0 = 100$ and the operating frequency is $\omega/\omega_T = 1/10$. Then we have $r_\pi/X_\pi = 100 \times 10 = 10$.

Bipolar Unity Gain Frequency

- Similar to a FET we have the following relationship

$$\omega_T = \frac{g_m}{c_\pi + C_\mu}$$

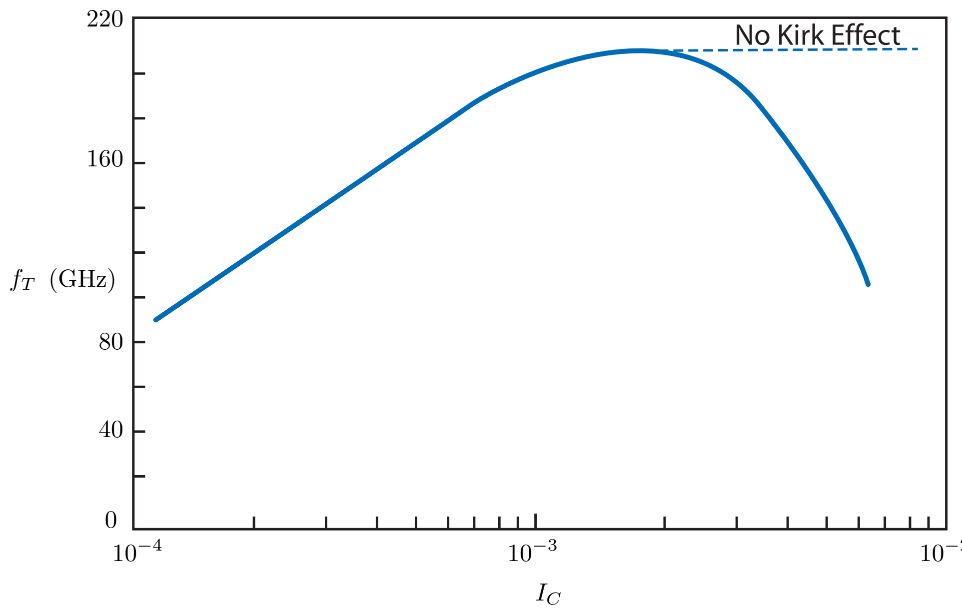
- Expanding the denominator term

$$\omega_T = \frac{g_m}{c_{bej} + C_d + C_\mu} \approx \frac{g_m}{2C_{je0} + g_m \tau_F + C_{jc}}$$

- The collector junction capacitance is a function of V_{bc}, or the reverse bias. To maximize ω_T, we should maximize the collector voltage. Re-writing the above equation

