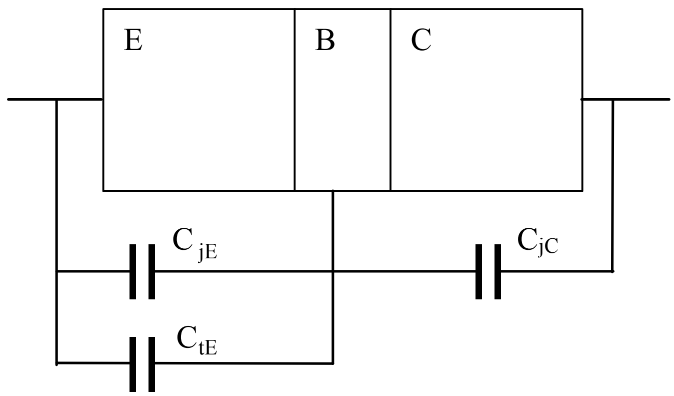
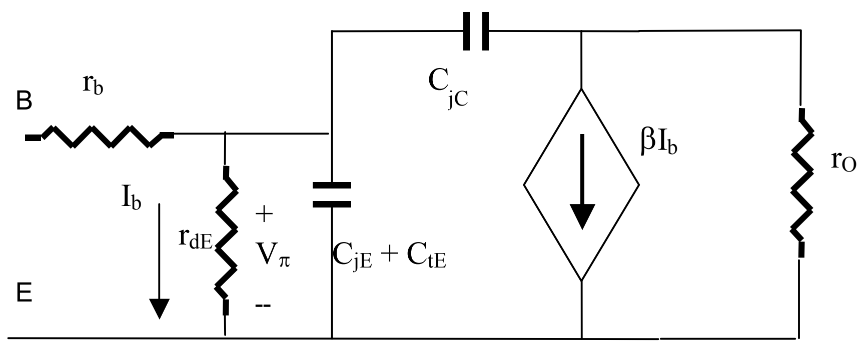
ECE 304: Bipolar Capacitances

The Bipolar Transistor: S&S pp. 485-497

Let's apply the diode capacitance results to the bipolar transistor. There are two junctions in the bipolar transistor. The BC (base-collector) junction is reverse biased in the active mode, and so it has only a junction capacitance contribution to the equivalent circuit, C_{jC}. The EB (emitter-base) junction, on the other hand, is forward biased in the active mode. Therefore, it will exhibit *both* a junction capacitance C_{jE} and a transit time capacitance C_{tE}. If we look at the simplified version of the bipolar, we can see how to hook these capacitors into the circuit:



Putting this into the regular hybrid π model we show how to obtain S&S Fig. 5.67, p. 487. The parasitic base resistance r_b has been added. This resistor is small (about 10Ω or less). However,



in circuits where the source driving the transistor has very low resistance, r_b is significant for frequencies high enough that C_{jE}+C_{dE} short-circuits r_{dE}. Then r_b is the only impedance in the base lead of the circuit. Also, the presence of r_b activates the Miller effect (discussed later) increasing the role of C_{jC} in the frequency response.

There are some changes in notation needed to obtain the standard circuit. First, we re-label a few items: $C_{jC} \rightarrow C_\mu$, $(C_{jE}+ C_{tE}) \rightarrow C_\pi$, $r_{dE} \rightarrow r_\pi$. Then we notice that the dependent current source depends on **only the part of the base current** that flows through $r_{dE}= r_\pi$ (labeled here by phasor amplitude I_b). That is, to use this circuit we have to worry about the current divider made up of $(C_{jE}+ C_{tE}) \rightarrow C_\pi$ and $r_{dE} \rightarrow r_\pi$. This worry is a nuisance, so we prefer to use the voltage V_π instead. Because the current in $r_{dE} = r_\pi$ is simply $I_b = V_\pi/ r_\pi$, the dependent source can be re-expressed as $\beta I_b = \beta V_\pi/r_\pi = g_m V_\pi$, where the **transconductance** g_m is defined by:

$$g_m = [\partial\, i_C / \partial\, \upsilon_{BE}]_Q = I_{CQ}/V_T = \beta I_{BQ}/V_T = \beta/r_\pi.$$

Making these changes, the circuit is as shown in Figure 1 below.

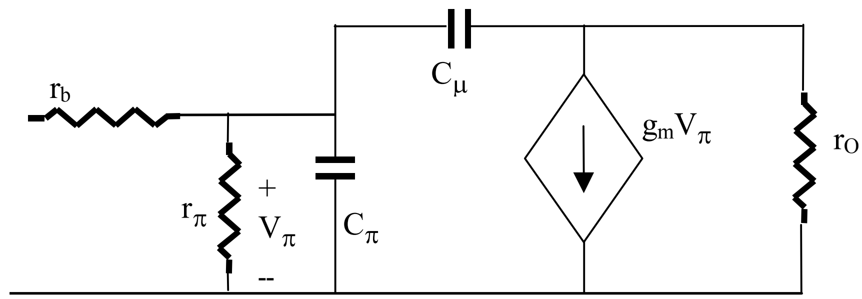


FIGURE 1

Hybrid π model with internal capacitances C_π and C_μ

Short-Circuit Current Gain:

Read S&S pp. 487-488. We will look at the circuit of Figure 1 with a resistive load in parallel with r_O and an *ac* current source as signal. There are two limiting cases with very different frequency response: the case when the load $R_L=0\Omega$ (current amplifier) and the case when $R_L=\infty$ Ω (transresistance amplifier).

Cut-Off Frequency of the Current Amplifier ($R_L=0\Omega$):

If we hook up our circuit with a short-circuit from the collector to the emitter and neglect r_b, we can find the *current gain* of the circuit A_i = (*ac* collector current)/(*ac* input current) (S&S h_{fe}, following Eq. 5.158).

EQ. 1

$$A_i = \frac{I_c}{I_b} = h_{fe} = \frac{g_m - j\omega C_\mu}{\frac{1}{r_\pi} + j\omega(C_\pi + C_\mu)} = \frac{\beta_{ac}\left(1 - j\omega\frac{C_\mu}{g_m}\right)}{1 + j\omega(C_\pi + C_\mu)r_\pi}$$

The **cut-off frequency** f_T of the bipolar transistor is defined as the frequency where this current gain drops to unity: that is it is the frequency where there is no gain. The time constant C_μ/g_m is very short, so the numerator is well approximated by 1. Setting $|A_i| = 1$, we can solve for the value of f_T. Taking $\beta_{ac} >> 1$ and $C_\mu/C_\pi << 1$ (valid if I_C is large, but not at low values of I_C) we find to lowest order in C_μ/C_π (S&S Eq. 5.163):

EQ. 2

$$f_T = \frac{\left(\beta_{ac}^2 - 1\right)^{1/2}}{2\pi r_\pi \left[C_\pi\left(C_\pi + 2C_\mu\right)\right]^{1/2}} \approx \frac{g_m}{2\pi(C_\pi + C_\mu)}\ .$$

Now let's put in the current dependence of all the parameters to see how f_T depends on the current level. We have $g_m = I_{CQ}/V_T$, $1/r_\pi = I_{BQ}/V_T$,

