6.772/SMA5111 - Compound Semiconductors

Lecture 14 - Heterojunction Bipolar Transistors

- **High frequency perfomance of HBTs** (cont. from Lec. 13)
  - Small signal linear equivalent circuit
  - High frequency limits - ω_T, ω_{max}, ω_v
- **Digital applications of HBTs**
  - Basic BJT logic families
- **Linear circuit applications of HBTs**
  - Examples from Vitesse VIP-1 InP HBT process

HFET i-v: for HEMT and HIGFET, cont.

As before, we write the drain current at y as:

$$i_D = -\left[-qN_{ch}(y)\cdot W\cdot \overline{s_y}(y)\right]$$

We just found $N_{ch}(y)$, and we can write the average net carrier velocity in the low- to moderate field region as:

$$\overline{s_y}(y) = -\mu_e F_y(y) = \mu_e \frac{dv_{CS}(y)}{dy}$$

Putting this together yields:

$$i_D = W\frac{\varepsilon_{WBG}}{t_{WBG}}\left[v_{GS} - v_{CS}(y) - V_T\right]\mu_e \frac{dv_{CS}(y)}{dy}$$

Integrating with respect to y from 0 to L, or equivalently, with respect to v_{CS} from 0 to v_{DS}, we have the result given on the next foil….

HFET i-v: for HEMT and HIGFET, cont.

The result of the integration is:

$$i_D = \frac{W}{L} \mu_e \frac{\varepsilon_{WBG}}{t_{WBG}} \left[(v_{GS} - V_T) v_{DS} - \frac{v_{DS}^2}{2} \right]$$

This expression is valid as long as $v_{DS} < (v_{GS} - V_T)$. When $v_{DS} > (v_{GS} - V_T)$, i_D saturates at:

$$i_D = \frac{1}{2} \frac{W}{L} \mu_e \frac{\varepsilon_{WBG}}{t_{WBG}} (v_{GS} - V_T)^2$$

These expressions should look familiar from MOSFETs. They are identical to the commonly used MOSFET expression.

Finally, the small signal transconductance, g_m, in saturation is:

$$g_m \equiv \left. \frac{\partial i_D}{\partial v_{GS}} \right|_Q = \frac{W}{L} \mu_e \frac{\varepsilon_{WBG}}{t_{WBG}} (V_{GS} - V_T)$$

HFET i-v: for HEMT and HIGFET, cont.

One last thing to look at for these devices is their high frequency performance. A good way to do this is to look at ω_T:, which we recall can be written:

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

We have not found C_{gs} yet, but we in fact already know it because C_{gs} for these devices will be the same as it is for a MOSFET:

$$C_{gs} \approx \frac{2}{3} W\, L\, \frac{\varepsilon_{WBG}}{t_{WBG}}$$

Using this, we find:

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

without velocity saturation

HFET i-v: for HEMT and HIGFET w. velocity saturation.

When we have severe velocity saturation, we have $s(y) = s_{sat}$, and we write simply

$$i_D = W \frac{\varepsilon_{WBG}}{t_{WBG}} \left[v_{GS} - V_T \right] s_{sat}$$

Calculating the small signal transconductance now, we have:

$$g_m \equiv \left. \frac{\partial i_D}{\partial v_{GS}} \right|_Q = W \; s_{sat} \frac{\varepsilon_{WBG}}{t_{WBG}}$$

with velocity saturation

These expressions should look familiar from MOSFETs. They are identical to the commonly used MOSFET expression.