$$\omega_T = \frac{1}{\tau_F + \frac{2C_{je0}+C_{jc}}{g_m}}$$

Bipolar Bias Dependence

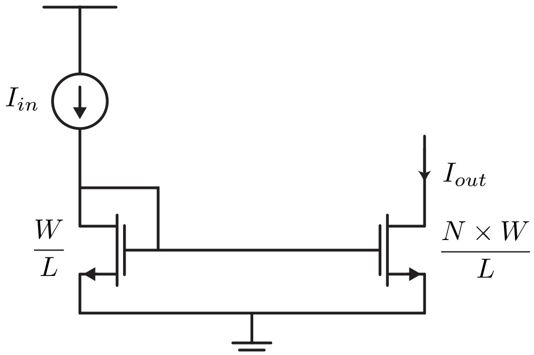


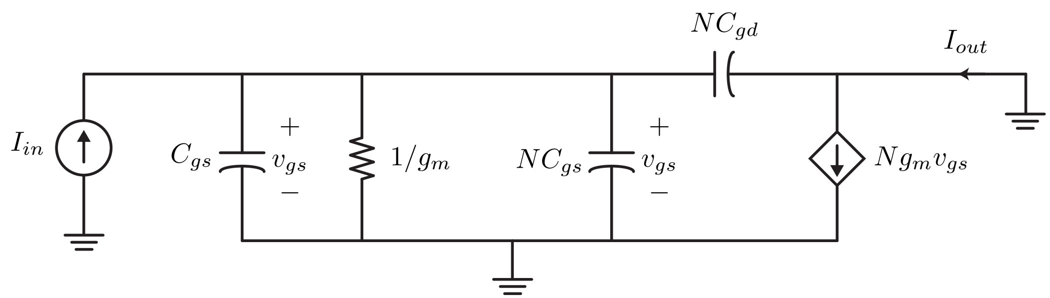
- We can clearly see that if we continue to increase I_C, then $g_m \propto I_C$ increases and the limiting value of ω_T is given by the forward transit time τ_F

$$\omega_T \longrightarrow \frac{1}{\tau_F}$$

- In practice, though, we find that there is an optimum collector current. Beyond this current the ω_T drops. This optimum point occurs due to the Kirk Effect. It's related to the "base widening" due to high level injection.

Example: A Current Amplifier





- A $1 \times N$ current mirror has broadband frequency response which can be illustrated with the equivalent circuit. The diode-connected device can be replaced with a conductance of value g_{m1} in shunt with the amplifier input capacitance C_{in}.
- If the current amplifier drives a low impedance load, the transfer function is given by

$$G_i = \frac{i_o}{i_s} = \frac{g_{m2}}{Y_{in}(s)} = \frac{g_{m2}}{g_{m1} + sC_{in} + N \times sC_{in}}$$

$$G_i = \frac{\frac{g_{m2}}{g_{m1}}}{1 + (N+1)\frac{s}{\omega_T}}$$

Current Amplifier Analysis

- Note that the transconductance of output device is N times larger since it can be thought of N devices in parallel. The complete transfer function is therefore

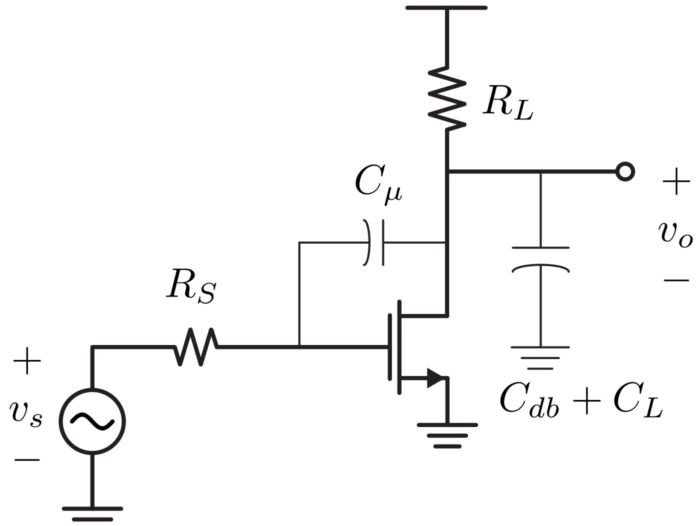
$$G_i = \frac{N}{1 + \frac{s}{\omega_T/(N+1)}}$$

and the gain-bandwidth product is given by

$$G_i \times \omega_{-3\text{db}} = \frac{N}{N+1}\omega_T \approx \omega_T$$

- It's important to note that the above analysis holds only if we assume the load impedance is extremely low, ideally a short. If we connect a physical resistor to the output, the Miller effect will produce a significant feedback current which invalidates our assumptions.
- Note amplifier is large signal linear.

CS/CE Amplifier Bandwidth



- Due to Miller multiplication, the input cap is usually the dominant pole

$$\omega_0^{-1} \approx R_s(C_{in} + |A_v| C_\mu)$$

$$\omega_0^{-1} = R_s C_{in}(1 + \mu |A_v|) \approx R_s C_{in} \mu |A_v|$$

CS/CE Amplifier Gain-Bandwidth

- Assuming the voltage gain is given by the low-frequency value of $g_m R_L$, we have

$$\omega_0^{-1} = R_s C_{in} \mu g_m R_L = (g_m R_s)(g_m R_L) \frac{C_{in}}{g_m} \mu$$

$$\omega_0^{-1} = |A_v|^2 \frac{R_s}{R_L} \omega_T^{-1} \mu$$

- The amplifier has a bandwidth reduction factor of A_v^2

$$\omega_0 \times |A_v|^2 = \omega_T \times \left(\frac{R_L}{R_s} \right) \times \frac{1}{\mu}$$