EQ. 3

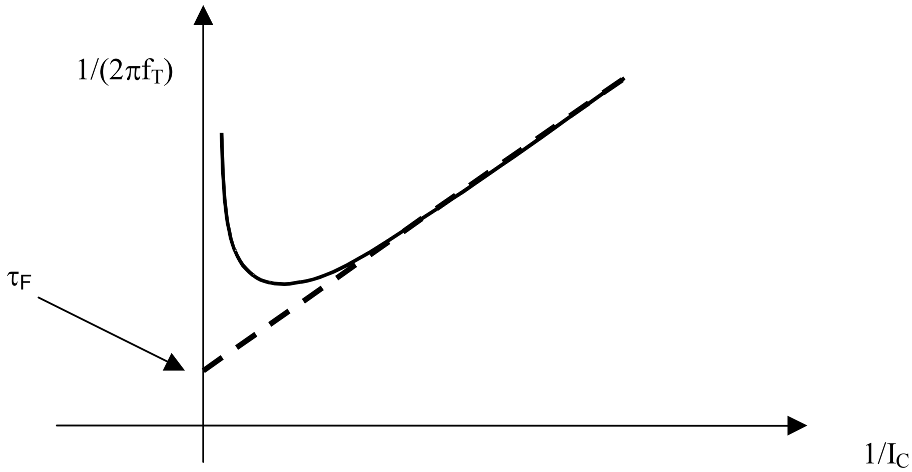
$$C_\pi = C_{jE} + I_E\,\tau_F / V_T$$

(where we have introduced the *forward transit time* τ_F to replace the transit time τ_T of the pn diode). Approximating $I_C >> I_B$ we find:

EQ. 4

$$\frac{1}{2\pi f_T} = \tau_F + \frac{(C_{jE} + C_\mu)V_T}{I_{CQ}} = \tau_F + \frac{(C_{jE} + C_\mu)}{g_m}$$

That is, if we plot $1/(2\pi f_T)$ vs. $1/I_{CQ}$ we should get a straight line with intercept τ_F. Actually we get a plot like Figure 2 below.



CjE

FIGURE 2

Sketch of EQ. 4 for $1/(2\pi f_T)$ (dashed line) and more realistic behavior (solid line)

In Figure 2, the dashed line is our EQ. 4, the one we will use in hand analysis. The solid line is closer to what is actually observed (and approximated by PSPICE). The departure from our analysis at low $1/I_C$ (large I_C) is due to the widening of the electrical base width at large current densities, a subject discussed in ECE 352.[1]

The point is this: **as the collector current varies, the frequency response varies**. If the current is low enough ($1/I_C$ is big enough) the variation is almost a straight line. But at large currents (small $1/I_C$) the frequency response begins to get worse (f_T drops, $1/f_T$ goes up).

PSpice Comparison with Formula

If we set the parasitic resistances of the collector r_b and of the base r_c to zero, we can easily obtain the PSPICE results for C_π as a function of DC collector current to compare with EQ. 3. For the Q2N2222 the results are shown in Figure 3.

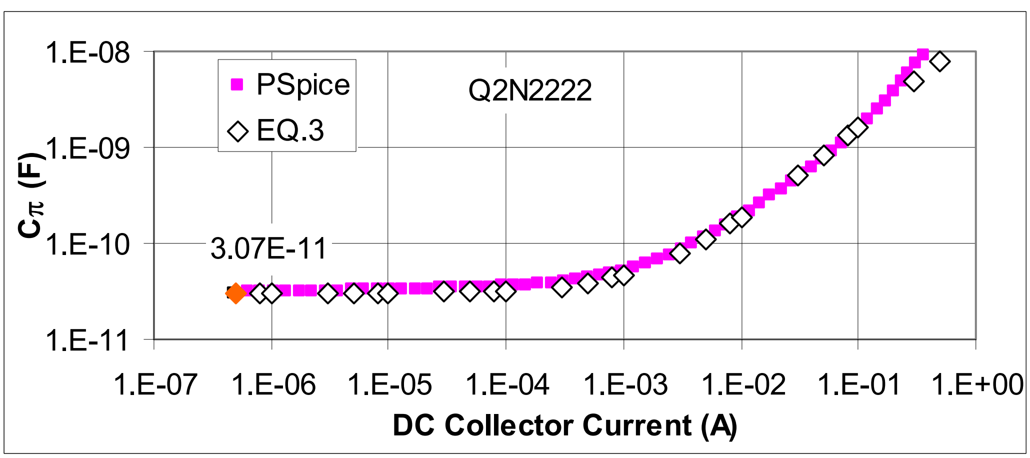


FIGURE 3

C_π vs. I_{CQ} for Q2N2222 with τ_F=0.411ns and r_b and $r_C \approx 0$

The results in Figure 3 are obtained by a small-signal AC analysis using the circuit of Figure 4

[1] Ask your 352 instructor about this phenomenon

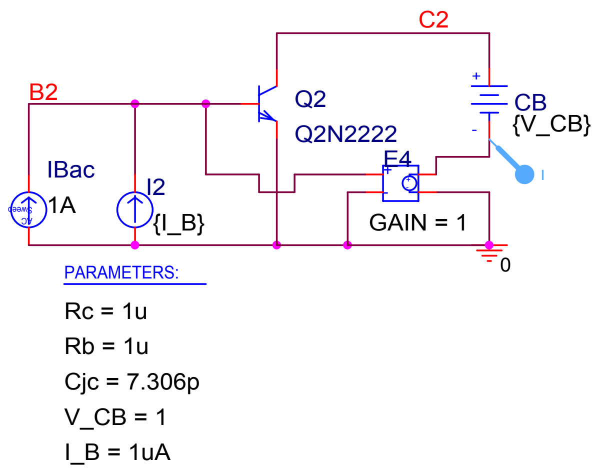


FIGURE 4

Circuit for finding C_π

In the circuit of Figure 4 an AC base current is driven into the Q2N2222. The resulting AC base voltage is fed to the collector by VCVS part E, so both base and collector move together, placing zero AC voltage across C_μ so it draws no AC current and does not affect the input impedance of the transistor. The AC base current is given by

EQ. 5

$$I_b = V_\pi\,(1/r_\pi + j\,\omega\,C_\pi) \rightarrow C_\pi = \mathrm{Im}(I_b/V_\pi)/(2\pi f)$$

We sweep the DC collector current by sweeping the DC base current. Then C_π vs. I_B and I_C vs I_B are pasted into EXCEL and Figure 3 is constructed.