Finally, the small signal gate-to-source capacitance, C_{gs}, in saturation is simply the gate capacitance, $WL\varepsilon_{WBG}/t_{WBG}$ and thus:

$$\omega_T \approx \frac{g_m}{C_{gs}} = \frac{s_{sat}}{L}$$

with velocity saturation

High frequency performance metrics: ω_T, ω_{max}

We introduced ω_T, which we defined as the frequency at which the magnitude of the short circuit current gain was unity, and showed that:

$$\omega_T \approx g_m / C_{gs}$$

It turns out that ω_T is a good metric for high speed switching, but for microwave applications, a better metric is ω_{max}, the unity power gain with matched source and load impedances. We call this frequency , and find that it is roughly given by:

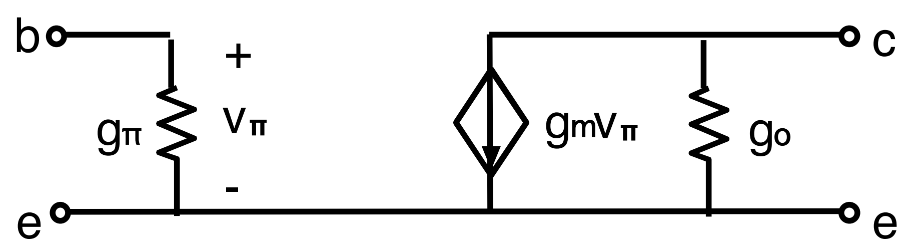
$$\omega_{\max} \approx \sqrt{\frac{\omega_T}{4 R_g C_{gd}}} \approx \sqrt{\frac{g_m}{4 R_g C_{gd} C_{gs}}}$$

This later metric shows the importance in microwave amplifier applications of FETs of reducing the gate resistance, R_g, and the gate-to-drain capacitance, C_{gd}.

Note: We'll introduce a third high frequency performance metric, ω_v, when we talk about HBTs.

Linear equivalent circuit (lec) models - HBT

The linear equivalent circuit of a BJT biased in its forward active region is the same as that of an FET, except that the input resistance is now finite, and the notation is different:



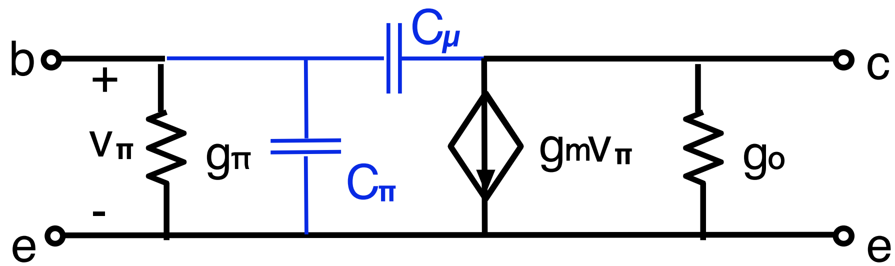
Using the large signal model for the BJT to evaluate the parameters in this model we find the following expressions:

$$g_m \equiv \frac{\partial i_C}{\partial v_\pi}\Big|_Q \approx \frac{qI_C}{kT} \qquad g_\pi \equiv \frac{\partial i_B}{\partial v_\pi}\Big|_Q \approx \frac{g_m}{\beta_F}$$

$$g_o \equiv \frac{\partial i_C}{\partial v_{CE}}\Big|_Q \approx \frac{I_C}{V_A} = \lambda I_C$$

HBT l.e.c. models - high frequency model

To extend this model to high frequencies we introduce small signal linear capacitors representing the charge stored on the gate:



We find:

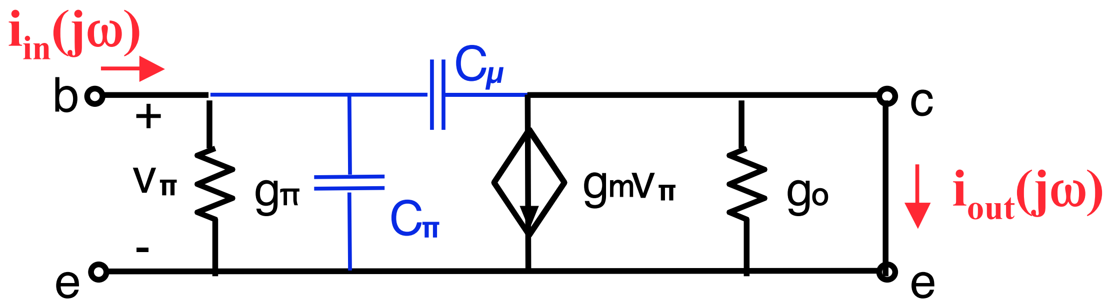
$$C_\pi \equiv \frac{\partial q_E}{\partial v_{BE}}\bigg|_Q = C_{eb,depl} + C_{eb,diff} \quad \textit{where} \quad C_{eb,diff} \approx g_m \tau_{tr}$$

$$C_\mu \equiv \frac{\partial q_C}{\partial v_{BC}}\bigg|_Q = C_{cb,depl}$$

More on this later.

HBT intrinsic high freq. performance metric - ω_T

As we saw when we looked at FETs, a measure of the high frequency performance of a transistor is obtained by calculating its short circuit current gain, $\beta_{sc}(j\omega)$, and finding the frequency at which its magnitude is 1. For a BJT we have:



$$\beta_{sc}(j\omega) = \frac{i_{out}(j\omega)}{i_{in}(j\omega)} = \frac{j\omega C_\mu v_\pi - g_m v_\pi}{g_\pi + j\omega(C_\pi + C_\mu)v_\pi} = \frac{\beta_F\left[j\omega(C_\mu/g_m) - 1\right]}{1 + j\omega(C_\pi + C_\mu)/g_\pi}$$

Thus:

$$\beta_{sc}(j\omega) \approx \beta_F \qquad for\ \omega << g_\pi/(C_\pi + C_\mu)$$

$$For \quad g_\pi/(C_\pi + C_\mu) << \omega << g_m/C_\mu, \qquad |\beta_{sc}(j\omega)| \approx \frac{g_m}{\omega(C_\pi + C_\mu)}$$

High frequency performance metrics: ω_T, ω_{max}, ω_v

We have ω_T for a BJT, and it is interesting to write it in terms of the bias point and to then take the limit as I_C becomes very large:

$$\omega_T = \frac{g_m}{\left(C_\pi + C_\mu\right)} = \frac{qI_C/kT}{C_{eb,depl} + \left(qI_C/kT\right)\tau_{tr} + C_{cb,depl}}$$

$$thus \quad \lim_{I_C \to \infty} \omega_T = 1/\tau_{tr}$$

We found a similar result for FETs, that is that ω_T is approximately the inverse of the transit time through the active part of the device. In our simple model for the BJT the transit time that appears here is the transit time through the base, but in fact we find that it should in fact be the transit time from the emitter to the collector, which includes the time to transit the emitter-base and base-collector depletion regions:

$$\tau_{tr,B} \approx \frac{w_b^{*2}}{2D_{eB}} + \tau_{tr,EB} + \tau_{tr,CB}$$

High frequency performance metrics: ω_T, ω_{max}, ω_v

As we said with FETS, ω_T is a good metric for high speed switching, but for microwave applications, a better metric is ω_{max}, the unity power gain with matched source and load impedances. For a BJT, we find that ω_{max} is roughly given by:

$$\omega_{\max} \approx \sqrt{\frac{\omega_T}{4R_b C_\mu}} \approx \sqrt{\frac{g_m}{4R_b C_\mu C_\pi}}$$