Bandwidth Example

- Say we need a gain of 60 dB ($A_v = 1000$) and $\frac{R_L}{R_s} = 2$. The technology has a capacitance ratio of $\mu = 0.2$:

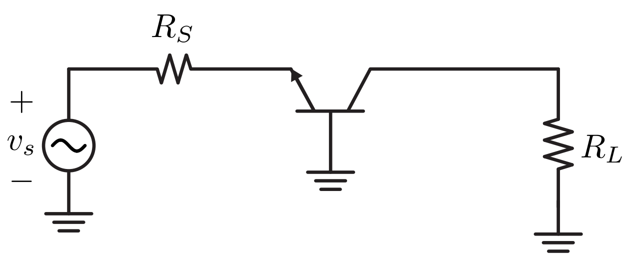
$$\omega_0 |A_v|^2 = 10^6 \omega_0 = \omega_T \times 2 \times 5$$

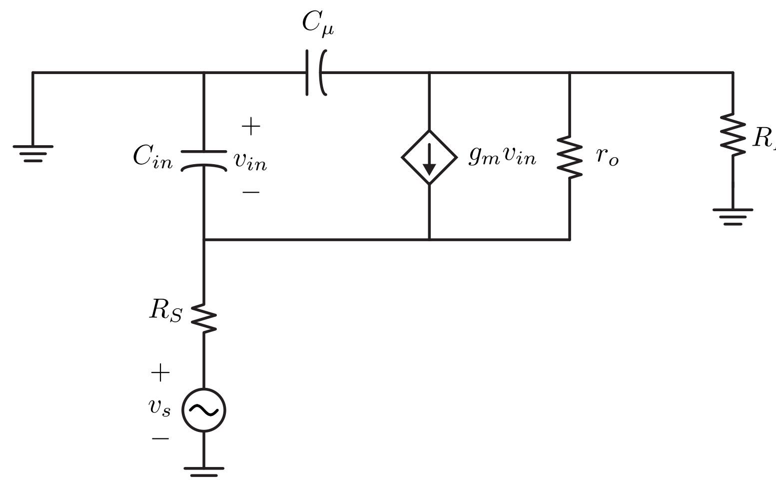
$$\omega_0 = \frac{\omega_T}{10^5}$$

- Compare this to a current mirror amplifier. When we follow the "normal" gain-bandwidth tradeoff, we have

$$\omega_0 = \frac{\omega_T}{A_i} = \frac{\omega_T}{1000}$$

Common Base Amplifier





- Write KCL at base node of circuit

$$\frac{v_s + v_{in}}{R_s} + g_m v_{in} + sC_{in} v_{in} = 0$$

$$v_s = -v_{in}(1 + g_m R_s + sC_{in} R_s)$$

Common Base Amp (cont)

- And write KCL at the output node

$$(sC_o + \frac{1}{R_L})v_o + g_m v_{in} + sC_\mu v_o = 0$$

$$v_o\left(\frac{1}{R_L} + s(C_o + C_\mu)\right) = -g_m v_{in}$$

- The voltage gain is a product of two terms

$$A_v = \frac{v_o}{v_s} = \frac{-g_m R_L}{1 + s(C_o + C_\mu)R_L}\frac{v_{in}}{v_s}$$

$$A_v = \frac{G_m R_L}{(1 + s(C_o + C_\mu)R_L)(1 + sR_s\frac{C_{in}}{1+g_m R_s})}$$

Common Base Bandwidth

- Note the transconductance is degenerated, $G_m = g_m/(1 + g_m R_s)$. Note that the input capacitance is also degenerated by the action of series feedback.
- Unlike a CE/CS ampilifier, the poles do not interact (due to absence of feedback capacitor)
- First let's take the limit of high loop gain, $g_m R_s \gg 1$

$$A_v = \frac{\frac{R_L}{R_s}}{(1 + s/\omega_T)(1 + s/\omega_L)}$$

where $\omega_L = ((C_o + C_\mu)R_L)^{-1}$ is the pole at the output.