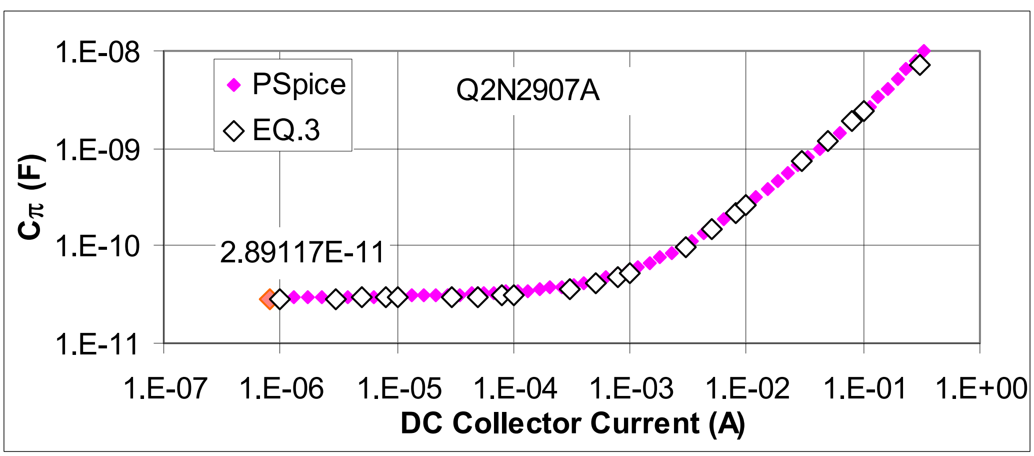


FIGURE 5

C_π vs. I_{CQ} for the Q2N2907A with τ_F =0.604ns and r_b and $r_c \approx 0$

The formula for C_π (Eq. (2)) for $I_{CQ} \approx I_{EQ}$ is $C_\pi \approx C_{jE} + I_{CQ}\,\tau_F\,/V_T$. For the Q2N2222 under forward bias at low current levels where C_{dE} is negligible, PSpice calculates $C_{be} \approx C_{jE}$ = 30.7pF. In Figure 3 we plot $C_\pi \approx C_{jE} + I_{CQ}\,\tau_F/\,V_T$. ≈ 30.7pF+ $I_{CQ}\,\tau_F/\,V_T$, with τ_F=0.411ns.taken from the

• model statement in PSPICE. (Please note: PSPICE allows us to put currents of 10-100A through the Q2N2222, but in reality it would burn up at these current levels). The comparable plot for the Q2N2907A is shown in Figure 5.

Notice that the junction depletion capacitance is dominant at low current levels, and the diffusion capacitance leads to a linear increase of C_π for higher current levels. This current level dependence of C_π causes a current level dependence of the cut-off frequency, f_T as shown in Figure 6 and Figure 7.

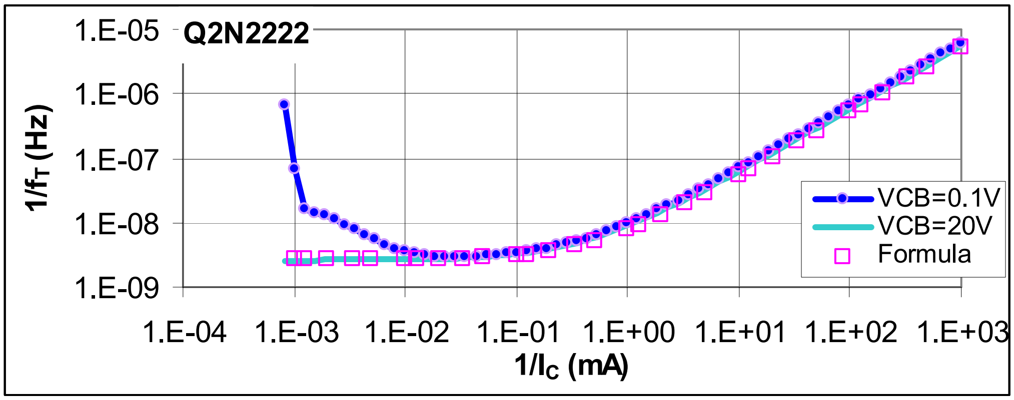


FIGURE 6

$1/(2\pi f_T)$ vs. $1/I_{CQ}$ for the Q2N2222 with τ_F=0.411ns

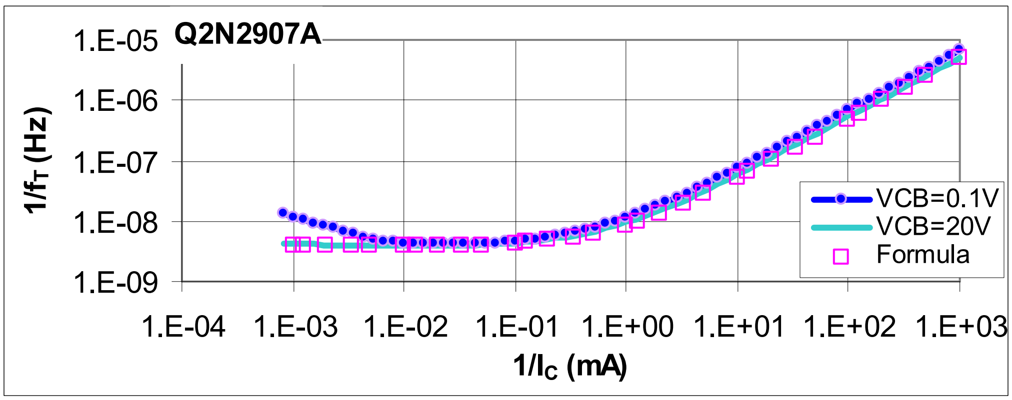


FIGURE 7

$1/(2\pi f_T)$ vs. $1/I_{CQ}$ for the Q2N2907A with τ_F= 0.604ns

Figure 6 and Figure 7 compare PSpice values for $1/(2\pi f_T)$ vs. $1/I_{CQ}$. For $I_{CQ} \leq 0.1$A (most of the useful current range) the fit in Figure 6 is fairly good with τ_F=0.411ns. Figure 6 is the equivalent of Figure 2, but on a log-log scale chosen to expand the region of large I_{CQ} (small $1/I_{CQ}$). The plot of $1/2\pi f_T$ for the Q2N2907A is shown in Figure 7.

Comparison of Current Amplifier Current Gain with PSpice

If β_{ac} were independent of current level, the gain would be independent of current level. However, Figure 9 shows how the current gain varies with current level according to PSpice when the Q2N2222.bipolar transistor is used in the circuit. (The physical origin of the current dependence of β is discussed in ECE 352). The β-variation with I_{BQ} also means that the *dc* value of β, given by $\beta_{DC} = I_{CQ}/I_{BQ}$, is different from the *ac* value of β, given by $\beta_{ac} \equiv dI_{CQ}/dI_{BQ} = \beta_{DC} + (d\beta_{DC}/dI_{BQ})I_{BQ}$ For the Q2N2222 the two are compared in Figure 9. Figure 10 shows β_{ac} as determined in PSPICE using a small-signal analysis instead of differentiating DC-curves. The two determinations agree, as we expect.