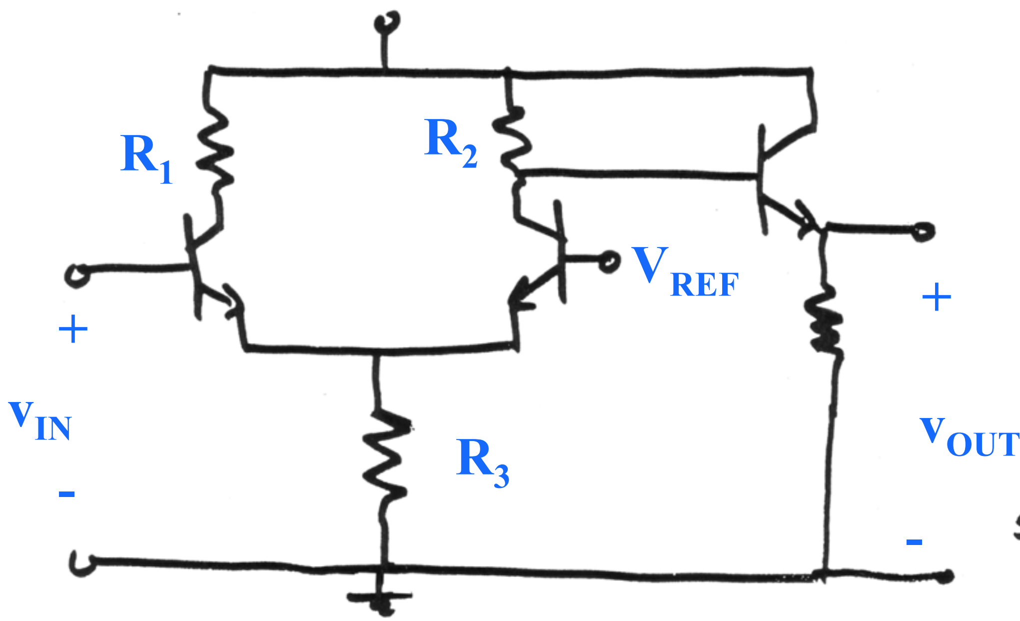
This metric shows the importance in microwave amplifier applications of HBTs of reducing the base series resistance, R_b, and the base-to-collector capacitance, C_μ.

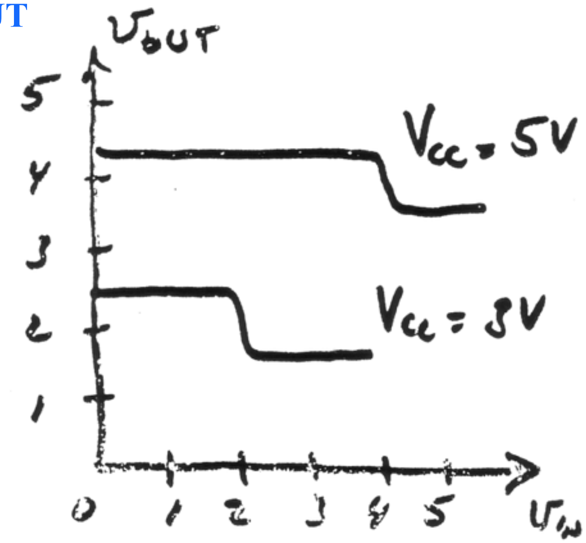
Recently an additional high frequency performance metric has been suggested, the 3dB breakpoint frequency of the effective transconductance, which is called ω_v:

$$\omega_v = \frac{\omega_T}{g_m R_b} = \frac{1}{R_b C_\pi} \qquad \left(Note: \ in\ an\ FET,\ \omega_v = \frac{1}{R_g C_{gs}} \right)$$

Bipolar Logic - ECL

Emitter coupled logic, ECL, is based on the differential amplifier and is the fastest bipolar logic:

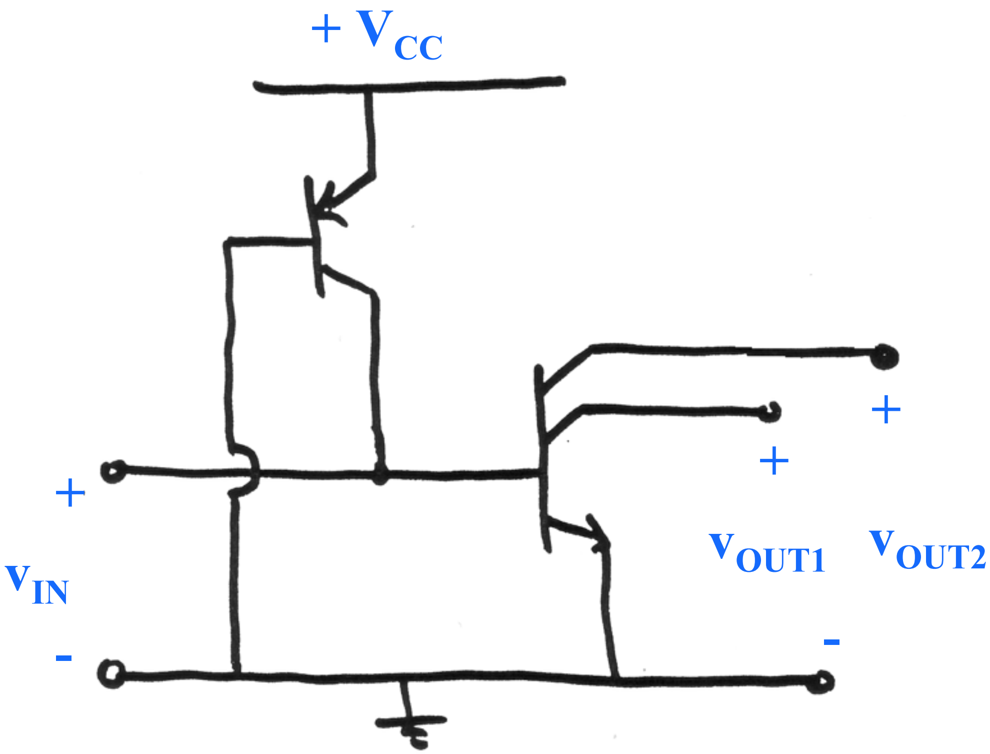




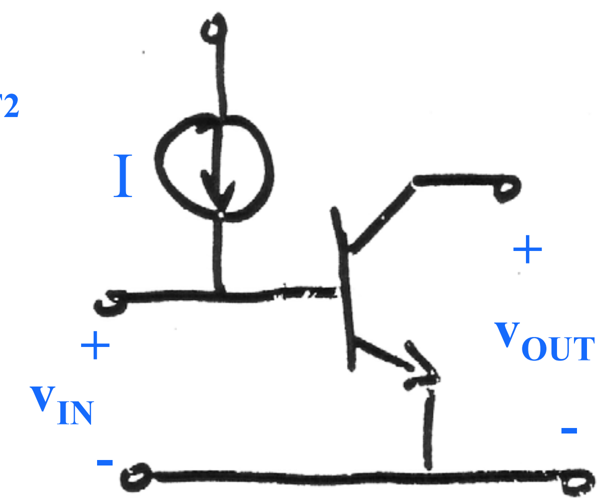
Small voltage swings
Low noise
Very fast
Complex circuit

Bipolar Logic - I^2L

Current injection logic, I^2L, is a very simple logic that was very popular a number of years ago (less so now):