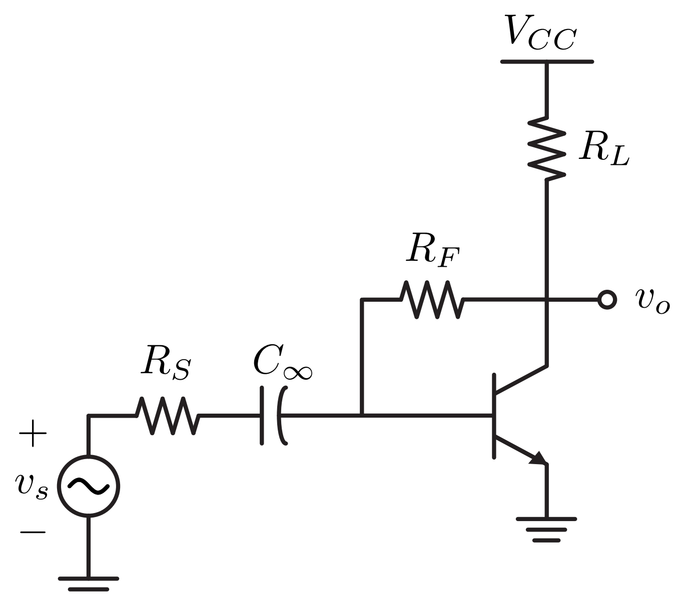
Matched Common Base Amp

- The common-base amplifier has the nice property that the input impedance is low (roughly $1/g_m$) and broadband, thus easily providing a termination to the driver (a filter, the antenna, or a previous stage). If we assume that $R_s = 1/g_m$, we have

$$A_v = \frac{\frac{1}{2} g_m R_L}{(1 + s/2\omega_T)(1 + s/\omega_L)}$$

- The 3dB bandwidth is thus most likely set by the time constant at the load.

Shunt Feedback Amp



- The shunt-feedback amplifier is a nice high-frequency broadband amplifier building block. The action of the shunt feedback is used to lower the input impedance and to set the gain.

Shunt Feedback Gain / Input Resis

- The in-band voltage gain and input impedance is given by (see homework)

$$A_v = \frac{-R_F}{R_s}$$

$$R_{in} = (1 + \frac{R_F}{R_L})\frac{1}{g_m}$$

- For an input match, $R_s = (1 + \frac{R_F}{R_L})\frac{1}{g_m}$, or $g_m R_s = (1 + \frac{R_F}{R_L})$
- Since the voltage gain sets R_F, the input impedance match determines the required transconductance g_m (and hence the power dissipation)
- A bipolar version will dissipate much less power due to the higher intrinsic g_m

Shunt Feedback Amp BW

- The amplifier is broadband and approximately obeys the classic gain-bandwidth tradeoff $A_v \omega_0 \approx \omega_T$
- A zero-value time constant analysis identifies the dominant pole

$$\tau_1 = C_{in} \left(R_s || r_\pi || \frac{R_F(1 + \frac{R_L}{R_F})}{1 + g_m R_L} \right)$$

$$\tau_2 = C_\mu \left(R_F || \frac{R_L(R_s || r_\pi)}{R_s || r_\pi || \frac{1}{g_m} || R_L} \right)$$

$$\omega_{-3\text{dB}} \approx \frac{1}{\tau_1 + \tau_2}$$

Shunt Feedback/CC Cascade

- If the shunt-FB amplifier needs to drive a low impedance load, a broadband voltage buffer is needed
- As shown below, an emitter follower (or source follower) provides the solution (note this is a fast pnp). Note the buffer is broadband (gain ≈ 1) and only loads the core amplifier by the degenerated input capacitance $C_{in2}/(1 + g_{m2}R_E)$