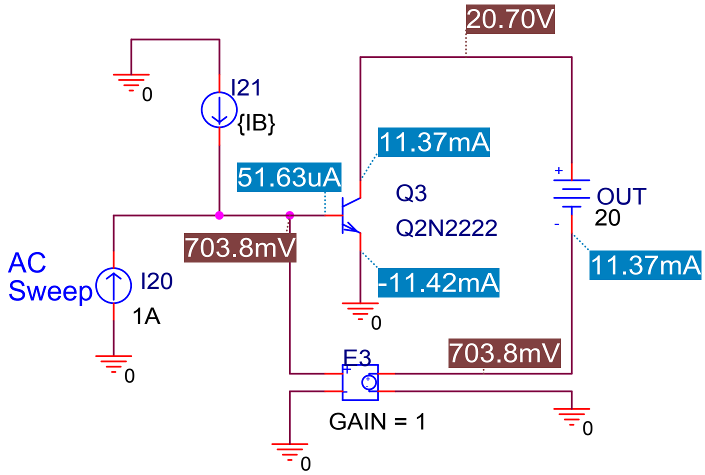


FIGURE 8

Circuit for determination of AC and DC β-values with fixed V_{CB}; the VCVS maintains V_{CB} = V_{OUT}

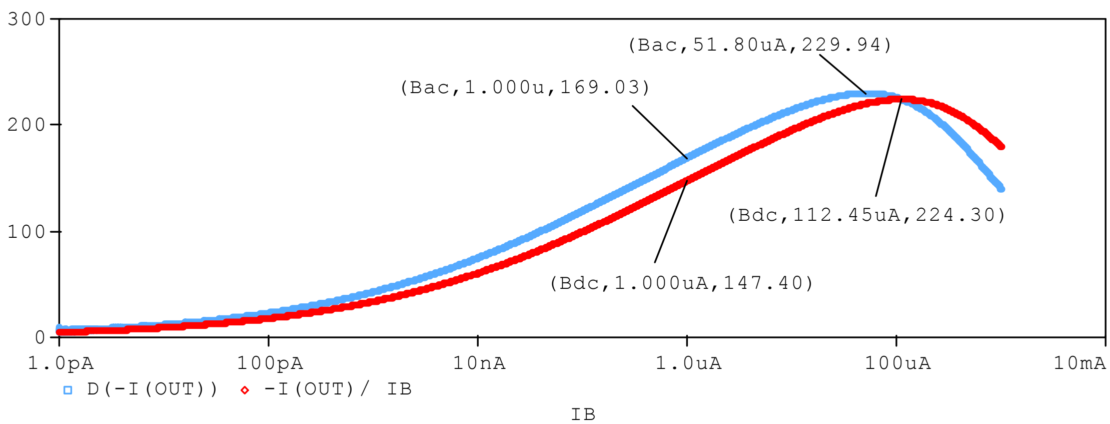


FIGURE 9

Comparison of *ac* and *dc* β-values for the Q2N2222 using the circuit of Figure 8 to hold V_{CB} = 20V; $\beta_{AC} = \beta_{DC}$ at the current level where β_{DC} has a maximum (zero slope); AC value is found as derivative of the DC plot of I_C vs. I_B and DC value is the ratio of DC I_C / DC I_B

The small-signal current gain of the amplifier in Figure 8 at low frequencies is simply $\beta_{AC}(V_{CB}$=20V), and is shown in Figure 10. It varies with current as already discussed.

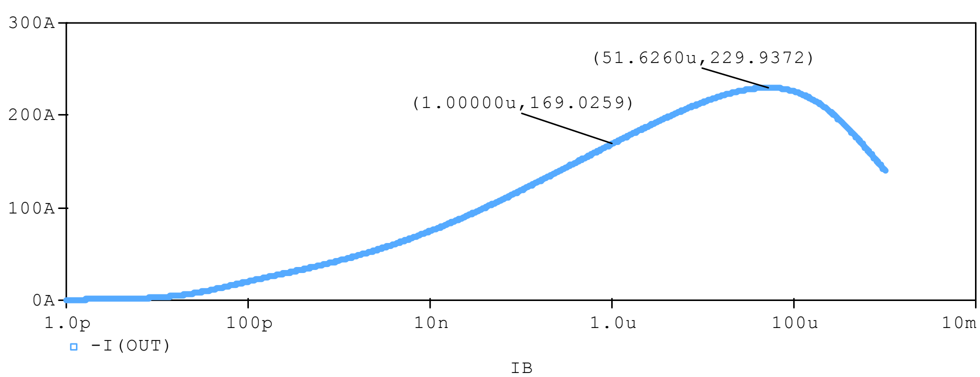


FIGURE 10

Small-signal current gain of amplifier o f Figure 8; this small-signal β_{AC} agrees with the β found by differentiation in Figure 9

Current Amplifier

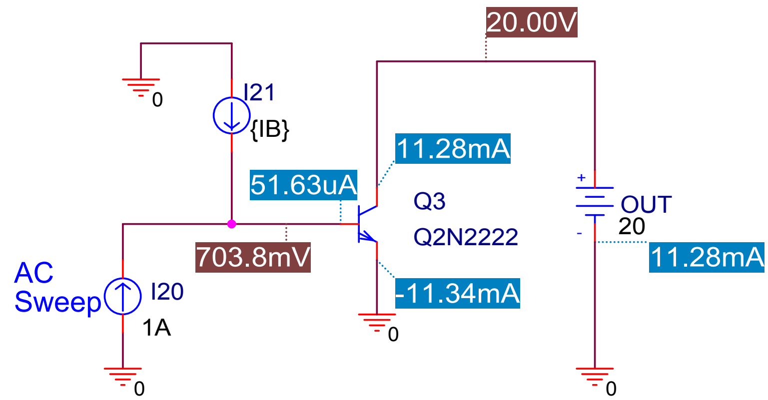


Figure 11

Current amplifier; output current is current in supply labeled Out

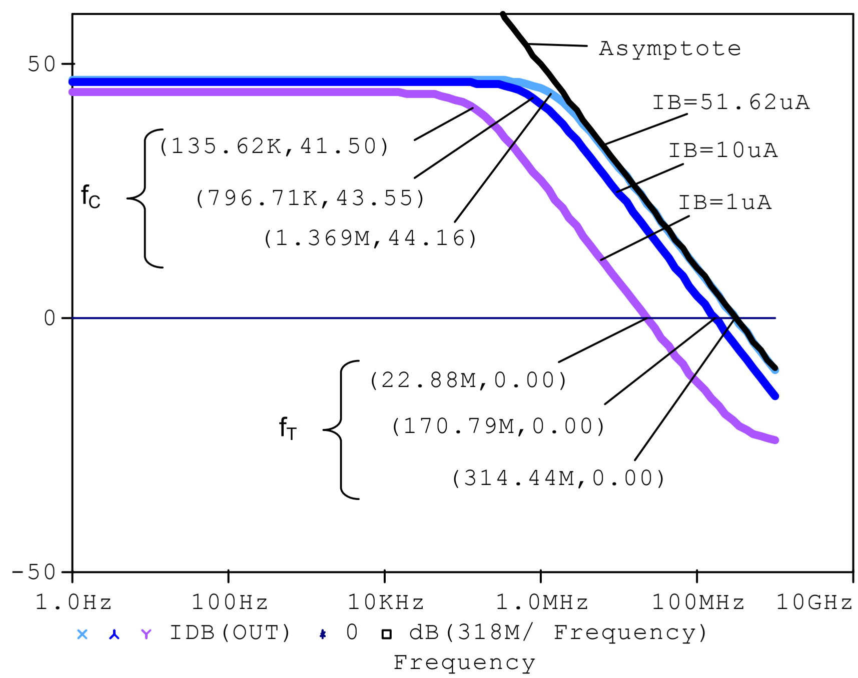


Figure 12

Current gain in dB for Figure 11 from PSpice for I_{BQ}=54μA (I_{CQ} =11.8mA, β_{ac} = 228). Also shown are the asymptote at high frequencies given by A_i=318MHz/f, and the curves for I_{BQ}=1μA and 10μA.