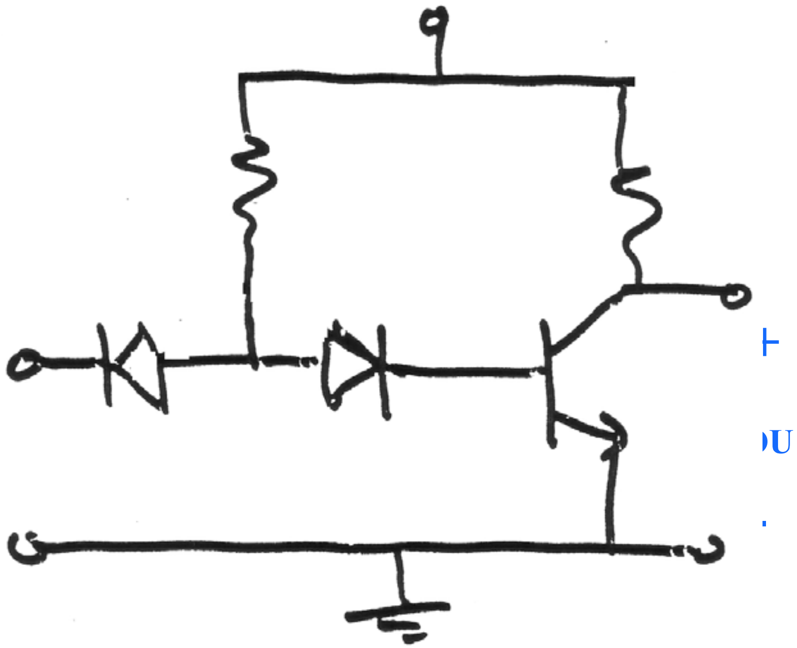


Best looked at as….

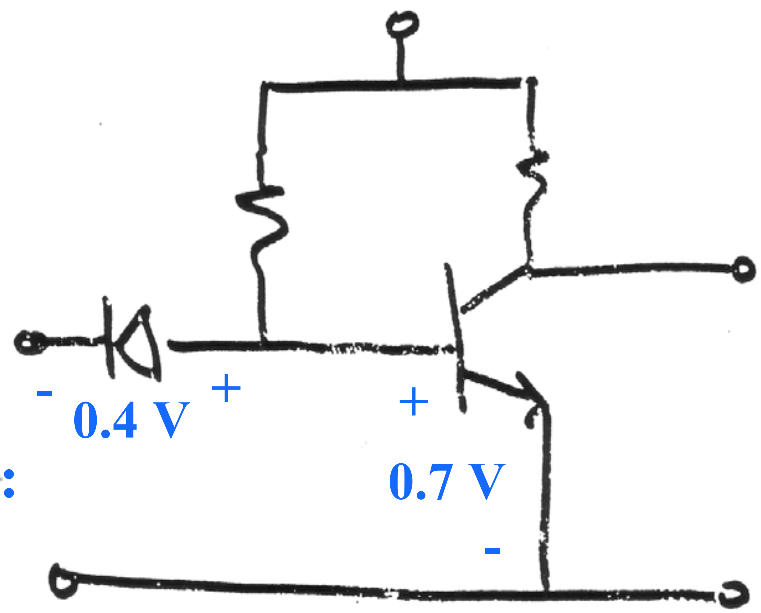


Bipolar Logic - DTL and TTL

I^2L is very similar to RTL, DTL, and TTL, and these are easier to implement (no pnp), especially with HBTs:



If we have diodes with different turn-on voltages available, we do not need a buffer diode (nor TTL):



Advantages of InGaAs/InP over other materials:

- High electron mobility in InGaAs (1.6 times that of GaAs and 9 times that of Si). The extent of transient electron overshoot is also greater than in GaAs. Hence higher f_t values can be obtained.
- The bandgap of GaInAs, being narrower than that of silicon and GaAs, produces indium phosphide HBTs that have a very low turn-on voltage (V_{BE}) and are therefore lower power dissipation.
- For a given level of doping, indium phosphide has a high breakdown field.
- Recombination velocity at InGaAs surfaces (10^3 cms^{-1}) is much smaller than that of GaAs surface (10^6 cms^{-1}). The base current due to surface recombination at the emitter periphery is reduced.
- Higher substrate thermal conductivity than in the case of GaAs (0.7 vs. 0.46 Wcm/K).
- The device is directly compatible with sources (lasers and LEDs) and photodetectors for 1.3 μm radiation, which corresponds to the wavelength of the lowest dispersion in silica-based optical fibers. It is therefore a good candidate for integration in OEICs.