Now let's look at the frequency dependence of the current gain of the current amplifier. Figure 12 shows the PSpice current gain in dB for I_{BQ} =52μA, the value of current that maximizes the gain. According to our formula, the corner frequency occurs when $\omega=\omega_C$ where $\omega_C(C_\pi + C_\mu)\, r_\pi = 1$. From Eq.(2) $C_\pi = C_{jE}+I_E\tau_F/V_T$ =23.91pF+ 11.8mA x 0.411ns/.026V = 210pF (compared to 220pF from PSpice). Also $C_\mu = C_{bc}$ =2.38pF from PSpice. Using the PSpice values for C_π and r_π (=515Ω), we estimate the corner frequency as $f_C = 1/[2\pi(222\text{pF} \times 515)]$ =1.4MHz, and the asymptote at high frequencies as $g_m r_\pi/(f/f_C) = \beta_{ac} f_C/f$ = 319MHz/f. These values show that the formula agrees well with PSpice provided we use the PSpice values for all the circuit parameters.

The current gain in dB is plotted in Figure 12 for several other values of Q-point I_{BQ}. Notice that both the low frequency (maximum) gain and the corner frequency increase as I_{BQ} is increased. The larger gain is related to the increase in β_{ac} with larger current level as shown in

Figure 10. The corner frequency can be related to the unity gain frequency f_T. The corner frequency satisfies $f_C = 1/[2\pi(C_\pi+C_\mu)r_\pi] = f_T/(g_m r_\pi) = f_T/\beta_{ac}$, where EQ. 2 has been used. That is, from EQ. 4,

$\frac{1}{2\pi f_C} = \beta_{ac} \cdot \left[\tau_F + \frac{(C_{jE} + 2C_\mu) \cdot V_T}{I_{CQ}} \right]$, which shows that **for larger current levels,** $1/f_C$ is reduced, or **f_C increases**, in agreement with Figure 11.

Summary:

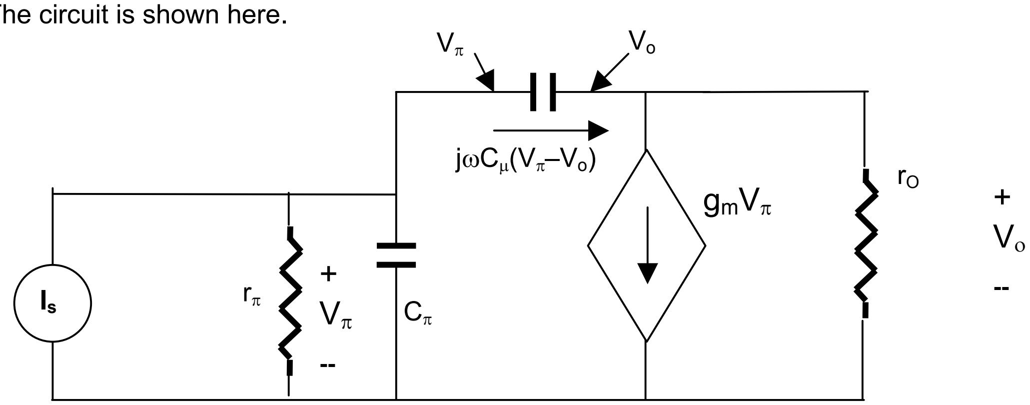
The frequency response of the current amplifier is limited by the transit time capacitance of the transistor BE junction at large current levels, and by the junction capacitors at low current levels.

Miller Effect: The Frequency Response of the Transresistance Amplifier (R_L=∞ Ω)

Read S&S, p. 107-112, p. 496 and pp. 588-597.

Let's look at the effect of C_μ on the frequency behavior of the bipolar transistor. The worst-case influence is found if we open-circuit the simple amplifier of Fig. 8 (make $R_L=\infty$). We will take the output as the voltage at the collector, and the input as the input current, so the gain A_r of the amplifier satisfies $V_o = A_r I_s$. In other words, the gain has dimensions of Ohms, so this amplifier is called a *transresistance* amplifier.

The circuit is shown here.

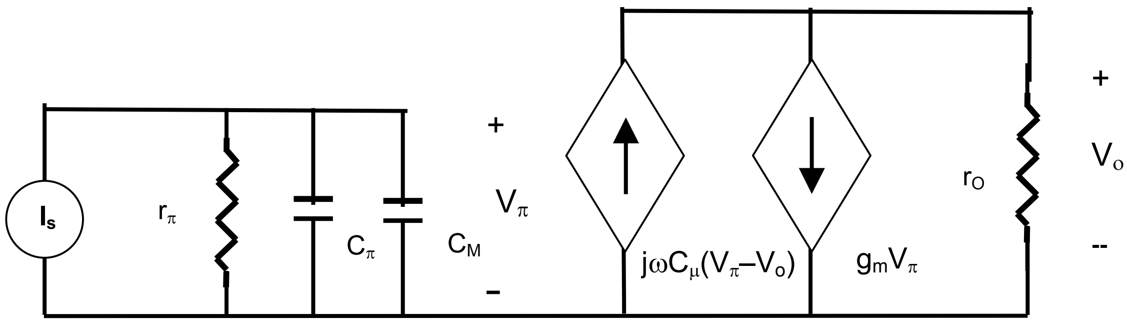


Notice that the capacitor C_μ has a *different time-varying* voltage on each side. In the current amplifier with the output shorted, this does not happen because one side of C_μ is at ground in that case. That is, the voltages on the two sides of C_μ are *different*, but on the grounded side the voltage is *not time varying*. We will see that in our circuit where C_μ has a *different time-varying* voltage on each side, the frequency dependence of the open-circuited can be much degraded.

The current through C_μ is shown as $j\omega C_\mu(V_\pi - V_o)$ and this current **couples** the input side of the circuit seen by the signal I_s to the output side of the circuit, complicating the analysis of the circuit. Miller suggested the following trick to decouple the input and output. He suggested that we rewrite the coupling current as:

$$j\omega C_\mu(V_\pi - V_o) = j\omega C_\mu V_\pi(1 - V_o/V_\pi) = j\omega C_M V_\pi$$

Now it looks like we have a new capacitance (the *Miller capacitance*) with a voltage V_π across it. (The catch is that we don't know V_o/V_π so we don't really know what C_M is …but we will fix this later). Then the circuit can be re-arranged as shown below.



The new dependent VCCS source on the output side of the circuit provides the same current as the original connection to the input through C_μ. On the input side of the circuit, KCL at the base node provides the same equation as the original circuit, and it is easy to see that the *input impedance* of this circuit is

$$Z_{in}=r_\pi // [1/j\omega(C_\pi + C_M)\ ].$$

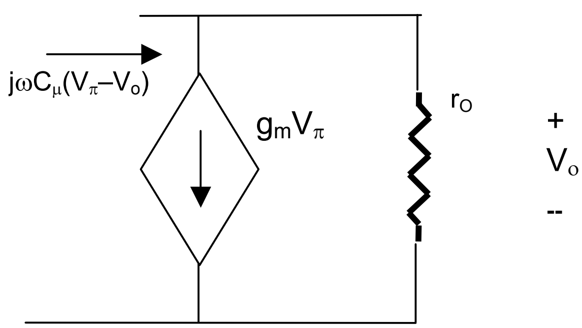
Also,

$$V_\pi = I_s\ Z_{in}.$$

Miller's Theorem states that we can decouple the input circuit from the output circuit in a case like this by introducing the *Miller capacitance* C_M, as we have just done, that is, by using

$$C_M \equiv C_\mu(1–V_o/V_\pi).$$

Is C_M really a capacitor? If we think of a capacitor as a simple circuit element that has no frequency dependence, C_M is a capacitor only as long as V_o/V_π has no frequency dependence. What we need to do now is check the frequency dependence of V_o/V_π. To do this, we look at the output side of the circuit below.



We find V_o by finding the current through the resistor r_O, which is $j\omega C_\mu(V_\pi–V_o) – g_mV_\pi$. Hence,

$$V_o = [j\omega C_\mu(V_\pi–V_o) – g_mV_\pi]\ r_O.$$

Re-arranging we find,

$$\frac{V_o}{V_\pi} = \frac{\left(j\omega C_\mu - g_m\right)\cdot r_O}{1 + j\omega C_\mu r_O}$$

If we substitute this result in the expression for C_M we find

EQ. 6

$$C_M = C_\mu\left(\frac{1+g_m r_O}{1+j\omega C_\mu r_O}\right).$$

Clearly, C_M *is frequency dependent*, and is not a true capacitor. However, if $\omega C_\mu r_O << 1$, the denominator will not depend on frequency. The condition $\omega C_\mu r_O << 1$ is enough to insure that in the expression for C_M we can use the *midband* value of $V_o/V_\pi = -g_m r_O$.

$$C_M \approx C_\mu g_m r_O$$

The use of the midband value of V_o/V_π to evaluate C_M is called the *Miller approximation*.

Let's evaluate $g_m r_O = (I_C/V_T)(V_A/I_C) = V_A/V_T$, where V_A is the Early voltage and V_T is the thermal voltage. For the Q2N2222, V_A = 74V according to PSpice. Then $g_m r_O$=3,591 and C_M = 3,591 C_μ! For the Q2N2222 with V_{CE} =20V, PSpice shows C_μ = CBC = 2.38pF, so $C_M \approx$ 8.5nF, which is a huge capacitance. Going back to $V_\pi = I_s Z_{in}$, we can find the effect of C_M on the gain of the circuit.

$$A_r = \text{Gain} = V_o/I_s = (V_o/V_\pi)(V_\pi/I_s) \approx -g_m r_O Z_{in} = -g_m r_O \{r_\pi/[1+j\omega(C_\pi+C_M)r_\pi]\}.$$

That is, the gain begins to roll off for a *corner frequency* ω_C given by $\omega_C(C_\pi+C_M)r_\pi = 1$.

PSPICE Comparison

For the case of a Q2N2222 with V_{CE}=20V and I_{BQ}=10µA, PSPICE calculates the corner frequency at 6.5kHz. To compare a hand calculation to PSPICE, we use C_π from the ANALYSIS/EXAMINEOUTPUT file as CBE = 67.2pF. This result is more accurate than $C_\pi = C_{jE}+I_C\tau_F/V_T$ = 23.91pF + 1.94mA τ_F /26mV because it includes non-ideal behavior that this formula omits. Likewise, taking r_O from the output file is more accurate than using $r_O = V_{AF}/I_{CQ}$. Then $C_M = C_\mu g_m r_O$ = 2.38pF x 74.5mA/V x 48.2kΩ = 8.55nF. Taking r_π from the ANALYSIS/ EXAMINEOUTPUT file as RPI=2.85kΩ, $2\pi f_c = 1/[(C_\pi+C_M)r_\pi] = 1/[(67.2\text{pF}+ 8.55\text{nF})\ 2.85\text{k}\Omega]$; f_c = 6.5kHz.
In this example the Miller capacitance C_M = 8.55nF is the major contribution to f_c. If I_{CQ} is increased above $I_{CQ}\approx$0.5A, the contribution of the Miller capacitance becomes secondary to the diffusion capacitance term in C_π. To clarify the current dependency, we substitute into $(C_\pi+C_M)r_\pi$ our simple formulas for the dependence of C_π, r_O, and r_π on the current, to obtain

$$\omega_C\ [\ (C_{jE}V_T+C_\mu V_A)/I_{BQ}+\beta_{ac}\tau_F] = 1.$$

That is, **the smaller I_{BQ} is selected in our circuit design, the lower the corner frequency.** Also, at the corner frequency $\omega_C(C_\pi+C_M)r_\pi = 1$, the magnitude of the gain is $|A_r| = g_m r_O r_\pi/|(1+j)| = (V_A/V_T)(V_T/I_{BQ})/\sqrt{2} = V_A/(\sqrt{2}I_{BQ})$. That is, the gain varies inversely with I_{BQ}, and directly as the Early voltage. PSPICE results are shown in Figure 14 using the circuit of Figure 13. Figure 13 uses a very large inductor that prevents the V_{CE} voltage source from short-circuiting the ac collector voltage to ground, while allowing the dc collector bias to be applied.

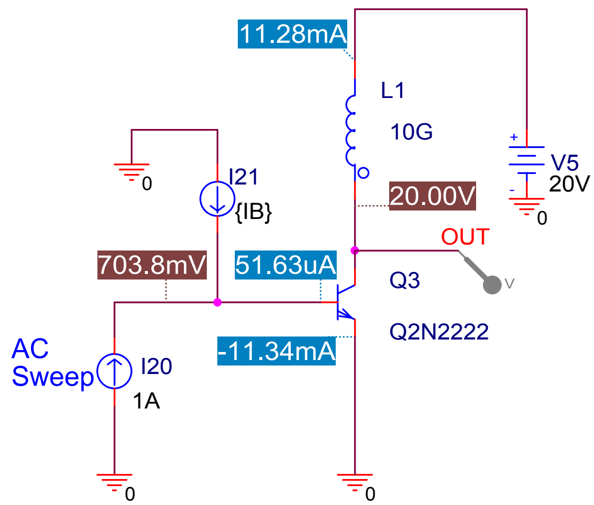


FIGURE 13
PSpice circuit for plotting frequency response of transconductance amplifier

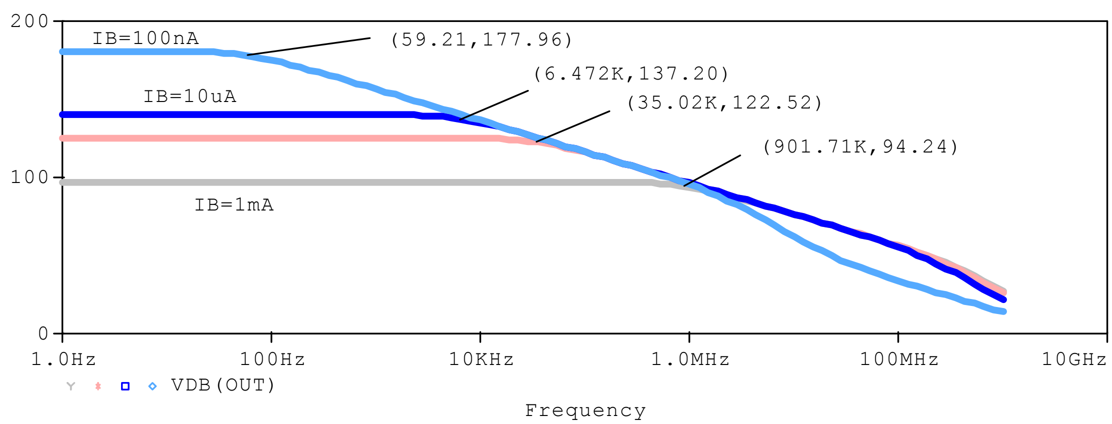


FIGURE 14
Bode plots of transconductance gain in dB at various I_{BQ}

GAIN-BANDWIDTH TRADE-OFF: S&S, P.93, PP. 558-559

. For an amplifier with a roll-off frequency determined by the Miller effect there is a **gain-bandwidth tradeoff**. That is, to get higher gain, one must reduce bandwidth and *vice versa*. In fact, the product of the maximum gain and the bandwidth (corner frequency) is a constant. For this particular amplifier, the **gain-bandwidth product** (Eq. (6.3) p. 558) is given by our formulas as $1/[2\pi C_\mu]=6.69 \times 10^{10}$ V/(A s). PSpice results are tabulated below.

Base Current	Gain at Corner (V/A)	Bandwidth (Corner f)	Max. Gain x Bandwidth Product V/(A· s)
100nA	7.91E08	59.21Hz	4.68E10 x √2 =6.62E10
10μA	7.24E06	6.472kHz	4.69E10 x √2 =6.63E10
1mA	5.15E04	901.7kHz	4.65E10 x √2 =6.57E10

Summary:

When a capacitor has a different time-varying voltage on either side of it, the Miller *Theorem* often is useful in simplifying the circuit. Provided the ratio of the voltages on the two sides of the capacitor is not frequency dependent near the corner frequency, the Miller *capacitance* can be estimated using the Miller *approximation* and then used **to find the corner frequency** of the amplifier. Depending on the gain, the Miller capacitance can be very large, and can be the main factor causing the roll-off of the gain. This **roll-off frequency will depend on the current** level selected for the circuit. The gain also depends on the current level and, when the corner frequency is dictated by the Miller effect, a **gain-bandwidth tradeoff** exists.

Appendix 1: PSPICE output file

NAME	Q_Q3
MODEL	Q2N2222
IB	5.16E-05
IC	1.13E-02
VBE	7.04E-01
VBC	-1.93E+01
VCE	2.00E+01
BETADC	2.19E+02
GM	4.23E-01
RPI	5.39E+02
RX	1.00E+01
RO	8.27E+03
CBE	2.12E-10
CBC	2.38E-12
CJS	0.00E+00
BETAAC	2.28E+02
CBX/CBX2	0.00E+00
FT/FT2	3.14E+08

FIGURE 15

PSPICE output file for the Q2N2222 showing notation for capacitances. $C_\pi = C_{BE}$, $C_\mu = C_{BC}$

The bipolar capacitances in PSPICE are named CBE ≡ $C\pi$ and CBC ≡ $C\mu$.; R_X = series resistance between base contact and r_π; capacitances CJS, CBX/CBX2. will not be used in this course.

Appendix 2: Using PSPICE to obtain f_T vs. I_E

This appendix is just an FYI, so glance at it but don't worry about it too much.

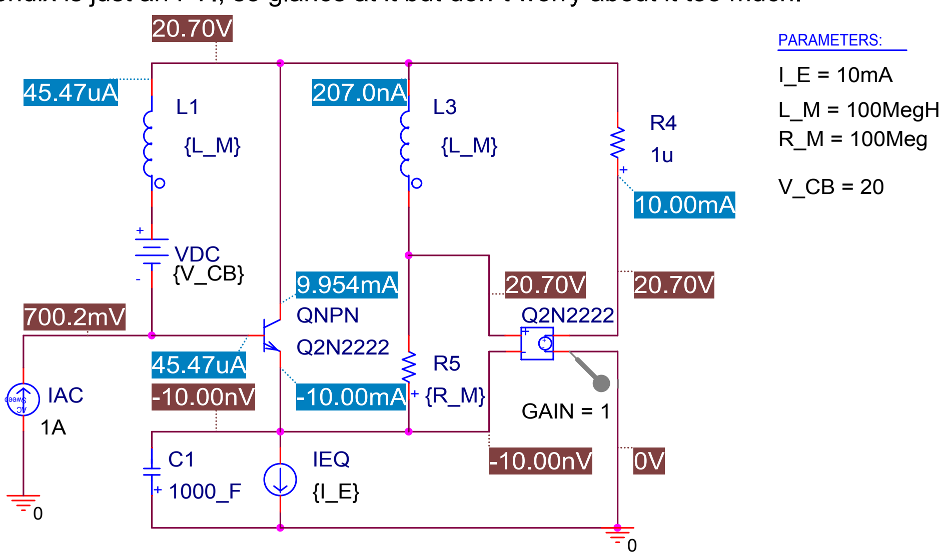


FIGURE 16

Circuit to find AC current gain for a set value of DC emitter current I_E

Although we would prefer to find f_T as a function of DC collector current, the circuit is simpler to set up for a sweep of I_E. Later a DC simulation can find I_C vs. I_E to convert to a plot vs. I_C. However, a plot of f_T (the unity gain frequency) vs. I_E is indistinguishable from a plot vs. I_C on a log-log scale.

Discussion of circuit

The circuit in Figure 16 contains large inductors to allow DC currents to bias the transistor while at the same time blocking AC current. Thus, the AC collector current is forced to flow through the voltage source in the VCVS labeled Q2N2222.[2] The current probe on this VCVS picks up this AC collector current. The purpose of the VCVS is to control V_{CB} at the same value regardless of the applied DC bias current I_E. In this way, as I_E is swept only V_{BE} changes, and V_{CB} is held constant. Constant V_{CB} means that C_μ does not change when I_E is swept. It also means that the transistor is held in the active mode, regardless of the value of I_E.

Circuit for the Q2N2907A

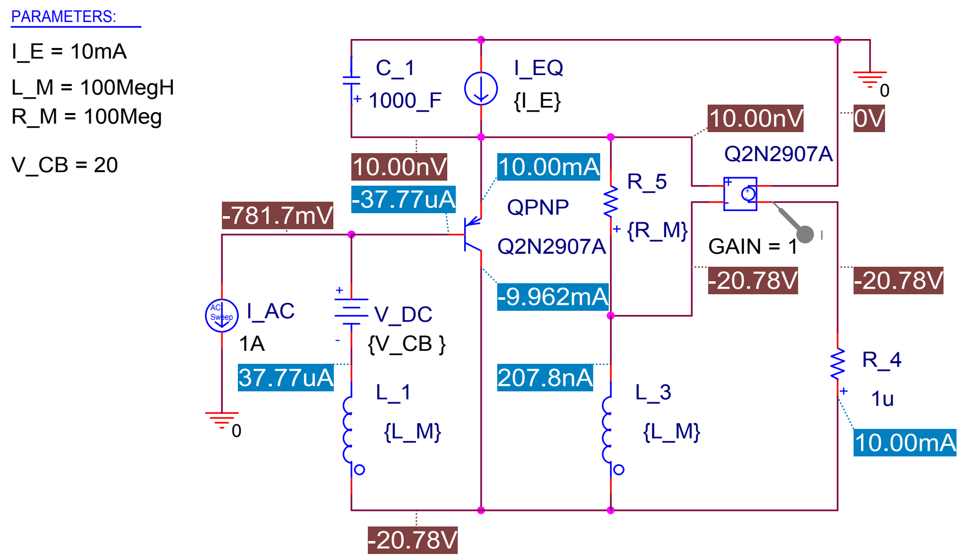


Figure 17

Circuit for the Q2N2907A

The circuit for the PNP transistor is a mirror image of that for the NPN, as shown in Figure 17

Results

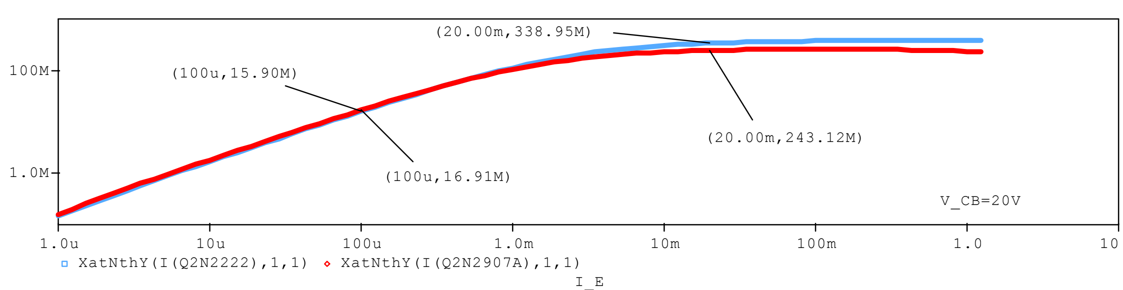


Figure 18

Comparison of Q2N2222 (top curve) and Q2N2907A unity gain frequencies vs. DC emitter current I_E at a V_{CB} = 20 V

[2] This label for the name of the VCVS is used so the Probe plot will label the current I(Q2N2222); likewise the VCVS for the Q2N2907A is labeled Q2N2907A. That way both plots are easily identified when made on the same graph.

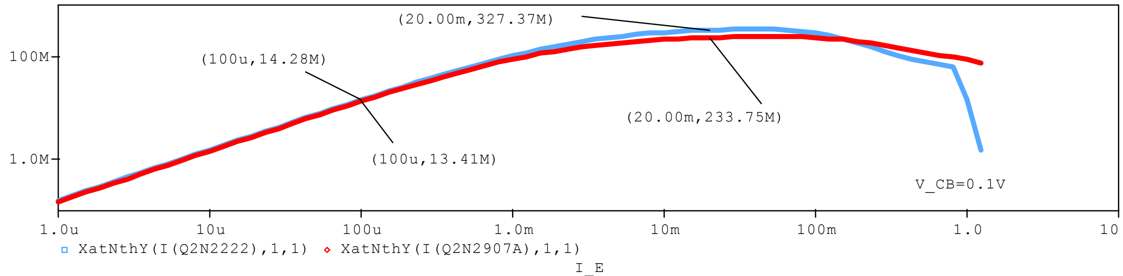


FIGURE 19

Comparison of Q2N2222 (top curve) and Q2N2907A unity gain frequencies vs. DC emitter current I_E at a V_{CB} = 0.1 V

I have no idea why the f_T-plot at low values of V_{CB} for the Q2N2222 shows such a dramatic drop at high current levels, while the Q2N2907A does not. Of course, the Q2N2222 is unlikely to be used at a current of I_E = 1 A, so it is improbable that the PSPICE model for the Q2N2222 has ever been compared with real data at these current levels. So the drop is to be taken with some skepticism.

CHECK ON THE CIRCUIT

As a spot check to see the circuit does what we expect, we compare a value of f_T from the circuit against the PROBE output value.

NAME	Q_QPNP	Q_QNPN
MODEL	**Q2N2907A**	**Q2N2222**
IB	-3.78E-05	4.55E-05
IC	-9.96E-03	9.95E-03
VBE	-0.782	0.7
VBC	20	-20
VCE	-20.8	20.7
BETADC	264	219
GM	0.382	0.375
RPI	692	613
RX	10	10
RO	1.36E+04	9.45E+03
CBE	2.65E-10	1.92E-10
CBC	2.47E-12	2.35E-12
CJS	0.00E+00	0.00E+00
BETAAC	264	230
CBX/CBX2	0.00E+00	0.00E+00
FT/FT2	**2.28E+08**	**3.07E+08**

FIGURE 20

PROBE output file for I_E = 10 mA, V_{CB} = 20V (NPN) V_{BC} = 20V (PNP); unity gain frequencies are labeled FT/FT2 at bottom

Figure 20 shows the PROBE OUTPUT FILE values for f_T as FT/FT2 = 2.28 $\times 10^8$, and 3.07 $\times$ 10^8. For comparison, the current gain obtained using Figure 16 and Figure 17 is shown in Figure 21. Agreement is excellent.

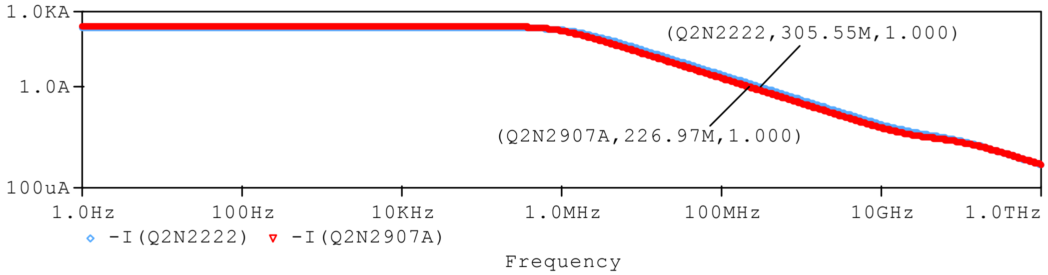


FIGURE 21

Unity current gain frequency for the Q2N2222 and Q2N2907A using the test circuits of Figure 16 and Figure 17

Reference

See the on-line PSPICE Reference Guide C:\Program Files\OrcadLite\Document\rel92pdf.pdf pp. 203 – 213.